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A COMPUTER DESIGN OF CATV DISTRIBUTION SYSTEMS

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Abstract

CATV distribution systems are expensive. To reduce hardware costs, we have developed a package of computer programs that generates complete layouts and designs for CATV distribution systems. Since the computer does not rely on "intuition", it is not restricted to using routine approaches but is free to select the best combination of components and layout which meet the system specifications.

To illustrate how money can be saved with network optimization by computer, we examine a number of results derived from actual computer runs. These results include the layout of feeder and trunk cable, the location of distribution amplifiers, and detailed assignment of amplifier locations as a function of cable sizing and coupler assignments. The computer's designs not only save money but are free of the approximations, rules of thumb and inadvertent errors introduced by human designers.

Introduction

The large number of subscribers and the requirements on signal quality make well-designed broadband cable television systems among the most difficult networks to achieve. For this reason we at Network Analysis Corporation have combined modern network analysis and computer methods to optimize CATV distribution system design. The result is a computer program which completely engineers a CATV distribution system. The computer-designed systems, when compared to manually-designed systems are produced faster, are more dependable and have significantly lower hardware cost. Furthermore, once a design has been developed, its details and specifications are already in a form suitable for computerized inventory, maintenance and replacement studies.

The Problem

The CATV system designer is faced with a host of competing variables and requirements such as cross-modulation, noise, bandwidth, temperature, alternate routes and component characteristics. Somehow, he must contend with all of these factors to produce his design. The design involves many crucial decisions.

TABLE 1

The Complex Decisions for CATV Design

- Selection of head end sites
- Location of messenger cable
- Selection of trunk distribution points
- Selection of components and manufacturers
- Selection of amplifier output levels and gains
- Location of trunk and feeder cable
- Sizing of cable, location of amplifiers and assignment of splitters and couplers
- Specification of tilt compensation, padding and settings for amplifiers
- Assignment of automatic gain and slope control, and temperature compensation
- Specification of subscriber taps
- Location of power supplies
- Provision for future system expansion

If any of these aspects are not given adequate consideration, the result can be a costly or a low performance design.

The result of the design process must be a complete design detailing: location and specification of all components including cable, couplers and amplifiers, signal levels, cross modulation and noise levels throughout the system, a bill of materials, and all of the other items shown in Table 1 above.

To cope with all this, the human designer must compromise and decide on many of the design parameters either independently or without examining the full range of interactions among them. These compromises can be costly. In all cases in which good manual designs produced by professional designers have been directly compared with designs independently generated by NAC's computer programs, the computer has produced substantial savings while obtaining equal or superior system performance. The savings on hardware have ranged from 8% to 40% of the cost of the manual design. Even for a small system with only 15 miles of strand, NAC's design was 8% lower than the best effort of a group of designers who designed their system as part of a competition among themselves.

Examples of Savings by Computer CATV Design
Since the computer does not rely on "intuition", it is not restricted to using routine approaches that were developed to handle similar but not identical cases. The computer is free to select the best combination of layout and components which will meet the system specifications. After studying the results of the computer's optimizations, it becomes evident that the computer designs are based on sound engineering principles applied in unique and original ways to each particular situation. The best way to illustrate how money can be saved using NAC's computer CATV design program is to examine results derived from actual computer runs. Component specifications for these examples are shown in Table 2.

The computer can, of course, use any components with any characteristics from any manufacturer or manufacturers. The simplified characteristics in Table 2 are representative and are used for the sake of illustration. The specifications, such as output level, which is normally chosen by the computer, are assumed to have been selected on the basis of overall system constraints on noise, cross modulation, intermodulation and performance under temperature variation. Thus, the examples are reduced to their simplest terms.

In the examples we assume the following system specifications and constraints:

1. The telephone poles are located 100 feet apart.
2. The minimum signal level at the termination of any feeder is 26 dBmV for an undedicated system. Amplifier gains have been derated to allow for subscriber tap losses.
3. There can be no more than two extender amplifiers in any cascade.

4. Cable size can change only at a splitter, coupler, or amplifier.
5. Amplifier output levels must be exactly as in Table 2. Variable gains and equalizers are included in the amplifiers.

6. Items such as AGC, power supplies and reflections are ignored for case of illustration.

● A cable television distribution system involves a vast number of possible structures or layouts--far too many to select by eye, experience, intuition, or by evaluating every possibility. For example, examine the strand map shown in Figure 1. Any CATV designer would consider this system trivial--there are only 4 blocks with 4 telephone poles per street. Yet for even this simple example, if every street is to be covered, there are 49,152 possible feeder cable layouts. One possible layout is shown by the heavy lines.

It is easy to see that even for a very small town with only 15 miles of strand, the number of possible layouts is so large that both intuition and brute force enumeration of all possibilities fail as optimum design methods. In fact, for most small systems if one were to cover the earth with computers each 1 square inch in area and each making one million evaluations per second, it would take more than the lifetime of the universe to examine all possible system layouts for the town. NAC's computer programs are able to avoid these problems to produce savings.

● Cable is available in certain fixed diameters. For trunks, designers usually select .412 inch, .500 inch, .750 inch or 1.00 inch coaxial cable. The discreteness of cable sizes can invalidate most insights the human may have while at the same time creating a huge and onerous selection from possible cable size combinations. Even for the small section of trunk shown in Figure 2, with cable sizes changing only at splitters or amplifiers, and allowing only two possible sizes for the trunk, there are 1024 possible cable size combinations.

Figure 3 shows how the computer's optimum choice of cable has reduced the hardware cost in one actual case of trunk design. We assume the location of the first trunk amplifier is given. The next trunk amplifier is moved 1.5 db left, a splitter combination is changed and as a result $78.00 is saved in cable cost.

• The steps required in creating a well designed CATV system are all extremely dependent on one another. It is usually impossible for a human designer to consider simultaneously even a small number of interacting problems, i.e., distribution point locations and the complete feeder and trunk layout. But to find the lowest cost system, many factors must be considered simultaneously. The human uses rules of thumb to reduce his problem to a manageable size. For example, one such rule used by some designers is: "keep the trunk as short as possible by pulling back the distribution points as far as possible." A design using this philosophy is shown in Figure 4a. The feeder design shown is the best possible one given the distribution point. A superior design, produced by the computer, is shown in Figure 4b. This design has a longer trunk and even has an extra trunk amplifier. But it costs $158 less.

• Changes in one part of a design can often have suprisingly significant effects on other seemingly remote parts of the system. Engineers have great difficulty in considering more than one local area at a time. For example, in the system of Figure 5a, a designer placed a 3-way hybrid splitter, the SP3W, on one output of the trunk bridger. He correctly judged the trunk bridger with 4 feeder outputs to be wasteful in this situation. Looking at the overall picture, the computer cascaded an SP8 and an SP3 to obtain the design shown in Figure 5b., using one less extender amplifier, with a resultant saving of $92.

• Small changes in design decisions can cause large changes in cost and performance. One of the most complex decisions is the location of distribution points. $140 was saved by moving the distribution point only 100 feet from the position in Figure 6a to the position in Figure 6b even though the layout was not affected. The money was saved by removing extender amplifiers and converting .500 cable to .412 cable.

In some cases it is undesirable to use more than one cable size for feeder cable. This may be due to the added cost of inventory or the added installation problems. However, often costs can be assigned to these factors. When they are added into the cost of cable, it is usually still worth while to use more than one cable size. Certainly, in many cases near the ends of feeder lines, small sections of cable of large diameter can eliminate many extender amplifiers. The computer can take these factors into account in its optimization.

- For the sake of simplicity, the above examples have been for undedicated systems. NAC's computer program integrates the assignment of subscriber taps into the overall design procedure to achieve large additional savings over conventional techniques. For example, the system in Figure 7. is a good manual design for a system designed with a flat loss allowance for subscriber taps of 6 dB between extender amplifiers. Required subscriber tap locations are indicated by darkened squares.

When the taps are added to this system, the resulting design will have four extender amplifiers. However, if the design procedure takes into account the actual tap losses rather than allowing a fixed flat loss, savings can be made. Thus, for the taps with characteristics shown in Figure 8 with a required signal of 11 dBmV at the tap output at 270 mHz , the design in Figure 9 is achieved with the given tap locations. The extender amplifier inputs can now go as low as 20 dBmV and the extender gains are 17 dB or 20 dB. Note that only two extender amplifiers are now required instead of four.

A Complete CATV Computer Service

As mentioned previously, the above examples were simplified for case of illustration. The computer program also performs temperature and AGC calculations, assigns equalizers, locates power supplies, and can add extra poles and strand where allowed and where economical.

In addition to the savings, speed and performance assured in performing these operations by computer, there are two other striking advantages.

a. Suppose a new line of components appears on the market. The human designer must begin anew to gain experience before he can produce efficient designs. NAC's program has no such problem. The computer has actually designed systems with components that do not yet exist but are being considered as possible new products. The computer program is simply fed the characteristics that the manufacturer would like his device to have and the computer program produces its design. The manufacturer can then judge whether the proposed device is worth producing. Among the system features the program has evaluated are integrated circuit components, two-way systems, new lines of equipment and specialty items.

b. Once a system has been built, the program is not through. It can be used to set up a data base for inventory maintenance and replacement schedules, and to monitor, study, adjust, alter or update the system throughout its lifetime. Its uses have included:

- Aging and replacement studies
- Modernization by using new equipment
- System expansion
- Expansion of capabilities
    - bandwidth
    - addition of two-way sections

Conclusion

The CATV industry stands at the threshhold of one of its most explosive and vital periods of growth. The design decisions and commitments made now will have long lasting effects on the cost, performance and ultimate capability of the vast cable television enterprise. It is essential that these new systems be designed efficiently and economically. NAC's computer CATV design program can play a vital role in this effort.

COMPONENT CHARACTERISTICS

COMPONENT	TRUNK OUTPUT LEVEL (DBMV)	MAXIMUM GAIN ON TRUNK (DB)	FEEDER OUTPUT (after all splits) (DBMV)	MAXIMUM GAIN TRUNK TO FEEDER (DB)
TRUNK AMPLIFIER	29	22.5	—	—
		SYMBOL: COST:	$350	
TWO OR FOUR FEEDER TRUNK BRIDGER AMPLIFIER	29	22.5	42	48
	TWO FEEDER SYMBOL: COST:	$600	FOUR FEEDER SYMBOL: COST:	$700
TWO OR FOUR FEEDER DISTRIBUTION AMPLIFIER	—	—	42	36
	TWO FEEDER SYMBOL: COST:	$400	FOUR FEEDER SYMBOL: COST:	$500

COMPONENT	FEEDER OUTPUT LEVEL (DBMV)	MAXIMUM GAIN ON FEEDER (DB)
EXTENDER AMPLIFIER (one in cascade)	40	14
	SYMBOL: COST:	$150
EXTENDER AMPLIFIER (two in cascade, must be used for both amplifiers in a cascade of two)	37	11
	SYMBOL: COST:	$150

CABLE

.500": 1.5 db loss/100' at 270 MHZ
SYMBOL: - - - - COST: $.095/ft.

.412": 2.0 db loss/100' at 270 MHZ
SYMBOL: ——— COST: $.065/ft.

SPLITTERS AND COUPLERS

$16: SP8 — 1.5 db loss; 8 db loss

$18: SP3 — 3.5 db loss; 3.5 db loss

$19: SP3W — 6.5 db loss; 3.5 db loss; 6.5 db loss

Table 2
Extender amplifier gains have already been reduced by 6 dB to allow for tap insertion losses in examples of designs of undedicated systems.

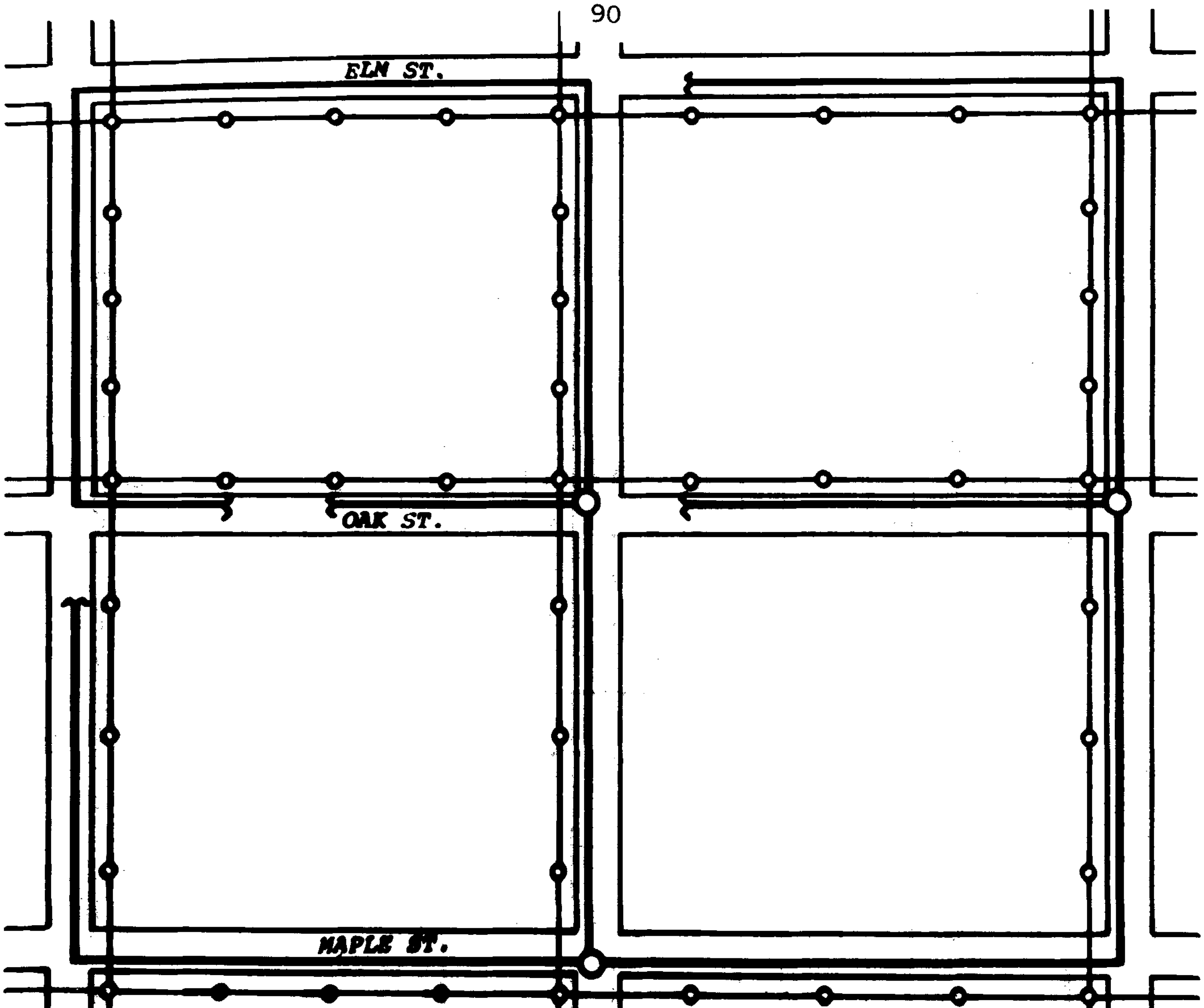


Figure 1
There are 49,152 possible feeder layouts for this four block strand map.

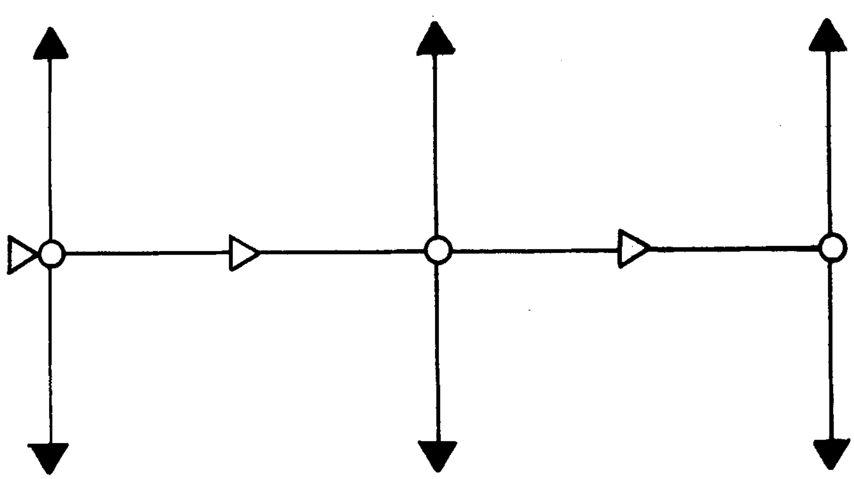


Figure 2
There are 1,024 possible cable diameter combinations for this layout.

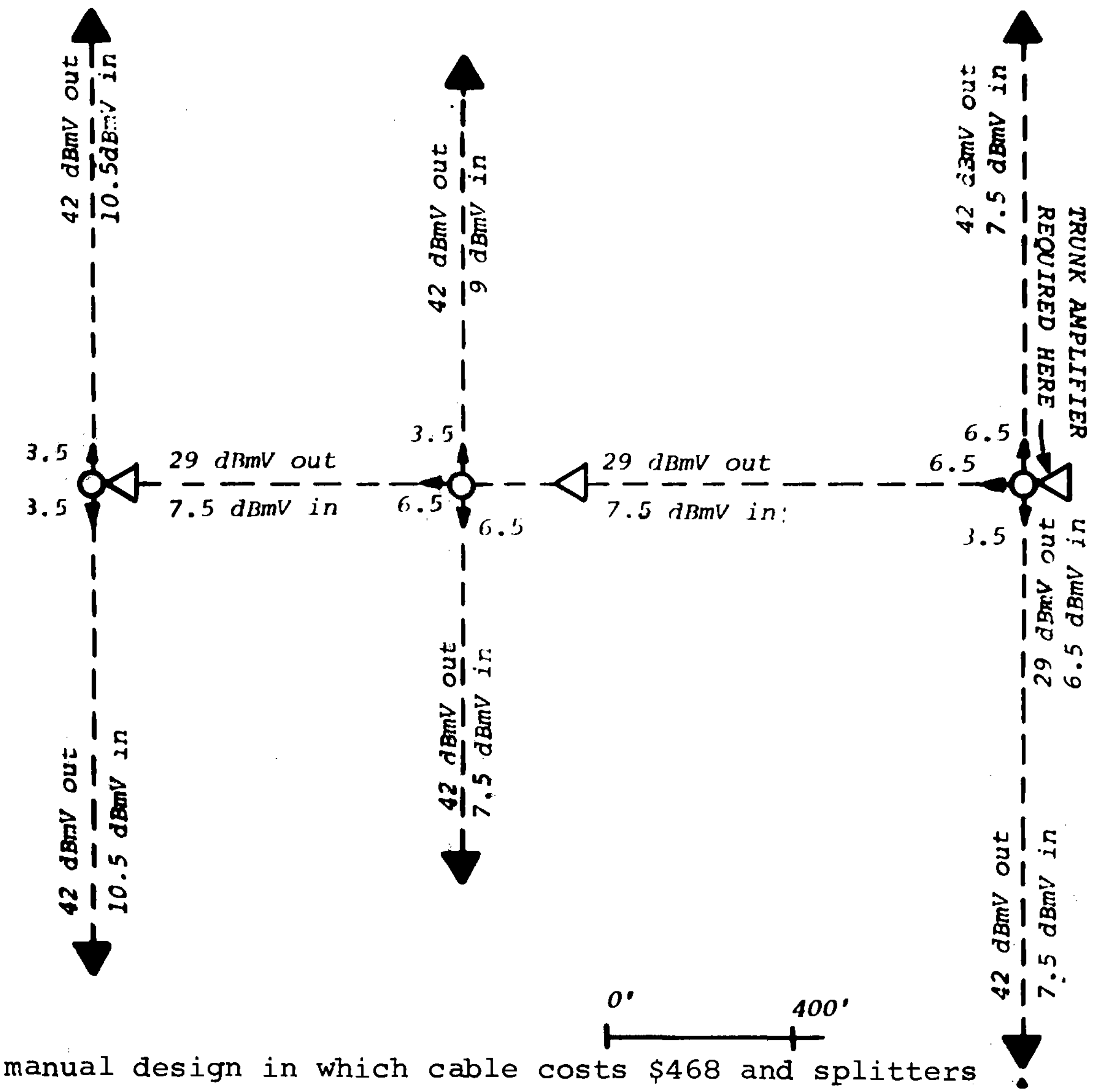
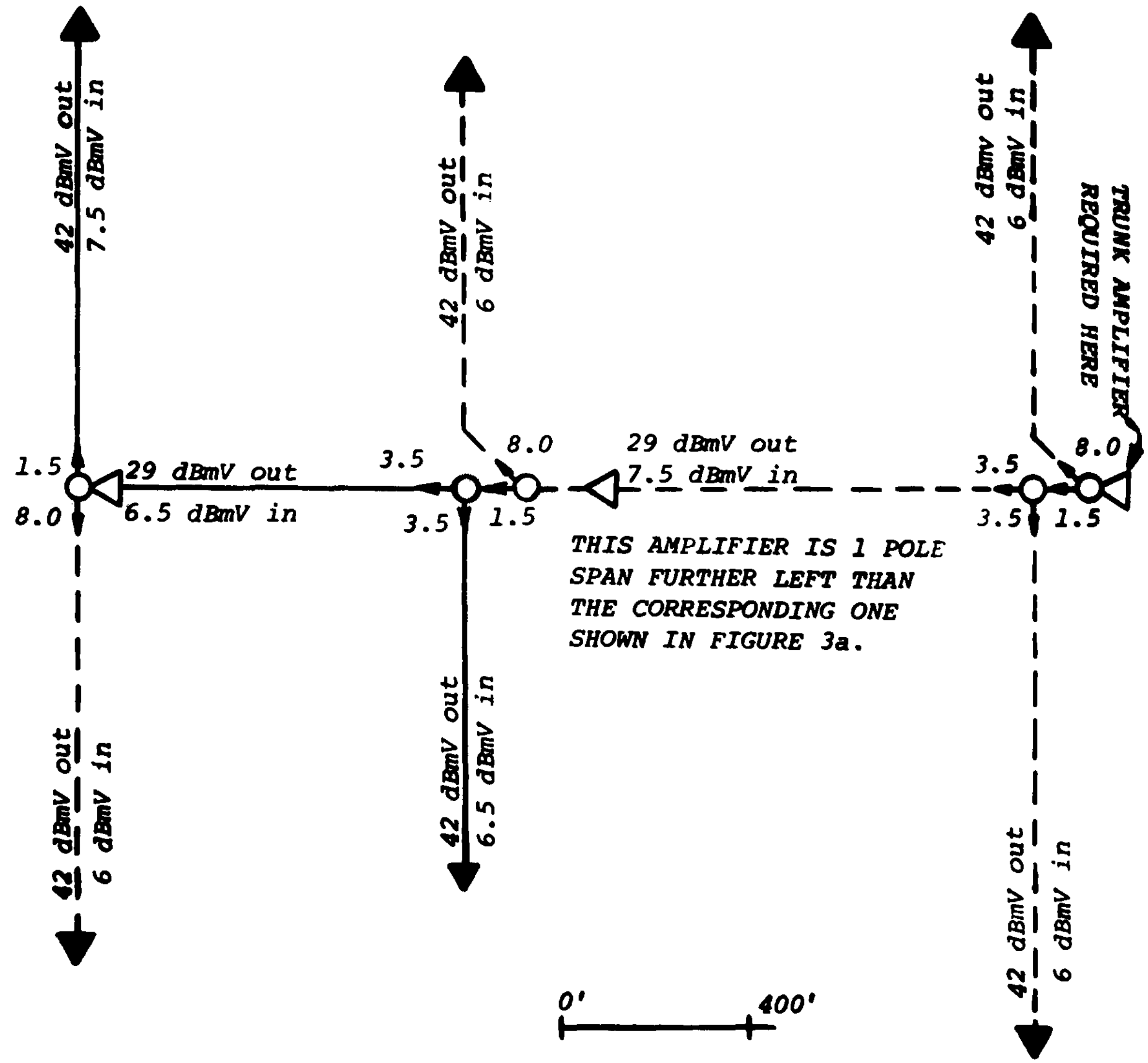


Figure 3a
An excellent manual design in which cable costs $468 and splitters cost $56.



<u>Figure 3b</u>
In NAC's computer design splitters cost $84 but the cable cost saving is $78.

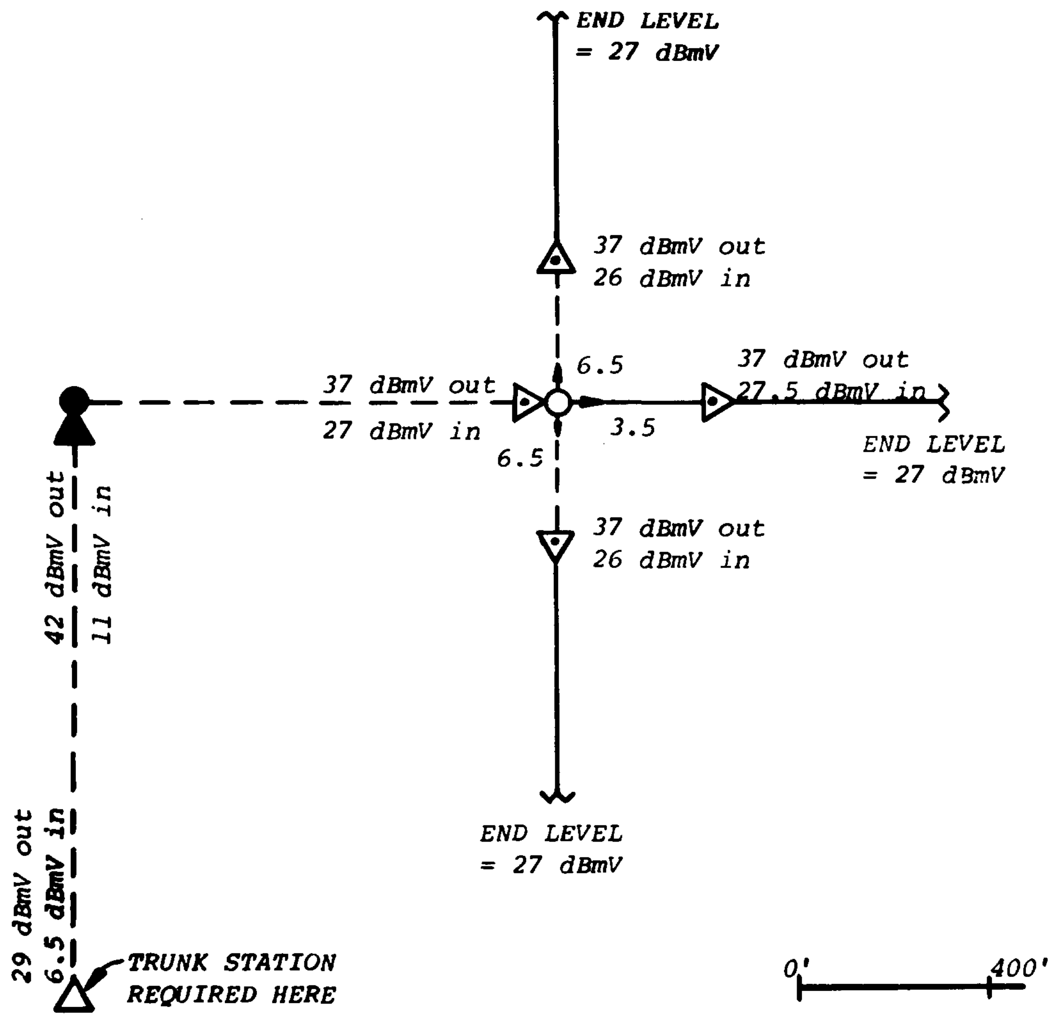
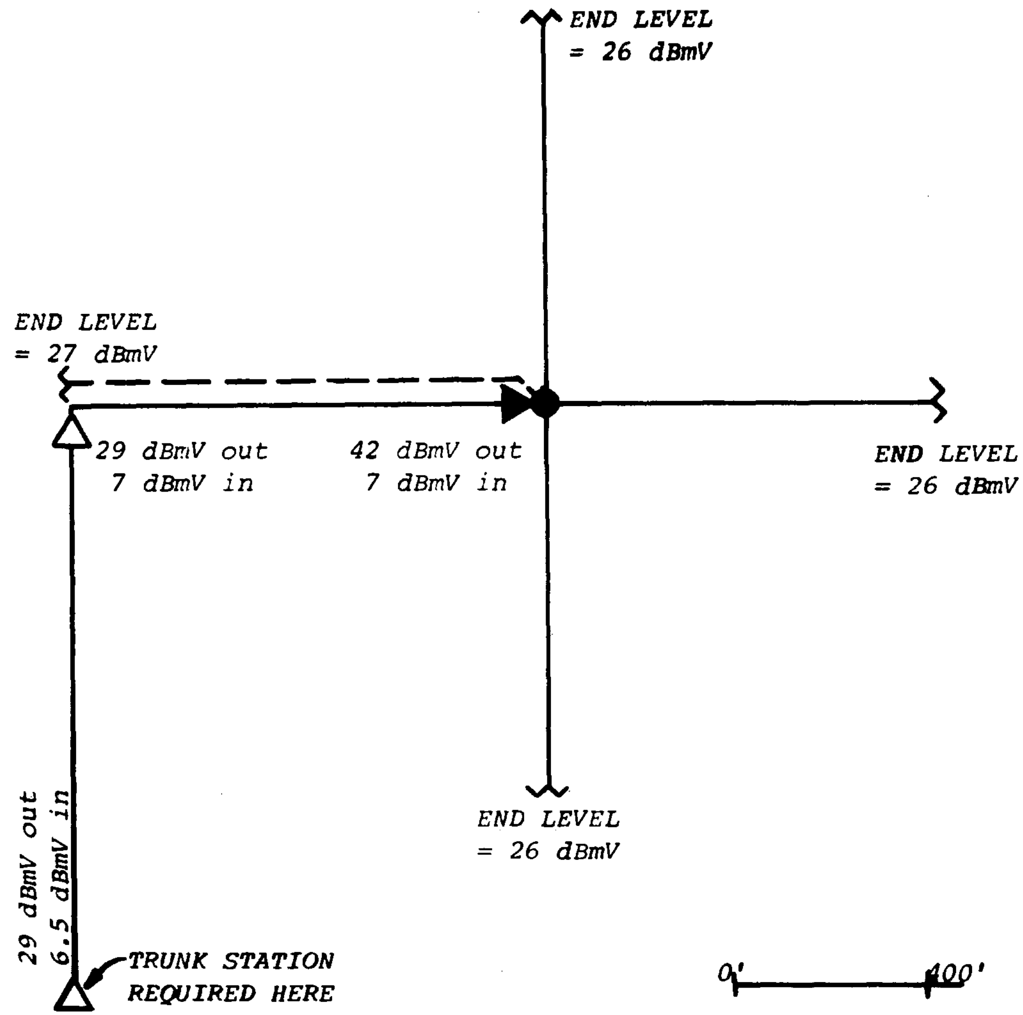


Figure 4a
Manual design.



<u>Figure 4b</u>
The computer design costs $158 less--even though the computer design (Fig. 4b) contains one more trunk amplifier and has more trunk cable than the man-made design (Fig 4a).

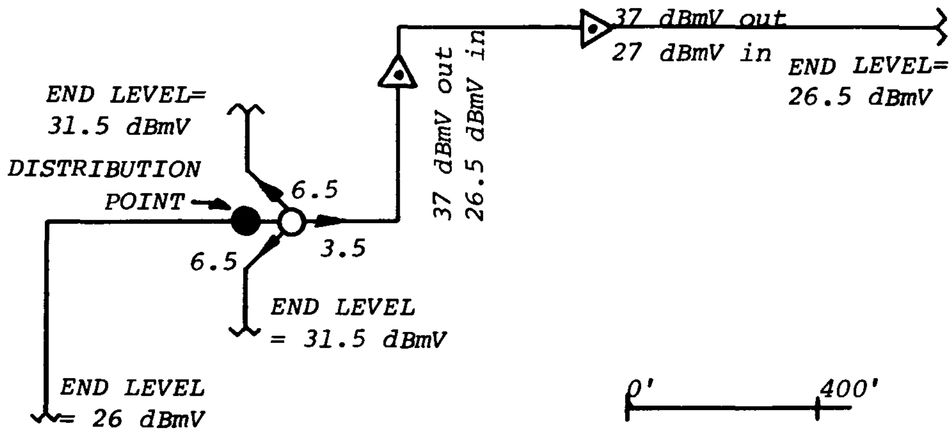


Figure 5a
Manual design.

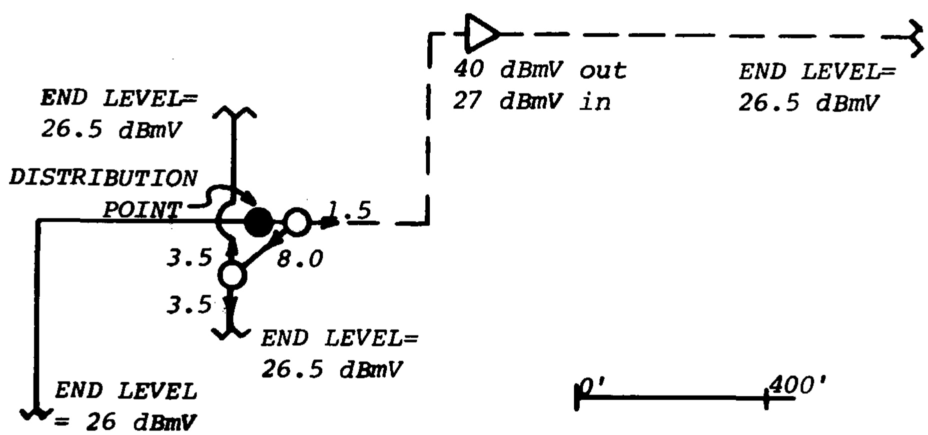


Figure 5b
NAC computer design. The computer excels at solving a tough problem--tailoring splitter losses to system needs. The computer design (Fig. 5b) saved more than 17% of the cost of the human designed system (Fig. 5a). It did this by using a directional coupler instead of a hybrid splitter at the distribution amplifier and by making better use of amplifiers and cable.

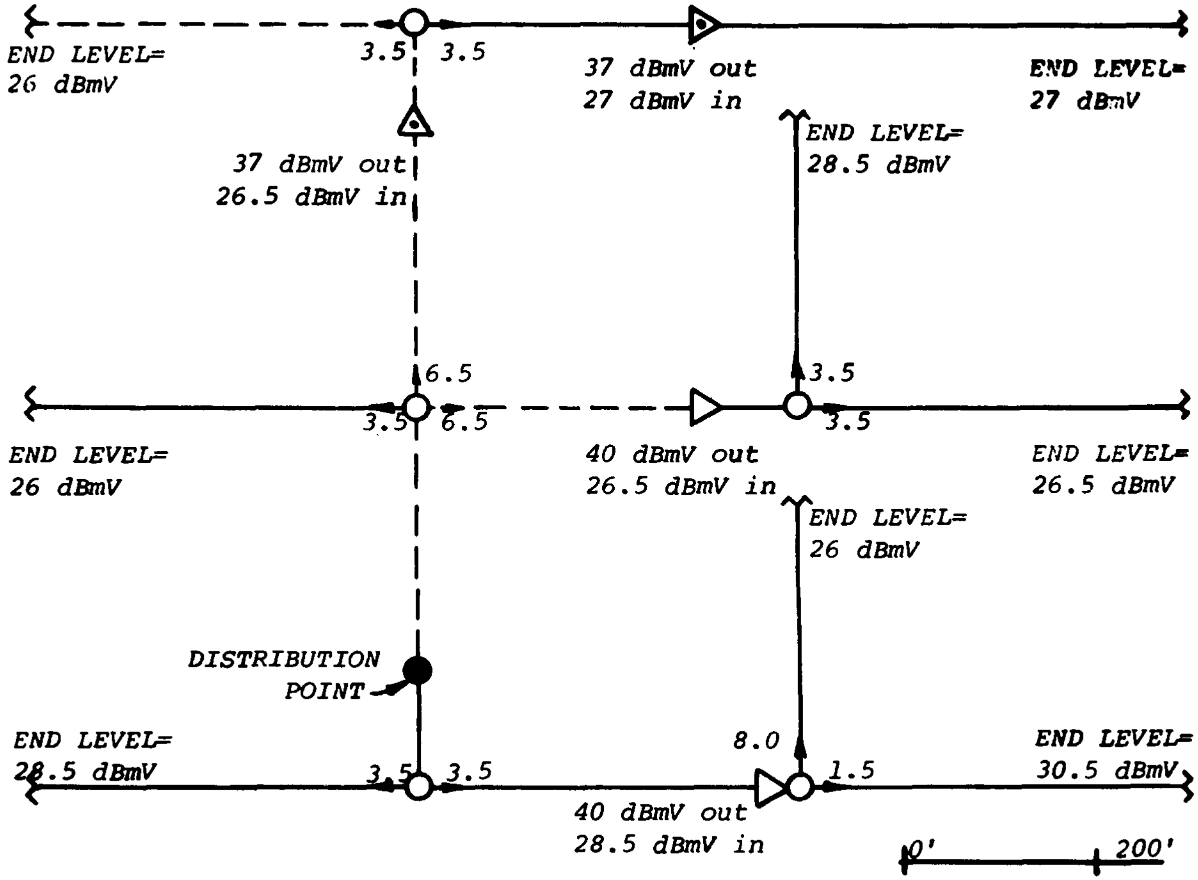


Figure 6a
Manual design.

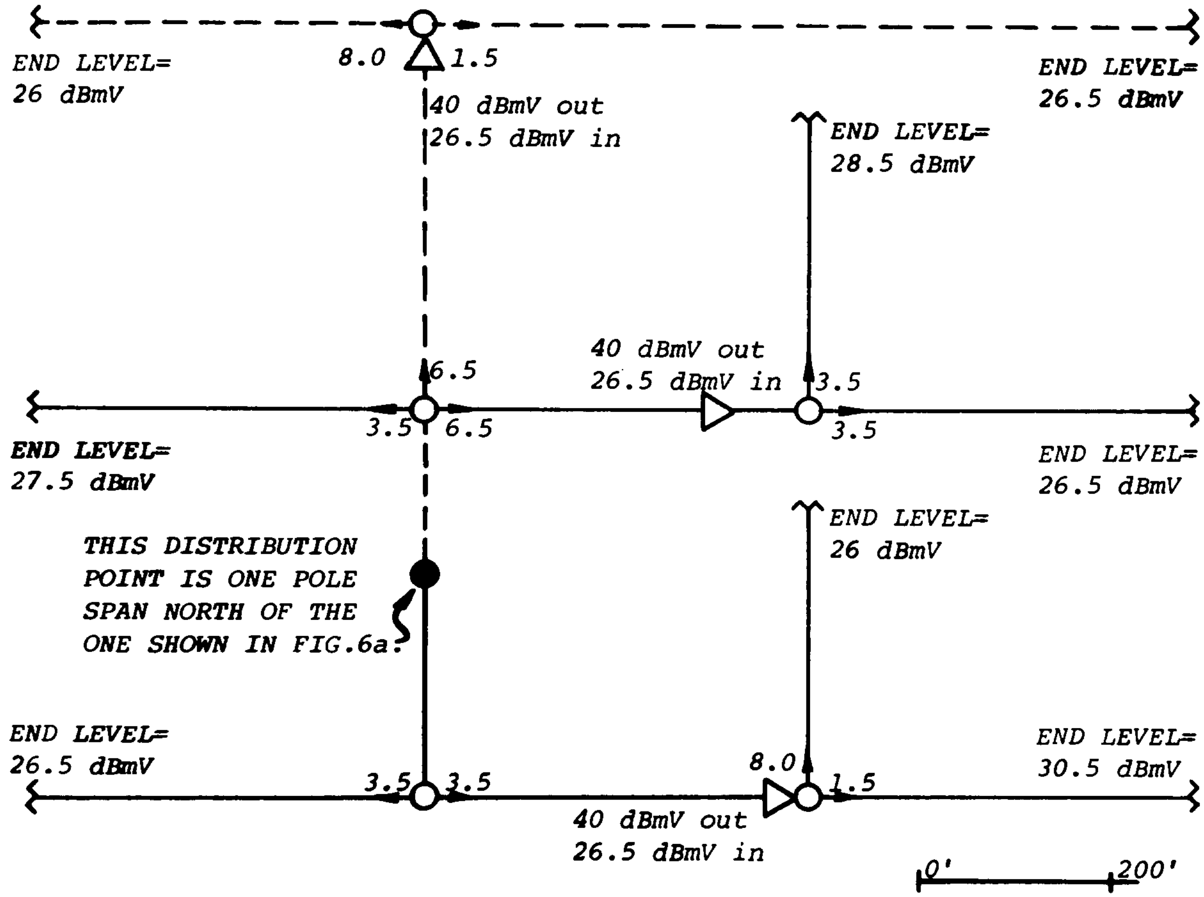


Figure 6b
In NAC's computer design, a 100 foot difference in distribution point location saves $140 or 13%.

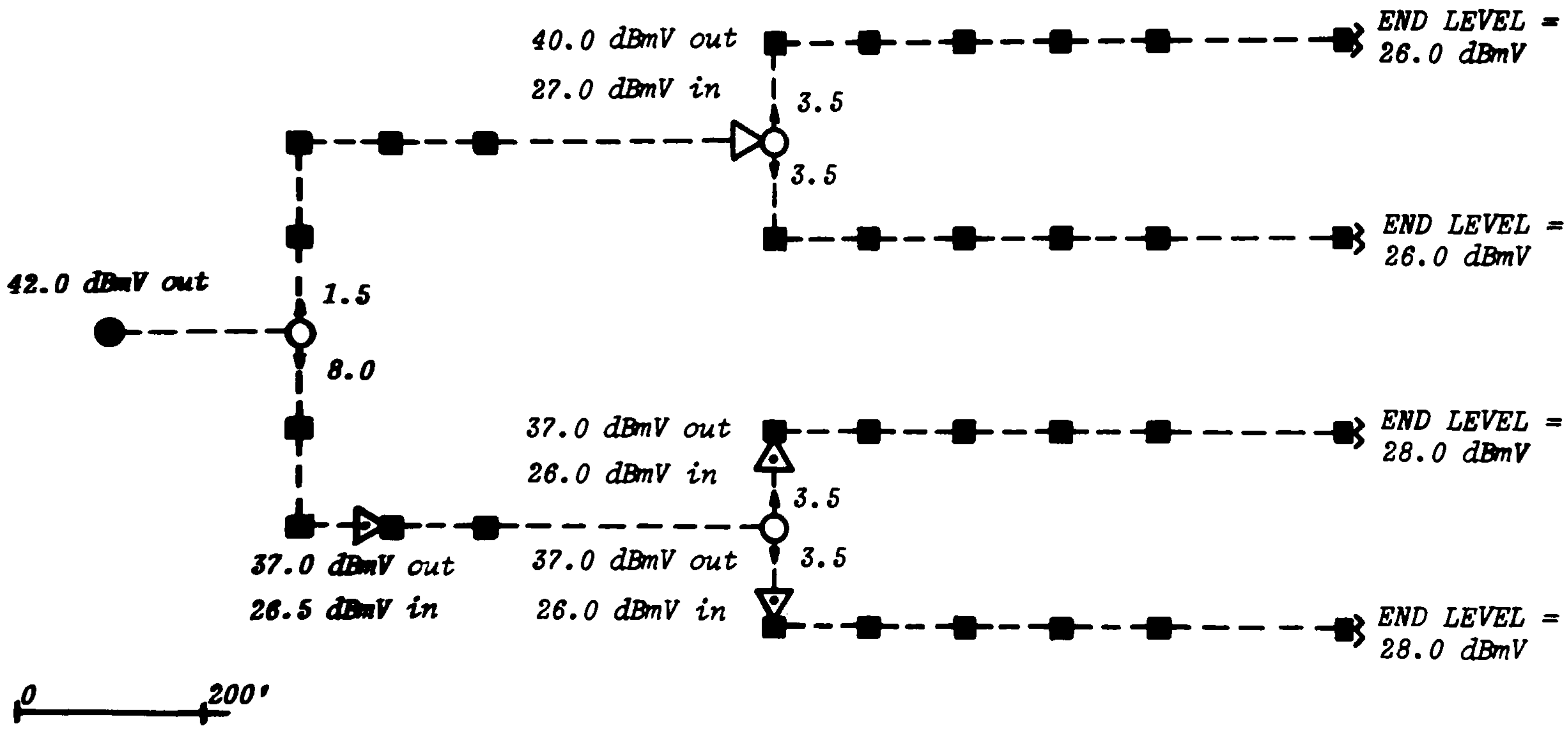


Figure 7
A manual design allowing 6 dB flat loss for taps.

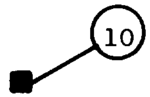
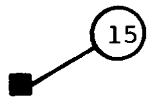
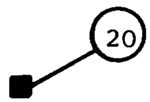
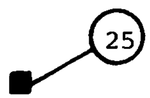
Tap Symbol	*Tap Loss at 270 mHz (dB)*	*Insertion Loss at 270 mHz (dB)*
	*10.0*	*1.5*
	*15.0*	*1.0*
	*20.0*	*0.5*
	*25.0*	*0.4*

Figure 8
Subscriber tap characteristics.

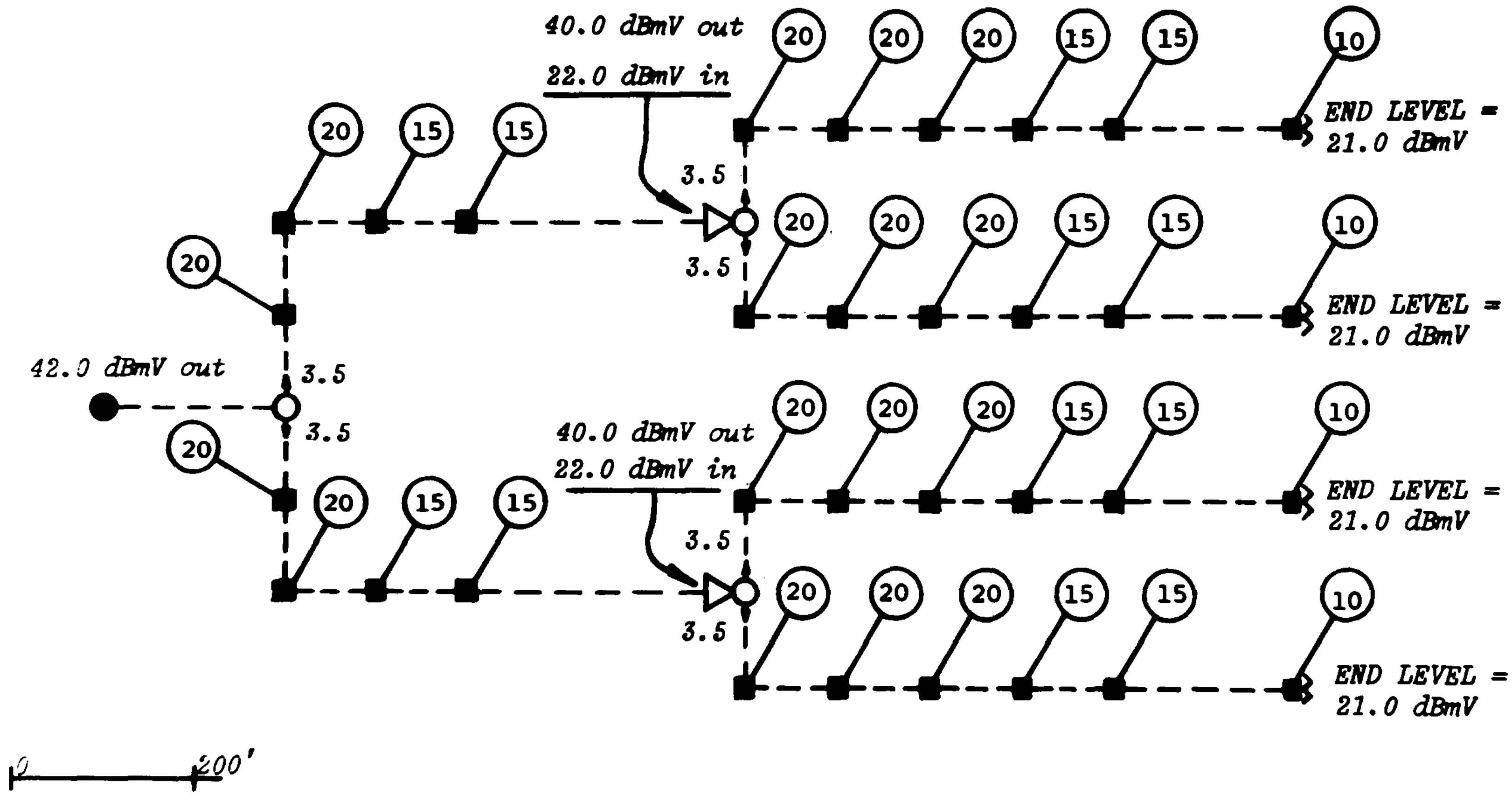


Figure 9
The computer design takes the tap characteristics into account in the optimization. The design above contains two less extenders than the manual design in Figure 7.

BIOGRAPHY FOR NCTA

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Mr. Mason is the Technical Director of MITRE's Systems Development Division. He has been with MITRE for ten years, working on a variety of civil and military systems projects.

Before MITRE, he worked thirteen years at Hazeltine Electronics Corporation on radar and beacon systems.

He has an ME from Stevens Institute of Technology and an MSEE from Brooklyn Polytechnic Institute.

BIOGRAPHY

KENNETH J. STETTEN

Associate Department Head, MITRE Computer Systems Department and Project Leader of MITRE's TICCIT (Time-Shared Interactive Computer-Controlled Information Television)

Mr. Stetten joined MITRE in 1968. He is the originator of the TICCIT minicomputer technical concept and is the co-inventor (patent pending) of the Television Home Computer Terminal. He has had 15 years of diversified experience in information systems and other technologies. His previous work encompasses the areas of CRT development and TV systems, space electro-optical instrumentation, infrared chemical warfare and airborne contaminant detection/identification and a vast range of related technologies. He served as an officer in the U.S. Air Force Special Weapons Center. He holds 5 U.S. patents.

Mr. Stetten holds a B.S. in Engineering Physics from Lehigh University and an M. A. in Physics from Boston University.

A LOW COST INTERACTIVE HOME TV TERMINAL

William F. Mason & Ken Stetten
The MITRE CORPORATION

As part of the design of a cable system for a large city, MITRE has had to take an objective look at the many ideas that have been proposed relative to providing new services into the home via cable. Part of the analysis involved identification of the variations in home terminal hardware that would be required to provide these services.

Types of Home Terminals

Figure 1 indicates one way in which the hundreds of services that have been proposed can be grouped into the eight fundamental hardware configurations that would be needed to provide them. Starting on the left of the figure with conventional black and white television, the cost rises with color or an A-B switch or converters, as additional channels are added. In group two we add the ability to encode transmissions for selective distributions to people who have unscramblers in their homes. Special distribution/capabilities on the network can also be "hardwired" to provide exclusive distribution to groups of subscribers, e.g., doctors.

Adding some form of a frame grabber (third category) allows distribution of another class of service wherein a single channel can provide many different displays, e.g., stock reports, ballgame scores, local activity schedules, etc. In this case, each frame is coded and the subscriber can set his decoder to choose any of hundreds of services on a single channel (more about this later).

Going now into two way services, we have the conventional voting or polling capability wherein a central computer polls the network and accepts the votes or selections input from each subscriber. This class of terminals ranges from inexpensive, where there are only a few voting options, to an elaborate alphanumeric keyboard capability for use in general two way communications.

In category five, a credit card checking device is added to the network capabilities of category four. This category is listed separately because of certain validation procedures that should be part of such a system. Meter reading, burglar and fire alarms, etc., are possible in the next category, wherein various types of sensors in the home are used to measure or monitor various phenomena and then report to the central system using the fundamental hardware provided in category four, but with special interface to the various sensing devices in the home. We include in this category the ability of the central system to control certain devices in the home if desired. For example, the second heating element in a hot water heater can be controlled by the utility company to help alleviate peak power problems. Utility customers have been offered lower rates for such cooperation.

Category seven provides interactive communications of the type that would be needed for sophisticated computer aided instruction (CAI) or computer mediated instruction (CMI) into the home. Finally, we have services that involve high bandwidth digital or video communications between terminals.

Demonstrating These Capabilities

In order to examine the hardware involved in each of these categories as regards both cost and technical feasibility, MITRE has installed in six homes in Reston, Virginia, a terminal system that is capable of providing most of the services in Figure 1. Reston is located about 10 miles from the MITRE facilities in McLean, Virginia, and our computer is connected via a microwave link to the Reston Transmission Company headend in Reston. Reston has a dual cable distribution system and Channel 13 of their A-cable is used to distribute computer interactive services. Figure 2 is a schematic of the system showing that the computer provides on a single channel, 600 different frames of information, any of which can be selected by any subscriber having the appropriate terminal equipment. We call this the "public service" channel because of the types of information we are putting on it. The subscriber simply selects with a thumb dial on the home equipment, Figure 3, any of the demonstration material that we are interlacing (Figure 4). The particular materials we are providing are simply to show the types of things that may be offered on such a system.

The demonstration terminals also have the capability to let the subscriber interact dynamically with the computer. Since the Reston cable system is not equipped for two-way services yet, we are using telephone lines for the up-link from the subscriber to the computer. (Within the MITRE facilities we use two-way cable.) This class of service allows the subscriber to telephone the computer and receive a response on his TV screen that is not seen by other subscribers. He first receives a frame that introduces him to the services available. The touch-tone telephone is used as an input device to select services. Our demonstration allows subscribers to take computer-aided instruction in math or to use the computer to perform mathematical calculations (add, subtract, multiply, divide, raise to powers, take square root, store, etc.) or, more simply, to use the computer to sort through a variety of information. For example, we allow the subscriber to look up telephone numbers using his TV screen.

The "home calculator" demonstrates how computational capabilities can be provided in the home. The educational and social materials illustrate more wide ranging possibilities.

Our system also has the capability to record programs off-the-air or to record movies addressed to a single address, possibly during the quiet hours of the night, for replay as desired. Although full movies by this method would be impractical, we plan to use this capability to send short sequences of frames that can be "stepped" onto and off the tape recorder in a manner that might be used for delivering newspapers or mail by TV.

In this demonstration MITRE has concentrated on the use of readily available equipment in the home, i.e., the TV set and the telephone. The only special equipment that has been added is the circuitry for decoding the material sent to the home and a video tape recorder, which we feel will be a common item in a few years. As a matter of fact, the technical services that we are demonstrating in Reston will have a tremendous influence on the popularity of these video tape recorders if the addition of the simple circuits to grab a frame and refresh a TV set are added, as in the models we are using.

Cost Information

Although this is not the platform for publicizing the information we have accumulated relative to cost, a few comments are in order. The Figure 1 chart indicates the general range of costs for home terminals of various types assuming reasonably large implementation. Two-way digital services are not all that "futuristic" but we have avoided putting bounds on what the terminals may cost because it is so dependent on bandwidths used, etc. On the other hand, a sufficient number of devices can be bought in the $150 to $350 range so that this category of services should not be considered impractical. Studies now being performed by and for various large companies indicate that a very practical system can be described for certain types of markets. Within each Figure 1 system classification there are of course cost variations, depending on the particular service offered, how fancy the terminal is to be made, production quantity, implementation density, etc. Each of you will have opinions relative to the particular category of services of interest to your company or provided by your system. MITRE would be interested to have your comments relative to our cost summary.

Now let's turn to the cost of providing the computer services. We have made a number of analyses that indicate that interactive service can be provided into homes at a cost of around 20 cents per terminal hour. Roughly it goes like this: a minicomputer center to provide services to a population of around 10,000 would cost about $150K. Amortizing this over four years gives $37K per year. Adding $18K maintenance, clerical overhead and floor space, gives $55K per year to provide to each of 10,000 subscribers one hour a day interactive with the computer. (The actual computer time is a very small part of this because we serve an average of 100 interactive terminals at a time; more during peaks.) This amounts to about $55.00 per year or $5.00 per month cost to the operator. This can also be considered as $5.00 for thirty hours of use, or about 20 cents per hour of use by the subscriber. He would probably pay several times this for the service. Billing would be handled by the computer. If the system operator provided $400.00 worth of capabilities from Figure 1, he would have to charge around $10/month, which seems very attractive for the kinds of services we are discussing.

Conclusion

A complete analysis of the types of capabilities discussed here will be published by MITRE in the near future, but sufficient information has been summarized in this paper to indicate our belief that the time to start large scale experiments with computer interactive TV is now.

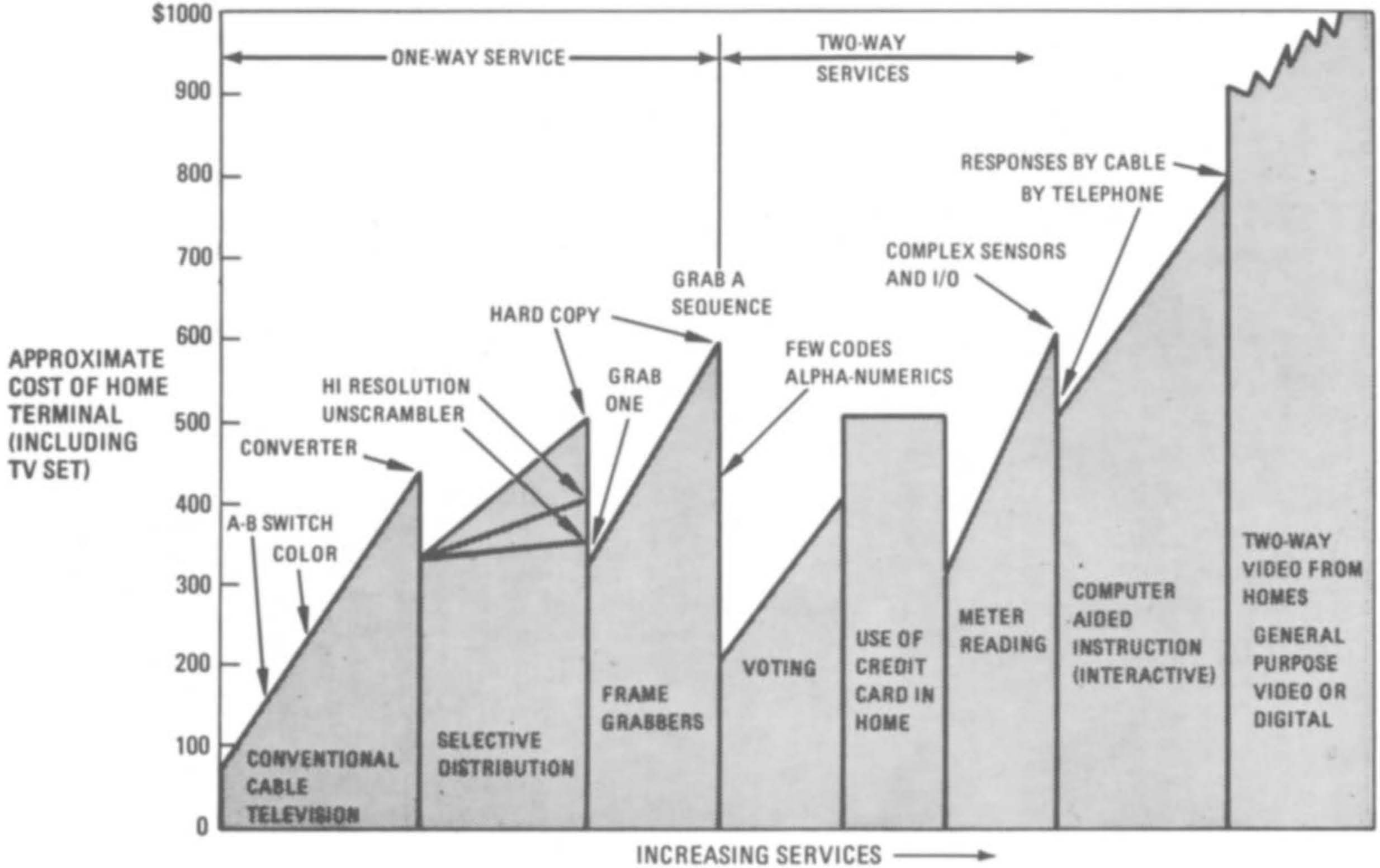

FIGURE 1
HOME TERMINAL OPTIONS

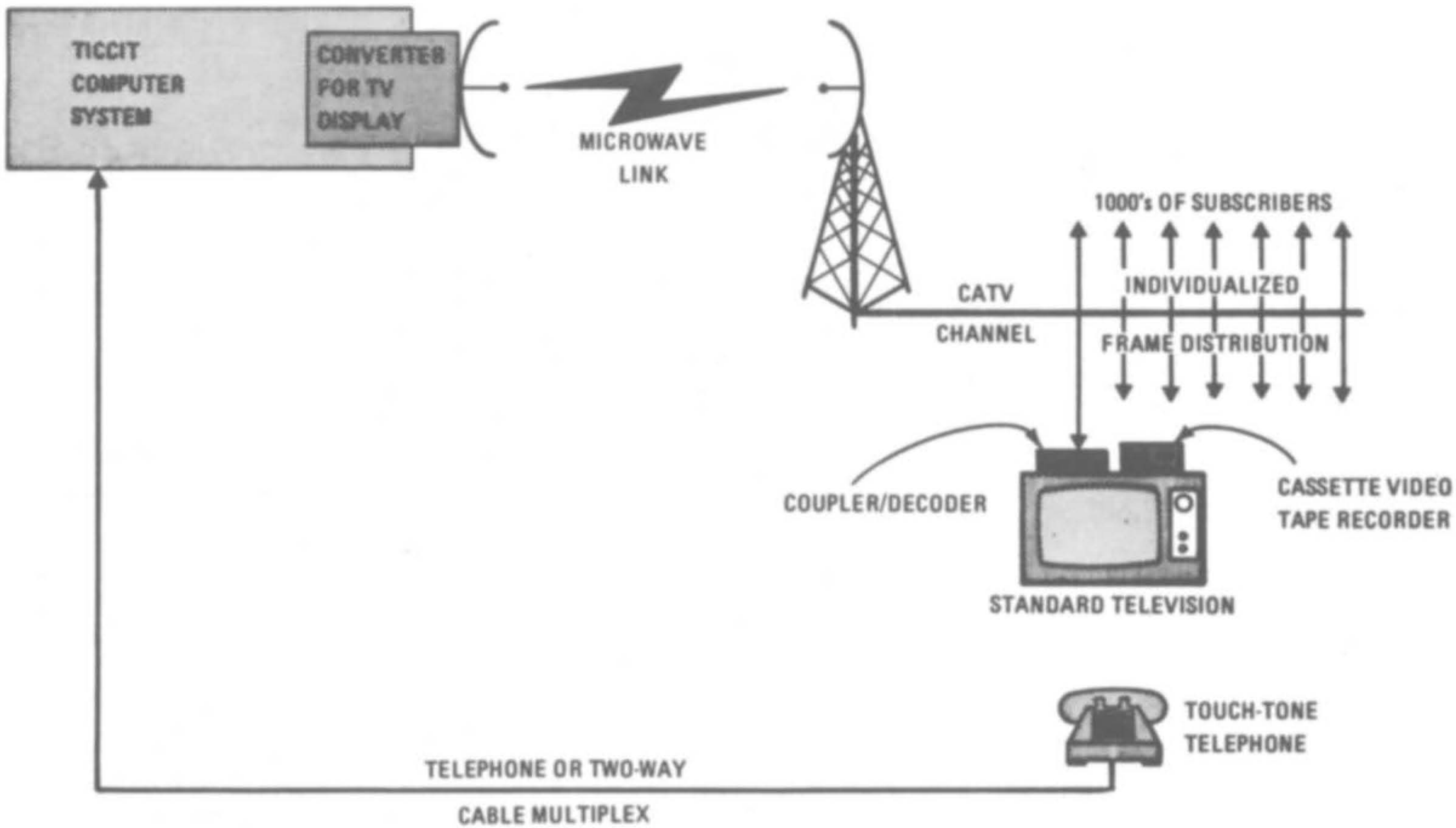

FIGURE 2
SYSTEM DIAGRAM

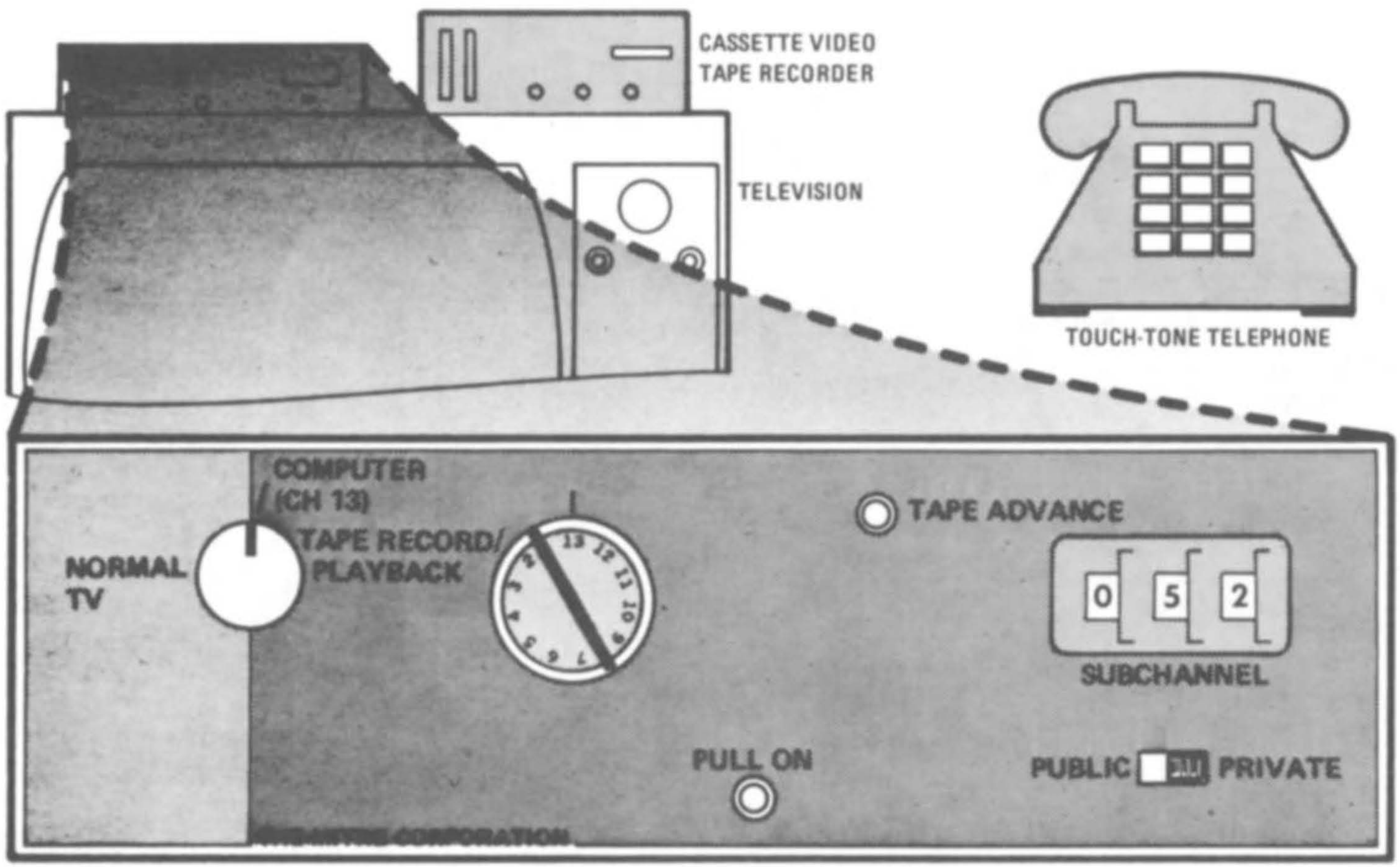

FIGURE 3
COUPLER/DECODER ("BROWN BOX")

DIRECTORY OF INTERACTIVE SERVICES

- RESTON COMMUNITY INFORMATION
  - DIRECTORY OF RESTON COMMUNITY ORGANIZATIONS – LIST OF 30 AND FILES ON EACH (SEE TEXT)
  - RESTON TELEPHONE DIRECTORY
- HOME CALCULATOR: ADD, SUBTRACT, MULTIPLY, SQUARE ROOT, RAISE TO POWER ETC.
- COMPUTER AIDED INSTRUCTION MATERIALS
  - ADDITION LESSON 2 DIGIT NUMBERS (CARRY)
  - ADDITION DRILLS (STANFORD U. CAI PROJECT)
- DIRECTORY OF NON-INTERACTIVE SERVICES (PUBLIC MODE)

DIRECTORY OF NON-INTERACTIVE SERVICES (PUBLIC MODE)

TIME OF DAY

ADVERTISEMENT

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WEATHER REPORT

DIRECTORY OF INTERACTIVE SERVICES (PRIVATE)

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FIGURE 4
LIST OF INTERACTIVE AND NON-INTERACTIVE SERVICES
AVAILABLE AT RESTON DEMONSTRATION

FIGURE 5
TERMINAL INSTALLED IN HOME

PAPER PRESENTED AT 1971 N.C.T.A. CONVENTION

Washington, D. C.

TeleMation, Inc.

Ronald S. Hymas

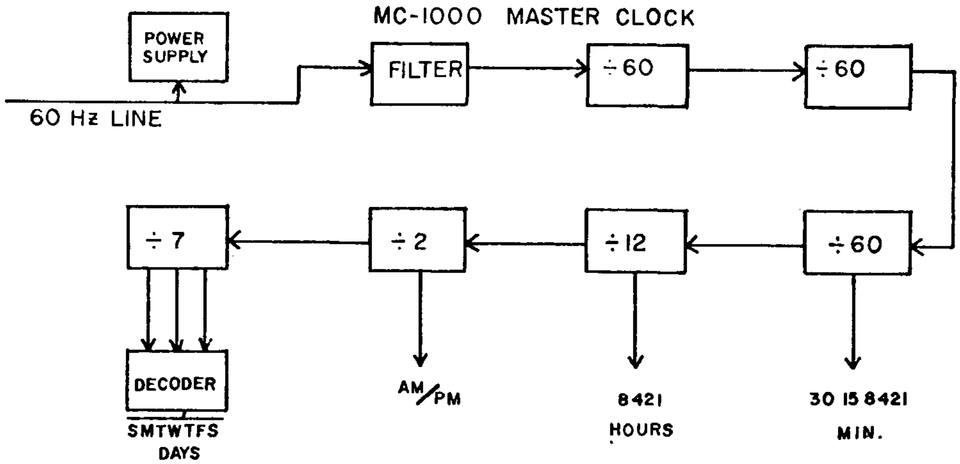
Salt Lake City, Utah

A NEW APPROACH TO ALL-DIGITAL NON-DUPLICATION SWITCHING

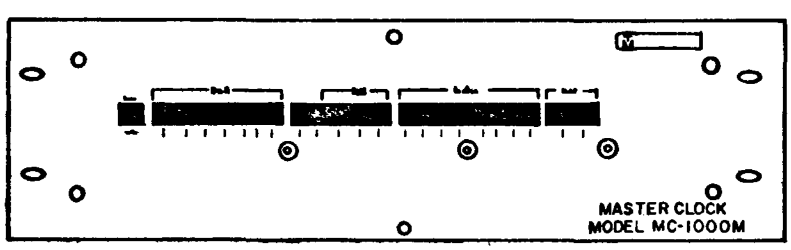
The TMP-1000 electronic programmer is designed to perform repetitive control functions such as CATV non-duplication switching when a schedule is repeated on a weekly basis. The programmer operates on the elasped-time principal. This means that the stop and start times of each event are programmed by the operator, rather than occurring at arbitrary intervals as with pulse-time systems. This permits programming the start and stop of each event to each selected minute throughout the day.

Digital electronics are used in a unique manner in the TMP-1000 to provide the benefits of larger, more expensive programmers that rely on disc, core, or tape storage, while maintaining a price competitive with the more limited pegboard programmers. Another important advantage of this design is in the elimination of moving parts which increases reliability.

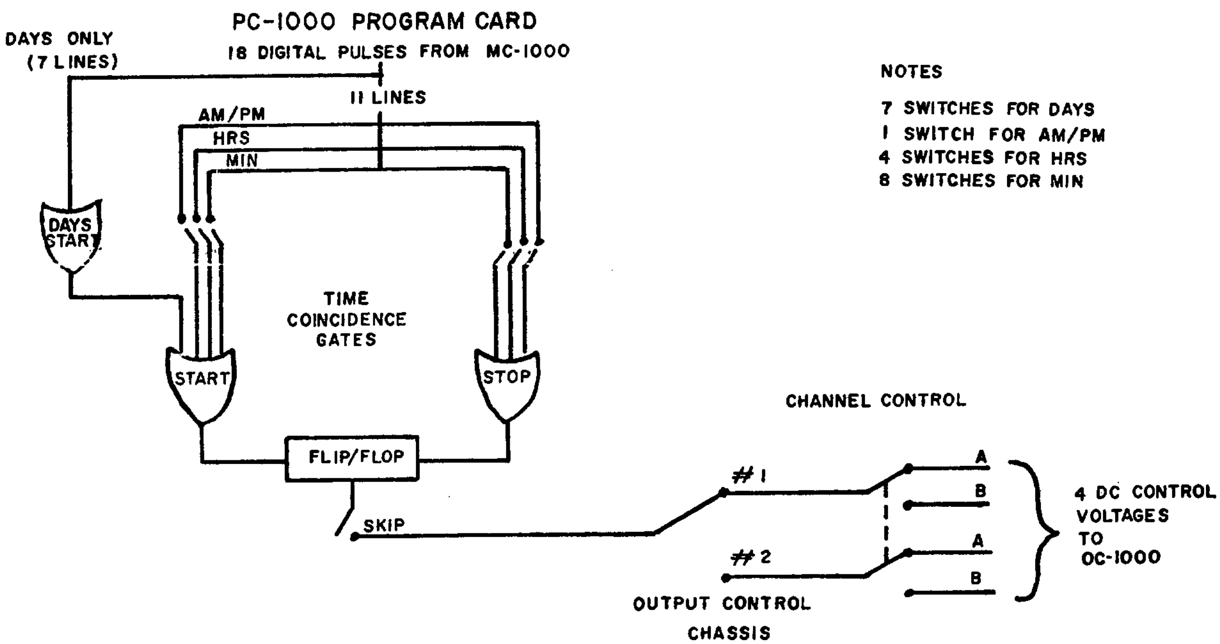
The TMP-1000 system consists of four basic components. They are: The MC-1000M Master Clock that provides timing signals and DC power for the system; the OC-1000 output control with its two-channel control relays; the EE-1000 event expander which expands the event capacity of the OC-1000; and the PC-1000 program control card which contains the programming and memory elements to start, maintain, and stop each event.



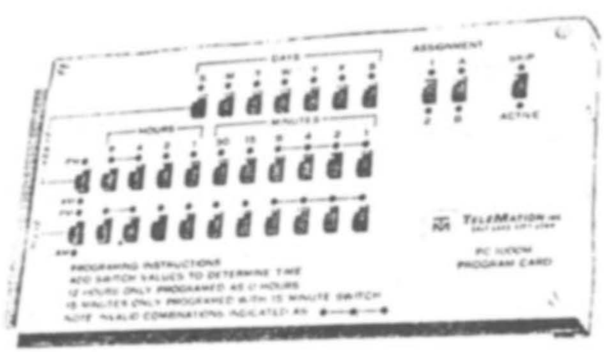
The heart of the system is a digital master clock, Model MC-1000M which counts down from the power line frequency. The frequency dividers are: Divide the power line frequency by 60 for a signal with a period of one second; then a divide by 60 gives you a minute period; a divide by 60 to obtain one hour period, and this frequency divided by 12 to get one-half day period. A divide by seven and decoder circuitry is required to obtain days. The output from the MC-1000 is then routed to the OC-1000 by ribbon wire, then to the PC-1000 Program Control Cards. The front panel controls for the Master Clock are three push-button switches which are used to set the days/hours, minutes, and seconds.



To set real time with the master clock, depress the day/hours control and advance the time to the proper day, AM or PM, and hours. Then depress the minute switch until proper time is indicated in the read out window. The "hold seconds" push-button is used to stop the digital clock in order to match the Master Clock to real time. This also clears the second counters so when released it starts at zero seconds. The indicator to the far left on the Master Clock is used to notify the operator of a momentary power failure; if so, the indicator light will flash and real time must be reset.

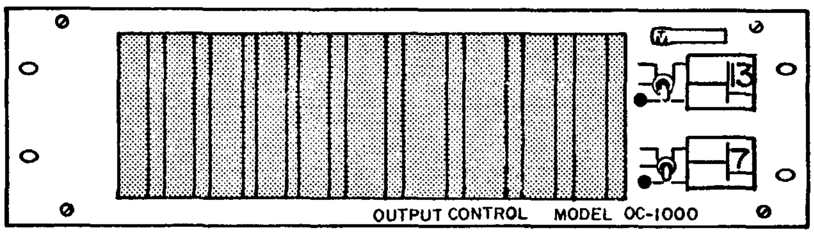


The PC-1000M Program Card controls one complete switching event from start to stop. This card also permits selection of day or days of the week that the switching event will occur. The 18 digital pulses from the Master Clock are routed to the days-only gate through the selected day switch. Eleven lines are then routed to the start time coincidence gates through the selector switches for AM or PM, hour, minutes switches. If one of the programmed days is the same as the day indicated by the Master Clock and the programmed start time is the same as the indicated real time, a pulse is fed to the flip-flop. The flip-flop acts as a storage device, and a DC voltage is then fed to the output control for the appropriate channel switching. This flip-flop will stay in the on condition until a stop time coincidence occurs. At this time the flip-flop will turn off. A pulse from the days gate is not required to achieve a stop time coincidence. A skip feature is built in to eliminate complete program card programming if a particular event requires no protection. If the skip feature is used, a front panel light will indicate to give you a record of the skipped event. An active light indicates on the front panel of the Program Card if a switching event is in progress.



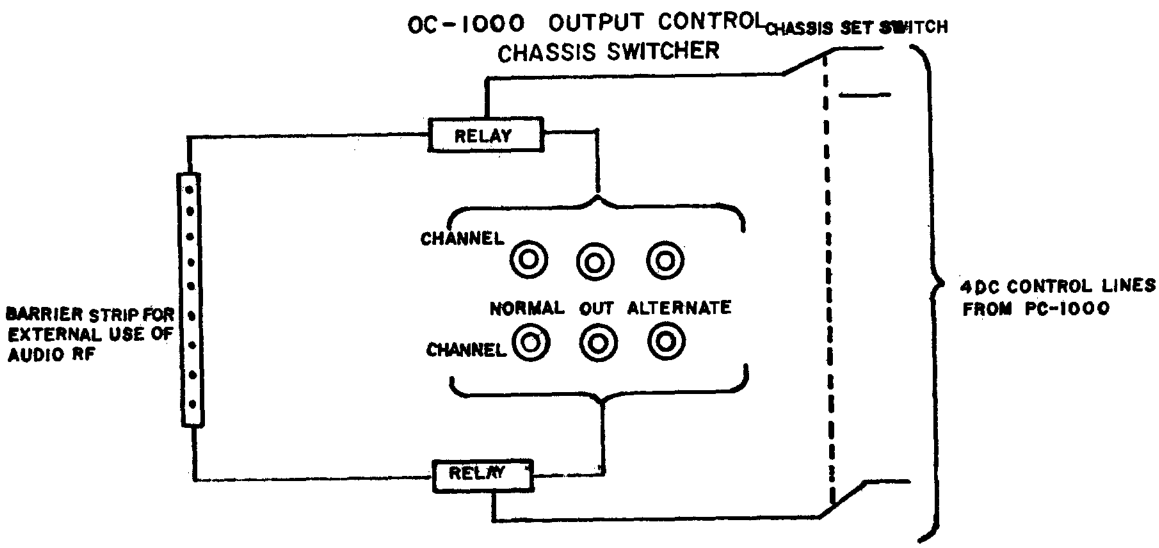
The output from the Program Card can be assigned to any of four different channels, depending on the placement of the assignment switches directs the output of the assigned output control chassis or OC-1000.

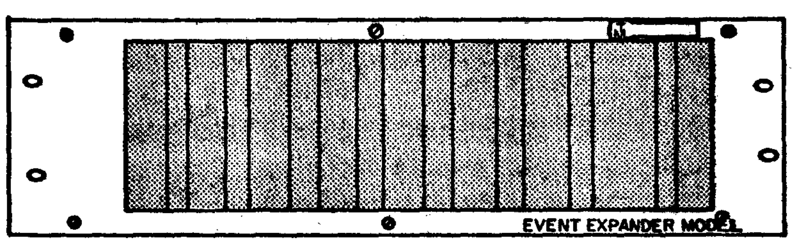
Each OC-1000 provides two channel protection, selected by the "A" and "B" switches. The four DC control voltages from the Program Card are then routed to the appropriate Output Control Switcher for relay switching.



The output Control Chassis, Model OC-1000, provides output switching capability. Front-panel switches are provided to allow operator manual override selection of either the "Normal" or "Alternate" input, or the automatic mode of operation.

The reference guide located by the selector switch gives you a log as to which sources can be switched to the output channel. In the alternate and normal mode these sources are routed to the output channel. In the auto mode you revert back to the program card. Circuitry is provided to directly switch video, with proper termination of the unused input. Additional contacts permit switching associated audio or RF control voltage. Video connection is by means of type BNC connectors, and barrier strips are supplied which switch either internally supplied +24 VDC or may be strapped to switch voltage from an external source. Each output control accommodates up to twelve program cards.





Event Expander Chasses, Model EE-1000, are available to accommodate additional Program Cards when the number of switching events designated to a given output exceeds the card slot capacity of 15 provided to allow a Program Card to be delegated to operate with either output in either of two associated OC-1000 Output Control Chasses. The Event Expander Chassis is a passive device, and may be connected as necessary to extend system event capacity.

Interconnection of the master clock with the output control and the Event Expander Chassis is made with a flat multiconductor ribbon wire. Sufficient length of this interconnecting wire is supplied with the Master Clock for future expansion. Connection of the ribbon wire to the Output Control and Event Expander Chassis is by means of a connector provided on the Output Control and Event Expander units. This convenient and easy-to-assemble wiring technique allows for rapid field connection of the component parts of a TMP-1000 System.

The TMP-1000 System has a capability of system expansion to six OC-1000s and fifty Program Cards which offer 700 switching events weekly for non-duplication.

A PRACTICAL APPROACH TO BI-DIRECTIONAL SYSTEMS

by

Perry L. Schwartz

MONMOUTH COMMUNICATIONS SYSTEMS

Presented at the
1971 NCTA Convention
Washington,D.C.

A PRACTICAL APPROACH TO BI-DIRECTIONAL SYSTEMS

by

Perry L. Schwartz

INTRODUCTION

In recent years there has been considerable discussion of Bi-Directional CATV systems. Most of these disucssions have directed themselves to a particular approach and spent most of the time innumerating the information and services that could be carried by the system.

I have spent considerable time in analyzing the various methods of providing Bi-Directional Transmission. The two techniques which become obvious are Dual Cable and Single Cable with diplexing filters. The techniques are not new or revolutionary, but simply show that today's system owner can build a good quality two-way system at minimal cost.

REVIEW OF VARIOUS TECHNIQUES

1. Closed Loop

In a closed loop system the system feeds out from Head End and physically returns to it in one continuous path (fig.1). This usually preclueds that the feeder system is not Bi-Directional. Needless to say this technique is extremely costly and sometimes impossible, due to available continuity.

2. Single Cable

A single cable technique, which I will discuss in more detail later, requires diplexing filters and sub-band amplifiers (fig.2). When using standard diplexing filters considerable care is required in design because a cascade of any length (10 or more) will cause bandwidth shrinkage.

3. Converters

This technique usually requires large and costly installation and the converter itself must be mounted within an outdoor enclosure. In addition to this, this system will usually utilize a single cable or dual cable technique for signal transmission. In some cases a multitude of cable are required with large switching centers and specialized converters at the home.

4. Dual Cable

A Dual Cable system is basically two separate systems connected only at the Head End and at the subscriber's home. This describes a total Dual Cable system; however, as I will discuss later, by carefully selecting feeders the Dual Cable technique becomes a versatile Bi-Directional system.

BI-DIRECTIONAL TRUNK:

Single Trunk Cable: It appears at first look that the single cable trunk Bi-Directional system would be to most, desirable from an operator's point of view. This is probably the case for existing single cable CATV systems. In order to upgrade the existing single cable system, a diplexing filter package can be used, which enables a system operator to make this conversion at a minimal expense. It should be noted however, that little is known at this point about the differential time delay of a single cable Bi-Directional system. In addition to the time delay question, the added noise of the sub-band amplifiers in the return loop must be added to the total signal-to-noise ratio of the Uni-Directional system.

Single Cable System - Diplexing Filters: By means of the diplexing filters and sub-band amplifiers, you can convert your existing CATV system to a Bi-Directional Trunk system. The package (fig.3) utilizes low loss diplexing devices, and a sub-band amplifier. The sub-band amplifiers utilize an automatic gain control to compensate for temperature changes in the coaxial cable below 54MHz.

DUAL TRUNK BI-DIRECTIONAL SYSTEMS:

The Dual Trunk method of providing a Bi-Directional transmission system is at this point in time, the most desirable. The features of a dual cable Bi-Directional system are:

1. Trunk Integrity - Adding devices to any trunk line can only degrade picture quality and add noise.
2. Band-Width (Channel Capacity) - When using a two cable system, it would be possible to carry as many channels in the forward direction as in the reverse direction.

Dual Cable : When constructing a new system using dual trunk cable the most desirable method would be to use a standard broadband push-pull 54-300 MHz amplifier in the forward direction, and an amplifier capable of 15 to 90 MHz for the return loop (fig. 4). It should be noted that return amplifiers are not needed at every location if the cable is the same in both forward and return loops. It can be shown that with careful design a return system using 0.500 cable and 90 MHz amplifiers on the return trunk is not very much more expensive than the basic uni-directional system. It should be noted that the return amplifier could be replaced by a trunk amplifier (54 - 300 MHz) and provide full return channel capacity. Alternately, the return amplifier could be a 6 - 90 MHz amplifier and provide added return channels below 54 MHz.

A unique feature of a system of this type is that the frequency range of 15 - 40 MHz can be used for return signals from selected feeders while the band from 54 - 90 MHz can be used for return signals originating along the trunk (fig.5).

BI-DIRECTIONAL FEEDER SYSTEMS

All of the techniques described above can utilize the standard line extender for Uni-Directional feeder systems. Under some conditions, it may be necessary to provide a return signal from a feeder leg. This return path can be provided by means of a diplexing filter and sub-band amplifier (fig.6).

This system utilizes a standard line extender coupled with diplexing filters and a return sub-band (15 -40 MHz) amplifier.

It should be noted that sub-band amplifiers are not needed at every station. The reason diplexing filters can be used so easily at these stations is that in a feeder system, the cascade does not exceed a maximum of 4.

In a Bi-Directional Feeder system the available bandwidth would allow for both video transmission and data; however, it does not appear to be feasible at this time to transmit video through all the passive devices because of the amount of input energy required.(fig.6) If a picture were to be transmitted from the home on a time-shared basis signals of the order of +60 dbmv would be required. At the present time I know of no RF modulator in the band below 40 MHz that can produce this type of output at a cost attractive to this application. Therefore, if only data is transmitted from the home, considerably less bandwidth is required, and much looser specifications can be tolerated on the return amplifier and filters.

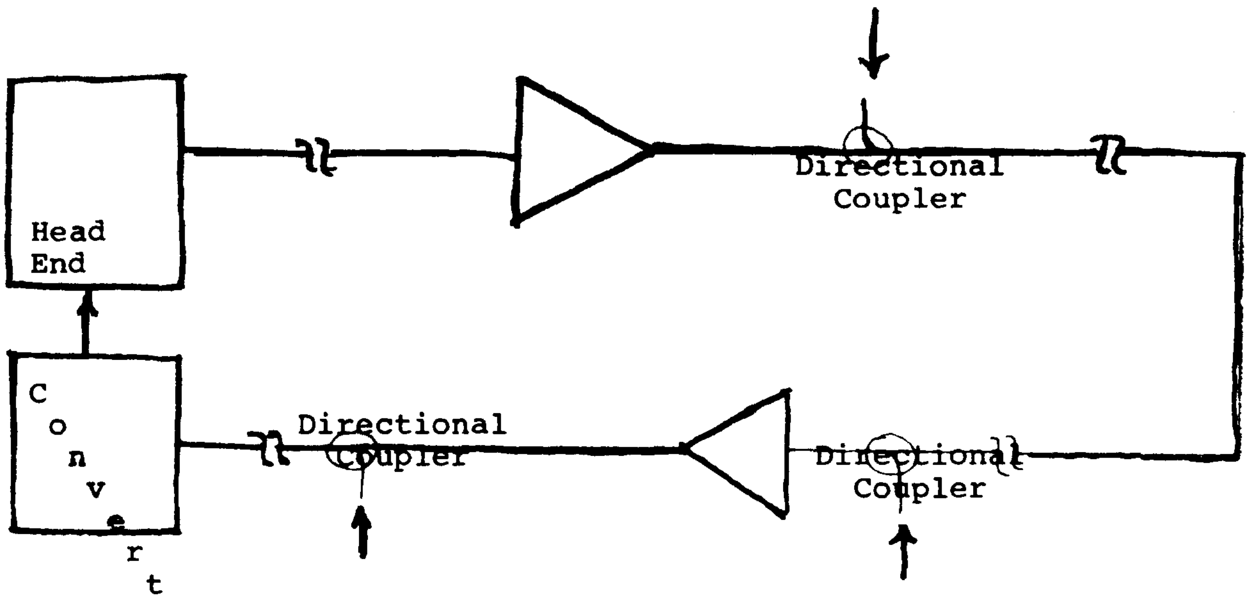


Figure 1 Closed Loop System

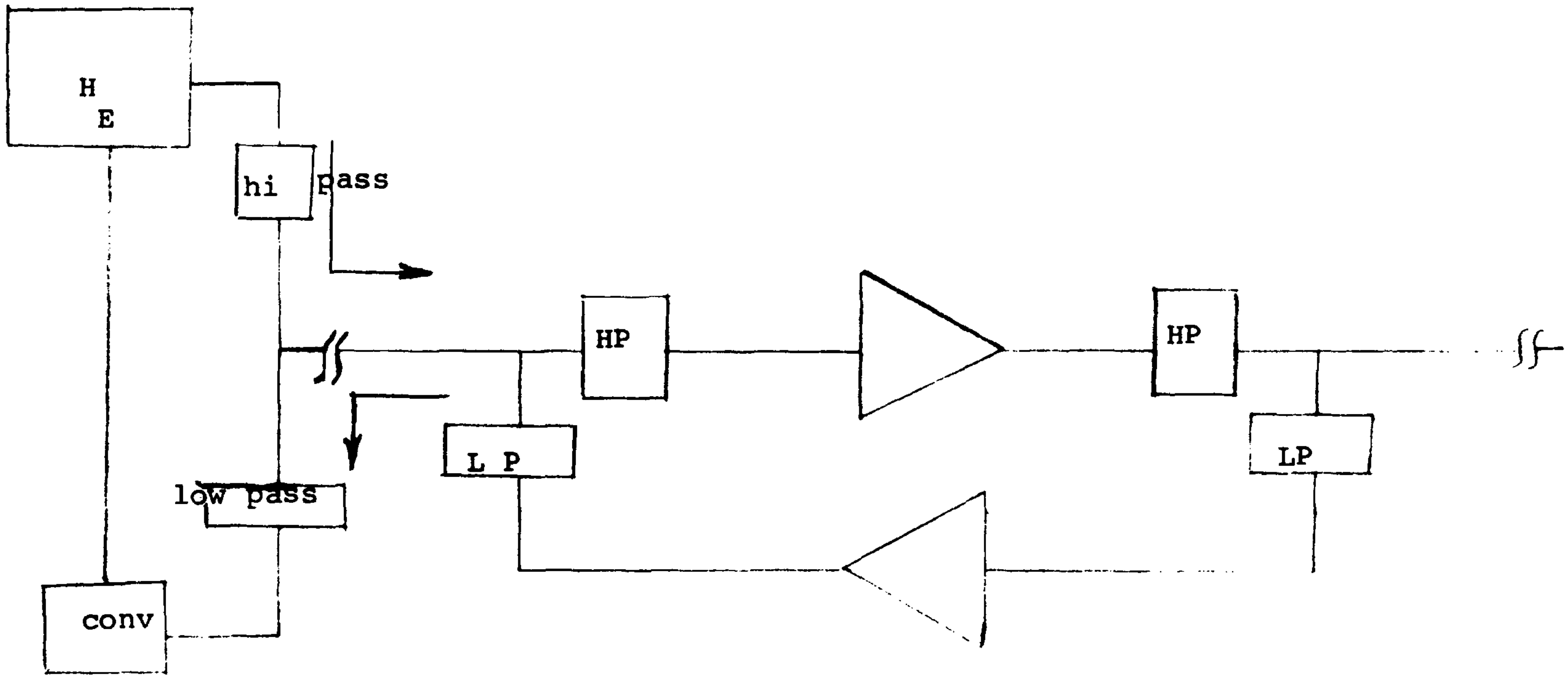


Figure 2 Single Cable System using standard diplexing filters

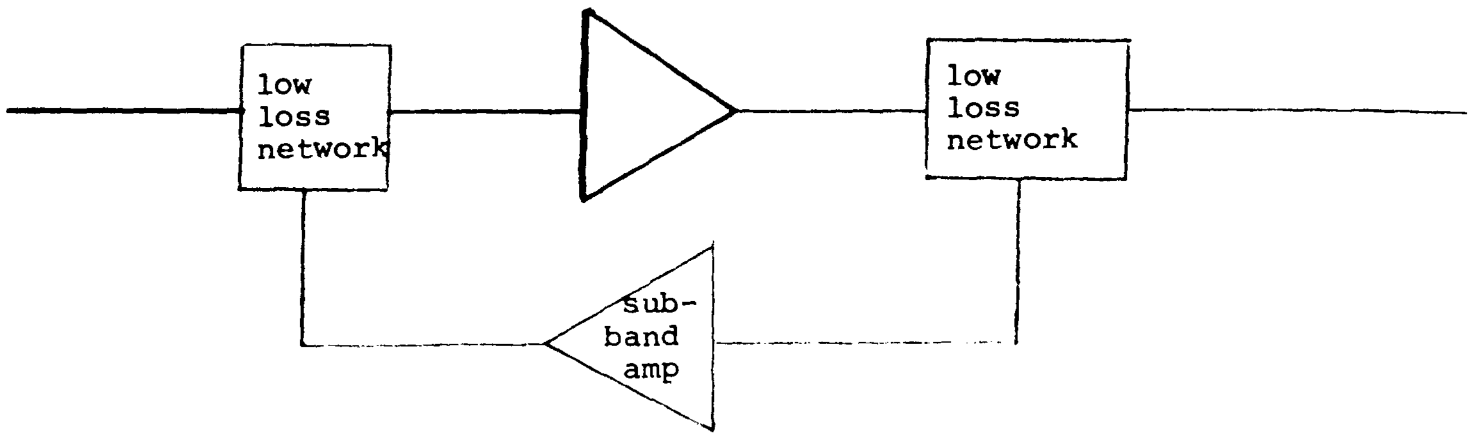


Figure 3 Single Cable System using loss filter networks

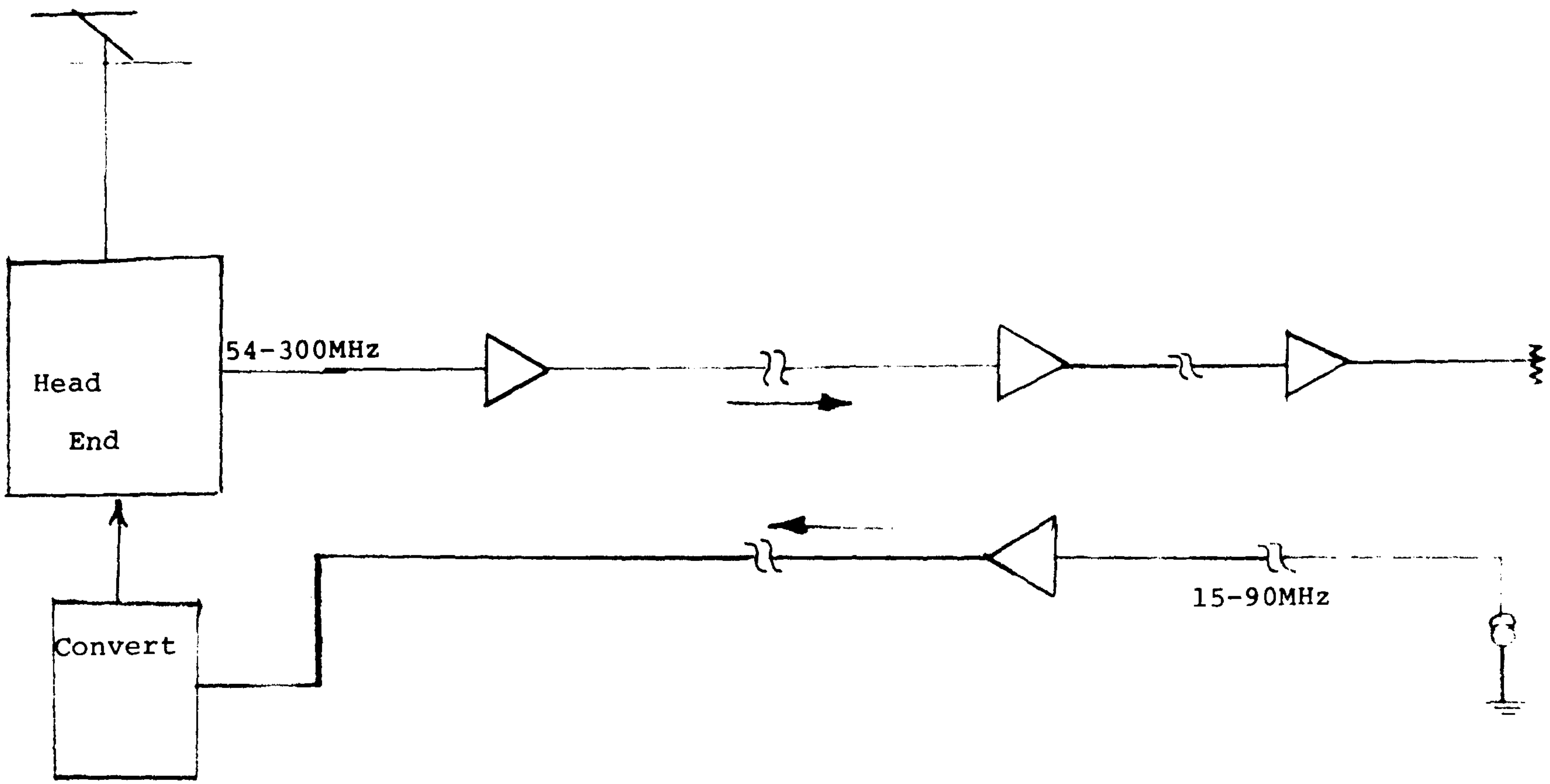


Figure 4 Dual Cable Bi-Directional System

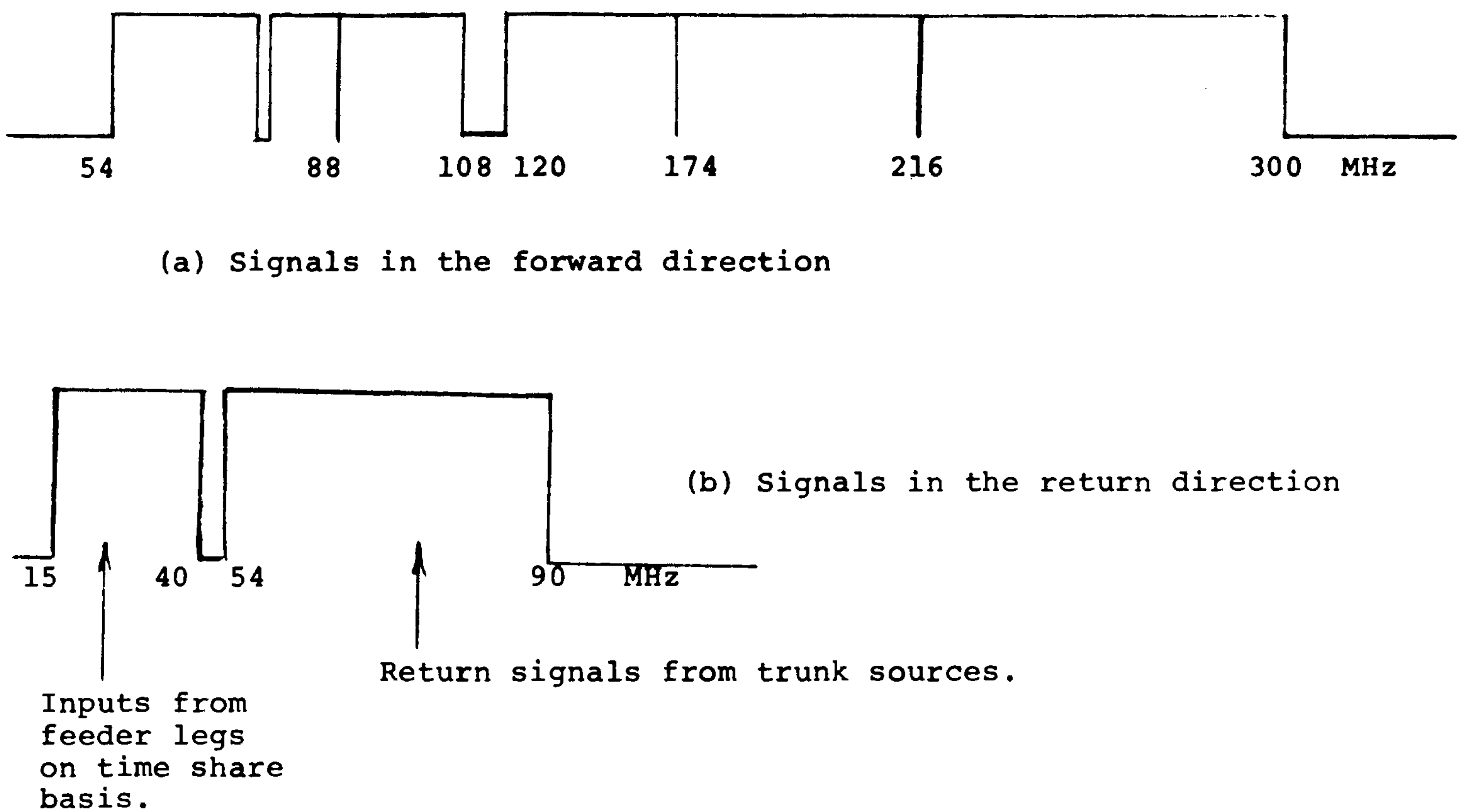


Figure 5 Band Pass of Dual Cable Bi-Directional System

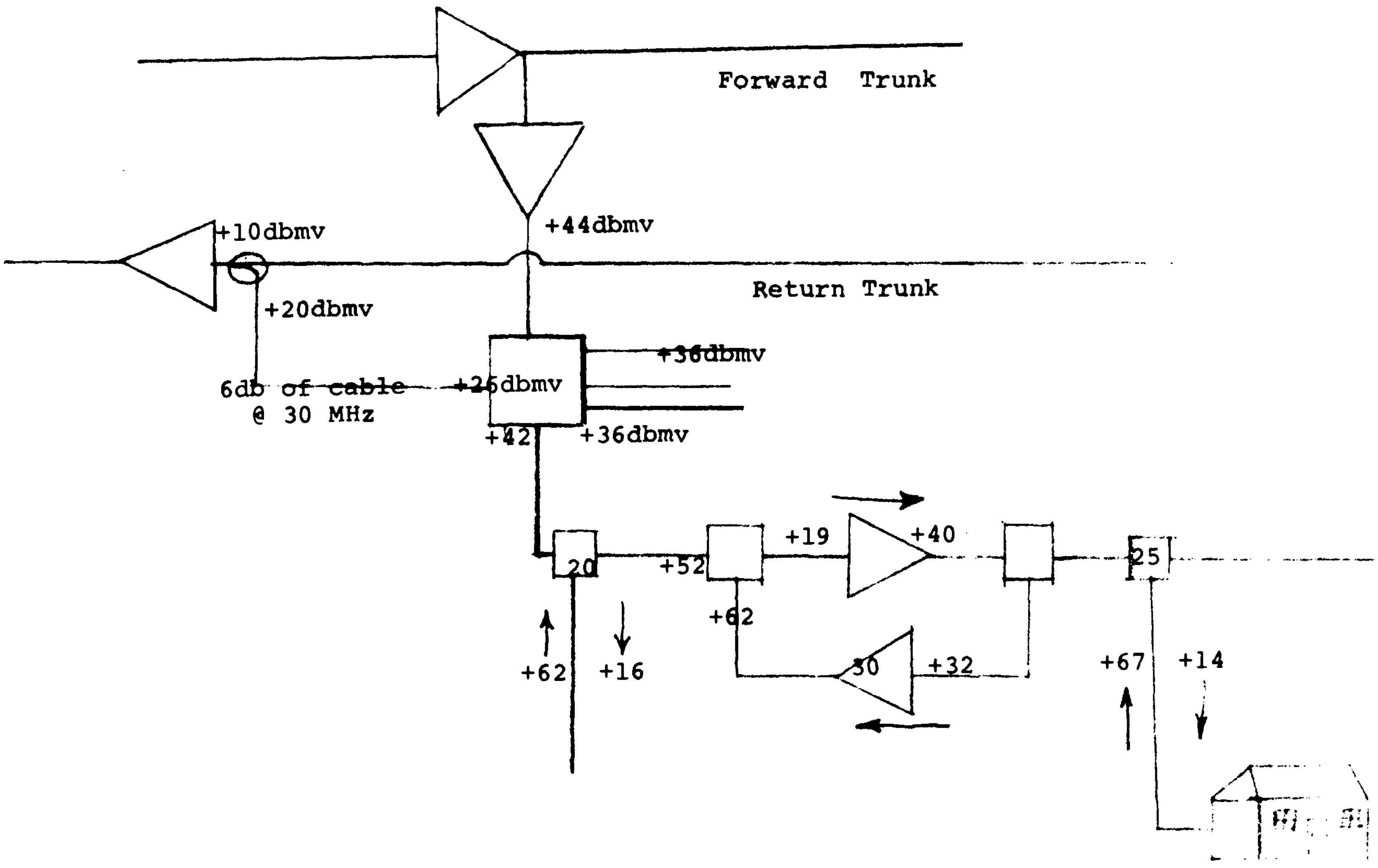


Figure 6 Bi- Directional Feeder System

A SECOND GENERATION CATV CONVERTER

BY

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OAK ELECTRO/NETICS
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PREPARED FOR PRESENTATION TO TWENTIETH ANNUAL CONVENTION
NATIONAL CABLE TELEVISION ASSOCIATION, WASHINGTON, DC
JULY 6-9, 1971.

A SECOND GENERATION CATV CONVERTER

INTRODUCTION

Set top converters have been on the market for over five years. For the most part, they have performed their intended function admirably. However, due to changing requirements and more stringent specifications, we are moving into an era in which second generation converters are required.

BACKGROUND

Set top converters evolved as a means of solving the direct pickup problem prevalent in strong signal metropolitan areas.

The method used was to down convert the cable signals to a 44MHz I-F and then up convert to a channel that was unoccupied by a broadcast signal. The RF circuitry to perform this function was enclosed in a well shielded metal compartment. The basic tuning methods were well established VHF TV tuner concepts and hardware.

It soon became apparent that the set top converter did not have to be limited to a meager twelve channels. To achieve more channels, the initial attempts were to either add another TV type tuner or to expand the coverage of the existing tuner. In either case, the 44MHz I-F was retained. As a result, the local oscillator signal emanating from the tuner antenna terminals fell into the CATV signal band. When only twelve standard channels are used, this is relatively unimportant. When mid and super bands are used, it results in converter to converter "cross talk".

GAMUT 26 - DESIGN PHILOSOPHY

In reviewing optimum approaches to second generation converters, a number of factors should be considered.

GAMUT 26 - DESIGN PHILOSOPHY (cont'd.)

Unlike broadcast, the cable signal levels operate within a relatively narrow range (-6 to +12db mV). Thus a dynamic, AGC able front end is not a requirement.

The future of cable is dependent upon providing additional channels and services to the subscriber. It would seem prudent, therefore, to avoid any energy emanating from the converter that could penetrate into the cable system.

The customer interface must be simple to operate and non-ambiguous.

The cost/performance factor must be equivalent or better than first generation converters.

DESIGN APPROACH

In developing the Gamut 26, we abandoned the use of a "TV tuner" as a tuning element. Also, the use of a high frequency I-F is mandatory if oscillator energy is to be prevented from entering the cable system.

Our approach can best be illustrated by referring to the block diagram. Figure 1.

The cable signal flows through a set of filters into a passive double balanced mixer (DBM). It is combined with a local oscillator frequency to form an I-F of 330MHz. Twenty-six channel tuning is achieved by simply switching in the appropriate oscillator frequency.

The I-F amplifier provides gain and selectivity to the up converted signal. Combining with a fixed tuned 2nd local oscillator, the signal is down converted to the desired output channel.

DOUBLE BALANCED MIXER

Little trouble from intermodulation, in the form of sums and differences of the signal or from its harmonics, are experienced when using the standard twelve channel VHF frequency allocation. Historically, these frequencies were carefully chosen to avoid this.

With the use of multichannel coverage, second order distortion becomes a severe problem. In particular, the low band can react with the mid, high and superband to create spurious signals which manifest themselves in the TV receiver as "beats". For example, the sum of a channel 2 and a channel A signal will produce an interference in channel 7. It has been our experience, and that of others, that the suppression of these beats should be in the order of 60db. See Ref. 2. It is for this reason that we have used a double balanced mixer as a front end.

Its operation can be described by referring to Figure 2. The RF signal is applied at the input port and combining with a strong local oscillator at the second port is converted into two side bands, equally spaced about the local oscillator signal. These side bands are available at the third port. In the case of the Gamut 26 the lower sideband (330MHz) is used as the I-F. It will be noted that the images (upper side bands) are in UHF, well out of the cable signal band.

By virtue of its inherent balance, the double balanced mixer (DBM) is a device that automatically provides a great deal of suppression to second order beats. Good port to port isolation and excellent cross modulation characteristics makes the device well suited for CATV applications.

BACK TO BACK OSCILLATOR

To tune the DBM, twenty-six discrete, ultra high frequencies are required (386 to 572MHz). These local oscillator signals must be high level, accurate and repeatable.

This is achieved by using the basic hardware from a VHF tuner transistor oscillator. Two oscillators were mounted "back to back" on a common shaft. Instead of thirteen positions common to VHF tuners, twenty-six become available. The memory fine tuning mechanism became a means of aligning the oscil-

BACK TO BACK OSCILLATOR (cont'd.)

lators in production. The B+ was fed through the tuning coils thus providing automatic switching to the appropriate oscillator. See Fig. 3.

As there is a separate tuning inductance for each frequency, the sequence of operation is completely independent. This is an advantage in the Channel Selector readout as the numeric channels of the low and high bands can be arranged in sequence as can the letter channels of the mid and super bands.

Detenting is provided by a large index wheel mounted mid-way down the shaft. By devoting careful attention to this detail we have achieved reset accuracies superior to VHF tuners even though the oscillators are working at UHF frequencies.

The circuits are basic, temperature compensated, Clapp type oscillators using high F_T transistors.

PASS/STOP FILTER

As well as the DBM functioned, we found that we could not maintain the proper balance in production to guarantee the 60db suppression of beats with an input signal level of +12db mv.

To ensure compliance to this specification we found it necessary to insert, what we term, a pass/stop filter into the circuit prior to the DBM. This device acts as a lowband filter when the converter is tuned to channels 2 thru 6. At any other channel the filter is switched to an elliptic function high pass with its zero of transmission tuned to the low band. Thus, any potential for forming sums and differences of the input frequencies is completely removed.

The low pass filter connected to the input terminal has a cutoff of 300MHz and is used to help prevent any oscillator energy from leaking out into the system.

PARTITION FILTER

The output of the DBM is fed to the I-F via a filter mounted in a metal partition wall. This broadly tuned bandpass filter has its center tuned to 330MHz. It primary function is to provide good selectivity at frequencies far removed from the I-F.

Another feature of this filter is the use of an open circuit quarter wave transmission line as a circuit element. At 330MHz the stub looks essentially like a capacitor. At the frequency of the 2nd LO, and its third harmonic, the stub behaves like a short circuit and provides a zero of transmission to these frequencies. This is invaluable in preventing the 2nd LO from penetrating into the DBM, and causing internally generated beats.

INTERMEDIATE FREQUENCY AMPLIFIER

The I-F video frequency is 331MHz and the sound is 326.5MHz. These particular frequencies were chosen so that the sums of the R-F signal frequencies that fall in this vicinity are converted into the sound trap and the adj.-channel sound traps in the TV receiver's I-F.

The input to the I-F is tied directly to the base of a common emitter transistor amplifier. The low noise transistor operates under self bias conditions with a large voltage drop in the collector circuit. This method was used so as to foil any instabilities due to feeding the collector with a choke - a problem with high FT devices.

The selectivity prior to the 2nd mixer is achieved with double tuned capacity coupled circuits. Attempting to achieve high selectivity (1.5% bandwidths) at 330MHz with lumped LC filters results in high insertion losses. To overcome these losses two stages of gain are used.

The mixer is a common emitter transistor whose output is tuned to desired output channel. In most cases this is channel 12. The output filter is again a two pole, high side capacity coupled circuit.

The 2nd LO is a Clapp type oscillator operating at 536.25MHz (when using Channel 12 as the output). Fine tuning is achieved

INTERMEDIATE FREQUENCY AMPLIFIER (cont'd.)

by varying the base current by means of a fine tuning pot. This changes the capacity of the depletion area of the p-n junction which in turn changes the frequency of oscillation to the point that fine tuning is achieved.

The output from the mixer tank is connected to the output terminal via an attenuator pad. The value of this pad is such that the overall gain is limited to an average of 6db.

The I-F amplifiers, 2nd LO and the mixer circuitry are all mounted on one printed circuit board. Although the frequencies used would normally preclude the use of a PC board, careful placement of parts and the optimum use of the ground plane has resulted in a circuit that is stable and uniform in production.

POWER SUPPLY

The power supply is a 24v, series regulated type drawing 35ma. To prevent direct pickup the entire supply is placed in a separate compartment and completely shielded from the rest of the circuitry.

An AC convenience outlet, controlled by the ON-OFF switch, enables the converter to be used as a remote control device.

FUTURE USES

We feel that we have fulfilled the design criteria and have produced a viable product that not only fulfills todays needs but it also can serve as a fundation for other services.

For example: By changing the output channel to 44MHz we would have a tuner for an All CATV receiver.

The addition of decoder modules would permit the use of pay TV channels.

The basic unit can serve as the subscriber interface in two way systems.

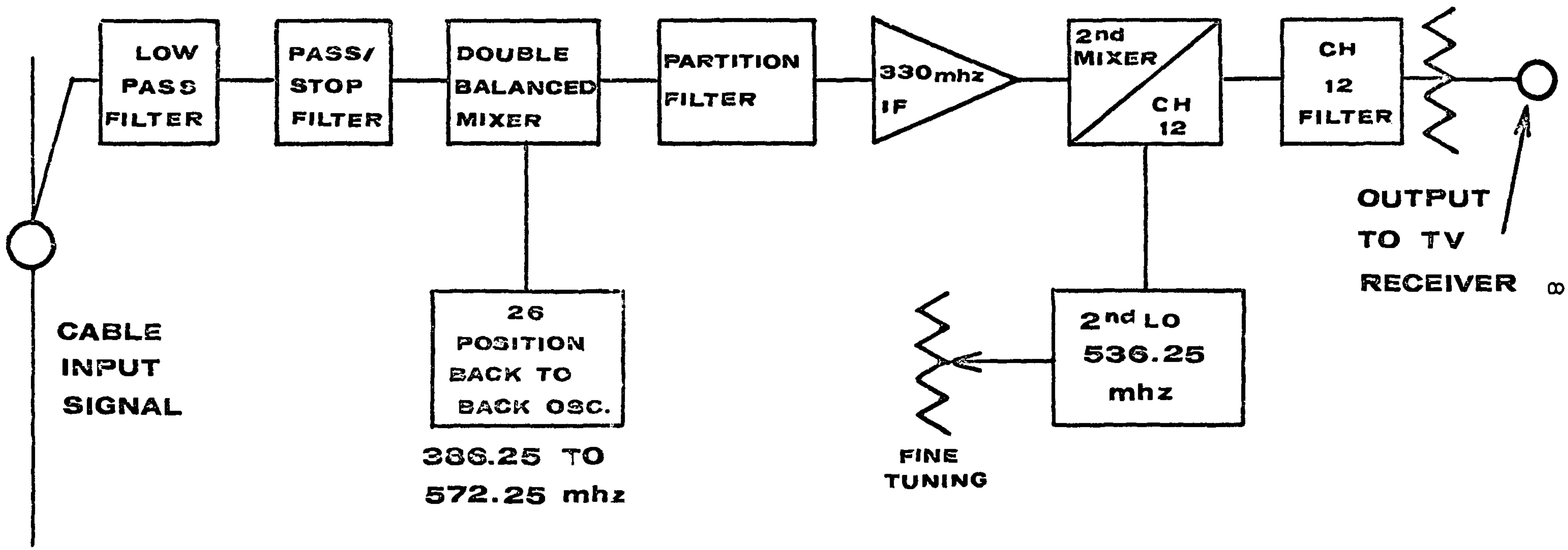

FIG.1

BLOCK DIAGRAM

GAMUT 26 CONVERTER
(CH 12 OUTPUT)

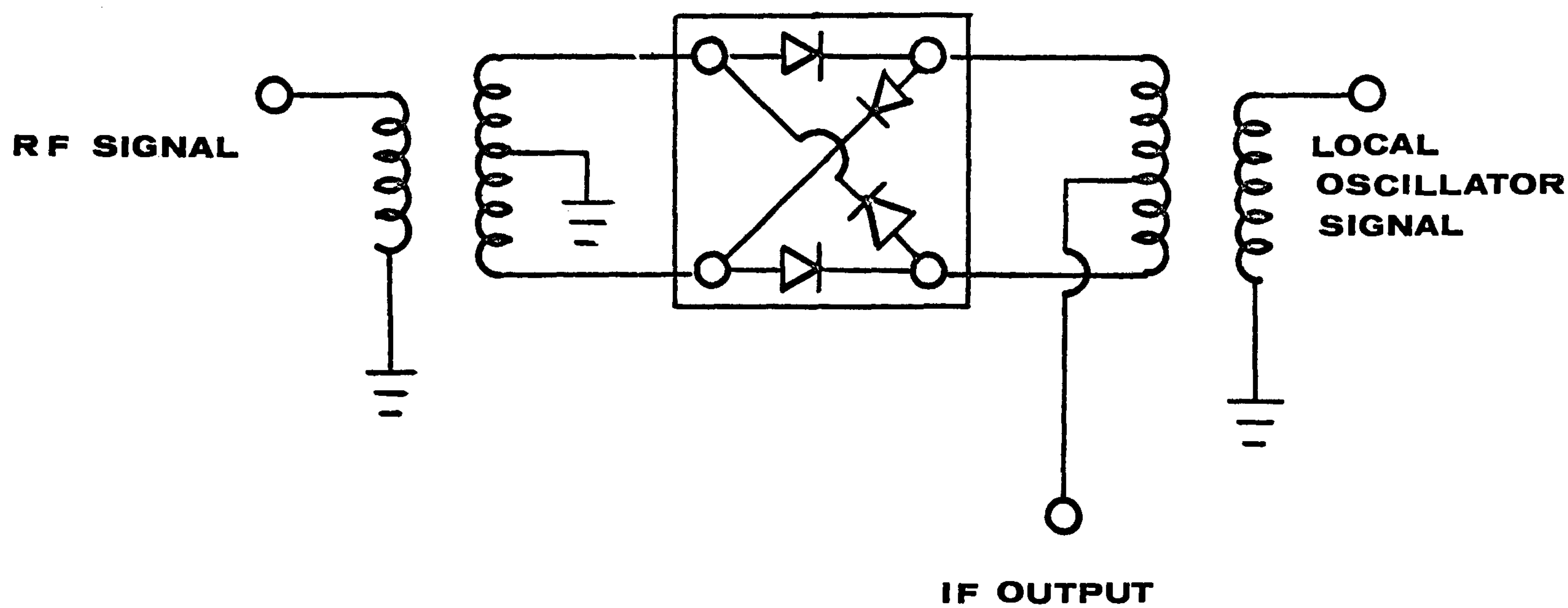

FIG. 2 DOUBLE BALANCED MIXER

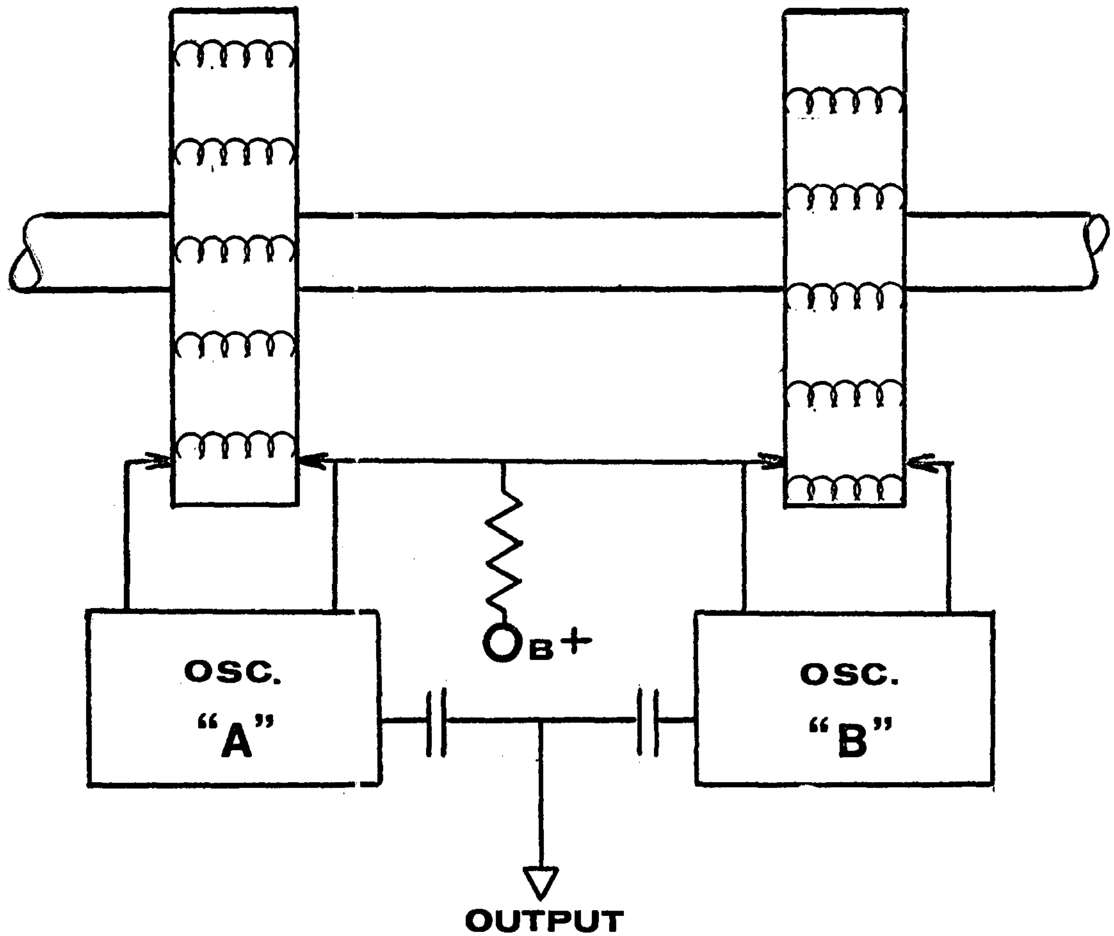

FIG. 3

BACK TO BACK OSC.

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2. "Technical Standards and Procedures for Cable Television Systems" Canadian Department of Communications, Release Date March 29, 1971.

3. Simons, "The Fundamentals of Distortion in CATV Amplifiers", Technical Handbook for CATV Systems, Jerrold Electronics Corp., 1968.

ABTO SYSTEM FOR CATV PROGRAMMING

By

Frank Marx
ABTO, Inc.
New York, N. Y.

IT IS WITH A FEELING OF HUMBLENESS THAT I SPEAK TO YOU. WHEN FIRST I WAS HONORED BY YOUR ORGANIZATION TO TELL YOU ABOUT ABTOGRAPHY MY WORDS WERE TO BE DIRECTED TOWARD A HIGHLY TECHNICAL AND ERUDITE APPROACH. PARENTHETICALLY FROM ME, PERHAPS A HIGHLY CONFUSING APPROACH AT THAT. THEN I STARTED THINKING. MY HUMBLENESS IS BECAUSE I HAVE BEEN ALLOWED AND PRIVELEDGED THROUGH THE YEARS TO SEE AND BE A PART OF THE GROWTH OF AM RADIO, FM RADIO, T.V., BROADCASTING, SATELLITE T.V. AND NOW THAT LUSTY AND VOCAL INFANT, CABLEVISION, WHO VERY WELL MAY OUTGROW THEM ALL. DURING EACH OF THESE PERIODS AND ALMOST ANNUALLY, NEW DEVELOPMENTS, FROM WHICH CAME NEW PRODUCTS, HELPED TO NOURISH THE INFANT. TO NAME A FEW. CONDENSOR AND DYNAMIC MICROPHONES TO REPLACE THE CARBON MIKE, VIDICON PICKUP TUBES TO REPLACE THE IMAGE ORTHICONS WHICH REPLACED THE ORTHICONS WHICH REPLACED THE ICONOSCOPE WHICH REPLACED THE FLYING SPOT DISC, THE REVOLUTIONARY TRANSISTORS NOW COMMONLY CALLED THE SOLID STATE INDUSTRY, TRULY BROAD BAND CABLE CARRYING MEGAHERTZ UPON MEGAHERTZ INTO AND CAPABLE OF FROM THE HOME. THIS LIST COULD GO ON AND ON. I WAS FORTUNATE TO HAVE BEEN PRESENT DURING LABOR PAINS AND PARTICIPATE IN THE DELIVERY OF AUDIO TAPE FOR BROADCASTING DURING 1944. ALSO THE 4 HEADED MONSTER WITH THE VORACIOUS APPETITE, VIDEO TAPE IN 1956. THEN LATER THE BABY BROTHER, HELICAL SCAN, PROUDLY SHOWN BY MANY EXHIBITORS AT THIS

SHOW. WHY THE NOSTALGIA AND ISN'T IT BORING? PERHAPS SO AND I APOLOGIZE. NOT REALLY THOUGH. CAUSE DURING THESE SEVERAL LIFETIMES IV'E SEEN MANY NEW TECHNICAL GLEAMS CONCEIVED, GESTATE, LABOR, BE BORN ONLY TECHNICALLY SOUND BUT WHICH DIDN'T REALLY SERVE A NEED. WHEN I FIRST LEARNED OF THE ABTOGRAPHIC PROCESS IN 1965 IT JUST COULDN'T AND INDEED WOULDN'T. I COULD SENSE THE NEED BUT COULD ALL, AND I EMPHASIZE ALL, THE TECHNICAL PROBLEMS BE SOLVED? COULD WE IN FACT USE ORDINARY BLACK AND WHITE FILM TO STORE COLOR INFORMATION? COULD THIS ORDINARY BLACK AND WHITE FILM BE USED IN AN ORDINARY CAMERA, BOTH 35mm OR 2 x 2 STILL AS WELL AS 16mm MOVIE CAMERAS. COULD THIS BLACK AND WHITE FILM, AFTER EXPOSURE, BE PROCESSED AS SIMPLY AND EASILY AS ORDINARY BLACK AND WHITE FILM? COULD IT IN FACT BE PROCESSED IN THE CAN. COULD IT THEN BE VIEWED THROUGH A STANDARD 16mm PROJECTOR AS AN EXCELLENT BLACK AND WHITE IMAGE? COULD THIS FILM BE THEN PROJECTED THROUGH THE SAME PROJECTOR WITH ONLY MINOR MODIFICATIONS AND SHOW COLOR? COULD IT REALLY OR WAS IT ONLY A DREAM. COULD ANYONE HONESTLY EXPECT TO RETAIN ALL THE ADVANTAGES OF BLACK AND WHITE FILM AND YET BE ABLE TO VIEW IT ON A T.V. SCREEN IN FULL NATURAL COLOR? WELL IT COULD <u>REALLY</u> AND IT DOES. IT'S CALLED ABTOGRAPHY. IT'S BEAUTIFULLY SIMPLE IN CONCEPT, BUT EXOTICALLY COMPLEX IN REALITY. THE UNDERLYING PRINCIPLE IS DIFFRACTION. FRAUNHOFER DIFFRACTION IF YOU MUST KNOW.

SLIDE No. 1 -

THIS REPRESENTS A SIMPLE EXPLANATION OF DIFFRACTION. THINK OF THE LIGHT SOURCE AS BEING THE SUN, A POINT SOURCE OF ILLUMINATION, A VENETIAN BLIND REPRESENTING A GRATING, THEN ON THE WALL OPPOSITE THE VENETIAN BLIND WILL BE A DIFFRACTION PATTERN WHICH REALLY IS A SERIES OF IMAGES OF THE SOURCE OF ILLUMINATION. THESE REPLICAS OR ORDERS OCCUR ONLY, AND THIS IS IMPORTANT, ONLY, AT RIGHT ANGLES TO THE SLATS OF THE VENETIAN BLIND OR THE GRATING.

SLIDE No. 2 -

THIS IS AN ARTIST'S SKETCH SHOWING THE WAY AN ABTO ENCODER RECORDS COLOR INFORMATION ON BLACK AND WHITE FILM. THE ENCODER IS IN FACT THREE DIFFRACTION GRATINGS PLACED AT DIFFERENT ANGLES, EACH GRATING REPRESENTING A PRIMARY COLOR. FOR EXAMPLE, ONE SET RED, THE SECOND GREEN AND THE THIRD BLUE. THE ORIGINAL COLOR SCENE PASSES THROUGH THE COLOR TAKING LENSE, THEN THE ENCODER, ON TO THE BLACK AND WHITE FILM. HENCE, THE ORIGINAL SCENE IS PHOTOGRAPHED IN BLACK AND WHITE WITH COLOR INFORMATION RECORDED BY THREE SETS OF DIFFRACTION GRATINGS.

PROCESSING OF THE EXPOSED BLACK AND WHITE FILM IS ACCOMPLISHED WITH NORMAL BLACK AND WHITE PROCESSING METHODS WITH CARE NORMAL ONLY TO REPRODUCING A GOOD BLACK AND WHITE PICTURE.

SLIDE No. 3-

THIS REPRESENTS A STANDARD PROJECTOR, EITHER 2 X 2 SLIDE PROJECTOR OR 16mm MOVIE PROJECTOR, MODIFIED TO PLAY BACK THROUGH A TV SYSTEM IN FULL COLOR, BLACK AND WHITE FILM. NOTE THE LIGHT SOURCE TO THE LEFT OF THE SCREEN. THIS IS A STANDARD XENON ARC LAMP USED TO OBTAIN A POINT SOURCE WITH HIGH BRIGHTNESS. THIS LIGHT PASSES THROUGH THE BLACK AND WHITE FILM WHICH, REMEMBER, REALLY CONSISTS OF THREE DIFFRACTION GRATINGS AT DIFFERENT ANGLES. WHEN LIGHT IS PASSED THROUGH THIS FILM THERE WILL APPEAR WHAT IS KNOWN AS A TRANSFORM PLANE, A DIFFRACTION PATTERN WHICH IS SHOWN IN THIS SLIDE.

SLIDE No. 4-

THIS PATTERN SHOWN ON THE SCREEN CONTAINS IN THE CENTER, ALL OF THE INFORMATION IN THE ORIGINAL PICTURE. FOR CONVENIENCE WE MAY CALL THIS THE LUMINANCE CHANNEL. PLEASE NOTE THAT THERE ARE IN ADDITION TO THIS LUMINANCE CHANNEL, THREE ADDITIONAL SETS OF INFORMATION SYMMETRICAL TO THE CENTER CHANNEL. THESE SETS CONTAIN THE INFORMATION NECESSARY TO RECONVERT THE BLACK AND WHITE IMAGE INTO FULL COLOR.

NOW BACK TO SLIDE THREE. IF AT THE TRANSFORM PLANE THERE IS PLACED AN ABTO DECODER AS SHOWN IN THIS SLIDE, A FULL COLOR IMAGE OF THE ORIGINAL SCENE WILL BE RECONSTRUCTED.

<u>SLIDE Nos. 5 and 6-</u>

NOW MAY I SHOW YOU TWO BLACK AND WHITE SLIDES IN SUCCESSION WHICH ARE PHOTOGRAPHS OF A 2 X 2 HONEYWELL PENTAX CAMERA AND A 70VR 16mm BELL AND HOWELL MOVIE CAMERA. BOTH OF THESE ARE CAPABLE OF PHOTOGRAPHING BLACK AND WHITE AND THE RESULTANT BLACK AND WHITE FILM BEING REPRODUCED IN BEAUTIFULLY NATURAL COLOR.

IF I MADE THIS EXPLANATION TOO SIMPLE I APOLOGIZE. IF IT SOUNDS COMPLEX PLEASE REMEMBER WHAT I PREVIOUSLY SAID, "THE SYSTEM IS BEAUTIFULLY SIMPLE IN CONCEPT BUT EXOTICALLY COMPLEX IN REALITY."

I THANK YOU.

AN ALL SOLID-STATE SSB-AM CARS BAND SYSTEM

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1. INTRODUCTION

Single-sideband amplitude modulation (SSB-AM) is well established as the signal processing method which assures minimum spectrum occupancy. Already four decades ago it became the generally accepted standard of multichannel telephone transmission in the form of frequency-division multiplex, first used on open-wire lines and cables, and later on microwave links. The last decade witnessed the transition to virtually exclusive use of SSB-AM on a world-wide basis in another segment of communications where frequency spectrum is at a premium, namely, short-wave radio telephony. This became feasible after an impressive arsenal of technological solutions for the rather difficult inherent problems of SSB-AM radio transmission became available. The December 1956 special issue of the Proceedings of the IRE on single-sideband techniques makes one of the most interesting readings on this subject. It might come as a surprise to some that one of the articles in that issue describes an experimental 24-channel telephone system using SSB-AM for beyond-the-horizon UHF transmission [1]. Not too long after that, in 1960, came a proposal for the use of SSB-AM on line-of-sight radio relay links [2]. This would quadruple the transmission capacity as compared to the most advanced FM radio relay telephone systems in use today. The main problem to be solved to this end is that of linear power amplification at microwave frequencies. Technological implementation, therefore, came first at lower frequencies where linear power amplifiers are available at higher output levels. SSB-AM 120-channel telephone systems operating in the 400-410 and 420-430 MHz bands were installed in the mid 60's to establish a high-quality commercial telephone link between West Berlin and the Federal Republic of Germany [3].

The first proposal for the use of SSB-AM for TV transmission at microwave frequencies was made in 1959 [4]. It envisaged TV broadcasting in the 12 GHz band, as a means of substantially increasing the number of TV programs in the Federal Republic of Germany*.

*In view of the population distribution the VHF and UHF channels offer satisfactory coverage for only three simultaneously broadcasted TV programs in that country and, as a matter of fact, in most of Europe.

Systematic studies and experimental investigations of this problem area have been carried out [5] in the course of which the feasibility of an SSB-AM microwave transmitter has been established [6].

The use of SSB-AM for microwave transmission of TV channel groups intended for CATV distribution started on an experimental basis in 1966 [7]. Frequencies in the 18 GHz range were used and transmission was on a group basis; i.e., the TV channels to be transmitted were multiplexed at VHF frequencies and transmitted with a single, broad-band microwave transmitter (refer to Fig. 2). Experience gained from these experiments and the subsequent development of a new system version for the 12.7-12.95 GHz CARS band [8] were instrumental in formulating the FCC rules for SSB-AM transmission in this band [9]. Thirty six regular and two "auxiliary" channels are assigned. This is the maximum usable capacity of links without intermediate repeaters. If one or more intermediate repeaters are needed, the transmission capacity is reduced to a total of nineteen channels.

The system design to be described in this paper fully conforms with the aforementioned FCC rules [9] and is based on the microwave solid-state technology developed at Fairchild for use in communication systems. As will be shown, the selected system configuration enhances transmission performance, flexibility of use, and reliability.

This paper is limited to the description of the overall system design. Technical data are specified in such a way as to facilitate their use for performance calculations of planned links using generally established procedures. Propagation aspects are treated only to the extent not covered in the existing literature on CARS band systems.

The feasibility of the described SSB-AM CARS band system design was experimentally verified in the Spring of 1970. A system model was since repeatedly demonstrated in the laboratory, simulating link lengths of up to 10 miles. FCC type acceptance tests are in preparation and the first field tests are to be carried out in 1971.

2. CHOICE OF MICROWAVE SIGNAL PROCESSING SCHEME AND SYSTEM CONFIGURATION

The signal processing scheme is selected for the advantages it offers. The reasons for preferring SSB-AM were as follows:

- This is the signal processing scheme with the minimum bandwidth requirement on a per channel basis. The occupied spectrum width in micro-wave CARS band transmission is the same as in VHF cable transmission. The 250 MHz wide CARS band can thus accommodate the entire 50-300 MHz bandwidth envisaged for future cable systems (refer to Fig. 1).

- When planning a radio system which is to be used with an existing cable system, it is advantageous to use the same channel multiplexing scheme in both cases. The SSB-AM CARS band system to be described takes the VHF channels directly from the frequency range required for cable transmission and puts them directly back into the same VHF frequency range (Fig. 1).

- It is always advantageous to use the simplest possible signal processing scheme consistent with the required transmission performance. The straightforward up-conversion/down-conversion scheme of SSB-AM (Fig. 1) is undoubtedly the simplest available.

Accordingly, SSB-AM was selected because it assures, on one hand, maximum spectrum usage and, on the other hand, the simplest possible transition from cable to radio transmission and vice versa.

The main difficulty in implementing analog AM systems, in general, lies in the non-linear transfer characteristic of the transmitter or amplifier output stage [10]. As a consequence, the tolerable amount of non-linear distortion determines the maximum usable power output which is in most cases one or two orders of magnitude below the saturated output power. This is a serious disadvantage of analog AM as compared to all other signal processing schemes which operate at saturated output power levels.

Figure 2 illustrates the most logical first approach to an SSB-AM CARS band link which consists of using a single, broadband transmitter-receiver pair for the simultaneous transmission of several VHF channels. The specific example used in Fig. 2 and for most of the following system considerations is that of a 6-channel head-end link with alternate channel transmission. The case of contiguous 12-channel transmission is treated later. However, all of the following considerations of multichannel transmissions apply irrespective of channel arrangement or transmission capacity.

Since the output power of the transmitter cannot be arbitrarily increased due to technological and economical constraints, the multichannel transmission performance must be considered for a predetermined available transmitter output level. It becomes immediately clear that under these conditions both the signal-to-noise ratio, S/N, and the relative cross-modulation level, -XM, deteriorate as the number of transmitted VHF channels, n, increases.

The deterioration of the S/N ratio with increasing transmission capacity is due to the fact that in a broadband group transmitter the available power output is shared by the simultaneously transmitted signals such that the available power per channel, P_{ch}, becomes

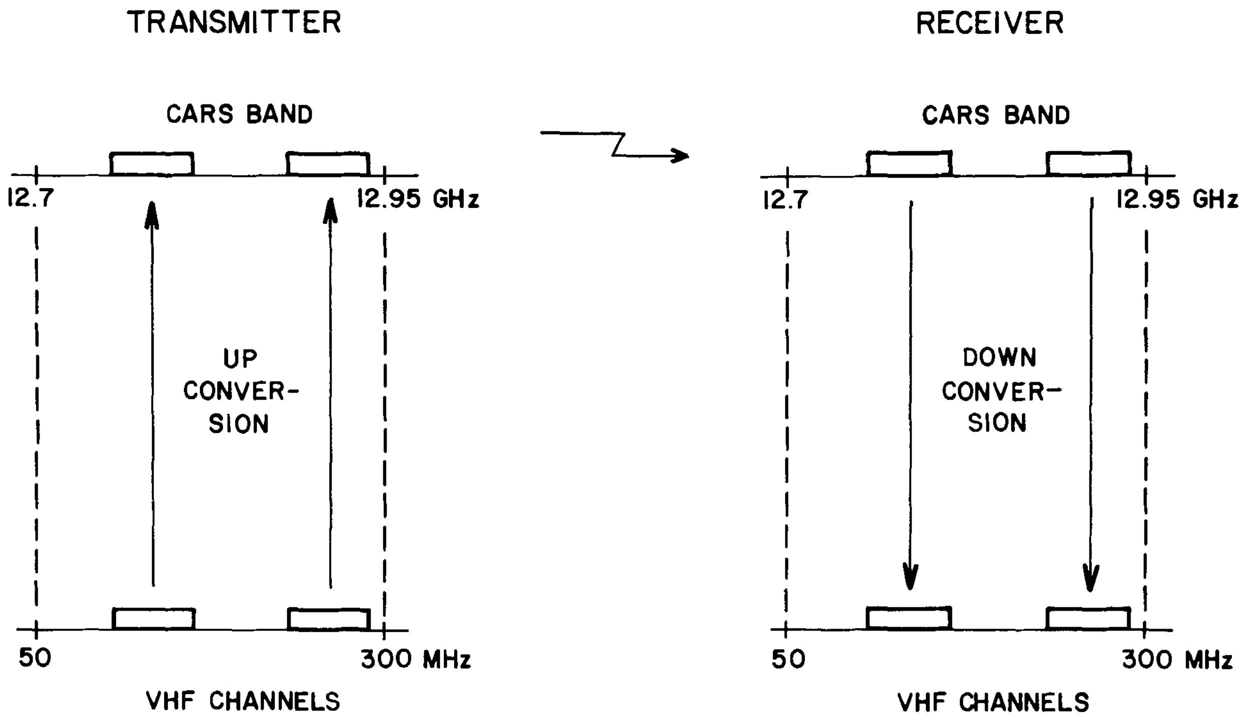

Fig. 1. SIGNAL PROCESSING SCHEME OF THE CARS BAND SSB-AM SYSTEM.

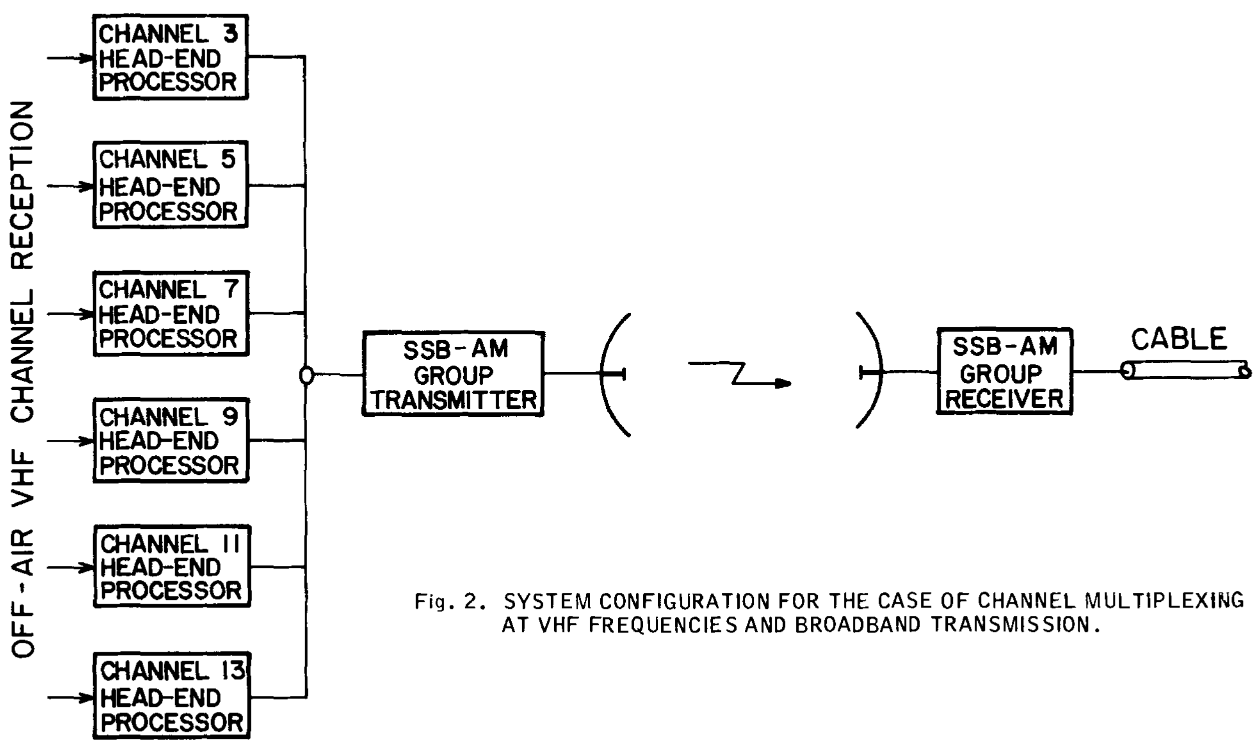

Fig. 2. SYSTEM CONFIGURATION FOR THE CASE OF CHANNEL MULTIPLEXING AT VHF FREQUENCIES AND BROADBAND TRANSMISSION.

$$P_{ch} = \frac{P_{tot}}{n}$$

where P_{tot} is the total available transmitter power and n the number of transmitted channels. The S/N ratio of each channel in a group of n channels, $(S/N)_n$, expressed in dB, thus equals

$$(S/N)_n = (S/N)_{n=1} - 10 \log n$$

where $(S/N)_{n=1}$ is the S/N ratio for single-channel transmission and n the number of simultaneously transmitted channels. As illustrative, numerical examples consider a 5-channel system and a 12-channel system. The S/N ratio in each VHF channel of the former system will be $10 \log 5 = 7$ dB lower than for a single channel system. The corresponding difference for the 12-channel system is $10 \log 12 = 10.8$ dB

The relative cross-modulation level, -XM, expressed in dB, increases with the increasing transmission capacity by the following amount [10]:

$$20 \log (n-1)$$

where n is again the number of transmitted channels. This means that the cross-modulation level in each channel of a 5-channel system will be $20 \log 4 = 12$ dB higher than in a 2-channel system; the corresponding difference for a 12-channel system amounting to $20 \log 11 = 20.8$ dB . Comparison with a single-channel system is meaningless since, by definition, there is no cross-modulation in that case. However, the 3rd order intermodulation specifications for single-channel transmission are much easier to satisfy than the cross-modulation specifications for even a two-channel system.

The above considerations lead to the conclusion that from the viewpoint of transmission performance it is preferable to use a separate microwave transmitter for each channel and multiplex them into a common antenna as illustrated in Fig. 3. The implementation of such a transmitter solution is described in the next section.

On the receiving end there is no need to use a separate microwave receiver for each channel because the group receiver approach, illustrated in Fig. 2, is satisfactory from the viewpoint of both S/N ratio and cross-modulation. The latter is, of course, more critical but the receiver operates at levels which are substantially lower with respect to saturation than is the case in the transmitter. A study based on considerations reported at the 1970 NCTA Convention showed that acceptable cross-modulation levels can be obtained with existing microwave mixer technology [11].

3. SYSTEM BLOCK DIAGRAM

3.1 Transmitter

Figure 4 shows the block diagram of the transmitter. The simplest possible solution has been adopted. There is only one active microwave circuit in the signal path, namely the upper-sideband parametric up-converter which directly delivers the required output power. Such a solution is undoubtedly the most advantageous one from the viewpoint of distortion because it minimizes the number of sources thereof.

The additional important advantage of the adopted transmitter configuration is that it lends itself best to an all solid-state implementation. The source of microwave power, the up-converter pump, is a CW oscillator with output power in the 1.5 - 2.0 W range. It consists of two cascaded silicon avalanche diode (IMPATT) oscillators with 1 W nominal power output each and a DC-to-RF conversion efficiency of better than 5% which is higher than the efficiency of commercially available klystrons in the same frequency range*. The avalanche diodes and the power combining scheme were developed at Fairchild. The power combining efficiency of the composite oscillator is virtually 100%. For a description of these techniques and more details on the obtained results it is referred to publications at the 1969 International Solid-State Circuits Conference [12] and at the 1971 International Microwave Symposium [13].

The advantage of the described transmitter over any other configuration can be fully appreciated only after the available alternatives have been considered in some detail. They all involve power amplification. Unfortunately, non-linear distortion in available microwave power amplifiers for CARS band frequencies, tube or solid state, is by no means lower than in a properly designed high-level up-converter. While it is true that there are several promising approaches for linearizing the tube and transistor amplifier transfer characteristics in the VHF to lower microwave frequency ranges, there apparently is as yet no practical solution available at CARS-band frequencies. This subject matter is, therefore, not treated here and references to publications are omitted.

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*The second generation up-converter pumps will use GaAs avalanche diodes whose efficiency is around 10%.

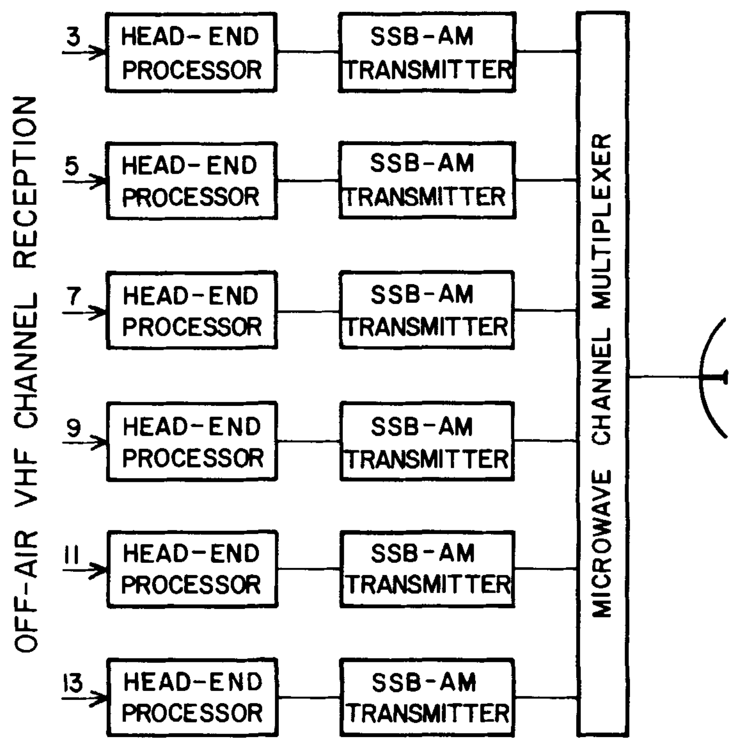


Fig. 3. TRANSMITTER CONFIGURATION WITH CHANNEL MULTIPLEXING AT CARS-BAND FREQUENCIES

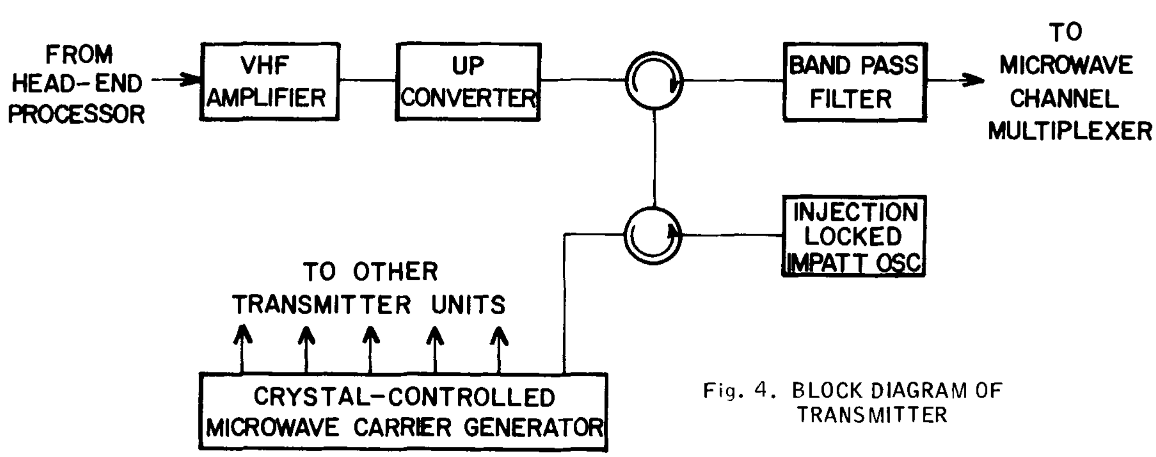


Fig. 4. BLOCK DIAGRAM OF TRANSMITTER

As can be seen in Fig. 4, a crystal-controlled microwave generator is used as the common carrier source for the entire multichannel transmitter group. It consists of a low-power crystal controlled multiplier source, a standard Fairchild product for use in microwave communication systems, followed by an injection locked avalanche diode oscillator which functions as power amplifier. Its output power is split to in turn injection lock all the up-converter pumps of the transmitter group. The purpose of this arrangement is to keep all the up-converter pumps in synchronism, which is indispensable because, as pointed out before, a group microwave receiver is used in the described system.

3.2 Receiver

The block diagram of the group receiver is shown in Fig. 5. It uses a conventional RF head consisting of a band-pass filter, a mixer, and a crystal-controlled multiplier source as local oscillator. Means for establishing synchronism between the latter and the microwave carrier generator of the transmitter, which is needed in the case of locally broadcasted channel transmission, can be added. A pilot signal derived from the microwave carrier generator is then inserted at the transmitting end in the frequency space the FCC reserved for this purpose [9]. At the receiving end, this pilot signal is filtered out and used to control the frequency of the local oscillator.

The receiver of Fig. 5 employs a separate AGC VHF amplifier in each channel. While the concept of a group receiver can be implemented all the way down to the VHF output, automatic gain control on a group basis is likely to prove impractical for many applications. This statement does not refer to problems of AGC circuit design but to multipath propagation effects in the form of frequency selective fading which would manifest itself as an irregular, frequency-dependent and time-varying amplitude distortion over the whole receiver bandwidth or portions thereof. These amplitude variations can amount to several dB. In general, the wider the occupied transmission bandwidth, the more pronounced this effect becomes. The bandwidth of a 12-channel SSB-AM CARS band system transmitting the standard VHF channels, for example, equals 162 MHz or, on a percentage basis, 1.25% approximately. This is by no means negligible as compared to frequency separations of frequency diversity systems operating in the microwave region which exploit the frequency dependency of multipath fading to reduce outage time due to such fading*.

*A discussion of multipath fading, published in the Lenkurt Demodulator [14] includes the statement that most frequency diversity systems have frequency separations of 2-5% of the lower frequency.

The problem of frequency selective fading at frequencies above 10 GHz has received very little attention, most likely due to the fact that the most severe fadings, which limit the usable path length at these frequencies, are caused by heavy rainfall. However, there is ample evidence [15,16] that the multipath fading problem in the CARS band is a real one whenever systems of considerable bandwidth are used. Systematic propagation studies would be necessary in order to determine a "safe" upper limit for the usable system bandwidth. An estimate based on the above quoted publications [15,16] leads to the conclusion that the danger of frequency selective fading must not be disregarded for SSB-AM systems occupying a substantial portion of the CARS band frequency spectrum. In practical terms, AGC on a group basis might be sufficiently safe only for small groups of contiguous channels, such as the five-channel group of VHF channels 2-6 (bandwidth: 34 MHz) or even the seven-channel group of VHF channels 7-13 (bandwidth: 42 MHz).

The above discussion, although limited to the SSB-AM system, should not be misinterpreted as being pertinent to this system alone. Severe multipath fading will significantly affect any wideband analog transmission system. The degree of this effect and the most suitable solution of the resulting problems will depend on the particular system. For the SSB-AM system under discussion, it is believed that the receiver configuration of Fig. 5 represents a technically and economically sound solution based on readily available technology. Of course, if signal processing is needed at the receiving site, as well, this can be easily accomplished by using head-end processors instead of AGC amplifiers.

3.3 Contiguous Channel Transmission

So far, the system configuration has been illustrated only in terms of alternate channel transmission (Figs. 2-5). To transmit a group of contiguous signals requires simply connecting two corresponding alternate-channel transmitter groups to a common antenna with two polarizations. The same applies to the receiving end. This solution is illustrated in Fig. 6 for a 12-channel capacity, but the principle applies also to other transmission capacities.

The adopted solution is attractive not only from the viewpoint of microwave channel multiplexing, which would be prohibitively complicated and expensive if a single polarization were used for the entire group of contiguous channels, but also from the viewpoint of suppressing adjacent channel interference due to third order intermodulation products. Figure 7 is referred to for an explanation. It shows the carrier frequencies of channel 10 and their third order intermodulation products as they appear in the CARS band. As can

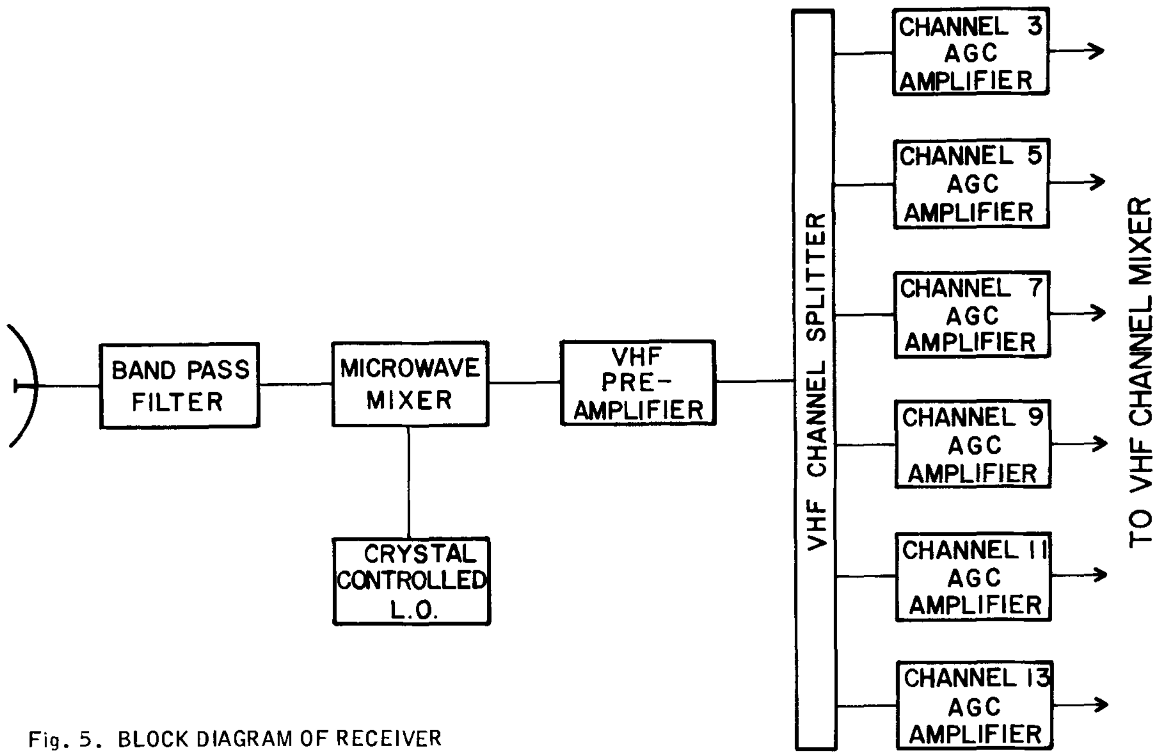
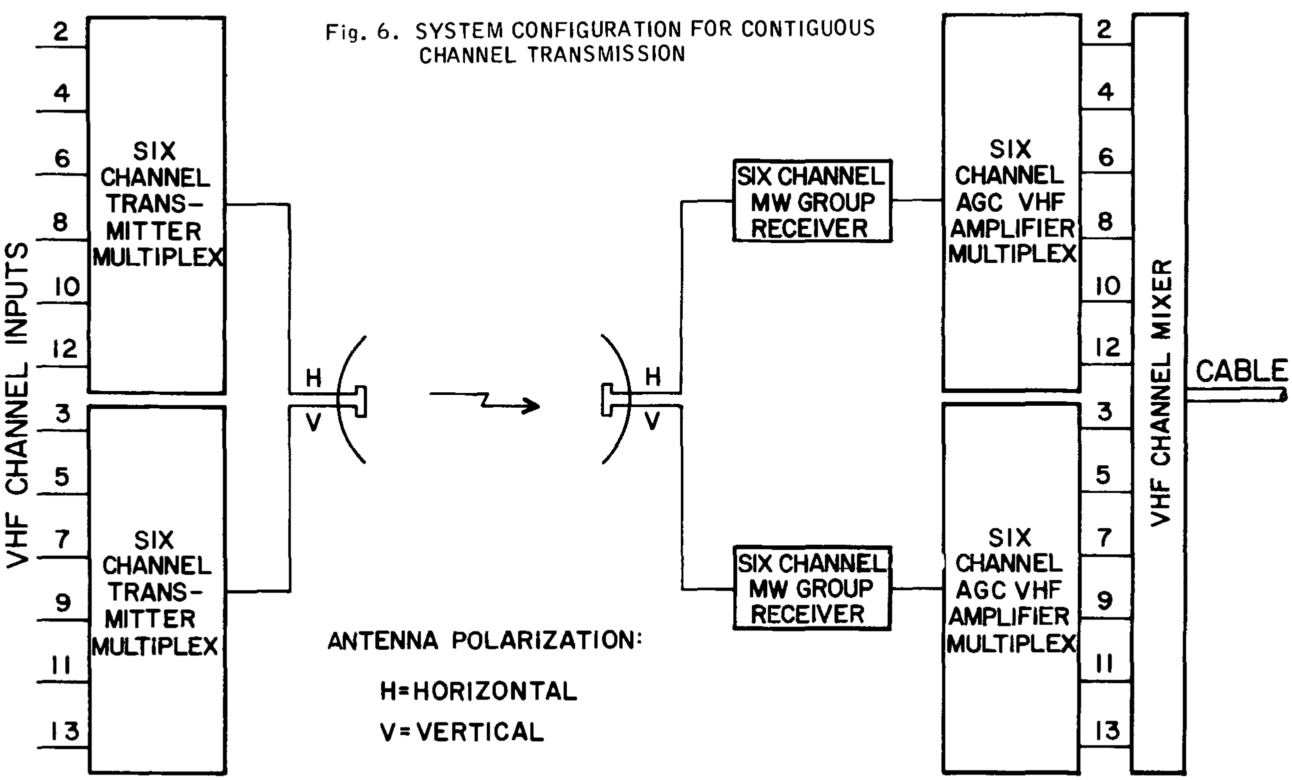


Fig. 5. BLOCK DIAGRAM OF RECEIVER

Fig. 6. SYSTEM CONFIGURATION FOR CONTIGUOUS CHANNEL TRANSMISSION



be seen, all the unwanted up-conversion products fall either within the transmitted channel itself or in the two adjacent channels. The FCC requires that the out of band products be attenuated by at least 50 dB below the peak power of emission [9]. It thus becomes clear that the isolation between the orthogonal polarizations of commercially available parabolic antennas for CARS band applications, whose order of magnitude is 20 dB, can significantly contribute to the suppression of adjacent channel interference when the transmitter multiplexing scheme of Fig. 6 is used.

The configuration of the receiving end is straightforward. Separate RF heads are used for each polarization and the VHF channels are multiplexed into the cable using conventional CATV techniques.

4. TRANSMISSION PERFORMANCE

4.1 Objectives

The described SSB-AM CARS-band system is conceived as a wireless substitute for cable trunklines. The transmission objectives depend, therefore, on the length of the cable trunkline the system is supposed to replace. The thermal noise and non-linear distortion allowance will be proportional to the length of the link in accordance with established cable system planning procedures [10].

The approach taken at Fairchild with regard to transmission performance objectives is twofold:

- Use the most effective system and component design for high transmission performance, consistent with reliable operation and competitive pricing.

- Closely cooperate with potential users of the SSB-AM CARS-band system in order to determine realistic transmission performance objectives for some immediate applications.

This approach turns out to be most beneficial to all the parties involved, which is understandable since a new class of systems is concerned.

It has been determined in this way that one of the most immediate needs is for multichannel radio transmission to replace the trunkline cable between the head-end and the distribution center. The described SSB-AM CARS-band system has been found to satisfy the requirements of such an application with a comfortable margin for link lengths up to 10 miles approximately, except in locations with extremely heavy rainfall which are limited to a few regions of the United States.

4.2 Signal-to-Noise Ratio

The following data are given for a 12-channel link in such a form as to facilitate their use in quick estimates of obtainable S/N values for link lengths of practical interest.

Transmitter output power at antenna terminal, per channel*	10 mW
Receiver noise figure	10 dB
Filter and isolator losses in receiver	2 dB
Feeder looses (transmitter and receiver)	2 dB

A 7 mile 12-channel link equipped with 10 ft. parabolic antennas will be assumed as illustrative example. The net transmission loss between the transmitter and receiver antenna terminals would amount to 45 dB for the case of ideal propagation conditions; i.e., no fading. A 4 MHz receiver bandwidth is used to calculate the ideal noise level, -108 dBm. Performing the straightforward arithmetic operation with these data gives

$$S/N = 65 \text{ dB}$$

which leaves a fading margin of approximately 20 dB for excellent reception (TASO Grade 1). The exact amount of the fading margin depends on whose system design criteria are used.

The quoted S/N performance has been verified in the laboratory using a transmitter-receiver pair whose electrical characteristics conform with the above data. Small horn antennas were used and up to 40 dB of fading was simulated with a variable attenuator.

4.3 Intermodulation

As pointed out before, intermodulation in the single-channel up-converter is the dominant form of 3rd order product interference in the SSB-AM system under consideration. Advice obtained from the Jerrold Electronics Corporation on how to carry out laboratory

*Higher output powers per channel are available for most transmitter configurations with less than 12-channels because of simpler microwave channel multiplexing.

tests of this particular transmission characteristic was invaluable in the absence of a standard test procedure. Three CW signals are used to simulate the color TV signal and sound. The amplitudes of the signals simulating the sound carrier and color subcarrier are -10 dB and -16 dB relative to the signal simulating the vision carrier. The most troublesome in-band 3rd order intermodulation product, V+S-C (refer to Fig. 7), must not be higher than -50 dB relative to the amplitude of the CW signal simulating the vision carrier.

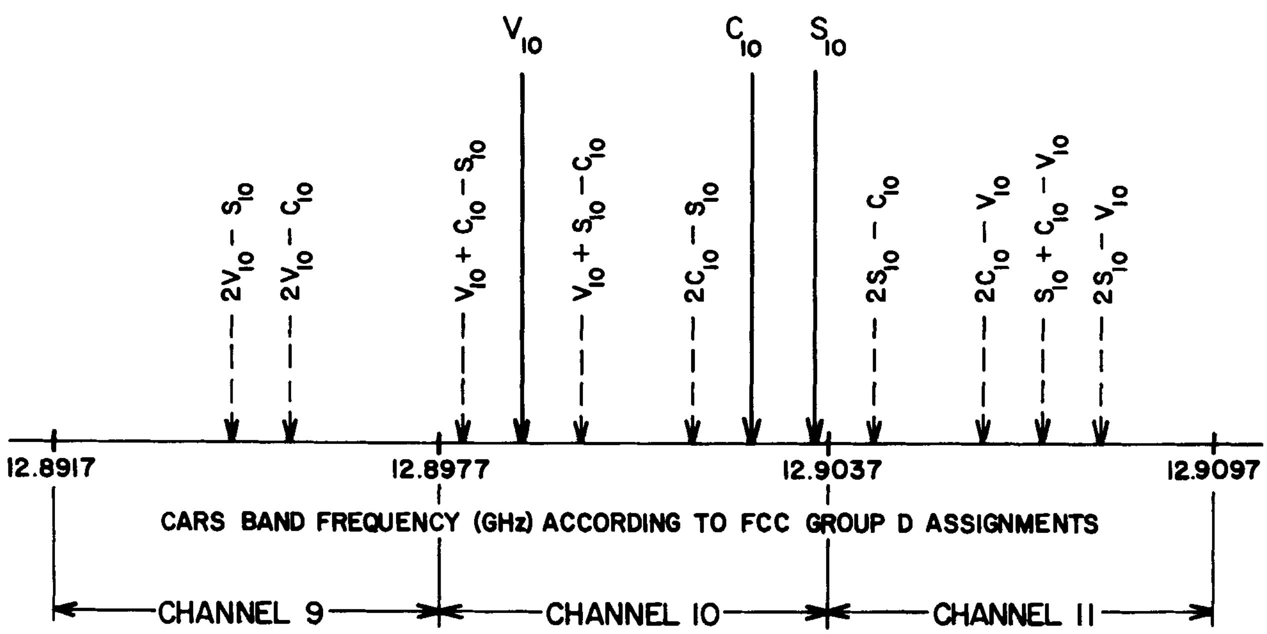
Figure 8 shows the result of this test for channel 10 up-converted into CARS band in accordance with the FCC Group D frequency assignments [9]. As can be seen, all the in-channel 3rd order intermodulation product amplitudes are more than 50 dB below the amplitude of the CW signal simulating the vision carrier. It should be mentioned, at this point, that the suppression of the out-of-band intermodulation products (refer to Fig. 7) which are not seen in Fig. 8 was satisfactory, as well.

V_{10} = CHANNEL 10 VISION CARRIER

C_{10} = CHANNEL 10 COLOR SUBCARRIER

S_{10} = CHANNEL 10 SOUND CARRIER

Fig. 7. THIRD-ORDER INTERMODULATION PRODUCT FREQUENCIES FOR UP-CONVERTED CHANNEL 10



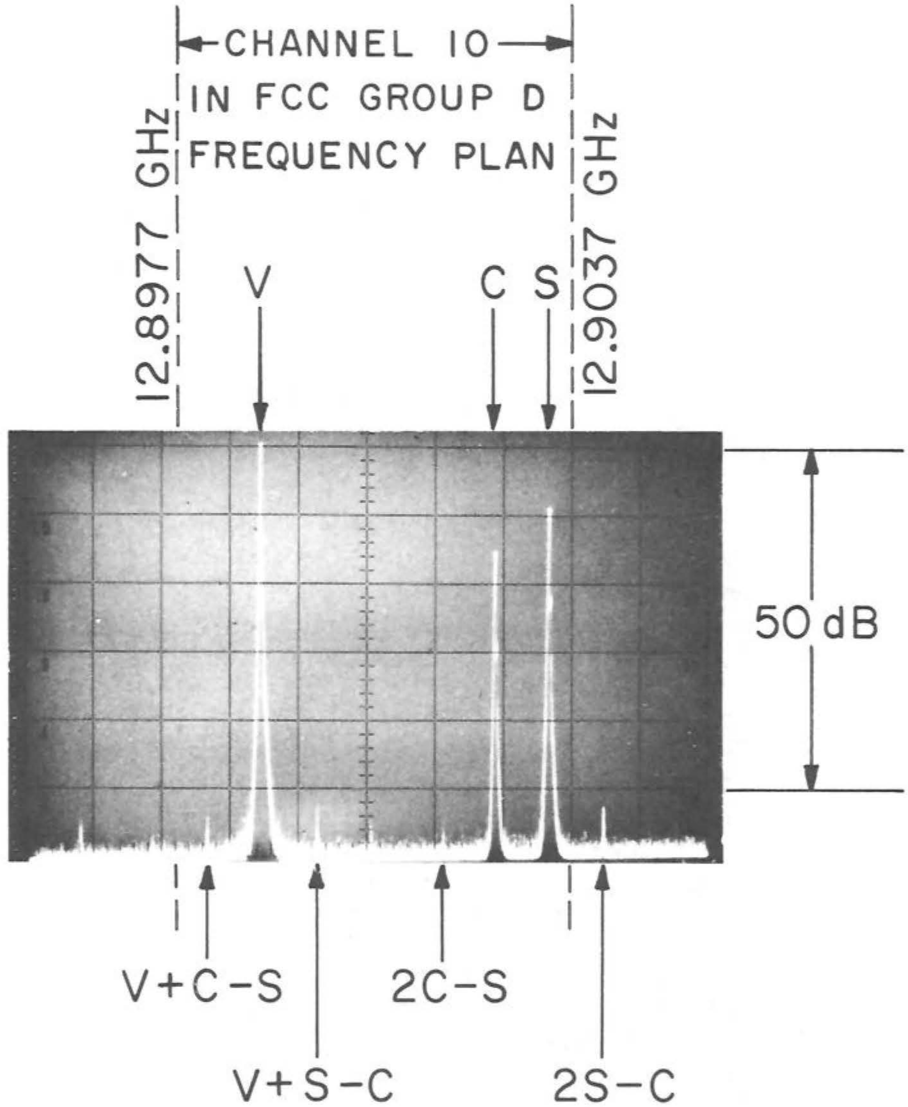


Fig. 8. MEASURED IN-BAND THIRD-ORDER INTERMODULATION PRODUCT AMPLITUDES OF CHANNEL 10 CARS-BAND TRANSMITTER

Of-the-air color TV signals were also used for subjective tests of the transmission performance. No difference could be observed between the quality of the of-the-air signal and the signal passed through the above described experimental link.

5. CONCLUSIONS AND COMMENTS

The main advantages of the described SSB-AM system are in the following design features:

- Channel multiplexing at microwave frequencies,
- simplest possible transmitter configuration, and
- all solid-state design.

These reflect on the operational characteristics of the system in the following way:

- Transmission quality is enhanced by using a separate up-converter for each channel and by avoiding subsequent microwave signal amplification.
- All three aforementioned design features enhance reliability.
- The modular transmitter design offers attractive flexibility and economy in system build-up whenever initial needs are below full transmission capacity. Plug-in transmitter modules can be easily added as needs grow.
- All solid-state system implementation became feasible only through adoption of the modular transmitter design without output power amplification. Avoiding the use of microwave tubes results in substantially lower power supply voltages (one order of magnitude) and in higher DC- to -RF conversion efficiencies which translate into lower power consumption.

ACKNOWLEDGEMENT

The authors deeply appreciate the interest in their development effort demonstrated by fellow engineers from the CATV industry whose comments and advice were of invaluable assistance.

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AN IMPROVED FREQUENCY MEASURING TECHNIQUE FOR CATV

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NEED FOR FREQUENCY MEASUREMENTS

Most specifications currently in force or proposed for CATV make some reference to the frequency accuracy and stability of the TV carriers in cable television systems. The specifications usually require carrier frequency accuracies approaching those applied to television broadcast transmitters, particularly in the case of adjacent channel operations.

Television channels processed through doub conversion, heterodyne processors using the same local oscillator for down and up conversion are not changed in frequency. Neither the visual carrier frequency nor the visual-aural intercarrier spacing is changed by such processing. Any frequency errors observed in such a case must have occurred in the broadcasting station.

Television channels converted from one channel to another, either by single or multiple conversion processes or by demodulation and remodulation are subject to possible frequency errors arising the processing system. Intercarrier spacing is usually not affected unless audio is reduced to baseband and remodulated for cable system use.

A convenient and accurate method of measuring carrier frequencies in cable systems has been developed and is described:-

ACCURACY REQUIRED

Broadcast standards for frequency accuracy are +- 1 KHz for both visual carrier and intercarrier. The universal use of intercarrier sound systems in television receivers makes the intercarrier spacing a more important parameter than the aural carrier frequency itself. It is desirable to have an accuracy of at least +- 100 Hz in a measuring system to check carriers specified to +- 1 KHz. This desired 100 Hz tolerance in measurement implies an accuracy of 5 parts in 10^7 or .00005% at 200 MHz. Such accuracies can only be conveniently obtained with digital frequency counter techniques.

VISUAL CARRIER MEASUREMENTS

Digital frequency counters operate by counting the number of cycles of carrier which pass through a gate in a precisely controlled period of time. The gate time is controlled by a counter derived from a "clock" within the instrument. This clock is a precision oscillator whose stablility and accuracy set the precision attainable with the instrument. Very low cost instruments use the power line frequency as a "clock". Better class instruments use temperature compensated or oven-housed crystals as the "clock". The "clock" can be periodically calibrated by comparison with more precise frequency standards. Most digital frequency counters have an upper counting limit of from 10 to 50 MHz. This operating range can be extended by the use of digital pre-scalers or heterodyne converters. Pre-scalers divide the incoming frequency, usually by multiples of ten, reducing the frequency to a range which the counter will handle. Heterodyne converters "translate" the incoming frequency into a range which the counter will handle using local oscillators derived from the counter's own clock.

continued...

Pre-scalers are currently available to handle inputs up to 500 MHz. Heterodyne techniques may be used up to microwave frequencies (tens of GHz). A frequency counter with 1 in 10^7 accuracy and with a 500 MHz range (usually achieved by built-in pre-scaler) is very suitable for CATV use.

Waveform input requirements for frequency counter instruments are quite critical. Higher frequency instruments usually have 50 ohm inputs and require about 100 millivolts of signal. The input signal must be relatively free of non-harmonically related spurious components. Since the counter operates by individually sensing and counting each cycle of carrier there must be enough of each cycle present to trigger the counter circuitry. Trigger sensitivity varies considerably from one model of instrument to another but very few counters will trigger on the highly modulated (80 - 90%) AM television visual carriers. Even if the desired carrier is separated from all the unwanted carriers present (associated aural carrier and other TV channels in the system) the high degree of amplitude modulation prevents proper operation of most digital frequency counters. This paper describes a method of separating the desired visual carrier from the unwanted carriers and reducing the amplitude modulation to a level which most counters will accept.

The desired visual carriers could be separated for counting by using a suitable band pass filter or by using a tunable band pass filter. We find it convenient to use a tunable television demodulator for this purpose, using the selectivity of the demodulator IF section to reject all unwanted carriers. Unwanted amplitude modulation is then removed by a limiter stage. The "limited" IF carrier is then translated back to the original input frequency using a mixer and the local oscillator from the tuner. This is merely double heterodyne conversion using the same local oscillator for both down and up conversions with a limiter stage inserted in the visual carrier IF. The output frequency is the same as the input but unwanted carriers have been removed and the amplitude modulation has been reduced to a level which most frequency counters will accept.

DETAILED DESCRIPTION

A practical prototype of a frequency measurement system of this type has been in use on Maclean-Hunter systems for some time. Figure 1 is a block diagram of the instrument. A Jerrold Model TD demodulator is used as the tuner and IF system. Other demodulators or television receivers can be adapted for this purpose. The TD demodulator has the local oscillator available at a suitable level from a connector right on the tuner. Sufficient local oscillator signal can usually be derived from the tuner by coupling through a small capacitor. The loading caused by the up-converter mixer causes the local oscillator to shift by only 15 KHz. Since the accuracy and stability of the local oscillator does not affect the output frequency, this is not important. The limiter stage which has been added is fed from the IF test point on the demodulator. It is a Motorola MC1330P integrated circuit, which is designed for use as a low level video detector in television receivers. It also has a limiter section designed to feed AFT circuitry in TV receivers. This output level is adequate for feeding the double balanced mixer which serves as an up-converter. A co-axial switch and two band pass filters are provided to reject the unwanted mixer images. The amplifier board from a small MATV type amplifier was used to amplify the output to about 500 millivolts for driving the digital frequency counter.

continued...

Effectiveness of the limiter was tested by modulating the TV carrier with a 15.75 KHz square wave. Figure 2 shows the modulation envelope of a TV carrier and its associated frequency spectrum. Modulation was set to about 80%. The first modulation side bands are about 7 db below carrier. Figure 3 shows the same carrier after limiting. The modulation envelope shows very little modulation. The associated frequency spectrum shows that modulation sidebands have been reduced by 17 db. The frequency counter accepted the limited carrier without any triggering problems. The 80% modulated carrier would not trigger our frequency counters and pre-scalers properly.

IMPROVEMENTS PLANNED

Some minor modifications are planned. A small "split-band" amplifier will take the place of the switch and band-pass filters. The present "bread-board" arrangement will be miniaturized and built into one of our TD demodulators.

We plan to experiment with a phase-lock loop in the IF which would replace the limiter stage. This would assure that we have a continuous signal available at all times without concern for limiter threshold levels. The present system will not count an overmodulated carrier properly. This may be considered advantageous because we have often detected overmodulated carriers in this way. It is inconvenient if one nevertheless wishes to make a frequency measurement on the overmodulated carrier.

COUNTER ACCURACY

Measurements made in this way are no more accurate than the clock in the frequency counter. We use a frequency standard system which is locked to Loran C transmissions. The Loran C receiver in our laboratory produces a 1 MHz signal which is phase-locked to the pulsed 100 KHz transmissions from the North Atlantic Loran C chain. The Loran C chain is controlled by a caesium beam atomic frequency standard. The Loran C transmission effectively transfers this standard to our laboratory with very little loss in accuracy. This Loran C clock is used to drive the counters in our laboratory and to calibrate counters before use in the field. Similar systems are available which lock to NBS 60 KHz transmissions.

INTERCARRIER MEASUREMENTS

Intercarrier measurements are relatively easy to make, using the same demodulator employed for the visual carrier measurements. Many demodulators have an intercarrier output available at a level suitable for direct counting by a digital frequency counter. Such an output may be easily derived from receivers or demodulators which do not originally have a 4.5 MHz intercarrier output available. The only problem encountered is the frequency modulated nature of this intercarrier sound signal. If the aural modulation can be stopped for the period of the measurement, there is no problem and the intercarrier frequency can be read by normal digital counter technique. If the carrier is modulated (FM), we can only expect to read the average carrier frequency since the counter will actually count the number of aural carrier cycles (actually intercarrier) during a fixed gate time. If the modulation is symetrical around the normal carrier frequency, and if we use a sufficiently long gate time, we will get an accurate measurement of intercarrier frequency. We have found that a 10 second gate time gives good results. With some counters this requires a pre-scaling of the intercarrier to a frequency range in which the counter uses a 10 second gating time.

continued...

The required gating time may be estimated by considering the normal frequency deviation used in TV aural carriers and the lowest modulating frequency likely to be used. Maximum deviation is ordinarily 25 KHz and the lowest modulating frequency is likely to be about 100 Hz. If we considered a "worst case" of square wave modulation by a maximum level 100 Hz square wave, the carrier could be at the deviation extremes for one half of each square wave (.005 seconds). The worst case in gating timing would be a situation in which the gate passed one more swing to one direction than it did to the other. With a square wave modulation to maximum deviation, this would pass about $4.5 \times 10^6 \times 5 \times 10^{-3} = 2.25 \times 10^4$ more (or fewer) cycles than the centre frequency. This 2.25×10^4 is a worst case error and is independent of the gating time. For this error to be less than 100 Hz in a 4.5 MHz measurement, it must amount to less than about 2 cycles in every 10^4 cycles counted. The total number of cycles counted should, therefore, be more than $2.25 \times 10^4 \times 2 \times 10^4 = 4.5 \times 10^8$ cycles. A 100 second gating time would pass this many cycles of a nominal 4.5 MHz carrier and guarantee the desired accuracy. This calculation has been based on worst case extremes and we have found that a 10 second gating time gives very acceptable results.

SUMMARY

A simple double conversion technique permits separation and limiting of television visual carriers so that they can be reliably measured with a digital frequency counter. The double conversion system uses readily available components and can be bread-boarded in a few hours. Frequency counters operating in the desired range of carrier frequencies are essential for practical measurements. Intercarrier measurements are easily made with the same equipment using the 4.5 MHz intercarrier output from the demodulator. A 10 second gating time is usually adequate to average the frequency swings caused by the FM modulation of the aural carrier.

July, 1971

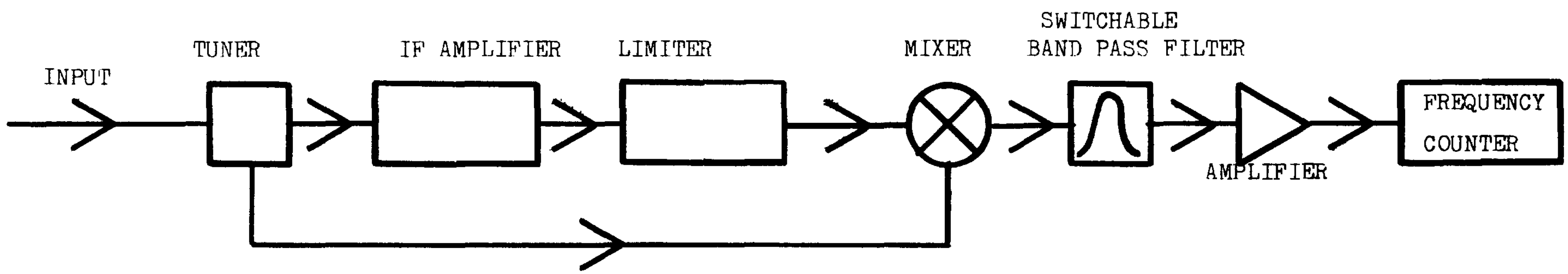
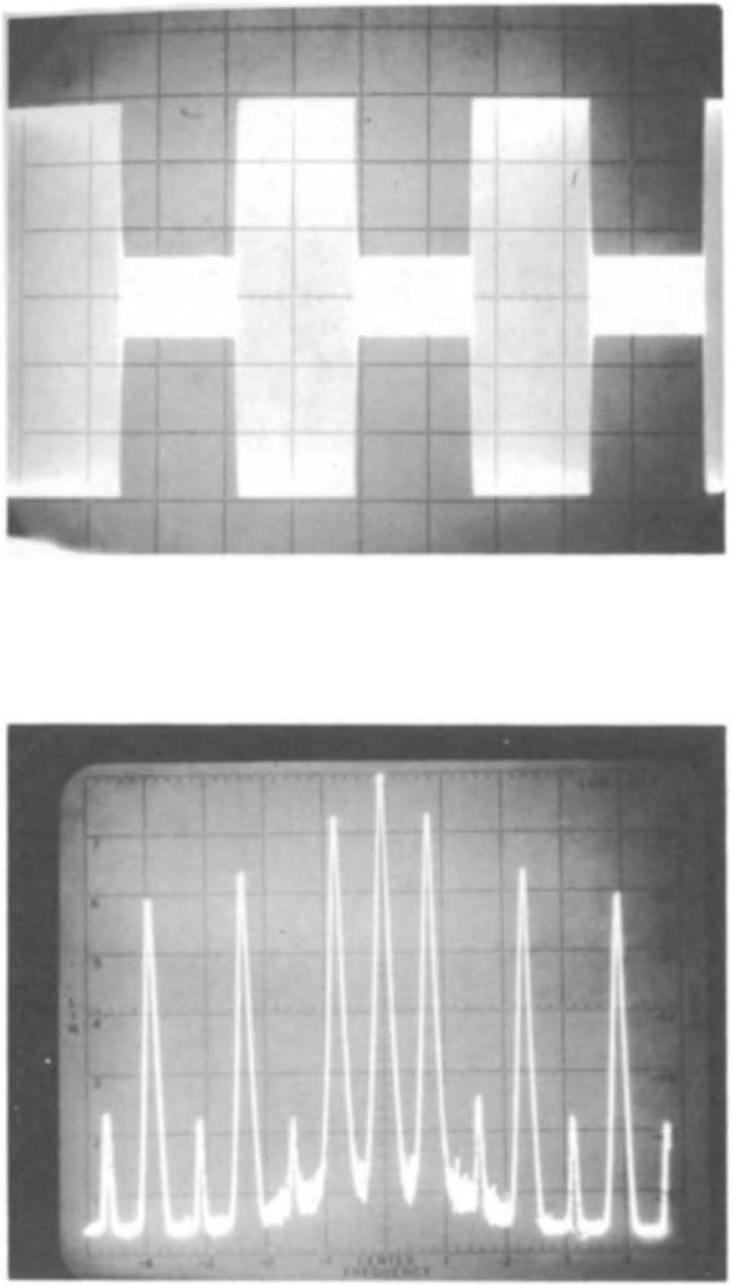
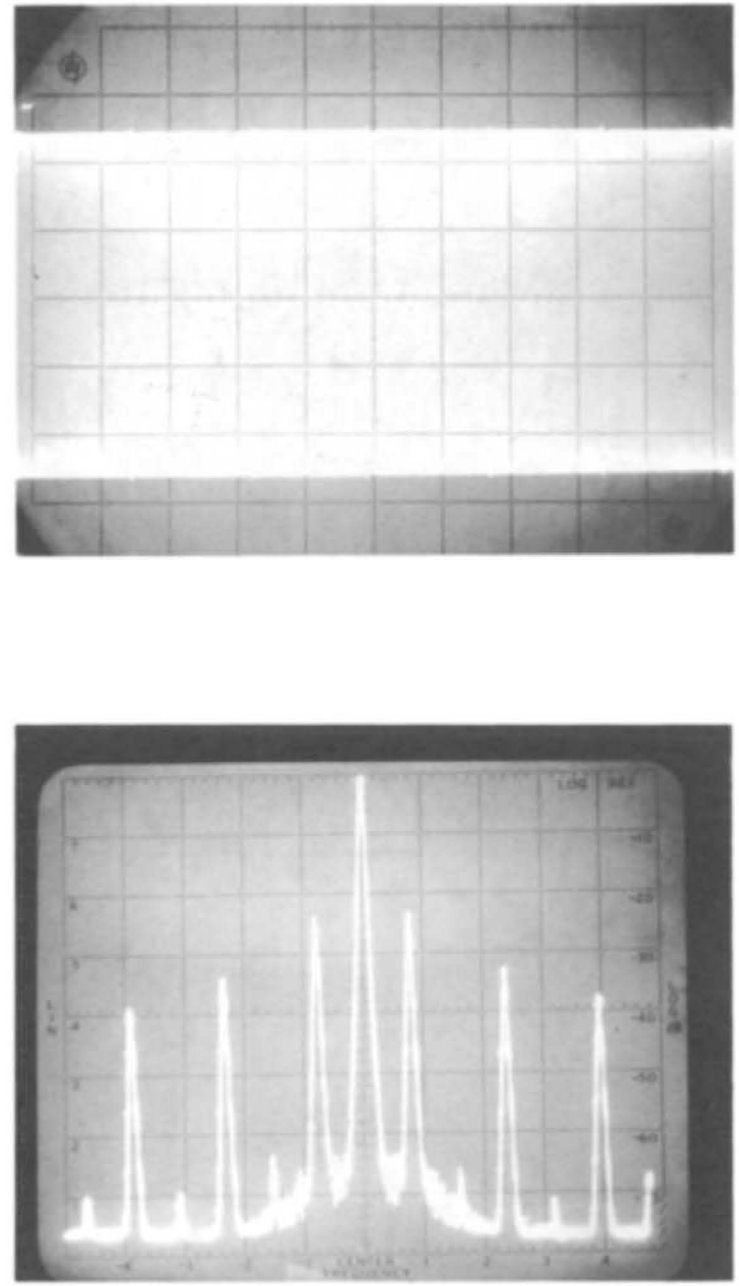

INPUT
TUNER
IF AMPLIFIER
LIMITER
MIXER
SWITCHABLE
BAND PASS FILTER
AMPLIFIER
FREQUENCY
COUNTER

FIGURE 1

80% modulation (15.75 KHz square wave)

after limiting



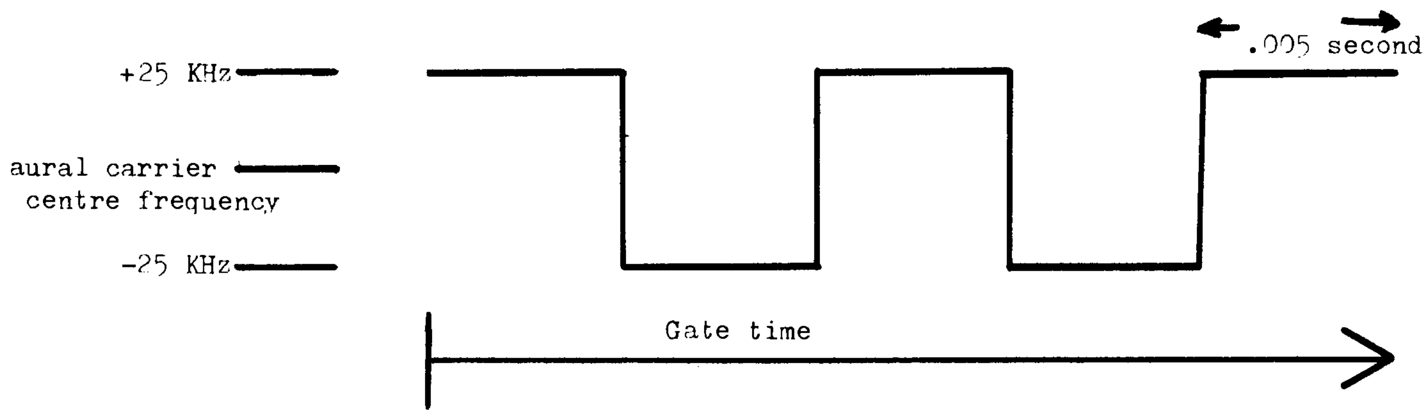


10 db/div

---------- 20 KHz/div ------------

FIGURE 2

FIGURE 3



Idealized FM (100 Hz square wave modulation)

Figure 4

AN OPTICAL LINK FOR CATV

by

R. T. Daly and M. G. Cohen

ABSTRACT

This paper studies the technical feasibility of using a free space cascade of optical (laser) links for CATV service. By employing available Weather Bureau visibility statistics for New York City, the effects of optical path attenuation on link reliability are determined. Based upon these results, a link is proposed which will provide 99.9% reliability with a 42db signal-to-noise ratio for transmission of 25 TV channels plus the FM band. Repeater spacings, required optical powers, detailed configurations, modulation techniques and costs are discussed.

quanTronix

AN OPTICAL LINK FOR CATV

by

R. T. Daly and M. G. Cohen

I. TECHNICAL DISCUSSION

The performance requirements and construction of link elements (transmitter and receiver) will be developed in this section based upon the following assumptions:

(a) The cascaded link elements are identical and are spaced by 80 meters. (80 meters is about the distance between East-West streets in Manhattan, New York City.)

(b) The cascade of links must achieve an overall reliability with respect to weather conditions of 99.9%. That is, for 0.1% of the time the carrier-to-noise ratio at the output terminal of the cascade may be below the minimum design value.

(c) Weather statistics correspond to those of New York City.

The cascade consists of K equally spaced outdoor intervals as shown in Figure 1. Atmospheric transmission path losses are uniform over the entire length of the cascade and correspond to 33 db/1000 ft. This value of optical path transmission is exceeded (on the average) in New York City 99.9% of the time.

As shown, the input to the link is a high-level, noise-free signal. At each stage in the cascade, the carrier-to-random noise power ratio, c/n, (measured at the output of each receiver) decreases, since each square-law photodetector load and associated preamplifier contributes additive white noise which can be treated as thermal noise at the effective noise temperature of the preamplifier. With a noise-free and high-level input signal at the head end of the cascade, the input to the second transmitter has a finite value $(c/n)_0$. At the output of the Kth receiver the carrier-to-noise ratio $(c/n)_K$ will be $1/K\,(c/n)_0$. Expressed in db, $(c/n)_0 = (c/n)_K + 10 \log K$. For a cascade one mile long consisting of 20 intervals, $10 \log K = 13$ db, so $(c/n)_0 = 55$ db is required for $(c/n)_K = 42$ db. This level must be maintained under the assumed optical path transmission conditions.

Before proceeding with a discussion of the equipment configuration, it is necessary to consider the significance of atmospheric transmission. The relation $T_a(R) = \frac{R^2}{R_c^2}$, where $T_a(R)$ = one way path transmission and R and R_c, the range and clear ($T_a = 1$) range respectively, establishes the fade margin required for the system. Statistical visibility data gathered at several U.S. cities for air traffic control planning permits the evaluation of the frequency and duration of the optical path fading[1]. Data given for New York (taken at JFK International Airport) show that over an 80 meter path, a fade of 17 db (referenced to a point in the receiver following the square law optical detector) is exceeded only 0.1% of the time. Thus, with this fade margin, each element of the cascade is within performance specification

99.9% of the time. Additionally, the data support predictions that fading over the cascade is correlated, i.e., the overall cascade availability is the same as that for any element.

With this information, the value R_c can immediately be established by noting that

$$20 \log T_a(R) = 40 \log (R/R_c) = -17 \, \text{db}$$

so that for R = 80 meters, R_c = 210 meters.

At the present state of technology, and for the signal bandwidth to be accommodated in this service, the optical transmitter must consist of a cw laser source in the wavelength range between about 1 μm (near infrared) and 0.4 μm (visible blue) and an external modulator capable of handling the signal bandwidth.

For the cascade element separation ($\approx$ 80 meters), the relatively high carrier-to-noise ratios required in this service and in the interest of cost, direct square-law optical detection is the practical choice. Depending upon the wavelength region selected, i.e., the laser transmitter choice, two more or less different detector devices can be considered. For those wavelengths corresponding roughly to the visible spectrum (0.7 μm to 0.4 μm) the combination of photocathode and electron multiplier (photomultiplier tube) is available. Alternatively, the use of solid state photodiodes for the near IR portion of the spectrum (1 μm to 0.6 μm) offers good performance. The lower cost, accurate linearity and excellent performance of available silicon PIN diodes as IR detectors strongly recommends their use, even though link performance may be degraded by a few db.

For a system consisting of a cw laser, optical modulator, solid state (diode) detector, optics and electronic circuits, the schematic diagram would be as shown in Figure 2.

It can be shown that the "clear" range, R_c, of a single transmitter-receiver link element, is given by

$$R_c{}^2 = \frac{R^2}{T_a(R)} = \left(\frac{m^2}{(c/n)_0}\right)^{1/2}\left(\frac{\eta_o P_L A_R}{2\Omega_T}\right)\left(\frac{S^2}{8kT_e C_T B B_o}\right)^{1/2} \qquad (1)$$

where (1) applies in the case of a system limited by detector/preamplifier thermal noise.

The parameters in Eq. (1) are defined as follows:

$T_a(R)$	=	one way atmospheric path transmission
R_c meters	=	maximum range separation at which performance level will be met with zero atmospheric path attenuation ($T_a(R) = 1$)
R (meters)	=	maximum range for $T_a(R) < 1$
m	=	peak modulation of optical carrier for any single channel signal
$(c/n)_0$	=	electrical carrier-to-random noise power ratio referred to maximum rms channel carrier
η_o	=	system optical efficiency, i.e., lens and filter losses
P_L watts	=	cw laser power generated
A_R (meters^2)	=	clear aperture of receiver
Ω_T (steradians)	=	effective solid angle of projected transmitter beam. $4\pi/\Omega_T$ would be termed "antenna gain" in microwave terms
B (Hz)	=	overall link bandwidth
B_o (Hz)	=	bandwidth of a single individual TV channel
S (amperes/watt)	=	responsivity of the optical detector element
k (joules/°K)	=	Boltzmann's constant
T_e (°K)	=	effective noise temperature of receiver preamplifier, i.e., $T_e = F \times (290)$ °K where F is the preamplifier noise factor
C_T (farads)	=	total capacity shunting the optical detector element

A good approximation to the details of the link can be gained by solving Eq. (1) using some typical values. For the moment we assume that a total of N independent but frequency-contiguous TV channels are to be carried and that the total depth of modulation is adjusted such that $m = (4N)^{-1/2}$* Other parameter values are as follows:

$(c/n)_0$	=	3.2×10^5 (55 db in the receiver circuits for a 20 element cascade)
η_o	=	0.7
A_R	=	1.3×10^{-2} m^2 (5" dia. optics)
Ω_T	=	1.5×10^{-6} ster (2×10^{-3} rad. linear angle or 0.12°)
B_o	=	4×10^6 Hz
B	=	$3/2\, NB_o$
S	=	0.4 amperes/watt (for silicon PIN diode at $\lambda = 0.63\,\mu m$)
k	=	1.38×10^{-23} joules/°K
T_e	=	600°K (3 db noise figure)
C_T	=	5×10^{-12} farads
R_c	=	210 m

Eq. (1) can be solved, using these values, to give

$$\frac{P_L}{N} = 110 \times 10^{-6} \text{ watts/channel}$$

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* The choice of this value for m results from a detailed calculation of the modulation scheme and will be published separately.

The value selected for S, above, anticipated that the laser wavelength was $0.63\mu m$ radiation from a helium-neon gas laser. Small helium-neon lasers are readily obtainable with power outputs of 3 milliwatts, sufficient to transmit over 25 TV channels.

A value S = 0.2 amperes/watt is appropriate for wavelength $\lambda = 1.06\mu m$ corresponding to a YAG laser. Eq. (1) would give, in this case, $P_L/N = 220 \times 10^{-6}$ watts/channel. Small YAG lasers have been built with outputs exceeding 500 milliwatts, which is one hundred times the required level for 25 TV channels.

Because the modulation/demodulation process generates beats among the modulated VHF channel carriers, it is not possible to cleanly transmit more than one octave on a "single" optical carrier. However, using the two orthogonal linear polarization states of a laser beam, two optical carriers can be created. These carriers may be separately modulated and then recombined for transmission. At the receiver, each beam produces its own photocurrents since no demodulated signal can arise from interaction of the orthogonal polarizations. Polarization beamsplitters and recombiners are not the most economical method of providing dual channel operation. It may be possible to employ acousto-optic modulators which act upon only one polarization with two crossed modulators positioned along the beam. Each modulator channel is imposed upon an orthogonal polarization component, with the transmission and demodulation proceeding as described previously.

Using the dual channel technique, substantially all of the presently desired CATV spectrum can be accommodated. Thus, 54 to 108 MHz feeding the first modulator covers channels 2 through 6 plus the FM channels, while 120 to 240 MHz feeding the second modulator provides coverage of channels 7 through 13 plus the non-standard midband and high-band channels A through M, resulting in transmission of a total of 25 TV channels plus the FM band.

II. SAFETY AND ECONOMICS

While there is general agreement as to the potential eye hazard created by intense laser beams in the spectral region from $1.5\mu m$ to $0.35\mu m$ wavelength, there is, as yet, no solid data as to the flux density levels which can safely be tolerated. At least two issued specifications[2][3] cite as a safe level 5×10^{-5} watts/cm^2 at $\lambda = 0.63\mu m$ (helium-neon laser) and 1.5×10^{-4} watts/cm^2 at $\lambda = 1.06\mu m$ (YAG laser). From a 5" diameter objective, therefore, about 7 milliwatts from a helium-neon laser and 20mw from a YAG laser will provide this flux level near the transmitter. However, these published values are generally regarded as highly conservative, perhaps by as much as 1000 times, and workers in the laboratory frequently receive higher exposures without apparent effect.

It has been reported that the cost of material and installation for one mile of underground CATV cable in New York City usually ranges between \$10,000 and \$20,000 and in some special cases, as high as \$40,000. Assuming the optical link to be fully equivalent to cable in function and performance, in order to be cost-competitive with cable, each repeater unit must sell for a maximum of \$1,000 (including installation) since 20 repeaters would cost \$20,000. At a higher selling price, the optical link would appear to find application only in special cases, for example, where its capability for quick installation with high recovery value on an abandoned route was important.

An estimate of the component and assembly costs for the types of equipment discussed in Section I shows that the helium-neon system with 25 channel capacity is just possible at $1,000/repeater if production volume were high. The YAG system, on the other hand, would be difficult to produce for under $2,500 - again, if volume were high.

III. RECOMMENDATIONS

The significantly lower cost of a helium-neon laser (about $100, in quantity) compared with a YAG laser (about $1,500 in quantity), is the major contributor to the difference in system cost. The helium-neon system would be an immediate choice on the basis of cost alone, but as the results show, a 3 milliwatt helium-neon laser provides little margin over the nominally required 2.6 milliwatts. Unfortunately, fundamental limitations exist in obtaining higher usable output from this laser unless the cost is significantly increased.

The YAG laser, on the other hand, while its cost is nearly $1,500/unit greater, provides a very significant margin. Thus, compared to a nominally required 5 milliwatts, the YAG can provide, easily, 500 milliwatts. This power excess can be advantageously used in several ways:

(a) Increased weather immunity - a reliability of 99.99% can be obtained with a 10-fold increase in power.

(b) Wider transmitted beam for more tolerance in mounting and aiming is obtained by increasing the projected beam to 3 milliradians.

(c) Greater tolerance to lens dirt, which reduces to a transmission to 70% instead of 90%, costs another factor of 1.8.

With these advantages, a total laser power of 180 milliwatts is required, still leaving a margin for component degradation. In view of these considerations, it is recommended that any system design which aims toward 25 channels utilize the YAG laser. For one or a few channels, however, the helium-neon laser remains the clear choice.

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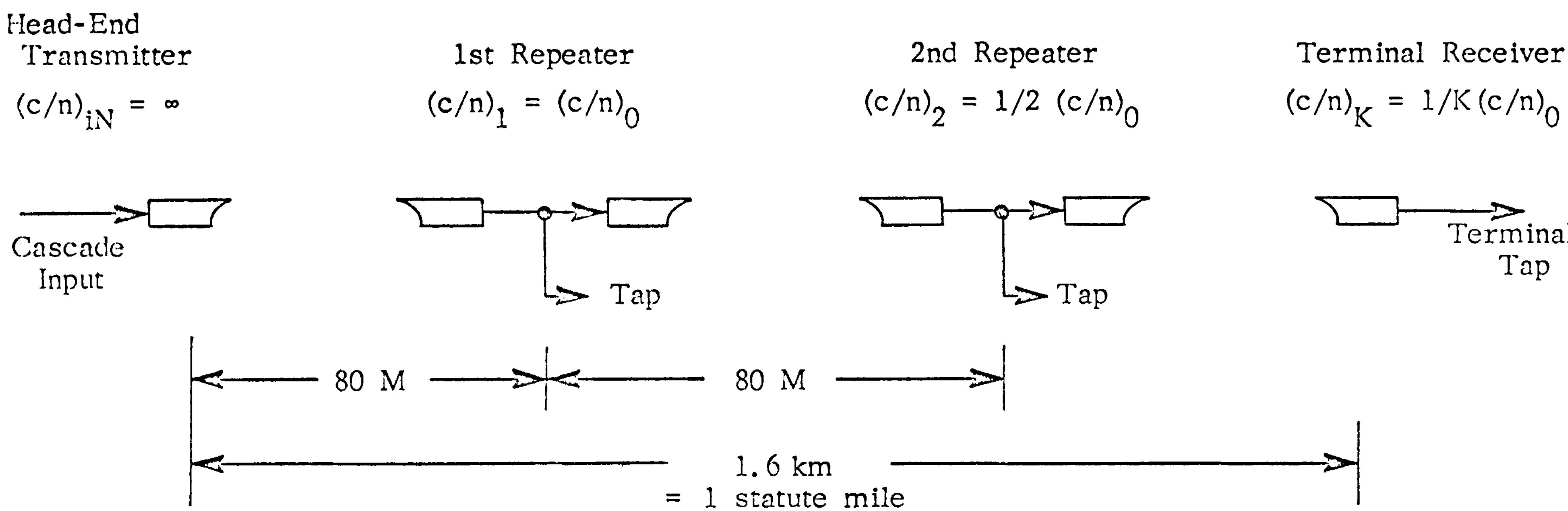

FIGURE 1. ONE MILE CASCADE OF K OPTICAL LINKS

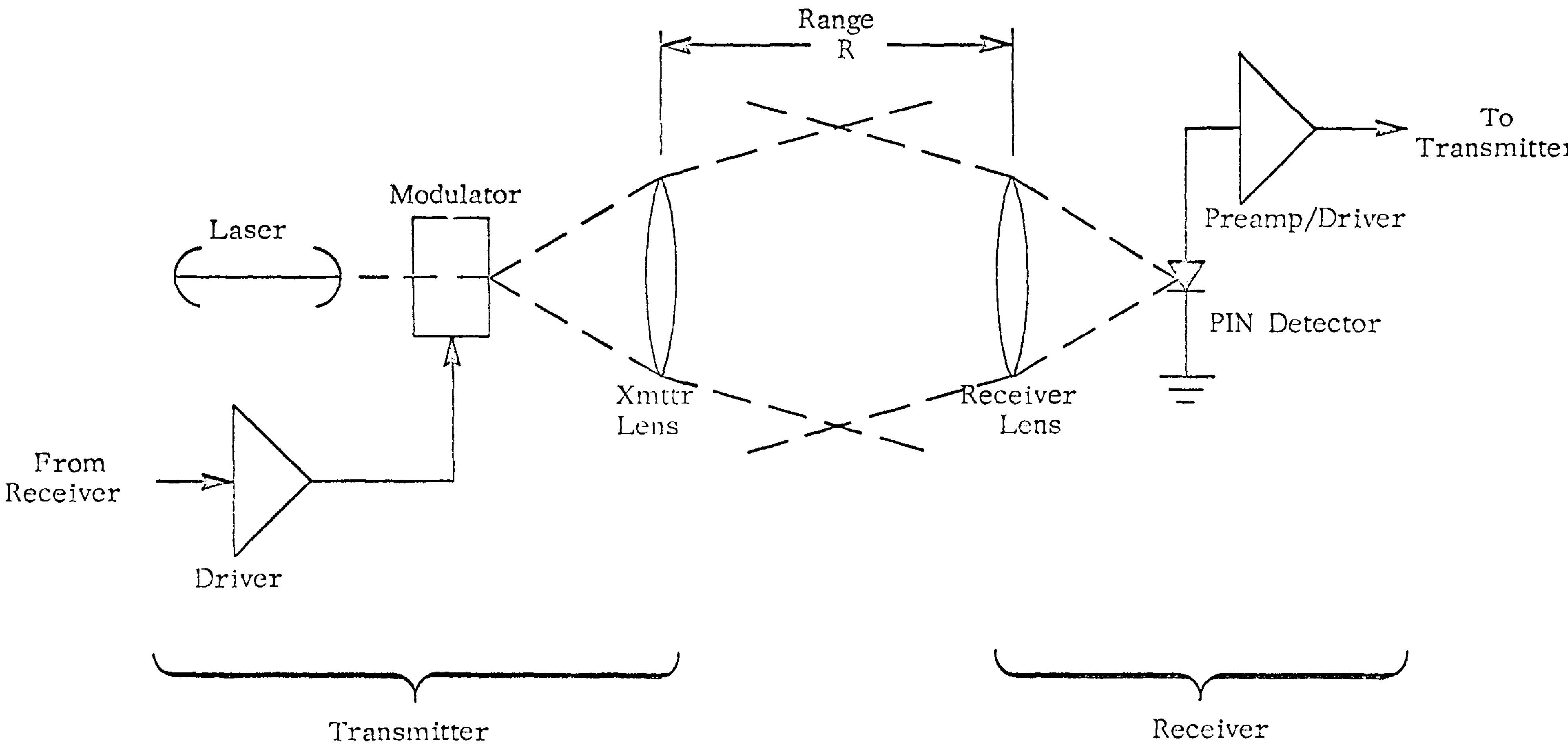

FIGURE 2. SCHEMATIC FOR ONE CASCADE LINK ELEMENT FOR CATV

ANTENNA RADIATION PATTERN ANALYSIS
&
CO-CHANNEL PROTECTION

by

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INTRODUCTION

Many CATV technicians, with otherwise well-rounded backgrounds, have been forced in the past and will be required in the future to make critical antenna or antenna-array selections to avoid annoying co-channel interference problems.

This paper has been prepared to clarify:

1. What are the basic terms and classifications of antenna radiation patterns.
2. Why radiation patterns must be taken on the antenna test range.
3. How to analyze antenna radiation patterns.
4. Why radiation pattern irregularities must be given special attention.

BASIC TERMS AND CLASSIFICATIONS OF RADIATION PATTERNS

Co-channel protection is basically an antenna performance problem which is characterized by the antenna specifications. Antenna manufacturers usually publish a more or less complete list of electric parameters of their products, such as antenna gain, input match, front to back ratio, beamwidth, etc. However, this qualitative information is not sufficient for co-channel protection evaluation. For a meaningful QUANTITATIVE evaluation we must have at our disposal the actual radiation patterns, as taken on the antenna test range.

Every antenna has a three dimensional radiation pattern because it is radiating into all angles of space. However, for co-channel evaluation purposes we can limit our investigation to the horizontal (E) plane. The different co-channel offenders arrive from different AZIMUTH ANGLES, thus a vertical radiation pattern would have no meaningful information.

The horizontal radiation pattern of an antenna describes the field intensity of the radiation as a function of the azimuth angle. The pattern may be presented in polar or rectangular coordinates. Polar presentation is preferred for popular publications. This information gives an easy to understand picture of the received or transmitted power distribution. By contrast, the RECTANGULAR radiation pattern permits a presentation of much finer detail including precise dB readings of peaks and nulls.

The radial deflections on the polar and rectangular charts may be arranged in:

* Linear scale
* Power scale
* dB scale.

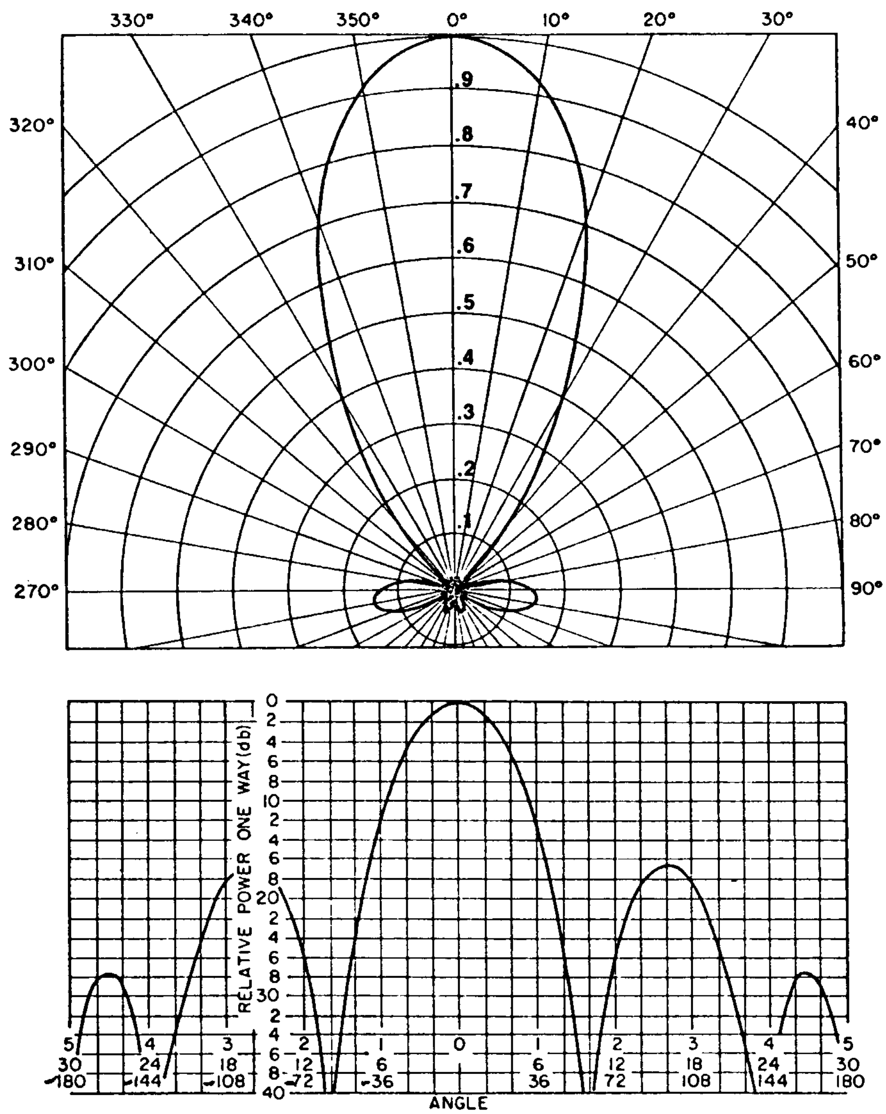


Figure 1

Compare the readability of a polar and horizontal radiation pattern presentation.

The dB scale, containing the logarithmic variations in the received signals, is the most beneficial for the examination of CATV antenna array radiation patterns.

Figure 1 presents polar and rectangular radiation patterns taken from the same antenna. It should be noted that while location, depth, and width of the null at 168° is somewhat fuzzy on the polar pattern, the rectangular pattern offers good readability and accuracy.

ANTENNA TEST RANGE AND RECORDING EQUIPMENT CONDITIONS

We realize that 99% of the CATV operators and technicians have not been involved and will not participate in antenna test range operations. But in order to comprehend the physics of antenna radiation patterns one must have a greater understanding about antenna test range and recording equipment conditions.

Usually a big, open field is selected for antenna test range purposes. The transmitting gear is established at one end and the receiving/recording gear is located at the opposite end. (Figure 2)

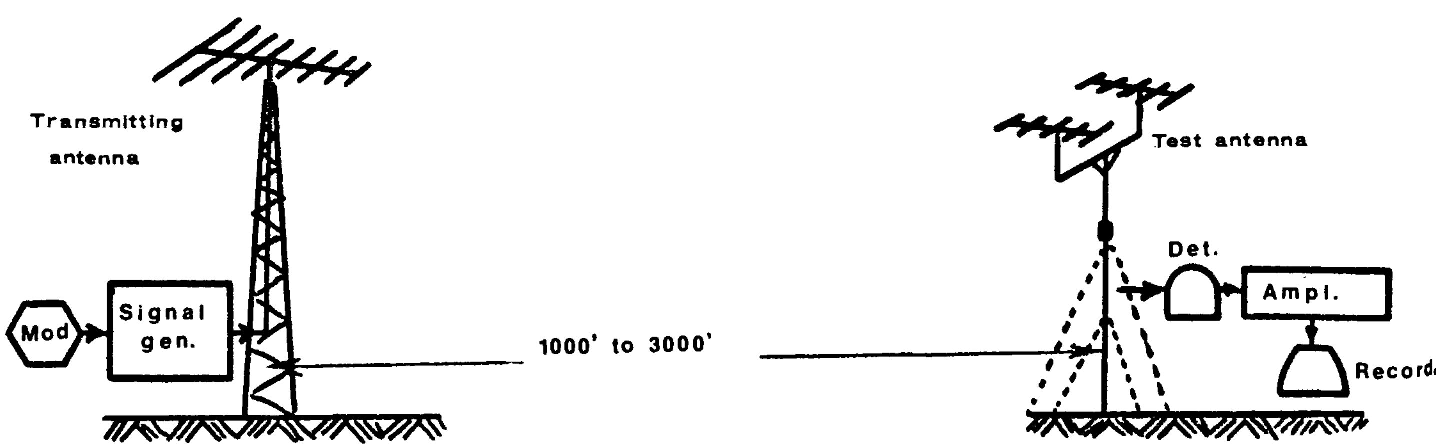


Figure 2

The modulated RF signal is beamed from the fixed high-gain antenna in the direction of the receiving (test) antenna. At the receiving end, the test antenna is rotated either manually or by motor. In both cases the rotor movement is syncronized with the recorder. The detected and amplified signal is processed through the recorder to produce a permanent chart of the radiation pattern.

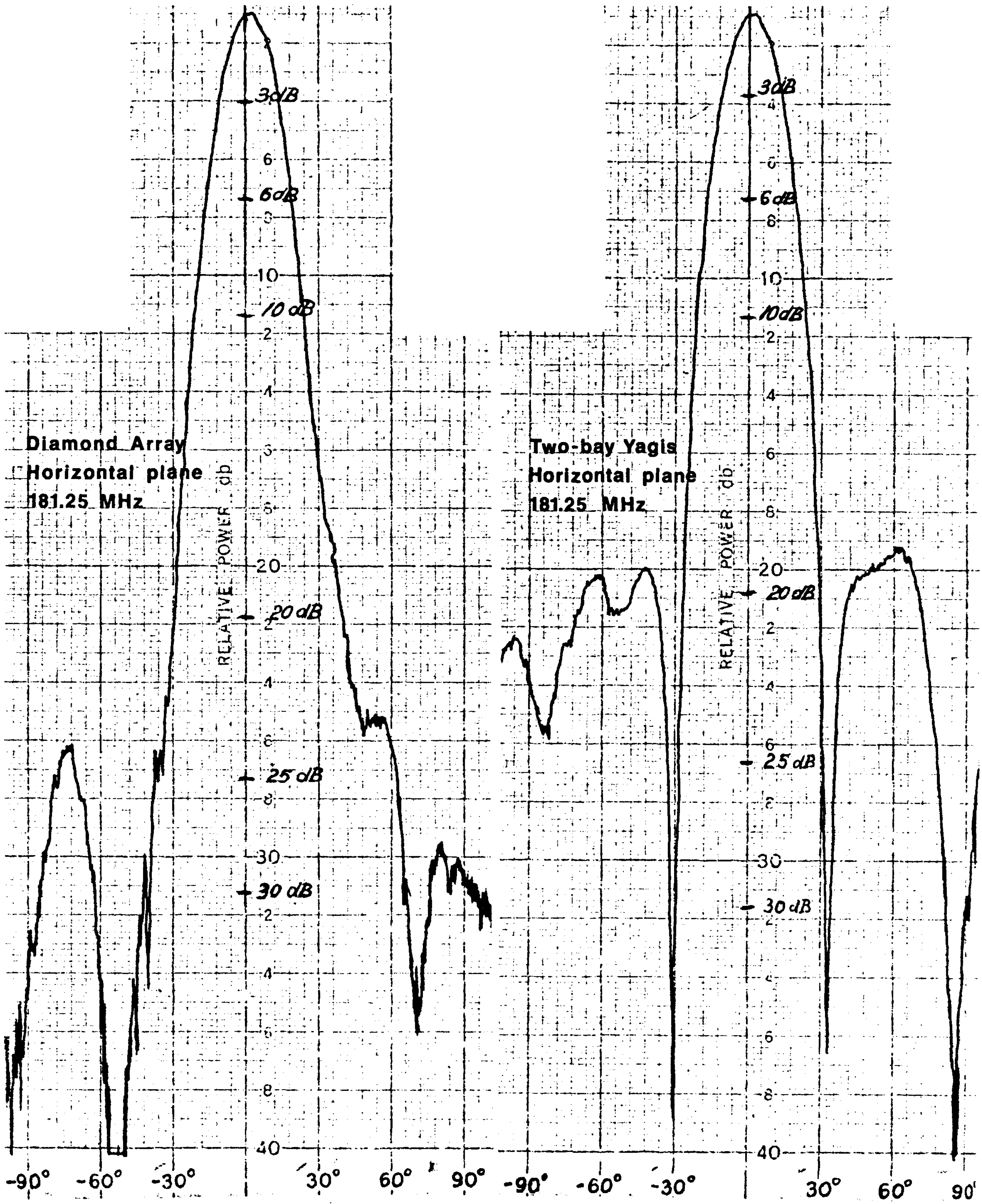
There are several critical antenna test site conditions which may adversely effect the accuracy of the obtained test information :

* Separation between the transmitting and receiving antennas (1000' to 3000' represents adequate separation)
* Directivity of the transmitting antenna (a minimum of 45° beamwidth is required)
* Height of the transmitting and receiving antennas above ground (50' to 100' is a sufficient height)
* The surface quality of the antenna test range (rough surfaces are preferred).

A combination of the last two conditions contribute to reflection problems, the most serious source of errors. Ground reflection is a function of the surface configuration, type of soil, content of moisture, vegetation, weather, transmitting frequency, polarization, etc. Increasing the transmitting and test antenna heights well above ground and the installation of conductive fences perpendicular to the line of propagation are two practical means to reduce ground reflections.

A number of recording equipment conditions may also contribute to the limited accuracy of the radiation pattern:

* The instability of the transmitter and receiving equipment
* Calibration inaccuracies of the output power
* Problems inherent in the detector characteristic.



Diamond Array
Horizontal plane
181.25 MHz
RELATIVE POWER db
3dB
6dB
10 dB
20 dB
25 dB
30 dB
Two-bay Yagis
Horizontal plane
181.25 MHz
RELATIVE POWER db
3dB
6dB
10dB
20dB
25dB
30dB
-90°
-60°
-30°
30°
60°
90°

Figure 3

Since detectors are non-linear devices, the detected signal depends largely on what portion of the characteristic is used for detection. This in turn is a function of the output power of the signal generator, the distance between the transmitting and test antenna, etc.

Good antenna engineering practices dictate the calibration of every chart before starting a series of radiation pattern tests. (See Figure 3) By inserting 3-6-10-20-25-30 dB pads into the transmitter's output, the detected voltages can be precisely marked on the chart. Figure 3 demonstrates the need for calibration. At the -22 dB chart-paper mark the calibration yielded an actaul -20 dB reading. In co-channel protection analysis the -20 to -30 dB region contains the most important segment of the chart, and calibration should not be omitted.

RADIATION PATTERN ANALYSIS

There are a number of paramount factors determining the co-channel protection capability of an antenna-array, all readable from the radiation pattern:

* The exact azimuth angle of the nulls
* The depth of the nulls
* The width of the nulls
* The shape of the main beam.

Potential co-channel offenders should be identified during the signal survey. The azimuth angles of these actual co-channel interference stations may then be determined accurately by a computer run.

Should a fixed structure array be employed, such as a diamond array of log-periodic antennas, its actual radiation patterns must be very closely examined: are the null directions coinciding with the azimuth angles of the co-channel offenders?

The other popular approach is to custom design antenna arrays to force nulls in those particular directions from where the interfering signals are arriving. In these cases actual radiation patterns may prove that the nulls are right on the target, or perhaps a few degrees off. In the latter case a slight reorientation of the tower mounted array may swing the null into the desired position without significantly decreasing the antenna gain.

THE DEPTH OF THE RADIATION PATTERN nulls may also be conveniently identified from a rectangular radiation pattern with a dB scale, enabling the CATV technician to affirm manufacturers' specifications or to discover overoptimistic claims.

Nulls exhibiting 20 dB depth cannot be considered adequate co-channel protection. 40 dB deep nulls are highly desired but seldom demonstrated on actual radiation patterns.

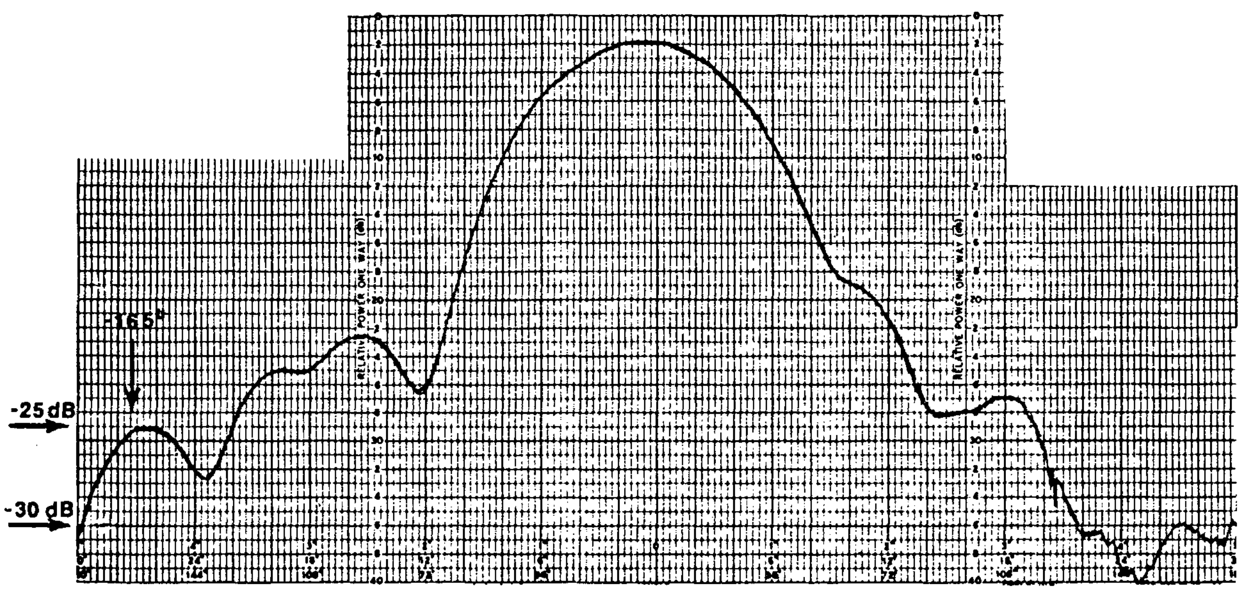


Figure 4

Figure 4 illustrates the need for F/B ratio specifications and radiation pattern comparison. The manufacturer's specifications stated 30 dB protection from the back. The radiation pattern indicates that indeed the diamond quad provides 30 dB protection from 180°. However, the co-channel offender arriving from -165° azimuth--and we can still call this "from the back"--will be attenuated only by 25 dB. This is a 5 dB deviation from published specifications.

THE WIDTH OF THE NULL is an additional important characteristic with which to be concerned. From a rectangular radiation pattern with dB scale, this parameter may be precisely measured at any null location.

Very narrow (1° to 3°) nulls should warn the CATV technician of two imminent problems:

1. It is difficult to orient a tower mounted CATV array with such accuracy under normal working conditions.
2. Under medium to heavy wind conditions the twisting of the CATV tower, combined with the movements of the antenna gates and pipes, could skew the the nulls of the radiation pattern by several degrees, thus significantly decreasing co-channel protection.

Nulls of 5° to 8° width are considered optimum for CATV application.

RADIATION PATTERN IRREGULARITIES

It is not uncommon to encounter asymmetrical radiation patterns. A re-examination of the diamond array pattern and horizontally stacked two-bay pattern (Figure 3) shows a number of asymmetrical features. These include a missing null at -90° on the two-bay pattern, or the development of a broad shoulder on the diamond array pattern at $+50^{\circ}$. These are warning signs indicating that either the antenna test range or the constructed arrays have hidden inadequacies. If co-channel conditions warrant the need for extreme caution, the radiation pattern testing should be repeated on a slightly altered test range or with a different mounting in order to identify the nature of the asymmetrical pattern performance.

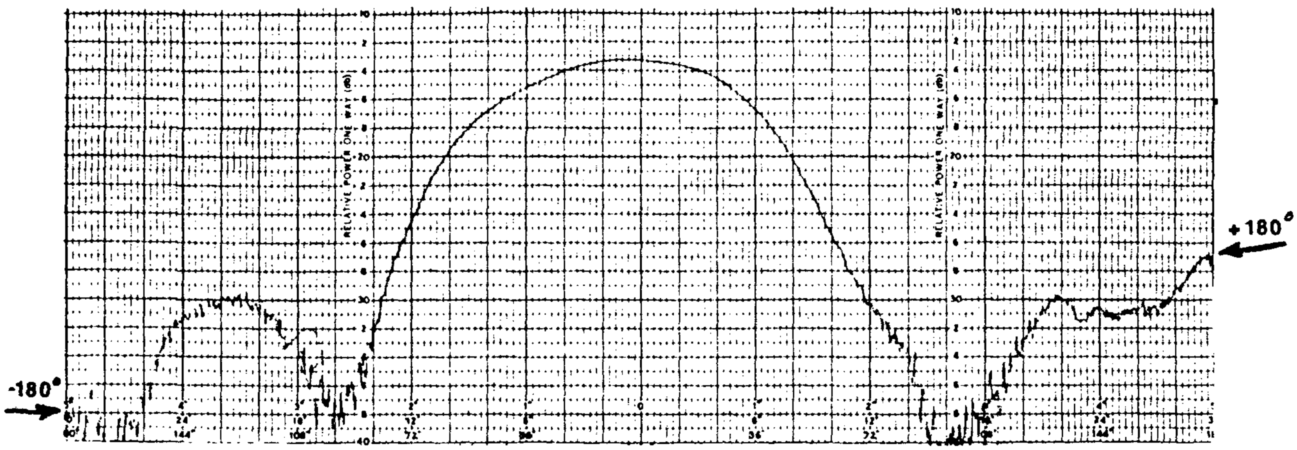


Figure 5

Figure 5 is a classical example of a faulty radiation pattern, not to be accepted for evaluation purposes. The radiation pattern of a log-periodic antenna is presented in this chart. Note that the signal level responses differ considerably at 180°. On the left side, the curve dips to -22 dB, while on the right side it levels off at -13 dB. The resulting 9 dB difference between the left and right side could be a serious impairment of the recording equipment or the result of transmitter/receiver instability.

The frequency of radiation pattern testing is also an important qualifying parameter. Co-channel beats are generated by the video carrier frequencies of two or more stations. Therefore, for co-channel testing purposes, the radiation patterns must be tested on the respective video carrier frequencies. The pattern response may or may not change within a couple of MHz; however, the bearings of the nulls, the depth of the nulls, and their width may shift considerably, warranting on-video-carrier testing.

Antenna arrays mounted on small diameter pipes on the top of the wooden test tower may exhibit perfect radiation patterns with deep nulls. But mounted on metal antenna gates, in the vicinity of a 48" face tower, they may not perform as well. Reflections from the horizontal braces of the tower and the long horizontal pipes of the antenna gates will generate phase sensitive cancellations: filling in the deep nulls in the pattern or causing null-shifts. There is little point for example to publish a "mast mounted" radiation pattern for a diamond array, if the array must be mounted on a CATV tower with 40" to 60" tower face.

CONCLUSION

It has been shown that co-channel protection can be evaluated by analyzing the actual radiation patterns of the array. What often stands in the way of such effective evaluation is the missing information: the properly taken radiation pattern itself. It is up to the CATV operator and technician to ask for and obtain that information.

Cable TV System Calculator

by J.Cappon.P.Eng.

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INTRODUCTION.

If someone who is not familiar with Cable T.V. would ask me what planning a Cable T.V. system is like, my answer would be: it is very much like playing chess. Before you make a move you have to consider all the consequences of that move, as well as the consequences of the moves you plan to take thereafter.
To illustrate this, fig 1 shows a typical situation.

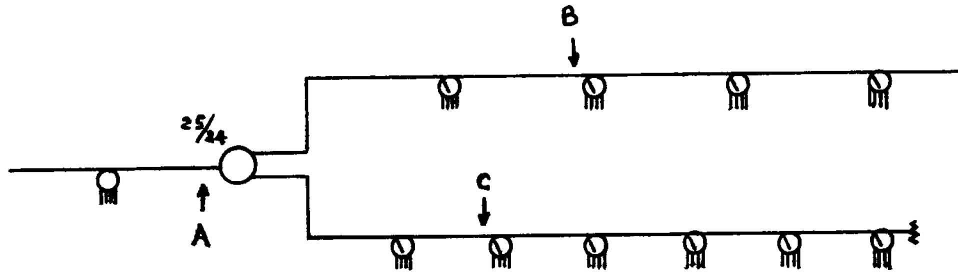

Fig.1

Signals arrive at the cable split with +25dBmV at ch 13 and +24dBmV at ch2. By merely splitting the signal amplifiers are required at point B and point C.
However, with one amplifier before the splitter at point A, this one amplifier is able to serve these two branches.
Fig. 2 shows the computation required to find out where the signal needs reamplification. Roughly three times as much effort is required to conclude that one amplifier would suffice.

CH 13 LINE	TAP	CH 2 LINE	TAP		CH 13 LINE	TAP	CH 2 LINE	TAP	
25.00		24.00			25.00		24.00		
3.30		3.30		SPLITTER	3.30		3.30		SPLITTER
-----		-----			-----		-----		
21.70		20.70			21.70		20.70		
1.70		.85		100FT .412	2.50		1.30		150FT .412
-----		-----			-----		-----		
20.00	10.00	19.85	9.85		19.20	9.20	19.40	9.40	
3.30		3.30		4-10 TAP	3.30		3.30		4-10 TAP
-----		-----			-----		-----		
16.70		16.55			15.90		16.10		
1.70		.85		100FT .412	2.50		1.30		150FT .412
-----		-----			-----		-----		
15.00	X	15.70	X		13.40	X	14.80	X	

Fig.2

From this example it is rather obvious that an aid to take over the time consuming calculations would be highly desirable.

COMPUTER ASSISTED DESIGN

A computer therefore, which in a split second could perform these calculations and come up with the optimum solution would be very helpful.
However, we must realize that the "Cable T.V. system computer" faces the same problems as his cousin the "chess playing computer".
The basic problem in programming a computer to play chess is in teaching the machine to be selective in the possible lines of play it considers. Where the human player is able to reject over 95% of possible continuations, the computer must labour through all variations before making a selection.
There are ofcourse more possible alternatives to consider by the "chess playing computer" than by the "Cable T.V. system computer".
To be precise there are six major alternatives to consider:

a. a splitter.
b. an amplifier followed by a splitter.
c. a directional coupler (-8dB) with the branch line pointing down.
d. a directional coupler (-8dB) with the branch line pointing up.
e. an amplifier + a dir.coupler (-8dB) with the branch line down.
f. an amplifier + a dir.coupler (-8dB) with the branch line up.

In a section of a distribution system with 16 cable splits, there are 6^{16} different combinations, (2.82×10^{12}). Allowing 1 m sec. for the computer to calculate the effect of one change we will have our answer in 89 years. This example illustrates that the computer, though fast, cannot consider all possibilities; it must therefore, as we stated earlier, be selective in it's choice removing the assurance that all possible alternatives were tried.

THE OPTIMUM SYSTEM.

In attempting to write specifications for an optimum design, four aspects need to be considered:

a. signal quality.
b. reliability.
c. maintenance cost.
d. initial cost.

A low cost system could have pieces of .412" and .500" intermixed, but for the sake of standardization this is not done. Each amplifier could be set to operate at nonstandard levels even while keeping an eye on distortion products. However, for the sake of standardization this too is unacceptable. House drops at -6dBmV could provide acceptable pictures, but out of consideration for older sets, safety factors, direct pick-up and possible second sets, this is not done. Even the most important cost factors such as amplifier operating and subscriber drop levels as well as the maximum number of "distribution line amplifiers in cascade" are compromises between these four previously mentioned aspects. Making these cost determining factors rigid, as one has to do when they are entered as design parameters in a computer program, may result in poor "trade offs".

Let me illustrate: Raising the minimum "tap Off" level by 1dB in a typical distribution line having seven 4 way tap off units spaced 100 ft. apart, will reduce the amplifier spacing by approximately 7%. Guaranteeing this new minimum level for 28 subscribers, based on an amplifier cost of $175.00 would cost approximately $\frac{7}{100} \times \$175 \times \frac{1}{28} =$ $ 0.44/dB/subscriber, which can be considered a reasonable compromise.
However, if at the end of a distribution line the signal would dip below its specified minimum, an additional amplifier would be required because 4 subscriber levels were .5dB below specification. Then the cost is $87.50/dB/subscriber, which should be considered a very poor trade off.
Had the "human touch" been involved this situation would have been spotted immediately, and one of the following possible alternative routes could have been taken:

a. permit the .5dB low subscriber level.
b. lower loss drop cable.
c. amplified tap.
d. indoor amplifier.
e. low power line extender.

Other areas where it is difficult to let the computer decide are where the requirements may alter because of:

a. new subdivisions.
b. possible rezoning of build-up areas.
c. difficulties in obtaining "right of ways".

To let the computer decide would require the programmer to establish probability factors which may be more difficult to determine than to solve the problem itself.

A CALCULATOR?

But up to now there has not been much choice. It is either a slow planner or a "fast" computer, and the cost per mile stayed somewhat the same. Yet a combination of a planner doing the design and a calculator performing the routine calculations appears to have merit. Such a calculator (preferably a desk type) must be capable of the following:

a. accepting input level information.
b. accepting cable type and length information between taps as well as from tap to TV set.
c. being programmed for required minimum level at TV set.
d. instantaneously providing tap off type and value.
e. working in reverse direction (from the end of a line).
f. retaining previously entered data when changes are made.
g. accepting a wide choice of dir.couplers and taps.
h. recording R.F. levels.

Almost all of these requirements could be met by a panel as shown in figure 3.

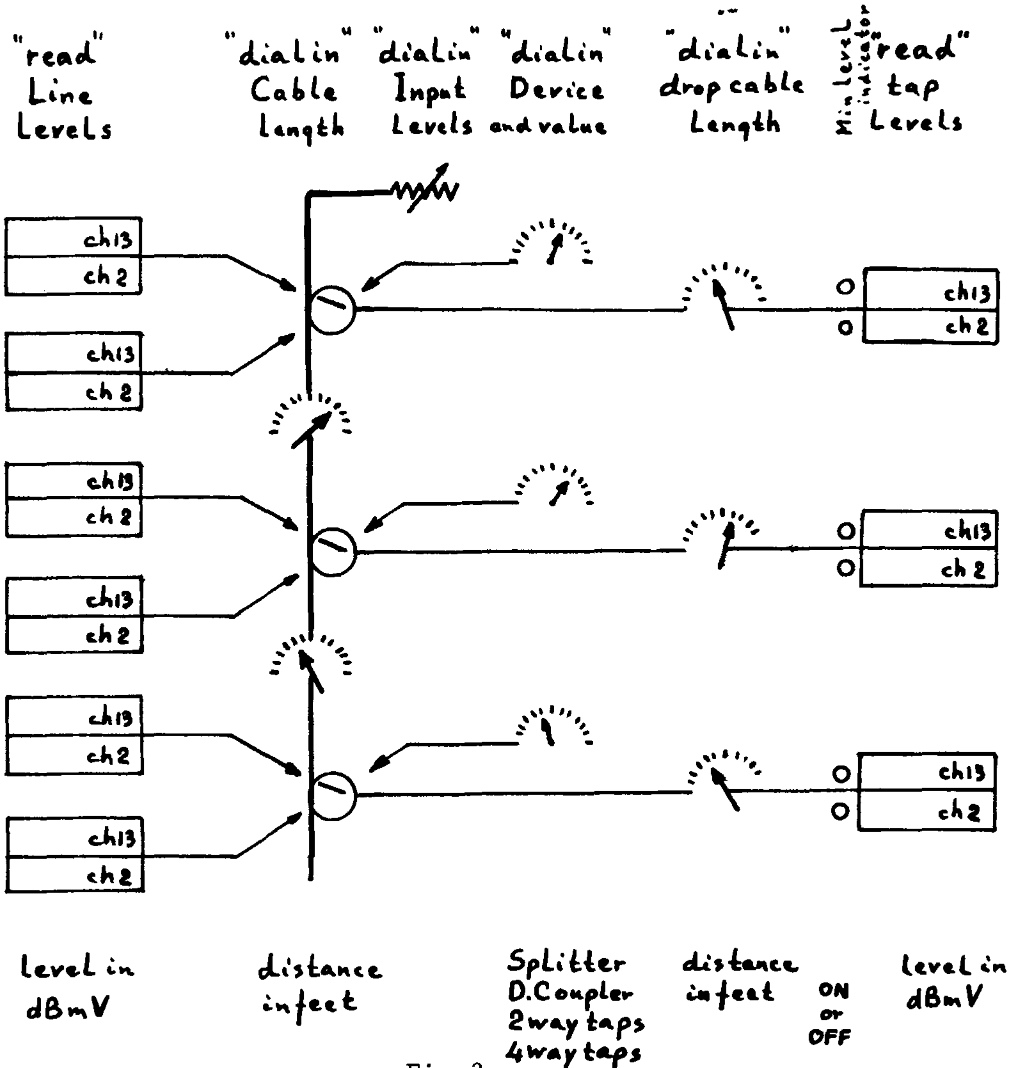

Fig.3

Input level is set with control and is displayed in the first window under "Line Levels". The largest drop cable length is "dialed in" for the first tap, and, subject to the setting of the "device knob" the tap level will now appear in the first window under "tap level". The "device knob" is set for the type of tap-off e.g. 2 way taps and then rotated to a value until both red lights have just extinguished. This process is repeated for each tap-off. When this calculator is requested to work in the reverse direction a feedback loop is required to maintain the last tap level at it's preset levels by controlling the input level.

a. In order to make a panel like this function, simulation can be achieved with RF through the use of high loss cable, switchable attentators, and digital read-cut RF level meters.

This possible solution is obviously too cumbersome.

b. Direct Current can be used. In order to prevent the wide dynamic ranges a current scale whould be chosen where each dB equals 10 mA.
All attenuating devices can now be represented by parallel resistors which are connected to ground,(providing that a constant voltage source is used as a supply) assuring a fixed current through each component according to the attenuation in dB of the device it simulates.
All contacts, plugs and sources need to be constructed in duplicate to simulate ch 2 & ch 13 operation simultaneously.

c. Similarly one dB can be represented by 1V, in which case a constant current source should be employed and each component be simulated by a series resistor, providing a fixed voltage drop according to the attenuation in dB of the device it simulates.

b and c are far simpler than the RF simulation technique; however these methods still leave much to be desired.

A MECHANICAL CALCULATOR?

Could it be that a mechanical-graphical method will out-perform electronics to serve RF distribution calculations?
The basis is a graph as shown in fig.4 with signal level at ch 2 p and ch 13 p plotted along the axes; each point on this graph will then represent a certain signal condition. The point marked A on this graph 41/34 respectively representing ch 13 p and ch 2 p carrier levels, is a typical output for a four way bridger amplifier.

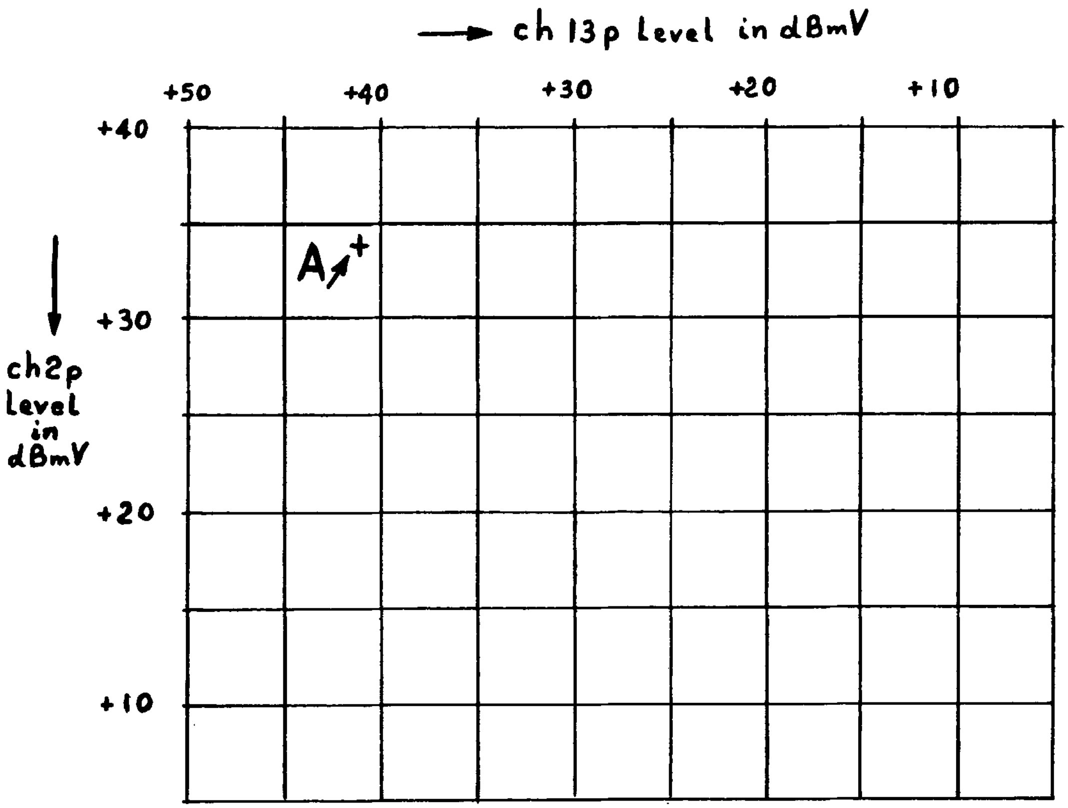

Fig.4

A cable run as shown in fig.5 can be plotted on the original graph. see fig.6

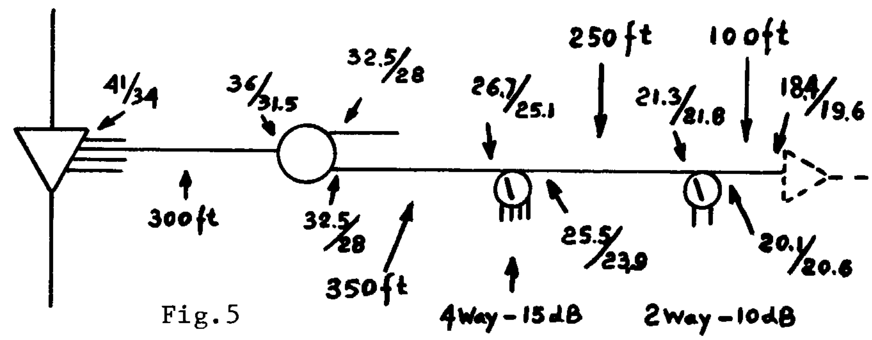

Fig.5

The first 300 ft. cable piece is represented by a straight line between it's input levels of 41/34 and it's output levels of 36/31.5. The 2 way splitter is represented by a straight line between it's input levels of 36/31.5 and it's output levels of 32.5/28. The other levels are represented similarly. It should be noted that the lines representing the flat losses of the splitter and taps have a 45 degree slope on the graph, while the lines representing cable are nearer 30 degree.

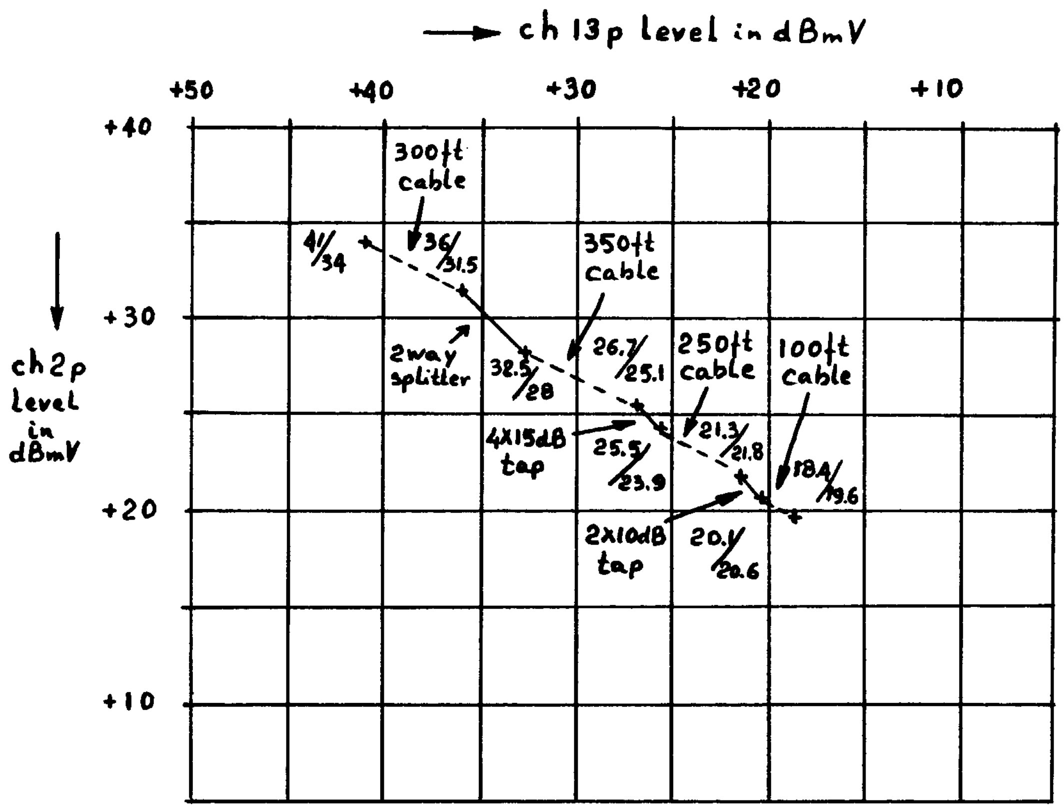

Fig.6

The graph can contain more information. Figure 7 shows the 2 way 10dB tap again; the horizontal dimension of that square represents the through loss at ch 13p; the vertical dimension represents the through loss at ch 2p.

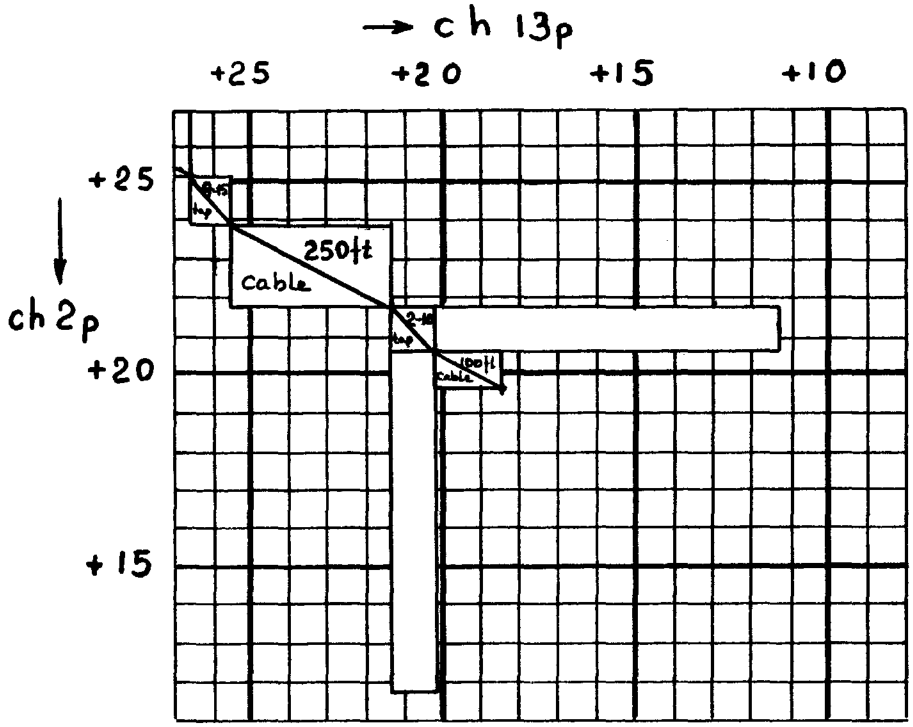

Fig.7

By changing the square to an L shape, through the addition of two rectangles as drawn, the tap off levels as well as the output levels can be read off the graph.
This is achieved by making the horizontal leg equal to the tap off loss at ch 13 and by making the length of the vertical leg equal to the tap off loss at ch 2.
A plastic module in the shape of an L can therefore be placed on the graph (see fig.8), indicating input, output, and tap off levels. Other L shaped modules representing other taps, dir. couplers and cables can also be placed on the graph and by sliding them together the signal flow and levels can be observed.

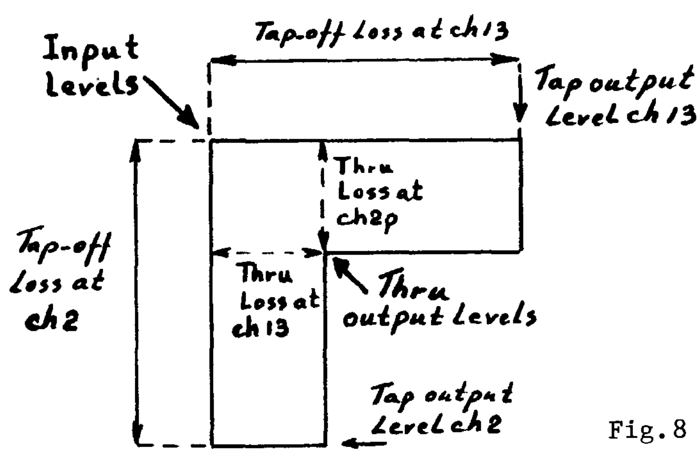

Fig.8

Once the required minimum tap off levels are established for the typical case (e.g. 100 feet of drop cable), it is possible to mark the area of levels on the graph where a 10dB device would be required, where a 15dB is required, etc. (see fig.9).

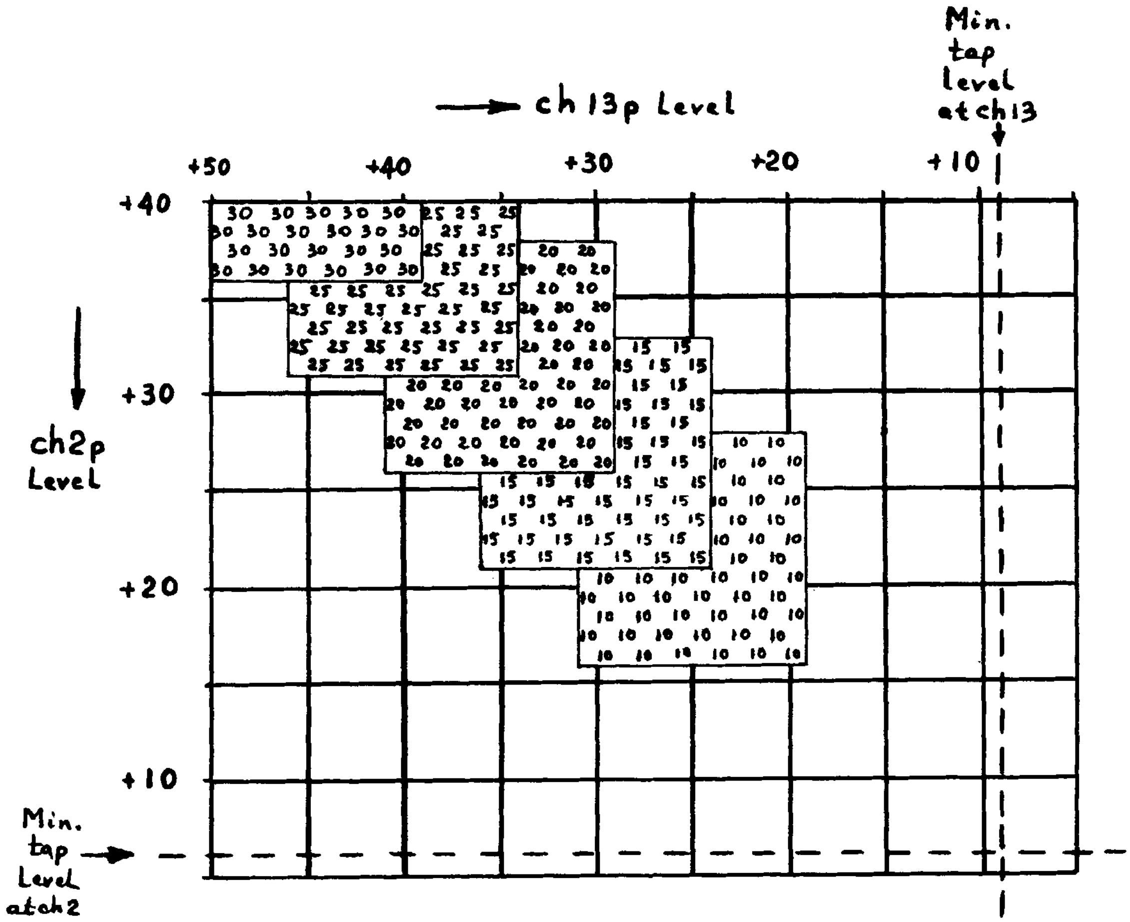

Fig.9

Since operators use different types of drop-cable and have different ideas as to what the tap levels should be, the area information and minimum required tap levels is printed on a transparent overlay and can be shifted to any operator's heart's content.

We have known for years that the tilt between ch 2 and ch 13 changes along the line between amplifiers and therefore a number of companies have marketed sloped taps.
However, it becomes difficult when designing in the conventional way, to decide if a sloped tap is desirable in a particular location and which frequency to choose first in order to determine the tap-off value. This problem can be solved because the overlay indicates whether a sloped tap or flat tap should be used and shows the value of the tap.(see fig.10)

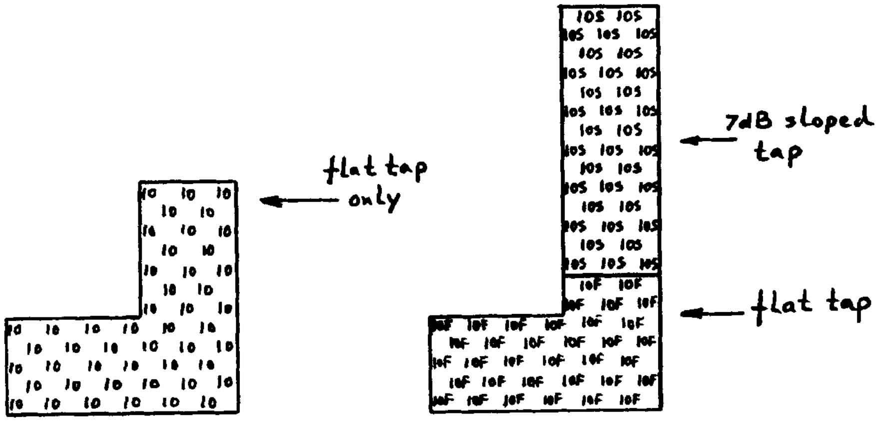

Fig.10

Yet there is more in this graph we can use. How does one determine which plug-in equalizer and, or pads to use in an amplifier when the input and output levels are known? Well, it takes some arithmetic. I have seen one techinician who carried a number of sheets around with all the tabulated data for two types of distribution amplifiers. It showed him what to do in regard to plug-in pads, switchable attenuators, and tilt and gain control for every conceivable input signal combination. Those days will soon be gone. Let us go through a little arithmetic again. Assume the desired output level is 40/36. A typical amplifier with all controls set for maximum gain provides 22dB at ch 13 and 20dB at ch 2. Minimum permissible input level therefore equals 18/16.(see fig.11)

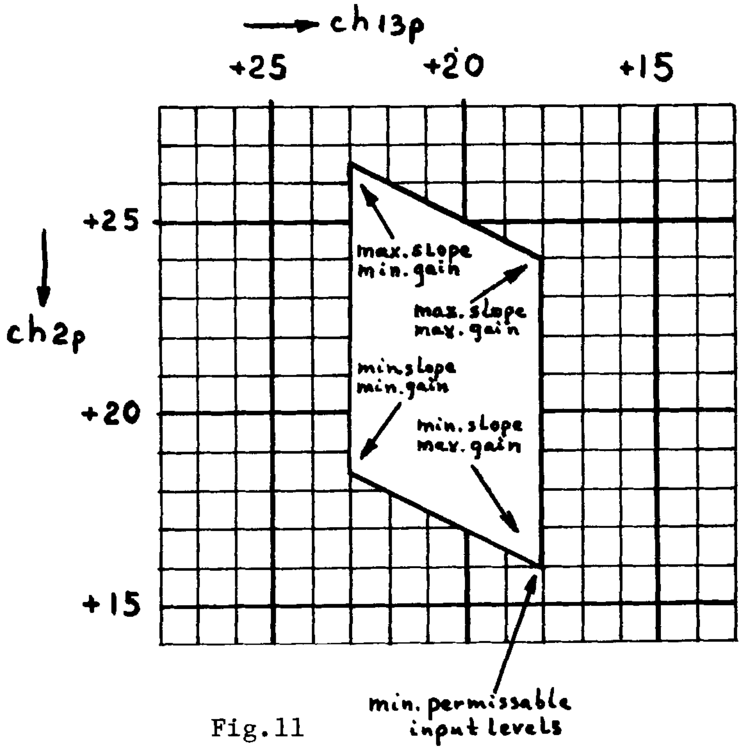

Fig.11

The internal slope control can reduce the gain at Ch 2 by 8dB. Therefore any point on the vertical line between 18/16 and 18/24 can be amplified to the desired output 40/36, by proper adjustment of the slope control. Similarly the gain control line between 18/16 and 23/18.5 indicates the levels which can be accomodated with the gain control.
Therefore any point within the drawn parallelogram can be amplified to the desired output levels
The shape of the usable area characterizes the behaviour of the controls, while the size indicates the control range. Fig.12 shows 4 possible amplifier characteristics.

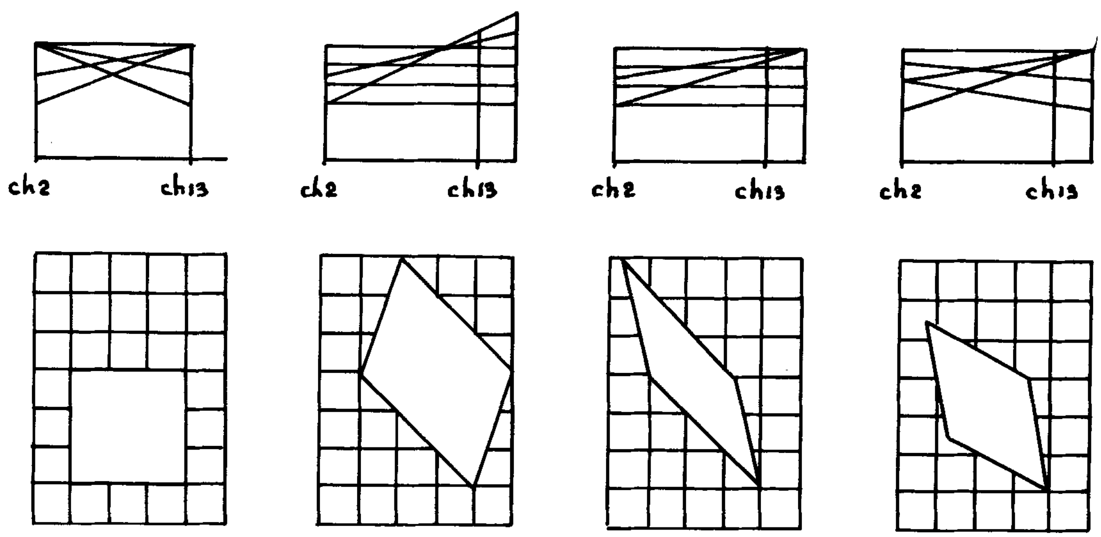

Fig.12

The addition of a switchable attenuator or "plug in" pads, results in a combination of parallelograms with overlapping areas. (see fig 13a) This permits the manufacturer to indicate a preference of one switch setting over another in these overlapping areas, resulting in one of the possible alternatives as shown in fig.13 b,c and d.

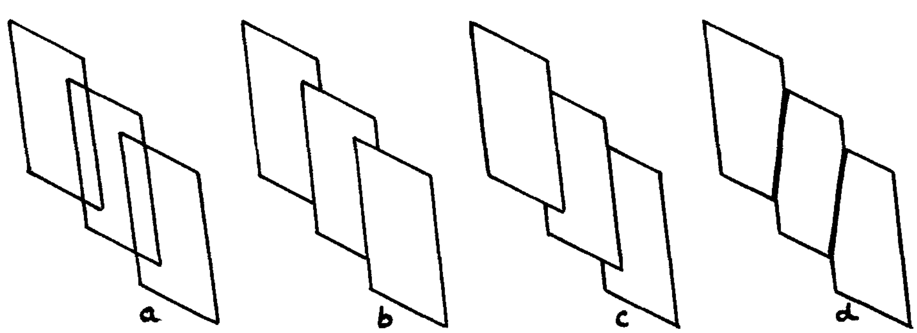

Fig.13

Similarly, switchable equalizers result in the effect shown in fig. 14b, while 14c shows the effect of the availability of plug-in equalizers and pads.
Here too it would be advantageous to know which setting to choose for optimum performance, especially where some input signals can be accomodated with 4 different combination "plug-ins".

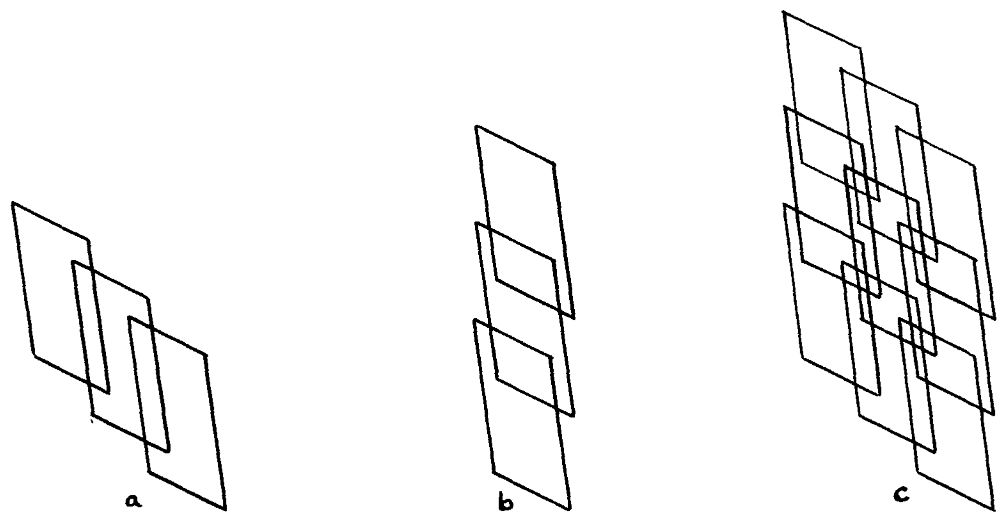

Fig.14

CONCLUSION.

As can be readily seen, there is an enormous difference in the cost and complexity of a panel constructed with RF or DC simulation techniques as compared to the graph and module method. The Cable TV calculator furthermore is portable and meets all the goals set out earlier.
It does not require leased telephone lines, terminal rental or expensive computor programs, it is completely self-contained.
The calculator has given the manual planners a new lease on life by increasing their efficiency by an estimated 50%-100%.

CALIBRATION OF CROSS MODULATION MEASUREMENTS

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Cross modulation measurement is an especially important parameter in the determination of CATV system performance. In the industry today several cross modulation test sets are available to fulfill the need to measure cross modulation. However, for accurate measurements these test sets must be calibrated in an accurate and repeatable manner. This calibration is accomplished by the insertion of a known cross modulation level, which is then used as a standard for comparison.

Since detectors used in these signal measurements are basically non-linear devices (which can be approximated with a square law expansion), it becomes essential to have the calibration point near the region of expected cross modulation readings.

When one talks about percent modulation in terms of dB and spectral component levels in dB corresponding to a given cross modulation level, there has usually been a considerable amount of confusion over these terms. Let us attempt to clarify these terms by a look at the facts. See Figure 1. As illustrated by Figure 1, the difference between the NCTA definition and conventional definition of an amplitude modulated wave lies in the way the modulation excursion on the carrier is defined. By equating the two definitions and solving mathematically (see proof in Appendix A) we then can see that M_t (NCTA modulation) is equal to M_c (conventional modulation) plus a correction factor T. See Figure 2. As shown in Figure 2, the factor T is equal to 0 dB at 100% modulation and 6 dB at 0.1% modulation.

Let us first consider several modulation levels using the conventional modulation definition. For a 100% square wave modulated carrier the first spectral component away from the carrier is 3.9 dB below the carrier. See Figure 3. For a 100% CW (sinusoidal) amplitude-modulated wave the first (and only) spectral components are 6 dB below the carrier. It is then apparent that a 0.1% square wave modulated signal has a first spectral component 63.9 dB below the carrier and a 0.1% CW amplitude modulated wave has its first (and only) spectral component 66 dB below the carrier level. See Figure 4.

Now considering the NCTA defined modulation, a 0.1% modulated signal has its first spectral component 69.9 dB below the carrier. It should be noted here that -60 dB cross modulation means that there is 0.1% induced modulation; not that the first spectral component is down 60 dB.

To produce an accurate calibration point, it is necessary to calibrate at very low cross modulation levels (at least 60 dB below the carrier). At this point then one must decide whether to use square wave modulation or amplitude modulation. Since it is difficult to accurately fix a modulation index, the choice of modulation then must be one which is accurate and repeatable.

One way of accurately simulating a low modulation index is to linearly add and envelope detect two CW signals that are closely spaced in frequency. If one of the signals is sufficiently small when compared to the other, then the detected output is the same as that of a conventional amplitude modulated signal having a modulation index equal to the voltage ratios of the two signals.

Consider the addition of two sinusoids of different and arbitrary relative amplitude M:

$$\mathrm{SIN}\ X + M\ \mathrm{SIN}\ Y = P\ \mathrm{SIN}\ \phi$$

where $P^2 = 1 + M^2 + 2\ M\ \mathrm{COS}\ (X - Y)$

and $\mathrm{TAN}\ \phi = \dfrac{\mathrm{SIN}\ X + M\ \mathrm{SIN}\ Y}{\mathrm{COS}\ X + M\ \mathrm{COS}\ Y}$

The derivation of these relations is shown in Appendix B.

If M = 1, the Sin X + M Sin Y reduces to 2 Cos ½ (X - Y) Sin ½ (X - Y), which is not conventional amplitude modulation. However, under the conditions stated before; that is, a large relative amplitude difference ($M \ll 1$) and close frequency spacing ($X \simeq Y$), then ϕ is approximately equal to X. Expanding the square root of $1 + M^2 + 2\ M\ \mathrm{Cos}\ (X - Y)$ in a Taylor series about 1 yields

$$P = 1 + M\ \mathrm{COS}\ (X - Y) + \text{Higher Order Terms.}$$

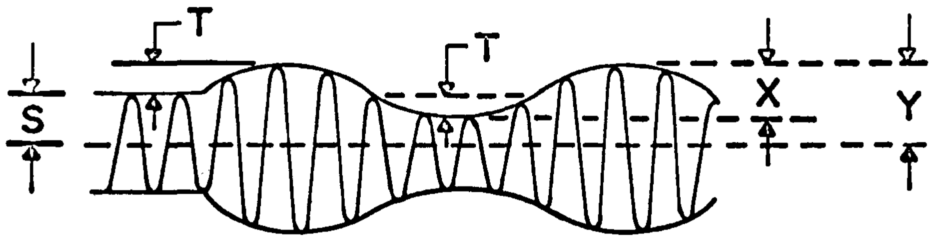
Therefore, for these conditions

$$\mathrm{SIN}\ X + M\ \mathrm{SIN}\ Y \simeq \left[1 + M\ \mathrm{COS}\ (X - Y)\right]\ \mathrm{SIN}\ X$$

This does not mean that new spectral components have been added by the linear addition of the two sinusoids; it simply means that this linear addition under the above conditions represents a conventional AM signal to a very good approximation.

Since we use this simulated AM signal to calibrate the cross modulation test set, we must include the 6 dB correction factor as shown in Figure 2. The simulated signal must be at the -66 dB level which corresponds with a -60 dB square wave (NCTA) modulated signal. One more correction factor must be considered; that of the audio analyzer used to read the detected signal. Since the audio analyzer is a narrow band device, one must consider the difference between the first side band or spectral component of the amplitude modulated and the square wave modulated signal. Expanding both signals in a Fourier series and comparing the first two spectral components reveals that the square wave modulated signal is 2.1 dB greater than an AM signal for the same modulation index as was previously shown in Figure 3. Therefore, +2.1 dB must be added to the -66 dB calibrating wave. Consequently, a -63.9 dB amplitude modulated signal is equal to a -60 dB square wave (NCTA) modulated signal. Again, looking only at the first spectral components we can see the relationship between the NCTA, conventional, and simulated modulation levels. See Figure 5 for a comparison of these levels.

C-COR has related this theory of calibration to actual operation of a unit which C-COR has built and proven in use for calibration of cross modulation test sets. A block diagram of the C-COR calibrator is shown in Figure 6. Its operation follows the theory described to provide a -60 dB cross modulation calibration point. The calibrator employs two crystal controlled oscillators with a difference frequency of 15.75 kHz. The outputs of both oscillators are fed to directional couplers one leg of which is used for a test point. A calibration on one oscillator is used to equalize the amplitude of that oscillator's output to the second oscillator's output while monitoring both outputs at the test points. The first oscillator's output then is fed to a precision 63.9 dB attenuator (which has been calibrated using a secondary standard). The attenuator output is then fed into a combiner along with the output of the second oscillator. At the combiner output then the signals provide a simulated -60 dB cross modulation calibration reference. This calibrator has proven itself as an instrument for consistent accuracy and repeatability in calibrating cross modulation test sets.

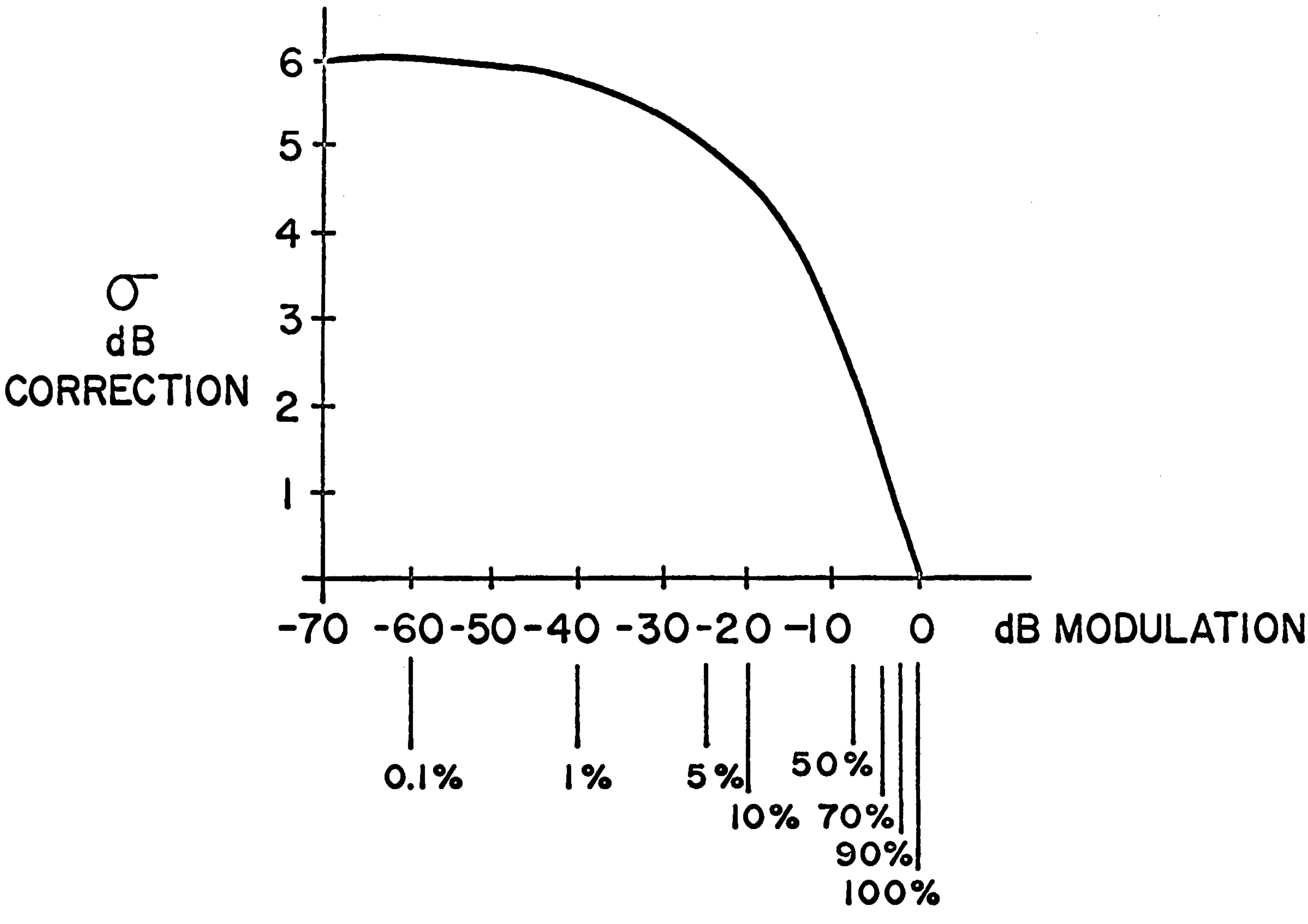


DEFINITION: CONVENTIONAL MODULATION $Mc = \frac{T}{S}$

DEFINITION: TELEVISION OR NCTA MODULATION $Mt = \frac{X}{Y}$

PERCENT MODULATION = $M \cdot 100$

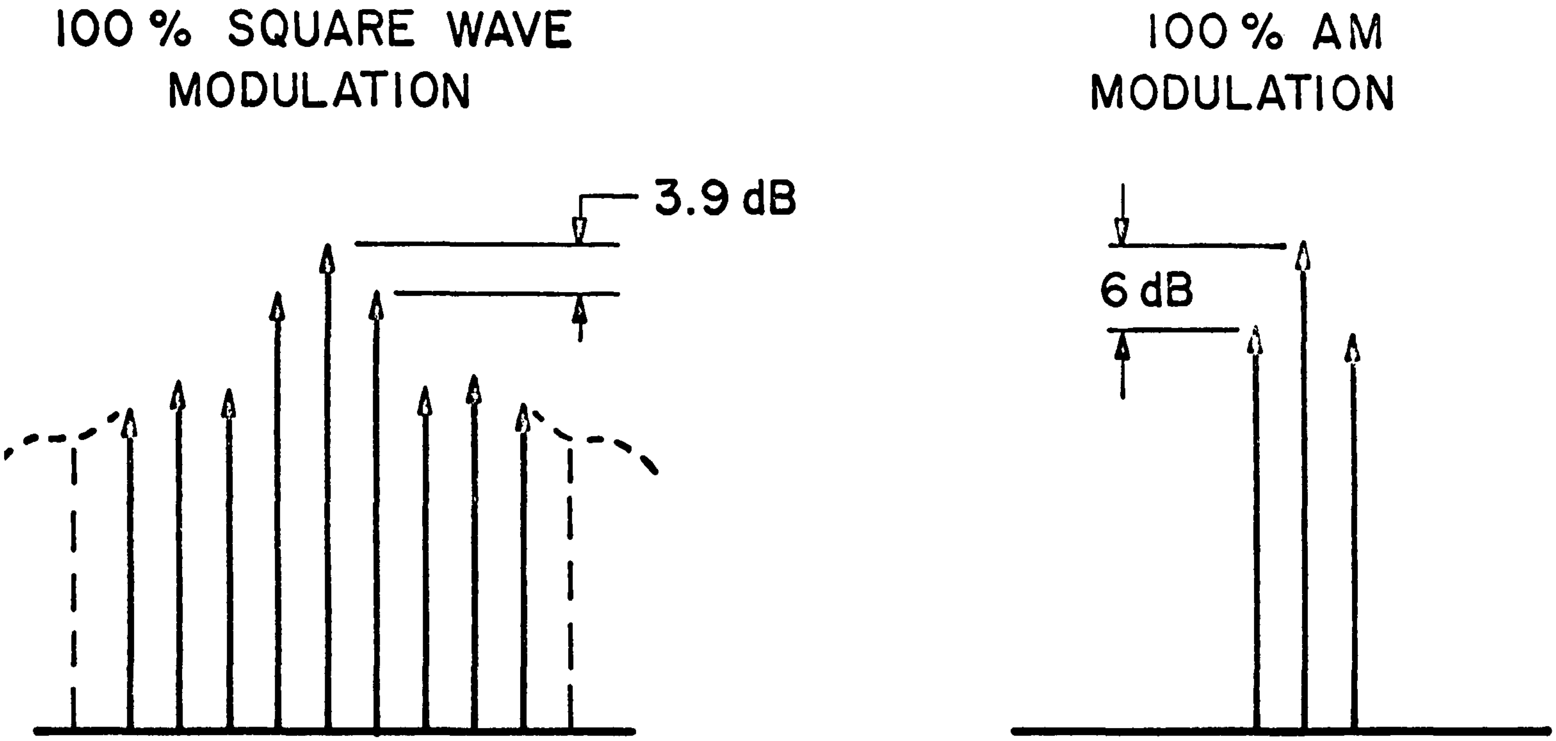
FIGURE 1



MODULATION LEVEL (NCTA)

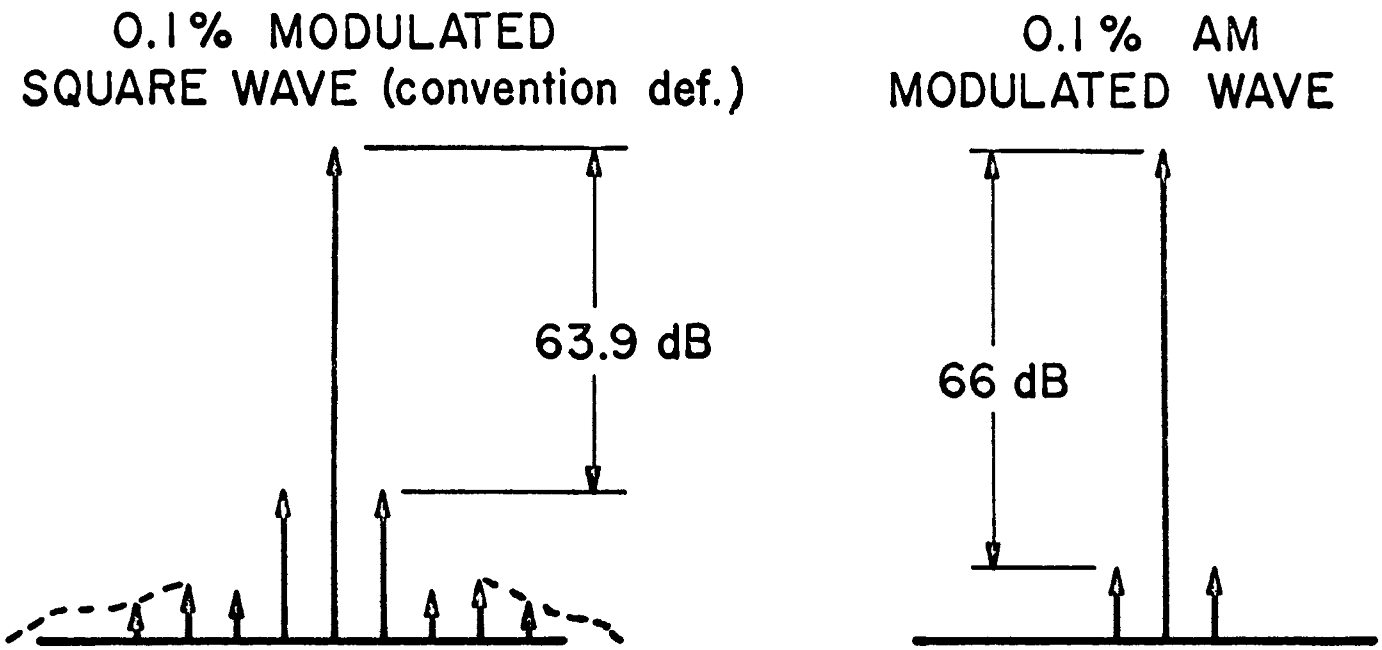
VERSES σ

FIGURE 2



SPECTRAL LINES OF A
CARRIER 100 % MODULATED
WITH A 15.75 KHz SIGNAL

FIGURE 3

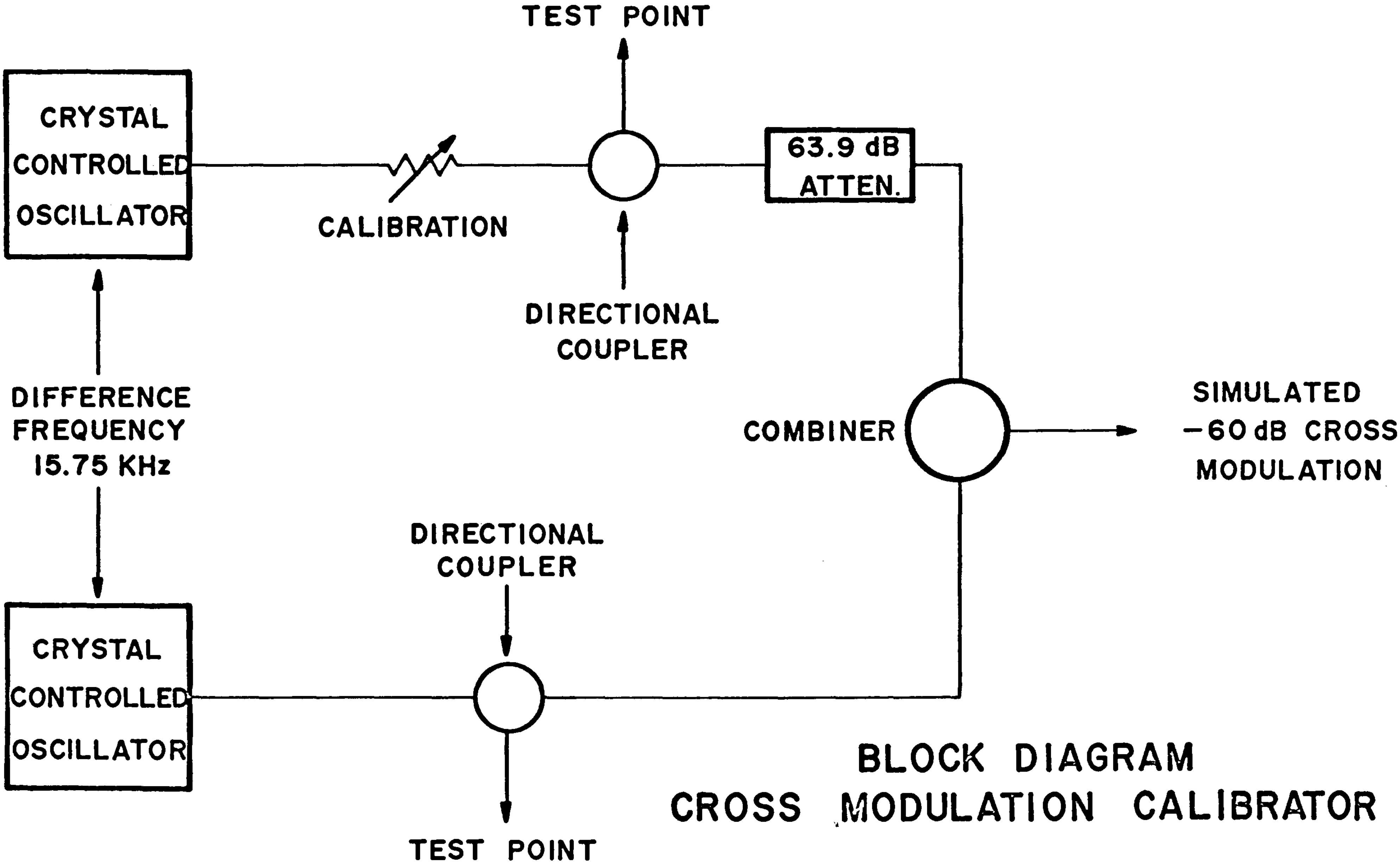


SPECTRAL LINES OF A
0.1% MODULATED SIGNAL

FIGURE 4

FIRST SPECTRAL COMPONENT LEVEL FROM CARRIER	
0.1% CONVENTIONAL PULSE MODULATION	− 63.9 dB
0.1% NCTA PULSE MODULATION	− 69.9 dB (−60 dB CROSS MOD)
0.1% CW CONVENTIONAL AM	− 66.0 dB
$\text{SINE X} + \underbrace{0.001}_{\text{60 dB ATTENUATION}} \text{SINE Y}$	− 66.0 dB (SIMULATED)
$\text{SINE X} + \underbrace{\text{M}}_{\text{CORRESPONDING TO 63.9 dB ATTENUATION}} \text{SINE Y}$	− 69.9 dB (SIMULATED)

FIGURE 5



BLOCK DIAGRAM
CROSS MODULATION CALIBRATOR

FIGURE 6

APPENDIX A

Conventional Modulation - $M_c = T/S$ See Figure 1

NCTA Modulation - $M_t = \frac{X}{Y}$

M - Modulation Level = 20 log m

Percent Modulation = 100 x m

Equate the modulation formulas by reference to Figure.

$X = 2T = Y M_t$ $\qquad T = \frac{X}{2} = S M_c$

$Y = S + T = X/M_t$ $\qquad S = Y - T = T/M_c$

$$M_t = \frac{X}{Y} = \frac{2T}{S + T}$$

$$M_t = \frac{2 S M_c}{S + S M_c}$$

By factoring and cancelling: $M_t = \frac{2\ M_c}{1 + M_c}$

Likewise: $M_c = T/S = \frac{X/2}{Y - T} = \frac{X/2}{Y - X/2}$

$$M_c = \frac{\frac{X}{2}}{\frac{2Y - X}{2}} = \frac{X}{2Y - X}$$

Again by substitution

$$M_c = \frac{Y M_t}{2Y - Y M_t}$$

Therefore $M_c = \frac{M_t}{2 - M_t}$

Taking the expression for M_t and expressing in dB yields

$$20 \log M_t = 20 \log 2 + 20 \log M_c - 20 \log (1 + M_c)$$

$$20 \log 2 = 20 (.3) = 6$$

Then let $T = 6 - 20 \log (1 + M_c)$

for which the original equation then becomes

$$\underbrace{20 \log M_t} = \underbrace{20 \log M_c} + T$$

$$M_t = M_c = T$$

APPENDIX B

$$\text{SIN } x + \text{M SIN Y} = \text{P SIN } \phi \qquad (1)$$

Using trigonometric relationships and considering the addition of two phasors which then becomes

$$e^{jX} + Me^{jY} = Pe^{j\phi}$$

Since this holds identically then,

$$\text{COS X} + \text{M COS Y} = \text{P COS } \phi$$

Thus $\text{TAN}\phi = \dfrac{\text{SIN } \phi}{\text{COS } \phi} = \dfrac{\text{SIN X + M SIN Y}}{\text{COS X + M COS Y}}$

Then $P^2 = (e^{jX} + Me^{jY})\ (e^{-jX} + Me^{-jY})$

which yields $P^2 = 1 + M^2 + 2M \text{ COS } (X - Y)$

Equation (1) now becomes

$$\text{SIN X} + \text{M SIN Y} = \left[1 + M^2 + 2M \text{ COS } (X - Y)\right]^{1/2} \text{SIN } \phi$$

* For M 1

$$\text{SIN X} + \text{M SIN Y} = \left[1 + \frac{M^2 + 2M \text{ COS } (X - Y)}{2}\right] \text{SIN } \phi$$

$$+ \left[1 + M \text{ COS } (X - Y)\right] \text{SIN } \phi + \frac{M^2}{2} \text{ SIN } \phi$$

Which equation is like conventional amplitude modulation where M has the same meaning as in amplitude modulation.

If M = 1 $\quad P^2 = 2 + 2 \text{ COS } (X - Y)$

$$\text{or } P = \sqrt{2}\left[1 + \text{COS } (X - Y)\right]^{1/2}$$

$$= 2 \text{ COS } \tfrac{1}{2} (X - Y)$$

and when applied to equation (1), the equation becomes

$$\text{SIN X} + \text{SIN Y} = 2 \text{ COS } \tfrac{1}{2} (X - Y) \text{ SIN } \tfrac{1}{2} (X - Y)$$

Since $\text{TAN } \phi = \dfrac{\text{SIN X + SIN Y}}{\text{COS X + COS Y}} = \text{TAN } \tfrac{1}{2} (X - Y)$

it follows that

$$\phi = \frac{X + Y}{2}$$

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NCTA-002-0267

CAN COAXIAL CABLE COPE WITH THE CATV SYSTEMS OF THE 70'S?

John C. Fan, Mgr., Research & Development Copper Clad Aluminum Wire
Frank A. Spexarth, Mgr., Communication Industry Products

Introduction

At a time when the cable television industry is absorbing such sophisticated new technologies as laser link and microwave transmission, it may seem unusual to suggest taking another look at a tool as familiar and comfortable as the coaxial cable. Unusual, that is, until you consider the demands likely to be made of this cable during the remainder of the 1970's.

Most observers predict that penetration of cable television into the top 100 cities during this decade will stimulate dramatic changes in both your services and systems. There is talk of 24, 48, 64, or more channels. You will be called upon to transmit not only video signals, but also data and facsmile reproductions. Two-way communications is likely. Average system size will jump from approximately 100 miles of cable per city to as much as 4,000 miles per city.

Such growth places heavy demands on the technical resources available to you. Utilization of laser link and microwave technology is crucial to providing economical service to large metropolitan areas. Yet even such advanced equipment will not eliminate the need to wire with coaxial cable in any system of the foreseeable future. In fact, it is the basic system you know today which will be called upon to provide the many new services upon which your industry will grow. There is little doubt that more sophisticated electronics hardware will become available as your industry requires it. The question is, can your cable keep pace?

The nature of a cable television system makes this a crucial question because noise-to-signal limitations are principally a function of the cable. The most effective way to improve system efficiency is to use cable with lower signal loss rather than to depend on high gain amplifiers which unavoidably are limited by attenuation characteristics of the system's cable. Thus, as the state of the art in active equipment improves, the electrical properties of the coaxial cable become even more significant to the system owner. This paper examines factors which must be considered in producing cable with low signal loss characteristics.

Attenuation

The coaxial cable is itself a system, a combination of materials fabricated together to transmit RF signals efficiently. Basically it is made of two metal conductors separated by an insulating (dielectric) material. (Figure 1) Attenuation, the loss of signal strength from one end of a cable to the other, is unavoidable and a function of eight variables: frequency, impedance, RF resistivity of the outer conductor, diameter of the outer conductor, RF resistivity of the center conductor, diameter of the center conductor, the dielectric constant and the power factor of the dielectric. (Figure 2) Only a few of these variables, however, can be altered in CATV cable to reduce attenuation.

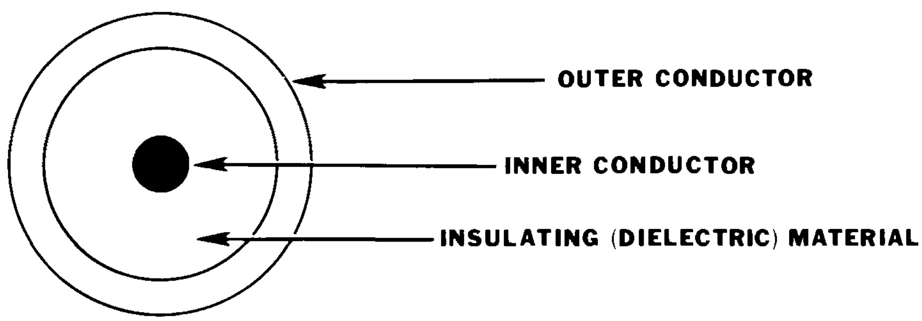

FIGURE 1

The basic attenuation formula:

$$\alpha = \underbrace{0.660 \times 10^{-3} \frac{\sqrt{\varepsilon f}}{Z_0} \left(\frac{\sqrt{R_1}}{d} + \frac{\sqrt{R_2}}{D} \right)}_{\text{Conductor Loss}} + \underbrace{2.77 \times 10^{-6} \, f \sqrt{\varepsilon} \, \tan \delta \, \frac{\text{dB}}{100 \text{ ft}}}_{\text{Dielectric Loss}}$$

FIGURE 2

Frequency (f)

Frequency currently is restricted to the 54 to 212 mhz band by the capability of the television receiver. However, plans are being made to employ the 5 to 300 mhz band to provide other capabilities such as 2-way communications.

Impedance (Zo)

There are two widely accepted cable impedances in use today, 75 ohm and 50 ohm. The two were developed to fulfill distinct needs: transmission with minimum attenuation and transmission of maximum voltage.

In designing coaxial cable for minimum attenuation, the optimum ratio of conductor diameters is determined from the attenuation equation as derived in detail by Dummer and Blackland . (Figure 3)

The conductor loss:

$$\alpha_c = 0.660 \times 10^{-3} \frac{\sqrt{\varepsilon_f}}{Z_o} \left(\frac{\sqrt{R_1}}{d} + \frac{\sqrt{R_2}}{D} \right)$$

For minimun α_c with air dielectric:

$Z_o = 75$ ohm with $\frac{D}{d} = 3.5$

FIGURE 3

In designing coaxial cable for maximum voltage rating (required, for instance, in microwave radar applications) the equation is as shown in Figure 4.

The peak voltage:

$$V = \tfrac{1}{2} E\, d \log_e \frac{D}{d}$$

where E is the voltage gradient with a cable spacing.

To achieve maximum V:

$\frac{D}{d} = 2.72$ thus $Z_o = 50$ ohm

FIGURE 4

Since for cable television transmission the principal concern is low attenuation, 75 ohm has been selected as the industry standard.

Outer conductor resistivity (R_2)

In CATV, aluminum is used almost exclusively for outer conductors since it provides the most suitable technical/economic selection of materials available. Outer conductor resistivity, therefore, can be considered a constant.

Outer conductor diameter (D)

The size of outer conductors has been fixed at various standard diameters to allow flexibility and interchangeability for connections and equipment hook up. This, too, may be considered a constant.

Dielectric (ε and tan δ)

Efficiency of the coaxial cable dielectric can, and does, change since the dielectric effect can be reduced by lowering the density of the insulating foam to approach an electrically optimum air dielectric. Foamed polyethylene, and more recently foamed polystyrene are the most common dielectric materials.

Center conductor diameter (d)

When lower loss dielectrics are used and the outer conductor size is not changed, as in the case of CATV coaxial cable, then the diameter of the center conductor must increase to improve performance.

Center conductor resistivity (R_1)

Copper clad aluminum and copper are the standard materials in use today. Due to the inherent characteristics of RF signals which are transmitted only on a conductor's surface, the RF resistivity of copper clad aluminum and solid copper are identical.

Cable Engineering

With the need for larger systems, improved cascadability and reduced system cost, cable designers are turning to dielectric materials having lower loss and lower density. Inherent in these changes however, is a reduction in the supporting capability of the dielectric. This creates a problem because there is a simultaneous need for larger--and therefore heavier--center conductors.

The solution to the dichotomy is found in copper clad aluminum wire--a material already used widely by your industry in current cable designs. Two inherent physical properties of copper clad aluminum provide the basis for this solution.

Weight

Because copper clad aluminum is 40% lighter than copper it effectively reduces the tendency of a center conductor to "drift" in the dielectric during its service life, thus preventing impedence discontinuities and shorts.

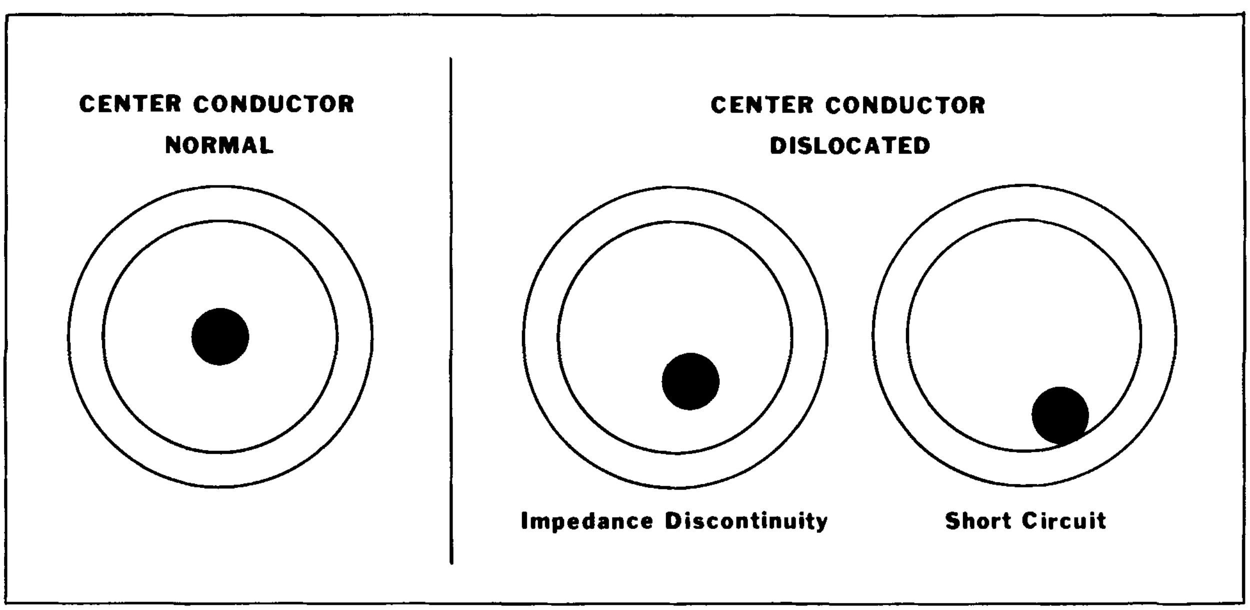

FIGURE 5

Bendability

In the installation of a CATV system, coaxial cable has to be formed into many bends. The bend may cause some impedance distortion if the center conductor is dislocated in an eccentric position within the cable. This can be caused by the resistance of the conductor to the bending forces.

Results of bending force tests verify that copper clad aluminum requires less force to bend or form into different shapes. It therefore reduces the chance of impedance discontinuity due to migration of the center conductor into the dielectric at the bend. The tests were performed with the bending fixture (Figure 6) mounted directly on a tensile testing machine to measure bending forces accurately. The comparison of the annealed copper conductor to annealed copper clad aluminum conductor is shown in Figure 7.

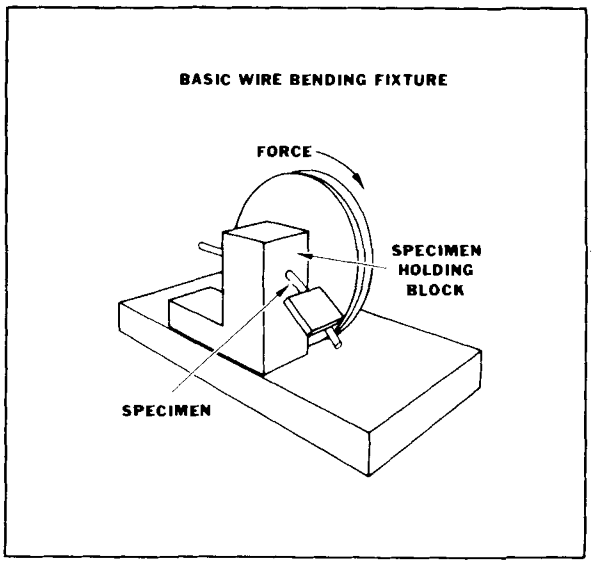

FIGURE 6

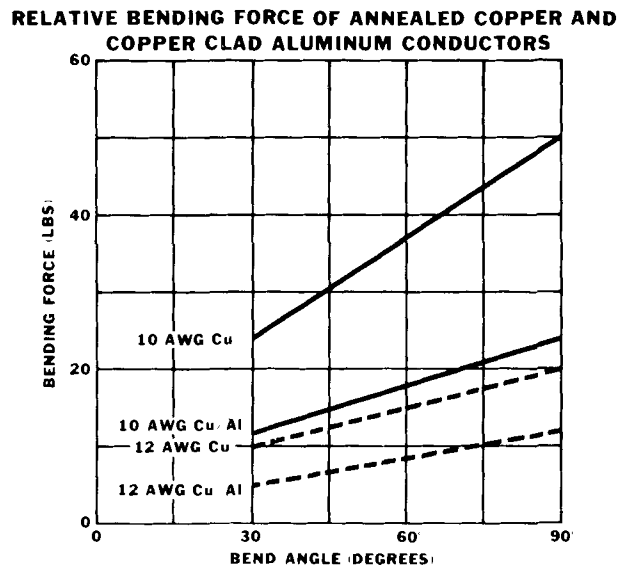

FIGURE 7

Frequency

With the predictable expansion of the usable frequency range to 5 to 300 megahertz in use in the '70's, the question comes to mind as to whether or not there will be any change in the attenuation properties of copper clad aluminum, especially in the lower frequency range. With the help of cable manufacturers, data was accumulated measuring the attenuation of copper versus copper clad aluminum cables using a foamed polyethylene dielectric. The results are shown in figures 8, 9, and 10. Note that copper clad aluminum in the 750 cable is equal to copper at the 500 khz frequency and shows only a loss of 10% attenuation as low as 50 khz. In the 500 and 412 cable with copper clad aluminum the attenuation is equal at 1 mhz and shows a 10% loss at 100 khz compared to copper. With the application of lower density, lower loss dielectrics, then, copper clad aluminum center conductors can be utilized for lower frequencies as the industry requires it. Copper offers no attenuation advantages over copper clad aluminum based on either current or expected demands in frequency range.

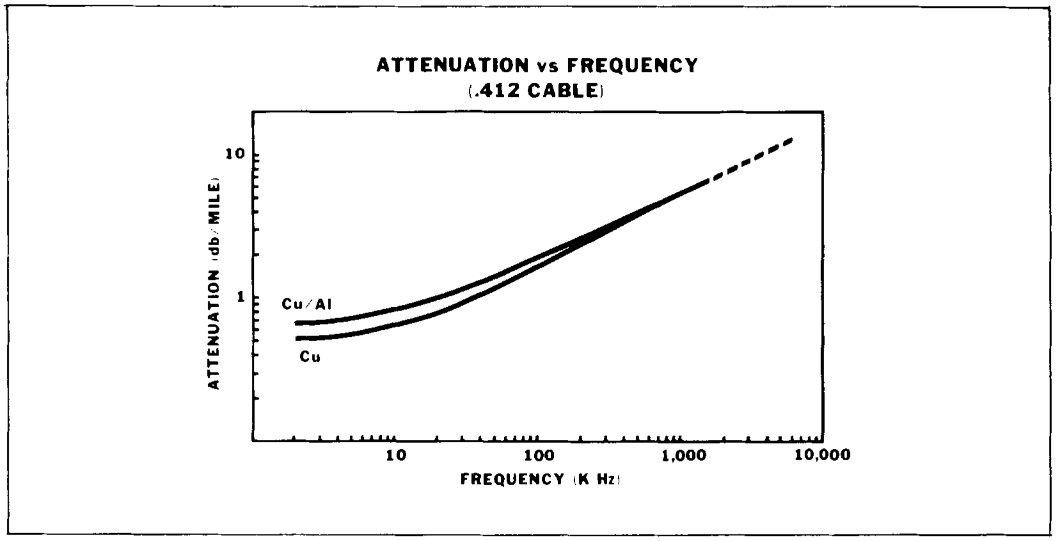

FIGURE 8

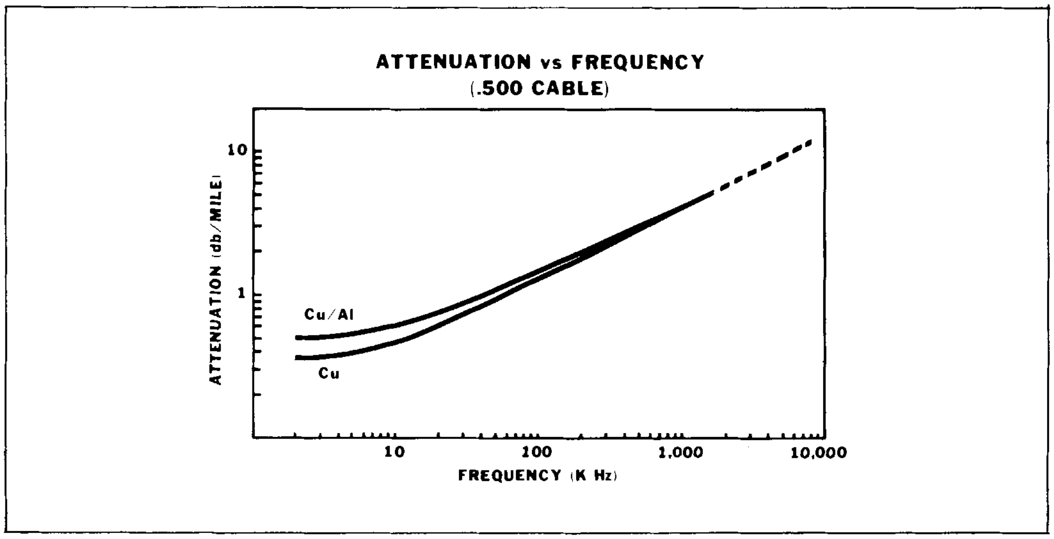

FIGURE 9

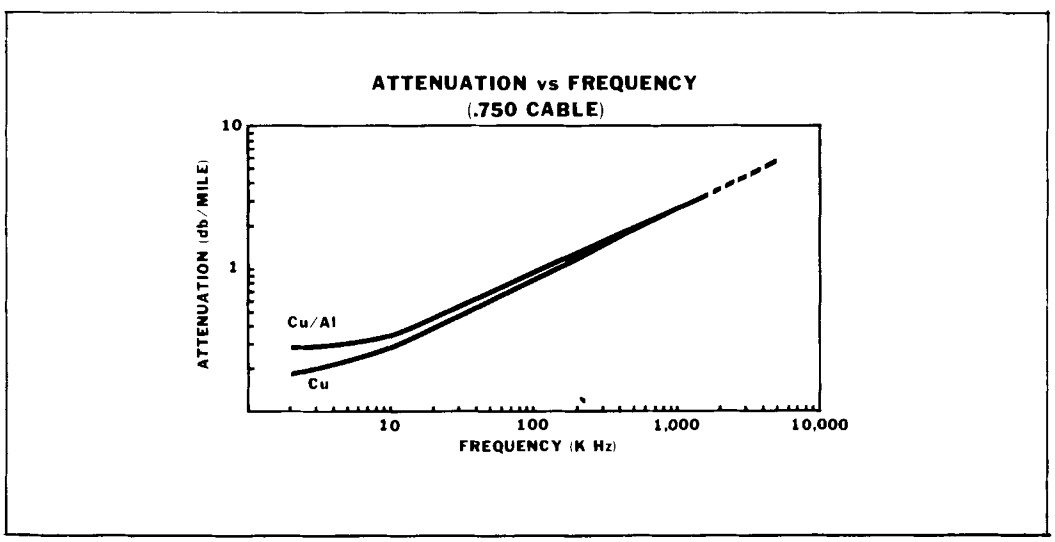

FIGURE 10

Power

It is a popular assumption that the 40% increase in loop resistance due to the use of copper aluminum center conductor requires a 20% increase in power supply. This is a misleading assumption and has caused some misunderstanding regarding the use of copper clad aluminum. Actually, experience indicates that over 80% of existing CATV systems using solid copper conductor cables could make direct substitution of copper clad aluminum center conductor cable without any additional power supplies. This is because:

1. The main consideration in the design of a system is the quality of the RF signal. This dictates that the layout be made of the amplifiers and the cable to the most suitable RF situation.

2. Cable powering is normally considered after this basic layout stage. The location of a power supply, therefore, is extremely flexible.

3. The use of marginal voltage is seldom seen in actual CATV systems. There appears to be sufficient voltage remaining for adding copper clad aluminum without adding power supplies.

4. Parallel powering is the common method used instead of series powering.

5. Geography seldom permits optimum power supply locations. The system must accommodate itself to existing streets and buildings.

In addition, many systems now under design, and most systems of the future, will use at least a 60 volt powering method. This virtually eliminates the question of loop resistance or marginal voltage. The economics in the system hardware, including copper clad aluminum coaxial cables, together with more efficient power usage, should be most attractive to the CATV system today and in the future.

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CASCADING OF INTER-MODULATION DISTORTION IN CABLE TELEVISION SYSTEMS

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INTRODUCTION

Expansion in channel carriage of cable TV systems so that channel allocations appear either between TV channels 6 and 7, or above channel 13, has resulted in new distortion requirements for cable TV systems and amplifiers.

These distortion requirements define the permissible second order distortion in the system and in each amplifier. This is an additional requirement to that of third order distortion. For the latter case, it is already possible to have third order inter-modulation "beats" occur as interfering signals in the standard TV channels. Third order distortion specifications have always been maintained in Cable TV systems rather vigorously to prevent this occurrence. This unique quality of low third order distortion was generated from the specification limitation on cross-modulation distortion; the latter distortion occurring from third order non-linearities. The maintenance of the magnitude of third order interferences to acceptable limits resulted as a by-product of the cross-modulation specification, since if the cross-modulation distortion for 2 channels was down a prescribed magnitude from 100% modulation, then theoretically the triple beat interfering carrier had to also be down from the desired carrier by a related prescribed magnitude. This comparison of the magnitude of third order inter-modulation products and the cross-modulation product has been analyzed and shown by Lotsch[1] and Simons[2]. Before the expansion of cable systems to more than 12 channels, little work had been done in ascertaining or maintaining the magnitude of second order inter-modulation products in Cable TV amplifiers, since the TV channels were originally spaced to preclude second order interfering products from developing. With expansion in the number of channels, attempts were made to use the amplifiers which had exhibited low third order distortion for those new systems in which requirements for second order distortion were severe. The amplifiers performed with varying success, since they had not been originally designed to proper second order distortion specifications.

This then lead to a new generation of amplifiers with improved second order specifications and with new techniques for minimizing the build-up of second order distortions in the CATV trunk cascades.

This paper will examine certain aspects concerning the increase of second order and third order inter-modulation products in CATV amplifier cascades.

A general theory of the manner in which second order distortion cascades in the system will be examined. It will be shown that the magnitude of the cascade effect depends upon the low frequency phase intercept of the phase shift vs. frequency curve of the amplifier. Calculations for several values of the phase intercept will be done to demonstrate the dependence of the second order cascading effect upon the phase intercept. Results of inter-modulation distortion products cascading in CATV systems will be shown as an indication of the manner in which the second order distortion practically cascades, and also as an experimental validation of the theoretical work.

Conclusions will show that second order distortion requirements for Cable TV trunk amplifiers can be fulfilled by state-of-the-art amplifiers, and that amplifiers which operate at higher levels (such as distribution and line extender amplifiers) require a similar magnitude of second distortion levels at their operating levels as those of the trunk amplifier. It will also be shown that in cascaded amplifier systems, third order distortion products cascade in magnitude more rapidly than second order distortions, and are therefore a limiting factor to the length of transmission systems. It will further be stated that the developed theory is applicable to transmission systems other than Cable TV systems.

SYSTEM SPECIFICATIONS FOR INTERFERING CARRIERS

In a Cable TV system, repeater amplifier gain together with cable attenuation is made equal to unity over the complete operating band. For systems of this type, in which the information is carried by amplitude modulation of the carriers, it has been shown that the overload-to-noise ratio of the channels decreases at the rate of 20 log n, where n is the number of amplifiers.[3] This is determined by the increase in noise by 10 log n (power addition) as the signal progresses through the amplifiers, and a required decrease in operating level by 10 log n, since the cross-modulation distortion (which increases by 2dB for each 1dB increase in signal level) increases by 20 log n (voltage addition) as the signal information is cascaded. System specifications for carrier-to-noise ratio, and allowable cross-modulation distortion, are the two most determining factors in defining amplifier appearance for achieving system objectives of Cable TV systems. In other words, repeater amplifier performance in regard to noise figure and cross-modulation distortion are determined by specific system requirements. The object in the repeater amplifier design is to achieve an amplifier which under cascaded type operation will perform in a manner so that system requirements are fulfilled. This amplifier design must achieve its performance at the most economical cost and with a required reliability factor. Over-design of performance factors which add to the manufacturing, installation or maintenance cost, or which degrades the maintainability or reliability of the system should not be considered as a positive factor in the amplifier design. Repeater amplifiers for state-of-art Cable TV systems generally exhibit specifications for noise figure of about 9-10dB and for cross-modulation of about -93dB at operating levels. This is sufficient for achieving cascade operation to system lengths of 1000dB at operating levels of 30-35dBmV. Although improved performance in these areas could be achieved by use of more expensive transistor devices and more expensive techniques (such as paralleling output transistors), the system performance objectives have dictated the amplifier performance, and amplifiers have thus been designed with suitable performance to achieve these objectives with suitable margins.

The criteria for allowable repeater amplifier inter-modulation distortion magnitude should also, therefore, be dictated by system requirements. In order to accomplish this, it is required to define the limits of inter-modulation distortion by the knowledge of the system requirements as to tolerable levels of interfering signals, and the manner in which second and third order distortions cascade in a transmission system such as that of Cable TV.

Several investigators have made subjective tests on observations of interferences in TV pictures in order to determine the tolerable level of single frequency interference in a TV picture. The results of these investigations are summarized in Figure 1. Fink considered an in-band carrier interference, such as that of co-channel interference, 55dB down from the video carrier to be a sufficient requirement for limiting interference to a non-observable level. Further work by CATV equipment manufacturers demonstrated that a value of 60dB could be a better requirement for certain worst-case type of situations. Bell Telephone Laboratories considered a worst case of -70dB for peak-to-peak signal vs. RMS interference. The Canadian BP-23 specification dictates a -57dB spec at the horizontal sync frequency and at the color sub-carrier frequency. The value of 60dB has become an acceptable value in the cable TV industry.

This single frequency interference, which is similar to co-channel interference, could result from spurious signals generated from either second or third order, or higher order inter-modulation products. For solid-state linear Class A circuits,such as those of Cable TV amplifiers, the second order distortions are predominate distortion products.

The above mentioned studies have considered only single frequency interferences. To the writer's knowledge, there have been limited and incomplete studies on multi-frequency interferences in TV pictures. Although much work is being done in examining multi-frequency interferences, published results are still sparse. This information is greatly needed since expansion of channel usage to 20 or 30 channels, in which the frequency band is from 50 to 270 MHz (present frequency limits of cable TV amplifiers) results in a multitude of interfering "beats" within each channel.

Evidently, a multitude of operating situations utilizing a multitude of interfering channels will have to be made and be subjectively studied in order to ascertain some criteria for the magnitude of disturbance which can be tolerated. Because of the indeterminate effects of multi-carrier interferences, it is not possible to fully define the tolerable magnitude of interference in a cable system cascade. If only the single frequency interference case is considered, then the figure of -60dB for video carrier to interfering carrier ratio would be an acceptable criteria.

EFFECTS OF AMPLIFIER CASCADING ON INTER-MODULATION DISTORTION

A broadband amplifier, such as the amplifiers used in Cable TV systems, has a linear phase curve within its pass-band. This is a necessary requirement for minimizing differential delay distortions which could otherwise occur. A typical amplifier phase shift curve is shown in Figure 2. An analysis of the variation of inter-modulation products in transmission systems utilizing broadband amplifiers having linear phase shift, has been made by Bell Telephone Laboratories[4]. Results are summarized as follows:

A. Second Order Distortion Components

1. The cascaded second order beat distortion signal A from Amplifier n-1 appears at the output of amplifier n as :

(1) $A_{n-1}(w_1 + w_2) = \cos [(w_1 + w_2)\, t + \theta_1 + \cancel{\theta}]$

Where $A_{n-1}(w_1 + w_2)$ is the relative magnitude of the sum frequency of carriers w_1 and w_2 , θ_1 is a fixed phase shift dependent upon the phase shift from one amplifier to the next of the carriers w_1 and w_2 , and $\cancel{\theta}$ is the extrapolated low frequency phase shift, determined by extending the phase versus frequency curve of the amplifier cable combination along its linear plot down to low frequency. This phase intercept is shown in Figure 2, which is a phase vs frequency plot of a trunk amplifier. It is the intercept at zero frequency of the tangent to the phase curve at any frequency. For an amplifier with linear phase shift, it becomes the extrapolated phase curve to zero frequency.

2. The developed distortion at the output of amplifier number n is

(2) $A_n (w_1 + w_2) = \cos [(w_1 + w_2)t + \theta_1 + 2\phi]$

where the symbols have the same meaning as previously.

B. Third Order Distortion of form $(w_1 + w_2 - w_3)$

1. The cascaded third order distortion signal from previous amplifier appears at the output of the succeeding amplifier as

(3) $A (w_1 + w_2 - w_3) = \cos [(w_1 + w_2 - w_3)t + \theta_2 + \phi]$

where

$A_{n-1} (w_1 + w_2 - w_3)$ is the relative magnitude of the triple beat product of carriers w_1, w_2 and w_3, where θ is a fixed phase shift dependent upon the phase shift from one amplifier to the next of the three carriers, and where ϕ is the zero-frequency intercept described previously.

2. The developed third order distortion at the output of amplifier n is

(4) $A_n (w_1 + w_2 - w_3) = \cos [(w_1 + w_2 - w_3)t + \theta_2 + \phi]$

where the symbols have the same meaning as described previously.

Examination of the second order distortion of equations (1) and (2) indicate that the distortions do not directly add from amplifier to amplifier unless $\phi = 0^{\circ}$. If ϕ is made equal to 180°, a direct cancellation of second order distortion in every other amplifier can occur. For other values of the phase intercept, the result is between total cancellation and total addition.

For the third order distortion case, equations (3) and (4) are shown to be equal, regardless of the value of ϕ (the zero-frequency phase intercept). Therefore, third order distortions of this type will cascade on a voltage addition basis, or as 20 log n, where n is the number of amplifiers in the cascade.

Other forms of second order distortions such as "beats" arising from the different frequencies, and second harmonics and third order distortions, such as those arising from third harmonics, and the sums of three frequencies, could also be studied from a similar analysis as that given in equations (1) through (4).

Analysis shows that the distortions can be summarized by the following:

1. All forms of second order distortions will cascade according to the formula given in equations (1) and (2).
2. Only third order products of the form $w_1 + w_2 - w_3$ and $2w_1 - w_2$ will cascade according to equations (3) and (4).
3. Other forms of third order distortion such as $w_1 + w_2 + w_3$, $3w_1$, and $2w_1 + w_2$ will cascade according to the following:

   Cascaded third order distortion signal from previous amplifiers appears at amplifier n as: (5) $A_{n-1}(w_1 + w_2 + w_3) = \cos[(w_1 + w_2 + w_3)t + \theta_3 + \emptyset]$

   The developed third order distortion at amplifier n is: (6) $A_n(w_1 + w_2 + w_3) = \cos[(w_1 + w_2 + w_3)t + \theta + 3\emptyset]$

It can be seen from equations (5) and (6) that for a specific value of $\emptyset$ that the generated distortion at amplifier n can be made out-of-phase with previously generated distortions, so that periodic cancellations of the distortion can be made to occur.

Figure 3 shows polar plots of the relative magnitude and phase of the growth of second order distortions from amplifier to amplifier as a function of the phase of the zero-frequency intercept $\emptyset$. These plots are calculated from summing the distortion contribution from amplifier n-1 and that contributed by amplifier n. The following can be noted from the polar plots:

1. For exactly a zero degree phase shift of the intercept, the distortion increases linearly with the number of amplifiers.
2. For exactly a 180° phase intercept, the distortion completely cancels every other amplifier.
3. For phase shifts between 0° and 180°, the distortion magnitude oscillates between a certain limit and the single amplifier value. The distortion magnitude and phase returns to that of the single amplifier value after the number of amplifiers in which $n = \frac{360^{\circ}}{\phi^{\circ}}$

From this figure, it becomes obvious that if a repeater amplifier and its cable combination could be designed so that the zero frequency intercept is close to 180°, then the absolute magnitude of the growth of the second order distortion would be not much greater than that from any one amplifier. This is a completely practical situation as will be shown in the succeeding paragraph.

PHASE RESPONSE OF CABLE TV AMPLIFIERS

Figure 4 is a phase vs. frequency plot of the Sylvania trunk amplifier set for flat gain. The phase shift is linear throughout the operating band of 50-270 MHz. If the linear portion of this curve is further extrapolated, as shown by the dotted line of Figure 4, it can be noted that at zero frequency the extrapolated phase shift is 180°. This is because this particular amplifier has an odd number of stages.

This unique quality of the Sylvania CATV amplifier fulfills the requirement demonstrated in the previous section, that if the repeater amplifier phase intercept is close to 180^{0}, then the second order distortion does not cascade appreciably. However, the flat aligned amplifier is not the actual operating condition of the amplifier. Phase plot in its actual operating condition is also shown in Figure 4. This is with the amplifier aligned for compensation of 23dB of cable, as measured at 270 MHz. It will be noted that the equalization introduces an additional phase shift so that the phase intercept now appears to be closer to 120^{0}. This is still a worthwhile condition, since Figure 3 indicated that a 120^{0} phase intercept results in the cascaded second order distortion still never becoming worse in absolute magnitude then that of any amplifier.

Amplifiers with phase shift type shown in Figure 4 have been used in amplifier cascades in order to test the cascade effects on second order inter-modulation distortion.

CASCADE TESTS FOR SECOND ORDER INTER-MODULATION DISTORTION

Figure 5 shows the results of measurements of a second order inter-modulation "beat" signal after each amplifier in the cascade. This particular beat signal was the sum of channel 5 and channel 6 video carriers. The "beat" frequency was 160.5 MHz. It can be ascertained from Figure 5 that no monotonic increase of the second order distortion occurred in the cascade. Partial cancellation of the distortion appeared to occur after every third amplifier.

Figure 6 is the result of another cascade using channels 4 and 13 as the carriers with the measured distortion being their difference beat. Again periodic cancellation becomes obvious.

Some rise in distortion can be noted. This is due to the fact that amplifier number 9 generated much greater distortion than the previous amplifiers. It, therefore, determined the magnitude of the distortion in the 10-amplifier cascade.

Figure 7 shows results of the cascading of a triple beat. It can be clearly seen that the triple beat cascades, according to theory, at the rate of 20 log n.

Figure 8 shows cascading of a second order distortion sum beat for amplifiers of different types of phase intercepts. For the amplifier with the zero-degree phase intercept, it is obvious that the distortion increases monotonically. With the same amplifier modified so that its phase intercept is 180^0, a periodic cancellation occurs.

CONCLUSIONS

It has been shown that second order distortion products in a transmission system can be controlled by judicious choice of the zero frequency intercept of the phase shift curve of the repeater amplifier. Although tests for validating this theory were made for a cable TV system, the theory is fully applicable to any transmission system.

Additional conclusions are:

1. Distribution and extender amplifiers have the same distortion requirements at their higher operating levels as the turnk amplifier, since they appear as additional amplifiers in the cascade. Their zero-frequency phase intercept should also be designed close to 180^0.
2. The method of achieving less than the required tolerable maximum of second order distortion for a single repeater amplifier should consider cost, reliability, maintainability, etc. Distortion reduction can be achieved by various means such as proper device

selection and circuit basing, feedback techniques, split-band techniques, proper selection of zero-frequency phase intercept, or by "push-pull" action. Results and performance and not the method should be the governing factors.

3. Certain third order distortions will cascade so as to increase monotonically. For this reason, the third order distortion becomes the predominant factor in limiting the lengths of cable TV cascades.

4. More intensive studies are required by the CATV Industry to more fully ascertain the effects of multi-frequency interferences in expanded channel systems. These studies should be coordinated with consistent criteria so that results can be more readily compared. Among some of the criteria to be defined are:

   a. Exact number of channels.
   b. Exact frequencies, tolerances, and stabilities of all channels.
   c. Levels of all channels.
   d. Effects of changing frequencies and/or levels of certain channels.
   e. Conditions of TV viewing.

   Results should then be noted on each and every channel.

Acknowledgement is given to Mr. Timothy Eller and Marty Zelenz of GTE Sylvania CATV Operations for their assistance in validating some of the analysis and in providing some of the test data.

REFERENCES

1 H.K.V. Lotsch — "Theory of Nonlinear Distortion Produced In A Semiconductor Diode"; IEEE Transactions on Electron Devices, May 1968, pp. 294-307.

2 Ken Simons — Technical Handbook for CATV Systems; Third Edition, March 1968.

3 William A. Rheinfelder — CATV System Engineering; Second Edition.

4 Bell Telephone Laboratories — Transmission Systems for Communications; pp. 224-228, Revised Third Edition.

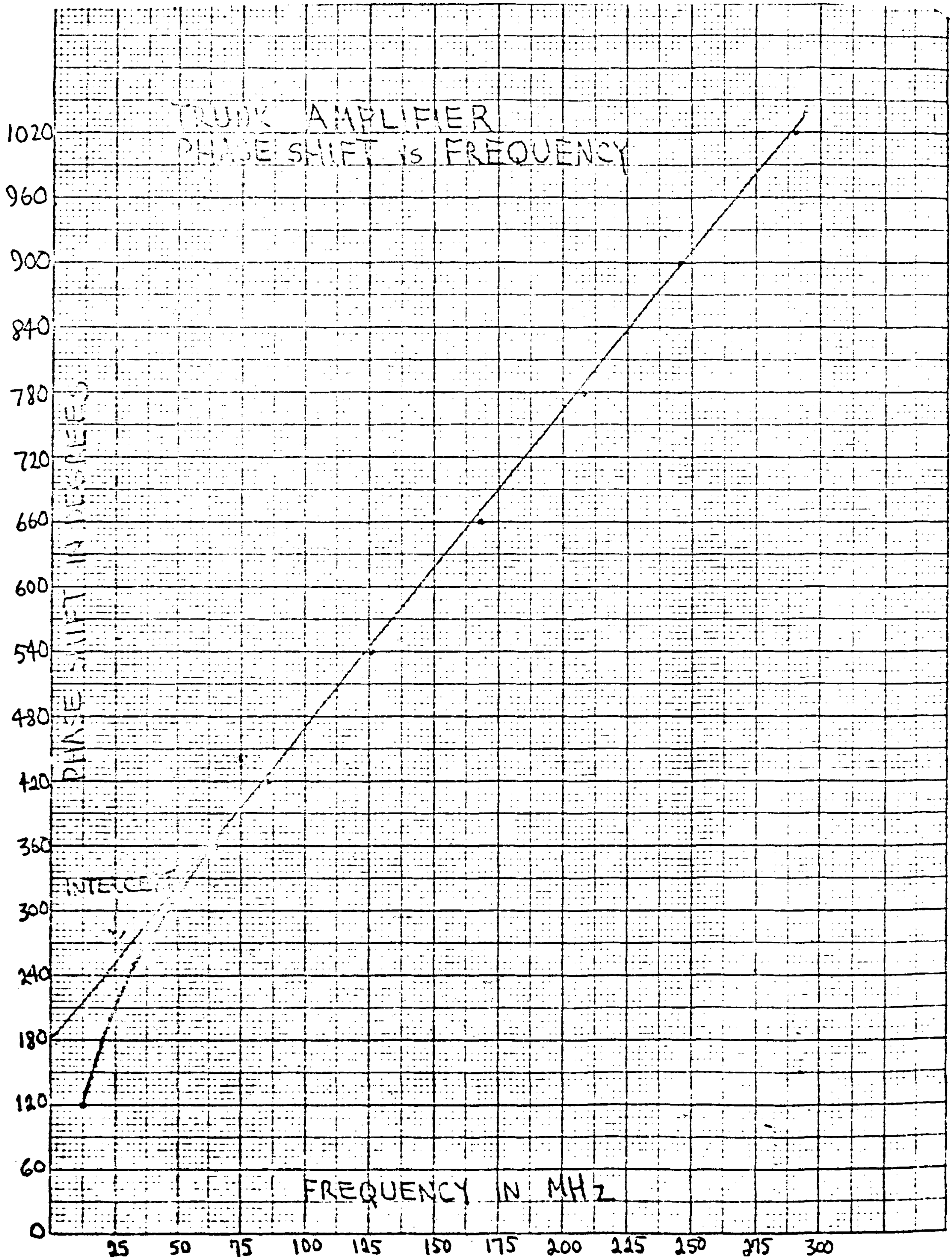
Figure 1

SINGLE FREQUENCY

IN-BAND INTERFERENCE TOLERANCE

INVESTIGATOR	SPECIFICATION
FINK-TELEVISION ENGINEERING	-55dB
CANADIAN-TELEVISION STANDARD BP-23	-57dB AT SYNC FREQUENCY AND AT COLOR SUB-CARRIER
ACCEPTED INDUSTRY STANDARD	-60dB WORST CASE
BELL LABORATORIES	-70dB AT SYNC FREQUENCY RMS INTERFERENCE TO P-P SIGNAL

Figure 2


TRUNK AMPLIFIER
PHASE SHIFT vs FREQUENCY
PHASE SHIFT IN DEGREES
FREQUENCY IN MHz
INTERCEPT
1020
960
900
840
780
720
660
600
540
480
420
360
300
240
180
120
60
0
25
50
75
100
125
150
175
200
225
250
275
300

2ND ORDER DISTORTION INCREASE AS FUNCTION OF PHASE INTERCEPT

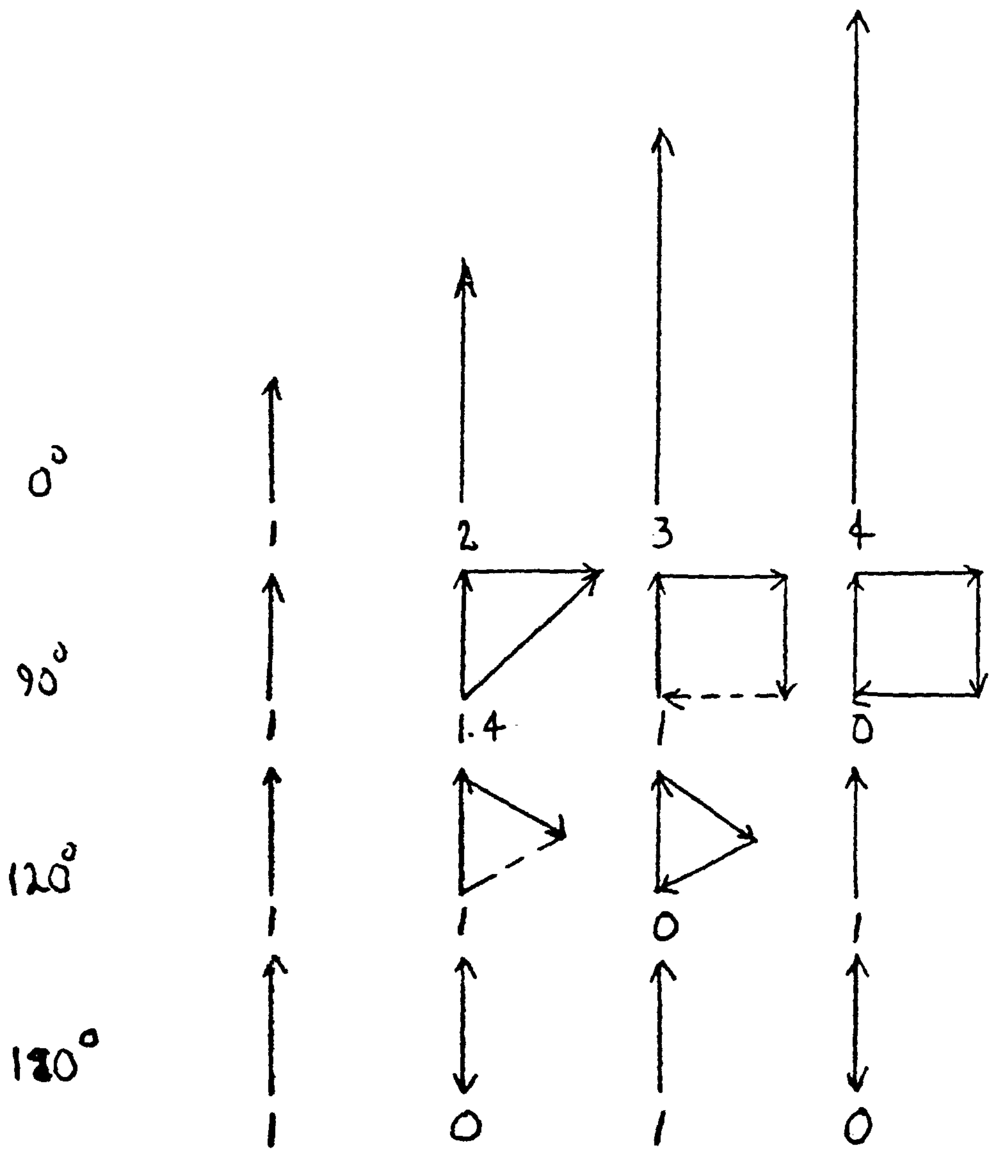
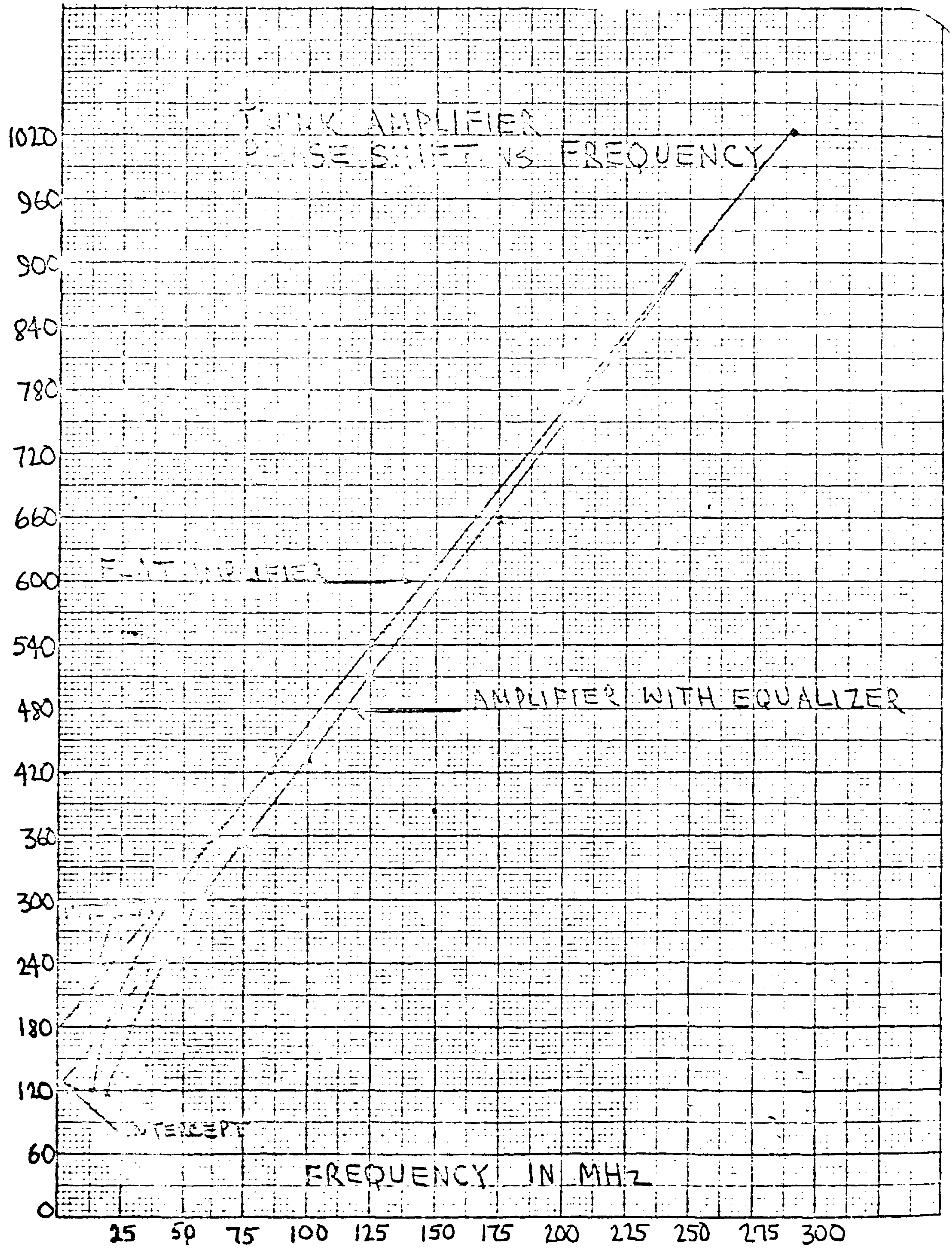
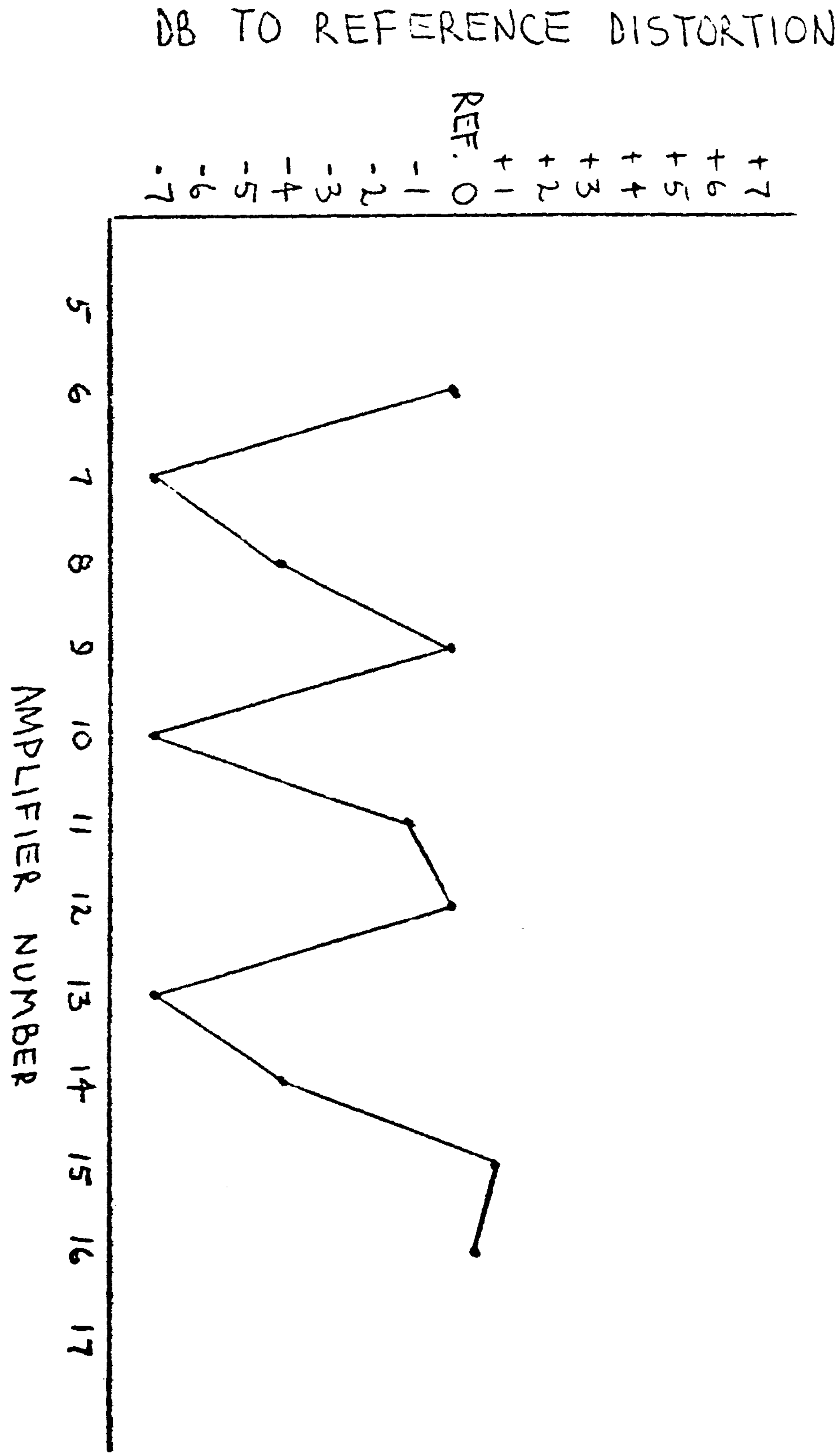


Figure 4



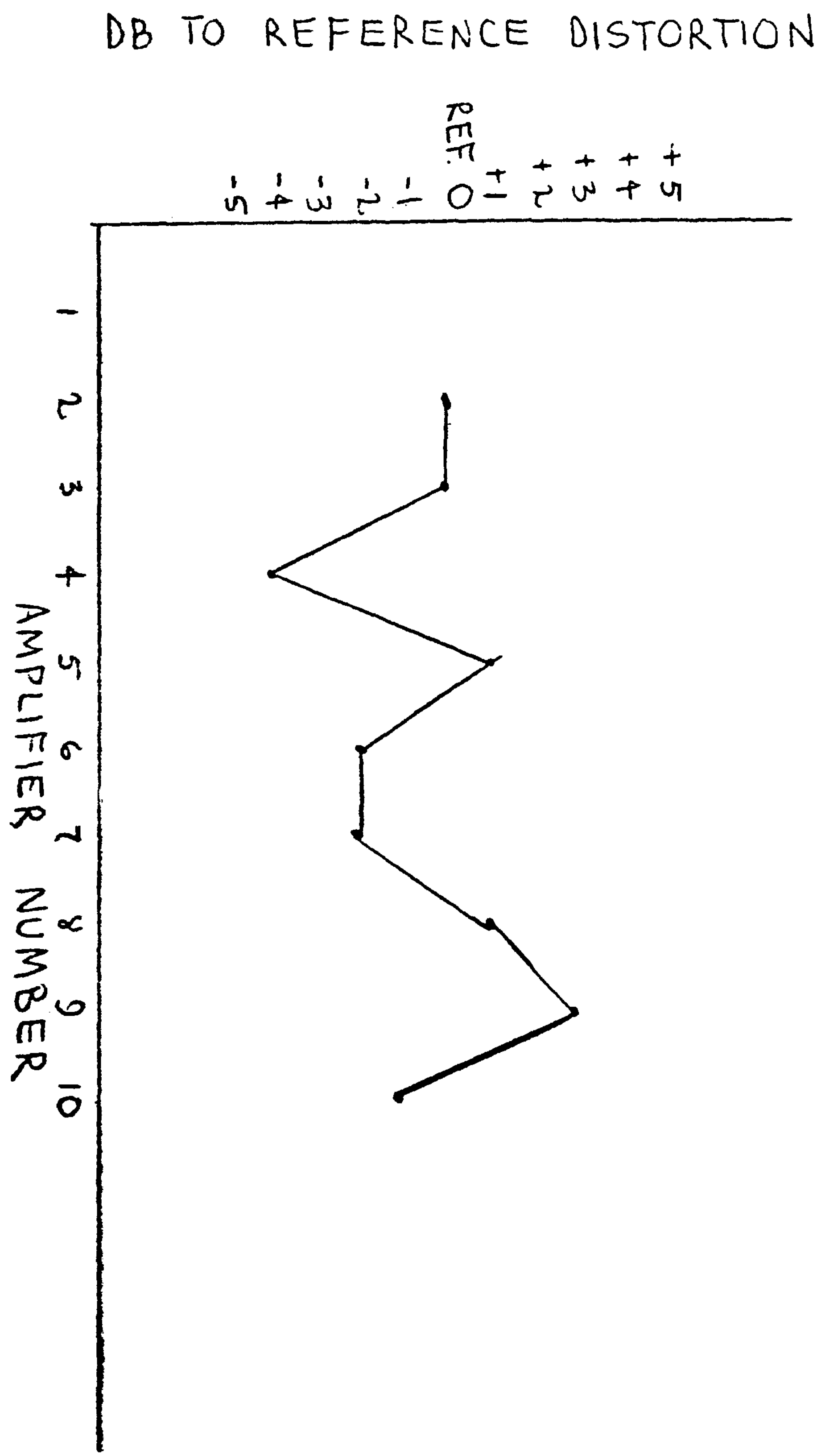
TRUNK AMPLIFIER
PHASE SHIFT VS FREQUENCY
FLAT AMPLIFIER
AMPLIFIER WITH EQUALIZER
INTERCEPT
FREQUENCY IN MHz
1020
960
900
840
780
720
660
600
540
480
420
360
300
240
180
120
60
0
25
50
75
100
125
150
175
200
225
250
275
300

Figure 5



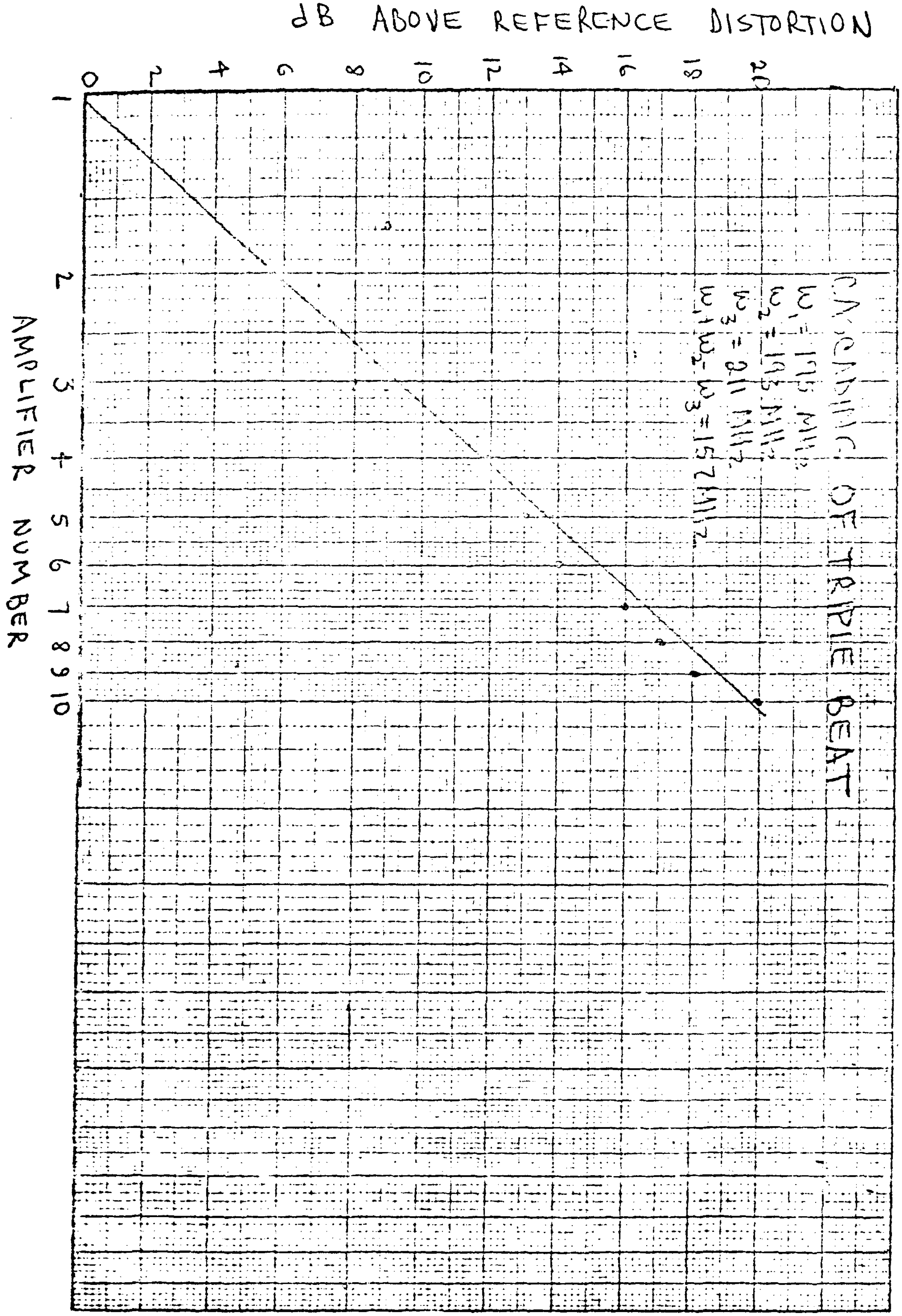
CASCADING OF 2ND ORDER INTER-MODULATION
CHANNEL 5 + 6 BEAT

Figure 6

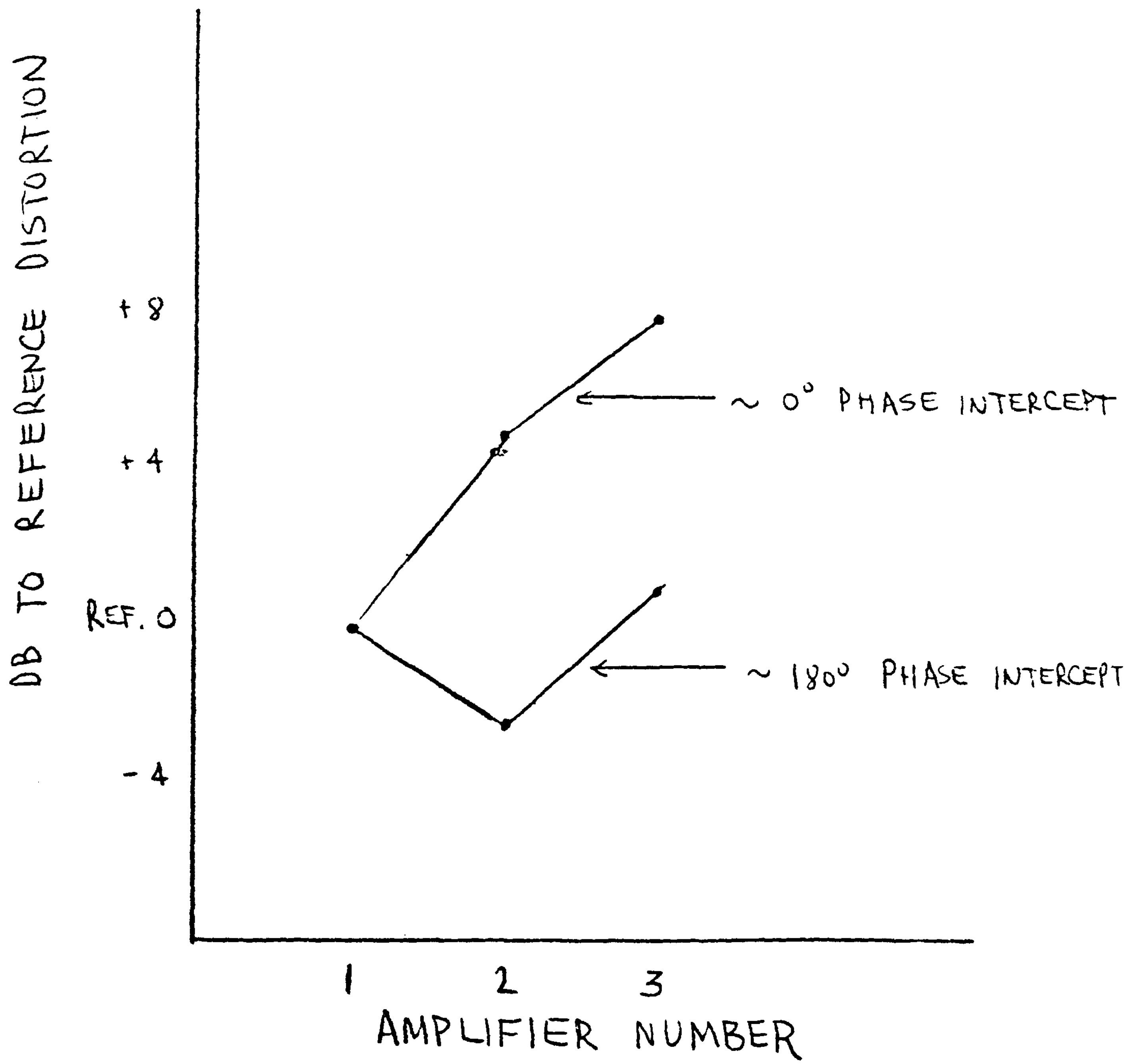


CASCADING OF 2ND ORDER INTERMODULATION
CHANNEL 4 + 13 DIFFERENCE BEAT

Figure 7


dB ABOVE REFERENCE DISTORTION
0 2 4 6 8 10 12 14 16 18 20
AMPLIFIER NUMBER
1 2 3 4 5 6 7 8 9 10
CASCADING OF TRIPLE BEAT
ω1 = 175 MHz
ω2 = 193 MHz
ω3 = 211 MHz
ω1+ω2−ω3 = 157 MHz

CASCADING OF DISTORTION
IN 3 AMPLIFIER CASCADE
CHANNEL 5+6 BEAT



ABSTRACT

CATV ANCILLARY SERVICES

H. J. Moeller
EDC Company
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The paper provides a description of future services to be offered, including telemetry circuits, secured closed circuits and communications channels and the basis upon which such services will be available.

Current state-of-the-art of terminal devices are presented in a definitive manner relating to services that will either simplify everyday social and economic activities and/or contributions to the entertainment and cultural needs of the subscriber. A base for broadband communications services in 1972, 1975 and 1980 are presented along with projected economics, hardware and software considerations.

DESIGNING INTRA-CITY AND INTER-CITY CARS BAND DISTRIBUTION SYSTEMS

By Dr. Joseph H. Vogelman, Sr. Vice President
and
Mr. Kenneth Knight, National Sales Manager
LASER LINK CORPORATION

This paper reports the practical applications of an FCC type approved development for Airlink CATV service.

The purpose of this article is to acquaint the industry with the practical applications of systems which have been filed with the FCC by cable operators and to show the economic and technical performance advantages of this new tool for cable television operators for both inter-city and intra-city CARS Band Distribution.

TWO-WAY MICROWAVE COMMUNICATION

It should be understood that Airlink CATV has two-way communications capabilities. Therefore the first system to be described will be one which we have in the design stage, shown on Figure 1, which we anticipate will be employed on a medical training program involving a medical center with member hospitals. The member hospitals participating will engage in bi-directional, interactive communications.

Figure 1 is a Laser Link System with two-way communications capabilities where the network illustrated shows ten (10) receiving stations, each of which is capable of transmitting back two (2) six megahertz channels in the same band, one of which could be used to provide specialized two-way or party-line communications between the receiving terminals.

In areas of community-wide communications services, specialized functions, such as Fire and Police Department reporting could be multiplexed over the Airlink System to each receiving location from which point it would be transmitted back as composite video to a Laser Link Transmitting Center, where messages could be relayed to Police and Fire Departments. Where such services are localized, each receiving point could serve as a contact with the appropriate local authorities.

THE LASER LINK SYSTEM

In the interest of first understanding the components of an Airlink system, we will review figures 2 to 6. Figure 2 is a block diagram of a cable system employing an air link equipment between the head end and the subscriber distribution system. Figure 3 is the typical configuration of the FCC type approved OLL 12T transmitter mounted on a 4 foot parabolic antenna. Figure 4 is a rear view of the transmitter with the cover removed showing the components. Figure 5 is the typical configuration of the receiver/converter.

Figure 6 is a view of the Laser Link Airlink System transmitter and parabolic antenna atop a tower at the Laser Link Laboratories, Garden City, Long Island, New York.

ACTUAL FCC FILINGS

The typical applications involved in distribution from a head end are:

1. A single head end to deliver multichannel signals to several remote population centers is shown in Figure 7. In this application, three communities at five, six and twenty-one miles will be serviced from an existing head end by three beams from a single transmitter.

2. To get signals over a mountaintop while simultaneously interlinking contiguous areas, the system design of Figure 8 is used. A 6,600 foot mountain hides the receiving site from the head end. This obstacle is overcome by using the mountain peak as a receiving point and simultaneously relaying the signals by means of a passive repeater consisting of two (2) ten foot antennas back to back.

3. The system of Figure 9 is used to economically jump over highways, railroad tressles and waterways. Three (3) sites at 4.6, 9 and 9.7 miles are connected to an existing head end which would otherwise require elaborate, expensive long runs of cable to get under highways and railroads and to go under or around major waterways.

4. Figure 10 shows a layout designed to optimize head end locations while avoiding "dry trunks." In addition to taking the signals from the head end to a new subscriber area, the design provides for future expansion to two additional sites by the addition of an independent amplifier, Type QLL-12TA to the transmitter.

5. In mountainous terrains a single head end on an isolated mountain provides service to many valley communities. The initial configuration provides service to three communities, with growth potential to ten (10) additional locations. One community, because its location is obscured by the terrain is reached by means of an active repeater operating at the intermediary receiving point. (Fig. 11)

6. Airlinks provide the capability to minimize underground distribution in urban areas. Since the subscriber populations are separated by non-revenue producing sections, the underground trunking costs are greatly reduced by the Airlink from the head end to each subscriber center. (Figure 12)

7. Figure 13 is an unusual application where we are taking an existing head end and beaming service to three additional cities, two via directional parabolic antennas line of sight to receiver/converter locations within twenty-five miles; and the third community serviced via a parabolic antenna from head end to a passive reflector on a mountain ridge, and thence deflecting the

signal to a receiver/converter six miles distant. Neighborhood service via input terminals at each localized Laser Link receiver/converter location are a standard capability to make available local community organizations.

This paper will describe systems which have been filed with the FCC which are in the seven categories listed above. It should be understood that in every instance where the Airlink has been applied there is a great economic savings as well as a technical improvement in system performance. Comparisons of typical configurations are shown on Table I. It should be noted by referring to the chart, (Table I), that even a single hop of eighteen miles has comparable signal to noise ratio (S/N) as a high performance cable trunk and we are comparing a $60,000.00 investment for Airlink against $230,000.00 capital cost for a cable system (not counting the annual pole rental charges). For "Seven Directions", with air distance from head end of twenty-five miles, the signal to noise ratio is comparable even for one (1) inch of rain, but the comparison of cost between Airlink and trunk cable at this distance is astonishing - $125,000.00 as compared to $2,226,000.00

All of these savings are achieved while making an improvement in the performance of the system under average weather condition.

FILING FOR AIRLINK CATV

For systems which are overgrown or where inter-city or intra-city Airlink connections satisfy a cable TV operator's requirements, the procedure for placing this system into service is as follows:

A. The cable operator submits his plan to our field engineering group who report on both the technical and economic advantages of the system as it should be applied to specific requirements and the most economic combination of Airlink and cable.

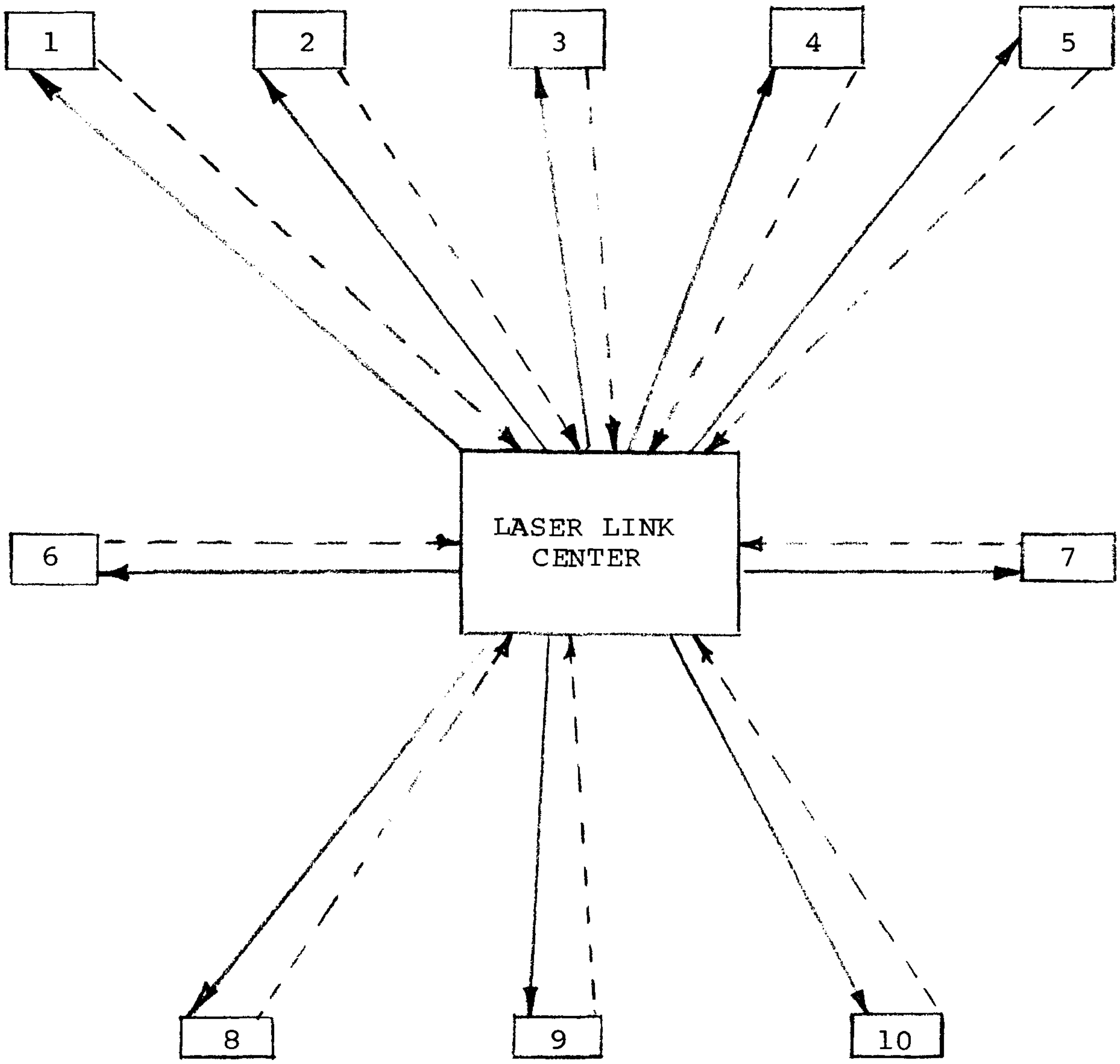
B. The cable operator takes the information received from field engineering and provides the responsive information needed on FCC Form No. 402. On this form, a cable operator is required to show the need for an Airlink service and for use of each channel as well as his planned location for transmitter and receiving sites.

C. Upon accepting the application, the FCC issues a BPCLD Number which acknowledges the suitability of the application for filing. The application is then in line for processing.

D. The FCC studies any possible interference and if all conditions are satisfactory, will issue a construction permit. The application of this system should make the Airlinking of areas with only a few hundred potential subscribers viable and economically sound; open up the urban markets so that the cable operator is not burdened by a leased cable association with the local telephone companies. "Dry trunks", railroad tressles, highways and the burden of obtaining right of way are eliminated to an important extent by the use of Airlink CATV.

The employing of Airlinks opens new markets for cable TV as operators as well as improving the standards in existing over-burdened systems. The equipment in this art has now been Type Accepted and will soon be adaptable for low-cost long haul CARS Band Microwave.

The future is great for those who have the wit and the means to bring new developments like this into early service.



1 CHANNEL

CENTER TO OTHER 10 LOCATIONS

12 CHANNEL

CENTER:

1 - QLL 12 Transmitter - 12 TV channels
10 - Beams 1 per location
10 - Standard Microwave Receivers & Modulators

SATELLITE LOCATIONS:

1 - QLL 12 Receiver - 12 TV channels	Total 10
1 - QLL 12 Receiving Antenna	Total 10
1 - Standard Microwave Transmitter	Total 10

With 11 TV sets at each location all 11 are in continuous communication with everyone else. Possible combinations are:

4 Two Way + 1 Three Way. Up To 11 Party Conference of all locations.

Fig. 1

FIG. 1

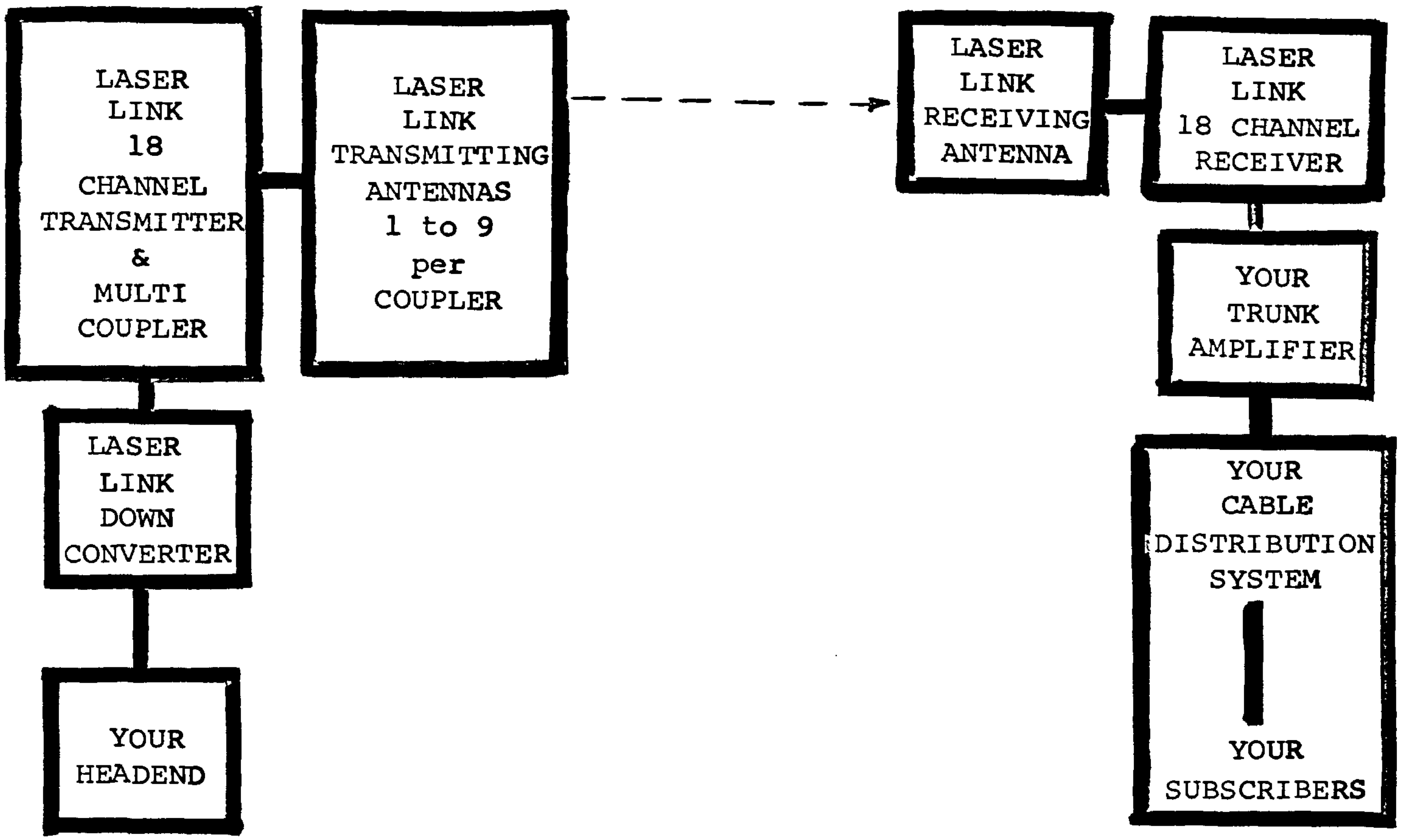

LASER LINK 18 CHANNEL TRANSMITTER & MULTI COUPLER
LASER LINK TRANSMITTING ANTENNAS 1 to 9 per COUPLER
LASER LINK RECEIVING ANTENNA
LASER LINK 18 CHANNEL RECEIVER
LASER LINK DOWN CONVERTER
YOUR HEADEND
YOUR TRUNK AMPLIFIER
YOUR CABLE DISTRIBUTION SYSTEM
YOUR SUBSCRIBERS

FIG. 2

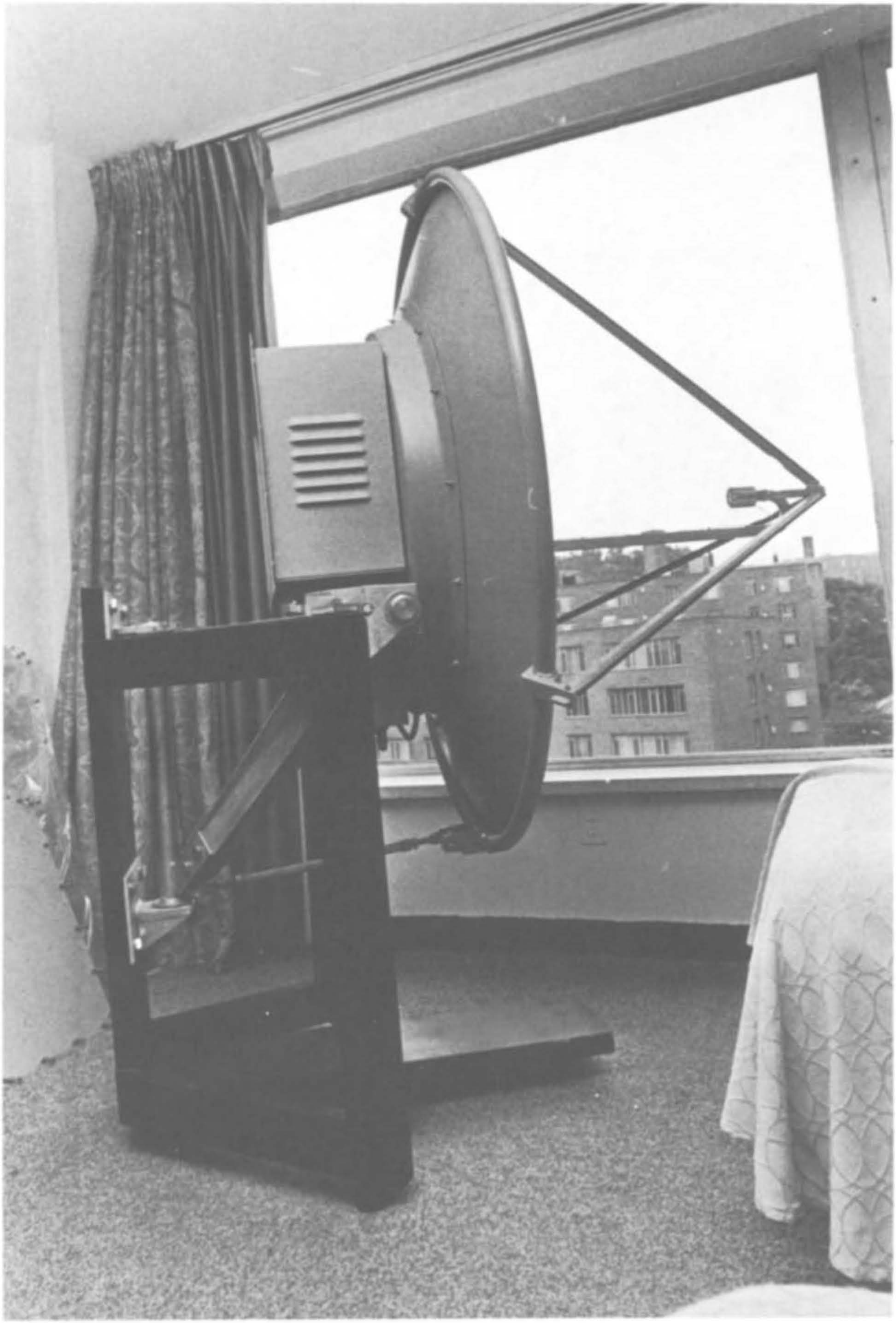


FIG. 3

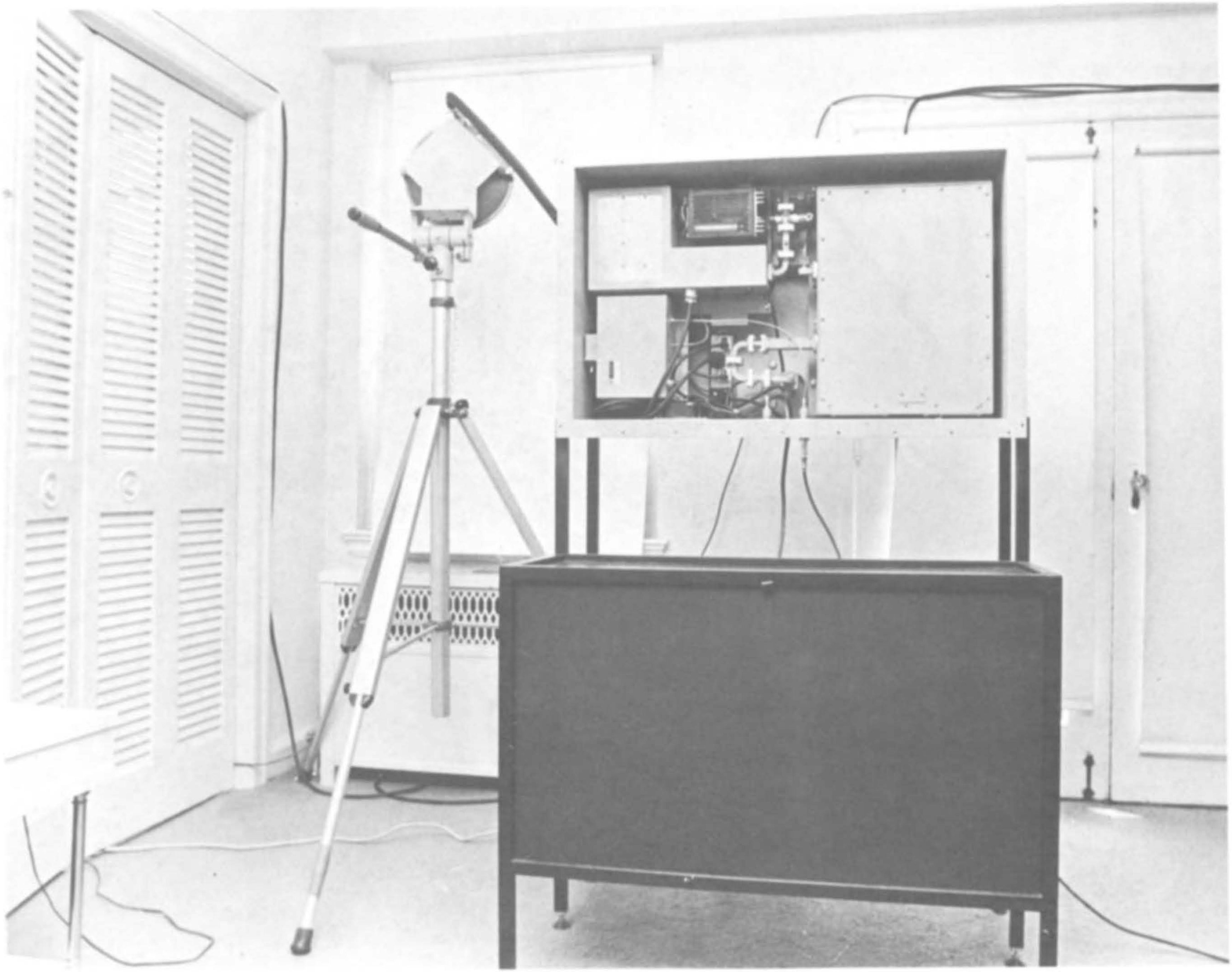


FIG. 4

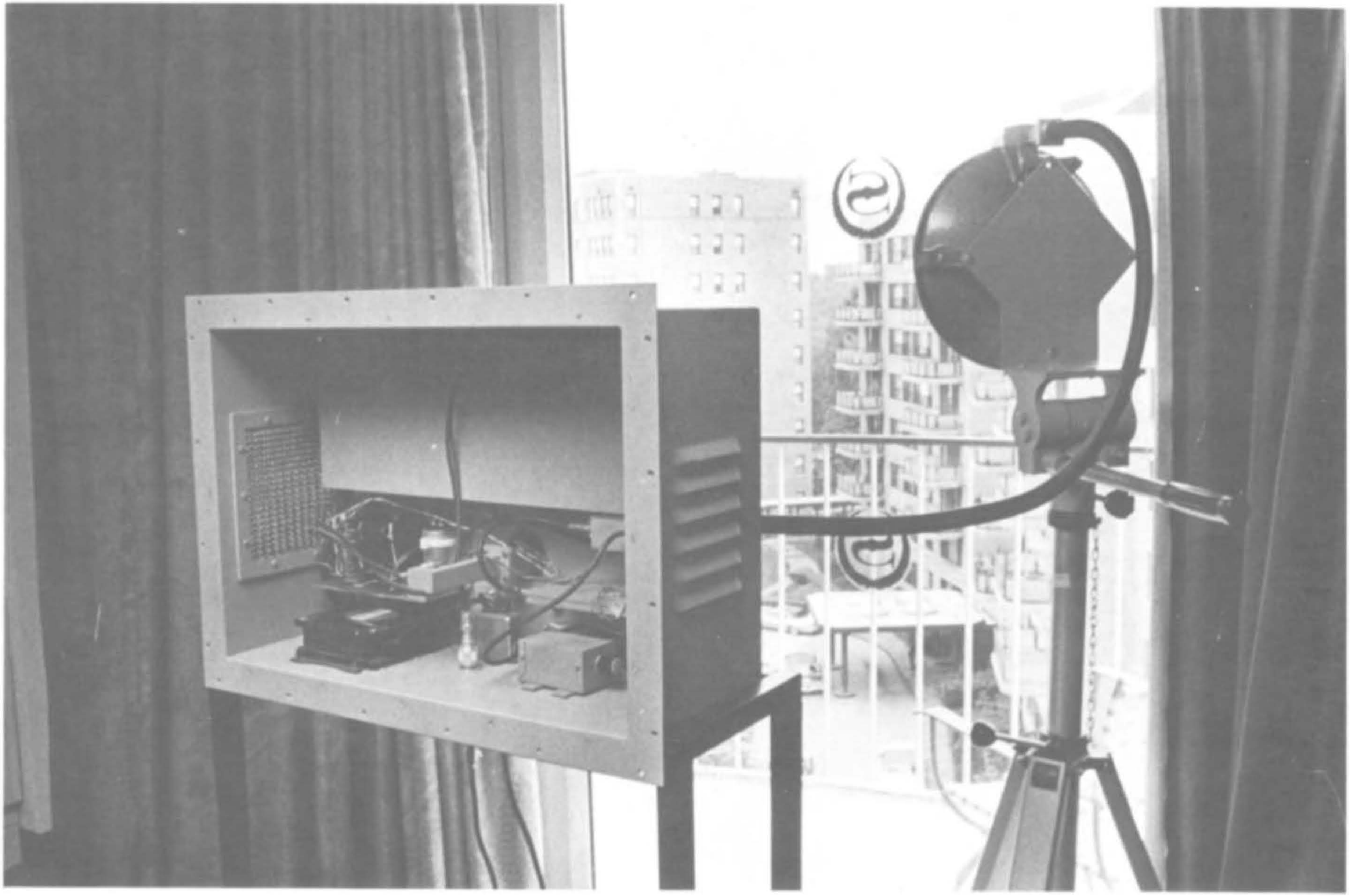


FIG. 5

The Laser Link cableless CATV transmitting-receiving terminal in Garden City, NY incorporates quasi laser link concept of antenna and parabolic reflector equipment at top. Back view shows transmitter housing containing modulation and amplification components.

FIG. 6

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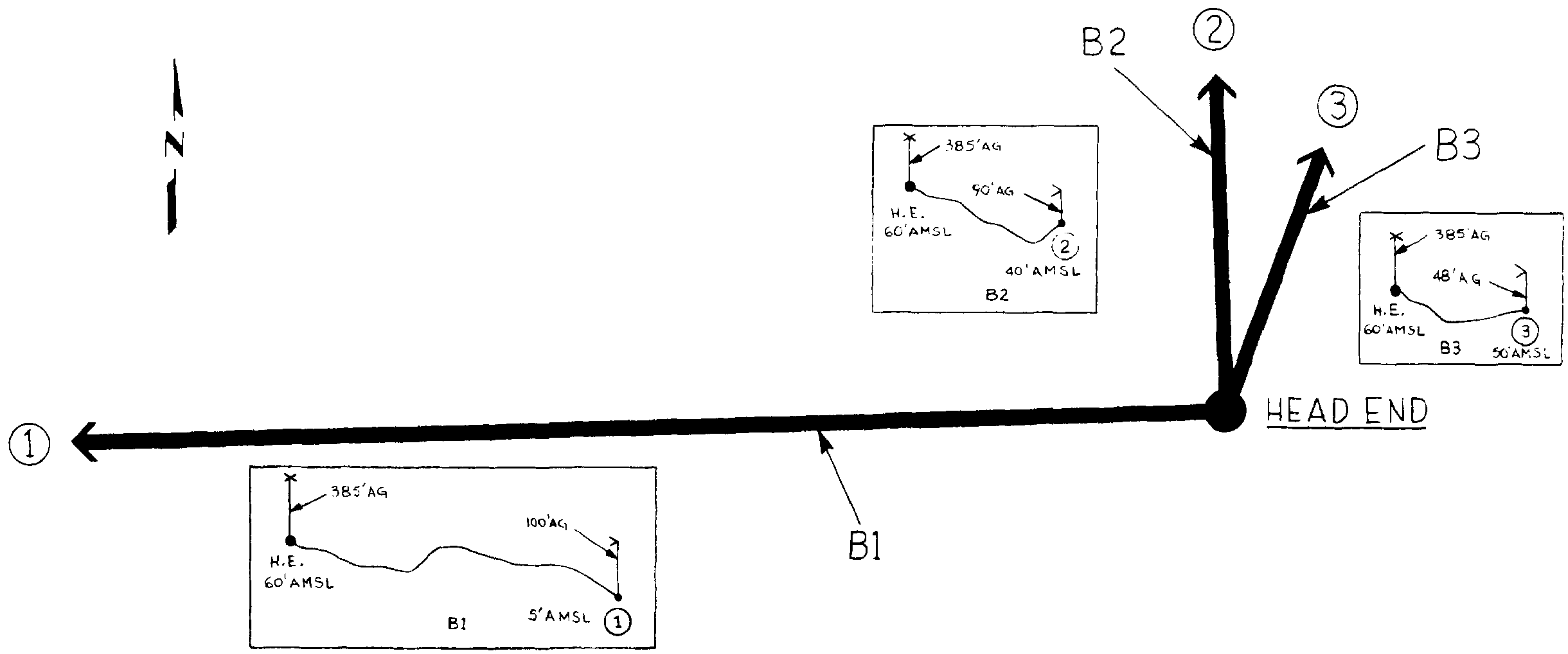
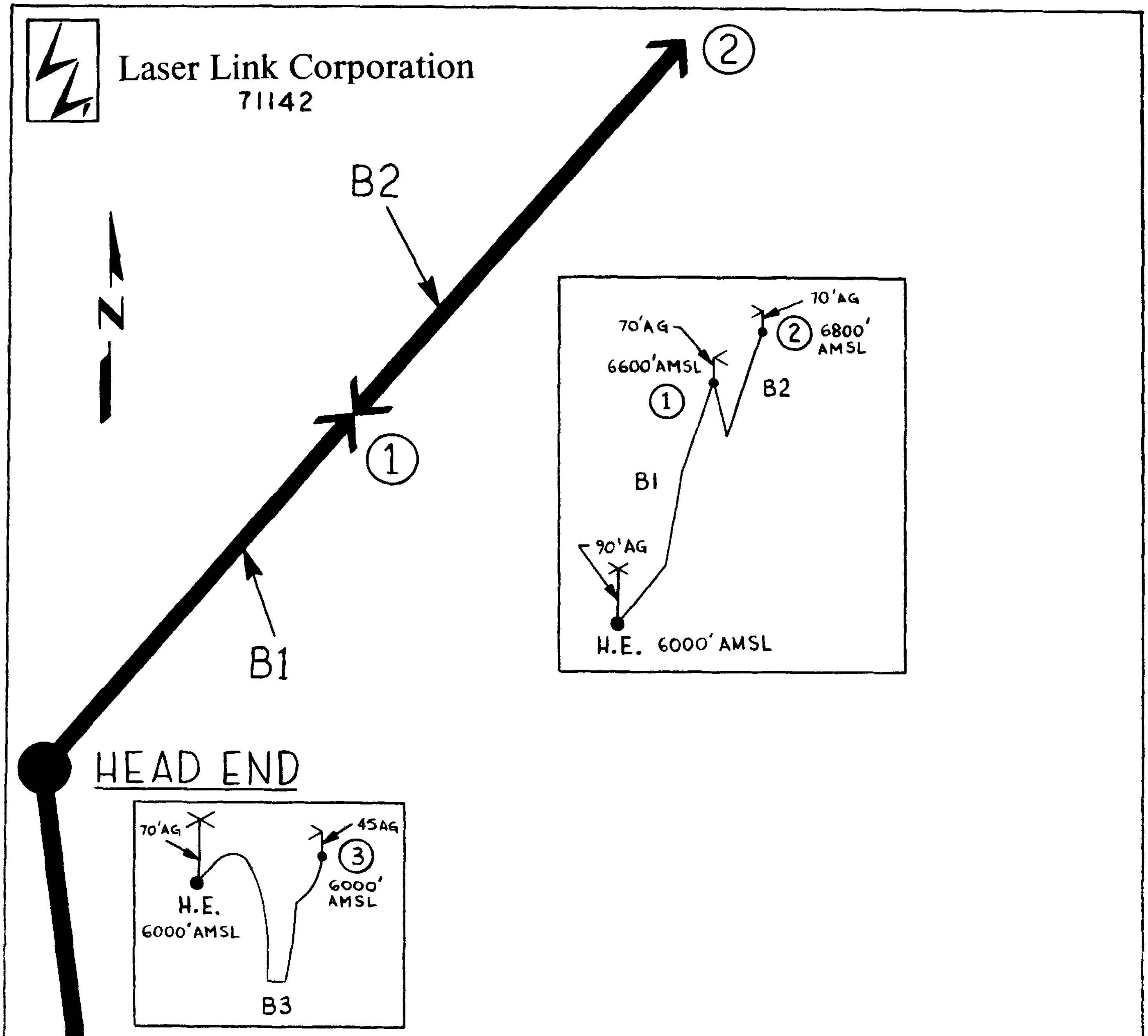


Site No.	Location	Miles to Head-end	Beam No.	Beam Power, W	Xmtr. or Rec'vr. Model [illegible]	Xmtr. Antenna Size [illegible]	Rec'vr. Antenna Size [illegible]	Other
1	SMYRNA [illegible]	21	B1	18.[illegible]	12RTDA	10	10	–
2	DE[illegible]ONA	[illegible]	B2	0.5	12R	4	4	–
3	DE[illegible]	5	B3	0.5	12R	4	4	[illegible]
Head-end	ORANGE CITY	–	–	19.0	12T	–	–	–

FIG. 7



Site No.	Location	Miles to Head-end	Beam No.	Beam Power, W	Xmtr. or Rec'vr. Model QLL-	Xmtr. Antenna Size FT	Rec'vr. Antenna Size FT	Other
1	PALMER PARK	3	B1	15	–	10	10	–
2	RATTLESNAKE HILL	4.5 (1)	B2	(2)	-12RTDA	10	10	–
3	STRATMORE HILLS	4	B3	0.1	-12R	4	4	–
Head-end	COLORADO SPRINGS							

Notes:
1. DISTANCE FROM RELAY SITE NO. 1
2. PASSIVE REPEATER OUTPUT FROM SITE NO. 1
3. RAIN RATE CAPABILITY = 1in./hr.

FIG. 8

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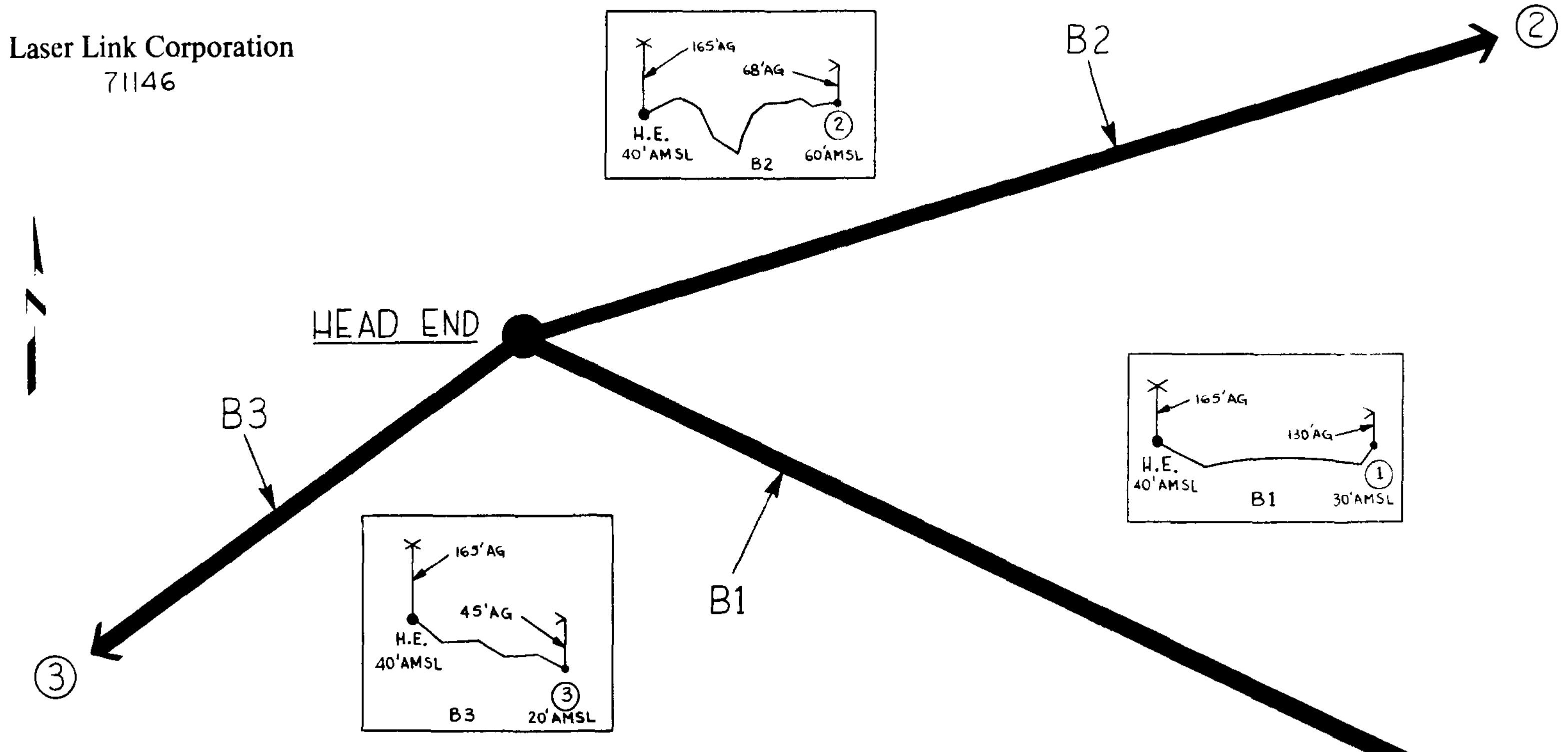
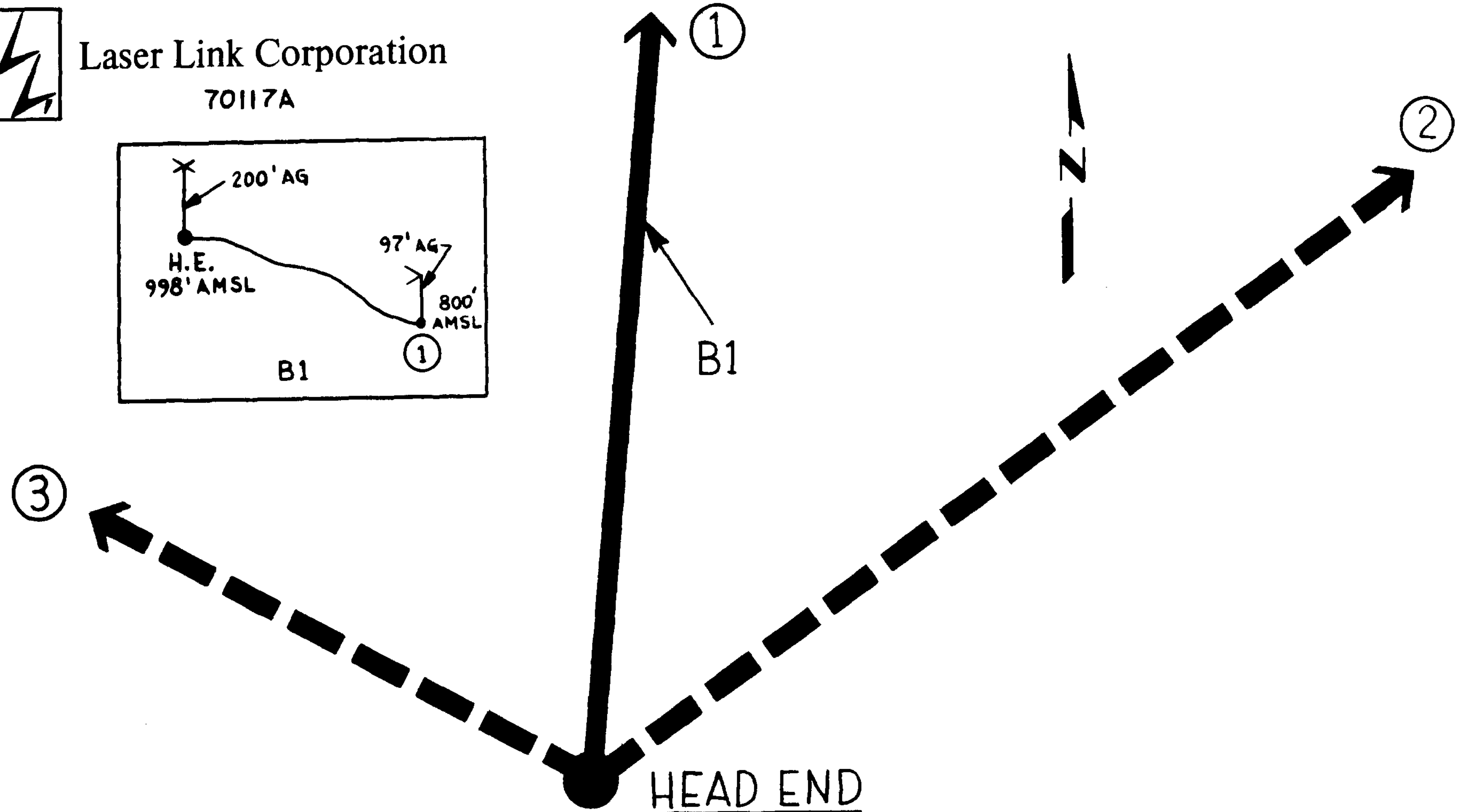


Site No.	Location	Miles to Head-end	Beam No.	Beam Power, W	Xmtr. or Rec'vr. Model QLL-	Xmtr. Antenna Size FT	Rec'vr. Antenna Size FT	Other
1	DESTIN	9.7	B1	4	-12R	10	4	–
2	EGLIN AFB, BUILD. 859	9.0	B2	14 (1)	-12R	4	4	–
3	EGLIN AFB, HURLBURT FLD	4.6	B3	1	-12R	4	4	–
Head-end	WRIGHT	–	–	19	-12T	–	–	–

Notes: 1. PROVIDES FOR EXPANSION CAPABILITY TO NELSON POINT WHEN REQ'D.

Laser Link Corporation

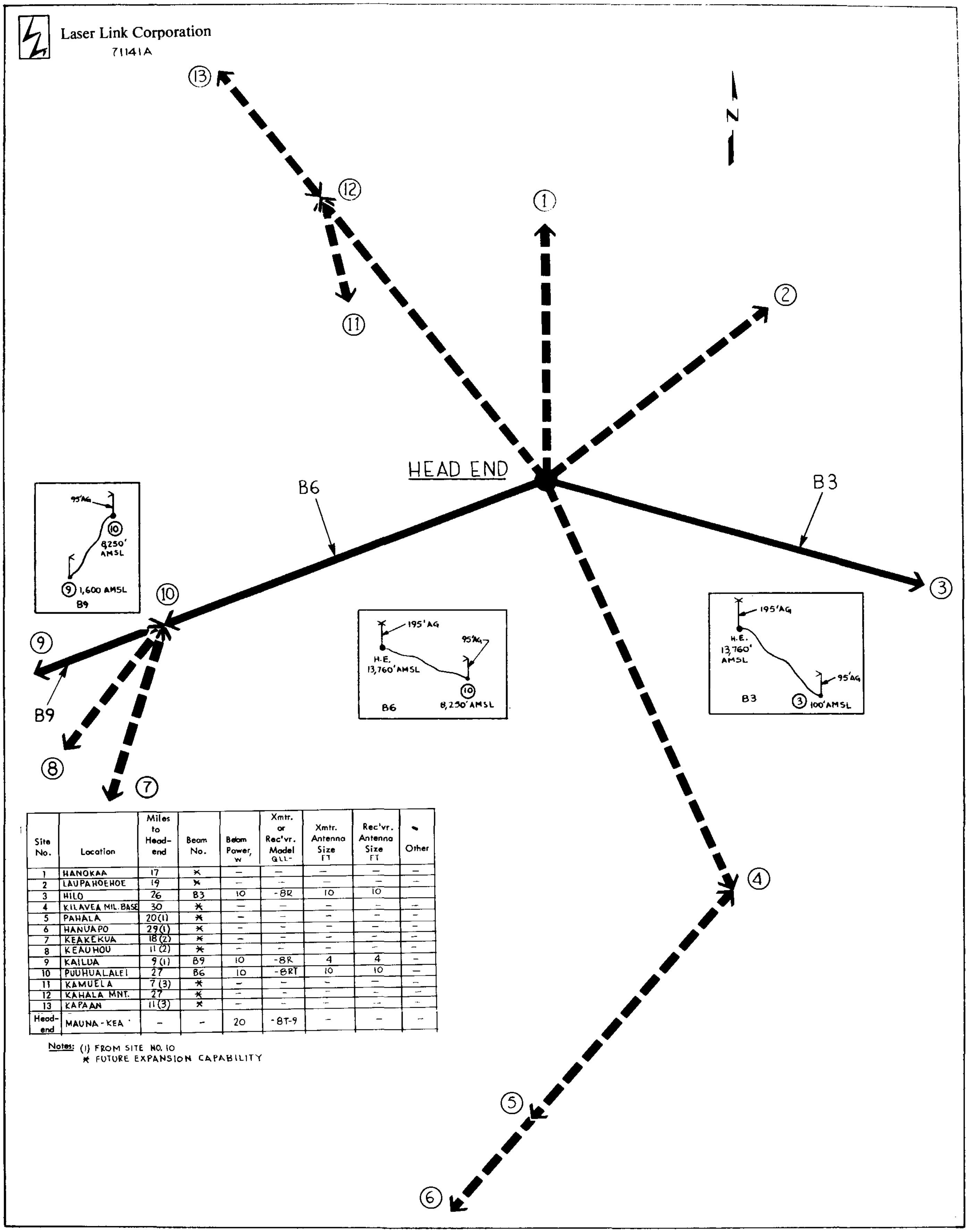
70117A



Site No.	Location	Miles to Head-end	Beam No.	Beam Power, w	Xmtr. or Rec'vr. Model QLL-	Xmtr. Antenna Size FT	Rec'vr. Antenna Size FT	Other
1	CALIFORNIA	21	B1	19	12 RTDA	10	10	–
2	JEFFERSON CITY	29	*	–	–	–	–	–
3	VERSAILLES	15	*	–	–	–	–	–
Head-end	ELDON	–	–	19	12T	–	–	–

Notes: RAIN RATE CAPABILITY = 1 in/hr
* FUTURE EXPANSION CAPABILITY

FIG. 10

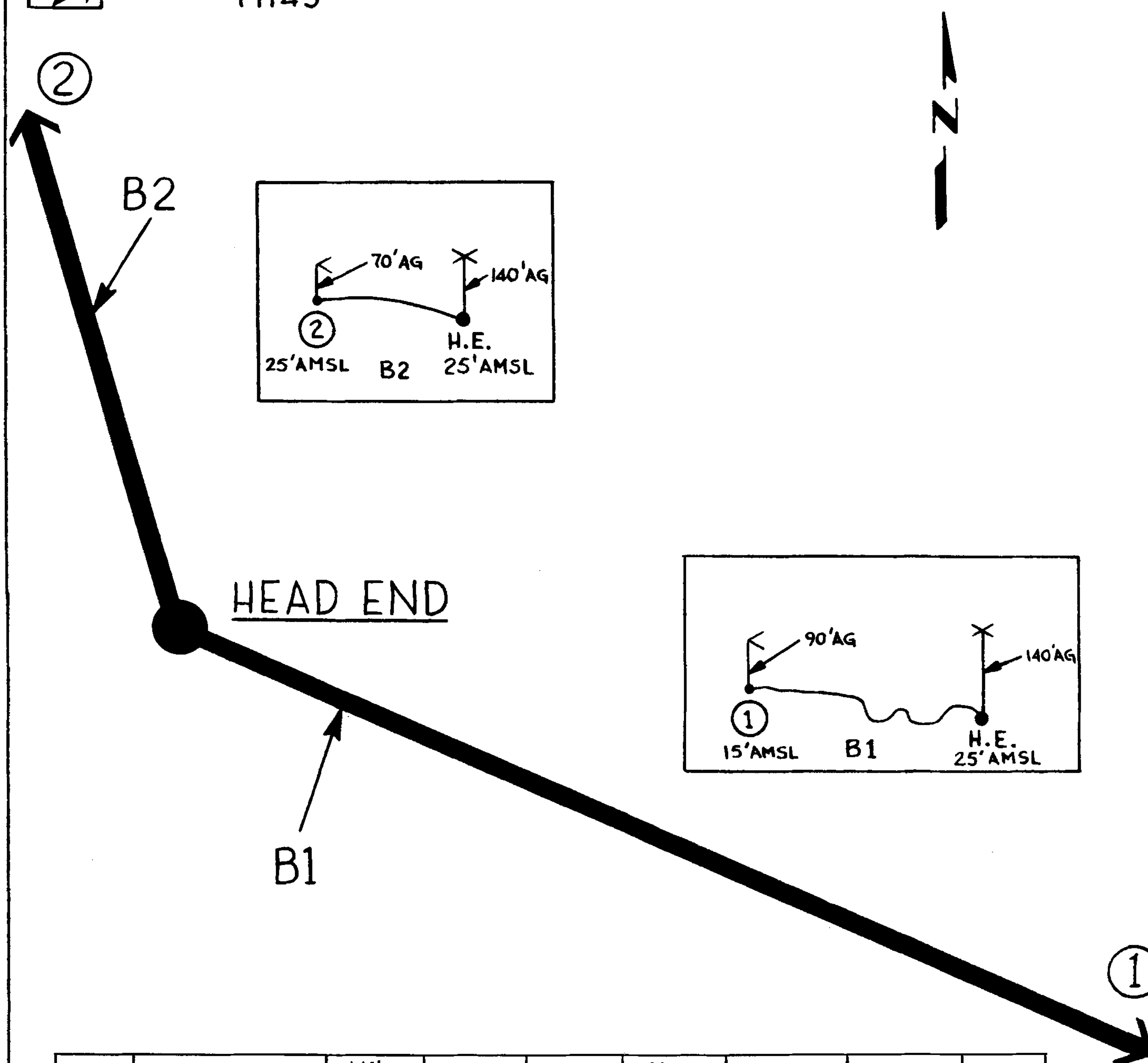


Site No.	Location	Miles to Head-end	Beam No.	Beam Power, W	Xmtr. or Rec'vr. Model QLL-	Xmtr. Antenna Size FT	Rec'vr. Antenna Size FT	Other
1	HANOKAA	17	*	–	–	–	–	–
2	LAUPAHOEHOE	19	*	–	–	–	–	–
3	HILO	26	B3	10	-8R	10	10	
4	KILAVEA MIL. BASE	30	*	–	–	–	–	–
5	PAHALA	20(1)	*	–	–	–	–	–
6	HANUAPO	29(1)	*	–	–	–	–	–
7	KEAKEKUA	18(2)	*	–	–	–	–	–
8	KEAUHOU	11(2)	*	–	–	–	–	–
9	KAILUA	9(1)	B9	10	-8R	4	4	–
10	PUUHUALALEI	27	B6	10	-8RT	10	10	–
11	KAMUELA	7(3)	*	–	–	–	–	–
12	KAHALA MNT.	27	*	–	–	–	–	–
13	KAPAAN	11(3)	*	–	–	–	–	–
Head-end	MAUNA-KEA	–	–	20	-8T-9	–	–	–

Notes: (1) FROM SITE NO. 10
* FUTURE EXPANSION CAPABILITY

FIG. 11

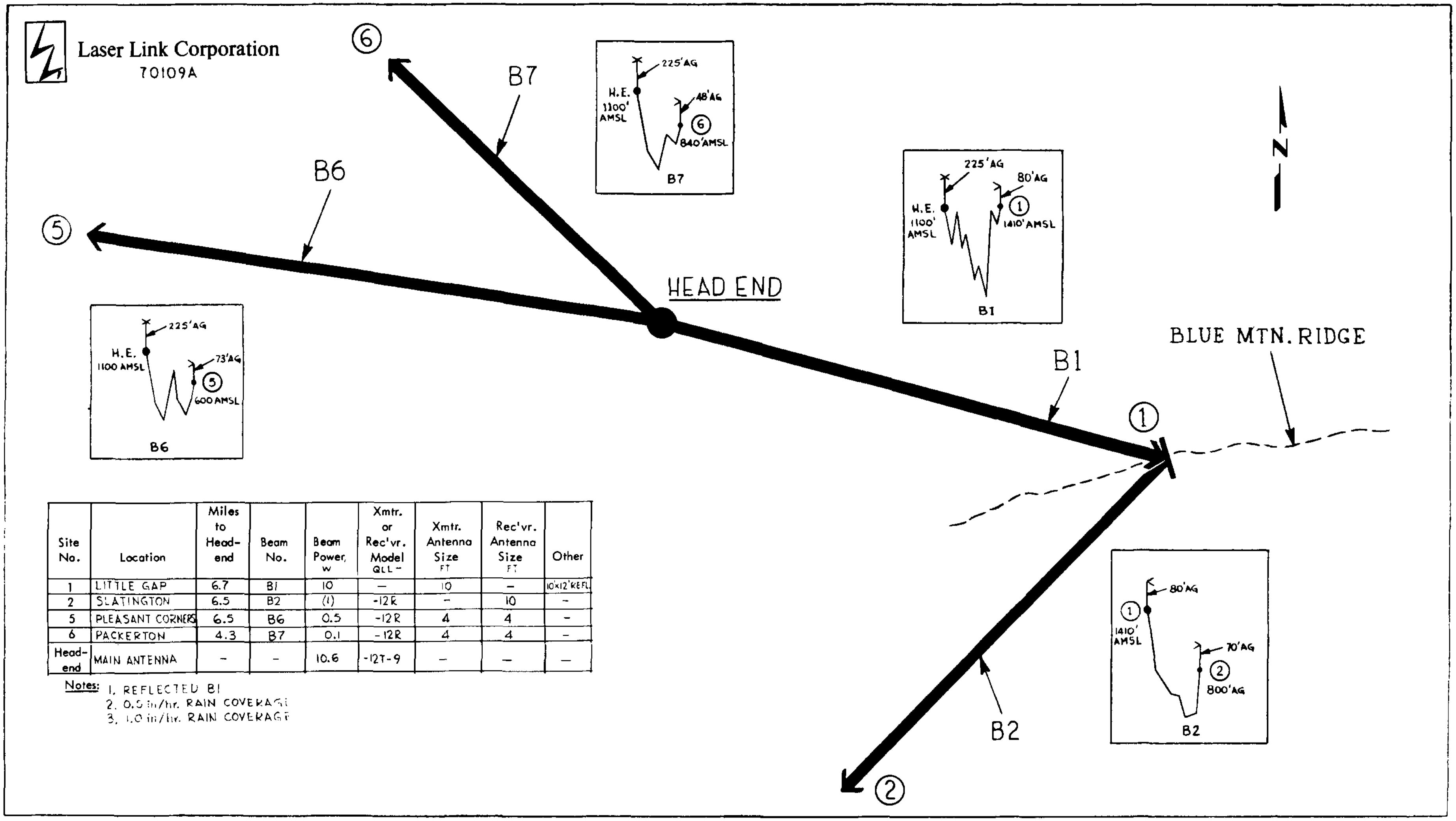
Laser Link Corporation
71143



Site No.	Location	Miles to Head-end	Beam No.	Beam Power, W	Xmtr. or Rec'vr. Model QLL-	Xmtr. Antenna Size FT	Rec'vr. Antenna Size FT	Other
1	FLOWER BLUFF	9	B1	5	-12R	4	4	–
2	N. CORPUS CHRISTI	4	B2	1	-12R	4	4	–
Head-end	CORPUS CHRISTI	–	–	6	-12T	–	–	–

Notes: RAIN RATE CAPABILITY = 1 in/hr

FIG. 12



Site No.	Location	Miles to Head-end	Beam No.	Beam Power, W	Xmtr. or Rec'vr. Model QLL-	Xmtr. Antenna Size FT	Rec'vr. Antenna Size FT	Other
1	LITTLE GAP	6.7	B1	10	–	10	–	10x12' REFL.
2	SLATINGTON	6.5	B2	(1)	-12R	–	10	–
5	PLEASANT CORNERS	6.5	B6	0.5	-12R	4	4	–
6	PACKERTON	4.3	B7	0.1	-12R	4	4	–
Head-end	MAIN ANTENNA	–	–	10.6	-12T-9	–	–	–

Notes: 1. REFLECTED B1
2. 0.5 in/hr. RAIN COVERAGE
3. 1.0 in/hr. RAIN COVERAGE

FIG. 13

TABLE I

COST COMPARISON OF LASER LINK AIRLINK AND TRUNK CABLE

TYPE	AIR DISTANCE FROM HEAD END	AIRLINK FOR VARIOUS RAIN RATES 0.5 inch/hr	1 inch/hr		HI PERFORMANCE CABLE TRUNK		
		S/N Ratio	S/N Ratio	Quoted Price	S/N	Capital Cost	Pole Rental
Single Hop	18 miles	52	45	61.6	45	230	3
Three Directions	18	46	46		45		
	20	45	43	94.8	44	742	10
	20	45	43		44		
Five Directions	18	49	46		45		
	8	57	45	107.3	49	638	8
	8	57	45		49		
	8	57	45		49		
	8	57	45		49		
Ten Directions	10		46	147.3	48	1198	16
Seven Directions	25 Maximum	46	-	125.2	44	2226	28
Urban High Density 100 Sites	within 20 miles	46	-	101.4	42	2500	35

Costs in Thousands of Dollars:

Excellent pictures requires more than 42db channel S/N

DYNAFOAM COAX COMES OF AGE FOR CATV

written and presented by:

John Arbuthnott, Jr., P.E.
NCTA National Convention
Washington, D. C.
July, 1971

SUBSIDIARY OF Insilco
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Wallingford, Connecticut 06492 Tel (203) 269-3381
Phoenix, Arizona 85005 Tel (602) 278-5576

INTRODUCTION:

Two years ago, Times Wire and Cable introduced a new low loss, high efficiency, semiflexible coaxial sheath cable designed primarily for use in Cable Television Systems. Since that time over three thousand miles of cable has been manufactured, delivered on site and installed. This paper is to further inform the industry about this, to our mind, important developement.

WHY SEARCH FOR LOWER LOSS CABLE?

CATV Cables are designed primarily to transport signals from one point to the other as efficiently as practical. In the typical Cable Television System, how well this is accomplished, all other things being equal, can be judged by a comparison of cable loss versus dollar cost.

What we are saying is that an increase in the efficiency of the cable is of no value if its cost increases at a rate equal to or greater than the rate of increase of efficiency. A convenient way to make this judgement is to multiply the cost of the cable by the attenuation. If a substitute cable, analyzed the same way, shows a smaller number it is possible to achieve a more cost effective system by its use.

A further stimulus to the developement of lower loss cable has been the desire to go to longer trunk runs. It is presumed that in these cases the economics are governed by front end location and construction costs rather than the less complicated analysis cited above.

This is best illustrated by assuming a trunk run designed on 22 db spacing and 36 amplifiers in cascade. The possible length of run is tabulated below.

JT-1750 (Polyethylene)	15 miles
JT-2750 (Dynafoam)	20 miles
JT-21000 (Dynafoam)	25 miles

The maximum length of run can of course vary with design criteria, but the comparison is still valid.

In summary, lower loss cables are developed to achieve overall reduction in system construction and maintenance cost.

DESIGN CONSIDERATIONS:

Dynafoam cable was developed with the intention to offer the system designer lower loss for the installation dollar and to increase the length of usable trunk run where, for one reason or another, it might be required; without having to resort to air dielectric cables, where pressurization costs can eat up any original economies attained.

Examination of the general formula for cable loss

$$\alpha = \frac{0.0174}{\text{Ln } D/d}\left(\frac{0.414}{d} + \frac{0.525}{D}\right)\sqrt{Ke}\ \sqrt{F} + 2.78 \text{ Tan } \delta \ \sqrt{Ke} \quad F$$

indicates that greater efficiency can be achieved by reduction of Dielectric Constant.

Furthermore when one considers that

$$Zo = \frac{60}{\sqrt{Ke}} \text{ Ln } D/d$$

it becomes apparent that the diameter of the inner conductor (d) must be made larger to maintain the proper impedance. This of course leads to a further reduction in cable attenuation.

Dynafoam cable accomplishes this by taking advantage of the greater foamability of Polystyrene over Polyethylene. Where Foamed Polyethylene limits the Dielectric Constant to values of 1.55 or greater, it is possible to achieve values of as low as 1.20 by using foamed Polystyrene.

For purposes of comparison the following table compares the overall diameters of cable having substantially the same attenuation against the insulation used.

Dynafoam	Foamed Polyethylene
0.340	0.412
0.412	0.500
0.500	0.750

Figure one shows the attenuation values of all Dynafoam cables versus Frequency from 10 MHz to 1000 MHz.

Simplified formulae are tabulated below for calculating the attenuation of all Dynafoam Cables.

JT-2340 α	=	$0.1010\ \sqrt{F} + 0.00074\ F$
JT-2412 α	=	$0.0827\ \sqrt{F} + 0.00060\ F$
JT-2500 α	=	$0.0671\ \sqrt{F} + 0.00051\ F$
JT-2750 α	=	$0.0444\ \sqrt{F} + 0.00040\ F$
JT-21000 α	=	$0.0336\ \sqrt{F} + 0.00040\ F$

When it is considered that the major cost of cable is material and that the amount of material in a cable is a function of at least the square of the diameter it becomes obvious that a smaller cable having the same attenuation as one that is larger will lead to greater economy.

MECHANICAL CONSIDERATIONS:

When first designed, the Dynafoam line was set up using the same mechanical criteria as the older Alumifoam Cables except that the inner conductors were adjusted to achieve the 75 ohm Characteristic Impedance required.

Subsequently it was found that the cables, particularly JT-2500 and JT-2750 were subject to mechanical change upon installation.

It was found that the 1/2" and the 3/4" cables had a tendency to flatten when bent on small radii.

The problem existed because the polystyrene air mixture used for insulation gives far less physical support than polyethylene foam. A bending jig was designed, Figure two, which maintains the circularity of the cable during bending.

Additionally it was found that JT-2500 and JT-2750 had some tendency to kink as it came off the reel. This problem, related to the same reduced support by the insulation, was effectively solved by increasing the wall thickness of the aluminum sheath. In both cases it was found that the problem was solved by a relatively minor increase in Aluminum Sheath wall thickness of only 0.005".

The following table shows the mechanical dimensions of Dynafoam Cable.

Cable Type	Inner Conductor O.C.	Insulation O.D.	Outer Cond. Wall	O.D.
JT-2340	0.075	0.300	0.020	0.340
JT-2412	0.092	0.362	0.025	0.412
JT-2500	0.114	0.440	0.030	0.500
JT-2750	0.172	0.666	0.042	0.750
JT-21000	0.227	0.890	0.055	1.000

INSTALLATION EXPERIENCE:

At the present time installations of Dynafoam Cable have been made in approximately thirty states ranging from Idaho to Florida. The cable has been supplied with and without jacket for aerial installation, and both flooded, jacketed and armored for underground installations.

As a matter of course, because the cable design represented a departure from what installation crews have been used to, several installations were checked out thoroughly, including return and insertion loss measurements both prior to and after installation.

It can be reported that, if installed with a reasonable degree of care and if the installer refrains from pulling around sharp corners, uniform cable prior to installation will be uniform after it is installed. The same can be said of foam Polyethylene cables.

What constitutes a careful method of installation is governed by the particular characteristics of the job at hand. Each man on the job can probably think of ways to install the cable without damage and we can not offer hard and fast rules for such a varied problem. We can offer some general rules that will help. I might add that these suggestions apply to the more conventional Aluminum Cables as well as to Dynafoam.

1. Some method of reel braking should be available.
2. Care should be taken to lead cable into the installation without excessive wall pressure.
3. Cable should not be pulled around right angle bends.
4. Cable pulling tensions should not be exceeded.

Some studies have been made to determine aging characteristics for Dynafoam cable. These studies, made on actual installations consisted of return and insertion loss measurements performed on installed cable after two years of service life. In no case was any deterioration in either transmission efficiency or cable uniformity noted.

Our studies have disclosed no indication of the various postulated possible faults which might develope.

1. We have found no indication of cracking dielectric. The material used in Dynafoam is sufficiently flexible for the task.
2. We have found no evidence of inner conductor wandering.
3. Insertion loss measurements after 2 years installation indicate no moisture absorption by the cable.
4. We have found the cable to react predictably to temperature changes.

ECONOMIC FACTORS:

In order to estimate the economic advantages of Dynafoam cable an analysis was performed using the following assumptions.

1. 100 mile system.
2. 70 miles of JT-1412 or JT-2340.
3. 15 miles of JT-1500 or JT-2412.
4. 15 miles of JT-1750 or JT-2500.
5. Substitution of the more efficient cable would not lead to a significant reduction in the number of amplifiers used in the system nor require any excessive variation in cost of installation.
6. Cost savings are based on Times Wire and Cable published price date for 100 mile quantities.

	Alumifoam	Dynafoam
Cost of 70 miles of JT-1412	$26,900.00	
Distribution JT-2340		$22,900.00
Cost of 15 miles of JT-1500	7,520.00	
Sub Trunk JT-2412		6,500.00
Cost of 15 miles of JT-1750	16,650.00	
Trunk JT-2500		9,350.00
Total Costs	$51.070.00	$38,750.00
Savings	$12,220.00	

FOR JACKETED SYSTEM:

70 miles JT-1412-J	$31,200.00	
JT-2340-J		$27,800.00
15 miles JT-1500-J	9,130.00	
JT-2412-J		7,690.00
15 miles JT-1750-J	18,700.00	
JT-2500-J		10,950.00
Total Costs	$59,030.00	$46,440.00
Savings	$12,590.00	

CONCLUSION:

It can be concluded that after over 2-1/2 years of installation experience and after more than three years prior experience in the factory and laboratory, Dynafoam cable represents a progressive step forward in the developement of better, more efficient cable system designs.

It offers the user basic economies for the construction of normal plant and in certain instances, where long trunk runs are required, a method of getting the best signal possible to remote areas.

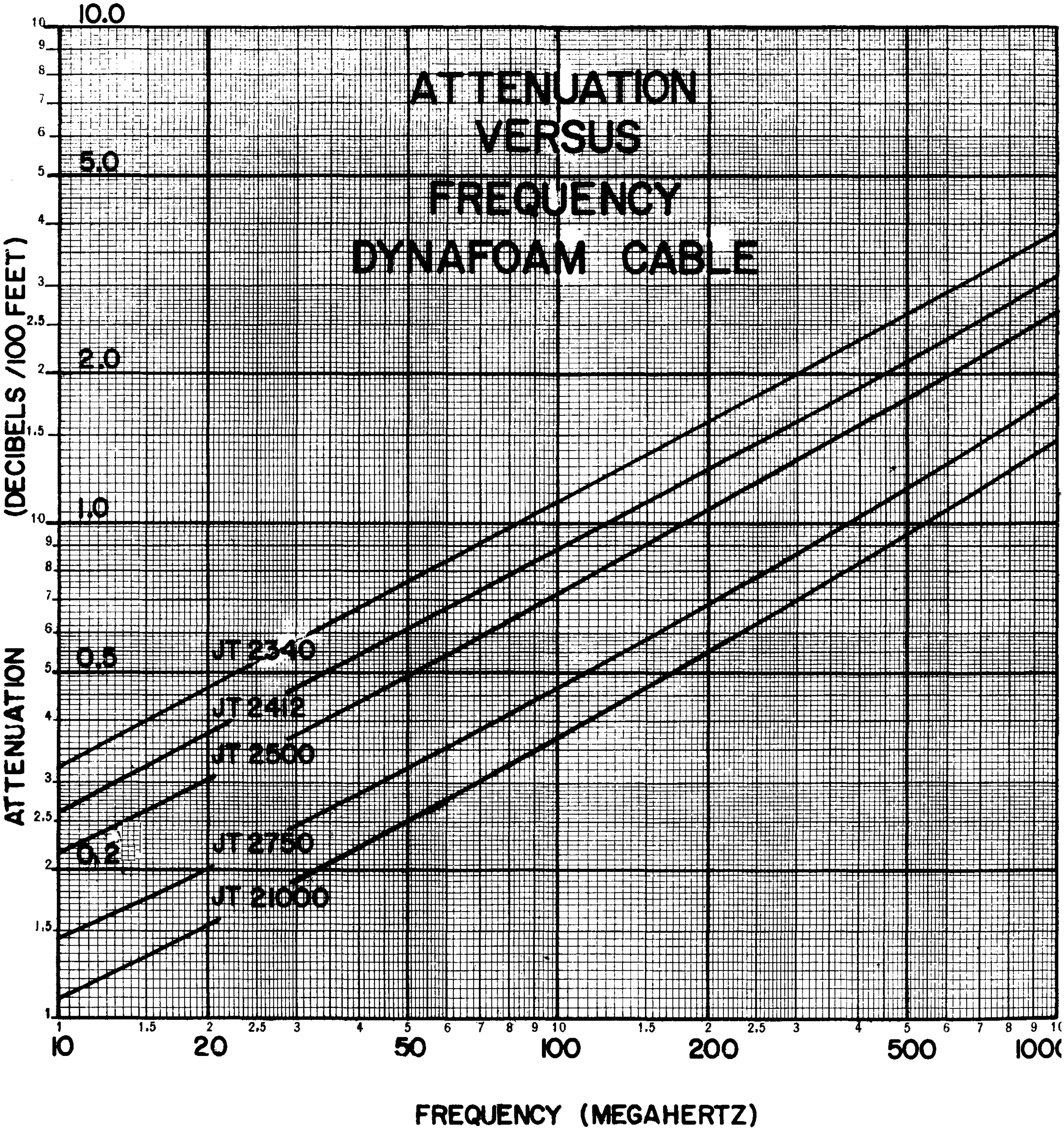
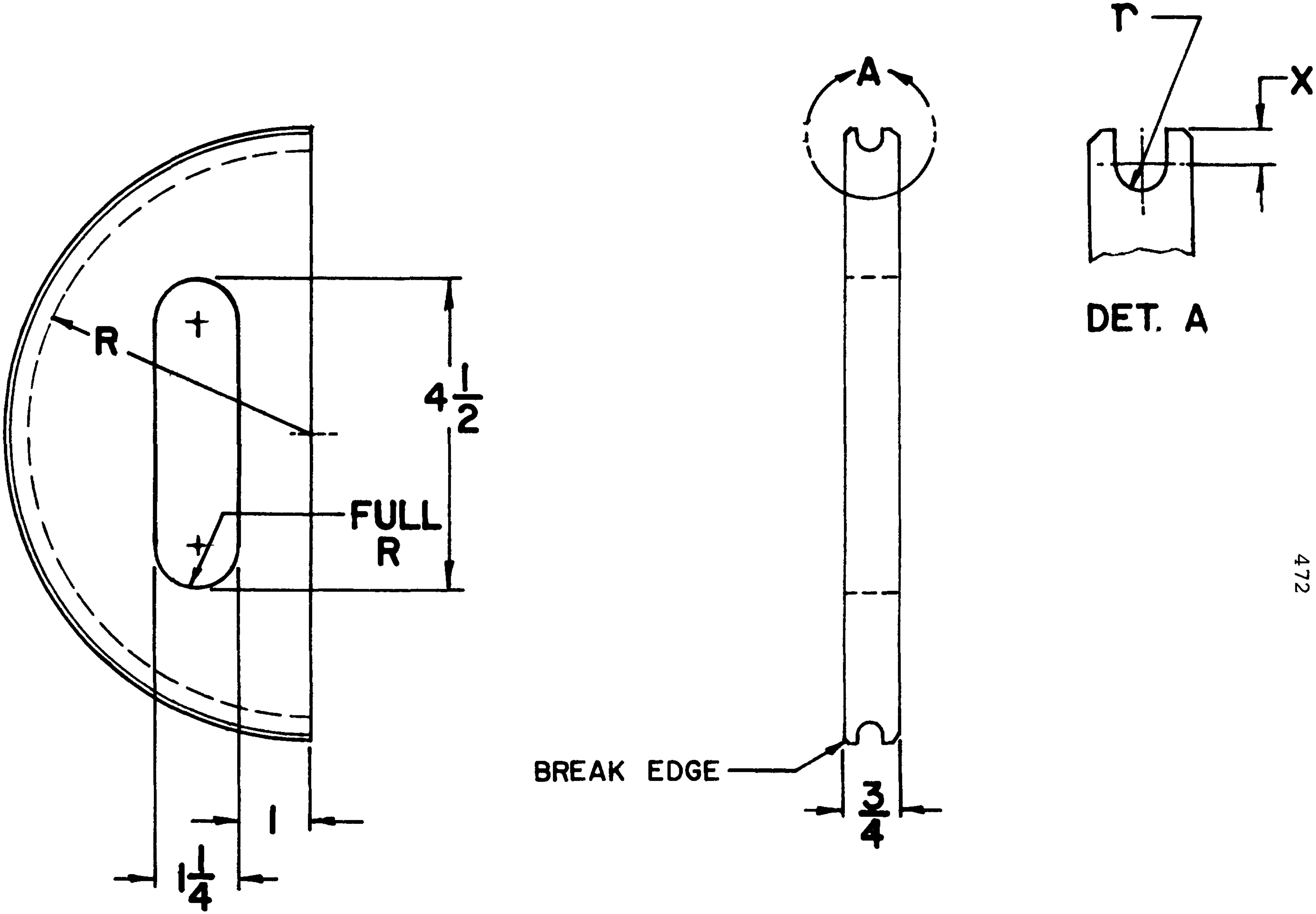

ATTENUATION VERSUS FREQUENCY DYNAFOAM CABLE
ATTENUATION (DECIBELS / 100 FEET)
FREQUENCY (MEGAHERTZ)
10.0
5.0
2.0
1.0
0.5
0.2
10
20
50
100
200
500
1000
JT 2340
JT 2412
JT 2500
JT 2750
JT 21000

Fig. 1

Fig. 2



R = DIA. OF CABLE X 12

r = DIA. OF CABLE +.010

$X = \frac{\text{DIA. OF CABLE}}{4}$

BENDING ARBOR
MAT'L. WOOD

EnDe-CODE

TELEVISION SIGNAL ENCODING AND DECODING

FOR

CABLE SYSTEMS

By

Abraham M. Reiter

Director of Advanced Engineering

Athena Communications Corp.

Prepared for Presentation to

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July 8-9, 1971

EnDe-CODE TELEVISION SIGNAL ENCODING AND DECODING FOR CABLE SYSTEMS

Abraham M. Reiter

Director of Advanced Engineering

Athena Communications Corporation

ABSTRACT

A low cost system for encoding and decoding the picture and sound signals on CATV channels is described. A simple encoder is connected to the modulator for each encoded channel and an all-channel set top decoder can decode all channels--standard and non-standard. No modifications to the cable system are required.

The heart of the system is the patented "gray blank" method of encoding video by replacing the synchronizing and blanking components of the composite video signal with a steady voltage midway between black and white. The color reference burst is superimposed on the steady voltage. Proper synchronization on a normal television receiver is lost because the receiver attempts to lock on the blackest portions of the video signal. The sound is encoded by positioning the FM sound carrier 1 MHz below the video carrier. Pulse signals that are used to restore the video to normal are amplitude modulated on the repositioned sound carrier. Decoding is accomplished without demodulating the video or audio.

SUMMARY

EnDe-Code is a system for transforming any 6 MHz channel into a private channel that is available only to those who have decoders. The system consists of an encoder that is connected to the modulator of the channel to be encoded and a decoder that is connected to the antenna terminals of the television set.

Composite video is modified by the encoder resulting in the gray blank video waveform shown in figure 1. Horizontal sync and blanking and vertical sync (not shown) are replaced by the gray level, i.e. the level midway between black and white. In the modulator, gray blank video is modulated on a 45.75 MHz carrier and converted to the desired output channel in the usual way.

The encoder also takes the frequency modulated 41.25 MHz sound carrier from the modulator, converts it to 46.25 MHz and amplitude modulates video restoring pulses. The 46.75 MHz is sent back to the modulator to be converted to the desired output channel along with the video carrier. The 41.25 MHz is suppressed.

The encoder is provided with a switch that returns the channel to normal service without the need to make any other adjustments or change any cables.

A standard television receiver that is tuned to the encoded channel is deprived of normal sync and receives no sound.

The decoder restores the sync and blanking to the video carrier (45.75 MHz) using the pulses on the 46.75 MHz carrier, and sound is restored by repositioning 46.75 MHz to 41.25 MHz. The restored carriers are converted to a convenient standard channel and delivered to the TV set. The entire decoding process is accomplished without demodulating the sound or video information.

ENCODER

Interconnections between the modulator and the encoder are shown in a general way by figure 2. The encoder is compatible with any modulator that generates standard IF carriers (45.75 MHz video and 41.25 sound) and converts these carriers to the output channel.

Video is connected directly to the encoder but audio is connected to the modulator where it frequency modulates the sound carrier in the normal way. The FM 41.25 MHz carrier from the modulator is diverted to the encoder.

The encoder does three things:

- -it generates gray blank video
- -it converts 41.25 MHz to 46.75 MHz at the same level
- -it amplitude modulates video restoring pulses at a modulation level of 30% on the 46.75 MHz

In the encode mode the encoder delivers gray blank video and 46.75 MHz to the modulator. The 41.25 MHz carrier is suppressed. In the normal mode the encoder delivers normal video and 41.25 MHz to the modulator and 46.75 MHz is suppressed. In either mode there are just two carriers.

The method of combining carriers for conversion to the output channel depends upon the modulator design. If a single mixer is used the lines carrying 45.75 MHz, 41.25 MHz and 46.75 MHz are combined and go to the mixer. If separate mixers are used for video and sound, the 46.75 MHz line may be combined with either of the other two lines.

Manipulation of signals within the encoder is shown in the block diagram in figure 3 and waveforms that aid in understanding encoder operation appear in figure 4.

Gray Blank Video

Gray blank video is generated in a switching circuit. The input video is clamped on the back porch of blanking to provide a reference voltage to which the gray level can be referred. When suppressing pulses are applied, the switching circuit output switches to the gray level. The color reference burst is superimposed on the gray level.

Figure 4 shows several lines of normal video, suppressing pulses and gray blank video starting just before field 1 and ending just after vertical blanking. The start of field 1 is defined by a whole line between the first equalizing pulse and the preceding horizontal sync pulse. All horizontal retrace intervals and the entire six line interval at the start of the field are gray.

Restoring Signal

The restoring signal, also shown in figure 4, is slightly modified sync that is band limited to 200 KHz before it is amplitude modulated on the 46.75 MHz carrier. Aside from band limiting, the difference between the restoring signal and sync is that the three half-line equalizing pulses following vertical sync are not present in the restoring signal. These pulses are left in the video as part of the gray blank video signal.

The band limiting filter delays the restoring signal and further delay occurs in the narrow band circuits in the decoder. To compensate for these delays the encoder advances the restoring signals approximately five microseconds with respect to the suppressing signals.

Front Panel Controls

There are three adjustments to assist in setting up gray blank video. They control the suppressing pulse width, suppressing pulse position and gray level, respectively. A video gain control is provided to set the depth of modulation.

The functional mode of the encoder is controlled by a three-station pushbutton switch. One station selects the normal mode, another the encoded mode and the third switches power off. Indicator lights display the selected operating mode.

DECODER

The decoder operates on signals in the standard television IF band, 41 MHz to 47 MHz, so it is necessary to convert the encoded channel to that band before decoding. After decoding, the restored channel is converted to a standard VHF channel and delivered to the antenna terminals of the TV set. If more than one channel is encoded, the same decoding circuits are used, but a tuner is required to convert each encoded channel to the 41 MHz - 47 MHz band. The output conversion remains the same regardless of the number of encoded channels or their frequencies.

Video Restoring

Figure 5 shows a block diagram of the decoder and figure 6 shows some waveforms that aid in understanding how the video is restored.

The augmenter is a three level RF switch, i.e. a circuit that can be switched to any one of two higher gain (or lower loss) states from a given reference state. If the gain is increased by 4.7 Db, the gray level will become the blanking level and an increase of 7.2 Db results in the sync level.

Two sets of pulses are applied to the augmenter to restore the blanking and sync levels, respectively. The pulses are derived from the restoring signal that is amplitude modulated on the 46.75 MHz carrier. After demodulation the restoring signal passes through a 0-200 KHz filter. Sync is made by clipping and squaring the restoring signal and blanking is made by stretching sync.

The restored video has a front porch that remains gray. Since the front porch occurs before retrace, it is not visible. If the receiver were underscanned, it would show as a gray, rather than black, border on the right hand side of the picture. The gray front porch is just as effective as black as a guard band, preventing the video content from affecting the position of sync in the sync separator circuits.

The color burst is keyed up to the blanking level and its amplitude increases as a result of augmenting. Anticipating this, the color burst is reduced in amplitude in the encoded video.

The bandwidth of the augmenter is 4.2 MHz and there are traps for 46.75 MHz and for adjacent high channel video, 39.75 MHz.

Sound Restoring

A simple heterodyne operation with a 5.5 MHz oscillator is all that is required to move the 46.75 MHz carrier to 41.25 MHz. The 5.5 MHz oscillator is crystal controlled to provide an accuracy of 550 Hz to the repositioned carrier. Tuned circuits in the output of the sound mixer reject 46.75 MHz.

Functional Integration

The decoder shown in figure 5 may be considered as a component that can be integrated with other equipment in a subscriber's home. For example, a decoding module could be built into a converter to provide a very attractive combination. The augmenter is on at all times and functions as an amplifier. The remaining circuits are tuned on for decoding only. A program is now under way to put a decoder module in the Gamut 26 converter manufactured by Oak Electro/Netics.

Another attractive possibility is a transponder with a decoding module. In addition to its other functions, the transponder could also sense when the decoder is on.

SYSTEM CONSIDERATIONS

Security

The security of the EnDe-Code system is adequate to block a number of possible approaches to unauthorized reception:

1. No adjustment of the television receiver can possibly restore the picture or the sound.

2. No commercially available standard equipment can be purchased to decode picture or sound.

3. Although the bill of material is simple, the construction and alignment of the decoder from a kit of parts requires sophisticated equipment and considerable electronic skill.

4. The proprietary nature of the system is an effective deterrent to unauthorized manufacture and distribution of decoders.

Interference

It is essential that adjacent channel interference from an encoded channel to a normal channel, from a normal channel to an encoded channel or between encoded channels should not be greater than that between normal channels.

One general observation can be made. A carrier modulated with gray blank video has a peak power that is limited to the black level which is about 3Db less than sync. On the average it is much lower than that because video signals rarely have sustained intervals of black.

As a result, encoded video carriers will interfere much less with adjacent channels than carriers modulated with normal video. For the same reason, encoded video carriers will contribute less to intermodulation and beats.

In figure 7, the positions of the carriers in four adjacent channels are shown. Two of the channels are encoded and two are normal. The case that appears to have the greatest potential for mutual interference involves an encoded channel and a lower adjacent normal channel because the sound carriers are only 0.5 MHz apart. That case will be discussed in some detail.

Consider first the effect of the normal sound carrier on the encoded channel. In the decoder, the 46.75 MHz amplifier must reject the adjacent sound carrier which appears at 47.25 MHz. Failure to reject 47.25 MHz will result in a 0.5 MHz beat on the restoring pulses. The design of the decoder is such that if the beat is smaller than the restoring pulses, sync will not be affected because the 200 KHz band limiting filter is adequate to remove it, after detection. Since the 47.25 MHz and 46.75 MHz carriers are about equal in level on the cable, very little attenuation of 47.25 MHz is required.

A more subtle, but far more serious possibility results from an undesired output in the sound mixer. The 47.25 MHz combines with 5.5 MHz yielding 41.75 MHz. That is just 420 KHz from the color subcarrier. To insure that it will not be visible requires that it be 40 Db below the video. Since it is 15 Db lower on the cable, the decoder must provide 25 Db of attenuation, which is well within its capability.

Consider next the effect of the encoded sound carrier on the normal channel. The sound in the normal channel will not be affected because the 4.5 MHz sound IF in the TV set can easily reject a signal of equal level that is 0.5 MHz away. To be sure that a visible beat does not appear the encoded sound must be no greater than -8 Db with respect to the video carrier in the normal channel, and the encoded sound satisfies that requirement.

Compatability with Existing Systems

EnDe-Code can be applied to any television channel and all the required signals are transmitted within that channel. This is why it can be installed in any CATV system without changing any hardware from the head end to the subscriber's matching transformers.

Cost

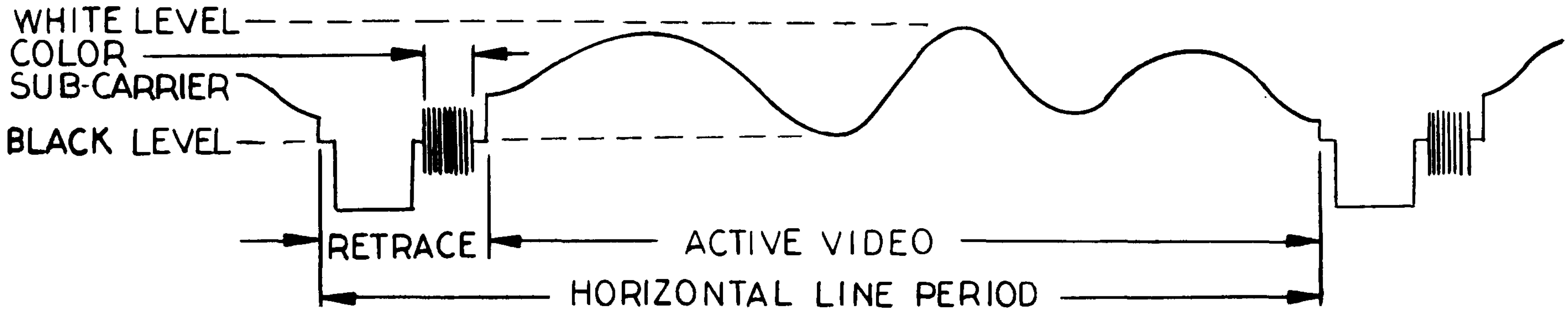
The cost of the decoder is clearly the most important element of the EnDe-Code system cost. The design was guided by the principle that manipulating video and sound signals on modulated carriers and at low level is simpler, less expensive and gives better results than manipulations that require demodulation. The decoder is low in cost because of success in adhering to that principle.

CONCLUSION

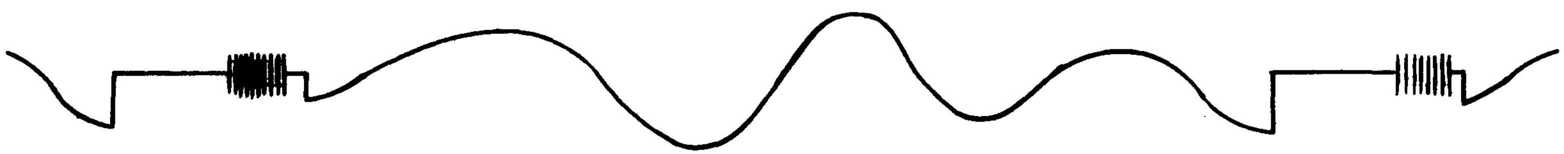
EnDe-Code, a low cost system of encoding and decoding television signals in CATV has been described.

It has been shown that it can be applied to any channel, standard or non-standard, to any number of channels and that a channel is readily switched between normal and encoded service in seconds.

It has also been shown that EnDe-Code does not interfere with normal CATV service or require any changes to the CATV system other than installation of encoders and decoders.



NORMAL VIDEO



GRAY BLANK VIDEO

FIGURE 1 - VIDEO WAVEFORMS

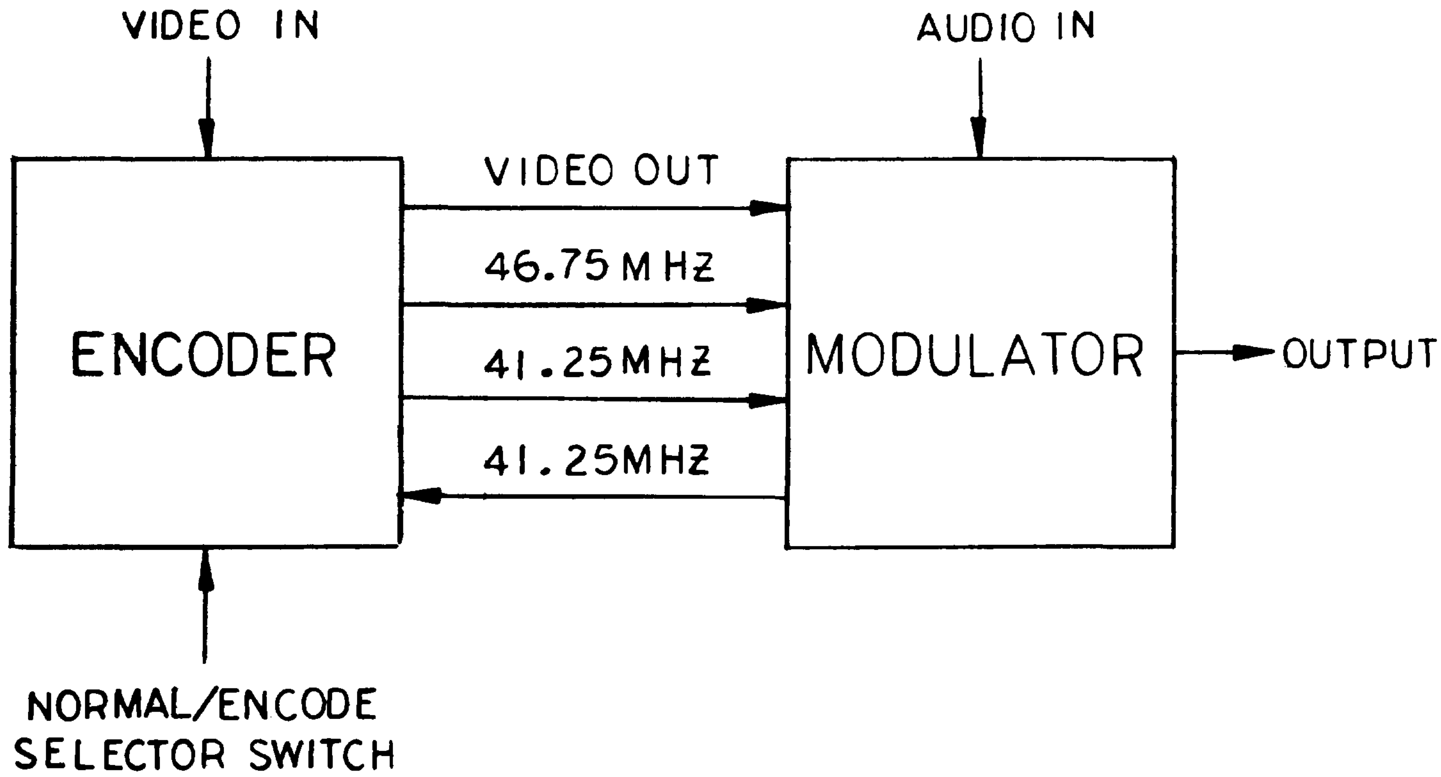


FIGURE 2-INTERCONNECTIONS BETWEEN ENCODER &MODULATOR

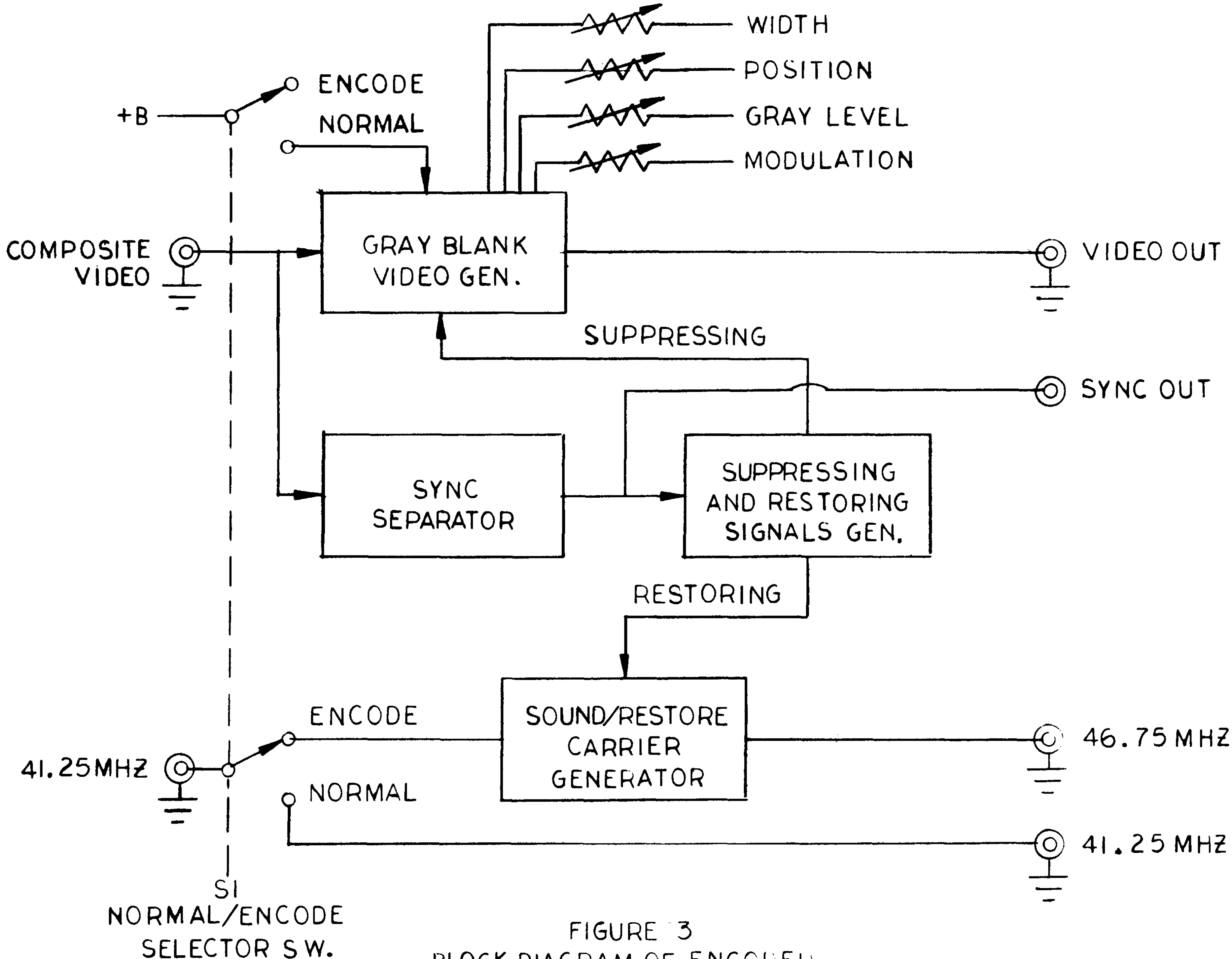


FIGURE 3
BLOCK DIAGRAM OF ENCODER

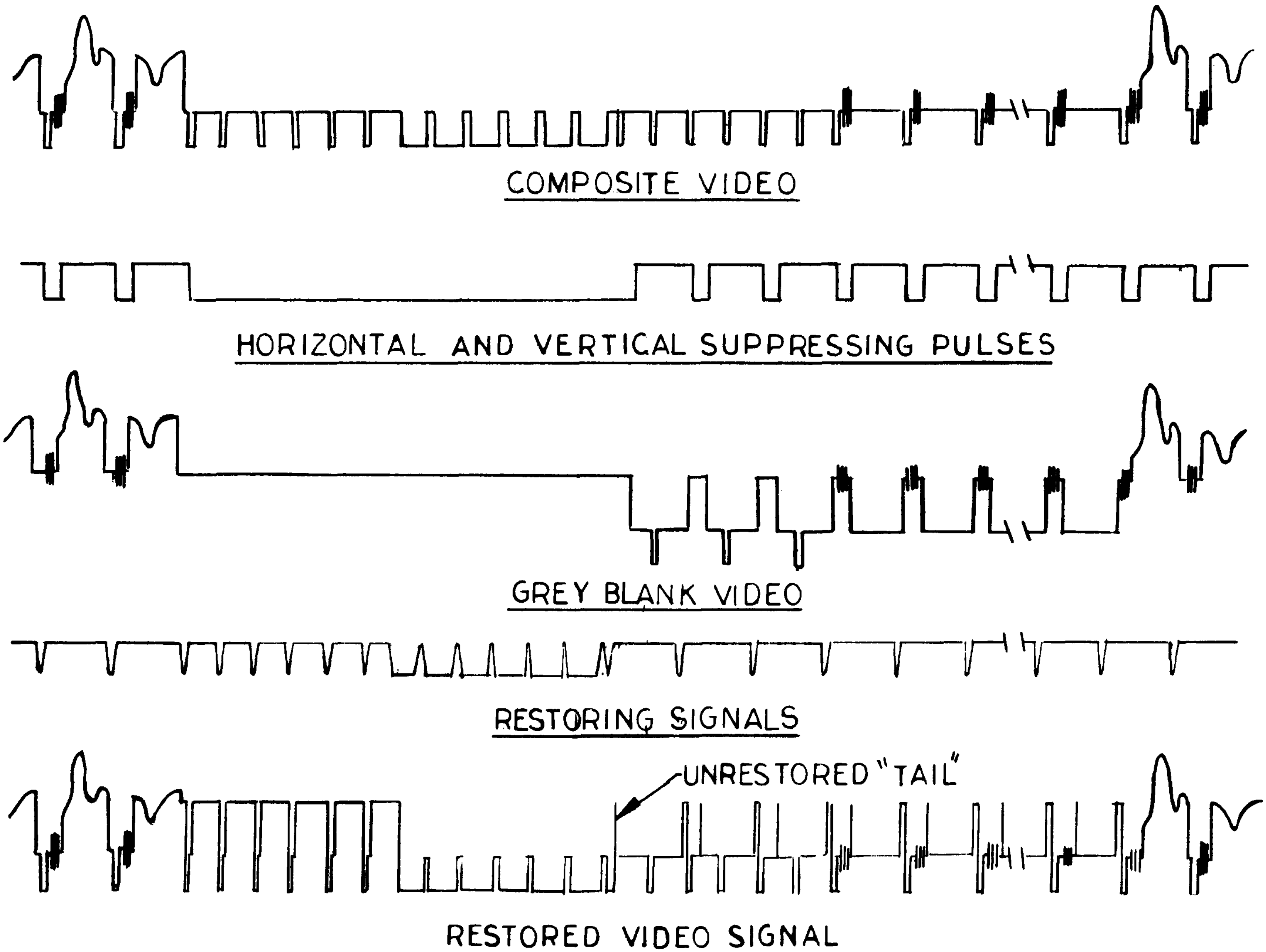


FIGURE 4 - WAVEFORMS

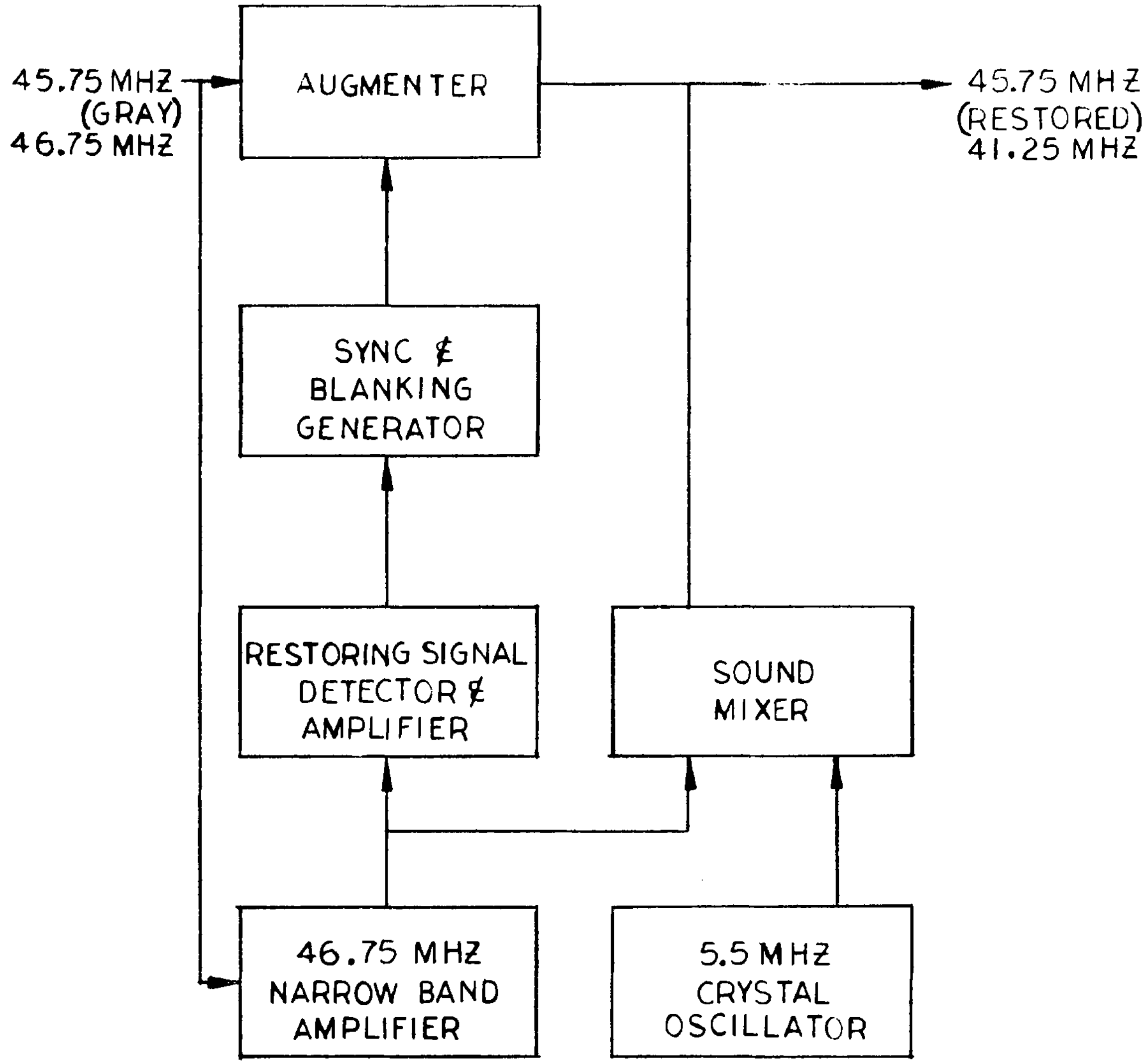
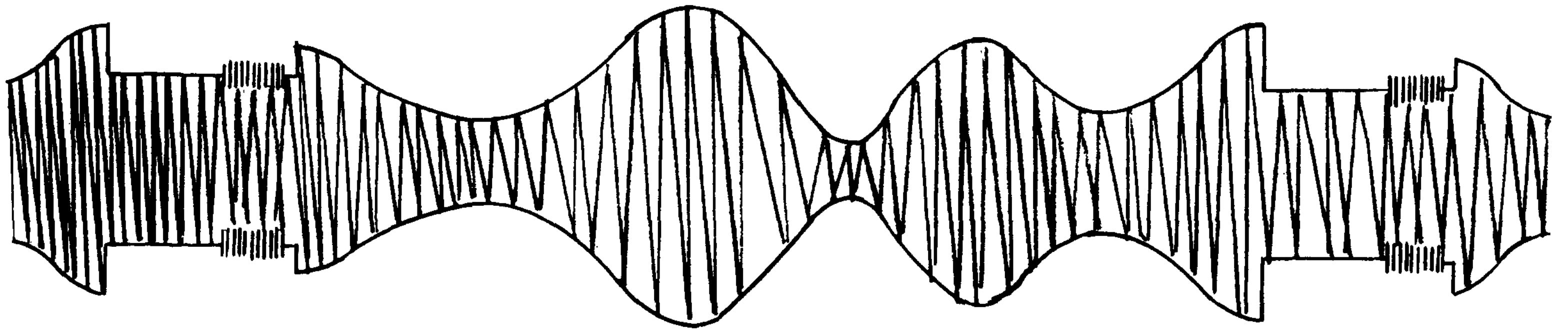
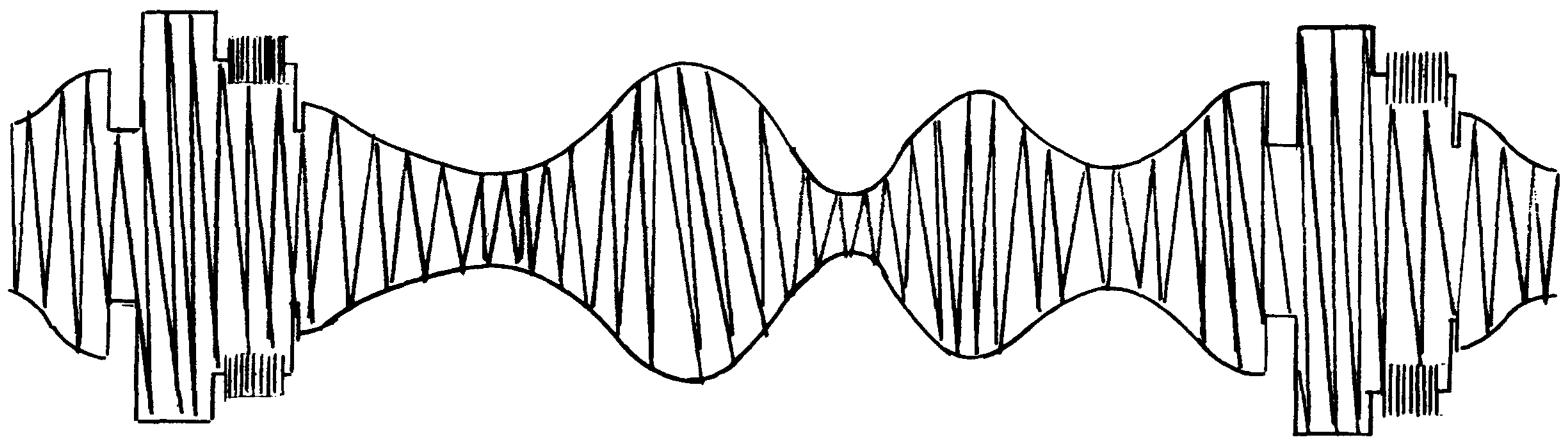


FIGURE 5
BLOCK DIAGRAM-DECODER



GREY BLANK VIDEO MODULATED ON AN IF CARRIER



RECONSTITUTED COMPOSITE VIDEO SIGNAL ON AN IF CARRIER

FIGURE 6-VIDEO WAVEFORMS

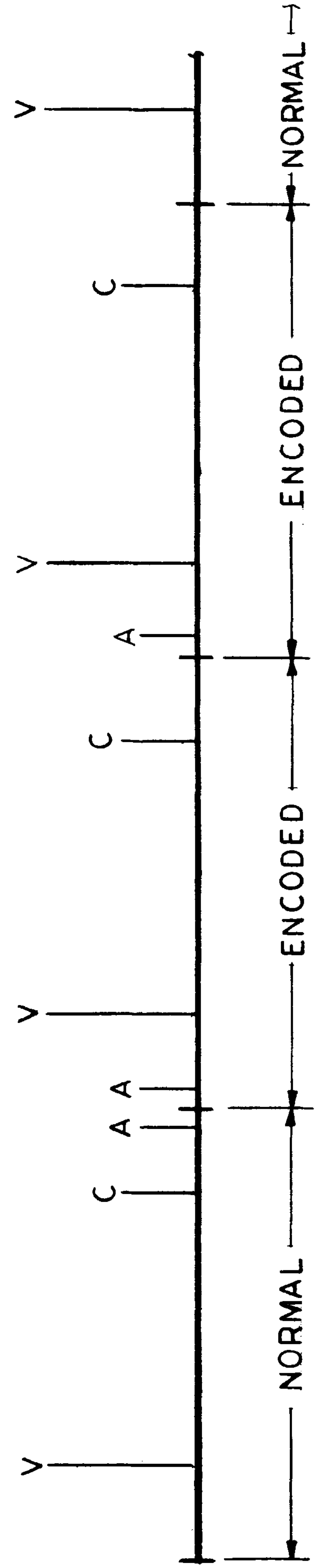


FIGURE 7-POSITION OF CARRIERS

ENVELOPE DELAY, PHASE DELAY, GROUP DELAY, CHROMA DELAY.... WHAT DOES IT MEAN, HOW IS IT MEASURED?

Fred J. Schulz, V.P. CATV Engineering
STERLING COMMUNICATIONS, INC.
43 West 61st Street
New York City, N.Y., 10023

Contents

1. Introduction
2. Types of distortion
3. Phase shift and delay
4. The make-up of waveforms
5. Waveforms through filters
6. Instruments to measure envelope delay
7. Chroma delay
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1. Introduction

The CATV industry is still growing, and in its path to maturity it is becoming more sophisticated technically. While in the early days of CATV it was sufficient to get a picture to the subscribers' TV set one is now confronted with meeting federal specifications. A number of new terms have crept into the CATV engineers vocabulary, some of which may be familiar, some of which may however have been misunderstood. The following presentation focuses on giving simple (and in some instances simplified) explanations of terms such as "envelope delay", "chroma delay", etc. An attempt will be made to explain the effects on picture quality. Appropriate test equipment, test methods and cost are explained. The explanations will be held in simple form so hopefully the average CATV engineer can follow; mathematics will be avoided.

The first part is devoted to the introduction of some necessary fundamental know how, and may be looked upon as a refresher for those already familiar with it.

2. Types of distortion

The function of a CATV system is to bring high quality undistorted pictures to a subscriber's TV set. The CATV engineer is usually familiar with distortions such as: noise, cross-modulation, second order distortion, echoes, amplitude distortion, etc. Cross-modulation and second order distortion are so called non-linear distortions. Amplitude distortion does usually not depend on signal levels; it is called a linear distortion.

Amplifiers and filters may also exhibit phase distortion, another linear distortion. Phase distortion is of prime concern in single channel and studio equipment, and will be explained here, since the CATV engineer seems to be least familiar with it.

3. Phase shift and phase delay

Phase shift is most readily explained with an experiment; let's feed a CW signal into a low pass filter and observe the input and output wave forms on a dual channel oscilloscope:

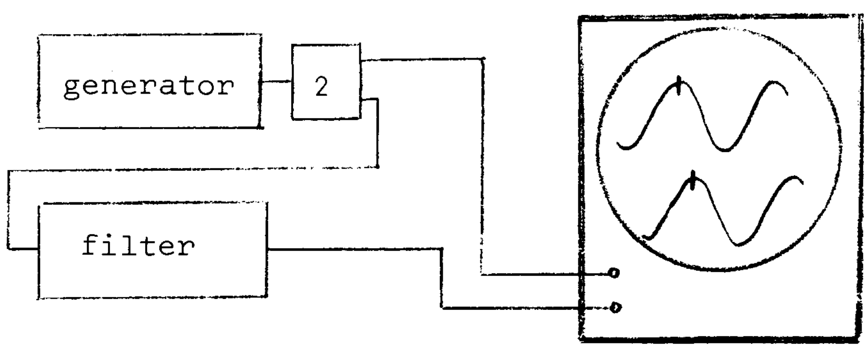

Fig.1

We find that the output waveform is not in phase with the input (the crests of the waves are shifted with respect to each other). We say there is phase shift. In this case the output lags the input wave. Phase shift is measured in degrees. One full cycle of a wave is 360°.

Phase shift occurs in any network containing capacitors and/or coils.

Another way of interpreting this effect is to say that the output is delayed compared to the input signal. Physically this makes sense, since it obviously does take a finite time for the signal to pass through the filter. If we express the phase shift as delay time we speak of phase delay. Phase delay = $\dfrac{\text{phase shift in degrees}}{360 \times \text{frequency}}$

TV waveforms are not as simple as the foregoing sinewave signal, but the basic principle still hold. We should be aware that the phase shift of a wave through a filter is usually not constant as the input frequency is moved through the range of the filter and particularly as the frequencies pass through the cut-off region. If we plot the phase shift versus frequency we may get a curve as follows:

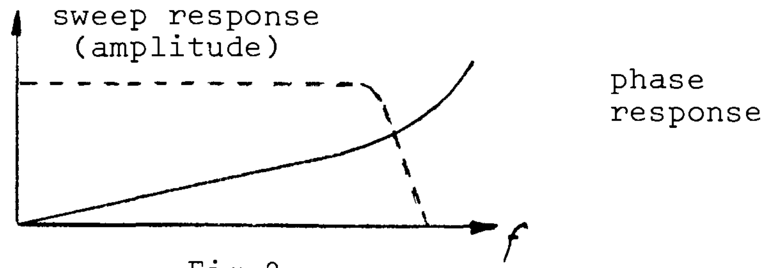

Fig.2

To be more specific we should label the sweep response amplitude response; after all the familiar sweep generator setup measures the amplitude versus frequency characteristic. The phase characteristic is seldom measured, but we believe this will become more common place in the future, particularly for single channel devices.

4. The make-up of TV waveforms

A square wave is a familiar waveform to all of us, if fed into a modulator a series of black and white bars will appear on the screen.

The square wave can be understood better if it is broken down into its components. This is called "Fourier Analysis", but let the word not frighten you, the basic principles can be well understood without the rather advanced mathematics usually associated with it.

Let us feed a square wave through a bandpass filter, one which will just pass the signal equivalent to the fundamental frequency. The resulting output is a sinewave.

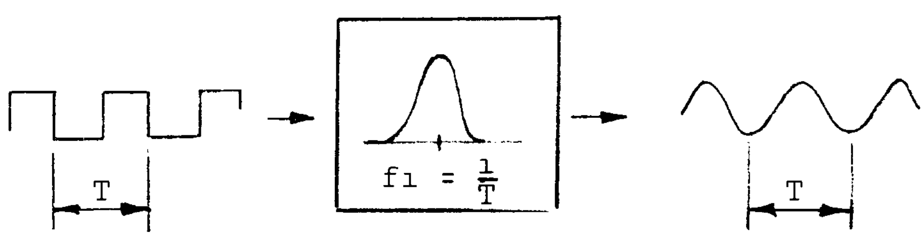

Fig.3

We also get a sinewave output if we feed the square wave through a filter with a bandpass for three times the fundamental frequency.

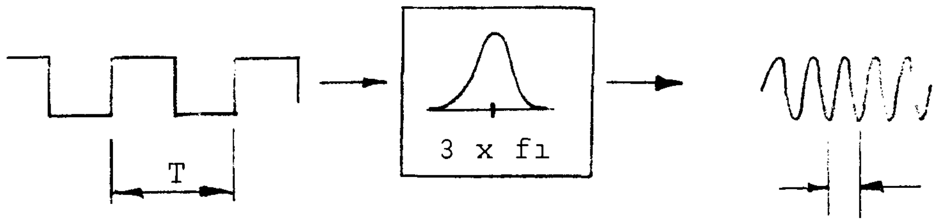

Fig.4

The same will happen if we use a filter for 5 times the fundamental, also for 7 times, 9 times, etc.

We can say that the square wave is made up of a fundamental and all odd harmonics. If the square wave has truly "sharp" corners and vertical sides we will get harmonics at <u>all</u> odd multiples of the fundamental up to infinity. The fundamental is the highest amplitude component, higher harmonics are of decreasingly lower amplitude.

We just "disassembled" a square wave into its sine-wave components. Conversely we can reassemble the wave from its sinewave components. Graphically this looks like this:

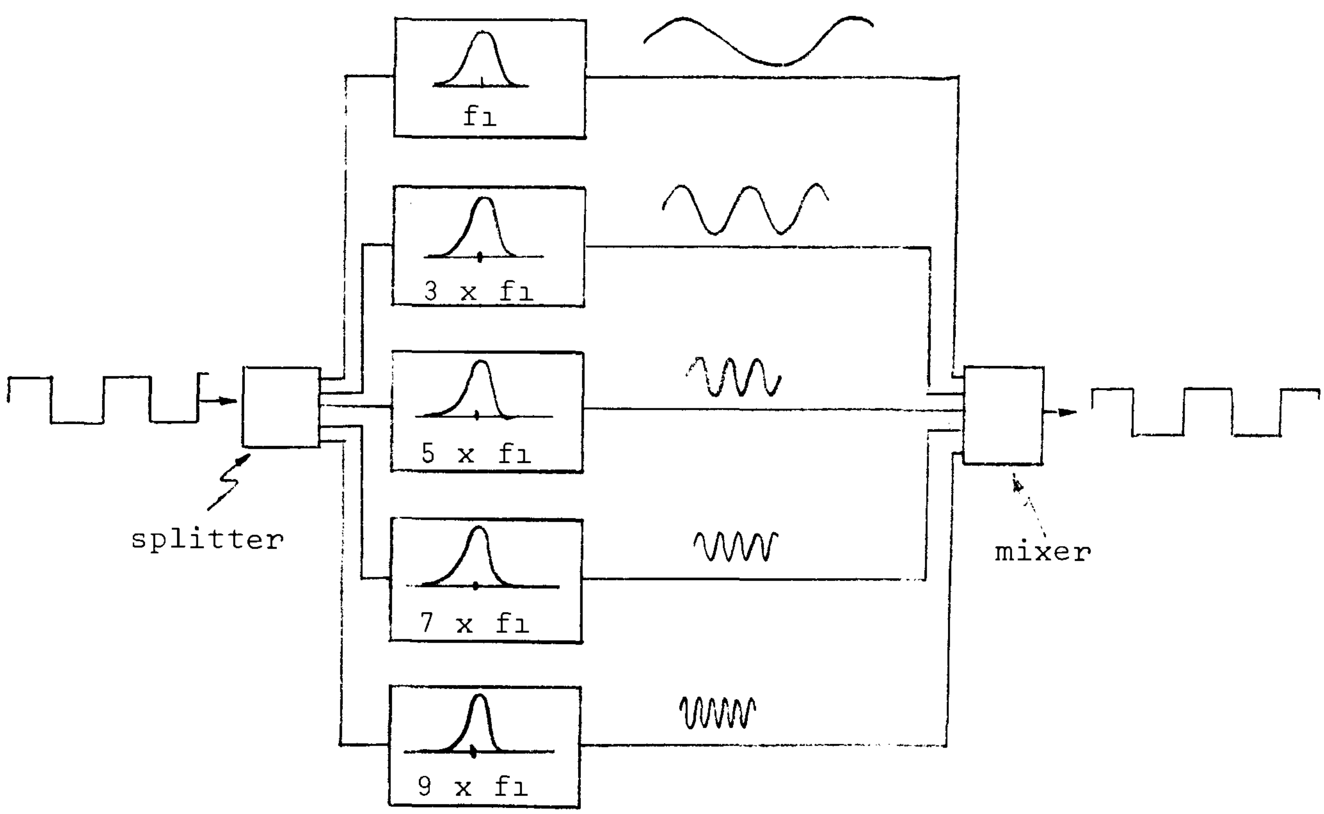

Fig.5

The following picture shows the fundamental and the first three harmonics of a square wave:

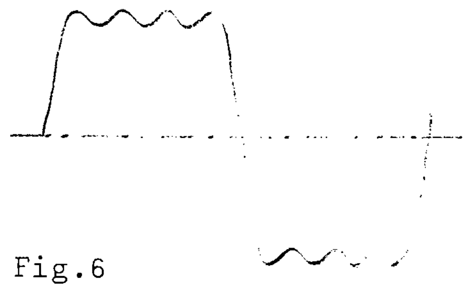

Fig.6

The result is not a very good square wave since one should also add the 9th, 11th and all other odd harmonics to make a true square wave. It should also be noted that a square wave does not contain any even, such as 2nd harmonics. The principle, however, holds.

Any waveform can be made up of harmonically related sinewaves. Another rule worth remembering is:

> " A waveform with very abrupt amplitude changes contains significant components up to many times its fundamental frequency."

5. Waveforms through filters

Let us look at the previous example again, but let us feed the signal through a low pass filter which passes only the fundamental, the 3rd and 5th harmonics, but cuts off the 7th harmonic:

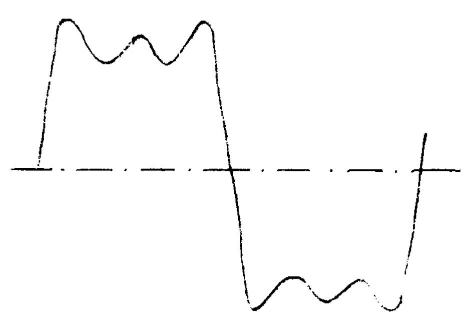

Fig.7

We see that to faithfully reproduce the input waveform one must pass all significant frequencies.

As the next experiment let us pass the signal through a filter which attenuates the 5th harmonic, such as a trap would:

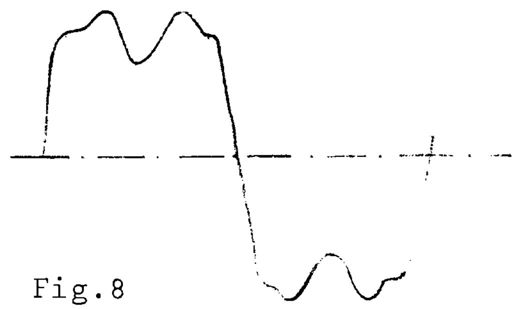

Fig.8

So far we concerned ourselves with changes in amplitude of the various frequency components only.

We indicated earlier that signals passing through a filter also get a change in phase or a delay, and further that not all frequencies may suffer the same delay. Let us pass the sample wave through a filter which shifts the 5th harmonic by 15° and the 7th by 30° with respect to the fundamental:

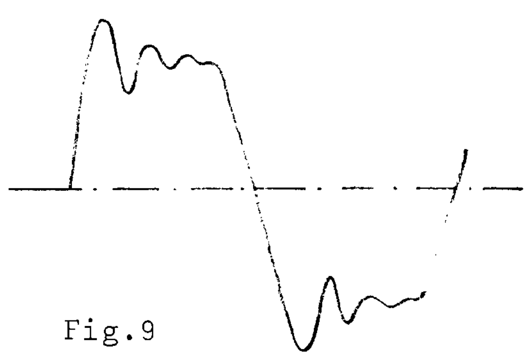

Fig.9

The output is non-symmetrical with a very pronounced ringing. As a further example let us shift the 3rd harmonic by 30°, the 5th by 60° and the 7th by 90°:

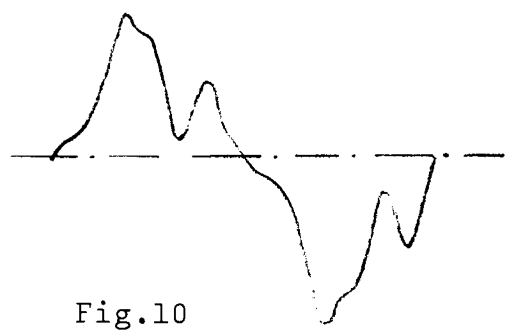

Fig.10

From these examples it is obvious that we <u>must retain the relative positions of the signal components or distortions will result</u>. The delay of the various frequency components, also called a group of frequencies, must be equal for distortion free transmission. If this is not the case we speak of group delay distortion or simply group delay. The group of frequencies, if applied to a modulator, will result in a carrier and associated sideband envelopes. The envelope which contains the modulation must be passed distortion free or we speak of envelope delay distortion. The terms envelòpe delay and group delay are generally used interchangeably.

For the sake of completeness let us also state the accepted definition of envelope delay:

> "Envelope delay is the rate of change of the phase versus frequency curve."

Mathematically expressed:

$$\text{Envelope delay} = \frac{d\phi}{d\omega}$$

Delays are usually so small that they are expressed in microseconds (millionth of a second) or nano seconds (billionth of a second).

The effects of envelope delay result generally in ringing (preshoot, overshoot) producing closely spaced ghosts particularly visible on vertical black/white transitions. Another effect, color misregistration will be treated later.

6. Instruments to measure envelope delay

Envelope delay measuring equipment is built to measure the change in phase a sample signal suffers when passed through a device under test. A block diagram of such an instrument follows:

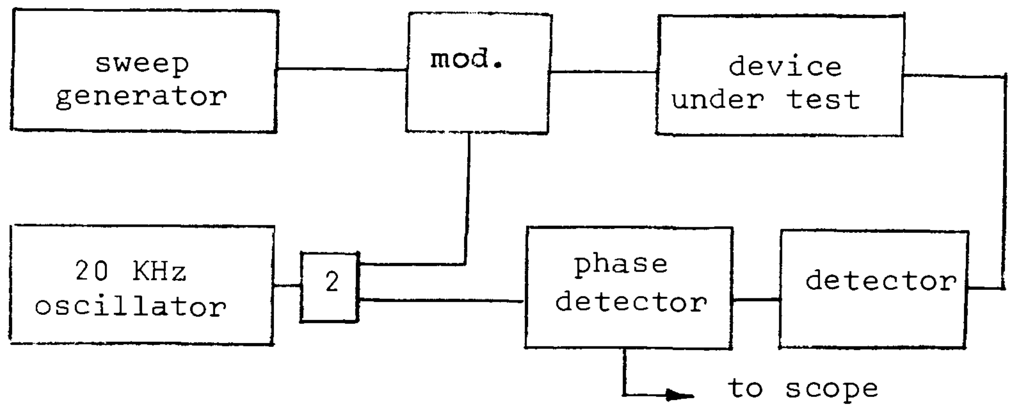

Fig.11

The following is a list of instruments now on the market with approximate prices (the list may not be complete):

1. Rhode & Schwartz - Model LFM 0.1 - 10 MHz	$6,300.- plus sweep gen. & oscilloscope
2. Wandel & Golterman - Model LO-1 0.1 - 14 MHz	$17,250.- complete setup
3. Datatek - Model D-700 0.05 - 50 MHz	$3,200.- plus sweep gen. & oscilloscope
4. RCA - Model BW-8A	$3,100.-
5. Hewlett-Packard - Model 3700 45-95MHz microwave analyzer (IF)	$7,500.- complete setup
6. Hewlett-Packard - Model 8405 Vector Voltmeter	$2,850.-

It should be noted that most listed equipment covers video frequencies only and additional modulators and demodulators must be used to cover RF frequencies.

It is not the purpose of this paper to give detailed information and exact cost of test equipment packages, but here are some comments on the various products.

The Rhode and Schwartz equipment package is by many regarded as the "Cadillac" of delay measuring gear. The full package to perform measurements from video frequencies to 250 MHz costs in excess of $10,000.-. Such a setup is primarily intended for lab use.

The Wandel & Golterman equipment is restricted in its frequency range to video, but can measure envelope delay where transmitter and receiver are separated such as in microwave systems.

The Datatek unit is a relatively new unit on the market and is aimed at the broadcaster since it covers the video and the new IF modulator frequencies. It contains a sync and blanking generator to make measurements in video amplifiers with DC restoration. It is also equipped to measure delay at various average picture levels and features 50 and 75 ohm impedance levels. This unit may find use in some CATV systems.

The HP microwave analyzer is aimed at testing microwave links and accomodates basically a 70 MHz IF frequency range.

The RCA unit is primarily intended for use with TV transmitters, it is one of the older designs.

The Vector Voltmeter needs additional external equipment to make it suitable for swept envelope delay measurements.

The listed equipment is intended to measure delay continuously with appropriate sweep generators. The results are displayed on an oscilloscope similar to the familiar amplitude sweep response. Most instruments allow for the simultaneous display of both amplitude and delay as shown below for a bandpass filter.

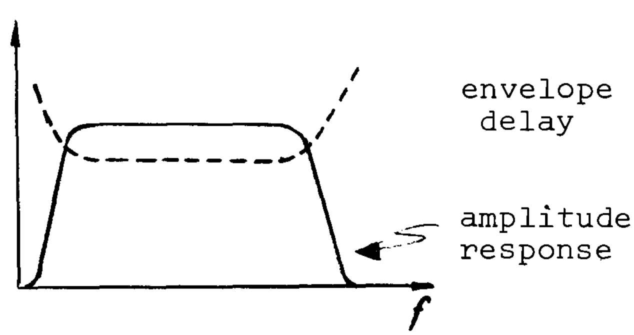

Fig.12

Delay measurements also require very good impedance matching all around; because mismatch causes reflections with resulting ghosts, and ghosts are a delay phenomena also.

To measure delay at RF frequencies one needs a modulator and a demodulator, which by themselves already exhibit some delay. The best modulator demodulator combinations available today have a residual delay of ± 20 nsec. over a 4 MHz video band pass.

7. Chroma delay

A standard color TV picture consists of a high definition black and white picture with a low definition color picture added to it. The color information is carried at the upper end of the video spectrum around a 3.58 MHz color subcarrier. The frequency distribution looks like this:

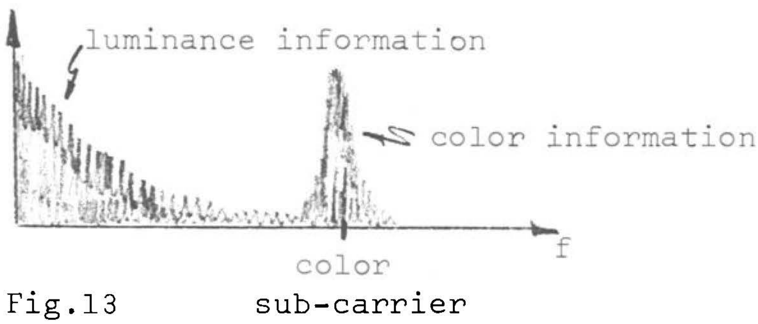

Fig.13

It is important that the color information arrives at the picture tube at the same time as the black and white picture or the color will be out of registration. This error can be called chroma delay. In reality it is simply the envelope delay between the low frequency black and white picture components and the delay of the color information around 3.58 MHz, as shown below:

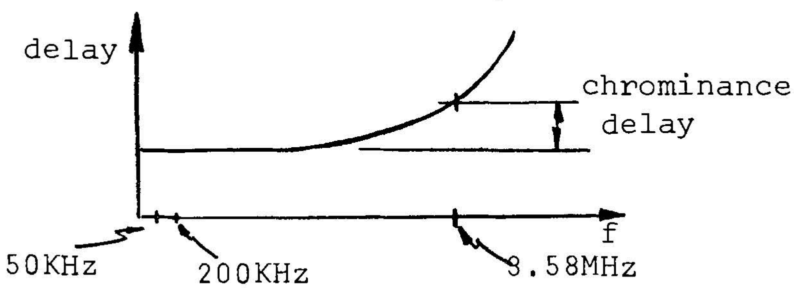

Fig.14

Chroma delay is one of the most visible effects of envelope delay and methods to measure it rapidly have been developed.

The 20T pulse has gained wide acceptance for measuring chroma delay; it is a low frequency pulse modulated with 3.58 MHz color subcarrier signal. Its shape and frequency components are pictured below:

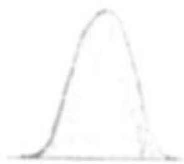

Fig.15

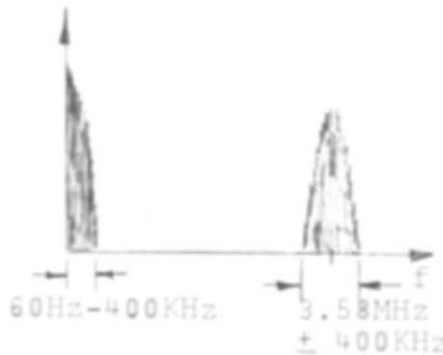

Fig.16

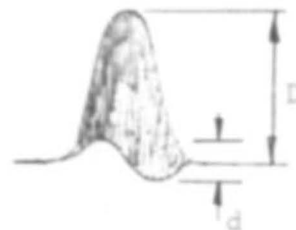

Fig.17

The base line of this pulse gets distorted when the signal is passed through a device with chroma delay. It is possible to determine the delay by measuring pulse height (as referenced to the bar signal level) and the base line excursion using graphs or formulas.

Several manufacturers offer instruments that allow the introduction of delay of opposite polarity and of 3.58 MHz gain/loss to straighten the base line; the introduced delay is then a direct measure of the chroma delay. Instruments to generate 20T pulses cost around $1,500.- usually coupled with generation of other test pulses. The cost of the special receiver described above is approximately $1,500.-. The 20T pulse was originated in Europe and is in wide use there; U.S. networks are not using it yet for transmission over the air.

Tektronix is now featuring a 12-1/2T pulse which yields a wider frequency spectrum around the color subcarrier to be more representative of the actual color information bandwidth and to yield easier computation of the chroma delay from the measured base line excursion.

Tektronix has introduced a way to measure chroma delay using the color bar test pattern. This method makes use of the fact that the chroma signal transition between the green and the magenta bars is easily identifiable because the phase changes 180°, and the luminance changes level as well. The delay can then be measured on a waveform oscilloscope. Luminance/chrominance cross talk effects are also eliminated if Tektronix generator, model 146, is used, since the chrominance and luminance signals are available separately. A scope presentation looks like this:

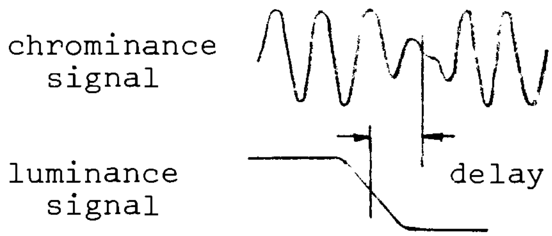

Fig.18

The 20T (or 12-1/2T) pulse measures the "average" delay between the low luminance frequencies and the color subcarrier region. Some people have reported good agreement between 20T pulse measurements and continuous measurements with swept envelope delay equipment, others have found discrepancies. Discrepancies may possibly happen when the delay curve is very non-symmetrical about the color subcarrier. Another effect of non-symmetrical delay is color quadrature cross-talk, which results in color boundary effects.

Color misregistration is also called funny paper effect. Opinions on how much chroma delay is tolerable vary widely in industry. A figure of 200 nano seconds results in a shift of the color by approximately 1/16" on a 23" TV set. At least one CATV manufacturer considers this tolerable. A major test equipment manufacturer feels that 250 nano seconds yields an untolerable misregistration. A member of the Philips research laboratories considers 50 nsec. chroma delay just perceptible under studio conditions. The subject matter is of great importance. Shift of red letters on white background are more readily visible than e.g. blue letters on white. A recent study made by Bell Laboratories, showing color slides on a video monitor to a group of trained observers found that 100 nsec. of flat delay or 180 nsec. of shaped delay was just perceptible. For no objectionable impairment the figures were 260 nsec. for flat delay and 480 nsec. for shaped delay. Shaped delay means delay gradually rising (or falling) from the low frequencies towards the color subcarrier region. Flat delay means constant delay over the subcarrier region; color processors, which separate luminance and chrominance signals may exhibit flat delay.

About the only thing everybody agrees to is that it is highly desirable to keep the delay as small as possible, since delays are additive.

8. Effect of delay

We already treated the special case of chroma delay. Delay effects at other frequencies result mainly in waveform distortions. We saw earlier that a waveform can be distorted by deleting higher frequency harmonics or by delays. A pulse which is representative of an actual picture waveform as produced by the scanning beam in a TV camera, and which does not contain frequency components above 4 MHz is the so called 2T sine squared pulse. This pulse is transmitted by many TV stations as a vertical interval test signal. The pulse is 250 nsec wide at half amplitude. When envelope delay is present, the pulse gets distorted as follows:

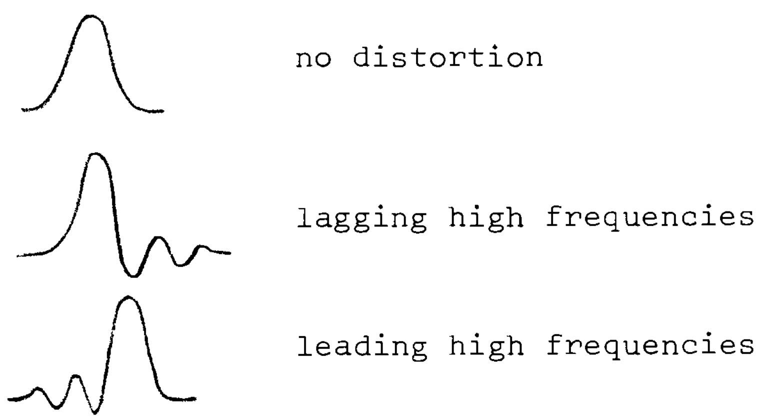

Fig.19

A multitude of distortion patterns are possible. To judge the acceptability of a distorted pulse, a frame or window has been devised, the so-called K-factor graticule.

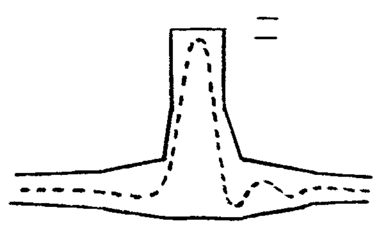

Fig.20

If a pulse is within the outlines of a 2% graticule it is considered very good, 4-6% is acceptable. There is no total industry agreement on the K-factor and a tie-in with actual numbers of envelope delay has not been possible. The K-factor has been treated extensively in the literature.

9. Delay effects in 2-way filters

Two-way filters are in essence a combination of a low pass and a high pass filter with a sweep (amplitude) response as follows (both LP and HP sections shown together).

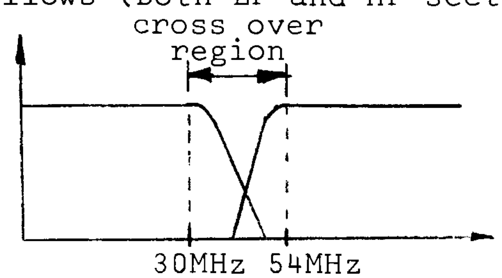

Fig.21

We stated earlier that sharp amplitude response changes are likely to bring delay changes with it. It is easily seen that the narrower the cross over region is made, the sharper the cut-off rate must be. A narrow cross-over region is desired to make the sub-channel return band as wide as possible.

A compromise between usable bandwidth and tolerable envelope delay at the edge channels must be made. There is some specmanship taking place, but most manufacturers feel that 30 MHz is the highest usable sub-channel frequency. Delay effects are cummulative, 50 amplifiers with filters at the input and output result in 100 filter delays all added. If we consider 200 nsec of chroma delay tolerable, each filter cannot contribute more than 2 nsec. The designers of two-way gear are well aware of this problem. To get more reliable measurements one tests a number of filters in cascade, commercially available delay test gear cannot resolve nano second delays very accurately. A fair number of systems have used sub-channels to feed signals back to the head-end by using either Blonder-Tongue model MSVM, Jerrold FCO-47 or similar filters. These filters have a very narrow cross-over region and should not be cascaded in large numbers. A B-T model MSVM was measured at 8 nsec chroma delay at channel 2.

Most of the cross-over filters are of a design with a smooth envelope delay curve so it seems that the chroma delay is a good indication of delay performance. This is not the case with sharp band pass filters however, where the delay may vary in a ripple like fashion.

10. Measured delay curves of some CATV equipment

Fig.22 shows amplitude and envelope delay of a sharp band-pass filter with a trap for the adjacent upper picture carrier. Fig.23 shows a band-pass filter without an adjacent channel trap. It should be noted that the amplitude response over the channel looks reasonably identical for both filters, the delay curves however, do not. The delay distortion below the picture carrier has not received as much attention as the delay at the upper band edge. There seems to be a new awareness of this fact and future equipment designs will take this into account. Delay errors below the picture carrier can produce leading pre-shoot and a following slow approach to final value at sharp video transitions, resulting in a smearing effect on vertical edges.

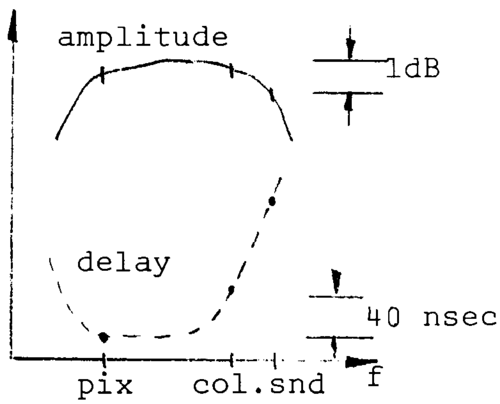

Fig.22

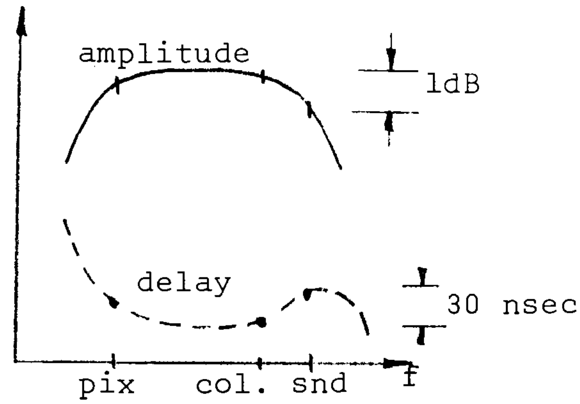

Fig.23

Fig.24 shows the amplitude and delay response of a strip amplifier with less selectivity than the previous band-pass filters. Envelope delay is very small. Fig.25 shows the delay of a narrow band trap tuned for maximum signal rejection (60dB down). Tests on this trap showed a great variation in envelope delay with very small changes in tuning. The same is true for sharp band-pass filters. Test equipment found in the average CATV shop is not good enough to align sharp filters reliably.

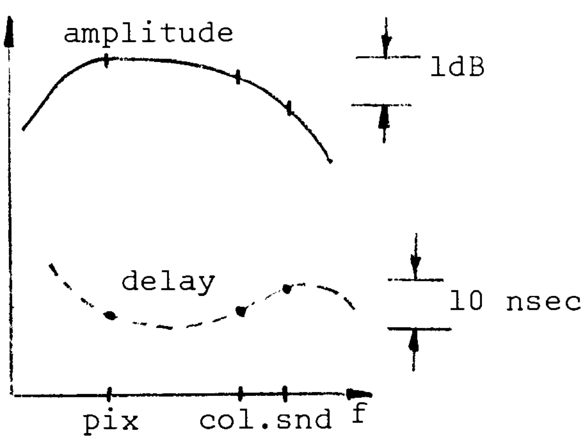

Fig.24

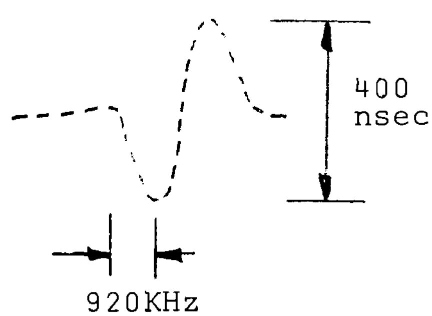

Fig.25

11. The broadcaster and envelope delay

Filters with sharp cut-off have particularly bad envelope delay problems unless compensated for by using extra filter sections. Compensated filters are costlier.

A TV set IF amplifier must be reasonably "sharp" to suppress adjacent channels, it also contains a sound trap, plus traps for the suppression of adjacent channel carriers. The IF does introduce a certain amount of envelope delay. When the currently used color system was proposed by the National Television Standards Committee (NTSC) in the early 1950's, it was proposed to pre-distort the transmitted signal rather than to include costly delay equalizers in every TV set. The recommendations resulted in the currently used transmitter delay curve shown below:

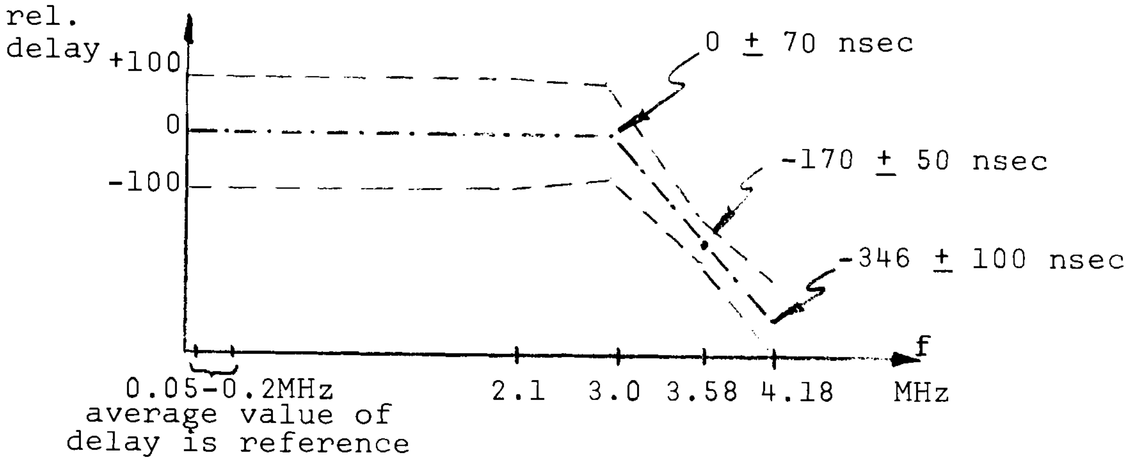

Fig.26

This pre-distortion is exactly the opposite of a "typical" TV set, which has a rising delay curve from approximately 3 MHz to the upper band limit of 4.18 MHz. There were relatively few color TV sets measured to arrive at the so called "typical" receiver delay curve. The response characteristic of a TV receiver is not covered by regulation nor by industry standards. Recent measurements on a number of receivers has indicated a wide range of envelope delay curves. TV receivers are usually designed for a good compromise between selectivity, transient response, color response, cost and other factors. Some people feel that the present transmitter delay curve should be changed; however, until so done one must adhere to it.

A broadcaster is obliged to pre-distort the radiated signal per FCC specs. This pre-distortion is taken care of by type accepted filters and by measurements during the manufacture of the transmitter. Proof of performance measurements, made when a transmitter first goes on the air do not require the measurement of delay. Many broadcasters use adjustable equalizers to make up for delay deficiencies in the transmitters. Equalizers of this kind cost several thousand dollars.

All video signals fed to the transmitter are automatically pre-distorted, however, the signal may suffer delay distortion in the studio or on its way to the transmitter. There are no rules on the maximum allowable delay distortion for signals leaving the studio. It is then perfectly possible that an over distorted signal is transmitted. Color processors used in studios often separate the luminance and chrominance signals with the possibility of introducing delay (or advance) in either luminance or chrominance. Video tape recorders can be a source of delay distortion. Many of you undoubtedly have observed delay effects only on some programs of a given channel and not on others. Efforts are currently underway by broadcasters to tighten color broadcasting standards.

12. Envelope delay in FM equipment

FM receivers should have linear phase response (no envelope delay) otherwise FM signals are distorted. Some recently introduced FM gear payed particular attention to delay response in the IF amplifier with resulting superior performance, such as low harmonics and intermodulation. The achieved performance would probably be noted by the real Hi-Fi bug, but not by the average listener. Present FM head-end gear is designed along more conventional lines with typical Hi-Fi gear performance.

13. Delays in trunk and distribution plants

It was pointed out earlier that delay distortion and rapid amplitude changes such as found in single channel equipment go hand in hand. A CATV plant is wide band and accordingly has little envelope delay. Jerrold showed the following delay curve of an 18-amplifier cascade.

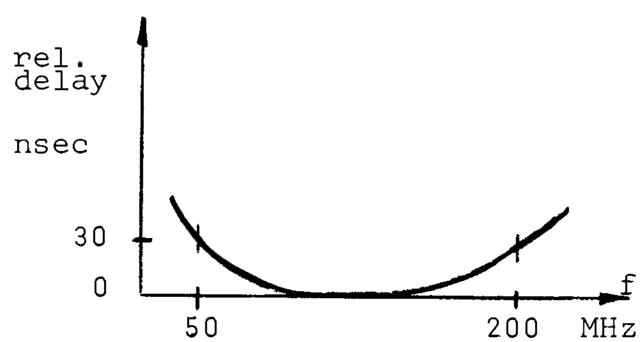

Fig.27

Hewlett-Packard in their appl. note #92 show the delay of a single CATV amplifier as 2 nsec from 54 to 120 MHz, and within 1 nsec from 120 MHz to 216 MHz.

14. Regulations

The proposed technical standards for CATV do not mention envelope delay specifically. The standards however, clearly indicate that the FCC wants the CATV system to faithfully reproduce the received signals at the subscriber's set.

The recently enacted Canadian standards require local origination signals to be pre-distorted per DOC (Canadian equivalent of the FCC) specs and that head-end processors must retain the original pre-distortion. Type approval of gear is considered since it is recognized that not every CATV operator will be able to purchase envelope delay measuring gear.

It is my opinion that it is only a matter of time before the FCC will take a similar stand.

Manufacturers will probably be required in the future to certify modulators, demodulators and signal processing equipment as to envelope delay and other types of distortion.

It may be of interest to realize that some TV studio specifications call for $\pm$ 25 nsec envelope delay.

15. Summary

An attempt has been made to explain "delays". It should be understood that delays can cause but one of the many distortions a signal can suffer from scene to TV set. For single channel equipment envelope delay measurements are a necessary complement to the more familiar amplitude response. Both amplitude and delay responses must be performed under all encountered signal level conditions.

16. Acknowledgements

Manufacturers of delay measuring and test equipment as well as many people in the CATV industry have been most helpful by giving freely of their time and knowledge. All such help is gratefully acknowledged.

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FUTURE PROMOTION OF THE CABLE TV SYSTEM APPLICATIONS

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SUMMARY

The cable TV is expected to play an important role in the near future as a major medium of live data transmission in the nation-wide or a confined "information society".

In order to meet such tomorrow's demand, the cable TV total system which has four capacity and their applications as follows, was developed.

1) Multi-channel Transmission: Full channel of 4FM, 12TV and etc.
2) Various Purpose Information Service :
   Local origination, Video transmission by light emitting Device, Video Responce System, Electronic newspaper by Facsimile and etc.
3) Wide Aerial Distribution :
   Multi-cascaded trunk line amplifiers by Metropolitan Cable Television standard.
4) Bi-Directional Communication :
   ITV Telephone using lower sub-channel.

INTRODUCTION

In Japan, small community antenna TV systems have been existing from about 10 years ago. NHK started to retransmit TV signals by means of UHF band wireless TV repeating system and VHF band cable TV system for helping poor TV reception in isolated areas.

Now in large cities, they can't receive clear color pictures because of gohst phenomena due to higher buildings increasing rapidly.

At first, Tokyo Cable Vision Foundation built the technical standard of Metropolitan Cable Television (MCT) system and equipments, and then they are beginning construction in large cities, such as Tokyo and Osaka.

In this process, near future, they are planning of an Information society by making cable TV system to be a transmission media.

This paper discribes the possibility of realizing simultaneously, the following four items with a CATV system :

1) Multi-channel transmission
2) Various purpose information services
3) Wide aerial distribution
4) Bi-directional communication

The system and its comprising equipment are based on the MCT standard and that of NCTA as well.

Various applications, such as optical transmission link using light emitting devices, electronic newspaper transmission by a high speed facsimile and bi-directional communication using ITV telephone, were tried with the system.

CONSTRUCTION OF AN EXPERIMENTAL TOTAL CABLE TV SYSTEM

Fig. 1 shows the experimental system composition where the four transmission services mentioned above are realized. This experiment was carried out at Shibaura area with high field intensity of Tokyo Metropolis. TV and FM signals received through VHF and UHF broad band log-periodic antennas are brought to the receiving terminal equipment, and then they are processed in bandwidth or converted to aother channel so that unoccupied adjacent channels can be used.

Local originating programs, facsimile, data communication, ITV telephone, pilot carrier, etc. are delivered on unoccupied channels or vacant frequency band, mixed with retransmitted TV signals, and then they are sent on to the distribution networks.

Those multiplex signals go through each ten of trunk cables and trunk amplifiers, and then they are dropped into the terminal equipment such as TV receivers, radio receivers along the distribution networks.

Bi-directional communication services take place between the head end and third trunk amplifier, and one channel of color ITV telephone and one pilot signal are transmitted toward reverse direction, using a band for reverse transmission services.

MULTI-CHANNEL TRANSMISSION

First requirement of cable TV subscribers will be "more channels". To satisfy the desire, the head end equipment are built so that each signal can be processed at a time and the total twenty signals are transmitted. The head end equipment are consisting of rack mounted type receivers and modulators, as shown in Photo. 1. The forward direction band (70 – 250 MHz) has eleven channels of color TV (seven of retransmission, one of U-V conversion and three of local origination), one of facsimile, four FM (two of retransmission and two of local origination and two of pilot carrier used for trunk amplifier's ALC and ATC.

The transmission frequency allocation and head end output spectrum are shown in Fig. 2 and Photo. 3.

The reverse direction band (20 – 60 MHz) having one channel of TV and one of phot carrier, is shown in Photo. 5. Thus, number of total transmission signals of both forward and reverse directions comes up to twenty. In order to realize this multi-channel transmission, the unique techniques of RF and IF BPF concept are applied for head end equipments, utilizing, signal processors for retransmission or channel conversion.

As the video signal is interfered with the audio carrier of adjacent down channel, the received signal is separated into video and audio in IF band (19.5 MHz), and the audio level is processed to be about 15 dB down compared with video carrier at RF output. In addition, input and output of the signal processor are connected to each helical resonated type BPF, which further improves the passband characteristics. Consequently, the total spurious level is attenuated to -70 dB or less compared with transmission signals.

In consideration of the reliability, the hybrid IC's developed by OKI are applied actively. In VHF band, MN series tantalum thin-film hybrid IC Amplifiers are used; in UHF band, microwave strip line circuits are applied for the wide band receiving mixer. Head end equipments are standardized for optimum design. An UHF band receiving unit of signal processor is shown in Photo 2.

VARIOUS PURPOSE INFORMATION SERVICES

In the second place, it must be considered about contents of multi-channel transmission. With the image of future cable television system in mind, an experimental transmission system is developed such as light telecommunication, electronic news paper, video response and ITV telephone system.

Symplified optical transmission

In the experiments, imaginating the case where the cables can't be laid between the head end and the studio separated by a river or buildings, a symbplified optical transmission system is applied as an example of wireless link system. As illustrated in Fig. 3, light emitting and photo diodes are used. The video signals are transmitted from the studio to the head end by pulse code-light modulation and lens convergent technique.

Video response and data transmission

In the future, subscriber's TV set will display many kinds of data when home terminals are linked to a computer through a cable TV system.

In this cable TV system experiments, subscriber's CRT displays the results of various questions and calculation from the key board of OKI-SCOPE, which is capable of taking memories out of OKITAC 4300 mini computer.

In addition, simultaneous directives are accomplished by using the data communication equipment.

As the results, it is confirmed that various traffic informations, shopping servics, etc. can be realized.

CATV–Facsimile transmission

Cable TV systems will also help in coping with the increasing labour cost in mail delivery system and also in satisfying the needs for quick delivery of news paper and mail. In this connection, it is considered that home facsimile systems are applied for cable TV distribution network.

In this experiments. 7th channel is used for facsimile transmission. The 7th channel is not available for normal TV transmission in Tokyo area because the 7th and 8th channel are overlapping partly in their channel allocations. Therefore the 7th channel is used for the transmission of high speed facsimile newspaper via Cable TV System. Modulate-Demodulation configuration of facsimile is illustrated in Fig. 5. Output spectrums of head end and that of 190 MHz Modulated signals are shown in Photo 6 & 7. The spectrum which is received after ten cascaded trunk amplifiers are shown in Photo. 8. Fig. 6 is a sample of a facsimile news paper received at home terminal (1/4 of actual size).

WIDE AERIAL DISTRIBUTION

From the stand point of cable TV enterprise, it is natural that they want the system to be extended and number of a subscribers to be increased. The trunk amplifiers is developed, which have functions of bi-directional and multi-channel transmission, based on MCT specifications.

In this experimental system, ten trunk amplifiers and ten drums of cable TC-10C TC-10CAF (Fujikura wire), are used. Coaxial cables are constructed by foamed polyethylen aluminum pipe. Fig. 7 shows the characteristics of distortion (2nd &3rd harmonic level of 90 MHz) after ten cascading amplifiers. Frequency allocation of Japan is different from that of U.S. In Japan the frequency range of higher TV channels twice as high the range of lower TV channels. The sum or differente beats between lower and higher TV channels generated in an amplifier fall in related TV channels. For this reason, the second harmonic distortion characteristics must be particularly improved.

Photo. 9 & 10 show the developed trunk amplifier. The latter is of underground type and techniques of undersea seismograph are applied in it, and so has complete waterproof construction. The system design chart is shown in Fig. 8. The trunk amplifiers have more than several tens of cascadability. Fig. 11 shows inter and cross modulation characteristics of trunk and bridging amplifiers.

Bi-DIRECTIONAL COMMUNICATION

In order to make the cable TV system of today applicable also for the bi-directional communication services of tomorrow, each unit of repeater equipment such as amplifier and filter units are optionized as shown in Photo 9 (b). This concept is based on the system design philosophy that the reverse direction transmission is realizable from the subscriber's terminals as illustrated in Fig. 9. Fig. 10 shows the characteristics of the dividing filter used for bi-directional truank amplifier.

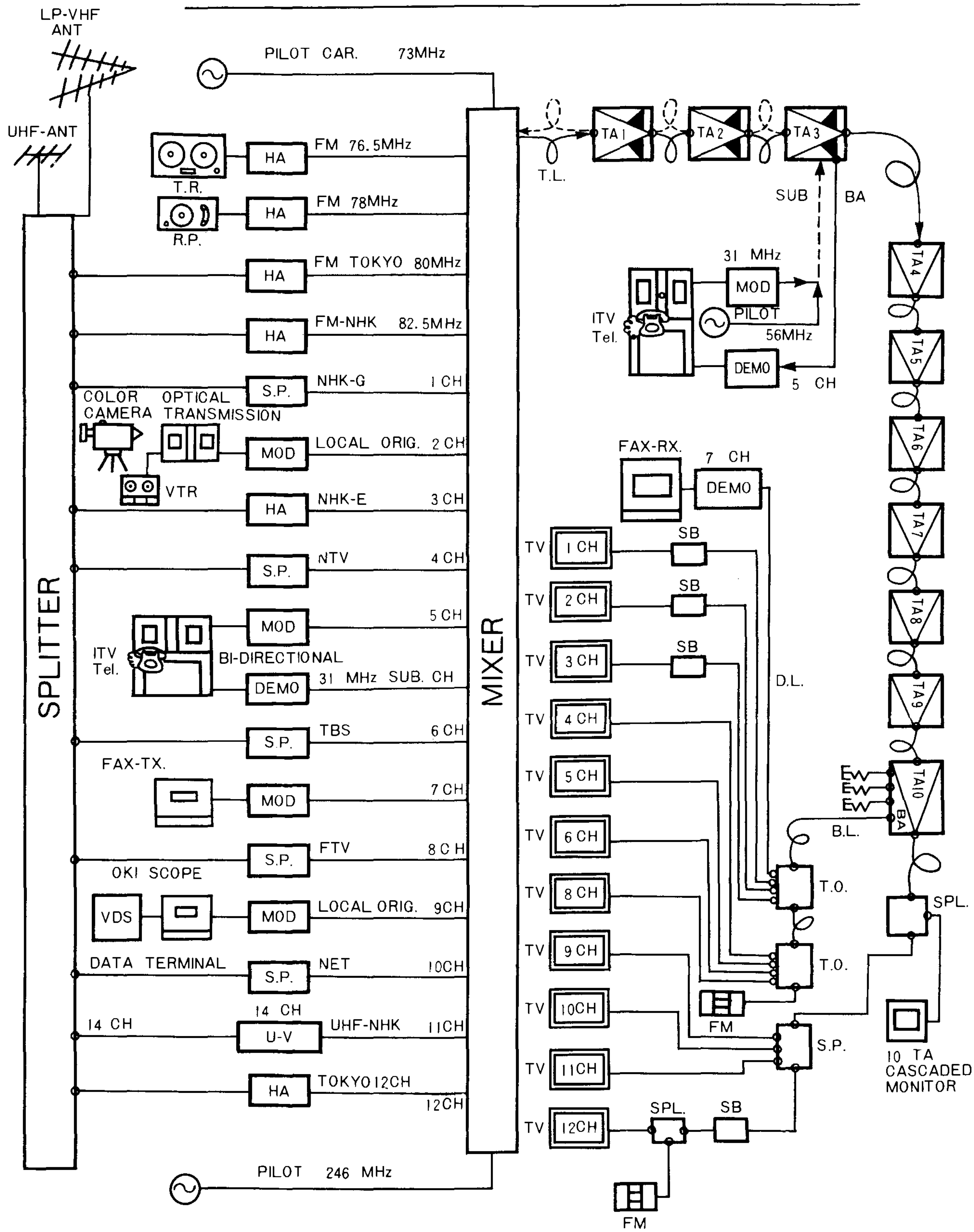
It is required that the envelope delay value be as small as possible (i.e. below 2.5 ns per channel), so that the phenomena of color shift may not occure due to multiple cascade. The echo loss occured in the loop of bi-directional amplifier is higher than 50 dB in passband, and 30 dB in the boundary area of both bands of forward and reverse direction.

In this experimental system, bi-directional services are demonstrated within three repeating section and color TV telephone shown in Photo 12 is used for end equipment. Photo 11 shows the distribution equipment such as tapoff and safety box (Border line of responsibility between CATV system and subscribers).

CONCLUSION

Now in Japan, they consider carefully and earnestly about "How the future CATV should be. They are now in the cradle including VP. Quick development and enlargement of software and video terminals for total picture communication systems are strongly required in accordance with social tends.

Fig. 1 CABLE TV SYSTEM EXPERIMENT (IN TOKYO)



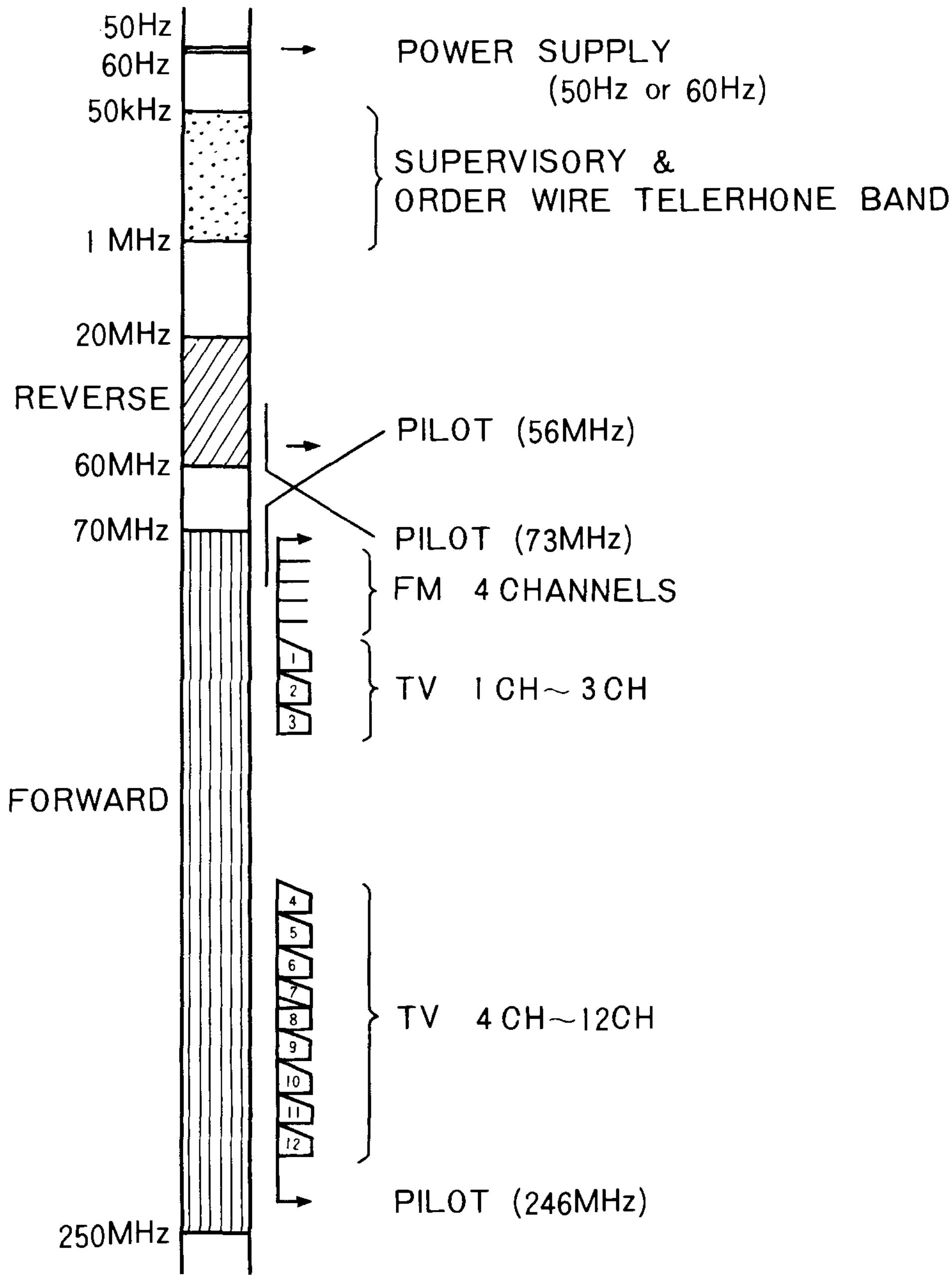


Fig. 2 FREQUENCY ALLOCATIN OF BI-DIRECTIONAL CATV SYSTEM

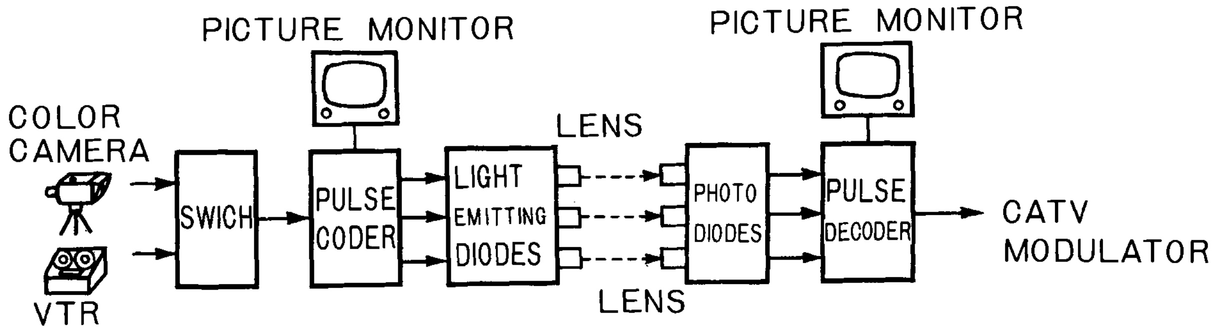


Fig. 3 OPTICAL COMMUNICATION LINK

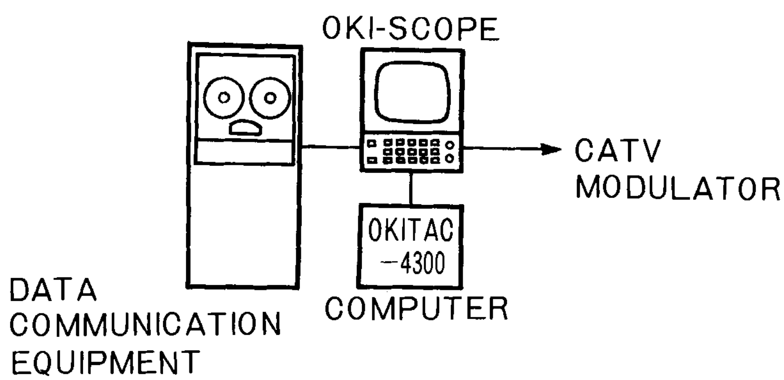


Fig. 4 VIDEO DISPLAY & DATA TERMINAL

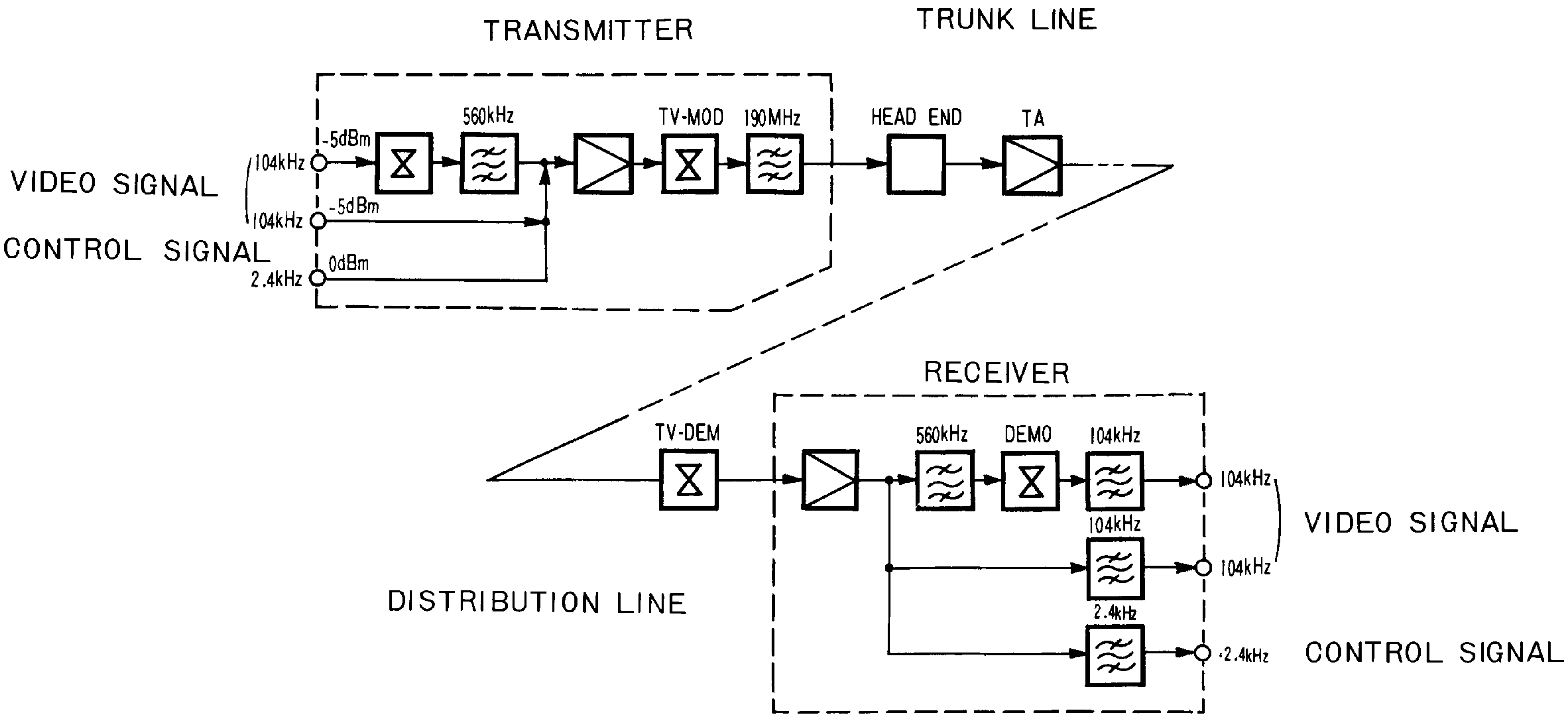


Fig. 5 CATV-FACSIMILE TRANSMISSION SYSTEM CONFIGURATION

冲と藤倉・電興の
CATV新技術
双方向通信や光伝送

読賣新聞
THE YOMIURI SHIMBUN
CATV・宣伝特集号
11月30日(月)−12月3日(木)

ボタンを押せば最新ニュース

ケーブルTVシステム

情報化社会に貢献する
ファクシミリ専門メーカー
松下電送

読売の長期連載「東京をどうする」

破滅か、再生か
激増する脱出希望組

20年後、東京を分散
官庁街も富士山ろくへ

Fig. 6　CATV-FACSIMILE RECEIVED PAPER　(SIZE 1/4)

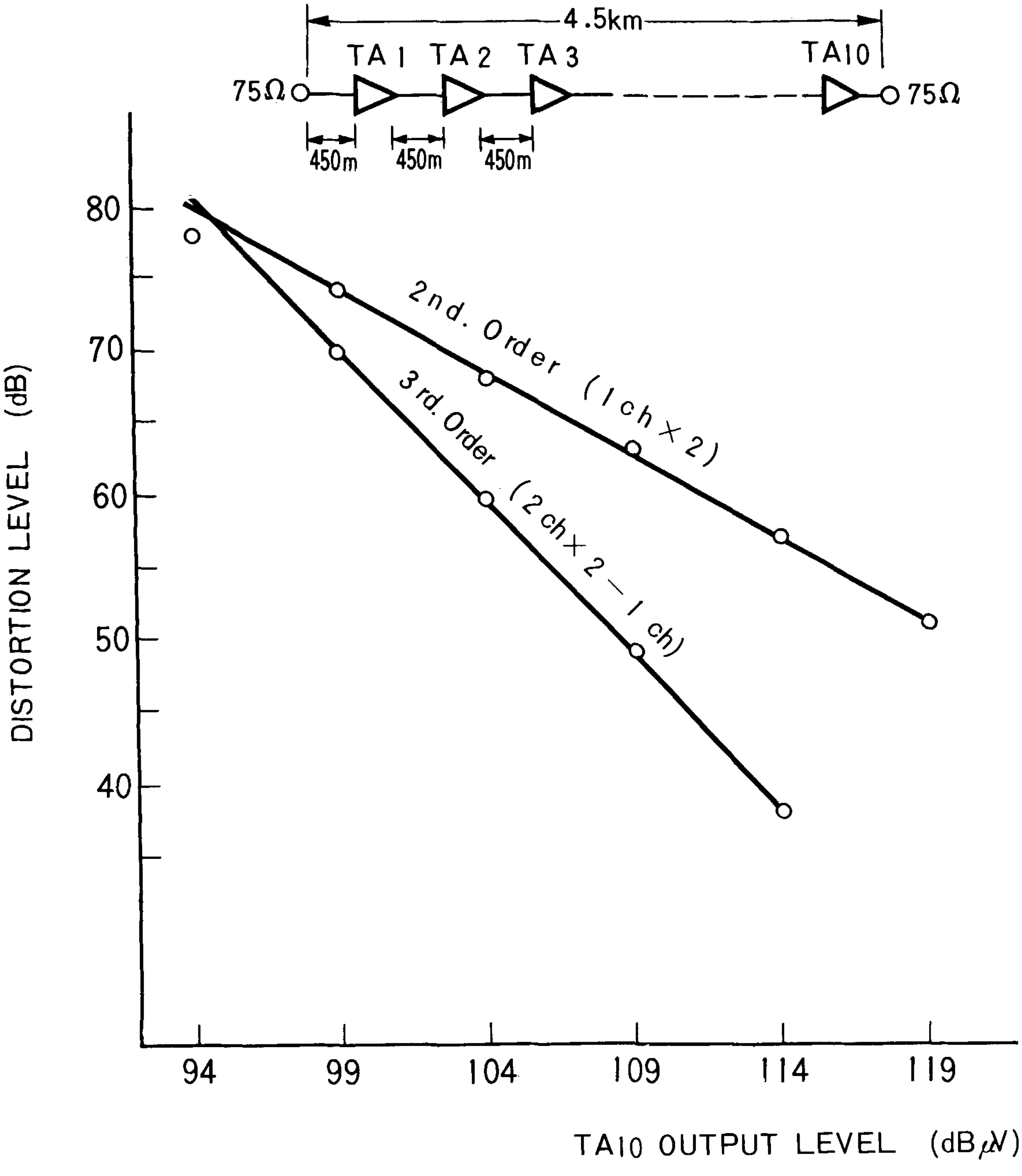


Fig. 7 MEASURED DISTORTION CHARACTERISTICS FOR 10 CASCADED TRUNK AMPLIFIERS.

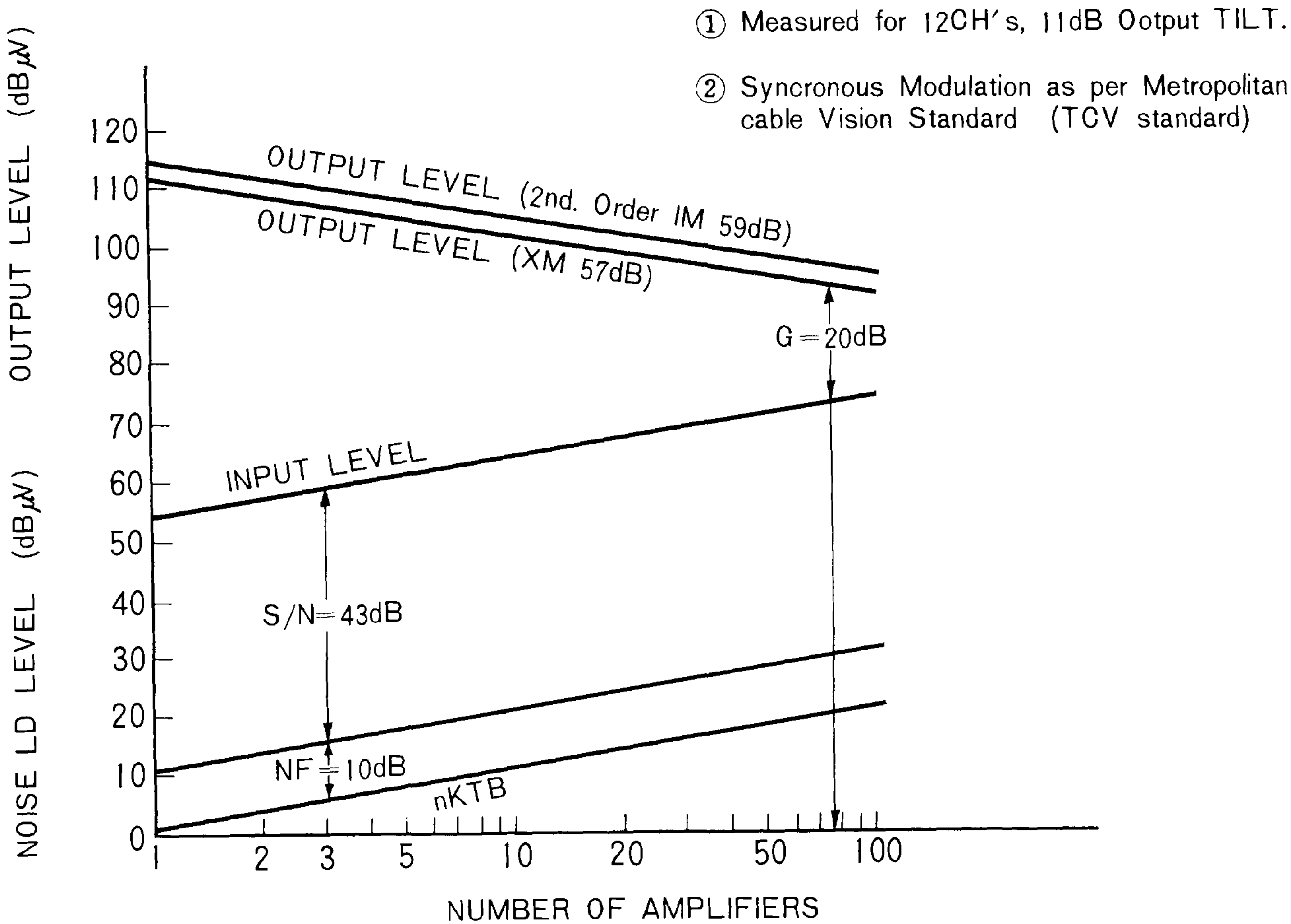


Fig. 8 CASCADABILITY CHART OF SYSTEM

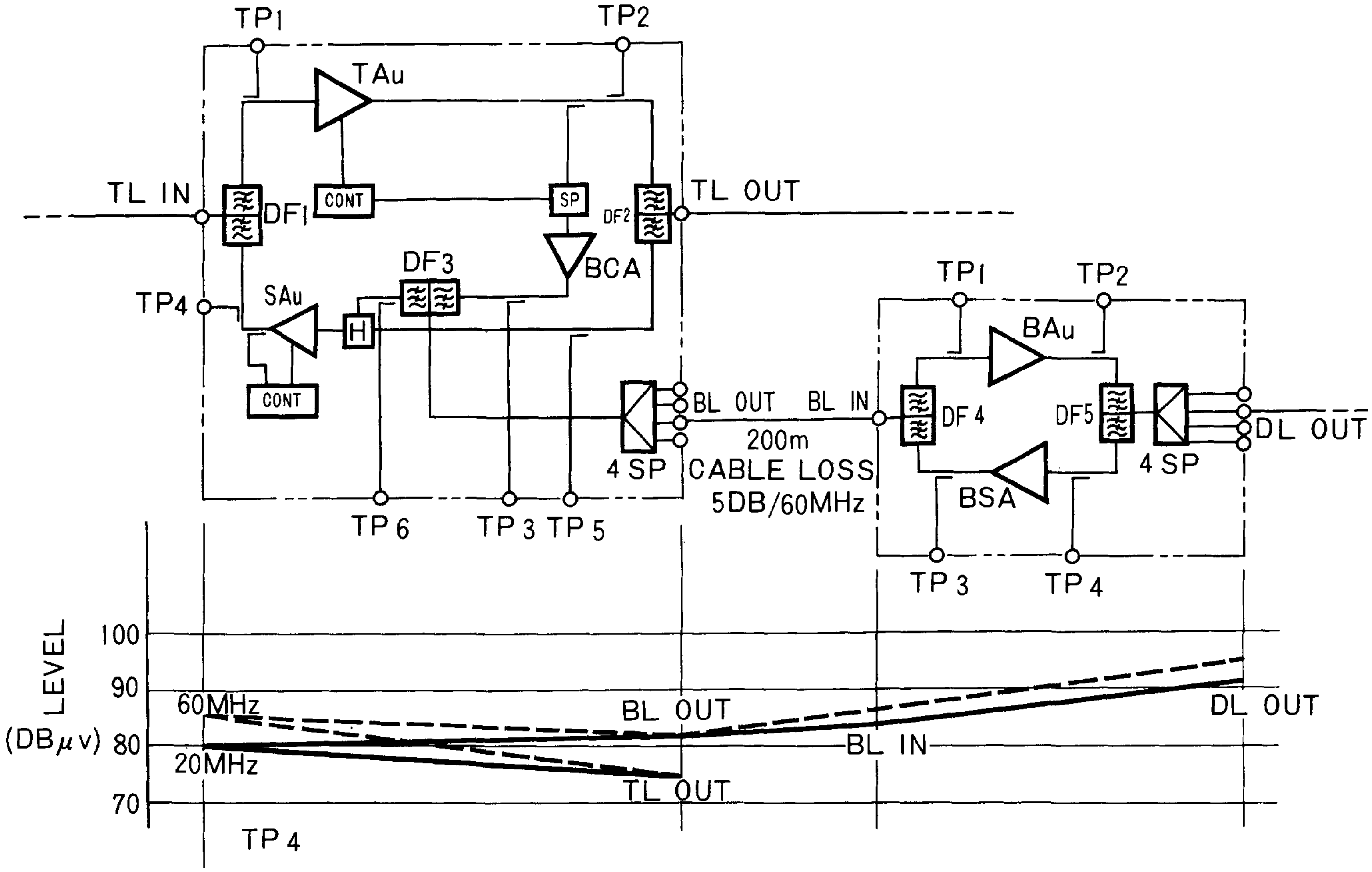
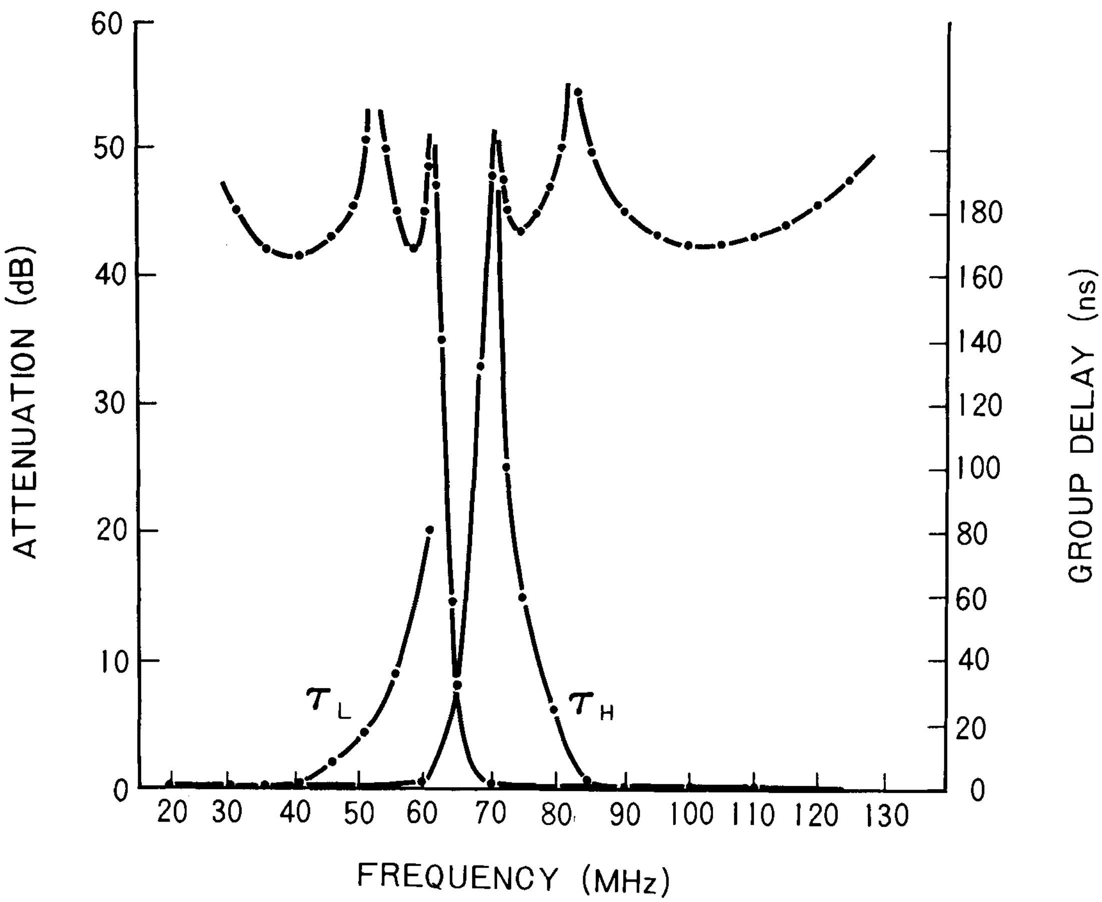
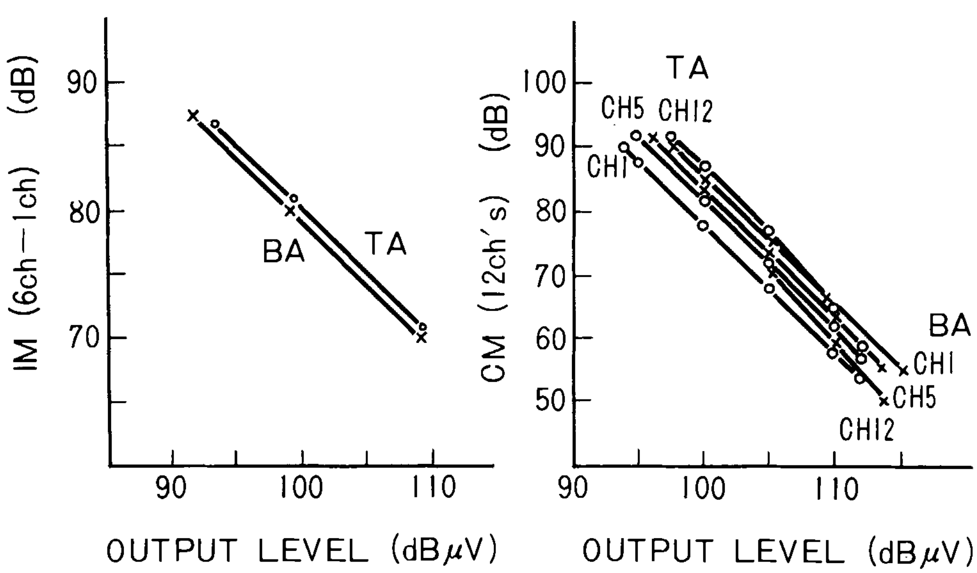


Fig- 9 BLOCK DIAGRAM & LEVEL OF BI-DIRECTIONAL TRUNK & BRIDGING AMPLIFIER

Fig.10 DIVIDING FILTER CHARACTERISTICS





Fiq.11 MEASURED INTER & CROSS MODULATION VERSUS OUTPUT LEVELS FOR TRUNK & BRIDGE AMPLIFIER.

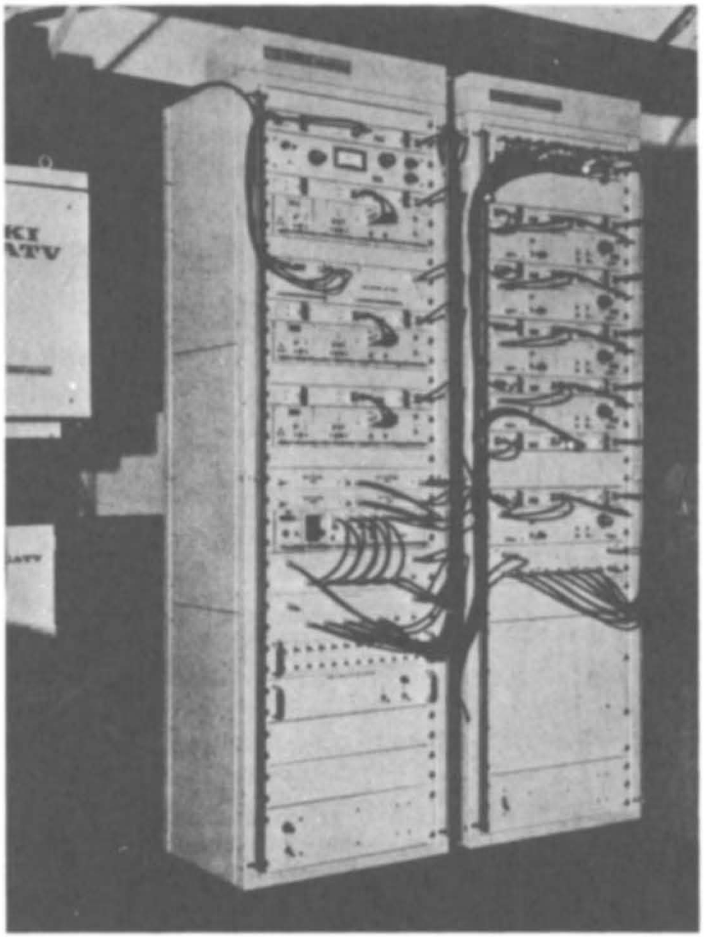


Photo 1

HEAD END EQUIPMENTS WITH FULL CHANNELS

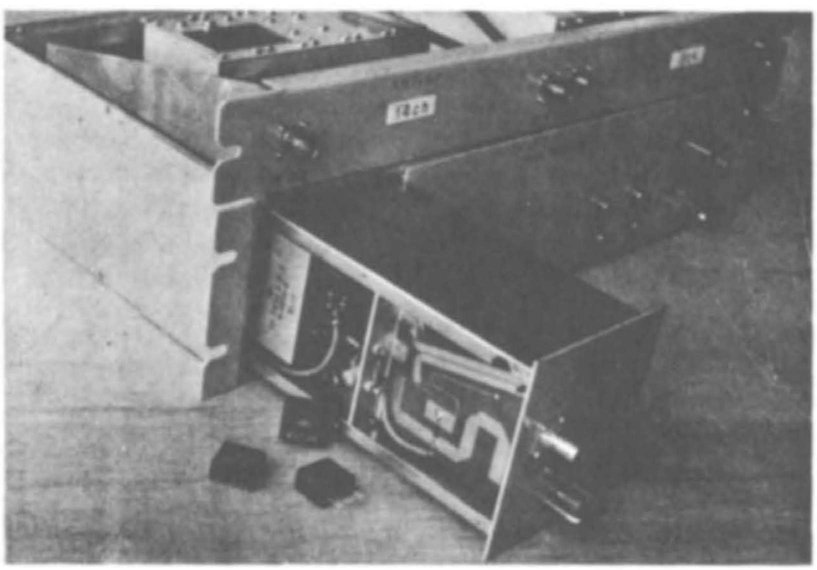


Photo 2 UHF RECEIVING UNIT OF IC DESIGN SIGNAL PROCESSOR, UHF MIXER USING MICROWAVE-IC & VHF AMPLIFIER USING TANTALUM THIN-FILM HYBRID- IC

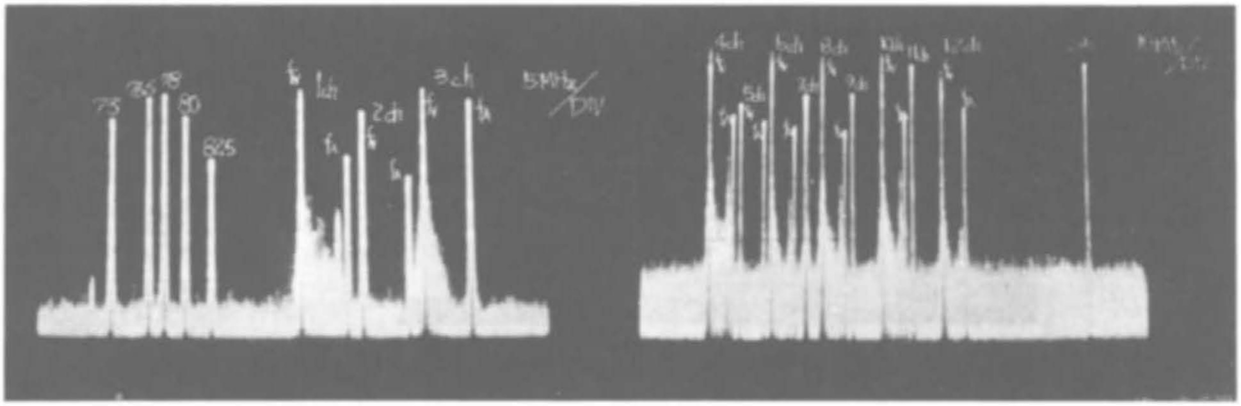


Photo 3 FREQUENCY SPECTRUM OF MULTI-CHANNEL TRANSMISSION AT HEAD END OUTPUT

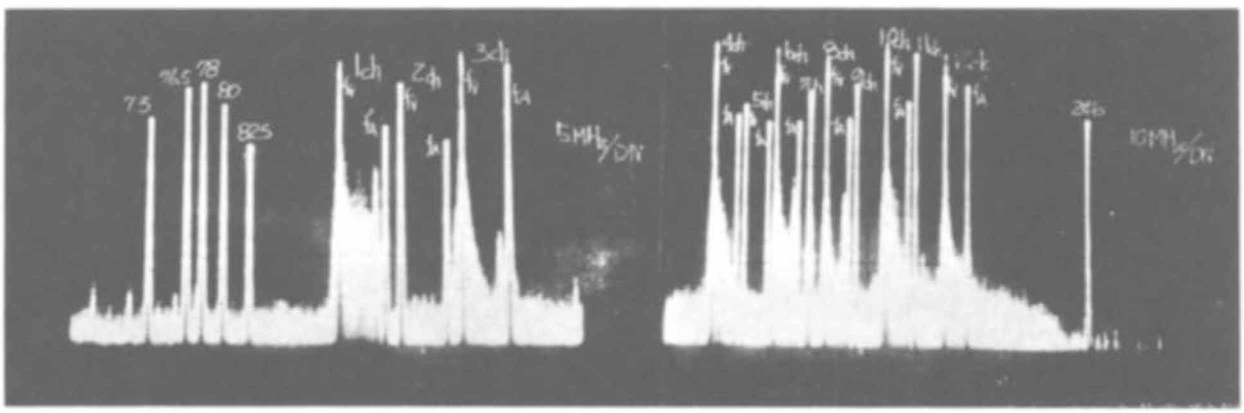


Photo 4 TRANSMISSION FREQUENCY SPECTRUM AT DROP TERMINAL (AFTER 10 CASCADED TRUNK AMPLIFIERS)

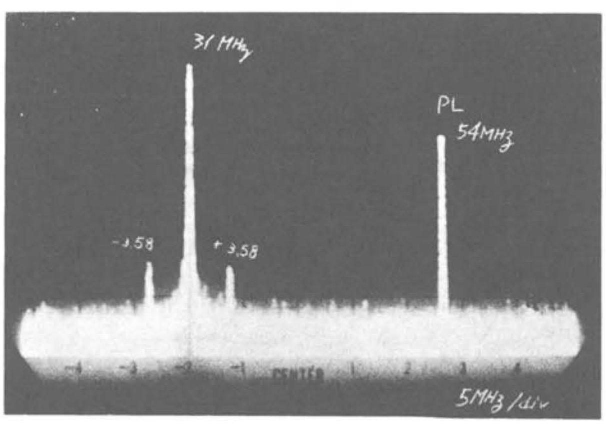
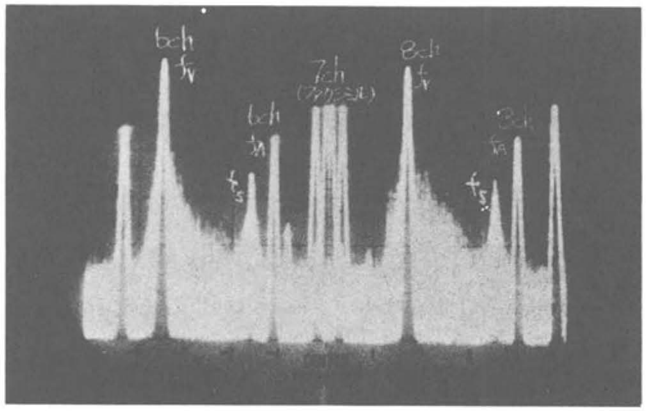


Photo 5 REVERSE BAND FREQUENCY SPECTRUM (31 MHz: Video carrier of color ITV)



V: 10 dB/div.
H: 2 MHz/div.

Photo 6 FREQUENCY SPECTRUM OF HEAD END OUTPUT. Facsimile signal is inserted in empty narrow CH.7 between CH6 and CH8.

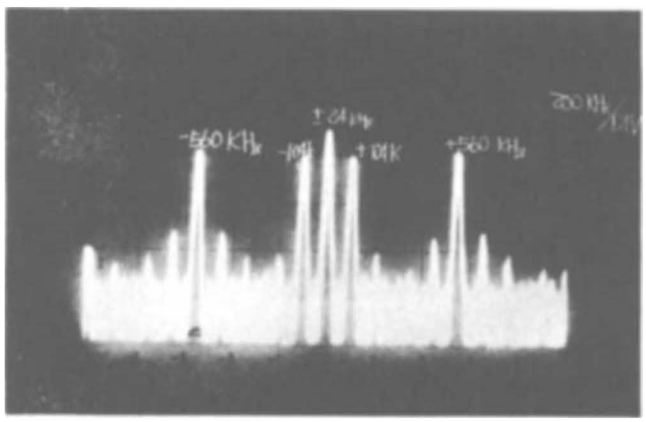


Photo 7 MODULATED FACSIMILE SIGNAL SPECTRUM AT HEAD END OUTPUT

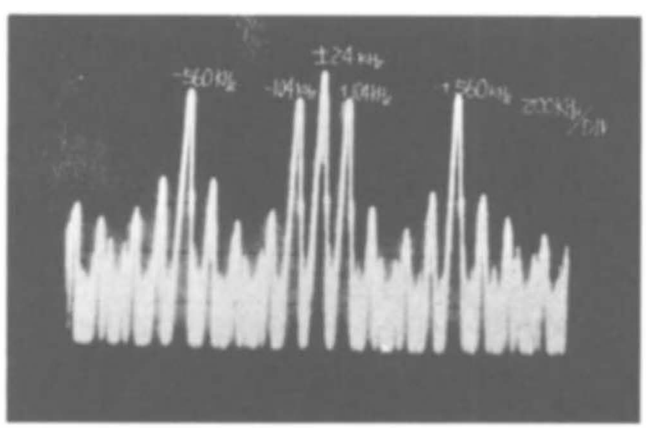
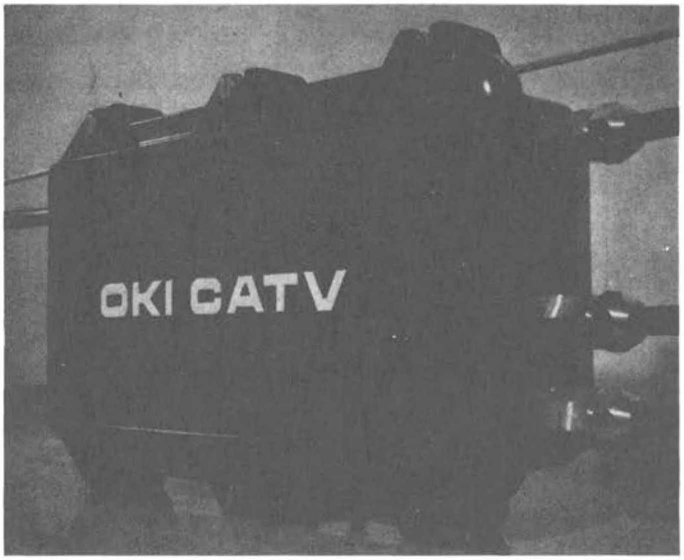
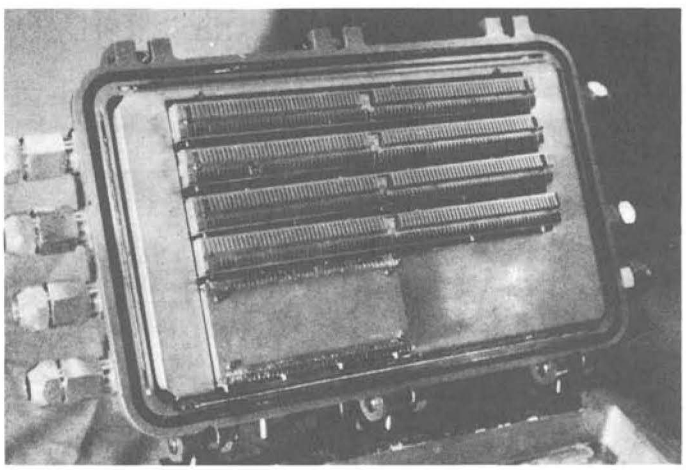


Photo 8 TRANSMISSION SPECTRUM AFTER 10 CASCADED TRUNK AMPLIFIERS



(a)



(b)

Photo 9
FEATURE OF BI-DIRECTIONAL TRUNK BRIDGE AMPLIFIER.

Plug-in unit is a signal processing equipment, such as TAU, ALC/ATC, BAU, SAU, & PSU

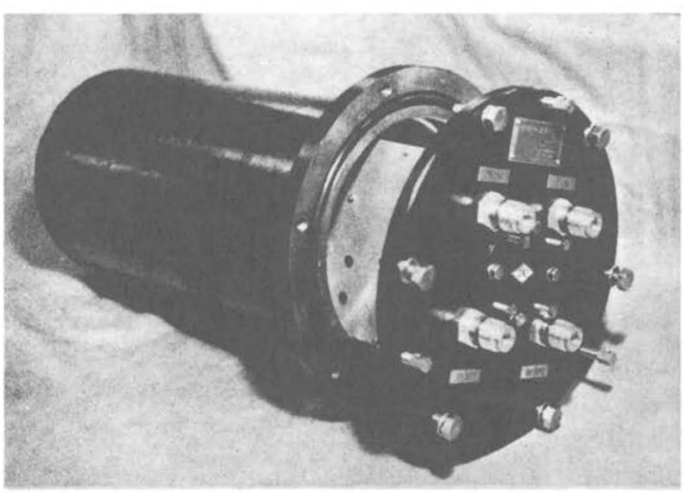


Photo. 10
UNDERGROUNDED TYPE TRUNK BRIDGE AMPLIFIER

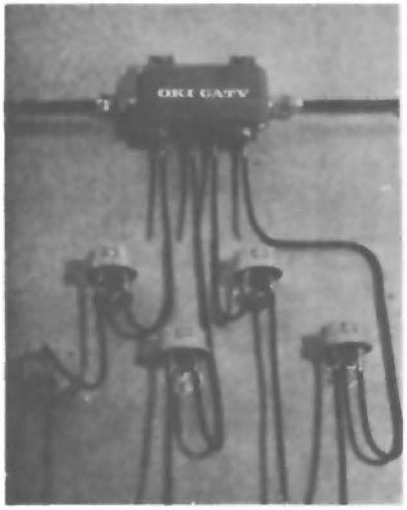


Photo 11

CONFIGURATION OF DISTRIBUTION SYSTEM–TAP OFF & SAFETY BOX

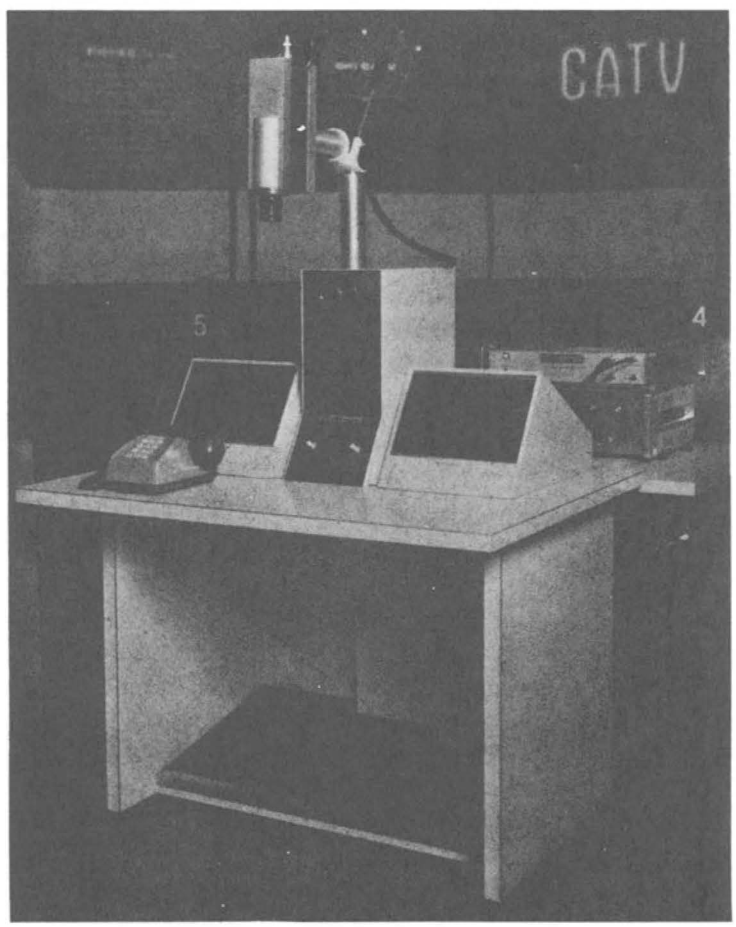


Photo 12

BI– DIRECTIONAL TERMINAL–ITV TELEPHONE EQUIPMENT

Helical Scan VTR's and the Cablecaster

By: Keith Y. Reynolds
VTR Product Manager

International Video Corporation

The average American Family is exposed to several hours of good quality color programming every day of the week. Eighty to ninety percent of this material is on video tape and originates from the major commercial TV networks, the educational TV network (NET) and the local TV station. If these viewers subscribe to the local cable system and the cable company is originating program material, they expect the same basic quality from the local origination channel that they see on other program sources.

High performance, reliable helical scan video tape recorders are capable of providing network quality color at a reasonable cost so most cable system managers are--or will be--interested in these recorders.

There are several VTR characteristics that are important to the cablecaster.

The first and most fundamental characteristic is the helical scan format. A format must be chosen which satisfies the requirements for color, wide bandwidth, good signal-to-noise ratio, good audio, time base stability and so on. Tape width, 1/4", 1/2", 1" and 2" is a primary consideration.

To date, 1/4" and 1/2" VTR's have not been designed which provide the stability and bandwidth required by the cablecaster. Two inch helical scan VTR's have all but disappeared from the market. This, then leaves the 1" format for the superior performance required for cablecasting. Several 1" formats are available and it is important that the cablecaster pick one that provides the best picture quality as well as one which is popular with the cable industry so that interchange with other cable systems is possible.

International Video Corporation manufactures a full line of 1" helical scan video tape recorders, all of which meet the stringent requirements of the cablecaster. These VTR's vary in price from under $3000 to over $30,000. Tapes made on the low cost VTR's can play on the more expensive recorders and vice versa.

A recent survey revealed that more cablecasters are using color recorders manufactured by IVC than by any other manufacturer. The primary reasons given were reliability and performance.

One important performance specification is video bandwidth. In order to properly record and playback the video bandwidth required for full color recovery and high picture resolution, the VTR's video frequency response must extend from 30 Hz to 5 MHz. A VTR with bandwidth significantly less than this will produce pictures lacking detail.

Video bandwidth is related to Horizontal Resolution. The term horizontal resolution is used to describe the ability of television equipment to reproduce fine detail. As a rule of thumb, with United States television standards, the relationship between resolution and frequency response is 80 lines of horizontal resolution for each 1.0 MHz of frequency response.

Therefore, if the VTR has a video response up to 5 MHz, the horizontal resolution is 400 lines.

If the VTR has a video frequency response of 5 MHz, it is possible to record the entire NTSC color video signal onto the tape. Color recovery, or stabilization, is then a function only of the playback electronics. The advantage of this approach is that you can record color signals with a monochrome VTR and either play the tape back on another VTR that is equipped with color recovery electronics, or purchase color electronics at a later date for updating the monochrome VTR. In either case, the taped program is in color.

Of course video bandwidth alone isn't the only requirement for high quality pictures. It is also necessary to have a VTR with good video signal-to-noise ratio. This is the ratio of amplification of useful information, or signal, to spurious information, or noise, and is expressed in decibels. A good video tape recorder should have a signal-to-noise of at least 42 dB peak-to-peak signal to RMS noise.

A VTR with a low signal-to-noise ratio will reproduce pictures with a noisy or grainy appearance.

All tape recorders introduce time base errors. They are present in audio, instrumentation, quadruplex and helical scan video tape recorders. Time base errors are significant if they cause an observable effect in the picture displayed on the monitor or television set. Various brands of receivers vary in the extent to which they are sensitive to these time base errors.

Video tape recorder time base errors can be caused by capstan servo instability, capstan eccentricity, drum instability and tape tension variations. These errors can be minimized by careful mechanical design as well as by electrical means.

IVC recorders are designed to exhibit very low time base errors. This is accomplished by the incorporation of a number of design features. The small scanning drum, with a diameter of 3.8", produces a relatively short 12" track length which minimizes the effects of tape tension variations For proper playback, the length of the video scan covered by one pass of the video head must exactly equal the length covered during playback or there will be a discontinuity in the reproduced time base at the time of the transition from one field to the next. Therefore, the shorter the scan length, the less it is effected by tape stretch caused by temperature and humidity variations.

Since the IVC capstan is located ahead of the scanner and meters the tape onto the scanner assembly rather than pulling it around, a tape tension of only 8-12 ounces is required. This also minimizes tape tension errors.

Conservative design and precision construction have minimized the possibility of mechanical imperfections contributing to time base instability.

Although all the parameters mentioned are very important, such things as good differential gain, or the amplitude change introduced by the video circuits; and good differential phase, or the phase change introduced by the video circuits are important too.

Good audio is necessary and it is important to choose a VTR that can reproduce audio frequencies up to 10 KC with minimal distortion and flutter and with maximum signal-to-noise ratio.

Although all IVC VTR's meet all of the requirements mentioned, each model has been designed with certain special features for different applications.

The newly announced IVC-700 Series is designed for the cablecaster who needs good quality, but who must operate on a very low budget. The IVC-700 Series has many of the quality features of the IVC-800 Series which has become the industry standard. They satisfy many system needs since tapes made on the IVC-700 Series VTR's are completely compatable with all other IVC-VTR's. The basic IVC-700 VTR is a recorder/reproducer with 5 MHz bandwidth.

The addition of a $500 color board provides color reproduction. A IVC-700-PB unit is also available. This configuration is a video tape player only. The IVC-700-PB is a valuable addition to any cable system because these units can free up the more expensive record/playback units for production recording schedules, remotes etc.

All IVC-700 Series VTR's are equipped with an advanced design, reliable transport mechanism similar to that used on the IVC-800 Series VTR's. This includes an optional remote control panel which controls all tape motion functions. These include: rewind, fast forward, play, record and stop.

The IVC-800 Series VTR's include several models designed for various special applications. Each IVC-800 model configuration is a rugged, field proven, reliable VTR, designed for monochrome or color operation.

The basic IVC-800A-SM is a recorder/reproducer with slow motion. The IVC-800A has a 5 MHz bandwidth and excellent signal-to-noise ratio. Next in the series is the IVC-820. This model includes the exclusive

Instant Video Confidence feature which allows playback of the video while recording. This is accomplished by locating an "I.V.C." video head in the scanner immediately after the video record/playback head. Annoyances such as head clogging, excessive dropouts and over-deviation can be detected immediately and corrected before the entire program is recorded.

The IVC-825 incorporates a capstan servo which insures precise program timing. In addition to this feature, the IVC-825 uses selected components to provide superior signal-to-noise ratio, lower flutter and better differential phase. This configuration is extremely popular with the cablecaster and others who require superior quality.

The IVC-870 features both assemble and insert editing. Assemble editing is used to produce an uninterrupted program tape from several separate segments. With this type of editing a completely new recording--video, audio and control track is made for each new segment and it is not possible to retain previously recorded information following the end of the assembled segment.

Insert editing is used to insert a new segment into a previously recorded tape without disturbing the information immediately before or after the inserted segment. Typically, it would be used to correct a mistake that has been made in the middle of an otherwise good program, or perhaps used to update a short segment in a pre-recorded program tape. With insert editing the original control track is retained while new video or audio and video are added. Precautions are taken in the video erase system so that previously recorded information either before or after the inserted segment is not disturbed.

The IVC-870 is the most sophisticated video tape editor in its price class. It is a vertical interval editor. This means that when you want to make an edit, the edit logic waits until the vertical interval before the edit can take place. When in the insert mode, the exit splice also takes place during the vertical interval. This results in a disturbance free edit every time since the electronic splice occurs between television fields.

Every cable system using video tape recorders should have at least one good editing VTR to handle program goofs or to consolidate program material on video tape.

The top of the IVC VTR product line is the IVC-900 Series video tape recorder. This VTR was designed from the gound up to be the finest quality helical scan recorder ever manufactured. Performance, such as time base stability, is the best in the industry. Since it meets all FCC and EIA specifications, it can be used as a broadcast recorder. An optional time base corrector manufactured by IVC not only brings the IVC-900 output stability to an undetectable ± 4 nanoseconds, it also permits direct color recovery and permits the VTR output signal to be mixed and faded with camera or other switcher input signals. In addition, for the first time, a helical scan VTR can dub to a quadruplex recorder.

All IVC-900 Series VTR's feature 3-1/2 hour playing time and Instant Video Confidence. A unique tension servo automatically corrects for tension errors during playback.

Three package configurations are available. A cased version, a rack mounted configuration and for the ultimate in operation convenience, an optional console. In the console version a standard IVC-900 Series VTR is mounted on a center pivot, providing complete access for maintenance. An eye level location is provided for a color or monochrome picture monitor, waveform monitor, audio amplifier and speaker, video display switcher and an optional IVC-4102 color time base corrector. Additional 19" rack space is provided below the recorder for other video equipment or tape storage. The basic IVC-900 configuration is a monochrome recorder/reproducer. Additional plug-in circuit board options include an NTSC color processor, a processing amplifier and a color drop out compensator. The IVC-960 includes insert and assemble editing capability. This editor is more sophisticated than the IVC-870 editor since it not only waits for the vertical interval, it also waits for the proper frame before making the edit. This results in a perfect frame to frame edit everytime.

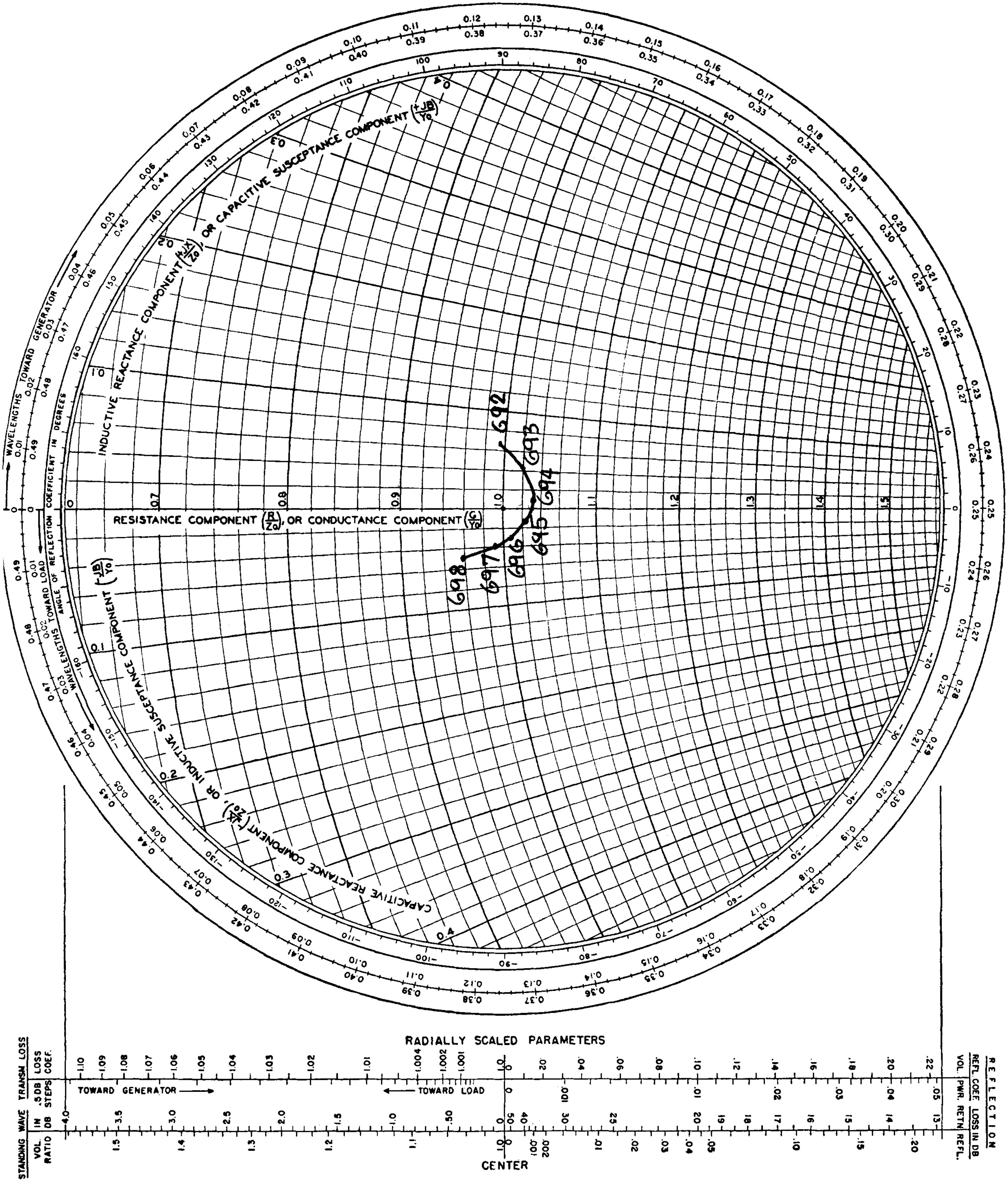
The Cablecaster can use the IVC-900 Series recorders to provide the utmost in picture quality to his subscribers. In addition, an entire evening of color local origination programming can be accommodated on one reel of video tape.

The IVC-700, IVC-800 and IVC-900 Series VTR's all completely compatable with each other. All designed for color. All with state-of-the-art performance specifications. All reliable, and all priced for every budget, large or small and all part of the IVC family of recorders that have become the standard of the cable industry.

* * * * *

NAME	TITLE CHANNEL 51 ZIG ZAG	DWG. NO.
SMITH CHART FORM 5301-7561-NE	GENERAL RADIO COMPANY, WEST CONCORD, MASSACHUSETTS	DATE

IMPEDANCE OR ADMITTANCE COORDINATES



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HIGH ENERGY TAPE FOR VIDEO RECORDING APPLICATIONS

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Over the past two decades, cable television has grown from small community antenna systems serving remote areas to large systems penetrating all types of communications. As cable television entered the "70's", new demands and opportunities represent fantastic challenges. Demands for program origination and better quality pictures must be met. Advances in technology such as two way cable, and more channels, offer new opportunities. Also, rumbles were being heard concerning improvements in television recording. This news of better things to come seemed to be all encompassing. From a hardware standpoint promises of new and improved video recorders were being backed up by introductions of sophisticated machines for both quadruplex and helical recording. The industry began using a new type of video tape that guards itself against damage. Several announcements and demonstrations have been made concerning revolutionary approaches to the duplication of recorded video material. All this is fascinating and important to our developing industry, but a common denominator that cuts across all of the topics that we have mentioned is the capability of the magnetic recording medium itself.

There have been vast improvements in the media during the last several years. Improvements that took us from what we thought were good black and white pictures, to the well defined, richly colored pictures that each of us expects to see on our monitors today. The low noise oxide introduced in the mid-sixties, coupled with advanced clean running binders, was an important breakthrough moving us toward that excellence that we now take for granted. Because of the improved signal-to-noise ratios attainable with the low noise oxide, multiple generation dubbing was not only possible, but became the accepted way to produce everything from a dog food commercial to a ninety-minute extravaganza. In the close analysis, the oxide on the tape has caused a lot of changes in the television industry, and it appears that these changes are not about to stop.

In exploring new ways to make even further improvements in the electro-mechanical properties of video recording tapes, it appeared that we had gone just about as far as was possible with our present day family of synthetic low noise oxides. If we wanted to see a meaningful improvement in the key recording characteristics of both RF output and signal-to-noise ratio, we would have to enter into some extensive research centered upon modifying the basic oxide particle.

HIGH ENERGY TAPE FOR VIDEO RECORDING APPLICATIONS

After many years of work, and extensive application evaluation, we were ready to announce the result of this basic research. We had succeeded in developing a new family of oxides that offered all the features indicated in the objectives of our tape design engineers. The products in which the new oxides are used are referred to as High Energy tapes because of the higher output that can be derived by proper application of this new recording media.

Since it is generally well known that a higher output can be predicted when the coercive force and remanence of a tape are increased, it was an adjustment in these parameters that made this breakthrough possible. We achieved the needed increase in coercivity and remanence by modifying the composition of the gamma ferric oxide and properly dispersing it in a binder system. A small amount of cobalt has been introduced into each particle of oxide in a manner that allows the control of the resultant coercive force to a pre-determined level. This technology of introducing cobalt into the particle to increase coercive force has been known for years by the chemists in the industry, but results in the past were disappointing in that the introduction of the cobalt detrimentally altered the particle size and shape. This altering of particle size and shape had an adverse effect on the signal retention of the tape, especially when it was heated or flexed. With the development of this new technology, it is now possible for us to produce High Energy oxide that has the same controlled particle size and shape as the gamma ferric oxide. As a result, it has the same signal retention ability as the gamma ferric oxide.

This new oxide technology enables the tape manufacturer to tailor make magnetic tapes with coercive force values from as low as 300 oersteds to as high as 1000 oersteds while maintaining retentivity values in the 1200 gauss range. Because of the ability to control the coercive force in the finished product, tapes can be specifically designed to accomplish a specific purpose.

We could now use this proprietary oxide to develop a family of tapes that would provide immediately noticeable benefits to the user. When used to manufacture video tapes, both improved signal-to-noise ratio and increased RF output are achieved. We are currently producing tapes for both quadruplex and helical recording with coercivities of 500 and 900 oersteds. The 500 oersted version is totally compatible with video recorders that are in use today, and the 900 oersted tape will find use in the field of advanced systems technology. It is interesting to note that the benefits just described were accomplished without the need for the usual technical tradeoffs that plague design engineers. These new High Energy tapes still incorporate the best features of present day tapes in the areas of physical handling, tape wearability and head life.

The classic means used to compare the capabilities of various oxides is the familiar hysteresis loop. An analysis of the loop provides us with three important parameters. These are the saturation value, referred to as B_s; the amount of retained flux, known as B_r; and the amount of energy required to reduce the retained flux to zero, called the H_c. These three properties, as well as the shape and size of the oxide particle, determine the magnetic capabilities of the finished tape.

Here in figure 1, we see three hysteresis loops. The small one represents the pattern seen for conventional video tape, such as "SCOTCH" Brand No. 361. The next largest is a trace of the loop formed by a compatible 500 oersted High Energy tape, and the largest described an experimental 900 oersted High Energy product. The three important points are labeled on each of the three curves. Saturation, or B_s, is seen in the first quadrant in the upper right hand corner. The amount of retained flux, B_r, is seen on the positive vertical axis and the H_c or energy required to reduce the flux to zero appears on the left. In this type of measurement, a great deal can be learned from the shape and size of the curve in the second quadrant between the points B_r and H_c. The amount of area that lies under the curve in this quadrant is a measure of the relative energy available from that oxide for recording.

Note the additional area gained by the two High Energy products. Inspection reveals that the 900 oersted tape yields nearly four times the area of conventional 300 oersted oxides. And the 500 oersted tape encompasses an area that is about twice as large. This extra energy that is available can, of course, be used to good advantage for television recording.

One of the things to investigate in evaluating a new video tape formulation is the amplitude of RF output seen when this tape is played back. There are three sets of curves in figure 2 that illustrate this. The curves next to the 300 oersted label will be used as a reference because they are typical of conventional tapes currently in use for quadruplex recording, such as "SCOTCH" Brand No. 400. Here, represented as a solid line, we see the traditional optimization curve showing the peak being reached at a point slightly above the level that represents 25% of the available RF drive used in recording the tape. The relative output on playback is seen on the scale as reaching 40 units. While this peak level of 40 on the vertical axis is important for making comparisons with tapes with greater coercive forces, noting the shape of the curve is important, too. We see here that the optimization curve rises rather steeply to the peak and then descends with equal steepness.

HIGH ENERGY TAPE FOR VIDEO RECORDING APPLICATIONS

When we compare the optimization curve for the 500 oersted High Energy tape, we immediately notice a 4 db increase in RF output. We notice also that the entire curve is positioned farther to the right, indicating that additional RF drive was used when the tape was recorded. We mentioned that the 500 oersted High Energy tape was compatible with today's recorders. This curve substantiates this, as the entire curve falls well within the available RF drive. The peak, or actual optimization point, falls at about 45% of the maximum drive utilizing a 1.5 mil gap depth. While a new head that has full gap depth requires more drive, the record drivers have more than enough range to easily obtain optimization.

The shape of the 500 oersted curve is quite different than that representing the 300 oersted conventional video tape. It is not as steep and this is a convenient advantage. The operator performing the optimization will notice a broader optimization range making it easire to adjust for the desired maximum. This broadened range also means that RF drive optimization need not be performed as often throughout the life of the video head assembly.

The broken lines in the figure indicate the chroma slope for each of the tapes, this represents the equalization change that is needed to compensate for wear of the video head pole tips. As the tape coercivity is increased, the slope of this line decreases, meaning that less correction is needed to maintain proper equalization. Just as was true when speaking about RF drive, the equalizers, too, would not require adjustment as often and such adjustment would be less critical.

The uppermost solid curve is representative of the 900 oersted experimental product. This tape is not designed to be compatible with recorders in normal use. As can be seen, the RF drive was set at maximum to obtain optimization. The head that was used was specially selected for this test to provide the efficiency needed to attain the optimization point. While we agree that this does not represent normal operating conditions for present equipment, it does clearly demonstrate the rather dramatic 7.5 db increase in RF output that this 900 oersted tape will yield, when compared with the 300 oersted tape that is the standard of the industry today.

The optimization curve is even more broad than it was with the 500 oersted tape and the chroma slope is approaching the horizontal. It is evident that when machines are readily available that will make use of tape in the coercivity range, the need for repeated RF drive and equalization adjustment will be minimal.

HIGH ENERGY TAPE FOR VIDEO RECORDING APPLICATIONS

Signal-to-noise ratio is a key parameter and a much discussed topic when speaking in terms of picture quality. This is especially true where video tape is concerned. Great strides in this direction have been made by camera designers, switching and processing equipment manufacturers and by the builders of the recorders themselves. With the advent of the high band color standard, companies such as ours introduced video tape manufactured with low noise oxide. Master tapes were clean and quiet. So quiet, in fact, that it was now possible to produce a second or third generation dub that looked as good as the master would have looked using the former tape. It wasn't long and multiple generation editing and dubbing was the standard way of doing business.

A meaningful improvement in the signal-to-noise ratio of the recording tape for today's as well as tomorrow's applications is the second very important benefit that is gained with the use of the High Energy oxide. In figure 3, we see a set of curves that, at first glance, look a great deal like the ones in the previous drawing. Once again, on the X axis, we see RF drive indicated in per cent of maximum. On the Y axis, however, we have shown signal-to-noise referenced to optimum conditions for the 300 oersted tape. The peak on the bottom curve, then, is zero db. The improvement in signal-to-noise ratio, when using the 500 oersted tape, amounts to an impressive 4 db. Once again, we should restate that this will be achieved on today's recorders without any modification. All that is necessary is a normal optimization adjustment.

On the same drawing, we have also shown a plot for the 900 oersted tape. With the specially selected head and maximum RF drive, we realize a 7 1/2 db improvement over the reference tape. This, of course, is not a compatible product, but does stimulate the imagination as we look to the future; to a time when machines are designed that can supply the additional RF drive needed to properly record it and sufficient erase fields so that it can be erased and used again.

The additional 4 db gained in signal-to-noise ratio, when using the 500 oersted tape, is particularly advantageous when one considers that many tapes used today are really fourth generation copies of a master. With the aid of the chart in figure 4, we can easily compare several combinations of multiple generation dubbing.

The data shown is for quadruplex recorders, however, helical recorders follow the same pattern. The same relative advantage is realized when High Energy tape is used for multiple generation copies on helical recorders.

HIGH ENERGY TAPE FOR VIDEO RECORDING APPLICATIONS

The first example, labeled Number 1, charts the progression of the four generations using 300 oersted tape in each step. This is the way things are being done now, using conventional video tape. Moving to the right through the first example, we note that the established signal-to-noise ratio of the master on the 300 oersted tape is 50 db. We have placed a circle around that number because it will be used for comparisons with the other examples. As we move into the second generagion that will also be on 300 oersted tape we encounter a loss in duplication of 1.5 db in signal-to-noise. The second generation copy will have a signal-to-noise ratio of 48.5 db. The duplication loss into the thrid generation is again 1.5 db, and the same is true for the fourth generation. This final copy has a signal-to-noise ratio of 45.5 db which is down 4.5 db from the original master.

Look now at the second example. Here, the master and each of the succeeding generations were made on 500 oersted, High Energy tape. With this product, the master reflects the 4 db improvement in signal-to-noise ratio so we begin with 54 db. Each generation of copying will again reduce the ratio by 1.5 db. This results in a fourth generation copy with a signal-to-noise of 49.5 db. When we compare this to the original master recorded on 300 oersted tape, we see only one-half db difference. We have now succeeded in producing a fourth generation copy that is as good, visually, as the traditional master made on conventional tape.

The third example indicates the use of the High Energy tape throughout the mastering and editing steps, but here conventional tape was used for the final fourth generation copies. The first three generations are the same as example Number 2, with 1.5 db duplication loss per step. As we move into the fourth generation, that will be recorded on standard 300 oersted tape, we will encounter a 2 db duplication loss. The net result is a final copy with a signal-to-noise ratio of 49 db. This is an insignificant one-half db down from a final copy made on the new High Energy tape and barely perceptible one db below a conventional 300 oersted master.

The numbers on the chart clearly suggest that for almost all fourth generation copies, it would be wise to master and edit on High Energy tape and economically sound to produce those final copies on traditional video tape. In those instances when the absolute ultimate in signal-to-noise is required, an extra one-half db can be gained by using High Energy tape throughout the entire process for either quadruplex or helical recorders.

Our discussion to this point has centered about the 500 oersted product. We have seen from the previous curves, however, that both RF output and signal-to-noise ratio are greatly improved as the coercivity is increased. Since the High Energy oxide lends itself so well to being tailored to yield a wide range of coercive force, it is our hope that future machine designs will tape advantage of this aspect. For a given application it may be that a system making use of 650 oersted tape would be ideal. For another use, 825 or 435 might render optimum performance. Tape is no longer the limiting factor in the recording process. High Energy tape is a reality. All that is needed now is the hardware to take advantage of this breakthrough.

We say this as a preface to our discussion on the application of High Energy tape to the field of helical video recording. Here, just as with the quadruplex systems previously discussed, increased coercivity of the recording tape has the capability of yielding an increase in both RF output and signal-to-noise ratio.

Figure 5 contains the optimization curves for the standard 300 oersted helical tape now in use as well as two High Energy constructions. Once again, RF drive is plotted along the bottom in terms of the drive that is available and relative RF output appears along the vertical axis. Note that with traditional tape optimum drive is about 50% of what is available. The 500 oersted High Energy tape will require about 70% of the total available drive and would deliver 6 db more RF output. This, of course, is compatible with present day equipment.

We have also shown a curve representing a 700 oersted experimental product. Even with the use of a specially selected video head, we were just able to reach the optimization point. You will notice, however, that the increase in RF output amounts to 8 db. Just as we noticed with the quadruplex examples, as we increase the coercivity the curve becomes more rounded and loses its steepness. This would again mean that optimization adjustments would be less critical and that they would be required less often.

Signal-to-noise ratio is also increased with the higher coercivity tapes. Figure 6 plots RF drive against signal-to-noise for the three tapes being discussed. Using the standard 300 oersted tape as the reference, we see that the 500 oersted compatible product offers an increase of 4 db, and the 700 oersted experimental tape yields a 6 db increase.

These curves, used in the last two figures, were generated on a one inch helical recorder. While this type of machine would not handle 700 oersted High Energy tape without modification, it does an excellent job with the 500 oersted tape. This, however, does not hold true across the complete line of helical recorders. In some cases the signal-to-noise ratio established by the machine electronics is very close to what is possible with traditional tapes. With most half inch recorders, even though we see a significant increase in RF output, the signal-to-noise ratio cannot be improved more than 2 db because of the electronics improvement in signal-to-noise, it is distrubing to note that the signal-to-noise of the tape playback is only 5 db below the EE (electronics-to-electronics) capabilities of the recorder. With the tapes used up until now these shortcomings were not really noticeable, but with the introduction of the High Energy family of tapes we can see that picture quality is seriously hampered by the limitation of the recorder electronics. Now that an improved tape is available we sincerely urge the hardware designers of our industry to develop the equipment to utilize the potential of this new oxide to complete what can amount to a great leap forward in video technology.

A summary of 500 oersted compatible High Energy tape performance on existing helical video recorders is as follows. All helical VTR's will give a noticeable improvement in signal-to-noise ratio (about 2 db) and RF output (about 50%) increase without making any recorder adjustments. Half inch recorders generally show no further signal-to-noise improvement when record drive is optimized for High Energy tape even though the RF output increases because of recorder electronic limitations. One inch recorders generally show another 2 db improvement in signal-to-noise ratio, or a total of 4 db, when record drive is optimized for High Energy tape. The RF output is then double that of conventional tape.

Recorded High Energy tapes can be played back on any recorder without adjustments with full signal improvement. In other words, to realize best performance, only the VTR used for record need be adjusted.

As we look to the future we can see constant attempts at miniaturization and a desire to place more information on a reel of tape. The High Energy tapes that we have been discussing have a greatly improved short wavelength response. This offers the possibility of operating at slower speeds. To many, this ability to operate at a reduced speed signals the gateway to practical video cassette recording. Up until now the drawback has been the need for an overly large cassette, an unduly short program or a serious sacrifice in picture resolution.

To demonstrate the slow speed capability of the three tapes we have been discussing, we modified a recorder with a 13 microinch head gap to run at half the normal speed. Figure 7 compares the results of this test with the same recorder operating normally. Here we see RF output plotted as a function of the recorded wavelength. Our zero db reference point is established at the one-eighth mil, normal operating point for the standard 300 oersted product. This is the vertical line on the left. By reducing the speed to one-half, we are then recording at one-sixteenth mil, and the standard product is seen to be nearly 5 db down in RF output as it crosses the vertical line on the right.

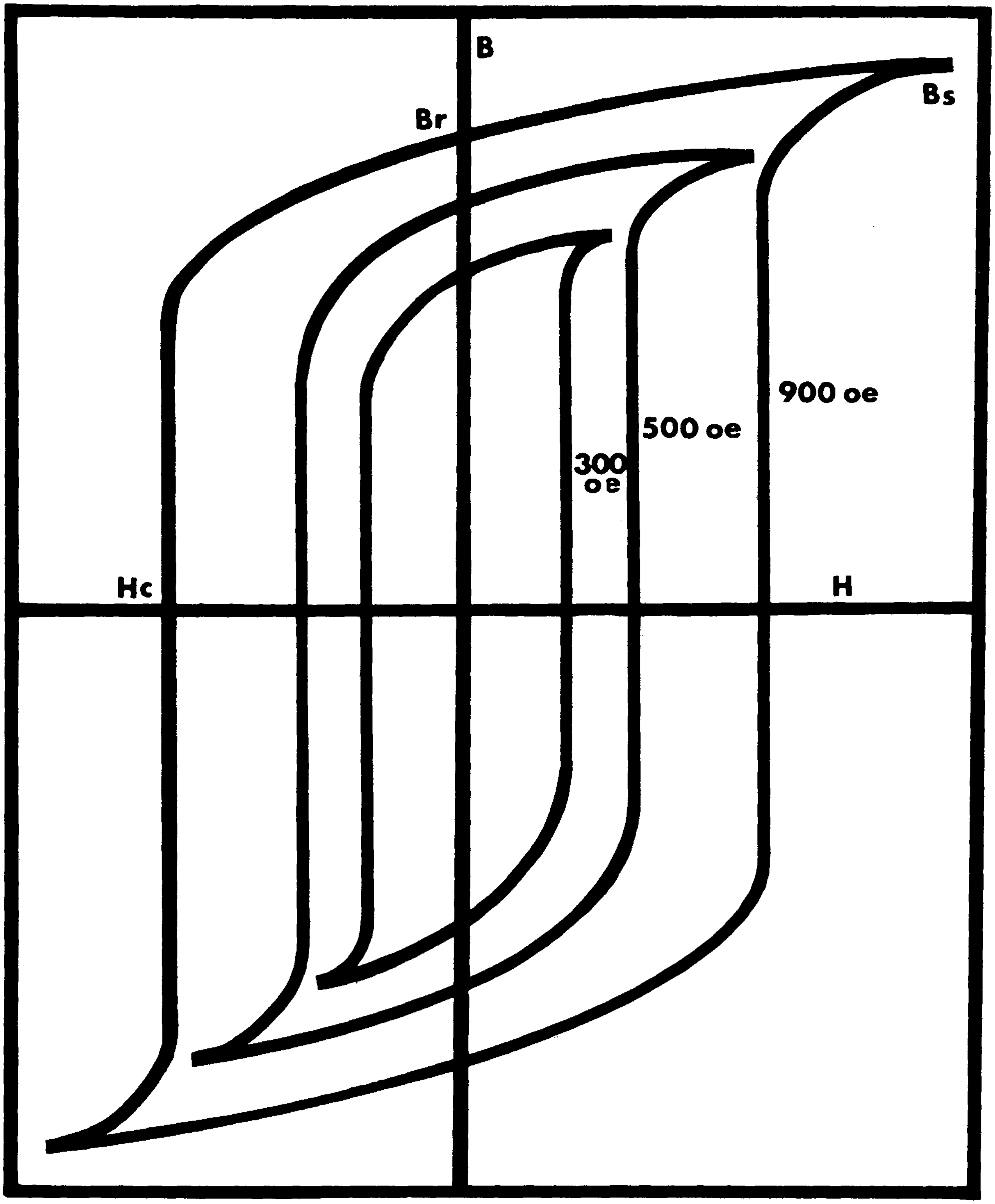
HIGH ENERGY TAPE FOR VIDEO RECORDING APPLICATIONS

The 500 oersted High Energy tape operates at a plus 4 db from the reference at the normal speed, maintains this output at the one-twelvth mil wavelength, and has an output at one-sixteenth mil -- the half speed point -- that is about 2 db better than the 300 oersted tape at the normal speed. Standard tape at normal speed produces a picture of excellent quality. It is now possible for equipment designers to obtain better picture quality on half the length of tape by reducing the head-to-tape speed by one-half. The era of half speed recording is here now with a readily available compatible tape product.

If we follow a similar plot for 700 oersted tape, we observe an interesting result. In this case the experimental High Energy product has an output at half speed that is actually 5 db better than conventional tape at the normal speed. It is apparent from this that not only can one achieve a comparable picture at half speed, but it is now possible to obtain a better picture at half speed than has been possible at the normal speed.

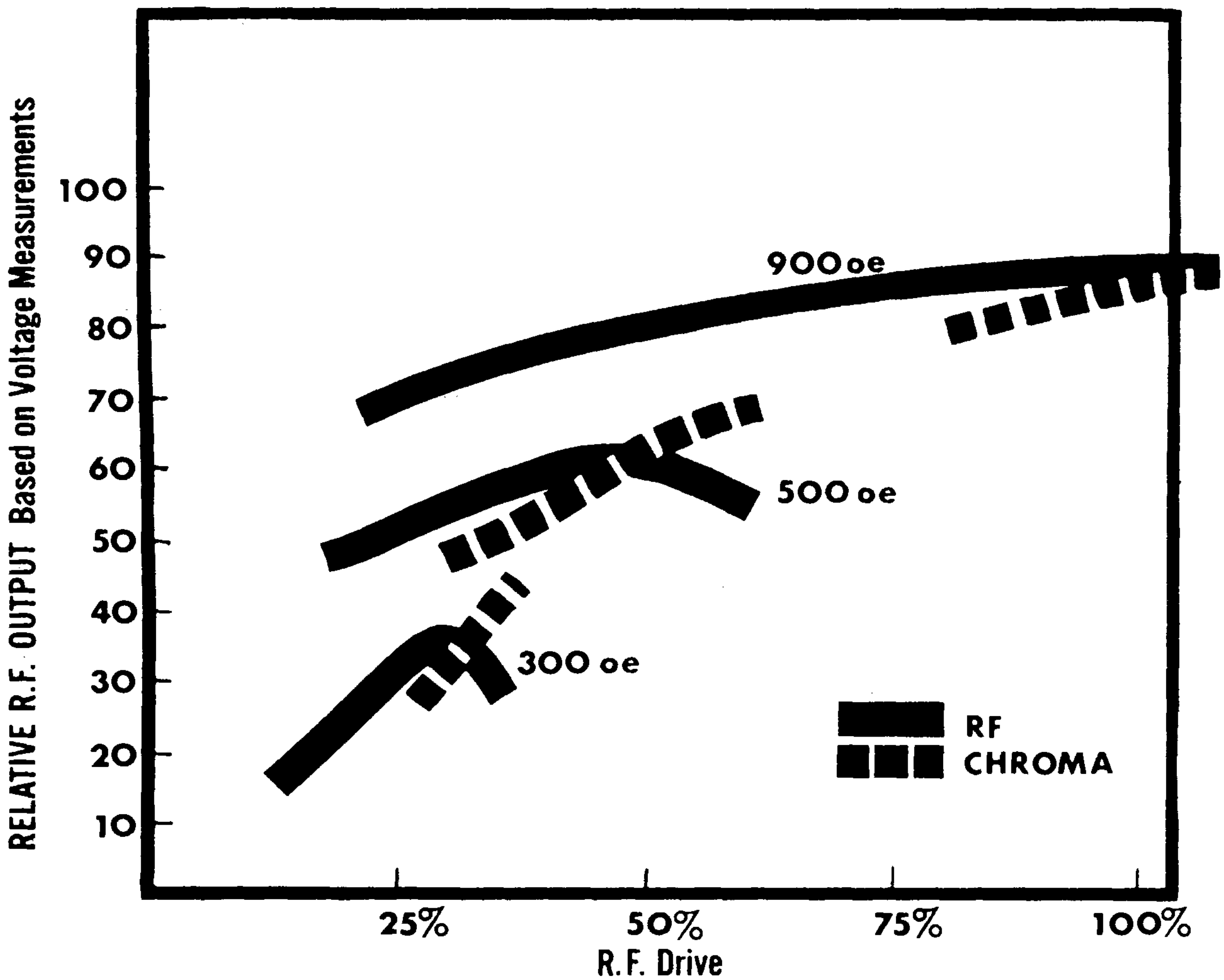
The latest breakthrough in oxide research has equipped us to accomplish many things in the immediate future. Increased RF output and improved signal-to-noise ratios are immediately achievable with the compatible 500 oersted tape and higher coercivity versions promise even further degrees of excellence. We can tailor coercive force of the finished tape product to provide the industry with whatever is needed to improve the quality of video recording. And the best thing about it is that this is not a laboratory dream; the tape is here today.

FIG. 1



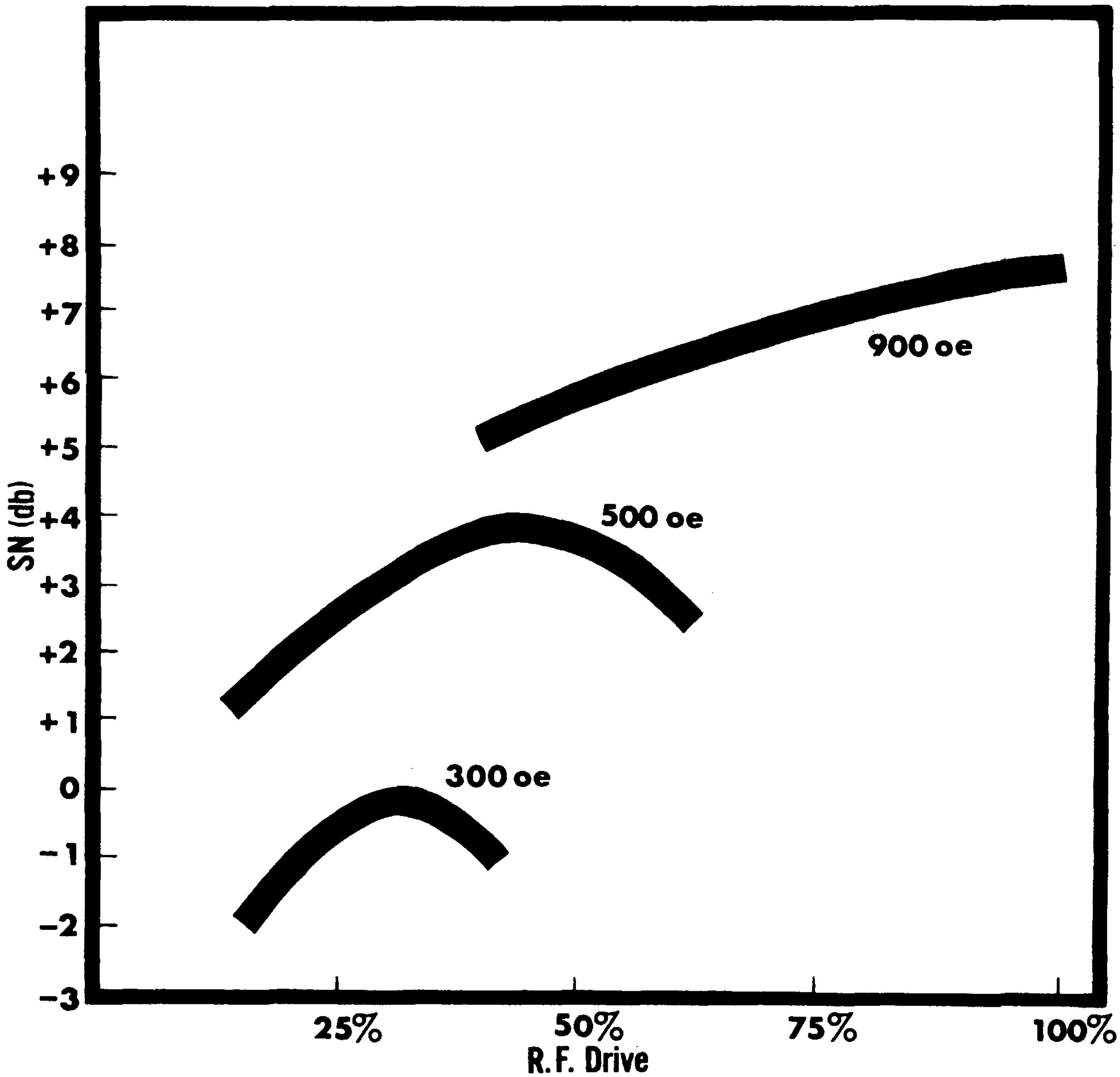
B-H HYSTERESIS LOOPS

FIG. 2



R.F. OUTPUT AND CHROMA RESPONSE VERSUS R.F. DRIVE QUADRUPLEX

FIG. 3



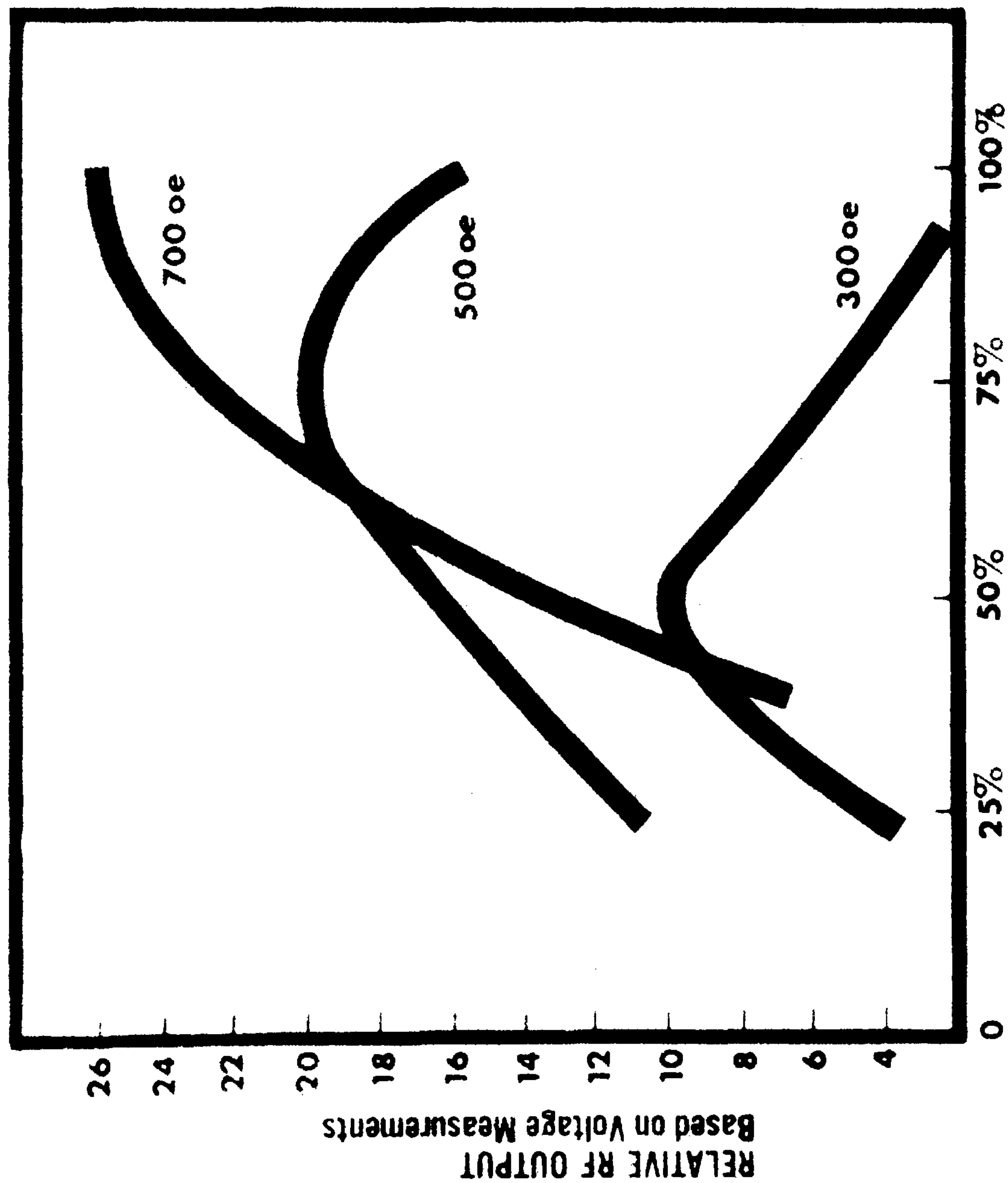
S/N VERSUS R.F. DRIVE
QUADRUPLEX

FIG. 4

	1ST GENERATION		2ND GENERATION			3RD GENERATION			4TH GENERATION			
	Master Tape Type	Established S/N	Dupl. Loss	Work Master Tape Type	Resultant S/N	Dupl. Loss	Edited Master Tape Type	Resultant S/N	Dupl. Loss	End Copy Tape Type	Final S/N	Change From 300oe Master
1.	300oe	50	1.5	300oe	48.5	1.5	300oe	47	1.5	300oe	45.5	−4.5
2.	500oe	54	1.5	500oe	52.5	1.5	500oe	51	1.5	500oe	49.5	−0.5
3.	500oe	54	1.5	500oe	52.5	1.5	500oe	51	2.0	300oe	49.0	−1.0

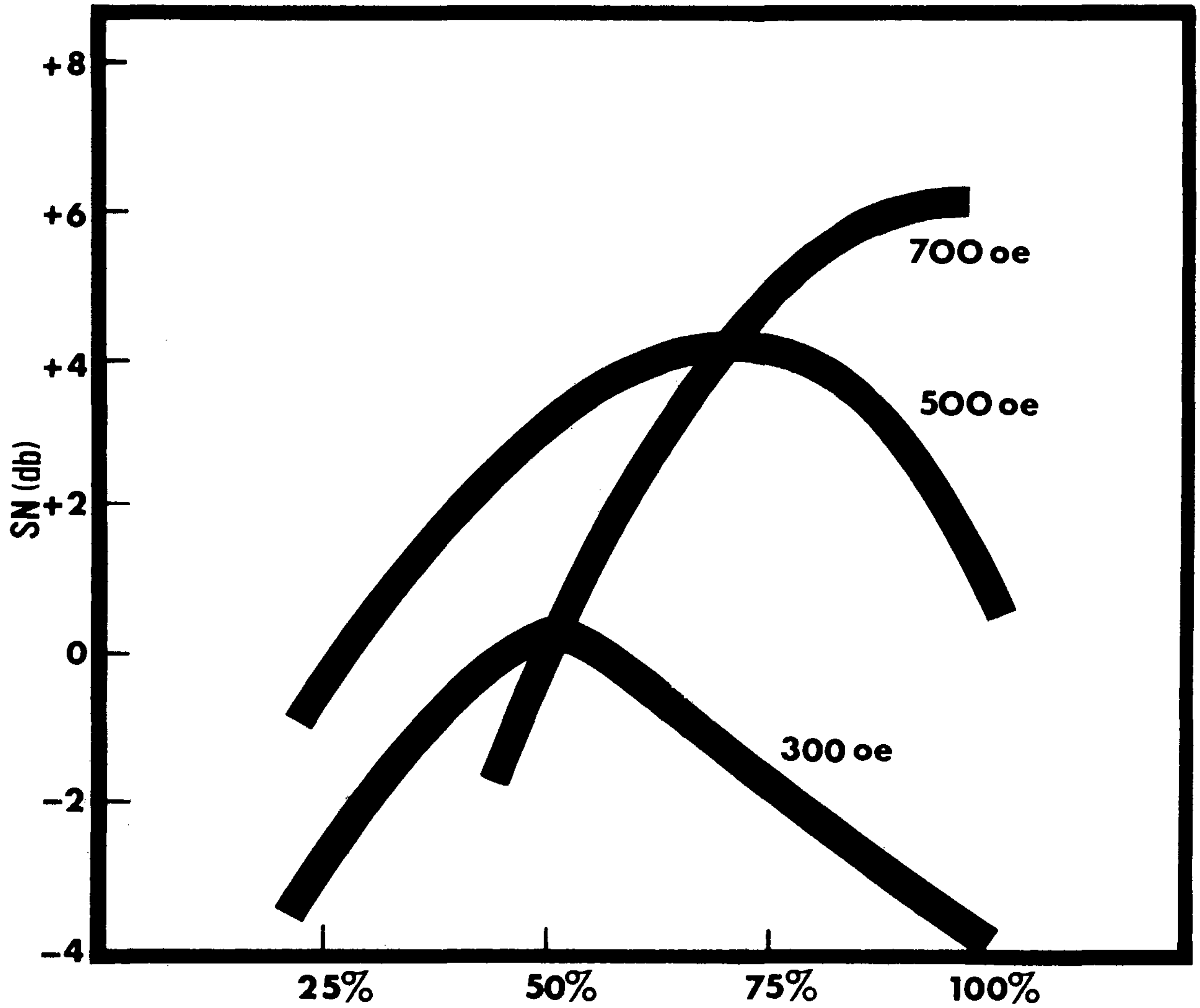
MULTIPLE GENERATION S/N COMPARISON

FIG 5



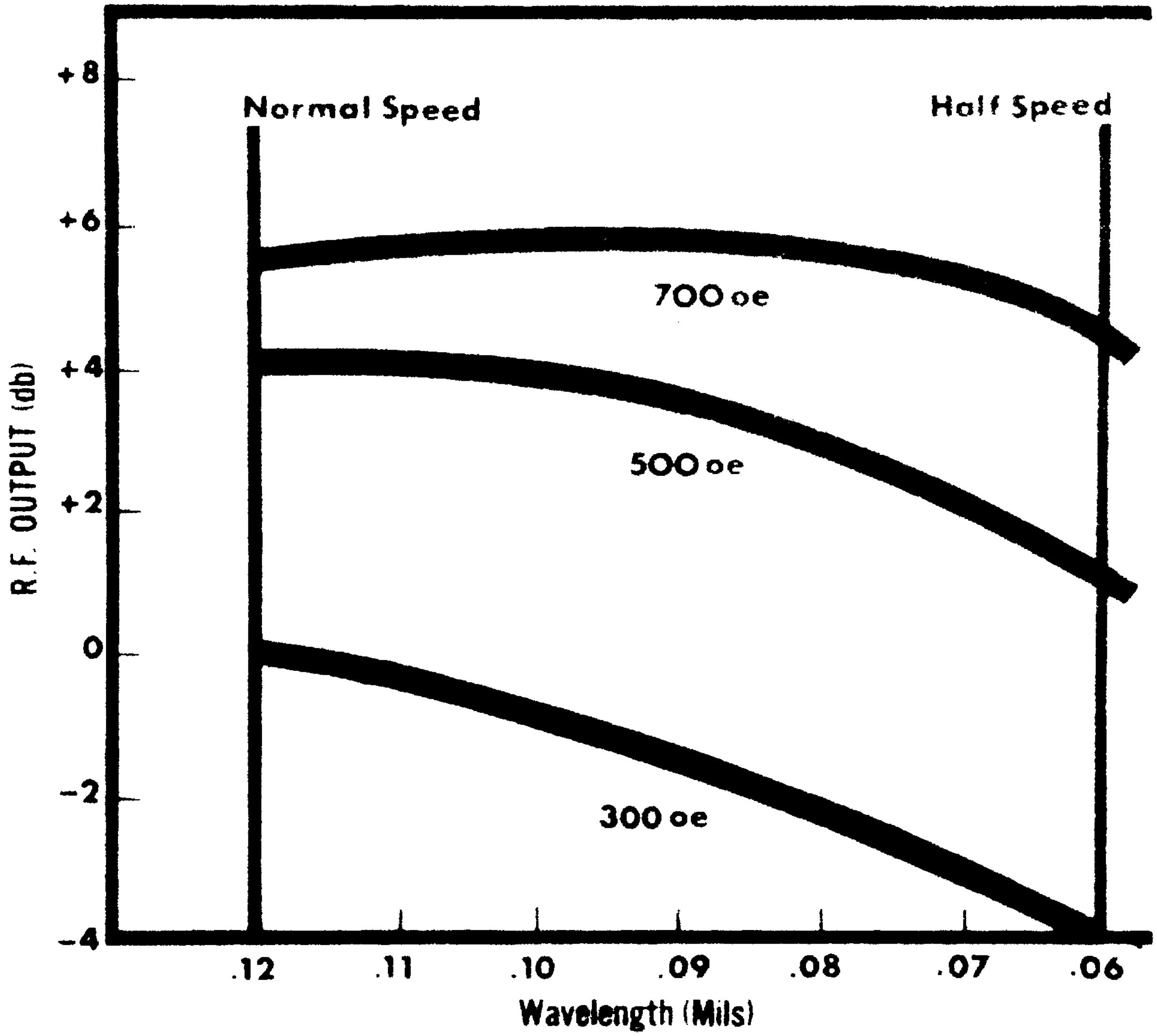
R.F. OUTPUT VERSUS R.F. DRIVE
1" HELICAL

FIG. 6



S/N VERSUS R.F. OUTPUT
1" HELICAL

FIG. 7



R.F. OUTPUT VERSUS WAVELENGTH OPERATING AT 2 SPEEDS

IN-SERVICE NOISE MEASUREMENTS
ON A
CATV SYSTEM

By

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Presented by the author
at the
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Washington, D. C.
July 6-9, 1971

Measurement of the signal-to-noise ratio in communications systems is of fundamental importance - both in evaluating performance and in planning preventative maintenance.

In television systems, the signal-to-noise ratio is measured in terms of the peak amplitude of the picture signal and the RMS noise amplitude. The picture signal amplitude with which we are concerned modulates the picture tube. That is, it is the blanking level-to-peak white level not the sync tip-to-peak white, i. e. $\sim$700 mV. Signal level is easily measured with a television waveform monitor.

Noise cannot be measured in terms of peak amplitude. This is because of the statistical nature of noise. That is, noise (being random) has peak amplitudes which vary with time in a random fashion. Theoretically, if one waits long enough, an infinitely large noise pulse will come along. Noise pulses could be several volts in amplitude, but these occur so seldom that they can be neglected.

Experience has shown that the subjective appearance of noise is related to the noise power level. It has long been the practice to measure noise levels in terms of average noise power, thus avoiding the occasional high noise peaks. Noise power may be expressed in RMS volts. Thus, the measurement of noise is simply the problem of measuring power or RMS voltages.

However, this isn't as easy as it may at first appear. In television, we are concerned with noise power extending to 4 MHz. As the noise voltage is a complex waveform, not a sinewave, we cannot measure peak or average voltage and know the RMS value. In communications engineering, the usual approach to measuring RMS voltage, or noise power, is to measure the heating effect produced by the unknown power dissipated in a known load. This can be carried out in the laboratory to very great precision. However, the circuit must be taken out of service for the test.

When one considers the power levels to be measured (a 40 dB S/N ratio = 6.6 mV in 75 Ω = 0.44 microwatts), it is at once apparent that the equipment must of necessity be delicate and perhaps not well-suited for rugged field use. The skill levels required are not generally available outside of laboratories. In short, measurement of RMS noise levels in the field is not practical. Furthermore, even if practical, it requires that the system is taken out-of-service for the test. In the case of CATV systems, this is not at all convenient.

The ideal test scheme permits accurate in-service testing and uses equipment suitable for field usage. In-service testing requires that the noise may be measured in the presence of the video signal, whose power level varies unpredictably. One scheme is to observe the video signal and noise on an oscilloscope. By observing the noise only on a line or two of the vertical blanking level, video information is eliminated quite easily.

On the other hand, the oscilloscope is fundamentally a peak reading instrument - it does not indicate RMS at all. For true "white noise" there is, however, a conversion factor which some observers have developed to relate apparent noise on an oscilloscope to its true RMS value. This factor has been reported as 14 to 18 dB, the higher value being favored.[1-2]

However, aside from the question of the correction factor, a much more serious source of error lies with the variations in observed noise between different observers, changes in the trace intensity on the waveform monitor, and the apparent noise level changes which occur with the changes in the brightness of the room in which the measurement is made. Variations of 6 dB have been reported due to this latter cause alone.[2]

Clearly, where preventative maintenance is to be determined by signal-to-noise level measurements, much greater repeatability in test methods is needed. As CATV system performance may be a matter of litigation, such large margins in the test results is undesirable.

Tektronix, Inc., has developed a new method which eliminates both sources of error. This method is suitable for field usage and in-service testing. It is based upon the comparison of the noise to be measured with a second, known noise source. Here, the observer needs only to make a comparison, not a

[1] L. E. Weaver, "The Measurement of Random Noise in the Presence of a Television Signal," BBC Engineering Monographs (No. 24), March, 1959.

[2] L. E. Weaver, "Television Video Transmission Measurements."

judgment. His comparison will, in our experience, be repeatable within ±2 dB in every case; and will be within 1 dB in most cases. Different observers will obtain the same results, which are independent of the waveform monitor's intensity or ambient illumination. The waveforms obtained are shown in Figures 1a and 1b, page 3. These figures illustrate the basic concept. Figures 2a and 2b show the results of small variations in noise levels.

In any comparison technique, it must be first determined that like is being compared with like. That is, two noise sources should both have similar distribution of energy over the same frequency spectrum.

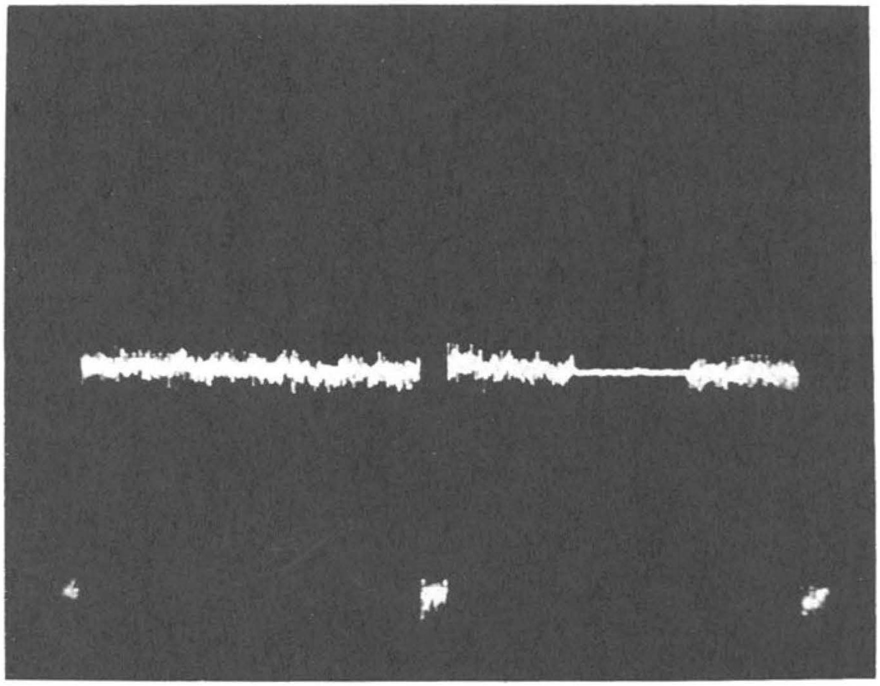
Fortunately, the random noise encountered in CATV systems (which accounts for snow in the picture) has approximately equal energy at all frequencies within the video band. This is called "white noise," and is readily generated by electronic means. Such a noise generator may be calibrated in the laboratory using a true RMS power measuring instrument. The long-term variations in RMS noise power will be small, hence frequent recalibration is not considered necessary. True RMS calibration in the factory avoids the question of the appropriate conversion factor mentioned above, with its spread of 4 dB. The only precaution in this measurement technique is that the noise being measured and the noise it is being compared with have the same frequency spectrum, or nearly so.

In many television systems, the bandpass extends considerably above 4 MHz. This is true of cameras and microwave radio relay links and may also be true of some demodulators. As noise above 4 MHz does not in general degrade the picture, it is desirable to exclude it from any measurements. White noise generators are inherently wideband. Their output includes significant energy above 4 MHz. Modern waveform monitors have only gradual roll-off in frequency above 4 MHz.

A low-pass filter, flat to 4 MHz, then exhibiting very rapid attenuation above 4 MHz, is required. Such filters are readily manufactured for 75 Ω circuits. Their design is given in the CCIR Volume 5 recommendation 451-1. These 75 Ω low-pass filters are essential to noise measurements. The filter must be placed in the measurement setup so that both noise signals are affected by it. A typical measurement setup is shown in Figure 3. The signal source shown is a CATV demodulator. The action of the Tektronix Type 147 is shown functionally in Figure 4.

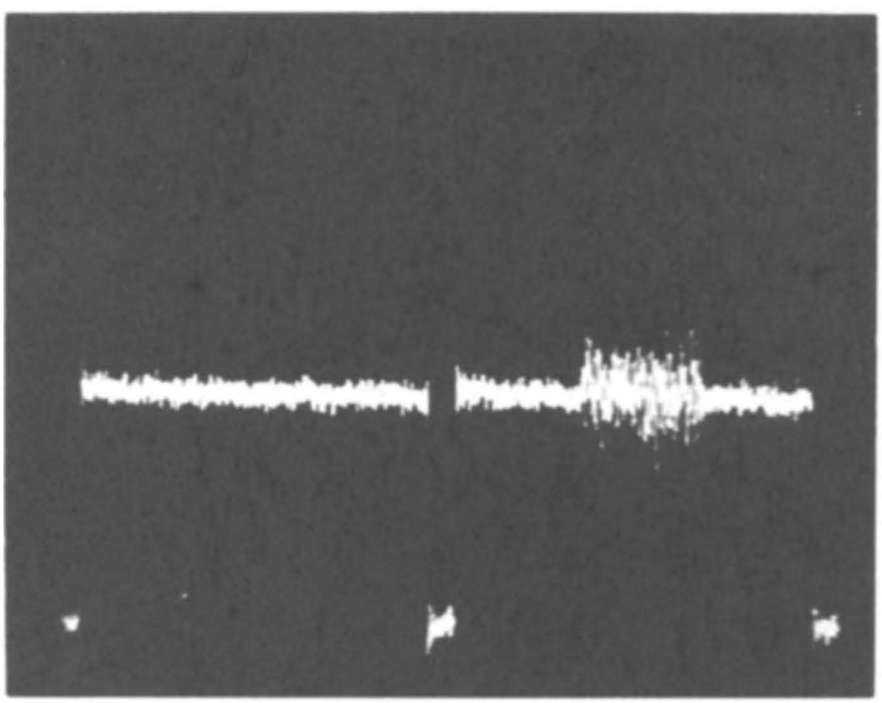
NOISE INSERTION MODE

Noise is deleted from center of chosen test line in vertical blanking interval. Noise from the noise generator in the Tektronix Type 147 is inserted in center of the chosen test line.



Inserted noise is much less than noise on signal

Figure 1a

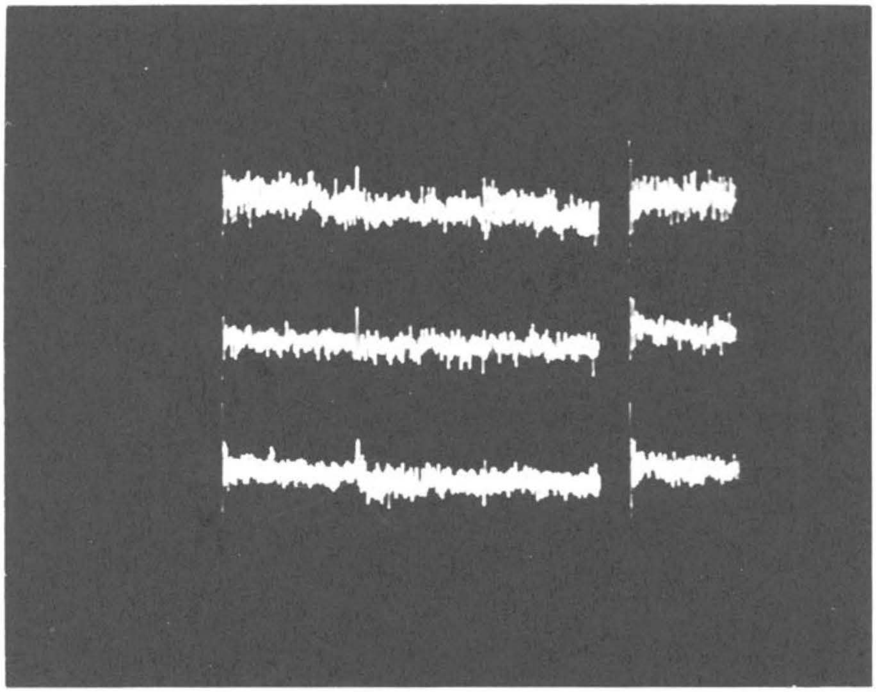


Inserted noise is much greater than noise on signal

Figure 1b

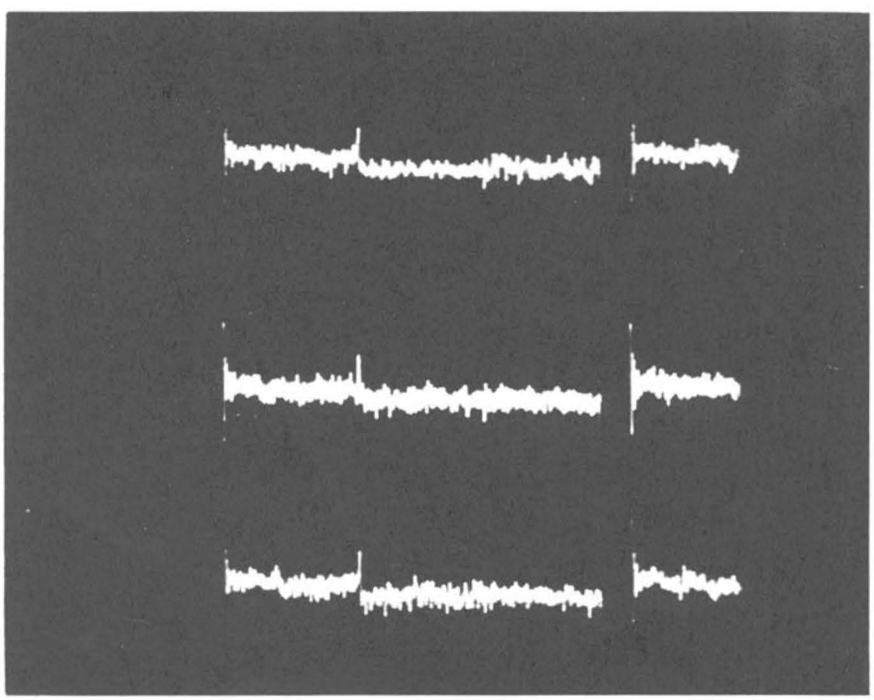
NOISE LEVEL DISCRIMINATION

Waveform Monitor Response - "Flat"
Noise Bandlimited - 4 MHz



±2 dB

Figure 2a



±3 dB

Figure 2b

IN-SERVICE NOISE MEASUREMENT

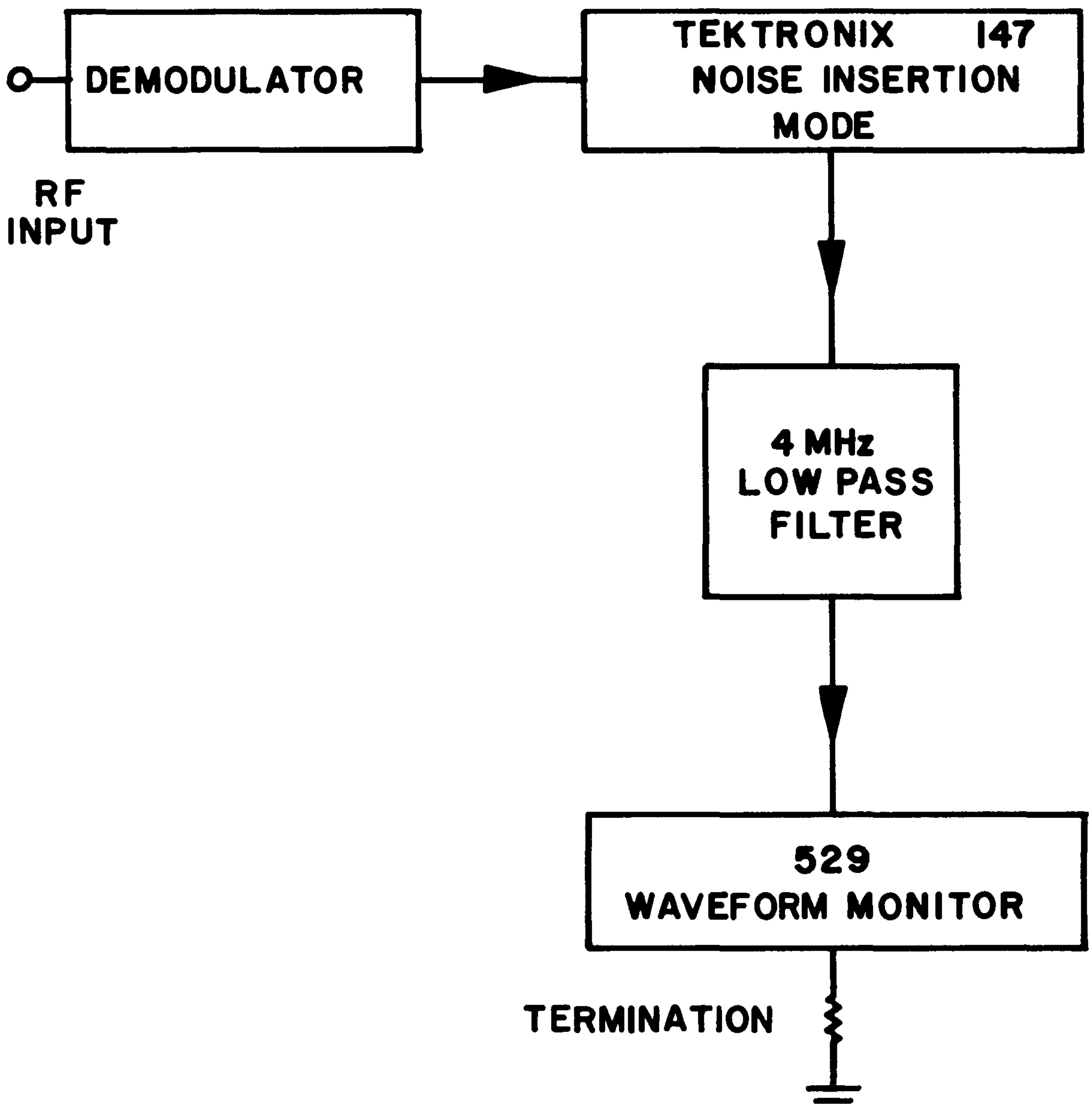
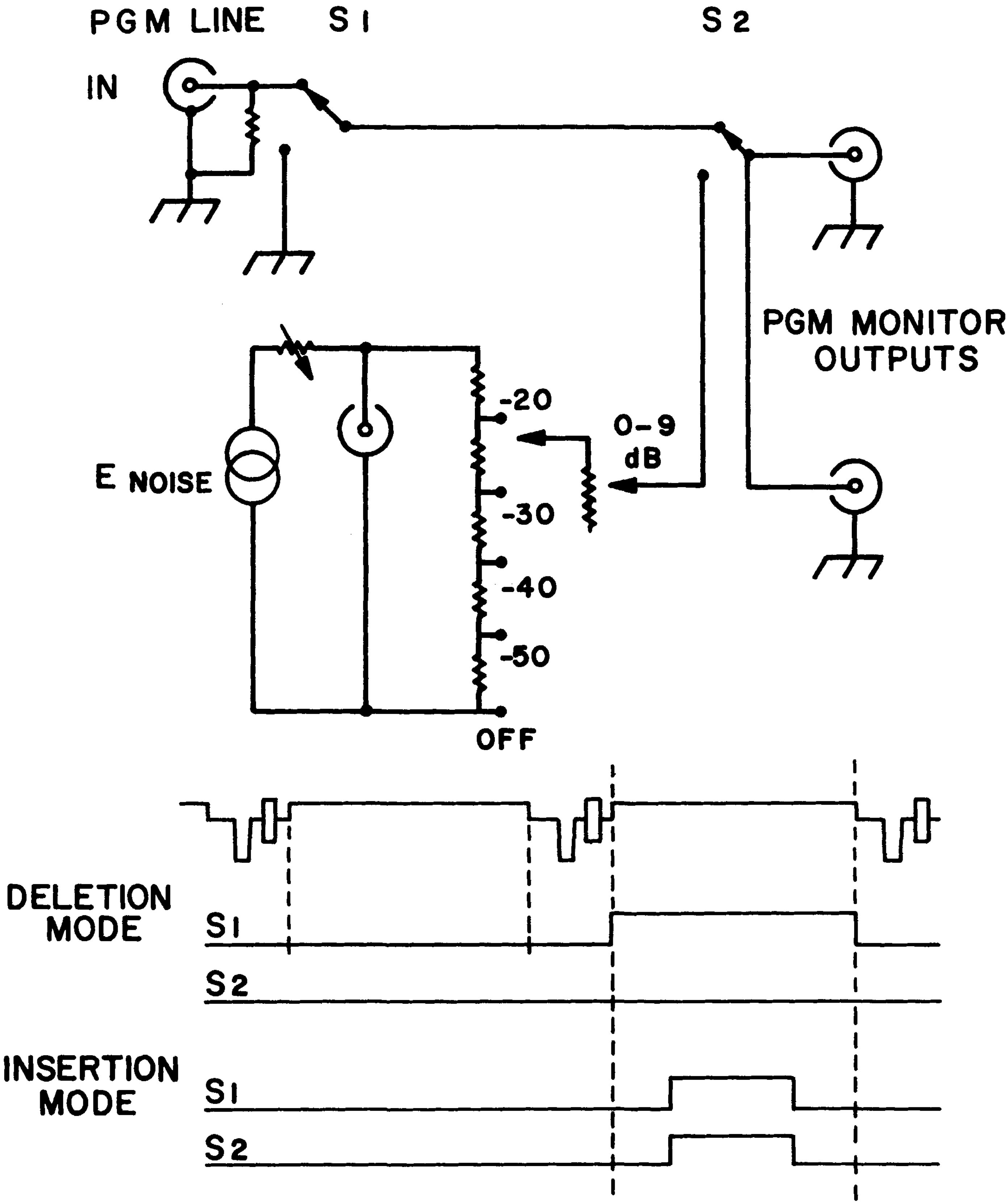


Figure 3



TEKTRONIX 147 NOISE MEASURING PLAN

Figure 4

Switches S_1 and S_2 are, of course, electronic switches. They operate during the chosen noise test line within the vertical blanking interval. S_1 is the deleter switch. When measuring noise, the instrument is said to be in the INSERTION MODE. Then, S_1 disconnects the incoming video signal from the input to S_2 during the middle half of the test line. At that time, switch S_2 substitutes the output of the built-in white noise source for the incoming video signal at the monitor outputs. One of these monitor outputs feeds the 4 MHz low-pass filter which removes all noise above 4 MHz from both incoming signal and locally generated noise. It is essential to terminate the filter at the waveform monitor. The effect of the 4 MHz filter is shown in Figure 5.

In the lower portion of Figure 5, the filter was connected incorrectly between the demodulator output and the Type 147 input. Here, the inserted noise observed on the waveform monitor is wideband, and the system noise (preceding and following the inserted noise) is bandlimited. The difference is much more apparent in viewing the monitor than is shown in the photograph.

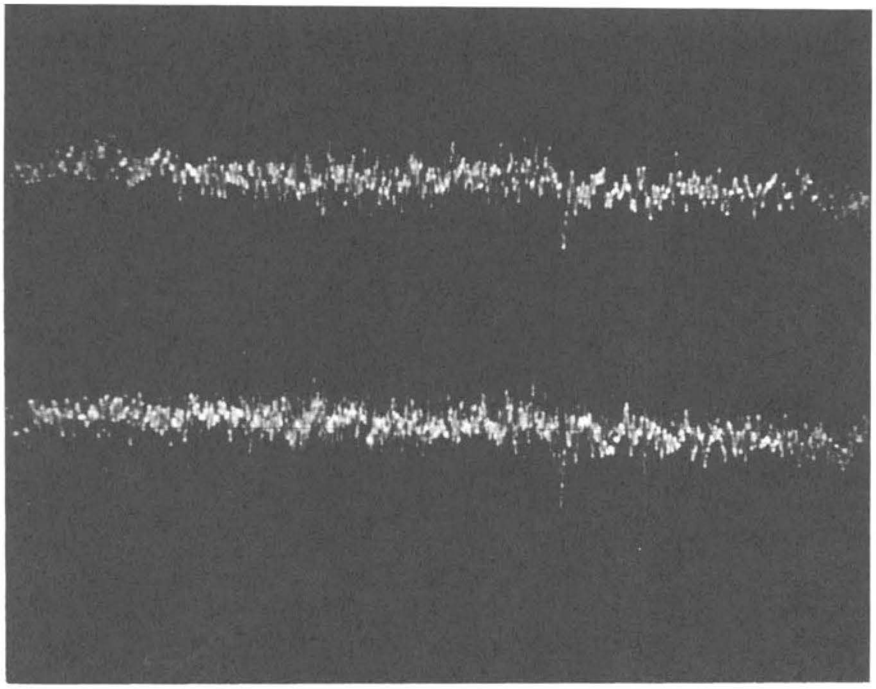
In the DELETION MODE, S_1 operates, disconnecting the incoming video during the entire active portion of the chosen test line. The action of S_1 and S_2 is so timed that no sync pulse or color burst is deleted. S_2 does not operate in the DELETION MODE. Video signals coming from the Type 147 in the DELETION MODE do not have any noise present during the active portion of the test line. This is shown in Figure 6.

The DELETION MODE is used where it is desired to measure the noise which is occurring within a CATV system. For example, at the head end site, where a microwave radio relay is fed baseband video from a demodulator, the Type 147 may delete all noise present at the output of the demodulator. In Figure 7, a 147, operating in the DELETION MODE, is shown connected to the microwave transmitter's video input.

At the output of the microwave link, a second 147 (see Figure 8), is operating in the INSERTION MODE. Noise is measured on the waveform monitor. Of course, the principle can be extended down the cable system to measure noise at the furthest subscribers' drop. This setup was shown in Figure 3.

Line 16 is well-suited for noise measurements. Under present FCC Rules, Line

EFFECT OF 4 MHz BANDLIMITING



Top

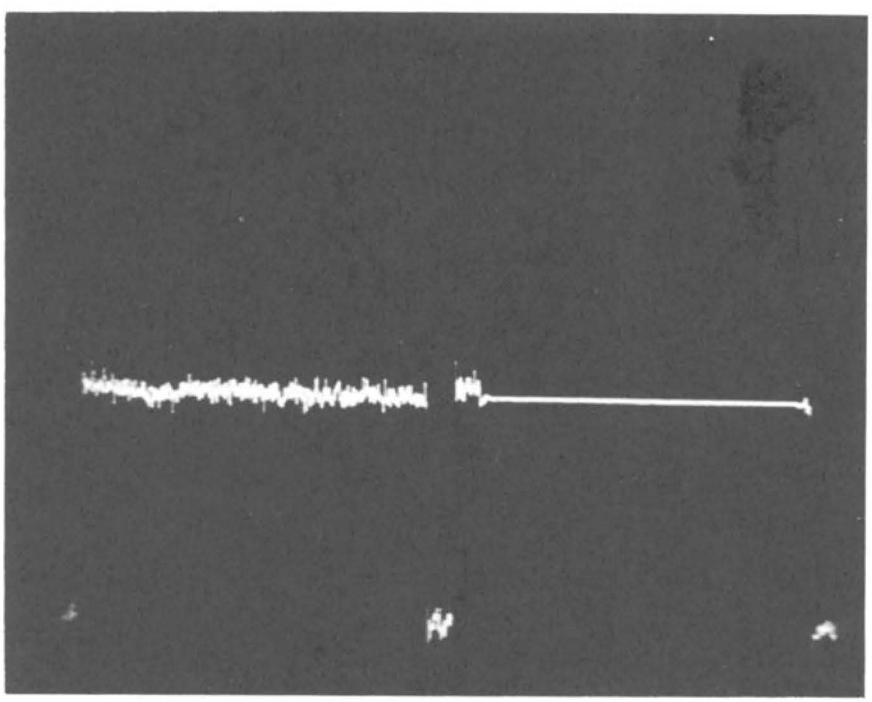
System and inserted noise both bandlimited to 4 MHz and matched at 40 dB.

Bottom

System noise bandlimited to 4 MHz, but inserted noise not bandlimited.

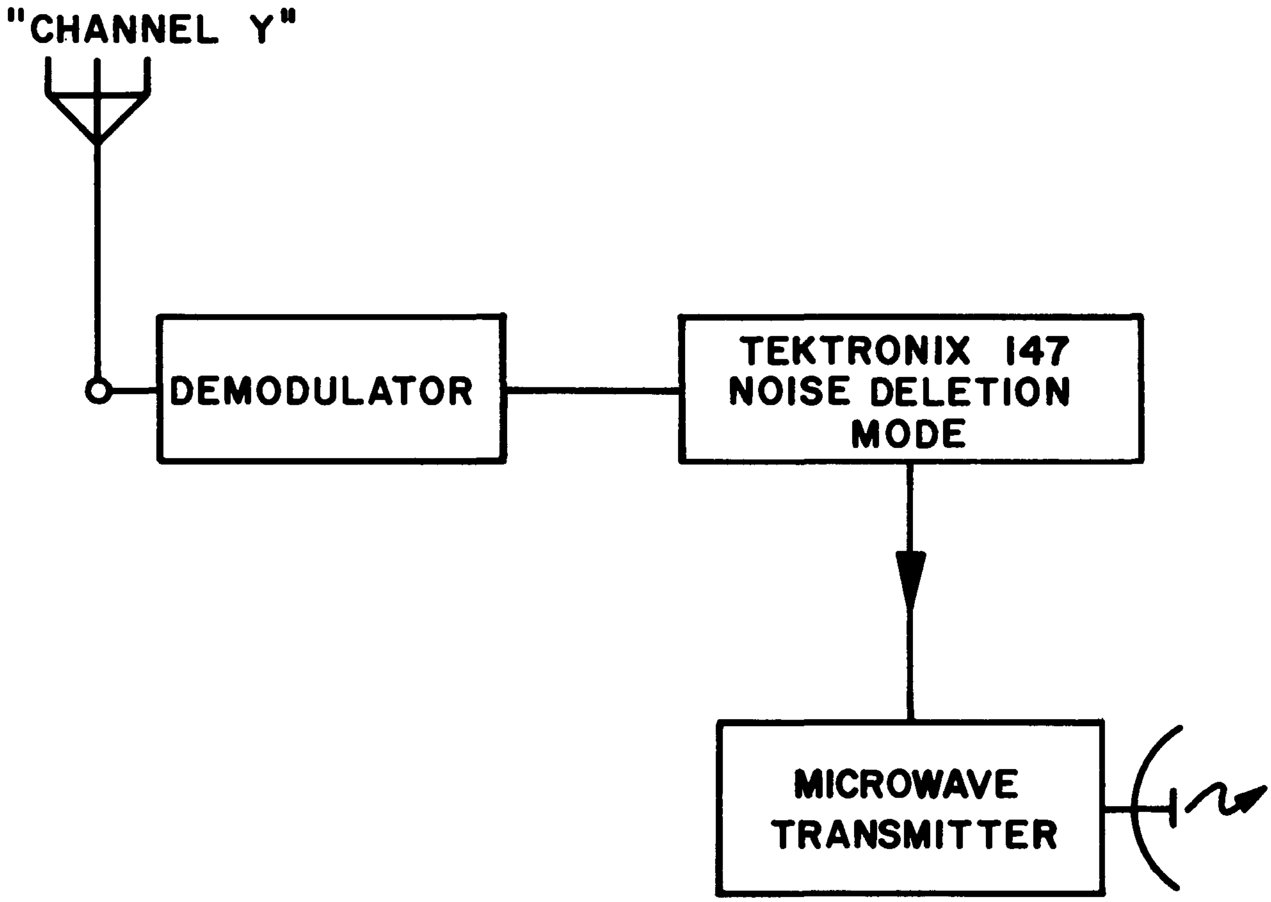
Figure 5

NOISE DELETION MODE



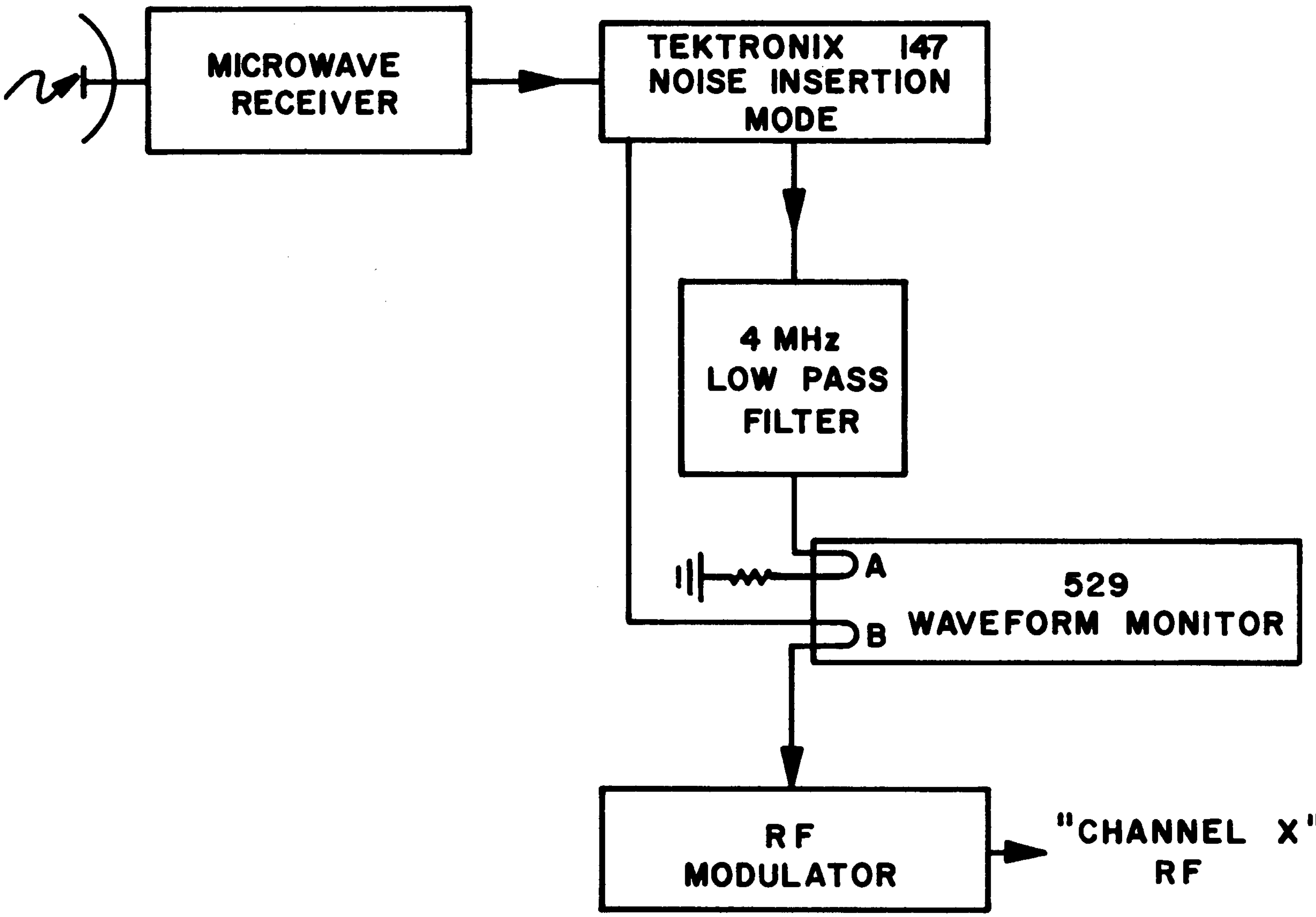
Display shows two successive lines in vertical blanking interval as shown on waveform monitor using the line selector.

Figure 6



IN-SERVICE MICROWAVE RELAY
NOISE MEASUREMENT
VIT SIGNALS MAY ALSO BE INSERTED

Figure 7



IN-SERVICE MICROWAVE RELAY NOISE MEASUREMENT

MICROWAVE VIT SIGNALS MAY BE MONITORED

Figure 8

16 cannot be modulated. Lines 17 - 20 may be used for test and other purposes in the near future. Line 21 may carry video at times. Noise levels on Line 21 may be higher than on Line 16 because Line 21 may carry noise from a video tape recorder. Hence, Line 16 is the optimum noise test line for CATV.

The fact that noise measured on Lines 16 and 21 may be different in measuring off-the-air signals is due to the broadcasters' practice of employing video processing amplifiers at the input of the transmitter. These processing amplifiers act as sync and blanking deleters, thereby removing noise distortion which the program sync and blanking have suffered. New sync, blanking and, sometimes, burst are reinserted by the broadcasters' processing amplifier. Noise on Line 16 will usually be deleted, leaving it a quiet line; while noise on Line 21 may not be deleted.

Noise is frequently measured through a noise weighting filter which is designed to attenuate high frequency noise components as the higher frequency noise components are less objectionable in the picture. If all sources of noise had the same power/frequency distribution, noise weighting would not be necessary. Noise in CATV systems is "white noise" and no weighting is necessary. Noise arising in microwave (FM) relay links is not white noise. Its noise spectrum is called triangular (noise rising 6 dB/octave from about 200 kHz).

If all the noise to be measured were "triangular," a filter could be included in the Type 147 noise generator to give triangular noise. However, in practice, the noise level in television signals is not all contributed by the FM portion of the system so a triangular noise spectrum is not suitable. One way to avoid the problem of "triangular" vs "white" noise is to measure the noise using both the FLAT and CHROMA bandpass characteristics of the waveform monitor. The 4 MHz low-pass filter is used in both cases. The two numbers give information about the frequency distribution of the noise being measured. This concept has yet to be developed and field tested. It is offered in the hope that CATV engineers may wish to pursue the matter.

Noise measurements may prove useful in planning preventative maintenance of the system. Routine noise level monitoring at well-chosen points within the system is not enough. Where careful records are kept of performance, the system's long term performance can be determined. Obviously, where noise levels are increasing with time, a condition is located which is going to require correction.

In summary, the noise level of standard 1.0 volt video signals may be accurately measured by deleting the noise on part of a vertical blanking line and inserting "white noise." By comparison, on a waveform monitor the level of inserted noise is adjusted to equal the noise being measured. The noise level is determined from the settings of the calibrated noise attenuator. Bandlimiting is required between the Tektronix Type 147 monitor output and the waveform monitor. The method is especially well-suited to the CATV industry.

LEVEL CONTROL CONCEPTS FOR MULTICHANNEL AND TWO-WAY SYSTEMS

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ABSTRACT

System level control design philosophy is presented for multichannel and two-way transmission systems. Techniques for channel selection are introduced. Appendix II and III covers these techniques. These techniques are used to reduce the effects of second-order and third-order distortion products. Reasons for not standardizing pilots are clarified in conjunction with multichannel requirements and pilot placement.

Amplifier design approaches follow to fulfill system requirements. A novel automatic gain control (agc) amplifier design incorporated at Anaconda Electronics Company is discussed. Design techniques are used which provide level stability and desirable transient response.

New amplifier design approaches are presented. These are termed reverse gain control amplifier and weighted agc amplifier.

I. INTRODUCTION

Today new and advanced technology is causing greater capability, flexibility and diversity for every new and existing cable TV system. Consequently new philosophies are emerging to control system parameters. Particularly important is the control of exogenous parameters such as channel placement, signal levels and pilot carrier placement. These are essential elements in the vector of control for each system design whether it is for one-way or two-way usage. Control of these elements is required for reliable system operation.

The system level control philosophy presented is in a general context so it will apply to a large class of system. Level control is required to maintain adequate signal to noise ratio and acceptable distortion levels. These parameters are related to channel selection, frequency response, and pilot carriers.

Two-way system level control is accomplished by use of open and closed loop control techniques combined with single and dual cable designs. This way, the resulting systems remain simple and economical. The system with short haul trunk lines originating at the head end or hub type of system uses reverse gain control amplifiers. For long haul trunk lines dual cable is used with agc amplifiers and combining networks.

II. LEVEL CONTROL DESIGN PHILOSOPHY

A properly designed system, be it one-way or two-way, has safety margins built in to maintain adequate signal to noise ratio and distortion at suitably low levels. These safety margins are required to accommodate variations in amplifier performance with input signal level variation, temperature, inaccuracies in desired amplifier placement, and amplifier control settings (gain and equalization). The input signal level variation at a particular amplifier is due to the frequency response of the preceding amplifiers being temperature sensitive as well as the attenuation of the preceding cables being a function of temperature. Cable attenuation in decibels is described by

$$A = q_o + q_1 f^{\frac{1}{2}} + q_2 f \qquad (1)$$

where f is frequency and the other terms are a function of temperature and a particular manufacturer. The optimum amplifier design in terms of noise figure, distortion, load capacity, gain distribution and other costs will not be considered here. A conjugate gradient optimization algorithm can be used in such a design. References for conjugate gradient techniques are [1] and [2].

The fundamental problem here is the control of amplifier output levels. The need to control these levels is readily seen from equations (2) and (3) for signal to noise ratio and cross modulation respectively [3].

$$10 \log(S/N)_m = S - G - F_{db} - 10 \log KTB - 10 \log m \qquad (2)$$

$$XM_m = XM_R + 2(S_R - S) - 10 \log(N_c - 1) - 20 \log m \qquad (3)$$

where
S = amplifier operating output in dbmv
G = amplifier gain in db
The other terms are defined in Appendix I. These equations, in a less general form, have been used in earlier papers [4], [5]. Second order distortion terms may be a factor as well as other third order products. Appendix II and III discusses distortion and channel selection respectively. For a particular channel allocation design, a means of evaluating intermodulation distortion is given by

$$IM_m = IM + (S_R - S) - 10 \log m \qquad (4)$$

where
IM_m = intermodulation at the output of the mth amplifier in cascade

IM = intermodulation at the output of a single amplifier

S_R = amplifier reference output in dbmv

S = amplifier operating output in dbmv

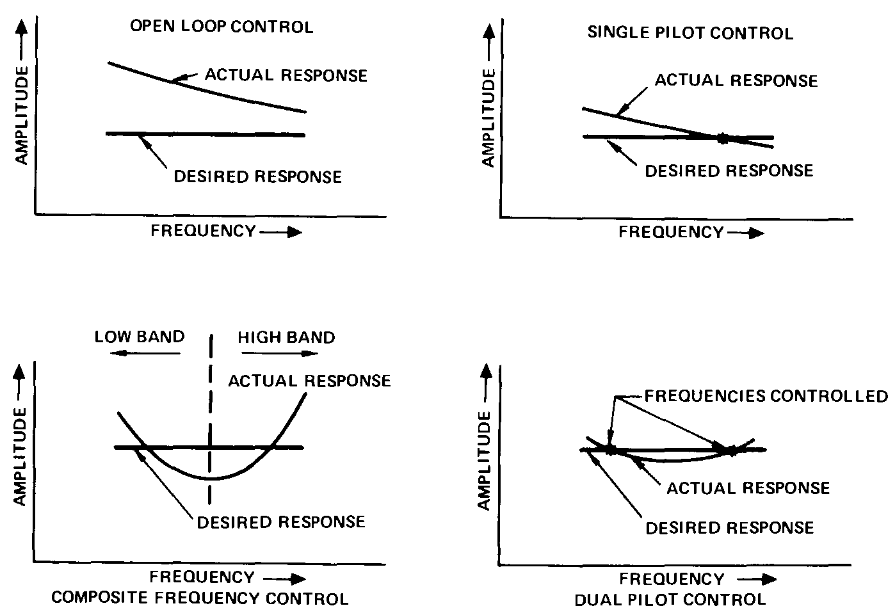
These equations are for what is termed a "well behaved amplifier" [5]. Furthermore, they assume all amplifiers have the same distortion and noise properties. Additionally, all amplifiers are supposed to have equal gain and output levels. Tests, over temperature, must be conducted to correlate computed results, from these equations, with actual test data.

Once correlation is completed, bounds on the system performance can be established from these equations. Thus, minimal standards of performance for signal to noise ratio and distortion are assured provided the output levels are controlled.

A suitable technique of level control advocated is the combination of open and closed loop control in the form of thermally compensated and agc amplifiers respectively. The ratio between the number of thermally compensated amplifiers and agc amplifiers in a cascade depends on the system noise and distortion requirements as they relate to temperature range and cascade length.

III. LEVEL CONTROL TECHNIQUES

The level control techniques presently used in one-way systems are reviewed in Figure 1. The shortcomings of open loop control and single pilot control are reviewed graphically. Any control scheme consisting of more than two carrier frequencies can be defined as composite frequency control. Usually the response is averaged over two bands of frequencies. In dual pilot control one pilot usually controls the amplitude at a single frequency while the other pilot controls the degree of equalization based on the amplitude at another frequency.



LEVEL CONTROL TECHNIQUES (CASCADE RESPONSE)

Figure 1

IV. CONTROL TECHNIQUE SELECTION

Since manufacturing tolerances exist between cables and uncertainties in response flatness exist with placement, temperature, taps, splitters, etc., an adjustment capability must be provided in at least some of the trunk line amplifiers. This further complicates the level control problem since we demand response stability with temperature changes for the multitude of adjustment possibilities.

With the amplifier frequency response changes over temperature reduced as well as practical, the next step is to select an agc approach with corresponding pilot placement which will provide for adequate level control. Also, pilot carrier amplitude and its effect on distortion must be considered and the pilot carrier(s) controlled must be representative of the process. This can be done by placing the pilot carriers at usable frequency band edges or within the band as shown in Figures 2 and 3. Here, frequency response, over temperature changes, is shown to be related to pilot carrier placement.

For a limited bandwidth and a short haul trunk line, thermal compensation combined with single pilot control may be adequate. Proper application of equations (2), (3), and (4) in conjunction with the channel selection techniques of Appendix III will generally answer this question.

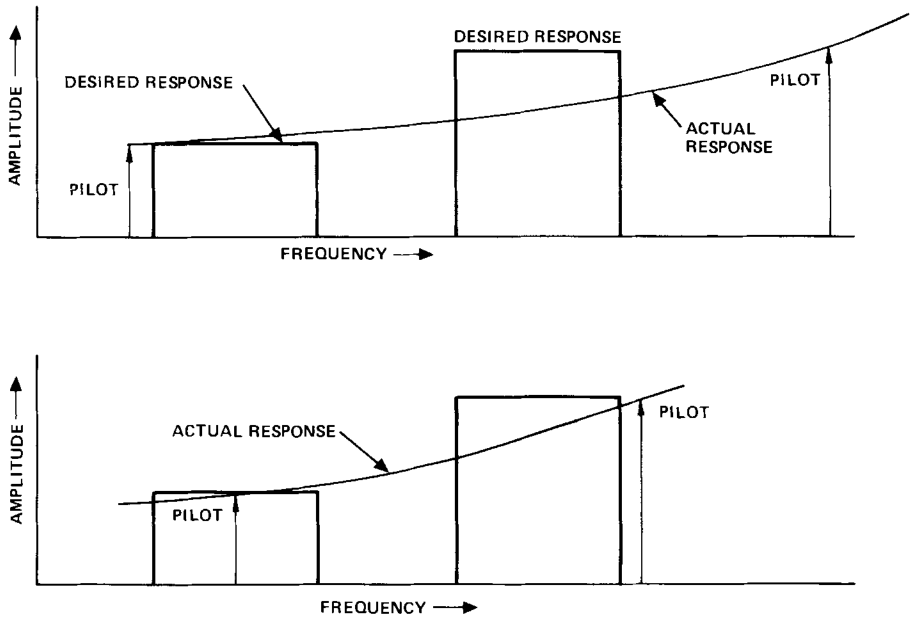
Since every equipment manufacturer's trunk line equipment has a different frequency response with temperature the techniques used or required for adequate control are different. Also for different systems with varying channel and bandwidth requirements it may be necessary to have pilot carrier frequency flexibility for improved level control.

There are also different points of view concerning dual, composite and modulated pilots within the industry.

Pilot carrier frequency standardization could unnecessarily limit system performance. Furthermore, present designs could be obsoleted by standardization rulings. Any plans to standardize pilots will have to satisfy individual system requirements in terms of bandwidth and channel selection.

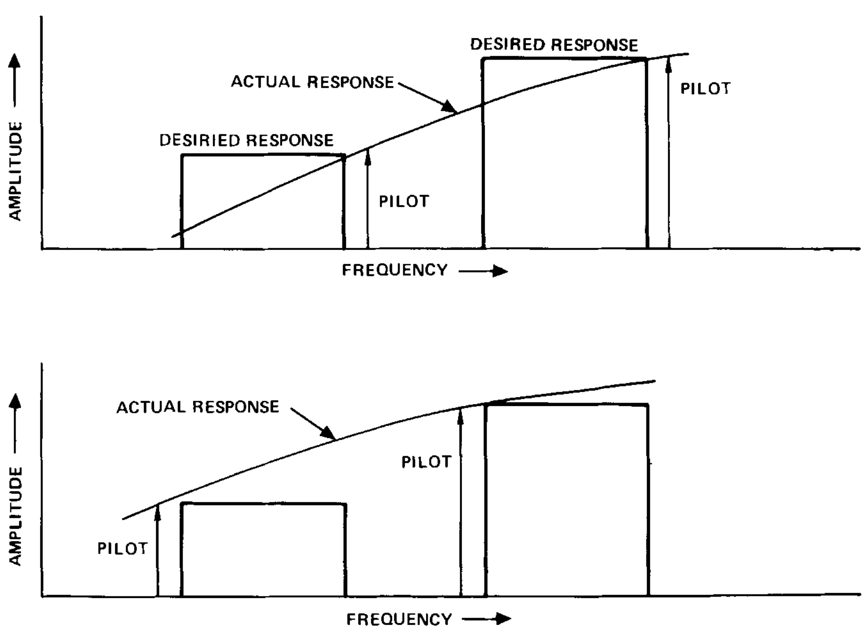
V. TWO-WAY SYSTEM LEVEL CONTROL HIERARCHY

A level control design philosophy, consisting of open and closed loop control techniques, can be applied to the reverse direction of a two-way system. Equations (2), (3), and (4) do not generally apply for the reverse direction. A correction noise term related to number of subscribers is required in equation (2). The information coding techniques incorporated will affect the system distortion requirements. Also access techniques to reduce queueing and prevent overload conditions are needed. Level control techniques may be considered independent of these factors. This viewpoint is pursued here.



PILOT SELECTION EXAMPLE (CASCADE RESPONSE)

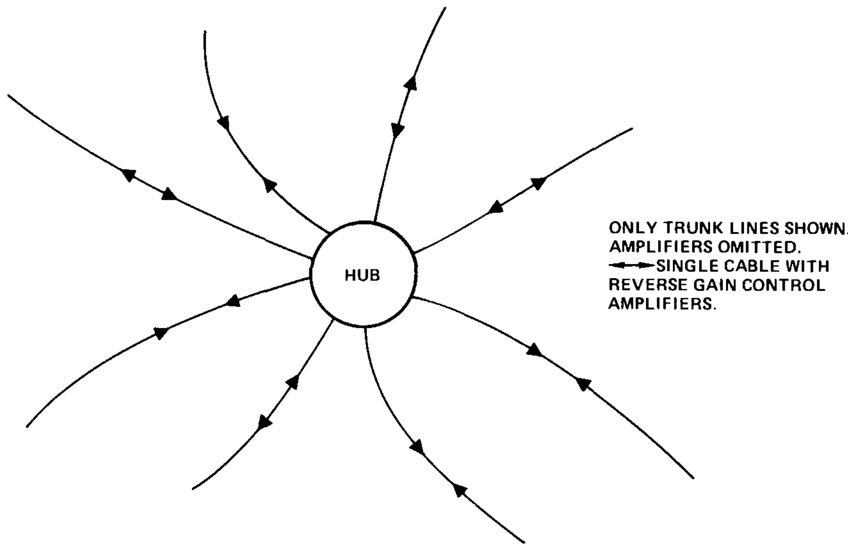
Figure 2



PILOT SELECTION EXAMPLE (CASCADE RESPONSE)

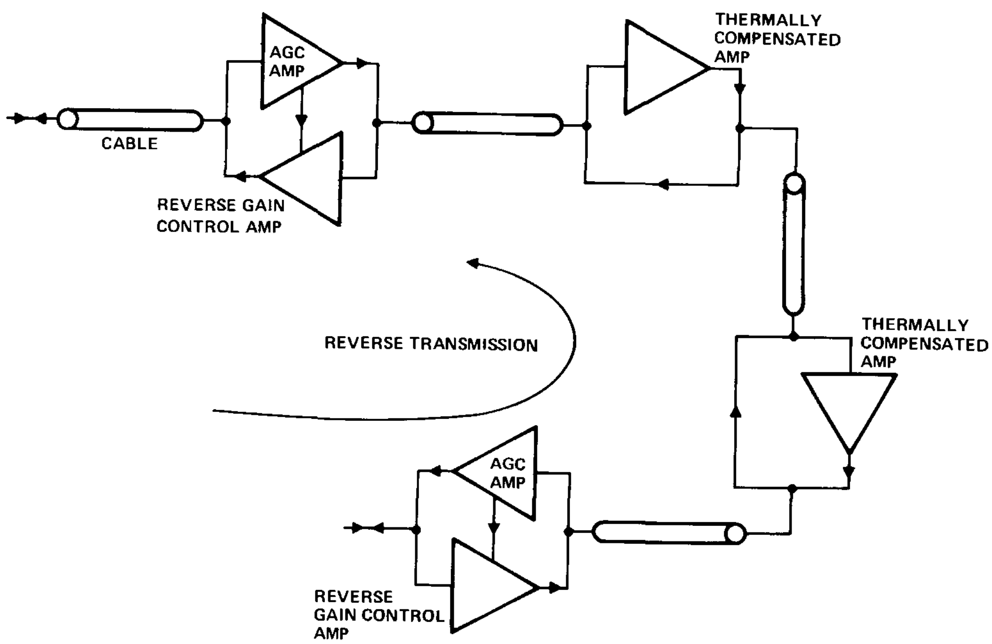
Figure 3

For short haul trunks or a hub type of system, open loop control can be applied. This approach is indicated in Figures 4 and 5 where reverse gain control amplifiers are used to control the gain.



TWO-WAY SYSTEM STRUCTURE (HUB SYSTEM)

Figure 4



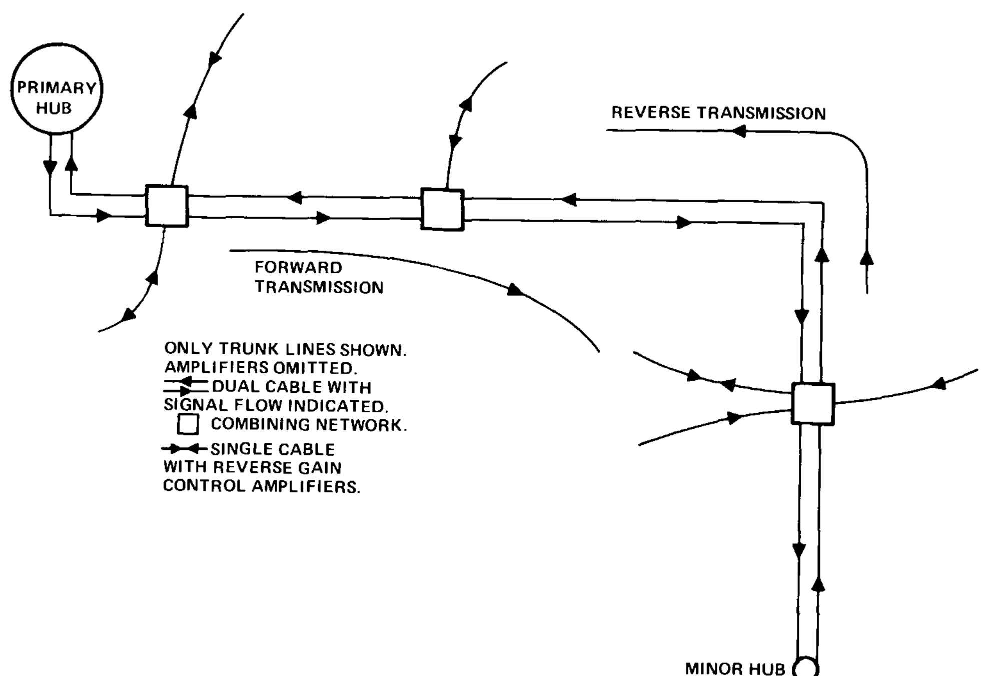
TWO-WAY COMMUNICATIONS WITH REVERSE AMPLIFIERS

Figure 5

Figure 5 illustrates this concept where the split band filters are omitted for simplicity. Reverse gain control amplifiers are used as needed and their gain is controlled by the agc amplifiers in the forward direction. Since this is open loop control its application is limited and for a long haul trunk a minor hub with a pilot carrier generator is required.

Since a system's structure is not made up entirely of long cascades, a combination of open and closed loop control techniques can be applied. Figure 6 shows such a system where a minor hub with a dual cable is provided. The dual cable portion of the system makes level control more easily achievable from an engineering standpoint. Added cost is the price paid for this approach. For short trunks which are branches of this long trunk line open loop control is used with reverse gain control amplifiers and split band filters. Since these cascades are relatively short, acceptable group delay and frequency response characteristics can be realized to achieve the desired system performance.

The combining networks of Figure 6 can also contain access electronics as part of a subscriber selection scheme. In this manner noise at the head end for the reverse direction can be maintained at acceptable levels.



TWO-WAY SYSTEM STRUCTURE (LONG HAUL TRUNK)

Figure 6

A savings in cost may be realized by exploiting the system map to find a hub type of structure where reverse gain control amplifiers can be utilized and long haul trunk lines avoided. This could also result in improved performance with a possible cost reduction for the forward direction. Due to the simplicity of the reverse gain control amplifier the major cost factor is the additional size of the housings and the addition of two-way filters. This is small compared to the extra cable, housings, level control electronics, and powering circuitry for the dual cable system. Clearly, a cost reduction is achievable using the reverse gain control amplifiers.

As two-way system design requirements become better defined, systems and manufacturers with flexibility and capability will be adaptable to the needs of the industry. Existing single cable systems that can incorporate two-way transmission will be needed. New system layouts incorporating dual cable will have insurance for future two-way usage. For these reasons a combination of open and closed loop control techniques are needed.

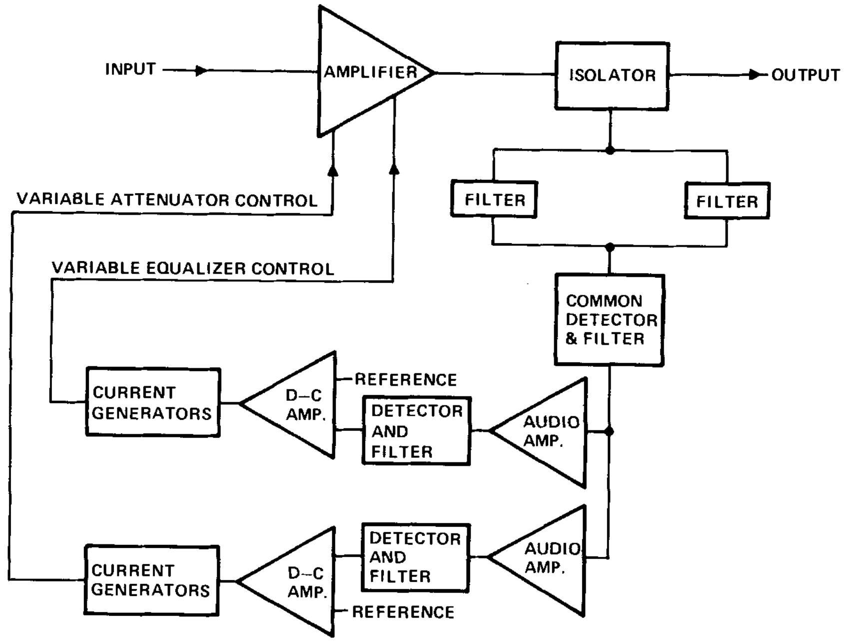
VI. ONE-WAY AGC AMPLIFIERS

Some of the hardware required in a one-way multichannel cable system is a dual pilot agc trunk line amplifier and a agc distribution amplifier or line extender.

The agc trunk line amplifier utilizes a dual modulated pilot and advanced design techniques to achieve long term stable operation and desirable transient response.

A block diagram explaining the design is presented in Figure 7. Here the signals from the two parallel bandpass filters enter a common averaging detector and filter. Economy and additional filtering of channels adjacent to each pilot carrier is achieved in this manner. At the detector output the unfiltered frequencies are designated f_1 and f_3. They are the modulating frequencies for the pilot carrier frequencies f_2 and f_4. Thus, low level d-c detection schemes requiring extremely stable components and references are avoided. Additionally, the audio amplifiers have bandpass responses for the frequencies f_1 and f_3. The signals at the audio amplifier outputs are large in amplitude compared to the signals out of the common detector. Now large signal detection follows and the resulting signals are compared to their appropriate references. The remainder of the diagram is fairly self explanatory with the exception of the isolator. This block consists of a stable low loss magnetic component which couples the amplifier output, at a reduced level, into a low distortion cascode amplifier.

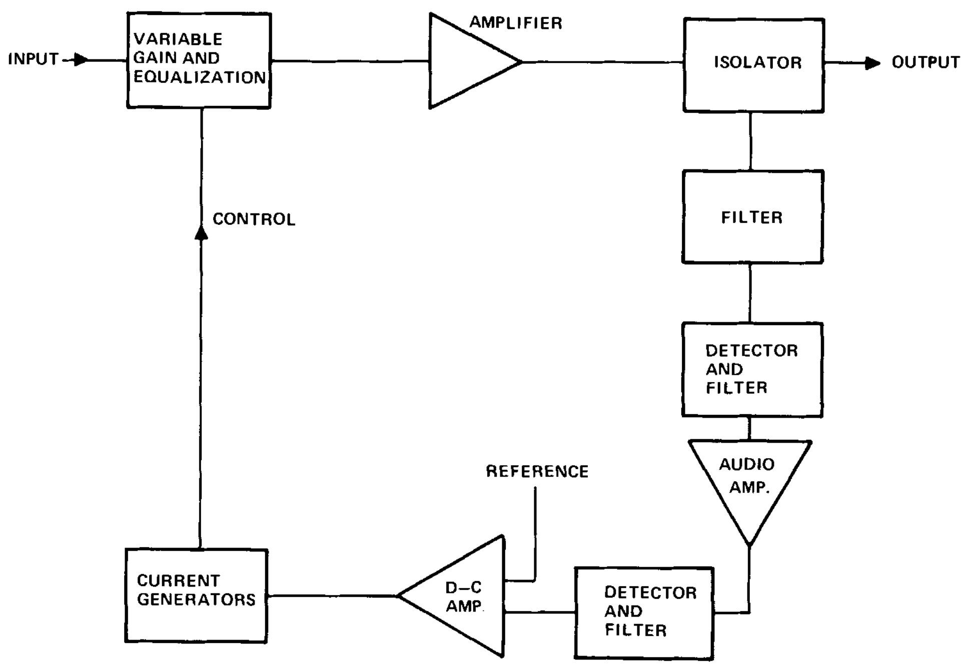
The transient response design is realized using state variable techniques. The detailed design formulation is found in Appendix IV.



AGC TRUNK LINE AMPLIFIER

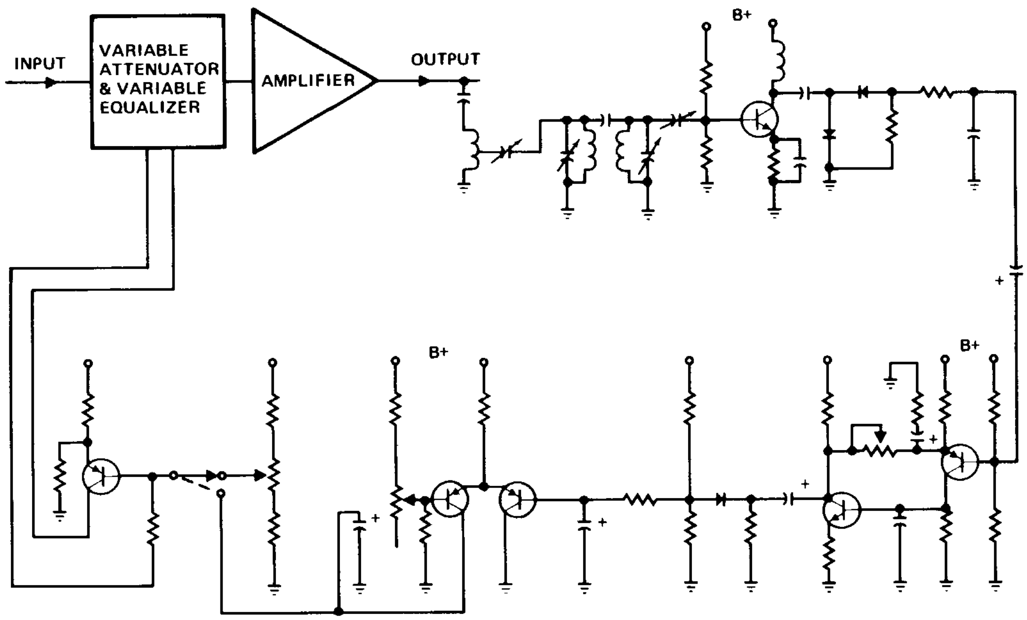
Figure 7

Since the agc line extender cascadability requirements aren't as stringent as those of the trunk line its design is compact while using only one modulated pilot carrier. This block diagram is in Figure 8; and Figure 9 shows the associated feedback circuitry.



AGC LINE EXTENDER

Figure 8



FEEDBACK CIRCUITRY FOR AGC LINE EXTENDER

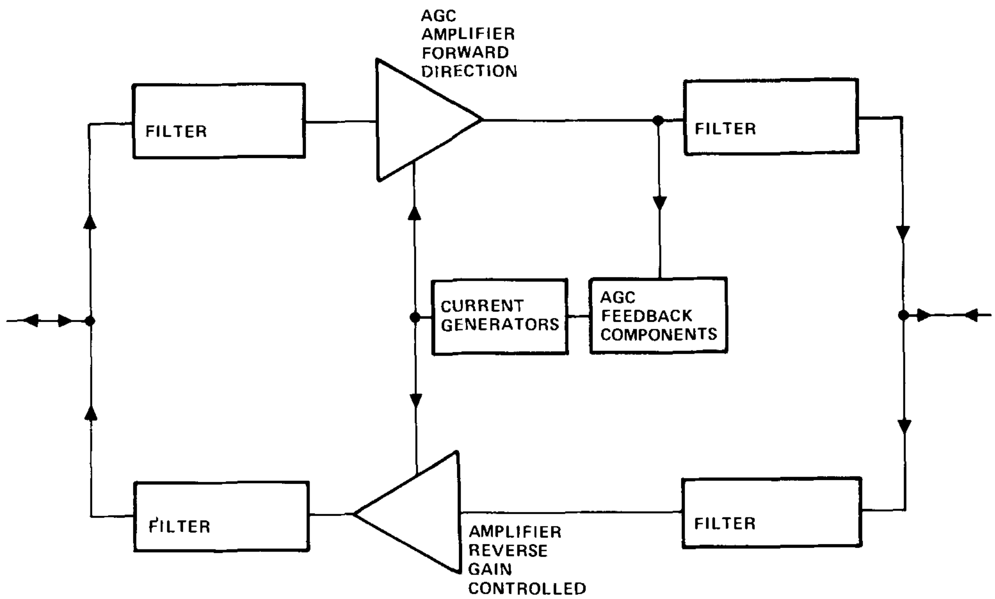
Figure 9

VII. NEW DESIGNS

A reverse gain control amplifier and weighted agc amplifier are new designs for level control applications.

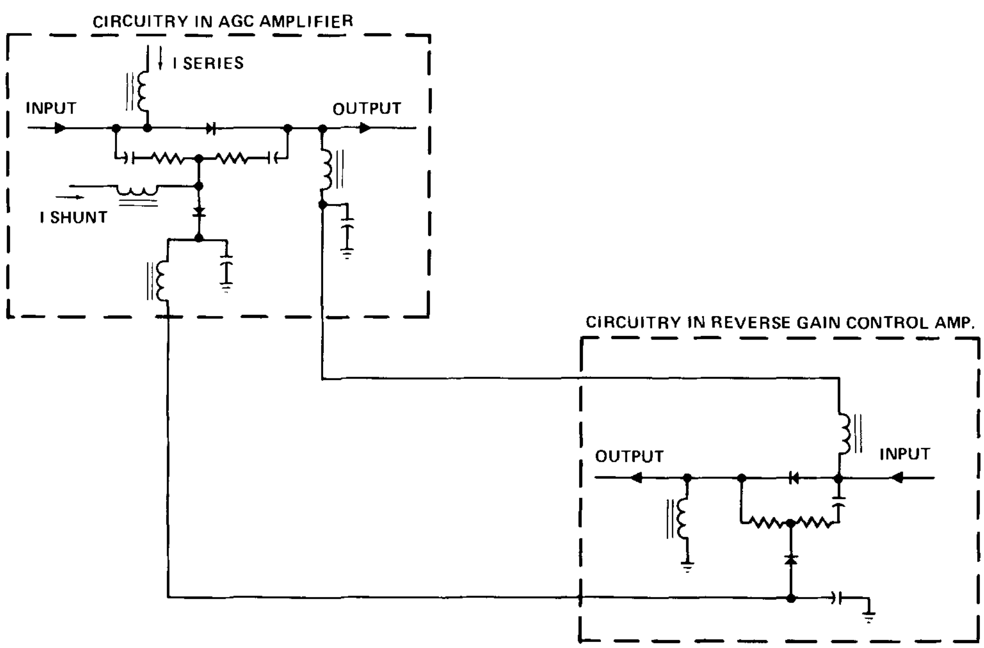
The reverse gain control amplifier is used exclusively for two-way system applications while the weighted agc amplifier can be used in either one-way or two-way systems. Figure 10 is a block diagram of the reverse gain control amplifier showing how it is used with a agc trunk line amplifier. Here the current generators used in the forward direction are shared by the reverse gain control amplifier. Additional electronics for agc feedback components are not necessary. A circuit example of this technique is found in Figure 11.

The weighted agc amplifier is similar in design to the agc trunk line amplifier of Figure 7. It has two distinct advantages overall. The primary difference in electronic components is the addition of a summing amplifier (See Figure 12). Functionally the two pilot carrier amplitudes are used to determine the setting of the variable attenuator. This complements the action of the variable equalizer and results in an extended agc range. Furthermore, redundancy is built into this design. For example, if one pilot carrier is interrupted some level control is maintained. In comparison with conventional designs, a pilot loss will result in no control and either the amplifier gain or equalization will go to its maximum value.



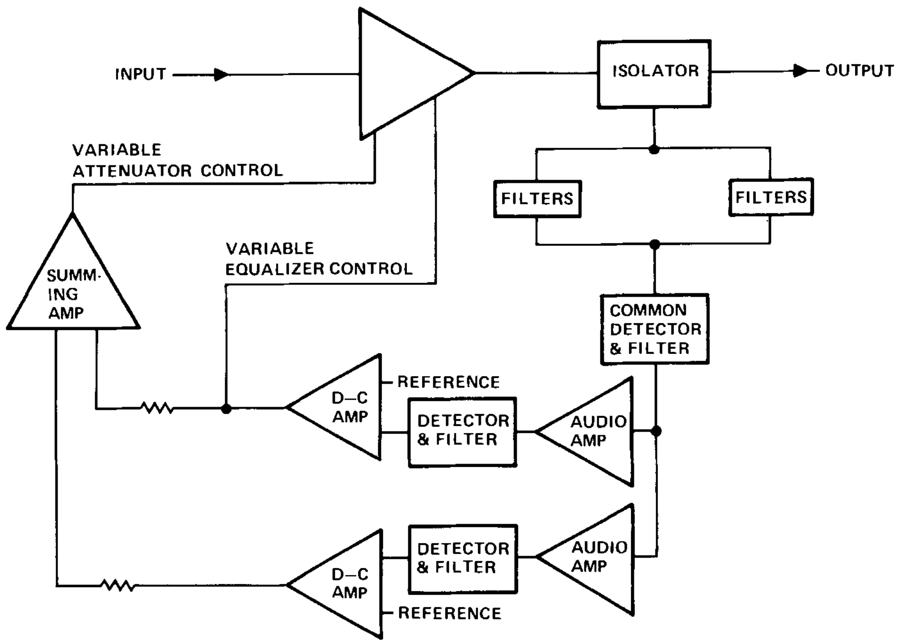
TWO-WAY COMMUNICATIONS WITH REVERSE GAIN CONTROL

Figure 10



CIRCUITRY FOR REVERSE GAIN CONTROL

Figure 11



WEIGHTED AGC AMPLIFIER

Figure 12

VIII. SUMMARY AND CONCLUSION

System requirements as they relate to level control have been emphasized. Additionally, level control philosophies were presented for both one-way and two-way systems. Finally, hardware descriptions were included for completeness.

The need for design techniques to minimize the total system cost while maintaining adequate performance was inferrred in the Two-Way System Level Control Hierarcy section. Procedures of this nature do not exist for the less complex problem of one-way systems. Such procedures could determine parameters such as head end location(s) as well as cable routes and equipment requirements.

APPENDIX I

Definition of Terms

For convenience, equations (2) and (3) are repeated [3]

$$10 \log(S/N)_m = S - G - F_{db} - 10 \log KTB - 10 \log m \qquad (2)$$

$$XM_m = XM_R + 2(S_R - S) - 10 \log(N_c - 1) - 20 \log m \qquad (3)$$

The terms are defined as follows:

$10 \log(S/N)_m$ = the signal to noise ratio at the output of the mth amplifier in decibels (db)

S = amplifier operating output in dbmv

G = amplifier gain in db

F_{db} = amplifier noise figure in db

KTB = available input thermal-noise power

m = the number of amplifiers in cascade

XM = 20 log(% normal modulation) / (% imposed modulation)

XM_m = cross modulation at the output of the mth amplifier

XM_R = cross modulation at the output of a single amplifier measured for two carriers at the S_R reference output magnitude

S_R = amplifier reference output magnitude in dbmv

N_c = number of channels in the system

APPENDIX II

Distortion Analysis

A Taylor series expansion is used to describe the nonlinearities in an amplifier output signal [6], [7]

$$e_o = \sum_{j=1}^{\infty} a_j e_{in}^j$$

The resulting second order and third order product terms are well known for $e_{in} = A\cos\alpha t + B\cos\beta t + C\cos\gamma t$.

For e_{in} consisting of n sinusoids at frequencies $f_1, f_2, \ldots, f_n$ a matrix $A_1 = (a_{ij})$ is defined where $a_{ij} = f_i + f_j$. References for matrix techniques are [8], [9], and [10]. Similarly a matrix $A_2 = (b_{ij})$ is defined where $b_{ij} = f_i - f_j$. The diagonal elements of A_1 are all the second harmonics generated and the elements from either above or below the diagonal are the sum frequencies of each pair of input frequencies. For A_2 the trace is zero and the off diagonal terms are the difference frequencies.

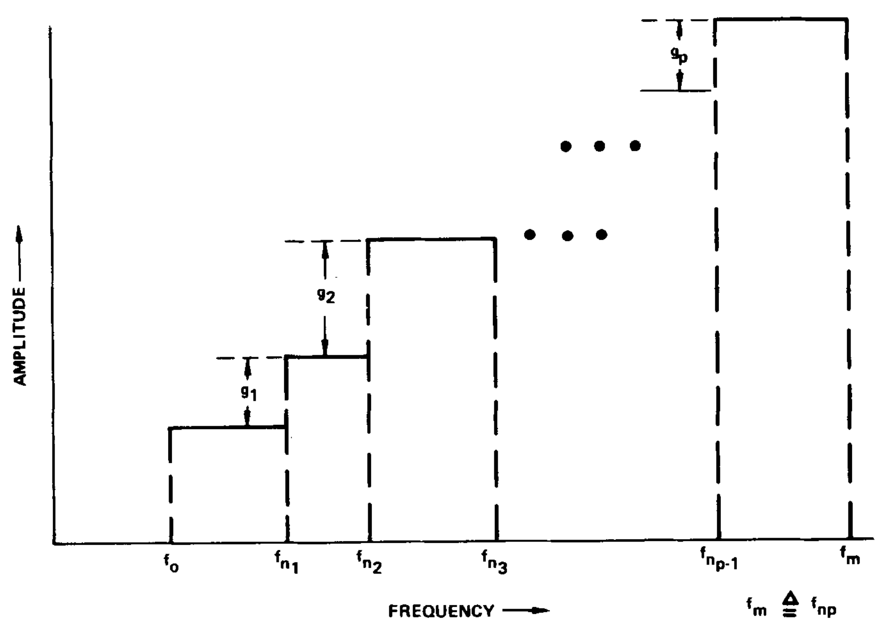
The third order product terms are three types; $3f_i$, $2f_i \pm f_j$ where $i \neq j$, and $f_i \pm f_j \pm f_k$ where $i \neq j$, $i \neq k$, and $j \neq k$. Appropriate matrices can be defined to analyze the resulting distortion terms.

Computer analysis programs have been written for general system designs using the above techniques.

APPENDIX III

Channel Selection

For a split band system corresponding to p = 3 in Figure 13 the carrier frequencies selected are within the frequency intervals (f_o, f_{n1}) and (f_{n2}, f_m). Here f_i, $f_k \in (f_o, f_{n1})$ and f_j, $f_l \in (f_{n2}, f_m)$. The index set



GENERAL MULTICHANNEL SYSTEM

Figure 13

I_1 contains i and k. Similarly $j,l \in I_2$. The number of carriers selected is n where n is the number of members in the sets I_1 and I_2.

The intermodulation sum frequencies must satify the following requirements to remain outside of the intervals (f_o, f_{n1}) and (f_{n2}, f_m).

$$\left.\begin{array}{l} f_{n1} < f_i + f_k < f_{n2} \\ \text{or} \\ f_i + f_k > f_m \end{array}\right] \text{— for low band carriers}$$

$$f_j + f_l > f_m \qquad \text{for high band carriers}$$

$$f_i + f_j > f_m$$

For the intermodulation difference frequencies the requirements are:

$$|f_i - f_k| < f_o$$

$$\left.\begin{array}{l} f_{n1} < |f_j - f_l| < f_{n2} \\ \text{or} \\ |f_j - f_l| < f_o \end{array}\right] \text{— for high band carriers}$$

$$f_{n1} < f_j - f_i < f_{n2}$$

or

$$f_j - f_i < f_o$$

The requirements for the $2f_q \pm f_r$ terms cannot be satisfied in general. Here $q,r \in I_1 \cup I_2$. The $2f_i - f_k$ term and $2f_j - f_l$ terms will fall at carrier frequencies for uniform video carrier frequency placement. The $2f_i - f_j$ and the $2f_j - f_i$ terms can be selected to reduce build up of distortion products at a particular frequency.

The desired requirements for the other terms are:

$$f_{n1} < 2f_i + f_k < f_{n2} \text{ or } 2f_i + f_k > f_m$$

$$2f_i + f_j > f_m$$

$$2f_j + f_i > f_m$$

$$2f_j + f_l > f_m$$

Requirements can be established for the $f_q \pm f_r \pm f_s$ terms and the $3f_q$ terms, where $q,r,s \in I_1 \cup I_2$.

It is not possible to satisfy all of the requirements simultaneously for a particular system. What can be done is to weigh most heavily those distortion terms which are characteristic of a particular system. For example, if the a_2 coefficient dominates in $e_o = \sum_{j=1}^{\infty} a_j e_{in}^j$ then the second order product distortion requirements are the major consideration.

APPENDIX IV

State Variable Design

The agc trunk line amplifier of Figure 7 is described by two differential equations of the form

$$\ddot{y} + b_1\dot{y} + b_o y = d_1 c_1 + d_2 x$$

$$\ddot{x} + a_1\dot{x} + a_o x = e_1 c_2 + b_2 y$$

where

y is the high frequency pilot amplifier output amplitude

x is the low frequency pilot amplifier output amplitude

c_1 is the reference signal for the variable attenuator control loop or gain loop

c_2 is the reference signal for the variable equalizer control loop or equalizer loop

Now define the following state variables:

$$x_1 = x$$

$$x_2 = \dot{x}$$

$$x_3 = y$$

$$x_4 = \dot{y}$$

State variable techniques can be found in [11], [12].

The system is now formulated in a state variable description and is compactly described by

$$\dot{x} = Ax + Bu$$

where

$$x = \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{bmatrix}$$

$$A = \begin{bmatrix} 0 & 1 & 0 & 0 \\ -a_o & -a_1 & b_2 & 0 \\ 0 & 0 & 0 & 1 \\ d_2 & 0 & -b_o & -b_1 \end{bmatrix}$$

$$B = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & e_1 \\ 0 & 0 & 0 & 0 \\ 0 & d_1 & 0 & 0 \end{bmatrix}$$

$$u = \begin{bmatrix} 0 \\ c_1 \\ 0 \\ c_2 \end{bmatrix}$$

The two poles in the equalizer loop are termed w_1 and w_2 expressed in radians per second and are determined from $a_1 = w_1 + w_2$. The loop gain here is determined from $a_o = w_1 w_2 + K_1$ where $K_o = K_1/w_1 w_2$ is the open loop gain of the equalizer loop.

For the gain loop, $b_1 = w_3 + w_4$ and $b_o = w_3 w_4 + K_3$ where $K_2 = K_3/w_3 w_4$ is the open loop gain of the gain loop. Here the two poles are w_3 and w_4.

The desired transient response design can be computed by solving for x with appropriate values for the elements in the matrices A and B.

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LOCATING AND ELIMINATING SOURCES OF POWER LINE INTERFERENCE

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With more and more cable systems being built in industrial, urban areas, and with distant signal importation restrictions reducing the incidence of cochannel interference, electrical noise is becoming one of the most important interference problems at cable system headends.

There are many sources of electrical noise interference. They include automobile ignition, welding equipment, electric fences, neon lights; in fact, any electrical device can cause trouble if it is not operating properly. Even the sun and milky way can, under certain circumstances, degrade a system's signal to noise ratio by 5 to 7 dB.(1,2)

The first step in an electrical noise investigation is to attempt to identify the source from the characteristics of the noise it generates. In fact, a great deal can be told about the source of noise interference by observing its appearance on a television screen, its pulse train on an oscilloscope, duty cycle, time of day of occurrence and any correlation to weather or other local events. For instance, noise that is limited to hours of darkness may be a lighting fixture, noise that increases with the humidity may be high voltage corona, noise that is only present when the local bar is open may be one of their signs, etc. Reference 3 contains an excellent discussion of the pulse characteristics for some representative examples of man-made noise.

In this paper, we will limit our discussion to the electrical noise caused by power lines. The most common signature of power-related electrical noise is the fact that it moves slowly through the television picture in bands. Two bands being characteristic of single phase power and six bands indicating three phase. It is at this point that most system technicians make their first mistake -- they jump to the conclusion that the cause is "the substation northeast of town", "the factory 15 miles south of town" or some other well-known power landmark. The technician makes his second mistake when he drives to this location, hears noise on his car radio and concludes that this must be the source. These two common errors result in wasted time for the utility company engineers as they try to eliminate all noise at that facility and frustration for the technician when his efforts do not remove his interference problem.

First, a search for a problem, any problem, should never be begun by deciding what the problem is and then setting out to prove it. A great deal of time can be wasted down the wrong path if an open mind is not maintained at the very beginning. Second, the car radio technicque wastes time since 1 MHz noise usually shows very little correlation to 100 MHz noise. Also, almost any power device, including probably a third of the poles in town, will radiate measurable noise within about 20-50 feet.

This brings us to Case History Number 1 and a much more efficient, systematic method of locating the source. In this Georgia system, the noise occurred only in channel 6, occurred in two bands moving through the picture and changed radically with rainfall. From this it could be concluded that the source was an outdoor, single phase power device.

The direction of the noise source from the headend was easily determined by using the system's search antenna. In this case the noise was strong enough to use a field strength meter for an indicator. In cases of weaker noise, or noise with a very short duty cycle, the noise peaks can be observed on an oscilloscope. Figure 1 shows the bearing taken from the headend (Triangle 1) and from three other locations, where a channel 6 dipole, field strength meter and compass were used to plot the directions. It is important to take readings from several widely separated locations to avoid being confused by local noise radiators.

Figure 2 shows two additional readings taken within the heavy black rectangle in Figure 1. These bearings pointed to the pole shown in Figure 3, approximately one and a half miles from the headend. The noise at the base of this pole measured +15 dBmV on the channel 6 dipole and varied considerably when the pole was shaken. The wiring configuration (a rather nice vertical antenna) on the pole explains why the noise peaked at one particular frequency.

Figure 4 shows the problem: a lighting arrester with a cracked cap. The power company said that these are notorious noise sources if their seal is broken and any rain water enters. When the arrester was disconnected, the noise ceased.

Case History Number 2 involves a high voltage noise source in Wyoming. Figure 5 shows the same direction finding technique used in Case 1, except that a truck mounted 10 element Yagi was used for the additional bearings because the noise was not strong enough for a dipole antenna. The problem in this case turned out to be dirty insulators on a high tension tower about one mile from the headend.

In this Case, we were lucky: high voltage noise can be very difficult to track down for several reasons. First, it is unavoidable. Good engineering practice allows 2 kw of corona per mile in good weather, with this going to as high as 200 kw per mile in heavy rain. Second, the periodicity of support structures causes standing waves, which, in turn, cause frequency spectra peaks and amplitude peaks at locations removed from the actual source. Third, a high tension line can act as a single wire above ground transmission line (with a characteristic impedance for a typical 500 kV line of about 425 Ohms) and propagate the noise several miles.

The power company themselves can be of tremendous help in locating the problem if approached with courtesy and diplomacy. They are anxious to do this since the generation of electrical noise interference is a public relations problem and, in fact, represents a loss of energy - their product. Their legal responsibility is quite clearly spelled out in Paragraph 15.31 of the FCC Rules and Regulations:

> An incidental radiation device shall be operated so that the radio frequency energy that is radiated does not cause harmful interference. In the event that harmful interference is caused, the operator of the device shall promptly take steps to eliminate the harmful interference.

If the source appears to be a high voltage line, their cooperation in locating the specific component causing the problem is essential. They can help to overcome some of the problems in locating high voltage line noise sources listed earlier by their experience and the use of highly sophisticated test equipment beyond the reach of the average cable system. They usually have access to such devices as sonic corona detectors and optical pyrometers. One optical pyrometer, used to detect overheating components, such as splices with internal arcing and high resistance connections, is so sensitive that it can detect a $3\frac{1}{2}^{\circ}$F rise in temperature in a 2" diameter object 20' away.

If the source of power line interference cannot be corrected, antenna methods similar to those used to reject cochannel must be used. Log periodic antennas and tapered amplitude distribution (diamond) arrays effectively reduce noise pick up from all but the desired signal direction. If the direction finding investigation accurately determined the direction of the noise from the headend, standard Yagi phased array techniques may be employed to create a null in the direction of the noise.

To summarize, use direction finding techniques to locate the source, make a visual inspection with binoculars for loose, dirty or cracked components, and, if necessary, return after dark to look for corona or arcing. If a high tension line is suspected as the source, get help from the experts - the power company themselves.

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CAPTIONS

Figure 1: Four triangulation locations. Location 1 is the headend site where the search antenna was used.

Figure 2: Enlarged view of the black rectangle shown in Figure 1. Large triangles are crossover points from Figure 1.

Figure 3: Radiating pole.

Figure 4: Close-up of lightning arrester. Arrow points to location on cap where a piece has been broken away.

Figure 5: Second noise hunt triangulations. Location number 1 is the headend search antenna.

MAINTENANCE OF LARGE CATV SYSTEMS

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Vice President-Engineering
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A Division of Cox Cable Communications

The San Diego CATV System began operation in 1962. Today it is the largest system in the United States with more than 49,000 subscribers and 660 miles of distribution plant. The system employs 64 technicians and installers and has a fleet of 51 trucks and vans. Three head-ends are needed to cover the area of the system. A Micro-wave link connects the system to the origination studio. Two-way radio is used extensively for communications but also controls non-duplication switching of the three head-ends from a central point.

In 1970 the City of San Diego passed an Ordinance governing the performance of CATV systems within the city boundary. Some of the technical specifications of this Ordinance are shown in Figure 1. While these specifications seem very reasonable, it must be realized that these are minimums for any point in the system including the subscriber's termination. This fact, coupled with long amplifier cascades, make exceeding these specification minimums more difficult. During the nine years it has been in operation, Transvideo has developed maintenance methods and procedures which make it possible to accomplish this. The main areas of maintenance fall into seven categories which I will discuss briefly.

HEAD-END MAINTENANCE

The three head-ends are routinely maintained by one highly trained technician. This man and his equipment are completely separate from the rest of the system so that he can give his undivided attention to this important function. Each week the head-ends are checked for signal levels, AGC and AFC action and the quality of video on each channel. Spurious frequency generation is investigated using a spectrum analyzer. A frequency counter is utilized to check the output carrier frequency of processors and modulators. Each month input levels to the antenna system and processors are checked for quality and level. Signal to noise and signal to hum are read and recorded. Every six months processors and other equipment are checked for alignment and response. Two forms have been developed for head-end maintenance. Figure 2 is used to record data at the various inspection times. Figure 3 is used in conjunction with the antenna system at time of installation and serves as a record of equipment configuration and signal condition on each channel.

TRUNK MAINTENANCE

The San Diego system is divided into service areas and technicians are assigned to each area. They normally work only in this part of the system and become very familiar with it. The condition of the trunk system is mainly determined by a series of test-monitor points strategically located in each of the three systems (Figure 4). Each maintenance day begins with a check of channel levels and quality at each of these points by the assigned technician. The result of these checks is relayed to the chief technician by radio who can then take proper remedial action. Most of the maintenance and trouble calls are dispatched by radio which eliminates delay. After reporting the monitor points the technician continues with routine balancing and system check out. Figure 5 shows the form used by the technician to report defective equipment in his area.

The test monitor points were established by inserting directional couplers in an output line of bridger amplifiers. This provides the highest signal level on the system so that the noise figure of the signal level meter will not be a factor. The signal to noise ratio of all channels at each monitor point is logged every three months using methods covered in NCTA Standards 005-C. This is done during non-broadcasting hours. Relative signal to noise readings can be taken during normal service hours by reading the noise above channel six or below channel seven. Care must be taken, however, that no FM or Commercial radio stations are present. Figure 6 is a compilation of some of the results obtained using these methods.

Monitor points are also used to determine cross modulation levels and system stability. Twenty-four hour recordings are made of one lowband and one highband channel at three months intervals. Figure 7 is a condensed recording showing proper action of the system. Figure 8 shows an abnormal system condition with improper AGC action or thermal control. Finally, the monitor points are used in conjunction with a spectrum analyzer to check for spurious products. (Figure 9) This form is used to record the above monitor point data. (Figure 10 and 11). These forms are used as records for each amplifier in the system.

SYSTEM RADIATION MAINTENANCE

In each service area routine radiation checks are made following the methods outlined in FCC Rules and Regulations Part 15 sub-part D.

Not only are non-subscribers protected, but system integrity to the high level signals of local broadcasters is maintained.

DIRECT PICK-UP PROCEDURES

Due to the number of Los Angeles signals carried by the system, it is necessary to carry the local channels 6, 8, 10, 12 on the system on frequency.

Much testing and evaluation is necessary to minimize the direct signal present at most subscribers' sets. Hilly terrain and the fact that two transmitters are located in Mexico prevent a uniform approach to the problem. Switches, better shielded drop cable, balanced transformers, high subscriber signal level, grounding and other methods are used to combat this problem. Naturally, none of these methods will work if the direct signal penetrates the distribution system. This makes the system integrity check used in radiation work doubly important. The form shown here in Figure 12 is used for both radiation and direct pick-up work.

DISTRIBUTION AND SUBSCRIBER MAINTENANCE

The mobility of the area trunk maintenance men is duplicated by the service technicians who cover the distribution plant and subscriber maintenance. These men are also assigned to specific areas and receive most of their calls by radio. The subscriber call is taken by a dispatcher who logs it in his Daily Work Report (Figure 13) and radios the call to the service man. At the home, all the channel levels are read and recorded as well as an analysis of the problem. A separate form is used (Figure 14) for each service call and turned into the dispatcher at the end of the day. Considerable importance is attached to subscriber level readings as these random samples of the system often serve as a good indication of conditions of the distribution plant. Any cases of direct pick up are also noted and passed on to a special group handling this work. A similar procedure is used for new installations.

COLLECTION AND ANALYSIS OF SYSTEM DATA

System data is derived from two sources--trunk maintenance and subscriber maintenance. Trunk data is derived from the form shown previously and is put into program form for computer analysis by the system chief technician. Subscriber data is derived by the dispatcher. When he receives the subscriber trouble call he verifies it against his work sheet and then fills out a Customer Service Call form (Figure 15). Data for the computer is taken from this card. Each month this information is fed to the computer which is programmed to analyze it by types of trunk and subscriber trouble, solutions to the problem, subscriber identity including phone number, technician identity etc. The computer tabulation allows us to determine the efficiency of maintenance being performed in an area and indirectly indicates plant conditions. The constant flow of data from these service areas is used to determine the overall system status and dictates what action is needed. Using these methods, we have been able to detect developing problems before they became the cause of widespread outage.

EQUIPMENT REPAIR

Complete records are kept on equipment from the time it is initially installed. Bench technicians work independently from the rest of system maintenance. In such a large system a constant program of equipment repair is necessary and vital.

SUMMARY

In conclusion let me point out that the degree and complexity of system maintenance obviously increases with system size. When large numbers of subscribers are involved, it becomes mandatory to keep ahead of developing system problems. If I were to pick the most important maintenance feature of the San Diego system it would have to be the establishment and full utilization of monitor points. An example of the effectiveness of this program is the decrease in subscriber trouble calls this past year. At the beginning of 1970 we were averaging a ratio if trouble calls/month to subscribers of over 3.5%. In May of this year the trouble call ratio was 1.2% or about 30% of the 1970 figure.

San Diego Ordinance Specifications

Minimum Subscriber Level	0 dBmV at 75 Ohms
Minimum S/N at Any Point	34 dB at 4 MH_z
Cross Modulation at Any Point	−46 dB at 32° F
Spurious Products	−46 dB
Hum Modulation	3 %
Multiburst Response	F.C.C. Sec. 73.687 (a)
Radiation	F.C.C. Part 15 Subpart D

Figure — 1

Cox Cable Communications, Inc.

ANTENNA SITE OPERATIONAL REPORT FORM

Location: **Date:**

Com. Ch.	Date Tube	Date Aligned	S/N Ratio	AGC Oper.	St. By Carrier	Operating Levels	Trap Al. Date	Beats, Problems or Component Change-Outs
2						in out		
3						in out		
4						in out		
5						in out		
6						in out		
7						in out		
8						in out		
9						in out		
10						in out		
11						in out		
12						in out		
13						in out		

Pre-Amp Ch.				Levels	UHF Conv. Ch.	Date Tube	Date Aligned	S/N Ratio	Output Level	Mixing Network
				in out						tube
				in out						align
				in out						sweep
				in out						
				in out	Hi CCG	Out	Tube			
				in out	Lo	Out	Tube			

Remarks:

Signed: **Title:** **Approval:**

Figure – 2

Cox Cable Communications, Inc.

ANTENNA INSTALLATION CHECK LIST

System ______________________

____

City ______________ State ________

Date ____________ Temp. ____

Channel ______________

Preamp Used Yes ☐ No ☐

Type of Downlead ________ Ft. ___

SIGNAL MEASURMENT	VIDEO	AUDIO
CALCULATED SIGNAL ________		
MEASURED CARRIER WITHOUT PREAMP ________		
MEASURED CARRIER WITH PREAMP ________		

TYPE OF CONVERTER USED ______________

REMARKS ______________

ANTENNA ARRAY

CONFIGURATION ______________

DIRECTION ______________

TYPE ____________ GAIN ______ DB

HEIGHT ________ FT.

RETURN LOSS ____________ DB

HORIZONTAL SPACING ________ IN.

VERTICAL SPACING ________ IN.

MIXING DEVICES

SPLITTER ______ PHASING LINES ______

PHASED YES ☐ NO ☐

PHASED FOR OFFENDING STATIONS

(1) ________ (2) ________

HOW WEATHERPROOFED ______________

TRAPS USED ______________

____

ENGINEER

Figure-3

CCC FORM 101-71


EL CAJON HEAD END
CHULA VISTA HEAD END
POINT LOMA HEAD END
1B
1C
1D
1E
1F
1G
1H
1I
1J
2A
2B
2C
2D
2E
3A
3B
3C
3D
MURRY CANYON ROAD
INTERSTATE 5
INTERSTATE 8
MONTGOMERY
LINDBERGH FIELD
GILLESPIE FIELD
PACIFIC OCEAN
NORTH ISLAND
POINT LOMA
MISSION BAY

Figure 4

TRUNK PROBLEM REPORT

Location ______________________ Date ____________

	In	Out		In	Out
Ch. 2			7		
3			8		
4			9		
AGC			10		
5			11		
6			12		
			13		

Pad ____________

Equal. ____________

Problem ____________

Type of Equipment ______________________

Remarks:

Turned in By:

Completed By:

Date ____________

Figure - 5

Location	Amplifier Cascade	Theoretical S/N	Actual S/N	NCTA 005-C S/N	Band Edge S/N					
					Ch. 2	73.5 MHz	Ch. 6	Ch. 7	Ch. 13	
EL CAJON										
1B	2	53	53.7	49.2	34	48	53	55	51	
1C	14	44.6	46.2	43.1	33	44	44	49	47	
1D	30	41.2	42.5	41.6	36	41	40	45	44	
1E	31	41.1	41.8	40.7	32	42	46	44	46	
1F	27	41.7	42.9	39.6	31	39	43	44	35	
1G	13	44.9	42.6	43.3	34	37	40	41	32	
1H	24	42.2	42	42	36	40	41	44	36	
1I	35	40.6	38	41.4	33	38	38	39	31	
1J	28	41.5	40.7	40.4	36	41	41	44	39	
CHULA VISTA										
2A	26	41.9	41.4							
2B	27	41.7	39.7							
2C	33	40.9	38.3							
2D	18	43.5	45.6							
2E	6	48.2	49.2							
POINT LOMA										
3A	2	53	49.6							
3B	8	47	43.3							
3C	9	46.5	42.5							
3D	16	44	39.3							

Figure – 6

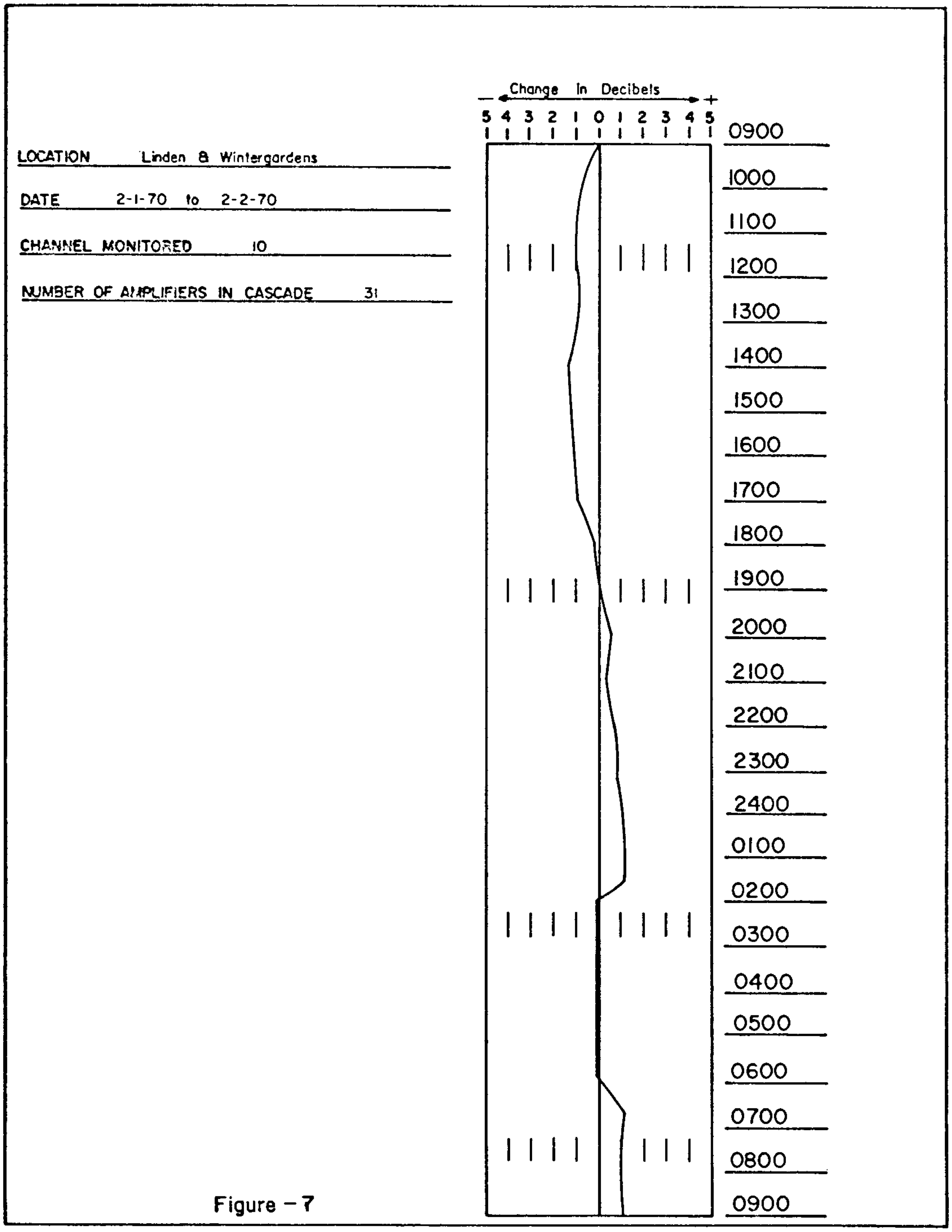
LOCATION Linden & Wintergardens
DATE 2-1-70 to 2-2-70
CHANNEL MONITORED 10
NUMBER OF AMPLIFIERS IN CASCADE 31
Change In Decibels
− 5 4 3 2 1 0 1 2 3 4 5 +
0900
1000
1100
1200
1300
1400
1500
1600
1700
1800
1900
2000
2100
2200
2300
2400
0100
0200
0300
0400
0500
0600
0700
0800
0900

Figure - 7

LOCATION 8712 Magnolia

DATE 2-10-70 to 2-11-70

CHANNEL MONITORED 12

NUMBER OF AMPLIFIERS IN CASCADE 24

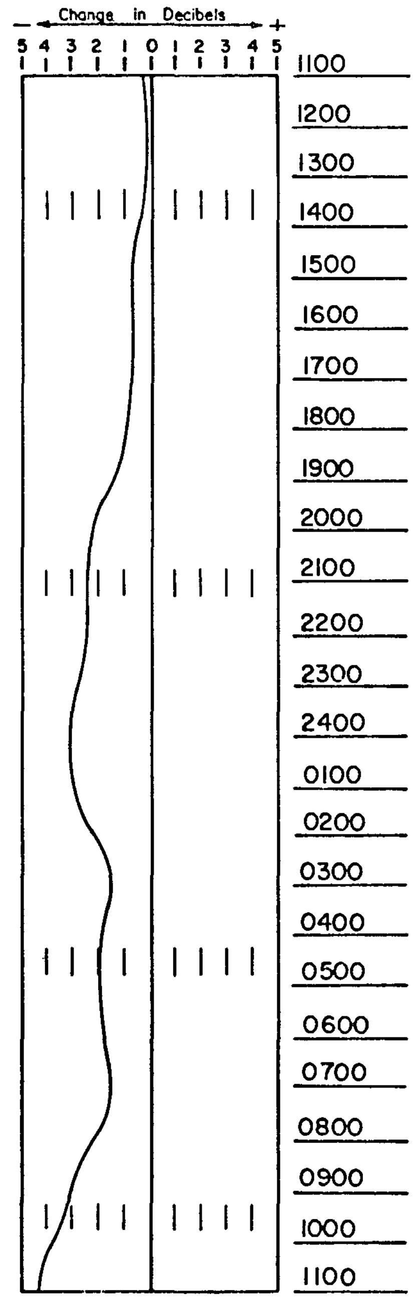

Figure -8

Cox Cable Communications, Inc.

Amplifier Extremity & Temperature Variation Test

System ______________________ Date __________ Temp. ________

LOCATION ______________________ CASCADE ________ AMPLIFIER NO. ________

CHANNEL	CARRIER LEVEL	NOISE LEVEL	SIGNAL-TO-NOISE RATIO	% HUM MODULATION	CROSS MODULATION	PICTURE QUALITY
2						
3						
4						
5						
6						
7						
8						
9						
10						
11						
12						
13						
P.C.						

CHANNEL	DATE ________ TEMP. ________ LEVEL (dbmV)	DATE ________ TEMP. ________ LEVEL (dbmV)	DIFFERENTIAL
2			
6			
7			
13			

____
ENGINEER

Figure - 9

CCC FORM 105-71

Cox Cable Communications, Inc.

TRUNK AMPLIFIER

System ____________________ Date __________ Temp. ____

SIGNAL READINGS (dbmv)

CHANNEL	LINE AMPLIFIER INPUT	LINE AMPLIFIER OUTPUT	BRIDGER OUTPUT
2			
3			
4			
5			
6			
7			
8			
9			
10			
11			
12			
13			
P.C.			

LOCATION ____________________

DC VOLTS ________ AC VOLTS ________

AMPLIFIER NO. ________ TYPE ________

PAD ________ EQUALIZER ________

EQUIPMENT DIAGRAM

REMARKS ____________________

____

ENGINEER

Figure -10

CCC FORM 103-71

Cox Cable Communications, Inc.

LINE EXTENDER AMPLIFIER

System ____________ Date ________ Temp. ______

Location ____________

AMPLIFIER NO. ________

PAD ______ EQUALIZER ______

AC VOLTAGE ______ DC VOLTAGE ______

SIGNAL READINGS (dbmv)

CHANNEL	INPUT TEST POINT	OUTPUT TEST POINT
2		
3		
4		
5		
6		
7		
8		
9		
10		
11		
12		
13		

EQUIPMENT DIAGRAM

REMARKS ____________

ENGINEER

CCC FORM 104-71

Figure - 11

NAME ______________________ ADDRESS ______________________

DIVISION ______________________ DATE ______________________

DIRECT PICKUP REPORT

1. System trunk and feeder cable radiation

SECTION ______________________

SLM TYPE ______________________

CALIBRATION DATE ______________________

ANTENNA USED ______________ ANTENNA GAIN ______________

CHANNEL MEASURED ______________________

DISTANCE FROM CABLE ______________________

RADIATION LEVEL IN MICROVOLTS ______________________

ACTION TAKEN: ☐ NONE ☐ SPLICE ☐ CONNECTOR ☐ CABLE BREAK

NEW RADIATION LEVEL AFTER REPAIR (IN MICROVOLTS) ______________

2. DIRECT PICKUP ON CUSTOMER SERVICE DROP

(a) Leakage found on cable ________ set ________

(b) Method used to test leakage: converter _____ shielded set _____

(c) Other methods ______________________

(d) Channels affected ______________________

ACTION TAKEN

(a) New house drop ______________ Type cable used ______________

(b) Transformer ______________ Type ______________

(c) Switch Transformer ______________ Type ______________

300 OHM side of switch connected to: ☐ Rabbit ears
☐ Built in antenna ☐ Outdoor antenna ☐ Nothing

(d) Pickup results ______________________

____

____

Signal 1st reading 2 ______ 6 ______ 8 ______ 10 ______ 13 ______

Signal adjusted to 2 ______ 6 ______ 8 ______ 10 ______ 13 ______

Figure – 12

DAILY WORK REPORT SERVICE **CABLE TV** CALLS

DATE

	Name	Address	Phone	Trouble	TBW Code	Appt. Time	Tech	Code	Time
1									
2									
3									
4									
5									
6									
7									
8									
9									
10									
11									
12									
13									
14									
15									
16									
17									
18									
19									

Figure-13

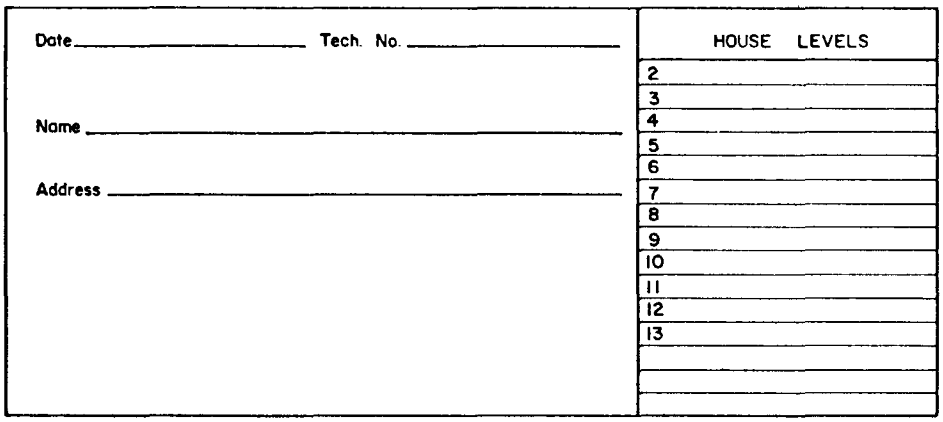

Figure – 14

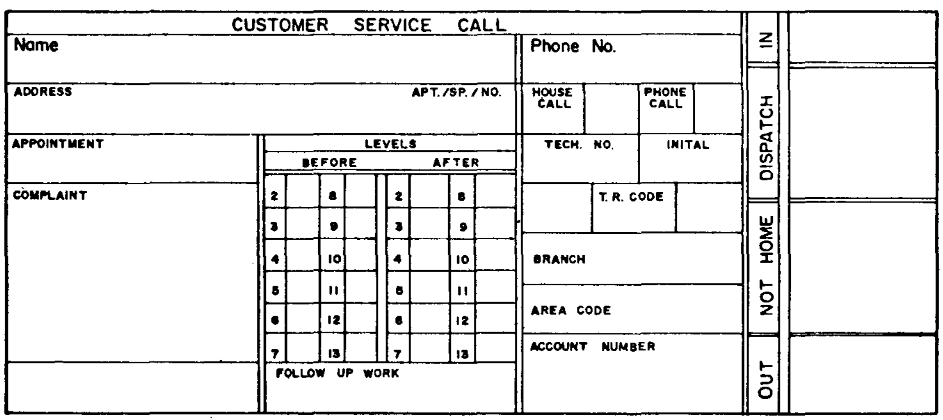

Figure – 15

MATCHING REPEATER AMPLIFIER PERFORMANCE CHARACTERISTICS TO CABLE SYSTEM LEVEL REQUIREMENTS

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PRESENTED AT THE 20th ANNUAL NCTA CONVENTION,
WASHINGTON D.C., July 7, 1971

INTRODUCTION

A cable television system provides a transmission path from a single signal processing center (traditionally referred to as head end) to multiple subscriber television receivers (Figure 1). The advent of transmission requirements from any point in the cable system back to the signal processing center results in a system with inputs from multiple locations in the cable system converging to a single output from the cable distribution system back to the head end.

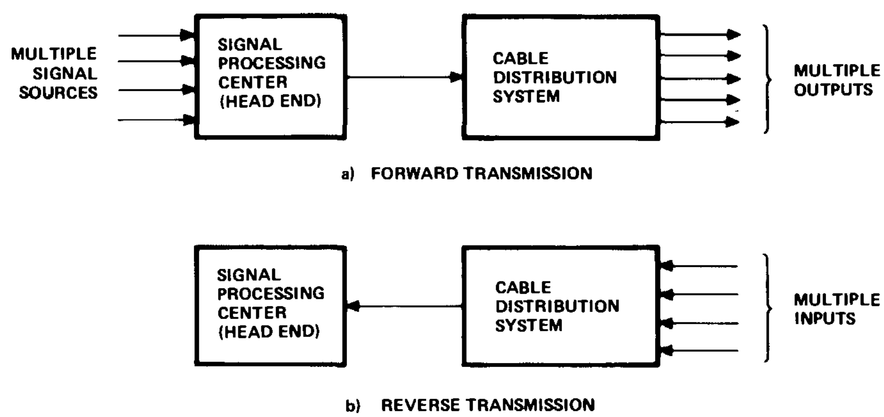


FIGURE 1

Numerous articles have been written on parameters and performance requirements of the cable distribution system. This paper will focus on the television signal level from cable distribution system input (head end output) to its output, the TV receiver input terminal.

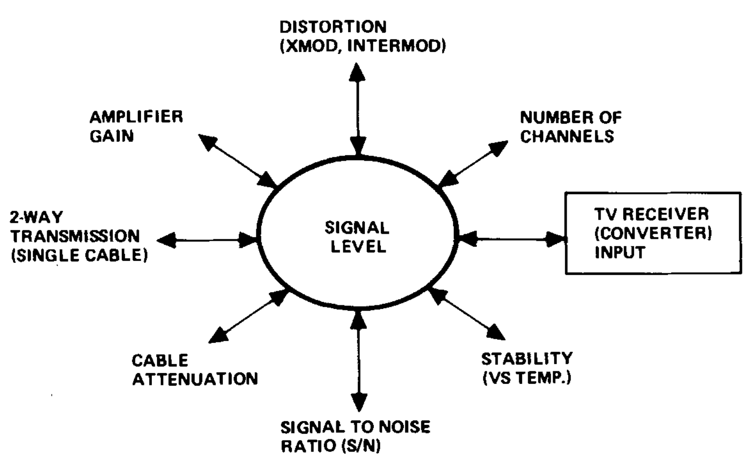
Knowledge of the signal level requirements in the cable distribution system guides the equipment designer to a hardware design which economically matches system requirements and likewise guides the system designer in selection of system components required to meet his system design objectives. Signal level, as the independent variable, determines many component (amplifier, cable, and directional taps) and system design decisions.

After considering the cable system signal level requirements, a coefficient system similar to that described by Carson (Reference 9) is used to exemplify sources of system performance limitations. The coefficient system is useful in weighting the performance of repeater amplifiers which have different operating characteristics.

Signal Level As Related to Equipment/System Parameters

A cable distribution system must deliver television signals to each subscriber's television receiver which produce acceptable pictures. Insight into cable equipment and system design results when the signal transmitted through the cable distribution system is distinguished by both level and quality. Signal level and quality specifications which are necessary to produce an acceptable picture are described in detail in such references as (4), (5), (6), and (7). The purpose of this paper is to show how certain cable amplifier performance characteristics are defined as a result of the television receiver input level requirements, and to relate the signal level to various signal quality performance factors.

A summary of parameters shown in Figure 2 may be viewed as defining cable system signal levels and/or as cable system signal levels defining cable system parameters. The value of these viewpoints will become apparent later.



CABLE SYSTEM SIGNAL LEVEL-CABLE SYSTEM PERFORMANCE PARAMETERS

FIGURE 2

For example, the television receiver or converter must be given multi-channel input signal levels within a given range, assuming acceptable signal quality, to produce an acceptable picture. Distributing a signal level above the minimum value required by the TV receiver or converter unnecessarily increases system cost. Some excellent work has been done (4), (5) in determining the range of levels which should be provided by the cable system to the TV receiver input.

Another example of the importance of selecting system signal levels relates to the system distortion and S/N parameters. The original source of the majority of noise and distortion in CATV equipment is the amplifying device, the transistor. Recent advances in microelectronic circuit technology (1), (2), (11) can allow efficient circuit designs to realize low noise and low distortion performance which are basically device limited. An example of a broadband microelectronic circuit currently used by Anaconda Electronics in production repeater station amplifiers is shown in Figure 3.

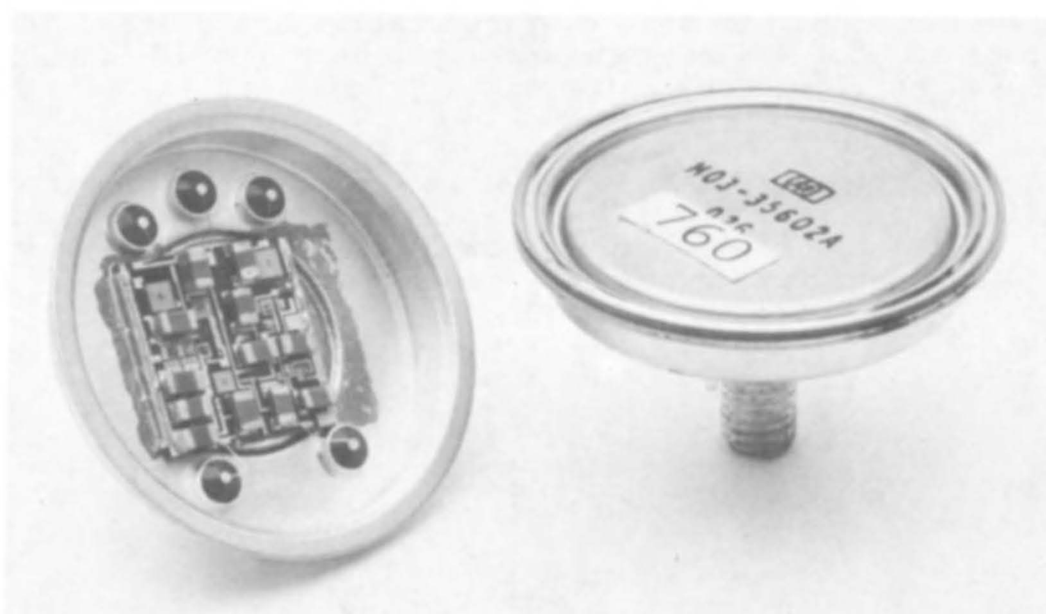


FIGURE 3

Trunk system signal levels may therefore be chosen to operate within the amplifier constraints of noise and distortion so that system design may be optimized. Alternatively, signal levels may be selected to provide added performance margin if system length is not a performance limiting factor.

Judicious apportionment of noise and distortion between trunk and feeder systems can be accomplished not only by choice of equipment but also by selection of system signal levels. This system design approach can result in a reduced system cost.

CABLE DISTRIBUTION SYSTEM

The cable distribution system provides the means of distributing head end signals to multiple subscriber locations in an area. The signal delivered to the subscriber's receiver terminals must be of such a level and quality as to produce acceptable pictures.

Definition of what signal characteristics produce acceptable pictures are covered in detail in (4), (5), (14). To design the most economical and efficient cable distribution equipment, the designer must consider in depth the requirements of the subscriber receiver equipment.

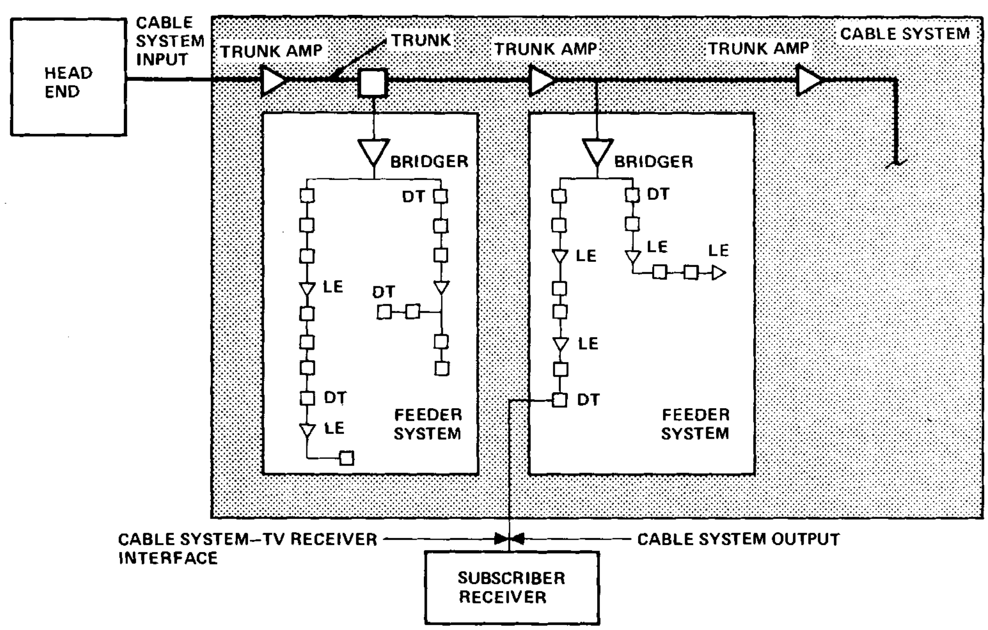


FIGURE 4

In Figure 4 is shown a representative block diagram of a cable distribution system. A trunk system is used for transmission of the signal from the head end to the areas in which the signal is to be distributed and does not directly feed signals to subscribers. A feeder system consists of equipment which is used to deliver signals from the trunk system to subscribers.

TRUNK SYSTEM

Trunk amplifiers are required at intervals along the trunk cable route to offset signal attenuation due to cable, power splitters, and intermediate bridging amplifier insertion losses. The amplifiers are spaced so that the signal level does not drop below the minimum value required to meet the system S/N objective nor exceed the maximum level consistent with the system distortion specification.

The laws of cascaded amplifier distortion and noise accummulation (6), (12), (15) provide guidelines for selecting levels on the trunk system needed to match trunk amplifier noise and distortion characteristics to the trunk system transmission objectives.

Some general principles applied to trunk system amplifiers which are helpful in meeting system noise and distortion objectives are listed below.

Automatic Gain Control (AGC) Amplifier

1. Should have the lowest noise figure and/or be short spaced to preserve the minimum S/N since its input level will tend to be the lowest in the system (with the possible exception of head end input).

2. Should be spaced at intervals such that its dynamic control range is not exceeded by that range dictated by cable attenuation variation from low to high temperature.

3. Should be spaced to maintain level variation to less than maximum value dictated by system S/N and distortion objectives in order to realize consistent year round service.

4. Should have distortion characteristic which does not change as a function of dynamic gain. The AGC amplifier should have consistent same crossmodulation, regardless of ambient temperature.

Manual Gain Control (MGC) Amplifier

1. S/N for MGC amplifier will vary as a function of temperature, but will always be greater than S/N of AGC amplifier (assuming amplifier with equal noise figure).

2. Distortion at cold temperature will be worst in the MGC trunk amplifier immediately preceding the AGC amplifier.

3. Open loop thermal compensation in the MGC will tend to minimize the amount of increased crossmodulation at low temperature.

Level Tilt in Cable Distribution System

The relationships between signal levels of each channel carried in the cable distribution system are established at the head end. The actual setting however, is determined by the design of the repeater amplifier and performance of other system components. It is essential that levels are set to match the characteristics of the repeater amplifier if the system S/N and distortion objectives are to be met in an optimum manner.

FEEDER SYSTEM

The amount of noise and distortion added to the television signals in the trunk system limit the amount of noise and distortion which may be introduced by the feeder system while meeting the system objectives. The minimum allowable signal level in the feeder system is established by the minimum level to be delivered to the subscriber receiver. Designing the feeder amplifiers (bridging and line extender) within this constraint establishes the feeder system S/N for amplifiers of a given noise figure and feeder leg cascade length.

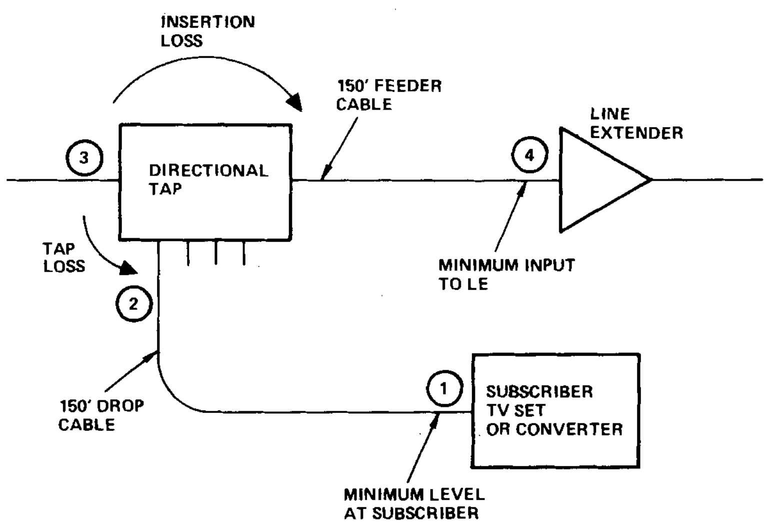
Noise and distortion accummulates along the trunk system cascade as stated previously. This fact means that noise and distortion in the trunk system near the head end is much less than at the trunk line extremities. Therefore, the feeder system fed from the trunk system consisting of a small number of trunk amplifiers can be allowed to operate in modes which tend to produce more distortion than would be allowed near the end of long trunk line cascades. Examples of these modes are: higher signal levels in feeder system, longer cascade of feeder amplifiers, lower output capability (assumed lower cost) line extenders.

Feeder amplifiers designed to serve the maximum number of subscribers require the highest possible signal level output with low enough distortion to meet system objectives. The maximum gain required for a feeder amplifier, and in particular a line extender, is therefore equal to the difference between minimum input and maximum output signal power. Examples which follow will clarify the requirements of a line extender amplifier.

MINIMUM INPUT LEVEL TO LINE EXTENDER AMPLIFIER

The minimum input to the line extender is a function of the minimum input signal delivered to the subscriber receiver terminals.

The determination of this level will be clarified by referring to Figure 5 and the following example.



MINIMUM INPUT LEVELS, SUBSCRIBER RECEIVER AND LINE EXTENDER

FIGURE 5

It has been shown (4), (5) that a minimum level of 0 dBmv or greater at location 1.0 in Figure 5 will result in picture quality which is system S/N limited and not television set limited.

To determine the minimum line extender (LE) input level, the attenuation between points 1 and 3 in Figure 5 will be calculated at the highest frequency supplied to point 1. Adding this attenuation in dB to the minimum subscriber level at point 1 establishes the input level to the directional tap. From this level is subtracted the sum of tap insertion loss plus feeder cable span loss to arrive at the minimum input level to the line extender.

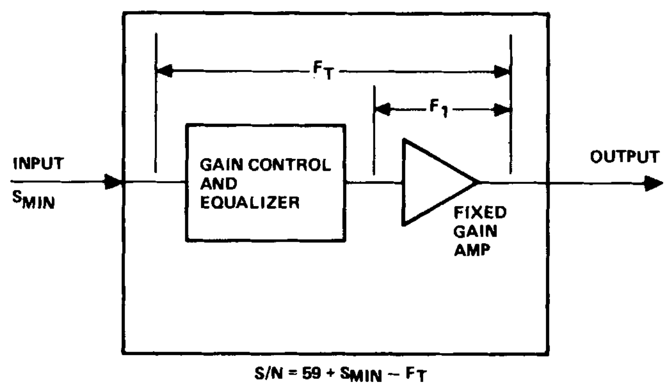
TABLE I LOSSES IN DB (Freq = 270 MHZ)

Feeder Cable Type Drop Cable Type	412 RG/59	412 *RG/59	500 RG/59	500 *RG/59
150' Drop Cable	8.4	5.4	8.4	5.4
Tap Loss (4-Output, 11 dB Tap)	11.0	11.0	11.0	11
SUM A	19.4	16.4	19.4	16.4
Tap Insertion Loss	2.5	2.5	2.5	2.5
150' Feeder Cable Loss	2.8	2.8	2.1	2.1
SUM B	5.3	5.3	4.6	4.6
Amount LE Input Level Above Subscriber Level (A-B)	14.1	11.1	14.8	11.8
Subscriber Level in DBMV	6	6	6	6
Minimum Input to LE in DBMV	20.1	17.1	20.8	17.8

*RG/59 (Type) Drop Cable Belden 8228

Feeder cable attenuation values are based on nominal catalog values at 70°F for Anaconda Sealmetic (SLM) 412 and 500. Drop cable values are extrapolated to 270 mHz from Belden Catalog No. 871.

The minimum level at point 4, the LE input, is 14.1 dB (for RG-59 drop cable and 412 feeder cable of length noted above) above the subscriber level in dBmv. For this example assume a subscriber level of 6 dBmv. Then the minimum LE input level is 20.1 dBmv.



LINE EXTENDER AMPLIFIER S/N

FIGURE 6

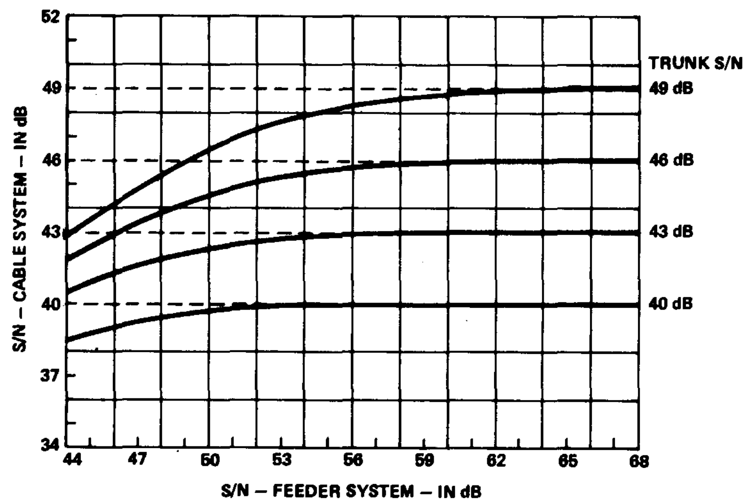
The affect of the line extender noise figure (F) on feeder system S/N (Figure 6) can now be determined by referring to the equation below:

$S/N = 59 - F + S_{min}$ dB
S/N for single line extender

$S/N = 59 - F + S_{min} + \text{Log } n$ dB
S/N for cascade of n identical line extenders

A noise figure of 20.1 dB results in a single LE S/N of 59 dB for the minimum input signal calculated previously.

The affect of feeder system S/N as a function of trunk system S/N on subscriber drop S/N is shown in Figure 7. Note that for a trunk system S/N of 43 dB, the most distant subscriber in the feeder system suffers almost no S/N degradation for a cascade of two identical line extenders, each with an S/N of 59 dB. <u>This fact is extremely significant in terms of LE amplifier circuit design, because an LE S/N of 59 dB for the minimum signal level calculated above means that the LE noise figure can be 20.1 dB.</u>



CABLE SYSTEM S/N APPORTIONMENT BETWEEN TRUNK AND FEEDER

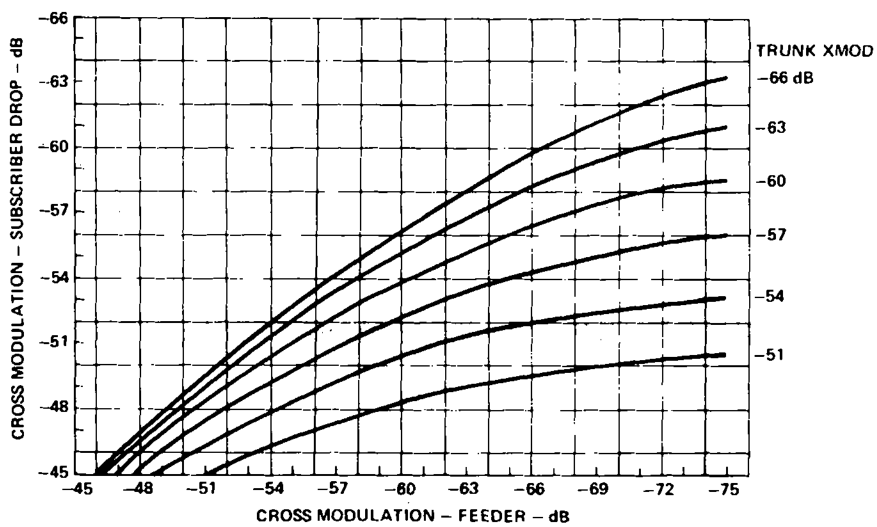
FIGURE 7

Integrated circuit broadband amplifiers can be designed and manufactured with a variety of frequency response shapes and with access to multiple gain stages. However, one of the most economical designs is a fixed gain block with flat frequency response. Permitting an LE station noise figure in the 16 to 20 dB range would allow use of a fixed gain, input-output integrated circuit amplifier of economical design. The line extender station block diagram is then accurately represented in Figure 6. <u>A high noise figure line extender requires education of the customer because a high noise figure does not necessarily mean a low S/N.</u> The individual system component or amplifier specifications must be related to system performance to realize the most economical design.

FEEDER SYSTEM DISTORTION

The feeder system is allowed to produce an amount of distortion equal to the difference between the distortion objective at the subscriber and the amount of trunk system distortion. The examples and discussion which follow mention only crossmodulation distortion, which to date has been a familiar system performance limiting parameter. However, similar and possibly additional analyses must be made to account for second and third order intermodulation products as well as triple beats in systems carrying more than 12 channels.

A relationship between crossmodulation at the subscriber drop as a function of trunk and feeder system crossmodulation is shown in Figure 8.



SUBSCRIBER XMOD APPORTIONMENT BETWEEN TRUNK AND FEEDER

FIGURE 8

The axes of this figure have been interchanged from those of Figure 5 in (5). The chart of Figure 8 is useful in determining the allowable feeder system crossmodulation distortion as a function of the crossmodulation objectives at the subscriber and the trunk system crossmodulation.

Feeder system distortion determines the maximum output level at which the line extender can be operated. From the method described previously, the minimum LE input level is defined. The LE gain required in the feeder system is now defined.

Some general considerations related to feeder system amplifier performance are summarized below.

Bridger Amplifier

1. System crossmodulation distortion performance can easily be limited by the bridger amplifier, since it operates at the highest level.

2. Cold temperature operation is the most critical for the bridger amplifier driven from an MGC amplifier or an intermediate bridger in the span preceding an AGC amplifier. The level will be the highest and therefore the crossmodulation distortion will be the worst.

3. There may be a system distortion advantage in operating the four (4) output bridger at a level <u>lower</u> than the line extender. Uniformity of feeder system levels is lost, but improved distortion performance results.

4. The bridger amplifier noise figure is relatively unimportant because its input level is relatively high, typically only 10 to 12 dB below the trunk output level.

Line Extender

1. For rigid feeder system level control, an AGC line extender (10) should be placed in each bridger amplifier leg where an MGC trunk amplifier drives the bridger, including intermediate bridger stations preceding AGC trunk amplifier stations.

2. Every other line extender in a feeder cascade should contain an AGC amplifier.

3. Level control to the subscriber home may be more critical in a converter system than a non converter system.

4. Open loop thermal compensation is advisable in each line extender.

5. Noise figure of line extender amplifier is relatively unimportant because it operates at level approximately 10 dB above trunk amplifier.

Channel Levels Across Band at Subscriber Receiver Terminals

The difference in levels between channels across the system bandwidth is becoming increasingly important for broadband (greater than 50-216 mHz) multichannel (greater than 12 channel) systems. Subscriber converters developed to date have a limited dynamic range so that the level spread across the bandwidth and the absolute level stability are critical.

The level spread from channel to channel across the band at the subscriber receiver or converter terminals is a function of the following:

1. Block tilt of channel levels (set at head end)
2. Type of feeder cable (loss per 100 feet)
3. Length of feeder cable preceding tap
4. Number of taps and splitters (and amount of flat loss in feeder line)
5. The magnitude of tap loss
   5.1 Frequency response of tap loss (flat and/or slope)
6. Type of drop cable (loss per 100 feet)
7. Length of drop cable

Amplifier Coefficient System

To further exemplify the sources of noise and crossmodulation distortion in the cable distribution system, use is now made of an amplifier coefficient system.

Carson (9) outlines a method for calculating amplifier cascade performance with amplifiers of different characteristics. Taylor, in reference (13), clarified the data obtainable from the principles outlined in (9), and his format is used in what follows. The different characteristics can be noise figure, distortion due to different operating level or distortion characteristics, number of channels (such as transportation amplifier), etc. The method basically normalizes the performance of each amplifier type in the cascade to a reference trunk amplifier.

After each component of the system which generates noise and/or distortion has been normalized to an equivalent reference amplifier (component coefficient), the component coefficients are summed to equal a total equivalent cascade. The system S/N and crossmodulation distortion are then determined by the following familiar equations:

S/N (System) = S/N (Ref Trunk) - 10 Log C_n

Where C_n is the equivalent cascade of reference trunk amplifier.

XM (System) = X_1 (Ref Trunk) + 20 Log C_x

Where C_x is the equivalent cascade of reference trunk amplifiers.

The following example is given to illustrate the principle involved:

Given: Trunk Amplifier

XM = -82 dB at level of 40 dBmv (12 channels)

X_1 = -98 dB at operating level of 32 dBmv

Line Extender

XM = -82 dB at level of 40 dBmv (12 channels)

X_3 = -78 dB at level of 42 dBmv

Line Extender Coefficient

L_1 = 10 raised to exponent $(X_3-X_1)/20$=10 amplifiers

In other words, a single line extender produces as much crossmodulation as a cascade of ten (10) trunk amplifiers operated at a level of 32 dBmv.

A separate coefficient for noise must be calculated. The resulting coefficient is the equivalent number of reference trunk amplifiers to which the line extender is equivalent in terms of S/N degradation. The method for making this calculation is contained in Appendix B.

Component Coefficients as Function of Temperature

Three sets of coefficients are required to characterize the system S/N and XM as a function of temperature. The coefficients for each component are initially calculated for nominal temperature. At the maximum temperature, a set of coefficients is required to determine the worst case S/N. At the minimum temperature, a set of coefficients is required to determine the worst case crossmodulation.

For every one (1) dB increase in level due to cable attenuation reduction at low temperatures, a 2 dB increase in crossmodulation results in the "well behaved amplifier". The level variation of each component in the system which produces crossmodulation is accounted for and a coefficient for low temperature operation assigned to each component. Total system crossmodulation distortion at low temperature is then calculated by summing each of the component coefficient. Since levels increase at low temperature, the S/N will not be reduced.

In a similar manner, a set of coefficients is required at the high temperature to determine the amount by which the S/N is reduced because of the increased cable attenuation. Crossmodulation distortion is less than at the minimum temperature due to cable attenuation change, because the increased cable attenuation results in lower levels. However, care must be taken in the amplifier design to insure that gain reduction due to AGC action does not increase the amplifier distortion.

The component coefficients at low and high temperature depend upon the amount of level control in the system. The spacing of AGC amplifiers and degree of thermal compensation are the controlling factors in wide temperature range operation for maximizing S/N and minimizing crossmodulation.

An example of the coefficient system follows and is made under the following assumptions:

1. Crossmodulation,noise figure, and gain values are those given in Appendix A and B.

2. Cable attenuation changes .12% per degree F

3. Trunk amplifier spacing is 22 dB with AGC amplifiers spaced every other trunk station position.

4. Bridger coefficients calculated with bridger driven from MGC trunk amplifier.

5. Line extenders spaced at 12 dB of cable.
6. AGC line extender spaced every other position.
7. Noise figure and crossmodulation constant as function of gain.
8. Trunk amplifier output level 32 dBmv.
9. Bridger amplifier (4 outputs) output level 38 dBmv.
10. Line extender output level 43 dBmv.
11. Crossmodulation changes 2 dB for 1 dB of output signal level change.

The amplifier coefficients and system S/N and crossmodulation distortion are calculated for temperatures 0°F, 70°F and 110°F and are shown below:

TABLE II

		S/N		Crossmodulation	
Cascade	Actual	Equivalent		Equivalent	
Temperature		110°	70°	70°F	0°F
Trunk	20	27.58	20	20	24.455
Bridger (4-Out)	1	.276	.225	19.953	28.842
Line Extender	2	.287	.225	25.178	45.972
Coefficient C_n		28.14	20.45		
System S/N (dB)		44.01 dB	45.39 dB		
S/N = 58.5 -10 Log (C_n)					
Coefficient C_x				66.13	100.27
System Crossmodulation (dB)				-61.59	-57.97
XM = -98 + 20 Log (C_x)					

The system S/N and crossmodulation in the table above could be calculated by a number of methods. Each method would result in the same answer if the identical assumptions are used for each method. The coefficient $C_x = 66.13$, at 70°F, is the equivalent number of reference trunk amplifiers which generate a total system crossmodulation distortion of -61.59 dB. Note that for the assumptions made, which are realistic, the amount of distortion generated in the 4-output bridger amplifier is equivalent to the distortion which would result from a cascade of 19.95 reference trunk amplifiers. Because of system level variation due to cable attenuation change, the crossmodulation distortion generated at 0 degrees F is equivalent to that generated by a cascade of 28.84 reference trunk amplifiers. Table II graphically displays the sources of system S/N degradation by the different types of amplifiers used in the system.

CONCLUSION

Cable distribution system performance as related to system levels have been described. Clear insight into cable distribution equipment and system requirements is provided by use of signal level as a vehicle for analysis and design. This approach can result in providing guidelines for more economical designs.

An amplifier coefficient analysis and/or the use of Figures 7 and 8 are helpful in determining the allowable apportionment of noise and distortion between trunk and feeder systems.

ACKNOWLEDGEMENT

Portions of this paper are the result of investigations prompted by technical discussions with Mr. George Milner, Engineering Vice President, Cablecom General.

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APPENDIX A CROSSMODULATION COEFFICIENT

Reference Trunk Amplifier

M1 = -82 dB   Xmod at reference level (12 channels)

Y1 = 40 dBmv   Reference Level

F1 = 9.5 dB   Noise Figure at 270 MHz

G1 = 23 dB   Gain of Reference Amplifier

Q1 = 20 (Log (N9-1)- Log (11))   N9 is actual number of channels

X1 is crossmod of reference amplifier

X1 = (M1 + Q1)-(2)(Y1-E1)   Where E1 is output level

Bridger Amplifier

M2 = -82 dB   Xmod at reference level (12 channels)

Y2 = 40 dBmv   Reference Level

F2 = 12 dB   Noise Figure of bridger amplifier

G2 = 27 dB   Gain of bridger amplifier

E2 = Output level of bridger amplifier

X2 = (M2 + Q1) -2 (Y2 - (E2 + 7)   Crossmod of bridger at level (E2 + 7) dBmv

The bridger coefficient or number of equivalent trunk amplifiers in cascade (in terms of Xmod) is:

B1 = 10 raised to the power (X3 - X1)/20

Line Extender

M3 = -82 dB   Xmod at reference level (12 channels)

Y3 = 40 dBmv   Reference Level

F3 = 12 dB   Noise Figure at 270 mHz

G3 = 22 dB   Gain of line extender

X3 = (M3 + Q1) -2 (Y3 - E3)   X mod at level E3

The line extender coefficient (or equivalent number of trunk amplifiers is cascade, in terms of Xmod) is

L1 = 10 raised to the power (X3 - X1)/20

For a cascade of N2 line extenders,

L1 = (N2) (L1)

Total System Crossmod Coefficient CX (at nominal temp)

The total cascade number is then -	EQUIVALENT CASCADE	ACTUAL CASCADE
Total trunk cascade	N1 amplifier	N1
Bridger Coefficient	B1	1
Line Extender	L1	N2

System Coefficient = N1 + B1 + L1 = C_X

System Crossmod = X1 + 20 Log (C_X)

APPENDIX B - SIGNAL TO NOISE RATIO COEFFICIENT

Reference Trunk Amplifier

S1 Trunk S/N Ratio

$S1 = 59 - F1 + (E1 - G1)$ dB Equation

Bridger S/N Ratio

$S2 = 59 - F2 + (E2 + 7 - G2)$ dB Equation

Equivalent Trunk Cascade (In terms of S/N)

N3 = 10 raised to the power (S1 - S2)/10

Line Extender S/N Ratio

$S3 = 59 - F3 + (E3 - G3)$

Equivalent Trunk Cascade of single LE (In Terms of S/N)

Q4 = 10 raised to the power (S1 - S3)/10

N4 = (Q4) (N2) Where N2 = number of LE in cascade

System Noise Coefficient	EQUIVALENT CASCADE	ACTUAL CASCADE
Total Trunk Cascade	N1 Amplifiers	N1
Bridger Coefficient	N3	1
Line Extender Coefficient	N4	N2

System Noise Coefficient $C_n = N1 + N3 + N4$

System S/N = $S1 - 10 \text{ Log } (C_n)$

"PCM SUBCHANNELS FOR VIDEO MICROWAVE"

BY

DONALD KIRK and MICHAEL PAOLINI

ABSTRACT

A PCM subcarrier system has been developed which will permit video microwave to carry twenty-four 6 kHz channels along with the normal video signal. The channels may be used directly or split into forty-eight 3 kHz voice channels. The subcarrier system uses a differentially coherent quadraphase approach and has a data rate of three megabits.

The paper outlines the frequency requirements of implementing a subcarrier in the presence of video. Then a system block diagram description is given for the channelizing and subcarrier equipment. Performance curves taken from experimental data are given. These include the error rate performance for the system in the presence of differential phase and gain along with thermal noise.

SYSTEM CONSIDERATIONS

Although considerable baseband spectrum is available above the video signal on most microwave systems, it has remained relatively unused. The most notable exceptions are order wire subcarriers, fault reporting, and the addition of a few FM subcarriers for program channels. The reason for this sparse utilization is that the available spectrum above 5 mHz does not lend itself readily to conventional AM or FM subcarriers operating in the presence of the video signal. The nonlinearities of the microwave equipment, principally the differential phase and gain as well as the second harmonic distortion of the video signal all severely limit the achievable signal to noise performance of such subcarrier systems.

Figure 1-A contains a graph of the relative 15 kHz sidebands that are produced on a subcarrier due to differential phase and gain in the microwave system. The presence of these sidebands will show as degraded signal to noise performance for a subcarrier system. The sidebands due to differential gain will limit the performance of an AM subcarrier, and the sidebands due to differential phase will limit the performance of an FM subcarrier. However, if a digitally modulated subcarrier is used, the data stream may be regenerated, and the cross-modulation effects removed. Also, if a phase modulated subcarrier is used and the data stream is recovered by differentially comparing the phase of the subcarrier, then the effects of differential phase can be minimized even further.

If a digitally modulated subcarrier is to be used, some consideration must be given to how many subchannels may be placed in the available bandwidth. For the quadraphase system selected, the available bandwidth was taken to be between 5.0 and 7.0 mHz. This permits a sufficient guard band between the subcarrier and the video on the low side, and on the high side the band edge is sufficiently below the second harmonic of the 3.58 mHz color subcarrier.

To fully utilize the available bandwidth, twenty-four 6 kHz channels were implemented. This number of channels was derived after several considerations. First, a channel should be capable of carrying a 5 kHz AM broadcast station, as this was in fact to be the first application of the equipment. Secondly, a channel should be capable of being split into two 3 kHz channels for standard phone circuitry applications. Therefore, a channel with frequency response slightly above 6 kHz was selected as a basic channel unit. If additional channel bandwidth is required for a specific application, this can be achieved by occupying more than one basic channel unit.

To meet the signal to noise requirements of 55 dB, each 6 kHz channel has its analogue input signal quantized into 256 levels (8 bits). This number of bits allows for a basic signal to quantizing noise of 59 dB. The eight bit code, together with the fact that the 6 kHz channel would have to be sampled at approximately a 14 kHz rate, indicated that each channel would require about 100 kilobits/sec. The quadraphase system selected could readily handle three megabits in the available bandwidth, so a basic channel number for the subcarrier system was established at twenty-four channels.

Once the approximate sampling rate was selected, the question arose as to whether an optimum sampling frequency near 14 kHz would have minimum interference with the video signal. Subjective tests were made by adding an interference frequency in the 14 kHz range to a video signal. It was found that frequencies which were separated from the 15.734 video sync frequency by an odd multiple of one half the line frequency (60 cycles) produced interference "nulls". These points were subjectively much more tolerable than any other frequency inserted at the same level. A sampling frequency of 14.624 kHz was selected. This frequency is thirty-seven (a prime number) times thirty cycles (one half the line rate) below the 15.734 kHz video sync frequency.

A further requirement of the system was to have the subcarrier frequency synchronous with the data rate. To achieve this, a 17.5488 mHz master oscillator was utilized and divided by three to generate a 5.849600 mHz basic subcarrier frequency. The basic subcarrier frequency is then divided by four hundred to obtain the 14.624 kHz sampling rate.

This basic sampling rate times the eight bits per sample determines the individual channel data rate which is 116 kilobits/sec. Since twenty-five channels are used on the system (twenty-four channels plus one channel for system synchronization purposes) the total data rate is 2.9248 megabits/sec.

To insert this information above the video signal, consideration was given as to which type of modulation is best suited to withstand the problems of the microwave system. After a tentative evaluation of several possibilities, a differentially coherent quadraphase modulation system was selected. This approach has several inherent advantages over other possible implementations. Since the detection process for this system is accomplished by comparing the carrier phase difference between two sequential data bits, the effects of differential phase are minimized. This is because the differential phase is occurring at a slow rate compared to the data stream,and although the total phase shift of the subcarrier may be large over a 15 kHz interval, the amount of phase shift between two adjacent data bits is considerably less.

Another advantage of this approach is that since the modulation information is on the phase of the carrier, it may be amplitude limited to remove the effects of differential gain.

The quadraphase approach was implemented instead of binary phase shift keying in order to meet the bandwidth requirements of the subcarrier system. The occupied bandwidth of the system may be limited to 1.5 mHz, which easily fits into the allotted two mHz band above the video.

SYSTEM DESCRIPTION

Figure one shows a block diagram of how the subcarrier system is implemented on a microwave path. The twenty-four 6 kHz inputs are time division multiplexed by the PCM transmitter. The channels are formed into two binary data lines, each carrying data at one half the overall three megabit rate. The PCM transmitter also generates the synchronization information and the carrier source for the subcarrier transmitter. These signals are supplied to the subcarrier transmitter which modulates the data onto the 5.84 mHz subcarrier. The modulated signal is band limited and added to the normal video signal. The composite signal is then supplied to the microwave baseband.

At the microwave receiver, the composite signal is connected to the subcarrier receiver which removes the subcarrier from the video signal, and demodulates the data stream back into two binary data lines. The binary data lines are connected to the PCM receiver which decodes the data back into twenty-four 6 kHz audio channels.

Figure two contains a block diagram of the PCM transmitter. The 17 mHz crystal oscillator is used to derive all of the timing signals, including the 5.84 mHz subcarrier. The basic data rate of one bit per 700 nanoseconds is exactly one fourth of the subcarrier frequency. A single channel occupies four sequential time slots. There are twenty five channel assignments (twenty four channels plus one channel for synchronization) which total one hundred data intervals per 14 kHz sampling interval. The timing generator divides the basic data rate with a seven bit countdown chain which resets on the one hundreth data pulse. The seven bit timing pulses synchronize the indivual channel encoders onto a common pair of data lines.

The block diagram for the subcarrier transmitter is shown in Figure three. The 5.84 mHz carrier source is split into four phase related sources of the same frequency. The signals are at a 90° spacing, and are supplied to the phase modulator. The arithmetic unit cumulatively adds the data bits prior to transmission. This is necessary since the differentially coherent detection process in the subcarrier receiver is subtractive. Although the cumulative addition could have been done in the subcarrier receiver, this is not the optimum location. If the arithmetic unit is placed in the receiver, a single error in the detection process will cause all of the remaining bits in that sampling interval to be wrong. Therefore, the arithmetic unit was placed in the transmitter which has the data stream at a much higher signal to noise ratio and is essentially error free.

The synchronization channel occupys four data time slots as in the normal information channels. During the first two time intervals of the synchronization period, the 5.84 mHz subcarrier is amplitude modulated to an "off" state. This modulation is AM detected in the subcarrier receiver and used to phase lock the local clock in the PCM receiver. During the third time interval 0^{o} phase reference carrier is transmitted alternately (on different sample intervals) with 180^{o} phase carrier. The fourth time interval is used to always transmit 0^{0} reference phase. In this manner, alternate sampling periods are uniquely identified. The third time interval corresponds to a channel split pulse that is used when a 6 kHz channel is split into two 3 kHz channels. Each 3 kHz channel is alternately sampled by the 14 kHz sampling pulse, which effectively divides the 14 kHz sampling rate in half for each channel.

After the data and synchronization information is added to the subcarrier, the modulated subcarrier is band limited to restrict the occupied bandwidth in the microwave baseband. The incoming video signal is also band limited to 4.5 mHz to remove any harmonics which would fall in the subcarrier channel. The subcarrier is then added to the video signal and the composite signal is connected to the microwave baseband.

Figure four contains a block diagram of the subcarrier receiver. The composite baseband signal from the microwave receiver passes through a splitting filter which removes the subcarrier signal from the video signal. The subcarrier signal is then split to drive both an AM detector and a phase detector. The AM detector recovers the synchronizing pulse that occurs during the first two time intervals of the synchronization channel interval. The detected pulse is used to phase lock the receiver clock.

The phase detection process is accomplished by comparing the phase of the subcarrier during the presently arriving data bit to the phase of the subcarrier during the previous data bit. This process is subtractive, as the recovered data stream is the difference between the two bits. To recover the data stream directly, the bits are cumulatively added in the transmitter, so that the difference operation in the receiver produces the data stream directly.

Figure five contains a block diagram of the PCM receiver. The detected AM sync pulse is phase compared with a 14 kHz sampling pulse generated by dividing down the VCXO in the PCM receiver. The VCXO is at 5.84 mHz and has a countdown chain similar to the one in the transmitter. The phase comparator is enabled only during the first two time slots of the synchronization channel. This prevents spurious sync pulses from entering the phase comparator while the remaining twenty four channels are transmitting. To acquire lock, the sync pulse derived from the local VCXO slowly advances in time through the different channels. When the local sync pulse passes through the synchronization channel, the phase comparator is enabled, and lock is achieved.

The 5.84 mHz VCXO in the receiver generates the same timing signals that were available in the transmitter. Individual channel decoders are enabled for a four bit interval corresponding to their correct encoding channel, and the recovered data is decoded into the original analogue input signal.

SYSTEM PERFORMANCE

To evaluate the system performance in the presence of thermal noise and microwave distortion, a pseudo-random sequence generator which could occupy a single transmitted channel was designed. At the receiver, a channel card which was programmed to accept the known sequence was implemented, and an error count was made between the locally generated sequence and the received sequence. It was found that additional channels did not change the error rate performance, with the exception of the two channels adjacent in time to the error test channel. Therefore, the data was always taken with these adjacent channels fully loaded.

A test set was made which would simulate microwave distortion, as well as add thermal noise to the subcarrier signal. Differential gain was produced by amplitude modulation of the subcarrier at a 15 kHz rate. Differential phase was produced by varactor modulation driven by a 15 kHz full wave rectified sine wave which produced parabolic phase distortion similar to that encountered in microwave systems. The test set was checked by using a video test set which uses a 3.58 mHz test signal. Good agreement was observed between the video test set and the sideband predictions of Figure 1-A when measured on a spectrum analyzer.

Figure six contains error rate performance curves for different values of signal to noise with different amounts of microwave distortion. For no microwave distortion, the subcarrier system achieves an error rate of 10^{-5} for a signal to thermal noise ratio of 17 dB. This would mean an individual 6 kHz channel would have approximately one error per second. Experimental results with 5 kHz music channels indicated that less than ten errors per second would pass for an undegraded music channel.

The outside curve of figure six (on the right hand side) is an error rate versus signal to noise curve where diffential phase and gain have been applied to the subcarrier. Differential gain sufficient to produce 20 dB relative sidebands on the subcarrier, and differential phase also sufficient to produce 20 dB relative sidebands were simultaneously impressed upon the subcarrier. This corresponds to approximately 23° of differential phase and 3.5 dB of differential gain. The microwave distortion has the effect of shifting the error rate curve to the right, that is, it makes apparent signal to noise degradation. To achieve a 10^{-5} error rate, the signal to noise had to be increased to 21 dB as compared to 17 dB without microwave distortion. Otherwise, there was no change in the individual 6 kHz channel performance. This indicates the digital system does eliminate the differential phase and gain effects of the microwave system, but must be operated at a slightly higher signal to noise ratio than would otherwise be expected.

For less than 20 dB differential phase and gain sidebands, the error rate curve shifts to the left toward the thermal noise curve. The two inside curves plotted are for 20 dB differential phase sidebands only, and for 20 db differential gain sidebands only.

Prototype equipment has been evaluated on a four hop back to back microwave system, and through a 500 mile heterodyne repeater system. In both cases, an error rate of 10^{-5} was achieved by operating the subcarrier approximately 17 dB below the composite video signal.

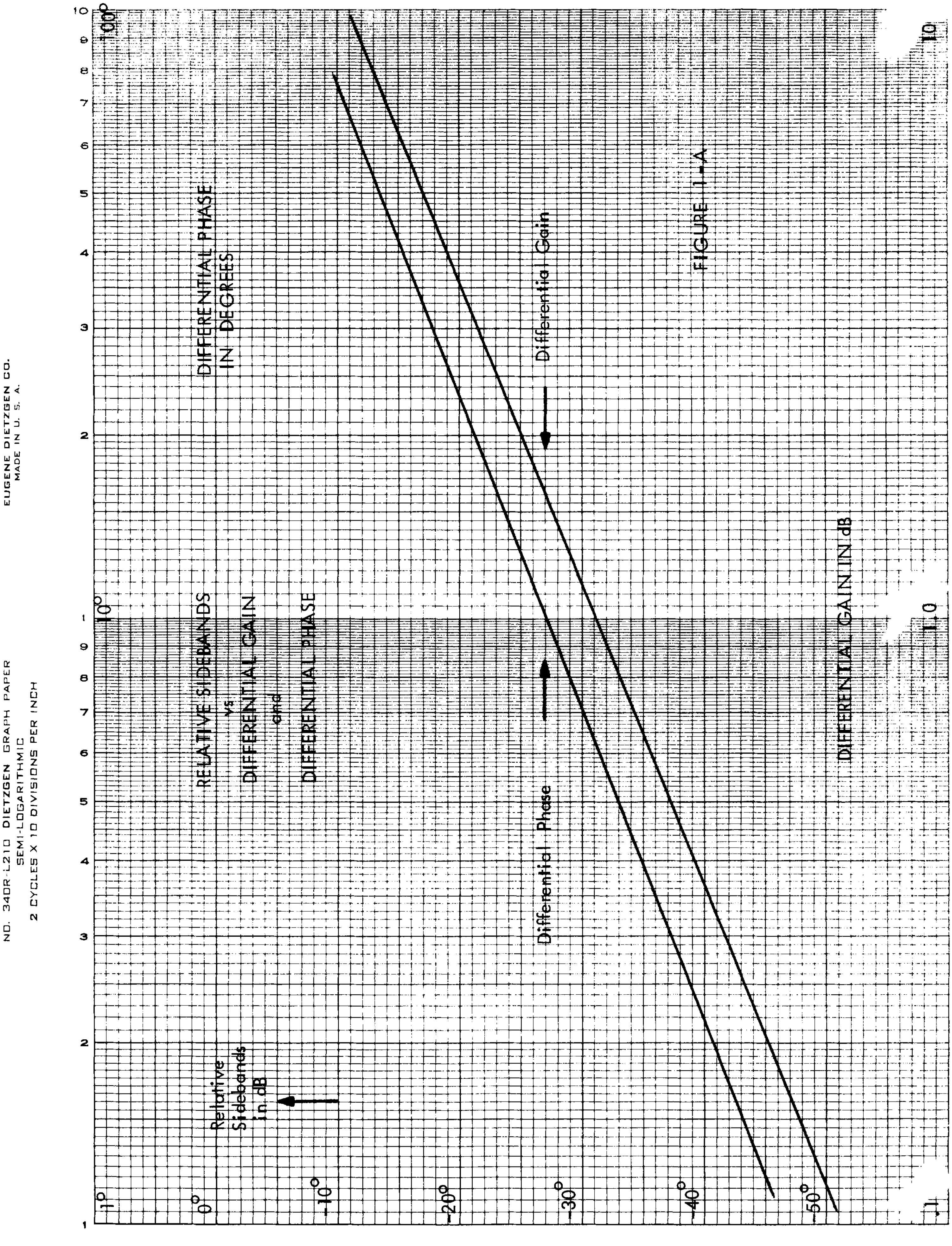
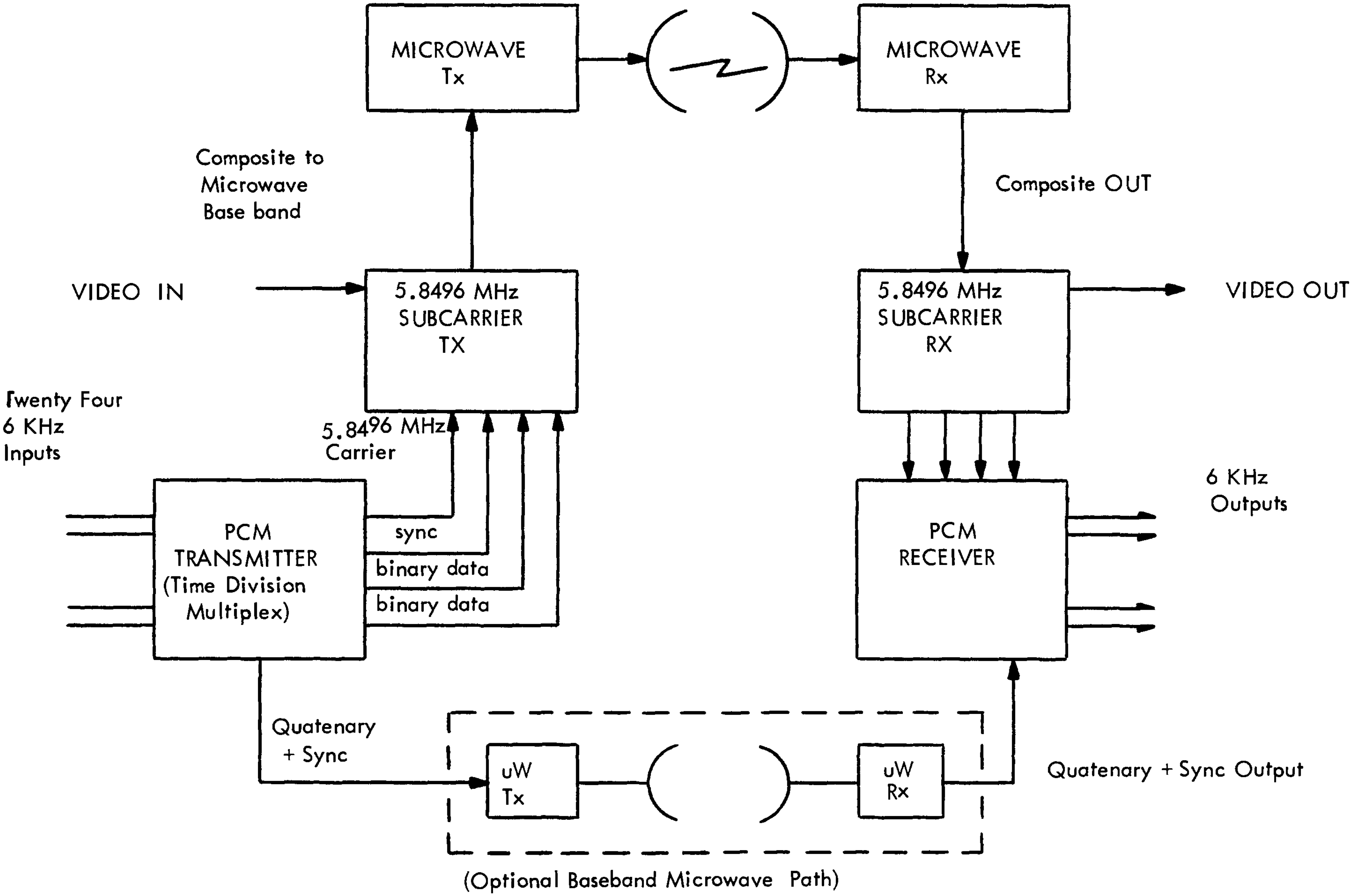

RELATIVE SIDEBANDS
vs
DIFFERENTIAL GAIN
and
DIFFERENTIAL PHASE
DIFFERENTIAL PHASE
IN DEGREES
Differential Gain
Differential Phase
Relative
Sidebands
in dB
DIFFERENTIAL GAIN IN dB
FIGURE 1-A
0°
-10°
-20°
-30°
-40°
-50°
1°
10°
100°
.1
1.0
10
NO. 340R-L210 DIETZGEN GRAPH PAPER
SEMI-LOGARITHMIC
2 CYCLES X 10 DIVISIONS PER INCH
EUGENE DIETZGEN CO.
MADE IN U. S. A.

FIGURE ONE


MICROWAVE
Tx
MICROWAVE
Rx
Composite to
Microwave
Base band
Composite OUT
VIDEO IN
5.8496 MHz
SUBCARRIER
TX
5.8496 MHz
SUBCARRIER
RX
VIDEO OUT
Twenty Four
6 KHz
Inputs
5.8496 MHz
Carrier
6 KHz
Outputs
PCM
TRANSMITTER
(Time Division
Multiplex)
sync
binary data
binary data
PCM
RECEIVER
Quatenary
+ Sync
uW
Tx
uW
Rx
Quatenary + Sync Output
(Optional Baseband Microwave Path)

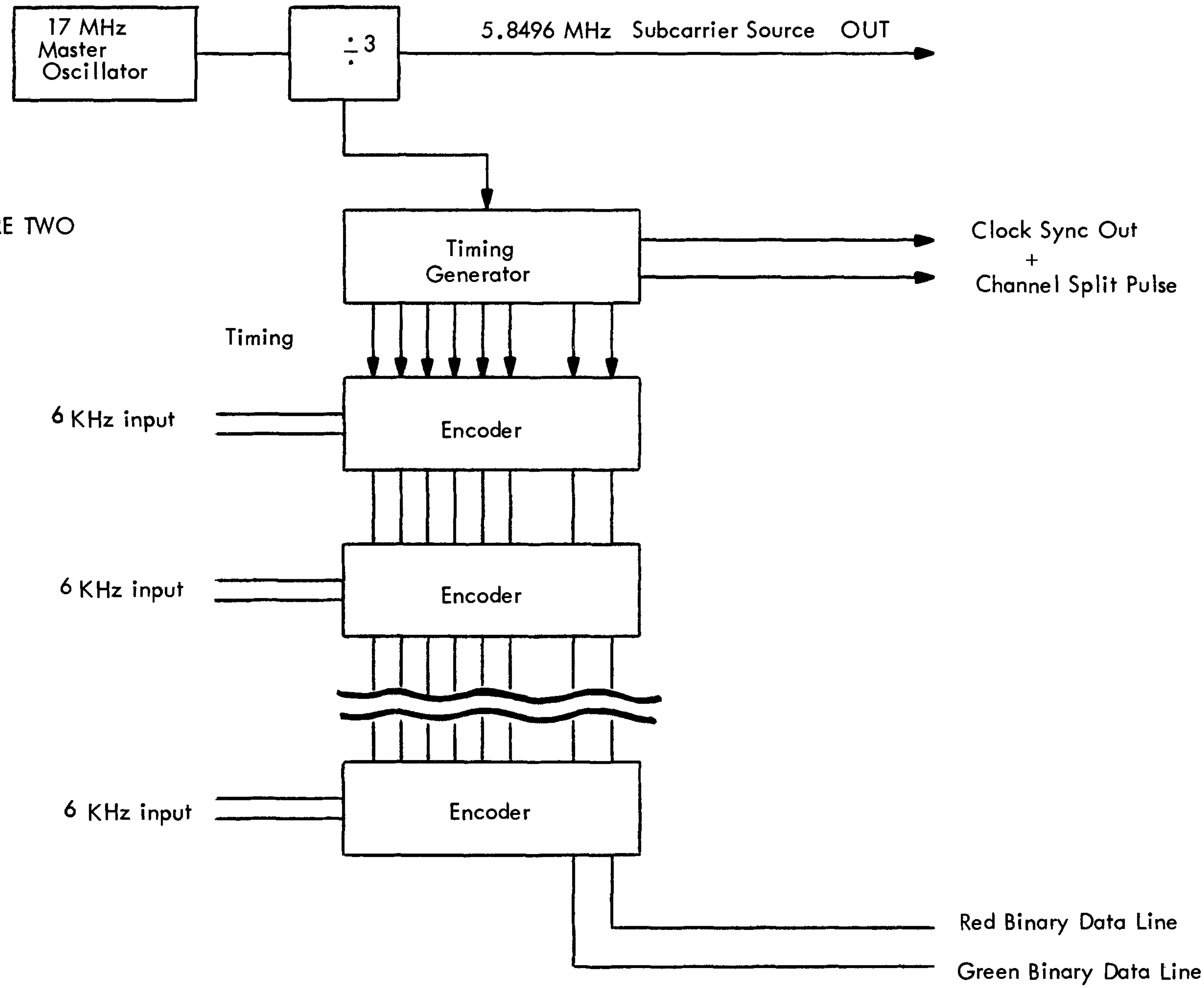
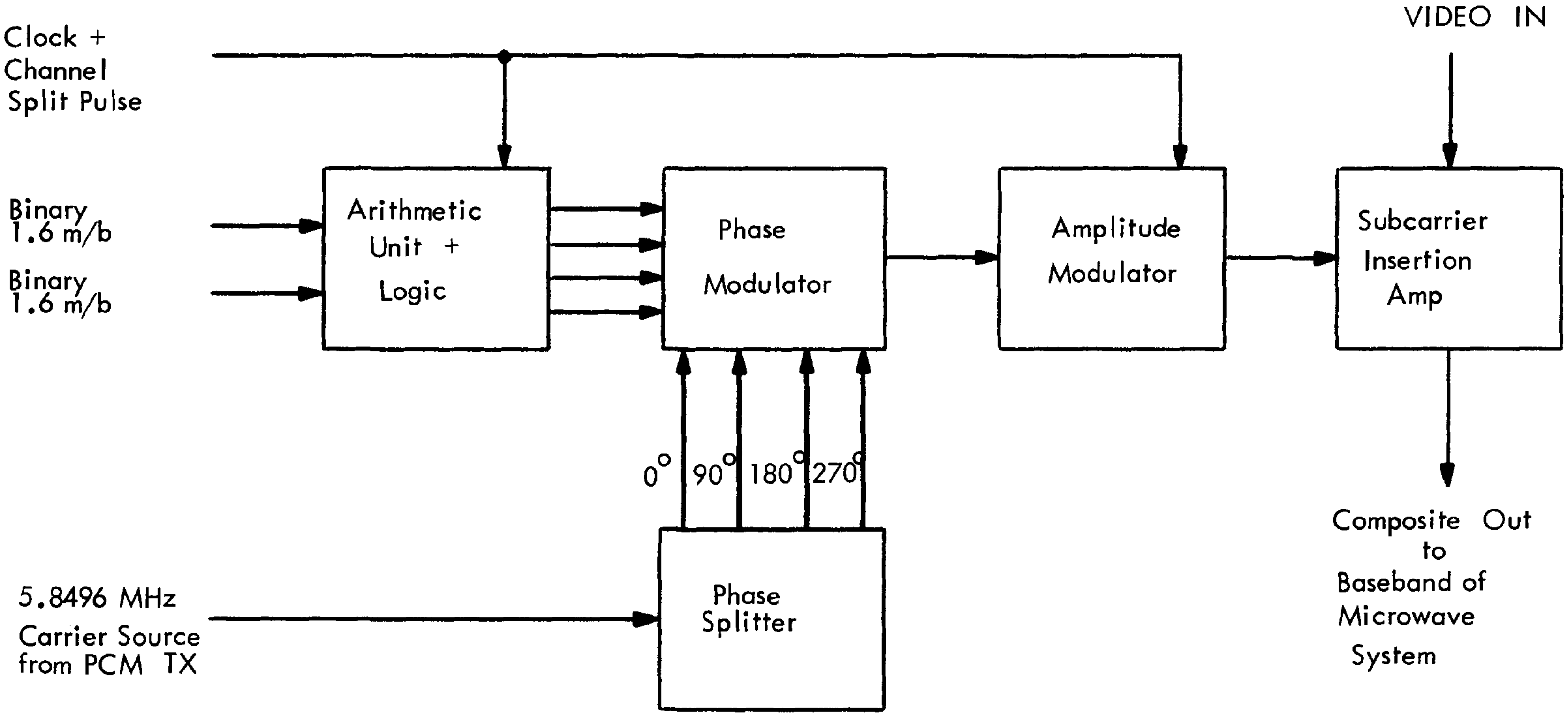


FIGURE TWO

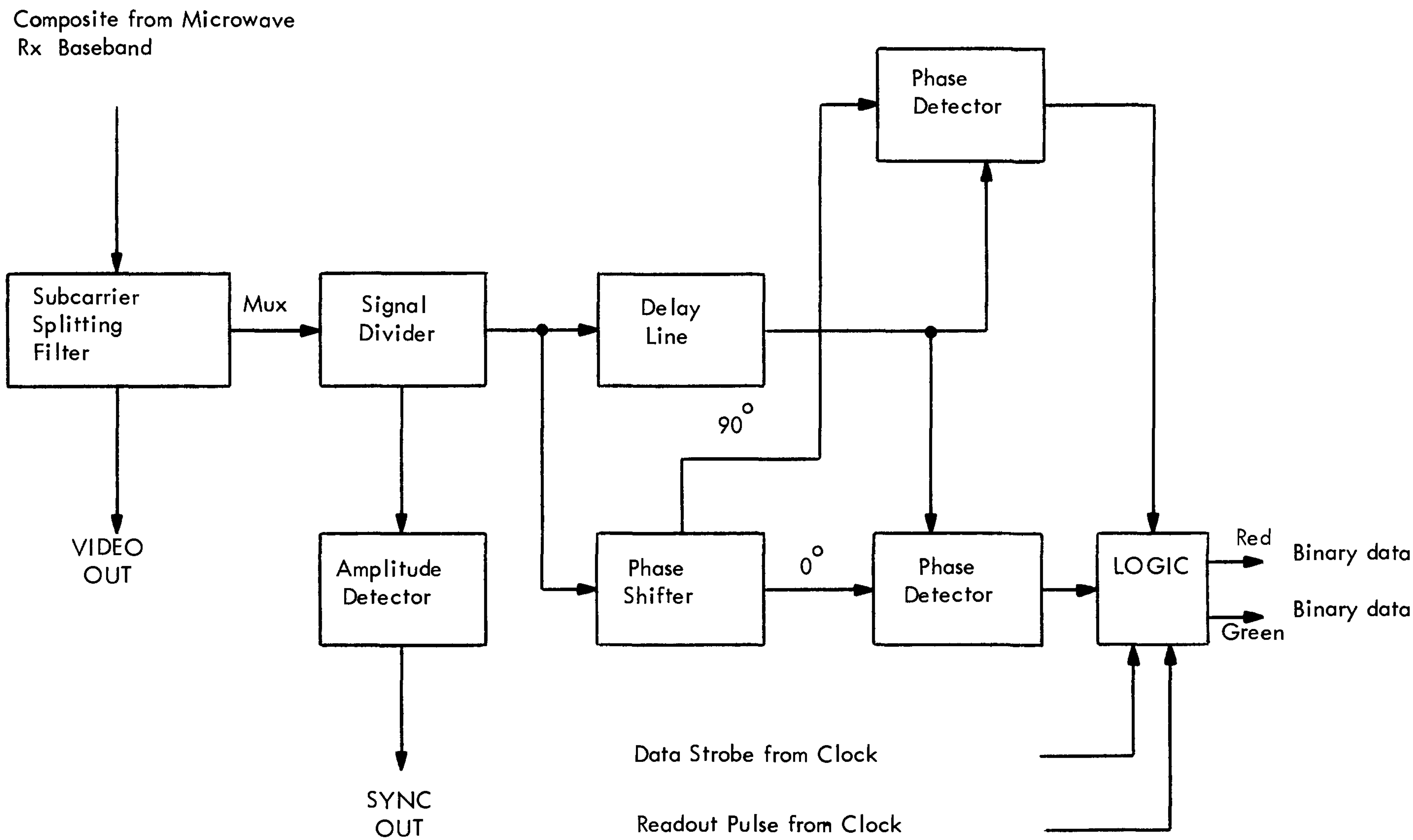
PCM MULTIPLEX TRANSMITTER

FIGURE THREE

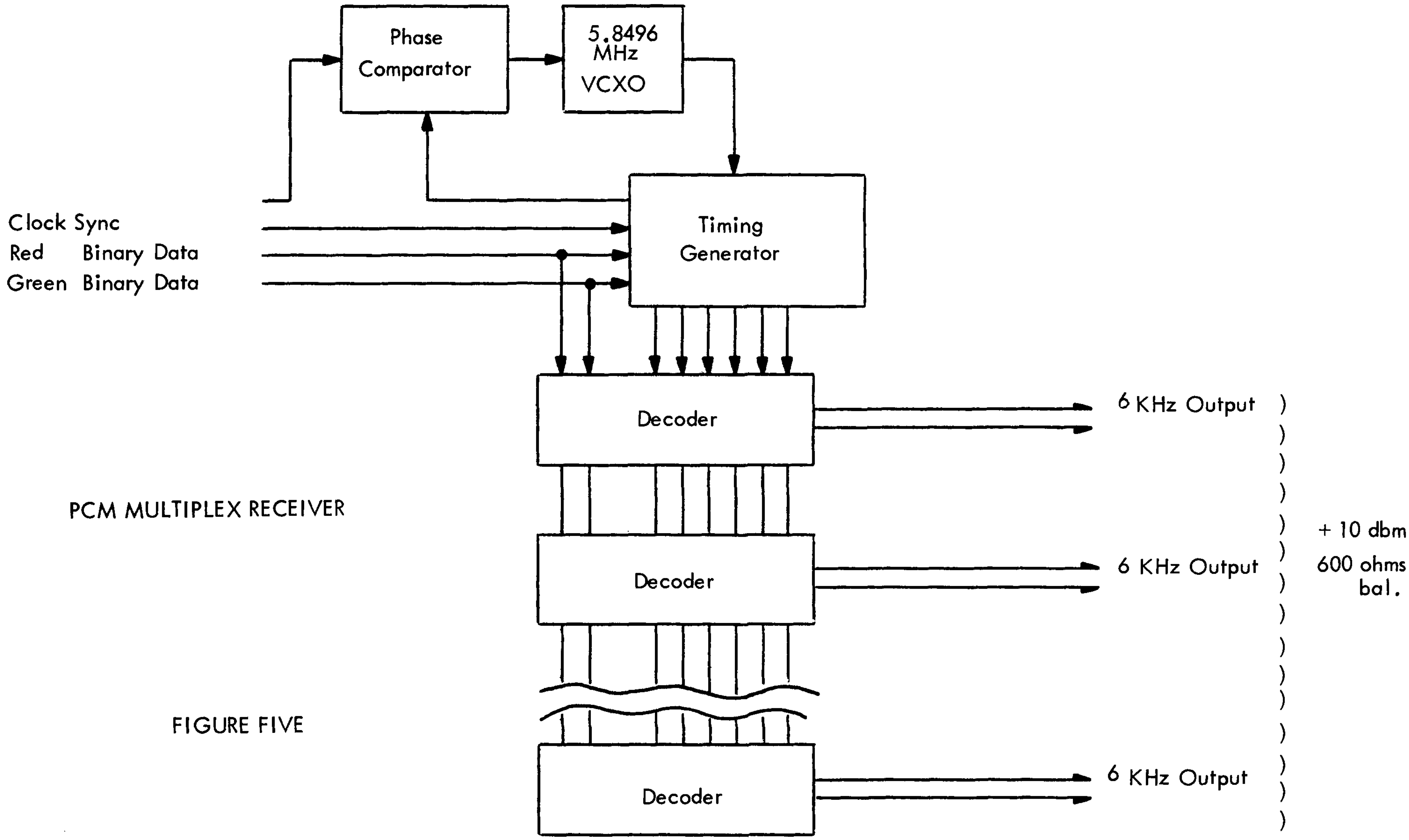


SUBCARRIER TRANSMITTER

FIGURE FOUR



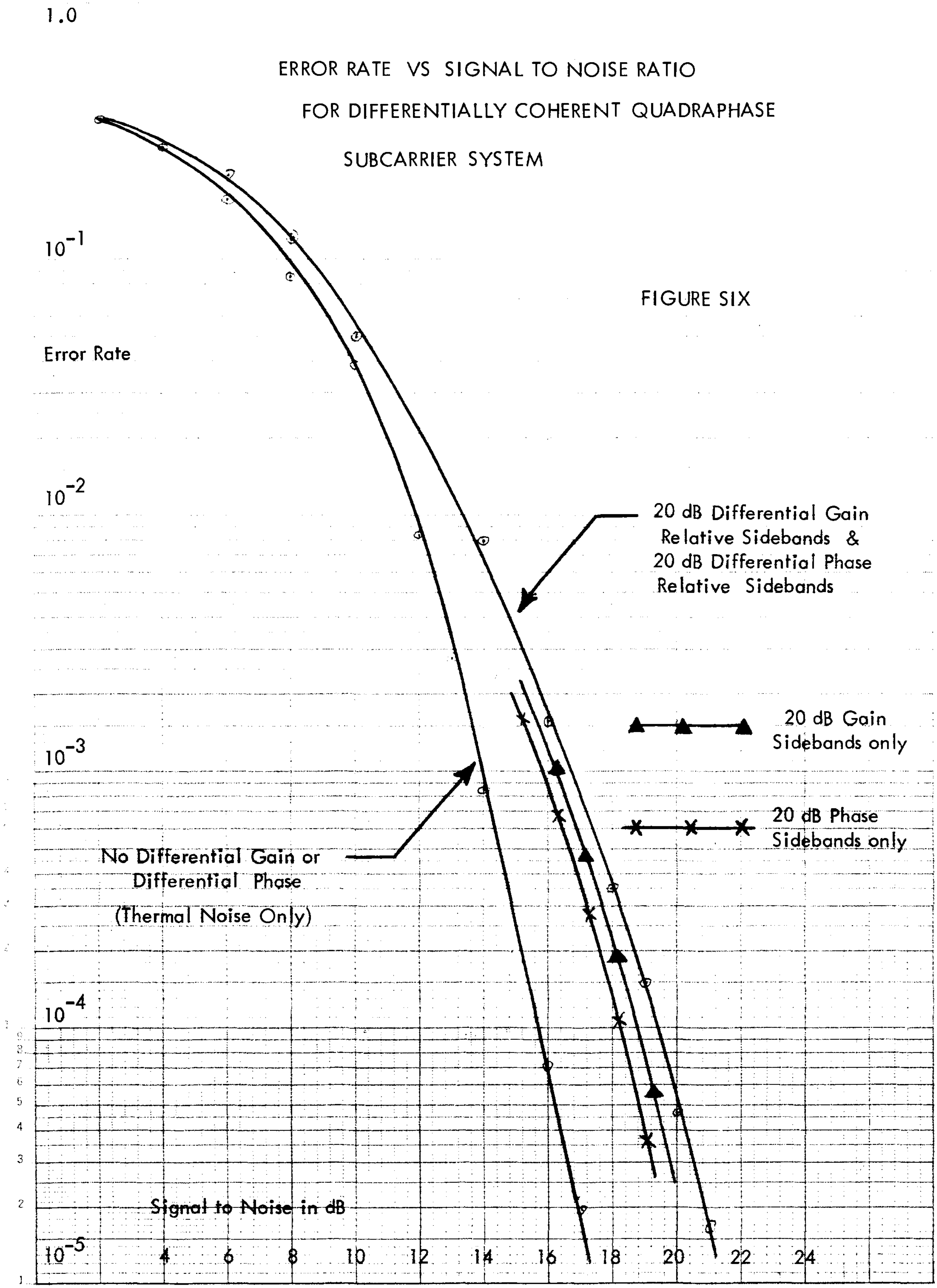
SUBCARRIER RECEIVER



PCM MULTIPLEX RECEIVER

FIGURE FIVE

Harmonic Distortion 2% max.
IM - 60/6000 2% or -34 dB


1.0
ERROR RATE VS SIGNAL TO NOISE RATIO
FOR DIFFERENTIALLY COHERENT QUADRAPHASE
SUBCARRIER SYSTEM
10^{-1}
FIGURE SIX
Error Rate
10^{-2}
20 dB Differential Gain
Relative Sidebands &
20 dB Differential Phase
Relative Sidebands
20 dB Gain
Sidebands only
10^{-3}
20 dB Phase
Sidebands only
No Differential Gain or
Differential Phase
(Thermal Noise Only)
10^{-4}
Signal to Noise in dB
10^{-5}
4
6
8
10
12
14
16
18
20
22
24

Security Alert
A Two-way Digital Communications System

Marvin Roth, Senior Engineer
Scientific-Atlanta, Inc.
Commerical Communication Div.

INTRODUCTION

For years CATV has provided a one-way communications system. Many people have proposed an increase in the services offered to the subscriber by providing two-way communications utilizing the CATV distribution system. The uses and applications of such systems are limited only by one's imagination. The technical problems of two-way data communications system have been solved; hardware has been developed and is now available. Scientific-Atlanta's Security-Alert system is such a system.

The Security-Alert two-way data communications system was developed primarily for use in monitoring and reporting the condition of many remotely located transponder units (subscriber stations) to a centrally located interrogation unit (central station). The information to be communicated is digitally encoded and modulated on an FSK (Frequency Shift Keyed) carrier and transmitted over the CATV cable distribution system.

The system consists of a central station (to be located at the headend or some central distribution point), the transmission path, (the CATV distribution system), and the subscriber stations (one unit for each subscriber).

A few of the many applications include the monitoring and status reporting of:

1. Home fire and intrusion alarms
2. Monitoring on-off conditions of remotely located equipment
3. Telemetering equipment to measure parameters such as voltage, current, power levels, temperature, pressure,etc.
4. Monitoring and reporting locations of faults along the CATV distribution line

The system can be easily modified to provide the capability for additional applications such as:

1. TV tuner monitoring and polling
2. Opinion polling
3. Mass audience participation game-playing
4. Remote control of equipment and machinery

SYSTEM DESCRIPTION

Security-Alert is an automatic sequential polling system. (Figure 1) Five bits of binary coded information (32 different messages or alarms) can be received by the central station from each of 8,192 subscriber stations. The output of the system is a printed record showing any change of status of alarm (message) from the subscriber stations. Refer to Figure 2. The information which is recorded by the digital printer contains:

1. A five digit OCTAL coded address
2. A two digit OCTAL coded alarm (message)
3. The time of day when the alarm (message) was received

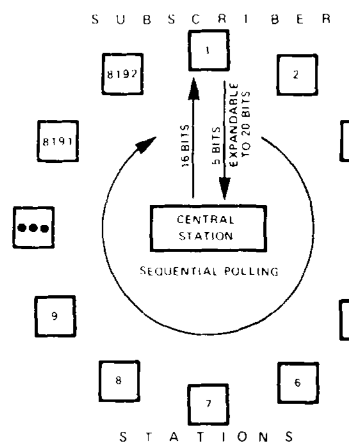


Figure 1 Sequential Polling

ADDRESS CODE	ALARM CODE	TIME HR/MIN
11657	04	1156
10615	07	1021
01253	25	0907

1ST Alarm occurred at 9:07 A.M.
Subscriber 01253 reported a Code 25.

2ND Alarm occurred at 10:21 A.M.
Subscriber 10615 reported a Code 07.

3RD Alarm occurred at 11:56 A.M.
Subscriber 11657 reported a Code 04.

Code 40 = Power Failure
Code 41 = Data Failure
Code 42 = False Alarm

All other Codes are assigned by the user.

Figure 2 Recorded Output

Fail safe provisions have been incorporated within the system. If a subscriber station is polled and an answer is not received within a reasonable time period, the system interprets the nonresponse and a "power failure" code is printed out for that subscriber station's address. A positive response must be received, or a power failure printout occurs. This feature can be used to easily locate equipment failures along the CATV distribution line as well as locating faulty subscriber units. Two other fault location indicating codes are also automatically interpreted and printed out when they occur.

The speed of transmission and processing information is of utmost importance in communications systems carrying information which may mean the difference between life and death, such as reporting fires. Security-Alert was designed with this thought in mind. The system cycle time (the time it takes to sequentially poll, examine, and report the response of 8,192 subscribers) is variable. The system is an adaptive one. For a fixed bit rate, and a time division multiplex system the cycle time depends on:

1. The total number of active subscriber stations.
2. The distance of each subscriber station to the central station
3. The number of alarm/messages and change of status of alarm/messages.

For example, let us consider the case of 8,192 subscriber units whose average distance from the central station is ten miles (20 miles round trip). If no alarms exist, the cycle time would be approximately five seconds. For each change of status of alarm/message, the cycle time would be increased by 0.5 seconds (the time it takes to print out the message).

Means have been provided in the basic system to easily change the system capacity. Two bits (four unique combinations) are not used in the basic system but are available in the interrogation word. These two bits can be used to:

1. Increase the maximum number of subscriber units sending five bits of alarm/message information from 8,192 to 32,768.
2. Increase the number of alarm message bits received from 8,192 subscriber units from 5 bits to 20 bits (1,048,576 different combinations).
3. Combinations of the two extremes of the above.
4. The two extra bits can also be used as command or instruction words to cause ar event to occur at any subscriber station (such as remotely turning equipment off or on).

BASIC OPERATION

The Security-Alert system operates on a master-slave principle. The master unit is the central station. All of the subscriber stations are slave units. Each subscriber station is assigned and programmed to recognize a unique binary coded address word. A subscriber unit is capable of transmitting information only after it has decoded the input address code, compared it with the preprogrammed address and recognized it as its own address. Refer to Figure 3.

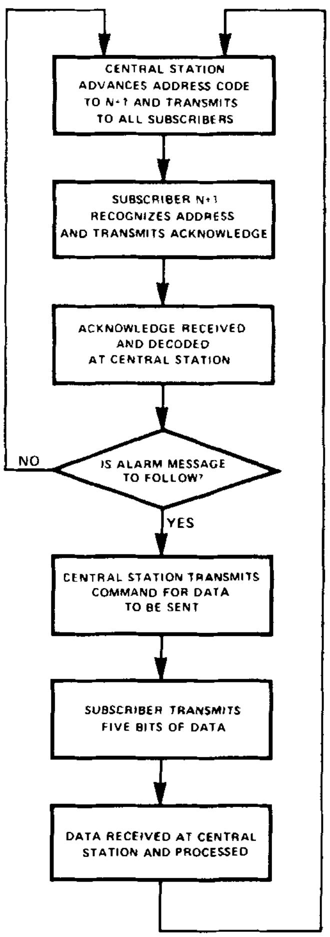


Figure 3
Flow Diagram of Basic Operation

The address code is transmitted by the central station to all subscriber stations. The subscriber station which is pre-programmed to recognize that unique address replies with an acknowledgement pulse. This acknowledgement pulse is coded to indicate whether an alarm message exists or not. If there is no alarm message the central station advances the address by one count and transmits the new address to all subscriber stations. If an alarm message is received the central station transmits a command causing the answering subscriber station to transmit a five bit binary word which contains the alarm message. This two-step method of alarm reception was chosen to decrease system cycle time. If no alarms are reported, only one bit is transmitted in the reverse direction. Five bits are transmitted only when an alarm condition exists. After the central station processes the alarm message the address word is advanced one count and the next subscriber station is interrogated.

CENTRAL STATION

The central station produces all timing and control signals used by the system. The timing signals are transmitted along with data to all subscriber units. The transmission of these timing signals eliminates the need for timing oscillators and complicated circuits to phase lock the incoming data in each subscriber station.

The central station is comprised of four major subsystems (refer to Figure 4). A digital clock provides a visual display of hours, minutes, and seconds. It also provides the proper signals representing hours and minutes for recording on the digital printer.

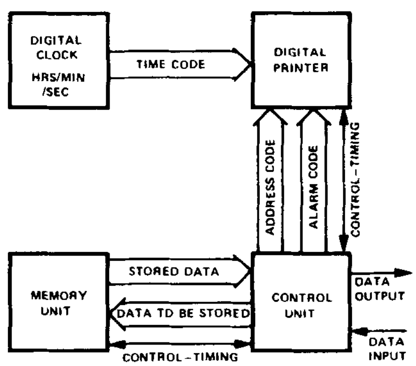


Figure 4
Central Station Block Diagram

The digital printer is the primary output device. As previously discussed, it prints the alarm code, the address code, and the time of day automatically upon command from the control unit.

The control unit provides all timing and control signals, transmits an interrogation word to all subscriber units, receives the subscriber station's reply, processes the information and causes an output when required.

The memory unit provides eight bits of storage for each of the 8,192 addresses. Addresses can be manually assigned in memory. If an address is assigned it is transmitted. If the address is not assigned, the control unit does not transmit that address code. The address register is advanced and the next address is processed. The memory is nonvolatile and remembers the last alarm printed for all addresses. The data in memory is compared with the input data and a printout occurs only when the incoming alarm for a particular address is different from that previously recorded and memorized. The system senses and prints a change of status of alarm rather than an alarm. It is obvious that this increases system speed by preventing the same information from being printed more than once.

Since the central station controls the entire system, let us start with the generation of an interrogation word and follow the signal as it travels to and from the subscriber station.

INTERROGATION WORD

The "interrogation word" is a polling command which is simultaneously transmitted to all subscriber stations. It contains data composed of 16 bits of binary coded information as well as timing signals. The data portion of the interrogation word is produced by generating a 13 bit binary code address word and applying these in parallel with three other bits to a parallel-to-serial converter. The output of the parallel-to-serial converter is a 16 bit NRZ (Non Return to Zero)word. Refer to Figure 5.

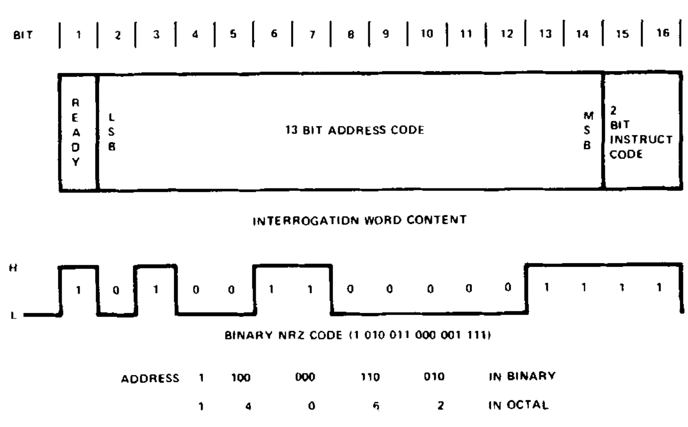


Figure 5 Interrogation Word

The first bit is used to ready all subscriber units by resetting a counter in each unit. The next thirteen bits contain the address code. The last two bits are reserved and can be used to issue instruction commands, to increase the maximum number of subscribers, or to increase the number of alarm/message bits from the subscriber.

The binary coded NRZ word has timing signals added to it by a bi-phase encoder. The output of the bi-phase encoder is then modulated and transmitted.

Figure 6 shows a binary NRZ code.

each clock pulse transistion. Data and timing information are combined and transmitted on one continous pulse train.

The clock rate of both Ø 1 and Ø 2 is 160 KHz. The data is derived from Ø 2 producing a bit rate of 160,000 bits/ second. Since the interrogating word contains sixteen bits, the length of the interrogation word is 100 microseconds. Since the bi-phase code consists of the interlacing both Ø 1 and Ø 2, the maximum transmission modulation rate is 320 KHz.

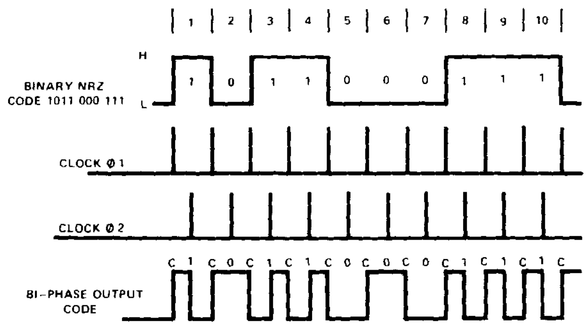


Figure 6 Bi-Phase Encoder Timing Diagram

Ten bits are shown for simplicity. A binary "one" is represented by a high level, and a binary "zero" is represented by a low level. Timing information such as the data rate is not contained in this code. To identify the binary code it is necessary to know the data rate so the word can be examined and the code understood. By adding timing signals to the binary NRZ code through an encoder as shown in Figure 7, a bi-phase code is generated.

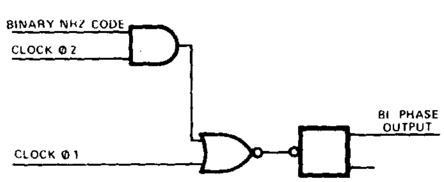


Figure 7 Bi-Phase Encoder

By examing the bi-phase code shown in Figure 6, it can be seen that the levels of "high" and "low" are meaningless and that data and timing information is contained in the transistion between the high and low levels. For each clock pulse Ø 1 a transistion occurs. When the input data is high a transistion occurs at a clock pulse Ø 2 and when the input data is "low" a transistion does not occur. The clock pulses occur at a regular interval, the clock period. The data always occurs half way between

FORWARD TRANSMISSION PATH

The bi-phase binary encoded interrogation word modulates an FSK (Frequency Shift Keyed) transmitter. Refer to Figure 8. A binary "one" is transmitted at a frequency of 113.9 MHz and a binary "zero" is transmitted at a frequency of 111.1 MHz. The two frequencies are generated by precision crystal controlled oscillators.

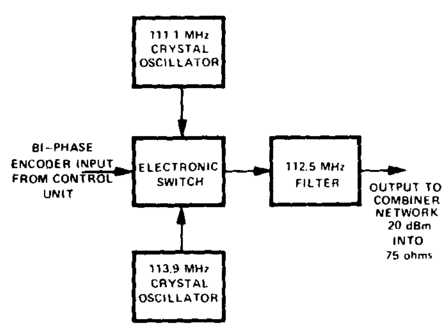


Figure 8
112.5 MHz Transmitter Block Diagram

The output of the oscillators are passed through bandpass filters with very steep slopes to keep unwanted sidebands out of other channels. Available power at the 75 ohm output terminals is +29 dBmV. This will be at least 15dB below the normal level of TV signals at the combiner output. The modulated interrogation word is added with the TV signal in a combiner network and travels down the normal CATV distribution system to each subscriber unit. The input range of the receiver in the subscriber set can vary in signal strength from -10 dBmV to +15 dBmV and still be within its operational range.

Propagation delay along the distribution system is an important parameter in calculating system cycle time (refer to Table 1). It consists of cable delay and delays through the various amplifiers and various amplifiers and passive devices along the distribution line. As a rule of thumb, a delay of 8.3 microseconds per mile is used when calculating system cycle times.

Table 1

FORWARD PATH PROPAGATION DELAY FOR INTERROGATION WORD	
INTERROGATION WORD LENGTH	100 MICRO SEC
FIXED TRANSMITTER-RECEIVER-DECODER DELAY	30 MICRO SEC
CABLE DELAY (INCLUDING 4 AMPLIFIERS/MILE)	8.3 MICRO SEC/MILE

SUBSCRIBER STATION

Refer to Figures 9 and 10.

The subscriber drop is always a single cable carrying two-way information, It enters the subscriber station unit where the input signal is split. Half of the signal is made available through a type "F" connector to provide signals to a TV set or other device. The digitally modulated signal passes through a passive frequency selective "tee" network and then into a 112.5 MHz receiver. The input sensitivity of the receiver is between -10 dBmV and +15 dBmV. The receiver consists of three stages of tuned amplification followed by a broadband limiter. The output of the limiter is fed to a discriminator circuit and the output level of the discriminator is sensed by a level detector which reproduces the transmitted bi-phase binary encoded word. The digital signal is then fed to a decoder and the timing signals are separated from the data contained in the bi-phase coded input word.

The first bit (always transmitted as a one) resets all counters and readies the transponder for more data. If a noise pulse were to provide this bit all subscriber stations would be inactive but ready, and would still interpret the first bit of the transmitted message as a ready pulse and not as data. Only a burst of noise having the same word length as the interrogation word would be interpreted as a transmission.

The timing pulses are counted in a binary counter located in the address comparator. The preprogrammed address is gated out and compared bit by bit with the input

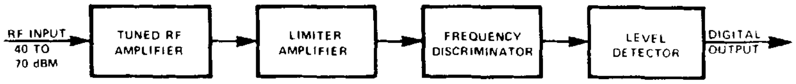


Figure 9 112.5 MHz Receiver

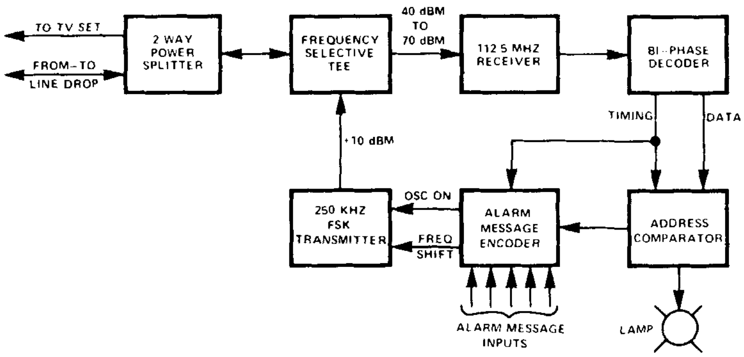


Figure 10 Block Diagram Subscriber Station

data. As soon as a subscriber station recognizes one bit difference between the input data and the preprogrammed address, it resets itself and waits for the next interrogation word. The one station which recognizes all input bits as being identical with its own pre-programmed address sets a bistable circuit. This causes two events to happen. A lamp on the front panel of the subscriber unit turns on signifying that this particular subscriber station is being interrogated. This provides a self-checking feature at the subscriber station.

At the same time the station's alarm message inputs are examined. If no alarm conditions exist (signified by open circuits to the five alarm message inputs) the 250 KHz FSK transmitter is instructed to turn on and transmit a 275 KHz signal for approximately 125 microseconds. If an alarm message exists (closure of one or more alarm message inputs) the transmitter is turned on and a frequency of 225 KHz is transmitted for approximately 125 microseconds. The signal is referred to as the acknowledge word and it is transmitted into the subscriber drop cable through the frequency selective tee network and power coupler.

The central station interprets the reception of 275 KHz as a no-alarm message and proceeds to advance the address register and interrogate the next subscriber station (see Figure 11). If the received signal is 225 KHz, then five 12.5 microsecond data strobe pulses spaced every 100 microseconds are transmitted sequentially to all subscriber stations. Only that subscriber station whose address had just been recognized allows the data strobes to be gated through to the alarm message encoder. The encoder converts the five input message code to a serial binary string of "ones" and "zeros." The 250 KHz FSK transmitter is turned on for a period of approximately 525 microseconds, and the transmission is FSK modulated between 225 KHz and 275 KHz by the data contained in the serially coded alarm message. The central station processes this information and then interrogates the next subscriber unit.

If two units in a time division multiplex system which uses one cable attempt to transmit at the same time and at the same frequency they will interfere with each other. A failure in one of the subscriber stations which would cause the transmitter to be turned on continously would render such a system useless. To prevent this, a simple fail safe feature has been incorporated in the 250 KHz transmitter. In the event of a failure of this type, the oscillator stage will turn itself off and remain off until the failure condition is corrected.

When this station is interrogated by the central station a "power failure" code will be printed out at the central station and the problem will be brought to the attention of the system operator.

Signals are available within the transponder to decode the two instruction word bits. These are used for optional features of the system.

REVERSE TRANSMISSION PATH

The acknowledgement pulse and the alarm message word are FSK modulated and transmitted in the reverse direction. A binary "one" is transmitted at a frequency of 275 KHz and a binary "zero" is transmitted at a frequency of 225 KHz. The power available at the transmitter output is +60 dBmV into a 75 ohm load. The input range at the receiver end is between +10 dBmV and +30 dBmV.

The system was designed for use with either a single cable or with a separate reverse transmission cable. The single cable system will require all components in the distribution system to be bi-directional. In an existing distribution system it may be desirable to install a separate low cost reverse transmission cable (RG59) instead of replacing all un-directional components. The loss along the cable (RG59) at the frequency of 0.25 MHz is approximately 20 dB per mile. To keep the signal level along the line above the minimum detectable level at the receiver, inexpensive low frequency amplifiers and combining networks are available. Since the loss per mile is low, fewer amplifiers would be required in the reverse than in the forward direction.

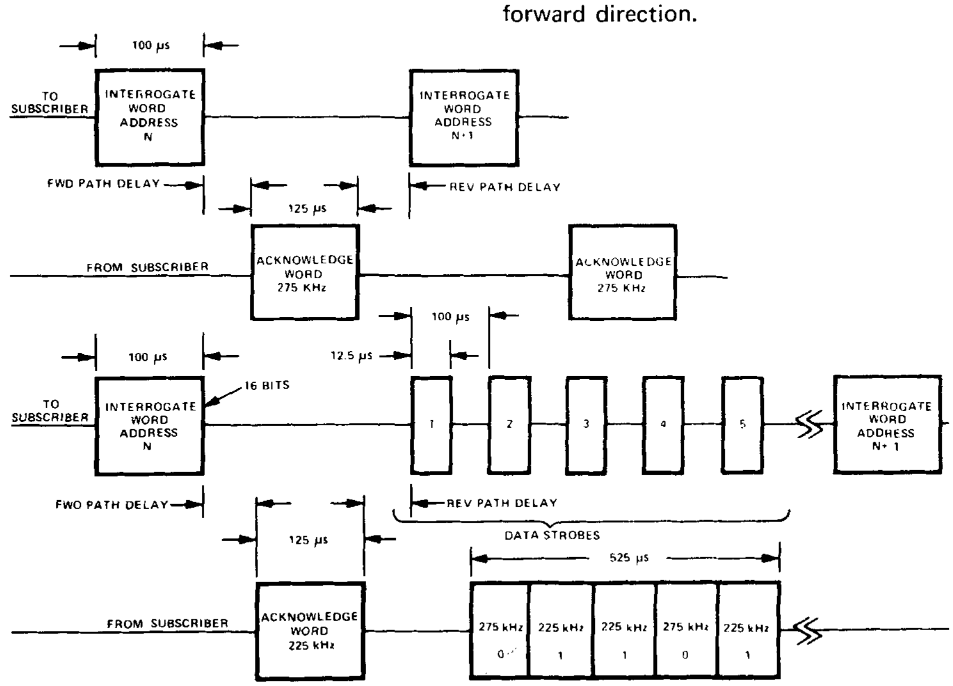


Figure 11 System Timing Diagram

The cable drop to the subscriber is always a single cable independent of the single or dual cable system.

In calculating the cycle time the forward transmission path propagation delay due to the cable and the components in the reverse transmission path must be taken into account. These are given in Table 2.

Table 2

REVERSE PATH PROPAGATION DELAY FOR ACKNOWLEDGEMENT WORD	
INTERROGATION WORD LENGTH	125 MICRO SEC
FIXED TRANSMITTER RECEIVER PROCESSOR DELAY	120 MICRO SEC
CABLE DELAY (INCLUDING 4 AMPLIFIERS/MILE)	8.3 MICRO SEC/MILE

DATA RECEPTION AND PROCESSING

The modulated acknowledgement pulse and alarm message which are transmitted in the reverse direction are received at the central station and applied to a 250 KHz, three-section Butterworth filter (see Figure 12). The filter outputs to a 250 KHz receiver which is composed of a limiter amplifier, a frequency discriminator, a level detector,

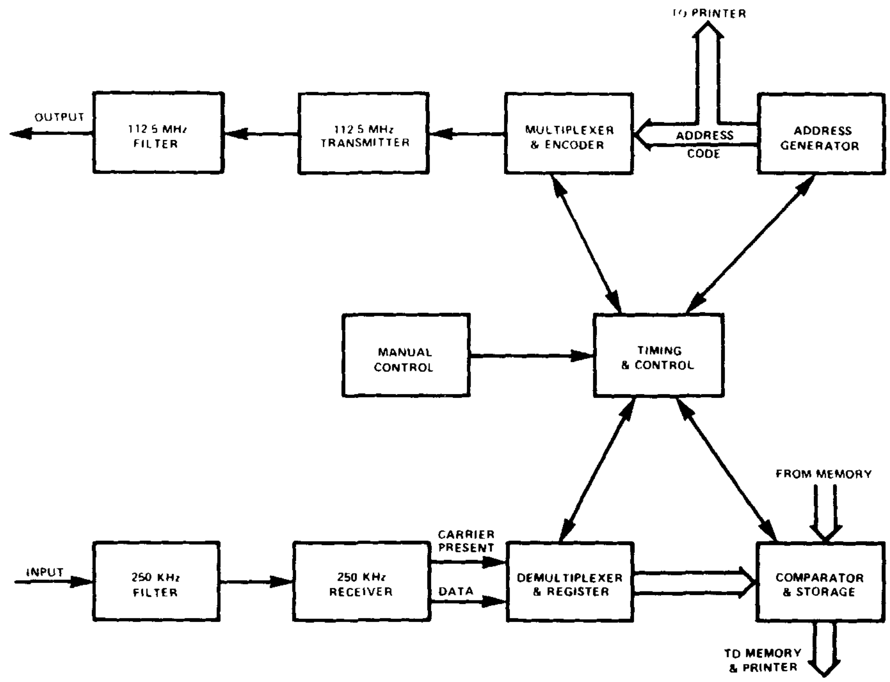


Figure 13 Control Unit Block Diagram

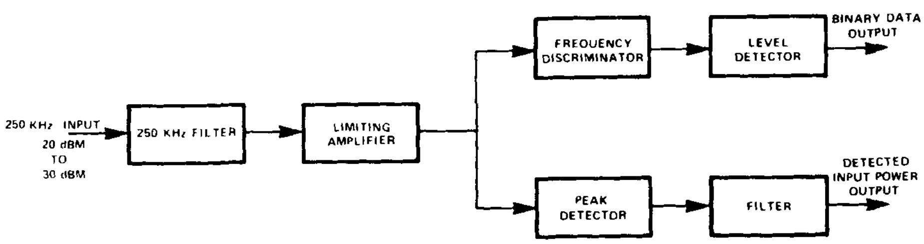


Figure 12 250 KHz Receiver & Filter

and a carrier presence detector. The input to the discriminator circuit and carrier presence detector is held constant by the limiter amplifier as the input signal level varies from 3 millivolts to 30 millivolts. The carrier presence detector provides an output (a binary "one") whenever a signal (either 225 KHz or 275 KHz) is received. The frequency discriminator and level detector provide a binary "one" only when the input signal frequency is 225 KHz. The two digital outputs are applied to the data processing portion of the control station.

Figure 13 shows a block diagram of the control unit. The control unit is normally automatic in operation; however, an operator may take control of the system manually. All timing signals are derived from a 1.28 MHz crystal controlled master clock oscillator. The output of the oscillator is divided down to provide the required timing signals. All signals (input, printer, manual commands) which are asychronous in nature are synchronized with the master clock before being processed. This may delay data processing up to one clock period, but the delay is small compared with other system delays.

ADDRESS GENERATOR AND MULTIPLEXER

A block diagram of the address generator and multiplexer is shown in Figure 14. The 160 KHz clock Ø 2 is applied to a 13 bit binary counter through NAND gate 1. The second input to NAND gate1 consists of inhibit functions derived in the timing and control circuitry. When data is being processed the address counter is prevented from being advanced. The address counter will advance only when both inputs to NAND gate 1 are high. Right after the advance of the address counter, the timing and control circuitry cause an inhibit function to occur until the data received from the new address is processed.

The timing and control circuits also provide inhibit functions to NAND gate 2. This gate sets a bistable circuit which allows the 160 KHz clock (Ø 2) pulses to be applied to a four-bit binary counter. The 16th count of the counter is detected and bistable circuit 1 is reset, and remains reset until the next uninhibited clock pulse is allowed to set it. The four output lines of the counter are applied, to and control the single output line of a 16 line to 1 line multiplexer. The 16 lines of input consist of the 13 bit address code and three other bits entered in parrallel. The bits are sequentially gated out of the multiplexer on the output line a bit at a time, as controlled by the 16

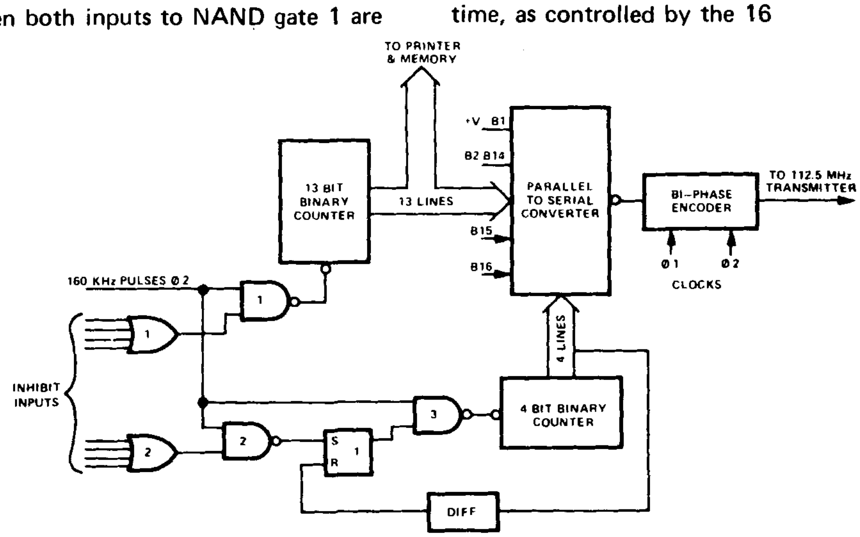


Figure 14 Address Generator & Multiplexer

combinations of the four bit control line. The sequential single line output is applied to the bi-phase encoder and then to the 112.5 MHz transmitter.

DEMULTIPLEXER AND DATA REGISTER

The 250 KHz receiver provides two output signals. One output signal occurs when a carrier signal of either input frequency is present. The second output signal is the detected data (refer to Figure 15). At the end of the transmission of an interrogation word, a pulse is generated by the timing and control circuitry that sets a bistable circuit starting a clamped time delay circuit. The time delay is set for the maximum round trip propagation delay time of the system. If a signal is received before the end of this period then the bistable circuit is reset and the input data is examined by NAND gates 1 and 2. A binary "one" is decoded by gate 1 as a "no alarm" message. A binary "zero" is decoded by gate 2 as an alarm message. The output of NAND gate 3 causes code 40 to be recorded and stored in memory if a signal is not received before the end of the time delay period, and if Code 40 (power failure) has not been previously printed for this address.

If the data received contains a "no alarm" message it is compared with data stored in memory. If a "no alarm" message was not previously stored in memory code "00" is printed out. If the received data was decoded as an alarm message then the data strobe generator is actuated and five sequential data strobes are transmitted to the answering subscriber station. The subscriber station will reply with a five bit binary coded message. When this signal is received the data trigger generator is actuated and five sequential triggers are generated. The triggers are used to sequentially gate the serial input data into a serial-to-parallel converter. Since the timing of the triggers is based on the reception of the data word, each bit will be read in and stored at the proper time. At the end of the fifth trigger a "cycle count trigger" is generated. The output of the serial-to-parallel converter (the data register) will hold the stored five bit input data word until the next data word is received by the system.

ALARM COMPARATOR (Figure 16)

When an alarm is detected the message is examined three times to prevent the processing of a possible false alarm. The five bits of input data are compared with the message previously stored in memory for the address being interrogated. If the output of Comparator 1 is a binary zero (an equal condition) at the time the "cycle count trigger" occurs, the timing and control circuitry cause the address generator to advance, and the next address is interrogated. If the output of Comparator 1 is a binary one, it is gated with the "cycle count trigger" in NAND gate 1 and bistable 1 is set. This causes one input of NAND gate 2 to go high and also causes the five bit input data word to be stored and held in a five bit "latch" circuit. The timing and control circuitry then cause the same address to be interrogated for a second time. The second set of five bits of input data is compared in comparator 2 with the first set of data previously placed in storage. If the two sets of data are not equal bistable two is placed in the set state. The same address is interrogated for a third time at the end of the "cycle count trigger". The third set of input data is compared with the stored first set in comparator 2. Bistable 2 is set if the comparison indicates a "not equal" condition.

Each time the cycle count trigger occurs, a modulus 3 counter is advanced one count. After the third trigger the two

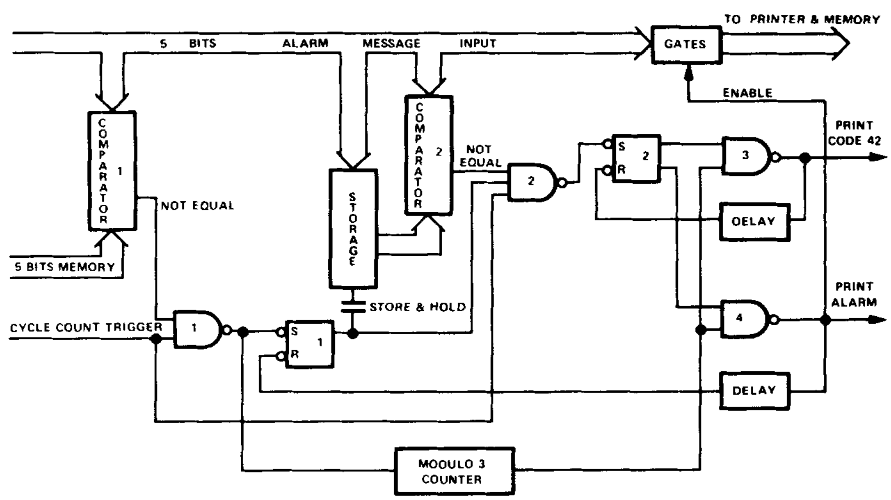


Figure 16 Alarm Comparator

gates NAND 3 and NAND 4 are strobed. If the input data was identical three consecutive times then gate 4 causes a "print alarm" command to occur. The five bit coded message stored in those data registers is printed by the recorder and stored in memory. A code "42" (false alarm) is printed if the data was not identical three consecutive times as sensed by gate 3 and stored by bistable 2. After printing the bistable circuits are reset and the timing and control circuitry cause the next address to be interrogated.

APPLICATIONS

Transponder Data Inputs

The five data input lines to the transponder are coded a binary-coded octal. Data input one represents $(01)_8$; input two represents $(02)_8$; input three represents $(04)_8$; input four represents $(10)_8$; and input five represents $(20)_8$.

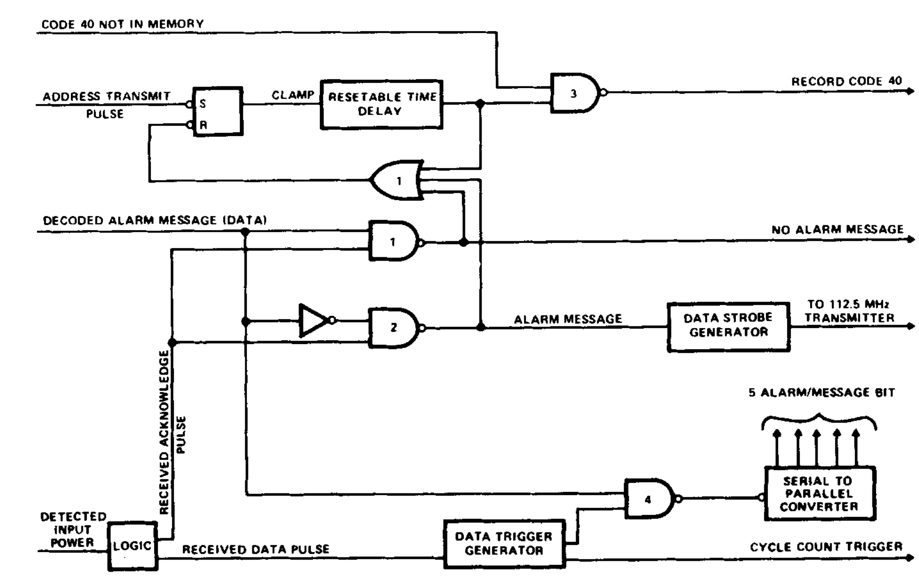


Figure 15 Demultiplexer & Data Register

The switch inputs are normally open contacts. Figure 17 shows a block diagram of five input switches S1–S5.

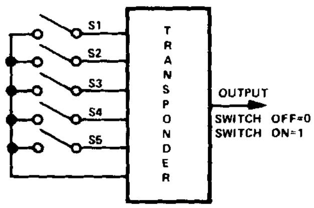


Figure 17
Five Independent Switch Inputs

Table 3 shows all possible combinations of the five switches. An "O" represents an open switch contact (an "off" condition) and a "1" represents a switch closure (an "on" condition). When all switches are off the transponder will transmit a $(00)_8$ code. When all switches are on the transponder will transmit a $(37)_8$ code. All other combinations are shown in the transmittal code column of Table 3.

Table 3 Transponder Input Codes

SWITCH	S1	S2	S3	S4	S5		TRANSMITTED CODE
LINE \ CODE	01	02	04	10	20		
1	0	0	0	0	0		00
2	1	0	0	0	0		01
3	0	1	0	0	0		02
4	1	1	0	0	0		03
5	0	0	1	0	0		04
6	1	0	1	0	0		05
7	0	1	1	0	0		06
8	1	1	1	0	0		07
9	0	0	0	1	0		10
10	1	0	0	1	0		11
11	0	1	0	1	0		12
12	1	1	0	1	0		13
13	0	0	1	1	0		14
14	1	0	1	1	0		15
15	0	1	1	1	0		16
16	1	1	1	1	0		17
17	0	0	0	0	1		20
18	1	0	0	0	1		21
19	0	1	0	0	1		22
20	1	1	0	0	1		23
21	0	0	1	0	1		24
22	1	0	1	0	1		25
23	0	1	1	0	1		26
24	1	1	1	0	1		27
25	0	0	0	1	1		30
26	1	0	0	1	1		31
27	0	1	0	1	1		32
28	1	1	0	1	1		33
29	0	0	1	1	1		34
30	1	0	1	1	1		35
31	0	1	1	1	1		36
32	1	1	1	1	1		37

GENERAL APPLICATIONS

The five data inputs may be used in two different ways. Each of five independent functions may be monitored to indicate an on-off condition. All combinations of the five independent inputs can be uniquely identified by the octal code as shown in Table 3. The switches represent inputs such as S1 for smoke or heat detection, S2 for intrusion detection, S3 for a "panic" alarm indicator. If fire occurs, code 01 is transmitted. If an intrusion occurs code 02 is transmitted. If fire and intrusion occur simultaneously then code 03 is transmitted.

The five data inputs may also be used to monitor 32 discreet levels of one input variable as shown in Figure 18. A remotely located transducer monitors a varying parameter. The analog output of the transducer is converted to binary-coded octal in an analog-to-digital converter. The range of measurements can be resolved into 32 discrete steps.

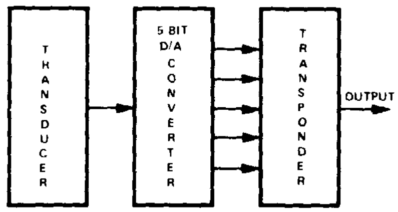


Figure 18 Using the Transponder For Monitoring One Input Parameter in 32 Discrete Measurement Levels

With optional circuitry (a multiplexer) the system can be expanded to transmit twenty bits of data. The increased system can be expanded to transmit:

1. Twenty independent on-off switch closures (Figure 19)
2. 1,048,576 discrete levels of one independent variable (Figure 20)
3. Combinations of the two above extreme cases.

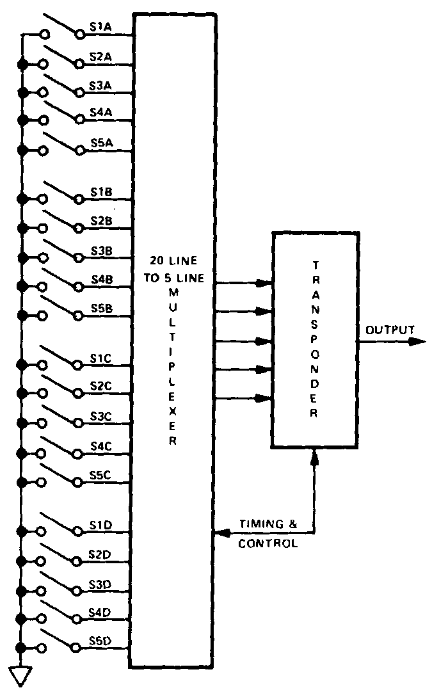


Figure 19 Using the Transponder for Sensing 20 Independent Switch Inputs

This is accomplished by using bits 15 and 16 of the interrogation word for byte control. There are four possible states of the two bits and each byte contains five bits of data. The subscriber station is interrogated four times to receive the twenty bits of data.

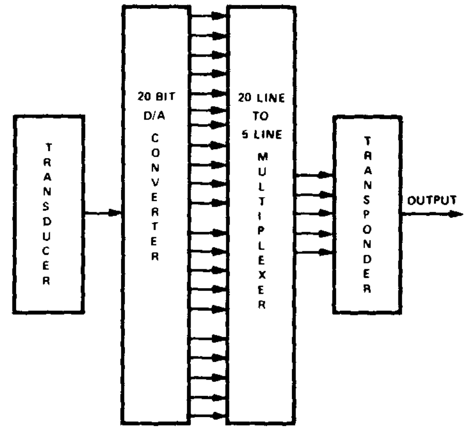


Figure 20 Using the Transponder For Monitoring One Input Parameter For 1,048,576 Discrete Measurement Levels

CONCLUSION

CATV distribution systems can be adapted for two-way data communications. The first type of communication system to be used will probably be a "polling" system such as the Security-Alert system discussed within this paper.

This system is available and can be used to increase the services offered to the subscriber.

Probably the first application of this system will be used to monitor and sense remote events such as fire and intrusion alarms. The system can monitor any switch closure and can also be adapted to monitor time varying parameters such as voltage, current or power. As more services are required by the subscribers the system can be expanded and adapted for uses such as:

1. Polling (program tuner and opinion)
2. Game playing
3. Remote control
4. Educational purposes
5. Information retrieval
6. Special program selection
7. Subscriber remote turn on - turn off service
8. Distribution system fault locating and reporting.

The future of two-way data communications utilizing the CATV cable promises to be an ever expanding field.

SPECIALIZED TEST EQUIPMENT FOR CATV DISTRIBUTION MEASUREMENTS

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INTRODUCTION

Historically, the test equipment used in setup and troubleshooting of the CATV distribution system has evolved from the early days of little or no test equipment to today's practices of adapting laboratory type equipment for field use plus a continually growing use of low priced signal level meters. It is true that test equipment manufacturers have recently "discovered" CATV and produced a rash of 75 ohm units. Now it is possible to put together an impressive array of test equipment and study in detail a system amplitude, phase and delay transmission characteristics, providing one can keep the equipment operating long enough and can cope with a lot of other practical problems associated with getting around and tapping into a CATV distribution system.

Over a period of years the authors have been "witness to" or participated in field excursions where the Ritual of a "count down" was observed to make sure that all of a multitude of pieces were safely stored aboard a vehicle before departing. More often than not it has been normal for a critical piece of gear to be left behind or to find that transportation vibration has killed some thing crucial to completion of the appointed task. The frustration of these experiences can be matched only by the problem of trying to interpret data taken on equipment with built-in errors, which mask the significance of that data.

These experiences have lead to belief that there is a crying need in the industry for a quality specialized test set capable of making basic field measurements with reliability and precision required by today's standard of performance. The authors contend that such a basic instrument makes much more economic sense than elaborate time domain reflectometry and spectrum analyzers which certainly have their place at today's level of sophistication.

In retrospect, it is somewhat surprising that the economics advantages of such test equipment have not been properly recognized. The true cost of errors and delays due to test equipment problems can be staggering. Hundreds of dollars per day in labor wasted, vehicle charges, customer aggravation - these items add up quickly to pay for a $2,000.00 item, which can save down time or prevent errors. Utility companies and the military have long recognized the need for "test sets" specifically designed to "set up" and adjust complicated systems in a rapid and foolproof manner. Perhaps the authors draw an unfair comparison with manufacturing where it has become normal to review such expenditures for tooling in terms of labor saving dollars. Certainly if this were done, the demand for such test equipment would have been much stronger.

GENERAL

With this previous discussion as background, it can be stated that this paper will describe a test set designed and slated to be produced on a "limited edition" basis at C-COR for the express purpose of making the "nuts and bolts" measurement of signal level, noise, and distortion throughout a CATV system. The major economic motivation in our case is the potential improvement in efficiency in our own Systems Engineering Department.

In addition, certain specialized production test units constructed and used by C-COR will be briefly discussed. Some of these represent the forerunner of the circuitry used in the test set; others are shown because they illustrate the similarity between field test and production test or may have some value around a maintenance laboratory.

A review of some of these factory test units, designed and built at C-COR will follow in the succeeding paragraphs.

FACTORY TEST EQUIPMENT

Some of the problems associated with factory testing are very similar to those encountered in the field. For instance, one of the most critical is the frequent calibration of selective RF voltmeters and the radio and frequent measurement of things like noise figure at spot frequencies. To facilitate these operations, we have constructed two basic test units. The first shown in Slide I is a Level Calibrator, which is simply a stable multiple signal source that is periodically calibrated by the Quality Control Department. Test personnel plug their meters into the calibrated ports for frequent calibration at commonly used levels but are not able to tamper with the calibrated unit. This latter statement is a recurring theme in this kind of testing.

A related item is a Switchable Fixed Tuned Converter shown in Slide II. This unit, where used in conjunction with a noise figure meter, provides quick and foolproof measurement at a number of spot frequencies. Adequate filtering is provided to avoid spurious responses. Likewise, levels at spot frequencies can be quickly measured without turning or adding calibration factors. Finally, the converter provides an acceptable means of extending the frequency range of other equipment.

As a matter of interest, it is worthwhile to take a quick look at some other pieces of specialized test equipment. A Cross Modulation Test Set of modular construction (Slide III) where an attempt has been made to design a self-contained work station, which is occupied for some 16 hours a day. The Hum Modulation Test Set and the Lightning Test Set shown in Slide IV and Slide V illustrate units built to fill a need not met on the commercial test equipment market.

One important point to be made at this stage is that many items similar to those shown can be constructed in an equipped CATV laboratory, if they have reasonably ingenious personnel and (1) learn to seek out "circuit modules" available commercially and (2) modify commercially available instruments.

CATV DISTRIBUTION TEST SET

As was previously mentioned, the test set evolved from circuitry which was initially designed for production test and from a prototype built for preliminary evaluation. This prototype is pictured in Slide VI. General characteristics of the revised unit now being designed are shown in Figure 1 with data on the feasible tests included in Figure 2.

Block diagrams of the internal components of basic test set and the transmitter unit are shown in Figures 3 and 4.

APPLICATION

For use by technical personnel in balancing, aligning, trouble-shooting, monitoring, evaluating performance before serious deterioration of distribution.

PHYSICAL

Self-contained, MIL quality with regard to ruggedness - suit-case format for airline travel.

ELECTRICAL

"Secondary standard" type stability with wide calibrated operating temperature range (-20, +120° F). Calibration where possible controlled by transmitter at antenna site.

Sensitivity sufficient to look at cascading effects. Internal modular construction, for repair and calibration check.

GENERAL CHARACTERISTICS CATV DISTRIBUTION TEST SET

FIGURE 1

SIGNAL LEVELS

Typical selective voltmeter usage except that no compensator knobs are needed. "Panoramic" operational mode to provide ready view of relative levels on six channels.

SIGNAL-TO-NOISE

An internal amplifier with 6 dB noise figure to make feasible measurements at low signal levels.

IM PRODUCT & SPURIOUS SIGNALS

Capable of measuring -60 dB under CATV conditions.

HUM MODULATION

At least -60 dB (.1%).

CROSS MODULATION

A relative measurement which can be related to -60 dB in a multiple channel system.

SOME POSSIBLE TESTS USING CATV DISTRIBUTION TEST SET

FIGURE 2

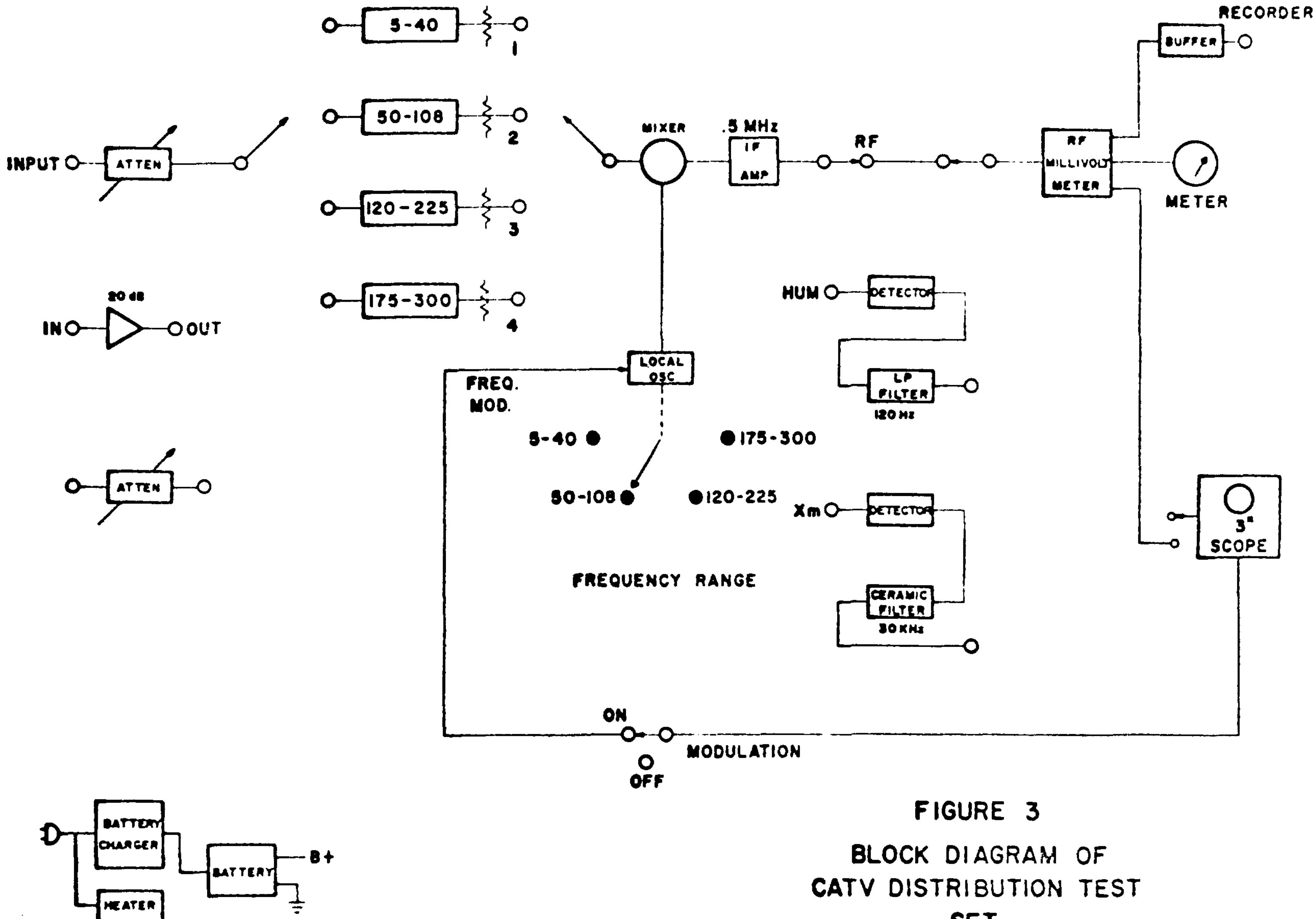

FIGURE 3

BLOCK DIAGRAM OF
CATV DISTRIBUTION TEST
SET

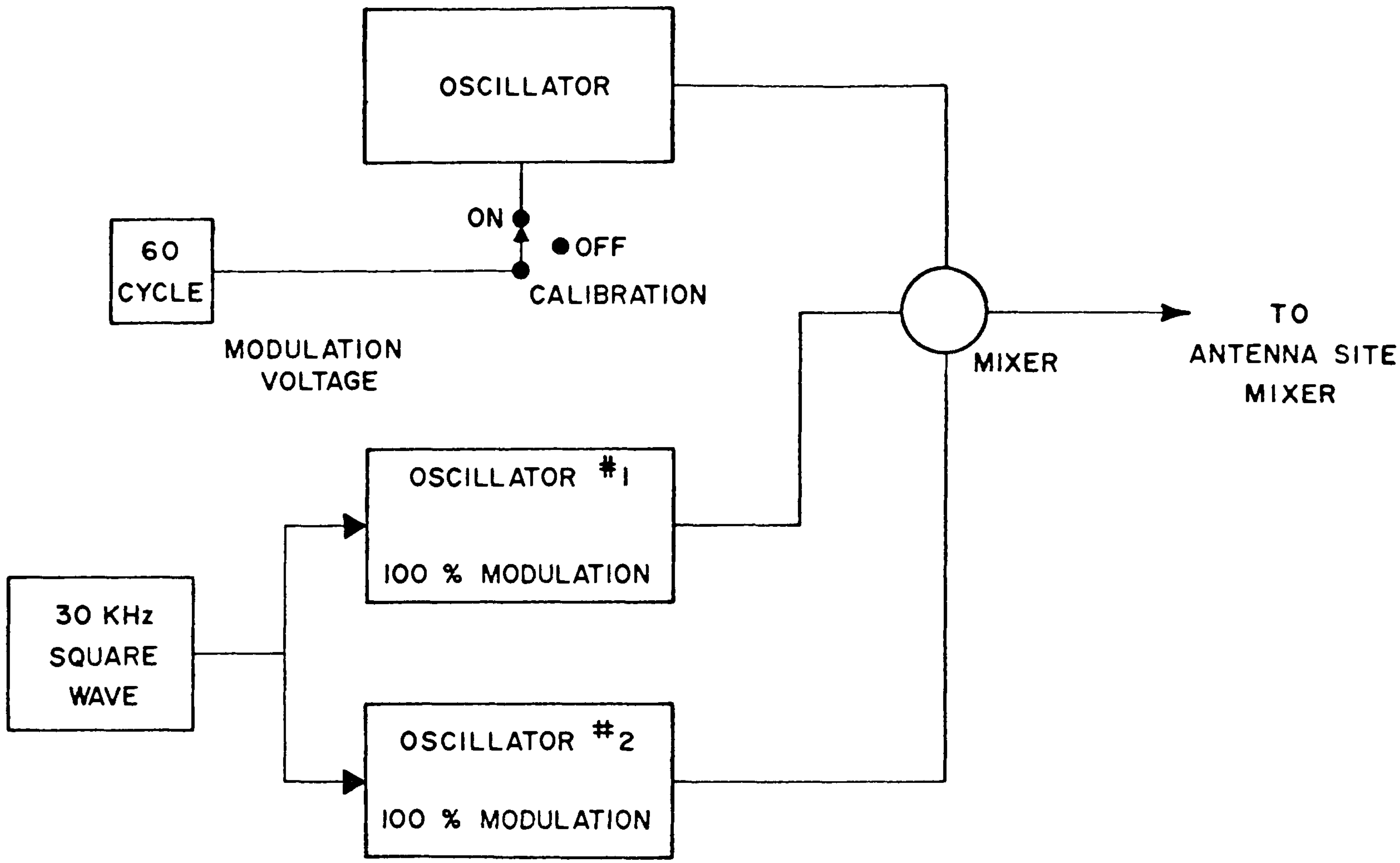

FIGURE 4

TRANSMITTER FOR
CATV DISTRIBUTION TEST
SET

SPECTRUM ANALYZER APPLICATIONS IN CABLE TELEVISION

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Spectrum analyzers are bound to find many applications in cable television systems because of their close family relationship to the familiar signal level meter. Signal level meters are heterodyne receivers tunable manually and are usually single conversion receivers, tuned manually, fixed IF bandwidth with a moving coil meter indication of input signal level. Spectrum analyzers are also heterodyne type receivers but have electrically swept local oscillator(s) and usually display the input signals on an oscilloscope type display signal amplitude versus frequency. The more sophisticated versions have logarithmic display over a range of 70 db or more, selectable bandwidths, and very stable sweep and tuning characteristics.

Our company has been using spectrum analyzers for more than four years, principally for observation of signal levels in various parts of the system and for finding and eliminating many sources of spurious signals in cable television systems. Our early spectrum analyzers had 5 KHz bandwidth, tuning range of 1 to 300 MHz in a single display and a 50 db dynamic signal display range. About six months ago we acquired a more sophisticated spectrum analyzer system consisting of a Hewlett-Packard 141T display frame with 8553B RF section, 8552B IF section, 8443A tracking generator/counter, and an 8554L RF section. The 8553B tunes 1 KHz to 110 MHz (in a single sweep if desired) and has a range of IF bandwidths from 300 KHz down to 10 Hz. The associated IF section provides 70 db display (10 db/division) or 16 db of display in a 2 db/division mode. Linear display is also available. The 8554L RF section tunes 1 to 200 MHz but is limited to 300 Hz bandwidth by the less stable oscillators in this RF section. The 8443A which operates with the 8553B RF section adds a tracking oscillator which acts as a sweep generator which frequency tracks with the associated receiver. A built in frequency counter operates with a "marker" on the display to permit 8 digit reading of any desired point on the display. The counter provides 10 Hz resolving power and has an internal clock specified to 3 parts in 10^8 accuracy. Detailed specifications are available from the manufacturer.

TRACKING GENERATOR APPLICATIONS

The tracking generator works only with the 8553B RF head and is consequently limited to a 110 MHz range. We use heterodyne conversion techniques to extend the range to higher frequencies but this can be done only over a narrow frequency range, e.g. one or two TV channels, because of flatness problems in the accessory mixers. Figure 1 is a simplified block diagram of the spectrum analyzer and tracking generator/counter. This combination makes an extremely useful sweep generator instrument. Spurious frequency components in the sweep signal output are ignored by the selectivity of the receiver section. The 70 db dynamic range in the display makes it possible to observe response over a wide amplitude range. This is particularly useful in working with channel band pass filters and with head end processing equipment. The associated counter permits easy direct digital reading of frequency at any point on the display. Dispersion calibration is reliable and this also assists in the examination of cable TV equipment.

continued...

Frequency range of the tracking generator/spectrum analyzer combination can be extended by use of a double heterodyne technique similar to that used in cable TV head end signal processing equipment. A local oscillator of the desired frequency drives two mixers, one acting as an up converter for the tracking generator signal and the other as a down converter to bring the signal back to the tuning range of the spectrum analyzer. The selectivity of the spectrum analyzer looks after image problems. A pad in the input to the splitter assists the hybrid splitter in providing isolation between the two mixers. A combination of good mixer isolation characteristics and good isolation in the splitter hybrid provides adequate overall isolation. Local oscillator drive must be adequate to make up for pad and splitter loss. The local oscillator stability and frequency accuracy affect any frequency measurements made with counter in this system. A laboratory signal generator is adequate for most purposes. Its frequency can be measured separately. For many purposes, we use a special local oscillator which we developed to extend the range of the 8553B RF section to 220 MHz.

We have developed a special purpose heterodyne "block converter" for converting the 110-220 MHz band down to 0 - 110 MHz so that it can be tuned on the 8553B RF section, see Figure 3. The 1 MHz clock signal from the 8443A counter is multiplied up to 110 MHz and used as the local oscillator in a block converter. This 110 MHz local oscillator is also available for use in the double heterodyne range extension system discussed above. At the present stage of development of this converter, the local oscillator is not as clean as we would like for use with the narrowest dispersion and bandwidths available on the 8553B and we may introduce a phase lock loop system to generate a cleaner local oscillator locked to the 110th harmonic of the 1 MHz clock signal.

The double heterodyne system was modified slightly to permit use of cable system carriers as frequency markers in examining the characteristics of a "high performance" channel band pass filter for channel 11. The filter had been ordered to provide maximum rejection of a strong local channel 10 signal. See Figure 4. Cable system carriers were mixed into the input of the channel 11 filter being tested using a hybrid directional coupler. Rejection of channel 10 visual carrier is seen to be only 26 db and channel 10 sound carrier is only about 6 db down! The wide dynamic range visible in a single display makes this a very useful technique for the examination of filters and processing equipment.

An old single channel amplifier strip (channel 11) was selected for demonstrating the use of the tracking generator in checking head end processing equipment. This strip was retrieved from a pile of obsolete equipment. Figure 5 shows the frequency response as observed with the tracking generator and with a conventional sweep generator and broad band detector. The strip is obviously very badly aligned. This is obvious even from the conventional sweep display in 5C. Figure 6 shows results after a preliminary re-alignment.

It is possible to check the alignment of processing equipment by injecting the tracking oscillator signal at reduced level through a directional coupler at the input of the processing unit and recovering it for the spectrum analyzer at the output of the processor. This technique is demonstrated in Figure 7. Generator level has been reduced to about 40 db below visual carrier so as not to interfere with the programme. Beat interference seems barely perceptible at this level. In Figure 7B

continued...

the analyzer gain has been raised and bandwidth reduced slightly to present a greater range of the processor response curve. Response is same as that observed without signals present. Any significant anomalies in processor frequency response can be detected in this way since dispersion range and centre frequency can be changed at will.

The tracking generator range is 10 KHz to 110 MHz and it makes a very good video sweep generator for checking the response of video equipment. It can be used to check the frequency response characteristics of modulators, particularly if a sample of the unmodulated RF carrier is available for use in a product detector system. See block diagram, Figure 8. We have also obtained usable results with the use of an independent local oscillator instead of the modulator carrier, particularly with higher modulating frequencies.

Figures 8B and 8C show the modulating frequency response of an IF modulator used in many of our cable systems. The unmodulated carrier was not available in this case and a signal generator operating close to the carrier frequency was used as the local oscillator. Response is fairly flat to just past 5 MHz.

The tracking generator and spectrum analyzer have been used as the sweep generator and tracking receiver for swept displays of group delay in RF equipment. The basic block diagram is shown in Figure 9. The group delay test set operated on the 20 KHz modulating frequency and group delay was displayed on another oscilloscope. The spectrum analyzer display showed the amplitude response of the system and provided frequency marker information.

SPECTRUM ANALYZER APPLICATIONS

The spectrum analyzer is convenient for investigation of system operating levels and has been used for that purpose by a number of systems for some time. The 8554L head is convenient for this purpose since it will display up to 1200 MHz in a single display. Low and high band may be displayed simultaneously. The 1200 MHz tuning range makes it convenient for use with UHF signals. The 300 Hz IF bandwidth and stability are adequate for most "general purpose" system applications. Co-channel interference can be easily recognized and measured with 8554L RF section. Most intermodulation products can be distinguished with this RF head, except those occurring at "near zero beat". Utility of the spectrum analyzer for distortion product measurement can be increased by use of a set of channel band pass filters as external preselectors or preferably a tunable band pass filter. These external RF preselectors reduce the risk of generating unwanted distortion products in the spectrum analyzer first-mixer.

The special bandwidth and stability characteristics of the 8553B RF section and associated 8552B add some additional applications to the cable TV engineer's bag of tricks. The 10 Hz IF bandwidth, 20 Hz/division dispersion and associated frequency counter make it possible to use the instrument for precise determination of carrier frequencies (to about 10 Hz precision). Low level spurious signals can be observed and "counted". Frequency determination helps considerably in determining the source of a spurious carrier. Many of the characteristics of the 8553B section can be used in the 110 - 220 MHz band by the use of the precision heterodyne converter previously described.

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Figure 10 shows a series of steps in the analysis of a co-channel interference problem on channel 2. Figure 10A shows the entire channel with visual and aural carriers and chroma information. Channel 3 carrier is also visible. In 10B the co-channel components are visible. The channel 2 carrier here operates with zero offset. Co-channels are visible with +and - offsets. The - offset is the stronger and is only 36 db below carrier. In 10C we examine the - offset co-channel more closely. It appears as a single carrier about 10 KHz below the carrier. The component about 6 KHz below the carrier is believed to be an intermodulation product. These observations were made after about 25 amplifiers in Cascade, but are valid with respect to the co-channel component. In 10D dispersion has been narrowed and bandwidth reduced to 30 Hz. Gain has been increased by 20 db. Three separate co-channel carriers are now apparent. In 10E we go to maximum resolving power and show the three co-channel interfering carriers. Their relative level has changed since the previous observation (interval about 5 minutes). Their relative frequency has also changed due to slight drift in the individual transmitters. The centre frequency in 10E has been marked as 55.24003 MHz. The individual carrier frequencies can be interpolated from the 50 Hz/division calibrated dispersion. The drift of the individual interfering carriers can be easily observed on the spectrum analyzer because of the exceptional stability of the analyzer. The analyzer drift can be noted by reference to the associated frequency counter. When used in our laboratory, we reference the counter to a Loran C derived frequency reference which is accurate to at least 1 part in 10^9. In field work, we rely on the crystal-in-oven in the counter which is specified to be about 3 in 10^8 after three hours warm up.

The 10 Hz bandwidth and exceptionally good shape factor permit using the spectrum analyzer to check hum modulation by observing and measuring the 60 Hz and 120 Hz modulation sidebands of the affected RF carrier. This is demonstrated in Figure 11. For demonstration, we used a small 53.25 MHz crystal oscillator energized by a battery. Figure 11A shows the spectrum of the bettery energized crystal oscillator. Figure 11B shows the spectrum after passing through a small MATV type amplifier. The 120 Hz sidebands (full wave power supply) are barely visible at about -70db. Figure 11C shows hum modulation in another small amplifier. Sidebands are about 60 db down. Hum modulation observations of this kind cannot be made on carriers with ordinary TV modulation. The hum modulation sidebands are completely obscured by the sidebands caused by the vertical sync pulse information.

The spectrum analyzer was used to analyze the characteristics of the 110 MHz local oscillator developed for the 110-220 MHz block converter. Figure 12 shows a succession of bservations on this oscillator. Figure 12A some spurious 10 MHz and 1 MHz components. The 10 MHz sidebands are about 70 db down. The 1 MHz components are about 65 db down. Figure 12B shows the 1 MHz sidebands more clearly. A significant noise component in the main carrier is now apparent. It appears to have the shape of the 110 MHz bandpass filter used to separate the desired 110 MHz harmonic component. Figures 12C and 12D show the noise more clearly. This noise level and other spurious products will have to be reduced below 60 db before the local oscillator is satisfactory for this purpose.

A spectrum analyzer is a convenient instrument for calibration of FM deviation using a carrier null technique. Frequency modulation by a sinusoidal modulating frequency can be analyzed in terms of Bessel functions which indicate periodic nulls of the carrier as deviation is increased. The second Bessel carrier null occurs for a modulation index of 5.52007 and a simple calculation indicates that a modulating frequency of 4.53 KHz and deviation of 25 KHz will produce the second carrier null in the case of standard TV aural carrier modulation. To set the FM

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modulation level on a cable TV modulator, a modulating frequency of 4.53 KHz is used and the modulation level increased until the second null is observed on the spectrum analyzer. The nulls are easily seen as the modulation level is increased. A VU or voltmeter can then be used to indicate proper audio level for 100% FM modulation. Figure 13A shows an unmodulated aural carrier. This is not a very good modulator as the unmodulated carrier should show less frequency deviation than is apparent here. In Figure 13B, FM at 4.53 KHz has been applied and modulation level adjusted to the second carrier null. Depth of mull is approximately 25 db. Figure 13C shows the same display at 10 KHz/division dispersion. Figure 13D shows the spectrum of the modulating signal (4.53 KHz). Number and amplitude of distortion products are immediately apparent. Principal distortion product is the third harmonic and it is about 20 db below carrier. The spectrum analyzer can be used for analyzing audio signals above 1 KHz. Note that the occupied bandwidth (to the -40 db points) is about 80 KHz. This is somewhat more than observed on broadcast station carriers. We suspect that most broadcast stations undermodulate somewhat.

The spectrum analyzer can be used to investigate the characteristics of a television modulator by direct observation of the RF spectrum when modulated by sinusoidal modulating signals of controlable frequency. The series of photographs in Figure 14 shows RF spectrum of an IF modulator, of a type often used in cable systems, when modulated by a good quality sinusoidal oscillator. Modulation was set to 80% at 1 MHz video by observing the RF modulation envelope. Modulating Signal was then reduced by 20 db to reduce the harmonic distortion in the modulator. The video oscillator was swept manually over the desired range while the spectrum analyzer swept at quite a fast rate. The display was built up using the storage feature of the oscilloscope and was photographed using a time exposure while the video oscillator was manually swept. Figure 14A shows the spectrum as the video oscillator is tuned from 1 MHz to 10 MHz. This is an IF modulator and the vestigial sideband is the upper sideband. The up-converter will invert the sidebands and remove most of the "out-of-band" components. In Figure 14B, the gain has been increased to bring the sidebands to the top reference line. The vestigial sideband characteristic is not correct. It is not cutting off soon enough. Figure 14C and 14D show this vestigial sideband more clearly. Two photographs have been taken with different exposures to optimize different parts of the display.

A multiburst signal at normal level was used as the modulating signal and the resulting RF spectrum observed. The multiburst used has bursts at 0.5 MHz, 1.5 MHz, 2.0 MHz, 3.0 MHz, 3.6 MHz and 4.2 MHz. The spectrum at video frequency is shown in Figure 14A. The photograph should have shown the 0.5 MHz component at same level as the others. The resulting RF spectrum is shown in Figure 15B. Vestigial sideband is not adequate, as previously demonstrated. A full field multiburst like this should be a good signal for checking the frequency response from a broadcast transmitter right through a cable system.

In Figure 15C the low frequency modulation characteristics are being checked. Figure 15C is a multiple observation with successive modulating frequencies of 50, 100, 200, 300 and 400 Hz. The photograph was taken from the multiple observation on the storage tube. Low frequency sidebands are not symetrical which probably indicate some FM'ing of the carrier at these frequencies.

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Figure 15D was taken to show the resolving power of a good spectrum analyzer. The photograph shows interlace of chrominance and luminance sidebands at about 1 MHz. The signal was a video "saturated colour bar". This observation was made on video components about 2.6 MHz below colour subcarrier. This kind of observation is within the power of the 8554L head since only 5 KHz/division dispersion and 300 Hz bandwidth were used. The high resolution head permits study of the "fine structure" by using the 10 Hz bandwidth.

SUMMARY

A few special applications of the spectrum analyzer and tracking generator/counter have been illustrated. Additional applications abound and the usefulness of these instruments is limited only by the ingenuity of the engineer using them.

How It Works

Both spectrum analyzer and TG/C mate to form a signal-analysis and swept-frequency measurement system that embodies versatility and precision. Basically, the TG/C generates a signal that coincides in frequency with the spectrum analyzer's tuning—and then accurately measures this signal's frequency on an 8-digit counter.

The spectrum analyzer is a triple-conversion receiver with three local oscillators (LO). For wide scan widths, the first LO sweeps while the third LO tunes to a fixed frequency. The second LO, a crystal oscillator is always tuned to 150 MHz. For narrow-scan widths (20 kHz/cm and lower), the first LO phase-locks to a crystal-controlled reference (at 100-kHz intervals) and the third LO sweeps. With a 2-MHz bandwidth, the 200 and 50 MHz IF's are

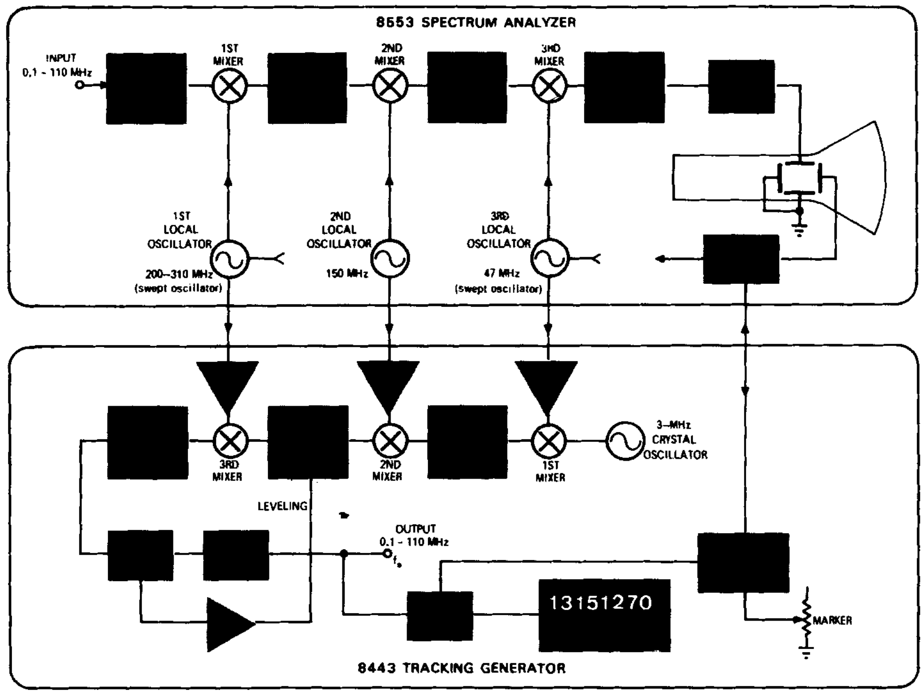


Figure 1 (courtesy Hewlett-Packard)

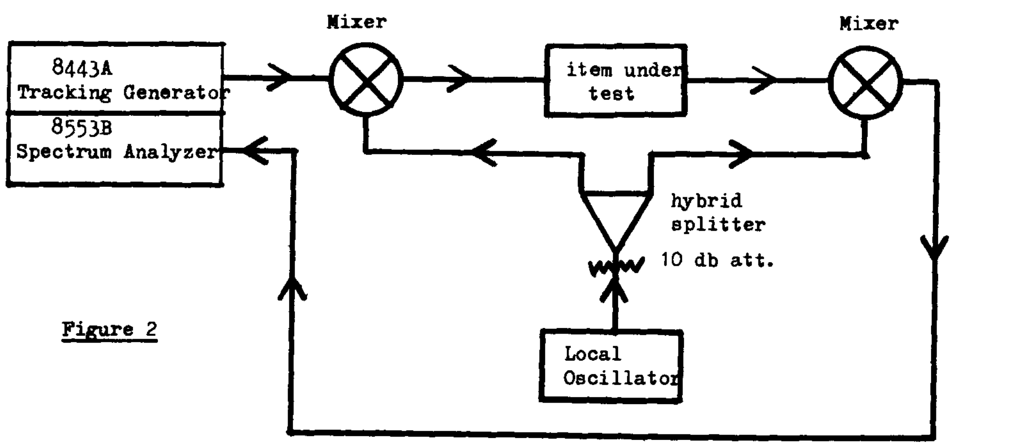


Figure 2

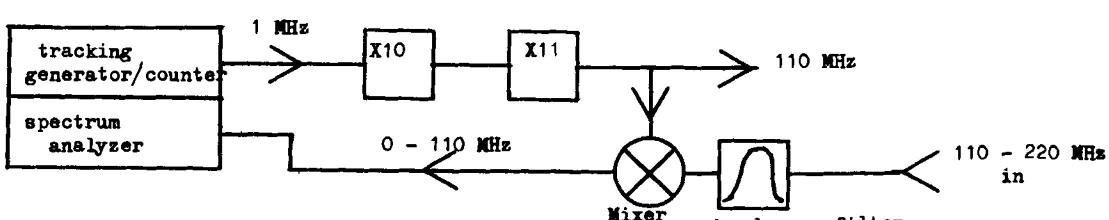
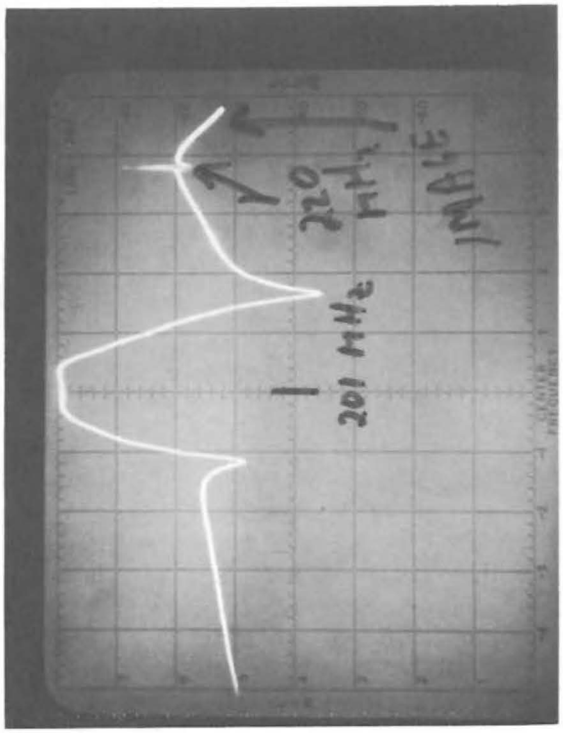


FIGURE 3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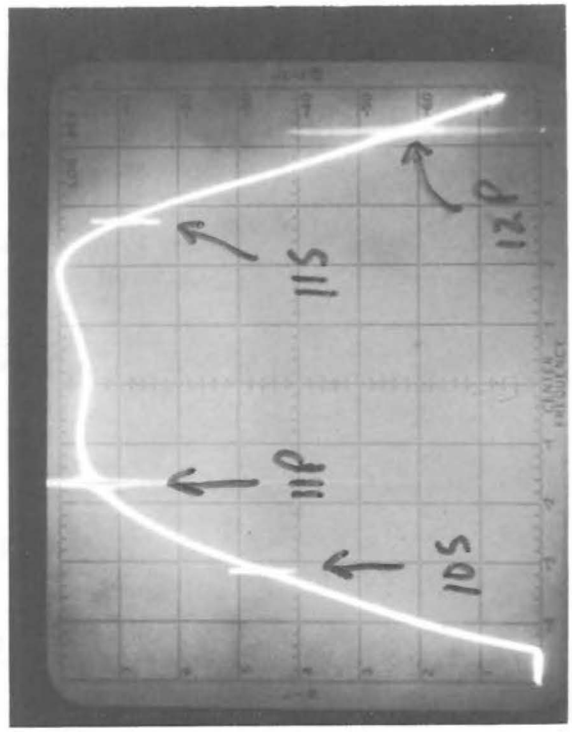
220 MHz
IMAGE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11S
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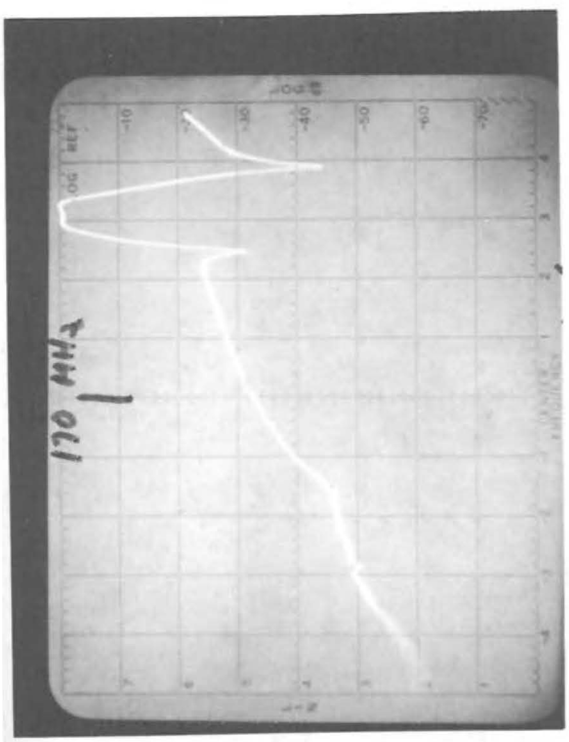
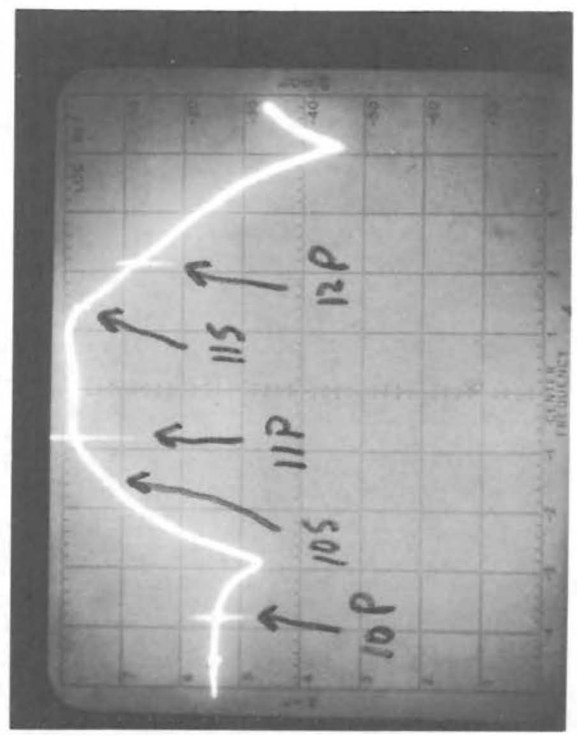

170 MHz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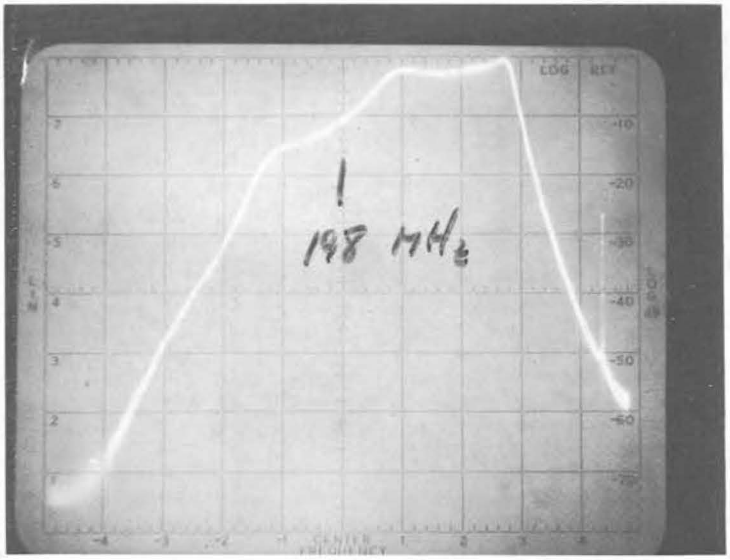
10S
11P
11S
12P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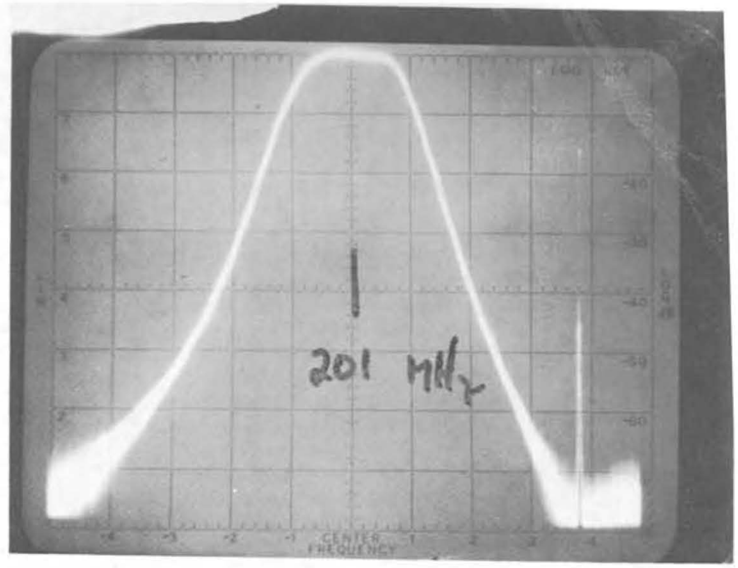
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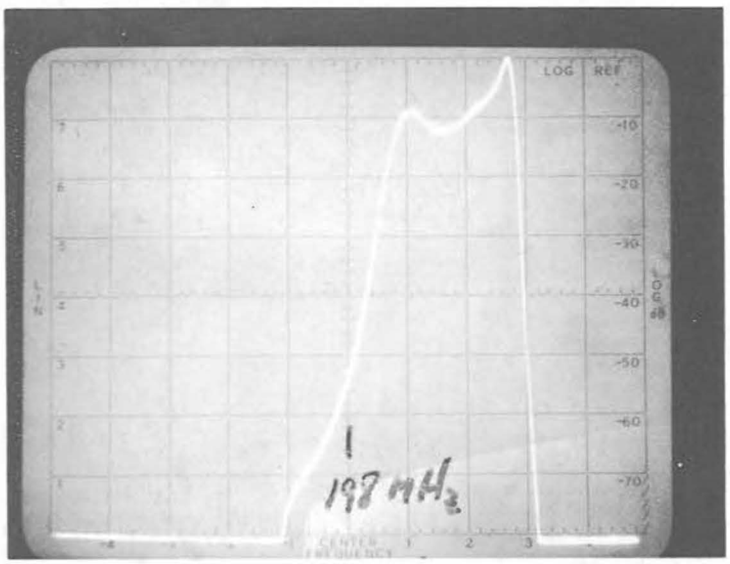
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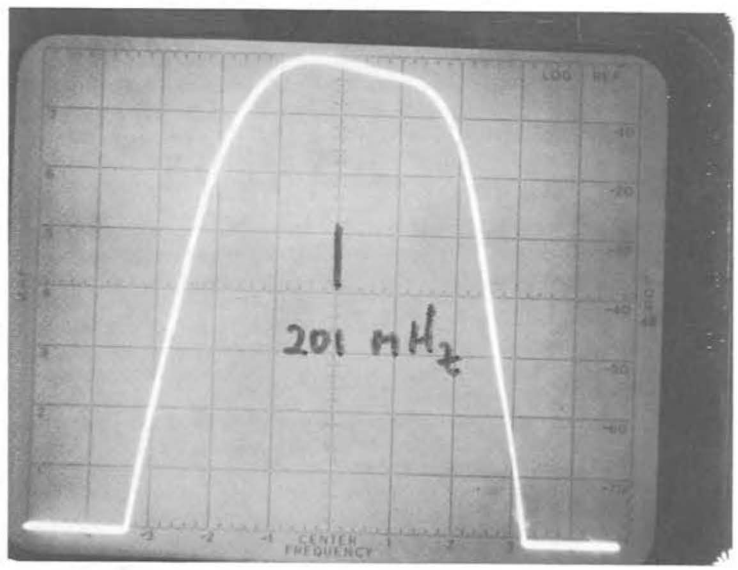
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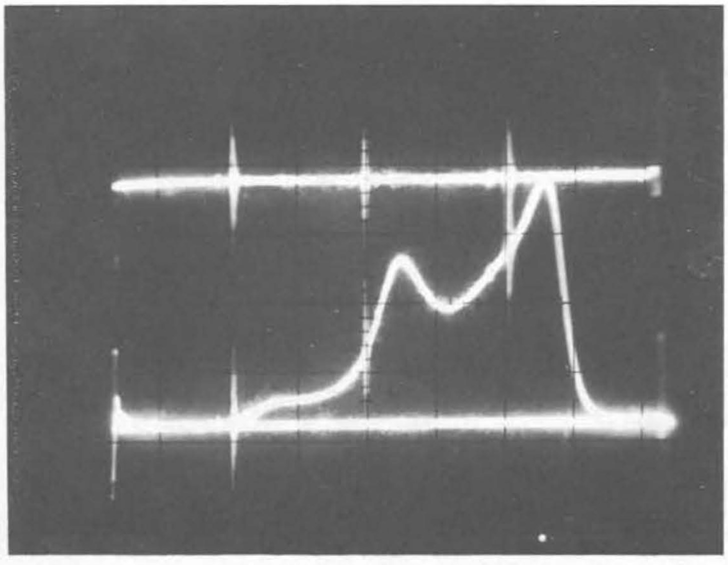

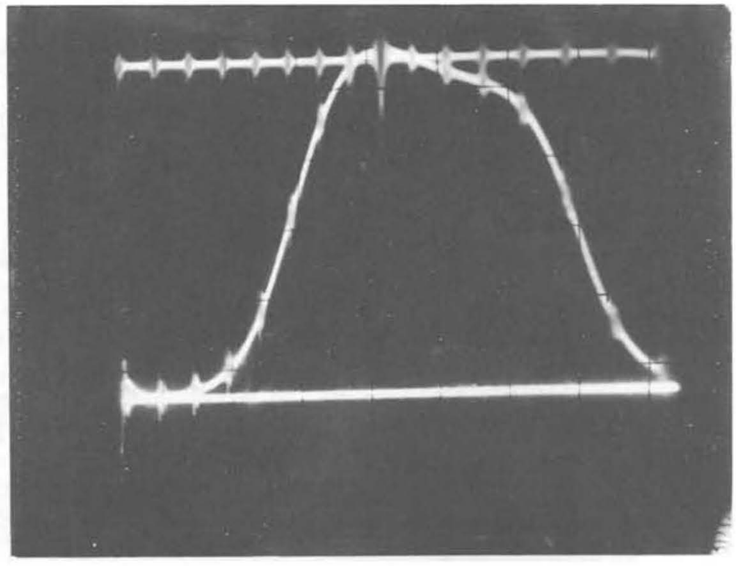

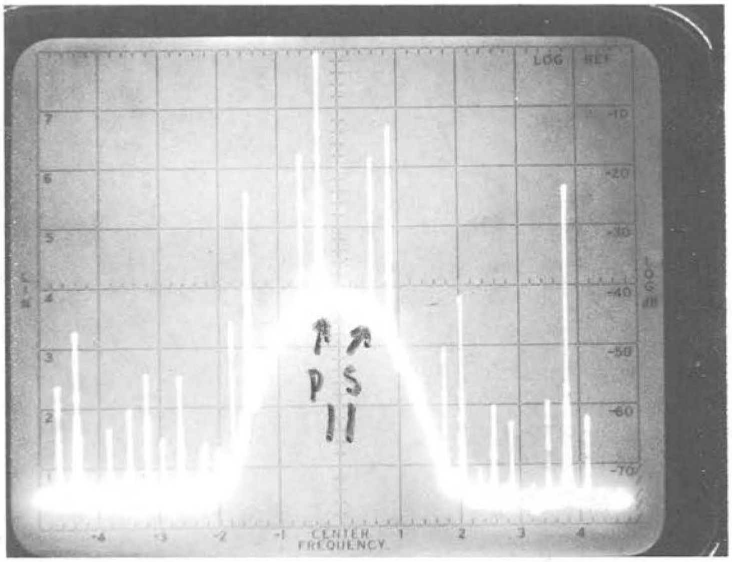
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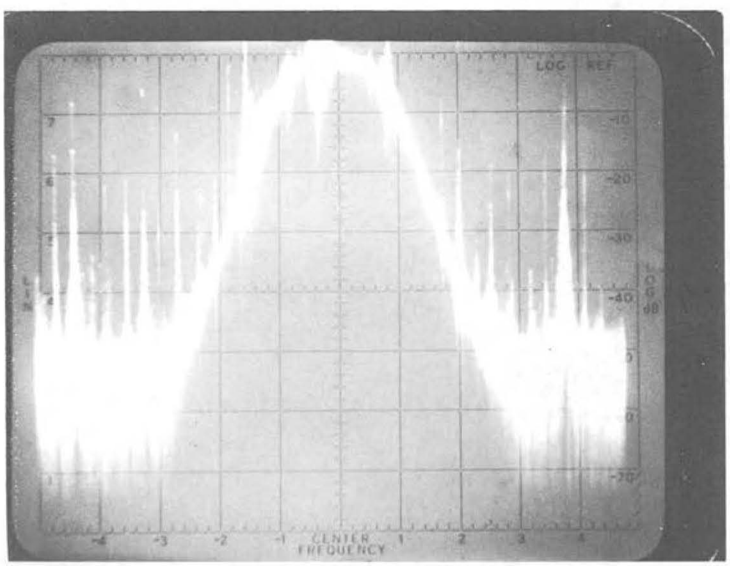
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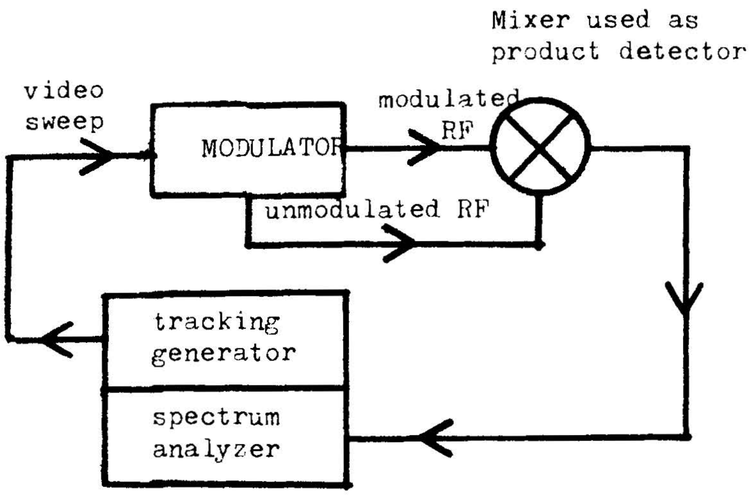
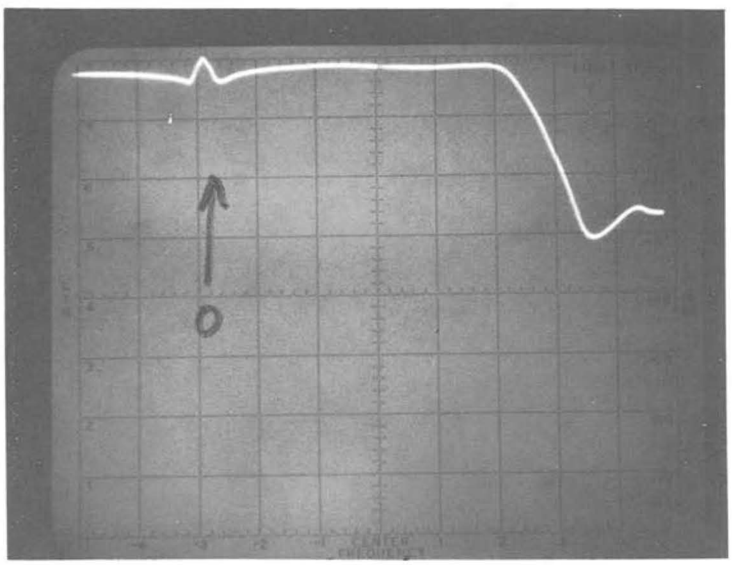

Mixer used as
product detector
video
sweep
MODULATOR
modulated
RF
unmodulated RF
tracking
generator
spectrum
analyzer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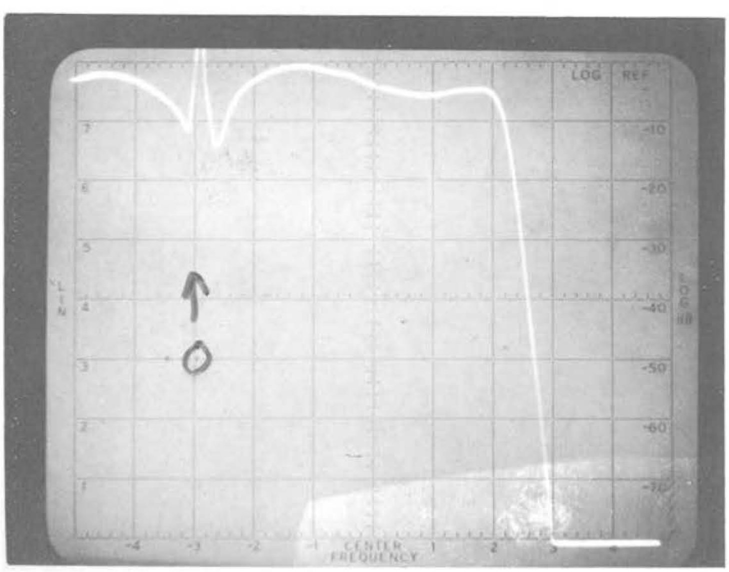
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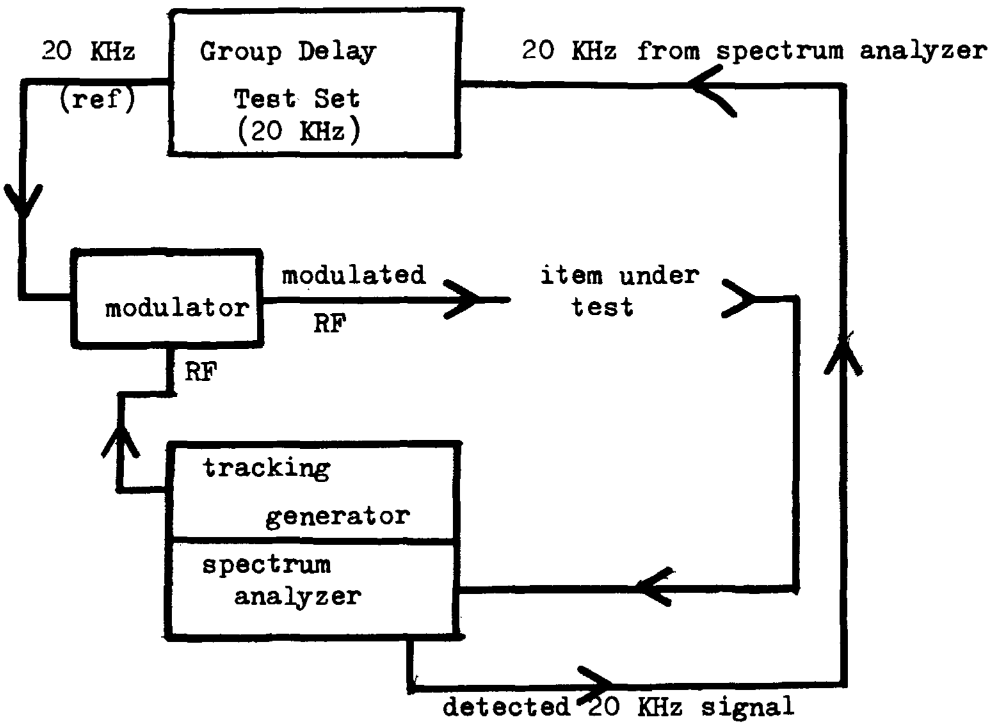
20 KHz
(ref)
Group Delay
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(20 KHz)
20 KHz from spectrum analyzer
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test
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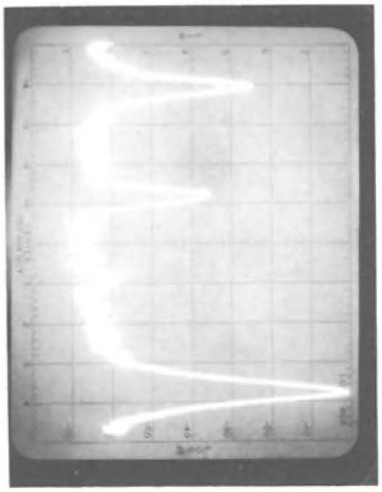
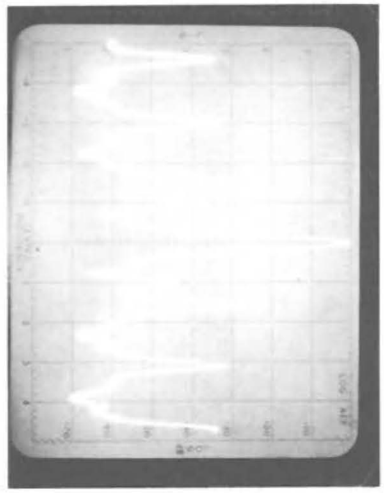
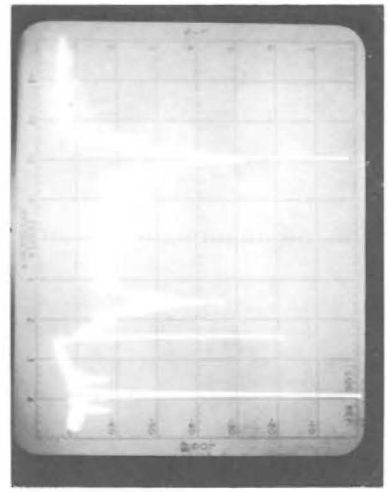
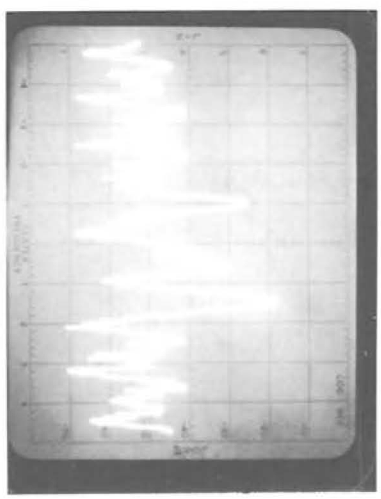
spectrum
analyzer
detected 20 KHz signal

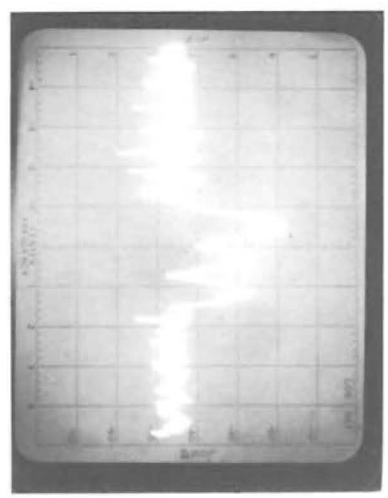
Figure 9

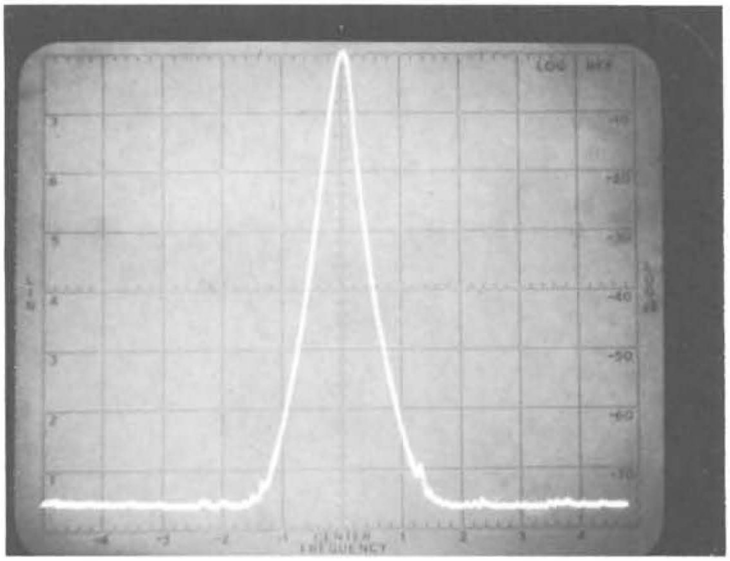








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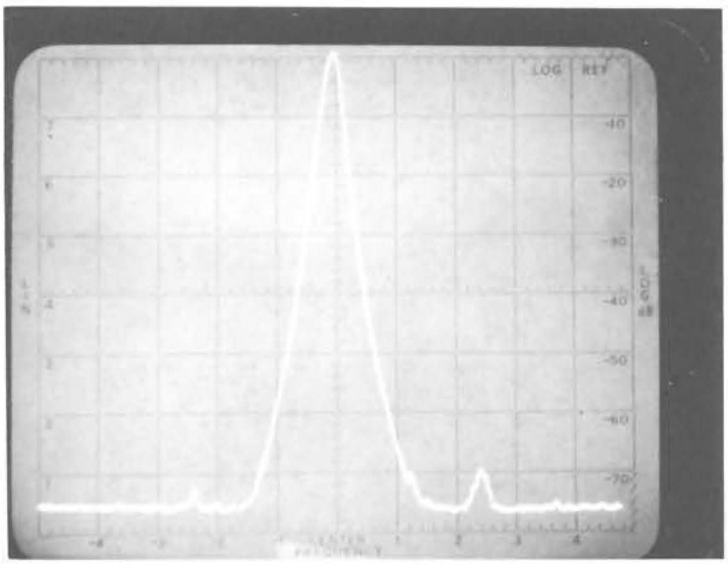
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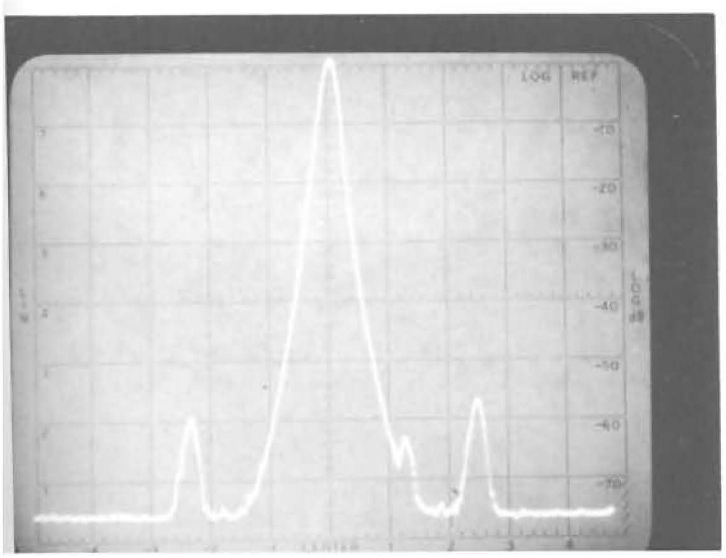
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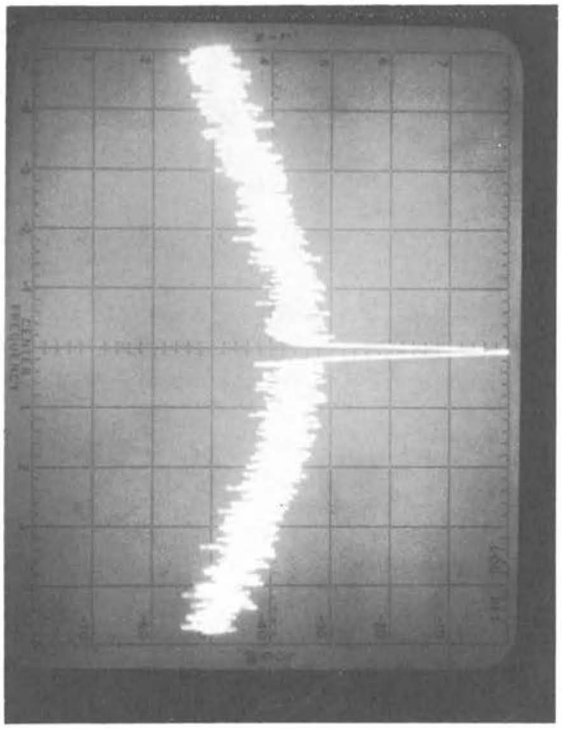

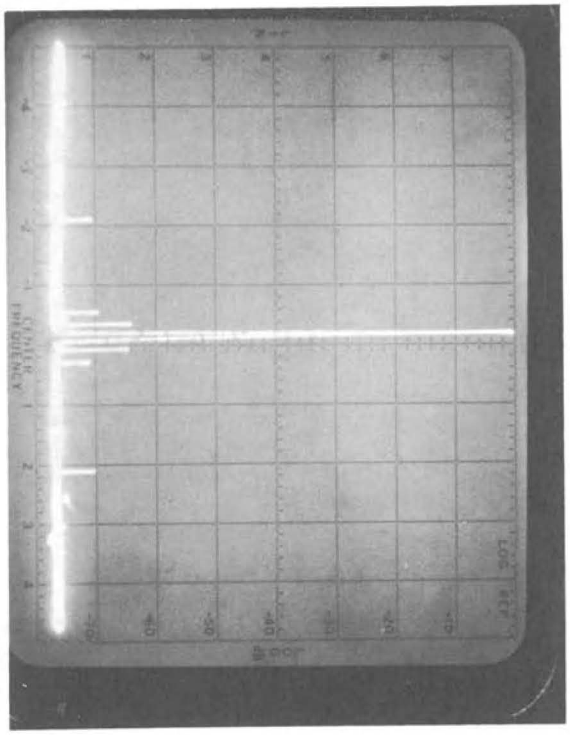

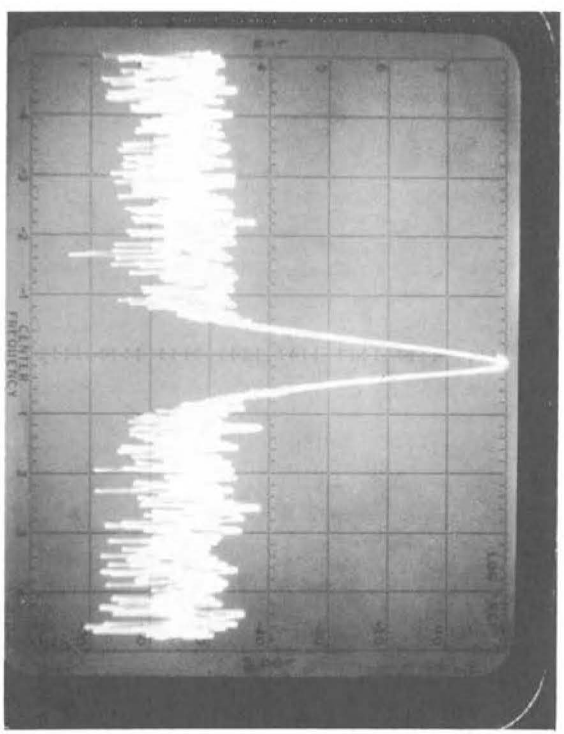

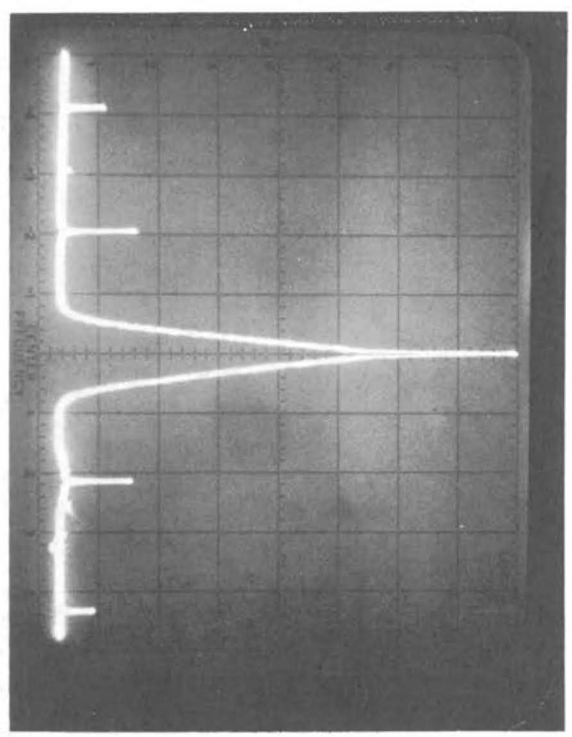

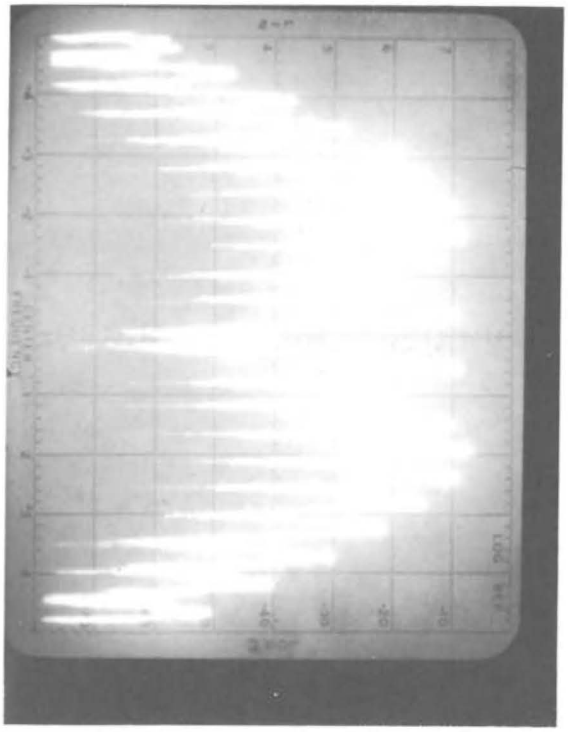

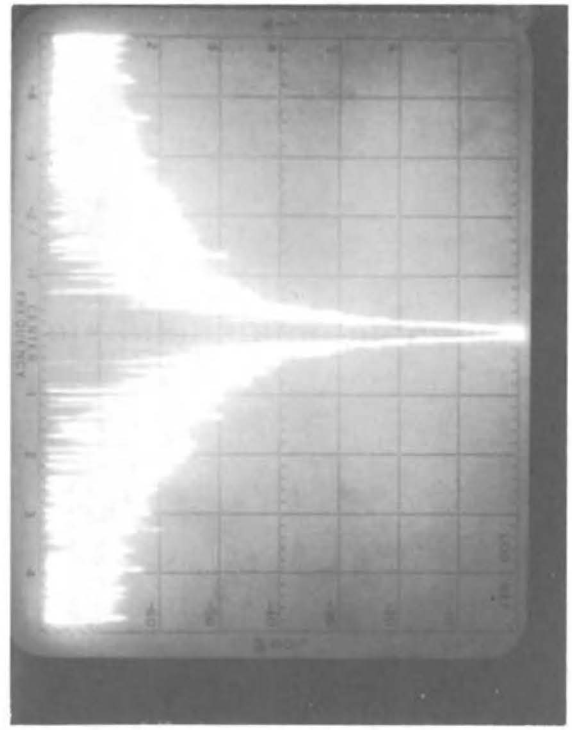

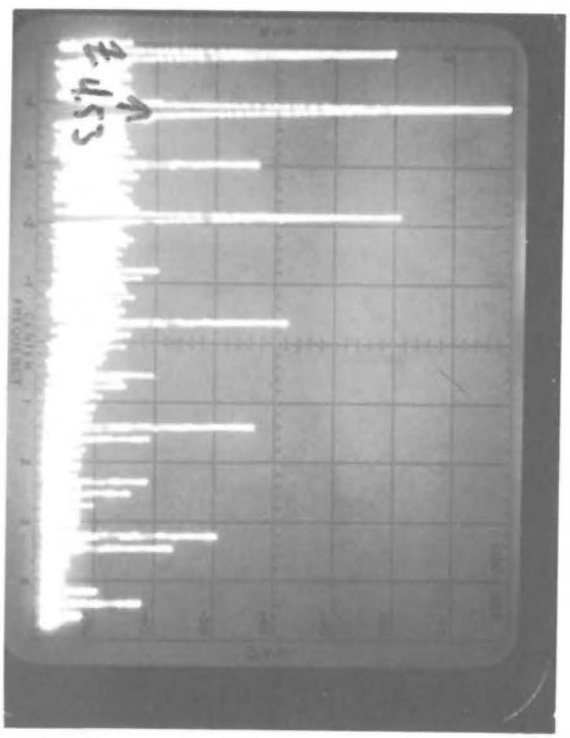
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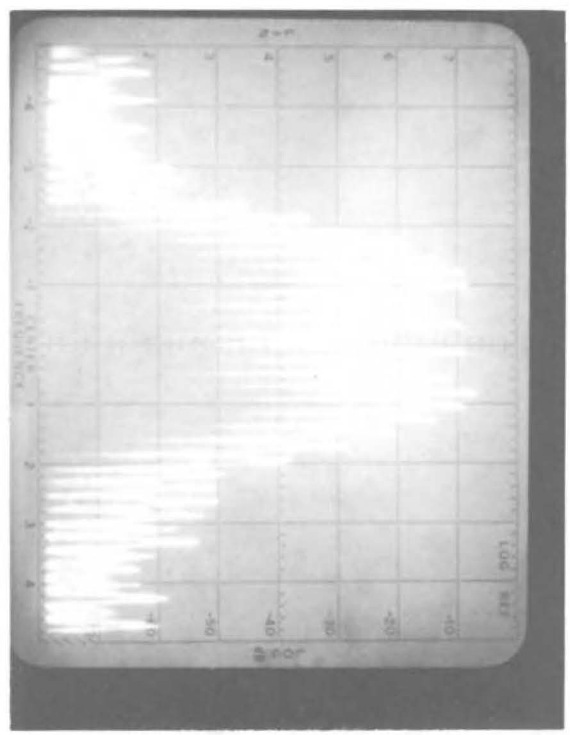
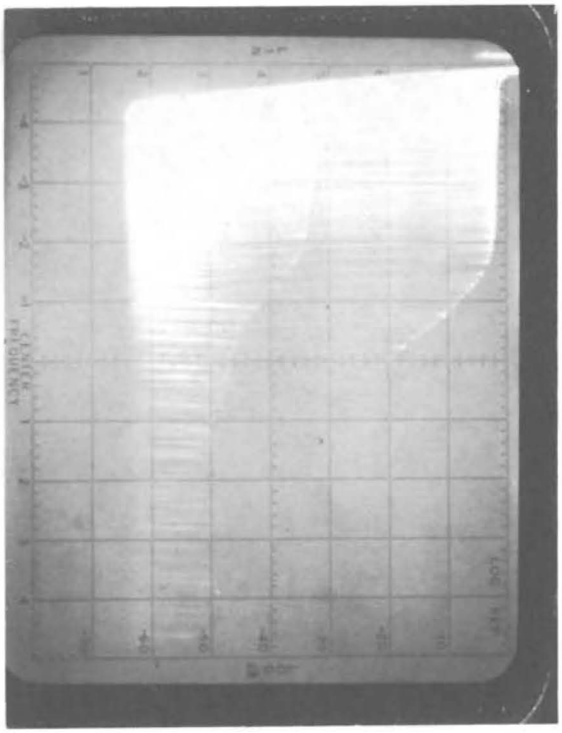
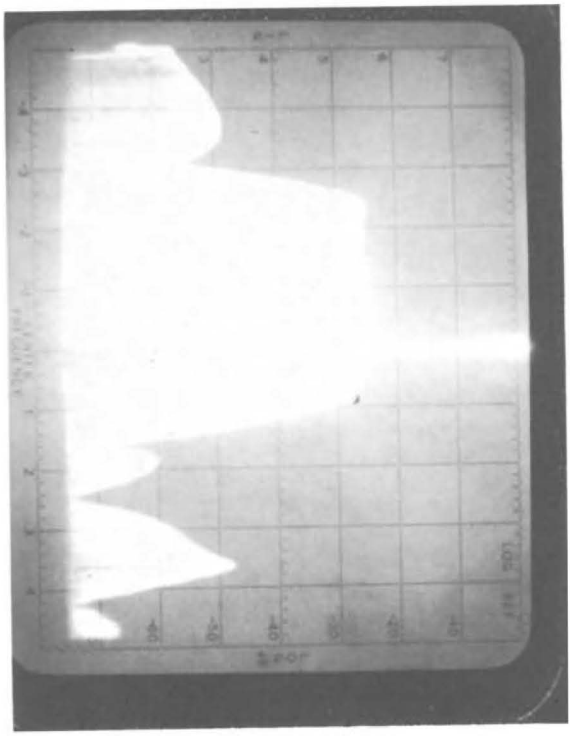
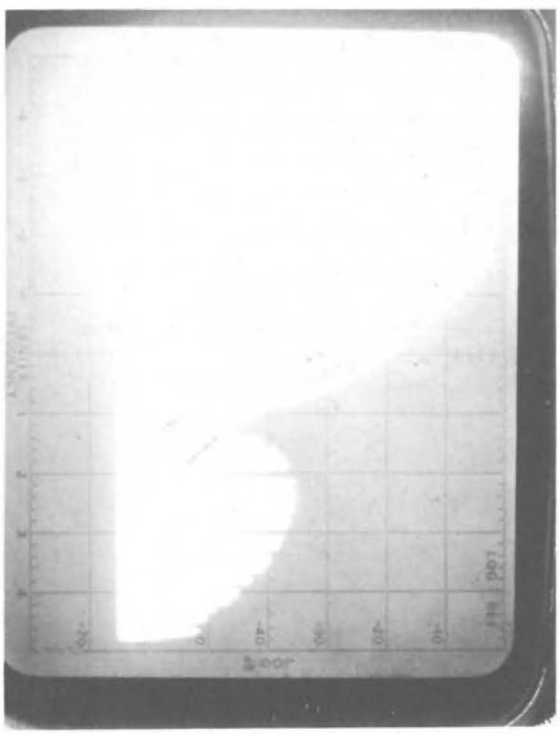
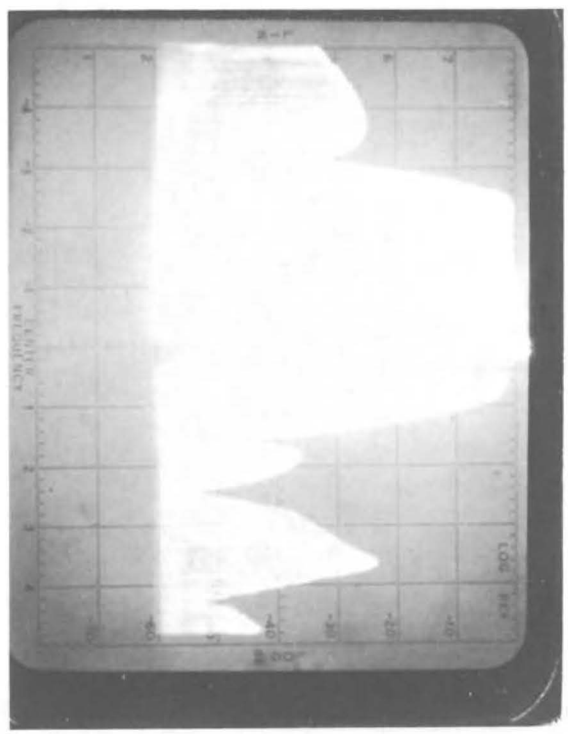
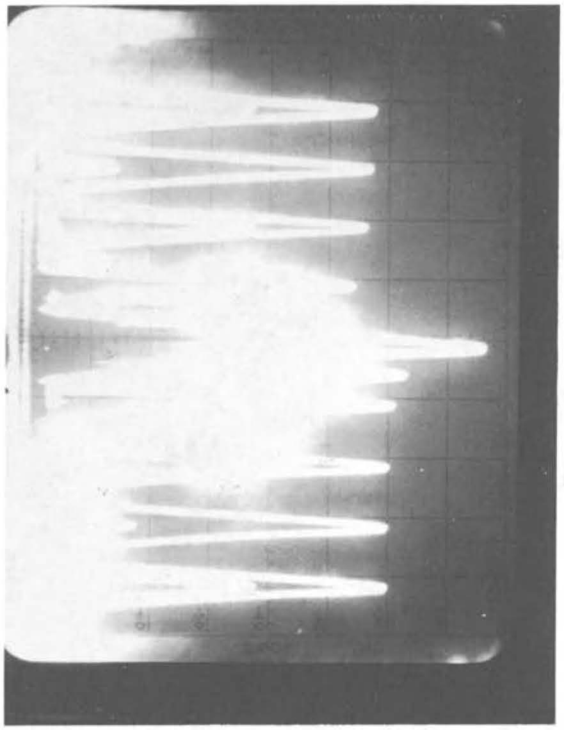


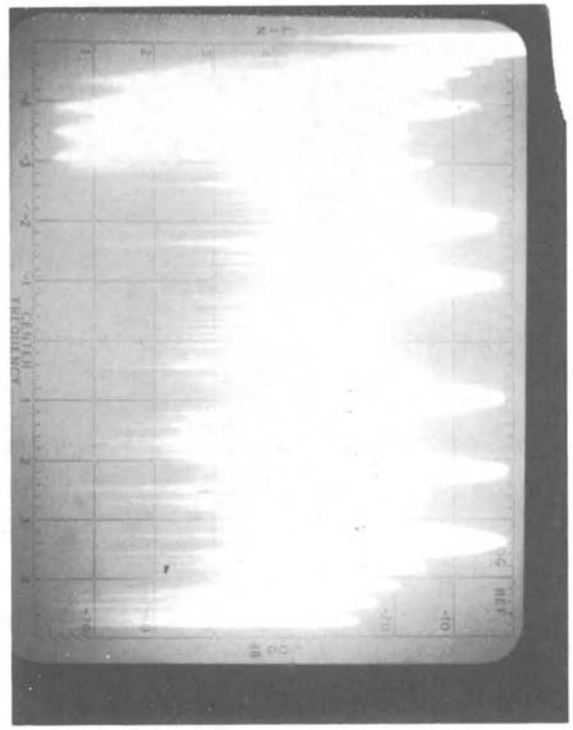


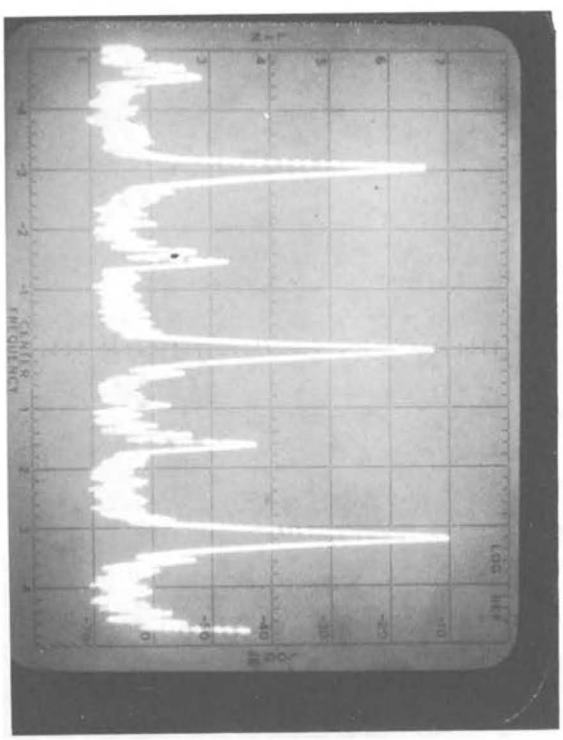


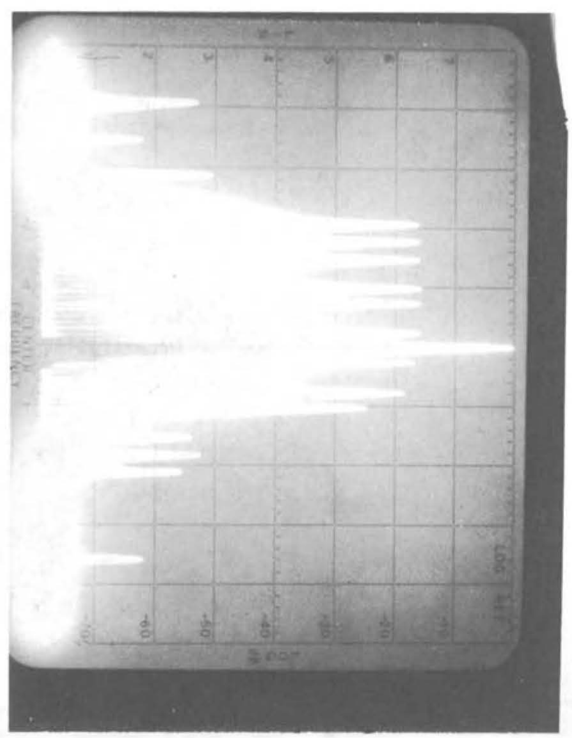












For presentation July 8, 1971,
NATIONAL CABLE TELEVISION ASSOCIATION,
Washington, D. C.

TELECINE SYSTEMS
FOR THE CATV ORIGINATION CENTER

by

Kenneth K. Kaylor

Philips Broadcast Equipment Corp.

Although "live" programming is considered a necessary part of the program origination services of a community-oriented CATV system, <u>telecine</u> facilities will hold the key to success or failure of such an operation from an economic standpoint. The word "tele-cine" was developed during the early days of television broadcasting to define those facilities devoted to the video reproduction of the various film media.

Since the original commercial telecine television camera was an "iconoscope" camera which had a very large sensitive surface (about 3" x 4"), a film projector was focused onto the sensor by using a standard projection lens as shown in Fig. 1. This technique was quite simple and optical alignment was very easy.

Eliminating the "Shutter Bar" effect

In the case of motion picture film, the theatre projector had to be modified in order to prevent a "shutter bar" effect caused by the difference in frame rates between television and motion picture standards. Standard sound motion picture film operates at 24 frames per second while the U. S. standard for television scanning is 30 frames per second. The intermittent mechanism had to be modified so that the length of time between "pull-downs" alternates between 1/20 and 1/30 second. The average of these two fractions is 1/24 second, or the time demanded by the standard 24-frame-per-second motion picture projection rate. When this particular intermittent mechanism is coupled with a five bladed shutter, the frame that remains for 1/20 second is scanned by three 1/60 second fields, while the one remaining for 1/30 second is scanned by two fields. It is not practical simply to speed the motion picture film speed up to 30 frames per second. Unnatural motion results, of course, and sound frequencies are distorted.

The introduction of new super-8mm sound film projectors with different frame rates further complicated the technical aspects of television film coverage. These commercial projectors also had to be specially modified for proper operation at commercial television standards.

Multi-projector systems

As the requirement for more and more film and slide origination developed at the television stations, it became apparent that techniques whereby more than one projector could be used with each camera were necessary. One such arrangement used at the WNBT film studio in New York (Figs. 2 and 3) was a track assembly whereby the camera could be rolled from port to port in the studio wall in order to pick up several slide and motion picture film projectors. A somewhat similar arrangement used at another early television station provided for the camera to be mounted on a pedestal with a panning mechanism. This camera could be turned by remote control to face a number of different projectors mounted on stands placed in a semi-circle. Both of these systems had the inherent disadvantage of a significant time lapse when changing from projector to projector.

With the introduction of the vidicon television sensor in later years, the image size of the sensor was reduced to 3/8" and 1/2" and optical alignment became much more difficult. "Uniplex" television cameras coupled to one projector were still used (Fig. 4); however, scan reversal was necessary when using the one-for-one system.

An economical method for using several projectors with one camera was needed. Furthermore, it was desired to reduce the time lapse required to alternate between projectors. These requirements were accomplished by using a prism or mirror "multiplexer" assembly which would direct the light from as many as three projectors directly onto the surface of the vidicon tube (Fig. 5). Mechanical "dousers" were inserted or removed from the light path of each projector to select the proper image.

Although this technique was relatively inexpensive, it was extremely difficult to align the images. Control of light levels had to be accomplished by use of "automatic target" controls or individual remotely controlled "neutral density disks" mounted in the light path from each projector.

A simple method to accomplish alignment of several projectors, as shown in this picture (Fig. 6), was to provide the camera with an objective lens which was focused upon the image plane developed at a "field lens" located between the projectors and the pick-up device. The use of this lens provided several advantages (Fig. 7). First and most important, optical alignment was considerably less complex. Secondly, the light could be controlled by inserting a neutral density disk in the camera optical path thus reducing the number of disks required. Light control could also be accomplished by using an automatic iris on the objective lens. In addition to douser operation with a prism multiplexer, selection of the proper image was accomplished by pneumatic insertion or removal of "first surface" mirrors as shown in Fig. 8.

Two-camera, multi-projector systems

With the introduction of the multiple surface mirror multiplexer head as shown in the left half of the illustration (Fig. 9), it became practical to use two cameras with three or four projectors as indicated in the layout on the right. A control system that would allow the image from any of the projectors to be directed to either camera was used. An intricate system of automatic controls prevented the image which was "on air" from being disturbed if a conflicting command was given for the alternate camera. Control of the light was accomplished by neutral density disks at the sources or located in the light path at each camera. Although these systems were quite versatile and economical for large television stations, it appears that the complexity is not required for most CATV installations.

Color telecine for cable TV

A typical color telecine film island for CATV use is shown in this "exploded view" (Fig. 10). Components of the system include a remote-controlled slide projector, a remote controlled super-8mm film projector, and a specially developed lightweight 16mm film projector. These units are mounted on a rigid framework and coupled through a mirror multiplexer system to a Norelco LDH-1 type Color Film Camera. Here (Fig. 11), light control is provided by an automatic iris on the objective lens coupled with automatic target control of the vidicon sensors. The use of a waveform monitor and a monochrome picture monitor is recommended if one desires to have the highest picture quality. A high quality system such as the one described will assure a noise-free picture which will meet or exceed the quality picked up from local stations or network originations and relayed to your customers. This system is compatible with either monochrome or color operation. The high quality color system shown in the diagram costs about $25,000.

Sometimes it is desirable from an economy standpoint to utilize one camera for both telecine applications and pick-up of cards, opaque objects or "live" action scenes. Two methods are reasonable when such cost reductions are deemed necessary.

One method is to use a live camera with a zoom lens that can be integrated with an optical multiplexer. A precision wedgeplate is coupled to the multiplexer in such a way as to allow the camera to be critically aligned to the field lens (Fig. 12). The camera can then be attached or detached at will. Since many of the new cameras, either monochrome or color, have zoom lenses with automatic iris mechanisms, the light control is still automatically corrected.

Rear-screen projection

The second method for utilization of a live camera as a film chain involves the use of the rear screen principle. This method reverts back to the early days of television film systems inasmuch as a single "live" camera can be used to pick up the images from several projectors as well as from cards, radar displays, meters, or "live" objects. You are all familiar with several of the "time n' weather" devices which form the simplest device in this category (Figs. 13 and 14).

One might expand this concept by using a large translucent screen and projecting several images from various projectors onto its image surface from a rear position. These images would be picked up by a camera located in front of the screen. Either one large screen can be used or several small ones with individual projectors as shown in Fig. 15. The advent of a special new type of rear projection screen which reduces "hot spots" and defocusing problems is the heart of the system. This screen allows a standard vidicon or Plumbicon* TV camera to be used for viewing images from practically any existing audio-visual projector and still be used to view a "live" scene (Fig. 16).

For this sample system, we have chosen to use four projection sources, a card rack and a live "video disk jockey" set (Fig. 17). We have provided for a "stop-motion" 16mm sound projector, a filmstrip projector, a 35mm projector and a television-modified super-8mm projector. Thus, one could accommodate almost any type of locally originated material as well as a vast storehouse of educational materials and "Freebies." In this case, two cameras are mounted on a pedestal (Fig. 18) located at the center of focus of a zoom lens that is provided with a close-up adapter. All pick-up points are located on an arc centered on the pedestal. Each projection console (Fig. 19) forms a chord of the circle. The card rack (Fig. 20) also forms an equal chord. Economy is the "name of the game" in the rear screen system. It is designed for minimum operator costs in the "disk jockey" format. All "start," "stop," and "change" controls for each projector are located in a convenient place on the operational console (Fig. 21).

Since the camera is on a pan/tilt device and employs a zoom lens, one can modify materials brought to the station by amateur photographers. That is, if slides are poorly framed, they can be edited by changing size and framing position simply by focusing on the desired portion of the projected image as illustrated here (Fig. 22). This, of course, is a capability unique to the rear screen system.

All other monitoring, switching, and camera control functions are located in this same desk so that your local "disk jockey-engineer" can have everything at his finger tips.

Titles

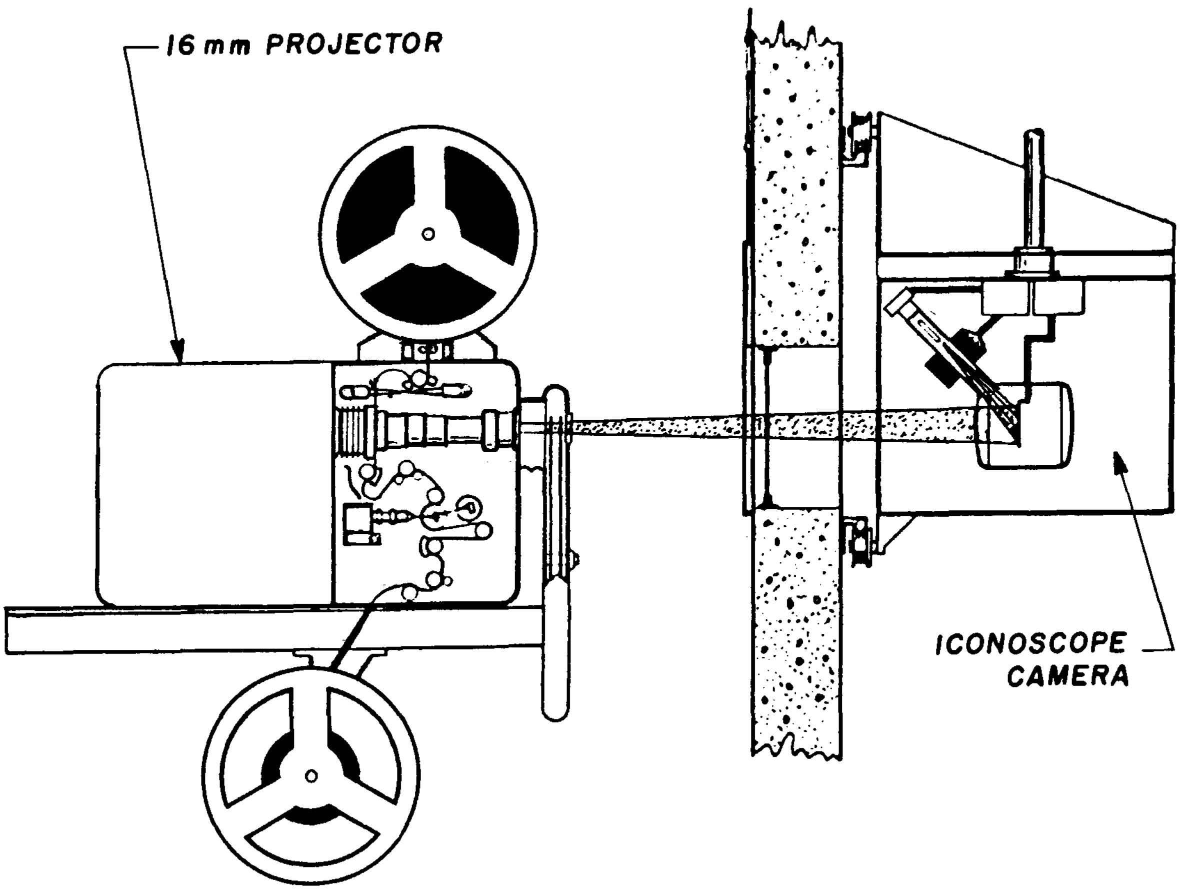
Several methods for title reproduction have been developed. In addition to the card rack previously discussed, both vertical and horizontal "crawl" devices have been fabricated from moving belt systems as shown in Fig. 23, or the rotating wheel principle shown in Fig. 24. The ability for a cable operator to use a standard typewriter, "rub-off" lettering, "paste-ups" or Polaroid* pictures is a real time and money saver for quickie commercials or promotional slides. A simple "document viewer" (Fig. 25) which again uses a zoom lens and appropriate lighting is ideal for this service. A "positive/negative" phase reversal switch on the camera control provides for image polarity to be reversed, thus allowing black letters to be reproduced in white. Electric zoom allows for size control, while "scan reversal" switches on the camera provide for special effects when desired. Either monochrome or color cameras can be used with this system.

"The telecine production center"--that's really where the action is in the money-making end of cable television origination. I hope that this review of the evolution of the telecine system and its variations, along with some of the tricks of the trade, will help each of you bring greater flexibility and imagination to your own plant to provide new versatility without mortgaging the ranch. If you want any help, give the boys in the Norelco back room a call. They might have just the solution for your problems!

# # #

Norelco
PHILIPS BROADCAST EQUIPMENT CORP.
A NORTH AMERICAN PHILIPS COMPANY
ONE PHILIPS PARKWAY, MONTVALE, N. J. 07645
TEL: 201-391-1000 TWX: 710-990-7301


TELECINE SYSTEMS
for the CATV
ORIGINATION CENTER



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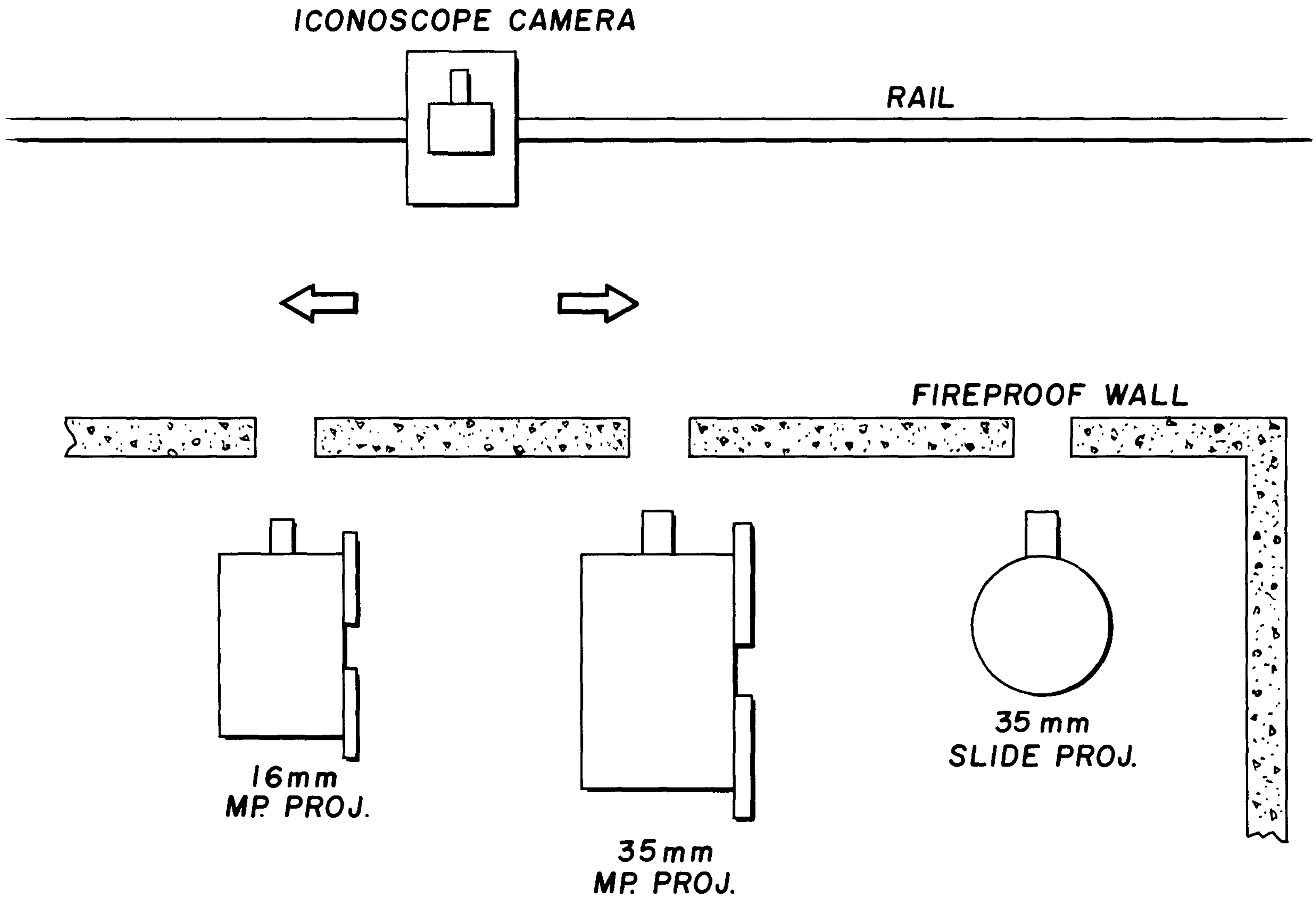
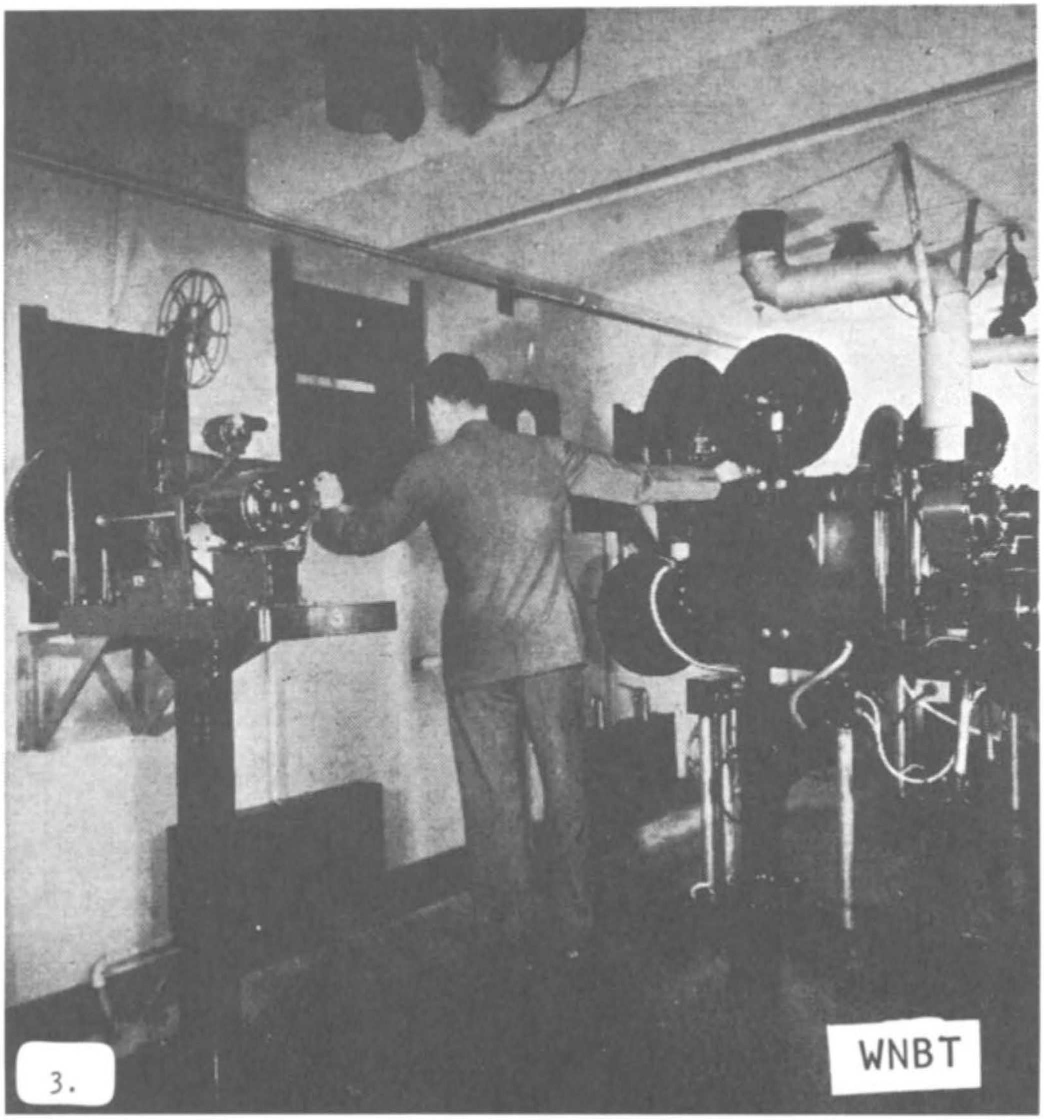


2. MOVING RAIL TELE-CINE SYSTEM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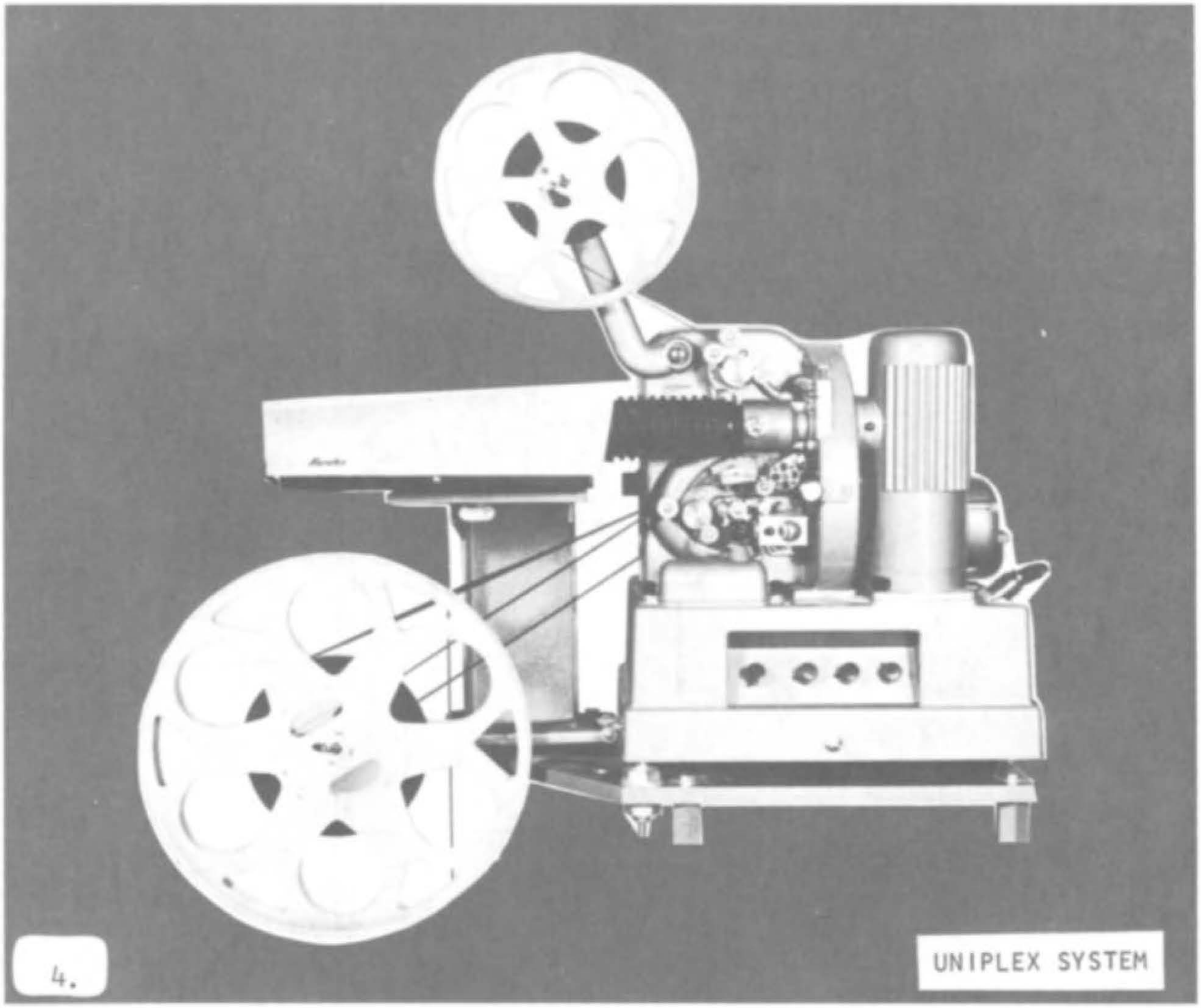
WNBT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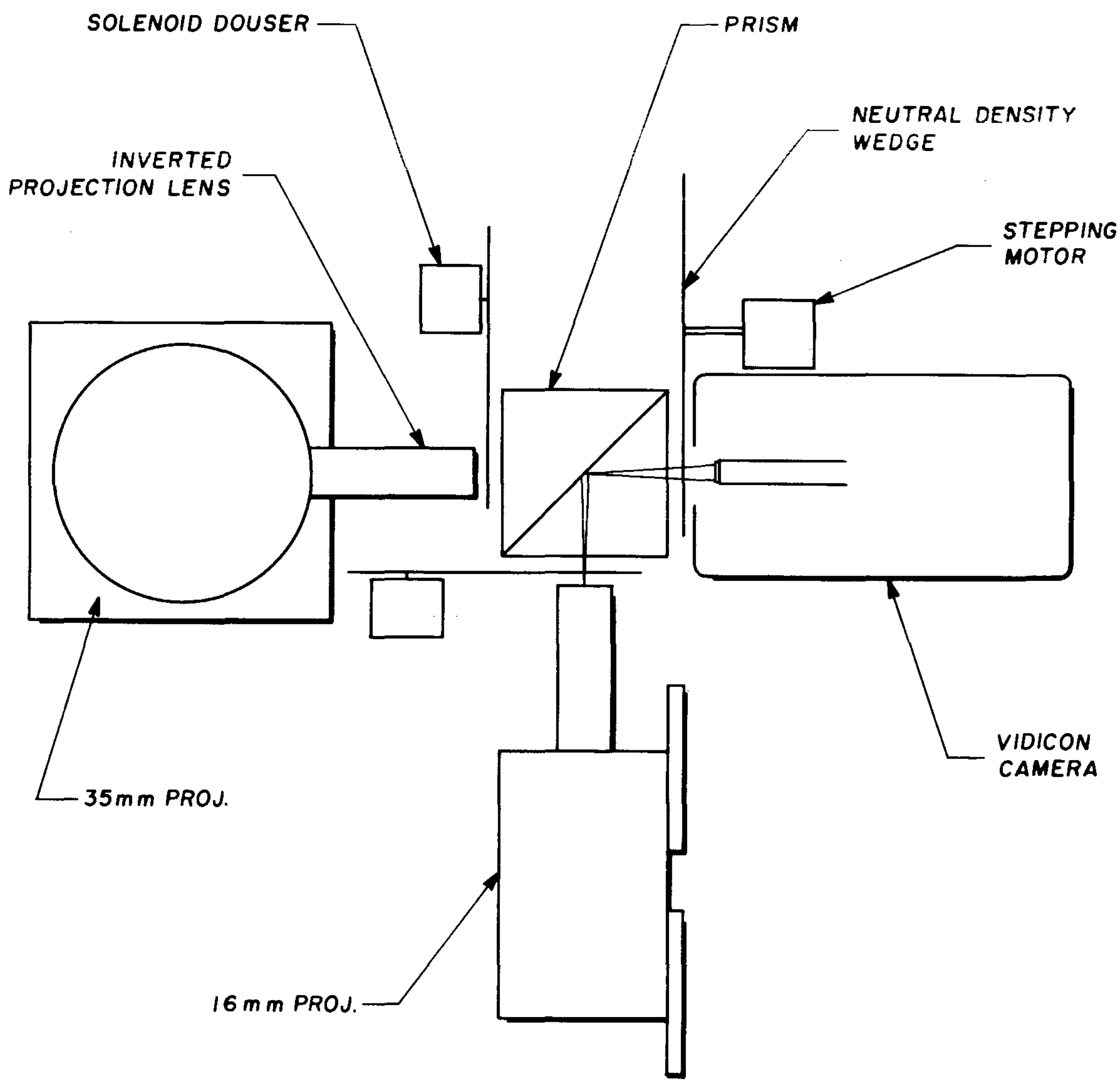
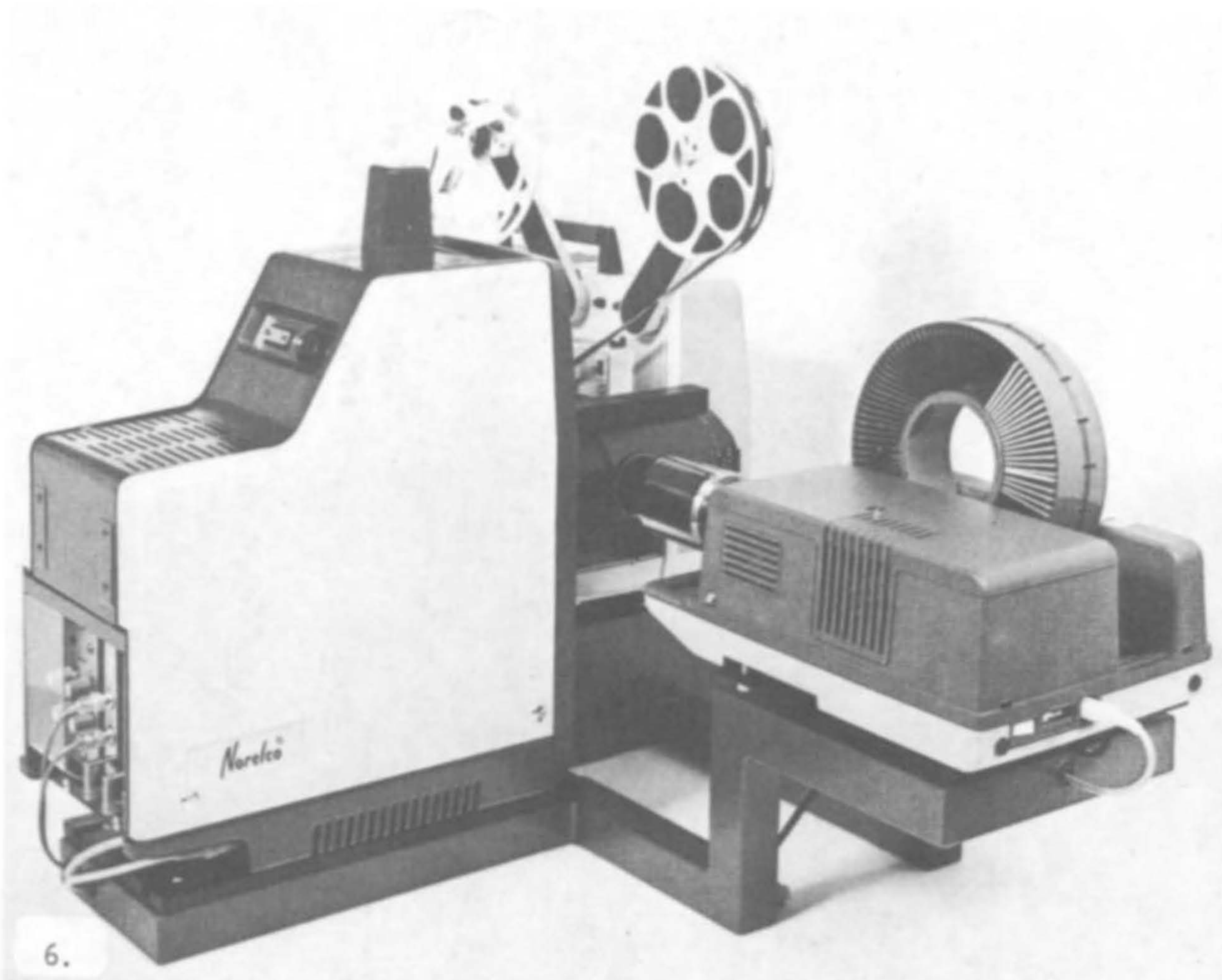
UNIPLEX SYSTEM



5. DIRECT PROJECTION MULTIPLEX SYSTEM



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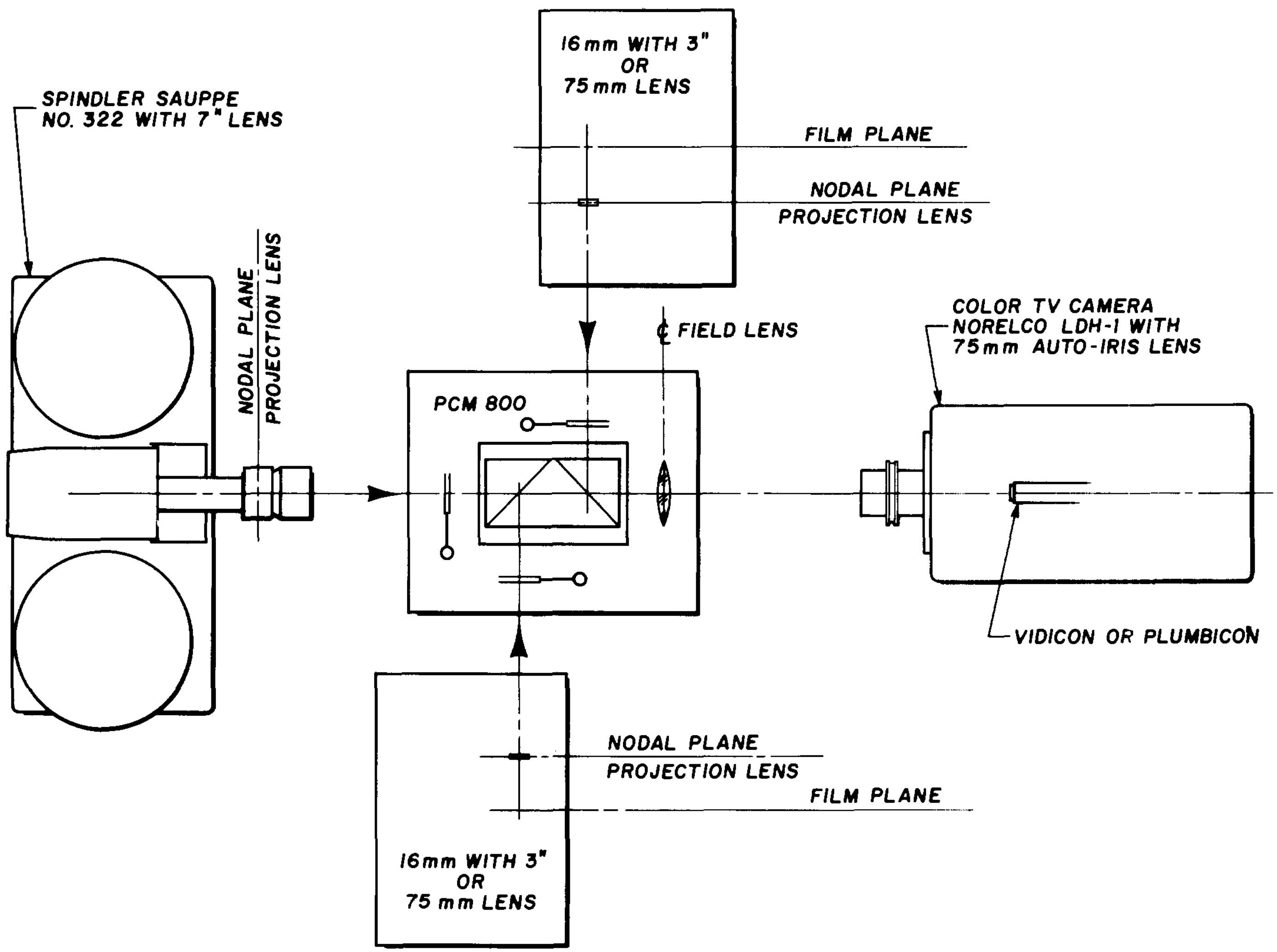
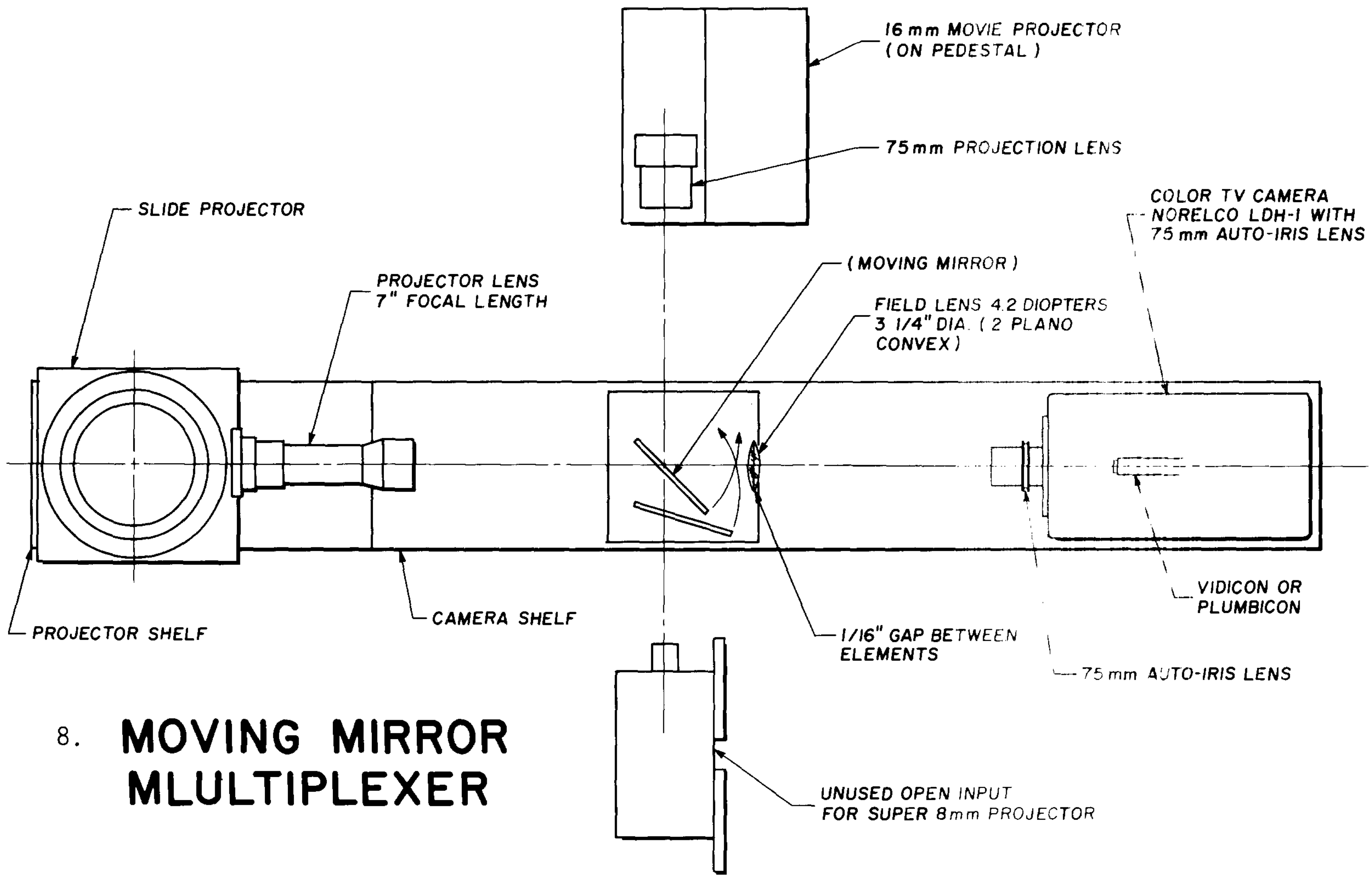
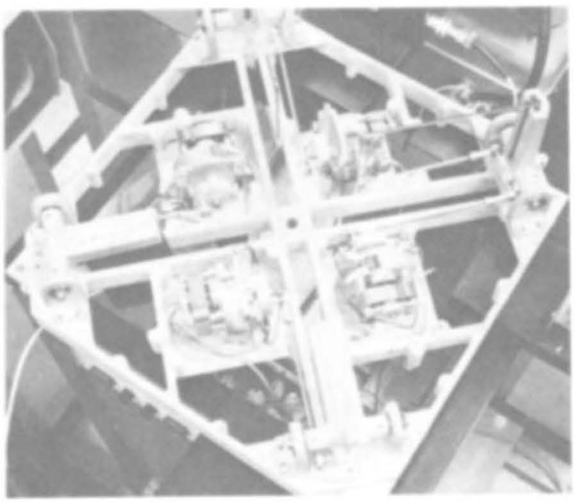


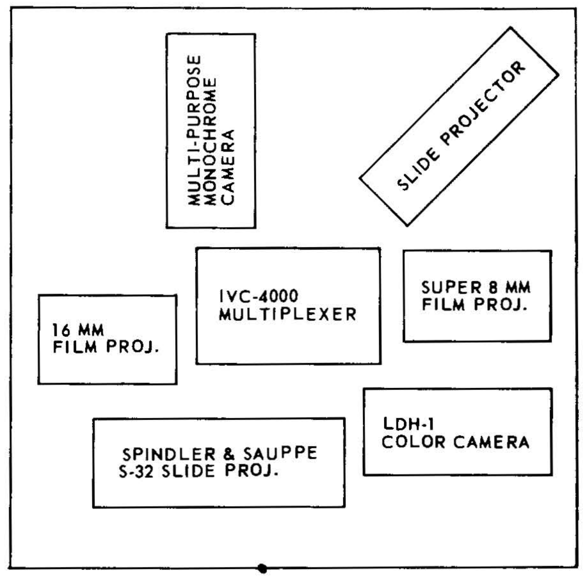
7. PRISM MULTIPLEXER



8. **MOVING MIRROR MLULTIPLEXER**

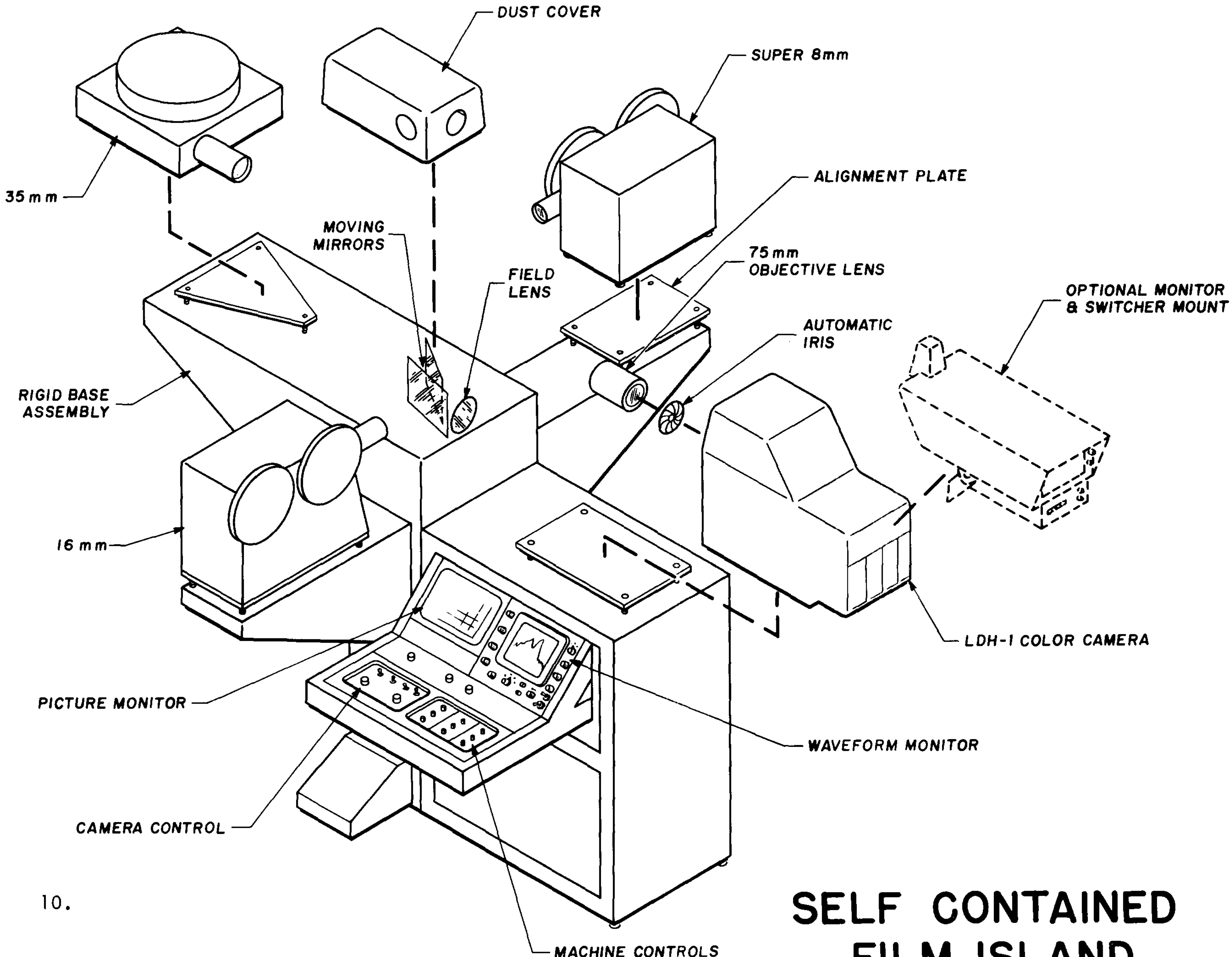


I.V.C. MULTIPLEXER HEAD



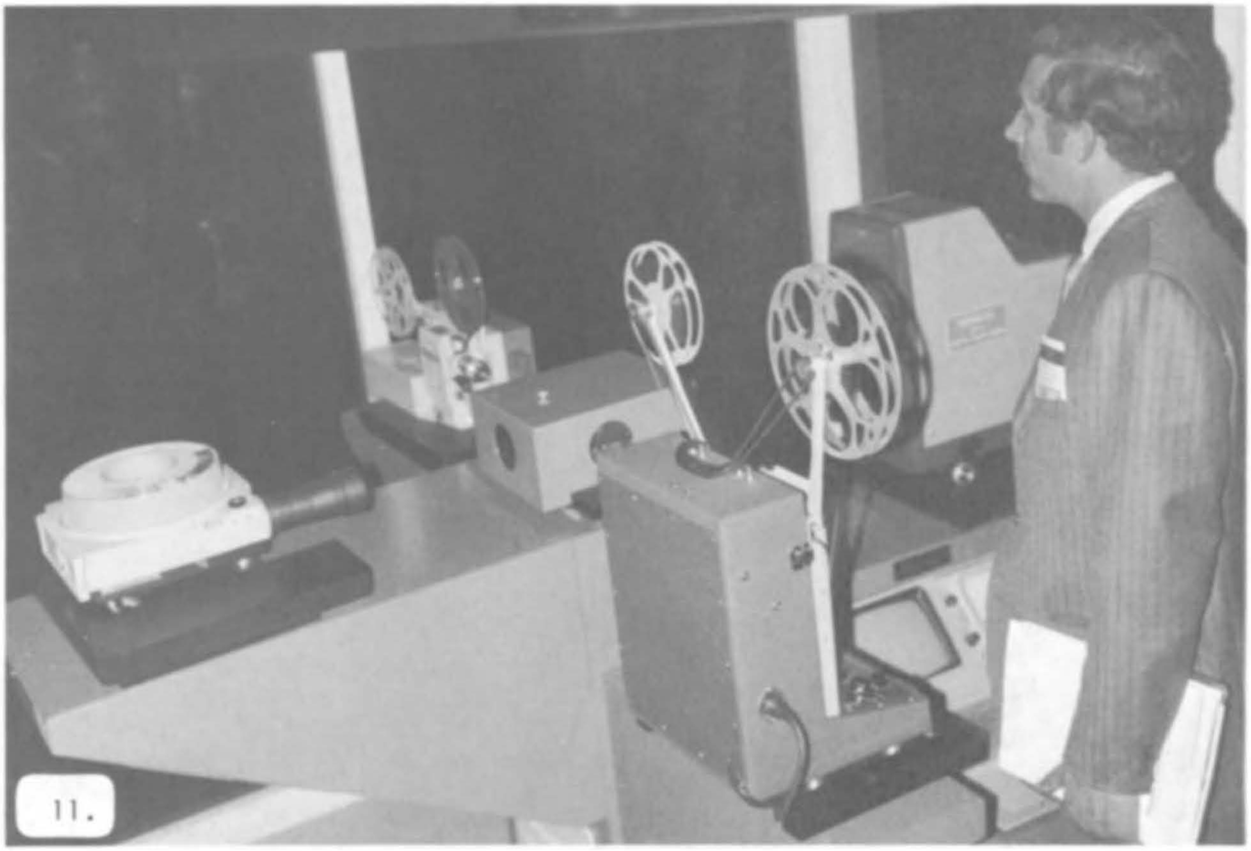
9.

TYPICAL 2-CAMERA TELECINE LAYOUT

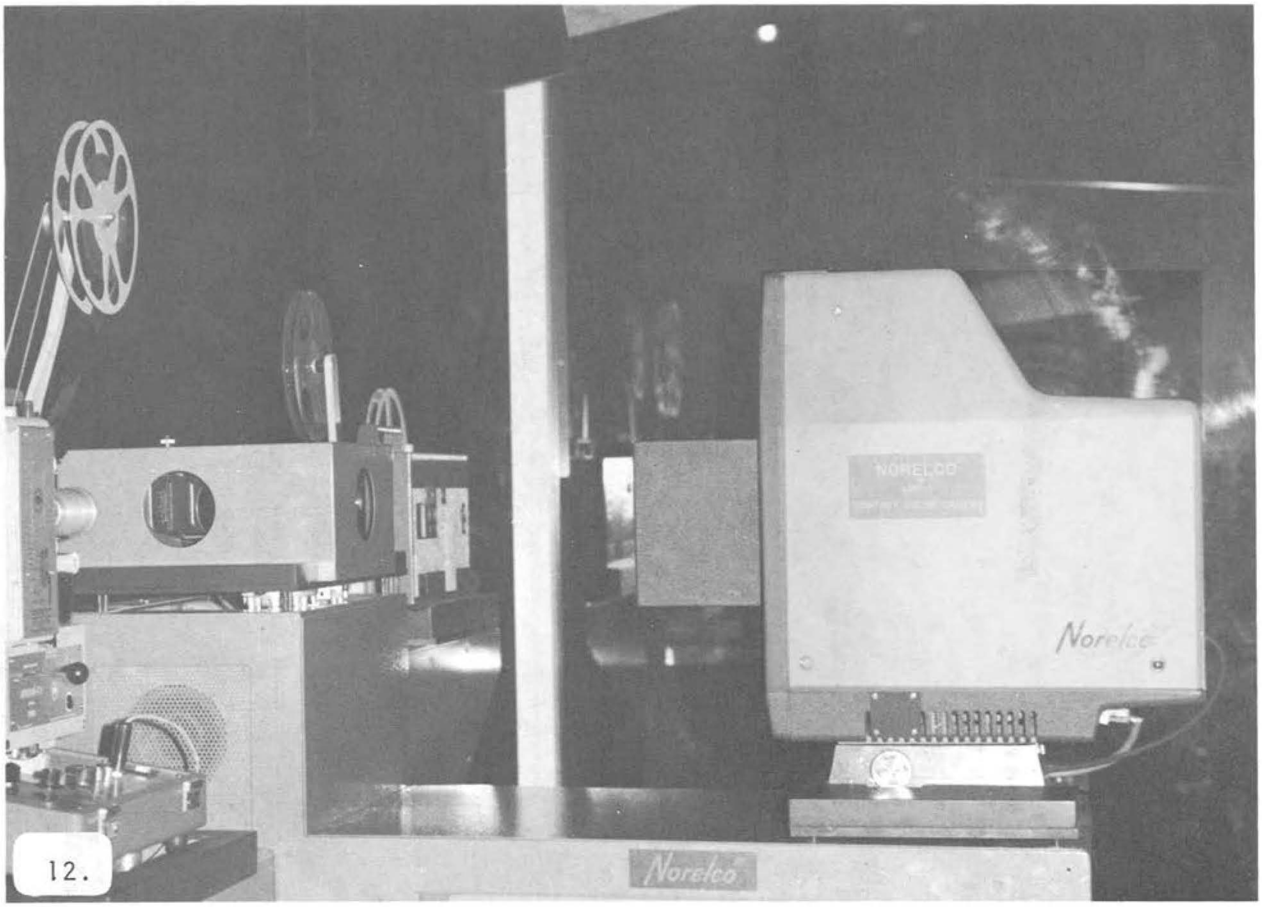


10.

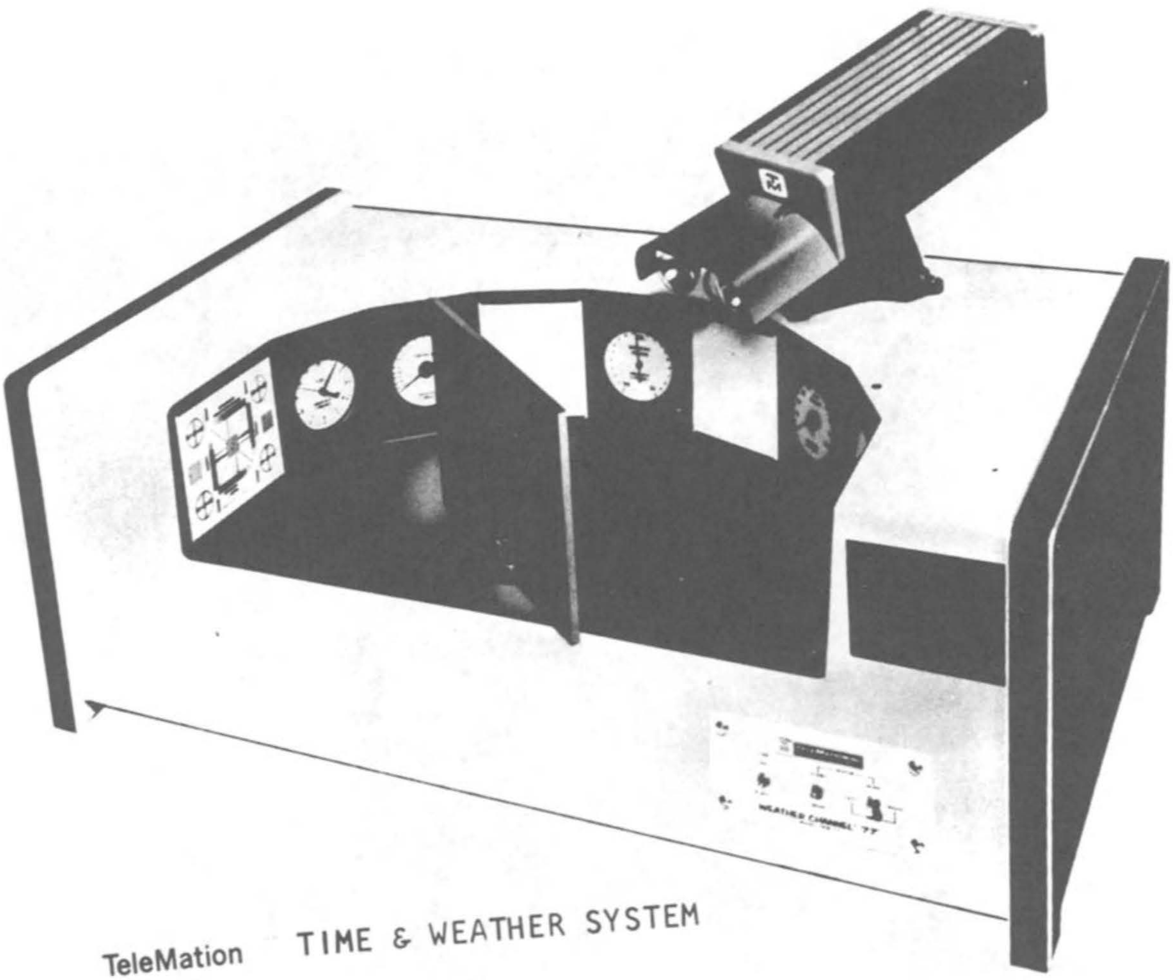
SELF CONTAINED FILM ISLAND



NORELCO LDH-1 COLOR TELECINE ISLAND

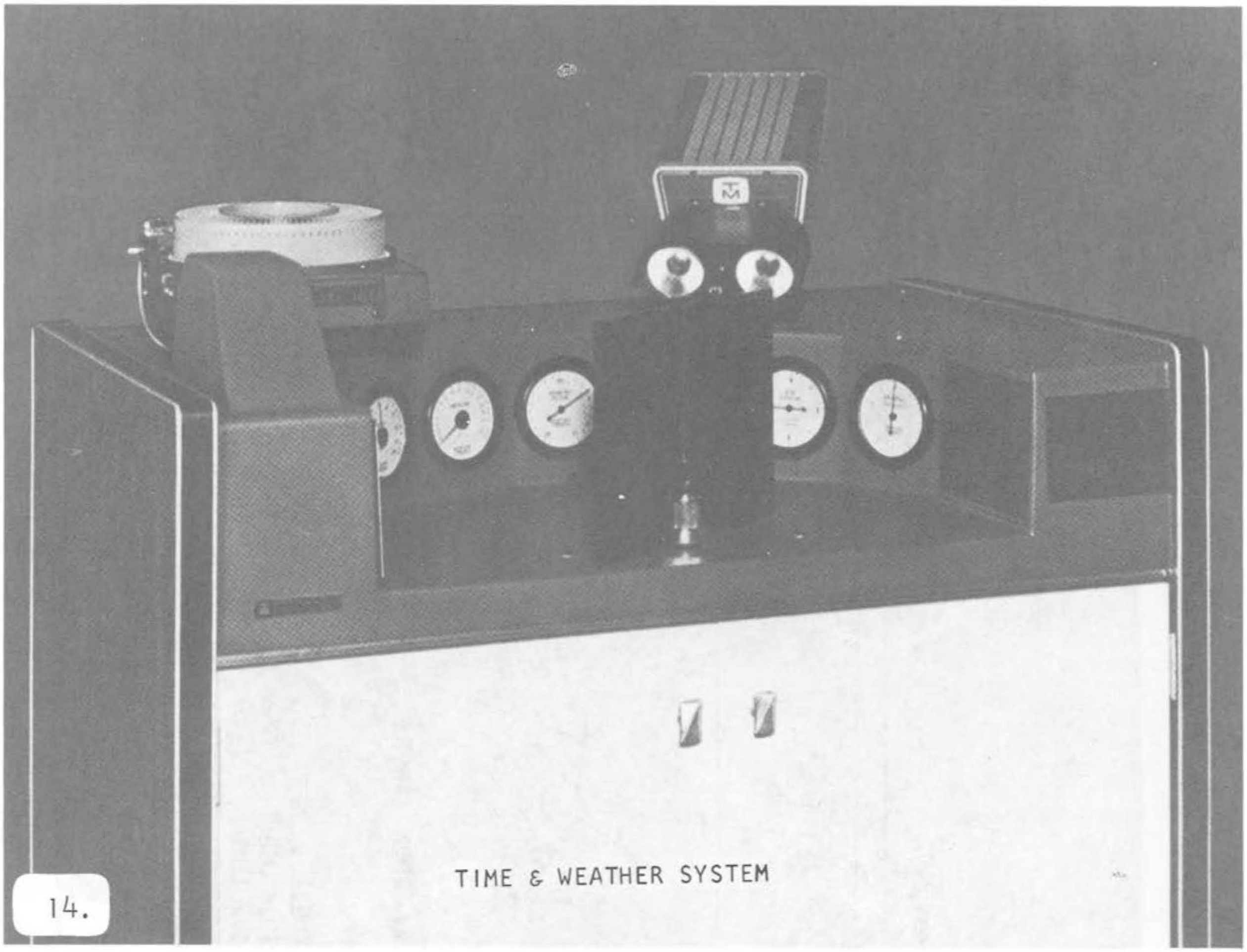


NORELCO COLOR CAMERA IS REMOVABLE FROM TELECINE ISLAND



TeleMation TIME & WEATHER SYSTEM

13.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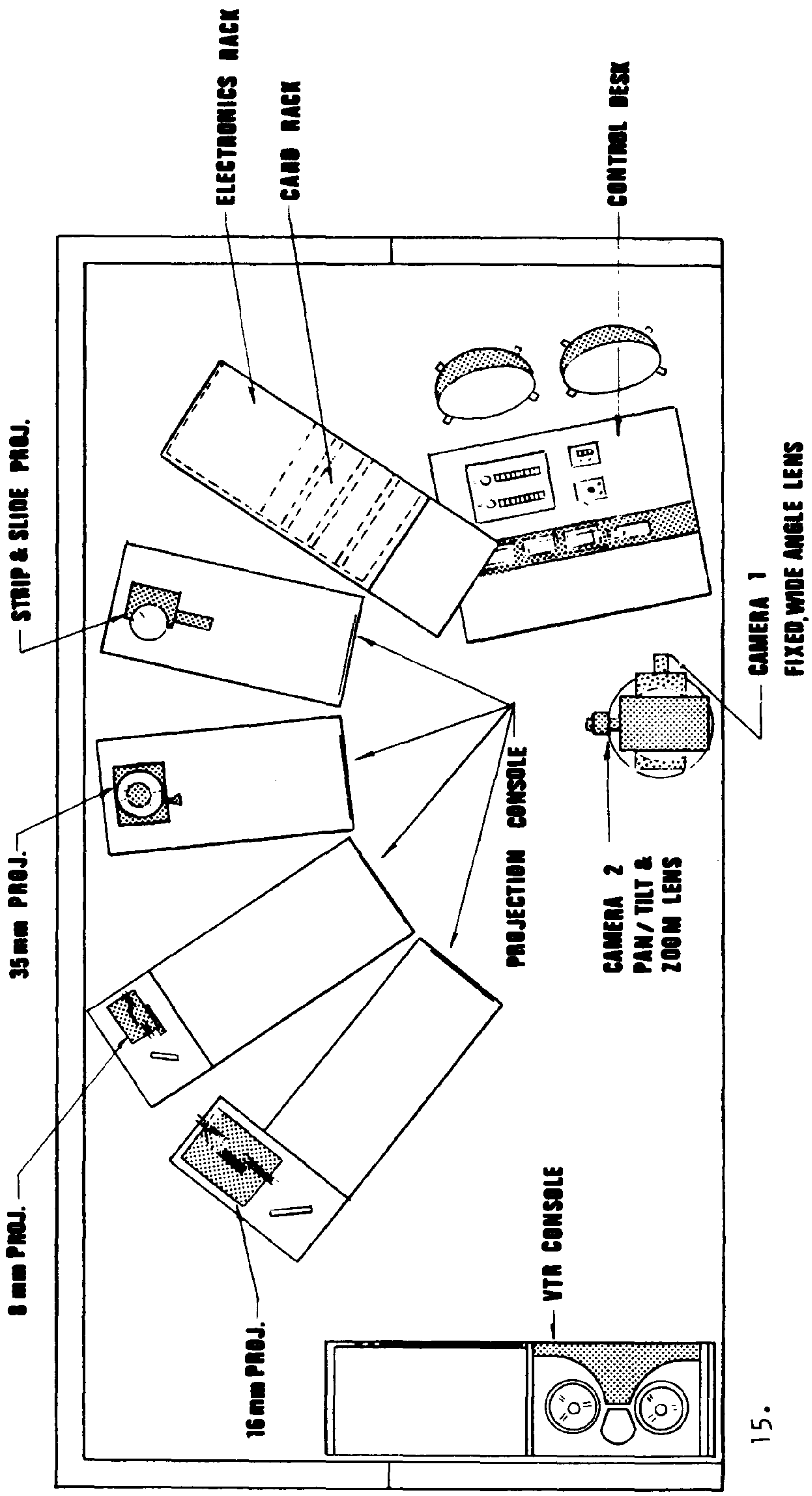
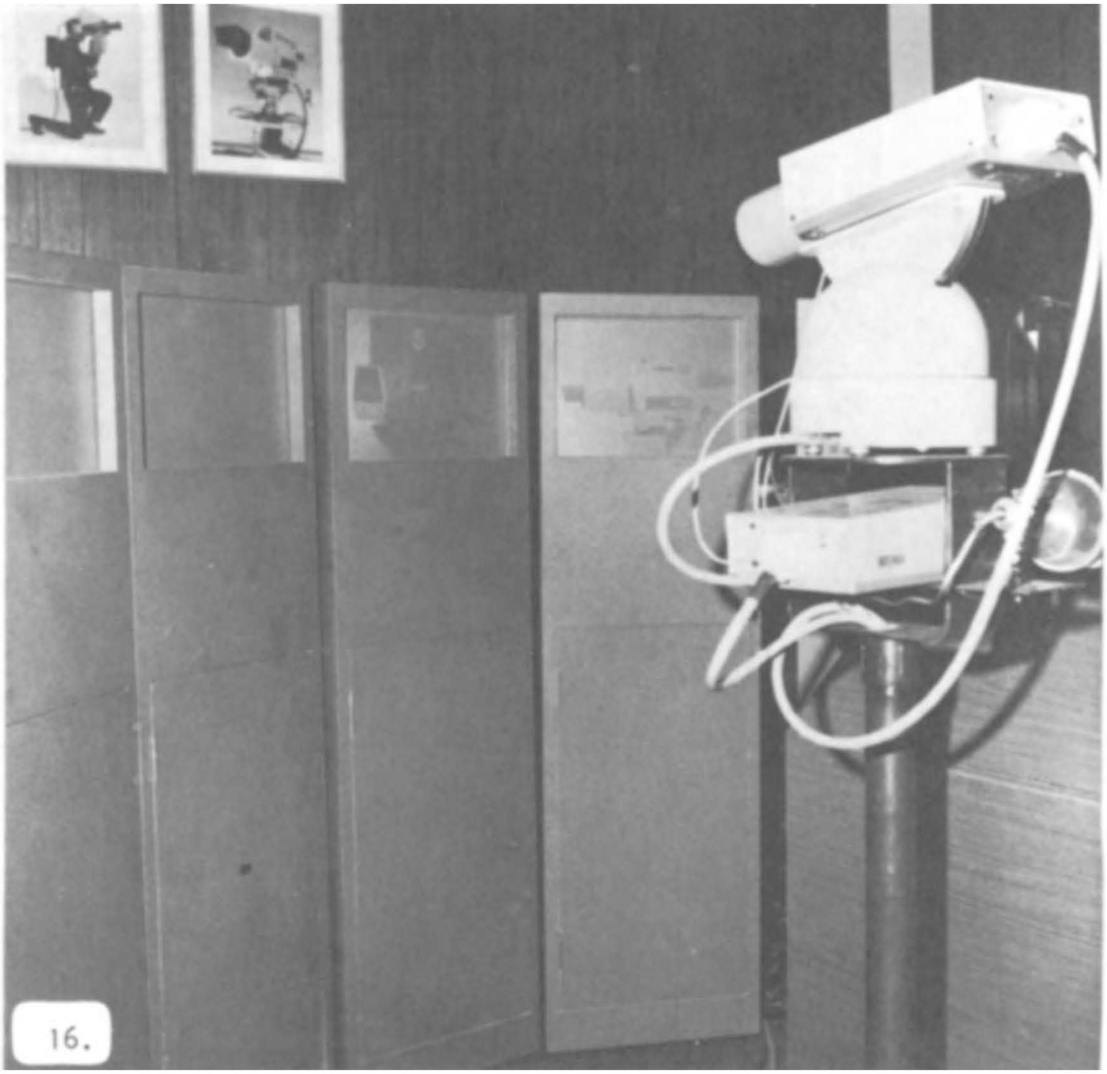
TIME & WEATHER SYSTEM


ELECTRONICS RACK
CARD RACK
CONTROL DESK
STRIP & SLIDE PROJ.
35mm PROJ.
8 mm PROJ.
16mm PROJ.
PROJECTION CONSOLE
VTR CONSOLE
CAMERA 1
FIXED,WIDE ANGLE LENS
CAMERA 2
PAN / TILT &
ZOOM LENS

15.



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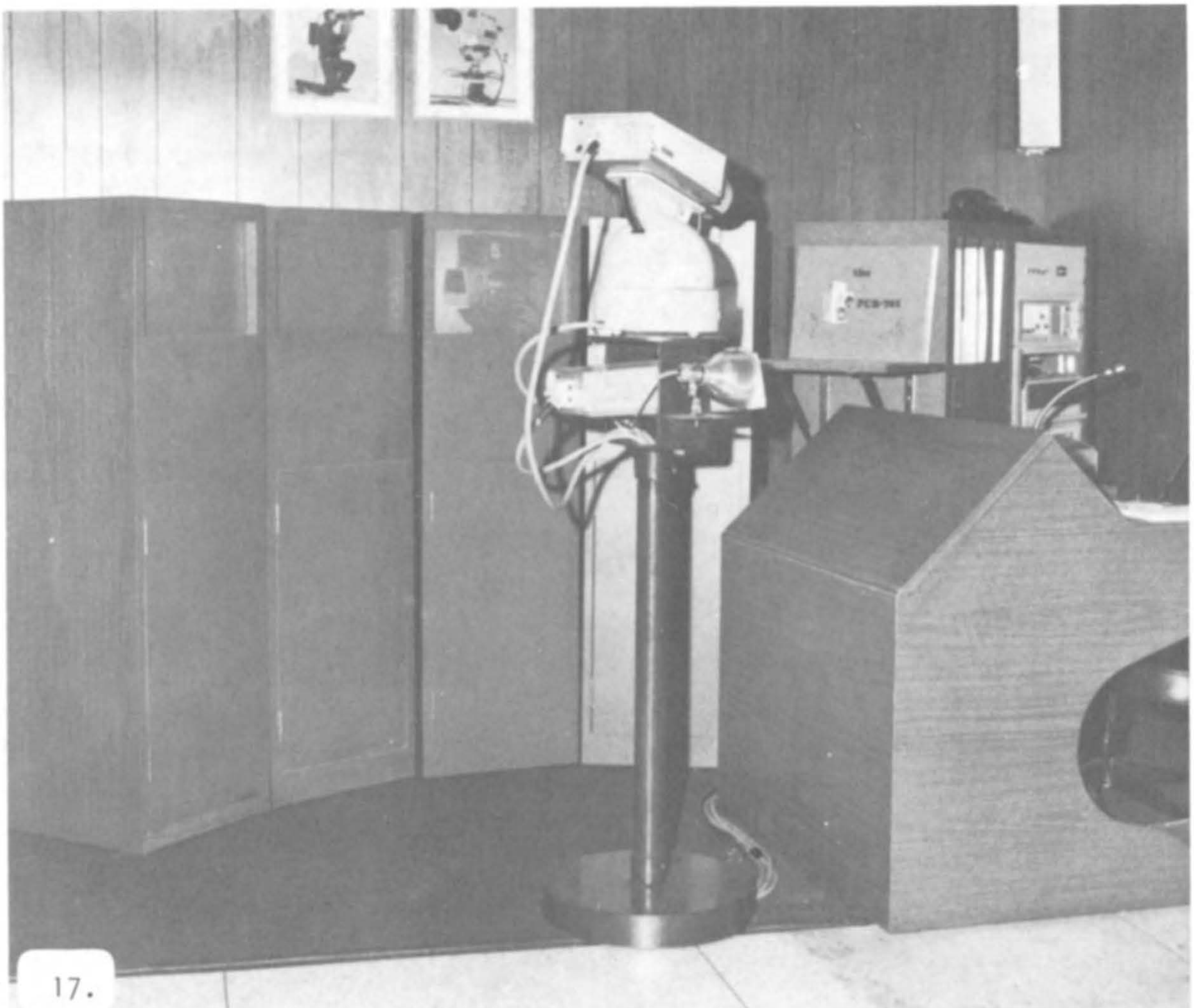
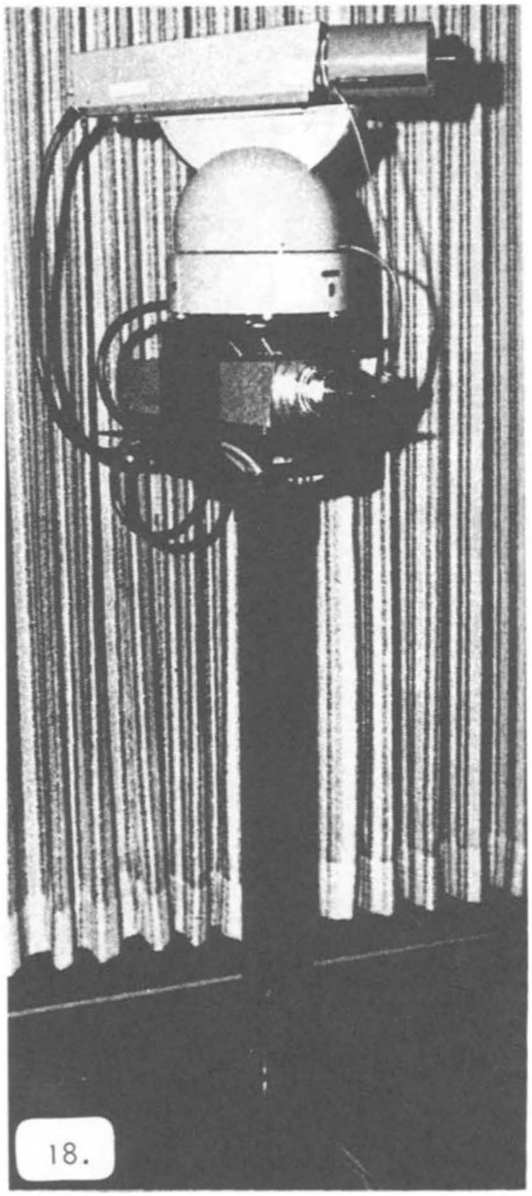
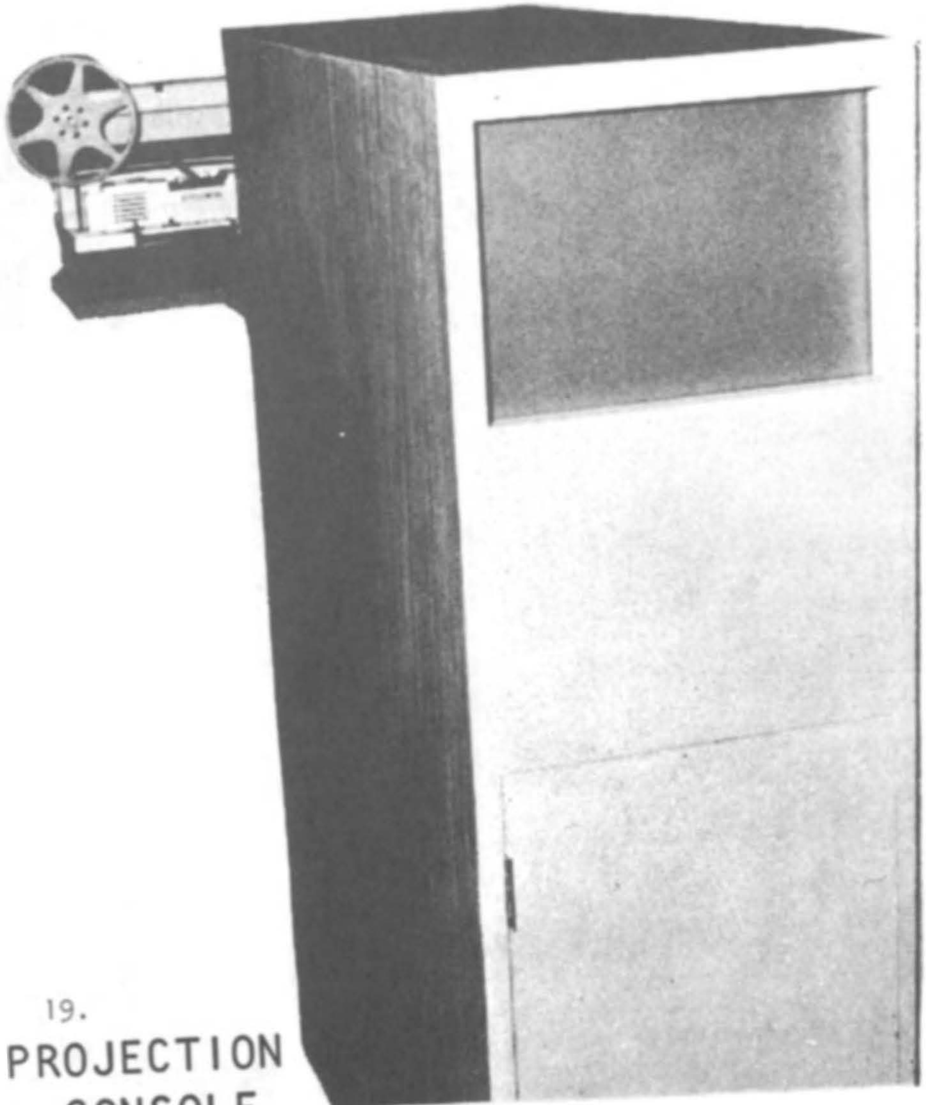
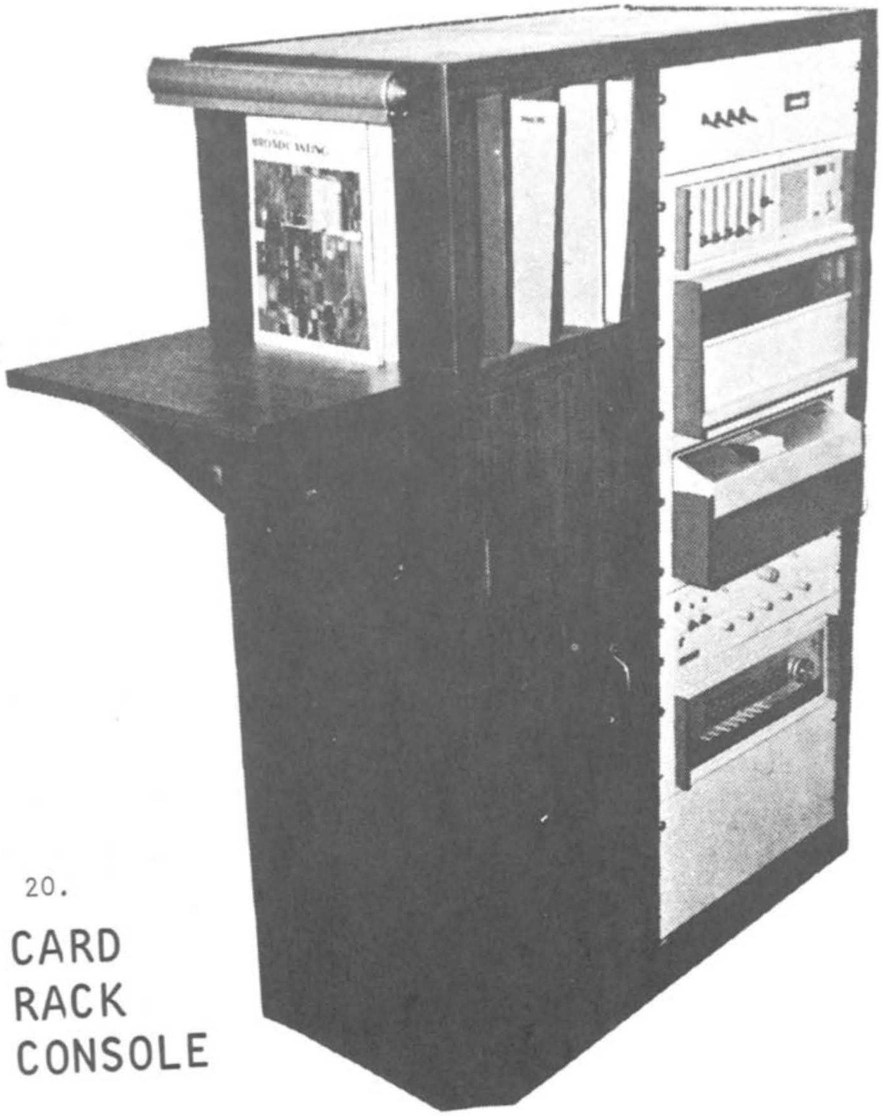


VERSATILE NORELCO "VIDEO DISC JOCKEY"
SYSTEM


18.



19.
PROJECTION
CONSOLE

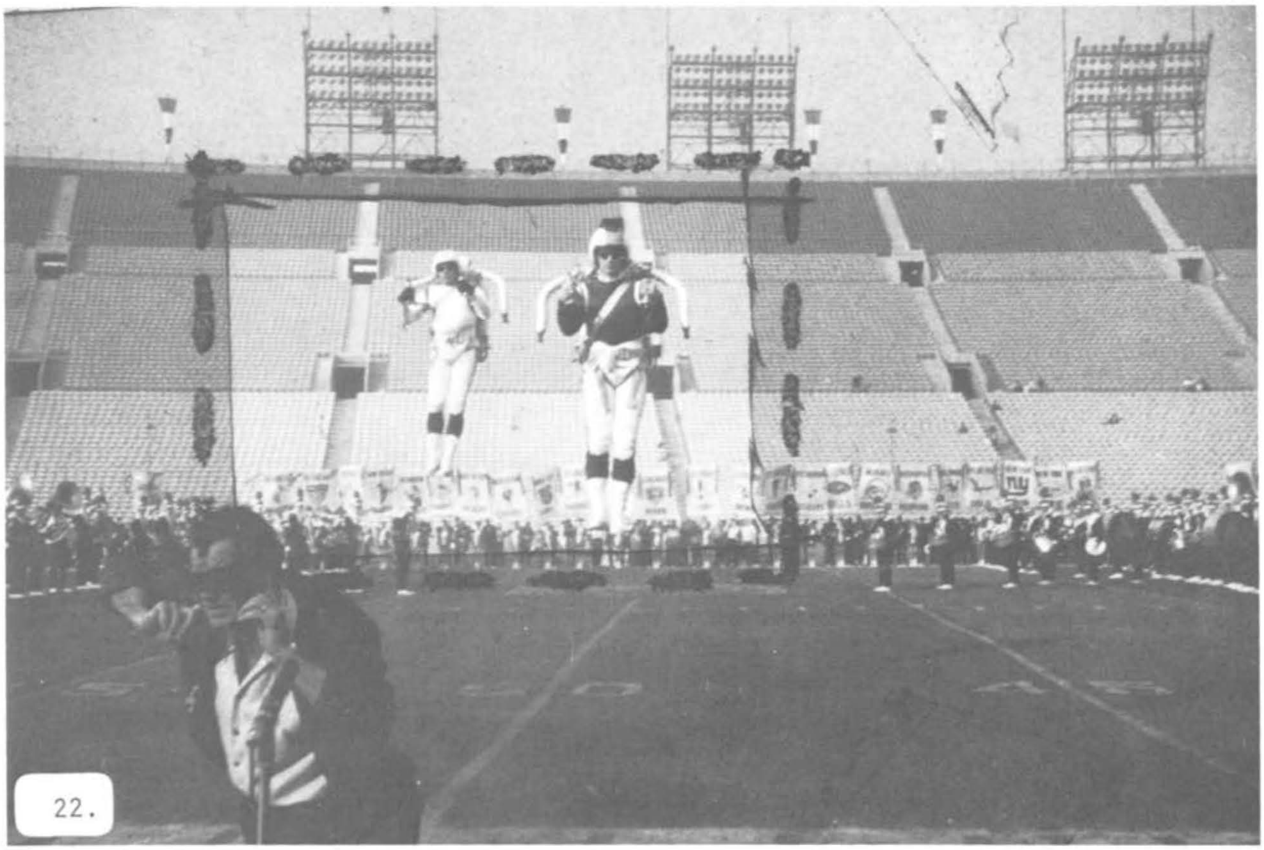


20.

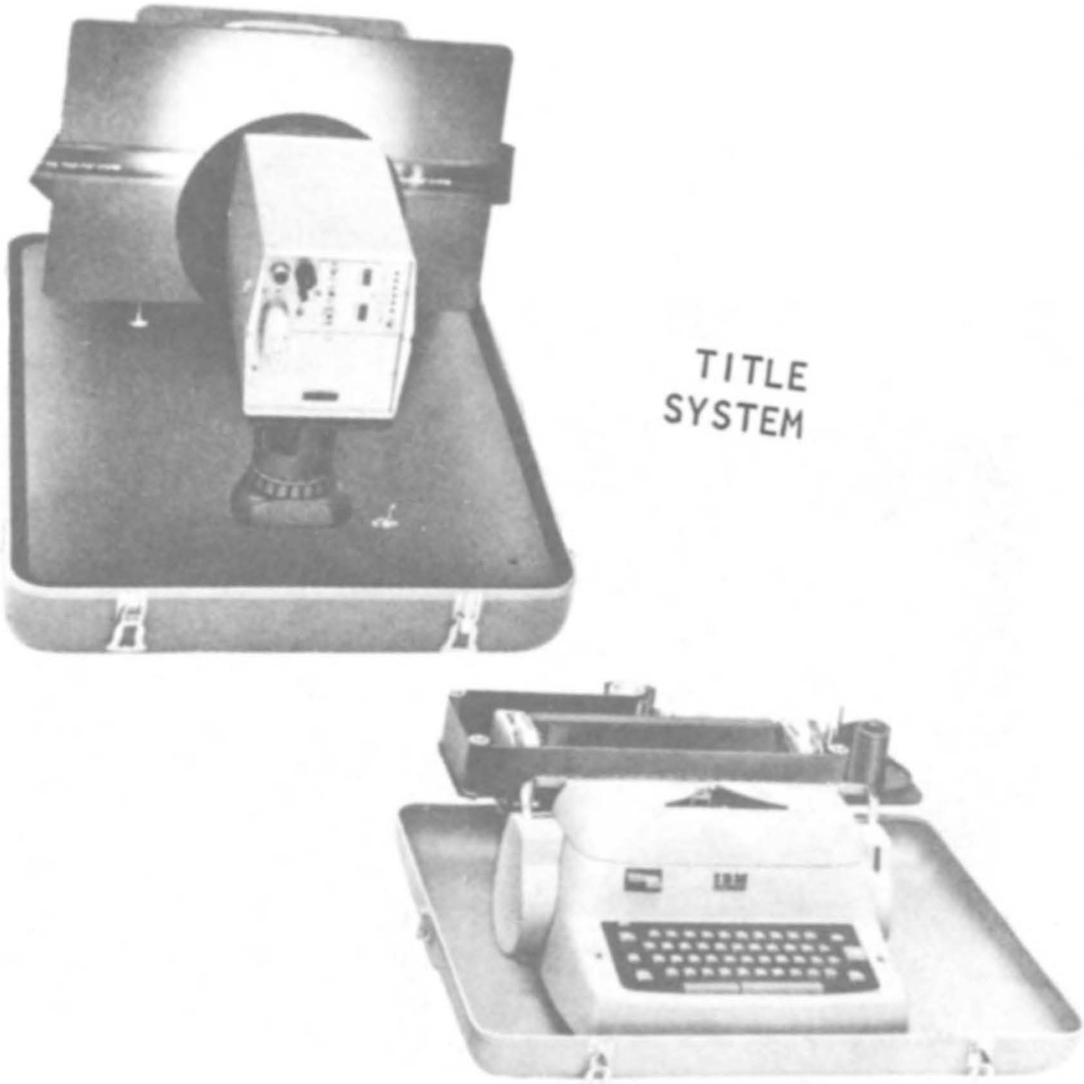
CARD
RACK
CONSOLE

CONTROL CONSOLE

21.



PICKING UP DESIRED PORTION OF SLIDE FROM
REAR-PROJECTION SCREEN



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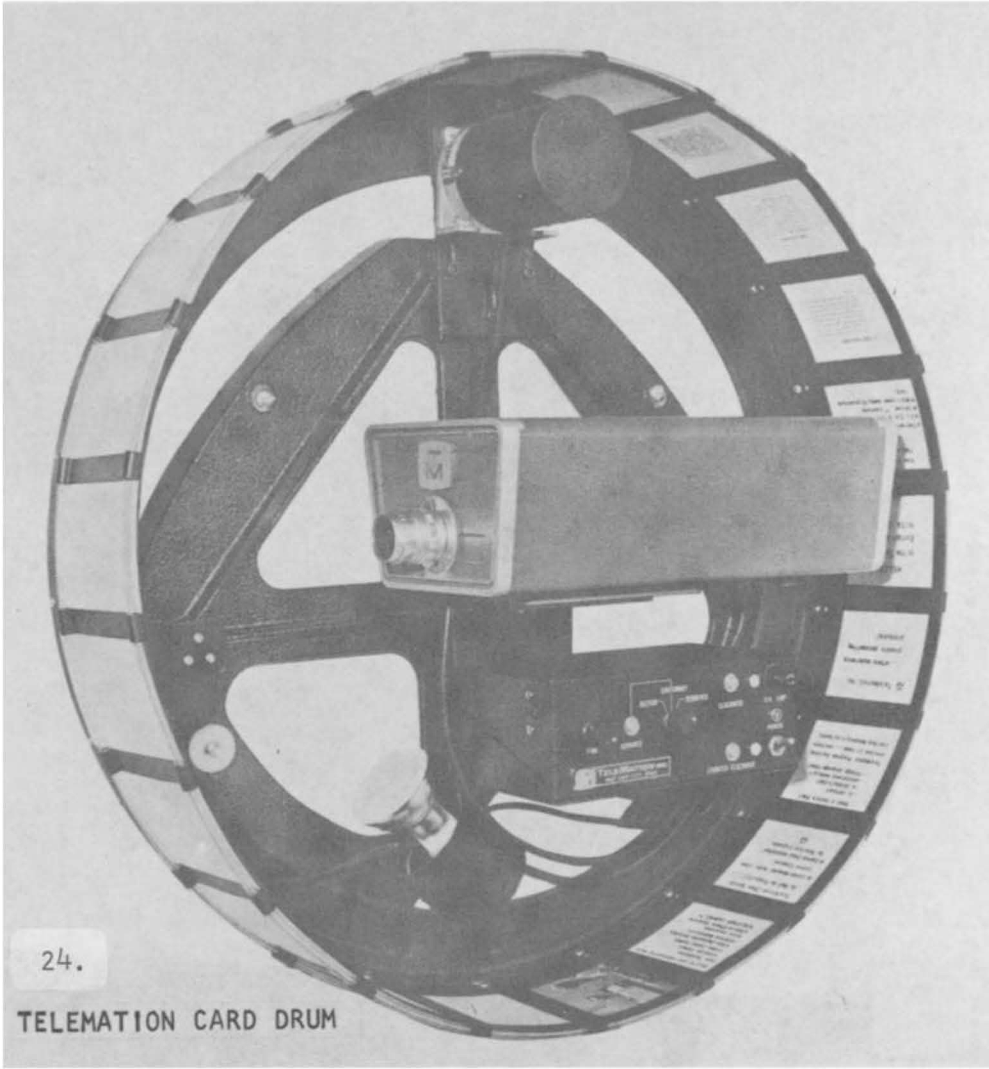
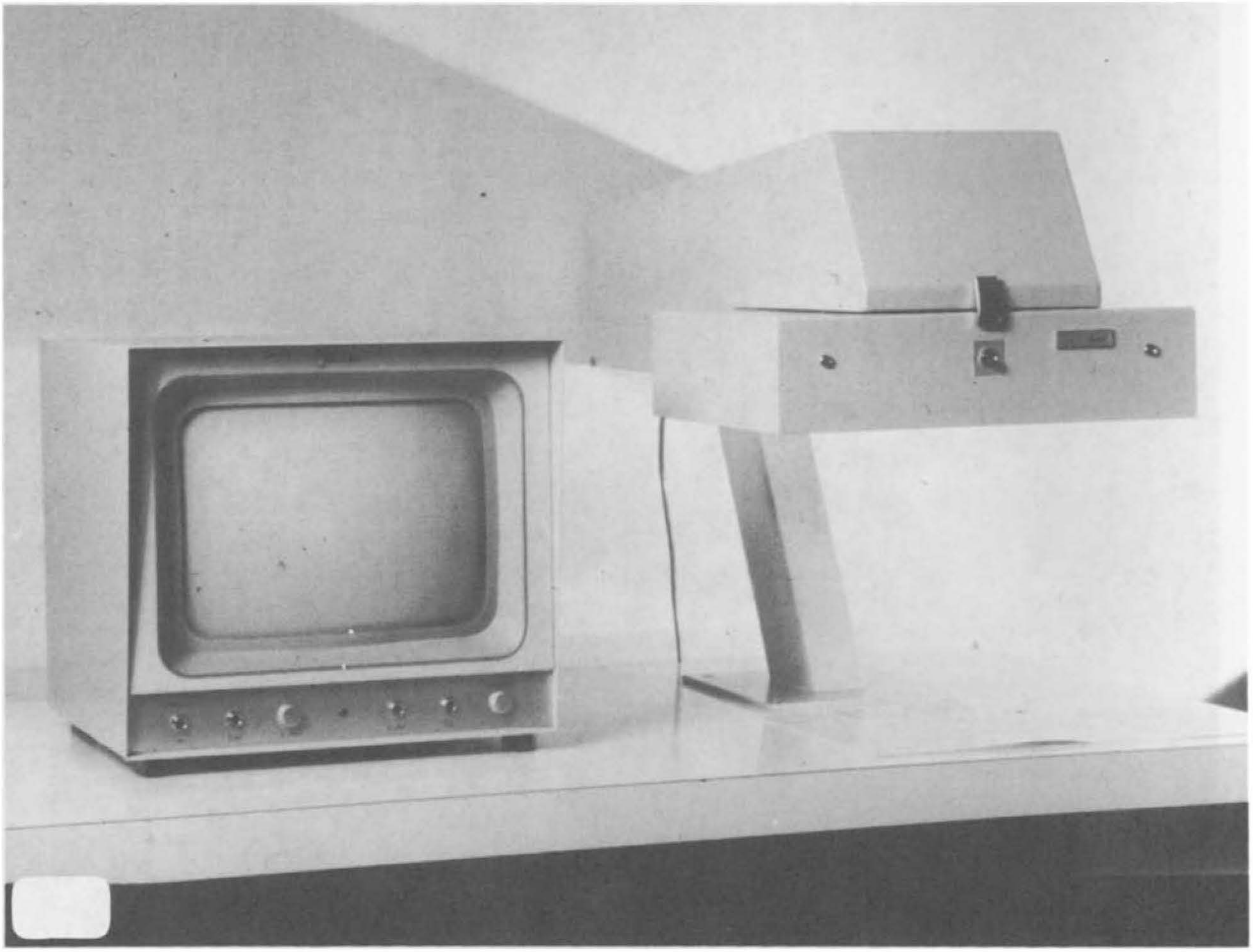


24.

TELEMATION CARD DRUM



NORELCO DOCUMENT VIEWER

Norelco®

THE SUBSCRIBER RESPONSE SYSTEM

R. T. CALLAIS AND E. W. DURFEE
HUGHES AIRCRAFT COMPANY

The Need

Communication between individuals or groups in today's world takes place in a variety of ways, each of which involves a particular medium. Many of our habit patterns and our general way of life are strongly affected by the communications media to which we are exposed to or choose to use.

The well-known mass communications media include radio, television, telephone, motion pictures, correspondence (mail), newspapers, and large circulation magazines. With the exceptions of the telephone and mail, these media are largely one-way communications systems from which the general public, may receive information but cannot readily communicate back in the same medium.

One-way communication media have serious drawbacks for those on both ends. The receiver cannot make his individual opinions and needs known without resorting to another medium. And the originator does not have access to those opinions and needs which could be exceedingly valuable. A two-way mass communication medium would, in contrast, not only allow the receiver to express his views, and possibly get what he wants, it also allows the originator to modify his operation for whatever effect he chooses.

The presently available two-way media (mail and telephone) have serious shortcomings for mass communication. The traditional distribution system for mail is slow and expensive, particularly where mass distribution is required; and the telephone is much better suited to communication between individuals than to mass communications.

The public is at a considerable disadvantage under a unidirectional mass communications system. The medium constrains the public to act in essentially a passive role. Reactions and responses to the "downstream" information are either absent, indirect or must get back "upstream" via telephone or mail, with the penalties noted. As a consequence, only a very small percentage of any mass audience responds in any detectable way, and it is extremely costly and time consuming to obtain anything like a precise determination of response.

The reaction of the public to advertising, entertainment, sports or other types of programming is ascertained presently by very limited surveys conducted after the fact. It is consequently difficult for private corporations or governmental agencies to know what the public accepts, rejects, or is willing to purchase until the acid tests of sample polls, sales reports or vote tabulations reveal the facts. Even reasonably rapid responses are not generally possible, and the validity of extending the results from selected population samples to the entire public may be questionable.

The explosive expansion of the freeway system and the wide acceptance of air travel has resulted in a complete change in business and social habits. It is commonplace for a businessman to fly cross country for a short meeting - whereas in earlier times the business would have been conducted by letter or telephone. In a similar manner, the housewife makes use of her automobile to travel relatively long distances for food and other shopping services.

Increasing public interest on the effects of pollution as well as the growing inconveniences due to congestion in the airways and freeways may well augur another change in life style, particularly if an acceptable substitute can be found. The key to a new life style could well be an adaptive two-way communication system in which people could perform a greater portion of their work, make more of their purchases and receive even more of their entertainment without leaving their homes or offices.

Another universal need which is not adequately satisfied by current media is emergency communication of medical, fire, intrusion, and other alarms to the proper agency without delay and at a reasonable cost.

Other examples could be cited in which presently available communication methods do not adequately satisfy human needs. Even more disturbing is the realization that technological limitations inherent in the present media not only prevent resolution of these difficulties and shortcomings but also allow for little foreseeable growth to answer needs which have been clearly recognized by both government and industry alike. Moreover, such communications requirements will in fact constitute the very corner stone upon which the life styles in this and future decades will be based. In its response to the F.C.C., (Docket #18397 - The Future of Broadband Communications) the IED/EIA stated that "--- The mushrooming growth in available information and the demands for access to this information is bringing about a revolution in communications which will produce a profound change in the very way society is structured and in the way we live."

The need, concisely stated, is for a widely available, broadband, two-way communications system that can rapidly handle large amounts of information in both directions. In other words, a mass communication system that can actually be used by the masses. It need not compete destructively with existing media but could, complement the impressive communication services we already enjoy.

Such a system would have to meet the present and emerging needs for sound, pictorial, and data transmission and allow for undefined future growth by economically viable modular expansion.

The Means

Cable television, the one-time stepchild of broadcast television, now appears as the leading candidate to solve the major communication needs cited. Here again the IED/EIA states that, "The terms "CATV" for Community-antenna television and "CTV" for cable television fail to do justice to the potential of the medium." The great natural assets of CATV is that its facilities are either in now or are planned and the new two-way services can be provided requiring only a relatively small incremental unit investment. Further, it can be expected that the additional revenues resulting from both these new services and associated increased market penetration will provide the system operator with greater financial resources necessary to keep pace with anticipated growth demands. The 300 MHz electromagnetic spectrum contained within the radiation-tight cable does not violate free space as does the emanations of off-the-air television. Once having exhausted the initial 300 MHz capacity to meet present needs, additional 300 MHz bandwidth units can be added by the expedient of additional small diameter coaxial cables.

The technology for installing a cable system is similar to that now providing universal telephone and power service to the entire country. It requires no technological breakthrough to conceptually substitute coaxial cable for wire and visualize a nation wired for broadband cable service limited only by customer demand. The next logical step is the expansion of such a service into two way operation.

Two-way use of cable television is already technological feasible: either by simply utilizing separate downstream and upstream cables or by simultaneous bi-directional signalling on a single cable with frequency multiplexing and two-way amplifiers and filters.

Viewed broadly, therefore, the solution to interactive, universal communication is in hand. There remains however, the choice of a specific implementation: the selection of a specific sub-technology to accomplish this two-way communication. But this is really a second priority decision. The first requirement is to define what the system is to do in a market which has not yet been proven. Which of the inadequacies of the present media and which of the foreseen needs should be addressed first in the choice of system design parameters and techniques?

Another fundamental question is that of design obsolescence. Should considerations of low cost installations in this fledgling market override the risk of relatively early obsolescence as public acceptance and usage grow rapidly? Or should a system be devised wherein growth capability and modular expansion are integral parts of the design plan.

Recognizing the uncertainties of customer acceptance and the time span required to fully exploit the services offered, Hughes has elected to approach the problem from a systems viewpoint. We have, designed a complete two-way communications system rather than a single device or service.

The total system approach has produced a design that can grow to meet the long term goals and does not attempt to prejudge the relative value or eventual marketability of services to be offered.

The System

The Subscriber Response System (SRS) is designed to permit the cable system operator to add two-way communication capability with a modest initial investment and yet be able to expand in a modular fashion as the number of subscribers, the traffic, and the demand for additional services increase - all without obsolescence of previously installed equipment.

The Subscriber Response System is shown in Figure 1 incorporated into a typical CATV system. The two-way communications take place between a computer complex termed the "Local Processing Center" (LPC) and the Subscriber Terminals located in the subscribers' residences or places of business. The LPC equipment shown in Figure 2 can be located at the Head End, at the Local Origination Studio, or remotely from the local CATV system.

Depending on the choice of location, signals between the LPC and the Head End are fed by cable or microwave relay. At the Head End, the downstream SRS signal is frequency multiplexed with the normal CATV video spectrum and sent downstream through the cable network, including the existing trunk and distribution system.

At the subscriber's home or business location, the composite signal at the normal drop line cable is routed to the Modem Unit of the Subscriber Terminal shown in Figure 3.

The Modem frequency converts a 26 channel TV spectrum and furnishes a fixed frequency signal to the TV set, normally channel 8 or 12, thus eliminating a separate frequency converter. The Modem performs all of the radio frequency modulation and demodulation and most of the digital signal processing required at the Subscriber Terminal. It also furnishes the interface for all accessories used in the system. The Modem requires no operating controls and is designed for installation at an unobtrusive location, nominally behind the TV set.

All operating controls for the terminal are located at the Subscriber's Console shown in Figure 4. The Console is interconnected to the Modem by a small diameter cable which allows approximately 50 feet separation between the units depending on the installation requirements at the subscriber's location.

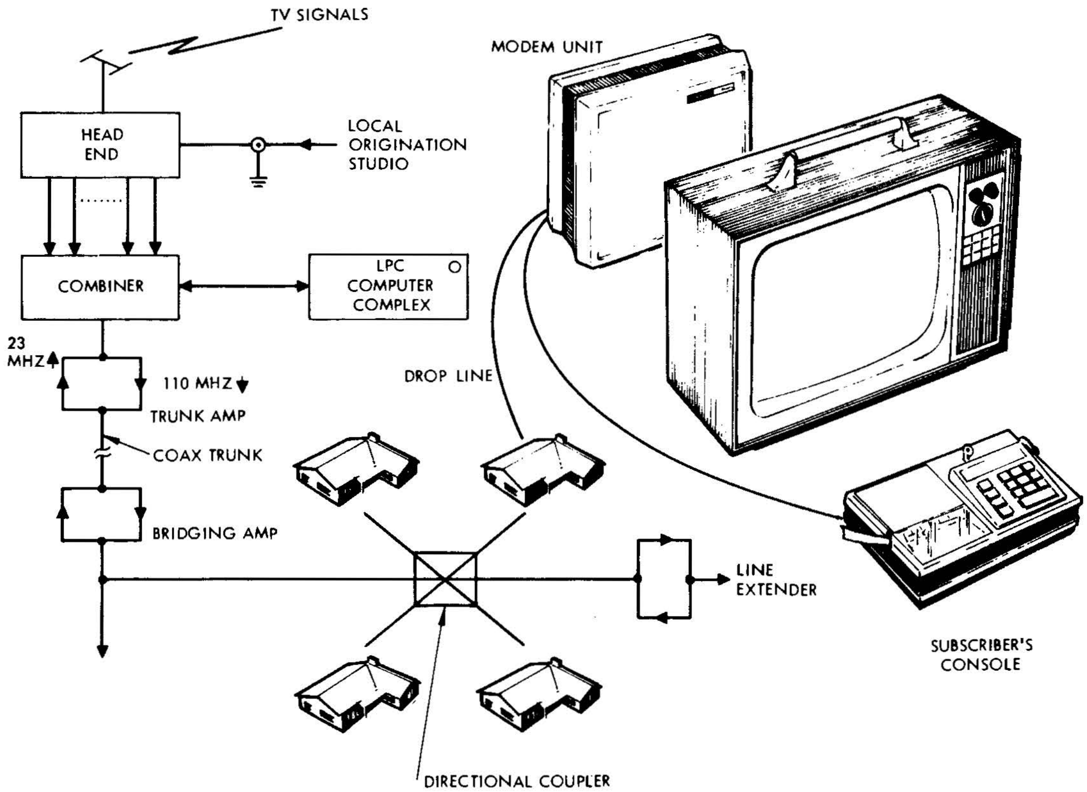


Figure 1. Overall CATV Two-Way System

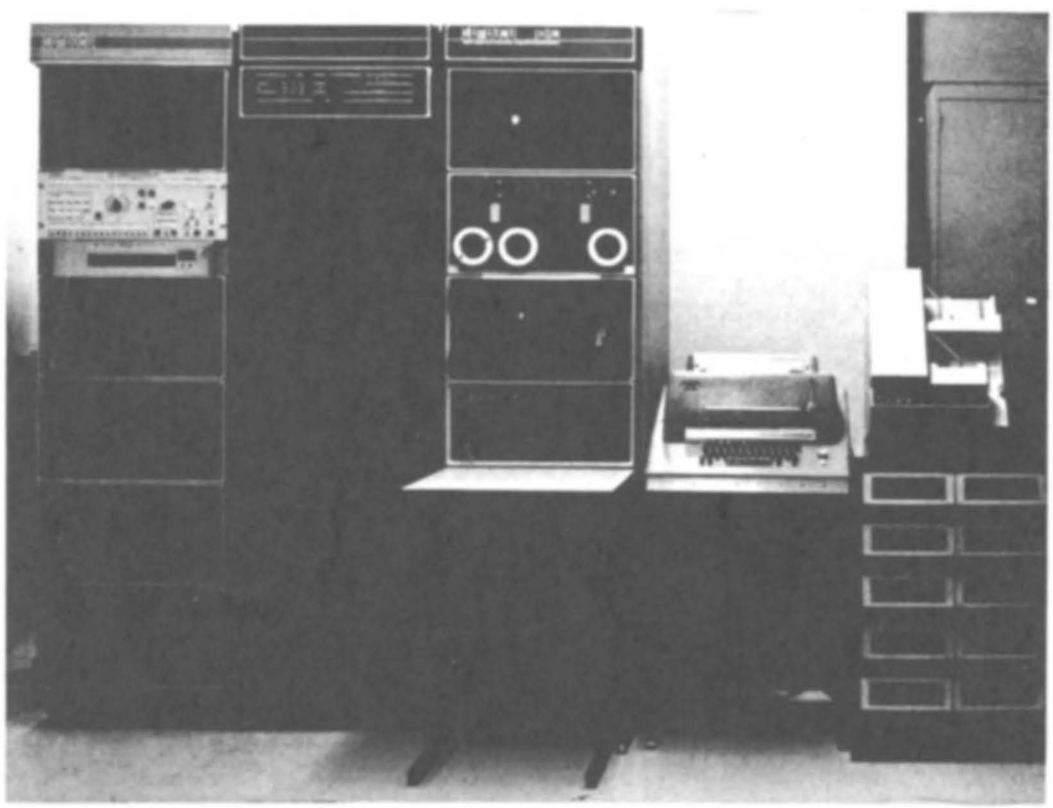


Figure 2. SRS Local Processing Center

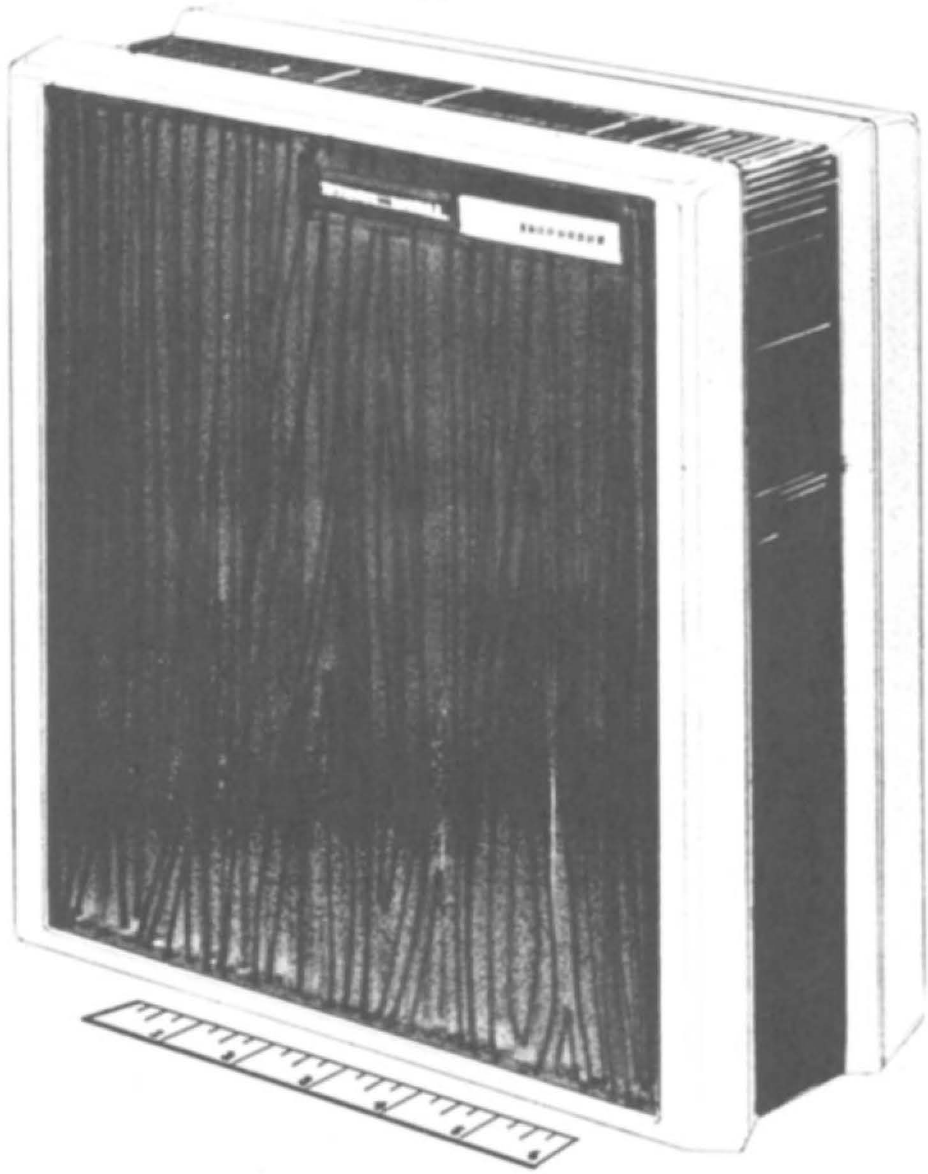


Figure 3. SRS Subscriber Terminal Modem

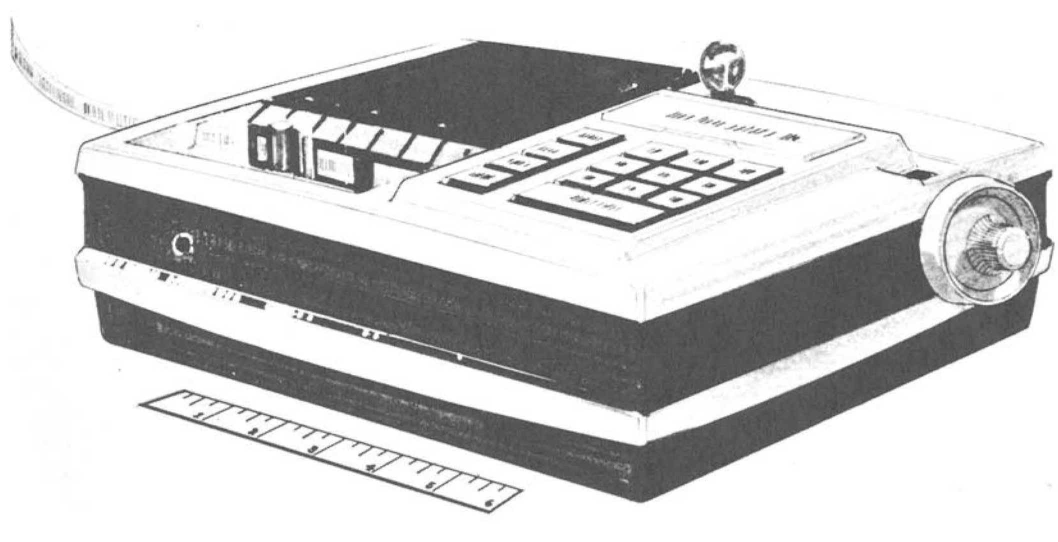


Figure 4. SRS Subscriber Terminal Subscribers' Console

In addition to a TV Channel Selector Switch the Console contains a keyboard and a small strip printer allowing the subscriber to engage in two-way communications with the Local Processing Center.

Communications upstream from the Subscriber Terminal to the Local Processing Center are transmitted back from the Modem either over the same cable network with suitable upstream amplifiers and filter networks to by-pass the existing downstream amplifiers, or over a separate cable in a two cable system.

The resulting spectrum of signals on the cable is shown in Figure 5. The downstream SRS signals occupy a 4 MHz bandwidth from 108 to 112 MHz. The downstream form of communication is digital pulse code modulation (PCM) at a 1 Megabit per second rate. The digital data is then used to frequency shift key (FSK) a 110 MHz carrier.

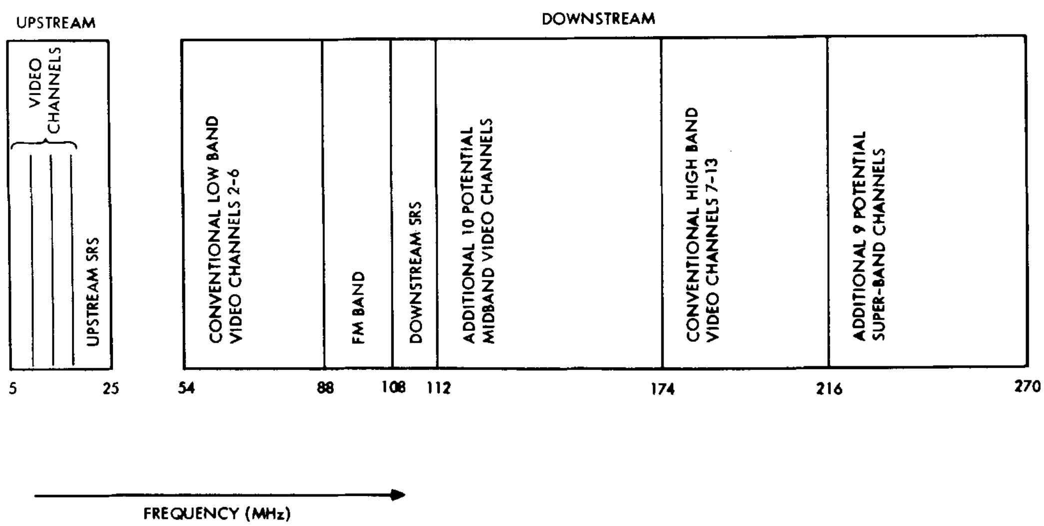


Figure 5. Cable Spectrum Allocation

The upstream signal occupies a 4 MHz bandwidth extending from 21 to 25 MHz. Again the communication is via digital PCM at a data rate of 1 Megabit per second. In this case the digital data is used to phase-shift key (PSK) a 23 MHz carrier.

A typical communications sequence that illustrates the basic principles of operation of the SRS system is shown in Figure 6.

All communications are initiated in the Subscriber Response System at the Local Processing Center. The LPC sends an interrogation message addressed to each subscriber in sequence at a periodic rate. The meaning of the interrogation message is basically the query "Do you have requests?" The Subscriber Terminal will always reply to the interrogation with any of a number of possible requests or statements. The subscriber's replies will be sent upstream

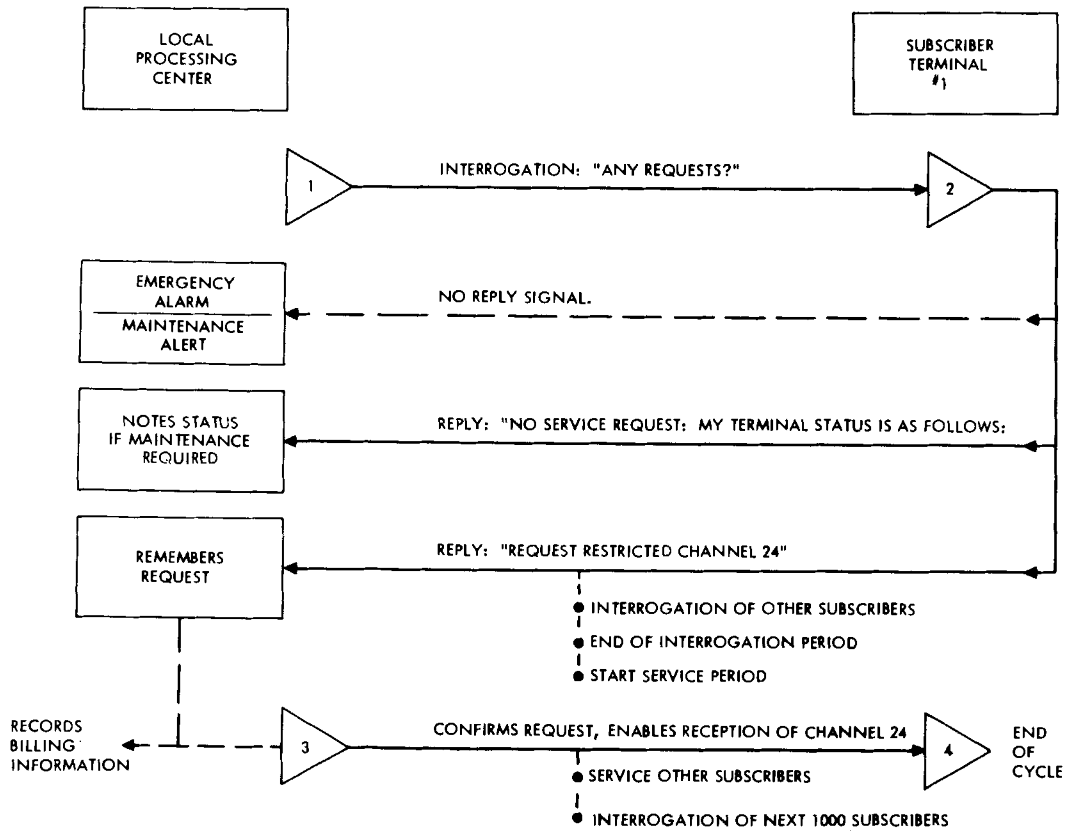


Figure 6. Typical SRS Communications Sequence

bearing the subscriber's address followed by a number of bits devoted to the content of the message. The absence of a return signal from the subscriber will indicate either a physical break in the cable path to his location or a defective Subscriber Terminal. The Local Processing Center will recognize the absence of an expected signal and take appropriate automatic action: it will post a maintenance alert for service personnel and will also flag a potential emergency alarm to cognizant police or protection agencies when such service is requested.

If a particular subscriber has initiated no requests when his terminal is interrogated, the terminal will automatically reply, giving a terminal status report. The terminal status report will indicate the state of the terminal with regard to proper functioning of the terminal circuitry, the condition of accessory devices, and other diagnostic information. The LPC will note the terminal status and take appropriate action.

When a subscriber has initiated a prior request his reply to an interrogation will indicate his address and the particular request, rather than the terminal status report. In the example shown in Figure 6, he requests permission to view a "restricted channel", which might be programmed at that time for a medical lecture restricted to eligible doctors or other eligible professionals. In this case, the LPC will check his request and eligibility to view this program on the particular channel at that time. If the subscriber is eligible, the LPC will remember the subscriber's address for future action. If he is ineligible, the LPC will take no further action.

Each subscriber is interrogated in turn until a group of 1000 subscribers has been processed. Following the interrogation period, the LPC then services the subscribers' requests. In the case illustrated in Figure 6 the LPC will send a downstream message to the subscriber enabling his TV video reception for the restricted channel requested; at the same time the LPC will prepare a billing record (assuming there is a charge for this program) on magnetic tape indicating the subscriber's address, the channel requested, and the time. (Alternate information could be substituted readily for differing requirements). At the end of the weekly or monthly billing period the magnetic tape could be used either at the LPC or another location to prepare actual billing statements for forwarding to the subscriber.

When 1000 subscribers have been interrogated and serviced, the process is repeated for the next 1000 subscribers and so on. The maximum capacity provided in the present system is approximately 65,000 subscribers per Local Processing Center.

For the larger capacity systems which will eventually be required in densely populated metropolitan areas, it may prove more efficient to centralize the Local Processing Center so that it can service a number of Head Ends. The centralized LPC would use a full-sized computer system rather than a minicomputer; it would be a faster unit with greater computing power, more storage capacity, and a greater selection of peripheral devices. Conversely the data handling equipment required at each Head End would be considerably reduced. The central LPC would be interconnected with the Head End two-way data interfaces by cable or microwave relay.

While it was previously stated that the Local Processing Center rather than the subscriber initiated all communication contacts, the communication sequence is so rapid that the subscriber subjective reaction is that he initiates all contacts. Typically for a system of 10,000 active subscribers the time required for a subscriber to receive a reply in response to his manually initiated request (i.e. depressing a key) will be less than 2 seconds, even in the prime times of heavy evening hour traffic.

The SRS system design provides functions necessary to provide a long list of potential services. Even services which appear to be purely visionary at the present time can be provided by later versions of SRS Subscriber Terminals without obsolescence of the earlier terminals or a major redesign of the system.

A listing of services that can be provided by existing designs of the SRS system is shown in Figure 7. As indicated earlier the SRS Console provides remotely controlled channel selection. At the present time, provisions have been included to select 26 possible channels by a voltage-controlled varactor tuner located in the Modem Unit.

- REMOTE CHANNEL SELECTION
- PREMIUM TV
- RESTRICTED TV
- CHANNEL POLLING
- OPINION POLLING
- EMERGENCY ALARMS
- METER READING
- ACCESSORY POWER CONTROL/TIMING
- TWO-WAY MESSAGE CAPABILITY

- SYSTEM DIAGNOSTICS
- SYSTEM CONTROLS
  - MASTER ENABLE/DISABLE
  - TRANSMIT ENABLE/DISABLE
- ERROR RESISTANT TECHNIQUES

Figure 7. SRS Present Services Capability

Six of these channels have been reserved for Premium Television usage, whereby the channel, at the control of the Local Processing Center, may be made available to the subscriber on a fee basis for premium cablecasting such as the showing of first run movies, live dramas, musicals or sporting events. The subscriber is required to indicate by a positive action (depression of a Premium TV key) that he wishes to purchase a program on the channel to which he is tuned. His request is immediately granted by an enabling of the video on his set, and he is automatically billed at the LPC. Provisions have also been included to allow free previewing periods on all Premium channels which can be individually varied in time of occurance and duration by the cable operator.

The restricted channel concept, alluded to earlier, is similar to that of Premium TV. It involves, however, eligibility as an additional requirement. The possible usages are numerous, including use by professional groups and societies, home educational programs, religious or other specialized interest groups, business meetings and home seminars, and adult video programming. Two channels are provided for restricted use in the present SRS design.

Both Premium and restricted channels, however, may be converted back to standard TV channels at any time by control signals from the Local Processing Center.

The restricted channel concept and implementation can be readily extended to include the enabling of groups of channels on a long term or permanent basis. This would allow the cable operator to base monthly rental charges on the number of channels to which the subscriber wishes access.

Channel Polling, somewhat analogous to the well-known Nielson type ratings is provided in the SRS system. In contrast to the limited sampling provided by currently available polling systems, the SRS system can provide polling of all subscribers in a CATV system on a program basis. It can also sample viewer response every few seconds if desired to obtain the response to political announcements, spot commercials, etc. The results of the channel survey can be displayed in a variety of ways. For example, within 5 seconds of the actual survey, a graphics display terminal shown in Figure 8 available as an accessory at the Local Processing Center, can be programmed to produce a histogram, illustrated in Figure 9. A hard copy readout is also available. The demographics of CATV system subscribers can also be correlated with the channel polling results or used to determine selective polling of particular subscribers or groups of subscribers.

Opinion polling in the SRS can range from simple "Yes", "No", or "Undecided" answers to TV video originated questions, to coded replies to more elaborate video or mailed questionnaires. Correlation and display of opinion polling results can be handled as flexibly as the techniques described for channel polling.

Any existing emergency alarm system - fire, intrusion, medical, etc. - can be monitored by the SRS system. The emergency signal is given priority over all other transmission, it is sent by the SRS Modem upstream and notifies the LPC of the address of the subscriber and the time and identification of the alarm or alarms. The LPC verifies the validity of the alarm, by immediate reinterrogations and sends a notification to the proper governmental, medical or protection agency - within seconds after the occurrence. The

Figure 8. Graphics Display Terminal

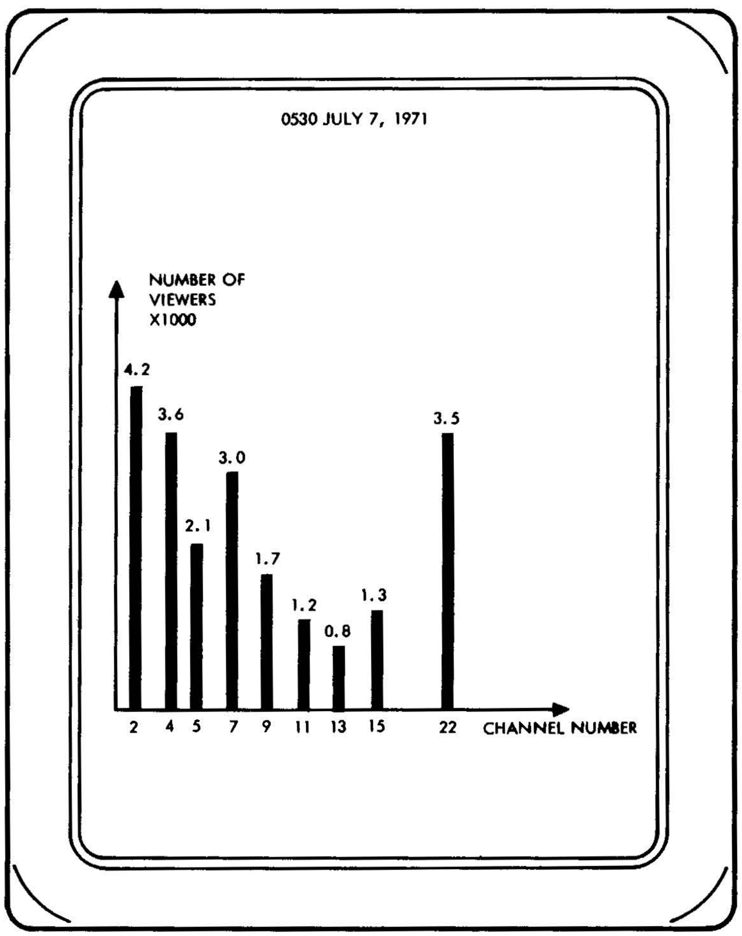


Figure 9. Channel Polling Display

LPC also verifies that proper action has been taken by sending a confirmation signal to the subscriber's terminal which basically resets the terminal releasing it from its locked condition and enabling it to be used normally. As mentioned earlier, a cut cable or non-functioning terminal would be detected at the LPC and treated as an emergency condition.

Various forms of data including meter reading can be automatically read out of the Subscriber Terminal by LPC command. Up to 20 digits per Subscriber Terminal can be transmitted upstream in a single burst which would provide for four-five digit utility meters for example. With an appropriate accessory switching unit, entire apartment dwellings could be read out through one SRS terminal.

By prior arrangement with the Local Processing Center, the subscriber can be provided with accessory power control or timing signals at any desired time. The potential application of this service include the control of power to accessories. For example, a video tape recorder could be turned on automatically to record a desired program. Other uses include wake up alarms, automatic sprinkler systems, and so on.

The SRS system is also provided with a two-way message capability. Upstream messages are initiated by the subscriber at the small numeric keyboard shown in Figure 4. The subscriber may enter messages in groups of up to twenty characters at a time. Assuming the message required more than 20 characters, the "Busy" indicator would light after entering the first 20. With a second or two the "Busy" lamp would be extinguished, indicating the LPC has received the message, and the remainder of the message could then be entered. As the subscriber enters the keyboard data, each character is printed on a half inch paper strip, allowing him to check for errors and providing a permanent hard copy record of purchases or other financial transactions.

Alphanumeric character messages may be transmitted downstream by the LPC, and would also appear on the strip printer. Alternatively, it is planned to offer as an optional accessory at the Subscriber Terminal a paragraph printer which would permit the downstream transmission of lengthy downstream messages at rates of 100 words per minute or greater.

The utilization of two-way message capability has manifold applications, some of which are listed in Figure 10. A full discussion of the applications is beyond the scope of this paper, but the generalized approach involves some form of coding and the insertion into the message of numerical data which specifies the request. The coding required could be furnished by mail in printed catalog form. Naturally occurring numerics such as credit card data, quantities, dates, time of day, etc., could be entered normally. Punctuation marks available from the keyboard would be used to separate logical message groupings for proper interpretation at the Local Processing Center. A typical message sequence using such coding is shown in Figure 11.

- HOME SHOPPING
- EDUCATIONAL INSTRUCTION
- RESERVATION SERVICES
- STOCK MARKET TRANSACTIONS, 'ORTS
- QUIZ SHOWS
- MAIL/ADVERTISING
- DATA BANK ACCESS

Figure 10. SRS Two-Way Message Capability

SUBSCRIBERS REQUEST: MR. R. WILLIAMS WISHES TO PURCHASE 2 TICKETS TO THE L.A. LAKER GAME ON 12/6/71 AT $5.50 EACH IN SECTION 21.

SUBSCRIBERS KEYBOARD ENTRY	<	15	:	120671	: 2	:	550	: 21	>	P
KEYS USED	START	LAKER	SPACE	DATE	QUANTITY		PRICE	LOCATION	END	PURCHASE

CONFIRMATION PRINTED AT SUBSCRIBER TERMINAL: MR. R. WILLIAMS (8705): 2 L.A. LAKER TICKETS HAVE BEEN RESERVED FOR YOU IN SECTION 21 ON 12/6/71 AT $5.50 EACH.

Figure 11. Typical Message Sequence

The discussion of services can be concluded with mention of services which are primarily useful to the CATV system operator, particularly with regard to the maintenance of high reliability of the CATV system.

The system design includes both hardware and software for self diagnosis of system malfunctioning. Provisions, have been included, as previously mentioned to diagnose individual terminal malfunctioning or outages, and cut or defective cables. In addition the terminal can detect and subsequently transmit indications of loss of power, noise bursts, loss of incoming carrier, and downstream parity errors.

The LPC will also detect loss of upstream carrier, noise bursts, power failure, upstream message parity errors, and malfunctioning in the computer and its peripherals. The LPC can also command off the transmitter of any suspected errant subscriber terminal, and can completely disable the terminal if desired for any reason.

Several other error resistant techniques have also been included in the system design which will drastically diminish the probability of error. Among these are dual transmission of all infrequent commands and the requirement for confirmational commands from the LPC to the terminal to indicate the satisfactory completion of particular message sequences.

Two models of the Subscriber Terminal have been planned at the present time. In addition to the SRS-102 model shown previously, a simpler, lower cost version, the SRS-101, is available without the strip printer and associated electronics. This model, shown in Figure 12, has four keys rather than the full keyboard furnished with Model SRS-102. The keys provided will enable purchase of Premium and Restricted TV, opinion polling, and TV shopping with a more limited coding capability. TV Channel Selection is of course retained in this model, and the Modem Unit remains substantially the same as in Model SRS-102. With the exception of the printer deletion, and reduced keyboard size, the SRS-101 can provide all other functions listed in Figure 7. The SRS-101 and SRS-102 are fully compatible and can be used interchangeably in the SRS system.

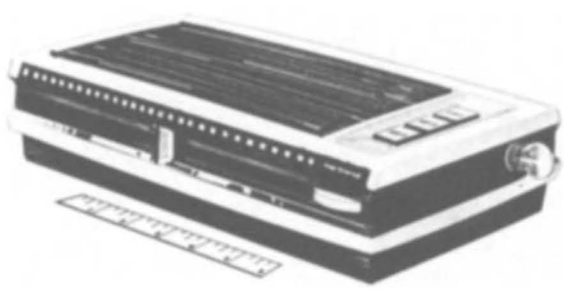


Figure 12. SRS-Model 101 Console

With regard to system growth potential Figure 13 illustrates some of the technically feasible increases in capability which can be modularly added to the SRS as the market demand warrants.

The Hardware

The SRS functional system diagram is shown in Figure 14.

The Local Processing Center described contains a standard PDP-11 mini-computer, including 24K words of core memory and a Model 35 Teletype. An Input-Output Processor has been designed to convert the parallel input-output of the PDP-11 into a serial form for transmission and reception of signals on the cable. For the prototype system the peripheral equipment includes a time of day clock, a 256 K word disc memory, a 7 track magnetic tape drive unit a card reader and a graphics display terminal together with hard copy readout.

PRESENT MODELS

SRS-101 FUNCTIONAL KEYBOARD

SRS-102 NUMERIC KEYBOARD WITH INTEGRAL STRIP PRINTER

OPTIONAL ACCESSORIES

100 WPM IMPACTLESS PARAGRAPH PRINTER

EMERGENCY ALARM INTERFACE

METER READING INTERFACE

EXTERNAL POWER CONTROL AND TIMER UNIT

POTENTIAL FUTURE CONFIGURATION

SRS-103 ALPHANUMERIC KEYBOARD WITH INTEGRAL PARAGRAPH PRINTER

OPTIONAL ADDITIONAL ACCESSORIES AND FEATURES

FRAME GRABBER

TWO-WAY AUDIO CAPABILITY

MULTI-TV SET CAPABILITY

Figure 13. SRS System Growth Potential

The required type and number of peripheral equipments for operational use will, of course, depend on the nature and number of the services provided, the number of subscribers in the system and the anticipated or actual traffic volume. Modular expansion capability is also provided in the LPC and software design to accommodate such growth situations.

Inputs to the computer can be made by punched cards or paper tape, or by manual operation of the teletypewriter.

The LPC system is designed to operate unattended except for routine maintenance. Manual inputs for special programming can be made by remote teletype interconnection or by use of the Model 35 Teletype at the LPC.

At the Subscriber Terminal, the downstream TV video is separated from the 110 MHz SRS signal and is routed to the varactor tuned frequency converter which furnishes the video input to the TV set. The 110 MHz signal is demodulated and 1 Megabit digital data processed in the Modem Unit to furnish signals to the frequency converter, external accessories or to the Operator's Console depending on the particular message sequence.

Upstream data originates at the Operator's Console, or at the accessory sensors and devices. It is encoded and stored at the Modem to await upstream transmission. In transmission the data is fed to a 23 MHz phase modulator, passed through a low pass filter and on to the subscriber's drop cable.

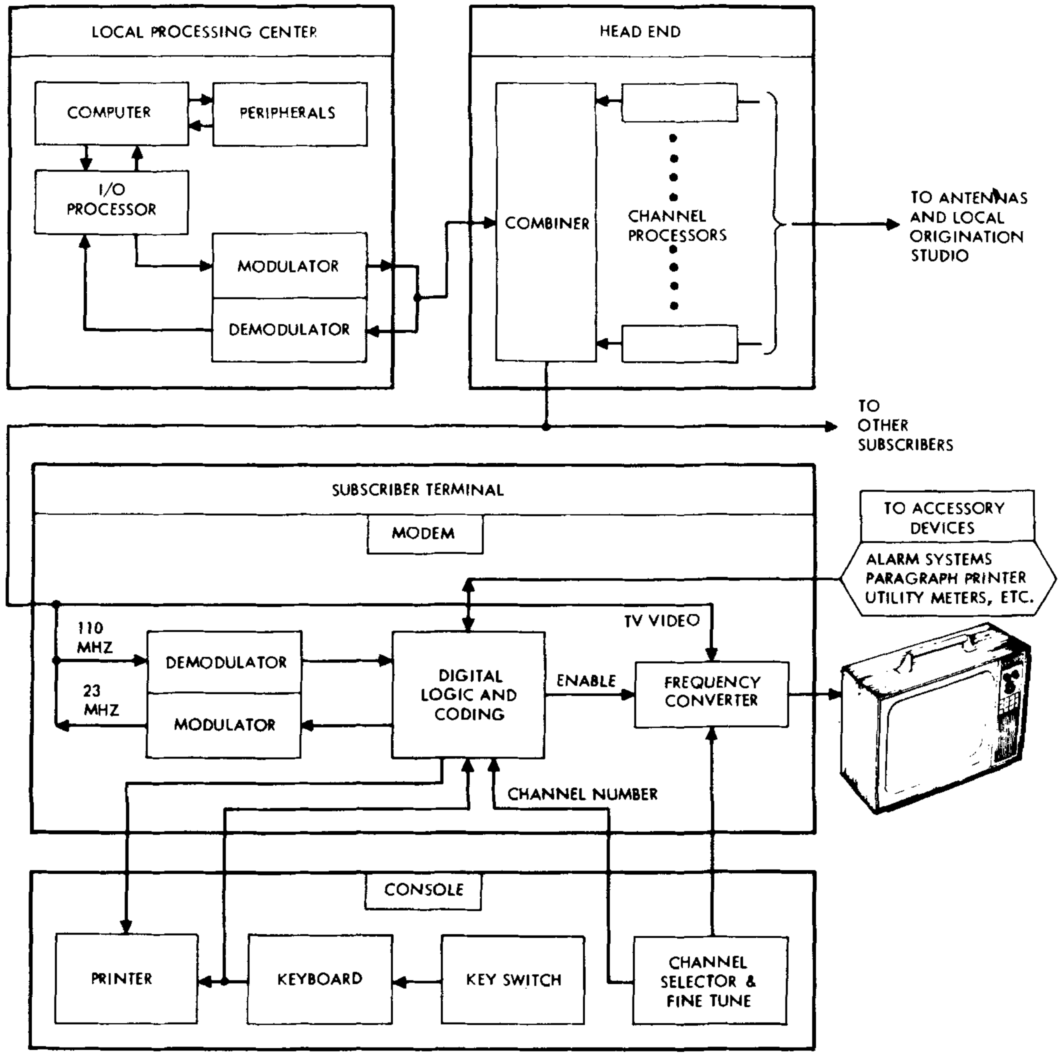


Figure 14. System Functional Block Diagram

Integrated circuitry of the TTL and MOS varieties is used widely in the Prototype Subscriber Terminal. Future plans for large quantity production include the use of large scale integration to further reduce size, cost, and power consumption and to increase system reliability.

Status

Some of the areas of technical concern in the design of any digital data communications system are:

1. A determination of the required bandwidths
2. The required bit error rate.
3. The effects of impulse and other noise forms.

4. The effects of various transmission deviations caused by the cable system and its associated components on system performance.

In designing the SRS system, these concerns were taken into account as were human factors considerations in designing the Subscriber Terminal.

In order to confirm design expectations and to answer questions that can only be resolved by actual trial, a demonstration system was completed early in 1971. This system consisted of a Modem and an Operator's Console shown in Figure 15 and 16. In addition, a small simulator (Figure 17) was designed to produce driving stimuli for the terminal to simulate desired features of the LPC. The demonstration system has been extensively tested both at the Hughes facilities in the Culver City area and on an experimental Tele-Prompter two-way cable system at Los Gatos, California. While this testing is continuing, the results to date have validated the design concept. All of the present system functions described previously herein were successfully demonstrated repeatedly even in the presence of near band burst transmissions in close proximity to the equipment. While measurements are still incomplete bit-error rates appear to be in the 10^{-6} to 10^{-7} range which correlates with expectations. Message errors which remain undetected in the system are expected to be orders of magnitude less than the bit error rates as a result of the error compensating message sequencing and verification techniques employed.

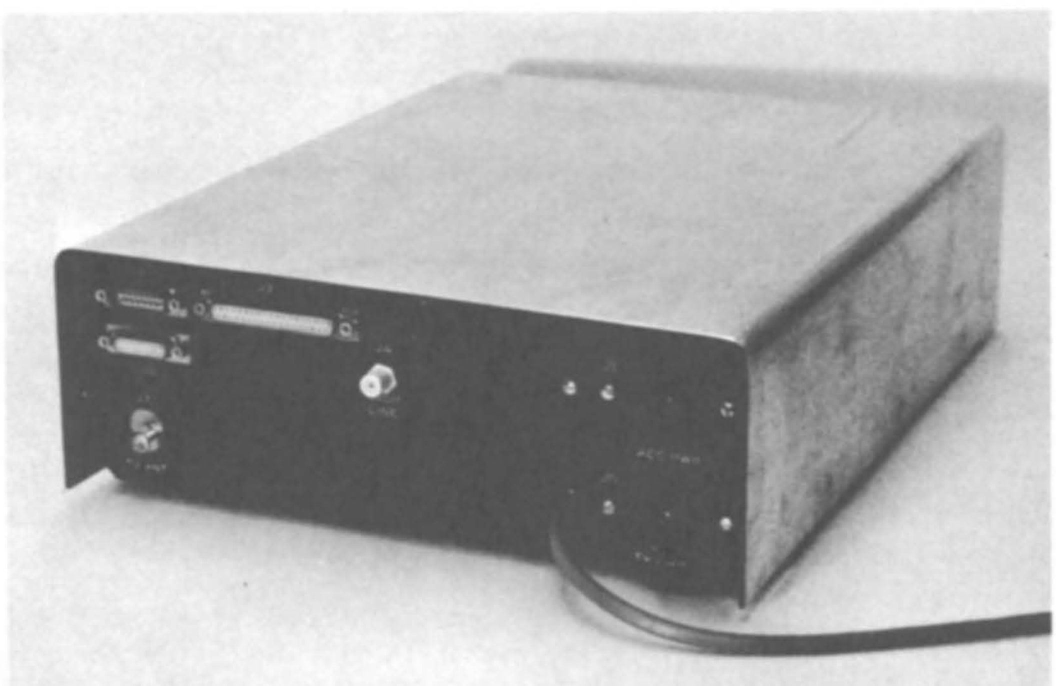


Figure 15. Subscriber Terminal - Demonstration Modem

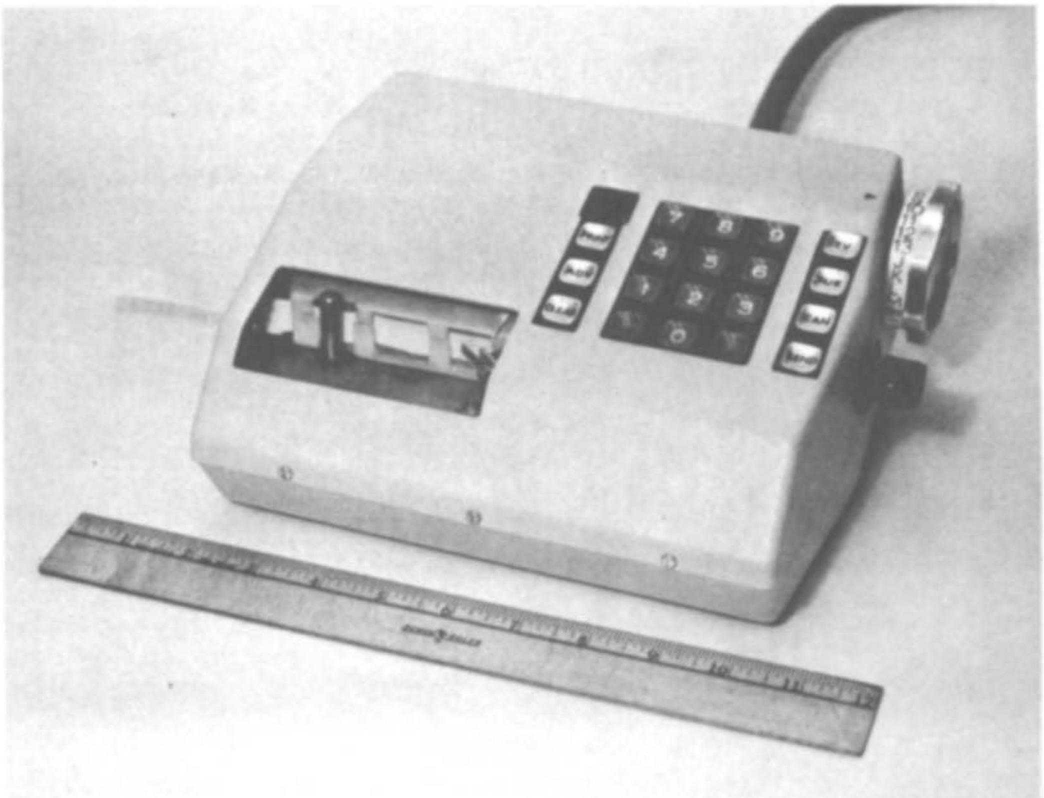


Figure 16. Subscriber Terminal - Demonstration Console

At the present time software designs are being completed and hardware is in production to permit a small scale but intensive field test with approximately twenty-five SRS Subscriber Terminals and a fully equipped Local Processing Center. The test will be conducted in the Los Angeles area on an operating cable system starting in the last quarter of 1971. It is expected that the test will include both technical and consumer oriented features to demonstrate the SRS system capability and to obtain the consumer reaction to the services offered.

Pending the results of the small sample to be tested later this year, and continued consumer surveys, plans are being formulated for the large scale production of SRS equipment by the Theta-Comm Company. This equipment will be used in a system wide test in 1972.

Conclusion

While it is impossible to predict the rate of public acceptance of two-way systems, we believe that they will be widely accepted. With this conviction, and a strong commitment to the future of communications, we submit that any two-way mass communication system worthy of serious consideration must meet the following vital requirement.

Figure 17. LPC Simulator

The system must have a well thought out approach allowing for the impossibility of predicting the market accurately.

- It must not be limited by prejudgement of the relative saleability of services.
- It must be flexible to change without becoming unwieldy or obsolescent.
- It must be economically viable as well as technically sound and reliable.
- It must deliver what is promised not only from a hardware standpoint but even more importantly from a software view.

We believe that the Subscriber Response System meets these requirements. It has not only the basic capabilities to meet a wide variety of existing needs with minimum cost but also the growth potential to mature with the market in whatever direction it may develop.

TOCOM SYSTEM

BI-DIRECTIONAL CABLE TELEVISION

INFORMATION AND CONTROL TRANSMISSION SYSTEM

Presented by

WILLIAM F. (BILL) OSBORN

CAS MANUFACTURING CO.

IRVING, TEXAS

ABOUT THE AUTHOR

Bill Osborn graduated from Texas A&M University in 1957 with a BSEE after serving for several years as an Army Officer. He is a Professional Registered Engineer and a member of Tou Beta Pu, Eta Kappa Nu, Texas Society of Professional Engineers and Rotary International.

After graduation, Mr. Osborn worked for Sandia Corporation and was responsible for establishing a secondary standards laboratory for the AEC. He then joined National Data Processing Corporation in Dallas and was involved in the design of one of the initial automatic banking and check sorting systems. In 1962, Mr. Osborn joined Arps Corporation as Vice President of Engineering. During his tenure with Arps he was responsible for the development of logging tools for the oil industry.

Mr. Osborn has been involved in numerous projects, including automation and control systems, instrumentation systems, computer systems, and communication and broadcast systems. Mr. Osborn is the inventor of several patents relating to instrumentation and the oil industry.

INTRODUCTION

TOCOM is the NOW TOTAL CATV COMMUNICATIONS SYSTEM developed by CAS Manufacturing Company. Bi-Directional Flow of information on a single cable, particularly the ability to transmit information from the subscriber to a central receiving point and to transmit control information to each subscriber location, is the end product of the TOCOM System.

TOCOM is a broad band, single cable, bi-directional CATV Communications System providing conventional 26 channel forward transmission, a 26 channel converter receiver, a crystal controlled subscriber identified digital transmitter, built into the converter, a hub located computer interface interrogator and master computer memory bank.

TOCOM is the vehicle to provide home protection systems, pay television, surveys for television rating service, controlled television channels, meter reading, amplifier level monitoring, and instant "subscriber response" polls, via automatic computer read-out and billing. Remote use of computers from the home and narrow band picture phone are possible future uses. Providing any one service would not economically justify a total communications system. Providing all services is an open door to profitably increasing subscriber revenues.

TOCOM SYSTEM - BI-DIRECTIONAL CABLE TELEVISION INFORMATION AND CONTROL TRANSMISSION SYSTEM

The TOCOM system consists of three primary elements - a Central Data Terminal, a Bi-directional coax amplifier system, and a large number of Remote Transmitter Receiver units. In general, the TOCOM system has the capability of transmitting from the Central Data Terminal, interrogation information to one or more selected Remote Transmitter Receivers. This causes the Remote Transmitter Receiving unit or units that have been interrogated by the interrogation signals to sample certain data and to transmit this information back to the Central Data Terminal, with all signals being transmitted on a single coax cable.

The system as it is presently designed, though expandable, has the capability at any Remote Transmitter Receiver location of interrogating seven words of information, each word containing 16 bits. In the present system, the seven words of 16 bits are so coded to return certain specific information; however, the seven words of 16 bits could be coded to return any information so desired. In addition to being able to obtain the interrogated information, the system also has certain other capabilities which will be discussed later in this paper.

The system as it is presently designed and presently coded at each Remote Transmitter Receiver location, will interrogate the following information. Referring to Figure 1, the system can determine if there is a fire, ambulance or police alarm activated at the location; it can determine to what channel the set is tuned; it can determine if the selected Pay TV channels have been authorized by the User; it can determine what opinion the user has (in conjunction with video signal the user is asked a question, he then will make his selection by pushing one of three buttons on the Remote Transmitter Receiver, thus indicating that his opinion is NO - YES - or NO OPINION). Also the system can determine if the TV set is on or off. In addition the Remote Transmitter Receiver unit, when connected to kilowatt hour meter, gas meter, water meter, or any other such type of device, can automatically read the meter or meters. In addition to being able to determine the specific information mentioned above, the system in conjunction with a digital computer that is located at the Central Data Terminal, can disable or enable select remote units, can determine where faults occur in the line, and determine if any specific Remote Transmitter Receiver unit has failed.

For the future of the system, with very little modification, we see the capability of not only interrogating information from the remote locations, but also the control of devices at remote locations. For example, to turn on or off heating system or air conditioning systems at a given time, we can feed the cat, wake up a person, turn on the coffee pot, or any number of things that would be desirable to have control of from a Central location on an automatic basis. In general, the system has the capability of control and interrogation of remote units be they located in a CATV system or an industrial complex. It is possible with very little modification for the system to interrogate more information if desired. For example, we can include parity bits for reliability purposes, we can measure analog signals, we can increase the number of words and bits, and if necessary, we can increase the sample rate considerably over and above the rate the system is now presently operating.

The system, as it is now operated, has the capability of controlling one thousand Remote Transmitter Receivers per group with the capability of 30 groups, that is 30,000 Remote Transmitter Receivers on each trunk line and further the capability of handling in effect any number of trunk lines. So in general we may say that the system, as designed, has the capability of handling roughly 120,000 – 180,000 Remote Transmitter Receivers. This is a round number; we can control considerably less or considerably more, if necessary, but we feel from a practical standpoint this large number is really in excess of what will be necessary on any one particular system. The speed of the system is such that to sample 1,000 Remote Transmitter Receivers, or up to 180,000 Remote Transmitter Receivers, will only require 30 seconds. What this really amounts to, is that we will sample on a simultaneous basis, more than one Remote Transmitter Receiver. It requires approximately 30 milliseconds to sample one Remote Transmitter Receiver, obtaining from that Remote Transmitter Receiver, one 16 bit word. We can readily see that if we sample more than one Remote Transmitter Receiver at any one time then, in effect, our sample rate goes up such that we still require 30 milliseconds per word sample but sampling more than one Remote Transmitter Receiver at any one time enables us to go up to the sample rate of 180,000 Remote Transmitter Receivers in 30 seconds. This can be seen by referring to Fig. 2, whereby a Central Data Terminal is controlling N number of trunk lines with up to 30 groups of Remote Transmitters on each trunk line and each group on each trunk line containing up to one thousand Remote Transmitter Receiver units. You will notice in the Fig. 2, that from the Central Data Terminal, we are indicating that we have information flow to police, fire, ambulance, power company, etc. In general, what this indicates is that from the computer located in the Central Data Terminal, on command from the computer based upon information received from the Remote Transmitter Receivers, we can automatically alert the police dept., fire dept., ambulance company; also, we can send data to the power company, water company, etc., such as to meter readings, etc.

In order to understand how this system operates, referring to Fig. 3, there is an indication of the format of the transmission of the system. The Central Data Terminal transmits interrogation information to the remote transmitter receivers with selected frequency coding in the 50 megacycle range. This information is received at each Remote Transmitter Receiver, operated on accordingly, and then transmits information back to the Central Data Terminal in the 6 to 30 megacycle region. The transmission back to the Central Data Terminal is as follows: Each group of Remote Transmitter Receivers on each trunk line is assigned a specific frequency; for example, 10 megacycles or 10.8 megacycles, 12 megacycles, 13.2 mgc., 14.7 mgc. (I might add that these frequencies have been selected such that harmonics fall in between the upper channels such that we don't end up with birdies in the video system.) Referring to Fig. 3, the operation or interrogation of any one Remote Transmitter Receiver unit is as follows: The Central Data Terminal transmits a master reset signal. This causes the remote transmitter receivers throughout the system to come to what we refer to as the initial state or reset state. We then transmit an ID Enable signal. This enables all the remote transmitters in the entire system to receive an ID code, which is 10 bits long, which is then transmitted to all of the Remote Transmitter Receivers. Each Remote Transmitter Receiver decodes these signals and, depending upon how it is decoded, in each Remote Transmitter Receiver, reacts to a particular ID code. For example, Remote Transmitter Receiver No. 1

in Group 1, and No. 1 in Group 2, and No. 1 in Group 3, etc., would all have the same ID code. When a Remote Transmitter Receiver receives it's code, it then, in effect, enables itself to say "OK I am the particular remote unit you are talking to, please send additional information." At this point, the Central Data Terminal transmits an additional signal which identifies the particular word that we would like to have interrogated from the Remote Transmitter Receivers which have received and identified themselves from the previous ID code. At this point, the particular switches, turned-on transistors, or whatever device we are interrogating, are enabled, or in effect, the information from these units, is transferred into the Remote Transmitter Receiver and at this point we then transmit 16 data shift bits from the Central Data Terminal. This causes the 16 bits of information which have been, in effect, brought into the Remote Transmitter Receiver from the combination ID code and word code and are caused to be shifted out, or in effect to be transmitted to the Central Data Terminal. The actual transmission is caused by the 16 bits turning on or off the appropriate 6 mghz oscillator, to 30 mghz oscillator, that is contained in the Remote Transmitter Receiver. This information is received at the Central Data Terminal and operated on accordingly.

In order to better understand the operation of the system, referring to Fig. 4, is a block diagram of a Remote Transmitter Receiving unit. At the remote unit, the information in the 52 mgc region and up is in effect bypassed through a by-pass filter and transmitted on to a conventional 26 channel converter which converts a particular channel, that is Ch. 2, 4, 6, 8, whatever it may be, to Ch. 12 and is transmitted to the users TV set. The 50 mgc interrogation information is brought to an RF section which is a portion of the Remote Transmitter Receiver unit. Here the interrogation information is decoded in such a manner as to cause the appropriate information to be sent either the ID register, the word code register, or the data register. As previously explained, the ID register, when it recognizes its particular code enables the output remote transmitter. In addition, the word code enables a particular set of switches or devices to which we are interested in interrogating, i.e., the alarm, channel numbers, etc. This information is transferred into the data register and then, in conjunction with the 16 shift pulses, operates on a transmitter such that the 6 to 30 mghz, whatever the transmitter frequency happens to be, is returned back down the co-ax to the Central Data Terminal.

At this point I would like to add a few comments regarding the features of the system which we feel are unique. One is in reference to the opinion polls. Located on the front panel of the remote transmitter receiver are 3 push buttons labelled No - Yes - or No Opinion. The sequence of events of operation of these switches is as follows: From the video portion of the program a viewer is asked a particular question - what is his opinion about this or that - at this time the Central Data Terminal transmits a particular code to each Remote Transmitter Receiver which, in effect causes the opinion circuits in all Remote Transmitter Receivers to be reset. The purpose of this is such that as each Remote Transmitter Receiver is interrogated, it will be necessary for someone to have pushed the opinion button, just prior to the time that the opinion poll is taken. The purpose behind this is to not obtain an opinion from every set but only those that are properly activated by one of the users. Following this sequence of events a little further, the time has just occurred where we have sent the reset information to each opinion circuit in each Remote Transmitter Receiver; the viewer has been asked the question, and at this time let's say for example he pushes the Yes Opinion button. Assuming a moment later, he decides No, I really meant No, then at that time he pushes the No button, this will cause the internal circuitry to, in effect, reset his Yes Opinion and to set his No Opinion, hence the person who has given his opinion has a short period of time to change his mind if he so desires. Also I would like to emphasize the condition that resetting this opinion circuit enables us only to obtain the opinion from people who actually activated the opinion circuits and not obtain an opinion that was, in effect, set by a child two hours before the program occurred. Referring again to Fig. 4, there is a certain portion of the remote transmitter receiver which is marked test. The purpose of this portion of the system is to enable the Central Data Terminal to send out certain selected information to each remote transmitter whereby we may in one condition disable every remote transmitter receiver throughout the system; we can, secondly, enable every Remote Transmitter Receiver on the system, or we can, by transmitting an identification code, along with a separate code, selectively enable or disable each respective remote transmitter unit. I will not get into the details of this but, from a maintenance point of view, this can be extremely helpful in operating on the system, if and when failures occur.

The Remote Transmitter Receiver units are so designed that the maintenance personnel, if a remote transmitter receiver fails, can in a matter of a very few moments change out a Remote Transmitter Receiver and have a new one in operation. To accomplish this simply required one to disconnect the coax coming into the unit from the CATV system, to disconnect the cable running to the antenna input on the TV set, and the connectors going to the remote alarm units, water meter, etc. The maintenance personnel simply reverses this process and has to do one other item, that is to place three wires that are in the remote transmitter receiver digital section to certain pins which, in effect, identifies that particular remote transmitter receiver with a particular identification code. Now, in most systems where it is necessary to, in effect, have an identification code, requires the soldering or connecting of as high as 48 to 64 wires in the proper location. We have devised a system here whereby only three wires need to be soldered in place. Further, the system is simple enough that anybody that can subtract can accomplish this very readily. I will not get into the details of this at the moment but I would like to emphasize that the system is designed to accommodate not only the user but to facilitate maintenance, of the system. In regard to the maintenance and repair of the systems, in reference to the Remote Transmitter Receiver unit, the console at the Central Data Terminal is so designed that a Remote Transmitter Receiver unit that has failed can be brought into the console, plugged in and it can be immediately determined if the RF section or the digital section of the remote transmitter receiver has failed. Depending upon which area of the unit has failed, there is a cook-book routine which can immediately isolate which portion of the system has failed and the repair technician can take the appropriate action at that time.

Referring to Fig. 5, which is a block diagram of the Central Data Terminal, the Central Data Terminal consists of really three major elements. 1. The Central Data Processor, 2. a Hard Wire Controller/Display system and the RF system, and of course, the normal

TV head-end system. There has been extreme care put forth into the design of the Central Data Terminal in order to obtain maximum use of equipment, ease of maintenance, ease of operation, etc.

The system, under normal operation, would be controlled from the Central Data Processing unit as follows:
The Central Data Processer tells the Hard Wire Controller, I would like to have the information from certain remote transmitter units whose ID code is such and such and I would like to receive words 1, 2, 3, 4 or whatever it may be. The Hard Wire Controller at this point takes over, and with this information, actuates the 50 mghz RF transmitter which sends out the interrogation information to the remote transmitter receiver. The information is returned then from the appropriate Remote Transmitter Receiver units and is brought into the display portion of the system. At this time the Hard Wire Controller tells the computer, I have the information - come get it. At this point then the Central Data Processor will bring in the 16 bit words from each remote transmitter receiver which was interrogated and then operate on this particular information as it so programmed.

Now, I would like to emphasize that the system was designed to operate normally on this basis, such that the Central Data Processor then has the maximum amount of free time to do other functions, such as bookkeeping, statistical analysis, etc.

However, it is possible to operate under two other modes of operation. The second is what we call the direct mode. In this type of operation, the Central Data Processor bypasses the Hard Wire Controller and directly controls the 50 mghz RF transmitter system, and can control the head-end system and the display system directly. Under these conditions, the Central Data Processor will then cause the 50 mgc RF transmitter to send out the interrogation pulses, the information will be received back and the Central Data Processor then obtains the information from the Display portion of the system. This design is to enable operation of the system if for some reason the Hard Wire Controller were to fail.

The third mode of operation of the system is a semi-manual mode. In this mode of operation an operator can take entire control of the system by going to the console, operating certain switches, which will indicate to the system that he desires to interrogate a particular Remote Transmitter Receiver and he desires certain information to be returned from that Remote Transmitter Receiver. This semi-automatic mode was designed into the system such that, in effect, the system could go off-line from the computer; monitor, for example, the alarm conditions on a continuous basis and leave the Central Data Processor free for other purposes so desired. In addition it gives a back-up if the Central Data Processor were to fail.

By using a Central Data Processor, i.e., a digital computer in the system, it is possible to do any number of things with the data received. In effect, we open up a pandora's box as far as capabilities. For example, the Central Data Processor can;

1. detect when an alarm condition occurs, and via an automatic system call the fire dept., police dept., or ambulance company and alert them to the fact that there is a fire, burglary, or whatever it may be at a specific location. In addition,

in the case of a fire, some additional information may be transmitted to the fire department. For example, we may alert the fire department that the fire is at a Paint Factory, an Old Age Home, or a Hospital and is a particularly critical situation. Or, for example, it may alert the police dept. that there is a burglary occurring at a jewelry store which would require a little more haste than if it were a warehouse containing newspaper.

Further, with the use of the Central Data Processor, we can determine any number of different types of statistical information. For example, in reference to opinion polls, we can determine by area who had what opinion. For example, does the south side of town have a different opinion than that of the east, north, west, etc. Since we are able to determine on a real time basis to what channel each set is tuned and whether the set is on or off it is easy to determine a TV rating type poll. For example, how many people are watching program A, how many are watching program B. I might add at this point, though we can't guarantee that certain people are watching the TV set but we can determine the TV set is turned on and tuned to a specific channel. It is quite feasible that by knowing this, and by use of the opinion poll portion of the system, we could determine more definite information. We could transmit an opinion poll asking people to respond by asking people to push the yes button if two or more adults are watching the program or push the no button if children under 12 are watching the program, etc. In reference to Pay TV, since we can determine who is turned to what channel on a real time basis, and that the Pay TV authorization switch is turned on we can automatically control Pay TV as far as billing. If customer 741 in Group 1 turns on his switch and turns to Ch 13, which is a Pay TV channel at 10:30 in the morning, then we will know within 30 seconds of when he turns his set on and we will know within 30 seconds of when he authorized and tuned to the Pay TV channel, and likewise we will know within 30 seconds of when he changed to a different channel and can, via the computer, automatically bill him for that particular Pay TV portion of the program.

In addition, with the ability to read the kilowatt-hour meter, water meters, etc., and in conjunction with the digital computer, it is possible for us to read the meters, simply store the raw date, transmit this data to the power company, water company, etc. or to calculate the bills, punch these out on punch cards, similar to what all of us receive in the mail every month. In effect, handle this portion of the reading of the meters and automatic billing in any way that the power company, water company, etc. so desires.

In general, and briefly speaking, it is possible, using the Central Data Processor, to operate on information in most any way we so desire. To bring this information out on telephone lines to remote locations, to another computer, or we can bring information out on punch tape, type it out on a teletype, we can put it out on disc memory, or mag tape, or punch cards. In effect, almost any manner or method that is so desired by the customer or by the operator, whichever the case may be.

I might further add that by programming the computer, it will act as a very powerful maintenance tool. If we interrogate certain Remote Transmitter Receivers, and we do not get any information back from a selected group, by proper programming we can readily determine that the probability is we have a line failure at a given point.

Likewise, it is true that if we interrogate a certain Remote Transmitter Receiver and have been unsuccessful in receiving any return signals from that unit, it is fairly obvious that that Remote Transmitter Receiver is malfunctioning. We can pinpoint this, and print out on a teletype and say to a repair man that the Remote Transmitter Receiver at 1701 Main Street has failed, or whatever the case may be. Further down the line we see the possibility and probability of placing Remote Transmitter Receivers along the coax system and monitoring such things as the AGC, the amplifier's gain, etc. Monitoring these measurements to determine if they fall within certain limits or detect changes such that if we see any type of failure we can alert the repair people, have them change out the failing amplifier, etc., before it actually fails.

Regarding the economics of the system, it is our opinion that with the extensive capability of this system, we feel that the economics are such that a good profit can be received by the operator in using the TOCOM system. For example, that the public is very ready for an effective, automatic, burglar alarm - fire alarm type system, because, in effect, we can monitor everybodys house that is on the system every 30 seconds and determine if they have a fire, burglary, etc. The TOCOM system is designed to readily accept the appropriate output signals from most fire alarms, burglar alarms, etc., or we can provide the system.

We feel that the ability to obtain opinions is an excellent revenue source from two points of view. 1. There are large quantities of money spent daily obtaining opinions from people - what kind of soap do you use, etc. Our system can easily be expanded - or used as it is to obtain such information. We can obtain this information, operate on it, and bring it out in a useable form in a matter of minutes where by in most systems the information is days, weeks and in some cases months old before it is ever reduced to a useable form. In regard to the opinion poll, we feel that this can be used effectively in the psychology of people in the effort to put the Remote Transmitter Receivers in every home. For example, if housewife A watches an opinion program every day at 10:30 and she is giving her opinions, you can bet that housewife B next door is also going to have to have a Remote Transmitter Receiver unit in order for her to give her opinion as is housewife A.

From the point of view of Pay TV, we feel that this is the first system that is not only economical but useable in regards to an effective pay TV system. Keeping in mind that we know on a real time basis, who is watching what and when. Further knowing who is watching what and when, it gives us for the first time an effective TV rating service. This portion of the system alone should, properly used, bring in considerable revenue to the operator of a TOCOM system.

I believe that it is essential that it be kept in mind that the TOCOM system, although it is presently designed and built to monitor specific functions, does have the capability of interrogating literally any information at remote points and transmitting this information back to a Central Data Terminal. We have given considerable thought to the various uses of this system, as applied to the CATV business, however, I feel sure that as time goes on, other uses will be brought to our attention and I am sure, as the system is designed, it can be easily modified to accommodate these needs.

I would like to further add that the TOCOM system, in addition to use in the CATV industry has a large potential in other areas, such as automation of oil fields, pipe lines, utilities, power plants, water plants, factories; any place where it requires the control of elements at a remote point from a control console. The TOCOM system has the advantage over presently used systems, if for no other reason than the ability to run one coax cable in lieu of running literally hundreds of twisted pair of telephone cables as is now the common situation.

In reference to the cost of the system, we have not at this point determined hard-set costs for the Remote Transmitter Receiver or Central Data Terminal, etc. However, we feel that the Remote Transmitter Receivers, which do include the converters, will cost in the neighborhood of $100.00 each (approximately) and will vary from that value up or down, depending on quantities. In reference to the cost of the Central Data Terminal, this cost can range anywhere from probably $80,000 and up, depending on the requirements, and in general will be dictated a great deal by the number of people to be controlled or interrogated and to the amount of programming necessary to program the processor in order to accomplish the functions desired by the operator.

SYSTEM CAPABILITY

GENERAL: INTERROGATE SEVEN SIXTEEN BIT WORDS FROM EVERY REMOTE UNIT.

PRESENT SYSTEM DETERMINES:

1. FIRE
2. AMBULANCE
3. POLICE
4. CHANNEL NUMBER
5. PAY TV AUTHORIZED
6. OPINION CONTROL WITH RESET CONDITION
7. SET ON/OFF
8. KW-HOUR METER READING
9. GAS METER READING
10. WATER METER READING
11. DISABLE OR ENABLE SELECTED REMOTE UNITS
12. FAULT ISOLATION

FUTURE:

1. CONTROL OF DEVICES AT EACH REMOTE LOCATION
2. MONITORING OF LINE CONDITION
3. IN GENERAL - CONTROL OR INTERROGATION OF REMOTE LOCATIONS

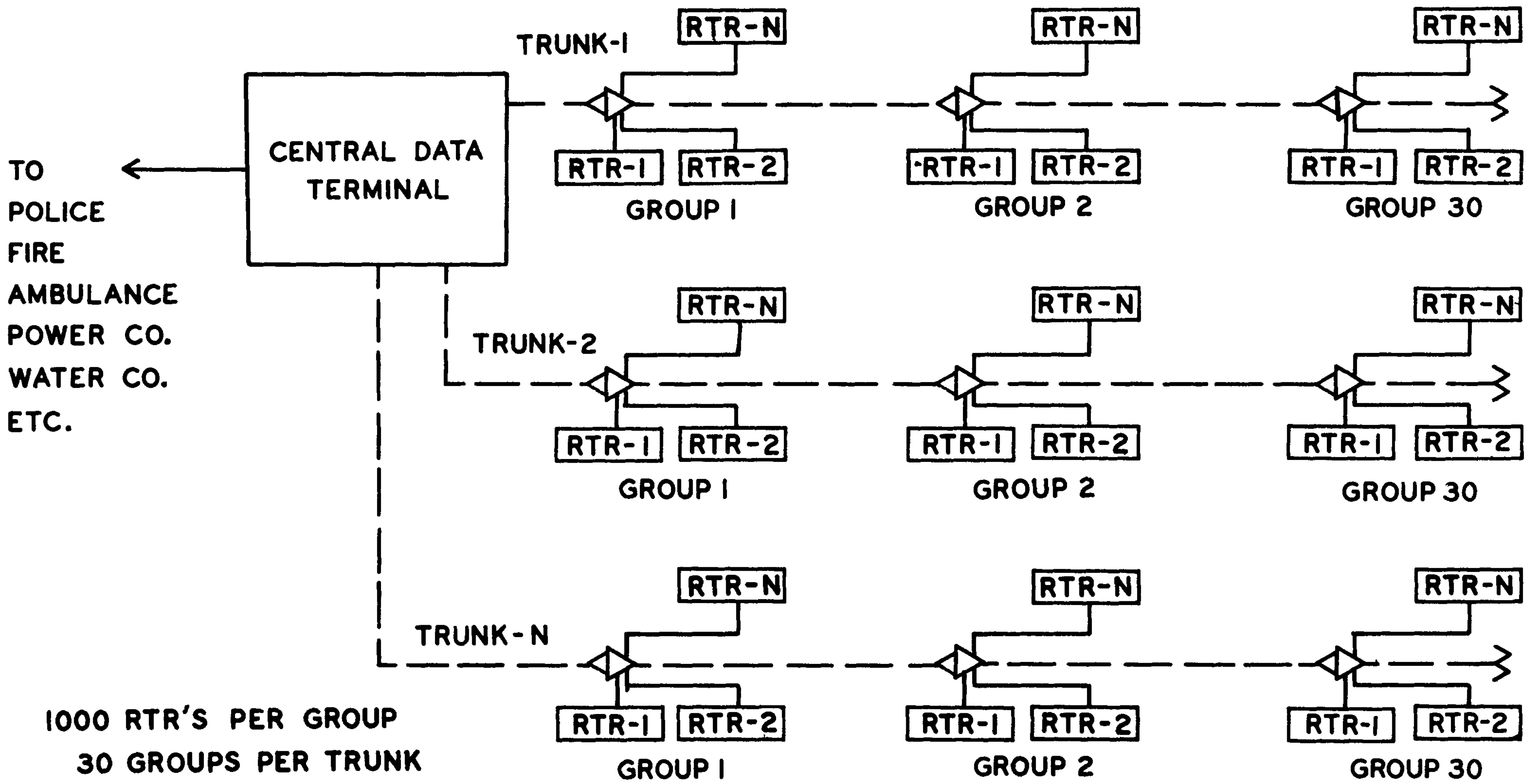
SYSTEM CAN EASILY BE MODIFIED TO:

1. INTERROGATE MORE INFORMATION
2. INCLUDE PARITY
3. MEASURE ANALOG SIGNALS
4. INCREASE SAMPLE RATE

FIGURE I

TOCOM

SYSTEM BLOCK DIAGRAM



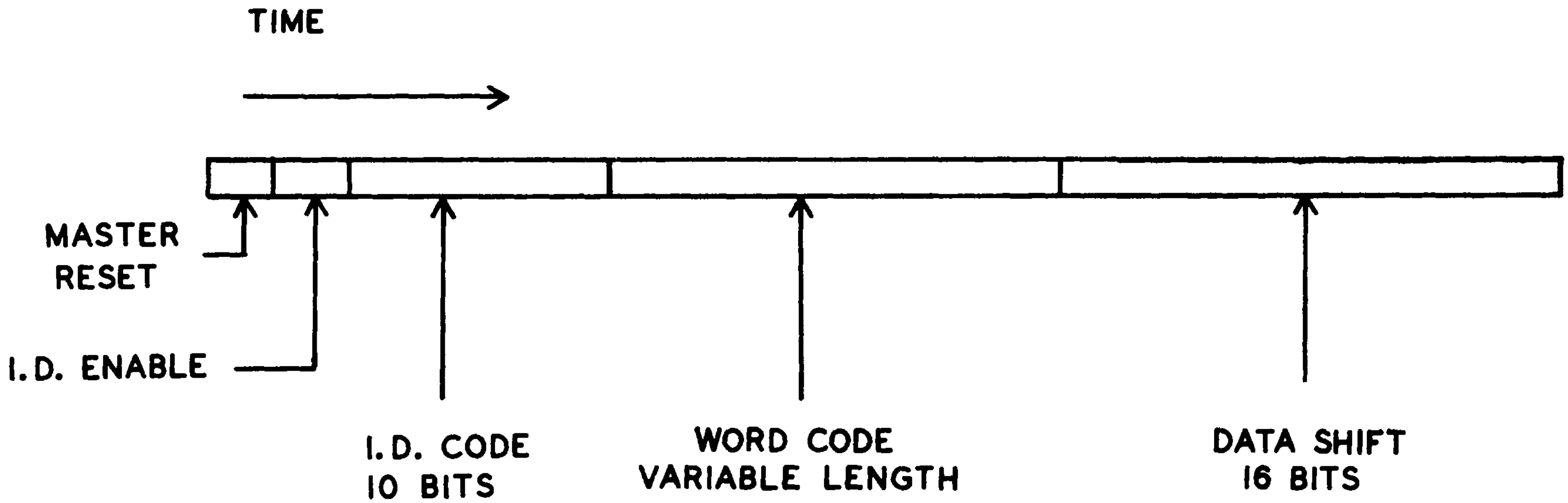
1000 RTR'S PER GROUP
30 GROUPS PER TRUNK
N TRUNKS

TOTAL RTR CAPACITY = 180,000

TOTAL INTERROGATION TIME 30 SECS.

FIGURE 2

TOCOM



SIGNAL FORMAT

FIGURE 3

TOCOM

REMOTE TRANSMITTER RECEIVER

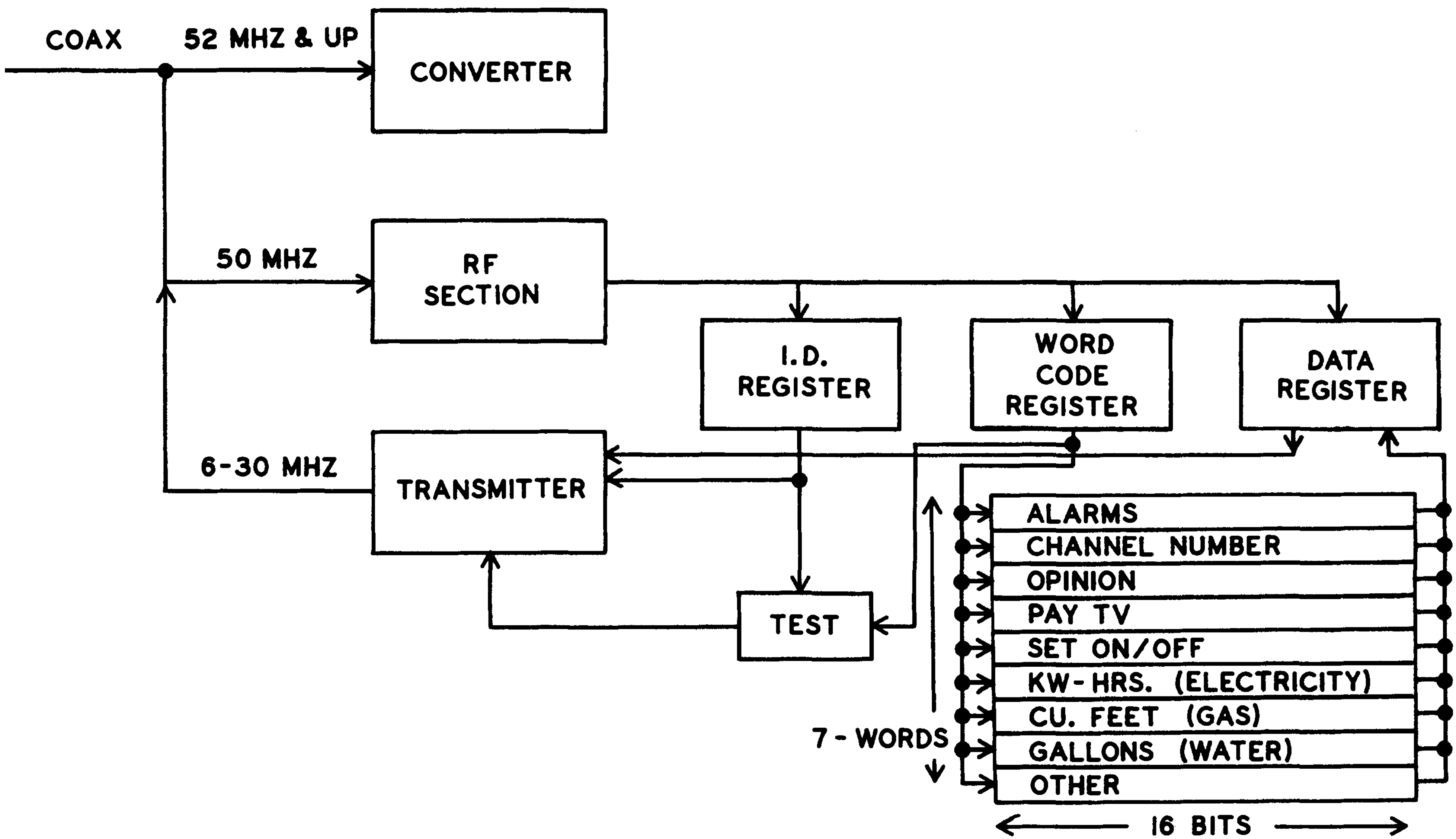


FIGURE 4

TOCOM

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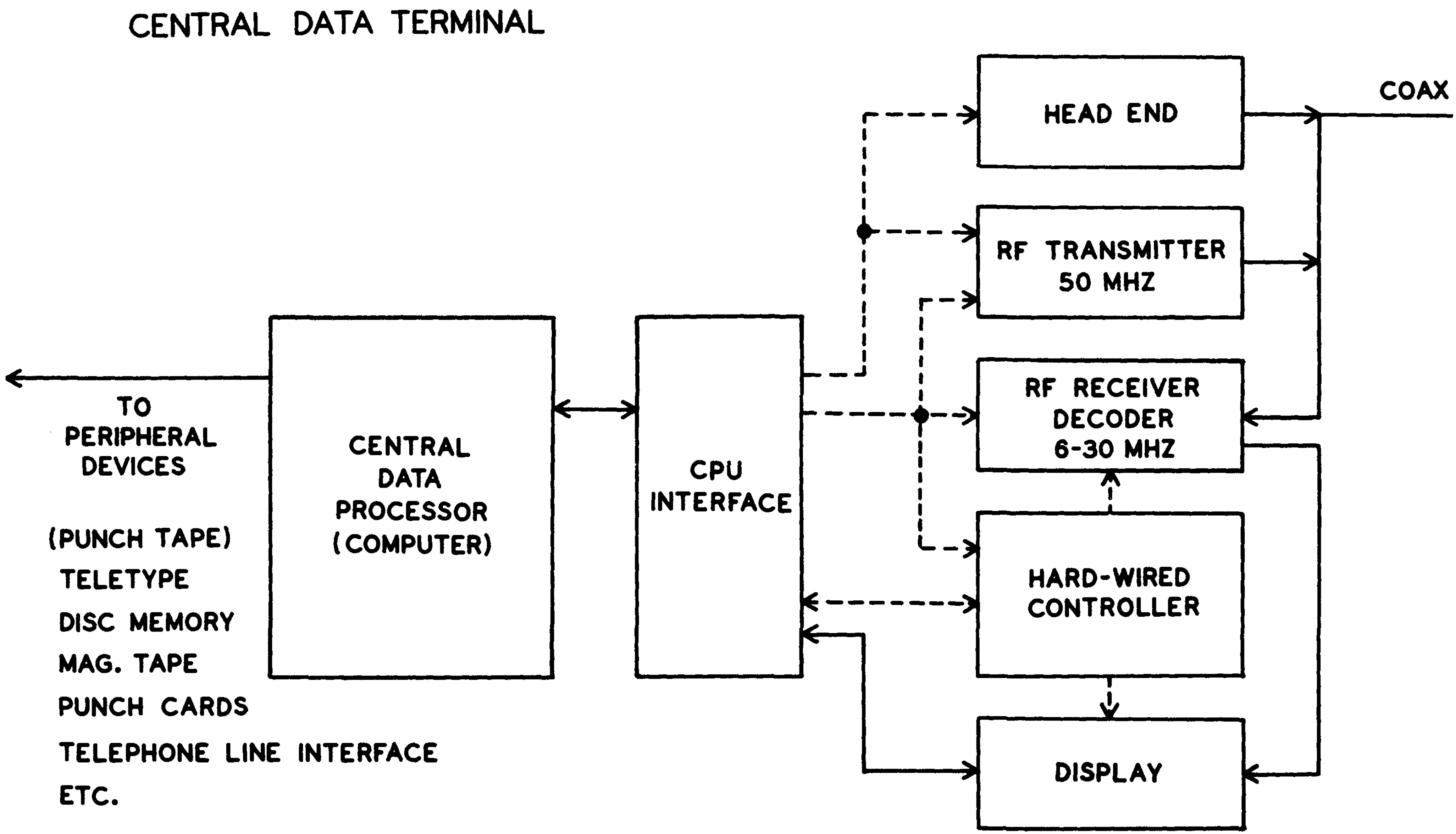


FIGURE 5

TOCOM

TWO-WAY CATV SYSTEMS PERFORMANCE

by ANDREW W. BARNHART

JERROLD ELECTRONICS CORPORATION

Presented at

NCTA SHOW

July 9, 1971

by Andrew W. Barnhart

Senior Project Engineer

Jerrold Electronics Corporation

TWO-WAY CATV SYSTEMS PERFORMANCE

Andrew W. Barnhart, Senior Project Engineer
CATV Development
Jerrold Electronics Corporation

I. INTRODUCTION

The advent of full two-way cable systems has generated many claims as to the various services made possible. To evaluate the full potentialities for two-way communication in a given application the correct design approach must be selected and the resulting performance estimated. There are several performance trade-offs unique to two-way transportation that must be considered in system design and specification.

There are many specifications of importance to two-way systems. This paper will be concerned with the three major performance criteria -- carrier-to-noise (C/N), cross-modulation (XM), and group delay -- and how these are expected to vary with the system size, crossover frequency, and channel loading. Many systems are considering two-way operation, each with its own specific requirements. This leads to many different two-way configurations, including multiple cable approaches, each suitable for a particular type of application. It is beyond the scope of this paper to treat all of these configurations; rather several specific two-way CATV transportation designs will be reviewed and the resultant system performance estimated. Many of the design values used for these systems were arrived at employing considerations discussed in a previous paper "System Considerations in the Design of a Two-Way Transmission System" cited in Reference 1.

II. SINGLE CABLE SUB SPLIT

A. SYSTEM DESCRIPTION

This elementary cable two-way TV system employs a single cable trunk and a single feeder. The frequency spectrum is split below Channel 2 with the return information, which is destined for the headend, occupying the band 5-30 MHz. The outgoing signals at Channel 2 and above are distributed to the subscriber using the conventional assignments. A diagram of the system showing the layout of trunk and feeder stations is given in Figure 1. The operating principles of this configuration and several of the equipment design considerations are treated in some detail by Reference 1. This approach is an economical choice for applications requiring only a limited return bandwidth. Equipment for a two-way system of this configuration was designed and tested, the performance of which will be described in the first example. The outgoing system employs push-pull amplifiers designed for 27-channel operation utilizing the 12 standard VHF channels, 9 channels in the midband region, and 6 supers above Ch 13 in the region 216-260 MHz. Figure 2 illustrates the spectrum utilized. The feeder return and the trunk return amplifiers are also push-pull and are assumed to be located at every outgoing location. It is convenient to rate the return equipment in terms of TV channel capability although a large part of the spectrum may be occupied by data. The trunk return amplifiers are rated for four TV channels. It is felt by the author that this is a conservative method of rating because of the XM loading effect of data will be much less than a TV channel.

When the feeders are used to gather the return information, it is unlikely that any one feeder will have a significant percentage of the total return spectrum. It is tempting to rate the C/N and XM of the feeders on the basis of a single TV channel. However, some mathematical difficulties arise when considering single channel XM. In this presentation the feeder is rated for two TV channels which implies that a single TV channel may be returned via the feeder -- for community service or on-the-spot coverage -- and the cumulative effect of the rest of the data on the feeder is less than or equal to that of another TV channel. Table I gives the performance that can be expected of these amplifiers.

B. SYSTEM TOLERANCE

With the equipment performance defined, it is possible to estimate the performance of a typical CATV system that includes two-way as a part of the design. Assume a system of the following characteristics.

Total Cable Bearing Strand	350 miles
Feeder-to-Trunk Ratio	4
Maximum Trunk Cascade	25 amplifiers
Total Number of Trunk Amplifiers	250 amplifiers
Maximum Feeder Cascade	2 amplifiers
Total Number of Line Ext. (3.6/mile)	1250 amplifiers
Channel Capacity Outgoing	27 channels
Channel Capacity Return Trunk	4 channels
Channel Capacity Return Feeder	2 channels

The design goal for the distribution is -54 dB XM. This may be achieved with the levels assumed in Table 2. They are based on equal contribution between the bridger and each line extender amplifier. The outgoing trunk amplifier level is determined in the usual manner -- halfway between the XM and C/N limits. The levels for the return amplifiers were selected to put the return system halfway between XM and noise. The XM is determined using the maximum cascade, and the noise is determined using the total number of return amplifiers that return signals on a given cable. This aspect is discussed in Appendix A. The level of the return feeders is also designed on an optimum cascade basis as is the trunk -- in contrast to the design method for the feeders. Table 2 summarizes the levels used and the performance of each group of amplifiers. The performance of the outgoing and the return systems can be estimated using the performance given in Table 2. The outgoing C/N will be nearly that of the trunk 43 dB, and the XM will be -51.4 dB (-63, -65, -67). The return system will have 43.5 dB C/N (45 and 49), and -66 dB XM (-69, -77). Although there are a large number of return feeder amplifiers (1250), their operation at optimum cascade level +45 dBmV enables them to have 4 dB better C/N than the return trunk.

Multiple Return Trunk:

Figure 3 is a symbolic representation of a CATV layout where the headend is shown between two roughly equal areas. Consider the case where the return trunks from each area are not combined but are brought into the headend separately. Two benefits occur: 1. If the return levels were to remain the same, a 3 dB improvement in C/N would occur because only half the number of return amplifiers would add noise to a given cable. 2. Twice the return bandwidth is available because signals can be selected from either cable. This results in a complete return system performance of -68 dB XM and 45.5 dB C/N. The worst case round trip performance -- a signal gathered by the return feeders at the maximum trunk cascade and then turned around at the headend and distributed to the farthest outgoing subscriber -- would be -50.2 dB XM and 41.1 dB C/N representing a 1.1 tolerance improvement due to the use of the return split. These performances are summarized in Table 3.

System Size:

In tolerance calculations for the outgoing trunk only the maximum cascade is required. However, for the return trunk and return feeder amplifier performance both the maximum cascade and the total number of amplifiers to be served by a single cable must be known. For a proposed system the total number of return amplifiers may be estimated from the estimated strand miles using feeder-to-trunk ratios and amplifiers-per-mile.

Thus, when considering two-way for a given application, both the maximum cascade and the system size should be estimated.

Alternate Return Amplifier Spacing:

Although the maximum single span loss at 31 MHz is only 6.6 dB, inclusion of flat losses within the station and those due to trunk cable splitting will increase the return module gain requirement to a range of 11-16.5 dB. The average single span return gain assumed previously was 14 dB. A return amplifier at every other trunk station requires a gain of 20-31 dB. Assuming a nominal module gain requirement of 25 dB for alternate spacing, the same output capability and noise figure, this return amplifier would have 11 dB less tolerance than the "amp every" station. However, since there is half the number of return amplifiers adding XM and noise, the net result is a return trunk system with 5 dB less tolerance. A return feeder level matchup problem may occur at trunk stations where the return amplifier is "missing" because the required insertion level will be one span loss higher.

C. FILTER CONSIDERATIONS

Two-way operation is obtained by frequency splitting with filters. Gain flatness, stability, and isolation requirements will determine the required filter amplitude responses (see Reference 1). For these considerations it is desirable to make the filter "look like a cliff". However, this type of amplitude response implies very large group delay and chroma delay near the cutoff frequencies. In fact, higher stopband rejection, better passband match, and narrow guardband all imply higher group delay. Even the type of filter response selected (i.e., Butterworth, Elliptic) is of importance. A quantitative treatment of these factors on group delay variations is given by Reference 2.

Amplitude Requirements:

The amplitude requirements for the trunk line filters are given in Figure 4. The contiguous type high/low split filter (equal 3 dB cutoff frequencies) cannot be used as the 6 dB per filter isolation will be insufficient to prevent oscillation with the trunk amplifiers installed in the station. The 44 dB filter floor requirement is sufficient to keep the amplitude gain ripple to less than ±0.03 dB resulting from closed loop feedback effects. Filters for this requirement were designed and built with nominal cutoff frequencies of 45 MHz for the high pass section and 35 MHz for the low pass. Practical considerations of alignment and amplitude rounding near the cutoff limit the use in cascade to 49 and 32 MHz respectively.

Group Delay Requirements:

In color TV transmission it is not the group delay as such that is troublesome but rather the variation of group delay over the TV channel bandwidth that is of concern. Chroma information modulated on the color subcarrier may arrive at a different time than the luminance information modulated on the PIX carrier. This time differential is termed chroma delay and results in color misregistration and blurring. Chroma delay and how it occurs from amplitude filtering is explained in Reference 1.

Two questions are of direct concern: How much chroma delay can be tolerated? and How much will a given system produce? The former question requires a subjective evaluation. This has been answered in part by a recent study under laboratory viewing conditions and is cited in Reference 3. The delay introduced by the CATV transportation system falls in the "shaped delay" classification of this reference, and from Figures 1 and 6 we may draw the following conclusions from the expected comment for "expert observers" viewing of color TV monitors: 1. at 500 ns of chroma delay most would find the reception impaired but none would find it "definitely objectionable". 2. at 230 ns most would find a perceptible effect but none would find it even "somewhat objectionable".

Filter Chroma Delay:

The measured group delay response of the pair of high/low split filters used in the trunk station is shown in Figure 5 (Curve 2). Table 4 gives the chroma delay for these filters for the worst case channels. For the 25-amplifier cascade these filters will introduce -155 ns (-6.2 ns x 25) of chroma delay in Channel 2, 435 ns in T10 (25 MHz PIX), and 158 ns of chroma delay in T9 (19 MHz PIX). Channel 2 was deliberately favored over Channel T10 in terms of chroma delay by the selection of cutoff frequencies for two reasons -- color TV on Channel 2 is a definite requirement in most applications whereas it is not for T10, and it is relatively simple to delay equalize T10 at the IF frequency since it will be returned to the headend. The Channel 2 group delay that is delivered to the subscribers can be halved by including +77.5 ns of compensating chroma delay at the output of the Channel 2 processor at the headend.

Group delay scales inversely with the cutoff frequency. It is, therefore, possible to calculate the group delay and chroma delay that these filters would have if they had been designed for different dutoff frequencies (see Reference 2). The actual high/low cutoff frequencies are 45/35 MHz. The calculated values of group delay are plotted in Figure 5, and the calculated values of chroma delay are listed in Table 4 for the measured filter data scaled to the cutoff frequencies of 41/32 MHz, 48/37.4 MHz, and 51/39.6 MHz. It is observed that the raising of the high pass cutoff frequency to 51 MHz in order to obtain more return bandwidth nearly doubles the chroma delay in Channel 2.

D. CASCADE PERFORMANCE

A complete line amplifier cascade of the above design with cable and a.c. powering was tested. Tracings of the swept group delay performance are shown in Figure 6. Chroma measurements in individual channels are listed in Table 4. The chroma delay measurements of the 9-amplifier cascade for Channels T9, T10, 2, 3 and 4 are in excellent agreement with the expected values based on the individual filter measurements. For these channels the system chroma delay is determined by the filters used. At frequencies above 200 MHz there is a slight group delay rise caused by the amplifier response rolloff past 260 MHz.

In the return amplifier system a pronounced rise in group delay occurs below 10 MHz causing significant chroma delay. This is not due to the high/low split filter, but can be attributed to the return amplifier and a.c. choking. The return amplifier employs many capacitors, both bypassing and interstage coupling. Although the amplitude response is flat at 5 MHz, there is a pronounced rolloff at 3 MHz which generates this group delay behavior. The a.c. bypass chokes used in power inserters and trunk stations also have the same effect.

Figure 7 is tracings of the swept amplitude response of the 9-amplifier cascade employing standard amplifiers and equalizers. The peak/valley is considered satisfactory as no special "mop ups" were required to obtain the performance. As expected, there was no evidence of any gain ripples due to closed loop feedback.

E. SUMMARY

The system configuration can achieve broadband two-way communication on a CATV network. The required trunk and feeder stations have been built and tested, the resulting performance has been projected to a two-way system comprising 350 strand miles. It provides 27 outgoing channels, 12 of which are the standard V's; the return bandwidth can provide one TV channel and 19 MHz of data. Since the passive components, i.e. power inserters and directional taps, are also available, the results indicate that this two-way system is viable with today's technology.

III. SINGLE CABLE MID-SPLIT

A. SYSTEM DESCRIPTION

The next system to be considered has the same configuration (Figure 1) as the previous but employs a frequency split in the midband (Figure 2). Fourteen TV channels are available for outgoing distribution, and 14 channels + 9 MHz of data are available in the return path. This spectrum utilization provides two important advantages-much greater return bandwidth is available and the wide range provided for the filter crossover results in low envelope delay distortion.

Because only the 7 high VHF channels are available for distribution to a standard TV set without converter, this system must be regarded as special purpose. It can be used independently for private channel communications for either video or data by the school, civil, or business communities; or it can be used in conjunction with a regular one-way CATV plant to provide both private channel communications and two-way capability for all subscribers.

B. SYSTEM TOLERANCE

The performance of the mid-split approach will be calculated employing the previous 350-strand mile system example. The total number of return amplifiers adding noise is halved as in the previous example, by using two return cables near the headend. The amplifiers used in this system have the ratings listed in Table 5, and their operating levels were selected in the same manner as for the sub-split. The lower channel loading of the outgoing plant results in better tolerance than the previous example, and the low round trip XM allows sufficient for distribution systems such as schools that will utilize private-channel capability. The increased channel loading on the return path requires a better quality amplifier, and for systems of this size it may be desirable to consider an additional return cable.

C. CASCADE PERFORMANCE

Equipment for this configuration was built and tested in a 15-amplifier cascade that includes all cable and a.c. powering. The chroma delay was expected to be minimal because it varies inversely with the square of the filter cutoff frequency. Figure 8 includes tracings of the swept group delay for the outgoing and return paths and lists the chroma delay of the worst case channels. Channels 6A, 6B, 6C can occupy what is normally the FM band. The rise of group delay below 10 MHz is caused by the amplifiers and a.c. chokes which results in Channel T7 having larger chroma delay than any other channel, although this delay is not likely to require delay equalization. The chroma delay of the standard VHF channels is so small as to be inconsequential in a 25-amplifier cascade.

Figure 9 shows tracings of the amplitude response for the complete 15-amplifier cascade. This response is satisfactory as it was obtained by employing only the normal amplifier and equalizer adjustments. The peak/valley of the return path is attributed to the equalizer error which was of an early design.

D. SUMMARY

This approach provides a viable solution to systems that require large two-way communication bandwidth of high quality. The return channels may be distributed either on the outgoing path for private use or on the regular CATV system for public viewing. The resultant low value of chroma delay will not be of significant concern for CATV transportation.

IV. MULTIPLE CABLE APPROACHES

A. DUAL TRUNK, SINGLE FEEDER

This system employs two trunk cables, with trunk stations at congruent locations, and a single feeder cable. With reference to Figure 10 trunk cable A is one-way only and uses the outgoing frequency spectrum of 54-260 MHz for carrying 27 channels. The A cable distribution is two-way carrying the 54-260 MHz spectrum from the A trunk in the outgoing direction, and returning 5-30 MHz from the subscriber locations to the B trunk station.

The B trunk cable is the mid-split two-way system previously considered. The 5-30 MHz portion of the B return spectrum is used by the A feeder return signals which are coupled over to the B station from a high/low split filter in the A station. The 30-108 MHz portion of the B return and the B outgoing have limited access and exit.

This approach has the following features:

Public Service:	54-260 MHz	27 Ch	Outgoing
(full distribution)	5-30 MHz	1 Ch + 19 MHz	Return
Private Service:	174-260 MHz	14 Ch	Outgoing
(limited access)	30-108 MHz	13 Ch	Return

The signals carried on the "A" cable suffer minimal chroma delay distortion in that the filters are only used in the feeders.

The high crossover frequency of the "B" cable filters result in low envelope delay distortion compared to the low frequency equivalent used in the "A" cable feeders.

Using the previous 350-strand-mile example the performance of this approach can be estimated. The private system will have the same performance as listed in Table 5, -56.5 dB XM and 42.7 dB C/N. The A system will have the same performance as in Table 3. The performance of a round trip channel using the B return trunk and distributed by the A system will be -49.4 dB (-63 and -51.4) XM and 41.3 dB (46 and 43) C/N.

The worst case path is the round trip "public service" TV channel that is gathered by the A return distribution and B return trunk and sent out the A outgoing system. Its performance would be -49.1 (-79 and -49.4) XM and 40.9 (51 and 41.3) C/N.

B. DUAL TRUNK, DUAL FEEDER

The last example is a dual trunk, dual feeder system. In this approach the A cable is a conventional one-way CATV system. The B cable is a full two-way system with 5-108 MHz trunk return, 5-30 MHz feeder return and 174-260 for the outgoing trunk and feeder.

This system has all of the advantages of the previous example with the feature of the additional channel capacity of the B outgoing distribution which is available to the subscriber by means of an A-B switch. In many cases enough channels would be available (19) by means of the switch that set converters would not be required for standard grade service.

Return:	5-30 MHz	1 Ch + 19 MHz	Public
	30-108 MHz	13 Ch	Private
Outgoing:	54-260 (A)	19 Ch	A-B switch only
	174-260 (B)	42 Ch	with converter

The tolerance of this system is the same as the previous example. (The B outgoing distribution is slightly better than A because of lower channel loading.)

REFERENCES:

1. H. B. Marron and A. W. Barnhart, "System Considerations in the Design of a Two-Way Transmission System", NCTA Official Convention Transcript, June 1970.

2. A. W. Barnhart, Group Delay Variations of Selected Filter Prototypes, Masters Thesis, University of Pennsylvania, Dec. 1970.

3. A. M. Lessman, Subjective Effects of Delay Difference Between Luminance and Chrominance Information of the NTSC Color Television Signal, Bell Laboratories, Homdel, N. J.

4. K. Simons, Technical Handbook for CATV Systems, Jerrold Electronics Corporation, Hatboro, Pa.

APPENDIX A

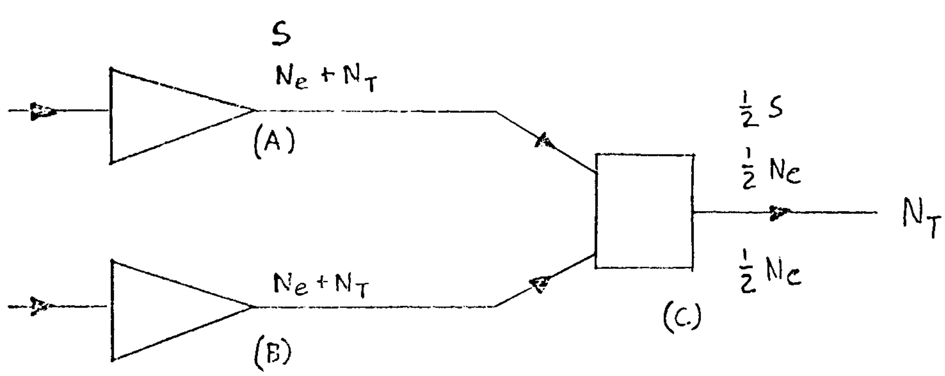
NOISE GATHERING EFFECT

Cable TV transportation systems are operated at unity gain; a signal introduced anywhere into the system, which is in the system bandwidth, will be maintained at approximately its original level. The output noise generated by any particular amplifier will also be transported by the system and maintained at its original level, and therefore the noise of all amplifiers returning signals to a given point will be present at that point. Since random noise adds in terms of power, the total noise will be the power sum of the noise output of each amplifier.

The noise referred to in the above discussion is the excess noise output, N_E, produced by the amplifier. Thermal noise, N_T, is ever present and does not build up. This is illustrated by the following example. Consider two trunk amplifiers whose outputs are combined by use of a 3-dB splitter, with a signal at the output of one amplifier having power S. The C/N ratio at this point (A) will be $S/(N_e + N_T)$. The C/N ratio at the output of the splitter (point C) will be $S/(2N_e + N_T)$. The excess noise at the output of the amplifier is usually at least a thousand times greater than thermal noise. We can thence say that the impairment in C/N ratio by combining two amplifier outputs is

$$\frac{(C/N)_C}{(C/N)_A} = \frac{S/(2N_e + N_T)}{S/(N_e + N_T)} = \frac{N_e + N_T}{2N_e + N_T} = \frac{1000+1}{2 \times 1000 + 1} = \frac{1}{2}\,\frac{1001}{1000.5} \doteq -3\ \text{db}$$

and therefore the total noise can be considered to be the sum of the noise output of the two amplifiers.



Amplifier	Operational Gain	Noise Figure	Output Capability for -57 dB XM	
Outgoing Trunk	23 dB	10 dB	+48 dBmV	27 Ch
Outgoing Bridger	28 dB	12 dB	+47 dBmV	27 Ch
Outgoing Feeder	24 dB	12 dB	+42.5 dBmV	27 Ch
Return Trunk	14 dB	8 dB	+52 dBmV	4 Ch
Return Feeder	16 dB	8 dB	+58 dBmV	2 Ch

SINGLE CABLE SUB-SPLIT AMPLIFIER PERFORMANCE

Table 1

Amplifier Group	Operating Level dBmV	Number For XM	Cascade XM dB	Number For C/N	System C/N dB
Outgoing Trunk	+31	25	-63	25	43
Outgoing Bridger	+43	1	-65	1	62
Outgoing Feeder	+39.5	2	-57	2	59.5
Return Trunk	+32	25	-69	250	45
	+31	25	-71	125*	47
Return Feeder	+45	2	-77	1250	49
	+44	2	-79	625*	51

*Return Split Assumed

Sample Calculations (Ref. 4):

XM Trunk Out $= XM_1 + 20 \log C = \left[-57 - 2\,(48 - 31)\right] + 20 \log 25 = -91 + 28 = -63$ dB

C/N Feeder Return $= C - N - 10 \log C = +45 - \left[-59 + 8 + 16\right] - 10 \log (1250) = 45 + 35 - 31 = 49$ dB

SINGLE CABLE SUB-SPLIT AMPLIFIER TOLERANCE

Table 2

SYSTEM	XM	C/N
Outgoing Trunk and Distribution	-51.4	43.0
Return Trunk and Feeder	-66.0	43.5
Complete Round Trip	-50.0	40.2
Return Trunk and Feeder/2 Return Cables	-68.0	45.5
Complete Round Trip/2 Return Cables	-50.2	41.1

SINGLE CABLE SUB-SPLIT SYSTEM TOLERANCE

Table 3

	Chroma Delay of Trunk Station Filters with Varying Cutoff Frequencies: High/Low (MHz)					
Channel	41/32*	45/35	48/37.4*	51/39.6*	Chroma Delay of 9-Amplifier Cascade	Chroma Delay Feeder Amp.
T7	2.2	1.4	0.9	0.6	-90	< 1
T8	5.3	3.1	2.1	1.4	~0	1.0
T9	10.4	6.3	4.4	3.1	58	2.6
T10	46.9	17.4	11.1	7.7	155	6.8
2	-4.2	-6.2	-9.2	-12.3	-55	-4.6
3	-2.8	-3.4	-4.0	-6.1	-31	-2.4
4	-1.9	-2.3	-2.7	-3.0	-21	-1.2

*Calculated Values

SUB-SPLIT CHROMA DELAY (NANOSECONDS)

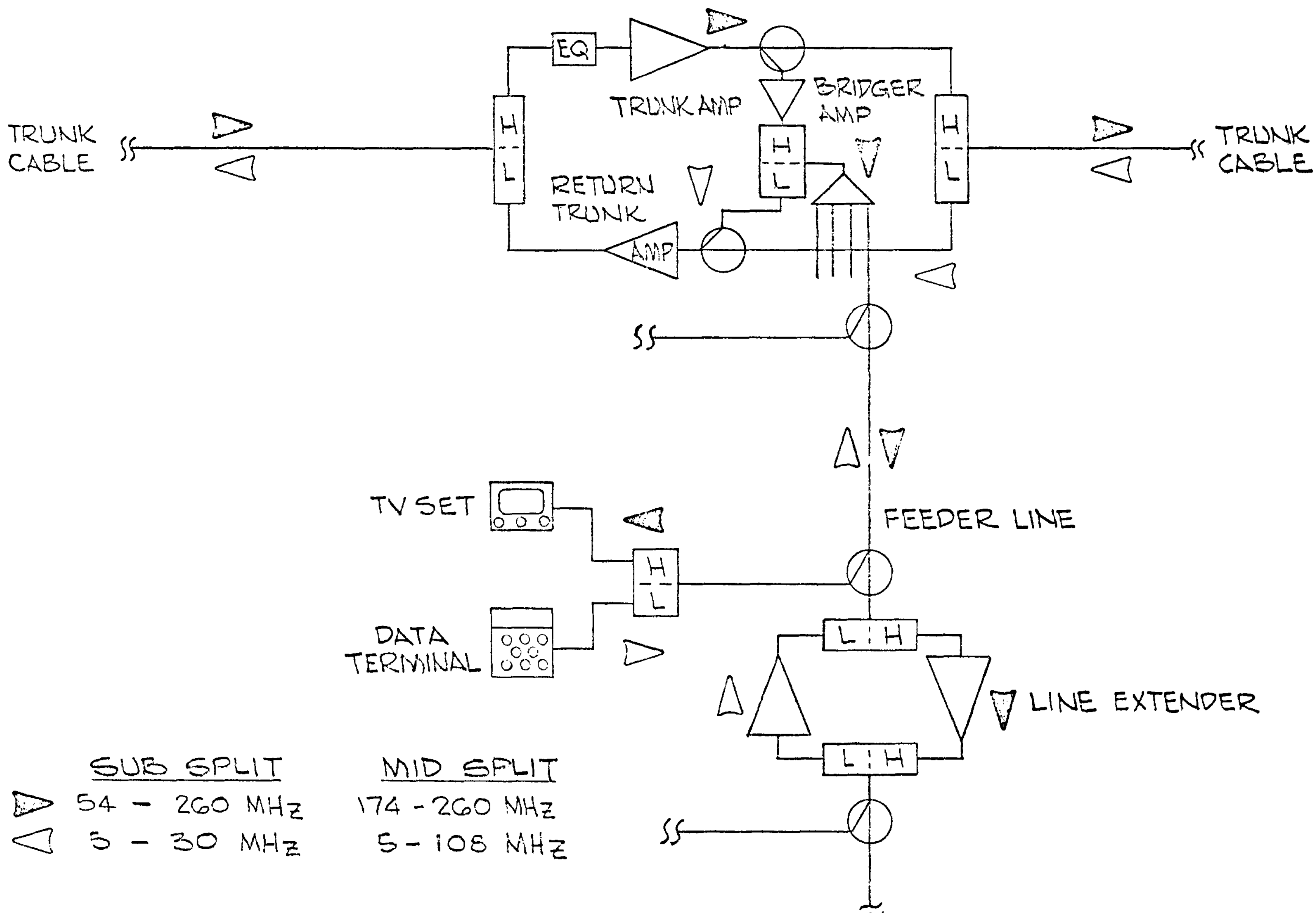
Table 4

Amplifier	Level dBmV	S_{MAX}* dBmV	Gain dB	NF dB	Cascade	Noise Total	XM dB	C/N dB
Outgoing Trunk	33.5	+50	23	10	25	25	-62	45.5
Return Trunk	35	+52	19	8	25	125	-63	46.0
Round Trip							-56.5	42.7

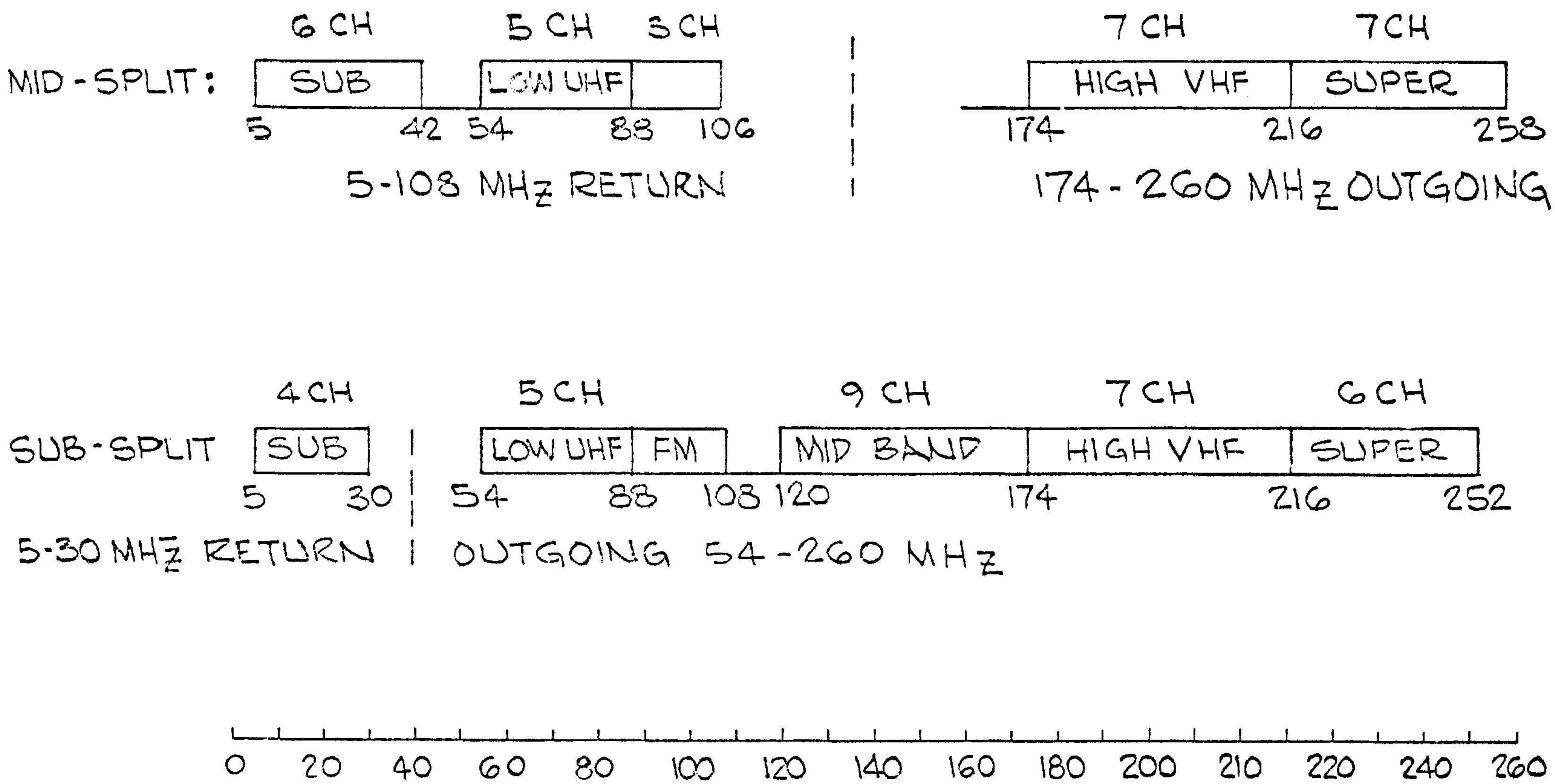
*Output Capability for -57 dB XM for 14 Ch $\left\{\begin{array}{l}\text{Low Band for Return}\\\text{High Band for Outgoing}\end{array}\right.$

SINGLE CABLE MID-SPLIT AMPLIFIER PERFORMANCE

Table 5

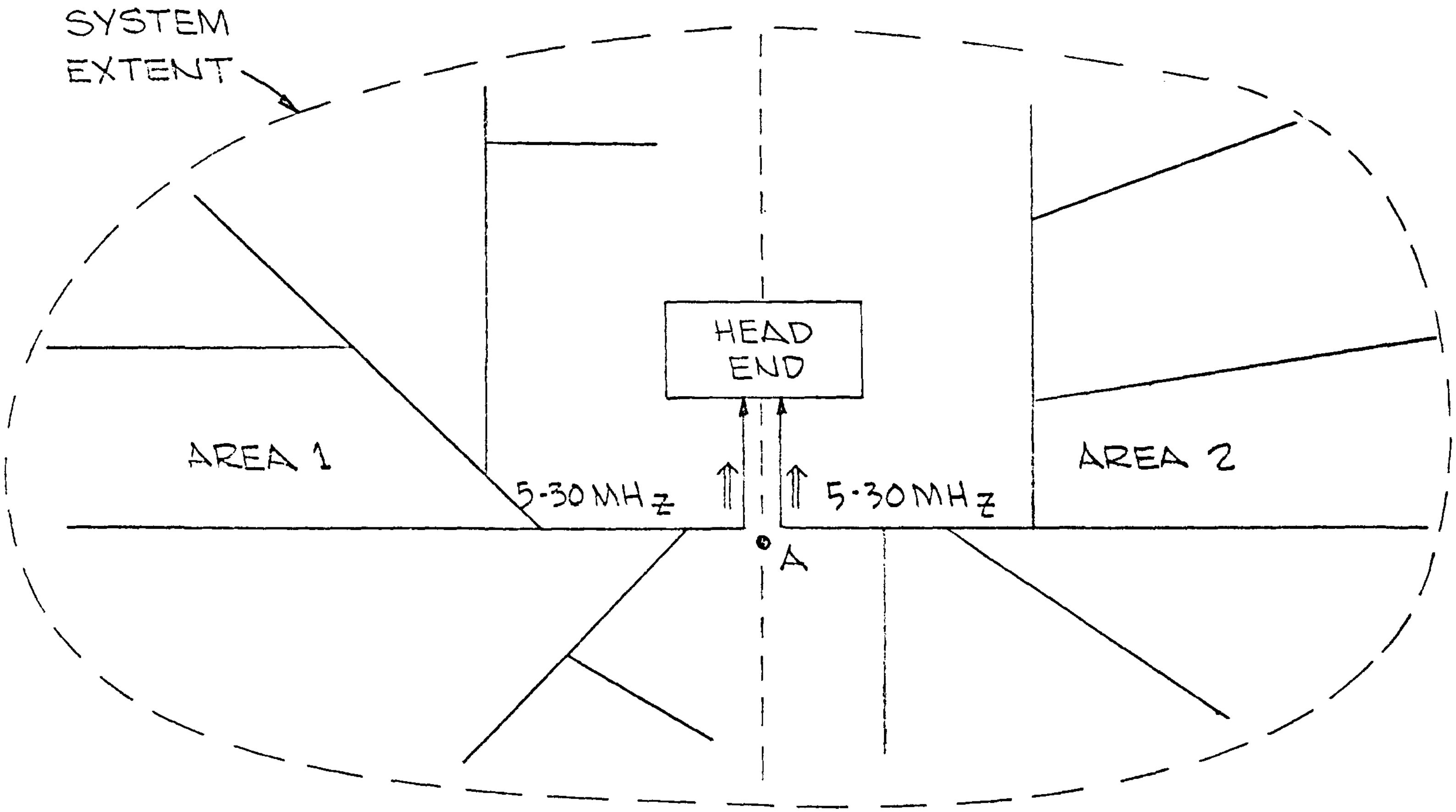


TRUNK STATION & LINE EXTENDER LAYOUT
FIGURE 1



SPECTRUM UTILIZATION OF SINGLE CABLE SYSTEM

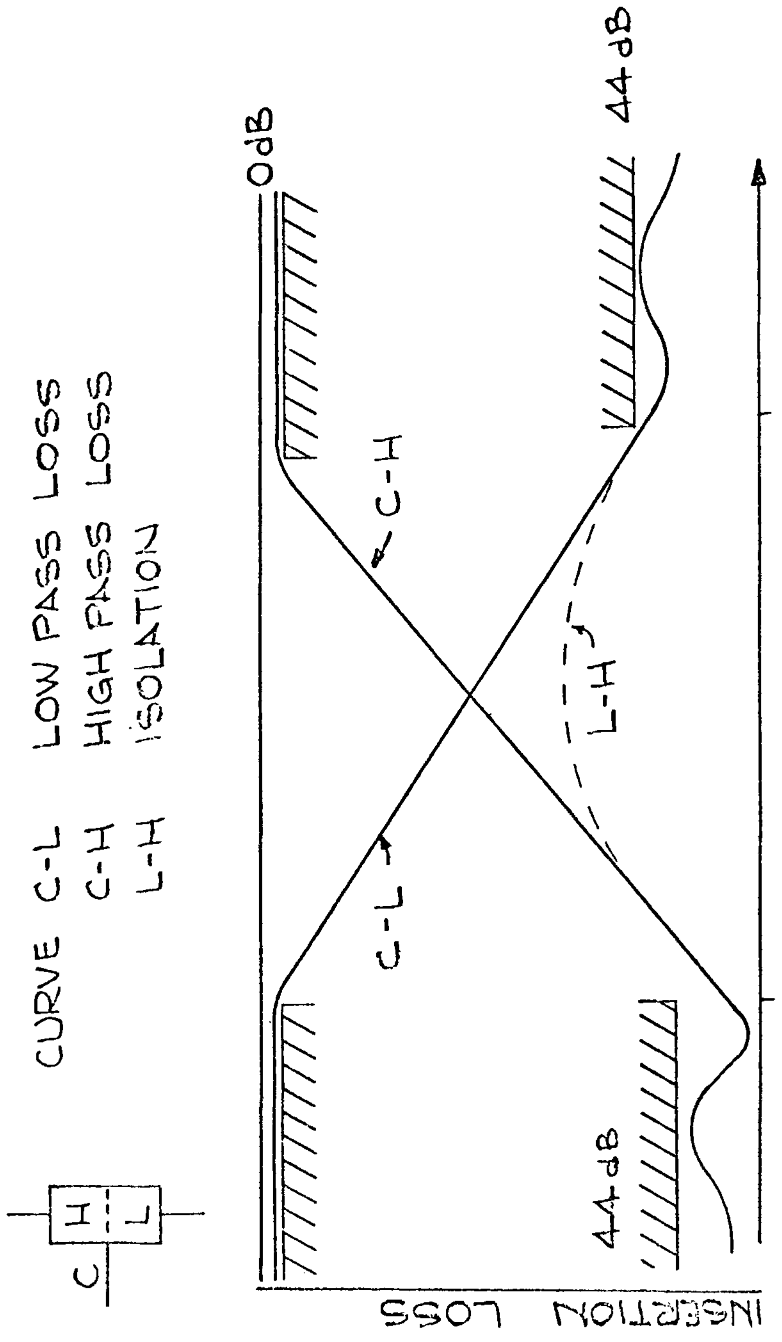
FIGURE 2



If two return cables are employed from point "A" to the Head End, and the areas are approximately equal, the C/N will be 3 dB better than if the return cables were combined at "A".

SYSTEM EMPLOYING MULTIPLE RETURN CABLES

Figure 3



FILTER AMPLITUDE REQUIREMENTS
FIGURE 4

GROUP DELAY OF HIGH/LOW SPLIT FILTERS
IN SUB-SPLIT TRUNK STATION

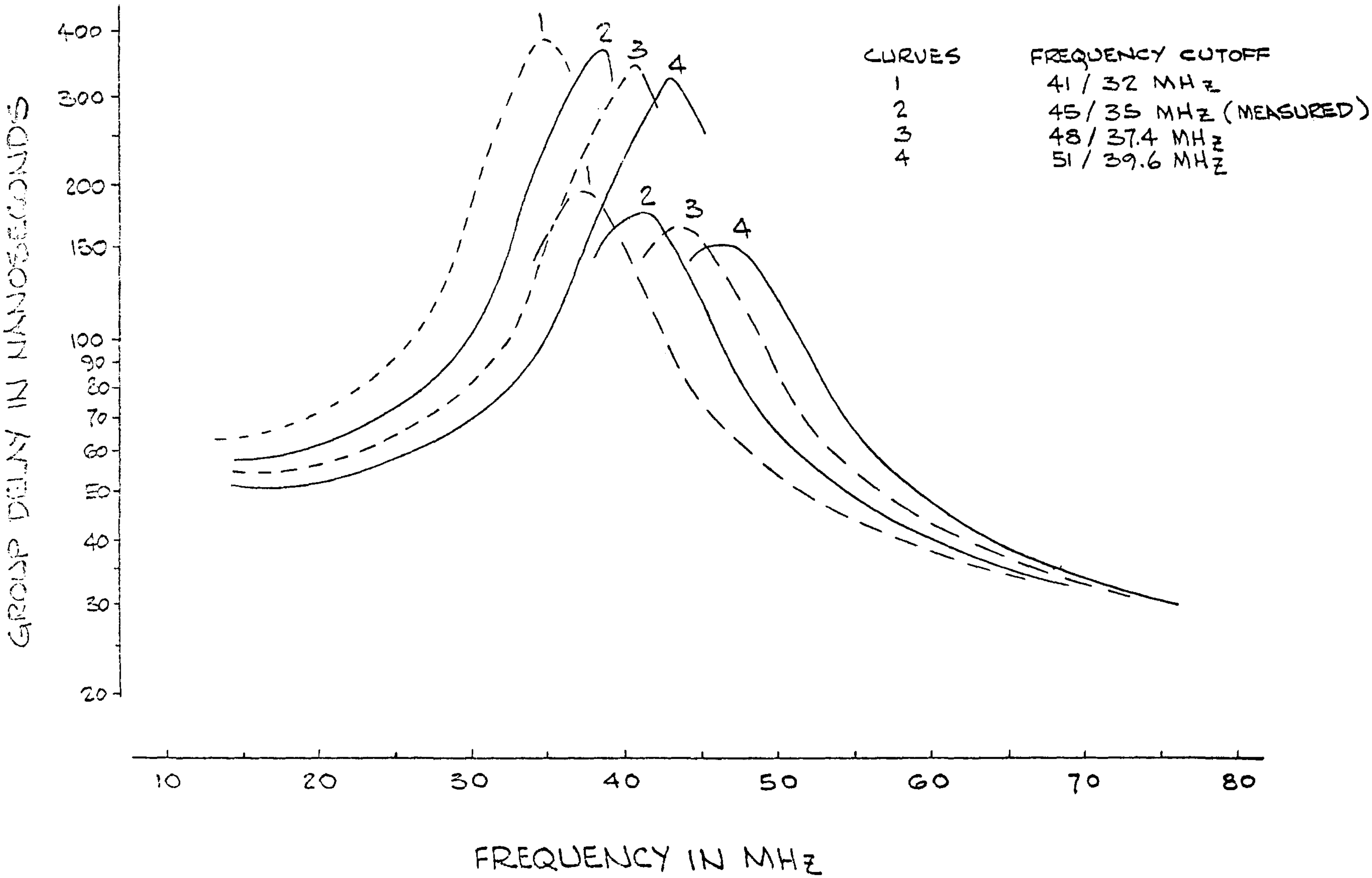
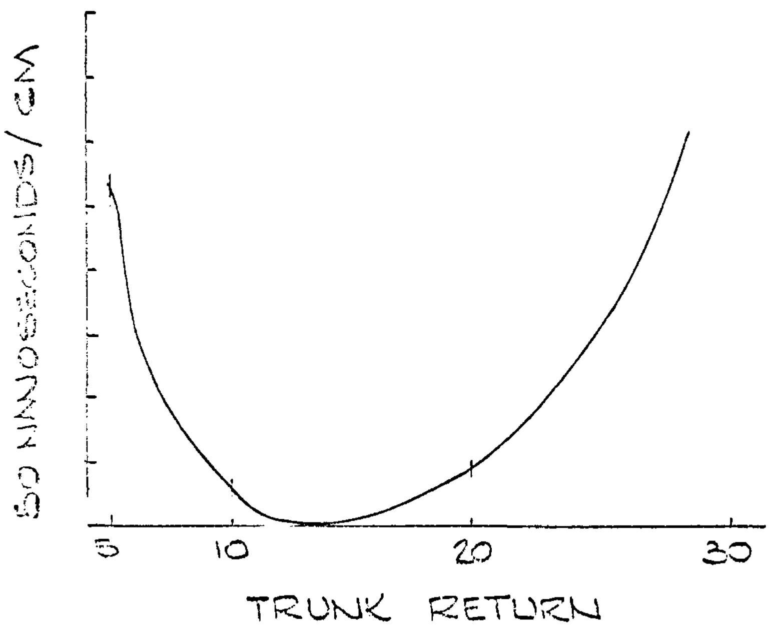
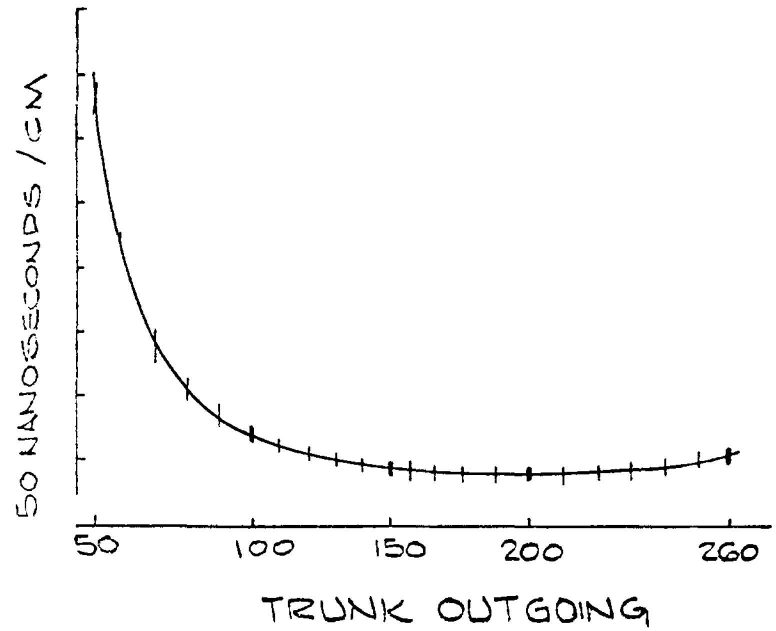


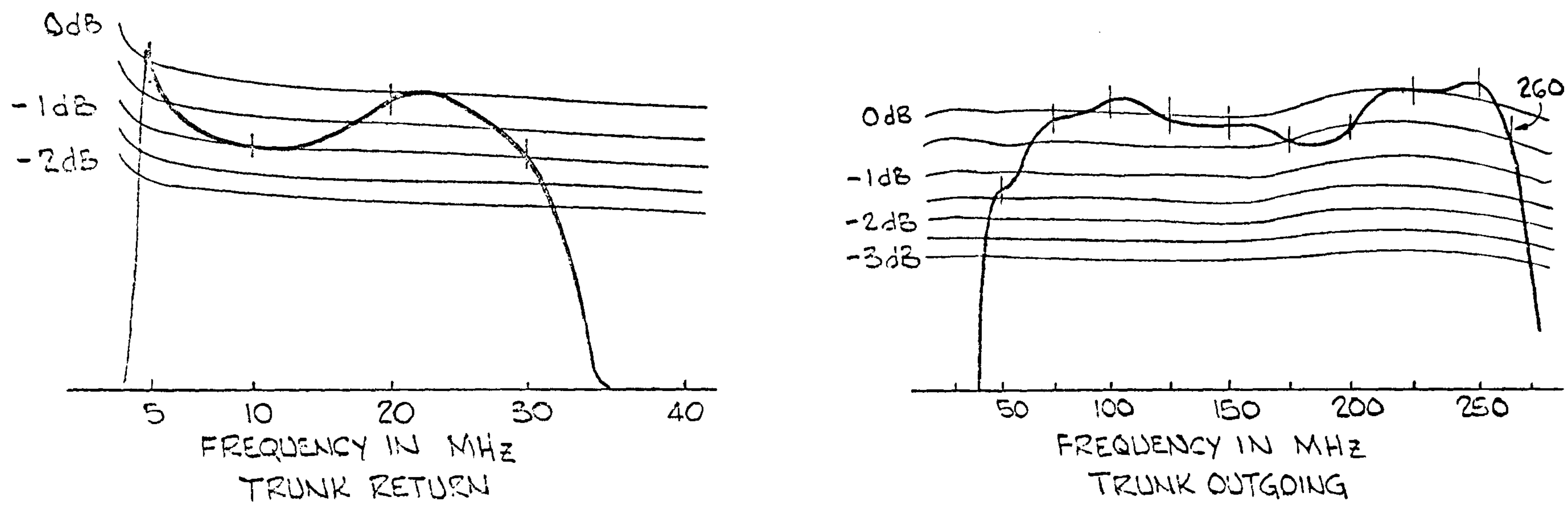
FIGURE 5



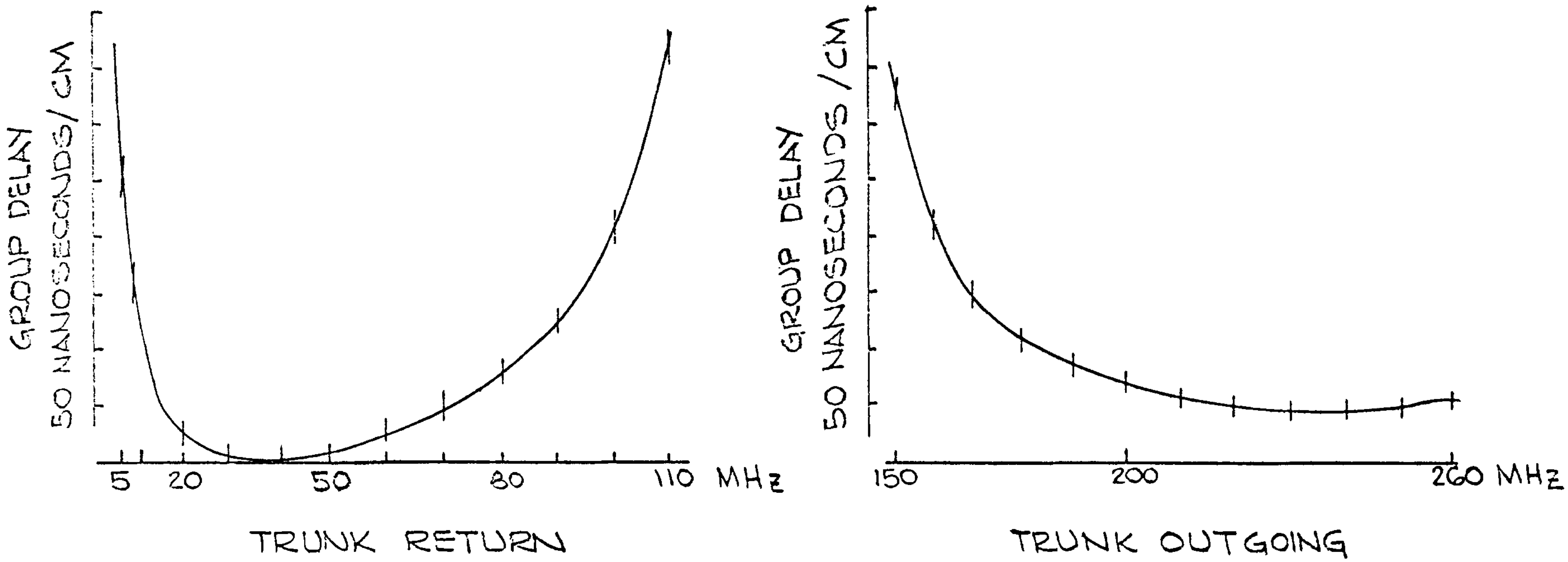


RELATIVE GROUP DELAY FOR COMPLETE CASCADE OF 9 TRUNK STATIONS (SUB-SPLIT)

FIGURE 6



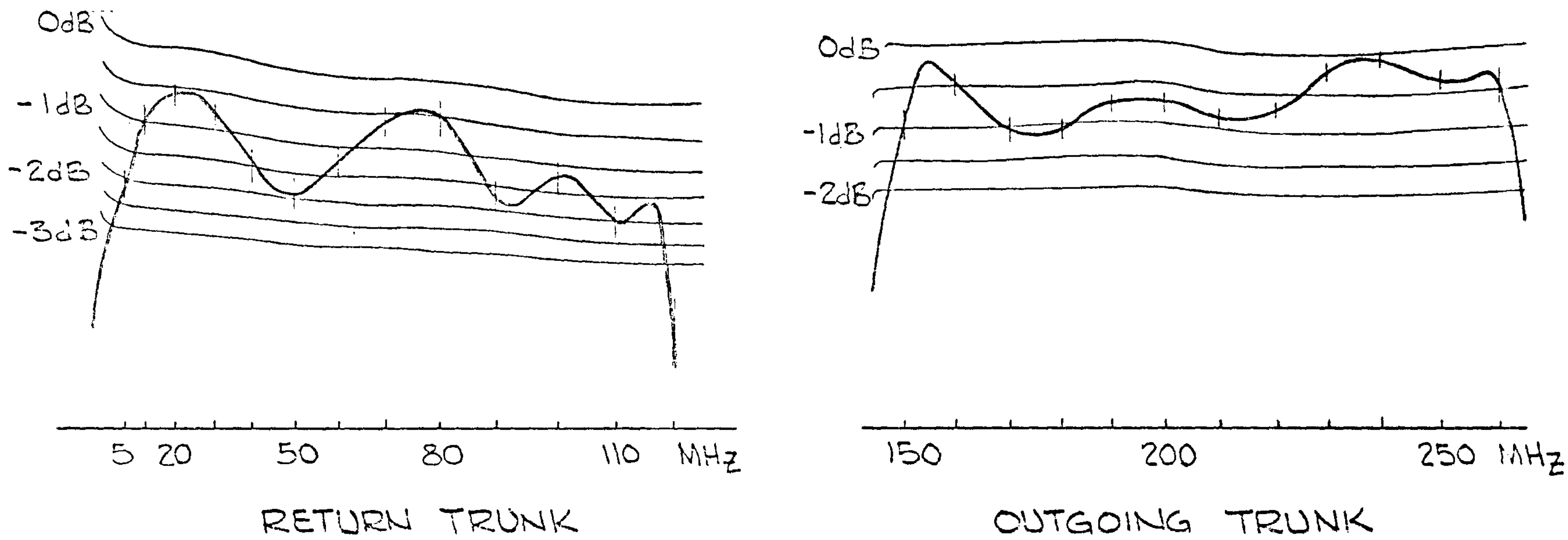
AMPLITUDE RESPONSE OF 9 AMPLIFIER CASCADE
FIGURE 7



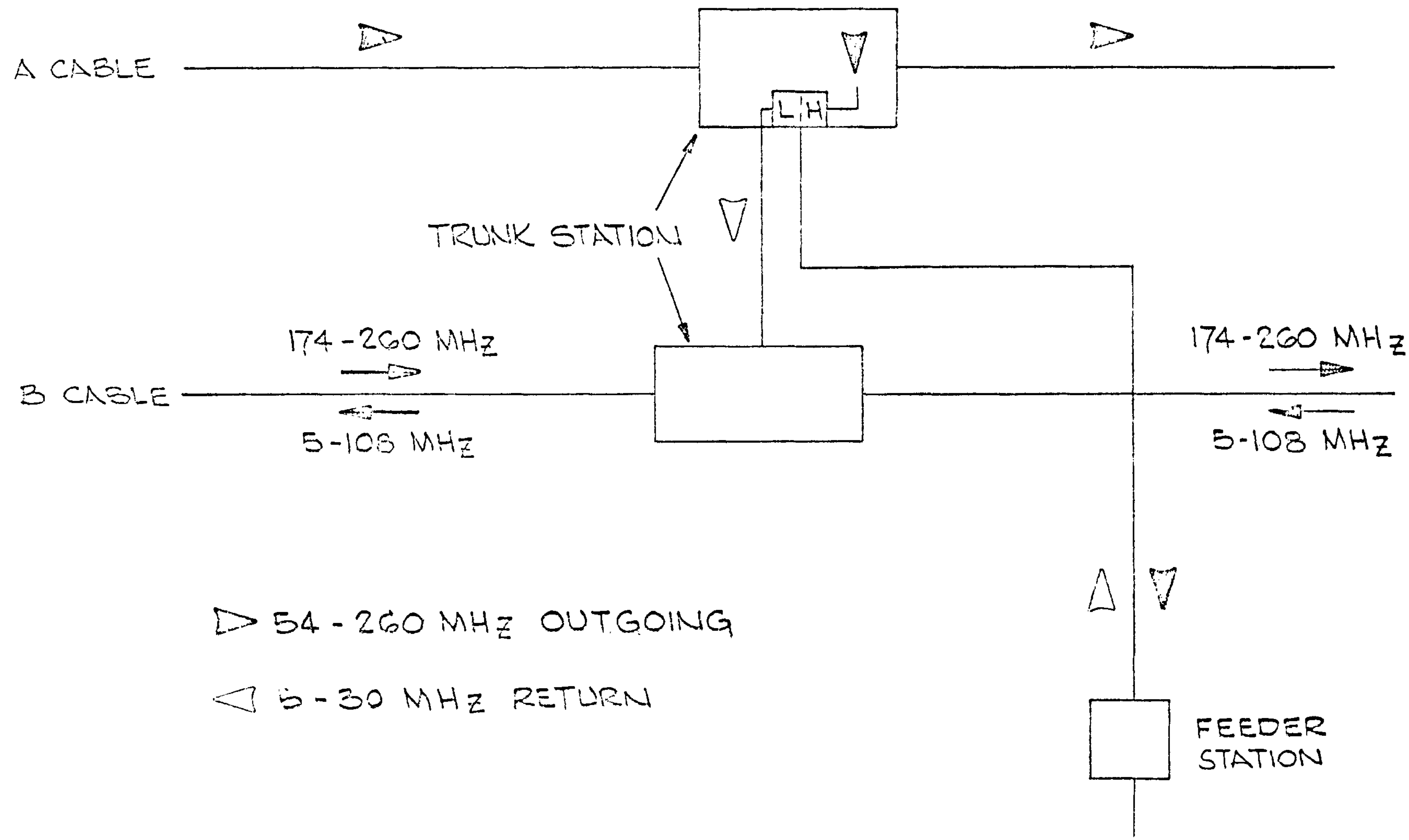
Channel #	T7	T8	T9	5	6	6A	6B	6C	7	8
Chroma Delay, nanoseconds	135	40	11	12	15	23	32	50	13	9

GROUP AND CHROMA DELAY 15-STATION CASCADE

Figure 8



AMPLITUDE RESPONSE 15 AMPLIFIER CASCADE MID-SPLIT
FIGURE 9



DUAL TRUNK SINGLE FEEDER SIGNAL FLOW

FIGURE 10

ISBN 0-940272-01-6; 0-940272-08-3; 0-940272-10-5; 0-940272-11-3; 0-940272-12-1; 0-940272-14-8; 0-940272-15-6; 0-940272-16-4; 0-940272-18-0; 0-940272-19-9; 0-940272-20-2; 0-940272-21-0; 0-940272-22-9; 0-940272-23-7; 0-940272-24-5; 0-940272-25-3; 0-940272-26-1; 0-940272-27-X; 0-940272-28-8; 0-940272-29-6; 0-940272-32-6; 0-940272-33-4; 0-940272-34-2; 0-940272-35-0; 0-940272-36-9; 0-940272-37-7; 0-940272-38-5; 0-940272-39-3; 0-940272-40-7; 0-940272-41-5; 0-940272-42-3; 0-940272-43-1; 0-940272-44-X; 0-940272-45-8; 0-940272-46-6; 0-940272-47-4; 0-940272-48-2; 0-940272-49-0; 0-940272-50-4; 0-940272-51-2; 0-940272-52-0; 0-940272-53-9; 0-940272-54-7