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NCTA Technical Papers

550 UPGRADES WITH FIBER: SELECTING COST-EFFECTIVE ARCHITECTURES

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Abstract

As a result of franchise requirements or marketplace demands, many systems will be upgraded to 550 MHz over the next few years. It is clear that fiber optics offers many cost-effective options that can be used in a plant upgrade plan. What is not clear is, which architecture will be the least costly for a given system. A secondary question might be, how the architecture that is selected today will influence the next upgrade.

This paper discusses key objectives that might warrant consideration when analyzing an upgrade plan. It is also suggests approaches to ease the task of comparing the options through design tests. An example of how different architectures might be combined in the same system is provided. The closing remarks address key issues that concern what one might strive for today to make the following upgrade as cost-effective as possible, with the least disruption to the plant.

GUIDELINES & OBJECTIVES

Overview

In order to focus on the task at hand, it may be useful to list some primary objectives of the upgrade process:

-The final approach must be cost-effective when measured against other available solutions

-The approach chosen should provide an easy, low cost, evolutionary path to further bandwidth expansion & new services.

-Improved plant reliability

-Lower plant maintenance requirements & easier plant maintenance

-Some improvement in picture quality

Cost-effectiveness

In evaluating an upgrade architecture, there are several key indicators which may be helpful in rating the overall cost-effectiveness of the approach. One such key indicator is the degree to which existing assets are re-used. The most valuable asset is the existing cable which; in the case of feeder cable, is present in nearly all locations where it will be needed. Other important assets that may not have been made obsolete by an upgrade include amplifier housings, AGC modules, existing power supplies (and their physical location), etc.

Another key indicator is the amount of projected variation in signal quality, or system performance, from one customer to another. Signal quality objectives should be established. Once fixed, the "constant distortion approach" of design should yield the most cost-effective plant. If the approach chosen results in a significant percentage of the customers receiving signals better than the minimum standard, the design is not fully optimized. In the past, constant distortion approaches to system design complicated plant maintenance activity, due to the number of different operating levels required for amplifiers. Later in this paper, an example of constant distortion design is shown where most amplifiers would operate with identical output levels.

Specifications for minimum tap levels also affect plant upgrade costs. If there is a single specification for the entire system, added cost savings can likely be found, since not all customers will have two television sets fed by a 150' drop (or conform to whatever criteria was used to establish a single spec). Efficient approaches would specify output levels for each of two or more drop length classifications. These level specifications would have been determined based on the number of outlets (the splitting loss) that would be fed without a drop amp, the minimum input level to the converter or customer terminal that yields the desired C/N ratio performance, and the length of drop cable to be used (perhaps in 30~50' increments). It is possible to minimize the work for the system designer by establishing tap levels based on average lot size, dwelling type, an entire street or subdivision, or a drop length classification and coding scheme noted for each pole or pedestal on the base maps.

The issue of minimum versus average tap port level is also important. If a minimum tap level is specified at 14 dBmV, 60% of the taps may actually have output levels of 16 dBmV. A money saving approach might be to specify a minimum tap level 2 dB below the target level. As a result, a small percentage of taps will have lower than target output levels. Calculations may show that it would be less expensive to equip those few customers, fed by low output taps, with a drop amplifier. In reality, the number of customers actually requiring drop amps may be less than the total number fed by these taps since some customers may have less than the expected number of outlets, or the drops may be shorter than the "average maximum" length estimated for that area of the system when the target levels were determined. It is important, of course, to add drop amplifier performance into the end-of-line performance calculations. A future option may be an architecture utilizing the active tap concept. [1] This concept uncouples the tap output level requirements from the feeder line design process.

Future Upgrade and Expansion

In order that a future upgrade would be possible for minimal additional investment, the path to reach the next bandwidth plateau must be examined today. A way to ensure low future upgrade cost is to devise a plan that will require the minimum disruption possible to the plant. That implies that the next upgrade should be considered while the current upgrade approach and architecture is being evaluated.

It is difficult to predict some of the requirements and options that may appear in the future. There are, however, some issues that we already understand quite well, such as cable attenuations and AM optical link performance at 860 MHz. Passive device performance, like that of taps and splitters, can be conservatively extrapolated from today's devices. The largest area of uncertainty is that of expanded bandwidth hybrid amplifier performance.

While 860 MHz amplifier products have been available in Europe for a decade, their performance specifications are not particularly useful given the disparity of the channel loading requirements between North American and European cable television systems. An approach that may be helpful in evaluating future plant upgrade options, is to design forward from the headend to the output of the optical node, and to design backwards from the customer's television set to an amplifier. Sensitivities to amplifier performance can then be evaluated and estimates made as to the performance improvement required over current 860 MHz European products. By assigning probabilities of success to the required improvements, a low risk plan to meet future performance and cost objectives can be developed.

Historical accomplishments in hybrid amplifier development provide some basis for conservative assumptions about future performance. While current 860 MHz hybrids are essentially single ended devices, it seems safe to assume that the development activity underway will be successful in producing an 860 MHz (or 1 GHz) true push-pull cascode hybrid. Once this activity has been completed, it follows that power doubling amplifier design can, at least, be duplicated by the "brute force" approach of physically using hybrids in a parallel configuration.

Discussions with hybrid manufacturers indicate that the development of 860 MHz or 1 GHz feedforward technology represents a significantly greater challenge. These manufacturers have also expressed doubt as to whether feedforward technology would be of value in high bandwidth systems of the future. They believe that cable operators will continue to reduce amplifier cascades through the deployment of fiber optic trunking. The complementary amplifier technologies would be ones with high output level capabilities, such as power doubling and quad power, but with lower distortion performance than is currently offered by feedforward. It is important to recall that a feedforward amplifier's output capability (compression point) is lower than that of a push-pull amplifier.

The following are guidelines that will maximize the chances that today's upgrade plan will be able to take advantage of future amplifiers.

-Where added reach from the node is needed, use 550 MHz 22 dB gain push-pull amplifiers spaced at a distance corresponding to 25 dB gain (power doubled) at 750 or 860 MHz.

-Use single cascade high output level line extenders (or distribution amplifiers) with an output split. In the future upgrade, these devices can feature dual active outputs.

-Use today's lowest technology in a way that results in high performance while meeting current cost objectives.

-Feedforward distribution amplifiers with 37 dB of gain should be used primarily where other, lower technology options are not cost effective.

If it becomes necessary to replace existing trunk amplifier or line extender housings, use new equipment featuring housing and platforms that have been designed for 860 MHz or 1 Ghz bandwidths.

Improving Reliability

The topic of improved reliability has been widely discussed in many industry forums during the last two years. One of the most straightforward means to improve perceived reliability is by shortening cascades. It is important to note that the value of shortening cascades applies not only to trunk amplifiers, but also to line extenders and even taps. The reliability improvements result from having fewer devices between the headend and the customers. The probability of an outage is proportionally reduced. Another parallel improvement in reliability comes by having fewer customers served by any critical device. A critical device could be defined as one whose failure would result in a total loss of cable television service to the customer.

Another way to improve reliability is to reduce the amount of plant where 60 Volt line power is present. Since a significant percentage of plant outages are related to powering or powering caused problems, by reducing the amount of plant required to carry line power, the probability of problems are reduced. In a short cascade node structured architecture, if power was removed from all tapped feeder lines, the reduction in the number of connectors that are required to pass power would drop to perhaps one half of the original amount.

Reducing and Simplifying Maintenance Needs

The following guidelines will reduce and simplify maintenance needs.

-Adopt a system architecture that features greatly reduced cascades (amplifiers, line extenders and taps). In this way the need for system sweeping activity can be effectively eliminated.

-Reduce the need for automatic gain control, and use exclusively amplifiers featuring plug-in pads and equalizers for level setting (ie., few or no remaining field adjustments). The need for system balancing and level set-up will be significantly reduced.

-Strive to power less of the plant. Power the rest of the plant more efficiently. With a reduction in the total number of power supplies, especially when stand-by power supplies are used (their contribution to perceived reliability is less with very short cascades), maintenance needs wil be reduced.

METHODOLOGY

Keep Options Open

In the upgrade planning process, there are no rules which require that a single architecture or approach be implemented exclusively throughout the area to be upgraded. It is important in the beginning not to exclude any options. A mix of two or three architectures, if careful planned, should not unduly complicate maintenance activities. In many systems, a single architecture simply cannot accomplish all the objectives stated earlier in this paper. The goal therefore might be to develop an upgrade plan that would result in as much plant as possible meeting all of the previously stated objectives. At the end of a 550 MHz upgrade project, if 60% of the plant can be further upgraded to 860 MHz in a simple manner (ie., a low cost amplifier module swap and a few added lasers), and the total project cost was competitive with all other 550 MHz upgrade options, capital funds will have been spent in an optimum manner.

It is useful to understand the strong points of each of the architectures listed below, and the degree to which each meets the previously stated objectives.

-Fiber Backbone (FBB)

This architecture was introduced by ATC in May 1988 at the National Cable Television Show in Los Angeles. [2] It involves the deployment of fiber optic nodes throughout the system in order to reduce amplifier cascades. A percentage of the trunk amplifiers upstream of the node location are turned around. Existing trunk bridger locations are usually retained. Feeder line rework will normally involve the re-spacing and addition of line extenders. This is probably the least costly approach if a drop-in module upgrade can be achieved that would retain existing trunk housings and locations. As a result of increased line extender cascades, and the use of feedforward trunk modules to provide the necessary gain and distortion performance to retain existing locations, this architecture will most likely require added fiber optic node locations, or major modifications to the coaxial plant when upgrading in the future.

-Cable area network (CAN)

This approach was introduced by Jones Intercable in late 1988. [3] It is similar to the Fiber Backbone Architecture except that all existing trunk amplifiers retain their original orientation (direction) in the cascade after the deployment of the fiber nodes. This gives the added benefit of an additional signal source to backup the fiber path at the node location. If the fiber optic cable feeding the node is cut, a switch in the optical node/bridger senses the loss of signal and switches from the optical detector output to the backup trunk RF signal. This architecture, which requires significantly more fiber nodes than a similar FBB approach for the same final number of trunk amplifiers in cascade, costs more to implement. Comments made regarding the ease of future upgrades to the Fiber Backbone Architecture apply equally to the CAN approach. Depending upon the distortion allocations as a result of future upgrades, the backup trunk feature of the CAN system may become ineffective due to severely degraded picture quality.

-Super distribution

This architecture was announced by Rogers Engineering in 1989. [4] A primary aspect of this architecture deals with the feeder line. It involves the adding of an express cable in parallel with the existing tapped feeder cable up to the last amplifier located at the end of the feeder line. The distribution amplifier used provides one output feeding the next amplifier in cascade and a second output for the feeder line. This second output feeds directly into a splitter which "backfeeds" and "forwardfeeds" roughly equal length tap strings. There are no requirements for power passing taps in this scenario since the feeder lines have no amplification beyond the amp located in the "express" line. The super distribution approach should allow for low cost module (or hybrid) upgrades in the future if the express line amplifiers are appropriately spaced to work with the higher bandwidth amplifiers of tomorrow. The cost for a 550 MHz upgrade using this approach may be higher than other possible solutions (550 MHz) since express cable would need to be overlashed on 60-70% of the total system distribution plant. The express line amplifiers used for 550 MHz can be relatively low performance/low cost since the spacings are quite modest at 550 MHz.

-Fiber to the feeder (FTF)

The Fiber to the Feeder architecture was first described by James Chiddix of ATC at the SCTE Fiber Optic Conference in Monterey California in early 1990. [5] The approach is a logical progression from the Fiber Backbone approach already described. An FTF architecture is a powerful option when existing trunk spacings consume a disproportionate amount of the "distortion budget". The coaxial reach after a FTF node can be increased by re-using existing trunk cable as express cable. Reach can also be increased by adding cable to allow the backfeeding of tap strings. The choice of distribution amplifier and/or line extender technology, and gain/output level capability also effects the reach from the node. The Fiber to the Feeder architecture is a very cost-effective solution for rebuilds, new builds, and those upgrades where simple drop-in amplifier module replacement is not possible. To facilitate further bandwidth expansions, without having to add additional nodes, a slightly different FTF configuration using up to three low gain distribution amplifiers in cascade followed by a single high output level distribution amplifier (DA) or line extender (LE), may be valuable. The latter configuration will be discussed in the analysis section of this document.

-Fiber to the service area (FSA)

This architecture, developed by Scientific Atlanta, [6] is similar to the FTF architecture previously described. It is essentially a multiple star approach. The homes served by a single optical node (the "service area") is limited to a specific maximum number. This number is based on future telecommunications service requirements. In the upgrade case, the optical node is the center of one distribution star. At least two other distribution stars are formed around existing trunk bridger locations. The bridger amplifiers are replaced with a distribution amplifier (with AGC) followed by an output splitter depending on the number of subsequent feeder legs. The former trunk cable is used to connect the node to the DA's at the centers of the distribution stars. As with all of the above mentioned architectures (except for super distribution and single cascade high output level LE/DA FTF), the requirement of additional fiber nodes exists when further expanding the bandwidth by a significant amount. Not unlike FTF, this is a cost-effective approach to 550 MHz upgrades.

Classify The Existing Plant

By classifying the existing plant of a given system into three categories before test design begins, the designer's time can be used more efficiently. The feeder line types are defined as follows:

-Short feeder lines, or long feeder lines where perpendicular access, in order to "break up" the feeder line into short feeder lines, is possible (the good)

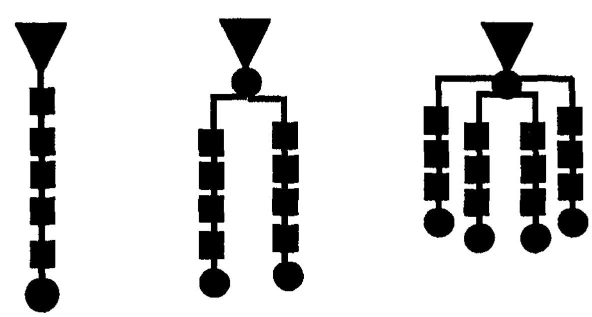
-Long aerial feeder lines (eg. already three plus line extenders in cascade at 300 MHz) with no perpendicular access (the bad)

The Benefits of Output Splitting

-This diagram shows how output splitting after the line extender lowers amplifier cost per mile. This savings can pay for added express cable using single cacade high ouput LE FTF architecture, or for added backfeed cable in other approaches.

Key Parameters

- **Cable- .412 Standard**
- **150' tap spacing**
- **Power Doubled Line Extender**
- **40/49 dBmV Output Level**
- **55/550 MHz**
- **11 dBmV min. tap output**



# Tap Ports	**20**	**36**	**56**
Total Footage	**750**	**1,500**	**2,400**
Amplifier Cost/Tap Port	**$ 11.50**	**$ 6.38**	**$ 4.10**
Amplifier Cost/Mile	**$ 1,619**	**$ 809**	**$ 506**

LE cost = $230 # tap ports/ mile = 126 ==> 100 homes/mile

Diagram 1

-Long underground feeder lines (eg. already three plus line extenders in cascade at 300 MHz) with no cost-effective perpendicular access (the ugly)

It is also important to note the amount of trunk cable present in different parts of the system. While a poor trunk to feeder ratio may have added to the initial construction costs, the presence of "extra" trunk cable today can further reduce the implementation costs of an FTF or FSA type architecture. All of the existing trunk cable can be used either as express cable, or in some cases, a low loss feeder cable.

Use Of A Building Block Approach

It may be useful in the analysis to segment each major component of the system into layers. Possible layers might include:

-The AML microwave link

-An AM fiber super trunk out to a secondary hub (or optical repeater site)

-The AM fiber optic distribution system

-The coaxial trunk, or dedicated express cable, and its amplifiers (if used)

-The distribution amps or line extenders

-The tapped feeder line

Once the layers have been defined, one should establish a first cut performance requirement. One can then confirm that each piece will fit together in a way that meets end of line objectives. Plant cost components may then be analyzed for each scenario, or architecture approach. In this way, cost sensitivities can be developed for the various plant components.

ANALYSIS

Design Observations

Diagram 1 illustrates the potential savings

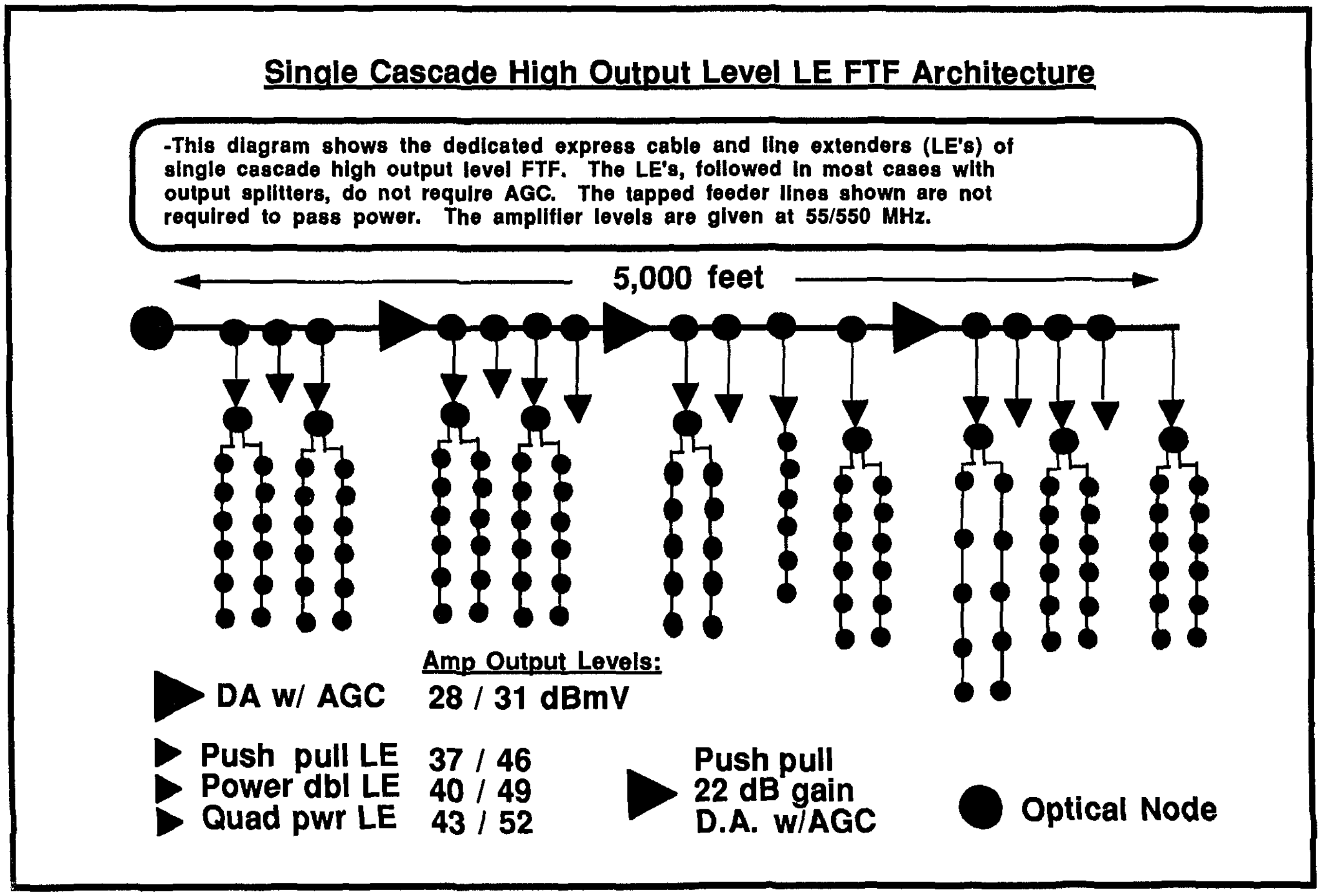


Diagram 2

offered by output splitting of a line extender. Feeder line backfeeds in an FTF or FSA architecture take advantage of this concept to improve reach, and/or minimize feeder electronics. This idea is a key component of the super distribution architecture. Diagram 1 shows three line extenders each operating with the same 49 dBmV output level at 550 MHz. The feeder cable used in each of the feeder strings was .412" standard cable, which has a 550 MHz attenuation specification (measured) of 2.9 dB/100'. The cost of the line extender was assumed at $230. The line extender costs per mile associated with each of the three scenarios are indicated in the bottom row of the diagram. The benefit of using output splitting is evident.

feedforward distribution amp plus a power doubled line extender. If complete stations must be replaced as part of the upgrade plan, there may be ways to save money using a different approach.

A factor to be considered is the price performance ratio of distribution amplifiers versus line extenders. Distribution amplifiers typically have added features and higher prices when compared to line extenders. Where ouput splitting is not required, and the amplifier does not need AGC, the added cost of a DA may offer little in return.

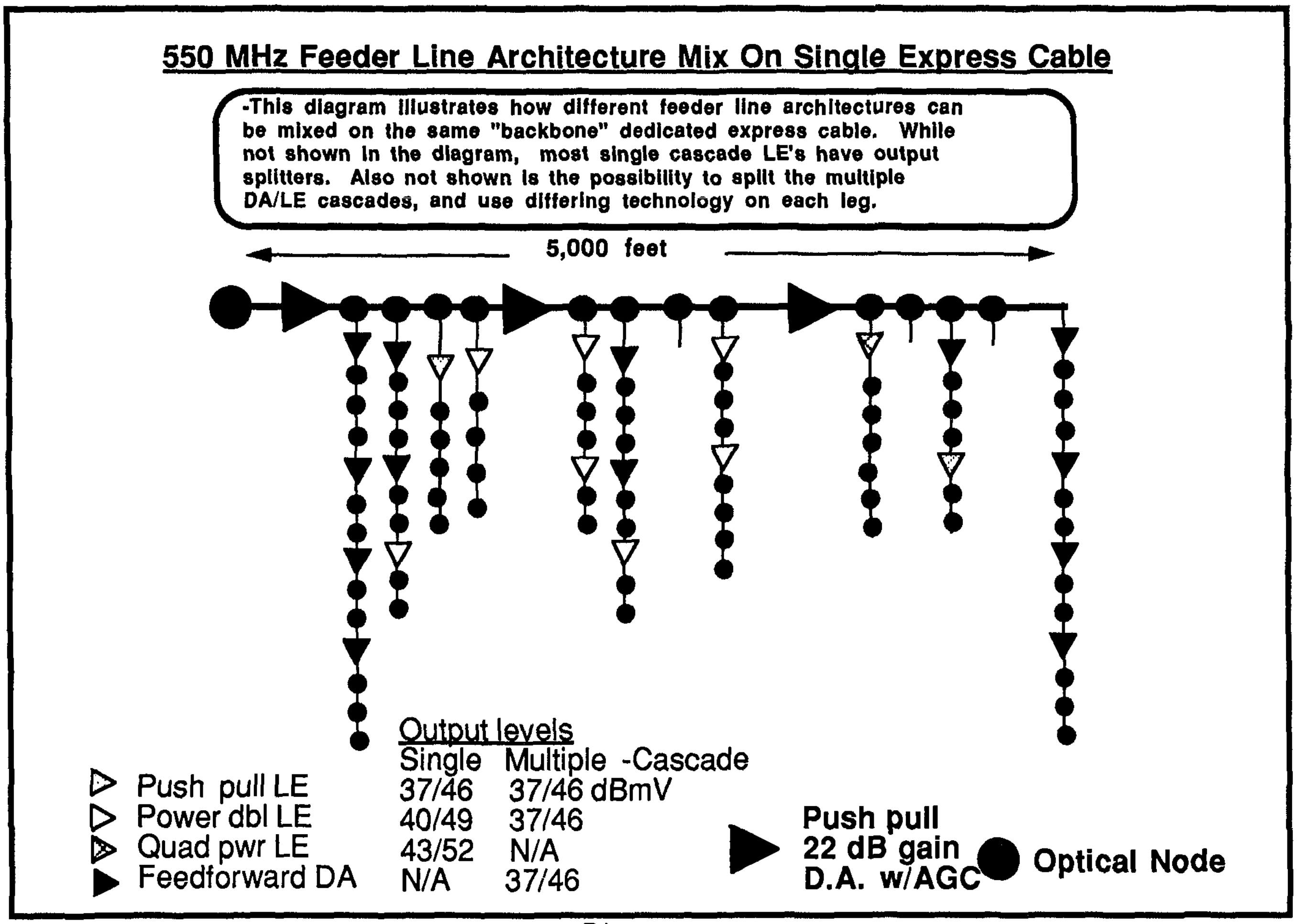


Diagram 3

Another observation concerns the cost of a trunk amplifier station, and its alternatives. With many vendors, the price of a feedforward trunk power doubled bridger station is approximately 50% more expensive than the combined cost of a

Express cable size can significantly influence the maximim reach from the node, there by driving the maximum amount of distribution plant fed by the node. It is important to understand all of the associated cost trade-offs when selecting express cable sizes or types.

In lower density construction, the potential cost savings as a result of using low fiber count optical cable, as compared to "trunk" size coaxial cable may be suprising.

Flexible Architecture and Design

After spending considerable time evaluating how different architectures would fit in a "severe" upgrade, it became apparent that while a given approach was cost effective for one portion of the system, something else worked better in other areas. A severe upgrade in this case is defined as one where even three power doubled line extenders could not reach the end of existing feeder lines. (Remember the good, the bad, and the ugly types of feeder lines mentioned earlier).

Conceptually, it seemed as though an architecture met all of the objectives outlined at the beginning of this paper. The architecture that held the most promise was the single high output level DA/LE version of FTF, shown in Diagram 2. While this approach worked well for the areas of plant consisting of "good" feeder lines, it became apparent that it would not work at all with long underground ("ugly") feeder lines. Implementing this approach on long aerial feeder lines where no perpendicular access was available, would result in a feeder line that resembled the super distribution approach. As previously mentioned, the super distribution approach is somewhat more expensive today as a result of the amount of added cable. Like single cascade high output level DA/LE FTF, super distribution allows cost-effective, minimally disruptive upgrades in the future.

To address the challenges presented by different types of feeder lines in close proximity to one another, the mixing of single cascade high output level LE's (or DA's) with multiple LE (or DA) cascades was considered. Both the single

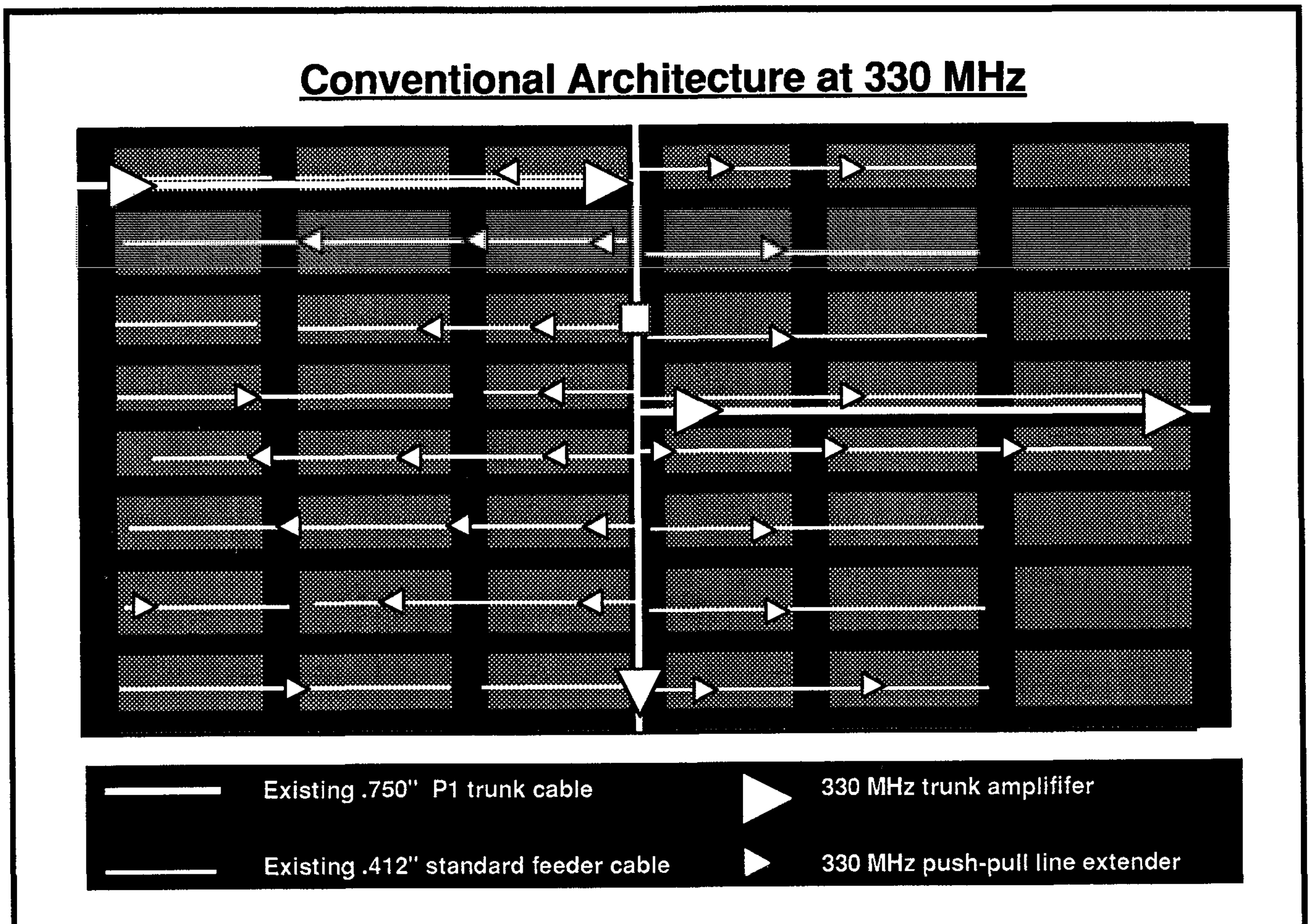


Diagram 4

Single Cascade High Output LE FTF Architecture at 550 MHz

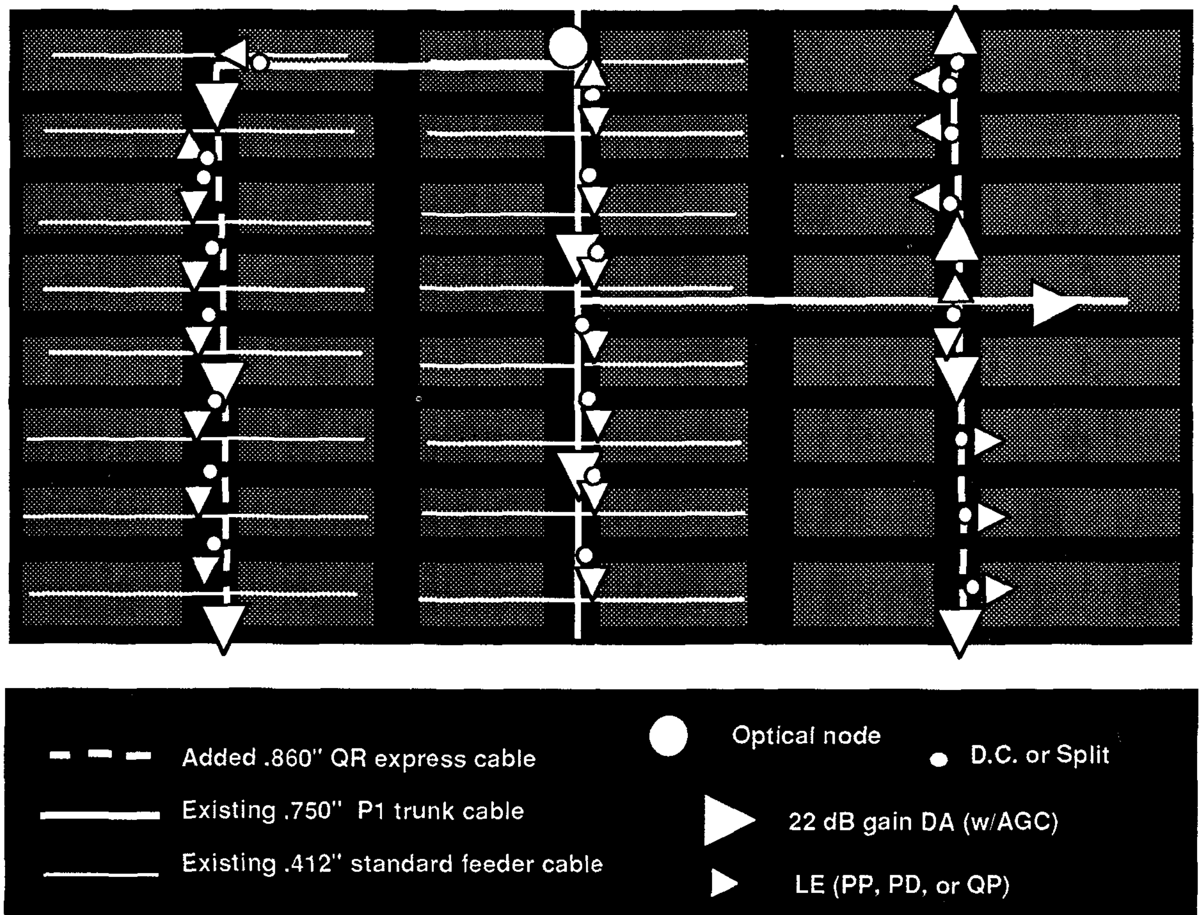


Diagram 5

cascade DA/LE and the multiple DA/LE cascades would be fed by directional couplers from a "backbone" type dedicated express cable (or former trunk cable). This approach is shown in Diagram 3. The option of having up to four LE/DA's in cascade would have allowed the longest underground feeder line (currently three power doubled LE's in cascade at 330 MHz) to be upgraded to 550 MHz in a most cost-effective manner.

In a significant number of cases, however, it was both cost-effective and possible to break existing feeder lines in half by adding an express cable perpendicular to the tap string. At the intersection of the express cable and the feeder line, a line extender (fed by a DC on the express cable) with a two way output splitter would have been installed. Diagram 4 shows the plant layout before the upgrade. Diagram 5 shows the plant after the implementation of single cascade FTF design.

In examining how to upgrade the long aerial feeder lines (the "good"), several conclusions were reached. Backfeeding offered a cost-effective means to reduce the number of line extenders required. In the absence of backfeeding, a cascade of four feedforward DA's would have been requied. Upon closer examination, however, the amount of cable to be added at each of the DA locations was equal or greater than the amount of added express cable required to "break-up" the feeder line. In some cases, strand was available on the future perpendicular express cable runs. In other cases, pole lines were available but stranding would have been required.

Diagram 6 compares the cost of a four feedforward DA cascade with two output split single cascade power doubled LE's. By using the latter of the two approaches, the savings in feeder line electronics cost would clearly pay for some additional express cable. The amount of express

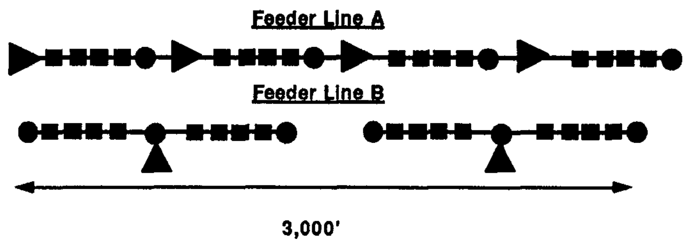
Feeder Line Electronics Cost Comparison at 550 MHz

Parameter	Feeder A	Feeder B
Reach	3,000'	3,000'
Tap Ports	72	72
Tap Output	11 dBmV	11 dBmV
Cable Type	.412 STD	.412 STD
Amp Type	FdFwd	Pwr Dbl
Amp O/P	37/46	40/49 dBmV
Qty. Used	4	2
Unit Cost	$ 450	$ 230
Amp. Cost	$ 1,800	$ 460
Cost/Mile	$ 3,168	$ 810

By using the distribution architecture depicted in feeder line B, as opposed to that in feeder line A (conventional approach), the reduction in LE/DA cost

= $ 2,358 per mile

The savings in electronics pay for the added express cable in single cascade high output level LE FTF designs.



▲ Pwr Dbl Line Extender ▶ Fdfwd Distribution Amplifier

Diagram 6

cable required to implement the single cascade FTF design must be determined in order to compare the total cost of each option. In one upgrade study, approximately 1,000' of .750" express cable per plant mile would have been required.

Design tests with both backfeeding and dedicated express cable/single cascade FTF should be performed on feeder lines in different areas of the system to determine if cost savings are available over conventional approaches.

The goal of the single cascade high output line extender was to feed taps located in the blocks on either side of the added express cable, as shown in Diagram 5. Designing backwards from the end of each block revealed the output level requirement for the line extender. The next step was to select a line extender/DA that would provide the output level required with the lowest technology and cost. The last three rows in Diagram 7 show the output level of a single cascade LE/DA based on a given technology. All output level and cummulative performance specifications shown in Diagram 7 assume an AM fiber optic link and a three DA cascade preceeding the listed amplifier(s). The optical trunk performance used was C/N = 50 dB, CTB = 65 dB, and CSO = 62 dB. The three distribution amplifiers used in the dedicated express cable were 550 MHz, 22 dB gain, push-pull type amplifiers with AGC, operating with an output level of 31 dBmV.

The first four rows of Diagram 7 show the output levels for some of the possible combinations of differering technology LE/DA's. It is important to note that an output level of 46 dBmV was possible for one single LE/DA as well as for two, three, and four DA's in cascade. In addition to the consistant output levels, the end of line performance specifications were almost identical. By selecting the lowest technologies possible for the required cascade, the cost of feeder line electronics (LE's and DA's) can be held to a minimum.

In the version of single cascade high output level DA FTF shown in Diagram 8, the dedicated express cable was fed by a high output level optical bridger. The maximum reach (between the node and the last LE), with the directional couplers installed to feed the single cascade LE's, was approximately 2,500'. As a result of this reach, the amount of plant fed by this node was less than required to be truly cost-effective. Another drawback was the significant

Mixed Feeder Line Architecture Amplifier Output Level Table (550 MHz)

Equipment cascaded	O/P 1	O/P 2	O/P 3	O/P 4	C/N	CTB
FF+FF+FF+FF	37/46	37/46	37/46	37/46	47	53
FF+FF+PD	37/46	37/46	37/46		47.3	53
FF+PP	37/46	37/46			47.6	53
PD+PD	37/46	37/46			47.6	53
PUSH-PULL (PP)	37/46				48	53
POWER DBL (PD)	40/49				48	53
QUAD PWR (QP)	43/52				48	53
FEEDFORWARD (FF)	N.A.					

The output levels shown for each type of amplifier and for its position in cascade, as well as the cummulative carrier to noise ratio (C/N) and CTB performance indicated, assume that the feeder line is attached to the end of a three DA cascade which is fed by a fiber optic node with the output specifications of C/N = 50 dB, CTB = 65 dB, and CSO = 62 dB. The three DA's in cascade are 22 dB gain, 550 MHz , push-pull type amplifiers with AGC, and with a nominal output level of 31 dBmV.

Diagram 7

cable spacing between the node and the last LE. Depending on the range of temperature variations, it may have been necessary to use a DA with AGC to keep output levels within the desired window. This cable spacing would also have created difficulties when upgrading to 860 MHz or 1 GHz. By using a trunk output level from the node, and up to three 22 dB gain DA's (with AGC) in cascade in the dedicated express cable, the requirement for AGC in the single cascade line extenders has been eliminated, and the cable spacing issue at higher bandwidths resolved.

CONCLUSIONS

It is hoped that the reader will have drawn two primary conclusions from this paper. The first conclusion being that it is possible to intermix at least two different types of feeder line architecture, fed by the same dedicated express cable, without requiring 20 different output levels. The resulting product can be one that is cost-effective today, while minimizing tomorrow's upgrade cost for a significant portion of the plant.

The second conclusion relates to the inherent advantages offered by the single cascade high output level DA/LE FTF architecture. To summarize the advantages offered by this architecture:

Unpowered, Short Tap Cascades

By removing the power passing chokes from current taps, the bandwidth can be increased to 860 MHz or 1 GHz with low development costs. In this process, the maximum tap insertion losses are expected to drop back to those specifications currently found at 400 MHz. These taps, if available today, would allow a future upgrade to 860 MHz with little disruption to the feeder line. By simplifying the taps, it is hoped that pricing will decrease, or at least, remain constant.

The Single Cascade High Output Amplifier

In the proposed FTF configuration, the requirement for amplifier AGC would be eliminated except for the few low gain express cable Distribution Amplifiers. Not only will this increase

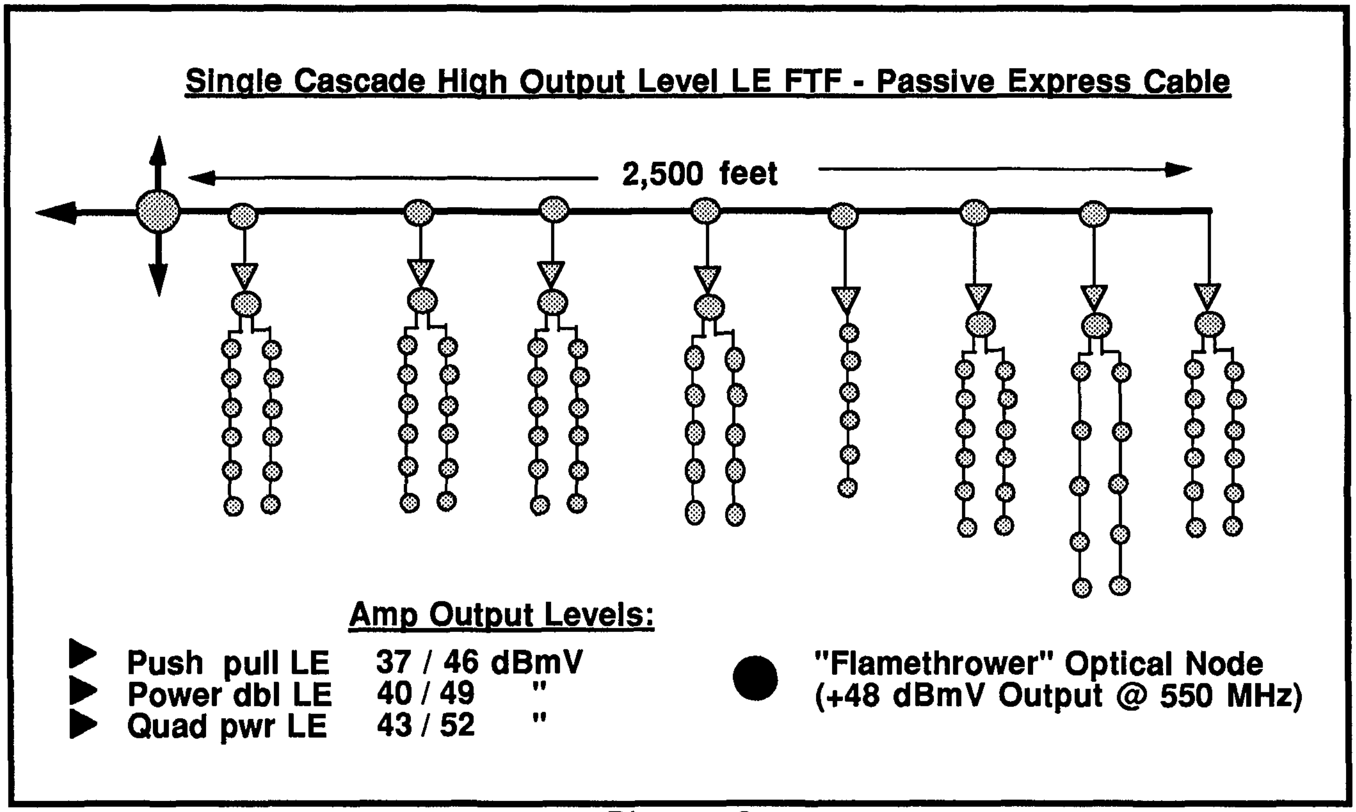


Diagram 8

amplifier performance and gain as a result of lower interstage losses, it will also improve amplifier stability thereby reducing maintenance requirements. By eliminating most LE/DA AGC requirements, the total cost of feeder line electronics can be significantly reduced.

With the LE's or DA's fed directly by a .750 or larger express cable, the power consumption should be less than in conventional plant. In addition, since relatively high output levels can be obtained from push-pull technology amplifiers, further reductions in power consumption can be obtained.

The fact that the express cable spacings are targeted at 22 dB (550 MHz) in addition to using low technology amplifiers whenever possible, this architecture ensures a low cost, minimally disruptive path to higher bandwidths.

Cable Use

Significant amounts of fiber optic cable would be installed when using this FTF architecture in a system upgrade. A moderate amount of coaxial cable, for dedicated express runs or backfeeding purposes, will also be added. By allocating more of the upgrade funds to the purchase of these "unlimited" bandwidth passive components, which can be reused for many years, the percentage of plant assets that may become technically obsolete (in the event of further upgrades) before being fully depreciated is reduced.

SUMMARY

When specifying how to best use available capital to upgrade a system to 550 MHz, the engineer will be faced with many options. Given the increasingly competitive nature of our industry, the long term impact of today's decisions must be carefully evaluated.

The challenge is to select an architecture that will assure the smooth, low cost evolution of today's cable television systems into tomorrow's high performance communications networks, while conserving the shrinking supply of capital funds.

References-

[1] J. A. Chiddix and J.A. Vaughan, "Upgrading Coaxial Distribution Networks with Amplified Taps," *NCTA '91 Tech. Papers*, 1991.

[2] J.A.Chiddix and D.M. Pangrac, "Fiber Backbone: A Proposal for an Evolutionary CATV Network Architecture," *NCTA '88 Tech. Papers*, 1988.

[3] Robert Luff, "The Broward Cable Area Network fiber model," *Communications Engineering Digest*, February 1989, Page 27.

[4] G. Hart and N. Hamilton Piercy, "Rogers fiber architecture," *Communications Engineering Digest*, July 1989, Page 44.

[5] David M. Pangrac, "Fiber trunk and feeder: ATC's new architecture," *Communications Engineering Digest*, May 1990, Page 28.

[6] Robert C. Loveless, "Fiber to the Feeder: A Star Architecture," *SCTE 1991 Fiber Optic Conference Tech. Papers*, 1991.

Biography

Jay A. Vaughan currently holds the position of Senior Project Engineer with American Television and Communications. In September 1990 Mr. Vaughan returned to the United States after a two year assignment in France where he was involved in the engineering and construction of 860 Mhz cable television systems.

Prior to his overseas assignment he held the position of Project Engineer with ATC. Mr. Vaughan has also worked for Rogers Communications, Jerrold Electronics, and others during his fourteen years in the cable television industry. He received his BSEE in Electrical Engineering from the University of Texas in Austin in 1981.

A DIGITAL VIDEO COMPRESSION SYSTEM FOR SATELLITE VIDEO DELIVERY

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Abstract

Compression Labs has developed a complete digital video/audio compression system for transmitting multiple NTSC video signals over a single satellite transponder. The system is operational and equipment production has begun. Using VLSI technology, very low-cost downlink equipment was developed for placement in consumer premises. Although the initial application was for direct broadcast satellite, the technology has application in many video media delivery systems, including cable and over-the-air.

System Description

The Compression Labs satellite communications system is composed of transmitter and receiver subsystems (Figure 1). The heart of the transmitter is the encoder, which compresses a video input to 1.8 Mbps, multiplexes this video with two channels of digital Dolby™ adaptive delta modulation (ADM) audio and a data channel, and outputs this multiplexed bit stream to a satellite QPSK modulator. The Dolby™ audio is coded at 200 Kbps per channel, for a total of 400 Kbps. A data channel of at least 19 Kbps can be sent along with the compressed video and audio. Video and/or audio bits may be replaced with data to yield data rates up to 2.2 Mbps. With forward error correction, total data rate output to the QPSK modulator is 3.0 Mbps. The encoder is controlled by a microcomputer or terminal which issues commands via a serial interface or by entry of commands via a front panel keypad.

The QPSK modulator receives the 3 Mbps multiplexed video/audio/data bit stream and digitally modulates a carrier within a satellite transponder bandwidth to an IF range of 52-88 MHz. The IF signal is then upconverted to the desired satellite GHz band (Ku or C). Carrier selection within the transponder bandwidth is made under computer control or via the front panel switches of the modulator. Multiple channels may be transmitted in a single transponder, with the exact number depending upon link budget parameters such as antenna size, transponder power and others. The system is a single channel per carrier (SCPC) type. A time division multiplex of multiple video channels onto a single carrier at a higher modulation rate is also feasible. The SCPC technique was selected to minimize consumer downlink equipment cost.

The receiver uses conventional antennas and low noise block converters (LNB) to acquire a signal from a satellite, amplify it and downconvert it from GHz to 950-1450 MHz. The Integrated Receiver/Decoder (IRD) contains a satellite QPSK demodulator and a video/audio decoder (Figure 2). The QPSK demodulator selects a given channel within this 500 MHz. Channel selection is via front panel buttons, from an optional infrared remote, or upon command from the uplink via the satellite. Since each channel occupies approximately 2 MHz, 250 channels are simultaneously available. However, the demodulator synthesizer operates in 125 KHz steps, permitting 4000 possible channel frequencies. The output of the QPSK demodulator is a 3 Mbps multiplexed bitstream containing compressed video, compressed audio, data and control bits.

The video/audio decoder forward error corrects the bit stream output by the demodulator and demultiplexes the command, video, audio and data information. The video is decompressed, converted to baseband composite video and output to a monitor. An optional channel 3/4 RF modulator output is available. The compressed Dolby™ Adaptive Delta Modulation (ADM) audio is decoded, converted into analog stereo signals and output to an audio amplifier.

Data at rates of 19 Kbps-2.2 Mbps is available at an external connector. The system operator can define the meaning of the data as desired. For example the data transmission capability can be used to broadcast text, either in ASCII format or facsimile formats. External add-ons interpret the data stream appropriately. For example a facsimile machine could receive a data stream representing Group III encoded fax.

Information such as menus and system status can be displayed on the video screen. Conditional access and hardware encryption are also available. VCR control signal outputs are available to control video recording at the decoder remotely from the transmission source.

Figure 1. Compressed Digital Video Satellite System Block Diagram

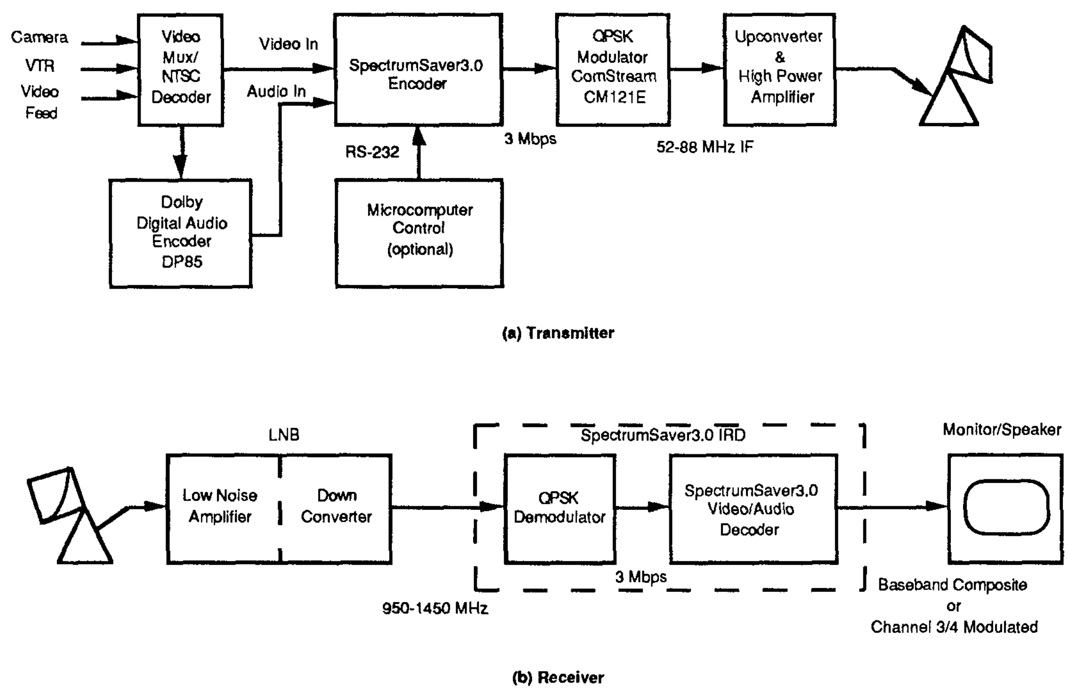
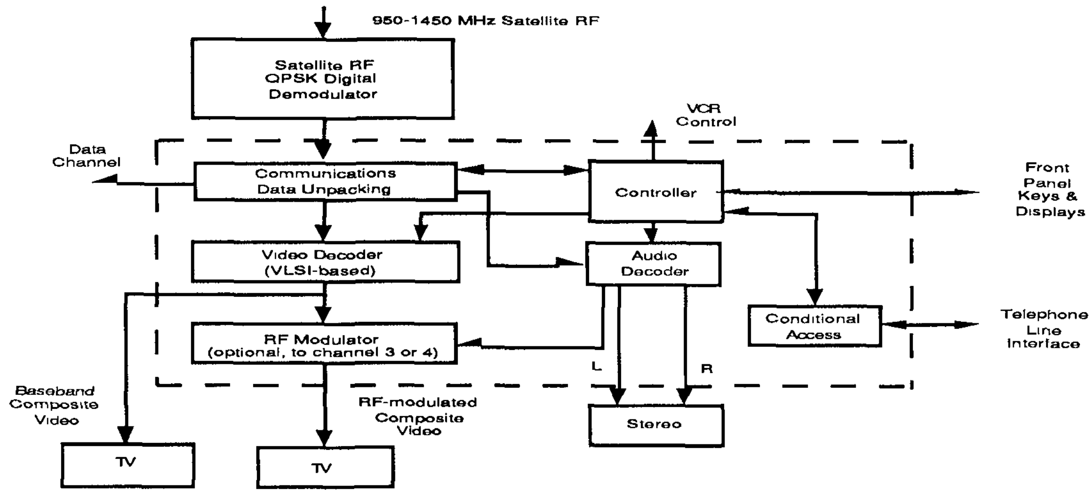


Figure 2. Compressed Digital Video Satellite Integrated Receiver/Decoder Block Diagram



System Design

From its 15 years experience coding many different types of video sequences, CLI made appropriate tradeoffs to optimize this satellite delivery system for its major objectives of:

- 8-16 channels per transponder
- "Better-than-VCR" quality (consumer tested)
- Affordable consumer price for customer premise equipment ($150-$300)

The "better-than-VCR quality" was determined using consumer focus groups viewing simulations at various bit rates. The quality judged acceptable by consumers over a wide range of video sources was realizable at a compressed bit rate of 1.8 Mbps for the video. The system is actually capable of operating at higher and lower compressed bit rates as well. An important quality advantage of digital video is the lack of transmission artifacts such as random noise, ghosting, etc.

Video Compression

The video compression algorithms are derived from proprietary CLI algorithms. They are based on the discrete cosine transform (DCT), variable-length coding, conditional replenishment and motion compensation. The various algorithm parameters were tuned to the requirement of "better-than-VCR quality" for satellite delivery:

- Spatial resolution of 480 lines with 368 pixels per line
- Full color
- Full 30 frames/second motion rendition
- Compression ratio of over 50:1

CLI selected DCT technology because:

- This technology can meet the requirements of the application (quality, bit rate, cost)
- The technology is aligned with international compression standards (Px64, JPEG, MPEG)
- The company has extensive experience with the technology

Key to the excellent performance of the algorithms is the extensive use of powerful custom VLSI to implement comprehensive motion estimation and compensation, as well as special adaptive pre- and post-processing to make appropriate tradeoffs matched to the visual system.

In the consumer IRD, video decoding is performed by three VLSI chips. Two of these are custom and one (a DCT chip) is off-the-shelf. Coupled with half a megabyte of video memory, these VLSI chips implement a low-cost solution suitable for sale to consumers.

Audio Compression

Dolby™ ADM audio compression was chosen because it has good quality, a reasonable compressed bit rate and very low cost. Two channels of compressed audio are provided. Each is compressed to 200 Kbps, for a total of 400 Kbps of compressed audio. At the time of system design decisions, this was the only solution available for less than $10. in the decoder. It is based on a single chip decoder supplied by Philips/Signetics, with the algorithm licensed from Dolby.

Digital Transmission

The RF QPSK modulation technology is based on well-known VSAT data transmission technology. An SCPC technique was selected over TDM because a 3 Mbps demodulator is less expensive than a 30 Mbps demodulator, allowing for the lowest cost of the consumer IRD equipment. A key objective of the system design was to allow use of very small receiver antennas, less than 1 meter. These are low cost and easy to install. To this end, the modem design has an E_b/N_o of 7 dB. Coupled with a proprietary forward error correction technology implemented in a custom chip in the IRD, this modem technology allows transmission of about 10 channels on a medium power satellite such as Hughes SBS-6, with less than a 1 meter receiver antenna.

Bandwidth limitations allow nearly 20 channels to be transmitted. However on medium power satellites the system is power limited. With the launch of high power satellites in future years, the system can be expanded to many more channels, without changing the existing IRD design. More channels can be sent with existing medium power satellites if larger receiver antennas are used.

Conditional Access

The system contains highly-secure conditional access mechanism. Each IRD has a unique address.and encryption key information. A telephone line interface is included to allow for feedback from each box and/or to update key information in the IRD. Key card update of keys is also feasible.

In addition, the complexity and proprietary nature of the compression technology, including the custom VLSI, makes it extremely difficult to reverse engineer the IRD.

Application of Technology to Cable

The technology used in this direct broadcast satellite system can also be applied in cable systems. Two applications are:

- Delivery of digitally compressed program video from programmers to cable headends via satellite
- Delivery of digitally compressed video directly from the headend to consumer homes over the cable

The video and audio compression can be very similar for cable. Delivery to headends for analog transmission to homes may use studio quality video compression to account for signal degradation down the analog cable. This requires higher bit rates. Approximately 4-6 studio quality video channels can be sent over satellites, thereby reducing transponder costs.

The major difference for compressed digital video delivery directly from the headend to the consumer is in the digital transmission technology. A different RF modem from that used in the satellite system is required.

A TEST SYSTEM FOR CONTROLLED SUBJECTIVE TESTING OF CABLE SYSTEM IMPAIRMENTS

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ABSTRACT

This paper describes an integrated test system for generating controlled CATV impairments. This system is being used in a program of subjective testing to evaluate the effects of these impairments on cable television transmission. This study is sponsored by CableLabs and is being conducted at the Jerrold Communications Applied Media Lab in Hatboro, PA. The system described herein is capable of generating controlled levels of five different impairments, singly or in combination. Provision is made for automated system operation, including acquisition of viewer opinions of picture quality.

INTRODUCTION

This paper presents a detailed description of the impairment test hardware originally described by Jeffers [1]. The test system described herein is capable of generating the following impairments to NTSC transmission:

- o Random Noise
- o Distortions
- o Phase Noise
- o Chroma/Luma Delay
- o Micro-reflections

The system is capable of generating these impairments singly and in any combination. The system is an automated system in which impairment levels, viewer response and subjective test data acquisition are all under computer control.

SYSTEM DESCRIPTION

General Description

A block diagram of the test system is shown in Fig. 1. For subjective testing, the input video source is a Pioneer LDV8000 Laser Disk Player. Source material for the tests consists of a series of still sequences which were recorded onto the disk from D2 tape. Weighted video SNR of the source is in the range 53 - 55 dB, depending on the scene selected. A mechanical routing switcher is inserted between the video source and the system in order to facilitate input of test signals for system setup and calibration.

All impairments are generated at RF with the exception of Chroma/Luma delay and phase noise, which are generated at baseband and IF, respectively. A 64 channel headend is used to generate distortion products. Cable ready receivers are used to view the system RF output. The demodulated output is also fed to an Anritsu MS6301B video signal analyzer for measurement of various video parameters. A D2 VCR is available for recording impaired output sequences.

A Jerrold Commander V frequency agile modulator is used for RF conversion of the baseband signal. The modulator output (47.5 dBmV) is attenuated, combined with the headend output and amplified prior to being input to the impairment generation circuitry at a level of 35 dBmV. The frequency of the reference channel modulator was set to Channel 38 (307.26 MHz).

IMPAIRMENT GENERATION

Chroma/Luma Delay

A series of cascaded allpass networks (Fig. 2) are used to generate chroma/luma delay. Each section of the cascade will produce a delay of about 50 nS. The cascaded sections are switched in binary combinations of 100, 200 and 400 nS to generate a maximum delay of 700 nS. System delay was calibrated by inputting a 12.5T pulse and measuring the C/L delay using the video analyzer.

Random Noise

A Noisecom Model 8110 noise generator is used as a noise source. This generator produces random noise in the frequency range 100 Hz to 1 GHz at an output level of 33 dBmV. Noise is added to all of the channels in the system prior to generation of distortion products. A programmable attenuator at the output of the noise generator permits control of the noise level.

Weighted video signal to noise ratio was measured using the Anritsu analyzer. The system is capable of generating SNR's in the range 24 to 50 dB.

Random noise can also be added to the baseband video signal as shown in Fig. 1. A second output from the noise generator is resistively combined with the video signal and the combined signal + noise is fed to the television receiver baseband input. Signal to noise ratios measured at baseband were essentially the same as those measured at RF.

Distortions

A 64 channel headend is used to generate distortion products. The headend uses a total of 16 video sources, each of which is split 4 ways at IF and then up-converted to RF. The headend is capable of being operated in standard, HRC and IRC modes. The headend output and the reference channel are combined and input to a cascade of four Jerrold XRIM-550 amplifiers to generate distortion products. Fixed attenuators are placed between each stage of the cascade and the gain of each amplifier is adjusted to produce 8 dB gain through the cascade. (The 8 dB gain figure is used in order to compensate for insertion losses of the external attenuators and couplers).

Composite distortions are produced by overdriving the cascade. The input level to the cascade can be varied from approximately 16 dBmV (no distortions) to 31 dBmV via a programmable attenuator. A second programmable attenuator is located at the cascade output. The cascade input and output attenuators are ganged so as to produce a constant level of 20 dBmV at the attenuated output of the cascade.

Using the above-mentioned headend, the number of triple beats falling into the reference channel is in excess of 1100. The reference channel receives about 18 second order beats. The range of distortion levels is controllable from -25 to -55 dB.

Phase Noise

Phase noise is generated by inserting a variable phase shift network in the IF loop of the modulator. The phase shift network is driven by a psuedorandom data generator (Fig. 3) whose output is low passed to 300 KHz and amplitude limited under control of the system computer. Phase noise is measured using the method described by Pike and Pidgeon [2]. The system is capable of generating phase noise in the range -59 to -93 dBc/Hz.

Micro-Reflections

The circuitry for generating micro-reflections is shown in Fig. 4. The signal out of the cascade is split into five separate paths: an undelayed path and four delayed paths. RG-11 cable, having an attenuation of approximately 2.5 dB per 100 ft. at 300 MHz, is used to produce the desired delays. Cable lengths of 50, 100, 200 and 400 feet are used to generate delays of 58.4, 116.8, 233.6 and 467.2 nS, respectively. The cables are trimmed to produce phase coherent delays at the combined output of the delayed and undelayed signals. Delay paths may be selected individually or in combination to produce single or multiple reflections. Fixed attenuators in each delay path are used to set equal signal levels through each delay path. A single amplifier and programmable attenuator are used to control the level of the delayed signals relative to the undelayed signal.

After combination, the delayed signals are fed through a variable delay network consisting of short pieces of cable which are cut to provide delays of 1/8, 1/4, 1/2 and 1 wavelength at the reference channel. This unit serves as a "digital trombone" to permit generation of in-phase and out-of-phase delays. The system will generate micro-reflections up to 0 dB relative to the level of the main signal.

CONTROL AND DATA ACQUISITION

All critical system functions are automated. An IBM compatible PC is used as the system controller. The computer is equipped with a National Instruments xxxxx IEEE-488 interface for instrument control. A Hewlett-Packard HP3497 Control and Data Acquisition unit serves as the interface between the IEEE-488 bus and the programmable attenuators and switches. These elements are controlled by contact closures from a series of HP44428A Relay Actuator cards in the HP3497 controller. Manual control of the system is also provided via a switch panel as shown in Fig. 5. An example of the control circuitry is shown in schematic form in Fig. 6.

The system control elements are Alan Industries Model 75MDA127 programmable attenuators and Tri-Lithic 7002F coax switches. The attenuators are programmable in 1 dB increments up to 127 dB. Insertion loss is about 3 dB and frequency response is withiin ± 0.7 dB over a frequency range of 0 - 1 GHz. The switches have a maximum insertion loss of 0.2 dB up to 650 MHz.

An H8568B Spectrum Analyzer serves as the principal instrument for RF measurements. The analyzer interfaces directly to the IEEE-488 bus.

The system is also capable of automated recording of viewer opinions of picture quality. This is done via a handheld device, known as the Subjective Quality Input Device (SQUID) which interfaces to the system via the HP3497. A photograph of the SQUID is shown in Fig. 7 and a schematic of the device is shown in Fig. 8. A linear potentiometer, biased to read from 1-5V in order to correspond to an impairment scale of 1-5, is set by each viewer to reflect his/her opinion of the picture quality. After a selection is made, the ENTER button is depressed, causing the SQUID's sense output to change state. The computer polls the sense line of each SQUID and, if the sense output of a particular device is high, that device's data output (i.e. - the voltage corresponding to the pot setting) is read by the computer.

Each SQUID interfaces to the computer via connections to an HP44421A Analog Multiplexer card in the HP3497. The SQUID outputs are read by the internal DVM in the HP3497 and the viewer assessment data are stored in the computer data base.

SYSTEM SOFTWARE

All of the control and data acquisition software for the system is written in the C language. The operator interface to the system control functions is menu driven. The operator interface menu is shown in Fig, 9. Selection of a particular test is made via the PC's function keys (F1 - F10).

Once a test has been selected, the program automatically steps through the test sequence. The system control functions include laser disk frame selection (via an RS-232 interface), setting of attenuators and switches and polling of viewer responses. These functions are repeated several times for each test sequence.

Test sequence control is accomplished via a series of test scripts which are called by the program. These test scripts are ASCII files which can be modified using a text editor to facilitate addition and/or changes to test sequences as required. A sample test script is shown in Fig. 10.

System re-calibration is also accomplished via a series of ASCII files. These files are simply lookup tables which relate attenuator settings to corresponding impairment values.

SYSTEM CONFIGURATION

A photograph of the impairment test system is shown in Fig. 11. The impairment generation circuitry is housed in two racks. The left hand rack contains the manual control panel, the video and RF circuitry and the system power supplies. The right hand rack houses the delay cables, the HP3497 controller, the spectrum analyzer and a 13" TV receiver which is used by the system operator to monitor picture quality and impairment levels.

RECEIVER MEASUREMENTS

27" cable ready receivers are used for all subjective tests. The receivers were purchased from local distributors and, presumably, exhibit typical product performance characteristics.

Signal to noise measurements were made on four receivers. The video source for these measurements was a Tektronix 1910 signal generator having a weighted video SNR of about 61 dB. The signal was modulated using the Commander V (SNR ≈ 58 dB) and fed to the receiver RF input. An attenuator in the receiver input line was used to adjust input signal levels. The input level was measured using the spectrum analyzer. Signal/noise ratios were measured at each receiver's baseband output using a Rhode & Schwartz UPSF2 video noise meter.

Fig. 12 presents the results of the SNR measurements. The knee for most of the curves occurs at about +5 dBmV with a spread of about 6dB in the individual receiver SNR's at this point.

CONCLUSIONS

An automated test system, capable of generating typical cable system impairments, has been developed for CableLabs' program of NTSC subjective testing. It is expected that this system will see increasing use for both subjective and objective testing in a simulated cable environment.

REFERENCES

[1] M. Jeffers, "Controlled Subjective Testing of Cable System Impairments to Picture Quality Using Psychophysical Methods", NCTA Technical Papers, 156--159, 1990.

[2] R. Pidgeon, D. Pike, " Oscillator Phase Noise and its Effects in a CATV System", NCTA Technical Papers, 1988.

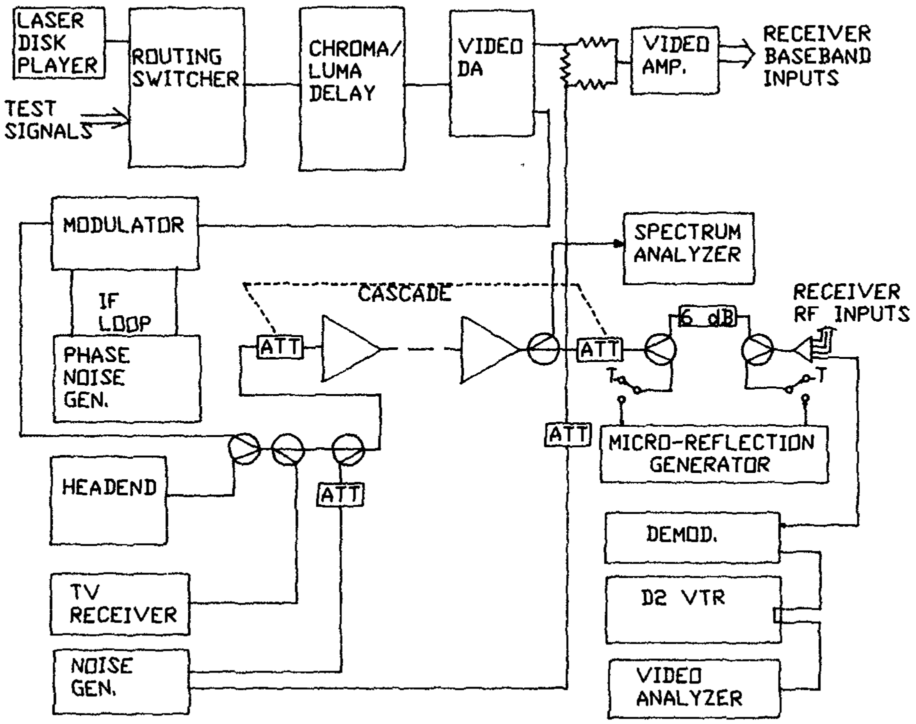

FIG. 1 SYSTEM BLOCK DIAGRAM

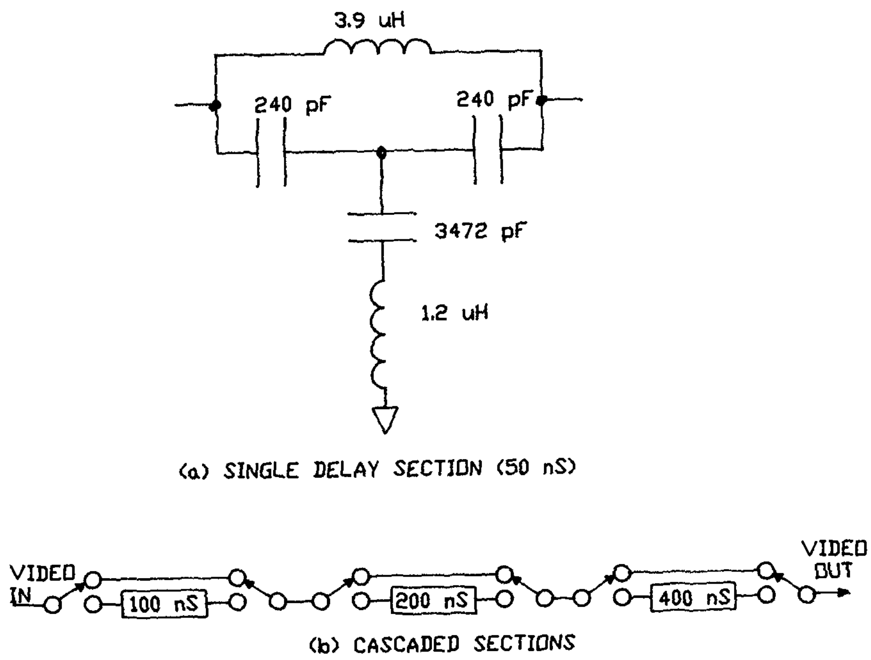

FIG. 2 CHROMA/LUMA DELAY GENERATOR

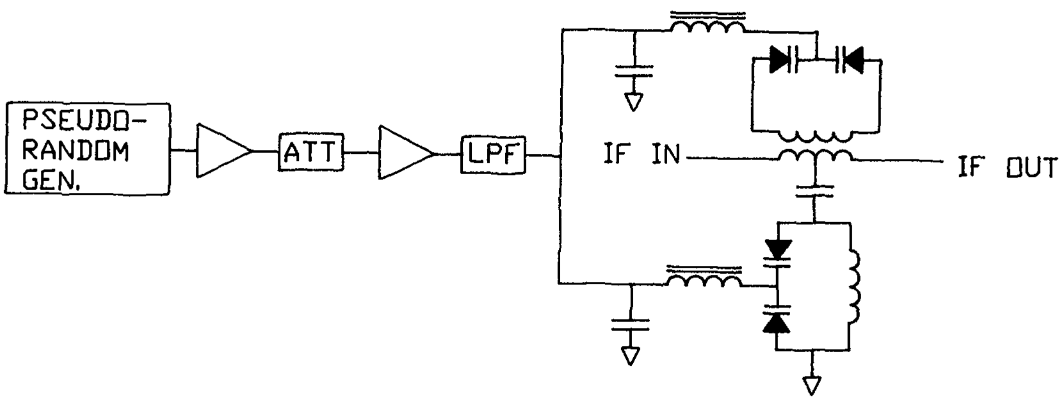

FIG. 3 PHASE NOISE GENERATOR

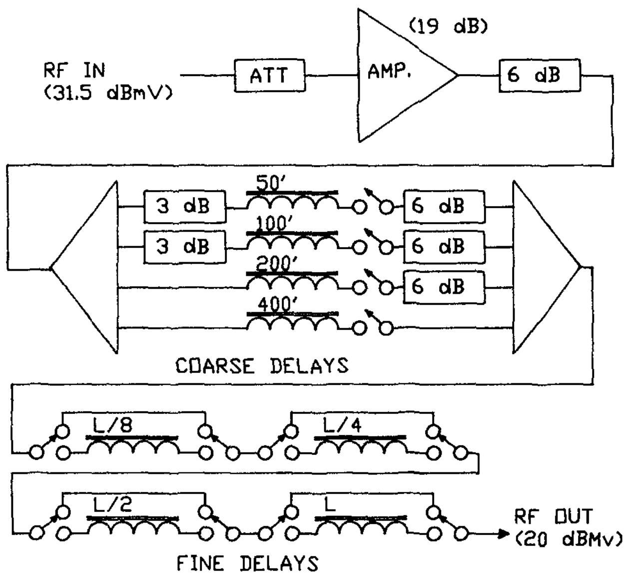

FIG. 4 MICRO-REFLECTION GENERATION

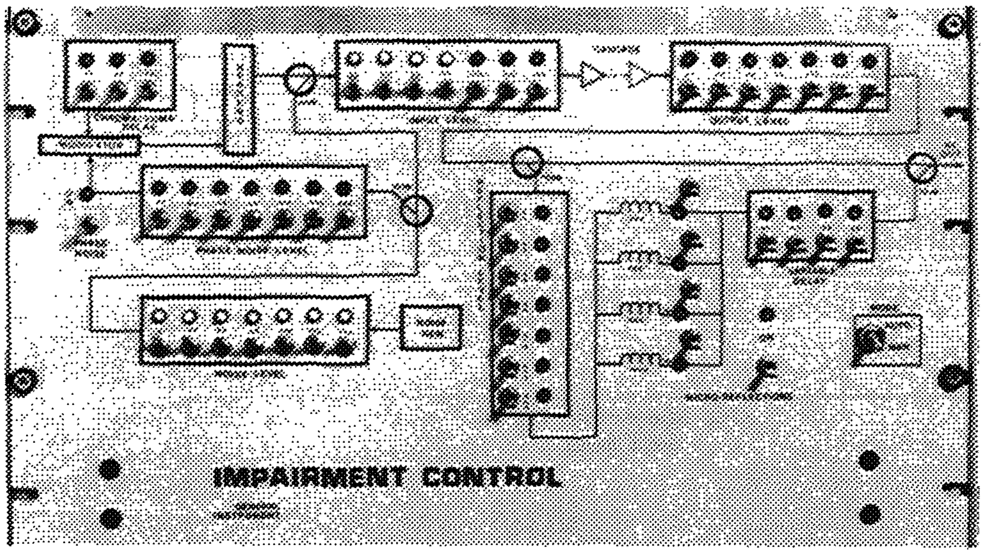

FIG.5 SYSTEM CONTROL PANEL

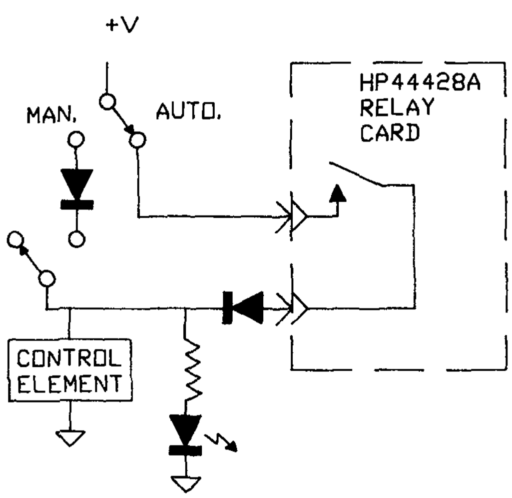

FIG. 6 CONTROL CIRCUIT SCHEMATIC

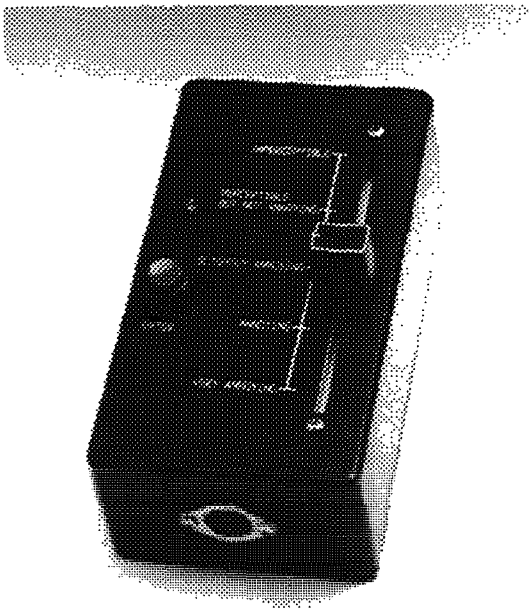

FIG.7-SQUID

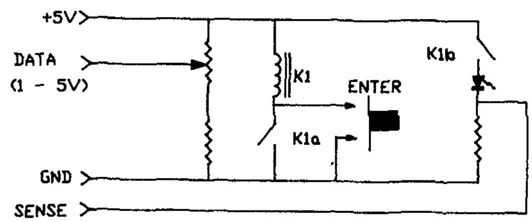

FIG. 8 SQUID SCHEMATIC

NTSC Subjective Tests
Main Menu

F1	Noise Test
F2	Phase Noise Test
F3	IM-2 Test
F4	IM-3 Test
F5	Envelope Delay Test
F6	Micro-Reflections Test
F7	Set Up
F8	
F9	
F10	Exit

FIG. 9 OPERATOR INTERFACE MENU

```
# Script file to control test setup

viewers          2      1      2      3
frames           2      35     230    3290            'balloons Shirley night_ext
presentations    1
noise            7 14 20 24 26 30 36 50               'attn settings
phase_noise      6 15 20 25 30 35 40                  'attn settings
im2              7   0 2 4 6 8 10 12                  'input attn settings
im3                                                   'not used
env_delay        5 50 100 200 300 400                 'ns
micro_refl_L      6 54   56   58   60   62   64       'attn settings
micro_refl_P      2  0   180                           'phase
micro_refl_D      4  58  116  233  467                'delays (ns)
```

FIG. 10 TEST SCRIPT

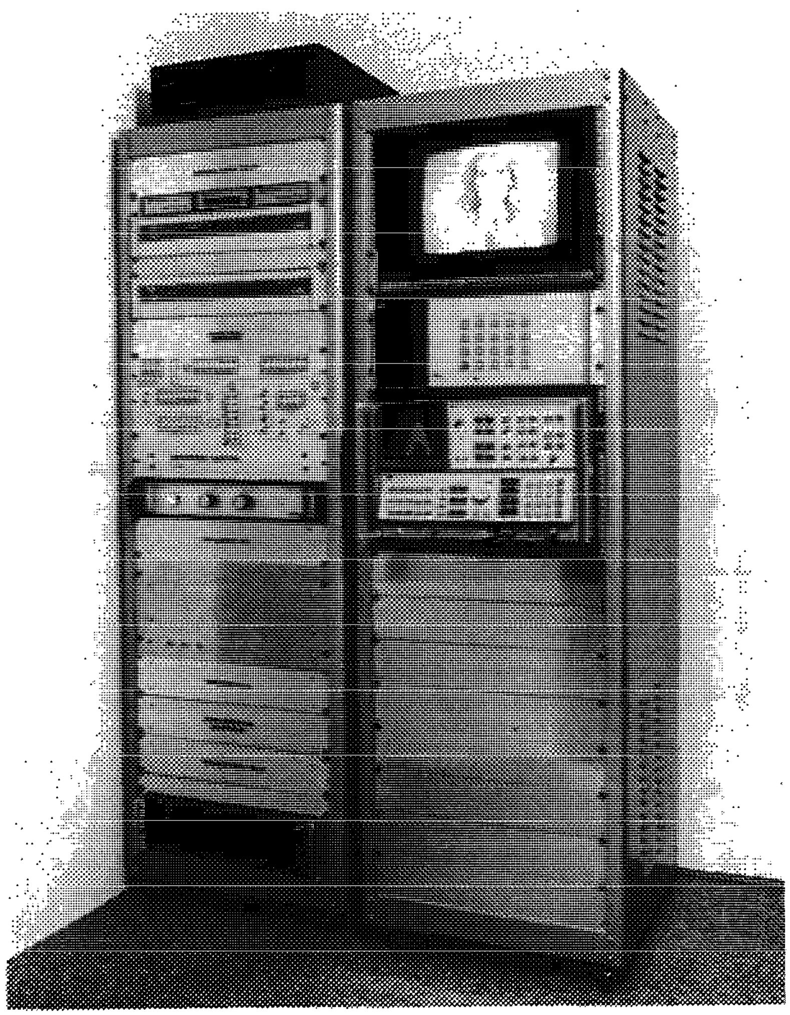

FIG.11 TEST SYSTEM CONFIGURATION

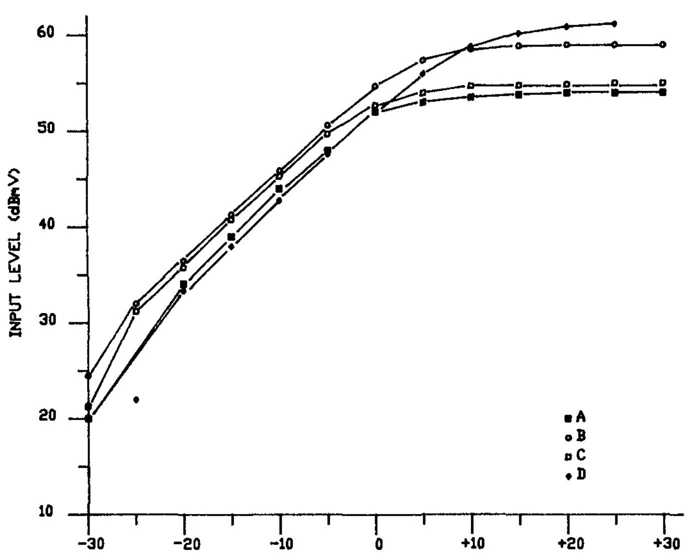

FIG. 12 RECEIVER SNR MEASUREMENTS

ADVANCED AUDIO FOR HDTV

Systems including Data Requirements

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ABSTRACT

The FCC Advisory Committee's Planning Subcommittee Working Party 1 defined the attributes of high definition video. The Commission has placed a channel bandwidth limitation of 6 mHz for the terrestrial broadcasting system. Working Party 1 has described HDTV audio quality requirements as being CD like. The HDTV testing laboratory in Alexandria, Virginia is measuring only two-channel stereo systems. Many papers describing studies that compare two-channel and multi-channel stereophony have been written by CBS in the U.S. [1], NHK in Japan [3], BBC in the United Kingdom [8], and IRT in Germany [7]. The studies agree that with two-channel stereophony the center signal (phantom or virtual) image is distorted by listening position. As screen size increases, the problem is exaggerated. All four organizations recommend that additional channels (speakers) are necessary to produce a stereophony image that can be viewed from a wide angle without distortion. This paper proposes that HDTV audio use a center channel to stabilize the audio image.

Introduction

The purpose of this proposal is to provide a starting point for the ATSC T3/S3 Specialist Group on Digital Services to define the non-video related objectives for HDTV systems. After committee review the document may be sent to proponents, existing video users and providers of ancillary services, and others for comment. This draft defines the program audio, other program-related audio services, audio program control, and ancillary services. The document does not address the ghost canceling training signal. This signal is part of the HDTV video system.

With the imminent introduction of simulcast high definition television, the North American television industries have the opportunity to introduce an advanced digital audio system that will match the viewing experience of the new video service. Recent advances in low bit rate coding techniques make it possible to achieve state of the art quality audio with bit rates as low as 128 kbit/s for each program audio channel. The low bit rate digital audio systems make it possible to transmit the number of discrete audio channels necessary to complement the high definition video.

Proposed Program Audio Attributes

1. The quality of the HDTV program audio will complement the full potential of the new video service.

2. The system will be designed to produce a stable audio image and enhance the ATV visual image.

3. Discrete audio channels should be used. The matrix process is not used in the four channel system. Linear transmission systems will be incorporated, preventing the distortion of the audio image.

4. Viewers will have the option of controlling the dynamic range of the reproduced audio.

5. The four channel simulcast audio system will also accept for transmission: three channel stereophony, two channel stereo, and monophonic programs.

6. The four channel program audio will be convertible to

BTSC stereo for NTSC simulcast.

7. A stereo second language audio program is provided (two channel). Alternately, these channels can be programmed as two separate monophonic channels.

8. A dedicated DVS channel for the visually impaired is provided.

9. The audio system should be more robust than the video to bit errors. Audio should not fail prior to video.

Ancillary Service

1. Teletext service is provided.

2. Conditional access channel is provided.

3. Expandable digital services is provided.

4. Closed Captioning is provided.

5. Program Guide channel is provided.

DESCRIPTION OF SERVICE

Main Program Audio

The main program audio consists of four discrete state of the art digital audio channels. The channels are center, left, right, and rear (surround). The center channel is a discrete channel similar to the center channel used by the film industry. The rear channel, a surround sound channel, is viewer optional. Audio production techniques are the same or similar to those used in the film industry. A programmer-originated control signal will allow the receiver to be switched to a viewer preselected mode of operation. Industry operating standards and recommended operating practices should be developed for the SAP switching nomenclature and logic to prevent the confusion that exists today with BTSC receivers.

Program Audio Expander (Viewer Optional)

To avoid viewer complaints of loudness, program providers have had to limit the dynamic range of the audio signal. A digital expansion channel is provided to allow viewers the ability to expand the audio to its original dynamic range. This channel will control a digital audio expander in the receiver that will restore the original dynamic range to the program channels. Care in implementing this service must be exercised to maintain integrity of the audio image. Twenty kbit/s of data is available per audio channel for the expander control.

Separate Audio Program (SAP)

A two channel full quality stereo service primarily for second language programming is provided. Alternately, the stereo service can be used for two separate monophonic program channels. With user preselection and a programmer-originated control signal, the receiver will switch on command from stereo SAP to one of the preselected monophonic SAP channels.

Descriptive Video Service (DVS)

DVS is a monophonic audio channel that provides program descriptive information to the visually impaired. Because of the growing interest in this service and of possible scheduling conflict with other SAP services, a channel dedicated to DVS is provided.

Teletext

Teletext service of 100 page cycle and 1000 characters per page (8 bits/character) transmitted in 20 seconds requires a bit rate of 40 kbit/s. It should be noted that in some situations teletext may be a program related service.

Other Digital Services

This channel is reserved for the possible interconnection with home digital devices, like computers. The channel could be used for educational or interactive services.

Program Guide

The proposed program guide should be capable of transmitting text accompanied by graphics. For viewer channel scanning, six lines of text identifying the on air show and the next up-coming program are provided. This receiver scanning menu will be refreshed at a one second rate. The data for a complete multi-page program guide is interleaved with the receiver scanning data.

Closed Captioning

Closed captioning is transmitted at a rate of 500 bit/s. To allow for the expansion of this system, a 2 kbit/s channel is provided.

Conditional Access

Because conditional access systems' specifications are usually proprietary for security reasons, detailed specifications on the signal may not be available. The ATSC Specialist Group on Interoperability and Consumer Product Interface T3/S2 is discussing standardization issues relating to scrambling and conditional access.

Capacity Objectives

It is clear that sufficient capacity needs to be allowed in Simulcast ATV systems for advanced audio and data services. When allocating the bit rates for the video, audio and data channels it may be necessary to make tradeoff in channel performance in order to stay within the 6 mHz. Based on information supplied by the ATV proponents, it appears that the 6 Mhz channel has the capacity for transmitting just under 20 Mbit/s of digital data. With overhead, the non-video service should represent less than 9% of the 6 Mhz digital channel's capacity.

The services can be broken down into two categories; program related and program unrelated.

Table 1. Advanced Audio Simulcast HDTV

	Service	DATA RATE kbit/s
PROGRAM RELATED SERVICE	Main Program, Four Channels	512
	SAP Stereo, Two Channels	256
	DVS, One Channel	128
	Expander Control Data	140
	Program Guide	10
	Closed Captioning	2
	Program Mode Control	2
	Conditional Access	400
PROGRAM UNRELATED SERVICES	Teletext Services	
	Other Digital Services	
	Overhead	To be determined by System Proponent
Total	Bits	1450

BACKGROUND ON FOUR CHANNEL SYSTEM SELECTION

Many papers have been written about the problems with two channel stereo television audio [1,2,3,4,5,6]. Papers from the BBC and IRT (FDR) [7&8] all characterize the image distortion experienced with two channel stereo, including recommendations they have submitted to the CCIR. From the data in the IRT paper [4], Figure 1 was drawn illustrating the image distortion that will be experienced when listening to stereo audio with two speakers spaced by 6 1/2' and listening at distances of 2' to 8'. This illustration shows that when listening to the stereo at a distance of 8', while 20" off the center line, the audio image will be shifted by 40%. At the price of separation most BTS receivers have the speakers installed adjacent to the screen to minimize image distortion as illustrated in Figure 2.

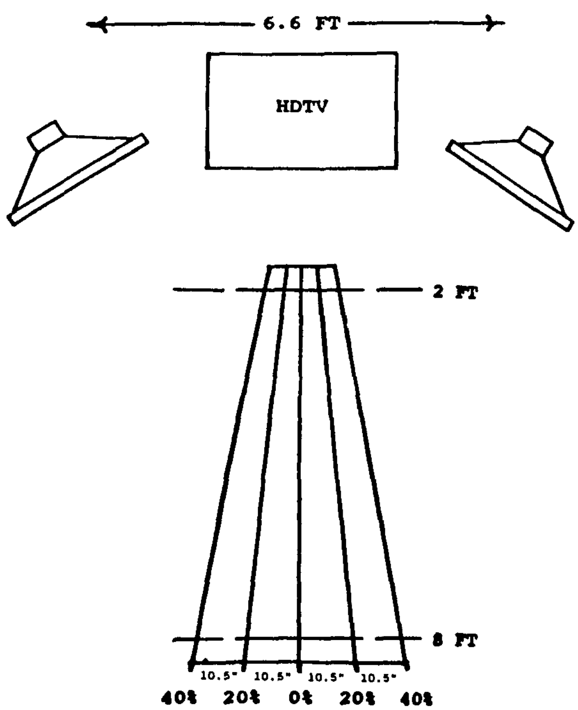

Figure 1. Stereo Image Distortion

The BBC in its recommendation to the CCIR suggested that a three channel system be used as the most practical approach to reduce the image distortion problem [6]. NHK in a study of eight different speaker combinations settled on a four speaker system; left, center, right and surround [2]. IRT in their papers concluded that four front and four surround speakers should be considered [4]. In a paper published by Torick of CBS, a three speaker system was recommended to resolve the image distortion problems [1].

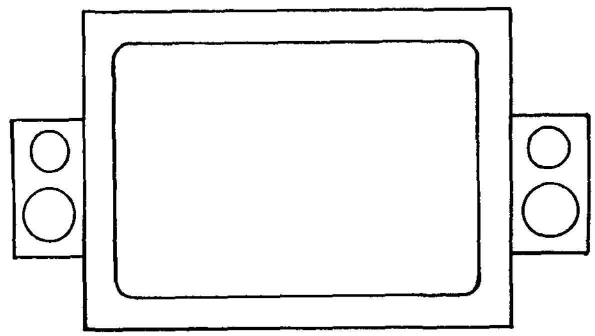

Figure 3. BTSC Stereo

Based on the literature published by the major international broadcasters and the experience of the film industry, a four channel system with a center channel and surround sound is proposed. Figure 3 illustrates the speaker configuration for this system. This proposal is a practical combination for image stabilization, viewing experience enhancement, and transmission channel numbers.

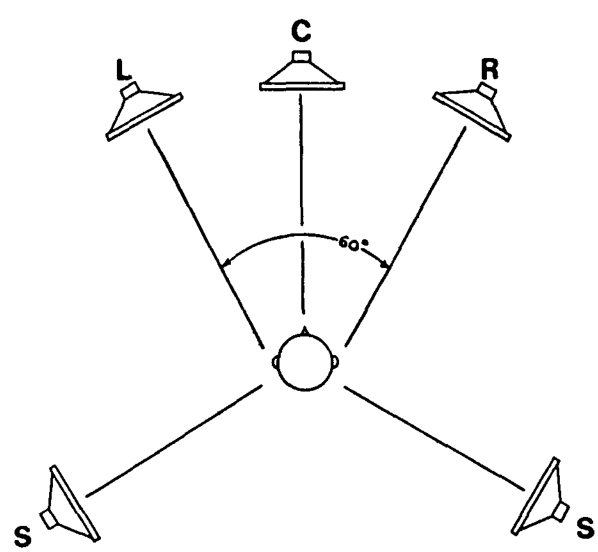

Figure 3. System Speaker Placement

TABLE 2. CHANNEL ACTIVATION WITH TRANSMISSION MODE

TRANSMISSION MODE	CH1	CH2	CH3	CH4	CH5	CH6	CH7
1 4 Channel	on	on	on	on			
2 3 Channel	on	on	on	off			
3 Stereo	off	on	on	off			
4 Monophonic	on	on	on	off			
5 SAP Stereo*					on	on	
6 SAP/1*					on		
7 SAP/2*						on	
8 DVS							on

CHANNEL DESCRIPTION

Ch-1 Center
Ch-2 Left
Ch-3 Right
Ch-4 Rear (Surround)
Ch-5 SAP Left or SAP #1
Ch-6 SAP Right or SAP #2
Ch-7 DVS

TABLE 3. LOUDSPEAKER ACTIVATION BY CHANNEL

TRANSMISSION MODE	CENTER	LEFT	RIGHT	REAR
1 4 Channel	on/ch-1	on/ch-2	on/ch-3	on/ch-4
2 3 Channel	on/ch-1	on/ch-2	on/ch-4	off/ch-4
3 Stereo	off/ch-1	on/ch-2	on/ch-3	off/ch-4
4 Monophonic	on/ch-1	on/ch-1	on/ch-1	off/ch-4
5 SAP	off	on/ch-5	on/ch-6	off
6 SAP/1	on/ch-5	on/ch-5	on/ch-5	off
7 SAP/2	on/ch-6	on/ch-6	on/ch-6	off
8 DVS	on/ch-7	on/ch-7	on/ch-7	off

References

[1] Torick (CBS), "A Triphonic Sound System for Television Broadcasting," SMPTE, 1983.

[2] Ohgushi Kengo (NHK), "Sound System Suitable for HDTV," NHK Laboratories publication, 1988.

[3] Komiyama (NHK), "Subjective Evaluation of Angular Displacement between Picture and Sound 'Directions for HDTV Sound Systems," J AES, 1989.

[4] Theile (IRT), "On the Performance of Two-channel and Multi-channel Stereophony, "AES reprint, 1990.

[5] Holman (Lucas film and USC School of Cinema and Television), "Surround Sound Systems Used with Picture in Cinemas and Homes," AES Conference Proceedings, 1990.

[6] CCIR 10/215 (United Kingdom), "Sound Systems for HDTV," 1989.

[7] CCIR 10/267 (Federal Republic of Germany), "Suitable Number of Sound Channels to Accompany Wideband HDTV," 1989.

[8] Meares (BBC), "HDTV Sound Systems: One Broadcaster's View," a contribution to the CCIR IWP 10/12, 1989.

ADVANCED TELEVISION RESEARCH ACTIVITIES AT CABLELABS

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Abstract

The broadcast and cable television industries in North America are on the threshold of a new era. Seven organizations are competing to have their advanced television system concepts adopted as North American standards for transmission of either High Definition Television (HDTV) or for emission of improved or extended definition NTSC to consumers' homes. This paper will describe overall activities in this area, as well as CableLabs' preparations to conduct the rigorous technical tests of these systems that will be needed to protect the cable industry's interests in the era of advanced television. CableLabs is also embarking upon efforts to further the development of digital transmission of compressed NTSC signals over cable systems, a capability now widely recognized as having major strategic importance to the continued growth of our industry.

Worldwide, there are two main areas of HDTV development underway. First, there is the activity targeted at developing a worldwide production standard, or at least a family of related standards, for the production of high definition television programming. This has been going on in earnest for about eight years.

Second, there are the more recent efforts, to develop transmission systems to get those advanced pictures to the home. Obviously, cable operators are most interested in the latter, and it is this activity that will be reviewed in this paper. I will also describe CableLabs' activities to coordinate development of the technology for the digital transmission of compressed NTSC over existing and future cable television systems.

The ATV Proponents

When it comes to transmission via cable and over-the-air broadcast, there are six independent organizations which are formally proposing systems for for adoption as an FCC standard for the United States.

They include systems from NHK (Japan Broadcasting), Philips Laboratories, MIT, Zenith, General Instrument, and the David Sarnoff Research Center.

There is also the SuperNTSC™ system from Faroudja Research Enterprises – an improved NTSC system that in 1990 was withdrawn from the formal FCC standardization process in the belief by its developer that it can be commercialized without FCC approval.

Some of these transmission systems are still being designed – some in fact are being recently redesigned as all-digital systems. CableLabs will conduct laboratory and field testing of these systems, with lab tests to begin in April 1991 and to finish during the second quarter of 1992.

The Market for Advanced Television

Whatever the outcome of those tests and the standardization to follow, the best estimates are that advanced television will happen slowly, with startup service during this decade, but at relatively low consumer receiver penetration. Beyond the year 2000, though, most observers expect major penetration of advanced television, and an eventual, total replacement of NTSC perhaps twenty-five or thirty years from now.

What Cable Must Do

For now, cable must make efforts to forward the interests of the industry and of our subscribers. We are the delivery mechanism for well over half the households in the U.S., and we need to influence the design of this soon-to-be-standardized advanced television system so that it works well over cable systems. The ATV system needs to take into account cable's particular set of transmission impairments, which are quite different from those of the over-the-air broadcast path.

Finally, we must prepare our cable systems, with respect to the quality of our channels and to the channel capacity of our systems, in order to be ready for the the era of Advanced Television.

If we are successful, we will be able to carry high quality ATV services with a minimum of disruption and difficulty, at minimum cost. On the more positive side, a redesign of the basic television signal carried on our systems may well improve the way cable systems operate, and will likely allow some new revenue opportunities.

The ATV Alphabet Soup

A review of some of the common acronyms in the advanced television field would be useful at this point:

ATV, of course, is a general term applied to almost any form of consumer television that is a step beyond NTSC (or PAL or SECAM, for that matter). There are a range of advanced television technologies: IDTV, EDTV and HDTV are the primary terms being used. The definitions I am about to review are in line with those formally agreed upon by the Advanced Television Systems Committee, and are therefore becoming standardized.

IDTV, or Improved Definition Television, refers to improvements to NTSC television which remain within the general parameters of NTSC emission standards and, as such, would require little or no FCC action to implement. Improvements may be made at the source and/or at the television receiver and may include improvements in encoding, filtering, ghost cancellation, and other parameters as long as what is transmitted and received qualifies technically as standard NTSC in a 4:3 aspect ratio. Faroudja's SuperNTSC™ claims to fall into this category.

At the moment, though, improved definition television has not been implemented by changes in processing at the transmission end. Practically then, IDTV today means a receiver that is specially equipped to process a normal television signal. These receivers generally reduce NTSC color artifacts via three dimensional comb filtering for separation of luminance and chrominance; they generally provide some video noise reduction; and, most obviously, they all double the number of scan lines appearing on the display tube.

Line doubling generally involves displaying 525 lines, progressively scanned; that is, 525 lines, every 1/60th of a second, as opposed to NTSC's 262.5 lines per 1/60th of a second. This reduces the visibility of the scan-line structure of the NTSC signal, and improves apparent vertical resolution, although there is actually no additional picture information being presented. (In the case of the Faroudja SuperNTSC™ system, line doubling creates a 1,050-line interlace-scan picture.)

Some estimates say that to date, only about 5,000 of the current-design IDTV receivers have been sold in the United States. They tend to have list prices in the $1,800 to $3,000 range for direct-view CRT models and come with their own set of picture artifacts along with their improvements. They are available from Philips, Sony, NEC, Panasonic, Hitachi and a few other major manufacturers. These IDTV receivers represent a transitional technology between today's NTSC and the true improved, extended or high definition television systems that we'll see in a few years.

EDTV, or Extended Definition Television, refers to a number of different improvements that modify NTSC emissions, but that retain NTSC receiver-compatibility.

In other words, additional picture information is inserted into the 6-MHz channel which will not disturb an existing NTSC receiver, but which can be decoded by an EDTV receiver and used to display an improved picture. These changes may include a wide screen image (16:9 aspect ratio vs. NTSC's 4:3 image), and some modest extension of resolution, to a level somewhat less than twice the horizontal and vertical resolution of standard NTSC. Digital sound is generally a feature of EDTV systems.

The David Sarnoff Research Center is preparing an EDTV system called ACTV, or Advanced Compatible Television. This system will be the first to be tested under the auspices of the FCC Advisory Committee on Advanced Television Service.

Finally, HDTV, or High Definition Television, refers to television systems with approximately twice the horizontal and vertical emitted resolution of standard NTSC and improved color rendition. HDTV systems are wide-aspect ratio systems, and they generally attempt to approximate the picture quality of projected 35mm motion picture film. They are designed to be watched on large screens, where the NTSC system has historically failed to please. Digital sound is being planned by all proponents of HDTV systems.

"Compatibility" and the Role of "Simulcasting"

By FCC ruling, HDTV systems must not obsolete existing NTSC receivers, and can avoid doing so in one of two ways: they can be truly receiver-compatible, although this appears to be an impossible constraint while fully delivering NTSC horizontal and vertical resolution twice. Alternately, they can simulcast, on a completely separate RF channel, a newly-designed, non-compatible HDTV signal.

It is this simulcast architecture that has been selected for standardization by the FCC, because it promises to deliver the highest quality of service to American homes. Simulcast transmission systems are being worked on by five separate transmission system proponents, as mentioned earlier: NHK, Zenith, General Instrument, MIT and Philips. The Commission has indicated it will not consider setting a standard for EDTV until it has selected a simulcast HDTV standard, if at all.

Given the FCC's preference for the simulcast approach, let's review it in some detail. Under this approach, HDTV program services would have to be broadcast simultaneously with their sister NTSC services, so that the existing NTSC receivers are not made useless. The Commission has ruled that the spectrum for these new channels will be found within the existing VHF and UHF bands. The obvious implication for broadcasters is that each television station may be allocated an additional channel for its simulcast HDTV service, and indeed there seems to be a possibility to attain something close to this. Thus, cable operators may someday see the need to literally double their capacity for broadcast channels, and perhaps for many cable program services as well, if they decide to offer service in HDTV.

In the long term, as the simulcast system takes over, the NTSC channels can be discontinued, and the spectrum, or cable capacity, recovered for other uses. But there will certainly be a period of many years where cable systems will need the channel capacity to carry all the desired NTSC services and the new ATV simulcast services as well.

Interference Characteristics of ATV Signals A Critical Issue

Since the FCC's interests and power lie in regulating broadcasters and in allocating the over-the-air spectrum, it is the broadcasters' concerns that are dominating the development of ATV systems.

Fundamental to the success of any simulcast HDTV system will be its ability to successfully transmit its service without interfering with existing NTSC stations – either on the same channel in a nearby market, or on an adjacent or near-adjacent channel in the UHF band (the so-called "UHF taboos").

The approach being taken in some of the simulcast systems to solve these problems involves spreading the channels' transmission energy uniformly across the 6-MHz channel, and limiting peaks of power at particular frequencies or at particular points in time. If these techniques prove successful in lab and field testing, spectrum can probably be found for the broadcasters, and introducing ATV signals on cable will be made very much less technically demanding.

Business Issues in Advanced Television

Beyond these transmission parameters, the business realities of the cable industry will impose additional requirements on ATV systems: We will need a seamless interface to consumer equipment to avoid the difficulties we now put our customers through and to avoid cable operators' investment in set-top converter/descrambling devices. There is time to plan this properly, if the cable industry makes its voice heard.

The subscription pay business and the pay-per-view prospects of advanced television are staggering, but only if we make sure the adopted system has conditional access capability built in from the start. This conditional access, by the way, must be under the control of the individual cable operator who markets to his customer base and not merely under the control of the national programming provider.

Ghost cancelling is highly likely to be a standard feature of any ATV system, since multipath problems and microreflections, in the case of cable, are possibly ruinous to high resolution pictures. Whatever the ghost canceling scheme, it needs to work as well over cable systems as it does over the air.

Cable, of course, has unique requirements for data transmission to its subscribers' homes, and the ATV systems need to take our present and future needs into account.

Finally, ideally, the broadcast ATV system should be structured to allow cable operators to augment the normal 6-MHz channel with some additional spectrum to provide a higher performance level, primarily, better picture resolution, and better motion rendition. We may need this capability to compete with home video media and DBS transmission of very high quality HDTV.

Beyond these basic capabilities, any ATV system will need to meet cable's economics for production; its satellite interconnection practices;

its headend design considerations; and ideally be able to survive its customary distribution practices, including FM supertrunking, AML systems, fiber links, and long cascades of trunk amplifiers, as well as the usual complement of feeder actives and passives.

Cable Industry Influence is Crucial

How to make sure the ATV system adopted meets these requirements? The answer is twofold: technical testing by CableLabs and participation by the cable industry in the deliberations of the FCC Advisory Committee on Advanced Television Service.

First, as I've mentioned, on behalf of our member companies, CableLabs will be conducting a series of tests on each of the proposed ATV transmission systems.

CableLabs' ATV Testing Program

Under a $2.5 million contract with the broadcasters' Advanced Television Test Center (ATTC) in Alexandria, Virginia, we will conduct cable transmission simulations with a test bed of our own design, operated by CableLabs' personnel, and according to test procedures developed by CableLabs and approved by the FCC Advisory Committee on Advanced Television Service.

An Overview of the ATV Testing Program

But to put the CableLabs tests in context, let me back up for a few moments to review the entire testing process.

Three Test Types - Three Test Sites

Overall, the ATV testing is a North American effort that will take place in a total of three venues:

- at the American broadcasters' laboratory, the ATTC, where basic picture quality will be tested along with simulated over-the-air impairments and interferences,

- at the CableLabs facility, where each ATV system will be tested under various conditions of cable transmission impairment,

- and at the facilities of the Advanced Television Evaluation Laboratory, in Ottawa, Canada. This subjective test operation will be operated by the Communications Research Center of the Canadian Department of Communications. A consortium of Canadian interests is sponsoring CRC's conduct of the subjective assessments of quality and transmission performance.

Basic Quality Tests

The ATTC will perform the basic picture quality tests in which both broadcasters and cable operators have a vital interest. These are evaluations of the picture quality that each ATV system is ultimately capable of, assuming an ideal transmission path. Test signals and pictures will be encoded by the ATV system's encoder, which in all cases involves some bandwidth compression processing, then modulated to RF, and finally demodulated for display on a high definition video monitor.

Quality attributes to be evaluated include luminance and color resolution, color rendition and motion rendition. Also to be evaluated will be performance with filmed program material and with electronic graphics material, each of which presents its own technical characteristics and potential problems.

The proponent systems' basic picture quality will vary depending upon the wisdom of the designers' choices of basic scanning format and bandwidth reduction and restoration techniques.

Impairment and Interference Tests

Certainly as important as the systems' basic picture quality is their performance in the face of transmission impairments and interferences.

Impairment/Interference Tests - Broadcast

As I've indicated, benign interference performance, in particular, will be of pivotal importance to selection of an ATV system that is practical in today's crowded over-the-air spectrum.

Broadcasters will test transmission performance in the face of various impairments, including random noise, impulse noise, multipath effects, and airplane flutter.

Broadcast interferences to be tested include co-channel and adjacent-channel interference, UHF taboo channel interference, and discrete frequency interference.

Impairment Tests - Cable

CableLabs will conduct tests of at least eight typical transmission impairments common to cable television distribution systems. Testing of the first system is currently planned to begin in April 1991. Testing of each system is expected to take a total of eight to 10 weeks in the broadcast and cable laboratories, and an additional period of perhaps six weeks for subjective testing in the Canadian laboratory, which will run concurrent to the lab testing of the next ATV system. Lab testing of the six systems should be completed by the end of April 1992.

Field testing will follow in the second and third quarters of 1992 according to procedures currently being developed by the FCC Advisory Committee. The Committee's final report and standards recommendation to the FCC is due by the close of the third quarter of 1992. The Commission plans to make its standards decision on a simulcast system during the second quarter of 1993.

Although HDTV may grow slowly in the marketplace, the next two and one-half years will involve intensive planning and decision making. The cable industry's considerable engineering and management talent needs to participate in the planning and deliberations of the FCC Advisory Committee on Advanced Television Service if we are to have an ATV system that meets cable's needs. I urge each of your companies to send representatives to these meetings and make your views known.

Digital Transmission and NTSC Compression

Just how cable might implement the digital transmission of compressed NTSC signals on its existing and future plants is perhaps a hotter topic than HDTV. CableLabs is studying these issues from both business and technical perspectives.

There are some enormous strategic implications in this topic:

- Digital transmission of NTSC offers the promise of uniformly high quality video and audio in each subscriber's home, regardless of the distance from the headend or the age or condition of the cable plant.

- Compression of NTSC signals (which in any case is needed to fit a single digital NTSC signal into today's 6-MHz cable channel) offers the promise of transmitting more than one NTSC program in each channel — perhaps two, three, or four programs simultaneously on a full-time basis.

Video compression's cost and complexity need to be traded off against the raw upgrading of a cable system's overall bandwidth. Bringing fiber optic cabling close to subscribers' homes, coupled with new wider-bandwidth RF amplifiers, can achieve increases in channel capacity in very cost effective ways. To be successful, video compression of NTSC will need to be low in cost.

As CableLabs investigates NTSC compression technolgies, we believe that any system ultimately deployed must provide a quality level that is at least as good, subjectively, as the best current analog NTSC transmission via cable. There is no reason to expect that viewers would, or should, tolerate a decline in delivered quality. Ideally, the usage of an NTSC compression system in a cable system should be both technically and functionally transparent. The goal is to provide new programming, and to provide it in a more flexible manner – not to "thrill" subscribers with new technology.

CableLabs' Technical Advisory Committee has formed a Subcommittee on Video Compression to study the issue of NTSC compression over the satellites that feed cable systems and over the cable plants themselves. This committee is chaired by Ed Horowitz, Viacom's Senior Vice President of Technology. The committee meets regularly with the staff of the Advanced Television Projects Department at CableLabs and with vendors proposing compression and transmission technology for use by the cable industry.

Whatever NTSC compression system is identified for use on cable, it is clear that the resulting signal will be a digital signal. Transmission of high-rate digital signals with low error rates over cable systems is a new field which requires research. CableLabs will conduct a series of digital transmission tests over laboratory simulators and real-world cable plants to determine exactly what modulation schemes will support what bit rates. This early research, much of which will be completed by CableLabs ATV staff this year, will give us a fundamental reading of cable systems' digital transmission capabilities. Pursuit of a particular digital transmission protocol and the compression system it carries will flow from this basic understanding.

In summary, we believe that the cable industry's twin interests in the future of television are in good hands. CableLabs is a key player in testing of the proposed HDTV and EDTV systems, and intends to lead the development of a digital transmission architecture for transmission of HDTV and compressed NTSC services as well.

ALL YOU EVER WANTED TO KNOW BUT WERE AFRAID TO ASK ABOUT MEASUREMENT ANOMALIES IN BROADBAND AM-VSB SYSTEMS

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ABSTRACT

Currently, increasing numbers of cable operators are adopting fiber optic distribution systems using AM-VSB modulated signals. While this technology has been shown to be cost effective in various hybrid fiber/coax architectures, the sizable deployment now taking place represents significant capital expenditures. As a result, operators are embarking on perhaps the most intensive product evaluations and cost/performance trade-off reviews in the industry's history. However, there are several sources of measurement anomalies that must be taken into account to obtain meaningful and reproducible results as these systems are moved from the lab environment to the field.

This paper will focus on several sources of these anomalies such as spectrum analyzer frequency response, external bandpass filter tuning, software design utilized in automated testing, impedance matching of the system under test, etc. Also, several accuracy improvement techniques will be considered.

INTRODUCTION

The cable television technical community is well-versed in the art of making noise and distortion measurements in broadband AM-VSB modulated systems. It has available to it such guidelines as the "NCTA Recommended Practices for Measurements on Cable Television Systems" to assist it in determining the proper test set-up, the correct test equipment properly applied, and to assist in the interpretation of test results with some suggested numerical limits for various measurements.

Competition in the home video entertainment market from various sources served to heighten the cable industry's awareness of picture quality and reliability issues. As a result, considerable attention is now being given to lightwave transmission as the technology most likely to successfully address these concerns. These events have caused the cable industry to set exacting system performance goals when implementing lightwave technology. This paper will discuss why there is a window of measurement ambiguity around the true data points that must be considered when analyzing test results from these systems.

Critical system performance measurements are generally made using a swept-tuned spectrum analyzer, either by itself or in conjunction with other test instrumentation. Some measurements, such as subcarrier ratios, can be adequately measured using a spectrum analyzer by itself. However, other performance measurements, such as carrier-to-noise and carrier-to-distortions, require more dynamic range than is possible to achieve from a spectrum analyzer alone. Bandpass filters followed by low noise preamplifiers are typically used to "enhance" the incoming signals. Although there are alternate techniques, this method is certainly the most common and is the one addressed here.

IMPACT OF SPECTRUM ANALYZER FREQUENCY RESPONSE

As with all broadband devices, spectrum analyzers have frequency response characteristics that are anything but flat. True, modern spectrum analyzers are very good, especially when sweeping small percentages of their total available bandwidth. A typical lab grade instrument might be as good as +/- 0.25 dB within any 10 MHz span while a field grade unit might be closer to +/- 0.5 dB.

Discounting the effects of all else, measuring adjacent carriers of the same power level can have this much uncertainty.

There is another less obvious problem caused by the response uncertainty of a spectrum analyzer. This occurs when we use an analyzer to "flatten" the output of a multiple frequency signal generator. This can be a good way to assess the frequency response of a device or system under test, but it may not be the best method when making critical distortion and noise performance measurements. Although the narrow-band frequency response errors of the analyzer may be small, its broadband errors can become quite large (as much as +/-1.0 dB). With this in mind, we can see that what would appear to be a flat output from the generator would actually vary across the band by the amount of the analyzer's response error. As these "analyzer-leveled" signals pass through a non-linear system under test, the harmonics generated will likewise vary accordingly. Also, these resultant distortions are not measured relative to their fundamental carriers, but are measured relative to the carrier in whose passband they fall. So, if the carrier of the channel under test was set at the -1 dB point along the analyzer's response curve while the fundamentals that created the distortions were set at the +1 dB point, then the ratio of the discrete second order product would be skewed by as much as 2 dB.

One way to minimize the effects of broadband response errors is to use a power meter to level the generator rather than a spectrum analyzer. Although this won't eliminate these errors entirely, if used carefully it can reduce these errors from several dB to perhaps a few tenths of a dB. Unfortunately, this can prove to be somewhat time consuming and we may be tempted to believe that since we're using the same analyzer "it will all come out in the wash". The dynamic ranges of active devices are not always constant across their bandwidth, and is one reason CATV active devices are measured at several places across their passband. To allow the test set to introduce this much potential error would risk compromising the tests.

EFFECTS OF AMPLITUDE MEASUREMENT INACCURACY

There can be no question that as spectrum analyzers continue to mature they become more accurate. One such improvement can be seen in their ability to measure relative power fairly accurately. However, both relative and absolute power measurements present a couple of subtle issues that, unless recognized, can cause us some problems.

Relative power accuracy is affected by a wide variety of things, but most notably analyzer flatness. As we would expect, this also has a big effect on absolute power accuracy as well. Often overlooked is the amount of calibrator error and its impact on absolute power measurements. If we think about this we can see that if an analyzer were perfect except for the calibrator error, any absolute power measurements would only be off by the amount of the calibrator error. Likewise, if the calibrator were perfect, we would only be off by the flatness error at the measured frequency. Since neither of these is in reality perfect, the amount of the error is the sum of the calibrator error and amount of the response error at the measured frequency. Depending on the frequency being measured, this can either reduce or increase absolute power error. In any case, calibrator error can increase measurement uncertainty by the amount of its error.

Fortunately, absolute power errors generally do not cause a problem since the vast majority of measurements in the CATV industry are relative. However, they can be very significant when measurements must be correlated with other instruments, or with the same instruments but at different times.

EFFECTS OF EXTERNAL BANDPASS FILTERS

External bandpass filters are used to reduce the risk of the test system adding distortions of its own into the measurements. By eliminating all but the band of interest from ever passing through active devices in the test set, the exposure to this potential error is limited.

One important characteristic of bandpass filters that we should be concerned with is their passband flatness. Depending on the type of filter used, the peak-to-valley response can be 0.5 dB, or more. Tunable filters can be much worse, especially if they are tuned at either end of their range. If, when measuring a noise or distortion ratio, the amplitude of the carrier and the amplitude of the associated noise and/or distortion is measured at the same frequency then the resultant ratio will be free from filter induced error. Furthermore, if these two data points are measured at different frequencies, then filter flatness will indeed disturb the measurement's integrity. This problem is worse with tunable filters because they tend to peak at the center of the passband and roll off on each side (see figure 1 below). This results in having the side bands suppressed by typically 0.25 dB or so. Although much less convenient, fixed filters can be characterized and their response irregularities minimized. Tunable filters are usually a single octave wide, and their passband is some percentage of the center tuning. So their useable passband may be too narrow (less than 3MHz wide) at the low end of their tuning range, while being too wide (greater than 5MHz) at the high end of their range.

IMPEDANCE RELATED PROBLEMS

As frustrating as it can be, we in the CATV industry must endure a 75 ohm life in a 50 ohm world. We are forced to measure a broad spectrum in very narrow portions that introduces additional sources of errors due to interconnecting cables, multiple impedance mismatches and transformations, etc. All these factors introduce response uncertainties of their own making the job of calibrating them out virtually impossible.

Thankfully, many analyzers are now available with 75 ohm inputs and with the amplitude scale calibrated in dBmV. Other instruments come with one 50 ohm input and one 75 but typically have only a 50 ohm calibrator. Additionally, digital analyzers generally have the capability of displaying amplitude data in a multitude of scales, including dBm and dBmV, but a word of caution is in order. It is not uncommon for these scales to be referenced to 50 ohms, which can offset the dBmV scale by 10*log(75/50), or 1.76 dB. This can be particularly tricky when using one of those analyzers that has both inputs and you switch

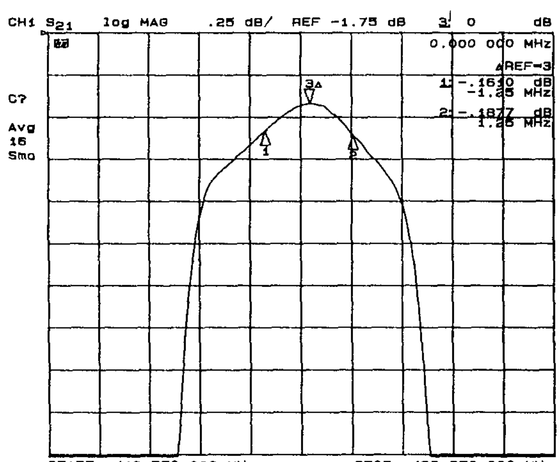

Figure 1
Typical Tunable Bandpass Filter Response Characteristics

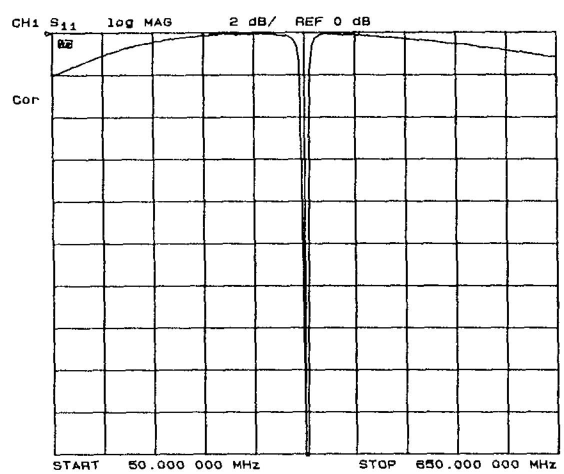

Figure 2
Typical Tunable Bandpass Filter Return Loss Response

between the two. A simple check for this is to set the reference level to 0 dBm, then switch the scale to read out in dBmV. If it says 48.75, you're all set; if it says 47, you'll know to adjust for it.

Bandpass filters are pretty close to open circuits out of band. In fact, 1 to 2 dB of return loss can be considered optimistic (see figure 2 above). While some devices under test may not be affected too badly by this, others may not be unconditionally stable. In these cases, poor terminations can cause unexpected results. Simply using a 10 dB pad as the terminating load for our device under test can go a long way in preventing this from becoming too much of a problem.

AUTOMATED MEASUREMENTS

In recent years highly programmable instruments have become commonplace. Test setups can be stored quickly in memory, trace data routed to memory cards, and even entire measurement sequences can be recalled and executed at the touch of a button. This is an extremely effective way to perform difficult measurement sequences with precision. By reducing the opportunities for the operator to make procedural or computational errors, we achieve a level of consistency and repeatability unheard of just a few years ago. In fact great strides have been made in standardizing automated testing methods based on the recommended practices for manual testing, further enhancing repeatability and correlation.

Although automated measurements can and should be used for all these reasons, there are a few things that should be kept in mind. First, digitally based instruments are well known for their high degree of numerical resolution. Unfortunately, this is often confused with accuracy. Simply because an instrument reads out in hundredth's of a dB does not mean that it's accurate to within a hundredth of a dB.

Another area of caution with automated measurements concerns dynamic range. A spectrum analyzer's dynamic range can be thought of as the amount of the CRT display for which the instrument's specifications hold true. Unfortunately, this is generally less than the total amount of display range we have available. Usually, dynamic range runs from 50-70 dB, while display range is normally 80-100 dB. This means that if we measure ratios directly from the CRT display that are greater than the dynamic range, our measurements can have fairly significant errors. Also, many instruments are prone to log scale fidelity errors and these can add up quickly. This can account for as much as 1.0 dB of error even when measuring inside the instrument's dynamic range window. Several instruments available today no longer suffer from significant log amp anomalies but these tend to have less dynamic range. These problems can be minimized by ensuring all signals and noise measurements are made within the first two graticule divisions.

Special care should be excercised when adjusting tunable filters. When set toward the low end of their ranges, they tend to get quite narrow. It may not be immediately apparent but the edges of the measurement band may become overly attenuated by the filter's skirts. Certain software routines for measuring carrier-to-noise ratios will search the entire display for the minimum amount of noise, and in these cases can be off by several dB. This issue is addressed in some software designs by always measuring noise at a fixed location near the carrier, while still others limit the allowable frequency excursion when searching for minimum noise.

Automating measurements offers many advantages to manual systems. Enhanced repeatability, as well as freedom from such things as operator and computational errors, etc., are compelling reasons to use this technology. Although digitally based instruments can sometimes software-correct repeatable errors that occur in hardware, they cannot improve a test system's base accuracy. In other words, if an instrument has an absolute power measurement uncertainty of 1 dB, all measurements made by that system will have at least that much uncertainty, whether a human is physically pushing buttons or a computer is doing it electronically.

Low Noise Preamplifiers

Finally, a note when using a low noise preamplifier for signal enhancement. Though it is sometimes overlooked, preamplifiers do add noise to the measurement system. Although this isn't always a problem, it can cause misleading results when a high degree of test system sensitivity is required. The exact amount of noise added to the system depends on both the noise figure of the analyzer and the noise figure of the preamplifier. Figure 3 shows the amount of noise added to a lab grade analyzer with 0 dB of input attenuation. The noise figure of this particular amplifier is approximately 4 dB. This is especially true when trying to measure very small amounts of noise and distortion. Therefore, whenever determining sensitivity, always terminate the input to the test system (typically the input to the bandpass filter) rather than the input of the spectrum analyzer.

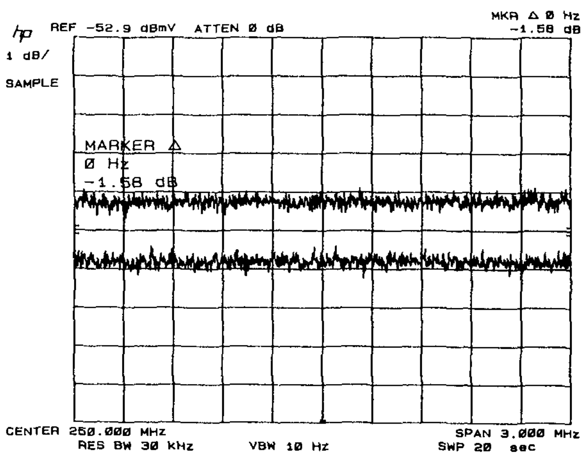

Figure 3
Test System Sensitivity with and without Low Noise Preamplifier

CONCLUSIONS

Our observations were based on many measurements made using a wide variety of makes and models of test equipment. The cautions suggested were not meant to discourage anyone from making measurements, but simply to point out that all measurements include uncertainties inherent in the process. By understanding some of the sources of these uncertainties, anyone making measurements on broadband AM-VSB systems can minimize their effects.

When making critical distortion and noise measurements, careful use of a broadband power meter can minimize the uncertainties caused by unleveled multi-frequency signal sources. Where applicable, the use of fixed-tuned bandpass filters with known characteristics can reduce the effects of the filter's passband ripple.

While automated techniques aid in making repeatable measurements, they do suffer from the limitations discussed above. Therefore, the test system operator must carefully monitor the process to ensure that the data is taken without exceeding the limitations of the test instruments.

This paper was not intended to be an all-inclusive treatment of measurement inaccuracies but to sensitize the industry to their existence. As test instruments and test methods improve, these anomalies will become smaller. As a result of ongoing efforts in our labs, we intend to present for future consideration a few specific test practices that may help narrow this window of uncertainty. For the time being, however, it appears that even with the best efforts applied, the window of uncertainty in these types of measurements is roughly 2 dB, and is independent of the test methodology used. Unfortunately, we are all subject to this degree of uncertainty so it is important that we maintain a proper perspective toward our test data.

AM OPTICAL BRIDGER NETWORKS FOR CATV

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Abstract

In its application of AM fiberoptic technology to supertrunks and to segmentations of trunk cascades, the cable TV industry has gained an appreciation for the performance and reliability features of this technology. It has become increasingly clear, however, that CATV distribution architectures other than the traditional tree and branch might be needed in order to fully exploit the advantages of fiber technology. This paper describes and discusses one of the most promising of these new architectures -- optical bridgers -- which permit widespread deployment of fiber in new-builds and substantial rebuilds without incurring a cost penalty. In an optical bridger network the CATV trunk is completely eliminated and each fiber receiver feeds a multiple-output RF bridger directly. With no trunk cascade, wide-area distribution can be made through four or five high-level amplifiers without compromising end-of-line performance. Using specific examples, this paper will review the properties of optical bridger architectures and will discuss hardware implications, such as selection of hybrid amplifier type.

BACKGROUND

Following the lead of a few key individuals, the CATV industry has embraced fiberoptic technology and has caused a remarkably rapid development of the components needed for AM transmission of video over glass fiber. Thus, at the beginning of 1990 there were only a small number of AM fiberoptic nodes in place within our industry, but by the end of the year the number in service was between 500 and a thousand. To a CATV industry accustomed to coax systems having RF amplifiers approximately every 3000 feet, the attractiveness of a transmission medium that offered ten to twenty mile unrepeatered spans was obvious. Prior to the development of the AM technology, however, those spans were achievable only by FM techniques, whose conversion costs were so high that widespread usage was unthinkable.

In the initial AM fiber transmission installations, of course, the equipment was also expensive, as is the case for most new technologies. Hence those installations tended to have specific attributes that made them more tolerant of high equipment costs. Typical of these first applications were supertrunks to combine headends in adjacent franchises or to break up conventional trunk cascades that had grown overly long and unreliable due to extensions. In addition a number of installations were pursued in part to test the emerging AM fiber technology and to develop experience, which in turn tacitly assumed that as time went on the AM fiber equipment would become more and more cost-effective.

In actuality, with increasing volume and manufacturing experience, the cost of an AM fiberoptic link has decreased dramatically during the past year. The price of an AM laser transmitter, which dominates the link equipment, has decreased from around $25,000 to the vicinity of $15,000. At the same time the distortion performance and output power of these transmitters has improved significantly, thus making high quality links more and more commonplace. Notwithstanding these strides, however, the costs of an AM optical link remain greatly out-of-line for an industry that generally measures equipment unit costs in tens and hundreds of dollars. Hence widespread deployment of AM fiber within the industry's traditional tree and branch network architecture was questionable.

Approximately one year ago, however, the industry began to give serious consideration to non-traditional architectures for CATV. In particular, multi-level star architectures were proposed by ATC at last year's SCTE fiberoptics conference in Monterey[1]. Soon afterwards our company announced its Flamethrower optical bridger and all at once it appeared that everyone was talking Fiber-to-the-Feeder, Fiber-to-the-Bridger or some other variant on the same theme. In any of these systems the conventional trunk amplifier cascade is eliminated and is replaced by multiple optical fiber links that feed coaxial cable stars, which provide extensive distribution to subscribers. At the heart of the excitement generated by these systems is the fact that the multi-level RF star distribution spreads the laser transmitter cost over a large number of households. Thus the per-subscriber cost of the optical link becomes much less of a barrier to deployment and the advantages of AM fiber performance and reliability become more generally accessable.

This paper will describe several optical bridger networks and will indicate the reasons for their cost-effectiveness. In addition, options for the choice of amplifier hybrids to be used in this equipment will be discussed.

OPTICAL BRIDGERS

Four-amplifier

As an introduction to the subject, we will consider a 4-amplifier optical bridger network, as shown schematically in Figure 1. One notes immediately the multiple-star arrangement: numerous optical fiber runs radiate out from the headend to various neighborhoods (A, B, C, etc), terminating in optical bridger node stations. Each optical bridger station consists of an optical receiver and a multiple-output rf bridger amplifier, which is a hub of one of the secondary stars. Multiple strings of up to three line extender amplifiers radiate out from each bridger.

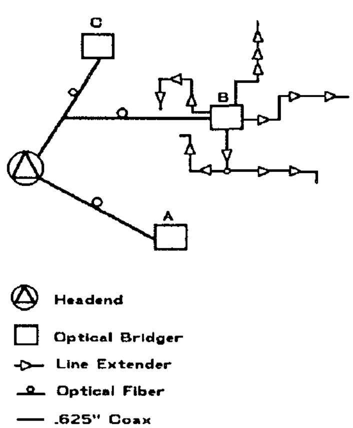


Figure 1. Four-amplifier Optical Bridger Distribution Network

In the 4-amplifier network no subscriber is more than four active devices away from the headend (one bridger and three line extenders). If we look in detail at one complete path between headend and end-of-line, as diagrammed in Figure 2, we can evaluate the performance of this network (Table 1). This performance assumes power doubler hybrid technology and does not utilize roll-back from the line extenders (using more than 150' of cable in the last tap span before a line extender). Since the optical link can provide 52 dB CNR over a 10 dB optical loss budget, the optical bridger node can be as much as 16 miles from the headend, which means that it can deliver end-of-line performance far in excess of that which could be achieved with a conventional trunk and feeder architecture -- even with feedforward trunk amplifiers (Table 2). Since the signal distortions resulting from the optical link are better than from a typical long cascade, one is able to feed three line extenders rather than the conventional two, thereby increasing the radial reach out from the optical node to approximately 3150 feet. In comparing Tables 1 and 2 one should note how the noise and distortion build-up is more or less equivalent from each segment of the optical bridger path, whereas in the long cascade tree-and-branch the trunk predominates in both signal degradations.

In this example one can see how the multiple-star architecture with optical fiber provides high signal quality and generous reach.

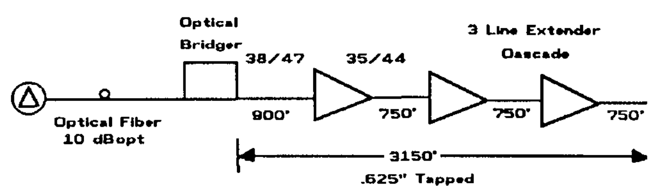


Figure 2. Four- amplifier String

Five-amplifier

In order to gain an added appreciation for the potential of the optical bridger type of architecture we can examine what happens when an additional star layer is inserted in the form of express feeder legs between each optical receiver and multiple-output bridger (Figure 3).

Table 1. Four-amplifier Optical Bridger Performance (550 MHz Power Doubled Amplifiers)

	CNR @ 50 MHz	CNR @ 550 MHz	CTB	XMOD
Fiber Receiver (10 dB loss budget, 40 channels/fiber	52	52	-65	-65
Optical Bridger	63	67	-64	-67
Line Extenders (3 Cascade)	56	56	-61	-64
End-of-line	50	50	-54	-56

Table 2. Trunk and Feeder (550 MHz Feedforward Trunk and Power Doubled Distribution)

	CNR	CTB	XMOD
Trunk Amplifiers (33 Cascade)	44	-58	-60
Bridger	65	-65	-66
Line Extenders (2 Cascade)	57	-67	-70
End-of-line	44	-53	-55

Table 3. Five-amplifier Optical Bridger Performance (550 MHz Power Doubled Amplifiers)

	CNR @ 50 MHz	CNR @ 550 MHz	CTB	XMOD
Fiber Receiver	52	52	-65	-65
Optical Bridger	64	62	-70	-71
High Output Bridger	60	62	-65	-68
Line Extenders (3 Cascade)	56	56	-61	-64
End-of-line	50	50	-53	-55

This is accomplished by operating the optical bridger node at somewhat reduced output levels, and feeding a second multiple-output bridger with nearly 2700 feet of untapped .875" coaxial cable (the "express feeder"). By comparing the performance of this 5-amplifier system (Table 3) with the 4-amplifier optical bridger (Table 1), one can see that the half-mile increase in reach has been achieved with nearly imperceptable degradation to end-of-line performance.

Since that reach extends in a number of different directions (limited, of course, by the local distribution of subscribers) one can see that each optical receiver node has access to all homes within a radius of approximately 1.1 miles. Alternatively one can view this system as covering areas of almost four square miles that are located as much as 16 miles away from the headend.

ECONOMICS

As has already been stated, one of the key aims of the optical bridger architecture is to minimize the cost impacts of the relatively expensive optical link by maximizing the number of end-users served by each node. Optical bridgers can do this well when the multiple output capabilities can be used to fan-out effectively. Thus a realistic economic evaluation of this architecture must be system-specific and can be done only with actual system maps.

In one sample system design performed recently for a 90 sub/mile operator, nearly 25 miles of plant (2200 subscribers) could be served from a single optical bridger node, with 50 dB CNR, 53 dB CTB and 59 dB CSO at the tap. If a single fiber link were devoted to that node (i.e., there were no optical splits to other nodes) then the optical transmitter and receiver would cost less than $7 per sub. Note, as well, that the amount of large cable for express feeder (750" in this case) was less than 9% of the cable used, so there were no unusual costs buried in the design.

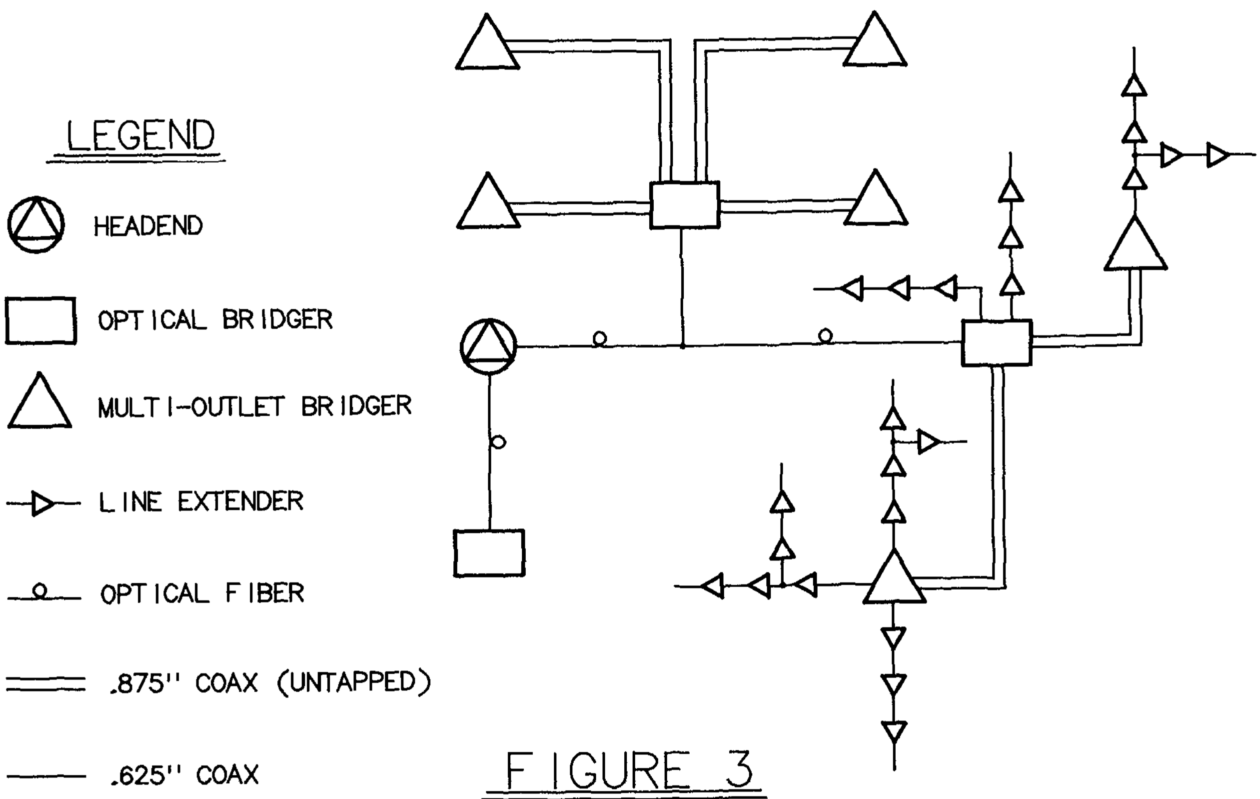


FIGURE 3
FIVE-AMPLIFIER OPTICAL BRIDGER DISTRIBUTION NETWORK

Clearly not all systems will be as fertile for optical bridger network designs as that one. Experience indicates that relatively dense operations (approximately 50 subs/mile or greater) offer the most ready opportunities for exploiting the multiple star achitectures. Furthermore one must recall that one of the early attractions of optical fiber was its ability to segment a CATV system for reasons of customer service and marketing. Thus there are extremes of fan-out that may be undesirable despite their attractively low equipment costs.

AMPLIFIER HYBRIDS

All of the optical bridger designs described so far have employed power double hybrid technology. In this final section we will discuss the relative advantages of using feedforward technology for the amplifiers in these multiple-star networks. Three single string designs are shown schematically in Figure 4, representing one maximal coax path in optical bridger networks employing different hybrid types. All of these stri ngs will deliver the same end-of-line performance: 50 CNR, 53 CTB and 54 XMOD. (The fiber link is omitted from the diagrams but not from the performance calculations.) In each case, the express feeder cable is .875" and all other cable is .750" . Tap separations are a uniform 150' and no roll-back is included.

The first diagram shows that a five-amplifier string using feedforward bridgers can attain a 6250' length (1.2 miles) with 25 tap locations. When power addition technology is used, as in Figure 4b, the maximum reach drops by 6.6% and has four fewer taps. Interestingly, Figure 4c shows that a four-amplifier string utilizing feedforward bridgers can actually reach nearly as far as the five-amplifier string in 4a -- by virtue of its very long express feeder leg -- but has only 18 tap locations.

In many actual system designs, however, single-string reach is likely to be a misleading measure of value. This is because the cost-effectiveness of the optical bridger architecture is based largely on the number of fan-outs that it makes possible. Because of the cost and the power dissipation of feedforward amplifier hybrids, it is not possible to install numbers of these modules within one CATV amplifier housing.

Thus multiple branches can be provided only by splitting the output of a single hybrid, which decreases the operating levels. Power addition hybrids, on the other hand, can be associated on a one-for-one basis for each output port, thus permitting multiple high-level outputs.

Furthermore, it has been shown[2] that distortion performance of feedforward hybrids degrades rapidly and unpredictably at levels approaching 50 dBmV (Figure 5). This is not surprising since those hybrid modules were designed for application to trunk amplifiers, which operate in the vicinity of 36 dBmV. The importance of this observation, however, is that the end-of-line distortion performance of the high level feedforward amplifiers will be very sensitive to operating level changes. This, in turn, may mandate the use of AGC circuits for the feedforward units. This is unfortunate because one of the features of the optical bridger architecture is the lack of a clear need for AGC, due to the short cable lengths involved. These considerations are summarized in Table 4.

Table 4. Comparison of Hybrid Amplifier Technology for Optical Bridgers

	Feed Forward	Power Double
Cost	≈$100 addition per hybrid	--------
Coverage	Longer reach in straight-line with single string	Larger span, due to multiple high level outputs
AGC	Needed	Optional

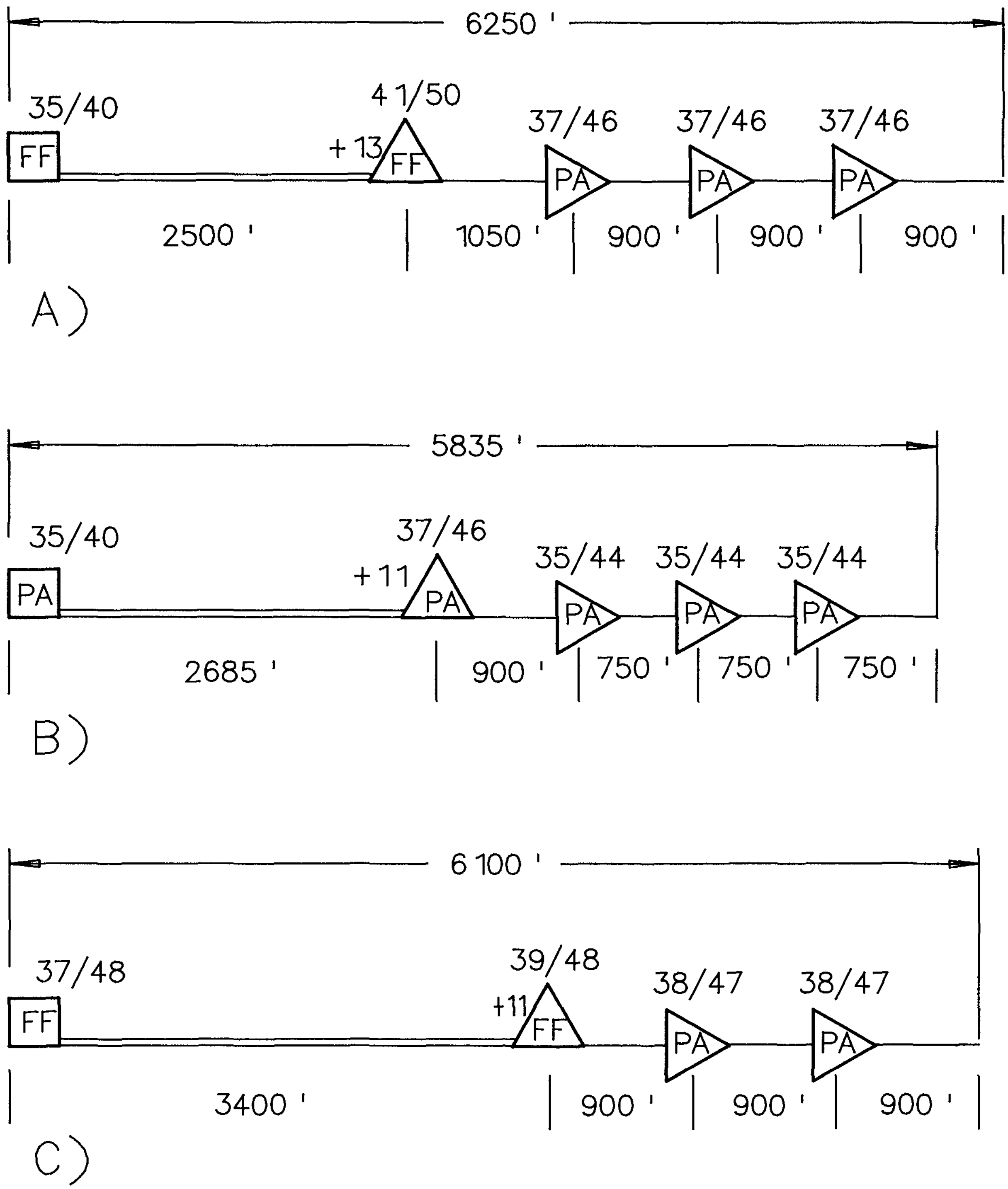
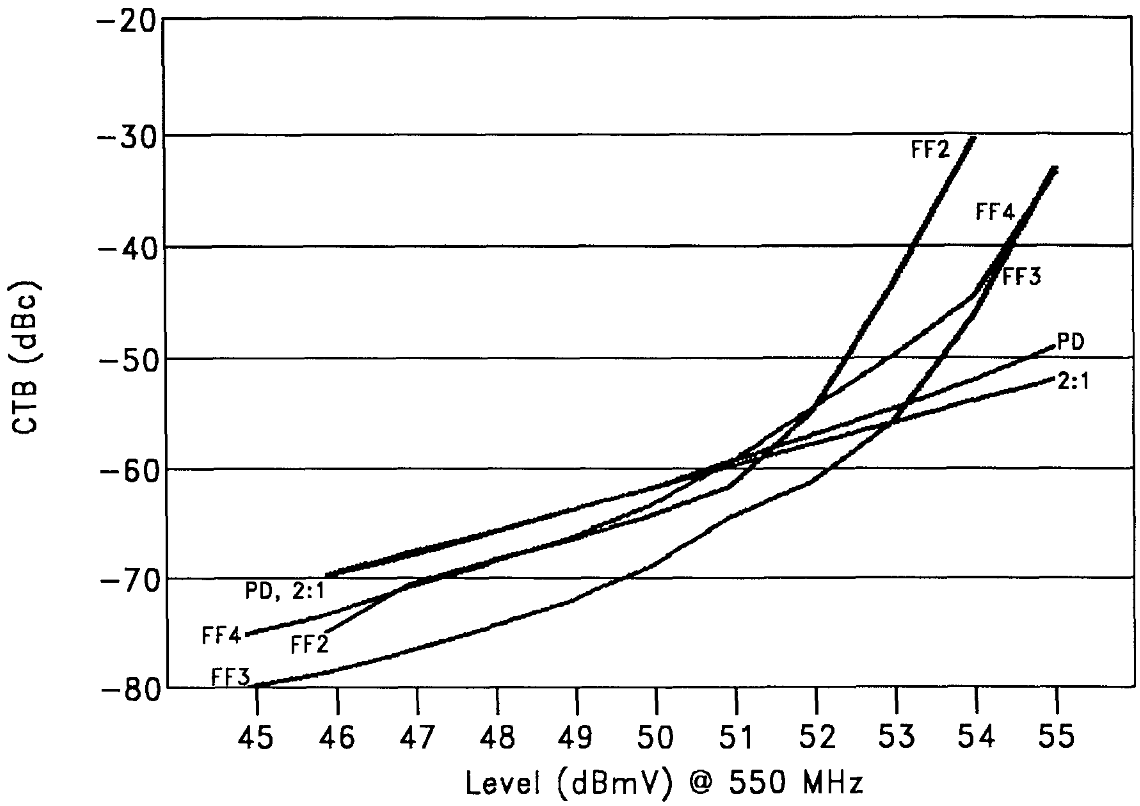


FIGURE 4

AMPLIFIER HYBRID SINGLE-STRING
MAXIMUM REACH

Measured Hybrid CTB
77 Channel, 8 dB Tilt, 547.25 MHz
Figure 5



CONCLUSIONS

The optical bridger type of distribution architecture provides a cost-effective means for deploying AM fiberoptic technology on a wide scale for cable TV. The multiple high-level output capabilities offered in the RF distribution equipment that is becoming available will be both a boon and a challenge to the system designer. This equipment provides the designer with new opportunities for efficient designs, but will also test his or her ingenuity and flexibility. The choices between specific amplifier units will depend on the details of the system maps, on cleverness of the designer and on the inherent costs of those units.

REFERENCES

1. J. Chiddix, "The Strategic and Business Value of Fiber," SCTE Fiber Optics 1990, Monterey, CA, March 1990.

2. D. Jubera and J. Hernandez, "CATV Hybrid Distortion in High Level Distribution Systems," Technical Report, Texscan Corporation, El Paso, TX, June 1990.

AMPLIFIERS FOR FUTURE NETWORKS

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Abstract

CATV distribution systems for the decade of the 90s will have new requirements placed on them to satisfy market demands for new services and higher reliability. These requirements will affect the architectures and layouts of distribution systems, and the type and performance of the RF amplifiers used in the RF portion of the distribution system.

Fiber's great bandwidth combined with shorter RF cascades opens up the possibility of using more bandwidth to provide more services to the consumer or different services to new consumers. This paper examines amplifier technologies that can be used to satisfy the requirements of new distribution systems. It also looks at migration paths for current systems so that they can satisfy future bandwidth requirements.

The performance of current amplifiers and the implications of their performance on the performance of extended bandwidth amplifiers is examined. Performance improvement opportunities within the RF amplifier will be identified. The resulting optimized amplifier performance will be analyzed.

I. INTRODUCTION

Future Networks for the delivery of Cable Television signals will almost certainly evolve from the fiber optic based architectures being implemented now. Gone will be the long cascades of amplifiers extending from the headend to the last subscriber. Most of the amplifiers will be replaced by fiber optics. Networks built now utilizing fiber optics to replace trunk amplifiers may use fiber even more extensively as they evolve into networks of the future.

Networks of the future will carry more information. The form that this additional information will take is not yet clear, nor is the final configuration of the RF distribution system; but almost certainly, amplifiers will still have a critical role.

The amplifiers that remain in these future networks must be able to operate effectively in these new architectures with greater channel loadings than systems are currently carrying. The changes in technology that will allow future networks to be practical will be available in the next five years.

II. FUTURE SYSTEM REQUIREMENTS

Many of the networks of the future will be based on the combined fiber and RF architectures, some of which are being built now. These combined fiber and RF architectures are often call Fiber to the Feeder (FTF). Both FTF systems and the future networks have the following common goals:

1. They must be cost effective so that they justify their capital investment.

2. They must deliver reliable service

3. They must deliver high quality pictures

In addition to the above, FTF systems can have an unique goal:

4. They must have a path to the future. The fiber and RF design and equipment of the FTF plant must lend itself to being reconfigured to meet the future network requirements without the need for a complete rebuild.

Future networks also have an unique goal:

5. They must deliver more information than systems currently carry.

The form this additional information will take will vary, but almost certainly it will require the provision of additional bandwidth in the distribution system. Bandwidths of up to 1 GHz have been proposed.

The exact mechanism of a 550 MHz FTF system upgrade to 1 GHz is not discussed in this paper, but because an upgrade appears to be the most cost effective means of ensuring a cost effective transition between 550 MHz and 1 GHz, some aspects of this paper are based on an upgrade scenario.

III. FUTURE AMPLIFIER REQUIREMENTS

The amplifiers used in future systems are essential for the achievement of these systems goals. The amplifiers used in these future networks must satisfy the following requirements:

1. They must have the highest output capability for the lowest cost to ensure cost effective systems.

2. They must operate reliably in these two areas:
   a. ability to operate without failure (MTBF)
   b. ability provide consistent performance (especially distortion) over a wide range of environmental and system conditions

3. They must amplify signals up to 1 GHz

4. They must perform at performance levels higher than those of current 550 MHz RF amplifiers

The future performance requirements of amplifiers are going to be influenced by:

1. Costs of fiber links

2. The performance of fiber links.

3. The maximum number of subscribers that can be served from one fiber receiver.

4. Cost of 1 GHz amplifiers

5. The type of signals to be carried.

The cost of 550 MHz fiber links may well decrease over the next five years but this decrease could well be offset by the increased cost of the fiber equipment required to carry the additional signals up to 1 GHz. Also performance improvements in 550 MHz fiber links over the next five years may be offset by the performance degradation caused by the increased channel loading. These two factors affect the depth of penetration into the network of fiber optics, and therefore the number of amplifiers in cascade.

The type of services being offered in future networks will also determine the number of amplifiers in cascade after the fiber receiver. To provide videostore like services, the number of potential subscribers fed from one fiber receiver may have to be limited. FTF systems today may feed up to 2,000 or more potential subscribers from 3 or 4 amplifier cascades. Future signal delivery requirements may limit this number to 500 subscribers and thus reduce the RF amplifier cascade.

Amplifier cost will go up as the bandwidth goes from 550 MHz to 1 GHz. Required features such as Automatic Level Control (ALC) along with other extended manufacturing bandwidth requirements will make 1GHz amplifiers cost more than 550 MHz amplifiers.

Two different signal type scenarios are being discussed for use in future networks:

1. Full loading of 50 MHz to 1 GHz with 6 MHz spaced NTSC channels (approx 151 channels).

2. Loading 50 MHz to 550 MHz with 77 6 MHz spaced NTSC channels and loading of 550 MHz to 1.0 GHz with digital or FM signals.

The net effect of all these factors on amplifiers is that RF amplifier cascades must be reduced. A one GHz RF amplifier will have to perform with better output capability than its 550 MHz counterparts so that the overall system cost and performance targets can be met.

To quantify this performance improvement an analysis of Composite Triple Beat (CTB) performance requirements follows:

<u>Amplifier Requirements for Full 1 GHz Channel Loading</u>

The 151 channel loading requirement represents an increase in the average amplified power level over that of a 77 channel loading. The effect of this increase on Composite Triple Beat (CTB) can be theoretically calculated using the following formula:

$$Loading\ CTB_{increase} = 20 \times \log\left(\frac{\#\ of\ 1\ GHz\ channels}{\#\ of\ 550\ MHz\ channels}\right) \quad (1)$$

Assuming a 77 channel 550 MHz loading and a 151 channel 1 GHz loading, the increase in CTB becomes:

$$\begin{aligned} Loading\ CTB_{Increase} &= \\ &= 20 \times \log\left(\frac{151}{77}\right) \\ &= 5.8\ dB \end{aligned} \quad (2)$$

This equation assumes amplifier behavior is predictable. The reality may be worse than this calculation shows. A beat pile up calculation shows a four fold increase in the number of beats in a 1 GHz system over the number in a 550 MHz system indicating a 12 dB increase in CTB. The maximum number of beats occurs in the area where hybrid amplifier's transistors will be most linear, so the actual measured change in CTB from 77 to 151 channels will be from 5.8 dB to 12 dB.

Higher amplifier output levels will be required at 1 GHz to overcome the higher losses that will occur in an upgrade from a 550 MHz system.

In an upgrade from a 550 MHz FTF system to a 1 GHz system, a 7.5 dB cable spacing between amplifiers at 550 MHz has been assumed. At 1 GHz this cable spacing would be 10.4 dB (for .500" PIII cable). The increase in amplifier output level to compensate for this would be:

$$\begin{aligned} \textit{Cable Output Level}_{increase} &= 10.4 - 7.5 \\ &= 2.9\ dB \end{aligned} \quad (3)$$

In addition an increase in flat loss of 2.6 dB has been assumed which would have to be overcome by an increase in amplifier output level. To fully compensate for the increased losses in an upgrade situation, an amplifier would have to operate at a higher output level. This output level would be equal to the sum of increase cable loss (3) plus the 2.6 dB increase in flat loss:

$$\begin{aligned} \textit{Output Level}_{increase} &= 2.9 + 2.6 \\ &= 5.5\ dB \end{aligned} \quad (4)$$

The CTB increase because of this increase in output level would be:

$$\begin{aligned} \textit{Output Level CTB}_{increase} &= 2 \times 5.5 \\ &= 11.0\ dB \end{aligned} \quad (5)$$

The amplifiers of a future network might have to compensate for the fiber link. If AM fiber is used, the increased channel loading at 1GHz will cause an increase in CTB. If the optical system itself does not degrade significantly with 151 channel loading, the RF portion will. AM light to RF receivers contain RF amplification. Assuming the optical system does not degrade significantly when the loading changes from 77 to 151 channels, the RF section will degrade as shown below:

$$\begin{aligned} \textit{Fiber CTB}_{increase} &= 20 \, x \log(\frac{151}{77}) \\ &= 5.8\ dB \end{aligned} \quad (6)$$

To compensate for these increases in distortions with 151 channels (from 77 channels) some system operating conditions would be changed. The RF amplifier cascade would be reduced, most likely by a factor of two. The improvement in CTB for a halving of the amplifier cascade (assuming amplifiers of equal performance) is:

$$\begin{aligned} \textit{Cascade CTB}_{decrease} &= 20 \, x \log(\frac{1}{2}) \\ &= -6.0\ dB \end{aligned} \quad (7)$$

FTF systems operate at 550 MHz with 9 dB of amplifier output tilt typically. Systems built for 1 GHz will operate with increased amplifier output tilt (this is possible because the RF amplifiers are not the CNR limitation) as much as 12.0 dB has been proposed. The improvement in CTB because of the 3 dB increase in amplifier output tilt would be:

$$Tilt\ CTB_{decrease} = \frac{\Delta\ tilt}{-2} = \frac{3}{-2} = -1.5\,dB \quad (8)$$

The total amplifier CTB change is the sum of the effects of the channel loading increase (2), the output level increase (5), the fiber link performance degradation (6), the reduction in RF amplifier cascade (7), and the increase in amplifier output tilt (8):

$$\begin{aligned} Loading\ CTB_{increase} &= 5.8\,dB \\ Output\ Level\ CTB_{increase} &= 11.0\,dB \\ AM\ Link\ CTB_{increase} &= 5.8\,dB \\ Cascade\ CTB_{decrease} &= -6.0\,dB \\ Tilt\ CTB_{decrease} &= -1.5\,dB \\ \hline Total\ CTB_{increase} &= 15.1\,dB \end{aligned} \quad (9)$$

To put this number in perspective, a 1 GHz amplifier would have to have 6 dB better CTB performance for a 77 channel loading than a current hybrid.

Amplifier Requirements for 77 Channel loading with Digital and/or FM Signals

While it may seem that loading the 550 MHz to 1 GHz spectrum with a combination of Digital and FM carriers would minimize the additional performance requirements of the RF amplifiers, the opposite is true.

Firstly, the 77 NTSC channels in the 50 to 550 MHz band will generate significant distortion products in the 550 MHz to 1 GHz band. The Digital and FM signals will be fundamentally more immune to these distortions than NTSC signals. However, the Digital and FM signals will operate at lower levels than the NTSC channels and so some improvement in amplifier distortion performance above 550 MHz will be necessary to ensure adequate signal quality.

Secondly the Digital and FM signals while operating at lower levels than the NTSC signals will add to the total amplifier output power and will produce additional distortion products. Whether these distortions appear as discrete carriers or as a general increase in the system noise floor will depend on nature, number and type of these signals (and will also depend upon picture content). Regardless of whether these distortions are discrete or widespread, the RF amplifiers will have to operate at higher levels of distortion performance than current 550 MHz amplifiers.

To understand how amplifiers of the future will be able to meet the performance criteria of future networks, it is important to understand what limits the performance of current 550 MHz amplifiers.

IV. CURRENT AMPLIFIER PERFORMANCE

Current 550 MHz RF amplifiers are not well optimized, they represent older technology that has been stretched from 300 or 450 MHz. Hybrid suppliers traditionally receive the blame for the performance limitations of our amplifiers. Some of this blame is well placed. Amplifier are presently being produced with power doubler hybrids with 77 channel CTB performance of 63 to 67 dB (at 44 dBmV flat output). As much as 5 dB of CTB is lost in the hybrid due to splitting and combining inefficiencies. The cost of RF amplifier hybrid production is directly related to the performance requirements and the hybrid supplier's yield of that product. Only 50% of transistors produced by a hybrid manufacturer can be used in hybrids to meet present performance requirements. New generation transistors for use in RF hybrids are designed more to improve hybrid supplier's yields than to advance the technology by increasing bandwidth or improving hybrid distortion performance.

However, not all of the blame for current amplifier performance lies with the hybrid suppliers. Amplifier manufacturers count on hybrid suppliers to improve the hybrids before we improve our amplifiers. For instance the input hybrid of an amplifier typically has a 5.5 dB noise figure. The amplifier that these hybrids are used in are specified with a noise figure of 9 dB. The 3.5 dB of carrier to noise performance is lost due to the input test points, diplex filter, output stage contributions, circuit inefficiencies within the station (see Figure 1.) and specification headroom taken by the amplifier manufacturer to ensure production yields.

Interstage losses (Interstage Gain Control and Pin Diode Attenuator) can be as much as 8 dB at 550 MHz. A reduction of these losses by 2 dB can reduce the gain requirement of the input hybrid resulting in better amplifier distortion performance.

Output losses directly affect the distortion (CTB, CSO etc.) performance of the amplifier. These distortions are primarily produced by the output hybrid U2 (see Figure 1.). An output hybrid has to overcome the output losses to deliver signals to the output ports, every dB of extra hybrid output level required to compensate for output losses represents another 2 dB degradation of the amplifier's CTB performance.

The losses of a typical 550 MHz amplifier used in FTF systems (with two outputs) are shown in Table 1. The insertion losses of the same devices at 1 GHz are also listed. Amplifier performance at bandwidths greater than 550 MHz will suffer from the same production limitations as does 550 MHz equipment.

V. AMPLIFIERS FOR FUTURE NETWORKS

If a 1 GHz amplifier were to be built using the 1 GHz losses shown in Table 1., internal losses would increase by:

$$\begin{aligned} &\mathit{Internal\ Loss}_{increase} \\ &= \sum 1\ \mathit{GHZ}_{losses} - \sum 550\ \mathit{MHz}_{losses} \\ &= 20.5 - 17.5 = 3.0\ dB \end{aligned} \tag{10}$$

In addition the amplifier would have to overcome the additional losses of the system at 1 GHz (4). This would mean a total increase in gain of the amplifier's hybrids of:

$$\begin{aligned} &\mathit{Hybrid\ Gain}_{increase} \\ &= \mathit{Internal\ Loss}_{increase} + \mathit{System\ Loss}_{increase} \\ &= 3.0 + 5.5\ dB = 8.5\ dB \end{aligned} \tag{11}$$

Therefore 18 and 22 dB gain hybrids used today for 550 MHz operation would have to be replaced with 22.5 dB and 26 dB gain hybrids respectively for 1 GHz operation.

The output loss increase alone of such an amplifier would be 1.5 dB which means that the output hybrid of the 1 GHz amplifier must have 3 dB better CTB performance than its 550 MHz counterpart. Combined with the required 5.5 dB higher output level required to overcome higher system losses (4), the total CTB improvement of the

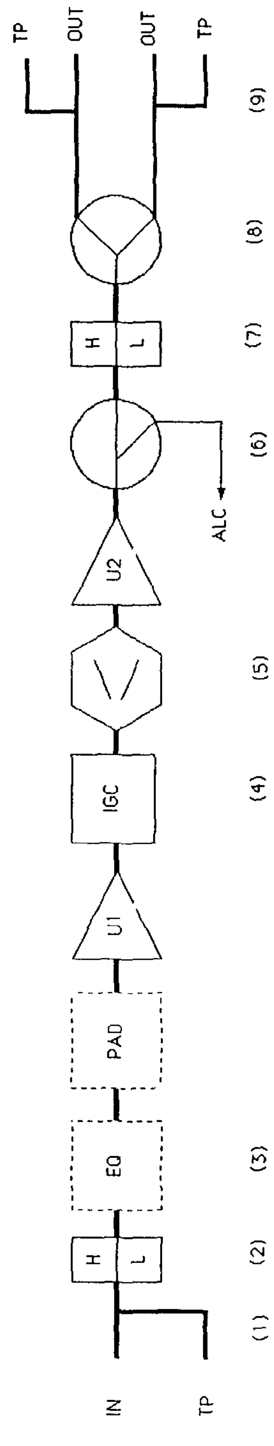

Figure 1. Amplifier Block Diagram

Table 1. Typical 550 MHz Amplifier Circuit Losses

Block Diagram Ref #	Component Description	dB Loss @ 550 MHz	dB Loss @ 1 GHz
(1)	Input Test Point	0.75	1.0
(2)	Diplex Filter	0.5	0.75
(3)	Equalizer	1.0	1.0
(4)	Interstage Gain Compensator	1.0	1.0
(5)	Pin Diode Attenuator	7.0	8.0
(6)	ALC Pick Off	1.5	2.0
(7)	Diplex Filter	0.5	0.75
(8)	Two way splitter	4.5	5.0
(9)	Output Test Point	0.75	1.0
	TOTAL INSERTION LOSS	**17.50**	**20.50**

Table 2. Optimized Amplifier Circuit Losses

Block Diagram Ref #	Component Description	dB Loss @ 550 MHz	dB Loss @ 1 GHz
(1)	Input Test Point (eliminated)	0.0	0.0
(2)	Diplex Filter	0.2	0.2
(3)	Equalizer	1.0	1.0
(4)	Interstage Gain Compensator	1.0	1.0
(5)	Pin Diode Attenuator	4.5	5.5
(6)	ALC Pick Off	0.6	0.8
(7)	Diplex Filter	0.2	0.2
(8)	Two way splitter	3.5	3.75
(9)	Output Test Point	0.75	1.0
	TOTAL INSERTION LOSS	**11.75**	**13.45**

Reference Numbers refer to Figure 1.

output hybrid would have to be:

$$Output\ Hybrid\ CTB\ \Delta = 3 + 11 \quad (12)$$
$$= 14\ dB$$

If 1 GHz amplifiers were built with the insertion losses listed in Table 1., the hybrids required would be very expensive and draw large amounts of power and could not be developed based on the next generation of RF transistors. The lack of optimization of current 550 MHz amplifiers, the performance required for extended bandwidth amplifiers can only be achieved by selecting and optimizing the active and passive devices used to construct them.

Optimized Amplifiers for Future Networks

The most significant area of optimization is the output losses between the output hybrid U2 (see Figure 1.) and the output ports of the amplifier housing. Every dB of reduction is nearly 2 dB of CTB performance improvement. A 2.5 dB reduction in these losses improves the amplifier's CTB performance by 5 dB. A reduction of 2.5 dB in interstage losses has the same distortion effect as replacing U1 (see Figure 1.) with a power doubler hybrid. In addition to improving distortion effects of the station pre-amp, interstage optimization reduces the power consumed by the station by as much as 7 Watts. Further optimization reduces gain requirements of the hybrids which results in further distortion improvements and increased gain stability versus temperature.

Table 2. represents optimized losses for the amplifier shown in Figure 1. The insertion losses are optimized for 1 GHz, and show the effect on 550 MHz losses. The input test point has been eliminated to help achieve this optimization.

Optimization is the process of maximizing the performance of each circuit in the amplifier. As can be seen tenths of dBs can add up significantly, 7.05 dB of loss can be removed from the amplifier shown in Figure 1. Achieving the circuit optimization described in this paper has been done by computer modeling of circuits in such a manner that the computer model functions exactly as the circuit functions when tested by normal sweep equipment. Computer circuit optimization is verified by bench performance test.

The selection of optimized active and passive components for a 1 GHz amplifier will result in increased cost and/or a loss of traditional features. A $0.75 transformer used in a splitter with a 5.5 dB loss at 1 GHz could be replaced by a $1.50 transformer and give an insertion loss of 4.0 dB at 1 GHz.

Applying the same techniques to the amplifiers active amplification stages would improve CTB, CSO and noise performance of the amplifier and also the RF portion of the fiber link.

VI. CONCLUSION

1. Future networks will need RF amplifiers with increased bandwidth and distortion performance capabilities regardless of the type of additional channel loading.

2. 1 GHz amplifiers built using techniques in use today will require active devices that are many years (perhaps a decade) away from being available.

3. Active and passive component optimization with an RF amplifier will allow practical 1 GHz amplifier designs to realized within a few years.

An Overview of the JPEG and MPEG Video Compression Specifications

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ABSTRACT

Digital Video Compression has become one of the fastest growing technologies in the last several years. As part of the development of this technology, several digital video compression standards are emerging. One of these standards JPEG, is intended for specifying the compression and decompression of single frame images. Another standard, MPEG, is intended for specifying the compression and decompression of motion video. These standards may have a significant effect on the CATV industry in the future. This paper describes these standards and gives an explanation of how they work.

INTRODUCTION

Over the past year, two digital video compression specifications have neared completion. The JPEG (Joint Photographic Experts Group) specification is being developed by a joint ISO/CITT committee. Also the MPEG (Motion Picture Experts Group) specification is being developed by a joint ISO/IEC committee. These specifications are important steps in the evolution of digital video systems.

The need for image data compression is apparent when we examine the data required to represent a single image. A 720 by 480 pixel image stored using 24 bits per pixel (eight bits per component for a red, green and blue color space) will require 1.036 Mbytes of memory. If this size image is used in a full motion video system with a frame rate of 30 frames per second, the channel data rate will be 248.8 Mbits/sec. The bandwidth required to transmit this signal using a modulation scheme which has a bit packing rate of 3 bits per hertz would be 82.9 MHz. This is too much bandwidth to be used in most practical systems. Some means of reducing the amount of data required to represent the signal must be found to make a digital video system viable.

DATA COMPRESSION

There are two general classes of data compression, lossless and lossy. Lossless data compression schemes rely on reducing the redundant information in the data while representing the data with as few logical indicators (bits) as possible. These schemes can be used for any kind of data since no information is lost in the data compression and expansion process. Lossy data compression schemes throw out information and rely on human psycho-visual properties in order to keep the distortions produced by data compression from being perceived. Both JPEG and MPEG are lossy compression schemes, however they both have lossy and lossless elements. Before describing the details of the JPEG and MPEG specifications, a discussion of some of these techniques is in order.

Three common lossless data compression techniques which are used in both JPEG and MPEG are Run Length Coding, Variable Word Length Coding and Predictive Coding.

Run Length Coding

Run Length Coding is used when the data tends to contain long strings of identical characters. Instead of transmitting the characters, the number of characters in the string is transmitted. For instance JPEG and MPEG use zero run length coding. When repeated zeros occur in the data, the number of zeros is transmitted instead of the actual zeros.

Variable Word Length Coding

Variable Word Length Coding, which is also called Huffman Coding, uses different length codes to represent characters. The length of codes used to represent characters is inversely proportional to the probability of occurrence of the character. In addition Huffman Codes have a prefix property which means that no short code group is duplicated as the beginning of a larger group(1). If the estimate of the probability of characters occurring is incorrect, Huffman Coding can increase the number of bits required to represent the data.

Predictive Coding

Predictive coding first estimates the present data value through some prediction algorithm. An error signal which is the difference between the actual data and the predicted data is then transmitted. The decoder uses the same prediction method as the encoder. By adding the error signal to the prediction, the original signal is exactly reconstructed in the decoder. If the prediction is a good estimate of the original data, the error signal will be small and can be represented by fewer bits than the original data.

Transform Coding

A common lossy coding technique which is used in MPEG and JPEG is transform coding. Transform coding takes image data in the spatial domain which tends to have high pixel to pixel correlation and transforms it in a manner which tends to group the energy into relatively few transform coefficients. The coefficients which are small or zero can be eliminated without any significant effect on the image(2).

JPEG SPECIFICATION

The JPEG specification describes two types of image compression algorithms for compressing and decompressing single gray-scale and color images. The first algorithm is the "Baseline/Extended" algorithm which is based on lossy transform coding and provides the greatest compression of the two algorithms. The baseline system is the minimum configuration for the transform coding based algorithm. The extended system includes the baseline system and some additional features. The second compression algorithm is the "Independent Function" system which is based on two dimensional differential pulse code modulation. The Independent Function system can be a lossless or lossy system and is intended for very high quality images. It does not provide large amounts of data compression. Because the JPEG system is intended for many applications, it supports numerous image color space representations and a wide range of image spatial resolutions. It is a symmetric algorithm in that the decoder performs the inverse process of the encoder and is therefore of the same complexity.

DCT

The block diagram of a baseline JPEG encoder is shown in figure 1. Each component of the image color space (i.e. R, G, or B for RGB, Y, U, or V for YUV color space) is partitioned into 8 x 8 pixel blocks. Each of these blocks is then coded. The compression algorithm is based on the two dimensional Discrete Cosine Transform or DCT given below(3).

$$F(u,v)=\left(\frac{1}{4}\right) c(u)\, c(v) \sum_{i=0}^{7} \sum_{j=0}^{7} f(i,j) \cos\left((2i+1)u\frac{\pi}{16}\right) \cos\left((2j+1)v\frac{\pi}{16}\right)$$

$$c(u)=\frac{1}{\sqrt{2}} \quad for \quad u=0$$

$$c(u)=1 \quad for \quad u\neq 0$$

$$c(v)=\frac{1}{\sqrt{2}} \quad for \quad v=0$$

$$c(v)=1 \quad for \quad v\neq 0$$

This transform takes a 8 X 8 pixel block of data which represents the horizontal and vertical spatial components of the image and transforms the data into a 8 X 8 block of coefficients which represent the horizontal and vertical frequency components of the

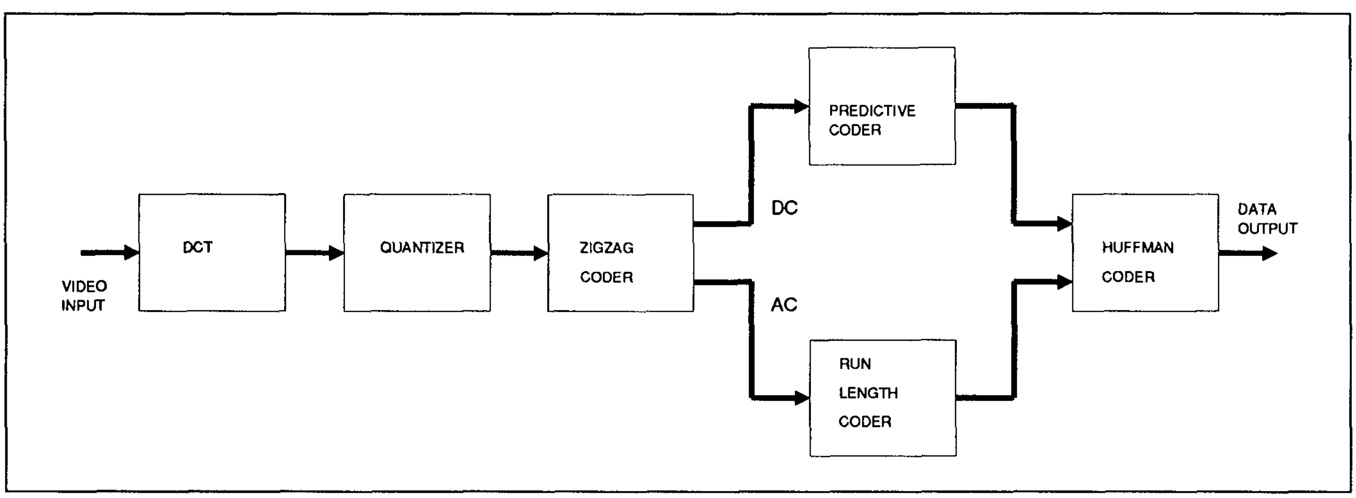


Figure 1 JPEG Encoder

10	20	30	40	50	60	70	80
10	20	30	40	50	60	70	80
10	20	30	40	50	60	70	80
10	20	30	40	50	60	70	80
10	20	30	40	50	60	70	80
10	20	30	40	50	60	70	80
10	20	30	40	50	60	70	80
10	20	30	40	50	60	70	80

Figure 2A 8x8Pixel Image Data

INCREASING HORIZONTAL FREQUENCY

INCREASING VERTICAL FREQUENCY

360	-182	-.2	-19	-.2	-5.7	-.1	-1.5
-1	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0
-.1	0	0	0	0	0	0	0
-.1	0	0	0	0	0	0	0
-.1	0	0	0	0	0	0	0
-.1	0	0	0	0	0	0	0
-.1	0	0	0	0	0	0	0

Figure 2B Transformed Image Data

data block. This has the property of concentrating the energy of the image into fewer terms.

Figure 2A represents data from an 8 X 8 pixel portion of the luminance component of an image. This block represents a portion of an image with increasing brightness from left to right. Figure 2B represents the DCT of this data block. Most of the higher frequency terms are very small or zero.

Quantization

The next step in compressing the image data is to quantize the DCT coefficients. Quantization is a process which breaks the possible DCT coefficient range into windows, then generates a code to represent each window. Each coefficient is linearly quantized (the quantization window size is constant) independently of the other coefficients. The quantization window or step size determines how

16	11	0	16	24	40	51	61
12	12	14	19	26	58	60	55
14	13	16	24	40	57	69	56
14	17	22	29	51	87	80	62
18	22	37	56	68	109	103	77
24	35	55	64	81	104	113	92
49	64	78	87	103	121	120	101
72	92	95	98	112	100	103	99

Figure 3 Quantizer Step Size

23	-17	0	-1	0	0	0	0
0	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0

Fiqure 4 Quantized DCT Coef.

many bits will be used to represent the data. The larger the quantization step, the greater the distortion produced by the quantization process. However, fewer bits will be required to represent the data. Since the JPEG specification allows each DCT coefficient to be assigned its own quantization step size, more important frequency terms can be represented more accurately than less important terms. Many of the higher frequency terms will be set to zero by quantization. Figure 3 shows a quantization matrix containing typical quantizer step sizes. The data in figure 2B is quantized using the quantizer step sizes in figure 3 to produce the data in figure 4.

ZIG-ZAG Scanning

The frequency terms are scanned or reordered according to increasing spatial frequency. Since higher spatial frequency terms are often zero or quantized to zero, there will tend to be many zero terms in a row. This prepares the data to be Run Length Coded. The scanning process is performed by zigzagging diagonally across the block of DCT coefficients, hence it is known as zigzag scanning. Figure 4 illustrates the quantized and ZIG-ZAG scanned data.

ENTROPY CODING

Entropy coding is a general term for lossless coding techniques which are used to code the quantized DCT coefficients.

Predictive Coding

Unlike the higher frequency terms that can be coarsely quantized without significant image degradation, the DC term (first term in the transformed data block) must be represented accurately. The DC term is predictive coded using the previous 8 X 8 block's DC term as the predictor. The difference between the predictor and the present block's DC coefficient is Huffman Coded to form the first portion of the block's output data.

Run Length Coding

The AC DCT coefficients (remaining 63 terms) are Zero run length coded and then Huffman coded. The number of zeros preceding a non-zero AC term is combined with the length of the nonzero term in bits. This number is then Huffman coded and the actual bits which represent the non-zero term are appended to the end of the Huffman Code forming output data. This process is repeated until the end of the data block is reached. If all of the terms to the end of the block are zero, then an end of block signal is coded. Each 8 X 8 pixel block in the image is encoded in this manner until the entire image is coded.

The data compression obtained using JPEG is varied by changing the quantization levels of the quantizer. There is no way to know what the compression ratio will be (and therefore know how much output data there will be) until the image is compressed. For this reason JPEG is a variable output data rate system. No quantization levels are called out in the specification, however communications syntax is specified which allows this information to be sent from the encoder to the decoder. Also the Huffman codes are not specified but can be sent from the encoder to the decoder. This allows the compression/decompression process to be optimized for a particular image or group of images.

The Extended system has several features which are not included in the baseline system. One of these is a progressive scan mode. The baseline system operates in a sequential mode which sends all of the image information in raster scan order (left to right, top to bottom). The progressive mode sends some of the information for the entire image so that a low quality image can be produced. More information is progressively sent until all of the image information has been sent. The Extended system also allows for arithmetic coding instead of Huffman Coding. Arithmetic coding uses the statistics of the images being coded which increases the amount of compression obtained.

JPEG produces images which are recognizable at entropy levels of .15 bits/pixel. Excellent quality images are obtained at entropy levels of .75 bits/pixel and images which are essentially indistinguishable from the original are obtained at entropy levels of 1.5 bits/pixel(4). This implies compression ratios of 160:1, 32:1 and 16:1 respectively (assuming 24 bits / pixel source images).

While JPEG could be used to encode video by compressing each frame of the video sequence independently, this would not achieve as high of a compression level as is possible since no temporal or frame to frame redundancy is removed. For instance, if there is no motion in a video sequence, then all but the first frame of video is redundant. Video compression which only operates within a single frame of video is termed intra-frame compression while algorithms which operate using information from more than one frame are termed inter-frame compression.

MPEG SPECIFICATION

The MPEG standard is a lossy inter-frame compression scheme based on the DCT. The specification itself is a specification for a decoder only, and leaves the implementation of the encoder to the system designer. The specification assures that if an encoder uses the proper syntax to encode image data, then a MPEG decoder will be able to decode it. Because the signal quality is a strong function of the encoder design, all MPEG systems will not have the same image quality. MPEG relies on Signal Bandwidth Reduction, Lossy Compression and Lossless Compression to obtain image data compression.

Bandwidth Reduction

The input video format for MPEG is 352 X 288 pixel non-interlaced video frames in a Y,U,V color space. Converting the video to this format from a CCIR 601 YUV 4:2:2 format requires reducing the amount of data (bandwidth) by almost 1/6 before the compression process starts.

MPEG COMPRESSION

The block diagram in figure 5 depicts a MPEG decoder and a typical MPEG encoder. The MPEG specification defines three different types of video frame coding: Intra-coded(I), Predictive-coded(P), and Bidirectionally Predictive-coded(B). Each of these types of frame coding requires a different amount of data to encode. An Intra-coded frame requires the most data to code with a Predictive-coded frame requiring the second largest amount of data. The sequence of these frames is determined by the encoder. Figure 6 shows how a typical MPEG video sequence might be constructed from these frames.

The Intra-coded frames are very similar to JPEG coding except in the details of the Huffman coding. With Predictive-coded frames, the previous frame is used as a prediction for the present frame. The encoder then uses this predicted frame to generate an error signal which is encoded like an Intra-frame. The same prediction frame is generated in the decoder and the decoded error signal is added to it to generate the final output. The same process takes place with Bidirectionally Predictive-coded frames except that either the next frame, the previous frame or both are used to make a prediction of the present frame.

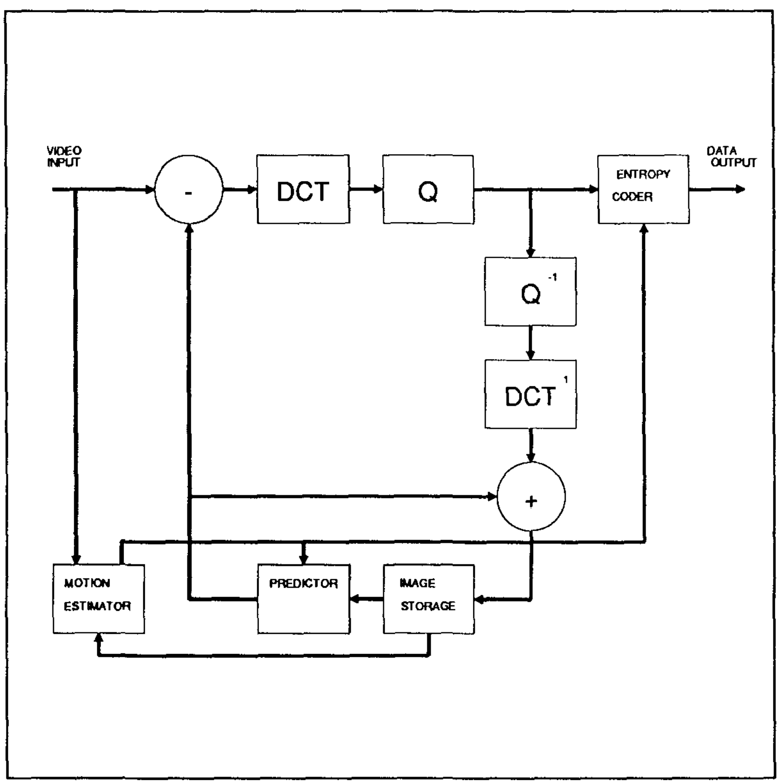
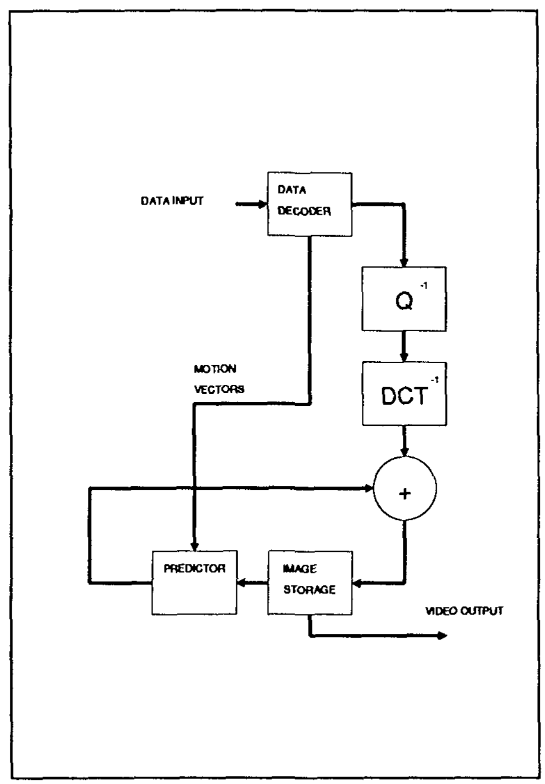


Figure 5 MPEG Encoder



MPEG Decoder

Motion Compensation
MPEG uses motion compensation to improve the accuracy of predicted frames. As with any predictive coded system, the closer the predicted frame is to the present frame, the less energy in the error signal and the fewer bits which are required to encode it. If there is no motion between the frames there is no difference to encode.

Motion compensation is performed by dividing the image to be coded into 16 x 16 pixel blocks (macroblocks). Each block in the image is associated with an optimally matched block in the prediction frame. A motion vector describes the location of this optimum block in the prediction frame. This vector indicates how far in .5 pixel resolution the optimum block is from the macroblocks original position. Thus by transmitting the relatively few bits required for motion vectors, the prediction can be greatly improved.

IMAGE QUALITY

There are several factors which determine the quality of an MPEG video sequence which are not specified in the specification. These are:

1. Motion Vector Generation
2. Image Pre and Post Processing
3. Image type sequence control
4. Rate control

All of the above items are determined by the encoder and are interrelated.

Motion Vector Generation
MPEG syntax describes how motion vectors are to be encoded and how the decoder processes them, however the manner in which they are generated is left to the system designer. The more accurate the predicted frame generated by the motion vectors, the less information that has to be encoded in the error signal. This allows more accurate coding of the error signal. Motion Vector generation is one of the most computationally intensive aspects of MPEG encoding. An inexpensive encoder might not generate any motion vectors and it would still be MPEG, however the image quality would be poor. Likewise for point-multipoint systems where few encoders will drive many decoders, highly elaborate motion vector generation might take place to obtain the highest quality images possible.

Image Pre and Post Processing
The manner in which the source video is converter into a format suitable for MPEG and the manner in which the MPEG decoder output is converted to a format suitable for the display device is not specified in the MPEG specification. It will however have a great bearing on the image quality.

Image Type Sequence Control
An MPEG decoder should be able to decode I, P and B frames. The specification does not indicate an algorithm for choosing which types of frames should be coded. The sophistication of the algorithm which makes this determination in the encoder will have an effect on the video quality. While it is possible to encode a fixed pattern of I, P and B frames, an algorithm which adjusts this pattern according to the image source material has the possibility of giving better results.

Rate Control
Like JPEG, MPEG has a coder data rate which is dependent on desired compression levels and the subject material of the image. In order to make the system operate through a fixed data rate communications channel, the output data from the

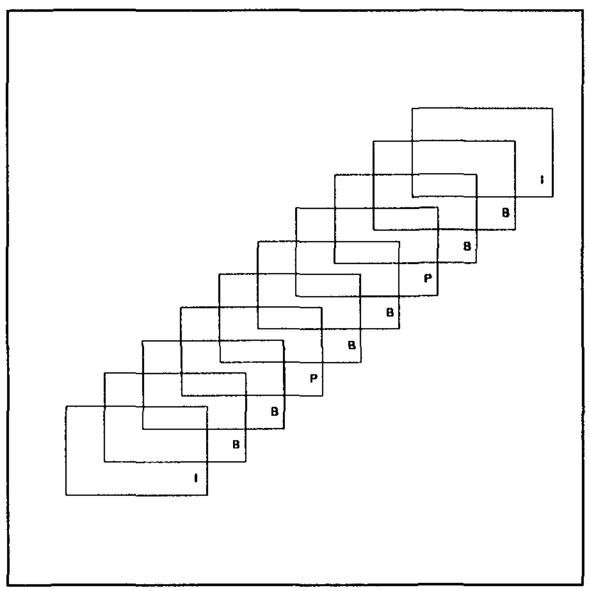


Figure 6 Typical MPEG Frame Sequence

encoder is buffered and the input data to the decoder is buffered. The compression level is dynamically adjusted by changing the quantizer scale factor or changing the type of frame which is coded so that the average data rate into the buffer is equal to the channel data rate. This allows the data to be read out of the buffer at a constant rate without over or under running the buffer. The output data rate for MPEG is constrained to 1.5 Mbits/sec. While the syntax will support higher data rates, a system operating at these rates would be considered some form of MPEG extension.

The MPEG Encoder is typically significantly more complex than a MPEG decoder. This is because most encoders will generate motion vectors which is a computationally intensive process.

CONCLUSION

This has described the JPEG and MPEG specification as they stand at this point. It is unlikely that any significant technical changes will be made before the specifications are approved.

A MPEG specification is presently being written for audio data compression which will provide for variable quality audio of up to Compact Disk quality stereo. In addition, there is an effort under way to write an MPEG II specification which will provide near broadcast quality video at a higher data rate and handle interlaced pictures. All of these specifications will provide a basis for many video systems of the future.

(1) D.A. Huffman "A Method for the Construction of Minimum Redundancy Codes," Proc IRE 40, 1089, 1952.

(2) W. K. Pratt, "Digital Image Processing," New York: Wiley, 1978.

(3) K. R. Rao, "Discrete Cosine Transform," New York: Academic Press, 1990.

(4) ISO/CITT, "JPEG Draft Specification," May 1990.

An 80-Channel High-Performance Video Transport System Over Fiber Using FM-SCM Techniques for Super-Trunk Applications

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ABSTRACT

Use of microwave subcarrier multiplexing (SCM) for transmission and distribution of a large number of video channels over optical fiber offers an attractive alternative to existing coaxial and multi-fiber systems. This paper describes a SCM system carrying 80 FM video channels, achieving a 60 dB SNR and an optical power budget of 14 dB. A transmission distance of 20 km with a 3 dB margin was also achieved without degradation. Such a system can be used in next-generation CATV super-trunk-type applications. Superior performance in terms of SNR and optical power was achieved by implementing the system based on optimization and judicious choice of parameters affecting its performance. The frequency plan, based on 9 MHz frequency deviation, uses the 3–6 GHz frequency band. The system uses high-frequency lasers having a 10 GHz bandwidth and RIN below –140 dB/Hz.

1. INTRODUCTION

Fiber-based super-trunk systems are currently used to provide high-quality FM video transmission between headends and several CATV hub locations.[1] Present-day commercially available systems,[2] using FM techniques, carry up to 16 video channels per fiber and offer hightransmission performance (SNR ≥ 60 dB). Transport of 80 high-quality video channels would require operating five of these 16-channel systems in parallel, necessitating use of multiple fibers and the associated optical transmitters and receivers.

FM-SCM (microwave subcarrier multiplexing of frequency-modulated video) has recently been shown to be an attractive alternative for transmission and distribution of a large number of video channels over a single fiber.[3–7] FM-SCM uses high-frequency lasers and photodetectors to access the frequency region above 1 GHz. Olshansky et al.[4–6] have reported SCM systems in the 2.7–7.6 GHz region carrying 60 and 120 FM video channels. For weighted SNR of 56 dB, the optical power budget was 14 dB. Way et al.[7] transported 90 FM video channels in the 1.7–6.2 GHz band over fiber and also achieved a weighted SNR of 56 dB with an optical power budget of 12 dB. The 56 dB SNR of most of the reported FM-SCM systems is less than the 60 dB preferred for super-trunk applications.

This paper describes a FM-SCM system for the next-generation high-capacity super trunk carrying 80 video channels over a single fiber with a 60 dB SNR that achieves an optical power budget of 14 dB. The first section develops an analysis of the fiber-optic system. This is followed by a discussion of the frequency plan and system implementation. The final two sections discuss the experimental results of laboratory measurements on a prototype system and the conclusions.

2. ANALYSIS

For super-trunk transmission systems, a high source SNR is required to ensure that a high-quality signal reaches the end-user. The SNR of any channel in a multichannel system can be described as shown in equation 1.

$$SNR_w = \{CNR_{if} + 10\ \mathrm{Log}(B_{if}/B_{bb})\} + A + 20\ \mathrm{Log}(1.6F_d/B_{bb}) \qquad (1)$$

Where:

- SNR_w = Weighted video signal-to-noise ratio
- CNR_{if} = Carrier-to-noise ratio in the IF bandwidth
- A = 20.37 dB — made up of weighting, de-emphasis, and conversion factors
- B_{bb} = Baseband filter bandwidth
- B_{if} = Intermediate frequency bandwidth (IF)
- F_d = FM deviation — sync tip to peak white (STPW)

This equation is frequently referred to as the FM advantage equation since the last two terms represent the improvement in the detected signal as a result of employing FM modulation. From this equation, the FM deviation required to yield the same SNR performance for an 80-channel system as that of the 16-channel system can be calculated. If it is assumed that the same amount of total modulation power is available in both cases, an FM deviation of 2.2 times that of a 16-channel system is required for operation of an 80-channel system for the same figures. Commercial fiber-optic systems use about 4 MHz FM deviation for a 16-channel system, implying that a super trunk employing 80 channels would require a 9 MHz deviation.

Another important factor in the specification of a super trunk is the optical power budget. It is simply the difference between the optical power required at the receiver and the laser-coupled power. To calculate the received power, the carrier-to-noise (CNR) equation must be solved, which includes the major noise terms that cause impairment in the signal. The CNR is expressed as shown in equation 2.[8] The terms of the equation have been arranged to show the dependence of CNR on optical power and modulation depth.

$$CNR_{if} \geq \frac{0.5}{\dfrac{kTBN_F}{R_f(mR_dP_s)^2} + \dfrac{2eB}{m^2R_dP_s} + \dfrac{B(RIN)}{m^2} + C_2m^2N_2 + C_3m^4N_3}. \qquad (2)$$

Where:

- m = Modulation depth per channel
- R_d = Photodetector responsivity
- R_f = Amplifier input impedance
- k = Boltzmann constant
- T = Absolute temperature
- B = IF bandwidth
- N_F = Noise figure of amplifier
- P_s = Received optical power
- e = Electronic charge
- RIN = Relative intensity noise
- C_2 = Second-order intermod coefficient
- C_3 = Third-order intermod coefficient
- N_2 = Number of second-order products
- N_3 = Number of third-order products

The first term in the equation represents the impairment due to thermal noise, while the second and third terms are the shot and relative intensity noise (RIN) impairments, respectively. The last two terms are the degradation due to the second-order and third-order intermodulation distortion.

Examination of equation 2 reveals some general trends in achieving a given CNR. Increasing optical power increases CNR for the terms involving thermal and shot noise. Increasing optical power does not change the CNR for terms involving RIN and intermodulation noise. Consequently, these latter impairments cannot be lowered by increasing the optical power. Increasing the modulation depth increases the CNR for terms involving thermal, shot, and RIN noise while decreasing the CNR for terms involving intermodulation noise. Thus, there is an optimum value of modulation depth where the thermal, shot, and RIN terms are balanced off against the intermodulation terms.

In wideband microwave systems, 50 Ω amplifiers are employed. Under these conditions, the shot noise will be small. In a thermally noise limited system with no intermodulation distortion, the optical power is proportional to the square root of the CNR. Equation 1 shows that a doubling of the frequency deviation yields a 6 dB SNR improvement. However, since the optical power is proportional to the square root of the CNR, only a 3 dB gain in received optical power is realized. This reveals that while increasing the frequency deviation results in a one-for-one increase in the SNR, only half the increase in the optical power budget is realized.

Figure 1 shows the effect of RIN as a function of modulation depth for a 9 MHz FM system. For this figure, a 50 Ω amplifier with a noise figure of 2.5 dB was assumed for the thermal noise model, and a PIN photodiode was used to predict the shot noise. In order to achieve the highest optical power budget, it is desirable to operate at as high a modulation depth as possible. If the modulation power is divided equally between all the channels, the modulation index is 100%/80, or 1.25%. This would produce only a modest level for the optical power budget while placing a severe constraint on the laser's RIN. Typical values for laser RINs are between −130 dB/Hz and −145 dB/Hz, depending on, among other things, the operating frequency. It has been reported that higher levels of modulation per channel are possible since not all the channels will carry the maximum power at one time.[6] Levels of modulation of 6% to 8% would negate the impairment and thereby maximize the optical budget. However, the higher modulation levels will be limited by the intermodulation distortion, as discussed in the next paragraph.

To determine the nonlinear impairments in the FM-SCM system requires the knowledge of the coefficients C_2 and C_3 used in equation 2. These coefficients have been calculated by the following equations.[9,10]

$$C_2 \approx 0.5\,(f/f_r)^2\,R(f) \qquad (3)$$

$$C_3 \approx \left[\{(f/f_r)^4 - 0.5(f/f_r)^2\}^2 + (2\pi\tau_p f_r)(f/f_r)^4\right]^{1/2} R(f)R(2f) \qquad (4)$$

$$R(f) = \{((f/f_r)^2 - 1)^2 + (f\Gamma)^2/(2\pi f^2)^2\}^{-1/2} \qquad (5)$$

Where:

$R(f)$ = Small-signal frequency response
f_r = Relaxation resonance frequency
Γ = Damping rate $\approx 0.32 \times 10^{-9}\,(f_r)^2$
τ_p = Photon lifetime

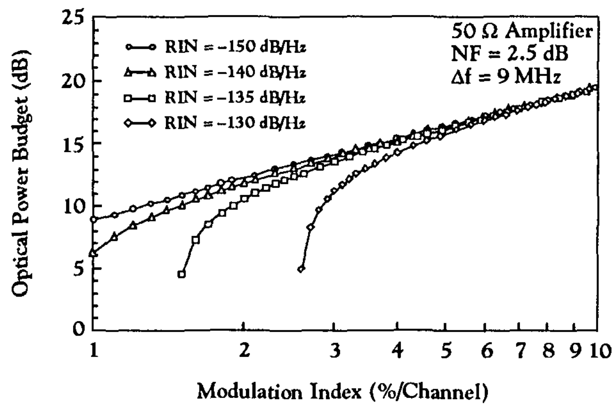

Figure 1. Optical Power Budget As a Function of OMI for Different RIN Values.

The values calculated by these equations were derived using a small-signal model for the laser. The results are expected to be optimistic since the modulation levels investigated here constitute large-signal behavior. Figure 2 shows a plot of second-order and third-order distortion coefficient as a function of carrier frequency for a laser with a 9 GHz resonance frequency. The third-order term peaks at the resonance frequency and one-half the resonance frequency. The second-order terms increase to a maximum at the resonance frequency. Notice, however, that the impairment due to intermodulation distortion is also a function of the number of intermodulation products at a particular frequency. Thus some flexibility is possible to minimize the distortion by choosing a plan that minimizes the product of the intermodulation distortion coefficients and the number of intermodulation terms.

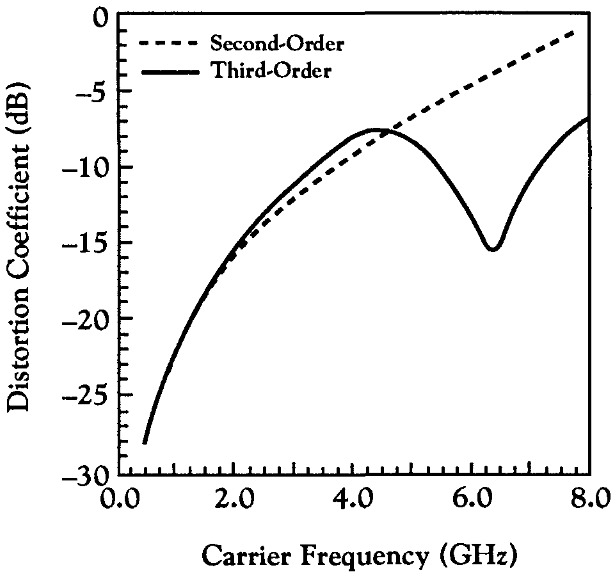

Figure 2. Laser Distortion,
Resonance Frequency = 9 GHz.

3. FREQUENCY PLAN

The selection of the frequency plan depends on many factors, including intermodulation distortion and laser frequency response. The use of an octave band of frequencies avoids the second-order distortion. The channel spacing must be selected taking into account the total bandwidth of the signal. The per channel bandwidth is dependent on the FM deviation. A 9 MHz deviation was specified as the minimum deviation required in order to achieve the required SNR (as discussed earlier). Larger deviations would use too much bandwidth. For this deviation, a channel spacing of 40 MHz was specified. Tests have verified that this spacing is adequate to pass the video signal with >60 dB SNR.[11] Assuming this channel spacing, a frequency span from 3.4 GHz to 6.7 GHz containing all 80 channels within an octave band was chosen.

4. SYSTEM IMPLEMENTATION

Figure 3 shows a simplified diagram of a FM-SCM video transport system. A number of FM-modulated video channels are up-converted to the microwave band and power-combined in the transmitter. The scheme uses block conversion of a number of channels to limit the component count needed to heterodyne the signals. The practical block size is limited by the filters that must reject unwanted images, local oscillator, and signal leakage. The composite microwave signal is used to intensity-modulate a wide-bandwidth laser. After transmission through a span of single-mode fiber and detection with a wide-bandwidth optical receiver, the microwave signal is down-converted and demodulated.

The system was implemented to verify the feasibility of using a fiber-optic link to transport 80 channels of video information. The unit consisted of a bank of 12 voltage-controlled oscillators modulated by 11 video sources and one test channel. The video sources were obtained from 11 satellite receivers receiving the standard satellite signals. The tests were done with no audio subcarrier. The test channel was modulated with a Tektronix 1910 video pattern generator. This bank of 12 channels was then block-converted into 7 contiguous frequency slots to produce an 84-channel system. No block conversion filters were employed for this stage of the tests. A notch filter was employed in the test channel to eliminate any impairment due to the microwave electronics. The sources were Ortel (10 GHz) lasers. An RF attenuator at the input of the laser was used to adjust the modulation level. An optical isolator was used after the pigtail of the laser to minimize reflections. A low-reflectance optical attenuator was used to evaluate the optical power budget. Selected tests were also conducted using optical fiber. The signals were detected with a wideband optical detector and a low-noise 50 Ω amplifier and down-converted to a satellite channel for demodulation by a satellite receiver. Signal analysis was performed using a Tektronix VM-700 video test set.

5. RESULTS

Figure 4 shows the test results of three lasers. For each level of modulation, the optical attenuator

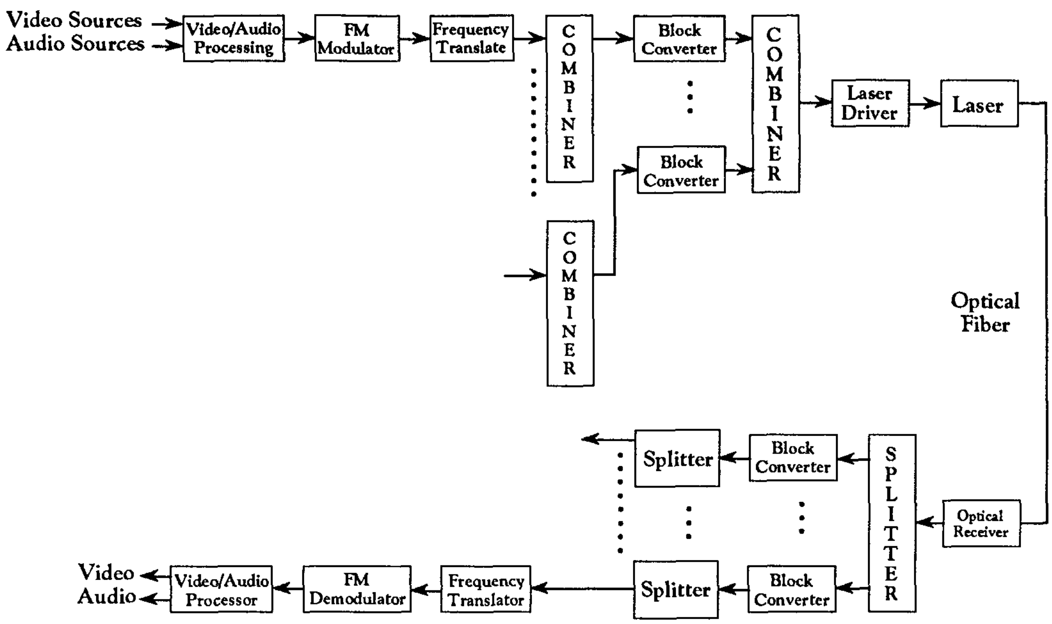

Figure 3. FM-SCM Transmission System Block Diagram.

was adjusted for a 60 dB SNR. The value of the optical attenuator plus the losses in the optical isolator and associated connectors were added to obtain the optical link budget. The curves were not normalized for laser power output, which accounts for the differing maximum values. The sharp fall-off at about 4%/channel modulation is attributed to the third-order intermodulation distortion. The dip in the center of laser 1 is considered measurement anomaly and was not a typical response. The computed simulation is shown as a dotted line. There is good agreement at low modulation levels, where thermal and RIN noise dominate. The agreement at high modulation levels is not good and suggests that further investigation into modeling laser large-signal behavior is warranted.

The system was also operated over 20 km of single-mode fiber. The fiber zero dispersion point was matched to within 10 nm of the laser central wavelength. Under these conditions, no dispersive effects were noted. The fiber attenuation was 0.5 dB/km. A 3 dB margin was recorded, resulting in an optical power budget of 13 dB exclusive of splices.

Another test was conducted to determine the sensitivity of SNR to optimum modulation index. Figure 5 shows the optical power budget as a function of modulation depth for three SNR ratios. As expected, the optimum modulation index is higher at lower SNR and the optimum range is broader. The optical power budget is also higher since the CNR requirement is lower, allowing more noise to be present in the system.

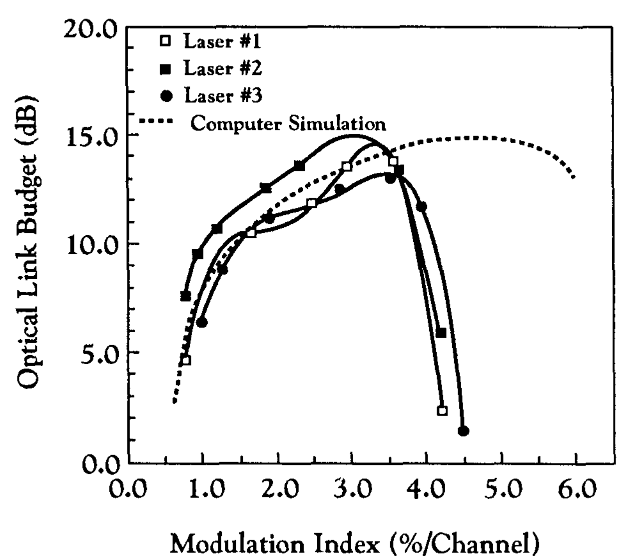

Figure 4. FM-SCM Performance, SNR = 60 dB.

6. CONCLUSIONS

This paper considers the use of FM-SCM techniques for the transmission and distribution of 80 video channels over fiber in a super-trunk-type environment. It identifies the effects of various parameters on system performance. The optimization and judicious choice of FM-SCM parameters offers an opportunity to enhance the system performance and facilitate its practical implementation. An 84-channel FM-SCM system was implemented based on parameter optimization and current state-of-the-art components. By operating over an octave transmission band, the system had an optical power budget of about 14 dB at a 60 dB SNR, for an optimum modulation depth of 3.5%/channel.

The performance of FM-SCM systems should further improve with the recent availability of lasers having higher coupled power and improved high-frequency characteristics and linearity (viz. Ortel type 1530B). Use of high-frequency low-noise APDs should increase the receiver sensitivity and hence the power budget. It is estimated these two improvements would add 6 dB to the achievable power budget.

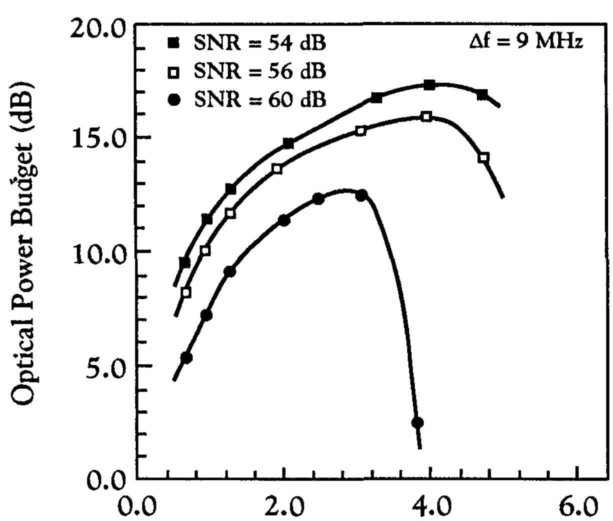

Figure 5. FM-SCM Performance.

ACKNOWLEDGMENTS

The authors thank R. Olshansky for many helpful discussions and A. Varallo for technical support and system construction and measurements.

REFERENCES

1. W.C. Brinkerhuff and I.M. Levi, *NCTA Tech. Proc.*, pp. 58–66 (1988).

2. For example, Catel, Synchronous Technology, American Lightwave Systems, and others.

3. T.E. Darcie, *J. Lightwave Tech.*, *Vol. LT-5*, pp. 1103–1110 (1987).

4. R. Olshansky and V.A. Lanzisera, *Electron. Lett. 23*, pp. 1196–1198 (1987).

5. R. Olshansky and V.A. Lanzisera, *OFC '89*, Houston, WC2 (1989).

6. R. Olshansky and V.A. Lanzisera, *Optoelectronics — Devices and Technologies 3* (2), pp. 143–153 (1988).

7. W. Way, C.E. Zahl, S.G. Menocal, N.K. Cheung, and T.P. Lee, *OFC '89*, Houston, WC3 (1989).

8. S.P. Walker, *Electron. Lett. 25*, pp. 1023–1024, 1989.

9. T.E. Darcie, R.S. Tucker, and G.J. Sullivan, *Electron. Lett. 21*, pp. 665–666, 1985.

10. R. Olshansky, et al., *IEEE JQE, Vol. QE-23*, No. 9, pp. 1410–1418 (1987).

11. Private conversation with J. Stilwell of CATEL.

Audio Compression Put To The Test

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ABSTRACT

Transmission of digital audio almost always requires some method of compression to reduce the required bandwidth and to supply larger numbers of services. Many audio compression systems exist, based on very different principles. The tests normally associated with measuring audio performance seldom challenge the best audio compression techniques. In this paper, one audio compression system, SuperSound, is studied. Tests are performed that are intended to evaluate its performance when dealing with complex signals. These tests provide a more realistic measure of actual audio performance than simple test tones. The "compression noise" that is measured quantifies the degree to which the original signal is altered by the compression/decompression process.

INTRODUCTION

Simple test tones consisting of one or two discrete frequencies have been used for many years to characterize audio systems. Tests such as harmonic distortion, intermodulation distortion and signal to noise ratio were sufficient to characterize linear audio systems. Linear systems use no processing other than gain and frequency equalization, and the testing was thus uncomplicated.

The inadequacy of such tests first became widely apparent with the common use of noise reduction techniques. Signal to noise ratio (SNR) measurement on a noisy system using noise reduction provides a good example. In a SNR measurement, a reference tone is first applied to the system. Often, when the test signal is present, a large amount of noise can be seen to accompany the tone. Yet when the reference tone is removed, the system reduces the gain, and with it the noise we are trying to measure. This familiar "noise pumping" that such systems cause defies simple SNR tests. Though often audible and sometimes objectionable, in most cases the excess noise is adequately masked by the presence of audio. It is the function of the noise reduction system to reduce the noise when the signal is absent, and thus improve the apparent SNR. It is usually preferable to the constant hiss that is the alternative.

But today the standards have been raised. Compact disks (CD) and uncompressed digital audio tape (DAT) have accustomed the audio consumer to noise floors and reproduction quality that are limited primarily by the studio recording equipment. Now, into this setting comes a host of new products. These products promise inexpensive recording of digital audio on cheap tapes and recordable disks. They enable broadcast transmission over terrestrial airwaves, from satellites and over CATV cable. They all promise "compact disk quality audio" and they all use some form of audio compression.

Audio compression is usually necessary to reduce the transmission bandwidth or storage requirements of the signal. We all know that "you don't get something for nothing". We feel that there must be some performance cost to audio compression. But in fact, it will be shown below that there is a large amount of wasted dynamic range at higher audio frequencies. To the extent that compression is achieved solely by taking advantage of such waste or redundancy in the signal, it is possible to perfectly reproduce the original signal and "pay no price". Computer data compression systems that achieve this goal are referred to a "lossless" compression systems. Unfortunately, this alone seldom results in sufficient data reduction for audio systems.

The compression system under study in this paper, SuperSound [1], is a system that does not attempt large amounts of data compression. It attempts to gain most of its compression by exploiting the waste and redundancy in the audio signal. As will be seen below, it comes relatively close to achieving lossless compression when processing real music signals.

Compression systems differ widely. They differ both in the amount of compression achieved, as well as in the price paid in audio quality. Compression systems can be designed that perform very well when subjected to simple test tones, yet can generate high "noise pumping", distortions and other less familiar artifacts when subjected to full loading. A method is needed to quantify signal degradations that occur in the presence of the signal. Ideally, a test could be devised that would yield a single number or "figure of merit". It could be used to characterize the extent to which a compression system achieves its compression at the cost of imperfectly reproducing the original signal at its output. Test signals should be used that are both representative of actual music, and that also fully load the system.

COMPRESSION NOISE

Fortunately, digital audio techniques have not only raised the standards of expected sound quality. They have also provided more sophisticated test methods than the simple test tones. Using digital techniques, it is now possible to take a digitized signal, compress it, then decompress it, and compare each sample of the processed signal with the corresponding sample of the original signal. Differences between the processed signal and its original represent an error due to the compression system. This error is the "noise" or degradation that the compression system generates. It is the cost of compression.

Compression noise is similar to the quantization noise that occurs in analog to digital conversion. As with quantization noise, compression noise can also be treated and measured in the same ways that we now study analog noise. Like other audio noise, the human ear is more sensitive to compression noise at some frequencies than others. Thus it is appropriate to use weighting when integrating the noise over the audio spectrum in noise measurement. The same noise weighting curves, such as "A" weighting [2], or CCIR/ARM [3] are appropriate.

In measuring compression noise using digital techniques, the test signals can be simple tones, as in analog measurements. More importantly, the test signal can be any audio waveform, from actual music to broadband noise. The system degradations can be very accurately measured while the system is fully loaded, without removing the test signal.

TEST SIGNALS

Though it is interesting to test a system loaded with actual music using the method described above, most music is quite variable in time. In actual measurements of the compression noise, this can result in readings that fluctuate with the instantaneous power in the music waveform. Much of the time, such a test would not present a sufficient challenge to the system. Also, the single-number "figure of merit" we seek would be a function of time and type of music. It would be highly subject to interpretation by the tester. These problems were overcome in two ways. The first was the use of a broadband noise signal with a spectrum that is closely equivalent to that of music. The second was the use of a "composite music signal".

In arriving at a useful test signal, first the typical spectrum of music was studied. The line output of a compact disk player was connected through an audio attenuator to the input of an HP 3588A spectrum analyzer. The top reference line was set to the level of a full-scale sinusoid. The full spectrum was swept every 614 milliseconds. A resolution bandwidth of 150 Hz was used. Peak hold was used to hold the highest level encountered at any given frequency. Entire disks [4][5][6][7] were played into the analyzer and the spectra were plotted as shown in Figures 1 through 4. Though the type of music varied widely, the spectra are quite similar. Note that in each case the spectrum rolls off with increased frequency. As mentioned earlier, much of the dynamic range available from 16 bit PCM is wasted at higher audio frequencies. The music simply does not demand it. This is good news for speaker manufacturers in that the small voice coils of tweeters never need to handle the same power levels encountered by bass drivers. Certainly it is possible with electronic music synthesizers to generate high levels at high frequencies, but in reality this is quite unpleasant. It occurs about as often as blown tweeters!

Figure 5 shows the spectrum of USASI (United States of America Standards Institute) noise. This noise spectrum consists of white noise filtered to peak at approximately 200 Hz, with a 6 dB per octave falloff below 100 Hz and above 320 Hz. it is available on the Audio Precision (AP) System One audio test equipment. It is also available on the National Association of Broadcasters (NAB) Broadcast and Audio System Test CD. It is used by audio broadcast manufacturers to simulate unprocessed audio program material. Note the similarities between the spectra of actual

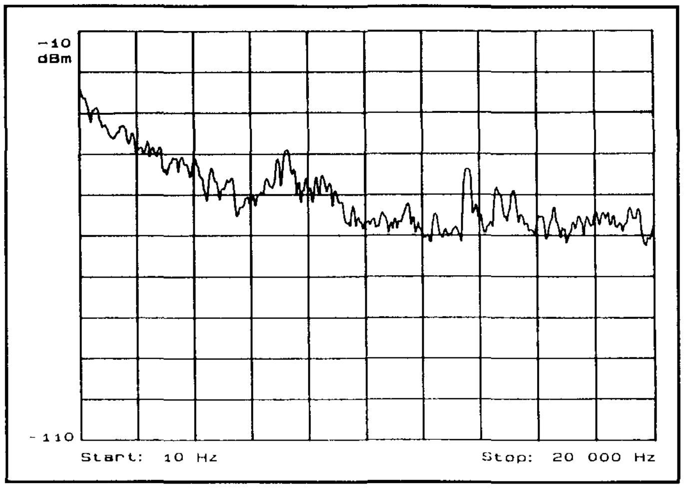

Figure 1 Spectrum of country music disk (full disk represented)

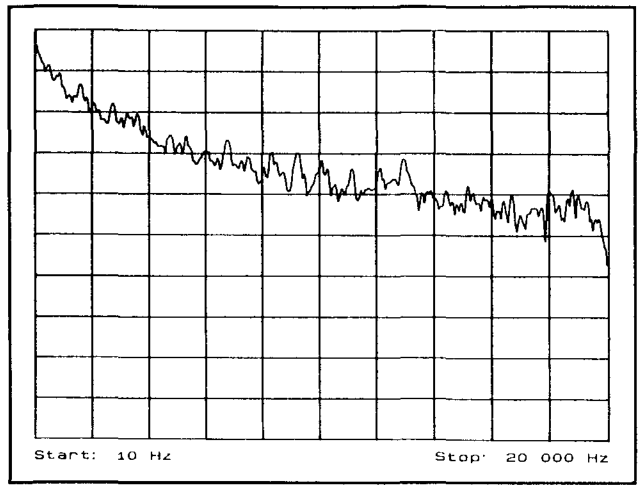

Figure 2 Spectrum of classical music disk (full disk)

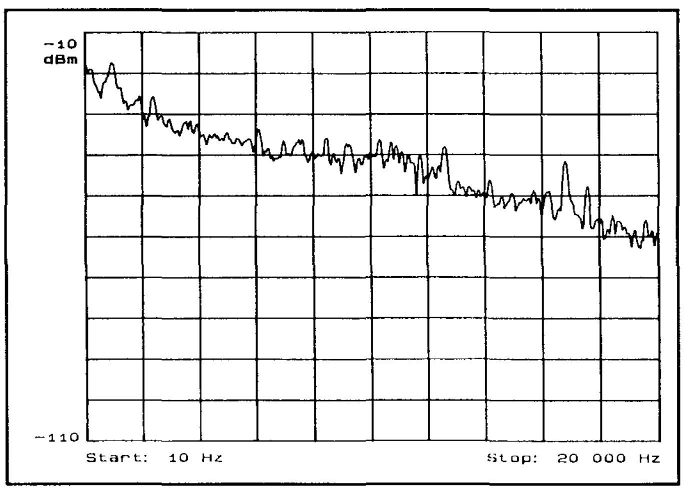

Figure 3 Spectrum of rock/Latin music disk (full disk)

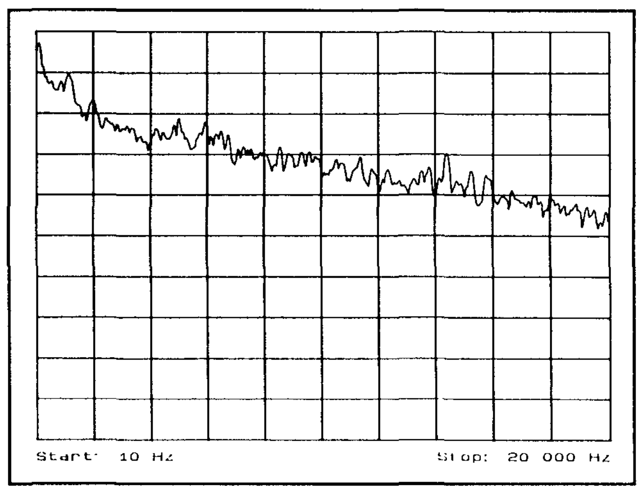

Figure 4 Spectrum of electric jazz music disk (full disk)

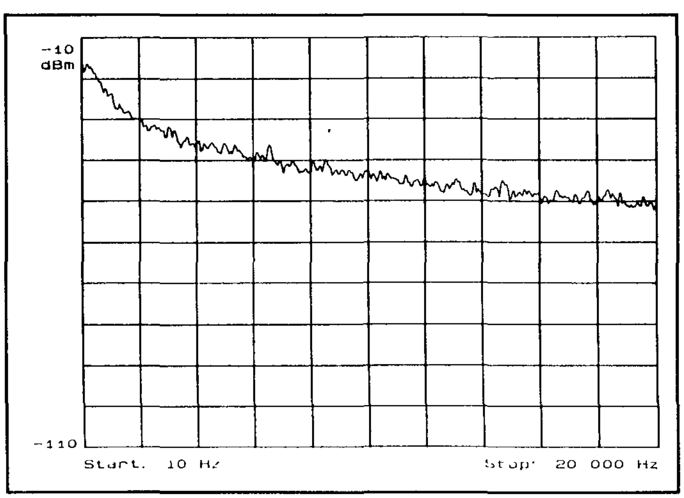

Figure 5 Spectrum of USASI noise (10 seconds)

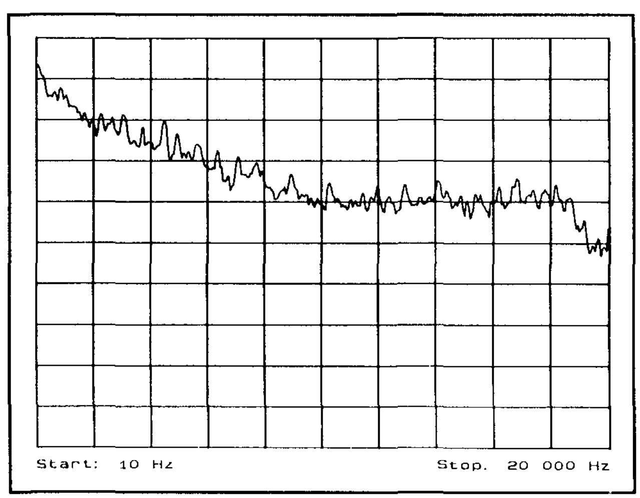

Figure 6 Spectrum of composite music mix (10 seconds)

music in Figures 1-4, and the USASI noise of Figure 5. The USASI noise allows loading the system with a realistic test signal, comparable to the peak music readings through the music spectrum, while providing a fairly constant power yielding noise measurements that fluctuate little. A time domain plot of USASI noise is shown in Figure 7.

Lest there be any concerns about the suitability of the noise waveform for audio testing, a second test waveform was generated. This time a "composite music signal" was created by mixing selected ten second sections [8][9][10][11] of country music, classical music, rock/Latin music, and electric jazz music from the sources used in Figures 1-4. The spectrum of this mix is shown in Figure 6. The ten second sections were chosen to get the largest number of instruments playing at the same time. The mix contained a full symphony orchestra, several synthesizers, two kinds of electric drums, "fuzzed" electric guitar, Latin percussion and brass, one female voice and two male voices all simultaneously. Well over 100 acoustic and electric instruments were playing at once. All were digitally summed without any scaling back of the levels. None of the highest peaks occurred at the same instants, and thus no clipping occurred. Peaks reached 92 % of full scale. The average level, however, did increase markedly and was sufficiently constant to yield steady readings. Though somewhat unrealistic, this signal provided a good "worst case torture" test. A time domain plot of this mix is shown in Figure 8.

TESTS

USASI noise from an AP System One was recorded onto an Apple MacIntosh IIcx computer hard disk using Sound Designer II and Sound Tools software and hardware. This sound file was then SuperSound compressed and decompressed. Figure 9 shows a portion of the time domain USASI waveform before SuperSound processing. Figure 10 shows the same after Supersound compression and decompression. The horizontal axis is calibrated by sample number. The vertical axis is calibrated with decimal quantization value, where 16 bits corresponds to a total of 65,360 values, or +/- 32,768 values. Display resolution is much poorer than any actual differences between the waveforms and thus it is not possible to compare the waveforms by inspection. Each sample of the original waveform was sign-inverted and added to each corresponding sample of the processed waveform. The resulting difference is the compression noise. It was stored in a sound file and is shown in Figure 11. Note that the noise is not visible on the same vertical scales as used in Figures 9 and 10. Thus the vertical scale in Figure 11 was expanded by a factor of 65.

All of the above processing was performed in the digital domain. The sound data was manipulated on the computer, and thus the difference calculations were mathematically exact. The samples that were compared represent the same sampling instants in time with zero time delay between them and zero amplitude calibration errors. This is only possible with fully digital processing, and on compression systems that accept digital input and produce digital output.

To quantify the compression noise, the file was passed digitally via the standard Sony Philips Digital Interface Format (SPDIF) [12] from the MacIntosh computer to the AP System One test equipment. Using fully digital signal processing, the noise was integrated over the entire audio spectrum and A-weighted. This resulted in a measurement of -87 dBFS (dB with respect to full scale).

The same procedure was repeated with the "worst case" composite music mix. Figure 12 shows a portion of the original time domain waveform. Figure 13 shows the SuperSound compressed and decompressed waveform. Figure 14 shows the noise using the expanded vertical scale. Again, the noise was digitally integrated and A-weighted and resulted in -80 dBFS, very respectable for such torture.
This level of noise is comparable to what the best studios can produce, and is certainly inaudible when such a high level signal is present.

CONCLUSION

It is possible to design compression systems that reduce the data rate or storage requirements to very low levels. Normally, larger amounts of compression results in more compression noise. But a measurement of compression noise does not indicate how audible the degradation to the audio will be. Many systems exploit the properties of the human auditory system to very effectively mask or conceal the compression noise. Though the concealment is not perfect, it does raise an interesting question. If the concealment were so good that the noise was completely inaudible, what value would a measurement of compression noise have?

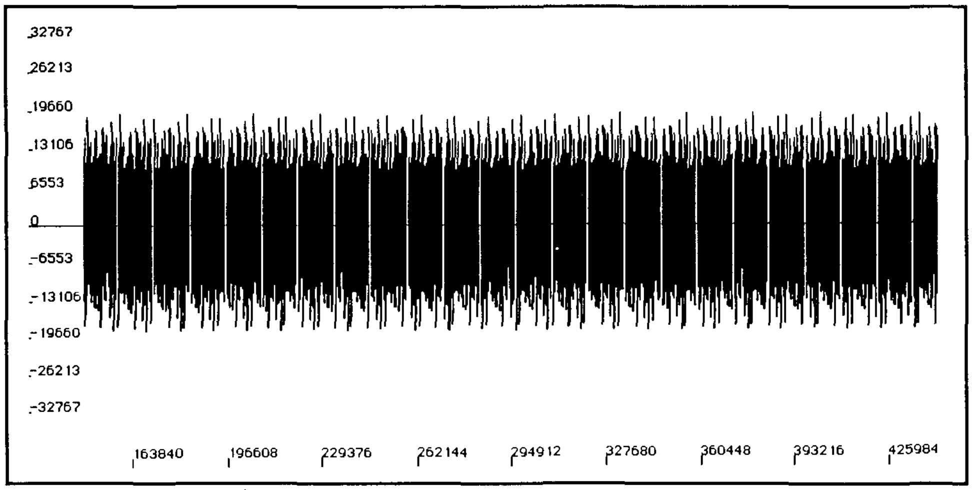

Figure 7 Time domain plot of USASI Noise
Approximately 7 seconds shown

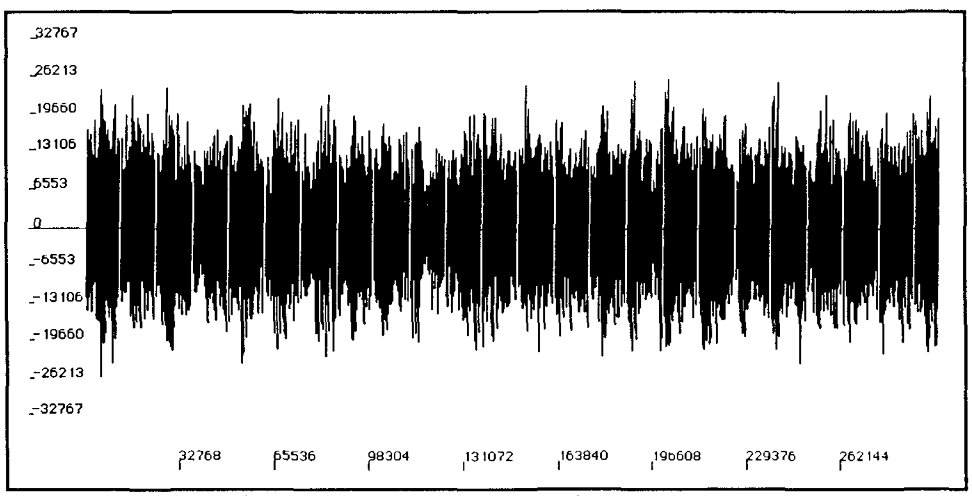

Figure 8 Time domain plot of composite music mix
Approximately 7 seconds shown

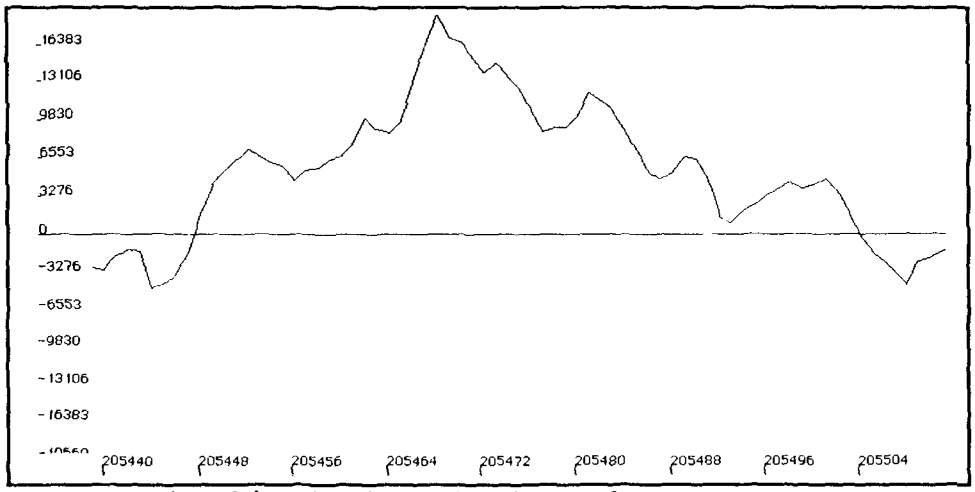

Figure 9 Section of time domain USASI noise waveform before SuperSound processing

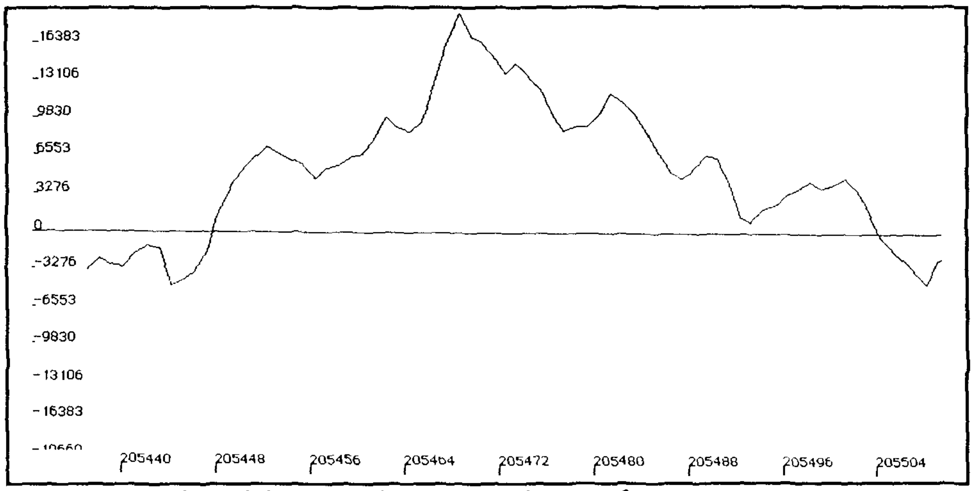

Figure 10 Section of time domain USASI noise waveform after SuperSound processing

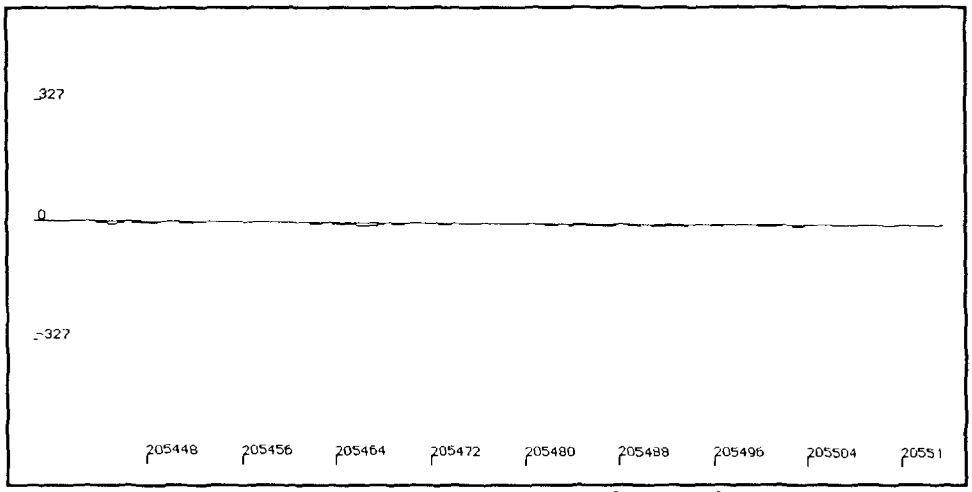

Figure 11 Compression noise for same USASI waveform section Vertical scale expanded 65x

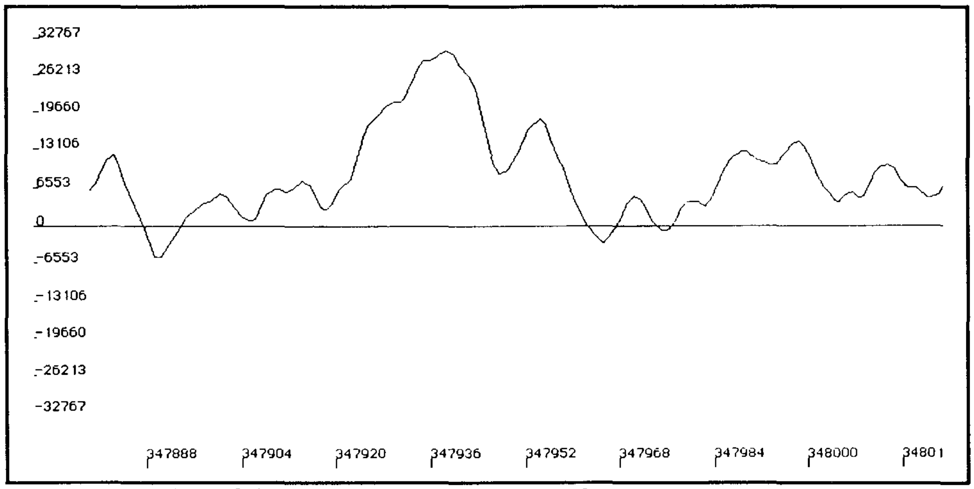

Figure 12 Section of time domain music mix waveform before SuperSound processing

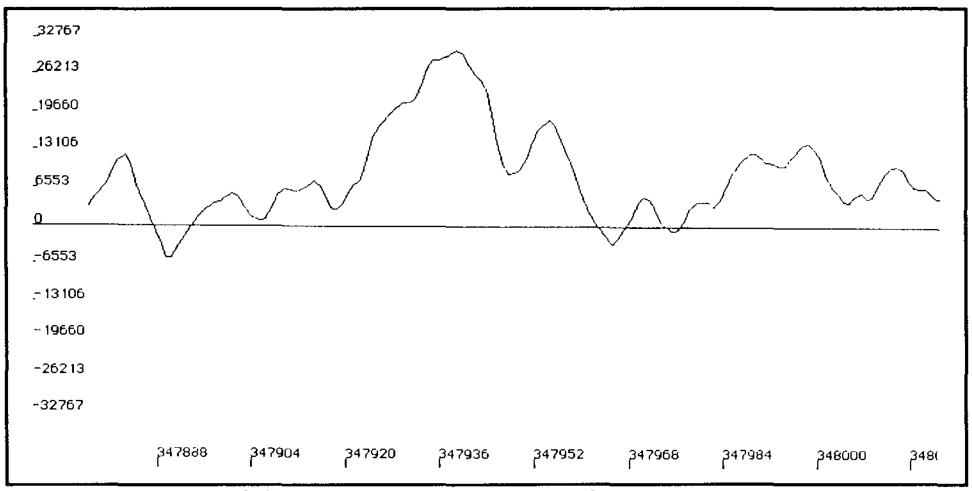

Figure 13 Section of time domain music mix waveform after SuperSound processing

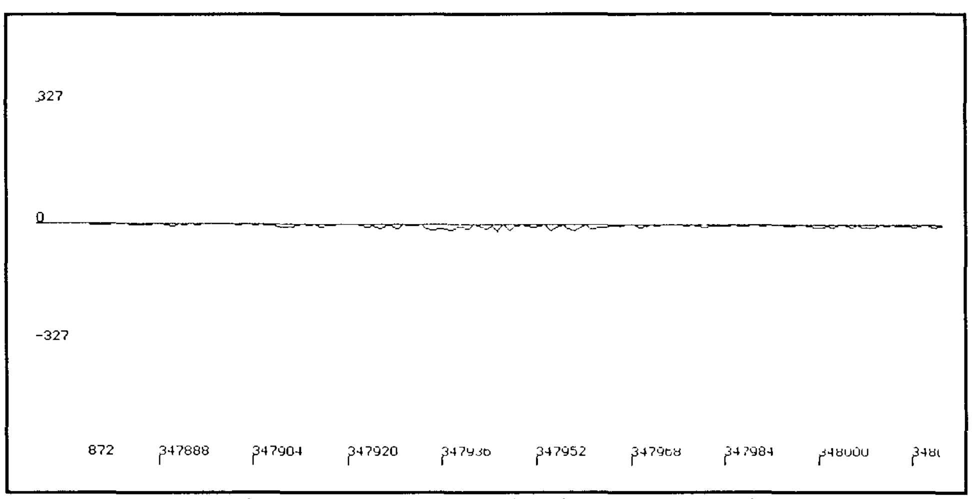

Figure 14 Compression noise for same music mix waveform section Vertical scale expanded 65x

The problem is that no one yet knows the effects of cascaded audio compression. As mentioned earlier, many products are being conceived and introduced that use audio compression. Each of these generates compression noise. The noise and artifacts generated by a transmission system may be undetectable to the human ear, but may be very real to the compression system in a recorder. In the recorder, the noise and artifacts will interact with the second compression system in unpredictable ways. The compounded artifacts generated may be much more objectionable than either of the two systems alone.

Short of not compressing the audio, the safest approach would be to use as little compression as possible, and thereby generate the minimum amount of noise.

We normally consider two pieces of digital audio equipment compatible if they use 16 bit PCM, share sampling rates, and interface using either SPDIF (consumer equipment) or AES/EBU (professional) [13] standards. But compatibility may mean more than that. Compatibility may be determined by the specific compression system used, and the degree to which compression is achieved by altering the individual samples as opposed to exploiting redundancy and waste in the data. Compression noise provides a measure of the degree to which the original signal is altered. As such, it should provide a good tool for evaluating the compatibility between audio systems using compression.

ACKNOWLEDGEMENT

I gratefully acknowledge Jay McMullan who wrote much of the software used in this work and whose help was indispensable.

REFERENCES

[1] Quinn, S.B. Jr., Frederiksen, J.
"A Compressed Digital Audio Transmission System," Paper presented at AES 7th International Conference, May 14-17, 1989.

[2] "Sound System Equipment," IEC Standard 268-1 (1968).

[3] "CCIR/ARM: A Practical Noise-Measurement Method," Journal of the Audio Engineering Society, Vol.27, No.3, March 1979.

[4] "Pancho & Lefty," Merle Haggard and Willie Nelson, Epic EK 37958

[5] Telarc Sampler Vol 1, used 17 classical selections by various composers and orchestras, CD-8001

[6] "Primitive Love," Miami Sound Machine, Epic EK 40131

[7] "The Chick Corea Electric Band," Chick Corea et. al. GRP D-9535

[8] ref. [2] Cut: "All the Soft Places to Fall," time: 1:40-1:50

[9] ref. [3] Cut: R. Wagner, "Riezi Overture," time: 0:05-0:15

[10] ref. [4] Cut: "Conga," time: 3:39-3:49

[11] ref. [5] Cut: "Elektric City," time: 2:43-2:53

[12] "Digital Audio Interface," EIAJ Standard CP-340 (1987).

[13] Audio Engineering Society Standard AES3-1985, also ANSI S4.40-1985.

AUTOMATIC LEVEL CONTROL ISSUES IN AM FIBER SYSTEMS

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ABSTRACT

AM fiber optic systems are playing an increasingly important role in the delivery of high quality signals in all types of CATV plant architectures. This paper discusses technical issues in the set-up, operation and maintenance of this equipment which are critical to maintaining system performance. Discussion is also given to the effects of various level control methods on the initial and ongoing system performance capabilities, particularly in light of real world headend and distribution plant operation. Technical and operational trade-offs are contrasted and recommendations are made for different areas of the installation. It is hoped that this paper will clarify some of these issues so that operators can make informed decisions to ensure stable and reliable operation of fiber optic systems, particularly as their use expands in CATV architectures.

INTRODUCTION

Amplitude modulation (AM) based fiber optic transmission systems have become increasingly attractive for transporting multichannel CATV video signals over substantial distances. Although the performance requirements and equipment types vary with different network architectures and applications, the basic means of providing signal transport is the same.

In typical CATV systems, as shown in Figure 1, individual frequency channels from multiple headend modulators feed combining networks. One or more combining networks gather the channels into a single band or multiple bands of frequencies to be transported by the distribution plant. In AM fiber optic systems these bands of channels also provide a modulation input to the laser transmitter(s). The AM laser transmitter conditions the composite headend signals to provide a signal which modulates the semiconductor laser bias current. This results in an intensity (power) modulated optical signal out of the laser. The optical signal is carried by a low loss single mode optical fiber to an optical receiver. A photodetector in the optoelectronic receiver converts the optical signal into an electrical signal current. The electrical signal is a replica of the headend composite channel band used to modulate the transmitter. The received composite signal is amplified and conditioned and fed to the distribution plant.

An important advantage of AM fiber systems is that the same multichannel NTSC, PAL, or SECAM signal format is maintained through the system. No format conversion electronics are required at either end of the optical link. This makes the AM fiber optic system "friendly" to the CATV system tie-in points. Because of this advantage, AM fiber optic systems generally require less equipment space in the installation. An AM system is also less costly to install, particularly on a per channel basis, than either FM or digital systems.

The single mode optical fiber used in AM fiber systems possesses attenuation characteristics which change extremely little with temperature variations, unlike coaxial cable. In most current AM fiber architectures, little compensation for the optical fiber response is required. However the carrier-to-noise ratio (CNR) and distortion performance (CTB,CSO) of AM fiber systems is tied directly to the relative level of the modulating multichannel carriers. Because of this, the issue of signal level control is important throughout initial equipment set-up, ongoing operation, and system maintenance. The practical, operational, and technical tradeoffs of level related performance issues are discussed in the following paragraphs.

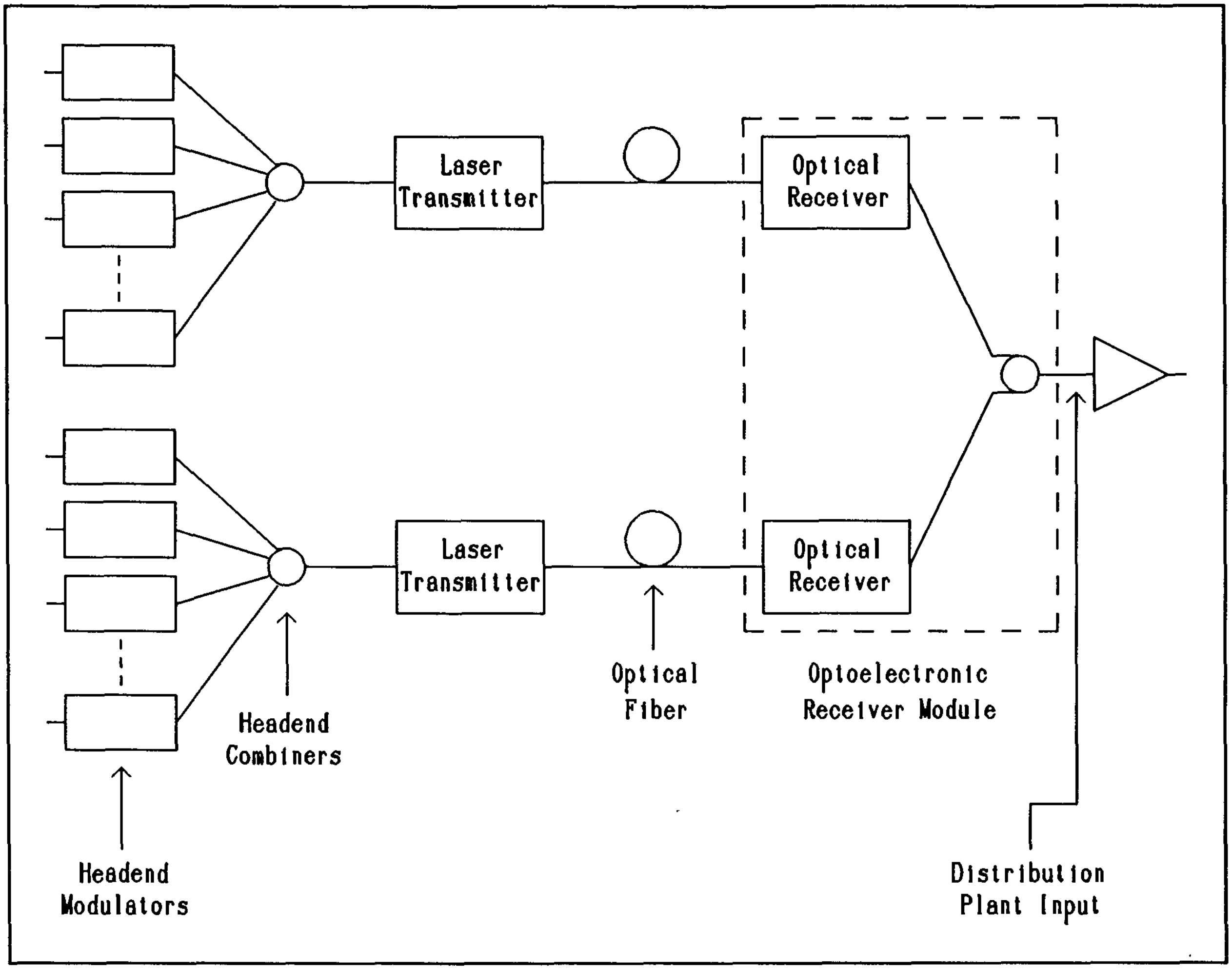

Figure 1. Typical Dual Band CATV AM Fiber Optic System.

LEVEL RELATED PERFORMANCE ISSUES IN HEADEND FEED AND AM TRANSMITTER

The composite FDM output from the headend which feeds the AM laser transmitter(s) can exhibit significant variations due to several factors. Since this headend composite output directly modulates the laser transmitter, variations in the modulating signal must be examined to understand their impact on overall system performance.

Figure 2 is a simplified block diagram of a laser transmitter. The signals modulating an AM laser have certain ideal requirements. The laser used in the transmitter exhibits optimum performance for a given application when operated at a specific composite modulation index. The RF drive level per channel modulating the laser is the determining factor in the modulation index of the laser. Ideally, it is imperative to precisely maintain the laser's modulation index at its optimum value to ensure specified system CNR and intermodulation distortion performance. If the laser modulation index is too large, the CNR performance improves, but the distortion performance is compromised. On the other hand, if the laser modulation index is too small, the distortion performance improves, but the CNR performance degrades.

In general, a larger composite modulation index is required to meet higher system CNR specifications. However, a maximum modulation index exists for each laser at which point laser distortion performance begins to deteriorate rapidly due to signal clipping. In high CNR performance systems, the laser is generally operating at or near its maximum modulation index. It is therefore critical that the modulation index and hence composite RF drive power be held constant.

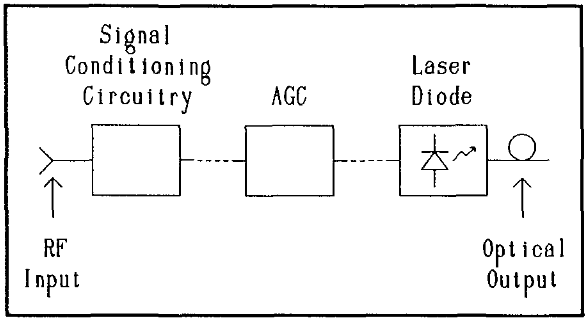

Figure 2. Simplified AM Laser Transmitter Block Diagram.

Channel loading also has an effect on laser modulation index. As channel loading increases, the laser composite modulation index increases, and the intermodulation distortion performance degrades. Thus, when increasing channel loading, the RF drive level per channel should decrease for every channel in order to preserve the composite modulation index and hence intermodulation distortion performance.

In practice, a laser transmitter is not modulated by ideal signals. The headend output RF level varies due to different factors. The addition or removal of a coupler, tap, or other equipment in the headend wiring scheme causes a change in the resultant headend RF output level. The headend RF output level also varies slightly with time, temperature, regular maintenance, and adjustment. Modulators may be added to or removed from the headend, thereby changing the transmitter's channel loading and the laser's composite modulation index. This changes the laser transmitter's distortion performance due to the change in laser modulation index.

The possibility of an RF overdrive condition from the headend feed must be addressed also. During headend equipment installation, or even during regular maintenance, the headend RF output level may inadvertently increase to an extremely high level relative to the specified RF drive level of the transmitter. This level may cause performance degradation, and may even irreversibly damage the laser. This is an extreme example of non-ideal RF signals modulating the laser.

Many situations exist in which the RF signals modulating the AM laser transmitter are not ideal. Therefore, appropriate circuitry may be incorporated into the transmitter to reduce the undesirable variations in RF level. An automatic gain control circuit (AGC) can be used to compensate for variations in RF level. The AGC circuit monitors the RF level feeding the transmitter and adjusts the gain through the transmitter in order to supply the laser with the appropriate RF drive level. The AGC has a range of input RF levels for which it can maintain a specified RF drive level to the laser. In this application, two alternative AGC implementations could be used.

One implementation is a pilot carrier type of AGC. This circuit monitors the RF level of only one or two pilot channels and adjusts the level of the entire spectrum of channels to compensate for the change in pilot levels. This type of AGC is useful if all channels vary in the same way and by the same amount. However, if additional modulators are brought online in the headend resulting in an increased channel loading to the laser, or if individual channels vary randomly due to time or temperature, the AGC does not change the RF drive level to the laser unless the RF level of the pilot(s) has changed. As a result, the composite modulation index of the laser changes due to the increase in channel loading or individual channel level variation, and the laser's distortion performance will in general degrade. Another disadvantage of a pilot carrier type AGC is that it cannot effectively be used to provide RF overdrive protection since it only adjusts the laser RF drive level if the pilot signal level changes.

A composite power AGC is another variation of the AGC circuit. This type of AGC monitors the composite RF power modulating the transmitter and adjusts the level of the entire spectrum of channels by the same amount to maintain a constant composite power. The advantage of a composite power AGC for an AM laser transmitter is that it maintains a constant laser modulation index with headend RF level variations and changes in channel loading. Therefore, optimum transmitter distortion performance is preserved. In addition, the composite power AGC can be used to guard against an RF overdrive condition. If the level from one modulator increases by a large amount, the composite power AGC detects the increase in transmitter input composite power and adjusts the RF level modulating the laser.

Considering the variations that can occur in the RF drive to the transmitter, if an AGC circuit is incorporated in the AM laser transmitter, the composite power type AGC is preferred. This type of AGC is effective in maintaining specified intermodulation distortion performance, and in preventing RF overdrive related damage to the laser.

LEVEL RELATED PERFORMANCE ISSUES IN THE AM OPTICAL RECEIVER

Ideally the optical signal illuminating the receiver's photodiode has a constant average intensity, constant modulation index, and single wavelength. The intensity of the received optical signal and the modulation index determine the RF gain required in the receiver for best performance. A single wavelength optical signal is less affected by fiber dispersion and other nonlinear effects in the optical plant.

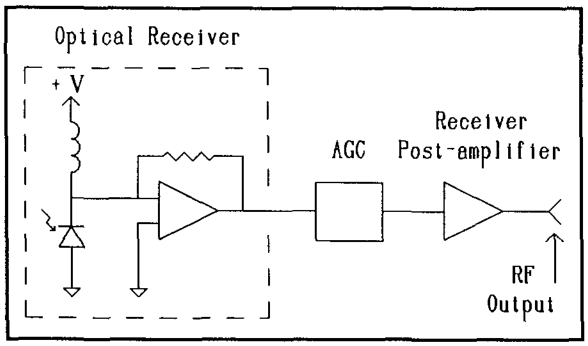

Figure 3. Simplified AM Optoelectronic Receiver Block Diagram.

In practice the quality of the received optical signal is affected by the fiber plant and the optical transmitter. The average intensity of the received optical signal may change due to maintenance or repair of the fiber plant. Additional splices to repair fiber breaks, replacing optical connectors with splices or vice-versa, or entirely rerouting an optical path will all affect the optical link loss. As discussed earlier several aspects of the transmitter design and RF signal source will also affect the optical signal. There may be variations of the laser diode output power due to aging or temperature. The composite modulation index may change as a result of additions or deletions of channels, variation of level or other changes to the laser drive signal.

Receiver performance as measured by carrier-to-noise ratio (CNR) and distortions, composite triple beat (CTB) and composite second order (CSO), is generally degraded by a non-ideal optical signal. All fixed gain optoelectronic receivers provide a specific RF output signal level and distortion performance for a specific optical receive signal. As the optical input power or modulation index increases, the receiver's CNR generally increases, but the receiver's contribution to the system distortion level also increases. Conversely, with decreasing optical input power or modulation index, the receiver's CNR decreases but its contribution to system distortion also decreases. Variations in received optical power from that expected can be due to several factors. The number and quality of connectors and splices used in the field installation may differ from the originally specified plan resulting in a different optical loss. If an OTDR measurement used to determine the optical loss is inaccurate then again the optical power will differ from that expected. Finally, fiber plant maintenance or repair can also alter the link loss.

If variations in optical loss occur, the optoelectronic receiver performance may be affected. If the optical loss is greater than expected the received optical power is lower than expected. Lower than expected received optical power results in a reduced RF output from the photodetector and optoelectronic receiver. Consequently the input level to the receiver post-amplifier is lower. This condition increases the significance of the noise contribution of the receiver post-amplifier to the system CNR. The final result may be a degradation in system CNR. If the optical loss is less than expected, then the received optical power is higher than expected. This results in an increased RF output level from the photodetector and optoelectronic receiver and generally improves the system CNR. However, with the optoelectronic receiver operating at a higher output level its contribution to system distortion is greater. The post-amplifier is also operating at a higher level and may contribute further to the system distortion. Consequently there may be a degradation in system distortion performance.

Changes in the modulation index can also affect the performance of a receiver. If the modulation index increases and all other system parameters remain constant, the RF output level from the photodetector and optoelectronic receiver increases. An increase in this output level causes an increase in the receiver and post-amplifier's contributions to system distortion as discussed earlier. Conversely, if the modulation index decreases and all other system parameters remain constant, the RF output level from the photodetector and optoelectronic receiver decreases. A decrease in this output level causes a degradation in system CNR by lowering signal level while the noise level remains unchanged.

Receiver performance is also affected by the channel loading. When a transmitter without AGC is used and the channel loading changes from the original channel plan, receiver distortion performance also changes. More channels create more beats and hence greater distortion but the receiver RF output level remains the same. For example, an increase in channel loading from 40 channels to 60 channels increases the receiver's contribution to system distortion by approximately 1.8 dB for CSO and 3.5 dB for CTB.

When a transmitter with a composite power AGC is used and the channel loading changes from the original plan, receiver output level and CNR performance may be affected. An increase in channel loading from 40 channels to 60 channels causes the transmitter AGC to reduce the RF input level to the laser by approximately 1.8 dB. This in turn, results in a 1.8 dB decrease in RF output level from the optoelectronic receiver and a corresponding decrease in receiver CNR. Distortion performance however remains unchanged.

In order to compensate for variations in optical loss, modulation index and channel loading, an AGC circuit may be implemented in the optoelectronic receiver. An AGC incorporated in the optoelectronic receiver provides automatic gain control of the receiver's RF level. Several methods can be used to implement an AGC. Each method has advantages and disadvantages.

A composite RF power AGC maintains a constant RF power at the output of the optoelectronic receiver. This method is advantageous from a distortion standpoint because regardless of variations in received optical power, modulation index, or channel loading, the composite RF power at the receiver output will remain constant. For a given input optical signal, the receiver generated distortion level is primarily a function of the RF power. Therefore distortion due to the receiver will not change. However, this method will not yield optimum receiver noise performance under all conditions. A variation in channel loading causes a change in the RF output level of the receiver because the AGC maintains a constant composite RF power at the receiver output. If the channel loading increases, the RF signal level to the distribution plant decreases and system CNR degrades.

A single carrier pilot type AGC maintains a constant RF channel level at the output of the optoelectronic receiver. The primary advantages of this method are a constant RF output level and a constant contribution to system CNR by the receiver amplifier and post-amplifier. The RF channel output level of the receiver remains constant with variations in received optical power, modulation index, and channel loading. In a multiple band system this AGC approach maintains consistent RF channel level between bands. The disadvantage of a pilot carrier type AGC is the fact that the receiver amplifier's distortion changes with variations in channel loading. If the channel loading increases, distortion due to the receiver amplifier increases and thus, system distortion may increase. However, when using a transmitter with composite power AGC, distortion remains unchanged.

In today's high performance AM fiber optic systems, high CNR is usually the performance specification most difficult to achieve. Proper circuit design can minimize the impact of the receiver amplifier and post-amplifier on system distortion performance. Consequently, in a receiver incorporating AGC, a pilot carrier type AGC is most desirable for use in the optoelectronic receiver. Careful implementation of the pilot carrier type AGC results in optimum receiver noise performance.

OVERALL SYSTEM REQUIREMENTS FOR LEVEL CONTROL

The key to the use of any technology is in assuring that it can provide the expected level of performance, consistency, and reliability necessary for the application. To date most applications of AM fiber optic systems have been in CATV systems where the distribution plant is upgraded, expanded or replaced. In current AM fiber optic system architectures each fiber link serves a large number of subscribers and directly affects the overall signal quality provided. Most CATV operators strive to extend plant service area, expand bandwidth, and improve signal quality and consistency. To achieve these goals the overall system performance must be consistent.

The bottom line of acceptable plant performance can be characterized by signal level stability, carrier-to-noise ratio, distortion performance, and consistent service availability. In the paragraphs above, the impact of various AM fiber signal level control implementations on overall system performance issues was discussed. A summary of the comparative advantages of each implementation is given in Table 1.

The importance of headend signal level maintenance on the performance of the CATV system, including the AM fiber link, is critical and stability can be augmented by the use of a composite power sensing AGC in the AM laser transmitter. At the optoelectronic receiver, control of signal levels feeding the distribution plant is also important. In single or multiple band AM systems the effects of temperature variations on the fiber plant and equipment can be minimized with proper equipment design. However, if an AGC design is to be used in the optoelectronic receiver, signal levels can best be stabilized versus optical loss variations, channel loading variations, and equipment changes by the use of a pilot carrier AGC.

AM fiber optic based CATV distribution systems have the potential to bring increased signal quality, additional channels, and increased service reliability to the CATV operator. The promise of this relatively new technology for CATV systems can be realized through comprehensive and careful development of system architectures and judicious design of equipment.

Table 1. System Performance Characteristics using Transmitter and Receiver AGC Combinations for Fixed Channel Loading.

Laser Transmitter AGC	Optoelectronic Receiver AGC	Consistent System Distortion	Consistent System CNR	Constant System RF Output Level	Effective Laser RF Overdrive Protection
pilot carrier	composite power	no	depends	no	no
composite power	composite power	yes	depends	no	yes
pilot carrier	pilot carrier	no	depends	yes	no
composite power	pilot carrier	yes	depends, but best compromise	yes	yes

CABLE AND THE CONSUMER ELECTRONICS INDUSTRY

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Cable Television Laboratories, Inc.

Abstract

Cable Television Laboratories, Inc. (CableLabs) has undertaken several very important strategies with the consumer electronics industry in an attempt to solve long-standing interface problems between the cable delivery system and the consumer display and storage equipment, and to coordinate future developments to enhance forward compatibility. Building on the foundation of relationships forged in the EIA/NCTA Joint Engineering Committee, CableLabs has taken the interface issues from the level of engineers to the industry executives and product planners.

Historically, the relationship between the cable and consumer electronics industries has not demonstrated much in the way of cooperation or communication. This has resulted frequently in unhappy customers and increased operating costs. The formation of the EIA/NCTA Joint Engineering Committee some years ago provided the first avenue of coordination between the two distinct industries. Issues and standards important to cable addressed by that committee have included the following :

EIA/ANSI-563: The Multiport Standard

IS-6: The Channelization Standard

IS-23: Tuner Design and Electromagnetic Compatibility Guidelines

Advanced Program Guide

This committee has also addressed numerous other compatibility and interface issues on an *ad hoc* basis. Most importantly, the engineering organizations within each industry have had the opportunity to discuss their concerns and goals. A brief status report of the above subjects follows. A subsequent treatment of CableLabs activities will show relevancy to the general issues.

EIA/ANSI-563 Multiport

The cable industry has continuously voiced a long-term strategic need for the capability to insert signal processing of various kinds between the tuner and display portions of the television receiver. Originally, the thought was to place the descrambling function at this point, thus obviating the higher capital requirements and the public dissatisfaction associated with the use of a set-top converter. This is still a viable concept for our industry, but offers little value to the consumer electronics manufacturer and is probably counter productive to the manufacturers of standard converters.

A deeper look at the strategic importance of having a signal breakout at that point in the receiver revealed several other important benefits, not just to cable but to the consumer electronics industry as well. For instance:

- Baseband interconnect from other consumer electronics hardware, such as VCRs and disk units.

- Insertion of after-market noise cancellation, decompression, or other signal enhancement devices.

- Interface to CE-Bus or other such home wiring systems.

- Applique closed captioning decoder as per recent federal legislation.

- Ancillary processors to facilitate the NTSC/HDTV transition period.

One large manufacturer has pre-production prototype receivers which were tested during January 1991 at CableLabs to determine compatibility with Multiport decoders. In spite of a considerable number of launch problems,there is still hope for Multiport.

IS-6 Channelization Standard

Cable and consumer electronics representatives have agreed upon an algorithm for determining channelization in excess of that previously set. It is anticipated that this interim standard will be canonized in the near future.

IS-23 Tuner Design Standards

This is the interim standard which has stirred the most controversy between the two industries. Of particular importance herein is a call by cable for a standard for direct pickup interference. At the present time, the United States has no standard for television receiver susceptibility above 30 MHz. Between 2 and 30 MHz, the so-called "Goldwater legislation" sets a standard of no visible degradation in the picture for ambient signals in the high frequency band of up to 1.0 volt per meter. This legislation was designed to protect ham radio operators from complaints by television viewers. At the extremes, Canada has a 0.1 volt per meter standard and Germany 4 volts per meter. Cable has asked for a minimum standard of no visible degradation in an ambient field of 1 volt per meter.

Other issues addressed in this standard involve tuner overload and local oscillator isolation.

Program Guide

Whether this effort is known as the "Advanced," "Automatic," or "Interactive" program guide, it is the same project and had its beginnings in Walt Ciciora's description of a smart, interactive program guide back in the early 1980s. The Program ID Subcommittee of the joint committee has the task of defining the possible transmission facilities required on the cable system to transport programming data from the cable headend to the consumer electronics in the home.

With this summary of on-going EIA/NCTA Joint Engineering Committee activities in mind, let us now turn to efforts in the consumer interface being pursued at CableLabs.

Direct Pickup Interference (DPU)

There are two vital pieces of information required for cable to make a case with either the consumer manufacturers or federal agencies supporting improvements in tuner susceptibility standards for TV receivers and VCRs. First, some quantification of susceptibility to direct pickup interference for currently manufactured TV receivers and VCRs must be developed. Secondly, a division of cable subscribers into off-air signal field strength isobars, showing percentages and estimated numbers of TV receivers served by cable as a function of ambient field strength must be made. The obvious blend of these two efforts will show the magnitude of the DPU problem nationwide, and the degree of tuner shielding required for a 100%, 90% or 80% solution.

Currently manufactured television receivers are not uniform in their susceptibility to off-air interference in the tuner, neither industry wide nor as a function of manufacturer or brand name. There is a degree of correlation among receivers utilizing the same common chassis design. The purpose of the proposed DPU testing of receivers is to develop statistical data on the general state of consumer products relative to tuner shielding and isolation, and to determine which particular model numbers have extremely severe problems.

The above data, when coupled with information which shows the number of cable-connected TV receivers and VCRs residing in different field strength isobars in typical cable systems in the United States, yields the information required for a long-term solution to this vexing problem.

It is doubtful that the consumer manufacturers would agree to, or that the federal government would require, a 100% solution to this problem, which could be very costly to the consumer. However, some level of remedy could be adopted which would solve essentially all DPU problems except those resulting from the worst case conditions, which are fortunately very rare, and would likely still require the application of a set-top converter.

CableLabs is being urged by its member companies to undertake the analyses required to develop the above-mentioned technical and statistical data to quantify the DPU impact nationwide.

A word of caution is required, however. Consider that the average television receiver sold in the United States has a life expectancy of 13 years, with approximately half of that time as the primary set in the home, and the other half in a secondary role. This means that a set sold in 1991, without the suggested improvements in DPU performance, is likely to still be operating in the year 2004.

Further, there does not seem to be any good way to retrofit existing receivers to improve their performance on a general basis. If we are successful in getting some resolution in 1991, then the problem will still exist on the main set of our subscribers through 1997, in decreasing numbers. CableLabs is also considering the possibility of some ancillary device or technique which would eliminate the effects of DPU without using the set-top converter; however this may only offer a low probability of success.

All of the above considerations apply also to VCRs. However, because of certain design features, they are generally less of a problem than the receivers. Also, their 3-1/2 year average life cycle means improvements penetrate our subscriber base much more quickly than those made in television receivers.

In our emerging relations with the consumer electronics industry, it is difficult to find a tuner designer who is not well aware of the DPU problem and who doesn't have a sincere desire to make the required improvements. The problem is that this industry, which works on high volume and extremely low per unit margins, cannot identify either a price or market share advantage in its spending of their profit margin to fix the problem. CableLabs has several efforts underway which should bring market pressure on the manufacturers to implement fixes to their equipment. More will be said of this below. If these efforts do not yield satisfactory results, the only alternative left to cable is to request rulemaking by the FCC.

The EIA-Japan has recommended that consumer manufacturers adopt a double conversion tuner, such as those used in cable converters, into both TV receivers and VCRs. CableLabs is following this development closely, as it could represent our best opportunity for an amicable and long-term solution to the DPU issue.

Advanced Program Guide

It has become increasingly evident that the existing types of program guides available to our subscribers are marginally acceptable for today's volume of cable programming, and even less viable for the greatly increased amounts and types of programming anticipated in the future. CableLabs has undertaken a major effort to provide a cable-unique program guide with very customer friendly features for the future.

The general description of the guide anticipated is as follows:

• Uses a data carrier, thus liberating the video channel normally consumed by the current cable scroller guides.

• Is interactive with the customer, providing for the sorting of programming by factors other than channel number.

• Can automatically control the VCR without direct intervention by the customer.

• Can make long-range conditional recordings. ("...record any John Wayne movie I don't already have archived.")

• Will index the customer's tape library with on-screen display to facilitate programming location.

• The in-home investment will be in the customer-purchased consumer electronics hardware.

• Can produce revenue streams via subscription, local advertising, and cross-channel promotions.

In 1990, CableLabs issued a Request for Information to 27 companies which had previously expressed an interest in television guides. Of the respondees, approximately six were considered to be serious in concept, in depth of understanding, and in resources. CableLabs will choose one or more of these companies to work with for the provision of the video guide of the future for cable's exclusive use.

The important efforts under development for this project are not only technological in nature, but also include the marketing and business aspects of the venture. Negotiations with the consumer electronics manufacturers have revealed a keen interest on their part in cooperating with cable on this project.

Retailer-Cable Joint Ventures

In our meetings with the consumer electronics manufacturers, it became clear to CableLabs that they consider their customer to be the local electronics retailer, and not the ultimate purchaser of the item. Therefore, the manufacturers respond carefully to the requests made of them by the retail stores.

Further, since the retailers and the local cable system do have the same customer to keep happy, and are not competitors, it would seem that some cooperation between these two entities would be useful. Several joint ventures have been tried around the country between a cable system and a local retailer and they have been very successful.

The basic scenario is that the cable system wires the store for full service, and, incidentally, must do a truly superb job of this and check it frequently for needed repairs. The store provides space for a fully functional cable business office on the premises. Service calls which possibly originate from a cable/consumer hardware interface problem, such as DPU, are coordinated with only one service representative responding. Further, an air of cooperation between the two industries makes for a much happier customer.

Cable has found that it is easier to sell both basic and premium services in the retail electronics environment and if the retailer is some distance from the normal cable business office, the customer appreciates the convenience of a closer facility. The stores have found that the cable customers who come in on cable business, but browse the store while there, can increase floor traffic from 10% to 20% by actual count. Several creative ways have been devised to reimburse the cable operator for the cost of wiring the store.

Consumer Electronics Liaison

As mentioned previously, a fairly good relationship has been established at the engineering level between cable and the consumer hardware manufacturers. However, in that industry, the engineers do not generally determine product features, so it became obvious that cable needed to develop a dialog with the executives and product planners as well.

CableLabs has worked diligently in this area over the last year and can report considerable success. We have been well received and have had in-depth discussions with the highest level executives at Goldstar, Matsushita/Panasonic, Mitsubishi, Philips, Sony, Thomson/RCA, and Toshiba. We discovered that they were generally unaware of both cable's success in terms of subscriber penetration in the United States, and of the several significant cable/receiver interface issues. We will continue our discussions with these decision-makers in the consumer electronics industry and have hopes for more cooperation in the future.

CABLE SATELLITES : THE NEXT GENERATION
Issues Facing Cable Operators and Programmers

Robert Zitter
Home Box Office, Inc.

ABSTRACT

The deployment of next-generation satellites in compliance with the FCC's uniform 2-degree spacing plan, together with the movement of cable programming, will occur during the next two years. A discussion of the transition scenario, the technical differences in the new satellites, and ground station requirements reveals that Cable TV and SMATV facilities will require reconfiguration and in some cases replacement may be necessary. The satellite movement and programming transfers in the early nineties compel the cable industry to examine the future performance of existing facilities.

o Introduction

C-Band satellites have become a reliable means of delivery for cable television programs and have played an important role in the phenomenal growth of the industry. We have seen numerous operational, technical and regulatory changes together with technological advances in satellite and ground station equipment that led to significant reduction in overall costs. The cable industry has demonstrated its ability to successfully deal with such changes over the years.

A development that will impact the cable industry is the FCC mandate to phase-in a uniform 2-degree spacing between U.S. domestic satellites. The intent of the plan is to alleviate overcrowding in the U.S. orbital arc. It times the deployment of next generation satellites with improvements in ground station receiving characteristics in order to control the ensuing increase in adjacent satellite interference.

This paper discusses the satellite and orbital changes that are expected to occur during the next two years, and presents some critical issues and challenges facing satellite programmers and cable operators .

o Time of Replacements

In the early days, 4-degrees of separation within the two segments assigned to U.S. domestic communication satellites -- 70 to 104 degrees and 117 to 143 degrees West Longitude (°WL) -- consisting of 15 satellite slots were found adequate. (The central portion from 104° to 117° has been reserved for Canadian and Mexican satellites.) But as the popularity and importance of satellite delivered services increased, the separation provided in the usable orbital arc for U.S. domestic satellites was found inadequate for planned spacecrafts.

Hence, in 1983, the FCC adopted a plan to essentially double the number of orbiting domestic communications satellites by gradually reducing the

Figure 1. Next Generation Satellites

CARRIER/ SATELLITE	FREQ BAND	ORBIT	LAUNCH DATE	
ALASCOM				
Aurora II (Satcom C5)	C	139	May 1991	
AT&T				
Telstar 401	C	97	May 1993	
Telstar 402	C/K	89	Mar 1994	
CONTEL ASC				
ASC II	C/K	101	Apr 1991	
ASC 1R	C/K	129	Sep 1993	
GE AMERICOM				
Satcom C4	C	135	Sep 1992	o
Satcom C3	C	131	Nov 1992	o
Satcom C1	C	137	Nov 1990	■
Satcom H1	C/K	79	1994	
GTE SPACENET				
GStar 4	K	64	Nov 1990	
Spacenet 1R	C/K	103	May 1993	
Spacenet 2R	C/K	69	Sep 1993	
GStar 1R	K	121	Jun 1994	
HUGHES				
SBS 6	K	72	Oct 1990	
Galaxy V	C	125	1991	o
Galaxy 1R	C	133	1993	o
Galaxy VII	C	91	1992	
Galaxy VI	C	91	Oct 1990	■
Galaxy IV	C/K	99	1992	
Galaxy IIIR	C	95	1994	
NATIONAL EXCHANGE				
Spotnet I	C/K	93	Mar 1993	
Spotnet II	C/K	127	Sep 1993	

o Cable Satellites
■ In-orbit Backup for Cable Satellites

spacing between satellites, so that over approximately ten years, a uniform 2-degree spacing would be achieved. This evolutionary approach was chosen because: It prevented early obsolescence of existing satellites; it granted manufacturers sufficient time to design and construct antennas that can better discriminate between adjacent satellites; and, it permitted cable operators to amortize their existing facilities. The FCC's 2-degree spacing plan guided most of the technical, regulatory and economic developments that we experienced during the past decade.

In the early nineties, Alascom, AT&T, Contel, GE Americom, GTE Spacenet and Hughes communications will construct and launch next-generation satellites as shown in Figure 1. The cable industry has committed to basically two satellite vendors -- Hughes and GE Americom -- involving four satellites. HBO and Turner Broadcasting are anchoring Hughes' Galaxy V and Galaxy IR, while Viacom Networks and a group of other programmers including HBO have chosen Satcom C3 and Satcom C4.

o <u>TRANSITION</u>

The preparations for launching the next generation cable satellites and the accompanying movement of programming actually began in 1988, as Galaxy I and Satcom 3R approached their designed end-of-life terms. Recognizing that approximately three years of lead time is required to design, construct and launch satellites, the FCC and satellite vendors developed a transition strategy with the goals of ensuring uninterrupted service and minimize inconvenience to users.

The transition timetable for the deployment of satellites is illustrated in Figure 2. Note that Satcom 3R (131°WL) has been temporarily replaced by Satcom 1R in 1991 and will remain there until Satcom C3 (131°WL) becomes operational in 1993. Satcom 1R (originally at 139°) moved its traffic to Satcom C1 (139°W) and then Satcom IR was repositioned to 131°W to take over traffic from the retiring Satcom 3R (131°W). Satcom C3 will take over when

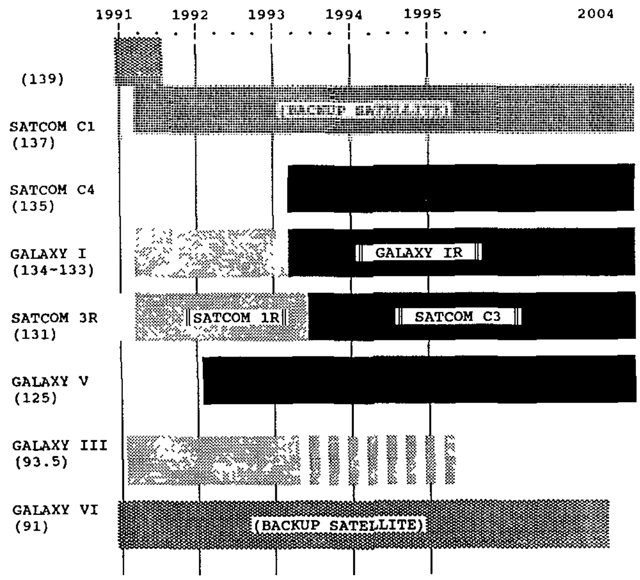

Fig. 2 Deployment of Next Generation Cable Satellites

Satcom 1R retires. Although the changeover and spacecraft maneuvers between Satcom 3R, 1R and C3 are complicated, most transfer activities will be transparent to cable operators because antenna repointing would not be required. Minor polarization skew adjustments needed to be performed when Satcom IR was relocated due to the finalization of the orthogonal interleaving of adjacent satellites, which was not supported by Satcom 3R.

Galaxy I will be moved by one degree from 134°WL to 133°WL during May or June of 1991 in compliance with the FCC uniform 2-degree spacing plan. This movement will require antenna repointing. The schedule will be announced to the industry ahead of time so that multiple-beam antenna feeds unable to be used in the 2-degree spacing plan can be replaced or new antennas with better specifications can be installed.

The move will most likely occur during a 7-day period, beginning on Monday. To minimize service disruptions, the bulk of the satellite movement would occur on weekdays. Only one or two repointing procedures would be normally required, depending on antenna size, antenna directivity and system noise margin.

Galaxy V (125°WL) will be launched during the last quarter of 1991 and will become operational in early 1992.

In the middle of 1993, Galaxy IR (133°WL) will replace the retiring co-located Galaxy I (133°WL) and remain there until its end of life in 2005

Another cable satellite, Galaxy III (93.5°WL) will remain in its current orbital location until its end of life in 1995. The broken line in Figure 2 depicts a possible transition of programming from Galaxy III to other satellites as transponder leases expire in 1993.

While these transfers are happening, both GE Americom and Hughes will maintain two C-Band fleet spare satellites Satcom C1 (137°WL) and Galaxy VI (91°WL) respectively. These in-orbit spare satellites were launched in advance (1990) to provide restoration in case of launch failures of the four primary cable satellites. The spare satellites will contain pre-emptible programming to accommodate protected cable services in the event of fleet satellite failures.

Figure 3. **PROGRAMMING ON CURRENT AND NEXT GENERATION SATELLITES**
(January 1991)

PROGRAMMING ON CURRENT SATELLITES			POSSIBLE PROGRAMMING ON NEXT GENERATION SATELLITES			
GALAXY 1	SATCOM IR (Formerly IIIR)	GALAXY III	GALAXY V	GALAXY IR	SATCOM C3	SATCOM C4
HBO E	HBO W	MTV E	HBO E	HBO E	Showtime	Nickelodeon
Comedy Channel	BET	MTV W	TBS	Cinemax E	Movie Channel	C-SPAN 2
Headline News	Request TV 1	Nick E	TNT	Turner Bdcst	VH-1	QVC
ESPN	USA W	HA!	Headline News	Disney W	E!	Nustar
Showtime W	CNBC	Viewer's Chce 1	CNN	USA W	QVC	Discovery
TMC W	Lifetime W	Viewer's Chce 2	ESPN	ESPN	A&E W	Prevue Guide
Disney E	TNT (to G1 3/91)	Weather Cha	USA E	Eternal Word	C-SPAN 1	Family Chan W
A&E	Cinemax W	C-Span II	A&E E	Univision	Family Chan E	Sci-fi Chan
TNT (3/91)	Bravo	VH-1	Disney E	Nostalgia	Home Shopping 1	Home Shop 2
CNN	E!	Lifetime E	Mind Ext. U	Group W	Learning Channel	Viewer's Choice
Cinemax E	Learning Channel	C-SPAN I	FNN	Inspirational	Lifetime E	Lifetime W
WTBS	FNN/Source	Nick W	BET		Viewer's Choice	WOR-TV
Showtime E	AMC/AVN	Family Net	Monitor Channel		Weather Channel	
TMC E	Request TV 2	Nustar	Group W-Nashville		Cinemax W	
USA E	Fashion	MEU	United Video-WGN		HBO W	
Disney W	TBN	EWTN	Trinity			
Discovery/video Mall	Shop TV	CVN	Viacom			
TNN	Univision	HSN				
Family Channel	QVC	ACTS				
WGN	VISN	CVS/PPV				
ESPN Blackout	CNN/feeds	ESPN Blackout				
CMTV	Travel					
WWOR	Home Shop 2					
Inspirational	Inspirational(4/91)					
Galavision						

o <u>Where will programming windup?</u>

As the new fleets of next generation satellites become fully operational, the programming lineup will transition as well. The transfer of programming will take place in 1992 and 1993.

HBO and Turner Broadcasting, joined by USA, ESPN, Disney and others will anchor Galaxy V and Galaxy IR. Viacom, joined by other programmers will anchor Satcom C3 and Satcom C4. HBO and Cinemax West feeds will be on Satcom C3. After the switchover to the next generation satellites, the cable programming lineup will most likely appear as in Figure 3.

To minimize service interruptions, the programming networks will most likely dual-feed each of their services for a certain period. Cable operators are encouraged to keep in touch with programmers to keep abreast with developments and to determine the exact timing and location of simultaneous feeds. It is imperative for cable operators to find out where the programs currently carried will wind up. In order to determine how the transition will impact individual cable system operation, you should become familiar with the future satellite plans of each programmer.

o <u>2° Spacing-Timing and Implications</u>

The process of retiring satellites and launching new ones has been carefully planned to ensure a smooth transition. Actually, the space segment procedures have already started. For example, the two spare in-orbit satellites have already been launched and now operational; new generation satellites are now either approaching final design stages or under construction; and, some spacecraft maneuvers have been accomplished. There are more activities that are planned to take place during the next two years, and more ground segment activities can be expected.

The eventual reduction of satellite spacing to two degrees will require careful examination of several ground segment technical issues. Of utmost importance is the earth station's ability to avoid interference from signals coming from undesired adjacent satellites. The parameters that affect interference are: Satellite transmit power (EIRP), antenna directivity, receiver sensitivities, signal formats, type of modulation, frequency offset, IF Bandwidth and filtering techniques. Of these, satellite EIRP and antenna directivity are major factors that determine acceptable or objectional levels of interference.

The majority of C-Band satellite receiving antennas currently serving the cable industry are of parabolic design, and pick up signals from a single satellite. Since 1983, antenna manufacturers have been improving their designs in anticipation of the 2-degree spacing scenario. Cable operators should find out from antenna manufacturers which designs need modification or replacement. In addition, proof-of-performance tests should be undertaken to ensure that antennas currently meet design specifications after many years of use. Corrosion, warping and misalignment degrade antenna performance.

Smaller diameter antennas such as those serving SMATV or smaller cable systems are particularly susceptible to increased interference from adjacent satellites due to their wider beamwidths.

Other antenna configurations such as those having multiple feed horns could be seriously affected if each individual C-Band feedhorns can not be physically moved closer to one another. To correct this problem, some antenna manufacturers devised new assemblies and/or feedhorn designs that are claimed to function under the 2-degree spacing plan.

In some cases, parabolic antennas that were retrofitted with dual or triple feeds might have to return to single feed configurations.

There is still sufficient time to identify potential problems and deal with them accordingly.

o <u>Technical Differences in the New Generation Satellites</u>

The technical parameters of next generation C-Band satellites have improved significantly. Traveling Wave Tube Amplifiers (TWTAs) have progressed from 5 Watts to 8 Watts, up to its new capability of 16 Watts. The latest in antenna beam-shaping techniques also allow a more uniform concentration of power to desired coverage areas. Furthermore, other spacecraft improvements such as better power subsystem design, better heat management, decreased intermodulation distortion (IMD), and improved transponder protection schemes ultimately yields a significantly improved next generation C-Band satellite system.

Figure 4. Comparison of Signal Quality

Ant. Size	Retiring Satellites	Next-Gen. Satellites
3.0 (M)	C/N=11.4 C/I=23.1 C/N+I=11.1 VSN=48.3	C/N=14.4 C/I=19.7 C/N+I=12.4 VSN=49.6
3.7 (M)	C/N=13.2 C/I=24.9 C/N+I=12.9 VSN=50.2	C/N=16.2 C/I=21.5 C/N+I=14.2 VSN=51.4
4.5 (M)	C/N=14.9 C/I=26.6 C/N+I=14.6 VSN=51.8	C/N=17.9 C/I=23.2 C/N+I=15.9 VSN=53.1
7.0 (M)	C/N=18.8 C/I=30.5 C/N+I=18.5 VSN=55.7	C/N=21.8 C/I=27.1 C/N+I=19.8 VSN=57.0

Where:
C/N = Carrier-to-noise ratio
C/I = Carrier-to-interference ratio
C/N+I = Total carrier-to-noise plus interference, calculated as power summations and taking into account slight EIRP differences among adjacent satellites.
The separation of 3-degrees were used for retiring satellites and 2-degrees for next generation satellites.
VSN = Weighted Video Signal-to-Noise Ratio

Given:
Antenna Efficiency = 65% Antenna Temp = 25 °K
LNA Temp = 70 °K IF Bandwidth = 28 MHz
EIRP(retiring) = 34 dBW EIRP(next gen) = 37dBW

These improvements are all well and good, but the forthcoming uniform 2-degree spacing environment will negate the benefits if appropriate steps at the ground segment were not taken.

Figure 4. shows how the signal quality would change if satellite spacing is reduced to 2-degrees using the next generation satellites. It can be seen that larger antennas (4.5M or larger) would be less affected by adjacent satellite interference and therefore would yield signal to noise ratios in the mid to upper fifties (dB) of video signal to noise ratio. These high values are desirable when feeding television to cable distribution systems, even more so as fiber reduces its distribution plant degradation and improves video signal-to-noise in subscriber delivery.

o Summary: What is the Impact on Cable Operators?

Satellite and cable operators are now faced with the inevitable challenges that next generation C-Band cable satellites, 2-degree spacing mandate, and the accompanying movement of cable programming present. Cable operators using antennas smaller than 4.5-meters or those using multiple beam feeds could be affected. Everyone will experience the inconvenience of antenna repointing, so break out the Liquid Wrench! Some will face equipment reconfiguration and perhaps in some cases, replacement. The transition plan is set, but there is ample time to prepare. The industry needs to work closely with each other to raise awareness of the need to ascertain the future performance of existing facilities. Specifically, there is a need to:
1) Determine the migration plans of programmers currently carried by the system;
2) determine the impact of 2° spacing on

existing facilities; and then,
3) come up with a system-specific technical plan and transition timetable.

By doing so, we will ensure a successful transfer of services to the next generation satellites and maintain the excellent signal quality that we strive to provide to all our subscribers.

* * *

REFERENCES

1. "Reduced domestic Satellite Orbital Spacings at 4/6 GHz," G. Sharp, OST Report FCC/OST R83-2 May, 1983.

2. "Fixed Service: No Surplus in Sight," P. Lambert, Broadcasting Magazine, July 30,1990, pp. 48-51.

3. Hughes Communications and GE Americom Press releases.

CABLELABS 1991 ADVANCED NETWORK DEVELOPMENT WORK

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Abstract

The cable network is often considered to be the portion from the headend to the subscriber's house. The network architect must widen his perspective to begin at the source of origination and extend to the customer premise equipment in the subscriber's home. This end-to-end assessment is necessitated by the advancements and migration of technology in production equipment and in consumer electronics. One of CableLabs' major tasks is to research Advanced Network Development (AND), to propose a model, and to test it. The intent of the AND process is to ensure that the fundamental principles of existing network topologies are not adversely impacted with the migration of new technology or functionality.

Advanced Network Development (AND)

The Network Architecture Model is a fundamental set of rules outlining a conceptual model of the cable network. In effect, the architecture defines the design of the network both in terms of functionality and physical deployment. The process of deriving the network architecture involves mathematical modeling, which includes network optimization. The tools used to constrain and optimize network solutions are the simplex method, Karmarkar and Multi-Frontal Method. Economic analysis and traffic engineering concepts are also employed to develop cost models, throughput capability, and points of contention. With these mathematical tools, we are able to develop and manage a complex set of parameters that allows us to constrain investment or technical attributes against an established set of objectives. These scenarios can then be evaluated in order to minimize the number of tests to be simulated or maximize investment opportunities.

Academic research plays a significant role in the development of a cohesive network architecture. While a great deal of the research is theoretical in nature, it serves to limit many of the network scenarios that are eventually evaluated. Network architecture is iterative in nature due to the ever-changing technology and functionality. Thus, it is obligatory to obtain a variety of perspectives and expert opinions to minimize any serious design flaw and to be in a position to optimize opportunities well in advance of the technology coming to fruition. The academic community provides CableLabs with this perspective. A list of a few of the universities or institutions CableLabs is working with is provided below:

University of Colorado, Boulder
University of Victoria, British Columbia
University of British Columbia
Simon Fraser University, British Columbia
Florida Atlantic University
Massachusetts Institute of Technology

Network simulation is an important component in developing a cohesive network architecture. The advent of the overlay of digital signals over analog to position cable to transport digital HDTV; switched video routing to provide selective programming; fiber to the hub or fiber to the bridger; the use of a hybrid fiber/coax network to provide bidirectional PCS services — all of these concepts must be simulated through a series of tests and configurations to validate the concepts derived from theory.

Finally, before the fundamental rules of a network architecture evolve into network design software, the concepts that have undergone rigorous testing and evaluation stages are rolled out into the Advanced Network Development Field Test. This is a controlled environment where the concepts are tested on a small area of a live cable system. The concepts evaluated here reflect technologies that have been simulated in the CableLabs' facility.

Network Design

Based upon the network architecture model, network design is a fixed series of parameters that specify the detailed engineering parameters for configuring a network. From this network design, installation crews can readily construct a new build or replacement/enhancement of an existing cable system. Typically, the use of network design software is constrained to specific parameters that preclude the network designer from conducting various "what if" scenarios. In general, the network solutions evaluated using these software systems include the coaxial-based cable system; fiber-to-feeder based system with or without a ring topology; and fiber to the bridger, with coax to the home. For each of these network designs, 300 MHz, 450 MHz and 550 MHz cable systems are developed.

The intent of the AND process is to ensure that the fundamental principles of existing network topologies are not adversely impacted with the migration of new technology or functionality. The tree/branch topology is an inherent strength of the cable industry's ability to provide cost-effective services and to compete with new entries into the cable business.

Standards Recommendations

With the convergence of technology being used by cable and the telco industries, the need to participate in standard-setting forums, particularly those that have long been the domain of telcos, is evident.

The State Department classification of CableLabs as a scientific institution allows CableLabs to participate and to make recommendations to the State Department. As a participant in the CCIR, CableLabs has access to CMTT, which also oversees the CCITT. The purpose of CableLabs' status as an observer is to ensure that the cable industry is in a position to interface with converging technologies now and in the future.

Traditionally, broadcasting standards have been managed by the CCIR. The CCITT has been governed predominantly by telcos and PTTs. However, in the last few years, the CCITT has become increasingly more involved in the development of digital broadband standards that include video services. To position cable, CableLabs has focused on key standards committees comprised primarily of telco participants to develop inter-operability of interfaces with various broadband networks. In October of 1990, CableLabs contracted a consultant to identify which committees would provide cable with the maximum exposure, with the least impact on dollars and resources. The results were recently discussed with the decision to begin with ANSI.

ANSI T1S1.5 is a subcommittee that focuses on digital broadband network architecture and video services. CableLabs has contracted a consultant, Dr. James S. Meditch, to represent the cable industry as an observer and eventually make recommendations to the State Department concerning interfaces to broadband networks. One of the objectives of this activity is to familiarize key individuals within the cable industry with the process of formating documents for submittal. This is an area that the industry can support by recommending key individuals to aid and facilitate this process.

The T1S1 subcommittee is responsible for development of SONET and ATM standards in the United States. As was mentioned in a report released by CableLabs in 1990, the SONET standard appears to be near completion. It is also a hardware-defined standard that is not necessarily well suited to the multiple bit-rate requirements of video. However, ATM is particularly well suited for variable bit-rate transport of video, since it is a software-defined standard that offers dynamic allocation of bandwidth. CableLabs involvement will ensure that operational practices are developed for the cable industry in accordance with ATM standards.

CableLabs is also involved in observing the development of international standards on PCN, which is governed by the CCIR working group on wireless communications. This working group is responsible for recommending spectrum and technical attributes for PCN at the WARC92 international conference.

Personal Communications Network

CableLabs initiated the study of PCN in May of 1990, as part of a Telephony Study. This study assesses the operational issues of providing telephony, traffic engineering design parameters, and equipment evaluation of those systems that operate over a cable system. It also addresses the technical and non-technical attributes of providing telephony services over a cable system for both wired and wireless networks.

With the increased public awareness of and response by the cable industry to PCN experimental licenses, CableLabs sponsored a conference on PCN in early January of 1991. The conference brought together senior management from the cable industry, along with Dr. Jerry Lucas of TeleStrategies, key consultants, manufacturers, cable PCN participants and regulatory representatives from the FCC and legal entities to discuss the opportunities and technical and regulatory ramifications of cable's entry into PCS services. The conference was attended by over 100 individuals from the cable industry.

CableLabs has just formed a PCN Technical Advisory Subcommittee to look into the technical attributes and economic viability of providing PCS services over a cable network. Issues related to PCN, such as upstream spectrum 5-30 MHz, PCN cable interface, PSTN interface, and switching and interconnection to the PSTN, were identified as areas that required CableLabs' immediate attention. The scenarios associated with cable interconnection to the PSTN will be defined by CableLabs. They include the following types of interconnection:

PABX or Class 5 to MTSO
PABX to LEC
Class 5 to LEC Class 5 or tandem switch
Class 5 to IXC

A major issue surrounding the timing of PCS is the availability of spectrum. This issue will likely determine the technology that will be used in the United States in the forthcoming years. The technologies now under development include:

Frequency Division Multiple Access (FDMA)
Time Division Multiple Access (TDMA)
Code Division Multiple Access (CDMA)
Packet Division Multiple Access (PDMA)

FDMA and TDMA technologies are dependent on spectrum availability that is not readily available in large market areas, such as New York City, Chicago and Los Angeles. CDMA and PDMA are the most promising in terms of the spectrum availability and utilization, however, they reside in spectrum that is largely unprotected by the FCC in cases of interference. CableLabs is focusing on strategic alliances and research associated with these technologies to pursue the technology that proves to be the most economical and available for deployment in a timely manner.

CableLabs is examining a number of network architecture issues: the ramifications of PCS on cable's existing topology; propagation; delay; cell size; the need for fiber-to-the-radio base station versus a hybrid fiber/coax configuration; and switch deployment, including centralized and distributed switching functionality. This work will include joint tests with equipment manufacturers and cable field tests and demonstrations. CableLabs will be contracting a PCN Systems Integrator to facilitate the process of defining the network design. CableLabs is also evaluating spectrum issues regarding the potential provision of PCS.

Conclusion

AND is an iterative process requiring continuous assessment of the end-to-end network infrastructure that delivers video source from the production end of the spectrum to the equipment that resides in the subscriber's home. Technology and customer demands will continue to change the complexion of the network. It is the task of CableLab's Advanced Network Development group to continually assess and optimize the network architecture to ensure the least impact and most functionality from the cable network design.

CABLELAB'S SCIENCE AND TECHNOLOGY FOR THE CATV INDUSTRY

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Today, more than ever before, the key determinant of the strength of any industry's economy is the efficiency of that industry's technology. The difference between success and failure in our marketplace today often comes down to who has the better technology. For the CATV industry to remain economically sound, we must achieve and maintain a leadership position in technology. In those areas where our activities have not historically defined the frontiers of technology, it is essential that we be close enough to those frontiers so that we are able to exploit new discoveries whenever and wherever they are made.

I maintain that the CATV industry, including cable TV systems operators, equipment manufacturers, and programmers, has and continues to build one of the strongest science and technology enterprises that the telecommunications industry has ever seen. CableLabs was established, in part, to give our business leaders a preview of technological change. To accomplish this, a technology time line had to be established, starting with technology assessment, moving promising technology to projects, and then weeding out the failures and turning the successes over to operations.

It is important to recognize that we live in an age characterized by change. The CATV industry has been aware for some time that a strategic technology entity, such as CableLabs, was needed to help manage the rapid pace of technological change. Even the best of leaders with the brightest strategic mind is reduced to dealing purely with tactics if a grenade is tossed in his or her lap with the pin pulled. Also, it became clear, that no single MSO or system operator could hope to gain a sufficient amount of the benefit from a basic R&D or engineering program to justify the investment on its own.

CableLabs has a major role to play in working with the CATV industry to take basic discoveries and move them through the pre-competitive development phase for a whole array of relevant technologies. By sharing the risks, we can speed up the process and enable our industry to compete on a more level playing field with the alternative entrants and competitors.

Of course, when the doors of CableLabs were initially opened, there was no time line. First executive staff had to be hired, space leased, and the hundred and one other details handled that are involved in starting a company from scratch. A board of directors and the executive board committee were elected. The technical advisory committee and the technical advisory steering committee were appointed. Liaison with existing technical groups and the manufacturers was established. Staff was hired and work began on timely projects that demanded immediate attention, primarily advanced television and consumer electronics issues. An extensive survey of the industry's technical leaders gave the Labs valuable insight into areas that required priority attention.

CableLabs started a technical newsletter (*SPECS Technology*) along with a clearinghouse that functions both as our eyes to the world and the world's eyes to us. Several projects were initiated, including building a developmental headend, scientific research on F-fitting failures, battery life testing, and NTSC subjective testing. An optimized systems operations (OSO) study is also underway.

An important activity which was assigned to the CableLabs' Office of Science and Technology was to define a list of critical technologies, to specify where we are with respect to business opportunities or possible threats in each of those technologies, and then to make a series of recommendations. Those include recommending action that should be taken to improve the cable industry's relative stance, to improve our utilization of these technologies, and to take more effective advantage of areas where we have a leading position.

As a result, a search for a top flight physicist with business training was undertaken. We looked for an individual who is able to understand business strategy and to communicate with the CEOs; who is able to understand the relationships between technology and business opportunities; and who is capable of developing a technology strategy responsive to the industry's business strategy. Now that this position has been filled, technology assessment is truly underway. Already, this effort is beginning to yield information that helps place important current and emerging technologies on the technology "S" curve. An "S" curve is a way of graphically displaying a technology's initiation, expansion and maturity.

For instance, we learned the following: that our existing silicon-based hybrid amplifier technology still has a significant amount of headroom both in performance and bandwidth. Large screen displays with very good resolution, brightness, and at reasonable prices are beginning to appear on the horizon. Fiber optic devices and even the glass (and other mediums) are developing much faster than originally anticipated. Fiber optic amplifiers are a reality now and will be a factor in the CATV business shortly. Coherent modulation schemes are much further along and are expected to be available for experiments within two to three years.

For CableLabs and the cable TV industry, fiber optics is a technology, not a business. However, the evolution toward fiber-rich network architectures, including the last frontier – fiber to the home – is a major driving force for the extension of broadband capabilities and fulfillment of the consumer's appetite for entertainment, a powerful marketplace incentive. Future improvements in lightwave capability may come from using coherent technology and optical amplifiers. Coherent systems will offer the advantages of greater receiver sensitivity and longer fiber reach, as well as greater receiver selectivity and ease of adding or dropping channels – in much the same way as FM radio stations are tuned-in, the ultimate optical radio.

Future progress in fiber optics depends heavily on photonics, or technology for making optoelectronics devices. Two key technological developments are making tomorrow's devices possible. One is quantum-engineered materials, which do not exist in nature and are tailor-made to have certain desirable optical or electrical properties. The other is the integration of a large number of individual optical and electronic components on the same chip. We are watching technological advances closely and are working with R&D facilities as well as manufacturers of components and equipment to assure development of devices and systems dedicated to cable-specific requirements.

What is the video outlook for America? This question is being asked often. Americans and the world have a love affair with multi-channel television. As the public demand and appetite for television increases, we must be able to deliver more and more programming. The way that we can do this is through video compression and digital modulation. It is not easy, it is not here yet, and it presents tremendous technical challenges in compression ratios and motion compensation.

However, the rapidly falling costs and increased processing speeds of semiconductor devices (microprocessors and memories) have already resulted in a heightened interest in accomplishing an advanced TV transmission standard using digital techniques. Without compression, it would take 50 or more cable channels to transmit a digital HDTV signal and about eight cable channels to send a normal, digitized TV channel. But new digital compression techniques have emerged to drastically reduce the digital television transmission bandwidth. Some concepts are well known. Other new approaches, like compression by fractals, are being developed. Which techniques are the best choice for cable applications remains an open issue.

As an example of some recent concepts, which have received attention in the trade press, consider a system utilizing data compression that transmits programming over cable systems or other distribution channels in short bursts to receivers that store information in memory, and then play it back in real time. A two-hour program may be viewed in real time , but can be sent in a short time (e.g., 15 seconds), allowing transmission lines to be freed for other uses. However, to accomplish this, even with significant compression ratios, it is necessary to have transmitters and receivers with several Gbits of RAM/video memory (based on standard memory chips) or other high storage capacity and very fast access devices. Also, this "video file" transfer will require high data stream rates (e.g., 2 Gbits/second). It is a very interesting concept, but the requirements and today's technology currently make any reasonable application cost prohibitive.

We need new approaches to video storage – reusable optical disks, organic molecule-based memory, or memory chip-by-inch. New HDTV sets will need video frame memories as well as digital signal processing. At CableLabs, we are particularly interested in the expected timing of device developments for the next generation of several integrated circuits (ICs). Memory chips (DRAM) are and will remain the worldwide "technology driver" in the future, and at the same time progress in their development is a benchmark for other families of devices such as microprocessors (CPU), digital signal processors (DSP), etc. All those ICs, by their manufacturing complexity and capabilities, are equal to or are some number of generations behind DRAMs. Today, a DRAM chip can store 4 Mbits of information, and the newest 16 Mbits chips are going through customer qualification. However, 1 Gbits chips are not expected to be qualified before 1998, and their availability is assumed for year 2004.

The new display technologies are focusing on reproducing the versatility of the CRT in larger viewing area-to-volumeratio and lower power devices. Flat screen technologies may give us a television set that can be hung on the wall like a picture. Both plasma and liquid crystal display technologies hold promise for the video flat panel of the future, but for now they are restricted to non-TV applications. One thing is for sure, the display of the future will convey information faster, with more resolution and color than ever before. The big question is what signal requirements and image quality perception will result from different distortion forms as compared to today's NTSC process. In other words, the question "what's good enough" is a moving target.

Technology advances and their implementation in cable systems' advanced network architectures are a very attractive arrangement, both from the perspective of quality/reliability and minimizing operating costs. And hopefully, the depressed economy and the need to tighten capital and operating expenditures won't jeopardize CableLabs' efforts and place the industry in an unprepared state to deliver advanced television signals as well as alternative services to cable customers when the time is right.

In the future, the technology "spoils" are going to go to the quick, the smart, the tenacious. What we need is greater flexibility and greater ability to match technology to changing customer demands. In the cable TV industry, long-range technological and engineering strategic thinking is not as well-developed as it should be. We are prepared at CableLabs to help, and we look on technological competitiveness as one of the most pressing of the CATV industry's problems. But, we deeply believe that we, as a technology-based consortium, are not nearly as well qualified to make the strategic business decisions for the industry as are each of system operators themselves. Nor do we believe that technological transfusions in the absence of such strategic plans are any solution at all.

In conclusion, let me say that the CableLabs' Office of Science and Technology plans to continue its diligent exploration of new technologies for our industry. CableLabs also fully intends to rely upon its highly effective Technical Advisory Committee, comprised of senior engineering and technical executives from our member companies. We intend to continue the work that has made the cable television industry the most successful communications industry in this country, bar none. The industry's multi-channel technology has long been admired by the international markets and to this day draws compliments from them.

CableLabs' Office of Science and Technology has rapidly come up to speed and is now positioned to help the CATV industry monitor the rapidly moving frontiers of technology. We clearly understand our role at CableLabs is to discover, capture, and mold promising technology to the benefit of the CATV business.

Thank you.

CATV CONSTRUCTION - EUROPEAN STYLE

Michael J. Jones
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Abstract

The following paper is an overview of the unique construction practices employed within the United Kingdom in order to build underground CATV and telephony systems. This paper discusses civil construction, cable installation and the different types of cables available in Europe.

The CATV market in the United Kingdom has been trying to emerge since the early 1980's. After many staggered starts it now seems to be "all systems go."

The total population of the United Kingdom is approximately 58.4 million people which equates to just over 21 million television homes.

Recently the Cable Authority in conjunction with the Department of Trade and Industry finished the franchising operation. They have issued, in total to date, 136 franchises covering 14 million television homes or a potential 14 million cable subscribers.

Build rates vary from 3 to 5 years depending on the franchise, but however long it takes, in order to cable the United Kingdom it will require digging up and re-instating approximately 140,000 kilometers or 86,970 miles of pavement and roads. Because of the UK housing, a large population in a relatively small land mass, for every mile of trench laid this opens up access to 160 to 220 homes per mile as potential CATV telephony subscribers.

With system being the operative word, let's look at what constitutes a United Kingdom CATV/Telephony System with particular regard to civil construction, cable installation and cable types available.

The most popular system architecture being used at present is the tree and bush configuration. This is a hybrid version of the tree and branch networks employed in the United States. The major difference between the tree and branch and the tree and bush configurations is the placement of the subscriber taps. In tree and branch we design the placement of the subscriber taps every 100 to 150 feet down the feeder cable depending on density and expect average drop lengths of 100 to 150 feet. In a tree and bush configuration, the node or multi-port taps are designed to feed up to 64 subscribers from a central point using an average drop length of 260 feet. It is not uncommon to reach upwards to 1000 feet per drop per subscriber depending on system design.

As mentioned earlier, the amount of civil construction is vast. The next section will explain in more detail typical methods of underground CATV/ telephony system construction.

Most pavements or sidewalks are finished in concrete slabs, solid concrete or tarmac. The main excavation is normally 18 inches wide and 18 to 24 inches deep. Disc cutters and small digging machines are used but the majority of the work is done by hand utilizing pneumatic drills, picks and shovels. Great care must be taken to minimize damage to existing underground services. Although survey maps exist, their accuracy is suspect. Damage to power, gas, water or telecoms can be costly and time consuming.

Once the trench has been opened, the next stage is the installation of the PVC or polyethylene duct. Sizes of the duct vary but the most common in use is the 90mm or 4" duct. The number of actual ducts placed in the trench can be anywhere from a single up to 6 ducts depending on system design and subscriber density. It could be many years before the local councils or authorities permit this type of massive construction project again. Many operators, (realizing this and also the future value of a conduit to basically two thirds of the households in the United Kingdom) are designing extra duct space into their networks, with an eye towards the future of either using the extra duct themselves or possibly leasing it as new technologies arise.

Once the duct is installed, in order to be able to enter and exit the duct system for construction and maintenance, a series of pits or underground chambers are used. Most installations employ two types of pit, main and intermediate. The main or larger pits are usually used in front of the equipment cabinets and at the corners of street crossings depending on the size and number of cables being pulled. The intermediate or smaller pits are normally spaced between the main pits at intervals of 75 to 100 feet to access the duct for subscriber installations, hard cable installation and plant maintenance.

There are three major reasons pits or chambers are used:

(1) The pavements are much narrower and do not easily allow the massive excavation required to install 45^0 or 90^0 rigid ducting.

(2) Because of the subscriber density constant access is required.

(3) Pits are used to facilitate 90^0 bends and road crossings.

Size and materials used for chambers or pits also varies. Some are plastic with metal flush-fit lids, others are pre-formed in concrete and some where necessary are hand built in brick. It really depends on cost restraints and individual engineering preference.

Once the duct work and pits are installed, temporary or permanent reinstatement follows quickly. Very rarely do you see long spans of open trench. The policy seems to be backfill and re-instate as you go.

Having completed their civil construction, these crews move on to make way for the hard cable pulling gangs. These gangs normally consist of four or five men.

Stationing themselves at a main pit, the cable pulling gang will assemble a drum or drums of cable on a stand behind the pit. They then open up all the intermediate pits between the pull points. Certain precautions need to be taken before the actual cable pulling begins. Such as the use of entry and exit rollers where the cable is fed into the pit or out of the pit. This prevents the coax from being deformed or the jacket rupturing on the lip of the pit. Another precaution which is utilized is the use of bell mouths into all exposed duct ends where the cable is to be pulled. This reduces the chances of damaging the hard as well as drop cable jackets on the sharp jagged edges of the duct ends exposed in the main and intermediate pits. Most trunk and distribution cable is pulled in by hand using draw ropes. Very few companies use power winches. Some of the shorter runs are installed using a Cobra.

The Cobra is a semi-rigid, limited reach, push/pull tube coiled on a circular frame which is fed into the duct. On the end of the Cobra is a swivel eyelet. The cable is attached to this and then drawn back through the duct.

Duct carrying capacity is critical as is good duct management. If the management is poor, full duct fill is not achieved. Thus causing wasted construction expense. As an example, the typical capacity of a 90mm duct is figured at 20 to 24 RG6/Sidecar cables. Thus one duct has the ability to contain the drop and telephony cables to wire 20 to 24 subscribers.

In many cases during a build, it is necessary to take the cable through a 90 degree bend. This is achieved by entering and exiting the pull pits and utilizing a sequence called "fleeting." During fleeting, the cable is pulled out of the pit and laid down the pavement in a large loop or figure of eight. The lead end of the cable is then fed back pit into the correct duct to accomplish a 90 degree bend. The advantage of fleeting over the conventional methods of using angled ducting is the ability to pull long lengths of cable in stages, pull pit to pull pit. This also reduces the maximum tension exerted on the cables. If fleeting is done correctly, the amount of 90 degree bends is virtually unlimited, no matter what size cable is being pulled. Obviously, great care must be taken during this process. The cable is extremely vulnerable to mechanical damage when laid out in this way. To lessen the possibility of damage, good construction practices must be adhered to as well as the previously mentioned use of entry and exit rollers, pit pads and bell mouths. Fleeting can be and is used to pull all types of cables into the CATV and telephony network.

All the system electronics are housed above ground in street cabinets. Cabinets housing the main trunk amplifiers also, in most cases, house the telephony equipment. Access for cabling is achieved through the base of these cabinets.

Similarly, distribution cables terminate into the street cabinets. This is normally the multi-port tap or node point, the bush in the tree and bush configuration. Depending on the requirements, it is possible to feed up to 64 subscribers from these points. Drop cable will vary in length from this point but a good average seems to be 80 meters or approximately 260 feet.

The drop installation crews gain access to the duct system via a subscriber point. These are normally set into the pavement next to the garden wall during initial construction and capped to prevent ingress of rubbish which could block or foul the system. The subscriber points, normally a 35mm or 1 1/2" duct, are connected to the main ducting by a directional coupling or "Swept T." The "Swept T" concept was developed exclusively for the UK market. As the Cobra is pushed down the duct from the subscriber point the "Swept T," being off-set, directs the Cobra towards the node.

The drum stand or "A" frame holding the drop cable or sidecar telephony cable is positioned over the pit nearest to the node cabinet. The cable is then connected to the swivel eyelet on the Cobra and drawn back through to the subscriber point in the pavement. If 90 degree bends are required, the same fleeting procedure is adopted as used with the larger trunk and distribution cables.

Most houses in the UK, be they private or council housing, have a small front garden which is normally enclosed. Permission must be obtained to dig a trench to bury the drop cables to the front wall of the house. Some operators use a small PVC conduit for the drop cables but the majority direct bury.

Once at the house a small plastic box is used to terminate both CATV and telephony drops. The wall is then drilled and the cable is taken to the isolator box inside the house.

Back at the node, the RG6/ Sidecar cable is separated. The drop is connectorized and joined to the multi-port tap for CATV service. The telephony pairs are prepared and connected for immediate use into the telephone exchange or are weather sealed for future use.

Even though the preference in the UK is towards North American style cables due to quality, availability and pricing; the actual range of cable types available for use is quite extensive.

European cable designs vary between manufacturers and also countries. Two main aspects remain common however.

(1) All primarily use some form of copper for both the inner and outer conductors.

(2) Semi-air spacing is the most commonly used dielectric construction.

Center conductors are usually made of solid copper. Copper coated aluminum or copper covered steel are rare.

Dielectric construction is varied but the primary styles are:

(1) Bamboo or fused disc
(2) Thread in tube
(3) Five cell or cartwheel
(4) Solid polyethylene
(5) Expanded foam

Bamboo or fused disc and thread in tube are typically Dutch or German design.

Five cell or cartwheel constructed dielectrics are typically British designed.

Solid polyethylene and expanded foam dielectrics are derived from old British military specifications and are manufactured by all countries.

Outer conductors or outer screens also differ. Once again the main conductor metal is solid copper but it now comes in the form of a tape which is longitudinally applied over the dielectric. No welding process is used. The copper tape is just over-lapped. Some manufacturers use a braid at this point as well. If the cable is to be used in an underground environment, an aluminum tape will also be applied to act as a water barrier. The whole assembly is then sheathed or jacketed in PVC or polyethylene. At no point is a flooding compound used. Bonding between components using high powered adhesives is also not common among European manufacturers. The over braid not only acts as an extra screen but also secures the copper tape in place.

North American cable manufacturers tend to rely on tried and tested cable constructions such as Parameter Three and the relatively new Quantum Reach products.

Both typically have a copper covered aluminum center conductor bonded to an expanded foam polyethylene dielectric, which is bonded to an outer conductor of solid aluminum. The aluminum outer conductor can be applied in one of two ways:

(1) Extruded aluminum tube

(2) An aluminum tape which is formed around the dielectric and then R.F. welded.

These methods are both acceptable from an engineering standpoint but the welded tape process enables the manufacturer to produce a more flexible product. This enables the cable to be installed in an underground system much easier. Saving both time and capital expense.

The cables are then coated with either polyisobutylene or tar flooding compound and over-sheathed in PVC or polyethylene.

European drop cables are constructed in exactly the same way as trunk and distribution but obviously scaled down.

One new addition to the drop cables being used in the United Kingdom has evolved through the necessity to provide both CATV and telephony services. We call it the "Sidecar."

It consists of a full specification RG, either 59, 6, 7 or 11 forming one leg and two twisted telephone pairs in either 24 or 26 AWG forming the other. The cable is then fully flooded with polyisobutylene and sheathed in polyethylene. Simple to install, it eliminates installation crews having to go back to pull another cable when further services are sold. This again saves both time and cuts capital expense.

I hope this transcript has given an insight into the United Kingdom and European CATV and telephone systems now being built to those in our industry not at present involved overseas.

Digital Video: Whatever Happened to Differential Phase and Gain

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Abstract - Digital video has moved from the background into operational systems and proposed systems which will be very important to cable television. The term digital video is applied to technologies including transmission of NTSC composite video via digital modulation, digital special effects and digital compression of images which may then be transmitted using digital or analog modulation. Most of us are comfortable with NTSC video transmitted via FDM/VSB-AM or FDM/FM. We understand the parameters of the baseband signal and the effect imperfections of the transmission path have on that signal. The parameters of digital video can relate to some of the parameters we understand but also introduce new ones. The same imperfections of the transmission path (noise,interference, linear distortion and nonlinear distortion) will affect digitally transmitted images differently than our familiar NTSC. It is often simplistically said that digitally transmitted images are "perfect" above some threshold and "crash" below that threshold. The process of digitizing affects the basic "perfection" of the image. When digital compression is used, the choice of compression algorithm affects image quality. This choice along with error correction and masking also determines the way transmission imperfections appear in the image.

INTRODUCTION

In this paper, we will first present a brief review of those transmission channel impairments that we have come to know very well in the NTSC environment. The review will include such linear impairments as frequency response, ghosting, and C/L delay inequality and how they affect the received NTSC image. We will then review some non-linear distortions such as differential gain and differential phase, and also take a look at the effect that interference and random noise have on the NTSC signal. The emphasis on this portion of the paper will be to relate the particular transmission impairment to the subjective effect on the NTSC picture. Once we have reviewed the familiar impairments of the NTSC environment, we will take a similar look at the effects that certain transmission channel impairments might have on digitally transmitted video signals.

The use of digitally encoded video signals, especially compressed video, has made great strides in the last couple of years. All the current HDTV proponents except NHK, have now proposed all digital systems at bit rates (including audio and data) of less than 20 Mb/s. This compares to 100 Mb/s for currently operating non-compressed NTSC digital systems. Compression and digital modulation schemes have also been proposed to place multiple NTSC signals within a 6 Mhz bandwidth using similar data rates.

How will the transmission channel's linear and non-linear distortions that we know about today affect these new digital signals? This question will lead us into a general discussion and definition of the measures that are commonly used to analyze transmission channel impairments, and their effects on a digital signal. In particular, we will focus on measurable quantities called "eye height" and Bit Error Rate (BER). Once we have an understanding of the BER and its interrelationship with the various channel impairments, we will discuss improvement of the BER by the use of Forward Error Correction (FEC) and improvement of the channel by adaptive channel equalization. Finally, since there are several different algorithms that have been proposed for the encoding of the digital video signal such as DCT, VQ, MVQ, etc., we will briefly examine the relationship between the type of algorithm and visible picture impairments resulting from certain transmission impairments.

THE EFFECT OF CHANNEL IMPAIRMENTS ON ANALOG NTSC

Any transmission channel can offer a multitude of potential impairments to an analog NTSC video signal. When you consider that in the satellite/cable environment, a video signal may pass through as many as 40 or more active devices including the uplink electronics, downlink electronics, headend processing, through the distribution plant, and into the home, each providing its own form of impairment to the video, it is amazing that the signal is viewable at all! (Some of our detractors may, in fact, argue that the signals aren't viewable.) This cascade of electronics, as well as the transmission channel itself, can cause linear distortions such as poor frequency response, C/L gain and delay inequality and ghosting, non-linear distortions such as differential gain and differential phase, as well as interference and noise. Since the transmitted waveform directly controls the picture display, any distortion will be related directly to a visible effect. Because we are so familiar with the NTSC environment, we are readily able to equate these various distortions to the visible picture impairments that we have come to know and love.

Linear Distortions

Linear distortions can be roughly defined as those distortions which may occur to a signal passing through a transmission channel that are independent of the amplitude of the signal. In other words, all signal amplitudes would be affected equally by the device causing the impairment. Inadequate filter bandwidths in the transmission path, for example, can cause poor frequency response which may result in soft pictures. Such filters may also exhibit an unequal envelope delay throughout the passband. If this is the case, we know that the chrominance information or luminance information may be delayed with respect to the other (C/L delay) resulting in the well known funny-paper effect. This may also cause video waveform transitions to be slow or ring resulting in a lack of definition or smearing of the video at transitions. Ghosting is another linear phenomena with which we have become very familiar as it is probably the most disturbing distortion imaginable and shows itself as a complete leading or trailing image superimposed on the original. Short delay ghosts, on the other hand, can cause picture softening or sharpening.

Non-linear Distortions

Non-linear distortions can be classified as those distortions whose magnitudes depend upon the signal level through the channel. Typically, such distortions will occur as a result of amplitude compression of the video waveform in active devices such as modulators, amplifiers, and demodulators. Differential Gain and Differential Phase are examples. Differential Gain can be defined as an unwanted change in the amplitude of chrominance signal with changes in the amplitude of the luminance signal---in other words changes in color saturation with changes in the brightness of the scene. Similarly, Differential Phase is an unwanted change in the hue (phase) of the chrominance signal with changes in the brightness of the scene. These distortions are not as easily recognizable on a monitor as some of the linear phenomena because they are more dynamic---varying with the content of the scene. Differential Phase is probably the most recognizable because hue (color) changes are more easily seen than saturation (depth of color) changes. A classic example of an extreme case of Differential Phase distortion might show up on a baseball players face as he stands out in the sun with his cap on such that the bill casts a slight shadow over half of his face. With differential phase, instead of a bright flesh--tone on one side of his face, and a shadowy flesh tone on the other, we might see either side of the individual's face to appear slightly green or magenta depending upon whether the shift in phase were positive or negative.

Interference

Interference, for purposes of this discussion, can be classified as any unwanted spurious signal, other than noise, which may affect the signal within the channel. Typically, the interference of which we speak in the cable environment consists of 2nd and 3rd order beat products caused by active devices in the distribution plant which may fall within the CATV band. In addition, such spurious beat products may be found at the output of CATV modulators and signal processors at levels close to 60 dB below the video carrier. These types of spurious signals, when they fall within a typical TV channel, manifest themselves in the form of clearly visible vertical, horizontal, or diagonal lines or beats in the picture. The magnitude of this effect is dependent upon both the magnitude of the interfering signal as well as its placement within the video channel. Beats that fall close to the video and color carriers are more readily visible than those farther away. Another factor that affects the visibility of the beat, is whether or not the interfering signal is itself modulated (or dispersed in energy). Spurious interfering signals which are unmodulated tend to show up as well defined beats in the picture, while those that are modulated tend to look more noise-like with lesser defined beats.

Noise

Random noise, of course, is a fact of life in any transmission system, and is one of the primary reasons that the CATV industry is working feverishly to limit the number of active devices in cascade in the distribution plant through the judicious use of fiber optics. Noise, of course, can have a devastating effect on an analog NTSC picture, with weighted signal to noise ratio (S/N) being its classical measure. S/N is defined as the ratio of peak-to-peak video (100 IRE) to rms noise present in the video waveform, after bandlimiting and weighting. Bandlimiting is used in the measurement to exclude irrelevant noise energy, while weighting is a form of low-pass filtering to weight the measurement with what is visible to the naked eye. This is done to account for the eye's inability to perceive high frequency noise. The subjective effect of random noise on the video signal of course, is a pronounced graininess or "snow" in the picture.

THE EFFECT OF CHANNEL IMPAIRMENTS ON DIGITAL TELEVISION SIGNALS

The review of NTSC analog transmission above shows that various channel impairments can be related directly to visible effects in the picture. While most of these channel impairments affect digital signals, this sort of direct relationship between impairment and visible effect does not occur with digital transmission. We will consider the various steps in the process of digital video transmission in order to understand how impairments in various parts of the system affect the picture.

FUNDAMENTAL CONCEPTS OF DIGITAL VIDEO

Quantizing

The first step in digital transmission is quantizing. Quantizing is the process of selecting a finite number of discrete values which the signal will be permitted to assume. Signal values other than those permitted are assigned one of the permitted values according to a specific algorithm or rule. Simply stated, the signal is assigned the nearest permitted value. The luminance portion of the NTSC composite baseband video, for example. can have any value within a 100 IRE range. The quantized signal would be constrained to a particular set of values, say integer IRE values. A luminance level of 72.68 IRE might be assigned the nearest permitted value, 73 IRE. A similar process is applied in time. Only the values of the signal at specific sample times are retained. An example of a luminance waveform before and after quantizing is shown in Figure 1. The quantizing steps have been made large for clarity in the illustration. We have retained the NTSC sync and line format because it is familiar. The quantization process described above is called Pulse Amplitude Modulation (PAM). The reverse process is applied to the received waveform. That is, the level at each sample time is compared to the permitted values and assigned the nearest permitted value. Unless noise or other impairments caused by the transmission channel are large enough to cause the received waveform to be quantized to an incorrect level, the impairment will not appear in the recovered waveform. In our example of quantizing in 1 IRE steps, the distortion or interference would have to be 0.5 IRE or greater to cause an error in the quantized level of the received signal. If the distortion level becomes larger, however, it will appear in the signal in much the same manner as it would if the signal were transmitted without quantizing. The digital video signal is not usually transmitted in this relatively direct manner.

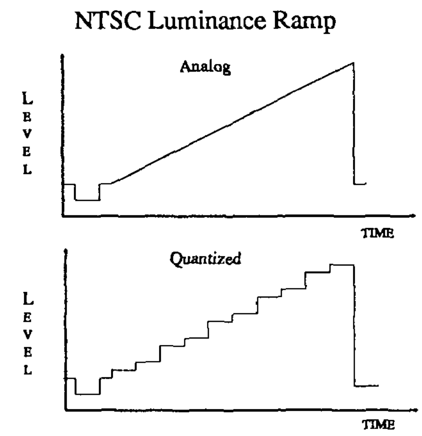


Figure 1 Analog and Quantized Luminance Ramp

Instead, the level of each sample is represented by a series of bits. The series of bits are then transmitted rather than the analog sample. The coding of the analog samples into a series of bits is called Pulse Code Modulation (PCM). Since each permitted signal level is exactly defined, the quantized waveform at the receiver can be a replica of the quantized waveform at the transmitter. For replication, the digital to quantized analog converter at the receiver must exactly match the quantizer at the transmitter and there must be no errors in transmission of the digital bit stream.

Modulation and Demodulation

Some form of modulation must be used if we are to transmit the digital bit stream on an RF carrier. Several different modulation techniques exist [1]. These include Bi-Phase Shift Keying (BPSK), Quadrature Phase Shift Keying (QPSK), Offset Quadrature Phase Shift Keying (OQPSK), Quadrature Amplitude Modulation (QAM) and Quadrature Partial Response (QPR) as examples. Each is different in detail and has particular advantages but most share certain basic characteristics. They can generally be broken down into double sideband suppressed carrier amplitude modulated (DSBSC-AM) channels with two or more discrete levels. These fundamentals are not so difficult to understand and we can begin by looking at the basic case of amplitude modulation.

Amplitude modulation can be expressed by the following general equation [2]:

$$M(t)=a(t)\cos(\omega_c t)$$

where

$M(t)$ = the modulated carrier EQ (1)

$a(t)$ = the modulating waveform

$\cos(\omega_c \times t)$ = the carrier

This equation can be used to describe NTSC luminance modulation. In that case, $a(t)$ is directly related to the luminance. The value of $a(t)$ at sync tip (-40 IRE) is 1 and at peak white (+100 IRE) is .125. As you can see, $a(t)$ has only positive values in NTSC VSB-AM. The carrier level varies from the peak value at sync to 87.5% depth of modulation at peak white and is never fully turned off. The equation can also be used to describe DSBSC-AM. In that case, the value of $a(t)$ ranges from +1 to -1. We can see that as $a(t)$ varies between the value of 0 and +1 the carrier amplitude varies from off to full on, but what of the negative values? Negative values simply mean a phase reversal of the carrier. In other words, between the values of 0 and -1 the carrier amplitude varies from off to full on but with a relative phase of 180 degrees.

When DSBSC-AM is used for transmission of digital information, discrete values of $a(t)$ are used to represent digital values. For example, two states +1 and -1 could be assigned. One state would represent a logical 1 and the other a logical 0. The carrier would be at full amplitude in both states but would have 180 degrees difference in phase. A modulation scheme of this type is often called Bi-Phase Shift Keying or BPSK. Here, the two modulation states represents "one" and "zero" states of a single bit. Actual data will consist of a stream of "ones" and "zeroes". Figure 2 shows a typical modulation waveform and modulated carrier. The bits occur at regular intervals (clock periods) and the change between states follows a sinusoidal shape. In ?, the sample time for each bit is at the grid line (10 bits are displayed). Note the relative carrier phase between sample points where the data is at -1 and +1. We have now described the essentials of a basic digital modulation format. More complex schemes which carry more bits within the same bandwidth can be employed but each can be built up from this fundamental model. All of the basic ideas remain the same.

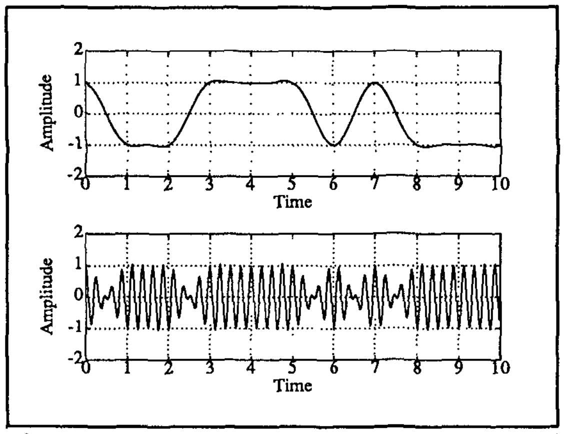


Figure 2 BPSK Data Waveform and Modulated Carrier

At the receiver, the signal is demodulated to produce an output level proportional to the phase and amplitude of the received carrier. Note that envelope detection cannot be employed since both the amplitude and phase of the modulated carrier must be detected. Synchronous detection is required. We can now form a simple diagram the entire digital transmission process Figure 3.

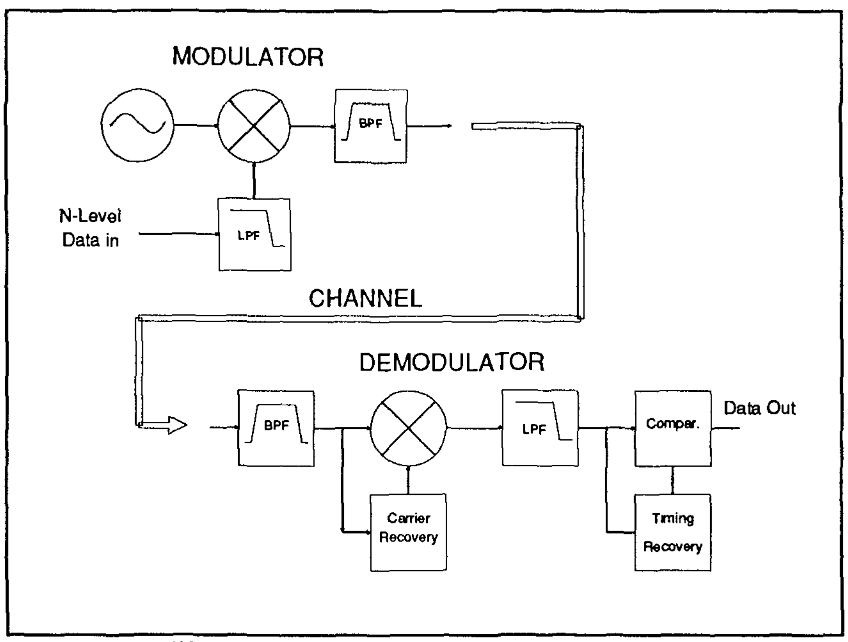


Figure 3 Simplified Digital Transmission System Diagram

Eye Diagrams

In the ideal case with no noise, interference or distortion, it is easy to determine each state. For

BPSK we simply sample at the correct time and determine whether the level is positive or negative. It is somewhat more difficult when the impairments are added. A look at Figure 4 will reveal why. Here we see a display of the demodulated BPSK (two level case) waveform such that two bit periods are displayed with successive bits overlaid in the same window. A sequence of bits containing many combinations of ones and zeroes are overlaid. This type of diagram is called an "eye" diagram because of its similarity in shape to the human eye. We can determine each bit value by determining whether the signal is positive or negative at a sample time. Note, however, that the time of sampling may vary somewhat according to our ability to recover timing information and the actual decision level may vary due to circuit variations. Because of this, the signal must be outside the box defined by timing variation horizontally and decision level uncertainty vertically. This will present little problem in this unimpaired case. The "eye" is said to be open when there is a clear open area between signal states as in this diagram. We will see later how the eye closes due to interference and noise. Other impairments due to distortions within the channel as well as imperfection of the filters which shape the data modulation waveform can also close the eye.

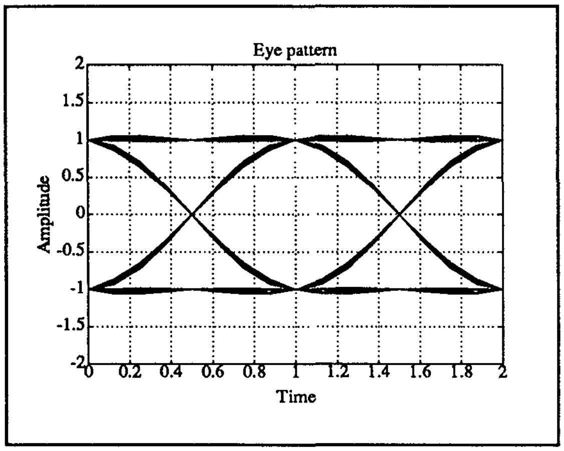


Figure 4 Two Level Eye Diagram (Without Impairment)

These same arguments will apply even when we consider a more complex modulation scheme. We can be more general by referring to the states as Symbols and each period as a Symbol period. A symbol can represent more than one bit value in some modulation schemes. For example, a second carrier may be added in quadrature with the first and similarly modulated. We now can transmit twice as many symbols. This type of modulation would be classified as QPSK or 4 QAM. Each carrier is independent and the analysis is identical to that above. There are simply two channels of data.

Alternately, the number of discrete levels can be increased. For example, we can let $a(t)$ have four discrete levels. An eye diagram for the four level case is shown in Figure 5. We have multiplied $a(t)$ by three to obtain scale lines at particular locations. One can see that the separation between levels has been reduced in both amplitude and time. The area free for the decision "box" is now smaller and the eye opening will be more easily closed by interference or other impairment than the simpler two state case. Each symbol (level) now represents a particular two bit sequence. There are four states representing 10, 01, 11, and 00.

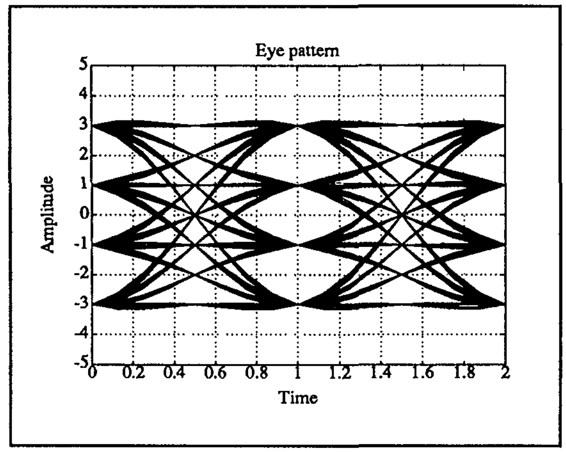


Figure 5 Four Level Eye Diagram (Without Impairment)

If we need more data throughput, we can add a quadrature carrier with the same four level modulation to double the number of symbols which can be transmitted. This would correspond to 8 QAM.

Interference and Noise

To understand how certain types of interference may affect eye closure, consider the effect of a CW interfering signal near the carrier frequency. This interference will alternately add to or subtract from the transmitted carrier, changing the level and/or phase of the received signal. The demodulated waveform under these conditions might look like that of Figure 6. The same "beats" seen as diagonal (or other) lines in NTSC video are visible in the demodulated data waveform. The "beats" cause the states to be less clearly separated. The open area between states has been reduced in both time and

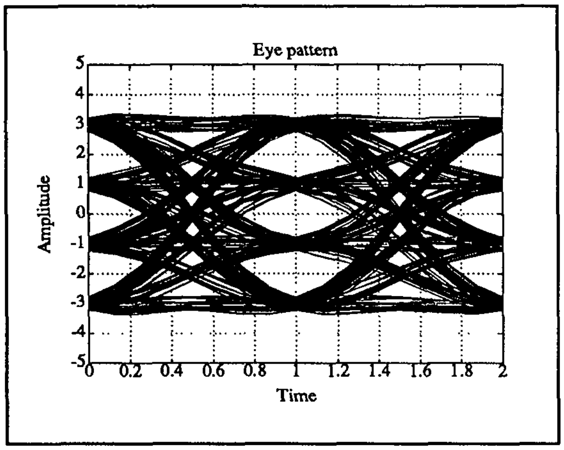


Figure 6 Four Level Eye Diagram With CW Interference

amplitude by the interference. Other impairments such as noise result in similar filling in of the open area as seen in Figure 7. If the distorted waveform passes through the decision "box", errors in the decoding may occur. When the eye closes, symbol errors will be generated and the quantized waveform at the receiver will differ from that at the transmitter.

Linear Distortion

Ghosting will also appear in digital transmission. The effect of longer delay ghosts will be to create lower level delayed replicas of the symbols. Short delay ghosts will be equivalent to frequency response errors. Quadrature ghosts, which sometimes appear in NTSC as image sharpening, can produce cross talk in some digital transmission systems. In all cases, the eye opening will be reduced. Ghosting can be a severe problem requiring ghost canceling, a form of adaptive channel equalization.

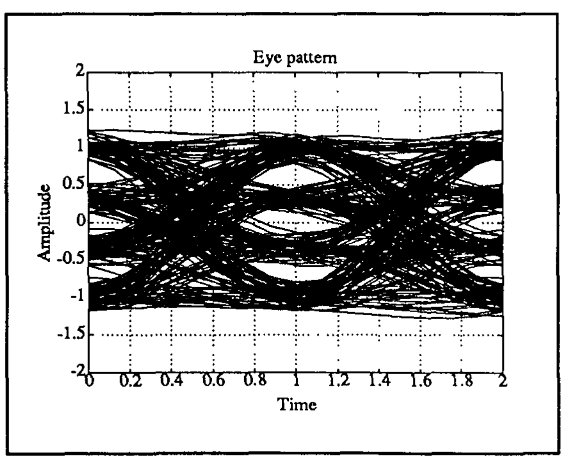


Figure 7 Four Level Eye Diagram With Noise

Adaptive equalization can also correct for amplitude and phase errors from various sources within the channel.

Summary of Channel Impairments

The important points to understand about the discussion to this point are:

1. The analog source is quantized to discrete levels.
2. These levels are represented by a sequence of bits.
3. The bits are represented by discrete modulation states i.e. carrier amplitude and phase state.
4. The demodulated received signal will exactly reproduce the transmitted bit pattern unless a combination noise, interference and distortion causes errors in decisions on modulation state.
5. Noise, interference and distortions close the decision "eye".
6. Less noise, interference and distortion can be tolerated as the number of modulation levels is increased.
7. In the absence of errors in the received bits, the reconstructed quantized waveform at the receiver will exactly match that at the transmitter to the extent that the digital to quantized analog converter at the receiver exactly matches the quantizer at the transmitter.

THE PICTURE IMPAIRMENTS OF DIGITAL VIDEO

We have already commented that a direct relationship between channel impairment and visible effect does not occur in digital video. Indeed, some types of impairments simply do not exist because of the way picture information is encoded. An example is differential phase and gain. These are distortions of color information in NTSC caused by nonlinear interaction between luminance and the color subcarrier. There is no subcarrier in digital video to suffer phase or gain distortion. Further, color and luminance are transmitted as separate components in the digital video systems of interest here. The video is typically divided into three analog signal channels, one carrying luminance and two others carrying color difference signals. The color difference signal may be the I and Q signals as used in NTSC or some other matrix. Each of these signals is digitally sampled separately and the three digital channels are time division multiplexed before modulation. Because they are handled separately, there can be no interaction. The same is true for other impairments related to NTSC encoding such as cross color and cross luminance.

Some new impairments such as quantizing noise are added. Quantizing noise is essentially the difference between the original analog signal and the analog signal recovered by filtering the quantized signal. This noise might appear as false contours in picture areas with low detail, for example. As a practical matter, quantizing noise can generally be made negligible in digital video system design.

Other impairments, such as luminance non-linearity still can occur in digital video but are limited to analog circuitry ahead of the digitizing process or following the digital to analog conversion. The picture information cannot suffer nonlinear distortion while in a digital format. Transmission channel, modulator and de-modulator nonlinearities can cause data waveform distortion and contribute to bit errors but cannot cause nonlinear distortion of the picture information. Similarly, noise in the circuitry preceding digitization will cause a different effect from that in the transmission channel. Noise ahead of the digitizing process can cause some disturbing effects, particularly when digital video compression is used. Noise in the transmission channel will simply contribute to bit errors. Cable system equipment can contribute to bit errors but will not cause or correct nonlinear distortions in digital video.

Bit errors certainly will cause picture impairment but how will the impairment appear visually in the picture? Even in the simple case of Pulse Code Modulation (PCM) with no compression, the effects can be varied. An error in the least significant bit representing the amplitude of a single sample may be nearly invisible. The same error in several adjacent samples might be more noticeable. The appearance could be a shift in hue, color saturation or luminance level of the affected pixels. An error in the most significant bit would likely be apparent. The effects of errors becomes more complex when digital video compression is considered.

There may be impairments caused by the compression algorithm itself. These impairments will depend on the source material rather than the transmission channel. For example, noise in the source material can cause false motion to be detected and can cause "blocking" artifacts. The video information is often processed as rectangular groups of adjacent pixels. Noise or other distortion within the block may affect the way the entire block is compressed. The individual processing blocks may not fit together, causing their boundaries to be visible. The degree of background detail and complex motion may also cause artifacts. An example of complex motion would be small foreground objects moving left with a detailed background moving right. All of these are akin to NTSC encoding artifacts. They are inherent in the algorithm used and cannot be caused or corrected by cable system equipment. The degree to which they are objectionable is dependent on psychophysical effects in perception.

There are a number of compression techniques which are in use [3]. These include predictive coding, transform coding, sub-band coding and vector quantization. Other processes which are used in one or more of these techniques are Huffman coding and run length coding. It is interesting to note that the bit errors which we have been discussing can affect more than a single sample in all the compression techniques mentioned above. In predictive coding, for example, certain sample values are determined from algorithms involving several other sample values and a correction value. One sample may therefore affect several others. In many compression schemes, spatial blocks of samples are processed together so that an error may affect an entire block of samples. The visual effect of errors can be quite different between compression schemes. The statistics of the errors are also important. Errors which occur in bursts will cause different effects than those occurring with a rather even distribution. In all cases, a fairly error free transmission is required for satisfactory performance. This leads us to a discussion of error correction techniques.

Error correction is a necessity in any practical transmission channel for digital video. Forward Error Correction (FEC) can be accomplished by adding some number of bits to the data. The values of these added bits are determined by algorithms involving various combinations of the data bits. By cross correlating various data and correction bits, errors can be discovered in the recovered data and the faulty bits corrected. Of course there are limits to the number of errors and the local frequency of errors which can be corrected successfully. These added bits also use some of the data channel capacity requiring a tradeoff between correction capability and the amount of image data which can be transmitted. Error masking can also be employed where errors can be detected in a block of data but the exact bit or bits cannot be determined. A simple type of error masking would be interpolation between known values. Digital systems are often viewed as working perfectly as error rate increases until the point that catastrophic failure occurs, causing a complete loss of signal. Error masking can afford a more graceful failure, allowing a less nearly perfect picture to be received under high error conditions such as found near the boundaries of coverage area in broadcast television. Hierarchical coding can provide a similar effect. A core picture of lower resolution can be carried by a lower data rate robust data stream with enhancement data carried in a higher rate less robust format. When the enhancement data is lost, the core picture can be displayed. Whether such techniques are needed will have to be evaluated.

Variable length coding can be damaged severely by errors. For example, Discrete Cosine Transform (DCT) coding often takes advantage of a large number of zero coefficients by transmitting the number of zero coefficients rather than transmitting each zero coefficient using the standard bit pattern for zero. This reduces the number of bits required for transmission and causes the number of bits required for each picture block to vary depending on content. The same is true for Huffman coding [4]. An uncorrected error in such a bit stream can cause all following data to be in error until a hard reset is encountered. Systems which use variable length coding schemes are therefore required to send periodic system resets to assure that the system can recover if an uncorrected bit error occurs. This periodic reset requires additional data to be transmitted.

We can see that the performance of the transmission channel (including its modulator and demodulator) can be characterized without regard for a particular compression algorithm. We have already discussed how the characteristics will depend on the modulation format being used. The number of bits which can be transmitted per second increases as the number of discrete levels increases but the amount of noise etc. which can be tolerated decreases. Applying a knowledge of the channel frequency response, noise and interference along with the modulator and demodulator characteristics will permit an analysis of Bit Error Rate (BER). The effect BER and its statistics on various digital video compression formats can be evaluated separately. Included in this evaluation would be the Forward Error Correction and any Error Masking which is a part of that system.

Summary of Picture Impairments

The important points to remember about this discussion of the impairments of digital video are:

1. Familiar impairments of NTSC video caused by the transmission channel such as Differential Phase and Differential Gain, etc., will not appear in digitally transmitted video.
2. Compression algorithms may introduce new inherent impairments which are independent of transmission channel characteristics. In other words, these impairments cannot be caused or corrected by the cable system equipment.
3. The source video supplied to most compression systems must be of very good quality and free of noise. Source video impairments can cause failure of the compression algorithm.
4. The effects of cable channel impairments can be characterized by Bit Error Rate (BER) and the statistics of these errors. These effects are dependent on modulation format but independent of the digital video system.
5. Digitally transmitted video can be designed to fail in a graceful manner if this is required.
6. The effect of BER and error statistics will depend on the digital video compression system employed.

SUMMARY

Digital video transmission of both HDTV and NTSC has become a key interest throughout the entertainment and information delivery industries. It is a subject which is likely to be new to many in the cable industry. The objective of this presentation is to provide a basic understanding of digital transmission. Impairments which occur in digital video are compared and contrasted with the familiar NTSC. We have briefly reviewed the effect of channel impairments on NTSC, relating the signal impairment to visible effect. The fundamental concepts of digital modulation and transmission have been presented. These impairments were related to measurable effects such as eye diagrams. The concept of Bit Error Rate was introduced and the visible effect of BER was discussed. We have briefly discussed how the compression algorithm affects susceptibility to errors and the use of Forward Error Correction and Error Masking. This discussion should prepare those unfamiliar with digital transmission for a better understanding of more detailed treatments of the various modulation formats and the effects of various impairments and compression schemes.

ACKNOWLEDGEMENT

The authors wish to thank Mr. Leo Montreuil for his kind help in preparing the simulated modulation and eye diagrams used as illustrations.

1. Dr. Kamilo Feher, *Advanced Digital Communications Systems and Signal Processing Techniques*, Prentice-Hall, Inc. Englewood Cliffs, New Jersey, 1987, Chapt. 7.

2. Members of the Technical Staff Bell Telephone Laboratories, *Transmission Systems for Communications*, Revised Fourth Edition, Western Electric Company, Inc., Winston Salem, North Carolina, 1971, p 97.

3. Joseph B Waltrich, "A Tutorial on Digital Video Compression Techniques", 1990 NCTA Technical Papers, PP 37-49.

F-connector Corrosion in Aggressive Environments - An Electrochemical and Practical Evaluation

Brian Bauer - Raychem Corporation

ABSTRACT

The drop has long been considered the weak link in the CATV system. As future systems require greater shielding due to CLI restrictions and higher bandwidth, the drop will receive increasing demands for sustained performance. At the root of many drop problems is corrosion at the F-interface.

As part of a CableLabs funded project, an investigation was conducted into the destructive dynamics of corrosion as it pertains specifically to the F-interface. The electrical performance profile of F-splice samples exposed to Copper-Accelerated Acetic Acid Salt Spray (CASS) was determined. These samples have been dissected, and much of the mystery behind the discontinuous nature of shielding degradation is explained through the use of Scanning Electron Microscopy and X-ray Spectroscopy. Relative corrosion rates of often used metals within a CATV drop interface, including port, connector, and cable materials are examined. The weakest material link in this connection, found to be the aluminum braid, is studied with regard to relative galvanic coupling to key materials. Corrosion data is offered that is pertinent to the design of future connectors/ cables/ ports.

INTRODUCTION

CATV's F-interface, has long been recognized as one of the weakest areas in a CATV plant. In the case of properly installed connectors, corrosion is probably the largest cause of interface failure. However, very little has been studied with regard to the effects of corrosion.

The F-interface, the focus of this paper, consists of the port, connector and cable. Corrosion is highly dependent upon the compatibility of mating materials and therefore these parts are to be thought of as a system. The way these interface materials interact is crucial to system survival especially in corrosive (coastal, humid, temperature varying, or high precipitation) environments.

This is the first in a three phase series of CableLabs directed F interface corrosion studies. The program objective is to investigate some of the primary material changes and corrosive dynamics of the CATV F-interface. The three phases are to 1) determine the effects of corrosion on the F-interface through the use of Copper-Accelerated Acetic Acid-Salt Spray Test (CASS) 2) investigate the corrosive effects of passing low currents through the F-interface, and 3) compare the above to that of actual environmental conditions in the field.

OBJECTIVES

The primary goal of the first phase was to investigate the electrical performance and material change due to corrosion of the F-interface and to investigate the dynamics involved therein. Particular objectives were as follows.

1) To determine the electrical performance versus time profile of mechanically well stabilized F-interface samples in a corrosive environment (CASS).

2) To determine material changes and key paths of corrosion within the F-connector. Methods used are

scanning electron microscopy and x-ray spectroscopy.

3) To compare the basic corrosive drive of copper acetic acid salt solution to that of deionized water and 3% salt water. This involves measurement of electrode potentials of CATV metals of concern.

4) To investigate the effects of metallic and bimetallic (galvanic) corrosion, due to CASS, on metals typically found within the F-interface.

The method of accelerated salt fog used is CASS ASTM B368 (Appendix A) due to its very aggressive nature of attack. It should be noted that many tests exist for replicating corrosive field environments. Other methods, such as the standard 5% salt spray of ASTM B 117, and cyclic humidity tests such as ASTM G 60, however, have not been effective in degrading the electrical performance in this sort of study within a reasonable time frame. By using CASS, key corrosive information can be attained in a more expedient manner, such as electrical performance profiles, corrosive deterioration paths, and other important patterns. Basic patterns are assumed to occur similarly in other tests of moisture caused corrosion, but results are obtained in a more timely manner using CASS.

ELECTROCHEMICAL STUDY

This section gives the results of the electrochemical corrosion tests performed on CATV F-interface materials. First a background of fundamental electrochemistry will be presented followed by a description of the kinds of electrochemical measurements performed and the results obtained. Then a brief introduction will be given as to how these measurements fit into a corrosion study. For further introduction to corrosion phenomena the reader is referred to an excellent book by Fontana and Greene [1].

BACKGROUND, ELECTROCHEMISTRY

Chemical energy can be converted to electrical energy by choosing the correct reactants. In the case of the familiar car battery, **Figure 1**, lead is the reactant at one electrode and lead dioxide is the reactant at the other electrode.

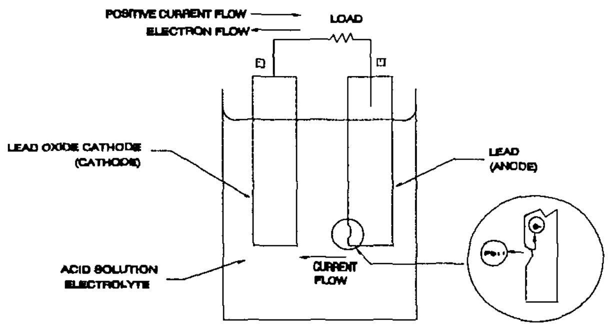


Figure 1: Battery Cell, analogous to the corrosion dynamic.

During the course of the battery discharge one of these reactants is "oxidized," which releases electrons. The electrons flow through the electrical circuits and return to the battery where the other reactant is "reduced." Since a transfer of electrons is involved in the electrochemical reaction, a voltage difference is measured between the two electrodes. It is noted that it is only a voltage difference that can be measured. An individual electrode does not have an absolute potential. In electrochemistry, the hydrogen electrode has been arbitrarily set to zero potential, and very often tables will list electrode potentials relative to this standard. The hydrogen electrode is difficult to use in the laboratory, however and in these cases a different "reference electrode" is used (**figure 2**). In this report we have used a saturated calomel reference electrode. For reference, its potential relative to the hydrogen reference electrode is 0.2425V.

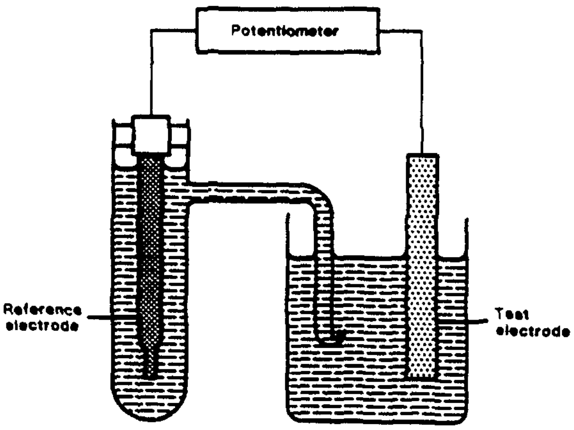


Figure 2: Use of reference electrode to determine electrode potential.

In the battery described above, one electrode was oxidized, one electrode was reduced, and the electrons flow in the outer electrical circuit. In freely corroding systems there is no outer electrical circuit. Therefore, the oxidation (which is the corrosion reaction) must occur in local proximity to the accompany reduction as in **figure 3**.

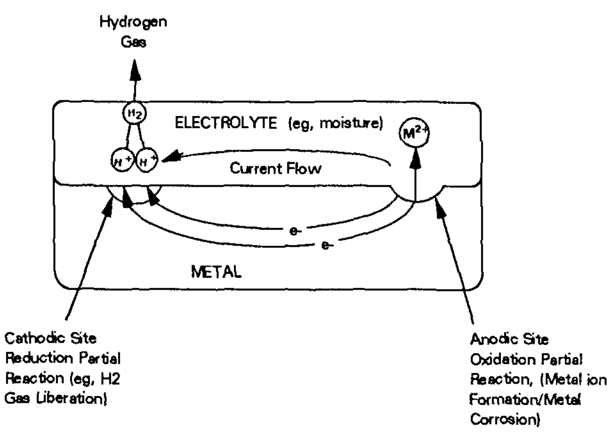


Figure 3: Corrosion of Single Metal in Presence of Electrode.

In the case of a single metal corroding, the oxidation and reduction can occur on separate surfaces (that are electrically connected). However, it is still very likely that the reactions will be in close proximity to each other.

Corrosion is an electrochemical process and as such it is driven by the voltage difference between the oxidation reaction and the reduction reaction, as in our battery above. An example of an oxidation would be the dissolution of iron to form iron ions, and an example of the concurrent reduction is the reduction of protons to form hydrogen. (If you drop iron in acid you will see a high degree of bubbling, which is hydrogen evolution.) Some metals have different driving forces (voltages) so it is expected that some metals will corrode less than others. We also see that the corrosion rate will depend on the reduction reaction. If no electrochemical reduction could occur, no corrosion could occur. Typical reductions that accompany the corrosion reaction are proton (or water) reduction to form hydrogen (as mentioned above), and the reduction of oxygen to form water.

One last point should be made concerning corrosion reactions. Since voltages are involved in the driving force for corrosion, we can change the voltage of a given metal to affect a change in the corrosion rate of the metal. One way to change the potential of a metal is to connect it to another metal that has a different potential. This is referred to as galvanic coupling.

In summary, we see that the rate of corrosion depends on the type of metal and the environment of the metal. The environment of the metal includes any coupling to one or more metals, the chemical composition, and the temperature.

DISCUSSION OF ELECTROCHEMICAL RESULTS

Galvanic Series:

In order to predict how metals might tend to corrode we first establish their potentials in solutions of interest. **Figure 4** shows the potentials of silver, nickel, copper, tin, cadmium, aluminum, and zinc when immersed in salt water, de-ionized water, and the CASS test solution.

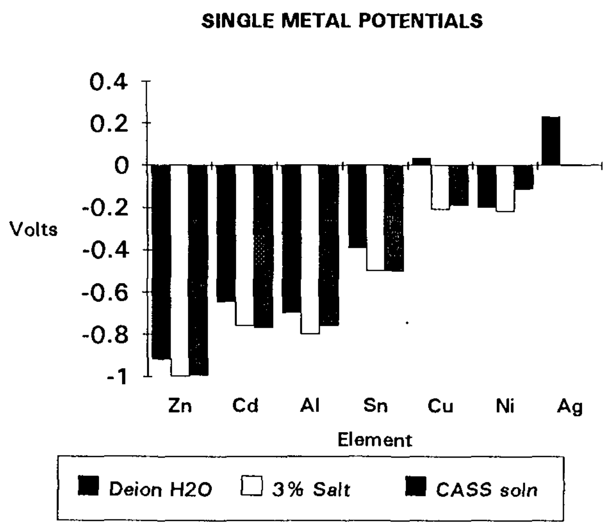


Figure 4: Single Metal Potentials with respect to calomel electrode, in various liquid immersions.

Potentials given are all relative to the saturated calomel reference electrode. The data show that the potentials for a given metal is relatively independent of the test solution. Metals that have relatively more positive potentials are considered "noble" and those with more negative potentials are more "active." More noble metals tend to corrode less than more active metals. From the chart, we see that zinc is the most active, aluminum and cadmium are next, tin is in the middle, nickel and copper are relatively noble, and silver is the most noble. It is very important to stress here that these potentials do not correlate directly into actual corrosion rates. It is entirely possible, for example, that tin and aluminum could corrode at the same rate in a given environment. The reason that the potentials do not correlate directly into corrosion rates is that voltage relates only to the driving force for the reactions. For any chemical reaction there are always activation barriers to overcome before the reaction can proceed. The actual rate of corrosion, therefore, depends not only on the voltage driving force but also on the magnitude of these activation barriers (called overvoltages). Each oxidation and reduction reaction has a different overvoltage and even the same reduction reaction occurring on differing surfaces has a different overvoltage.

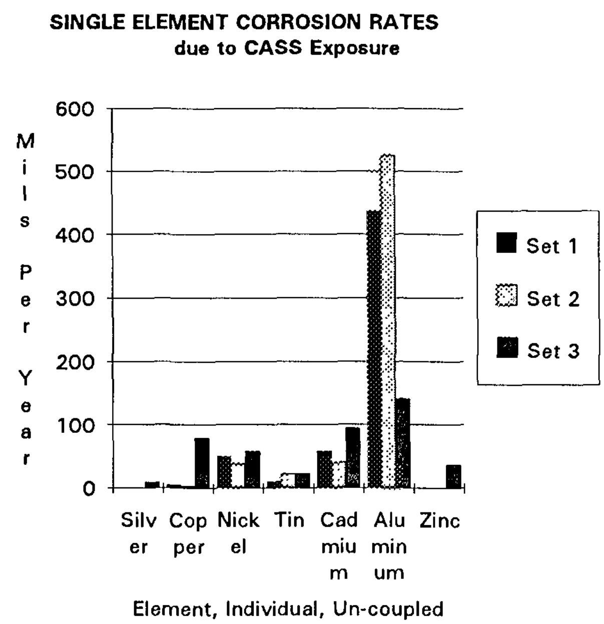


Figure 5: Single Element Corrosion Rates, CASS exposed, three different configuration sets.

Electrochemical techniques can be used to measure actual corrosion rates, but these measurements are not reported on here. What we have done here, instead, is to subject these samples to CASS solution and measure their corrosion rate by monitoring their weight loss with time (**figure 5**).

Galvanic Couples:

For galvanic corrosion to occur, two or more metals must be electrically connected and some form of electrolyte (eg. moisture) must be present. It is generally true that when two metals are connected the more noble metal will corrode less and the more active metal will corrode more than their respective free-corrosion rates. The solution potential of the metal couple will also fall between the respective potentials of the two metals. If more than two metals are connected, the analysis becomes more complicated. In a conventional F-fitting (**figure 6**) many metals are in contact in actual bimetal couples and in many cases more than two metal contact is involved.

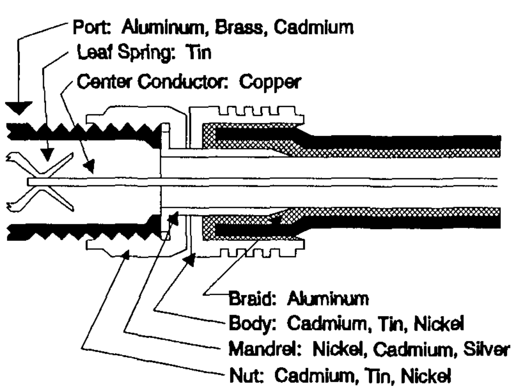


Figure 6: F-interface. The materials shown are those typically found at the surface of the given parts. Base metal of connectors is usually brass (copper, zinc, lead).

Figure 7 shows the potentials we measured in CASS solution of galvanic couples between Al-Cu, Al-cadmium, Al-tin, Al-nickel, cadmium-nickel, Cd-tin, and copper-tin. In the case of aluminum bearing couples, the potential often ends up close to the individual potential of the Al by itself. This does not necessarily mean, however, that Al is corroding at the same rate as it would in the absence of the other metal. Recall from the above discussion that it is the potential and the activation barriers that control the rate of corrosion. The activation barrier for the accompanying reduction reaction could be lower on the metal couple than on Al itself, thereby increasing the corrosion rate. We have tested Al galvanic couples in CASS solution. Unfortunately we have obtained a wide degree of scatter in the results (averages shown in **figure 8**). Qualitatively, the corrosion rates follow the predictions from the potentials. That is, aluminum corrodes fastest when coupled with copper, nickel, or silver, significantly less with tin, and less still with cadmium.

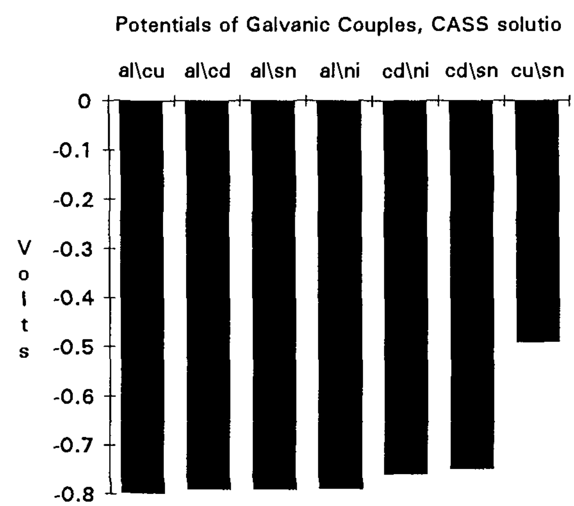


Figure 7: Potentials of Galvanic Couples when immersed in solution used to create salt spray in CASS test.

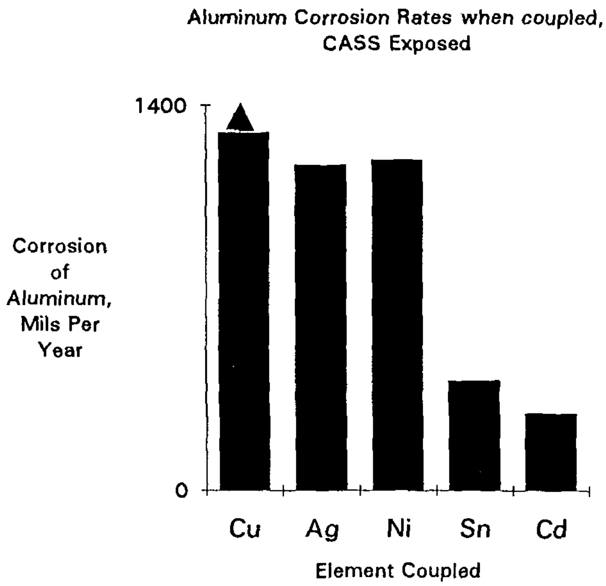


Figure 8: Corrosion Rates of Aluminum when coupled to typical CATV metals, CASS exposed. Average of wide scatter.

These findings also follow results obtained from previous work done by Mansfeld et al using 3% salt water immersion tests [4]. Similar to results here, those tests showed aluminum corroding with elements in the following descending order, silver > copper > 4130 steel >> stainless steel = nickel > Inconel 718 >> Ti-6Al-4V = Haynes 188 > tin > cadmium.

The data gathered and findings here show that both individually and more so when coupled, aluminum corrodes at a much higher rate than these other elements. It is clearly the weakest link in the interface, as is often witnessed when a cable is slit open at the rear of a corroded F-connector in the field. It is therefore imperative to keep moisture from reaching the cable braid. If this is not possible, the use of more corrosion resistant materials and combinations of materials is desirable. It is often inappropriate to use noble materials, as contacting materials must be electrochemically similar to minimize deterioration.

CASS ELECTRICAL PERFORMANCE TEST PROCEDURE

The practical evaluation of this report consists of exposing samples to the CASS environment and evaluating the results. First, the electrical performance of the F connector interface versus time in an aggressive environment (CASS) is determined. This is followed by an analysis of the corresponding physical material degradation.

Twenty samples of standard hex crimp connectors were tested for electrical degradation vs time of CASS exposure. The sample configuration, shown in **figure 9,** consisted of two hex crimp F-connectors, spliced together with an F-81 barrel splice. Each splice assembly was mounted securely to stabilizing boards. The splice was connected to five feet of quad cable on either side each of which extended outside of the CASS environmental chamber. These non-conditioned cable ends were used to measure the electrical characteristics of the internal sample.

CASS testing was conducted over a 56 day period, with electrical performance measurements taken at graduated intervals. Data was collected after 1, 2, 4, 7, 14, 28, and 56 days respectively. Electrical characteristics examined were signal egress, contact resistance, and signal transmission. The line diagram for each of the measurements are shown in **figures 10, 11, and 12** respectively.

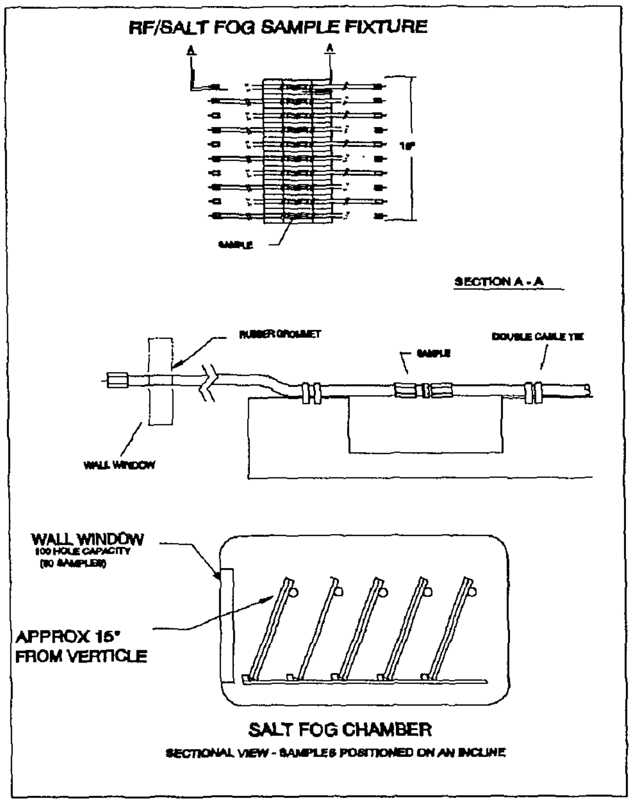


Figure 9: Layout of CASS chamber and configuration for sample stabilization.

Signal Egress

Signal egress measurements, **figure 10**, were taken using an HP 11940A close-field probe. The probe is a balanced magnetic field sensor which provides an output voltage proportional to the strength of the magnetic field at its tip. When the probe is held close to the sample, the radiated field magnetically couples to the probe and produces a larger output voltage. This frequency-preserved voltage is amplified by the pre-amplifier and is read by the spectrum analyzer for display.

Egress measurements were probed at the rear of each of the two connectors, A and B, where the jacket just meets the connector. The chamber was temporarily turned off and the lid removed for ventilation while taking egress readings. Samples remained otherwise undisturbed through the duration of tests. Other measurements were conducted outside of the chamber whereby the cable terminates externally.

The voltage values were read into a PC using a program which reads, stores, displays, and charts results. The frequency range taken was between 0 and 1000 Mhz. Egress levels are displayed in dB micro volts. The levels shown are good relative indicators and are not calibrated to absolute values. Typically an extremely corroded sample exceeds 20 dB microvolts higher than when pre-exposed. Nearly all samples remained just within this level up to 56 days.

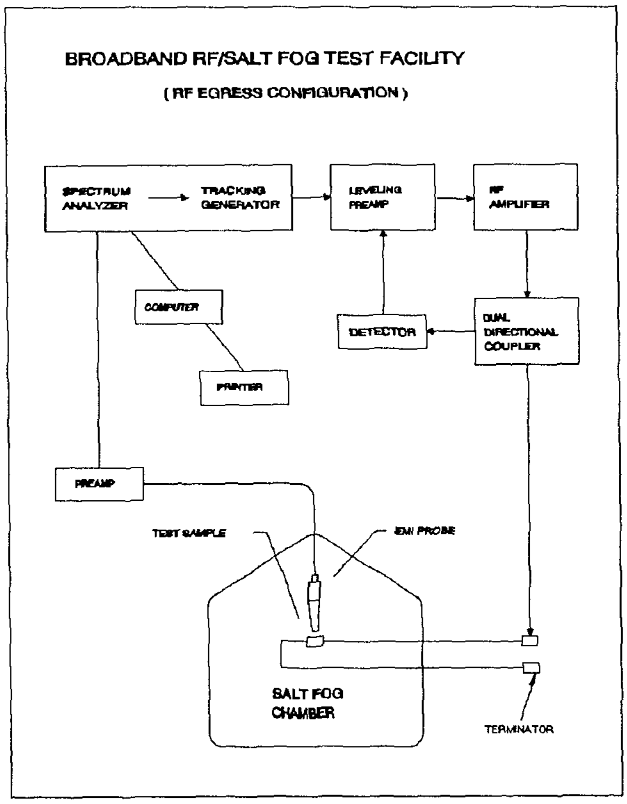


Figure 10: Configuration of equipment for signal egress measurements.

Contact Resistance

Contact resistance measurements, **figure 11**, are taken at the externally located cable ends. The Cambridge Technology Model 510 Micro-Ohmeter was connected to each of the two outside connectors and the contact resistance was displayed and recorded.

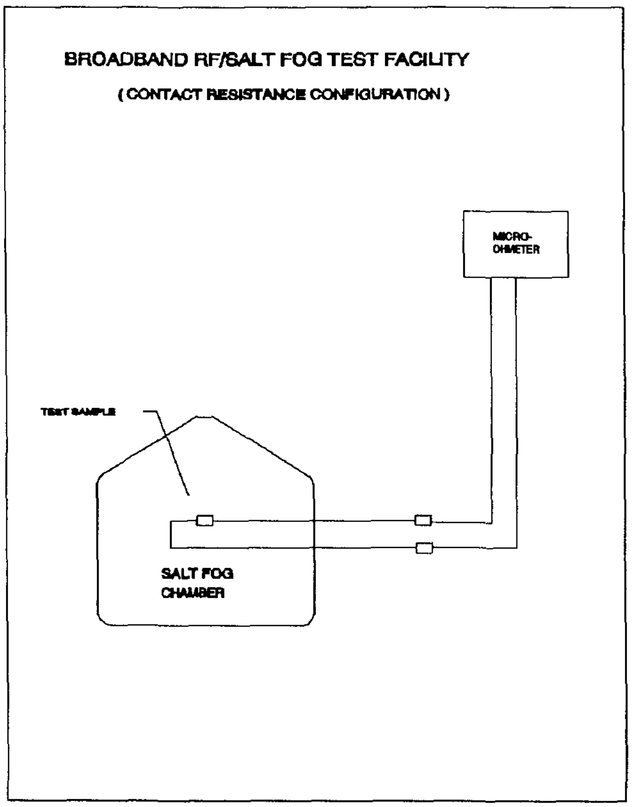


Figure 11: Contact Resistance measurement setup.

Signal Transmission

Signal transmission is measured with the equipment shown in **figure 12.** Radio Frequency signal is provided by the internal tracking generator of the HP 8590B Spectrum Analyzer. The signal was subsequently amplified by the Amplifier Research Amplifier and pre-Amp. A feedback loop was constructed to keep the swept signal in range and consistent. This signal transmitted externally into the drop sample and received directly by the spectrum analyzer. The results were automatically read, stored, and displayed by computer. Again the frequency range was from 0 to 1000 Mhz. Signal level is shown in dBmV.

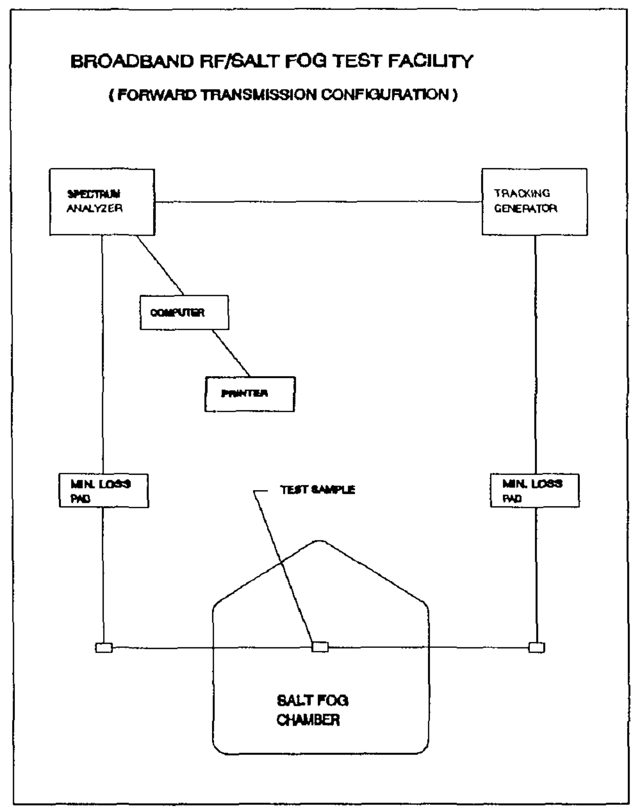


Figure 12: Setup for signal transmission measurement.

MICROSCOPY AND MATERIAL CHARACTERIZATION PROCEDURE

After samples were tested for electrical performance degradation during CASS exposure, samples were examined for corresponding moisture paths and material change.

SEM

The scanning electron microscope was used to take electron generated micrographs for the purpose of showing paths of moisture salt spray product deposition. Note that the presence of salt spray products doesn't necessarily mean that significant corrosion has occurred. It does, however, show that an electrolyte is provided. Actual corrosion is evident when material has been extracted (ie plating goes away exposing the base metal).

X-Ray Spectroscopy

X-ray spectroscopy was used to characterize surface materials. This method determines elements through analysis of irradiated electron wavelengths. Elements found on a surface are represented by the peaks of the energy spectrum shown in **figure 13**.

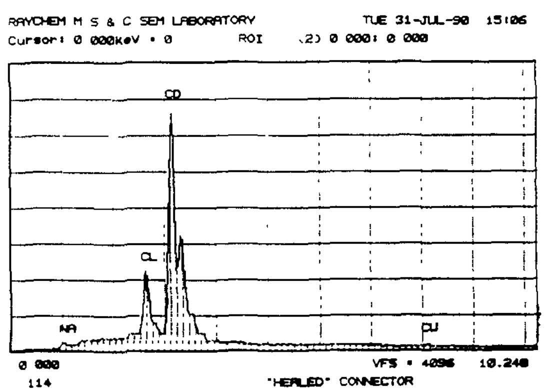


Figure 13: X-Ray Spectum of elements at surface of an F-connector.

This spectrum shows the deposited CASS products on a 3-day exposed sample. Similar technology, Energy Dispersive X-Ray Spectroscopy, was used to map out the presence of a particular element within a chosen surface area **(figure 14)**. This allows one to determine the locations of material degradation (actual corrosion), particularly where plating has been removed and base metal is exposed.

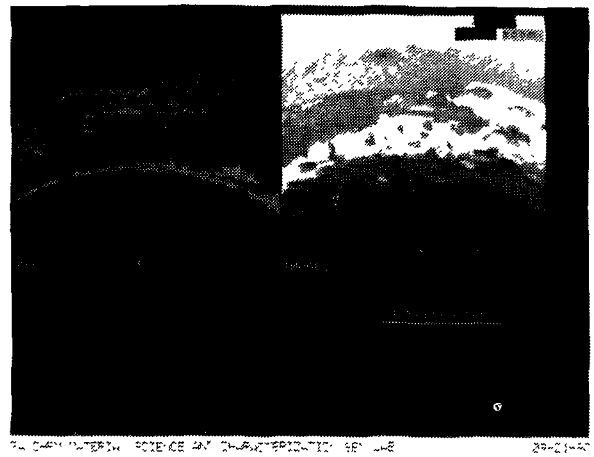
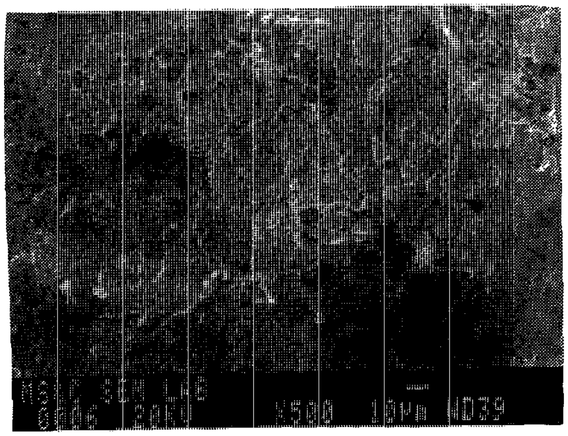


Figure 14: Energy Dispersive X-ray mapping of connector mandrel surface. Bright surface on left represents element of concern, in this case Copper, existing at surface.

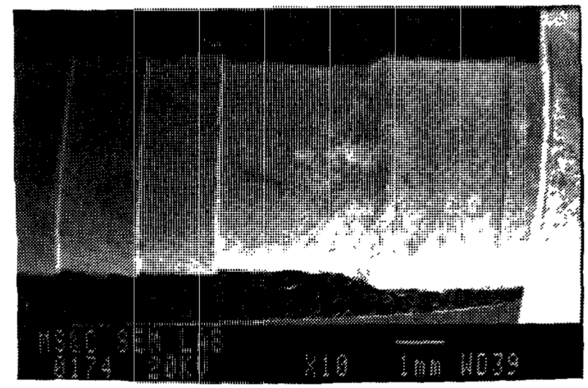
The SEM micrographs give an excellent representation of surface discontinuity and depth of field, **figure 15**. By initially making cursory observations of many contact areas of the corroded samples, key areas of material degradation were determined. This was followed by low magnitude, broad x-ray mapping of these areas to give a useful material profile.

Areas found to be of greatest concern 1) the mandrel post, 2) the mandrel face which contacts the F-81, 3) leakage paths (the internal wall of the crimped component, the nut threads, and the swivel joint). The two former areas are those which make and must preserve outer conductor contact.

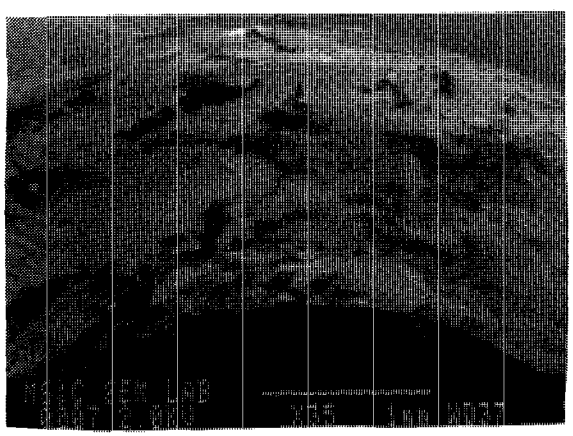
Figure 15
Scanning Electron Micrographs
Connector Parts, 56 Days



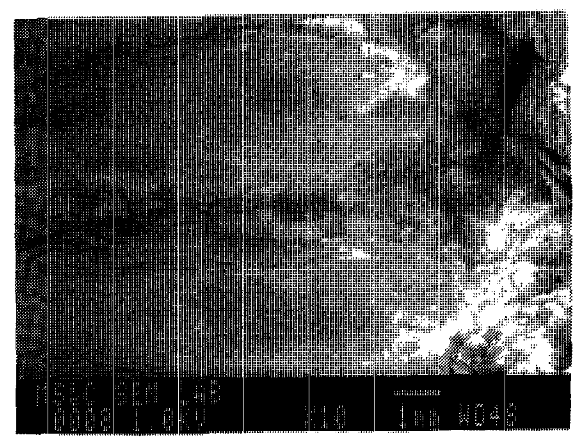
Mandrel Post, At Midlength, 500x



Mandrel Post

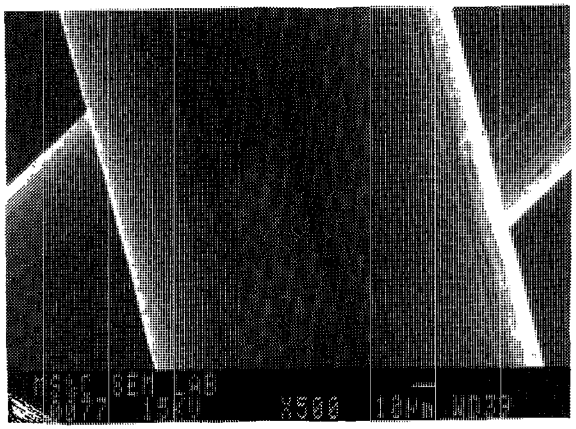


Mandrel Face

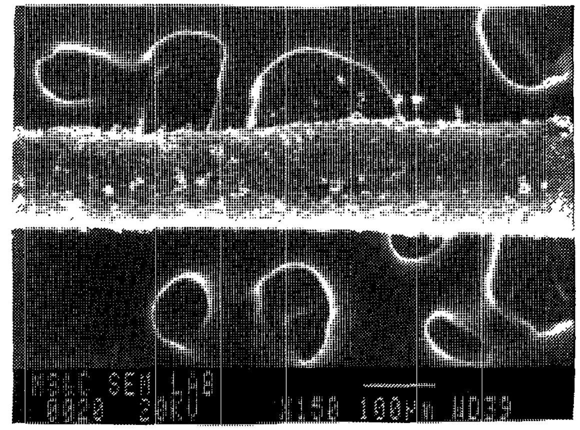


Inner Crimp Wall

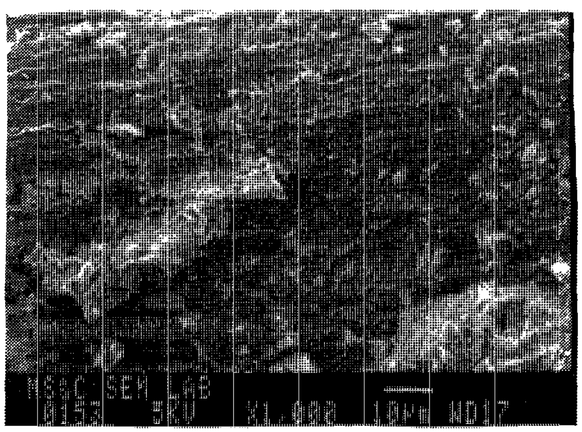
Figure 16
SEM, Aluminum Braid
Pre vs 56 Day Exposure



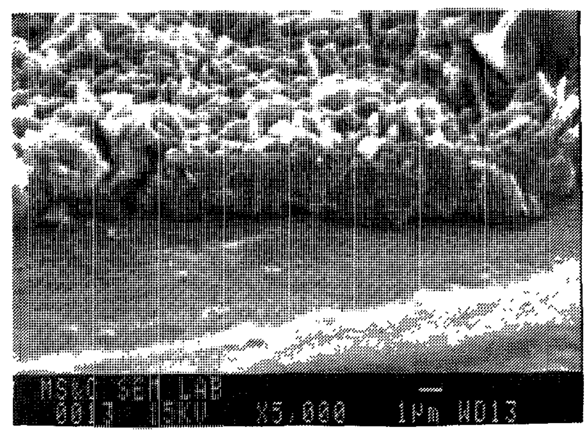
Braid, Pre Exposure 500x



Braid, After 56 Days CASS, 150x



Braid, After 56 Days CASS, 1000x



Typical Crystal Structure Due To CASS, 5000x

Rear of Connector

Micrographs of connector components, **figure 15**, show that much of the moisture and salt spray deposition into the rear of the connector has occurred along the crimp wall where the metal has been bent during crimping. These crimp bends have provided a channel for moisture to migrate. SEM's suggest that the moisture spends little time in these 'channels', depositing generously into the base of the post and moving beyond, towards the post/braid. Upon electrolyte arrival to the aluminum braid, **figure 16**, the largest degree of corrosion is observed.

Front of Connector

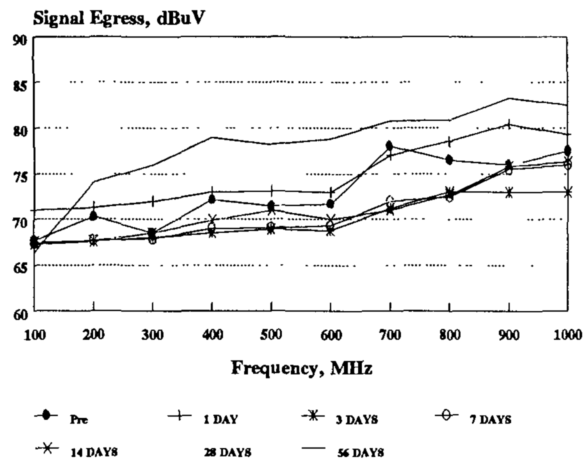
Little moisture has succeeded in migrating through the front of the connector during these tests. As seen in the SEM's of the nut, moisture has migrated approximately to the middle of the nut threads but little beyond. A significant amount of salt spray solution migrated to the mandrel face via the swivel joint. Although the mandrel face showed a fair degree of chloride buildup from incoming moisture, it was not significant enough to break contact or migrate to the center conductor. As a result the center conductor itself in samples of this test showed no signs of moisture or corrosion.

DICUSSION OF RESULTS: ELECTRICAL AND MATERIAL

The electrical characteristics of the samples varied with respect to time but did not show continuous degradation, (**figure 17**). The samples initially showed a decrease in signal leakage after the first 24 hours. This initial improvement may be explained by 1) settling of the contacting materials, providing an improved contact of metals, 2) electrolytes migrating within the swivel and other areas, **figure 18**, improving shielding due to better electrolytic continuity within interstices, and as a result 3) an improved impedance match is attained.

Signal Egress, CASS Exposed Samples

As Measured with Near Field Probe



Signal Egress, CASS exposed Samples

As Measured with Near Field Probe

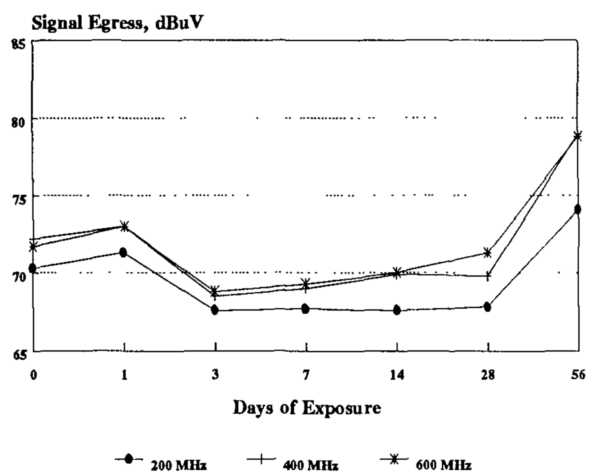
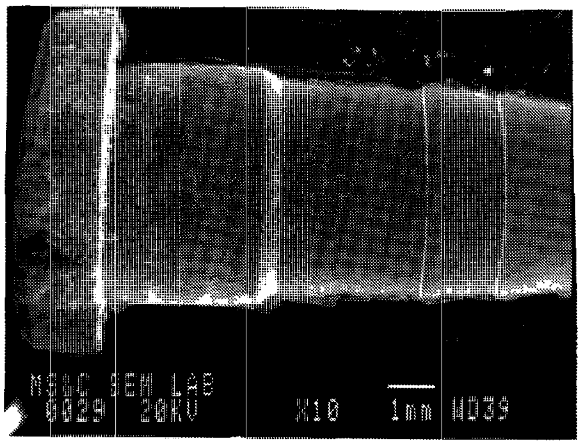
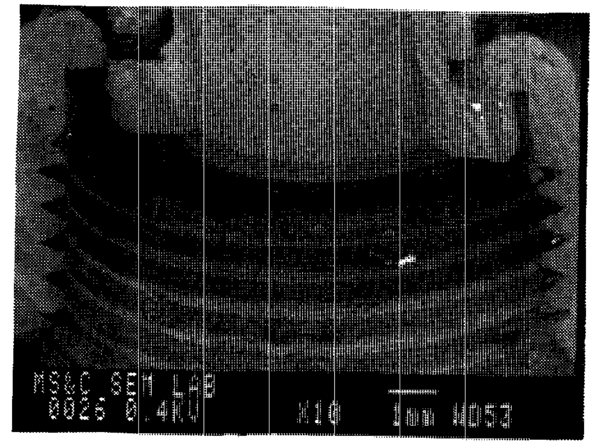


Figure 17: Profile of signal egress; above categorized by frequency, below by length of exposure.

Figure 18
SEM, F-interface Components, Improved Shielding
3 Day Exposure

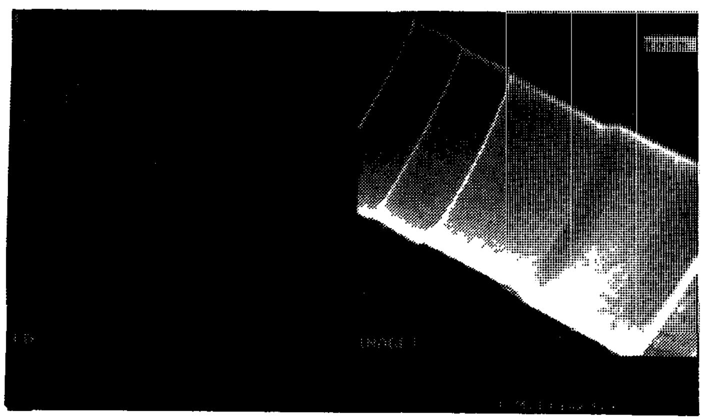


Mandrel Post

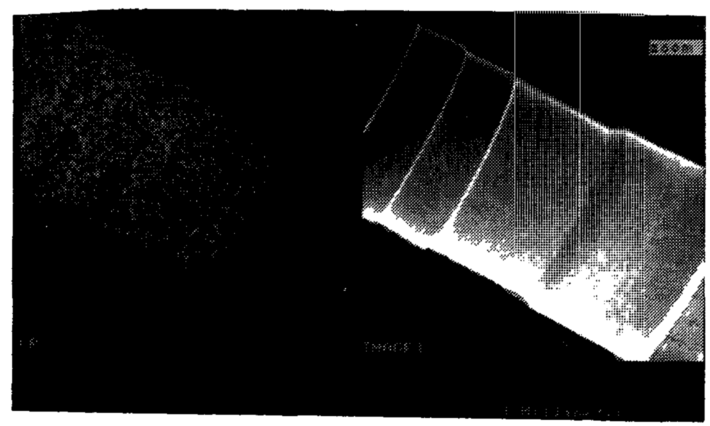


Post/Nut Interface

Figure 19
X-Ray Map Material Characterization, Connector Mandrel Post
Pre-CASS

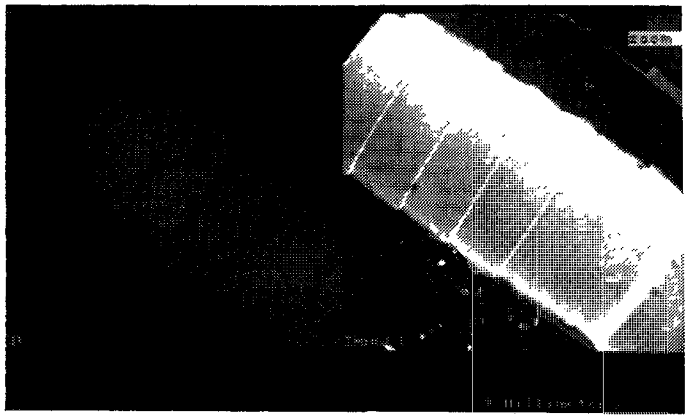


Cadmium

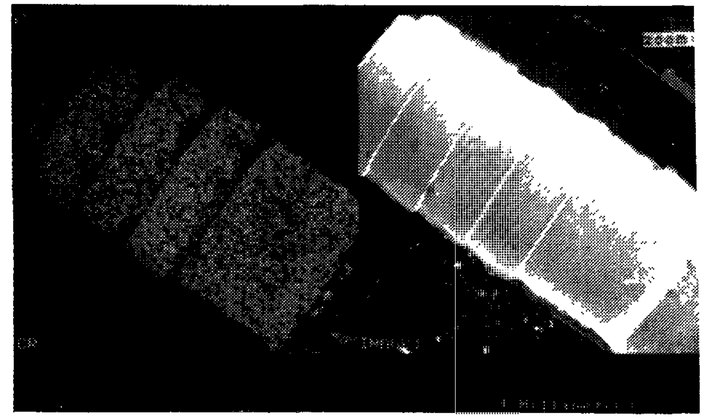


Chromium

Figure 20
X-Ray Map Material Characterization, Connector Mandrel Post
3 Days CASS

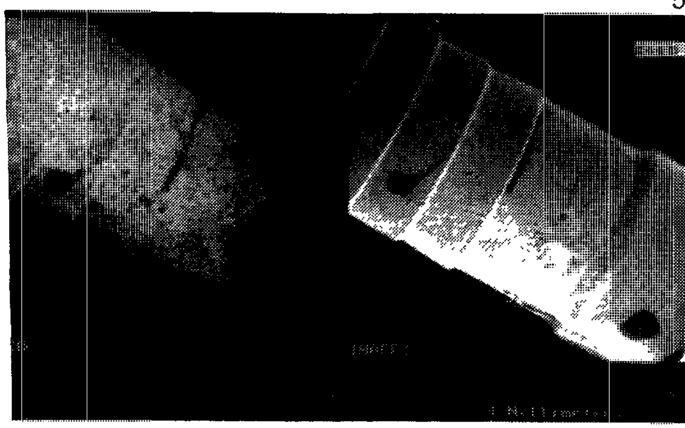


Cadmium

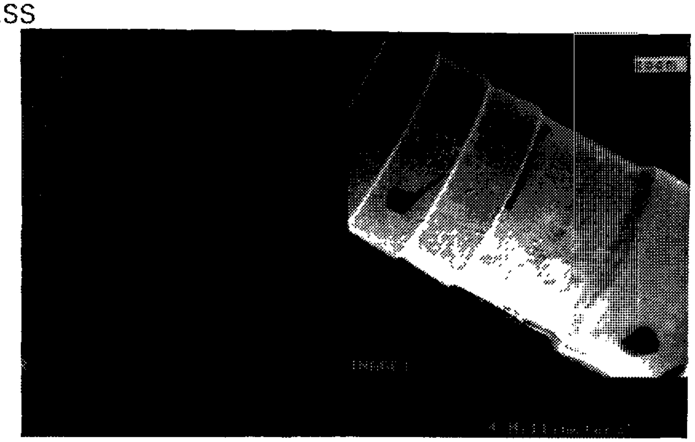


Chromium

Figure 21
X-Ray Map Material Characterization, Connector Mandrel Post
56 Days CASS



Cadmium



Chromium

The impedance, directly related to shielding effectiveness, is inversely proportional to capacitance. The capacitance is affected by any changes in the dielectric constant of substances between current carrying surfaces.

Results suggests that the moist CASS solution enters the interstitial voids in the F-interface, greatly reducing the dielectric constant between metal surfaces (eg. cable braids mandrel post). Hence the impedance, and resulting signal egress decreases significantly.

This improved shielding occurs during the very initial stages of corrosion, that is, before significant amount of material has left the key metal components such as the mandrel post, **figure 20.** From the x-ray mapping we see that the post has maintained material character, good cadmium and chromium plating coverage, similar to the pre-exposed sample, **figure 19.**

As the metal components continue to corrode at key junctions, any improvements in electrical performance due to initial material settling and electrolyte deposition begin to be counter balanced by loss of contact due to corrosion at dissimilar metal interstices. This crossover occurs somewhere between 28 and 56 days, **figure 17.**

Finally, signal egress has degraded significantly after 56 days. Upon inspection of samples exposed to this time, **figure 21,** we begin to see that a significant degree of the cadmium plating on the connector post has gone away.

The corrosion of the aluminum which interfaces with the post, as is expected based on data from electrochemistry experiments, was quite extreme. When examining the x-ray maps of the corroded post, we see some of the plating on surfaces has gone away after 56 days. However, a relatively great amount of the aluminum braid has degraded in this time. This is witnessed by the micrograph, **figure 16,** which shows a braid section from one of the CASS-exposed connectors. The sample has corroded from an original diameter of approximately .0062 inches to less than .0052 inches (a volume decrease of 30%) at some points. Even if the entire plating depth of the connector post (cadmium and chromium) had corroded, it represents less than 1% of the aluminum's corrosion penetration. This follows quite well with what we would expect based on the bimetallic and corrosion potentials performed in electrochemical tests, which show aluminum as the most corrosive element in the system.

With such severe material degradation, continuous longitudinal and circumferential contact is lost. The effect was a substantial increase in signal egress figure 17 by 56 days. The outer conductor contact resistance, **figure 22,** degraded less dramatically, as some degree of contact, at mandrel face and somewhat at braid/post, was maintained.

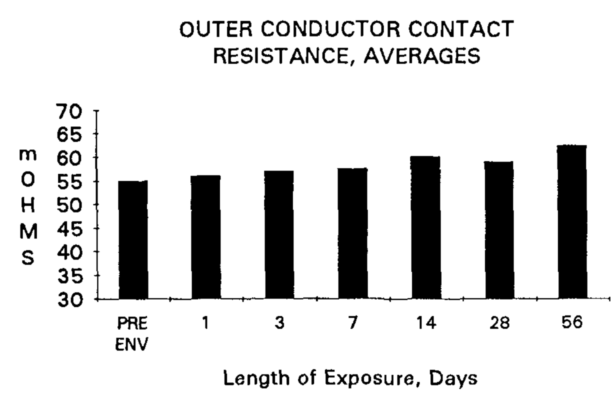


Figure 22: Outer Conductor Contact Resistance.

The mandrel did show some corrosion, **figure 14,** but near field probe measurements show less signal leakage at that junction.

No sign of corrosion degradation was found at center conductor/leaf spring interface. As a result, the signal transmission and inner contact resistance integrity was nearly maintained **(figures 23 and 24).**

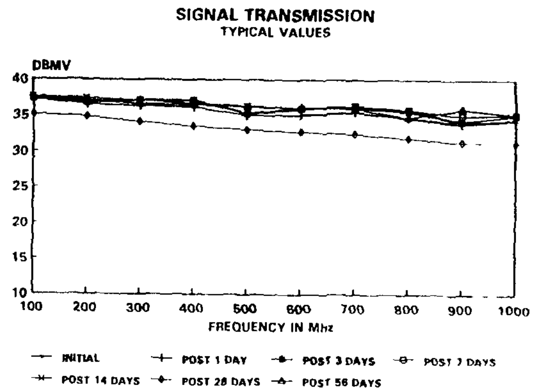


Figure 23: Signal Transmission

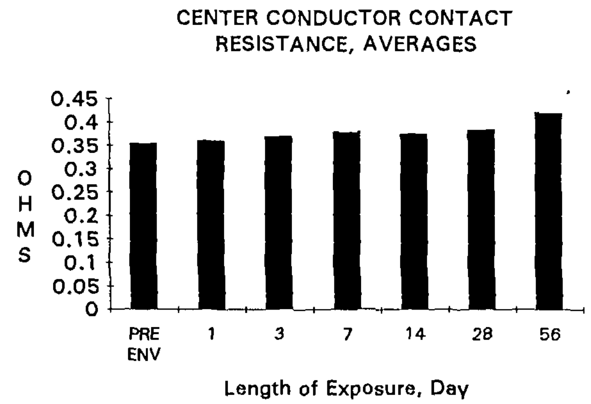


Figure 24: Inner (Center) Conductor Contact Resistance

As compared to connectors tested in previously unpublished CASS results, samples in this program corroded and degraded at a much slower rate. This can be attributed to lower fretting corrosion [2] as is known to occur on many electrical contacts which are typically not held stable.

One of the phenomena which can be attributed to fretting is the nature of samples to 'heal' over time. This has been witnessed often in the field, whereby a sample begins with a particular shielding effectiveness, the signal degrades, improves again, and so on in a discontinuous fashion. Fretting corrosion, and the removal of fretting corrosion products, has been suggested as a way in which discontinuity in signal degradation can occur [2], **figure 25.**

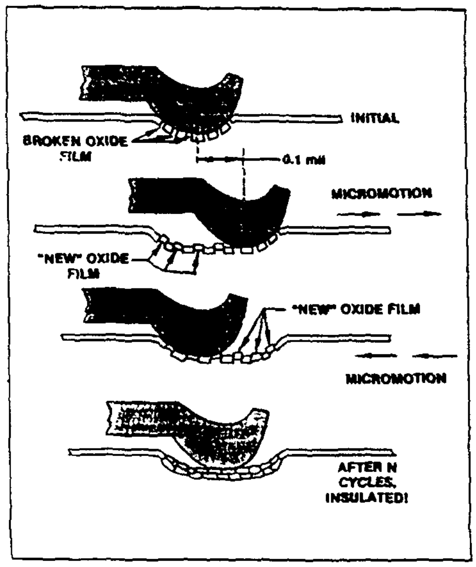


Figure 25: Process of Fretting Corrosion

Fretting occurs when a metal contact area corrodes, micromotions re-expose corroded surfaces, these areas develop additional oxide films/ corrosion products, cycling until contact is lost. The 'healing' experience, unfounded in this test program, occurs in the field when 1) removed upon larger vibration or impact, creating better shielding as metal to metal contact is almost restored and 3) signal degrades again through general and fretting corrosion. This may cycle until contact can no longer be restored. In this series of tests, healing has been successfully eliminated through means of stabilizing samples.

SUMMARY/CONCLUSIONS

1. Stabilized samples show an initial slight decrease in signal leakage followed by gradual increase through 56 days CASS exposure. Signal Transmission and Contact Resistance more gradually degrade over this time frame.

2. Initial decreased signal leakage, occurring after 24 hours, corresponds to wetting of the many contact surfaces, with no significant sign of corrosion. Samples gradually corrode and are quite degraded by 56 days. Key paths of moisture ingress and corrosion, are along the crimp bends and into the swivel. Moisture is deposited at the base of the post, which finally wicks into the braid/post interface. Some post material was degraded and a large amount of the aluminum braid was corroded after 56 days.

3. The CASS chamber solution, when used as an immersion bath for corrosive drive (potential) measurements, is qualitatively quite comparable to that of 5% salt water, and deionized water. The metal elements of concern fall within the same ranking for each of the solutions.

4. The corrosion of aluminum is significantly higher than other F-interface elements. The single metal corrosion rates in CASS are, in decreasing order; aluminum >> cadmium > Nickel~Tin > Copper~Silver. In the CASS enviroment, the aluminum corrosion rate decreases with element coupled in the following order; Copper > Nickel ~ Silver > Tin > Cadmium.

RECOMMENDATIONS

In environments for which the F-interface may be exposed to moisture, the following is suggested. These recommendations are ideal from a scientific point of view, but do not take into account other factors such as economics and tradition.

1. Use braid materials which are both less generally corrosive and are more compatible with contacting surfaces.

2. Use moisture sealing methods in order to repel electrolytes and hence minimize corrosion rates.

3. Mechanically stabilize the cable surrounding the interface to limit fretting corrosion.

4. Implement all of the above while using materials which are both more noble *and* are electrochemically more compatible (eg; Copper braid/ nickel plated post or silver plated copper braid/silver plated post).

ACKNOWLEDGEMENTS

The author wishes to acknowledge the contribution of Gary Trost, PhD. Electrochemistry, of Raychem Corporation, in providing guidance through the laboratory electrochemical investigation.

REFERENCES

1 **Corrosion of Resistance of Metals and Alloys**, LaQue and Copson, see lit search.'Corrosion Resistance of Metals and Alloys', LaQue and Copson, Reinhold Publishing Company, New York.

2 **Corrosion of Electrical Connectors**, Davis,Abbott, and Koch, Battelle Columbus Division; Metals and Ceramics information Center; Columbus, Ohio; 1986. p. 29.

3 **Corrosion and Shielding of CATV Connectors**, "The Evaluation of Ingress and Egress Problems in the CATV Sub Low Frequency Spectrum", Cablesystems Engineering, Broadband Communications Engineers, Section II, #18ST-36100-8-1367, 1980.

4 **Galvanic and Pitting Corrosion-Field and Laboratory Studies**, "Laboratory Studies of Galvanic Corrosion of Aluminum Alloys", Mansfeld, Florian and Kenkel, J.V., ASTM STP 576, American Society for Testing and Materials, 1976, pp. 20-47.

5 **Corrosion Engineering**, Fontana, M.G., and Green, N.D., McGraw-Hill, New York, 1967.

6 **CATV Distribution Equipment Corrosion**, McCaughey and Rogeness, Technical Papers NCTA 22nd, Anaheim, 1973.

7 **Measurement of Intermodulation Products Generated by Corroded or Loose Connections in CATV Systems**, Kellar, Bradford S., Technical Papers NCTA 33rd, 1984. NCTA Washington DC, p 23-28.

8 Personal Communication from Bradford S. Kellar, Raychem Corporation, Menlo Park, CA, 1990.

9 **The New Metals Handbook, Volume 13, Corrosion**, ASM International, Metals Park, Ohio, 1987.

10 **The Corrosion of Copper, Tin, and Their Alloys**, Leidheiser, Henry, Robert E. Drieger Publishing Company, Huntington, NY, 1979.

11 **Atmospheric Corrosion of Metals**, ASTM Special Technical Publication 767, ASTM PCN 04-767000-27, Philadelphia, PA, 1980.

12 **Corrosion Handbook,** Uhlig, H.H. (editor), Wiley, New York, 1948.

13 **Corrosion**, Shreir, L.L. (editor), Volume 2, Corrosion Control, Wiley, New York, 1963.

14 **Handbook on Corrosion Testing and Evaluation**, Ailor, W.H. (ed), Wiley, New York, 1971.

15 **Aluminum**, Van Horn, K.R. (ed), American Society for Metals, Metals Park, Ohio, 1967.

16 **Recommended Metal Combinations to Avoid Galvanic Corrosion**, Veilleux, E.D., Machine Design Data Sheets, 1960.

17 **"CLI and Its Effect on Aeronautical Navigation"**, Warren, Larry, CED, CLI Handbook, 1989.

FIBER OPTIC CABLE DESIGNS ADVANTAGES AND DISADVANTAGES

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Comm/Scope, Inc.

ABSTRACT

Although fiber optic cable is relatively new in the CATV industry it has been a commercial venture in the telephony industry for over ten years. This mature fiber optic cable industry offers a number of cable designs for different applications.

This paper presents an objective view of the different fiber optic cables being offered to the CATV market and the advantages and disadvantages of each. Different basic designs such as loose tube, central core tube, slotted core, and tight buffer are discussed. The advantages and disadvantages of dielectric vs. armored, and steel bearing cable are also examined. In addition a short discussion on future developments in fiber optic cable design is presented.

The methodology used in this paper is to examine, in detail, published specifications and papers and then attempt to present a one to one comparison of the different cable designs. Issues such as environmental, mechanical, and physical specifications are presented as well as field issues and how these pertain to the mostly aerial outside plant of this industry.

The results show that for different applications different optimum cable designs exist. Therefore, at this point in time, there seems to be no one optimum cable design for the CATV industry.

FIBER OPTIC CABLE

Design Objectives

The design objectives in fiber optic cable are fairly simple. The first concern of the cable designer is to protect the glass fiber from the outside environment. The fiber must be protected from the physical rigors of being installed and placed for up to 20 years in the outside plant. These include forces such as impact, tensile, twist, and compressive loads. In addition the fiber must be protected from any moisture. The fiber itself is degraded by moisture and if water were to get into a cable and freeze, it could physically crush the fiber. Probably the most critical design parameter is temperature performance.

The typical specified operating temperature range of fiber optic cable is from -40 to +70 degrees Celsius. The design problem is that the fiber has a coefficient of expansion on the order of 10^{-7}, while the majority of the plastics used in fiber optic cable design have coefficients of expansion on the order of 10^{-5}. Therefore, when the cable is subjected to temperature extremes the plastics expand and contract 100 times more than the glass fiber. If the fiber optic cable is not designed correctly this coefficient of expansion differential could impart forces onto the fiber which would manifest as drastic increases in attenuation or, in the extreme case, fiber breakage.

The cable designer offsets this differential in coefficient of expansions with high modulus, low coefficient of expansion materials such as fiberglass reinforced plastics and steel. The cable designer gives room for the fiber to collapse and expand like a spring by placing it in a loose tube.

In addition to the above technical design problems the fiber optic cable must be easy for the craftsman to work with. It should be easy to access and identify the fibers, as well as lightweight and small.

Fiber Optic Cable Designs

There have been a number different solutions to the design problems discussed above. For the purpose of this paper, tight buffer, slotted core, loose tube and central core cables are discussed. But because loose tube and central core cables are the products being offered to the CATV industry, the comparison sections of the paper will be limited to those two designs.

Tight buffer cables are called so because a layer of plastic is extruded directly onto the acrylate coated fiber, thereby creating a tight structure around the fiber. The advantages of such a design are in handling. Each fiber unit is larger, and less sensitive to handling mishaps because individual fibers have a relatively thick plastic protection covering them. Although these products have been used in outside plant environments, largely in the past by the Japanese, they are not well suited for those applications. The first problem is that whatever compressive or tensile forces are experienced by the cable are also experienced by the fiber. This means a large amount of high modulus, low coefficient of expansion materials, such as steel and aramid yarn, must be used in order that the fiber not see high strain levels. In addition, tight buffer cables become comparatively large and difficult to design when fiber counts exceed 24.

Slotted core cable is used great deal overseas and was used initially to some extent in North America. Slotted core cable consists of a cylindrical plastic core with longitudinal slots cut into it. The fibers are then placed into these slots. (See figure 1). After the fibers are placed into the core any number of a variety of armors and jacket layers can be applied. In some applications this design fell out of favor because of the difficulty in handling when the jacket was stripped off. Also the difficulty in filling and placing the fiber into the slots made for an expensive product. Slotted core cable combined with fiber ribbons are again gaining some popularity, especially in Japan, due to the high density of fiber that can be attained in such a configuration.

Loose tube cables are one of the two most popular designs offered in North America. One to 12 fibers are placed within a gel filled tube for protection. The tubes are then stranded around a dielectric or metallic strength member. The combination of the loose tube around the fibers and the stranding pitch of the tubes creates a tensile and contraction window. This window allows for the cable to contract and be elongated on the order of .3 % while imparting no stress on the fiber. The cable can therefore be designed such that at specified temperature extremes and tensile loads little or no strain is experienced by the fiber. This fiber optic cable core can then be protected by any number of different sheaths, depending on the application. This product has been very successful because of performance in the field and handling issues for fiber counts over 72.

The central tube fiber optic cable is the other popular design in North America. In this design the fibers are all encased in one large tube. The fibers are separated into groups either by ribbons or fibers bundled by colored ID threads. The ribbon design is applicable for high fiber count cables that are being put into systems that do not require low splice

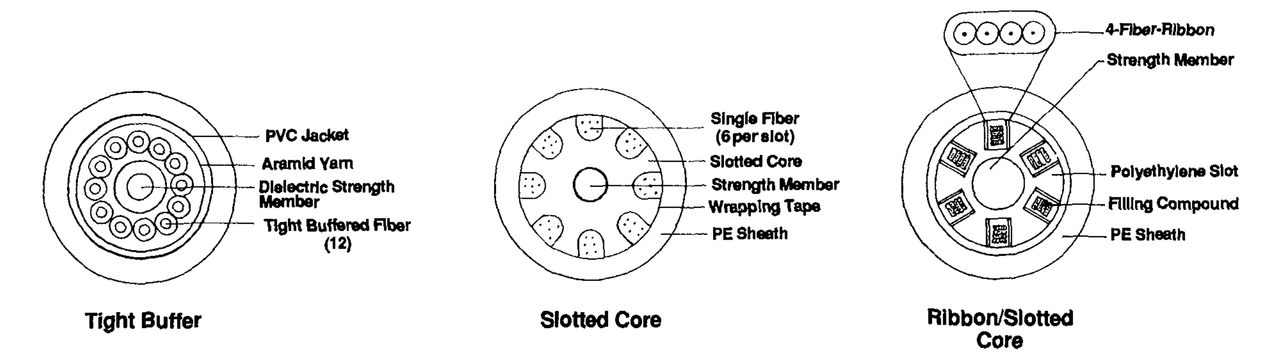


Figure 1

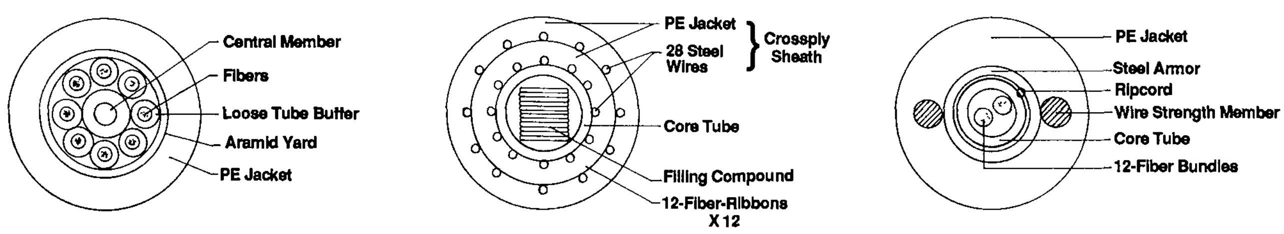


Figure 2

losses. These cables can be shipped pre-connecterized with easily used array splices, although the losses of the array splices can be sporadic and relatively high for single mode fiber. The fiber bundles have up to 12 fibers per bundle. Each individual fiber and binder thread is color coded. In fiber counts higher than 72 it can sometimes be difficult to manage all the fibers in one tube. In some sheath designs for this core a number of steel wires or small dielectric rods are used for strength and temperature compensation. These "crossply" sheaths are very environmentally stable but also very difficult to enter. A recent innovation to make these cable designs more user friendly is to armor the core and place either six dielectric or two steel strength members 180 degrees from one another longitudinally along the tube. After jacketing the cable, additional armoring and jacketing can be applied. This design, like the loose tube design, allows a contraction and tensile window for the fiber. Again, the cable can contract or elongate on the order of .3% with no effect on the fiber.

Because of field performance and ease of handling, the predominant cable designs being offered to the CATV industry are the loose tube and central tube type with bundled fibers. These two cable designs are themselves offered in a variety of different configurations. The remaining sections of this paper will compare and contrast these two cable designs and the different configurations of each.

Existing Fiber Optic Cable Specifications

There are a number of existing fiber optic cable specifications for telephone industry that are used for the CATV and other industries. The most common specifications are written by GTE, Sprint, REA, and Bellcore. Bellcore's TR-20 is in most cases the more comprehensive and difficult specification to meet. TR-20 covers cable qualification tests, material qualification, mechanical and environmental tests with allowable decreases in performance for each test. It is important to note that all measurement methods in Bellcore TR-20 are referenced to an ASTM or EIA-455 test procedure standard. Some tests that to date have no standards such as lightning and rodent tests are spelled out in detail in the document. A summary of the mechanical and environmental tests with their corresponding allowances are listed in the following tables.

Two of the more important tests mentioned above have no standard testing procedure per se: the lightning and rodent tests. Although specifications do not require that certain test levels be met, they do require that the tests be performed and the level of resistance reported. Each cable construction of the two designs being discussed must be sted in every one of these tests because the result depends upon the core and sheath construction. Typical classifications of the results of these two test are listed below.

It is important to note is that all suppliers of fiber optic cable to the telephone industry must meet these specifications in order to be a supplier. Consequently, the performance of any cable that meets Bellcore TR-20 will be about the same as any other

Mechanical & Environmental Tests			
Test	**EIA -455 Specifications**	**Mechanical Requirement**	**Optical Requirement**
Tensile Strength	FOTP-33	600 lb.$_f$ Bend Radius = 20x Cable O.D.	$\leq$.1 dB increase @ 1550 nm
Compressive Strength	FOTP-41	1000 lb.$_f$ Total Load	$\leq$.1 dB increase @ 1550 nm
Cable Twist	FOTP-85	± 180° Twist, 10 Cycle	$\leq$.1 dB increase @ 1550 nm
Low and High Temperature Bend	FOTP-37	Bend Radius = 15x Cable O.D. 4 Wraps ea. at -20° F, 140°F	$\leq$.1 dB increase @ 1550 nm
Cyclic Flex	FOTP-104	Bend Radius = 15x Cable O.D.	$\leq$.1 dB increase @ 1550 nm
Impact Resistance	FOTP-125	52 ft.- lb.$_f$ Impact, 25 Cycles	$\leq$.1 dB increase @ 1550 nm
External Freezing	FOTP-98	1 hr. min. freeze at -2° C	$\leq$.1 dB increase @ 1550 nm
Temperature Cycling	TR-20	-40 to +70° C 4 Cycles	100% $\leq$.2 dB/km increase 80% $\leq$.1 dB/km increase
Temperature Aging	TR-20	+85° C, 5 Days	100% $\leq$.2 dB/km increase 80% $\leq$.1 dB/km increase

Table 1

Lightning & Rodent Testing			
Design	**Construction**	**Lightning Resistance**	**Rodent Resistance**
Loose Tube	Steel Core No Armor	80KA	Poor
	All Dielectric	N/A	Poor
	Dielectric Core Armored	150KA	Excellent
Central Tube	All Dielectric	N/A	Poor
	Dielectric Core Steel Armored	105KA	Excellent
	Dielectric Core Copperclad Steel Armored	150KA	Excellent

Table 2

Cable Construction Comparisons				
Design	**Construction**	**Lightning Resistance**	**Rodent**	**Cost**
Loose Tube	Steel Core, No Armor	Poor	Poor	1
	All Dielectric	Best	Poor	2
	Dielectric Core, Armored	Excellent	Good	3
Central Core	All Dielectric	Best	Poor	2
	Dielectric Core, Armored	Excellent	Good	1

Design	Fiber Count	O.D. In.	Weight (lbs/kft)	Fiber ID
Dielectric Core, Armored	48	.49	105	Excellent
Loose Tube	96	.59	150	Excellent
Dielectric Core, Armored	48	.63	170	Excellent
Central Tube	96	.74	230	Good

Table 3

cable that meets that performance standard. Characteristics of handling, weight, lightning, and rodent resistance that are dependent on the construction of the cable should be considered, but each supplier has an option available to satisfy these requirements.

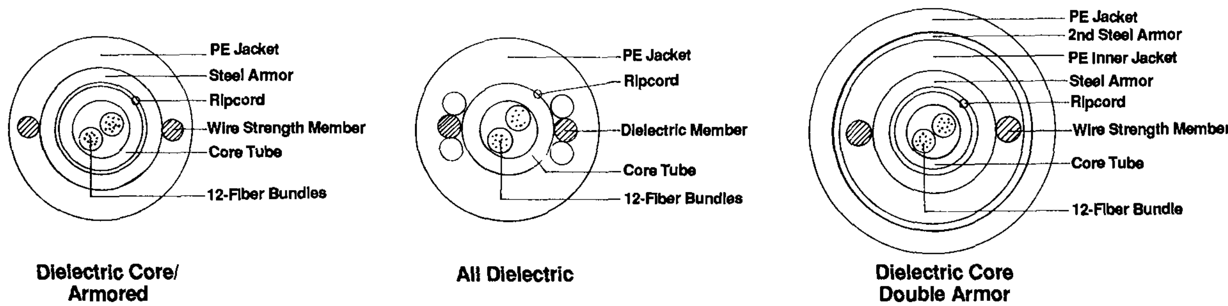
Comparison of loose tube and single tube constructions vs. specifications.

For each of the two designs being discussed there are a number of sheath designs. Each sheath design has cost/benefit trade offs.

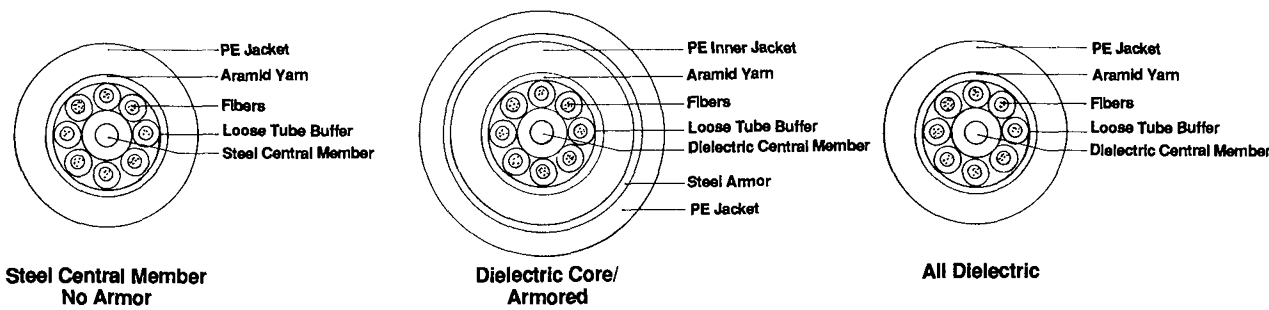
For the central tube cable there are basically two different cable constructions. Both constructions, by the definition of this design, have a dielectric core. This is important in the case of lightning protection. If the purchaser of the cable is concerned about lightning protection it is important that no metallic member be within the fibers such that a high current surge could short to that member and destroy the fiber in its path. The cable can have no metal in it at all if lightning is a serious concern or if grounding of any metallic members could be a problem. In this case the strength members in the cable would be some combination of aramid yarn, fiberglass roving and fiberglass reinforced plastic. (GRP or FRP)(See figure 3) On the other hand, an all dielectric cable has almost no protection against rodents.

When rodent protection as well as some lightning protection is desired, an armored version is available. In this case the dielectric core is surrounded by an armor. Strength members inside the armor are generally dielectric and those outside the armor can be metallic. If additional rodent or lightning protection is needed different configurations of armors and jackets can be used to give the necessary protection.

In the case of loose tube cables solutions to the above listed problems also exist. The most inexpensive loose tube cable made has a steel central member and no armor. This is a dangrous design in that it yields both poor lightning and rodent protection. (See figure 4) In a loose tube cable a dielectric core should be specified when lightning



Center Core Tube Constructions
Figure 3



Loose Tube Construction
Figure 4

is a concern. An all dielectric construction is completed with aramid yarn for strength and a PE jacket for protection. When rodents are a concern, an armor and additional jacket can be added. If both lightning and rodent resistance are desired, an armored cable with dielectric central strength member should be specified.

Since the performance of all cables meet the same specification, the only comparisons to be made between the two types of products are a comparison of what construction is best suited for each individual application. Table 3 shows a summary of the best options available and their relative costs based on material usage for the loose tube and single tube designs. Handling issues are essentially a matter of fiber identification, sizes, weights, and personal preference.

SUMMARY

All suppliers of fiber optic cable to the Bell system must meet TR-20 specifications. The product they sell to markets other than Bell companies do not necesarily meet all TR-20 specifications. Therefore, it is important that either a well written specification be submitted or an existing specification such as TR-20, or equivalent be referenced in a request for quotation. When another existing specification is referenced, any special considerations that may be required for CATV installation must be included, since all existing specifications have been written for the telephone industry with digital transmission in mind.

If the fiber optic cable meets the specification then the important issues are attenuation levels, lightning resistance, rodent resistance and personal preference. All of these issues are addressed equally well by different methods.

There is no one design best suited to the CATV market. Both central tube and stranded loose tube products meet the same specifications and each design has a construction that can meet the demands of almost any environment.

REFERENCES

Bellcore TR-TSY-000020 Issue 4

"Compact Lightguide Cable Design", P.D. Patel, C.H. Gartside, III, International Wire & Cable Proceedings 1985.

"Optical Stability of Single Mode Fiber in Loose tube cable designs", K.A. Emig, et al, IWCS Proceedings 1985.

"Qualitication Testing of Fiber Optic Cables", R. Ohlhaber, et al, IWCS Proceedings 1984.

"Multi-Hundred-Fiber Cable Compared of Optical Fiber Ribbons Inserted Tightly into Slots", S. Hatanoetd, IWCS Proceedings 1986.

"Lighting Considerations in Optical Cable Design", J.A. Olszewski, J.B. Masterson, IWCS Proceedings 1986.

"The effect of temperature dependent materials on Fiber Optic Cable Design", K.L. Coupe, et al, IWCS proceedings 1986.

"Qualification Procedure for Fiber Optic Cable Design", G. McKay, et al, IWCS proceedings 1986.

"A High Performance non metallic sheath for Lightguide Cables", C.J. Arrogo, A.C. Jenkins, P.D. Patel, IWCS proceedings 1987.

"LXE - A Fiber Optic Cable Sheath Family with Enhanced Fiber Access", P.D. Patel, M.R. Reynolds IWCS Proceedings 1988.

Fiber Optic CATV Transportation using combined PCM and VSB-AM Transmission
-ABSTRACT-

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1. Introduction

Fiber optic CATV transportation is achieving higher performances standards and lower cost with the utilization of the latest break throughs in VSB-AM and PCM technological. Studies have shown that VSB-AM fiber optic transmission provides an economical means for the transportation of high count CATV channels. Fiber Optic CATV transporation using VSB-AM trunk distribution in conjunction with PCM super trunking can provide a cost effective method of distributing high quality CATV signal to a large distribution area.

To support this hypothesis Sumitomo investigated the technical aspects of this transportation concept.

This paper summarizes the following experimental tests;

1. Multi-drop distributions of baseband PCM video signals from PCM fiber optic backbone.
2. Long distant PCM transmission of high quality video signals through multiple PCM repeaters.
3. End of line performance of a cascaded PCM, VSB-AM, and coaxial CATV transportation network.

We are also proposing high count base band video signal distribution using PCM equipment operating at 1.2 Gbps and 2.4 Gbps providing un-compressed transmission of high channel capacity per fiber distribution.

2. Features Of The PCM And VSB-AM Distribution Network

2-1. PCM Network

PCM transmission equipment time division multiplexes baseband video and audio signals. This technology provides:

(1). Repeater less transmission exceeding 50 km.
(2). Branching capability of optical signals through the use of optical couplers
(3). High transmission quality of video signals (60 dB weighted), independent of transmission distance and the number of repeaters.
(4.) Small size and lower consumption through the use of GaAs-LSI technologies.
(5). Easy installation and alignment.

2-2. VSB-AM Network

Optical VSB-AM transmission equipment intensity modulates multiple television channels. This technology provides:

(1). Direct modulatation, compatible with existing coax CATV distribution signals
(2). High performance over a wide bandwidth, up to 550 Mhz
(3). Easy installation and alignment. Configuartuions include strand mount, rack mount and pole mount equipment.

3. PCM and VSB-AM Equipment

3.1. PCM Equipment

Table 1 PCM Video Transmission Equipment

ITEMS		PCM 400 Mbps	PCM 1 2 Gbps	PCM 2.4 Gbps
Total bit rate		400 Mbps	1 2 Gbps	2 4 Gbps
Optical loss budget		25 dB	23 dB	20 dB
Optical devices		1,300nm, 1,550 nm LD and InGaAs PD		
Video	Number of Channels	4 chs 8 chs (WDM)	12 chs 24 chs (WDM)	24 chs 48 chs (WDM)
	Frequency response	20 Hz ~ 4.2 MHz ±0.5 dB, 4 2 MHz ~ 4 8 MHz +0 5/-1 0 dB (BTSC)		
	Coding	8 bits composite coding		
	DG/DP	<3 %, 1 3 °		
	S/N weighted	>56 dB	>60 dB	
Audio	Number of channels	8 chs 16 chs (WDM)	24 chs 48 chs (WDM)	48 chs 96 chs (WDM)
	Frequency response	20 Hz ~ 18 kHz ±0 5 dB 18 kHz ~ 20 kHz ±1 0 dB		
	Coding	16 bits linear coding		
	S/N	>60 dB		
Data channel (Option)		30 Mbps	60 Mbps	60 Mbps
Size		(W)435 mm (H)221 5 mm (D)500 mm	Video, (W)483mm (H)355mm (D)330mm Audio (Two sets of), (W)483mm (H)177mm (D)330mm	
Environment		0° C- 40° C		

3.2. VSB-AM Equipment

Table 2 VSB-AM Transmission Equipment

Bandwidth	50 MHz ~ 550 MHz
Number of channels	40 Channels
Optical loss budget	11 dB
C/N	51 dB
CSO	60 dB
CTB	65 dB
XM	60 dB
Size(Rack-mount type)	480mm(W), 99mm(H), 350mm(D)
Environment	0° C ~ 40° C

4. PCM Repeater Performance and Multi-Drop Distribution

The large link budget and optical repeater configurations associated with PCM transmission systems affords the opportunity for multiple drops access from the primary signal transmission path.

Figure 1. shows the evaluation system tested, which is composed of ten (10) PCM repeaters and three (3) optical couplers. Table 1 outlines the evaluation results. Test result verify that there is no degradation in picture quality.

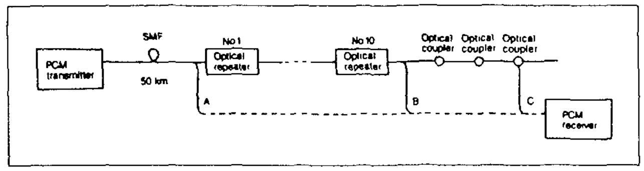

Figure 1. PCM evaluation system composed of PCM repeaters and optical couplers

Table 3. Evaluation results of PCM video signal performance

Test point	Video system				Audio system		
	S/N	DG	DP	Frequency response	S/N	Distortion factor	Frequency response
A After 60 km transmission	63 0	1 0	1 0	±0 30	70 8	0 03	±0 24
B After transmission through 10 optical repeaters	52 8	0 8	0 9	±0 30	72 2	0 03	±0 23
C After transmission through three optical couplers	63 0	0 6	1 0	±0 30	74 0	0 02	±0 23

5. Hybrid CATV System

PCM provides high quality long distant transmission, and VSB-AM provides economical distribution of video signals. The combination of these technologies provides flexibity in the design of fiber optic CATV networks.

Figure 2 is a block diagram representing a hybrid network composed of a, PCM transmitter, ten (10) PCM repeaters, three (3) optical couplers drops, a PCM receiver, VSB-AM transmitter, VSB-AM receiver, and twenty-five (25) coaxial amplifiers. Table 2 outlines the C/N (unweighted) evaluation results for each media of transmission. The end-of-line network performance achieved the minimum desired requirement of a 42 dB C/N.

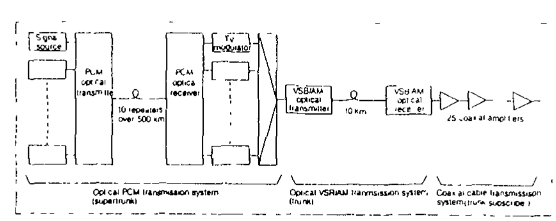

Figure 2. Total CATV system configuration; PCM, VSB-AM, and Coaxial

Table 4 C/N Evaluation results

After PCM transmission (through 10 optical repeaters)		56.9 dB
After VSB-AM transmission (5 dB transmission loss)		51.3 dB
After transmission through coaxial amplifiers	5th step	45.8 dB
	10th step	45.3 dB
	15th step	44.7 dB
	20th step	43.9 dB
	25th step	42.9 dB

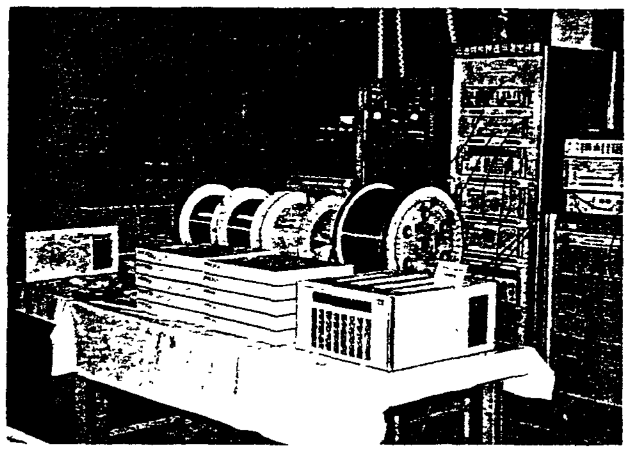

Photo 1 Scene of CATV testsystem

FIBER TO FEEDER DESIGN STUDY

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Abstract

The Fiber-To-Feeder (FTF) design approach is analyzed in comparison to tree-and-branch design for rebuilds. Data from numerous design studies is used to develop a "typical" FTF design. Total material cost is compared using several alternative FTF design approaches, and an optimal solution, known as Fiber to the Serving Area (FSA) is recommended. System performance and material costs are compared to conventional designs for varying plant densities. Additional benefits of FSA design are also discussed, including improved picture quality, increased reliability, reduced operating costs, and more compatibility with future services.

THE FIBER-TO-FEEDER DESIGN

The Fiber-to-Feeder architecture is dramatically different from the traditional tree-and-branch design in a rebuild situation. It should be noted that this paper is confined to the study of rebuild architectures only. In a Fiber-To-Feeder (FTF) design, AM fiber links comprise the trunk portion of the system, performing the function of a cascade of trunk amplifiers in a tree-and-branch design. The feeder portion of the cable system in an FTF design is similar to a conventional plant in that it is made up of RF amplifiers, which perform as bridgers and line extenders. A schematic of a generic FTF design is presented in Figure 1. From a system engineering standpoint, FTF offers significant improvement over tree-and-branch by eliminating cascades of trunk amplifiers, thereby removing a major source of noise. In fact, in an FTF design the target for end-of-line carrier-to-noise ratio is generally around 48 to 50 dB.

The only way to achieve this type of performance using a tree-and-branch architecture is to run AM fiber nodes to every bridger location. At today's AM fiber prices, however, this approach is not practical from a cost standpoint. Therefore, it is necessary to reconfigure the feeder plant in order to improve the design's economics. The key features which separate FTF from tree-and-branch designs, and in fact determine the success of a particular FTF design, are the technology and architecture used in the feeder.

FTF Design Principles

In order to illustrate the important aspects of FTF, a sample FTF design will be used as a reference point. The system under study is essentially a "typical" cable system, which is the result of averaging the data from a number of FTF designs. Figure 2 summarizes the basic elements of the sample design system. The comparative results of various design approaches are presented in Figure 3.

The simplest approach, followed in many of the early FTF designs, is to parallel the tree-and-branch approach by locating a bridger at the output of the AM fiber node and then cascading two or three line extenders from there. (1) As designers became more proficient, it became apparent that the critical factor in producing an

FIGURE 1

FIBER-TO-FEEDER DESIGN

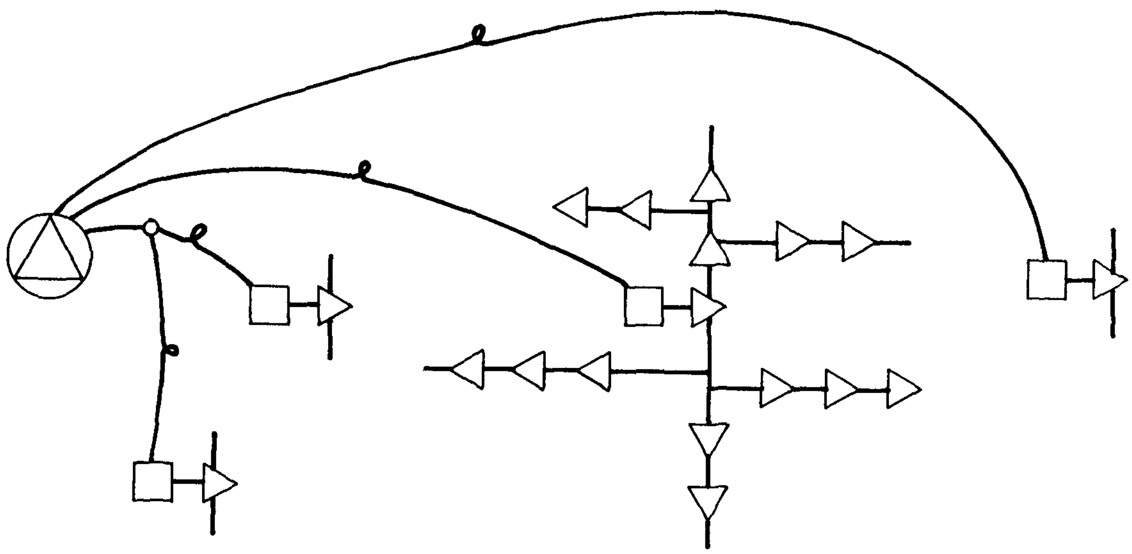


FIGURE 2

SAMPLE DESIGN PARAMETERS

Project Type:	Rebuild
Plant Size:	750 Miles
Bandwidth:	550 MHz
Density:	100 Homes/Mile

Conventional Design

Trunk Cable:	0.875" P-3
Trunk Electronics:	Feedforward
Feeder Cable:	0.625" P-3
Feeder Electronics:	Power Doubling
Fiber Cascade:	Node + 12 Trunks
Headend Cascade:	16 Trunks
Feeder Cascade:	Bridger + 2 LEs

economical FTF design is the miles of plant served by each AM fiber node. As the "serving area" of the AM fiber node increases, the FTF design becomes more and more economical. The approach described above, essentially copying a conventional feeder layout, is the least efficient in terms of the size of the serving area, covering generally 1 to 3 miles of plant.

In order to maximize the serving area of the AM fiber nodes, and thus realize the most efficient design, it is necessary to utilize feedforward technology in order to maintain the desired distortion performance level over the maximum distance. Since the amplifier cascade will by definition be relatively short, the noise contribution of the amplifiers will be minimal, and distortions become the critical parameter.

When push-pull amplifiers are used, the total cost is highest, and the average size of the serving area is about four and one half miles. Moving to power doubling amplifiers reduces the total cost and increases the size of the serving area to five and one half miles. Using feedforward amplifiers results in the lowest total cost, with the srving area expanded to seven and one half miles. By using feedforward technology, higher feeder levels can be maintained, thereby reducing the quantity of amplifiers needed. In short, although feedforward amplifiers are more costly, the use of fewer total amplifers will more than offset the cost difference and yield the lowest total cost.

To increase the size of the serving area still further, and thus make the design more economical, it is desirable to incorporate some trunk design principles into the feeder. Working from the output of the first amplifier after the AM fiber node, "Express Feeder" runs, or essentially supertrunks, are used to extend the reach of the amplifier cascades. The serving area is further subdivided into mini-serving areas, each of which is fed by an Express Feeder. The use of the combination of Express Feeders and feedforward amplifier technology in the feeder plant is referred to as Fiber to the Serving Area (FSA). The resulting design, which is illustrated in Figure 4, yields the lowest total cost, with a serving area of about twelve miles on average. The typical serving area contains approximately 2,000

FIGURE 3

SAMPLE DESIGN COMPARISON

	CONVENTIONAL TRUNK & FEEDER	FIBER-TO-FEEDER CONVENTIONAL FEEDER P-P	FIBER-TO-FEEDER CONVENTIONAL FEEDER PHD	FIBER-TO-FEEDER CONVENTIONAL FEEDER FF	FIBER TO SERVING AREA EXPRESS FEEDER FF
END-OF-LINE CNR (DB) :	46	49	49	49	49
SERVING AREA SIZE (MILES) :	2.0	4.5	5.5	7.5	12.0
<u>MATERIAL COSTS</u> (PER PLANT MILE)					
<u>COAX MATERIALS:</u>					
COAXIAL CABLE:	$2,275	$2,000	$1,950	$1,900	$2,050
ACTIVES & PASSIVES:	$3,225	$3,250	$3,050	$3,000	$2,850
TOTAL COAX:	$5,500	$5,250	$5,000	$4,900	$4,900
<u>FIBER MATERIALS:</u>					
OPTICAL CABLE:	$ 175	$1,000	$ 725	$ 550	$ 300
ACTIVES & PASSIVES:	$ 325	$3,500	$2,775	$1,950	$1,200
TOTAL FIBER:	$ 500	$4,500	$3,500	$2,500	$1,500
TOTAL MATERIAL:	$6,000	$9,750	$8,500	$7,400	$6,400
% OF CONVENTIONAL:	100%	163%	142%	123%	107%

NOTES: 1) Conventional trunk-and-feeder design required some fiber overlay to achieve end-of-line performance required.
2) All systems 550 MHz bandwidth.
3) AM fiber is dual tier.

homes, and this has become the key parameter in identifying serving areas. (2)

Each serving area will be fed by an AM fiber node. An optical loss budget is selected for all of the AM links, which is usually equal to the path loss to the most distant nodes. The system optical budget is generally set in the range of 10 to 12 dB, which allows optical splitting to be used in feeding the nodes which are closer to the headend. The result is that all of the nodes are virtually identical in terms of loss budget, and costs are reduced by sharing transmitters across multiple nodes. Figure 5 shows an overview of a CATV plant, with concentric circles around the headend representing the number of AM fiber nodes fed per transmitter. Figure 6 provides a more detailed look at a section of AM fiber trunk, showing the configuration of transmitters, nodes, and optical splitters.

FSA DESIGN PARAMETERS

The preceding discussion has been based, as mentioned above, on average

FIGURE 4

FIBER TO SERVING AREA DESIGN

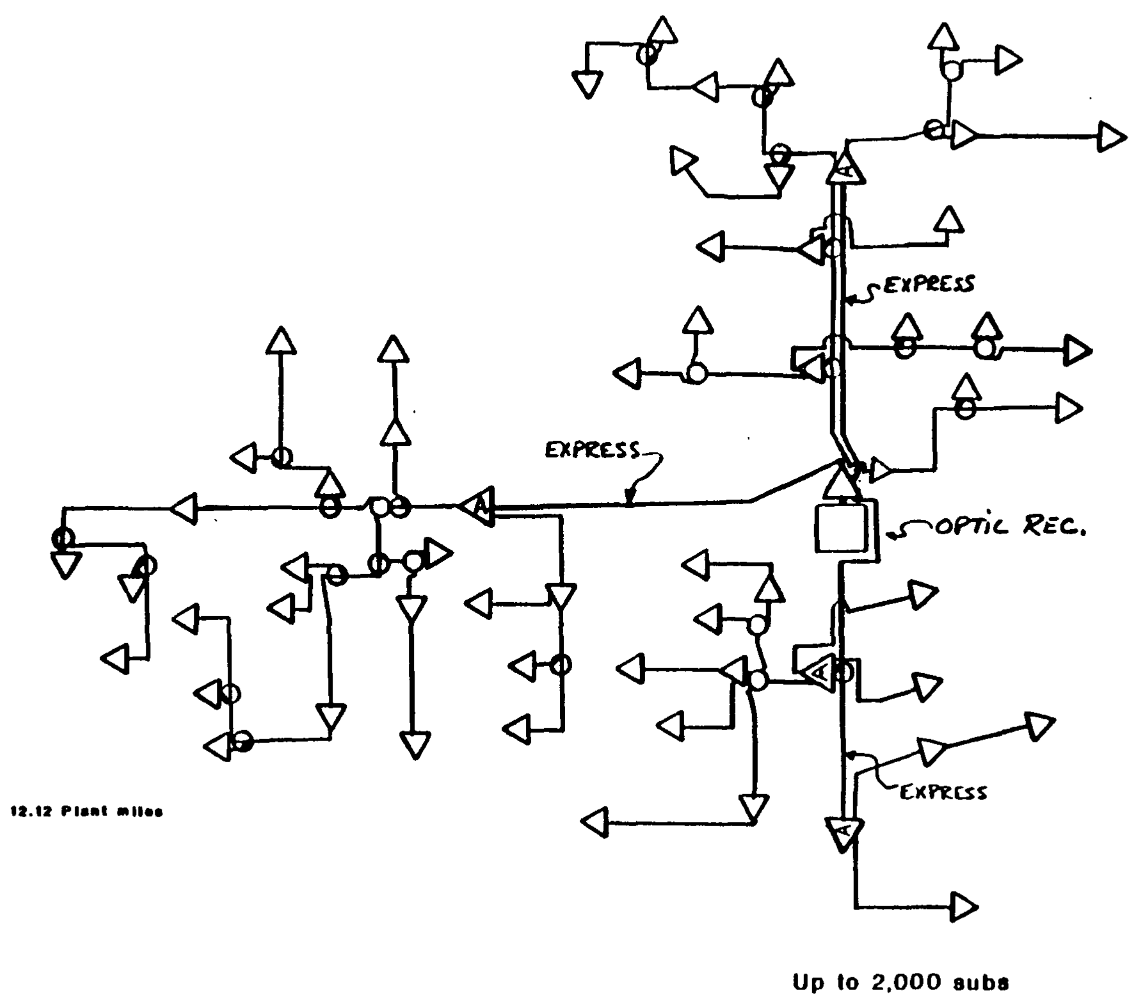
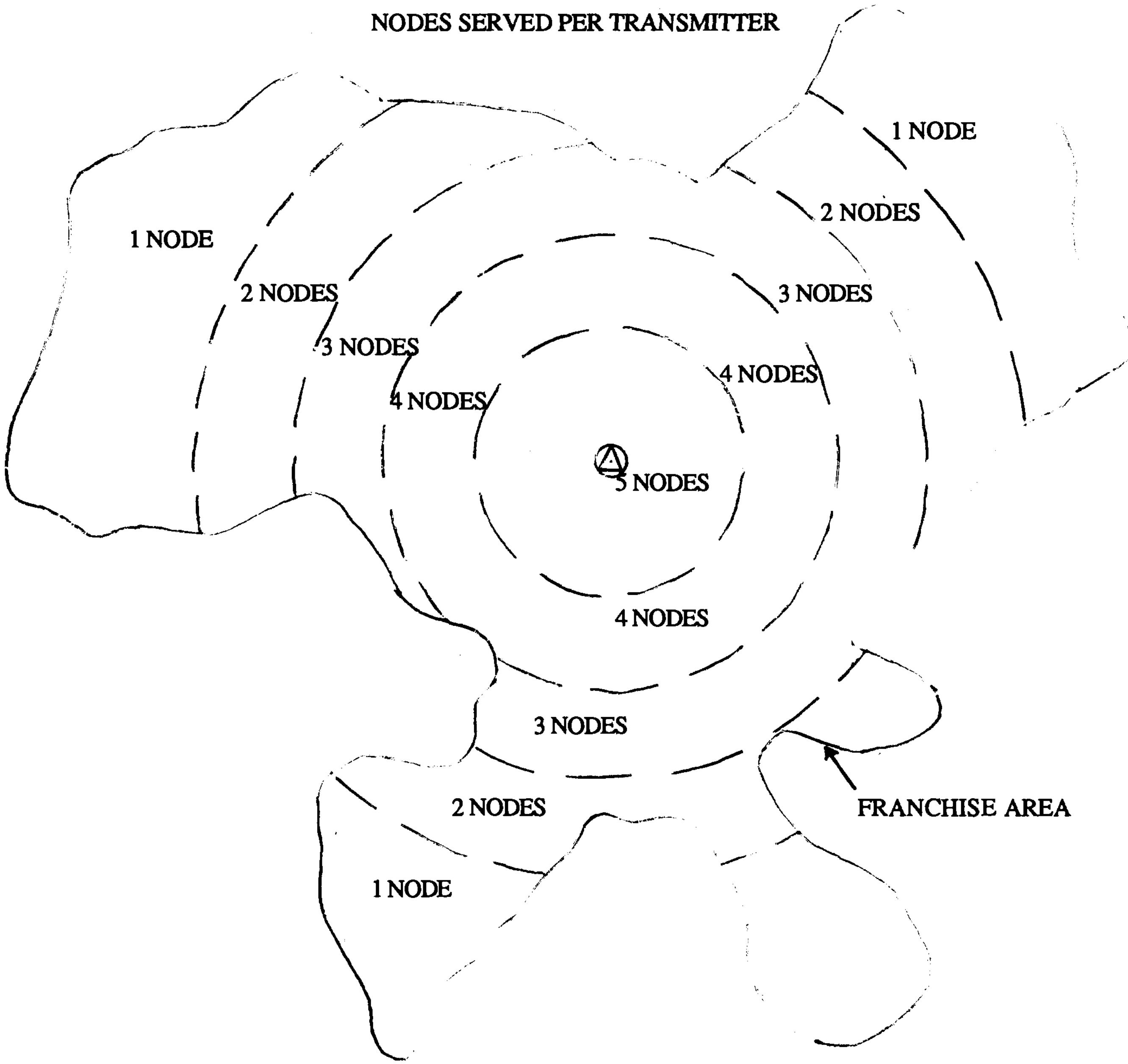


FIGURE 5

AM FIBER RANGE FROM HEADEND

NODES SERVED PER TRANSMITTER



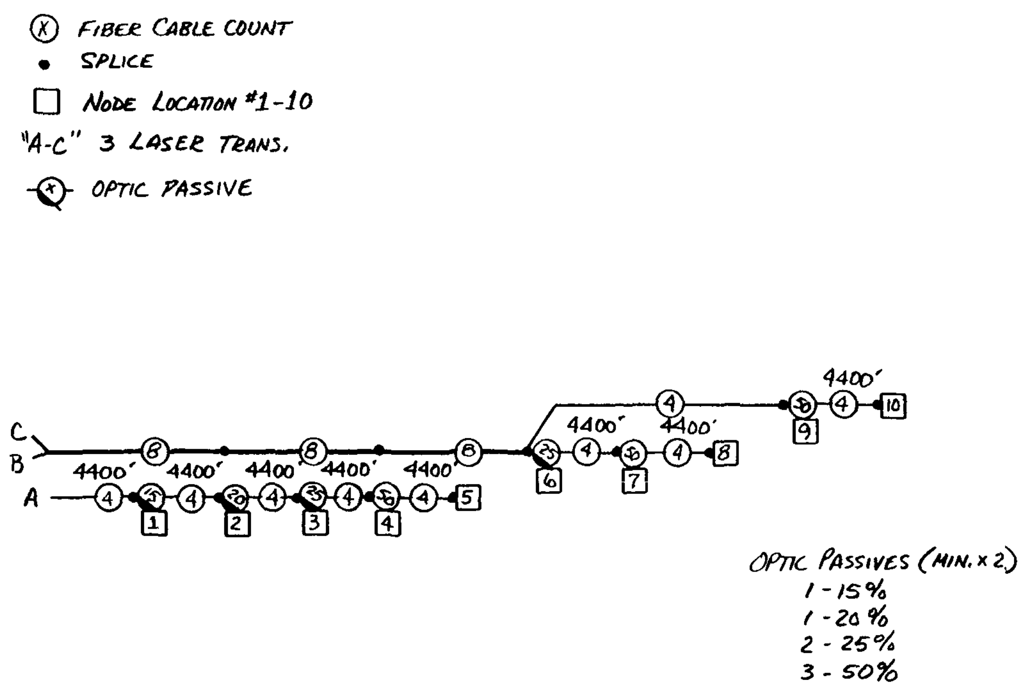
data generated from numerous individual designs. In looking through all of the specific cases, it is possible to identify some of the key design parameters, as well as looking at how the design differs from system to system. In all cases, it is desirable to divide the plant into serving areas of 2,000 homes. Smaller areas may be selected, but a cost penalty will be incurred. In any event, Express Feeders will be used from the output of the first amplifier, which support further subdivision of the serving areas. The design is most cost effective when the serving area is equal to or greater than 12 miles of plant.

The most significant factor impacting an FSA design is the density of the homes in the system being designed. As the system density decreases, the length of the amplifier cascade in each serving area increases. However, even in the densest systems, the minimum number of amplifiers in cascade is four, so at least every third amplifier in each cascade must be equipped with Automatic Gain Control (AGC) to maintain constant output over

FIGURE 6

AM FIBER TRUNK SECTION

FIBER TO SERVING AREA



plant. The carrier-to-noise ratio at the last tap is targeted in the range of 48 to 50 dB, as opposed to 45 to 47 dB in a comparable tree-and-branch design. In an FSA design, the fiber links add virtually no distortions and the amplifiers add very little noise. As demonstrated in Figure 8, the relatively short amplifier cascade sets the composite-triple-beat performance of approximately 56 to 57 dB, which in combination with the essentially tranparent fiber distortions results in end-of-line distortions of about 54 dB. In the same fashion, the fiber link sets the noise limit at about 50 dB, with no more than 1 dB noise addition in the amplifier cascade.

Reliability

The number of outages experienced by each subscriber is dramatically reduced as compared to a conventional plant. In an FSA design, the total number of active devices between any subscriber and the headend is reduced to a maximum of five in the typical case. Even compared to a fiber backbone with four trunk amplifiers in cascade, the number of hybrids in cascade is reduced by 40%. This means that any single failure affects a much smaller group of subscribers than in a conventional plant.

In addition, the number of outages

FIGURE 7

MATERIAL COST COMPARISON

	LOW DENSITY 60 HOMES/MILE	MEDIUM DENSITY 125 HOMES/MILE	HIGH DENSITY 200 HOMES/MILE
TRUNK-AND-FEEDER:	$6,200	$5,975	$6,785
FIBER TO SERVING AREA:	$6,625	$6,155	$7,070
FSA PREMIUM:	6.9%	3.0%	4.2%

NOTES: 1) All systems 550 MHz bandwidth.
2) AM fiber is dual tier.

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FIGURE 8

FSA PERFORMANCE COMPARISON

	FEEDER LEVEL	FIBER C/N	FIBER CTB	COAX C/N	COAX CTB	TOTAL C/N	TOTAL CTB
LOW DENSITY : (4 AMP FEEDER)	47	50.0	65.0	60.2	57.0	49.6	54.1
MEDIUM DENSITY : (5 AMP FEEDER)	46	50.0	65.0	58.2	57.0	49.4	54.1
HIGH DENSITY : (8 AMP FEEDER)	44	50.0	65.0	54.2	56.9	48.6	54.0

NOTES:
1) All systems 550 MHz bandwidth.
2) AM fiber is dual tier.
3) Optical loss budget is 10 dB.
4) 4 and 5 amp cascades: every third amp operated in AGC mode.
5) 8 amp cascades: every other amp operated in AGC mode.

in the plant as a whole is reduced. The total number of active devices employed in the system is reduced, so the number of system outages is lower as well. In fact, the FSA design uses 10% fewer hybrids plant-wide than a comparable fiber backbone or trunk-and-feeder. In previous AM fiber design approaches, while the reliability experienced by indivdual subscribers has improved, the overall system reliability has actually been degraded. This phenomenon is intuitive, since fiber cable and electronics were added to the existing cable plant. However, in FSA designs there is a net gain in reliability, since fiber cable and electronics replace trunk cable and amplifiers. (3)

Operating Costs

The plant operating costs in an FSA architecture are lower than a tree-and-branch design. The total power consumption of an FSA plant is lower than conventional trunk and feeder or fiber backbone designs. There are two primary reasons for this: 1) fewer hybrids are used, and 2) a standby power supply can be used for the optical receiver and surrounding amplifiers while the remainder of the feeder is served by much lower cost non-standbys. The maintenance requirements are lowered by at least an order of magnitude. It is relatively simple to balance and align the short amplifier cascades in an FSA plant. The combination of fewer failures and simpler maintenance procedures reduces the number of truck rolls and simplifies the tasks required, with corresponding reductions in spare parts inventories, technical training, employee turnover, etc. (4)

Compatibility with Future Services

The FSA architecture supports future services. The overall quality of the signals delivered to the subscriber is compatible with HDTV standards. Bandwidth expansion can be accomodated by electronics upgrades, and digital compression offers even more dramatic expansion capabilities. The double-star configuration is compatible with telephony type services; the 2,000 home serving areas are roughly parallel to those of the local telephone system. The use of Express Feeder lends itself to further overlays of fiber nodes. By locating future nodes at the termination of the Express Feeders, mini-serving areas of 500 homes each are established, and the design becomes a triple star. Cells of this size are compatible with switched services, such as video-on-demand. In fact, the advent of Personal Communications Network (PCN) technology makes even the transmission of voice and data services a real possibility.

CONCLUSIONS

In a rebuild situation, Fiber to the Serving Area brings cable operators a number of benefits for about the same capital investment as a conventional trunk and feeder plant design. The picture quality viewed by subscribers is greatly enhanced: the carrier-to-noise ratio at the last tap is from 49 to 50 dB. The number of outages experienced by each subscriber is dramatically reduced, and the number of outages in the plant as a whole is reduced as well. The plant operating costs are lower; the total power consumption of an FSA plant is lower than conventional trunk and feeder or fiber backbone designs. Finally, the FSA architecture supports future services, including High Definition Television, switched services such as video-on-demand and telephony-based services such as Personal Communications Networks.

Fiber optics has the potential to revolutionize the landscape of cable as we know it today. Once in a great while, a technology and a market come together to create a dramatic opportunity. In 1975 the marriage of cable television and satellite transmission technology opened up a whole new world for the CATV industry. Today AM fiber optic technology, combined with the vision embodied in FSA, positions the cable industry to serve the video entertainment and information needs of the twenty-first century.

REFERENCES

1) D. Pangrac and L. Williamson, "Fiber Trunk and Feeder -- The Continuing Evolution," NCTA Technical Papers, May 1990.

2) R. Loveless, "Fiber to the Serving Area: A 'Star' Architecture," SCTE Technical Papers, January 1991.

3) R. Loveless and J. Mattson, "A Fiber Optic Design Study," NCTA Technical Papers, May 1989.

4) K. Casey, "Performance and Economic Implications of Fiber Optics in Cable Television Transmission Systems," SCTE Technical Papers, January 1991.

FIELD TESTING OF FIBER OPTIC CABLE SYSTEMS

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Abstract

Cable operators need to perform certain field tests on their passive fiber optic cable systems in order to provide for the long term operation and maintenance of those cable systems. This paper will discuss the field testing required to properly characterize the optical condition of that plant so as to ensure the reliability of day to day operations. This paper discusses the relevancy and interrelationship of the data obtained from each of these different field tests normally conducted before, during, and after the installation of the passive fiber optic cable system.

INTRODUCTION

The process used to manufacture optical waveguides and subsequently to cable those fibers into a robust product suitable for installation in some of the harshest environments, is incredibly complex, technically sophisticated, and necessarily requires the strictest quality controls for both manufacturing and testing. The fiber optic cable that arrives at your warehouse door has been thoroughly tested and re-tested both as individual material components and as an assembled product. These rigorous testing procedures, now becoming standard for all fiber cable, ensure that the cable you are receiving meets the very highest standards established by regulatory bodies.

Nevertheless, the day that the finished cable product arrives on the operator's loading dock is the time to begin field testing in order to continue maintaining these high standards of quality and to ensure the technical excellence of the finished product -- the passive fiber optic cable system. See Figures 1 and 2. The largest investment of most operators is in their passive fiber transmission system. The amount of technical, logistical, and financial resources required to build the passive cable plant, mandate that extraordinary care be taken to verify the quality of the finished transmission system. Field testing by the operator or his contractor will provide verification of compliance with certain specifications required contractually of the manufacturer and contractor.

Fiber Optic Transmission System

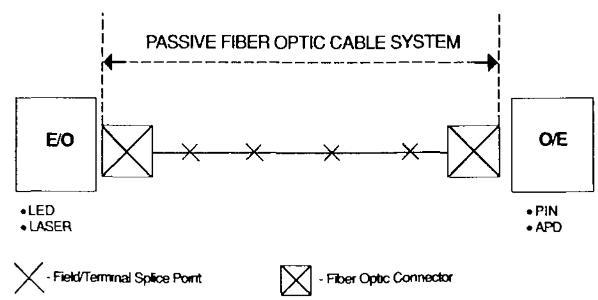


Figure 1

Typical End Termination Detail
Not To Scale

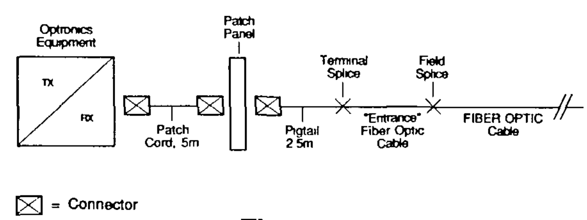


Figure 2

Reels should be unloaded in accordance with the manufacturer's recommendation. Normally, this means keeping the reel upright. Any lifting or movement of the reel should be done using appropriate materials handling equipment designed to handle the large reels typical of fiber cable. Any rotation of the reel should be done as recommended by the manufacturer. Most manufacturers place an arrow on the reel showing direction of rotation. The first field test to be done by the operator is simply a visual inspection of the reel to ensure that no obvious physical damage has occurred during shipment.

GENERAL

The operator is now ready to begin optical field testing of the fiber cable. Testing can be divided into three phases. They are as follows:

1. Pre-Installation
2. Installation/Splicing
3. Post-Installation/Final Acceptance

Optical characteristics that can be measured in the field include attenuation and dispersion. Dispersion characteristics are very closely controlled during the manufacturing of the optical waveguide and the test equipment is extremely expensive; therefore, dispersion tests are rarely performed and generally not required in the field.

Two methods of measuring attenuation employed in the field are the back scattering and two-point methods. More common terms sometimes used to describe these methods are an Optical Time Domain Reflectometer (OTDR) test and the Insertion/Cutback Loss test, respectively. Additionally, the two-point method is often referred to as an Optical Power Test. Both types of tests are required in order to completely and accurately characterize the overall optical condition of a fiber cable system.

Depending on cable and system specifications these tests may be done at both the 1300 and 1550 nanometer wavelengths as well as from both directions. Normally, the Optical Power Test is only performed as part of the Final Acceptance Testing. These test methods provide data which is complementary but may vary slightly as a result of equipment, technique, and skill of the technician performing the test. The key to successful characterization of the fiber cable plant is following correct procedures consistently. This will improve the correlation between measured and calculated optical losses.

TEST METHODS

Backscattering Method

The first and most common method used is the backscattering method. This test is performed using an Optical Time Domain Reflectometer (OTDR). The OTDR is primarily used for locating faults, discontinuities, and anomalies, and general attenuation checks. The OTDR measures distance and loss in an optical fiber by transmitting an optical pulse through the fiber and measuring the optical power reflected back to a sensor in the OTDR instrument. A block diagram of an OTDR is shown in Figure 3.

The OTDR calculates distance along a fiber by measuring the interval between the pulse's launch and the time the reflected energy is detected at the sensor. Loss is calculated by comparing the received power to the transmitted power. As the pulse moves along the fiber, thousands of data points corresponding to distance and relative power are gathered and displayed on the CRT. The result is a linear trace of the fiber displayed as distance from the source (horizontal axis) versus relative power (vertical axis).

The power reflected back from the fiber is caused mainly by Rayleigh scattering. This occurs when light traveling in a fiber strikes a non-uniform part of the core's molecular structure, or a non-silicon dioxide particle and is scattered in a spherical pattern. The part of the scattered energy which travels back towards the optical source is called the "backscatter" and is the basis for the OTDR measurements. Every fiber has a unique "backscatter coefficient." This is a measure of how well optical energy is reflected back to the source. The backscatter coefficient depends on the fiber's relative refractive index difference (the difference between the indices in core and cladding), the core refractive index, and the core diameter.

General attenuation checks include cable reel acceptance, splice loss verification, and final end-to-end measurements. Signature traces should be made of all fibers after connectors are installed, depicting the entire cable route. See Figure 4. These traces will be invaluable should trouble develop on the passive cable plant. Most important, the OTDR is essential for fault locating in the outside plant cable.

OTDR PRINCIPLES & TERMINOLOGY

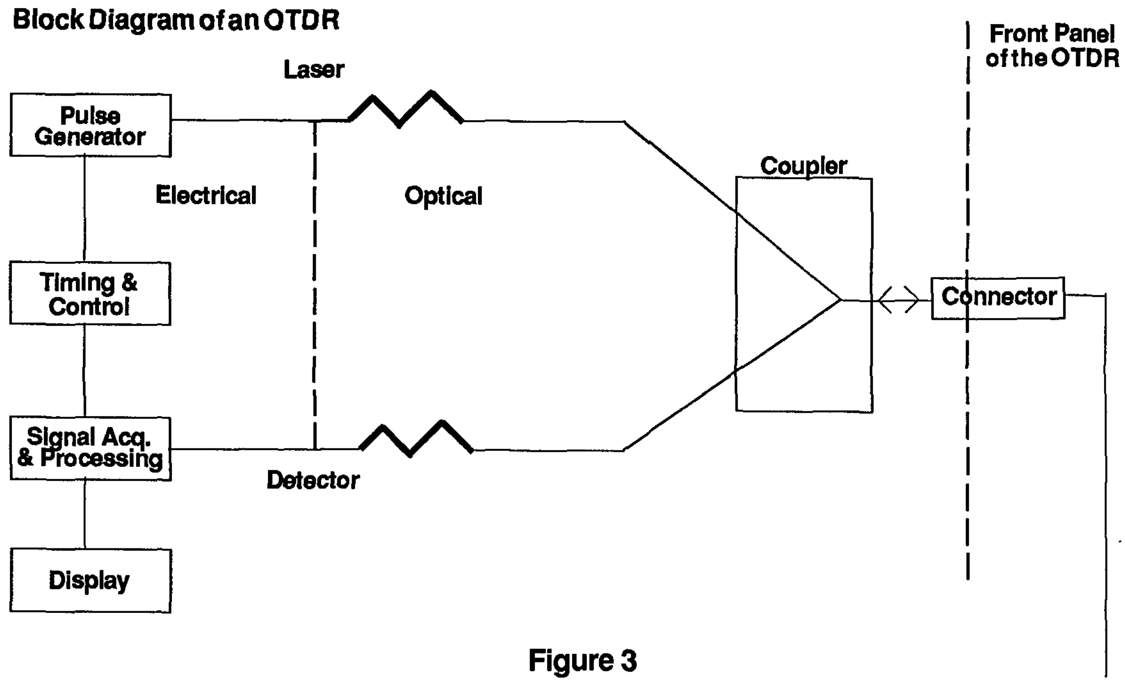


Figure 3

The OTDR has several significant advantages over other test methods. It is an extremely versatile instrument that can be operated by the technician with a minimal amount of training. It is the optical device normally used to locate fiber discontinuities in the cable plant. Through periodic comparison with the initial signature traces, the OTDR may provide early warning of a potential catastrophic failure by indicating points of stress in the cable. When the end configuration depicted in Figure 2 is used, the OTDR can be readily used to access the fiber system through the connnectors.

Although the OTDR is a very accurate and important test instrument, several inherent factors make the OTDR the least accurate of the attenuation methods that may be employed in the field. For very precise attenuation measurements, the cut-back or insertion loss method should be used.

Typical OTDR End-To-End Trace (Courtesy of ANRITSU)

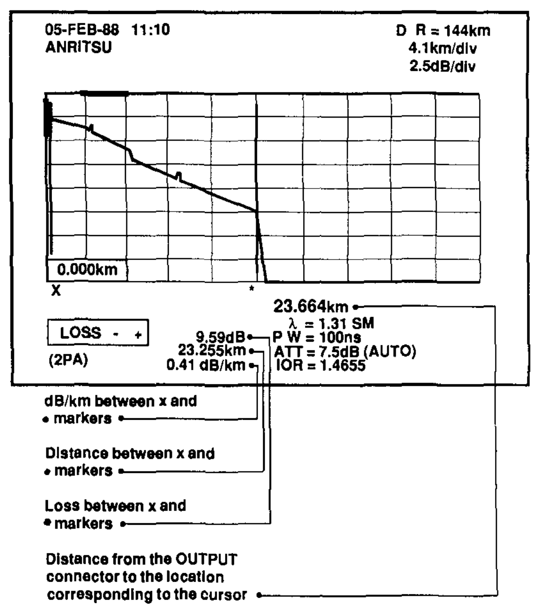


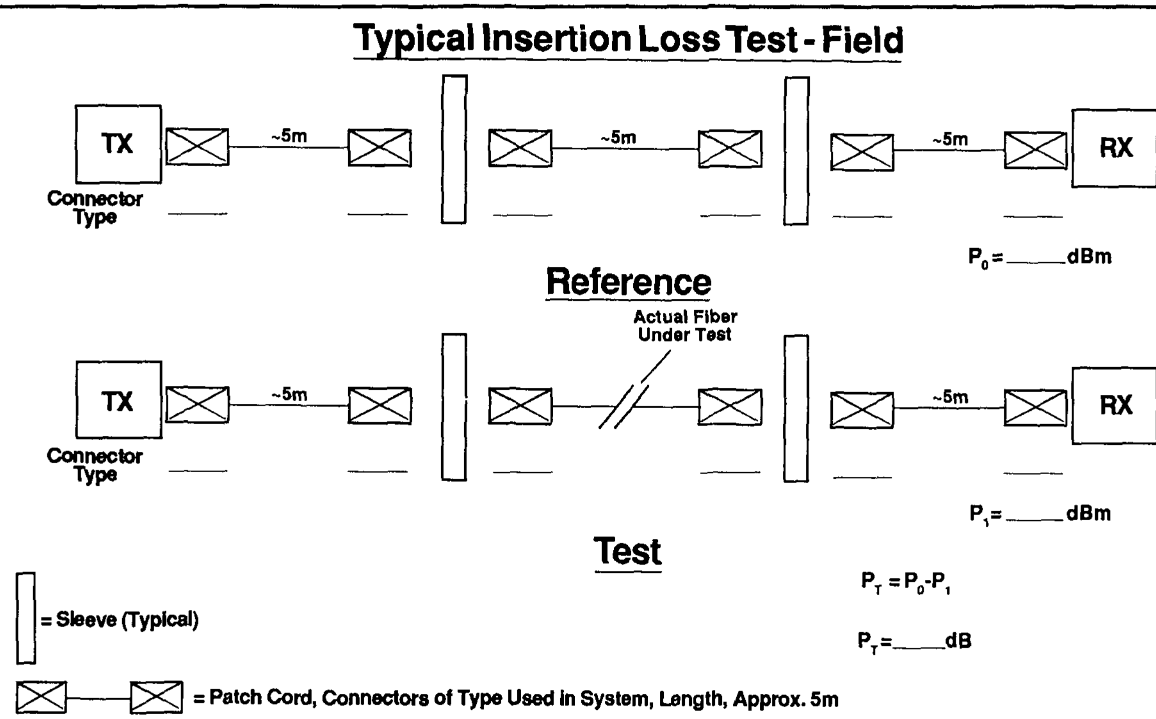
Figure 4

Cut-Back or Insertion Loss Method

By definition, the cut-back method is a destructive test and is not normally used for a fully concatenated fiber optic cable syste. It is not commonly seen in the field, but because of its accuracy and precision, it is often specified and used in cable manufacturing for acceptance test. Since this test is seldom used in the field, it will not be discussed in this paper. For additional information the reader is invited to see EIA Standard, EIA-455-78, Spectral Attenuation Cutback Measurement for Single Mode Optical Fibers.

The insertion loss test method is very similar to the cut-back method. It is also called the Attenuation by Substitution Method. This test method is defined in EIA Standard, EIA-455-171, Attenuation by Substitution Measurement -- for Short-Length Multimode Graded-Index and Single-Mode Optical Fiber Cable Assemblies. The equipment required to perform this test will include a stabilized light source, normally a laser, an optical power meter or receiver, and various patchcords or jumpers used to emulate and to access the fiber cable system.

With this method the first step is to establish a "reference"value to characterize the transmitter, receiver, and connector assemblies. This consists of inserting a short reference cable between the transmitter and receiver of the attenuation measuring (optical power meter test) set and in measuring the transmitter output power P_0 coupled into the fiber/jumper assembly. In the second step the reference cable is replaced by the cable link to be measured and (inserted) attenuation is obtained by comparing the level available at the output side of the cable link P_1 with the reference level P_0. Figure 5 is a depiction of an Insertion Loss Test.



EQUIPMENT DATA	TRANSMITTER (TX)	RECEIVER (RX)
Manufacturer		
Serial Number		
Type		
Wavelength		N/A
Output Power		N/A

Figure 5

This method has the advantage in that it is simple to perform and it can be used for fully connectorized/concatenated cables. This method has one disadvantage advantage in that the varying coupling conditions on the transmitter or receiver side may cause the reference level to deviate from the level coupled into the cable link resulting in minor measuring errors.

FIELD TESTING

Pre-Installation Field Testing

Pre-installation field testing consists of cable acceptance from the manufacturer by the operator. The operator conducts this test with an OTDR. The objective of pre-installation is to assure the operator that no damage has occurred during the transit from the manufacturer to the end-user's dock. The technician should be verifying optical length, attenuation, anomalies, and continuity. This data should be recorded either electronically or with a chart recorder. The majority of cables now being specified are for dual wavelength operation. Because of the sensitivity of fiber to attenuation related difficulties at the 1550 nm wavelength, it is recommended that this pre-installation test be conducted at that wavelength assuming an OTDR with that wavelength is available.

All fibers must be checked in order for this test to be effective! Random testing can result in less than desirable assurances of delivered product. This is a result of one of the innovative and very positive features of most fiber cable designs. These designs will allow damage to occur to fibers within the cable in a random and unpredictable manner. This means that it is possible to have an incident such as an air driven nail inadvertently driven into a cable during the lagging process that may damage only a few fibers in a multiple count fiber cable. At this point, our good friend Murphy steps in and the technician is guaranteed not to select the damaged fiber in his random checks. It is only after that particular 30,000 foot reel is installed in a very troublesome area that the broken fiber is discovered and at the very least a splice is added or worse, a reel of cable is replaced. Therefore, if pre-installation tests are conducted, then it is recommended that *all* fibers be tested.

This pre-installation check can provide assurances to the operator and to his construction group as to the quality of the fiber cable prior to installation. Many times these checks are conducted jointly by the contractor and the owner. This cooperation may preclude difficulties later should a cable be damaged during the construction operation.

Installation/Splicing

Installation or splicing field tests can be categorized into five main categories. They are as follows:

1. Optical Time Domain Reflectometer (OTDR)
2. Optical Power Monitoring
3. Local Injection Detection (LID)
4. Profile Alignment Systems (PAS)
5. Some Combination of the Above

There are several advantages and disadvantages to each of the five approaches. Detailed papers are available for each of the techniques that the operator may use. The major points to consider are availability of skilled technicians, equipment, and the logistics of communications during the splicing operation. In all cases the resulting splice loss should be recorded so that this value can be used when calculating the expected system loss. See Figure 6.

Typical OTDR Splice Loss Measurement (Courtesy of ANRITSU)

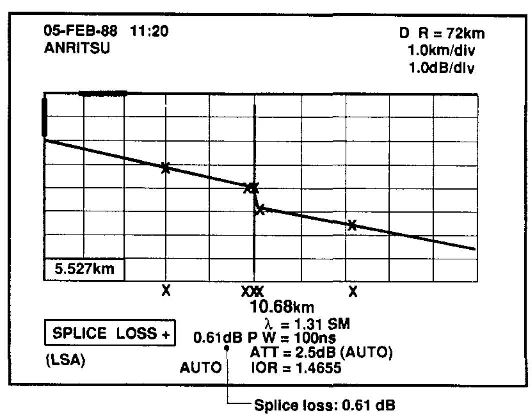


Figure 6

Post-Installation/Final Acceptance

Post Installation and Final Acceptance testing should consist of end-to-end OTDR test made from both directions at both 1300 and 1550 nm wavelengths. Additionally, an Optical Power Test using the Insertion Loss Method should be done at both wavelengths from both directions, if connectors are used. By performing these tests, the operator is assured of having all the necessary data required to characterize the passive optical system.

DOCUMENTATION

The minimum documentation required for a fiber optic cable system should include the following:

1. As Built Drawings
2. Splice Loss
3. End-to-End Optical Loss Measurements
4. End-to-End OTDR Signature Traces

The purpose of this data is to provide historical references for maintenance and emergency restoration. By maintaining this data, the operator is assured that he will be able to quickly respond, identify, locate, and repair, any problem that may occur within his passive cable plant. Examples of a Splice Log, End-to-End Loss Measurement, and Summary of Field Data-Attenuation Loss, are included as Examples 1, 2, and 3.

RESULTS

Analysis and Correlation of Test Results

The data that has been acquired during the field testing of the fiber cable is then analyzed. This measured field data is compared to that data that was obtained by summing cable loss, splice loss, and pigtail splice loss. See Figure 7 and Sample 3. It is not unusual to see variations on the order of ±5% between the results calculated or measured

Calculated Cable Loss

PROJECT:	ANYTOWN, USA							DATE.	2-Oct-89	
SPAN:	SOUTHPORT HEADEND TO NODE AT ELM STREET							TECHNICIANS:		
OPTICAL LENGTH	18,579									
SHEATH LENGTH:	18,616									
WAVELENGTH:	1300 NM									
FIBER NUMBER LOCATION LENGTH (KM)	END SPLICE HEADEND SOUTHPORT	CABLE LOSS *890108 3.295	SPLICE LOSS MAIN STREET SP *1	CABLE LOSS *890206 5.569	SPLICE LOSS OAK STREET SP*2	CABLE LOSS *890207 4.351	SPLICE LOSS HWY 428 SP*3	CABLE LOSS *890308 5.364	END SPLICE NODE AT ELM STREET	CONCATENATED SUMMARY LOSS (dB)
1	0.08	1.02	0.05	1.73	0 14	1.48	0 03	1.88	0.08	6.48
2	0.08	1.25	0.04	1.95	0.12	1.57	0.02	1.72	0.08	6.82
3	0.08	1.12	0.07	1.89	0.05	1.52	0.11	1.66	0.08	6.59
4	0.08	1.29	0.11	2.17	0.06	1.74	0.12	1.88	0.08	7.52
5	0.08	1.05	0.14	1.73	0.03	1.35	0.11	2.09	0.08	6.66
6	0.08	1 32	0.06	1.89	0.09	1.48	0.05	2 04	0.08	7.08
7	0.08	1.19	0.00	2 12	0.08	1.61	0.04	1.82	0.08	7.01
8	0.08	1.22	0.12	2.00	0.14	1.57	0.09	1.88	0.08	7.17
AVERAGE SPLICE LOSS (dB):		0.08								
AVERAGE TOTAL LOSS (dB):		6.92								

CABLE *890108		CABLE *890206		CABLE *890207		CABLE *890308	
FIBER NUMBER	LOSS db/km	FIBER NUMBER	LOSS db/km	FIBER NUMBER	LOSS db/km	FIBER NUMBER	LOSS db/km
1	0.31	1	0.31	1	0 34	1	0 35
2	0.38	2	0.35	2	0.36	2	0.32
3	0 34	3	0.34	3	0.35	3	0.31
4	0 39	4	0.39	4	0.40	4	0 35
5	0 32	5	0 31	5	0 31	5	0 39
6	0 40	6	0 34	6	0 34	6	0.38
7	0 36	7	0 38	7	0 37	7	0 34
8	0 37	8	0 36	8	0 36	8	0 35

Figure 7

from these multiple tests. Values outside of these ranges should be investigated. These differences can be caused by any or all of the following:

1. Optical Loss Calculation Methods
2. Physical Layout of Pigtails in Fiber Distribution Panels
3. Testing Procedures
4. Cleaning Procedures
5. Condition of Optical Connectors
6. Differences in Test Equipment

The values derived from the OTDR test and the Insertion Loss test are complementary. The OTDR provides an overall "picture" of the condition of the installed fiber optic cable system. The Insertion Loss test provides an absolute value of optical power through all connectors, sleeves, splices, and fiber.

SUMMARY

Three (3) sets of field data need to be collected during the installation process in order to maintain complete records about the optical properties of the passive cable system. This data is as follows:

1. Calculated Data-Obtained from Cable Reel Data Sheets and Splicing Logs
2. OTDR Data-Obtained from end-to-end cable test
3. Insertion Loss Data-Obtained from end-to-end cable test

This data then becomes part of the permanent record for both the customer and the cable manufacturer's files.

It is essential that any operator of a fiber transmission system maintain adequate information about that system for maintenance, trouble-shooting, and emergency restoration procedures. Information required to maintain that system include results from Optical Time Domain Reflectometer (OTDR) and Optical Power tests. The two measured and one calculated tests provide complementary information that accurately and completely characterize the optical condition of the passive fiber cable plant.

By periodically verifying the attenuation loss of the cable system with either an OTDR or Optical Attenuation Test Set, the cable operator may be able to preclude difficulties with the transmission medium at some future date. The objective is to ensure by field testing that the installation of the pasive cable plant was completed at the same high levels of quality as both the manufacturing and cabling operations.

REFERENCES

Anritsu, *Electronic Measuring Instruments,* 1990, pg. 69.

Heckman, Siegfried, "High Precision Attenuation Measuring Method," *IWCS Proceedings 1986,* pg. 322-328.

Grande, Philip; Hayback, Charles; "Improving the Correlation Between Measured and Calculated Optical Loss, *"NFOEC Papers 1990,* pg. 5.4.1-5.4.16.

Wilson, Paul; Bonnyman, Al; "Practical Considerations in the Testing of Fiber Optic Cables at the System Level, *"SCTE Fiber Optics Papers 1990,* pg. 33-37.

Matthews, James, "Q & A," *Guidelines,* Volume 5, Number 2, pg. 13.

Matthews, James, "Measuring Fiber Transmission Systems," *CED,* April 1990, pg. 52-61.

ADDITIONAL SOURCES OF INFORMATION

EIA Standard, EIA-455-57, FOTP-57, Optical Fiber End Preparation and Examination.

EIA/TIA Standard, EIA/TIA-455-60, FOTP-60, Measurement of Fiber or Cable Length Using an OTDR.

EIA/TIA Standard, EIA/TIA-455-61, FOTP-61, Measurement of Fiber or Attenuation Using an OTDR.

EIA Standard, EIA-455-78, FOTP-78, Spectral-Attenuation Cutback Measurement for Single-Mode Optical Fibers.

EIA Standard, EIA-455-95, FOTP-95, Absolute Optical Power Test for Optical Fibers and Cables.

EIA Standard, EIA-455-171, FOTP-171, Attenuation by Substitution Measurement -- For Short-Length Multimode Graded-Index and Single-Mode Optical Fiber Cable Assemblies.

Splice Log

PROJECT:				DATE:		
SPAN:				TECHNICIANS:		
OPTICAL LENGTH:				WEATHER CONDITIONS:		
SHEATH LENGTH:						
SPLICE POINT LOCATION:						
OTDR LOCATION:						
	FIBER	SPLICE ATTEMPT (LOSS IN dB)				
IDENTIFIER	NUMBER	1	2	3	4	COMMENTS

Example 1

Comm/Scope, Inc.

End To End Loss Measurement

PROJECT:				DATE:	
SPAN:				TECHNICIANS:	
OPTICAL LENGTH:					
SHEATH LENGTH:					
TRANSMITTER LOCATION:					
RECEIVER LOCATION:					
WAVELENGTH:					
TRANSMITTER SN:					
RECEIVER SN:					

IDENTIFIER	FIBER NUMBER	MEASURED POWER (dBm)	REFERENCE (dBm)	ACTUAL POWER (dB)	COMMENTS

Example 2

Comm/Scope, Inc.

Summary Of Field Data, Attenuation Loss

PROJECT:				DATE:	
SPAN:				TECHNICIANS:	
OPTICAL LENGTH:					
SHEATH LENGTH:					
WAVELENGTH:					

IDENTIFIER	FIBER NUMBER	CALCULATED LOSS (dB)	OTDR LOSS (dB)	INSERTION LOSS (dB)	COMMENTS

Example 3

Comm/Scope, Inc.

Future Hybrid AM/Digital CATV Systems

David Grubb III
Jerrold Communications

ABSTRACT

Digital video compression technology is being developed for NTSC and HDTV applications. In addition to the benefits to the subscriber in terms of picture quality, digital transmission has a number of implications for the distribution plant. This paper will detail some of the effects of digital signals on the CATV system.

INTRODUCTION

Digital video compression technology is being pursued by a number of R&D organizations around the world for a number of applications. Potential DBS operators are interested in increasing the number of program selections that they can offer with their limited transmission bandwidth. Compression ratios as high as 10 video channels per transponder are anticipated[1]. Bellcore is interested in developing the technology to allow the transmission of video programs over existing twisted pair telephone drop wires[2]. The computer industry is interested in integrating full motion video with audio and data (desktop video). The consumer electronics industry will likely be interested in offering digital video products for home use. In addition to all of the above activity on digital NTSC delivery, three of the current proposals for the U.S. HDTV transmission standard are all-digital. Because of all of this activity it seems inevitable that some video signals will be transmitted over cable systems in digital form in the 3 - 6 year timeframe. This paper will look at how this move toward digital transmission will affect the performance of CATV distribution systems.

APPLICATIONS

Basic services

Over the past several years, there has been an explosion in the number of basic services available to the system operator. A recent listing showed 61 satellite delivered basic services[3]. Add to that 5 - 10 local broadcast channels, regional basic services, local origination and text services and it is easy to envision a 550 MHz cable system that is completely filled by basic service in the not too distant future. While basic services could certainly be transmitted digitally, many operators will likely reject this option because of the costs associated with installing multiple converters in every home that receives basic service.

Premium and IPPV services

The use of digital compression for satellite transmission may help to fuel growth in the availability of IPPV programming because it will lower the programmer's costs. Some type of converter or on/off premise device is required to provide the IPPV functionality and signal security. A digital compression converter could provide signal security as well as offering the benefits of consistent signal quality and video program capacity expansion.

HDTV

A number of the candidates for the U.S. HDTV transmission standard are proposing all-digital systems. If one of these systems is chosen, then even if digital compression is not used for any NTSC services, cable systems will be required to transmit digital HDTV signals to their customers.

OBJECTIVE

The objective of this paper is to explore how a mix of AM and digital signals will affect distribution plant performance. As outlined above, future systems will probably require 80+ channels of basic service. When you add premium services, multi-channel IPPV and HDTV, system bandwidth requirements of 750 MHz, or higher, seem likely. Due to the limited availability of 750 MHz hybrids at the present time, the data presented in this paper is based on 550 MHz devices.

ASSUMPTIONS

All of the data presented in this paper is based on one simple assumption. The assumption is that medium speed (20 Mb/s) digital data can be transmitted through a 6 MHz channel and accurately recovered if the channel has a carrier to noise plus distortion ratio greater than roughly 25 dB. This is a dramatically different requirement than that of a VSB-AM video signal. For high quality pictures, a VSB-AM signal requires a carrier to noise ratio of 47 to 52 dB and carrier to distortion ratios of 53 to 57 dB. What this means is that the digital carriers can be transmitted through a cable system at lower levels than conventional VSB-AM channels. This reduces the loading on the RF and optical devices in the system and improves system distortion performance.

SYSTEM PERFORMANCE

The simplest case to analyze would be that of a system that transmits all channels digitally. In the case of an AM fiber trunk link, the C/N requirement might drop from 50 dB to 28 dB. This would increase the loss budget from 12 dB to roughly 26 dB. If the AM link served an average of 4 receivers from each transmitter, the digital link would serve roughly 64 receivers from each transmitter. This would obviously improve the system cost significantly. The effect on the RF distribution plant would be that far more taps could be served from each active device.

While the prior example was interesting, it does not really fit with most operator's strategies. As outlined above, most operators will likely prefer to transmit basic cable services using VSB-AM modulation. This means that when digital technology becomes available, the typical CATV system will have to carry a mix of AM and digital signals. The effects of this mixed signal environment on the CATV network are not as straightforward as the all-digital scenario, but they are still significant.

For this paper, a mix of actual measurements and computer simulations were used to quantify the effects of digital signals on RF amplifier performance. The performance of a typical push-pull CATV hybrid is shown in Figure 1. The data points show measured CTB at several frequencies when the amplifier is loaded with 77 channels at 44 dBmV per channel. The curve fit to the measurements describes the behavior of the amplifier over the frequency range of interest. Note that even though the number of beats is highest in the center of the band (roughly 2000 beats at 331.25 MHz), the amplifier CTB performance is worst at 547.25 MHz (roughly 1360 beats). This is because the amplitude of the individual triple beats increases with frequency.

Figure 2 shows the projected performance of this amplifier when operating with a tilted (cable equivalent slope) output. These data were calculated using a program that accounts for the amplitude of each of the individual triple beats that make up CTB at each frequency. The flat curve is the same as in Figure 1. In both tilted cases, the output level at 547.25 MHz is 44 dBmV. This graph shows that as the output tilt is increased, the amplifier's CTB performance improves significantly at high frequencies. There is not as much net improvement at low frequencies because the carrier levels are decreasing as well as the distortion levels.

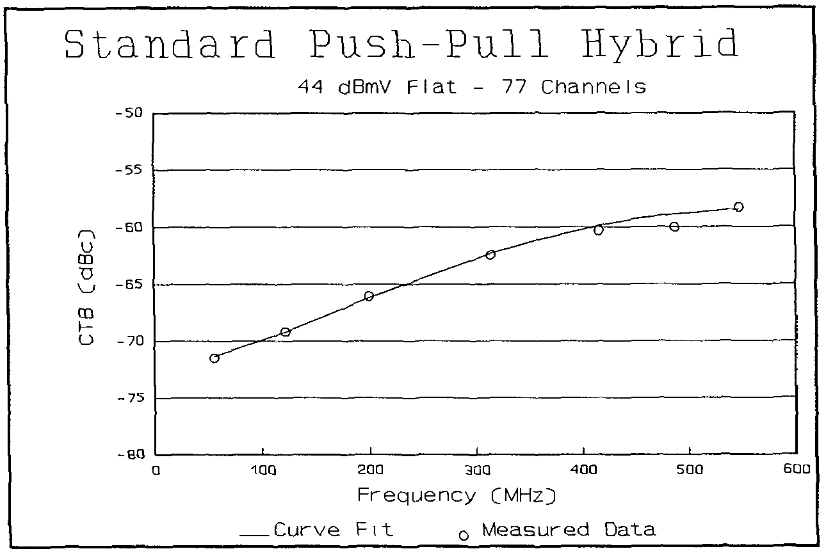


Figure 1

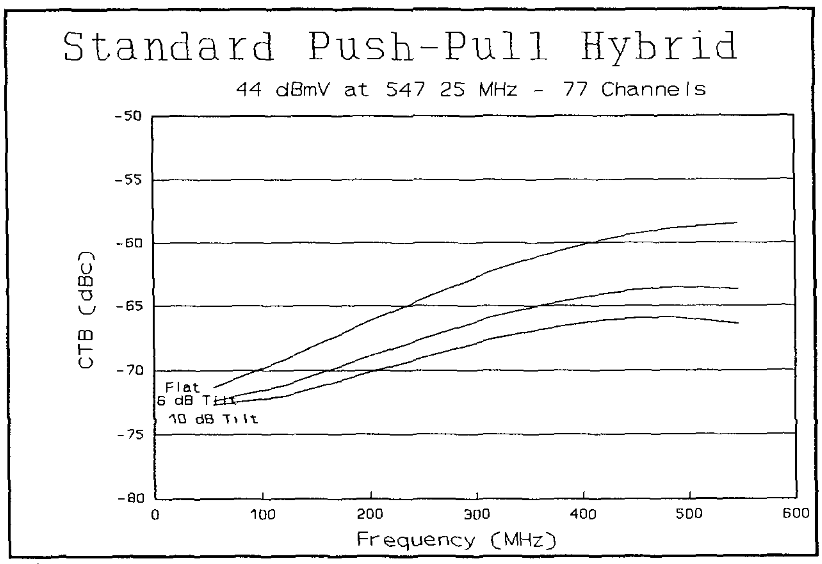


Figure 2

Because digital signals can tolerate higher levels of noise and distortion, it was assumed that the digital carriers would be transmitted through the system at a level 10 dB lower than for a conventional AM channel. It was also assumed that the digital channels would be located at the same frequencies (6 MHz spacing) as conventional AM signals. Two 550 MHz scenarios were investigated. The first assumed that standard AM channels would be transmitted in the band from 50 - 350 MHz, and digital carriers would be transmitted in the band from 350 - 550 MHz. This gives a total capacity of 44 AM channels and 33 digital channels. Each digital channel can support 1-5 video programs depending on the compression system employed. The second scenario is similar except the split point between AM and digital is located at 450 MHz, yielding 60 AM and 17 digital channels. Figure 3 shows the spectrum for the first scenario with no tilt. Figure 4 shows the predicted and measured amplifier CTB when loaded with the signal shown in Figure 3. The measured data points agree fairly well with the predicted performance. The predicted worst case CTB for an AM channel is -67.5 dBc at 349.25 MHz. This represents a 9 dB improvement in performance when compared to the worst case CTB for 77 AM channels. This is a dramatic improvement in device performance that can have a significant effect on system cost and performance. The performance improvement when using a tilted output is even more dramatic as shown in Figure 5 and Table I. Part of the reason that the distortion improvements are so dramatic is the fact that the hybrid distortion is worse at higher frequencies. For a device with a flat distortion vs. frequency characteristic, the improvement factors would be 5.0, 7.4 & 9.0 dB for 0, 6 & 10 dB of tilt, respectively. One example of a device with a flatter distortion characteristic is the Darlington hybrid. Another factor affecting the CTB improvement factor is that in a typical CATV amplifier the levels might be flat at the preamp output and tilted at the postamp output. To the extent that the input hybrid contributes to the overall amplifier CTB, the improvement factors will be decreased slightly. Because of these issues, the improvement factor for an amplifier or AM fiber product depends on the configuration and selection of components.

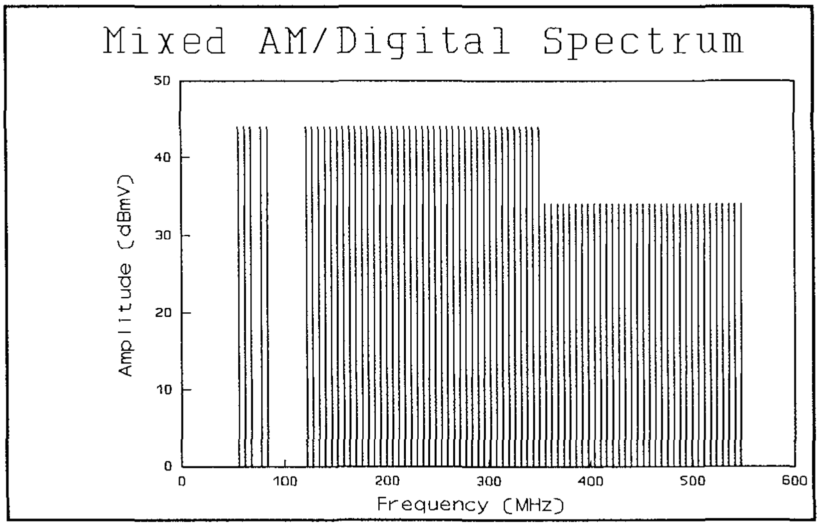


Figure 3

Figures 6 & 7 and Table II show measured data and projected performance for the same hybrid with AM video below 450 MHz and digital channels above 450 MHz.

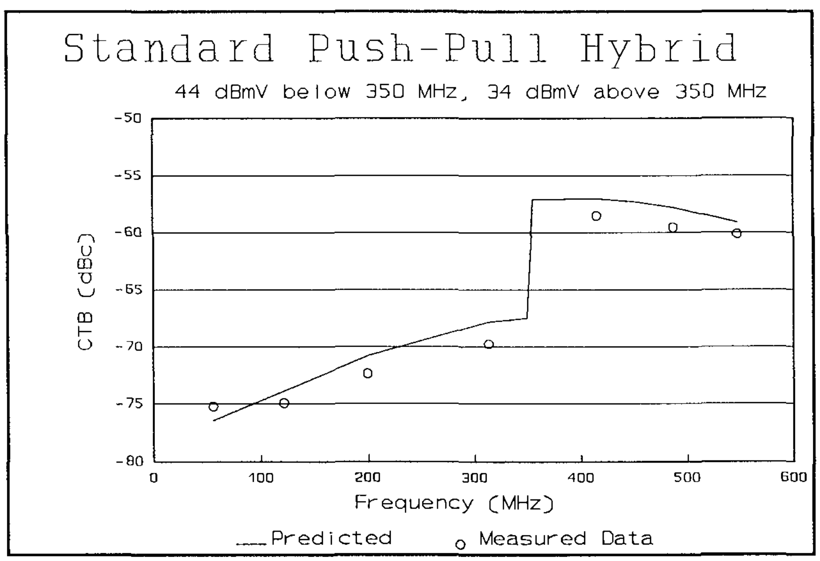


Figure 4

Figures 5 & 7 also show that the worst case CTB for a digital channel is about 10 dB worse than the worst case AM channel. If a distribution system was designed with a target CTB spec of -53 dBc for the AM channels, the CTB for the digital channels would be on the order of -43 dBc. This should be well below the threshold of degradation for the digital channel.

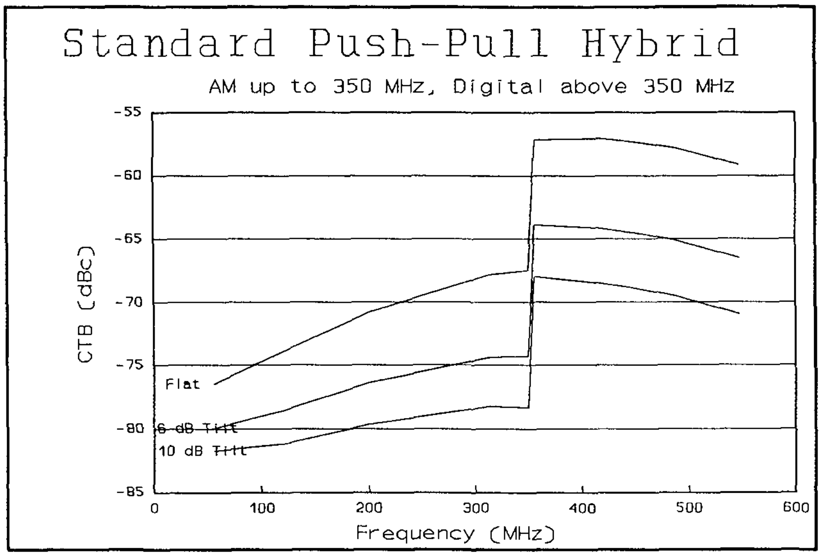


Figure 5

COST IMPLICATIONS

There are many ways that the effects of digital transmission on system performance can be exploited, depending on the individual system requirements. In a system upgrade, the use of digital transmission may make the upgrade more economical or allow a greater increase in performance or bandwidth. In a new build or rebuild digital transmission can make the system design more economical or allow higher performance levels for the same plant design.

A few small trial designs were performed to assess the value of the performance improvements outlined in Tables I & II. The designs assumed a new or rebuild situation and both conventional tree & branch and fiber to the feeder alternatives were investigated. The designs were done on a system with 100 homes per mile, with end of line performance (AM channels) of C/N = 47 dB, CTB = -53 dBc. The result of the study showed that the cost savings, mainly in RF and fiber equipment, are on the order of $50 - 100 per mile per dB of performance improvement. In other words, the equipment cost for a 350/550 MHz AM/Digital system is roughly $450 - $900 per mile less than a conventional 550 MHz AM system, assuming a 9 dB performance improvement factor. The savings works out to $4.50 to $9.00 per home passed or $7.50 to $15.00 per subscriber (60% penetration).

While these cost savings are significant in terms of total plant cost, they do not justify the installation of digital converters in every home for basic service in a new or rebuild scenario. The one exception would be operators who plan to scramble all channels, including basics, anyway. In their case the incremental cost of the digital converter

Table I

Worst Case CTB for AM Channels
350/550 MHz Split
Push-Pull Cascode Hybrid

Tilt	All AM	AM/Dig	Improvement
0	-58.5	-67.5	9.0
6	-63.5	-74.3	10.8
10	-65.9	-78.3	12.4

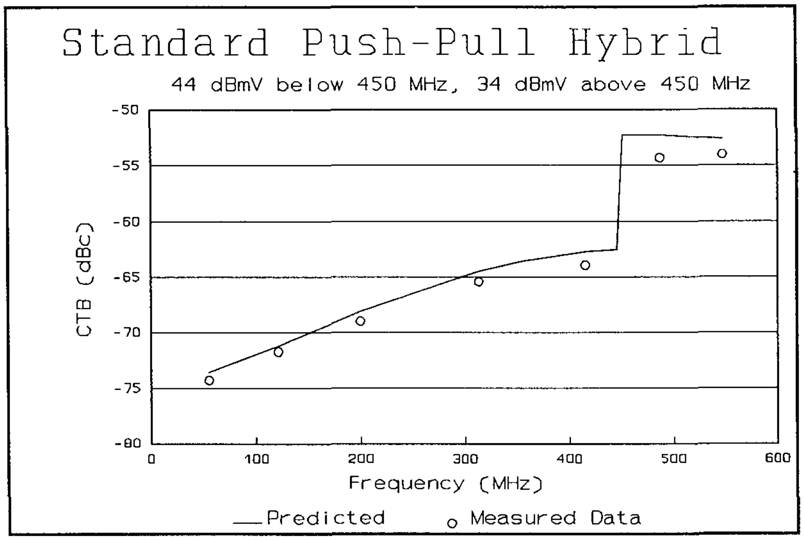


Figure 6

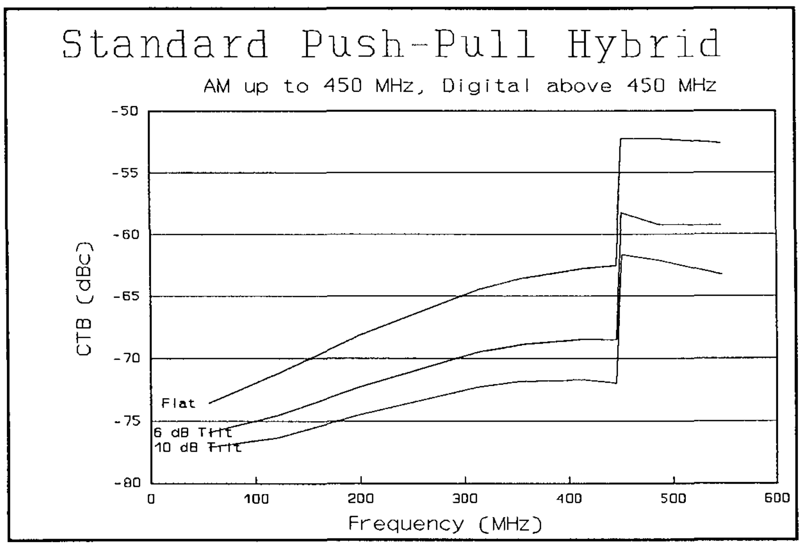


Figure 7

Table II

Worst Case CTB for AM Channels
450/550 MHz Split
Push-Pull Cascode Hybrid

Tilt	All AM	AM/Dig	Improvement
0	-58.5	-62.5	4.0
6	-63.5	-68.5	5.0
10	-65.9	-71.7	5.8

would have to be traded off against performance improvements and plant cost savings.

The prior analysis was for a new or rebuild or situation. In the case of a system upgrade, the tradeoffs will be different for each specific case. Digital transmission could be used to expand system program capacity without actually modifying the existing plant. For example, the top 5 channels of an existing system could be converted to digital channels. These 5 channels could then be used to provide 20 video programs (assuming 4 digital video programs per 6 MHz). If the additional program capacity is needed for premium or IPPV services, then this type of upgrade would probably be cost effective because the digital converters only go into homes buying premium services. However, if the system needs the additional capacity for basic services, it will likely be more cost effective to upgrade the distribution system.

EXPANDED BANDWIDTHS

Significant performance improvements are also expected for AM/Digital systems at higher bandwidths. The improvement factor will probably be in the range of 4 - 6 dB for a 550/750 MHz split system, compared to an all-AM 750 MHz system. The precise improvement factor will depend on the distortion characteristics of the devices used in these systems. Another important factor in the design of extended bandwidth systems is the high tap levels that will be required at 750 MHz for AM signals. The use of digital carriers above 550 MHz would ease high frequency tap level requirements because the digital signals can tolerate more noise.

CONCLUSIONS

When digital compression technology becomes economical, from a distribution plant perspective it will be beneficial to broadcast premium services in digital form. The digital technology can provide the desired signal security, and at the same time performance can be improved for the remaining AM video channels. Alternatively, the cost of the distribution plant can be reduced for a given desired performance level on the AM channels.

The use of digital transmission for signals above 550 MHz will enable more cost effective plant upgrades to higher bandwidths.

REFERENCES

1) "SkyPix cooks up home satellite dish for $700," EE Times, January 21, 1991, pp. 1-14.

2) "AT&T tests megabit data delivery over twisted-pair wiring," EE Times, December 3, 1991, p. 8.

3) Cablevision, December 17, 1990, p. 41.

GHOST CANCELLING AND CABLE

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Abstract

This paper provides a pragmatic approach to ghost canceller applications in a cable television system. The experience of Rogers Cablesystems in the deployment of ghost cancellers in its Vancouver operation will be shared with other cable operators and a check list of what precautions to take when installing a Ghost Canceller is presented. Further field tests are planned to assess the Ghost Canceller's ability to further recuperate an echo contaminated signal at various depths into the cable network. The ultimate goal of this exercise is to deliver the best possible picture.

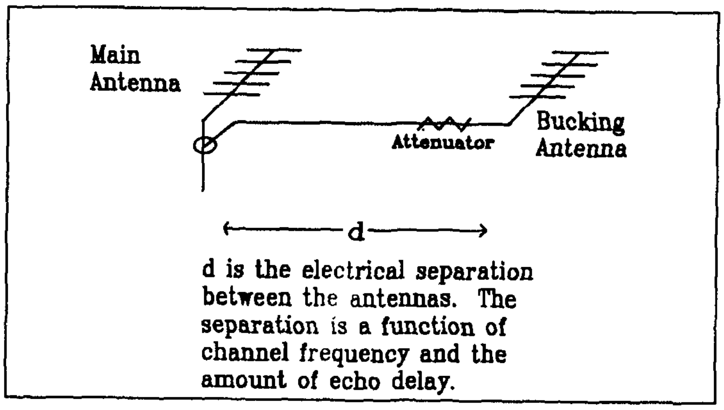
Introduction

Signal reflections from terrain and large structures has plagued broadcast television reception since broadcasting's inception. TV set manufacturers have been studying ways to correct for the resulting "ghosts" for decades but the circuit complexities required to accomplish this have been very costly. More recently with the advent of high speed integrated circuits microprocessors and low cost memory, stand alone "ghost cancellers" have been made commercially available but these are still quite costly for most consumers. Eventually these circuits will be embedded within the TV receiver itself especially within receiver intended for the reception of advanced television formats.

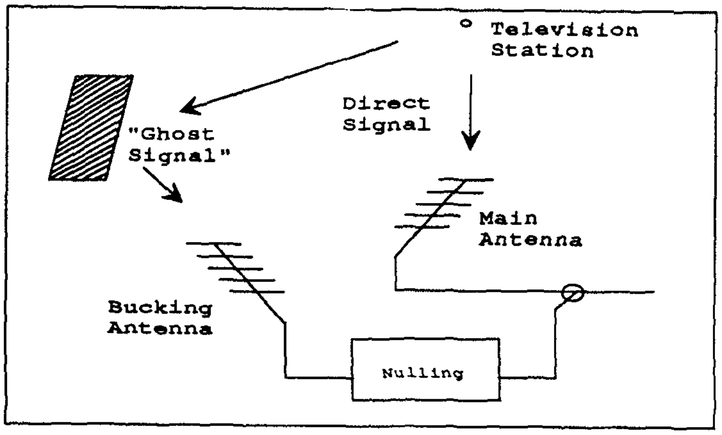
Consumers are demanding better quality reception now. It's going to take many years before a significant number of "ghost cancelling" receivers are in the market place. However, placing ghost cancelling equipment at cable television headends allows all television receivers connected to the cable television network to benefit from the resulting ghost free picture. With the rapidly increasing number of television sets obtaining their broadcast TV reception from cable there is little need to burden each television set with complex and costly echo cancelling circuits.

Pros and Cons about using antenna systems to eliminate "Ghosts"

Cable operators have long been battling the problem of multipath reflections contaminating their received headends signals. Historically, an RF technique (sometimes baseband technique) has been used to cancel the ghost (See Figure 1a, 1b and Figure 2). These techniques require a high degree of precision in trimming long lengths of cable to the appropriate delays or adjusting sensitive phasing devices and matching the amplitude of the reflections using active or passive devices. These techniques, although time consuming to set up, have been quite effective. Once they are set up, little maintenance is required. They usually don't achieve 100% cancellation. In particular, the colour burst signal is somewhat difficult to cancel precisely in phase. In addition, signal-to-noise ratios are sacrificed during signal recombination and any change in



(a)



(b)

Figure 1: RF Cancellation Technique

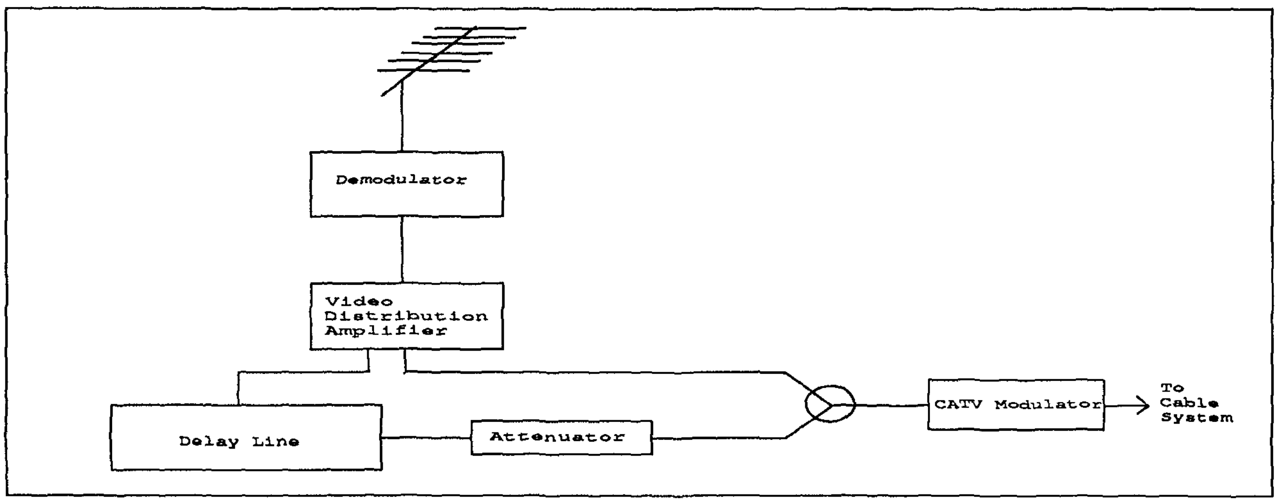


Figure 2: Baseband Cancellation Technique

propagation conditions affects cancellation performance.

Chronology of the Rogers Ghost Canceller Project

In early January, 1990, Rogers Cablesystems embarked upon a project in Vancouver, B.C. to improve the reception quality of the three Seattle based television stations. The three transmitters are located near downtown Seattle. Thus, the signals received at a remote headend site on Salt Spring Island (see Figure 3) were severely impaired by several reflections created by the Seattle downtown core. The situation was deteriorating as developers built additional commercial office towers.

The objective of the exercise, was to improve picture quality through the application of Ghost Canceller Technology. In early 1990, Electronic Ghost Canceller equipment was not commercially available in North America. A few technical journals had published articles about the availability of Ghost Canceller technology and equipment in Japan but none of this equipment was yet available in North America. Rogers was able to obtain some of these units from these Japanese manufacturers. With the cooperation of one of the Seattle broadcast stations to insert special signal in the VBI and the loan of insertion equipment from equipment suppliers, the units were placed in test at Salt Spring Island immediately following the NAB convention. The Ghost Cancellers were put into full operation on the first channel May 11, 1990. Shortly thereafter, the second and third units were put into service.

Fundamentals of Electronic Ghost Cancelling

Two configurations of ghost cancellers are now commercially available in Japan: a stand-alone set top unit and a version integrated within high-end television set. The set top unit 76 is essentially a television tuner demodulator and echo cancelling circuitry. The integrated version incorporates the ghost cancelling circuitry as part of a television sets own base band circuitry. They both use the same principle of operation relying on a reference signal in the vertical blanking interval. The Ghost Cancel Reference (GCR) signal was developed for this application by BTA of Japan. The format of the signal is a sinx/x bar signal (see Figure 4). The reference signal is inserted at the transmitter on lines 18 field 1 and 2 in the vertical blanking interval. It is transmitted in an eight-field sequence. The eight-field sequence is derived as a result of the four-field sequence of NTSC signal and the ease of extraction of the GCR signal by a simple subtraction. The GCR signal is able to cope with multiple reflections up to 44 microseconds of delay. Figure 5 is a generic block diagram illustrating the principle of operation of a ghost canceller. The echo is eliminated by the transversal filters. The tap coefficients of the filters are derived from the output signal and the reference signal. Based on the distorted GCR signal received, the microcomputer applies some mathematical manipulations to transform the distorted GCR to its original shape. The mathematical transform is then converted to the transversial filter tap coefficients.

There are two ways of calculating the tap coefficients of the transversal filters. The first method employs a Fast Fourier Transform with which all tap coefficients are calculated in one single shot. The second method is an iterative method

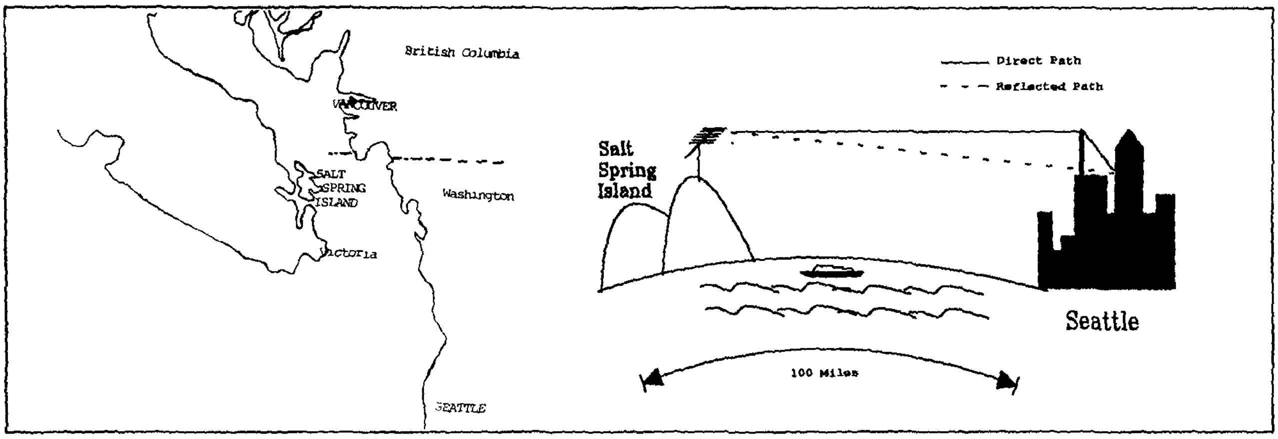


Figure 3: Transmitter and Headend Location

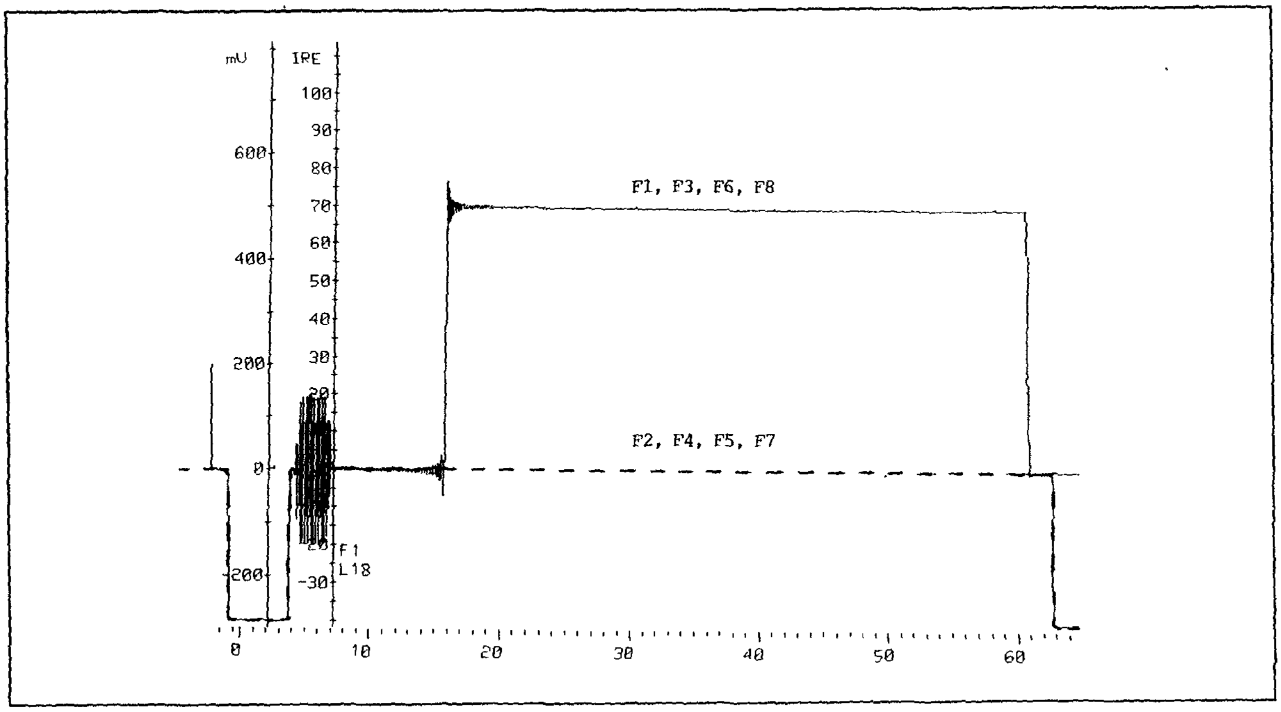


Figure 4: Format of a BTA Ghost Cancel Reference (GCR)

which requires less processing power but demands a longer calculation time. Most consumer grade Ghost Canceller units employ the iterative algorithm and an optimum number of taps for reduced cost while still being able to reproduce a relatively echo free picture. Some versions have differing training method to achieve a faster convergence time. Figure 6 shows a block diagram of a typical Electronic Ghost Canceller.

Installation of Ghost Canceller Equipment at the Headend

To most cable television technicians, installation of a television tuner is simple and considered straight forward. However, implementing Ghost canceller capabilities proved not to be a trivial matter. One first has to obtain cooperation from the broadcaster to insert the GCR signal on line 18 in the vertical blanking interval. One of the broadcast stations in Seattle was already using line 18 for automatic station monitoring. This had to be relocated.

Apart from freeing up line 18, attention must be given to the line preceding a GCR signal. Information on the line can be any test signal except it must be time invariant. Teletext, closed captioning and active video are examples of time varying signals. Such signals compounded with long delays may confuse the ghost cancelling iteration process and adversely affect the convergence.

Another complication with present equipment is channel tuning. Since existing Ghost Cancellers are made for the Japanese market, the tuners' frequency assignments are of the Japanese standard. The units will not tune any low band channels or the first four high band channels. This inconvenience may be overcome by modifying the units to accept input at video baseband. Some manufacturers do provide a video baseband input feature as an option. A problem associated with using the video input is that the cable demodulator feeding the Ghost Canceller must operate in synchronous detection mode. Its' tuner/demodulator characteristic must also be very linear for the ghost cancellation to function properly. Demodulator non-linearity greatly affects the Ghost Canceller performance. Figure 7 outlines an acceptable level of demodulator performance. When AC powering these early Ghost Cancellers, it is handy to note they may have been configured to operate with 100 volts ac, the Japan powering standard. Using 120 volts may seem to be no problem initially but that running a device constantly with a 20 % higher rated voltage will have an adverse effect on its reliability and performance.

Another concern is the 4.5 MHz subcarrier. The Ghost Cancellation process occurs at baseband and the video signal is digitized and processed in the digital domain. The processed video will likely contain digitally generated spurious products which fall above 4.2 MHz. This spurious above 4.2 MHz is of little concern when the Ghost Canceller is used in its consumer application feeding directly into a television set. Cable transmission requires a clean 4.5 MHz aural subcarrier. The spurious outputs resulting from the digital to analog conversion, may create audio distortion described as "a mysterious crackling effect on the main audio" by the customers.

Operating Experience with Various Makes of Ghost Cancellers

A few problems were experienced at the Salt Spring Island antenna site. For example, one of the

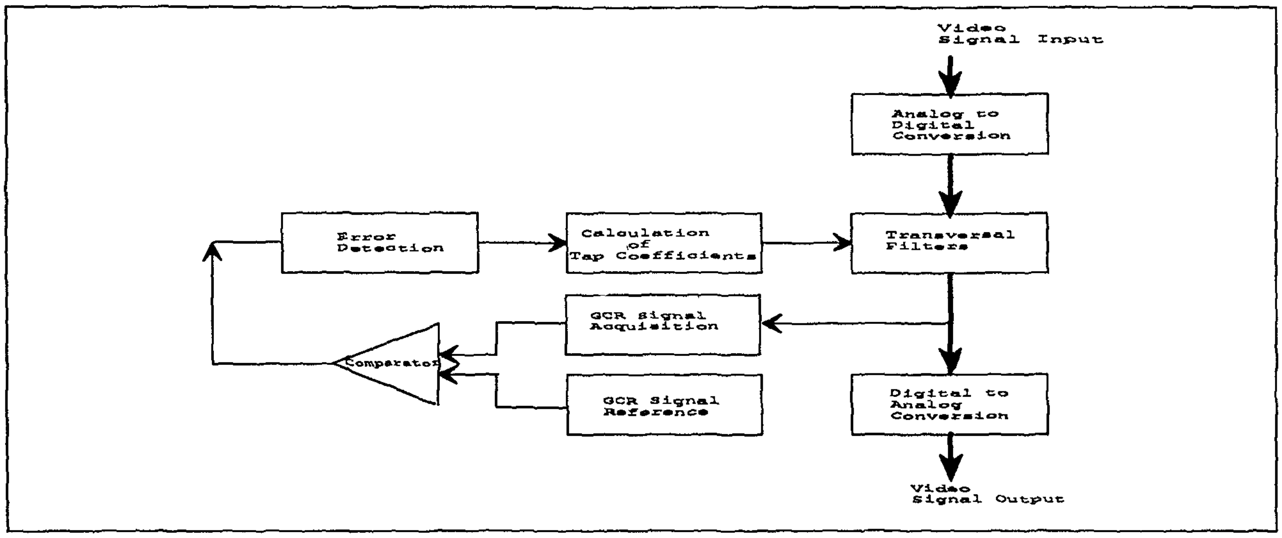


Figure 5: Diagram of Principle of Operation

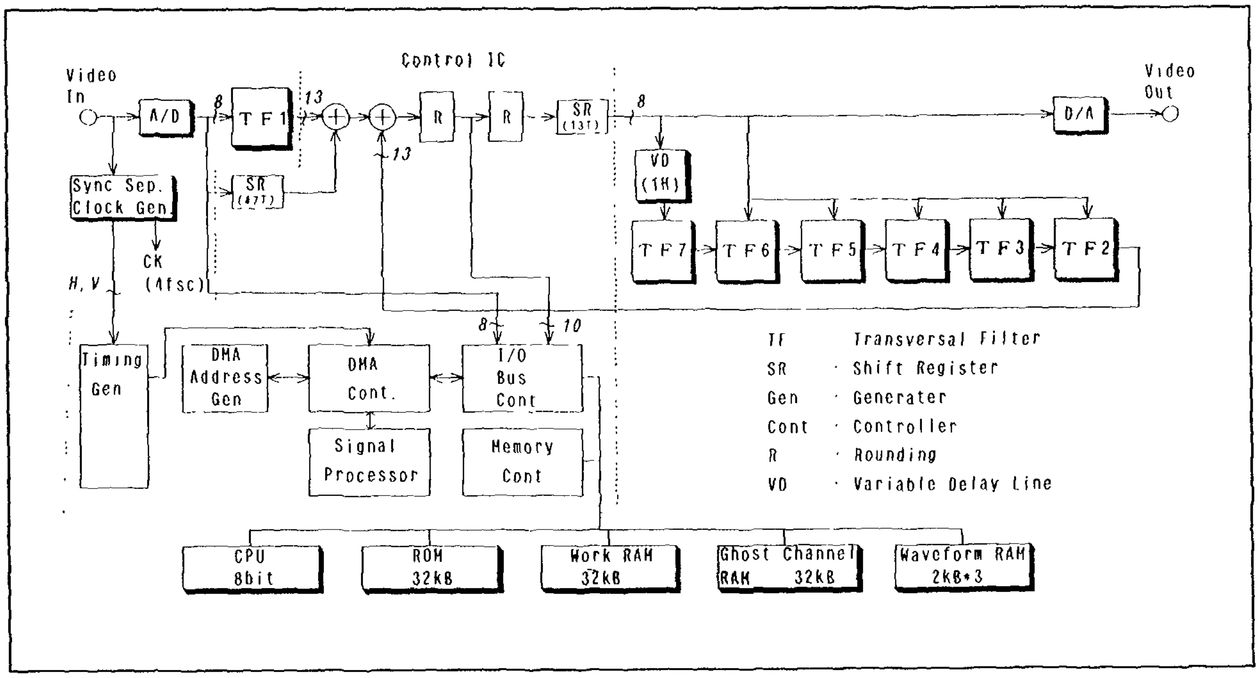


Figure 6: Block Diagram of a Typical Ghost Canceller

(Courtesy of Toshiba Audio Video Engineering Co.
Ltd. --- IEEE Transaction on Consumer Electronics
Vol. 38 No. 4, Nov. 1990)

ghost canceller units would occasionally unlock causing a ghostly picture to appear for a few seconds then correct itself. Another example was that the ghost cancelling iterative process locks-up. Consequently, the unit fails to track the change in ghost condition when propagation condition vary. The latter problem was resolved, in the short term, by installing a remotely controlled switch that would reset the unit when dispatch noticed the condition. In the longer term it is anticipated the manufacturer will correct these defects.

Ongoing Research into Ghost Cancellers and their Application

Rogers Cablesystems and CableLabs are jointly conducting further field tests on a variety of ghost cancellers. Various manufacturers and the three Broadcasters continue to provide their support. In addition to the three ghost cancellers mentioned above, the performance of a more sophisticated Ghost Canceller is being evaluated. This particular unit, although an engineering prototype, is tailored for the professional market. Along with baseband input and output, this Ghost Canceller features interframe processing techniques to achieve received signal noise reduction.

Initially, all the Ghost Cancellers will be tested in a laboratory environment. Each unit will be carefully characterized for its video and/or RF performance. Subsequent to the laboratory testing, a series of over-the-air tests will be conducted at the antenna site in Salt Spring Island to assess each units reaction to varying propagation conditions. The findings will be presented in a report through Cable Television Laboratories, Inc. in Boulder, Colorado.

The second phase of the tests are aimed at assessing the benefits of ghost cancelling equipment in improving the quality of transmission in a cable television network. GCR signal insertion equipment will be installed at Rogers' Vancouver central headend. A full set of propagation tests over coaxial cable plant will be conducted. Two test channels will be chosen, one from the low band and one from the ultraband. Four test points from the trunk will be chosen and correspondingly, four test points from the feeder plant will be selected. Each test point will be monitored over four months of periodic testing. Further test bench analysis will then be made on each individual Ghost Canceller. Each unit will be subjected to a range of noise and non-linear distortion products to determine thresholds of inoperability in the presence of interference. This will assess the ruggedness of the GCR signal to cable system transmission impairments. A final report will be prepared detailing the various findings and results of the analysis. Again, the second part of the report will be presented through Cable Television Laboratories, Inc.

Too Early for Conclusions

Ghost Cancelling equipment available today is mostly aimed at the consumer marketplace for domestic TV off air reception. From a cable operators' perspective, a unit at the headend immediately allows all customers to benefit from improved reception. The most intriguing notion is whether further improvements are possible by migrating the location of the units further into the network. The additional reflections and noise induced by this part of the network would be removed if indeed the Ghost Canceller does

Synchronous Detection	
Noise Figure	<7dB
Differential Gain	<2%
Differential Phase	<1°
Image Rejection	>60dB
Adjacent Channel Rejection	>60dB
Frequency Response	<0.5dB
Envelope Delay	<50 ns

Figure 7: Acceptable Demodulator Performance

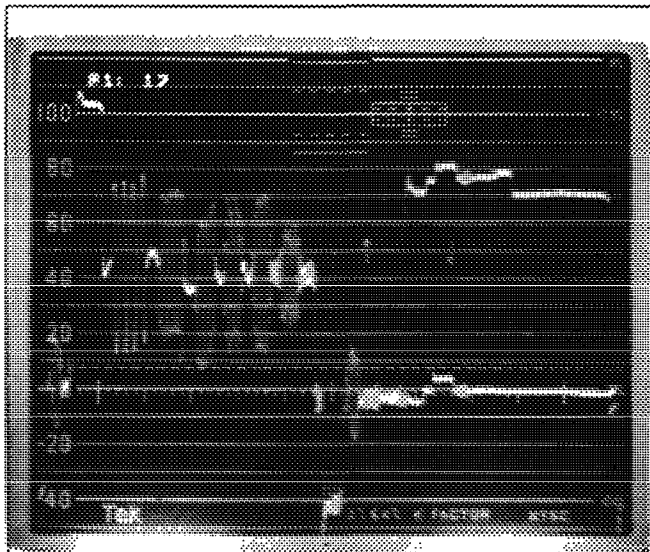


Figure 8a: Video Waveform without Ghost Cancellation

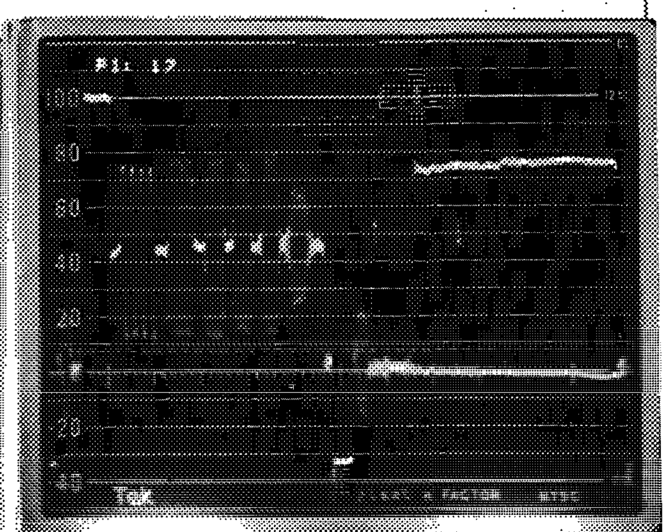


Figure 8b: Video Waveform after Ghost Cancellation

recuperate the original signal. However, with the inclusion of various active elements in the transmission path, will the performance of Ghost cancellation be sacrificed? What will be the tradeoffs?

Experience has already shown that the Ghost Cancellers do not work very well with a clamping amplifier in an FM microwave link. But, is the Ghost Canceller capable of eliminating multipath associated with FM or AM microwave or supertrunking systems?

Most existing ghost canceller equipment is a consumer electronic product and is not designed for the professional environment. How well will it stand up to the 24 hours a day and 7 days a week operating conditions of a headend application? What would it take to produce a unit having the professional quality suitable for broadcast and cable television applications?

What about the possibility of exploiting the Ghost Cancelling concept to tackle co-channel interference?

It is the intention of Rogers Cablesystems and CableLabs to assess the strength and weaknesses of the Ghost Canceller system and to study the full implication of this technology for the cable industry. It is hopeful that partial results of these Ghost Canceller studies will be available at the presentation of this paper.

Acknowledgement

The authors would like to thank Rogers Cable TV in Victoria, in particular, Mr. John Foss and Cor Maas for their assistance in conducting the initial field tests. The ongoing evaluation of the Ghost Cancellers is sponsored by CableLabs. The cooperation of Tektronix, KOMO, KIRO and KING stations and the support of various Ghost Canceller manufacturers is vital to this ongoing project. Their help and support is most appreciated. Currently, Rogers Cable TV has four Ghost Reduction Devices (JVC, NEC, SONY and Toshiba) in daily headend operations at the Salt Spring Island antenna site.

References

New Ghost Reduction Tuner Using a Training Method by Yutaka Miki et al. Matsushita Electric Industrial Co. Ltd.; IEEE Transactions on Consumer Electronics, Vol. 36, No.3, August 1990.

Development of a Ghost Cancel Reference Signal for TV Broadcasting by Hiroshi Miyazawa et al.; IEEE Transactions on Broadcasting, Vol. 35, No.4, December 1989.

Ghost Clean System by Hiroyuki Iga et al; IEEE Transactions on Consumer Electronics, Vol. 36, No. 4, November 1990.

Training Signal and Receiver Design for Multipath Channel Characterization for TV Broadcasting by J.D. Wong et al; IEEE Transactions on Consumer Electronics; Vol. 36, No. 4, November 1990.

Biographies

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Nick Hamilton-Piercy is the Vice-President, Engineering and Technical Services for Rogers Cablesystems Limited, his duties include managing the corporate engineering group, Rogers Engineering, as well as being responsible for the engineering and capital programs within the Company operating systems. Nick has been with Rogers for 15 years.

Prior to this, Nick was with the Canadian Marconi Company where he managed an analogue and R. F. engineering department responsible for the design development and manufacture of a wide range of telecommunications equipment including microwave radio systems, multichannel UHF tactical relay systems and microwave R.F. hardware.

Nick graduated in Engineering from the Medway College of Technology in Chatham, England, during 1961 where he specialized in electrical engineering and qualified for full Chartered Engineering status. He is a member of the Institute of Electrical Engineers (IEE) U.K.; a member of the Association of Professional Engineers of Ontario (APEO); a Senior Member of the Institute of Electrical and Electronic Engineers (IEEE); a Senior Member of the Society of Cable Television Engineers (SCTE); Chairman - Futures Committee, Canadian Cable Television Association (CCTA); Chairman ATV Subcommittee of Cable Television Laboratories Technical Advisory Committee, and Chairman of the National Cable Television Association Engineering Committee - HDTV Subcommittee.

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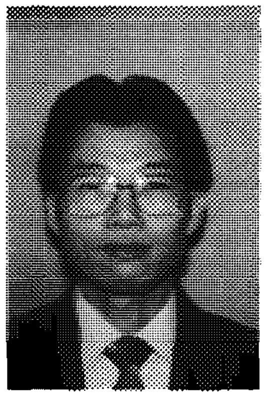
Gary Chan is a Staff Engineer at Rogers Engineering, Toronto, Canada. He received his B.Sc.E.E. degree in 1980 from the University of Manitoba and Master of Engineering (Electrical) degree in 1986 from the University of Alberta. Since his graduation in 1980, he has engaged in CATV design, operation, maintenance, research and development of advanced cable television technologies. He is a member of the IEEE and a registered Professional Engineer of Ontario.



IMPACT OF DISPERSION ON ANALOG VIDEO TRANSMISSION

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American Television and Communications Corporation

Douglas E. Wolfe
Corning Inc.

ABSTRACT

As the development of 1550 nm AM technology progresses, the impact dispersion will have on analog transmission has become an important issue. This paper discusses the effects of dispersion on AM video signals and possible solutions that will allow for the utilization of 1550 nm AM systems on standard single-mode fiber.

INTRODUCTION

As the development of 1550 nm AM technology continues, it is imperative that we evaluate and understand the effects that dispersion, which is greater at 1550 nm than at 1310 nm on standard single-mode fiber, will have on signal quality. Recent testing has discovered that high dispersion will cause a significant increase in the composite second order (CSO) distortions. CSO degradation was found to be one of the limiting factors on 1310 nm AM systems and with increased CSO degradation occurring at 1550 nm, there appears to be some uncertainty within the cable TV industry on the future of current 1310 nm fiber systems and their potential for 1550 nm upgrades.

CAUSES OF DISTORTIONS

There are two characteristics of today's lasers and fiber that contribute to CSO degradation of analog video signals: laser chirp and fiber dispersion.

The laser characteristic that contributes to the delay in analog transmission is called laser chirp. Laser chirp is caused by changes in the refractive index in the cavity of the laser due to changes in the current applied to the device. As modulation is applied to a laser, the current through the laser increases and decreases as a function of the applied modulation. When the current is at its peak, the wavelength of the laser is increased. When the current is at its minimum, the wavelength of the laser decreases.

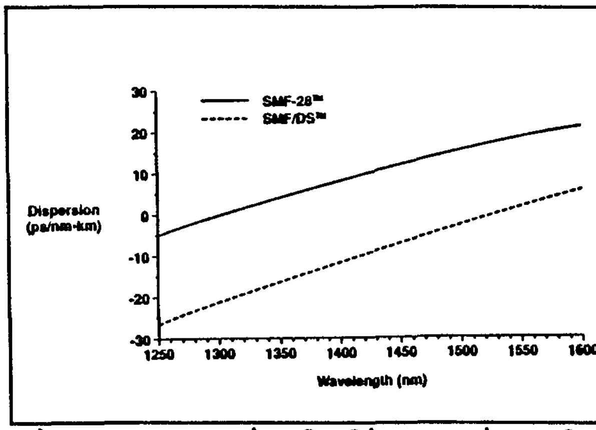

Figure 1 Typical dispersion for Corning's SMF-28™ and SMF/DS™ single-mode optical fiber.

The fiber characteristic that contributes to the delay is dispersion. Standard single-mode fibers are designed to have zero dispersion near 1310 nm. However, dispersion increases in a single-mode fiber with wavelength due to changes in the fiber material index of refraction. By examining Figure 1, it can be seen that as the wavelength of the laser increases, there is an increase in the delay caused by the fiber.

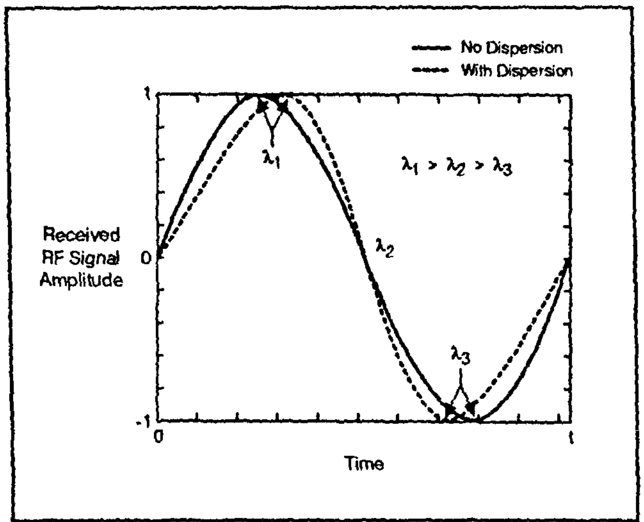

Figure 2 Delay distortion produced when a sine wave is transmitted on high dispersion fiber.

Together these two characteristics cause an amplitude dependent signal delay. This delay causes a transmitted sine wave to distort its shape, thus generating CSO distortions that significantly degrade signal quality. If you were to analyze a single RF carrier, the effects of this delay would cause the positive portion of the carrier to travel faster than the negative portion of the carrier, thus distorting the carrier's shape as it propagates through the fiber cable (see Figure 2 from [1]).

ANALOG DISPERSION EFFECTS

Several tests have been conducted to determine what the effects of high dispersion would be on an analog signal. In one test, a 1310 nm laser was operated on dispersion shifted single-mode fiber to create a high dispersion optical path. This test was conducted using 60 channel loading. Tests were also conducted by Synchronous Communications Inc. using a 1550 nm laser on standard single-mode fiber. This test was conducted using 35 channel loading.

The CSO of the system was affected by the high dispersion path as a function of the individual lasers. Figure 3 graphically depicts the amount of CSO degradation for a 1310 nm AM system on standard single-mode fiber and on dispersion shifted single-mode fiber, and for a 1550 nm system on standard single-mode fiber. Both the 1310 nm and the 1550 nm systems show a substantial decrease in the CSO performance when operated on fiber that is not optimized for that wavelength. The 1310 nm system operation was virtually unaffected when operated on standard single-mode fiber. There was also a frequency dependence observed. As the carrier frequency increased, the effects of high dispersion were more pronounced (Figure 4).

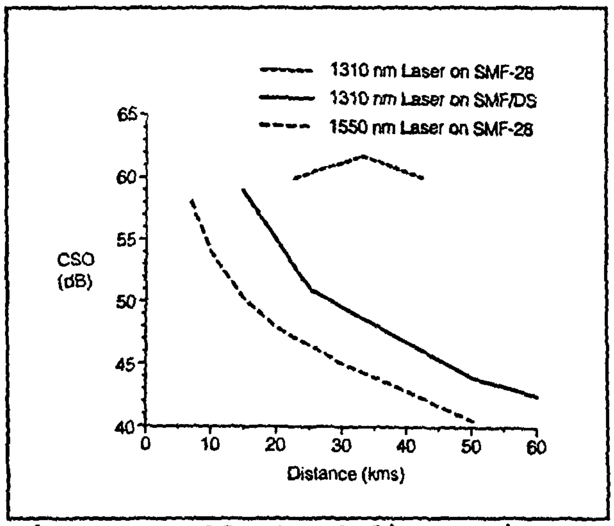

Figure 3 Effect of dispersion on CSO.

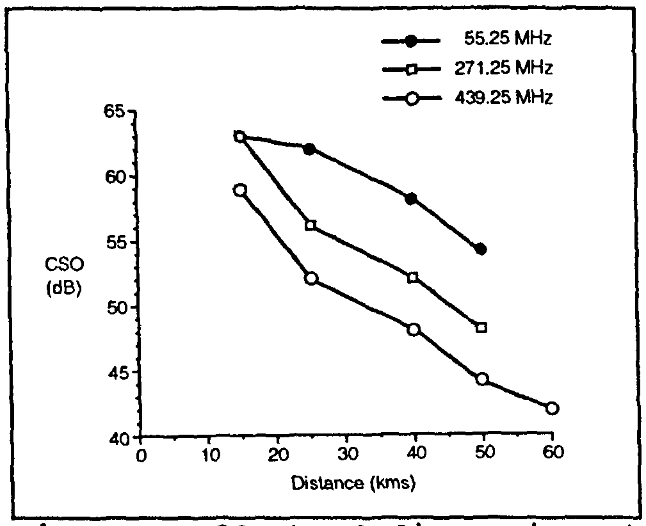

Figure 4 Effect of dispersion at multiple frequencies.

SOLUTIONS TO THE PROBLEM

The results clearly show that the performance of 1550 nm analog systems on standard fiber will be degraded, but there are several things that can be done to minimize the problem. These include delay compensation, operations in octave bandwidths, improved laser technology, the use of digital modulation, and the use of dispersion shifted fiber.

Synchronous Communications Inc. made a presentation at the SCTE Fiber Optics 1991 conference in Orlando [2] that stated the delay distortions, that are experienced at 1550 nm on standard fiber, could be compensated for with an electrical distortion compensation circuit. The compensation circuit produces the opposite effect of the fiber and laser delay characteristic. The circuit, which has been patented by Synchronous, produces voltage variable delay by using the applied modulation to vary the voltage across a varactor diode. The varactor, which changes its capacitance as a function of voltage, is used in conjunction with an inductor to form the voltage control variable delay circuit. Tests with the compensator circuit installed on a transmitter and optimize for 12 km of fiber, achieved CSO levels of 60 to 64 dB over 5 to 17 km of standard single-mode fiber. Since the delay distortions are distance dependent, the circuit may need to be installed at receivers since transmitters will be shared among multiple receivers over various distances.

Another method for dealing with the CSO problems will be to place all of the signals into a single octave of bandwidth. When all of the channels are placed in an octave of bandwidth, no combinations of channels will produce second order products that will fall within this octave. This may be the most practical solution for dealing with the expanded channel capacity issue that concerns most operators who have already installed fiber. Since our present 1310 nm AM systems are capable of delivering 50 to 450 MHz on a single fiber, when channel expansion is desired, the additional channels up to 900 MHz can be added to a 1550 nm AM system without any second order products. With the addition of wave division multiplexers at transmit and receive sites, the fiber system can be easily upgraded. As 1310 nm lasers improve, this technique should allow expansion to 1 GHz bandwidth.

Alternate modulation methods may also be used on the signals that are transmitted on 1550 nm systems. ATC believes that most of the bandwidth past 550 MHz will be used for pay per view (PPV), near video on demand (NVOD), and other services that may be non-entertainment. If the services are PPV or NVOD, then VSB-AM may not be an appropriate modulation technique since it is difficult to secure VSB-AM signals without causing visual impairments. Digital modulation for these services would be a more likely choice and would solve two problems at one time. It would avoid the dispersion problems associated with AM transmission and at the same time it would make transmission of these pay services more secure. However, digital modulation is not without its problems. The useable bandwidth (bit rate) of a digital system will be determined by the same dispersion that causes CSO problems in AM transmission. Without the development of compression, the number of digital channels that will be available on a single laser at 1550 nm will be limited.

Another method of dealing with the problems of dispersion at 1550 nm on standard single-mode is to install dispersion shifted fiber. There will be applications that require cable operators to deliver signals over longer distances than are possible with today's 1310 nm systems. Using 1550 nm lasers and dispersion shifted fiber, a system radius could be extended 1.5 to 2 times the distance at 1310 nm. When these scenarios arise, dispersion shifted single-mode fiber and 1550 nm will probably be the lowest cost alternative when compared to FM and digital super trunking.

Many of the new laser structures that are in the laboratory today will also minimize dispersion effects. Most of these new structures will result in narrower line widths and a significant reduction in laser chirp. Since dispersion is proportional to chirp, the dispersion effects will be reduced proportionally to the reduction in these items. External modulation of lasers could also substantially reduce the delay distortions. External modulation eliminates laser chirp and the lasers that are used have very narrow line widths which makes them a viable alternative.

SUMMARY

There are problems associated with the use of 1550 nm on standard fiber, but there are also many methods for dealing with the problems. The solutions that are described above will solve most of the problems caused by dispersion and any one or a combination of them may be used to allow the continued expansion of existing and future fiber systems.

REFERENCES

[1] H. Gysel et al., Electrical Pre-Distortion to Compensate the Combined Effect of Laser Chirp and Fiber Dispersion. Electronics Letter, Issue 5, February 28, 1991.

[2] H. Gysel: CATV AM Optical Transmission Links Using the 1550 nm Window, SCTE Fiber Optics 1991, Orlando, Florida, January 1991, pp 161-166.

[3] L. Williamson et al., Effects of Chromatic Dispersion on Analog Video Transmission. SCTE Fiber Optics 1991, Orlando, Florida, January 1991, pp 177-183.

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IMPACT OF MICROWAVE TECHNOLOGY DEVELOPMENTS ON CATV SYSTEM DESIGN

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Abstract

Recent microwave circuit developments have led to the realization of a third order intercept point of +62 dBm in a broadband block conversion type 13 GHz solid state transmitter. The 5 dB increased power output that this latest development allows adds to previous CARS band microwave block conversion transmitters, which in total now span a power capability ratio in excess of 200: 1. A summary explanation of the steps culminating in the dual-feedforward circuit utilized in this highest power transmitter is provided. The impact of improved third order intercept on noise and distortion in LDS microwave operation is reviewed. The resultant range capability is examined for various geographical areas. Comparison with fiber optic transportation systems is made and the advantages of CATV architectures employing parallel fiber and microwave paths is explored.

AML® TRANSMITTER DEVELOPMENTS

Seven years ago, CATV systems which utilized AML transmitters in microwave local distribution systems operated either with what was known as "high power" AML or " low power" AML. The former employed one klystron per channel and the latter had one klystron per eight channels, but both transmitters were based on a channelized design. These 13 GHz transmitters which were developed in the early 1970s were vastly superior to an even earlier 18 GHz design which was based on a high power broadband traveling wave tube (TWT) carrying the then standard 12 TV channels. The essential difference between these two approaches is that in the channelized approach the power output is limited only by the video-audio intermodulation product (which falls in the lower adjacent channel), while in the broadband transmitter the limiting third order distortion is the far more demanding composite triple beat (CTB) between video carriers. In the case of the TWT this severe distortion limitation is further aggravated by the large amount of noise generated within the TWT. With even a modest number of multiple carriers, the power capability of the TWT comes to naught because of the large amount of backoff in power dictated by CTB considerations, while the output noise floor looms ever larger. The squeeze between carrier-to-noise ratio (C/N) and C/CTB is all too familiar to CATV system designers.

The key microwave technology advance which permitted the development of reasonably well performing broadband transmitters was the advent of the high power field effect transistor (FET) amplifier. These early solid state devices, while not able to match the power output capability of the TWT, had a huge advantage in terms of noise performance which is critical to system performance. Nevertheless, the first solid state broadband transmitters were not even in the same performance ballpark as even the "low power" channelized equipment. However, in one critical respect, namely cost, the broadband transmitters were very attractive. Many applications, for which classical channelized AML could not be economically justified, were suitable for broadband microwave utilization.

High power FET technology has improved significantly during the past seven years. Saturated power output capability has increased from 1 watt to as high as 10 watts. Reliability has been proven by hundreds of 5 watt amplifiers which have been performing flawlessly for several years. However, this in itself could not close the 18 dB performance gap between the initial broadband transmitters and a single output of the channelized transmitter. A second microwave technology advance had to occur to bridge the gap. This was achieved in 1989 through the application of feedforward linearization.[1]

It must, of course, be noted that the comparison between the transmitters does not take into account that the channelized transmitter has multiple outputs, each of which have the same power per channel and the full complement of channels. Thus, while the IBBT-116 transmitter, operating at a level corresponding to 65 dB C/CTB, matches the power capability of the channelized MTX-132 transmitter (the comparison is reasonably accurate over a wide range of channel loading), the comparison is made in terms of the power available at the output port, which for the channelized transmitter is by no means the total output. Indeed, for a 64-channel system the standard multiplexing arrangement leads to 16 outputs. Often, however, these outputs are not all utilized. Moreover, some of the receive sites may be located relatively close to the transmitter so that the full output of that transmitter port, which feeds the close-in receiver, is not required. On the other hand, the broadband transmitter can easily incorporate a customized power splitting network which matches the needs of the application. Thus, in a typical application, the *total* power comparison gap is more like a factor of three than a factor of 16. This gap now has been closed by the latest broadband development.

AML-HIBT-118 TRANSMITTER

The performance of the HIBT-118 is summarized in Table 1. If one were to compare this capability to the IBBT-116, it would be seen

TABLE 1
HIBT POWER OUTPUT AND C/N FOR 65 dB C/CTB AND C/CSO

Number of Channels	P_o (dBm)	C/N (dBm)
12	20	66
21	18	64
35	15	61
60	12	58
80	11	57

that the output is 5 dB greater. Thus, the HIBT-118 output is also 5 dB greater than any one port of the MTX-132 and only 5 dB less than the output port of a high power STX-141 array. To describe the performance in yet another way, it would be accurate to say that, if it were to exist, a typical 200-watt FET amplifier, with 62 dBm third order intercept point, operated at the same output power and the same channel loading, would yield the same C/CTB.

The technological innovation utilized to achieve this performance is illustrated in Figure 1. The circuit consists of a feedforward amplifier imbedded within a feedforward correction loop. Although the concept is not entirely new[2], we believe this to be the first instance of application of this concept to a commercial product. In any case, the resultant performance breakthrough can have important consequences in CATV microwave Local Distribution Service (LDS) design. Figure 2 shows a photograph of the HIBT-118 transmitter. One aspect is immediately evident from this photo: the space required to house an 80-channel reasonably high-powered transmitter in the headend is now quite minimal. A close inspection of the photo shows the four heatsinks corresponding to the four 5-watt FET amplifiers included in Figure 1. These

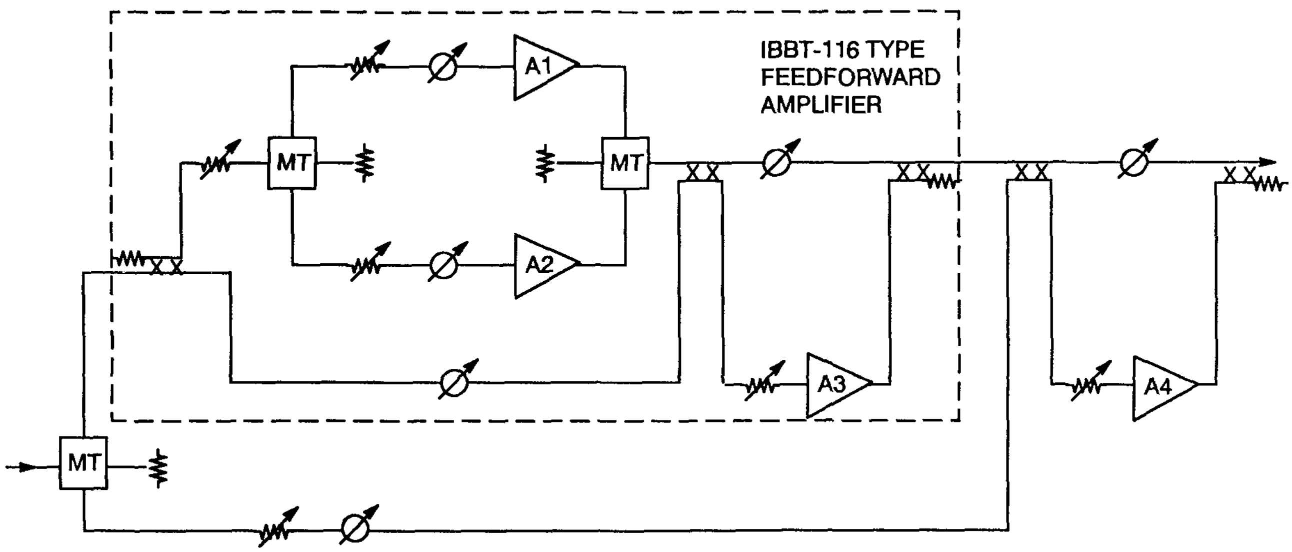

Figure 1 HIBT-118 double-feedforward output amplifier.

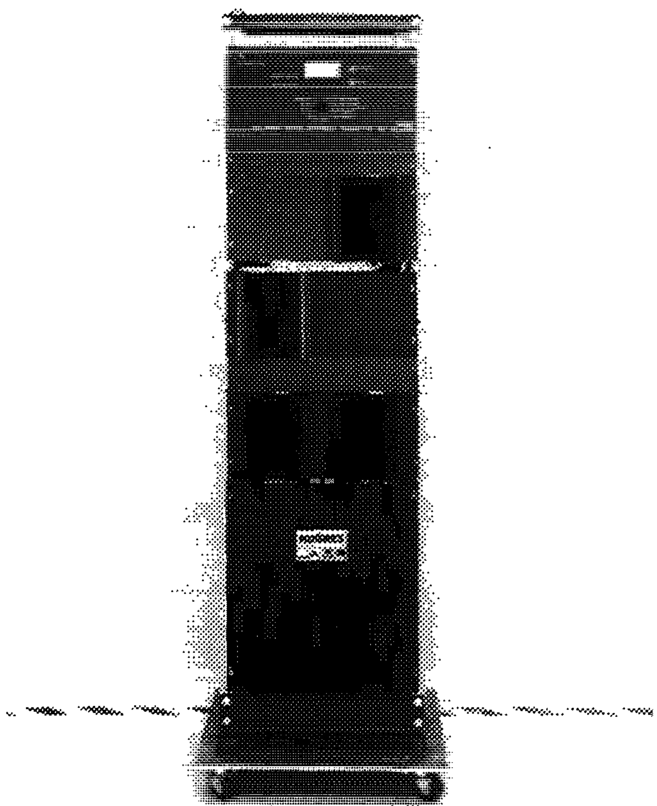

Figure 2 High power indoor broadband transmitter.

heatsinks operate only slightly above room temperature, the total transmitter power dissipation being less than 380 watts. Needless to say, the transmitter cost is also much less than a channelized system.

LDS SYSTEM DESIGN IMPLICATIONS

It may be of interest to consider the relative cost trend that the increase in broadband power output has followed. Certainly the higher power transmitters cost more, but the cost per unit of *total* power declines rapidly as power increases. Interestingly enough, the channelized transmitters are still the most cost effective choice if their full total output power capability is required. Such is the case in major metropolitan areas where up to 30 or more AML receive sites may serve as CATV distribution hubs. However, more typically there are only about four hubs, and then the HIBT-118 is far more cost effective while at the same time having the capability of servicing an intermediate size area. Even large metropolitan areas with multiple receive sites could be serviced through an array of HIBTs as shown in Figure 3.

The question of the range over which the HIBT can operate is certainly of interest. It is, however, very much complicated by all the associated parameters which may vary greatly. To illustrate the effect of just one variable, the location of the LDS system, consider a single microwave link with 10-ft antennas at each end. Assume further that a clear weather C/N of 56 dB and a C/CTB of 65 dB is required for 40-channel loading. Finally, assume that there are a total 4 dB of miscellaneous waveguide losses at transmitter output and receiver input. Table 2 summarizes the path distance for various multipath conditions and CCIR rain zones if only 1 hr/yr of fading below 35 dB C/N is permitted. The low and high distance extremes could for instance represent locations in Florida and Idaho.

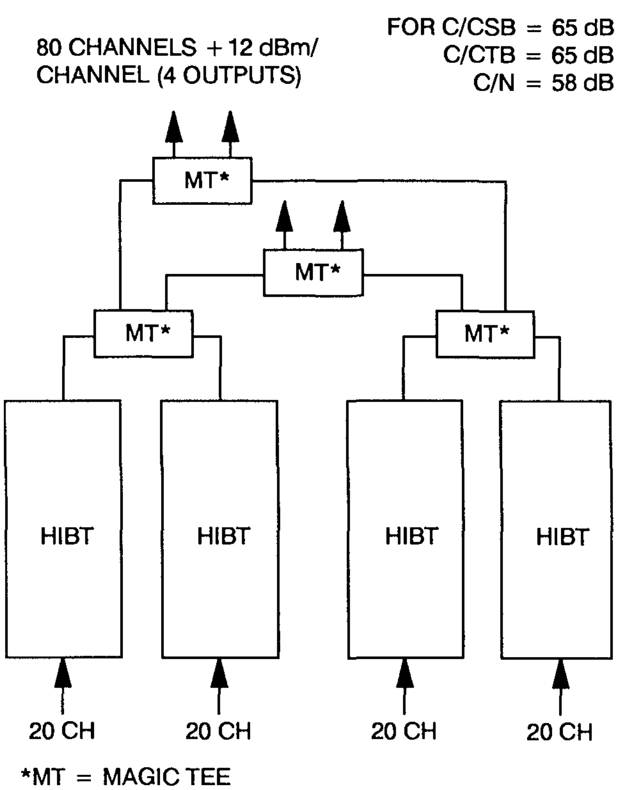

Figure 3 Multiple HIBT redundancy configuration.

Another aspect of the HIBT which could be of considerable interest to cable operators is illustrated by Figure 4. Here it is assumed that an existing 32-channel MTX-132 transmitter services four communities and that a channel expansion to 67 channels is desired. Referring to Table 1, the 35 additional channels can all be accommodated with the HIBT at an output level of +15 dBm. This closely matches the capability of a MTX-132 transmitter, having six channels in a circulator string, after the first level of magic tee splitting and combining. Therefore, the desired channel expansion can be accomplished with only 1 dB reduction in power out due to rearrangement, as indicated by Figure 4, of the original circulator channel multiplexing arrangement, and only one rack of equipment (the HIBT) is added. This compares with a 3 dB reduction and five additional MTX-132 racks at considerably greater cost if the channel expansion were implemented in the traditional manner.

Figures 3 and 4 are only two of almost innumerable combinations taking advantage of the flexibility of the high power broadband transmitter characteristics. If desired, an extra decibel of power output can be traded against a 3 dB reduction in C/CTB. In some circumstances, the HIBT, either alone or in combination with an IBBT, could entirely replace an old MTX-132 installation with

TABLE 2
RANGE OF 40-CHANNEL HIGH QUALITY
HIBT-118 LINK*

Multipath	CCIR Rain Zone	Distance (km)
Worst	E	12.1
Average	E	12.1
Average	D2	24.9
Average	B1	31.7
Best	B1	45.4

*See text for further definition.

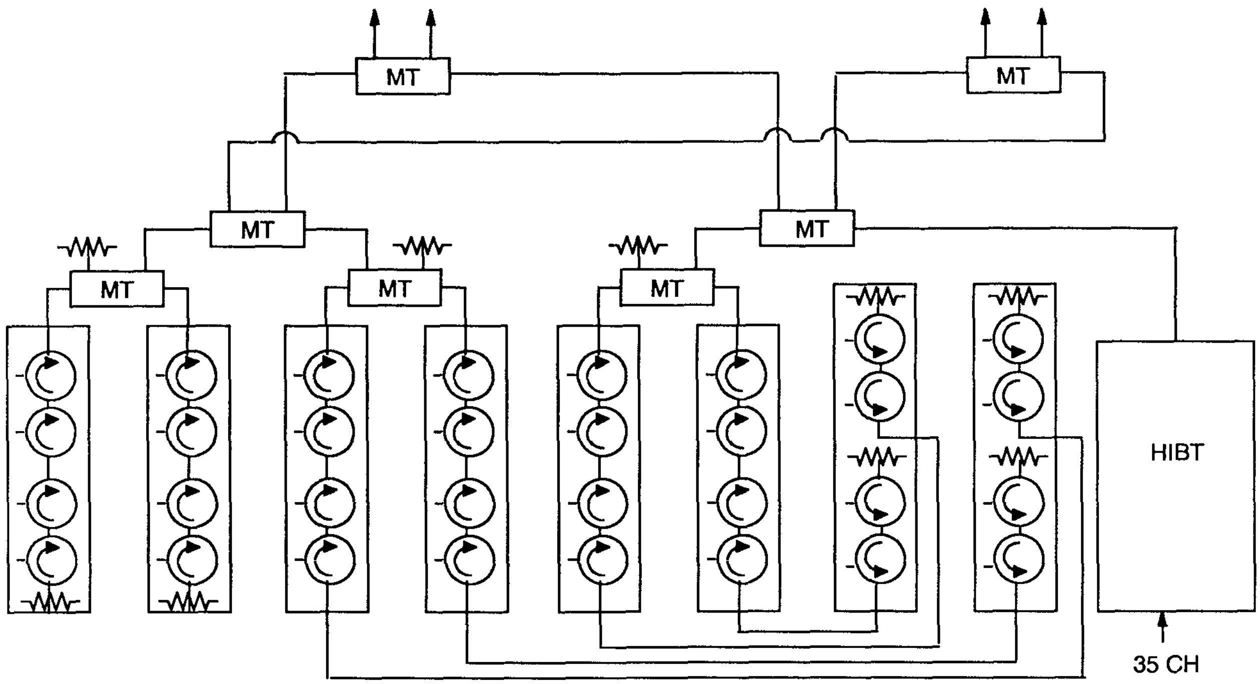

Figure 4 MTX-132 channel expansion using HIBT.

benefits to performance and additional headend space. Other rearrangements involving STX-141 arrays are also conceivable.

CATV SYSTEM DESIGN CONSIDERATIONS

Broadband microwave technology developments have enabled performance improvements which maintain a substantial edge over state-of-the-art AM fiber optic systems.[3] More importantly, the technologies can often complement rather than compete with one another. An obvious example is the case of a redundant supertrunk arrangement in which the microwave is backed up in case of rain fades and the fiber is insured against a cut. At the same time both are backed up in the case of equipment failure. For such applications a wide range of microwave broadband transmitters, summarized in Figure 5, are now available to fit the needs of the application.

One other aspect of broadband systems is worth mentioning. With the advent of digital compression techniques which potentially allow for even better spectrum conservation than traditional VSBAM, CARS band microwave will be able to carry many more channels than is presently possible. In addition, very large fade margin improvements would accrue if the system carried only digitally formatted signals. Although channelized systems would also benefit in both regards, the broadband systems can accommodate any carrier spacing without any modifications of existing filters.

SUMMARY

Advances in microwave technology have resulted in a wide array of products with which CATV system designs may be optimized. The highest power transmitter in this category is capable of exceeding by 5 dB the per port output of traditional "low power" channelized AML. This breakthrough in performance capability enables a wide variety of economical

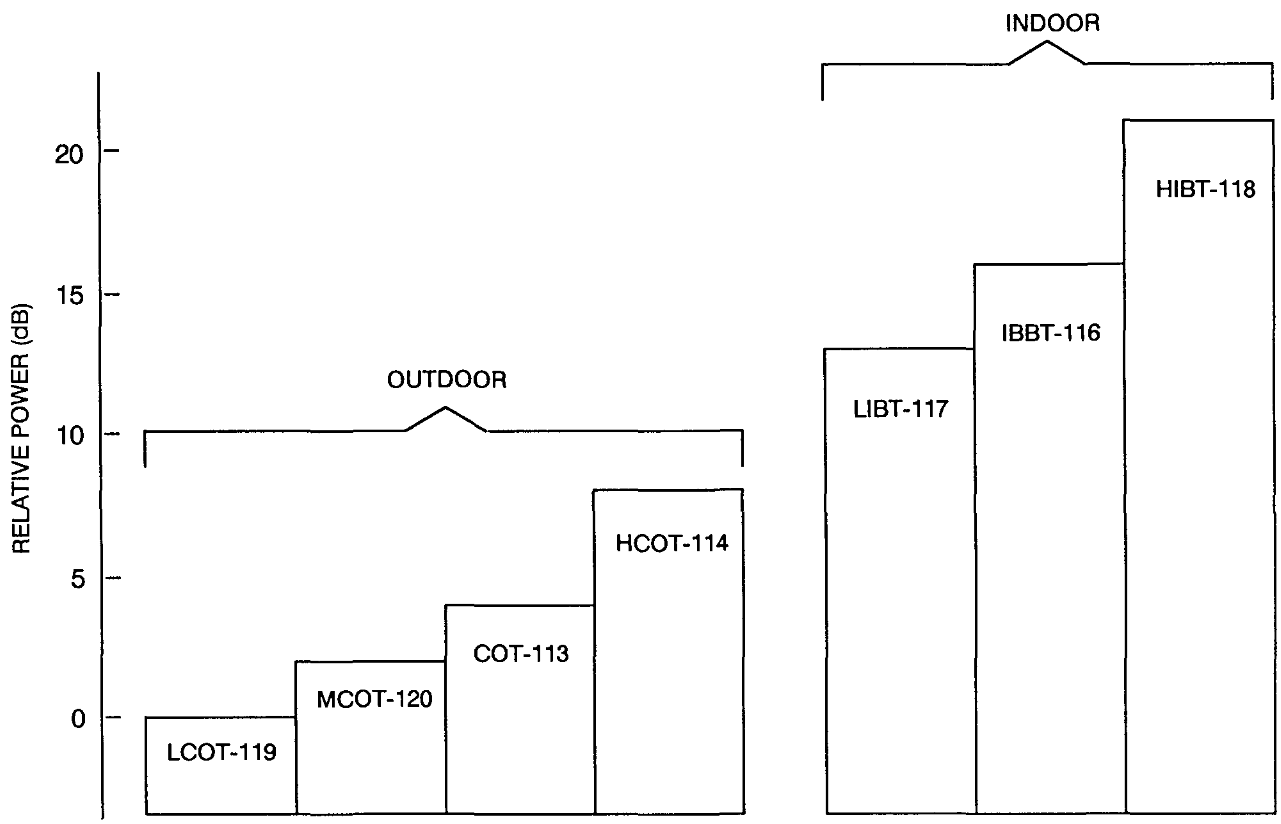

Figure 5 Broadband transmitter comparison.

LDS microwave configurations which heretofore have not been possible. The implications of this and future developments for system rebuilds could be of major benefit to CATV systems.

REFERENCES

1. T.M. Straus and R.T. Hsu, "New Higher Power Microwave Broadband Transmitter Design and System Application", *CCTA Technical Record*, 1989 (Toronto)

2. H. Seidel, "A Microwave Feedforward Experiment", *BSTJ 50*, No.9, November 1971.

3. J. Lipson, C.B. Roxlo, and C.J. McGrath, "AM Fiber Links: Performance Limits and Reliability", *CED*, November 1990.

Lightwave Multi-channel AM-VSB Video Systems: Technology Trends and Limits

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ABSTRACT

AM-VSB lightwave technology has advanced rapidly to meet the requirements of the CATV industry. This paper discusses the present state and projected progress of the most promising of many evolving technologies. Presently available fiber systems, based on 1.3 μm wavelength DFB lasers, can be implemented today with considerable cost and performance improvements over coaxial cable. These systems offer performance approaching fundamental limits and, with continued modest increases in output power (3 dB), offer opportunities for limited (4 way) splitting. Externally-modulated YAG lasers could offer similar capabilities, but have not yet been commercialized. Fiber amplifiers offer a possibility for increasing systems margins to the extent required for large scale (> 16 way) splitting, but serious technical hurdles must be overcome. Implementation of such future possibilities would follow naturally from an established base of today's fiber technology.

1. Introduction

With the recent availability of high-performance DFB lasers and the promise of high-power diode-pumped YAG lasers modulated with linearized external modulators, lightwave technology is beginning to make radical changes in the technology base of the CATV network. The ultimate form of this new capability is not yet clear. Advances in system performance have been rapid over the past two years. Laboratory demonstrations of linear optical amplification using Erbium-doped fiber amplifiers (EDFAs) may open the way for passively-divided fiber networks. The system designer is now faced with the choice between implementing currently-available components and waiting for even better future technology. Unfortunately, no-one can be certain of how each of these technologies will progress.

Although one is tempted to wait and watch the competing technologies evolve, the evolution of each must be driven by a real market. Continued improvements are a certainty, but new systems and rebuilds designed using today's lightwave technology offer significant opportunity for reduced cost and improved performance as well as a platform for tomorrow's services. These advantages can be realized now, and should supercede fears of today's technology becoming obsolete. With linearity close to fundamental limits and numerous vendors emerging with comparable state-of-the-art capabilities, the emergence of a new panacea is unlikely.

In this paper we present an overall perspective on the feasibility and merit of the numerous technology options. We show that although advances will continue within each technology, these advances will be incremental. Optical power levels could continue to increase and noise powers decrease, resulting in modest (< 3 dB) improvements over each of the next few years. Technology breakthroughs realized over the past two years have cleared a path for lightwave CATV, but continued improvements require optimization (< 3 dB), not revolution (> 10 dB). This optimization requires extensive engineering, which must be driven by a volume market. Widespread acceptance of lightwave technology makes sense today and is necessary to drive further improvements.

2. Fundamental Limits

In order to scrutinize claims from aggressive players in a highly competitive market, one would benefit from clearly defined bounds on what is and what is not technically feasible. Unfortunately, although such bounds have been defined, feasibility, at least among the key competitors, must be qualified by device yield or system reliability. These proprietary intangibles cannot be clarified merely by screening for fundamental feasibility. Nevertheless, the fundamental limits play a crucial role in understanding the overall status of the technology. Noise performance within a few dB of the quantum limit and modulation depths approaching the clipping limit suggest "state-of-the-art" performance. If such systems can be delivered with acceptable prices and in large volumes, then one can assume that the supplier has satisfactory control of the proprietary intangibles. One can then install a "state-of-the-art" system with confidence knowing that a technology breakthrough will not render it obsolete. Two fundamental bounds have been defined; the quantum (shot noise) limit[1] and the clipping limit.[2] Shot noise is generated whenever current flows in a diode and limits the carrier-to-noise ratio (CNR) in accordance with:

$$CNR_o = \frac{I_o m^2}{4qB}, \tag{1}$$

where I_o is the total (average) received photocurrent, m is the modulation depth per channel, q the electronic charge and B is the noise bandwidth of the channel (4 MHz).

Clipping limits the allowed modulation depth for systems using direct or external modulation. This limit exists even if one assumes that the light-versus-current (L-I) characteristics, for a directly modulated laser, or light-versus-voltage (L-V), for external modulators, are perfectly linear within a limited operating range. Infrequent excursions of the modulation current, for a directly-modulated laser, beyond this operating range (clipping) results in a total interference that can be described by:

$$CIR = \sqrt{2\pi}\,\frac{(1\,+\,6\mu^2)}{\mu^3}\,e^{\frac{1}{2\mu^2}}, \tag{2}$$

where

$$\mu = m\sqrt{N/2}. \tag{3}$$

N is the total number of channels. Eqn. (2) predicts the total interference from all orders of distortion, which is approximately the sum of all second- and third-order products within each channel. However, since clipping results in a rapid increase in higher-order distortions, the sum of the composite second order (CSO) and composite triple beat (CTB) will generally be less than this CIR. For external modulators one must divide the right-hand side of Eqn. (2) by a factor of two, to account for clipping at both zero and maximum light intensity.

These fundamental limits are plotted in Figure 1, for direct modulation and various values of m, I_o and N. The 3dB reduction of the clipping limit for external modulators makes little practical difference, because of the steep slope of the clipping curves.

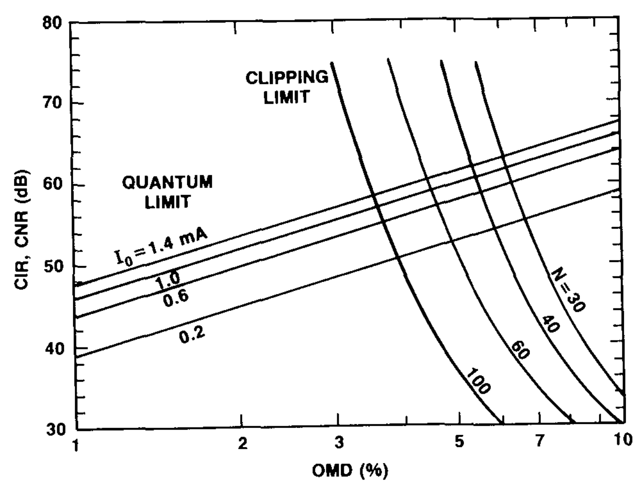

FIGURE 1

Fundamental limits of Carrier-to-noise ratio (CNR) (solid) and carrier-to-interference ratio (CIR) (broken) as a function of modulation depth per channel, for different received currents and number of channels.

State-of-the-art systems have receiver noise and laser relative intensity noise (RIN) contributions such that the CNR is within a few dB of the quantum limit. High performance DFB lasers can be operated with modulation depths very close to the clipping limit. Linearized external modulators, although not commercialized at this time, have been demonstrated[3] with modulation depths approaching that of the DFBs. For both types of systems, further improvements in linearity, RIN, or receiver design will have little impact. The only unbounded parameter is the transmitted power, which translates directly into additional span or source sharing. Thus the continued improvement in system capability will be realized almost entirely from optimization for transmitted power. This priority is discussed in detail in each of the following sections.

3. Direct Modulation

Systems using directly-modulated lasers are, at this time, the only proven and available commercial alternative. Impairments such as nonlinearity and RIN have been under investigation for years for digital and subcarrier applications. Rapid reduction of these impairments has been possible since no real shift in emphasis in device fabrication has been required. Increased linearity and decreased RIN are both consequences of high output power, and high output power has always been a priority in the development of lasers for all types of systems. The recently available high-performance lasers are the product of this continued development for high power, combined with added motivation and a few special considerations.

3.1 Current Confinement

The key to high-power operation, hence high linearity and low RIN, is in fabrication of a laser structure that effectively confines the injected current to the active layer. The active layer is the small region (2 μm X 0.2 μm) into which electrons and holes are injected from the surrounding p- and n-doped layers of the laser diode. Stimulated recombination of these electrons and holes provides an output optical power that is an extremely linear function of the current injected into the actual active layer. However, not all of the current injected into the device goes into the active layer. Leakage current, or current that is shunted through the semiconductor material adjacent to this active layer, as shown schematically in Figure 2,[4] reduces the fraction of the total current that is injected into the active layer. The net result is a sub-linear light-versus-current (L-I) characteristic.

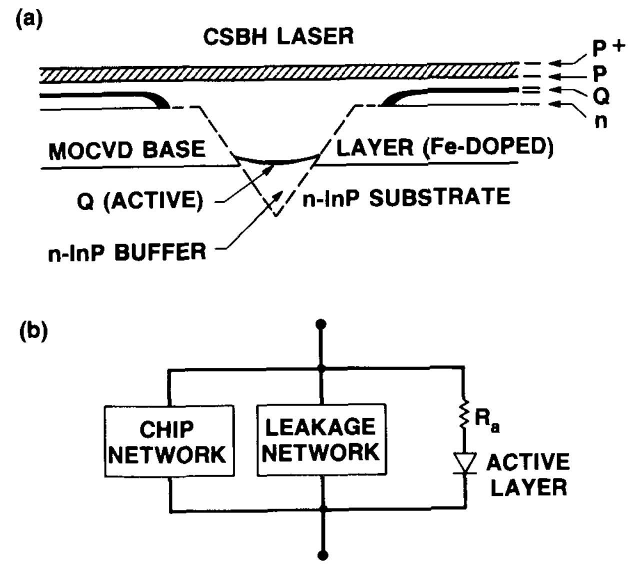

FIGURE 2

Cross section of channel-substrate buried - heterostructure laser (a) showing current through active layer and parallel current leakage path. Circuit model for laser chip (b) showing active layer diode with effective resistance R_a. The leakage network and a chip network amount for imperfection introduced by current leakage and chip parasitics, respectively (from Ref. 4).

Considerable work has been done to reduce leakage current.[5] The laser structure can be modeled, as shown on Figure 2b, by an active layer laser diode in series a small resistance R_a, and two generally problematic networks. Ideally, R_a would be zero and the leakage and chip networks would present infinite impedances. The leakage network generally contains a combination of nonlinear elements, corresponding to diodes or transistors unintentionally formed by design or fabrication imperfections. Elimination of this leakage, to the degree required for CATV linearity specifications, requires a serious commitment to laser fabrication technology.

The chip network includes the total capacitance of the laser chip, and the resistance and inductance of the bond wire through which current is applied. These parasitics are rarely a problem for the bandwidths required for CATV systems.

3.2 Resonance Distortion

In lasers that have been properly designed and fabricated to reduce current leakage, several other nonlinear mechanisms become important.[1] The most important of these is resonance distortion (RD). RD is a result of the nonlinear coupling between injected carriers and photons, in the active layer of the laser. This is the same nonlinear coupling that results in the relaxation resonance in the frequency response of the laser. For second-harmonic distortion, RD reaches a maximum when the modulation frequency is half the relaxation resonance frequency (f_r). The magnitude of sum or difference product generated by RD is maximum when the sum or difference frequency is equal to f_r. For third-order distortion, RD has maxima whenever combinations of two or three of the participating frequencies equal f_r. Fortunately, RD is zero at zero frequency and f_r is generally greater than 5 GHz, for good lasers. It is important to note that, although modulation bandwidths (resonance frequencies) need not exceed 500 MHz for signal response, f_r must be greater than approximately 6 GHz to eliminate unacceptable levels of RD.

3.3 Longitudinal Mode Control

The best system performance is obtained using single-frequency distributed feedback (DFB) lasers. These lasers use a Bragg grating along the length of the active layer to suppress laser oscillation at all but one optical frequency. Fabry-Perot (FP) lasers, which are much more easily fabricated, oscillate because of feedback from optical reflections at the cleaved ends of the active layer. Hence FP lasers oscillate in several (10) closely spaced (typically 1 nm separation) longitudinal modes of the FP resonator cavity. Power in each of these modes fluctuates randomly such that the total output power is constant. However, since fiber dispersion causes each mode to travel through the fiber with a different velocity, the power received after several kilometers of fiber fluctuates randomly. The resultant intensity noise makes FP lasers unacceptable for high-performance CATV systems utilizing AM modulation techniques. Properly fabricated DFB lasers operate in only one longitudinal mode. Mode-partition-induced intensity noise is therefore not a problem.

3.4. 1.3 Versus 1.55 μm Wavelength

The high-performance DFB lasers that are currently available meet the linearity and noise requirements as a result of considerable effort on the part of committed laser manufacturers. As of this date, this effort has been focussed on devices that operate at a wavelength of 1.31 μm, where the fiber loss is near 0.4 dB/km. Operation at 1.55 μm, where the fiber loss is typically 0.22 dB/km, has been demonstrated.[6] At this time it has not been determined whether the reduced fiber loss at 1.55 μm can make up for an intrinsic reduction in laser output power.

This reduction has two sources. The efficiency of current-light conversion is reduced by several mechanisms that consume injected carriers without creating photons.[5] One such process is Auger recombination, the effect of which increases rapidly with increasing wavelength. The second source of reduced efficiency at 1.55 μm is coupling into the fiber. Typical coupling efficiencies at 1.55 μm are 0.5 to 1.0 dB smaller than at 1.3 μm.

By far the most convincing argument for 1.55 μm laser development is for compatibility with Erbium-doped fiber amplifiers, as discussed in Section 5. These amplifiers may develop into essential components for future systems, but this is not certain. If they do, then committed 1.55 μm laser development will be required.

4. External Modulation

Recently available YAG lasers, pumped with high-power GaAs laser diode arrays, are capable of output powers in the vicinity of 200 mW. Unfortunately, dynamics within the YAG material make it impossible to modulate the output power, at frequencies required for CATV, by directly modulating the pump intensity. External modulators are essential. Unfortunately, external modulators have two serious problems; high insertion loss and poor linearity. Overcoming these problems has proven difficult, and although laboratory demonstrations have been successful, a commercially viable YAG-laser-based CATV system is not yet available.

Of the many types of modulators proposed, $LiNbO_3$ modulators are the only choice for CATV purposes. With proper attention to fabrication detail, these devices can be sufficiently stable and capable of handling YAG optical power levels. The insertion loss comes from several sources. Since the performance of the modulator depends on the state of polarization of the input light, polarization-preserving fiber is generally used to couple into the modulator. Efficient coupling from the laser into this fiber is more difficult than coupling into standard fiber. Power is also lost coupling from fiber into and out of $LiNbO_3$ waveguides and from waveguide bends that form the modulator. The total fiber-to-fiber insertion loss of a very good device is near 3 dB. Also, the device must be biased at a voltage corresponding to 50% transmission, at which point the linear intensity modulation is most linear. Combining these losses reduces the total coupling efficiency, between the laser and the output of the modulator, by at least 8 dB. Of the 23 dBm available power, approximately 15 dBm could be expected, at best, to be coupled into the output fiber. Reported values are somewhat less.[3,7]

Although this power level is higher than all but the best DFB lasers,[8] linearity must also be considered. $LiNbO_3$ transfer characteristics are intrinsically nonlinear. This nonlinearity arises because of the interferometric conversion of a relative phase difference into intensity (For Mach-Zehnder interferometer modulators). If biased at the point of inflection in the L-V curve (50% transmission) the second-order distortion is small, but the third-order distortion is approximately 30 dB worse than for directly-modulated lasers.[9] Without a linearization technique, the modulation depth per channel must be limited to less than 2%, which is less than half that used for direct modulation. The corresponding reduction in signal power generally eliminates any advantage that could have been gained by the high power and low noise of the YAG laser.

Linearization techniques change this state entirely, but are difficult to implement. Successful demonstrations have been reported[3] with modulation depths near 3.3%, for a 42 channel load. However, complications like poor coupling, noise from the linearizer,[3] or inadequate drive power for the modulator[10] always prevent externally-modulated systems from performing with the high power and low noise levels theoretically possible from the YAG lasers. Also, the reliability of the diode pump lasers has not yet been proven. These factors, combined with difficulties in manufacturing and packaging, has limited the commercial success of externally-modulated systems.

5. Optical Amplifiers

It is hoped that optical amplifiers can be used to improve the limited loss margins available from lightwave technology, making it possible to passively divide the optical signal between many fibers. Recent experiments[11,12,13,14] have demonstrated such possibilities, but many unknowns remain. Problems arise because of the high optical power levels required for high CNR. A lightwave CATV receiver must receive more than 0.5 mW of optical power. Unfortunately, the saturated output power for most types of optical amplifiers, defined by the output power for which the gain is reduced by 3 dB (optical), is typically less than 10 mW. In order to get reasonable increases in system margin, the amplifier must be operated in deep saturation. This generally results in increased noise and distortion.

Three types of optical amplifiers could be considered for CATV applications. Bulk semiconductor traveling-wave amplifiers are fabricated by removing (using anti-reflection coatings) the reflecting facets on Fabry-Perot lasers. This leaves a gain region in which the power grows exponentially as the optical signal propagates along the device. These devices have saturated output powers of several mW and can operate at any wavelength. The problem with these devices is nonlinearity. Gain is provided by stimulated recombination of electron-hole pairs. If the optical input signal is modulated, then the rate at which carriers are consumed is modulated, resulting in a modulated gain. The optical output is then a product of the modulated input and the modulated gain, which results in a nonlinear mixing. This nonlinearity is especially bad when the amplifier is close to saturation.

Much higher saturated output powers are available from multiple-quantum-well (MQW) semiconductor amplifiers. These devices are similar to the bulk amplifiers described above, except that the gain layer is replaced by several extremely thin layers (100 Å). Injected carriers are confined to discrete quantum states within each layer, such that stimulated electron-hole recombination occurs over a narrow frequency range, compared to bulk amplifiers. Saturated output powers near 50 mW have been measured.[15] Although these amplifiers suffer from the same nonlinear processes as described above, they could be operated well below saturation. The resulting distortion could be small, but this remains to be determined. These devices are strictly in the research phase and are not considered as near term possibilities. At present there is no alternative for CATV-compatible optical amplification at the 1.3 μm wavelength.

The most promise for CATV-compatible amplifiers is with Erbium-doped fiber amplifiers (EDFAs). EDFAs are fabricated using the same processes as standard fibers, but the core is doped with a high concentration of Erbium. Light from high-power pump lasers, at wavelengths of 0.98 μm or 1.48 μm, excites Erbium ions into an upper state. This upper state decays rapidly to an intermediate state, at which the ion remains awaiting stimulated or spontaneous decay. Decay stimulated by an input optical signal provides gain. The wavelength corresponding to this final transition is near 1.55 μm. Other dopants may provide optical gain at wavelengths near 1.3 μm.

Although EDFA-amplified CATV systems have been demonstrated, many unanswered questions remain. Characterization of the linearity is difficult since few linear DFB lasers exist at 1.55 μm. Noise characteristics depend strongly on the power and frequency of the pump lasers. Nevertheless, an amplified 1.55 μm 42 channel signal, from a DFB laser, has been amplified and split 16 ways,[13] with a CNR of 50 dB and composite distortions better than 60 dB. This demonstration suggests that these amplified systems will eventually be available.

The experiment used to demonstrate the 16 way split is shown in Figure 3. The EDFA consists of 45 meters of EDF, 2 high-power 1.48 μm semiconductor pump lasers, 2 wavelength multiplexing couplers, and several optical isolators. Each pump laser supplies 25 mW of optical power to the amplifier. The reliability of pump lasers operated at these power levels must be investigated thoroughly. This pump power results in a saturated output power of 6 dBm. The couplers allow the 1.48 μm pump light to be combined with the 1.53 μm source laser with little loss. Isolators are required because the EDFA, like all optical amplifiers, provides gain in both directions. Without the isolators, reflections from fiber connections can produce frequency dependence in the gain, or even laser oscillation.

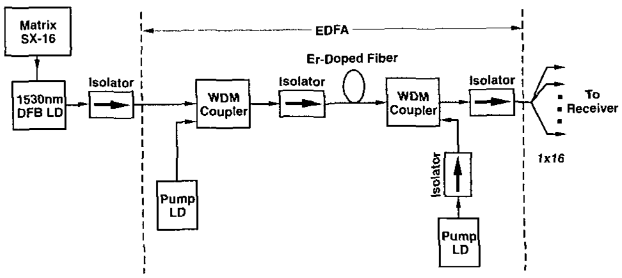

FIGURE 3

Schematic for 16-way split using Erbium-doped fiber amplifier (from Ref. 13).

By operating the amplifier well into saturation, a total output power of 12.2 dBm was achieved. The unsaturated gain of more than 30 dB was then reduced to less than 20 dB. Although this apparently caused little nonlinear distortion, the noise figure for the EDFA was 11 dB, which is much worse than generally obtained for unsaturated operating conditions. Getting an adequate CNR with such high noise figures requires an excessively high modulation index, and input signal powers greater than 1 mW,[16] hence placing further burden on the design of good 1.55 μm DFB lasers.

To summarize the status of EDFAs for CATV applications, it is apparent that the technology can be made to work, but the noise performance is questionable. For future systems, in which a high CNR must be delivered, noise from just a single EDFA may be unacceptable. The reliability of the pump lasers, operated at high power levels, and the optimum pump wavelength need to be clarified. Linearity does not appear to be a problem, but the nonlinear mechanisms have not been investigated thoroughly. The feasibility of such systems also hinges on the availability of linear low-noise 1.55 μm source lasers.

EDFAs may be an integral part of the future CATV network, but they do not obviate enthusiasm for currently available 1.3 μm technology. The cost of each amplifier, given the component count shown in Figure 3, is likely to be attractive only if a high degree of splitting (> 8) is possible. The additional noise introduced by the amplifier may hinder the high CNR (55 db) desired in upgraded trunk systems. Implementing a network based on 8 or 16-way fiber splitting requires design flexibility that does not exist in established systems. Upgrading existing systems may require the flexibility and incremental growth offered by the 1.3 μm systems, which offer limited splitting opportunities (2 or 4 way). Improvements will also continue with these 1.3 μm systems, including the possibility of 1.3 μm fiber amplifiers. One must compare the cost of the amplified system with that of systems based on future high-power DFB lasers. Introduction of whatever future technologies become available will be simplified if an established base of existing lightwave technology is in place.

6. Fiber/System Impairments

Optical fibers are as close to being a perfect transmission media as could be expected. The low loss, high bandwidth, immunity to electromagnetic interference, material stability and high strength have driven the explosive growth of digital fiber systems. For CATV applications, the only problems arise from reflections and Rayleigh backscatter. Dispersion is a problem if multi-longitudinal-mode sources are used, but not for DFB or YAG lasers operated at wavelengths close to the dispersion minimum. Dispersion can also introduce nonlinear distortion, for lasers operated far from the dispersion minimum.[14,17] Optical power levels are not near the magnitude at which fiber nonlinearity becomes a problem.

Fiber reflections, if coupled back into the laser, increase the intensity noise and nonlinear distortion generated by the laser.

Any high-performance system available today uses an optical isolator to eliminate this problem. At least 30 dB isolation is needed to remove feedback problems in systems that have been carefully installed so as to minimize reflections.

Another problem with backscatter and reflections is that the direct signal mixes with delayed reflected signals on the photodetector. The photocurrent generated by the detector is proportional to the square of the total optical field. This generates a beat product between the direct and reflected signal that can generally be described by an effective intensity noise. The amount of noise generated in the CATV band depends on the optical spectrum of the modulated source. Measured source spectra[1] suggest that two fiber reflections of -30 dB (optical power) can limit the CNR to 57 dB. Since one can anticipate reflections from many connectors within the system, reflections from each must be considerably less than -30 dB. Other reports[15] indicate that the intrinsic Rayleigh backscatter of the fiber may produce effective relative intensity noise levels of -150 to -160 dB/Hz. This noise, which may be higher than that generated by the laser, is intrinsic in that the light attenuated by the fiber is scattered by microscopic refractive index variations in the fiber.

7. Conclusions

AM-VSB lightwave systems have been developed that meet the cost and performance required for widespread application within the CATV industry. This capability is the result of rapid development, over the past two years, of high performance 1.3 μm DFB lasers. These systems offer spans greater that 10 km and limited passive division. Further refinements of this technology will yield incremental improvements only. Hence, obsolescence of today's "state-of-the-art" equipment should not be feared.

Several alternative technologies may emerge within the next few years. Of these, the most significant potential for improvement is offered by fiber optical amplifiers, which may provide sufficient transmission margin for signal division between tens of fibers. Since this splitting capability may not conform to many present trunk topologies, presently available DFB systems should remain in high demand. Widespread implementation of available fiber technology will provide not only the needed improvements in quality and reliability, but a platform on which a future can be built.

REFERENCES

1. T.E. Darcie, G.E. Bodeep, "Lightwave Subcarrier CATV Transmission Systems," *Joint issue - IEEE Trans. on Microwave Theory and Techniques and IEEE J. Lightwave Technology,* September 1990.

2. A.A.M. Saleh, "Fundamental Limit on Number of Channels in Subcarrier Multiplexed Lightwave CATV System," *Electronic Letters,* Vol. 25, No. 12, June 1989, pp. 776-777.

3. R.B. Childs, "Predistortion Linearization of Directly-Modulated DFB Lasers and External Modulators for AM Video Transmission," Optical Fiber Communications Conference, San Francisco, CA, January 22-26, 1990, Technical Digest, paper WH 6.

4. T.E. Darcie, "Subcarrier Multiplexing for Lightwave Networks and Video Distribution Systems," *IEEE J. Select. Areas Communications,* September 1990.

5. G.P. Agrawal, N.K. Dutta, Long Wavelength Semiconductor Lasers, Van Nostrand Reinhold Co., N.Y., 1986.

6. T.E. Darcie, G.E. Bodeep, "Lightwave Multi-channel Analog AM Video Distribution Systems," International Conference Communications, Boston, MA, June 11-14, 1989, pp. 1004-1007.

7. Orchard Release, Multichannel News, Febuary 12, 1990.

8. J. Lipson, L.C. Upadhyayula, S.Y. Huang, C.B. Roxlo, E.J. Flynn, P.M. Nitzche, C.J. McGrath, "High Fidelity Lightwave Transmission of Multiple AM-VSB NTSC Signals," submitted to *IEEE J. Lightwave Technology.*

9. G.E. Bodeep, T.E. Darcie, "Semiconductor Lasers versus External Modulators: A Comparison of Nonlinear Distortion for Lightwave Subcarrier CATV Applications," *Photonics Technology Letters,* Vol. 1, No. 11, November 1989, pp. 401-403.

10. R.M. DeRidder, S.K. Korotky, "Feedforward Compensation of Integrated Optic Modulator Distortion," Optical Fiber Communications Conference, San Francisco, CA, January 22-26, 1990, Technical Digest paper WH 6.

11. K. Kikushina, E. Yoneda, K. Aoyana, "6-Stage Cascade Erbium Fiber Amplifiers for Analog AM and FM-FDM Video Distribution Systems," Optical Fiber Communications Conference, San Francisco, CA, January 22-26, 1990, post deadline paper PD22.

12. W.I. Way, M.M. Choy, A. Yi-Yan, M. Andrejco, M. Saifi, C. Lin, "Multichannel AM-VSB Television Signal Transmission Using an Erbium-Doped Optical Fiber Power Amplifier," *IEEE Photonics Technology Letters,* Vol. 1, No. 10, October 1989, pp. 343-345.

13. S.Y. Huang, T.W. Cline, L.C. Upadhyayula, R.E. Tench, J. Lipson, "Point-to-Multipoint Distribution of 42 Channel VSB-AM Video Signals Using an Erbium-Doped Fiber Amplifier," Submitted to LEOS Summer Topical Meetings, Monterey, CA, July-August 1990.

14. M. Shigematsu, K. Nakazato, T. Okita, Y. Tagami, and K. Nawata, "Field Test of Multichannel AM-VSB Transmission Using an Erbium Doped Optical Fiber Amplifier at 1.55 μm wavelength Range in the CATV Network," IEEE/Lasers and Electro-Optics Society/Optical Society of America, Optical Amplifiers and Their Applications, Monterey CA, August 6-8, 1990, 1990 Technical Digest, Vol. 13, pp 240-243.

15. G. Eisenstein, U. Koren, G. Raybon, T. L. Koch, J. M. Wiesenfeld, R. S. Tucker, B. I. Miller, "A 1.5 μm Quantum Well Amplifier with High Saturation Output Power," Seventh

Intl. Conference on Integrated Optics and Optical Fiber Communication (IOOC '89), Kobe, Japan, 1989, post-deadline digest, pp. 62-63.

16. I.M.I. Habbab, L.J. Cimini, Jr., "Optimized Performance of Erbium-doped Fiber Amplifiers in Subcarrier Multiplexed Lightwave CATV Systems," Optical Fiber Communications Conference, San Diego, CA, Feburary 18-22, 1991.

17. M.R. Phillips, T.E. Darcie, D. Marcuse, N.J. Frigo, G.E. Bodeep, "Nonlinear Distortion from Fiber Dispersion of Chirped Intensity Modulated Signals," Optical Fiber Communications Conference, San Diego, CA Feburary 18-22, 1991.

18. A. F. Judy, "Intensity Noise from Fiber Rayleigh Backscatter and Mechanical Splices," Technical Digest - European Conference on Optical Communications (ECOC '89), Gotehnburg, Sweden, September 10, 1989, paper TUP-11, pp. 486.

MEASUREMENT DIFFERENCES WITH VARIOUS CHROMINANCE TO LUMINANCE GAIN AND PHASE TECHNIQUES

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ABSTRACT

When evaluating chrominance to luminance gain and delay inequalities, there are various techniques that can be used. Chrominance to luminance gain can be evaluated using a multiburst signal, (Sin x)/x signal, or 12.5T modulated Sin^2 pulse signal. However, each of these techniques will yield different results for similar parameters tested. This paper will present an empirical and analytical study on the differences between the various C/L measurement techniques.

INTRODUCTION

Chrominance to luminance (C/L) gain and delay inequalities are classified as linear distortions. Linear distortions are those which are not affected by signal amplitude. Basically, C/L gain inequality is defined as the difference in gain of the chrominance components compared to the gain of the luminance components as they pass though a television system. Similarly, C/L delay inequality is defined as the difference in time between the chrominance and luminance signal components as they propagate through a television system.

When viewed on a television receiver, delay distortion will cause color smearing or bleeding at the transitions in a picture. Figure 1 depicts C/L delay distortion in which the chrominance components are delayed with respect to the luminance components. Visually, C/L gain distortion can be seen as incorrect color saturation. Chrominance to luminance gain error's are caused by color peaking or attenuation in the video signal.

In television receivers C/L delay is on the order of approximately 170 nanoseconds. This delay is a result of the modulated video signal passing through the television's SAW filter and other supporting circuits. By design (compliance with FCC standards), video modulators used in cable television systems compensate for inherent C/L delays in television receivers by advancing the chrominance components 170 nanoseconds with respect to the luminance components.

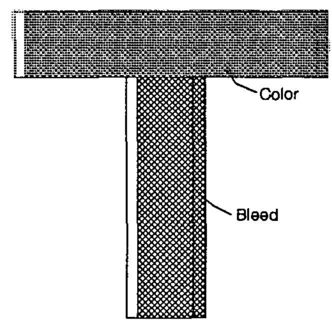

Fig 1. Visual Effect of C/L Time Delay

The standard cable transmission media used for transmitting video signals from the video modulator to television receiver usually does not contribute to C/L distortion. However, when a settop converter is in line with this media, C/L inequalities could exist. As a result of the video processing being performed in baseband converters, chrominance to luminance inequalities may be higher in baseband converters than in RF converters. Settop converters need to be designed for minimum C/L inequalities.

Verification of chrominance to luminance gain and delay distortion can be evaluated using a modulated 12.5T Sin^2 pulse, (Sin x)/x signal, or multiburst signal (gain only). However, for the same measurement the three techniques will produce three different results.

The reason for these differences can best be understood by reviewing the measurement methods. Afterwards a theory for these differences will be formulated.

MEASUREMENT METHODS

Modulate 12.5T Sine-Squared Pulse

The modulated 12.5T $\sin^2$ pulse is probably the more common method used for evaluating C/L inequalities. The main advantage for using the modulated 12.5T $\sin^2$ pulse is that it allows for easy evaluation in the time domain of both C/L gain and delay inequalities with a single signal. The magnitude and direction of either C/L gain or delay distortion can be determined by evaluating the baseline distortion of the 12.5T pulse. The baseline of the pulse will be flat if no distortion exists. However, as gain and delay distortion is introduced, the baseline will be affected in different ways. A single peak either above or below the baseline indicates the presence of C/L gain error only. Symmetrical peaks above and below the baseline indicates the presence of C/L delay error only. A combination of both C/L gain and delay error will result in non-symmetrical peaks above and below the 12.5T pulse baseline. Figure 2 depicts the effects of several types of C/L distortions.

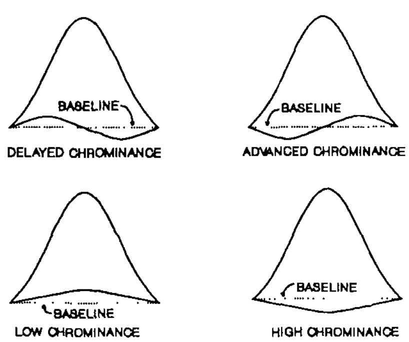

Fig. 2 12.5T Pulse C/L Gain And Phase Distortion Variations

Mathematically, the 12.5T modulated $\sin^2$ pulse can be expressed in the time domain as follows,

$$g(t) = \sin^2(w_m)t \cdot \sin(w_c)t + G \cdot \sin^2(w_m + \phi)t \tag{1}$$

Where: $g(t) = 0$ when $g(t) > 3.125$ microseconds, $wc = 2 \cdot \pi \cdot 3.58 \cdot 10^6$ rad/sec (chrominance carrier), $wm = 2 \cdot \pi \cdot 160 \cdot 10^3$ rad/sec (luminance modulation), G = C/L gain error, and ϕ = C/L delay $\cdot$ wm.

Expanded, equation 1 becomes,

$$\begin{aligned} g(t) = {} & .5 \cdot \sin(w_c)t - .25 \cdot G \cdot \sin[(w_c + 2 \cdot w_m)t + 2 \cdot \phi] - \\ & .25 \cdot G \cdot \sin[(w_c - 2 \cdot w_m)t - 2 \cdot \phi] + \\ & .5 - .5 \cdot \cos(2 \cdot w_m)t \end{aligned} \tag{2}$$

The wonderful aspect regarding this signal is that a low frequency reference pulse is transmitted with the modulated chrominance signal. This low frequency reference pulse is expressed in equation 2 as $.5 + .5 \cdot \cos(2 \cdot wm)t$. Inspection of equation 1, clearly shows how the composite signal's baseline is affected as C/L gain or delay vary. For example, figure 3 shows equation 1 plotted for a C/L gain and delay error equal to 0dB and 0 seconds respectively. Figure 4 shows equation 1 plotted for a C/L gain error equal to -2dB and a C/L delay error equal to 150ns.

The spectral distribution of a modulated 12.5T $\sin^2$ pulse can be determined by taking the Fourier integral of equation 1. The Fourier integral is given by,

$$g(t) = \int_0^{\frac{1}{2f_m}} [\sin(\omega_m t)^2 \sin(\omega_c t) + \sin(\omega_m t)^2]\, e^{-j\omega t} dt \tag{3}$$

Where the upper limit of integration, $1/(2 \cdot fm)$, is used to limit the time function's signal to only one $\sin^2$ pulse.

Figure 5 shows the spectral distribution of this Fourier transformation. Notice that the total luminance signal spans a weighted bandwidth from 0 to 640 KHz. Similarly, the total signal distribution around the chrominance signal spans a weighted bandwidth of ± 640 KHz, either side of 3.58 MHz. Mathematically, C/L gain can be determined the following equation,

$$\begin{aligned} C/L\ gain = {} & \int_0^{.64\ Mhz} |A(f)|\, \left| \int_0^{\frac{1}{2}f_m} f(t)\, e^{-j\omega t}\, dt \right| df \\ & - \int_{2.94\ Mhz}^{4.22\ Mhz} |A(f)|\, \left| \int_0^{\frac{1}{2f_m}} f(t)\, e^{-j\omega t}\, dt \right| df \end{aligned}$$

$$Where:\ f(t) = \sin(\omega_m t)^2 \sin(\omega_c t) + \sin(\omega_m t)^2 \tag{4}$$

Where $|A(f)|$ is the magnitude of the frequency response of the device under evaluation.

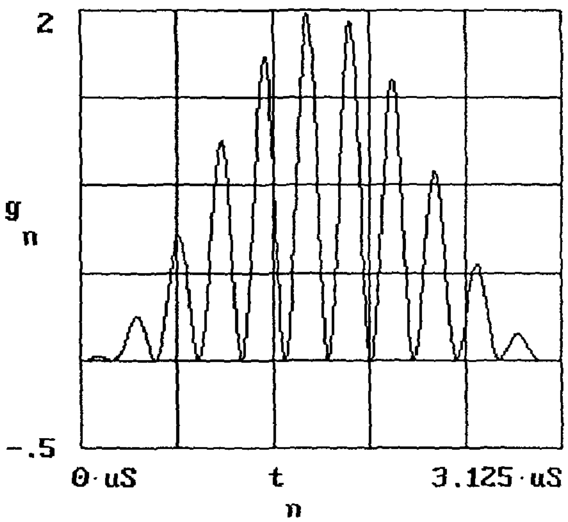

Fig. 3 Simulated 12.5T pulse With No C/L Gain Error

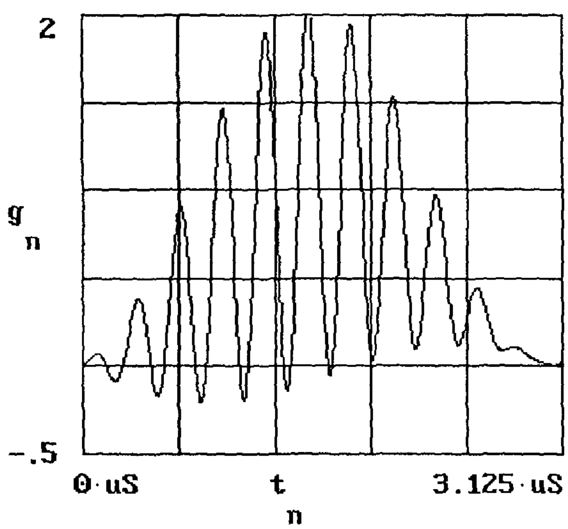

Fig. 4 Simulated 12.5T Pulse With C/L Gain And Delay Error

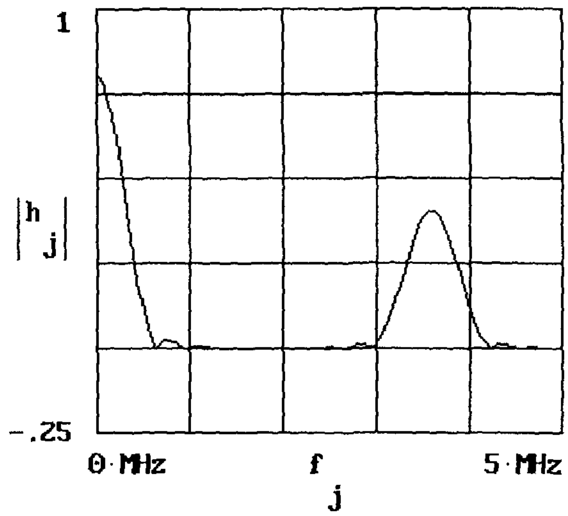

Fig. 5 Spectral Distribution Of A Modulated 12.5T Sin^2 Pulse

(Sin x)/x Pulse

The (Sin x)/x pulse has an extraordinary property which allows one to analyze the complete frequency response of a communication system with a single pulse. The (Sin x)/x pulse produces a flat and continuous frequency spectrum from 0 Hz to approximately the reciprocal of the pulse width. The Fourier transformation of this signal can be shown by the following equation,

$$h(t) = \int_{-\infty}^{\infty} \frac{\sin(\omega_m t)}{\omega_m t} e^{-j\omega t}\, dt \tag{5}$$

The (Sin x)/x pulse is shown in both the time and frequency domain in figures 6 and 7 respectively. If a 210 nanosecond (Sin x)/x pulse is transmitted at intervals of the horizontal scanning frequency (f_H), the frequency spectrum will contain equal amplitude harmonics of f_H up to 4.75 MHz. The described (Sin x)/x test signal which is available on several video generators is shown in figure 8.

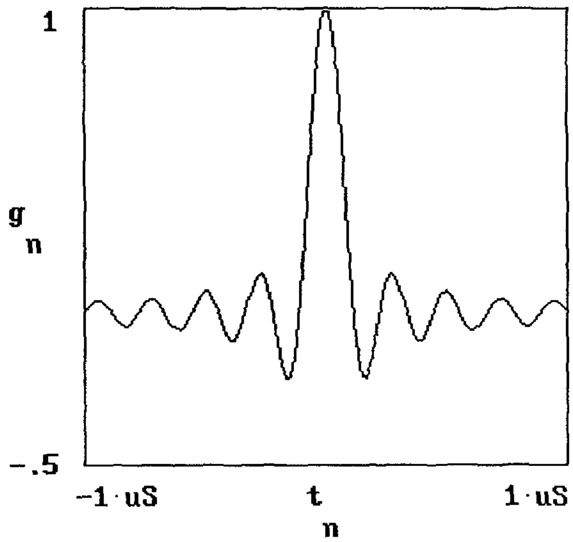

Fig. 6 Time Domain Response Of A (Sin x)/x Pulse

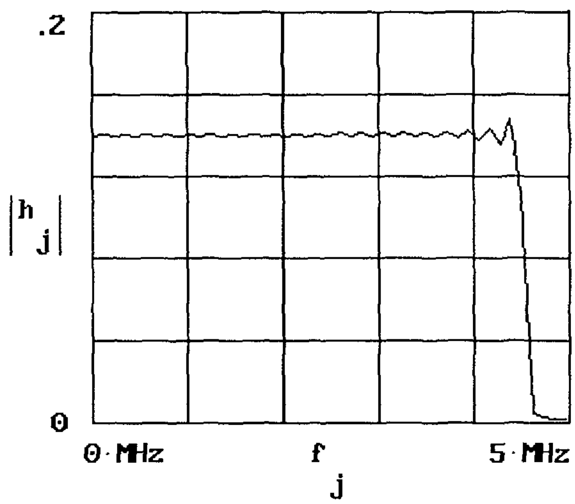

Fig. 7 Frequency Domain Response Of a (Sin x)/x Pulse

Analysis of response using the (Sin x)/x signal is done with a spectrum analyzer or special video analyzer. When using a spectrum analyzer, the only information obtainable is amplitude vs. frequency. However, there are instruments available that will display both amplitude and group delay as a function of frequency from approximately 220 KHz to 4.5 MHz. Using these instruments and choosing the proper frequency points, C/L gain and delay distortion can be determined.

Mathematically, C/L gain when using the (Sin x)/x pulse is determined as follows,

$$C/L\ gain = \int_{.22\,Mhz - BW}^{.22\,Mhz + BW} |A(f)|\ \left|\int_{-\infty}^{\infty} \left[\frac{\sin(\omega_m t)}{\omega_m t}\right] e^{-j\omega t}\,dt\right|\,df - \int_{3.58\,Mhz - BW}^{3.58\,Mhz + BW} |A(f)|\ \left|\int_{-\infty}^{\infty} \left[\frac{\sin(\omega_m t)}{\omega_m t}\right] e^{-j\omega t}\,dt\right|\,df \tag{6}$$

Where |A(f)| is the magnitude of the frequency response of the device under evaluation and BW is equal to 1/2 the bandwidth of the measurement system. Similarly, C/L delay can be determined as follows,

$$C/L\ delay = \frac{d}{df_{220\,KHz}} ARG(A(f))\ ARG\left[\int_{-\infty}^{\infty} \left[\frac{\sin(\omega_m t)}{\omega_m t}\right] e^{-j\omega t}\,dt\right] - \frac{d}{df_{3.58\,MHz}} ARG(A(f))\ ARG\left[\int_{-\infty}^{\infty} \left[\frac{\sin(\omega_m t)}{\omega_m t}\right] e^{-j\omega t}\,dt\right] \tag{7}$$

Where Arg( ) is a function for the phase response, which is equal to the arc tangent of the imaginary part of the Fourier integral of f(t) divided by the real part of the Fourier integral of f(t). Similarly, Arg(A(f)) is the phase response as a function of frequency for the device under test.

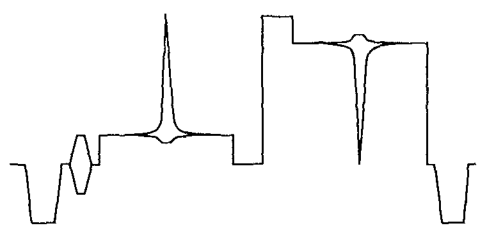

Fig. 8 Typical (Sin x)/x Signal

Multiburst Signal

The multiburst signal is unlike the 12.5T pulse and (Sin x)/x pulse which took advantage of their Fourier components and/or modulation theory to generate the frequency components at which gain and delay are mathematically analyzed. The multiburst signal consist of 6 frequency packets ranging from .5 MHz to 4.1 MHz. A typical multiburst signal is shown in figure 9. Chrominance to luminance gain measurements are made by comparing the response of the 3.58 MHz packet to the response of the 500 KHz packet. Alternatively, all packet amplitudes may be measured with respect to the bar amplitude, yielding a lower frequency reference. Automated measuring equipment using the multiburst is inconsistent in the reference used, even among different software revisions of the same equipment.

Although, not as intriguing as the 12.5T $\sin^2$ pulse or (Sin x)/x signal, C/L gain for the multiburst signal can be shown as,

$$C/L\ gain = \int_{.5\,Mhz - BW}^{.5\,Mhz + BW} |A(f)|\ \left|\int_0^{\frac{4}{f_m}} \sin(\omega_m t)\, e^{-j\omega t}\, dt\right|\, df - \int_{3.58\,Mhz - BW}^{3.58\,Mhz + BW} |A(f)|\ \left|\int_0^{\frac{16}{f_m}} \sin(\omega_c t)\, e^{-j\omega t}\, dt\right|\, df \tag{8}$$

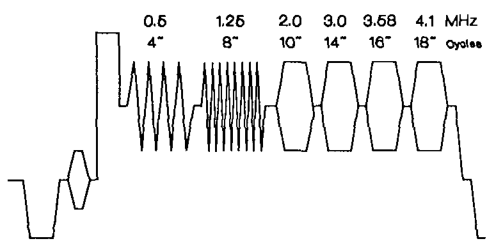

Fig. 9 Typical Multiburst Signal

MEASUREMENT DIFFERENCE THEORY

Figure 10 shows the spectral distribution of 3 C/L gain measurement methods superimposed on one plot. It should be noted that the amplitude of the individual responses are not absolute, but only relative values. In addition only the 500 KHz and 3.58 MHz components of the multiburst are shown.

The 12.5T pulse measurement is a comparison of the weighted average of the total response, A(f) · Fourier integral of the 12.5T pulse, around the color subcarrier, with the weighted average of the total 12.5T pulse response in the first 640 KHz of the luminance spectrum.

The (Sin x)/x waveform has a spectrum that is flat over the entire TV spectrum. The measurement made with it is more of a spot measure of the total response, A(f) · Fourier integral of (Sin x)/x pulse, at the color subcarrier, compared with the response at about 220 KHz.

The multiburst measurement is made by comparing the response of the color subcarrier with the response at 500 KHz, or with the low frequency bar. Other frequencies are also produced, but are not shown in figure 10.

If the gain, $|A(f)|$, of the device under test is flat from 0 Hz to 640 KHz and flat ± 640 KHz around the color subcarrier, all three measurement methods would produce the same results for C/L gain. The same reasoning is true for C/L delay if the delay is linear over the same frequency bands mentioned above. Now consider what happens when the overall baseband response of a typical baseband converter is taken into account. It is not unusual for a baseband converter to have the response shown in figure 11. Particularly, notice the response ± 640 KHz on either side 3.58 Mhz, expanded in figure 12. Since, the 12.5T pulse uses this entire spectrum for C/L gain determination, the results will obviously be different than (Sin x)/x's point measurement at 3.58 Mhz. The same will be true when comparing the results of the 12.5T pulse against the multiburst method. Using the example of the STT response shown in figure 11, the 12.5T pulse will measure less chrominance than either the (Sin x)/x or multiburst approach. The actual difference between the 12.5T method and (Sin x)/x method around 3.58 MHz can be shown as,

$$C/L\ gain = \int_{3.58\,Mhz - BW}^{3.58\,Mhz + BW} |A(f)|\ \left|\int_{-\infty}^{\infty} \left[\frac{\sin(\omega_m t)}{\omega_m t}\right] e^{-j\omega t}\, dt\right|\, df - \int_{2.94\,Mhz}^{4.22\,Mhz} |A(f)|\ \left|\int_0^{\frac{1}{2f_m}} f(t)\, e^{-j\omega t}\, dt\right|\, df$$

$$Where:\ f(t) = \sin(\omega_m t)^2 \sin(\omega_c t) + \sin(\omega_m t)^2 \tag{9}$$

Equation 10 does not take into account the device's response within the first 600 KHz. It is important not to ignore this spectrum. Even though, not as great a contributor as the response around 3.58 MHZ, both the test equipment demodulator and device demodulator showed a fair amount of roll off from DC to 640 KHz. Therefore, the actual difference between any C/L measurement method must use the complete C/L gain or delay equations discused in the previous sections.

After substituting the measured STT response for |A(f)| the predicted difference equals -1.2dB. This delta is very close to the actual measured delta shown in table 1.

It is very important to note, that depending on the response of the converter or demodulator being evaluated, the difference in magnitude and polarity of the C/L inequalities measured will vary. Table 2 shows actual measured results of an off air demodulator and two baseband converters. In this evaluation, the 12.5T method measured higher chrominance gain for the demodulator than for the baseband converters.

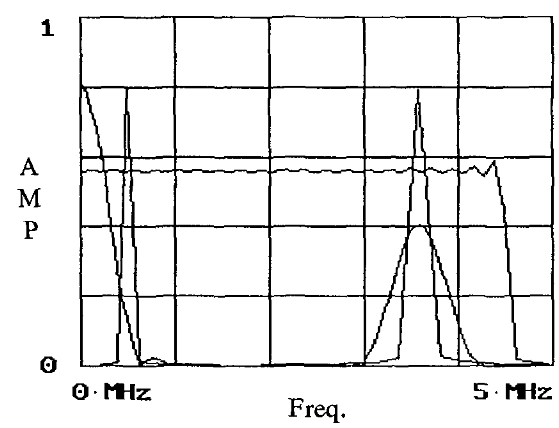

Fig. 10 Measurement Comparison

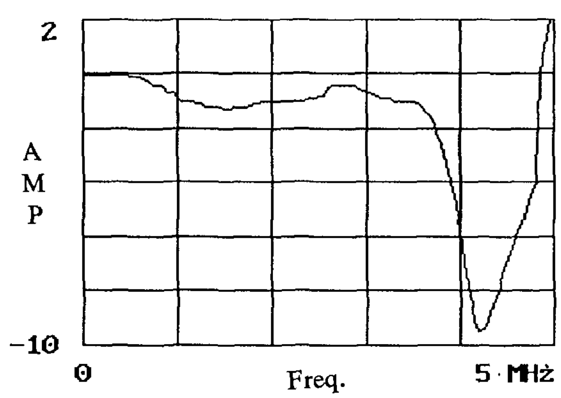

Fig. 11 Example Demodulator Response

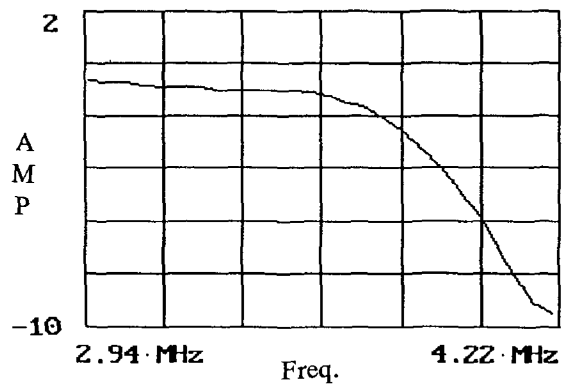

Fig. 12 Expanded Example Demodulator Response

Measurement Method	Gain (dB)	Delay (nS)
(Sin x)/x	-0.3	73
12.5T	-1.6	34
Multiburst	-1.0	
Δ(Sin x)/x & 12.5T	-1.3	-39

Table 1 Video Demodulator And Baseband Converter

CONCLUSION

As was shown mathematically and empirically, the C/L results obtained between the different C/L measurement techniques will vary depending on the response of the system being measured. This investigation is not suggesting that one method is better than another, only that they are different. However, when analyzing converter C/L inequalities, it is important that consistency in the measurement technique be used.

REFERENCES

1. Craig, Margaret. "Television Measurements NTSC Systems", Tektronics Inc, 1989

2. "VM700 Video Measurement Set Operator's Manual", Tektronics Inc, 1988

3. "NCTA Recommended Practices For Measurements On Cable Television Systems", NCTA, 1989

4. Van Valkenburg, M.E. Network Analysis. N.J.: Prentice-Hall, Inc., 1974

Measurement Method	Demod		Converter #1		Converter #2	
	Gain (dB)	Delay (nS)	Gain (dB)	Delay (nS)	Gain (dB)	Delay (nS)
(Sin x)/x	-.5	1	-.6	44	-1.0	79
12.5T	+.19	34	-1.45	34.7	-1.17	62.5
Multiburst	-.12		-.87		-.87	
(Sin x)/x - 12.5T	.69	33	.85	9.3	.17	16.5

Table 2 Comparison Between C/L Measurement Methods

OUTAGES: THE ISSUE OF THE 90'S

Bradley L Johnston
Warner Cable Communications, Inc.

I. ABSTRACT

Outages are probably the cable industry's most important operating issue. This paper will examine the following:

* *Importance of outages and the impact on business.*
* *Customer acceptable performance.*
* *Outage definition and detection.*
* *Opportunities for significant reductions.*

In 1990 CableLabs established an outage reduction task force. Much of the information presented in this paper comes from the work of this task force and its very capable members.

It is the author's point of view that significant reductions, 50% or greater, can be made in outages in the next 12 months with minor modifications to the equipment and plant configurations now in use. Outage reductions of some 50% are required to gain customer loyalty and a rating of excellence. Additionally, it is reasonable to anticipate even greater reductions in the 1992-93 time frame by isolating trunk powering from commercial power and ensuring that all equipment is highly reliable.

II. CUSTOMER SATISFACTION: IMPORTANCE OF OUTAGES

Customer satisfaction, or customer service, is our biggest problem and opportunity. Franchise renewal difficulties are almost exclusively preceded by customer satisfaction problems. The problem in Washington D.C. could not exist if customers rated cable TV service as excellent. In addition, customer service problems create high phone traffic, high service call rates, and can result in employee turn-over which all result in higher expense levels.

Recently Viacom completed a study in which individual customers were tracked over a number of months. Data were collected and analyzed to understand how controllable disconnect and pay downgrades compared to the customer's rating of service. The results, shown in Figure 1, show customers with *Fair-Poor* service ratings disconnect and downgrade almost twice as often as customers with *Good-Excellent* service ratings; 8.4% to 4.9% for downgrades and 6.9% to 3.5% for controllable disconnects. Excellent customer satisfaction increases revenues, reduces expenses, and helps keep local and national politics on a positive and manageable level.

Customer satisfaction can be characterized with a number of attributes such as:

* Phone service; degree of access to problem resolution and information.

* Installation; availability of convenient schedules, quality of work and courteousness of personnel.

* Sales; courteousness and competency of personnel.

* Repair; quickness of resolution, availability of convenient schedules if home visit required, and competency and courteousness of personnel.

* Outages; frequency and duration.

* Pictures; lack of noise and other distortions, and sound levels constant across all channels.

* Office; convenient location and hours.

* Billing; accuracy, understandability, easy problem resolution.

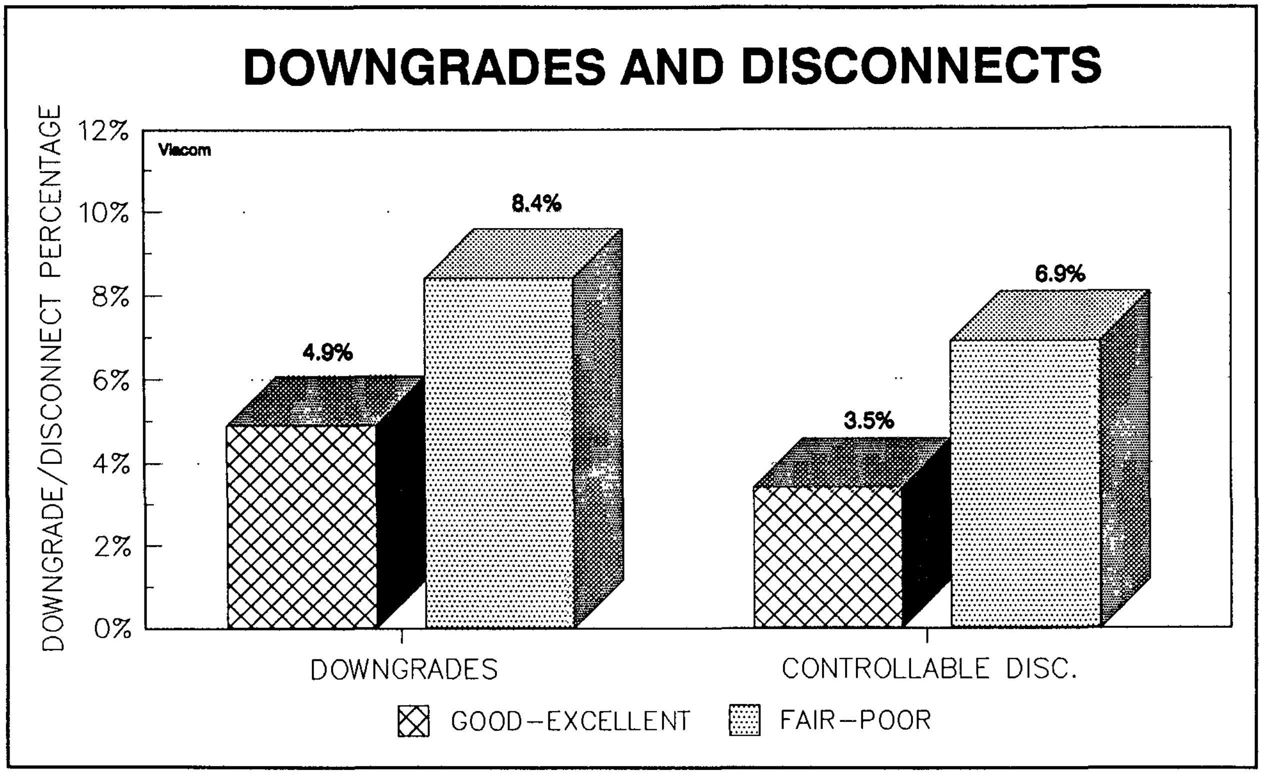

Figure 1

An effective technique for problem solving is to determine the key cause and bring focus to it. Studies have identified Outages as the most important customer satisfaction issue. Figure 2 shows data that correlates customer perception of service to the actual number of outages experienced. Clearly a strong correlation exists between satisfaction and outages.

One Warner Cable system with historically high Customer Satisfaction ratings had a drop of 7% in late 1990. Outages in this system over this same period increased by 52%. Another System that historically had low customer satisfaction scores had a 15% jump late in 1990, while outages decreased by 44%. In both cases the analysis concluded that outages are the principal cause for the change in customer satisfaction. Figure 3 shows Warner Cable's 1990 Customer Satisfaction data versus outages.

Outages directly impact the customer. Additionally, outages create activities that degrade performance in phone service, repairs, and pictures. Outages are clearly the priority issue.

Warner Cable, like many operators, has made significant service improvements in all areas except outages. We have been disappointed to find that ratings for Overall Customer Satisfaction have only marginally improved. Eliminating outages is the key to achieving ratings of Excellence.

III. OUTAGES: DEFINITION, DETECTION, ACCEPTABLE PERFORMANCE

A. Outage Definition

There are many definitions of outages varying from "all channels out to 20 or more customers, not counting loss of power or maintenance outages", to "one or more channels out to more than one customer for any reason." However, the definition must come from a customer's point of view.

If a customer watches a program and it goes out, the customer calls that an outage. The customer does not care if the Power Company is at fault, if the plant was taken down for maintenance, or if other channels remain on. The bottom line is the customer paid for certain channels, elected to watch a specific channel but now cannot watch the channel; that is an OUTAGE. I suggest that an outage is:

> **"Loss of one or more channels to four or more customers."**

C. Outage Detection

I recently had the opportunity to ask technical operations experts from a number of MSOs how they detect outages. Outage detection today is an imprecise process. Systems today most often use the över-load created on the CSR phone switch as a way of detecting outages.

An outage detection system is needed that will:

* Automatically, quickly, and accurately, create a flag when there is an outage.

* Allow 24-hour-a-day outage detection and automatically call standby personnel.

* Provide data to track performance; frequency, repeats and duration.

* Provide data conducive to analysis leading to action to prevent outages.

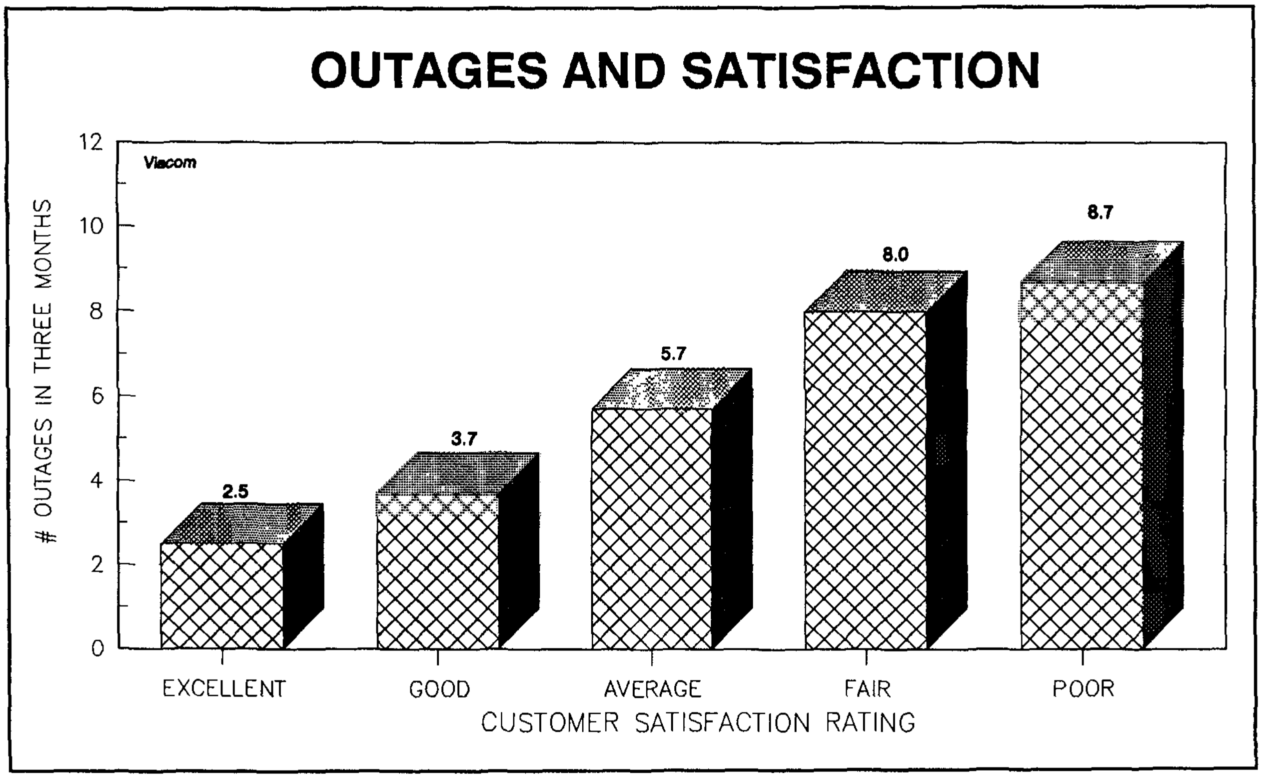

Figure 2

B. Customer Acceptable Outage Frequency

Figure 2 shows results of a study performed by Viacom which links frequency of outages to customer satisfaction levels. Note outage frequency shown in Figure 2 is over a 3-month period. On a monthly basis the data suggest that no more than about 1 outage per month can occur for the service to be rated as Excellent.

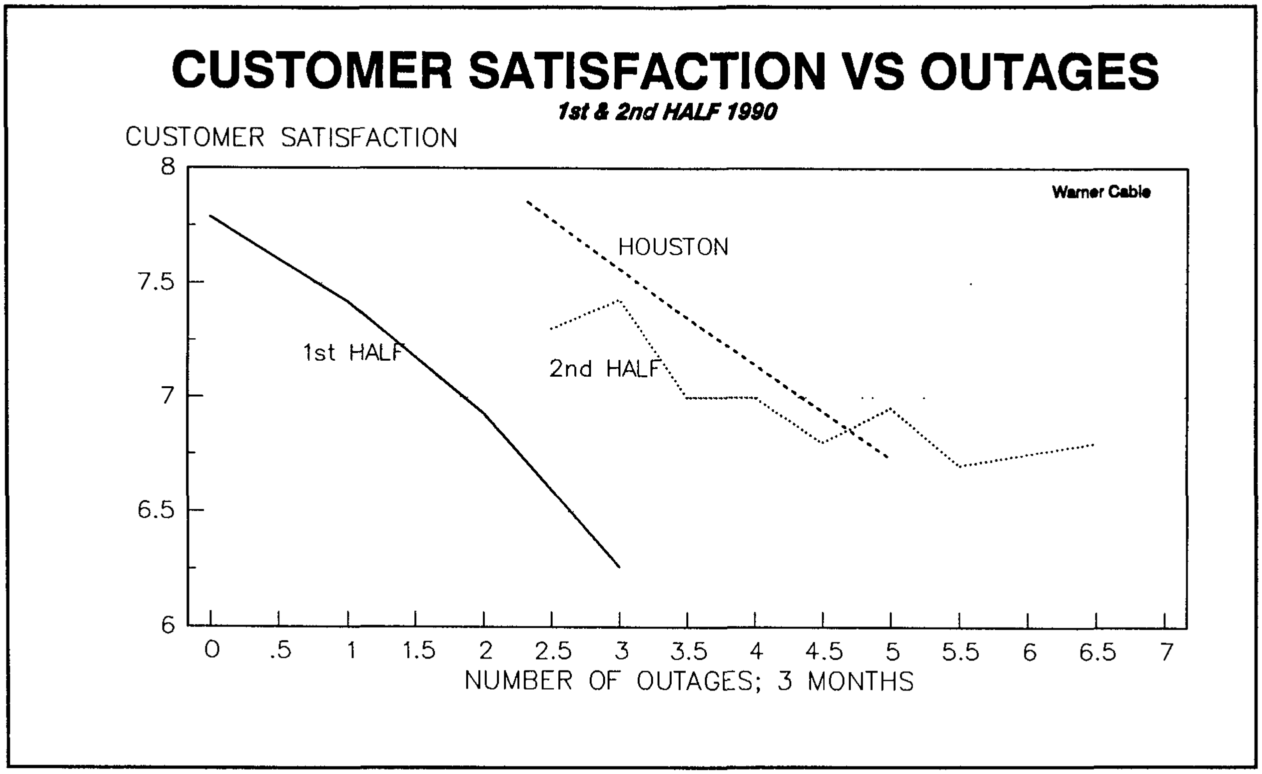

Figure 3

* Detection must be added and made operational for a nominal cost.

* Customers must be used for detection since customers are the only source for outage detection in systems with one-way plant.

Some billing systems have outage detection modules. The billing system with its customer data base and on-line tie to the CSR operation must be an integral part of outage detection. The problem most operators have with present detection systems is they require the creation of a data base that ties customers to specific amplifier locations. Creating this data base can be a significant and expensive effort.

Let me share with you an approach the Hampton, Virginia system recently implemented that avoids the large data base efforts. The system uses ZIP-plus-4 data to organize customers into geographic groups of 60 to 200 homes. Since customer addresses, phone numbers and ZIP-plus-4 data is readily available the data base is easy to establish. The billing system declares an outage whenever two or more customers call with an outage from a ZIP-plus-4 group within 120 minutes. This approach has proven to be extremely accurate to declare outages.

I am pleased to report that outage detection systems should be available by year's end to meet our needs as outlined above. The billing system vendors, ARU vendors, a number of MSOs and CableLabs are working to insure our detection and tracking needs can be met.

A reliable means to measure on-going results and to ascertain cause and effect is a prerequisite to making real progress. While cable systems have been operating nearly blind, it looks like the industry is close to bringing this important issue to closure.

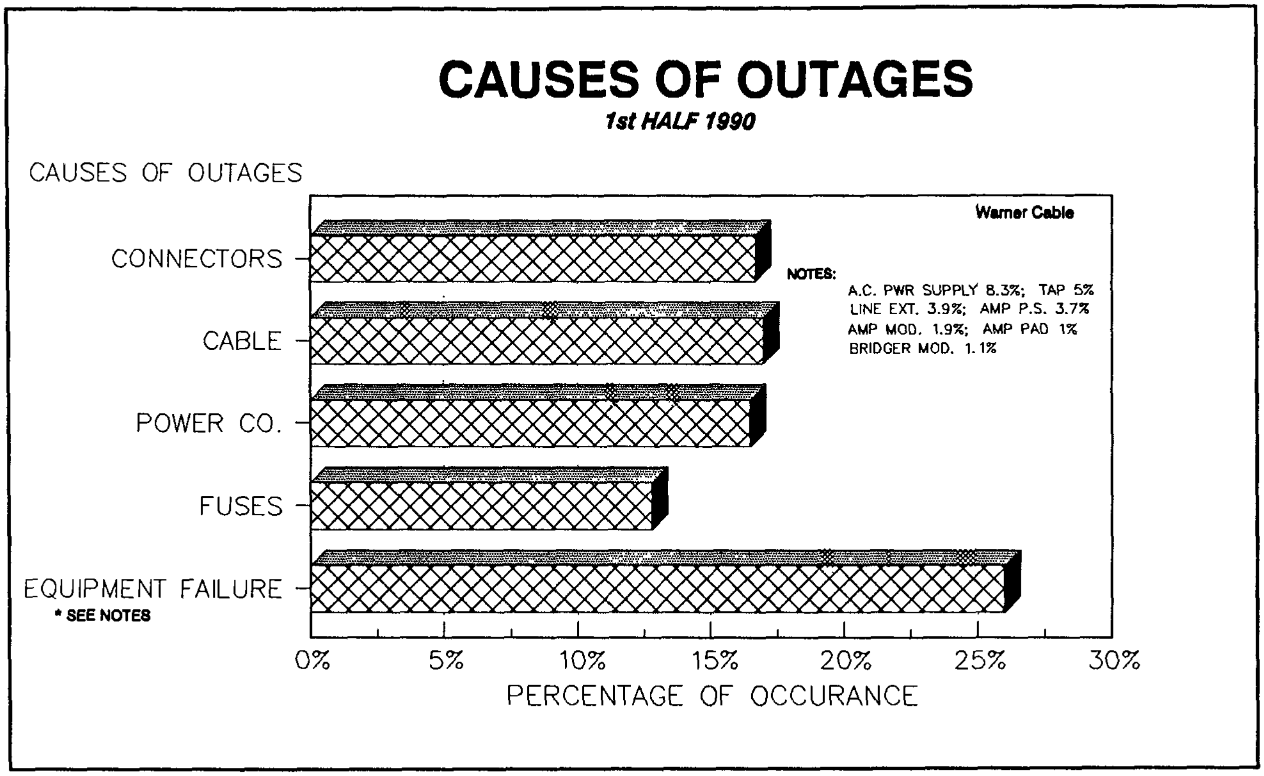

Figure 4

IV. CAUSES OF OUTAGES

A. GENERAL

Figure 4 shows the breakdown of causes of Outages and percentage of occurrence which can be summarized as:

* 25%+ Equipment Failure.

* 15% Cable Problems

* 15% Connector Problems

* 15% Power Company; Loss of A.C. Power

* 13% Nuisance Fuse Blowing

The data comes form Warner Cable over the first half of 1990. I have reviewed this data with a number of MSOs. While there are some variations in percentages, everyone agrees that these are the causes and are in the right order of importance.

One cause not listed is an outages created by system's repair and maintenance activities. Some systems do not consider plant taken down for repair or maintenance to be an outage. However, as we have mentioned earlier, the final judge, the customer, clearly considers this an outage.

Maintenance outages usually cannot be avoided but the impact can be reduced. The CSRs must know when these outages are to occur and for what duration. The impact on customers is greatly reduced if we tell them that we know about the outage because we are doing maintenance and tell them when service will be restored.

There is often latitude as to when maintenance is performed. Maintenance that creates outages should be done during low viewership periods. Some systems today require maintenance on the trunk to be performed during early a.m. hours. Naturally, plant that is in back yard easements should not be worked on during early a.m. hours.

It has been estimated, but without substantiating data, that up to 35% of all outages are from repair and maintenance. Here is an area where a significant improvement in customer perception can be obtained by simply implementing good operating practices.

B. EQUIPMENT FAILURE

Equipment failure can be categorized into two broad causes:

* Voltage surges.

* Poorly designed/manufactured equipment.

Equipment exposed to high voltage surges, unless protected will have high failure rates. Recent work, led by Roy Ehman with Jones-Inter Cable, has concluded that equipment can be adequately protected. It is interesting to note that these surge voltages often get to the equipment via sheath currents. Therefore, surge protection at the a.c. power supply is not sufficient protection.

The other category of equipment failure is poorly designed/manufactured equipment. The cable industry has gone through its share of equipment with excessive failure rates. It is time that we, as an industry, establish acceptable equipment failure rates. Most manufacturers provide reliable equipment most of the time.

Local cable system management needs a precise understanding of what failure rates to expect. The manufacturers need to know what their customers, cable operators, need. I cannot think of one complicated and reliable electronic system without established failure-rate budgets for the system components.

Outages due to equipment failure should be cut at least in half if the equipment is adequately protected from voltage surges and the equipment meets a 3-5% failure rate. There are systems operating with line equipment failure rates in the 3-5% range. Since this is our largest cause of outages, achieving significant reductions here is mandatory if we are to achieve an overall improvement.

C. CONNECTOR AND CABLE FAILURES

Outages caused by cable and connectors are usually a cable system operations issue today. Generally connector failures are workmanship/craftsman issues which can be effectively addressed with a focused training and certification program and routine quality control.

Cable problems are often due to cable cuts, an age-old cable operational issue. Obviously, the protection against cable cuts is to work closely with local groups to ensure notification before digging and to make others aware of the existence of cable plant.

Cable and connectors represent some 30% of the failures. These should be reduced by some 50% by insuring that all personnel are sensitive to the outage issue, work is done right the first time, and people who may cut cable are motivated to first advise us of their activities.

D. POWER COMPANY; LOSS OF A.C. POWER

If there is one message I would like to get across it is the following:

> The Power Company is not responsible for most of our outages; however, the outages they cause can be significantly reduced.

Cable TV power system are designed with little regard to outages. Cable systems design are done with great attention to the number of amplifiers in cascade since this is a key factor to meeting picture quality specifications. Similar attention and design rules are needed for power supply cascade; the number of power supplies in series to feed a customer. A customer fed from 15 different power supplies will experience more outages than a customer that has a 7 power supply cascade.

The Nashua, NH system recently analyzed their powering and found the maximum number of power supplies feeding a customer was 14, and customers on a 15 amplifier cascade were at a 7 power supply cascade. Nashua is now reconfiguring the power system so the maximum power supply cascade is 7 while customers on a 15 amplifier cascade will be on a 4 power supply cascade. The reconfiguration did not reduce the number of power supplies, only changed their location. These power supply cascade reductions will significantly reduce the outages to the customers on the longer cascades where a large number of customer outages occur.

Placement of power supplies is usually done with very little insight into the reliability of the power system at the tie points. Recent discussion with Power Companies has indicated there is differences in the power system reliability depending on a number of factors, including distance from the sub-station.

The number of outages seen by the customer can be reduced if power outages to the trunk were eliminated. Today standby power supplies are an option to insulate trunk from commercial power outages but considerable expense and operational difficulties are encountered with this approach.

Recently an investigation was undertaken looking into using higher voltages to significantly reduce number of power supplies with an eye toward powering all trunk amplifiers directly from the headend in which there is stand by power. There is a cable system in Minnesota using 300 volt supplies which has been operating without problems. There are issues concerning codes, safety, and acceptability to local utilities that need to be addressed before the industry could move to 200-300 volt supplies. But it is very intriguing when you realize that a 200 to 300 volt cable trunk power system could power all trunk amplifiers from the headend where standby power is already in place.

Reducing power supply cascades to customers, working with utilities to insure the most reliable tie points are used, and avoiding outages during planned utility outages by using generators (utilities also have scheduled repair and maintenance) should result in good reductions in power company outages. If a design can be developed that insulates the trunk from commercial power then power company outages should drop to more like 10% the present rate.

E. FUSES: "FUSES ARE OUTAGES WAITING TO HAPPEN"

Some 10% to 15% of outages are "nuisance fuse blowing." A nuisance fuse blowing is when the outage is restored by replacing only the fuse; i.e., no other problem had to be corrected. From discussions with most of the amplifier manufacturers, there is no specific rationale as to when to use fuses, how to size them, and when to use regular, slow blow, etc. Given no specific direction and their desire to protect the equipment, the vendors have often implemented a fusing approach detrimental to outage reduction and errs toward questionable equipment protection.

Fuses are used to protect circuits from high current conditions which arise from a shorting situation. Shorts, while they do occur principally due to repair, are a rare event in a cable system. The condition experienced often is over-voltage due to voltage surges. The over-voltage condition, just as a short, results in a higher current condition for the period of the over-voltage.

Since over-voltage is the condition the plant equipment needs to be protected against then fuses or any type of current sensing device is not appropriate. There are clamping circuit devices available that can very quickly sense the presence of voltage surge and short this surge to protect the circuits.

Specific suggestions will be made in the next section on providing surge protection and the use of other types of circuit protective devices. The basic approach is:

* Provide absolute protection from voltage surges.

* Use fuses only to protect from shorts; not over-voltage.

* Use slow blow fuses.

* Do not use any other current sensing protective devices; they create outages.

It seems reasonable to expect that outages due to nuisance fuse blowing can be essentially eliminated representing in excess of 10% of all outages.

V. REDUCING OUTAGES

A. General

In 1990, CableLabs established an Outages Reduction Task Force. The task force identified 6 areas for reducing outages. These 6 areas are:

1. Plant & Head-end Protection; Bonding, Grounding, Surges, Lightning, and Fuses.
2. Equipment Reliability; MTBF.
3. Outage Detection, Definition, and Acceptable Performance.
4. System Reliability Model.
5. Plant Powering.
6. Outage Prevention Through Operating Practices.

The CableLabs Task Force has established working groups for each of these areas and will be publishing recommendations in the coming months.

B. Plant & Head-end Protection; Bonding, Grounding, Surges, Lightning, and Fuses

There are a number of plant and head-end configuration issues including the use of particular devices that impact equipment reliability and thus outages. The approach is to configure plant and headends so voltage surges do not create equipment failures. It is the opinion of a number of cable engineering experts that this approach can be successfully implemented.

Specific recommendations which will be issued by CableLabs can be summarized as follows:

* Voltage clamping/crowbar circuits used at the power supply and at trunk amplifiers. CableLabs will publish qualification tests for these devices and list devices known to meet these tests.

* Slow blow fuses are the only recommended current sensing protection devices. Remember if over voltage conditions are removed then the only current protection is for short circuit conditions. Fuses are to be used at amplifier AC input to the DC power pack, output of the AC power supply and for AC routing at trunk distribution legs.

* Amplifier coupling and by pass capacitors are to be at 1000 volt rating.

* Gas diodes (Siemen's Diodes), MOV's and circuit breakers are not recommend.

* Bonding is a safety issue and should be done only to the extent required by code. Bonds should not be located on the same pole as spark gap.

* A ground is to be placed at each active that has a clamping/crowbar circuit. The ground should not be placed where there is a power company vertical. At the time of writing this paper, there was not agreement on the need for this ground so watch for specific CableLabs recommendations on this point.

C. Equipment Reliability; MTBF

For cable systems to operate at acceptable low levels of outages and meet the customer expectations, equipment must experience low failure rates. A fair amount of work is needed to ascertain what failure rates or MTBF (Mean Time Between Failure) are acceptable for the various equipment types (trunk amplifiers, line extenders, head-end modulators, etc.) to meet the customer expectations.

As part of the Outage Reduction effort, a group has been established to develop recommended equipment MTBF's. The group will solicit input from equipment manufacturers to address questions of definitions, etc. and seek concurrence on goals to be established.

At the time of writing this paper, activity was just getting underway but is expected that MTBF's for various equipment types will be published by CableLabs in the second half of 1991.

D. SYSTEM RELIABILITY MODEL

A cable system is comprised of a large number of pieces of equipment from TVRO's, receivers, modulators, descramblers, encoders, trunk amplifiers, line power supplies, line extenders, etc. The reliability budget needed from each equipment to achieve a certain customer outage level requires some level of mathematical modeling and study. One of the task force sub-groups is developing this model. It is expected results will be available in mid year and the details of the model and analysis leading to establishing equipment reliability goals should be published in the second half of 1991.

E. PLANT POWERING

As discussed earlier, there are a number of actions that can be taken to reduce the impact to cable customers of losing commercial power. Another working group under the CableLabs Outage Reduction Task force is investigating specific approaches with special focus on:

* Minimizing power outages by working with local Power Company; develop specific areas for Systems to consider.

* Methods to reduce power supply cascade with power system design guidelines.

* Method to insulate trunk from commercial power; including the use of higher voltage.

F. PREVENTION THROUGH OPERATING PRACTICES

There are a number of opportunities where prudent operating practices can reduce impact of outages. One prime area is maintenance outages, which can account for up to 30% of all outages. As mentioned earlier, the scheduling of maintenance outages and insuring that CSRs are cognizant are simple but very effective operating practices. Also as mentioned earlier, doing maintenance at very low viewing times, typically early a.m., greatly reduces the number of customers affected.

If standby power supplies are being used, then operating practices dealing with the process to dispatch a generator before batteries run down, quarterly battery checks, yearly battery replacement, will all greatly impact the effectiveness of the standby units.

While a group leader was not established for this area it is expected that there will be a series of CableLabs recommendations dealing with reduction through smart operating practices later in 1991.

PCN IN CABLE TV'S STRATEGIC PLANS

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ABSTRACT

Personal Communications Networks (PCN) will be an important new service for cable operators when commercial licensing begins in the mid-1990s. The technological and marketing synergies with cable TV make these services an inviting opportunity for cable companies in telephony. The potential synergy of cable TV and PCN has prompted several multiple cable system operators (MSO) to seek authority from FCC to conduct technical and marketing experiments. PCN is a network of advanced digital tetherless telephones, technologically similar to the mobile cellular services but using much more closely spaced microcells, and small cordless handsets suitable for pedestrian use. Low power and spread spectrum techniques are proposed to minimize interference. The disappointing experience in the U.K. with the more limited CT-2 is believed to be primarily a result of excessive haste in granting the four licenses, and in launching the service in April 1990 without fully developed, compatible infrastructure. Three PCN licenses have been awarded in the U.K, with commencement of service not expected before 1992 or 1993. EMCI has conducted extensive research on the market potential for CT-2 and PCN in the U.S. The phenomenal growth of cordless telephones, paging, and cellular telephones in the U.S. suggests significant public demand for truly tetherless personal communication facilities. The capital requirements and operational characteristics of microcell services will differ significantly from services such as cellular due to different system configurations.

INTRODUCTION

Cable TV has long claimed to be an important telecommunications medium. Twenty years ago, the Chairman of the Federal Communications Commission, Dean Burch, told the NCTA:

> *...that it's up to you whether cable is going to be just another way of moving broadcast signals around (hardly worth the ulcers involved) or whether it is going to become a genuinely new and competitively different medium of communications...*

It has been a long time coming, but, in answer to Dean Burch's challenge, cable TV appears finally to be at the threshold of the brave, new world of Personal Communications.

Personal Communications means people communicating with people, without being tethered to a phone jack, wherever they may be: at home, in the backyard, at the office, at the beach, in the grocery store, at the airport, in the car, or just walking down the street.

WHAT IS PCN?

PCN stands for Personal Communications Network, today's hot buzz-word in telecommunications.

PCN is a subset of the family of Personal Communications Services (PCS) that includes the entire gamut of tetherless telephony services.

PCN is a network of advanced digital tetherless telephones, technologically similar to the mobile cellular networks now operating throughout the world. The digital mode assures much higher quality voice transmission and substantially greater privacy than is possible with the present analog cordless telephones.

PCN is being touted as a *radio drop* that is much cheaper to install than standard copper access lines.

Personal Communications Services

VEHICULAR CELLULAR

Mobility:	*Use in moving vehicles on streets and highways almost everywhere in the U.S.*
Information Content:	*Voice communication - incoming and outgoing calls*
Price per Month:	*$80*
Size:	*Briefcase*
Battery Life in Use:	*Indefinite (uses vehicle battery on continuous charge)*

LARGE HANDHELD CELLULAR

Mobility:	*Use in or out of moving vehicles, almost everywhere in the U.S.*
Information Content:	*Voice communication - incoming and outgoing calls*
Price per Month:	*$100*
Size:	*Standard phone*
Battery Life in Use:	*1 week*

SMALL HANDHELD CELLULAR

Mobility:	*Use in or out of moving vehicles almost everywhere in the U.S.*
Information Content:	*Information Content: Voice communication - incoming and outgoing calls*
Price per Month:	*$100*
Size:	*Man's wallet*
Battery Life in Use:	*1 day*

WIDE AREA PAGING

Mobility:	*Use in or out of moving vehicles and carry on belt almost everywhere in the U.S.*
Information Content:	*Beeper or displayed information - receive only*
Price per Month:	*$40*
Size:	*Cigarette lighter*
Battery Life in Use:	*1 month*

STANDARD PAGING

Mobility:	*Use in or out of moving vehicles and carry on belt almost everywhere in the U.S.*
Information Content:	*Beeper or displayed information - receive only*
Price per Month:	*$20*
Size:	*Cigarette Lighter*
Battery Life in Use:	*1 month*

CT-2

Mobility:	*Use with any compatible base station in home and/or office and specified locations on the street*
Information Content:	*Voice communication outgoing calls only*
Price per Month:	*$60*
Size:	*Man's wallet*
Battery Life in Use:	*1 week*

PCN

Mobility:	*Use in home and/or office and in downtown areas*
Information Content:	*Voice communication " incoming and outgoing calls*
Price per month:	*$ 70*
Size:	*Man's wallet*
Battery Life in Use	*1 week*

COMMON CORDLESS CT-1

Mobility:	*Limited to less than 100 feet from its associated base*
Information Content:	*Voice communication - incoming and outgoing calls.*
Price per month:	*$ 5*
Size:	*Standard phone*
Battery Life in Use:	*indefinite, recharge*

Figure 1

Most cellular telephones are installed in automobiles where size and weight are not critical, and where plenty of power is available from a high capacity, continuously charged storage battery. Hand-held cellular phones are still too large and heavy for a shirt pocket or purse, and generally rely on rechargeable NiCad batteries to achieve reasonably long life in use.

The PCN handset, on the other hand, should be no larger or heavier than a wallet or small pocket calculator. It should be able to operate for a reasonable time in use with a small expendable dry-cell battery.

Cellular telephones are designed to communicate with a base station at distances up to several miles. To accomplish this may require up to 50 watts effective radiated power (ERP), more or less, at the cell site, and 1/2 to 3 watts ERP, at the mobile transmitter. When the mobile unit moves out of range of one base station, it is automatically handed off to another base station that is within range.

In order to achieve the objectives of PCN, however, battery capacity limitations mean that handsets could transmit only a few milliwatts ERP. At that level, the effective communication range would be limited to a tenth of a mile, more or less, from a microcell site. To cover just the entire downtown area of a major city would require hundreds of cell site base stations, perhaps even 1,000 to 10,000.

Because of the short coverage range of PCN microcells, vehicles travelling at highway speed might pass in and out of range too quickly for the handoff to be accomplished. For this reason, PCN is expected to be used primarily with handsets that are stationary, or perhaps moving slowly, at walking speed, for example.

PCN TECHNOLOGY

No frequency bands have yet been allocated to PCN, in the U.S. It is virtually certain, however, that spectrum will have to be shared, initially at least, with other radio services, possibly the private operational-fixed microwave service in the 1850-1990 MHz band. For this reason, spread spectrum technology is being seriously considered, here and abroad, to minimize the risk of interference between PCN and other services.

Spread spectrum is a technique developed for military use to neutralize the effect of enemy jamming. The two principal spread spectrum techniques are frequency-hopping (FH) and direct sequence (DS). As the FH term suggests, the carrier frequency is *hopped* about in accordance with a predetermined but repeatable *pseudo-random* sequence. In order to receive the signal, the receiver must be able to replicate the precise hopping pattern as it tracks the desired signal. The receiver will be tuned to an interfering signal for only very brief intervals, if at all. Moreover, the transmitter will rest for such a short time on any particular frequency that it is unlikely to cause interference.

PCN Topology

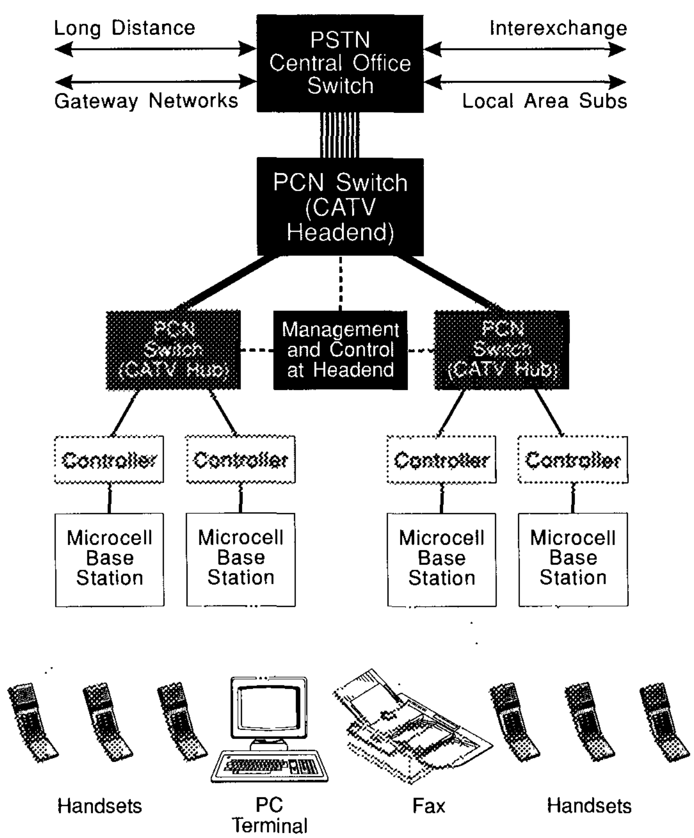

Figure 2

For *direct sequence,* the data stream is modulated with a very wideband, pseudo-random noise signal whose frequency and amplitude distribution is switched in a predetermined but repeatable manner. The data stream is recovered at the receiver by synchronously demodulating with an identical pseudo-random signal.

Access for a multitude of channels can be created in three ways. Frequency Division Multiple Access (FDMA) is comparable to the familiar FDM multiple channelling used exclusively on cable TV networks. For direct sequence spread spectrum, multiple channels can be transmitted either by Time Division Multiple Access (TDMA) or Code Division Multiple Access (CDMA).

In TDMA, each signal channel is assigned a specific time interval. The receiver is synchronized to the transmission so that it can respond only during the time interval designated for the particular desired channel.

In CDMA, a different pseudo-random noise pattern (or *code*) is assigned to each message channel. The receiver responds only to the particular channel for which the pseudo-random code matches the transmitted code. Any signal that has been modulated with a different pseudo-random noise signal, or none at all, will be completely obliterated by noise that is indistinguishable from the truly gaussian distribution. Thus a fair number of digital messages, can be transmitted simultaneously in the same frequency band, using CDMA, without mutual interference.

Reception of the spread spectrum signal, whether using TDMA or CDMA requires knowledge of the precise frequency and amplitude distribution of the pseudo-random noise pattern. TDMA also requires appropriate time synchronizing data. Encrypting this information is a particularly effective way to assure the security of the desired signal against unauthorized reception without the encryption key. Privacy is an important feature of PCN not provided by analog cordless telephones.

In June, 1990, the FCC issued a Notice of Inquiry (NOI) seeking suggestions on frequency allocation and technical standards for PCN. NCTA and many cable operators filed comments, and at least eight MSOs have filed petitions for authority to conduct technical and marketing experiments. Some experimental authorizations have been granted.

THE U.K. EXPERIENCE

In the United Kingdom, PCN licenses have been awarded to three consortia including, as participating members, Pacific Telesis, U.S. West, Mercury Communications, Telefonika of Spain, Deutsche Bundespost Telekom, Millicom, Sony, and Motorola, among others. All three consortia proposed to use the GSM (Global System for Mobile Communications) standards developed by the European Conference of Postal and Telegraph administrations (CEPT) for a pan-European digital cellular network. However, they have recently funded a special committee of the European Telecommunications Standards Institute (ETSI) to decide what changes may be needed for PCN. Commencement of PCN service in the U.K. is not expected before 1992 or 1993.

Two years ago, the U.K. Department of Trade and Industry (DTI) awarded four licenses for CT-2, a limited Personal Communications Service described as an *advanced digital cordless telephone.* CT-2 is an improvement over the conventional cordless telephone (CT-1) in several important respects. First, it is digital, using FDMA. Secondly, handsets are identified for billing rather than the base station, as in conventional cordless phones. This allows calls to be originated through any compatible base station. However, CT-2 cannot receive incoming calls because it does not have the capability to scan all base stations in order to locate the particular handset. Moreover, CT-2 does not have handoff capability.

In addition to use in the home or office, subscribers can use their CT-2 handsets to make outgoing calls from any location within about 600 feet of a compatible public base station, a service called *telepoint.* PBX base stations are also being developed with which multiple handsets could access multiple office telephone lines.

Early experience in the U.K. was disappointing, primarily as a result of excessive haste in granting the licenses and bringing the product to market. The Office of Telecommunications (OFTEL) believed that rapid roll-out would secure a national network of base stations, create demand, and decrease costs.

Three of the four licensees were operational by April 1990. Yet even in the most heavily travelled pedestrian areas of London, telepoint subscribers found it difficult, if not impossible, to locate areas for CT-2 handset use. To make matters even worse, the three carriers implemented telepoint service with different and incompatible proprietary protocols, because the compatible Common Air Interface (CAI) standard protocol was not ready in time for launch. A fourth licensee postponed introducing its product until CAI was ready. It seems almost axiomatic that competing personal communications services must be compatible to succeed.

While the first year of telepoint in the U.K. appears to have been a disaster, there are reasons to believe that the market will strengthen. Base stations must be compatible with the CAI standard protocol in 1991. Recently introduced base stations for PBX and home use are expected to create significant new demand. Major investment is being made to assure easy accessibility to compatible base stations.

CABLE TV SYNERGY

Applications to FCC by several of the top 10 MSOs, and others, for experimental PCN authorizations clearly highlight the synergy with PCN perceived by cable TV network operators.

PCN microcell base stations might be located on a lamppost, or telephone pole, or building at every other street corner. Initially, they could be as large as a battery standby power supply or a small refrigerator; eventually, however, they might be no larger than a cable TV line extender amplifier.

The optical fiber star trunk topology has many advantages for cable TV networks. Coincidentally, however, it is particularly suitable for providing the multiple duplex digital voice circuits required between PCN microcells and cable TV hubs, fiber optic nodes, or headends. Whereas a telephone company might have to install new sub-carrier or broadband plant virtually everywhere, cable TV already has broadband plant in place. The 18 MHz in each direction suggested in the Cablevision Systems experimental proposal can easily be provided in sub-split cable plant. Much more could be provided with the mid-split arrangement in the Institutional Networks.

Cable TV needs only to enlarge and adapt its customer billing, customer service and maintenance infrastructure to accommodate PCN transactions. The enhanced reliability that may be demanded by PCN subscribers can be provided by installing status monitoring facilities, providing automatically switched routing and equipment redundancy, backup power (including UPS where necessary), and by improved preventive maintenance procedures.

Residential PCN Layout

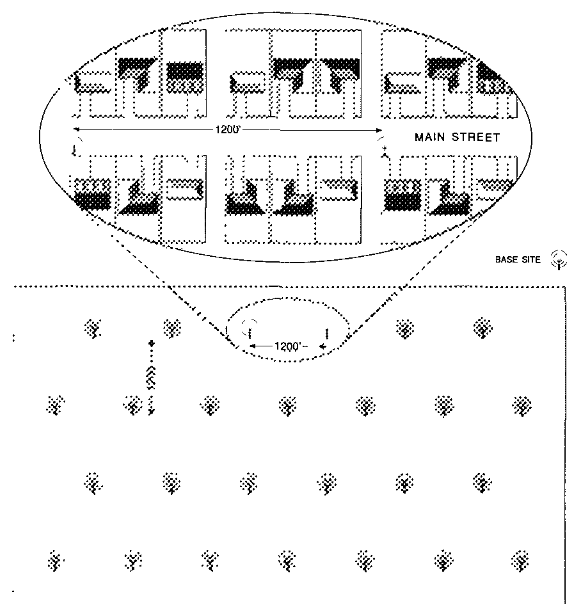

Figure 3

TELCO COMPETITION

The possibility that cable TV operators may be able to compete with the established telcos for Personal Communications subscribers is substantially enhanced by the probability that restrictions on telco entry in cable TV will be at least partially relaxed. Clearly, the trend, not only in the U.S.A, but all over the world, is to encourage competition in virtually every formerly monopolistic institution. It is simply inconceivable that telcos

Comparison of Selected Mobile Communications Technologies

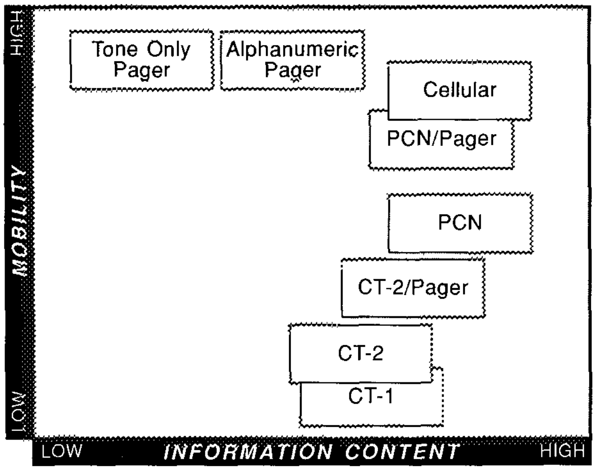

Figure 6

tions of CT-2's demand and operational characteristics are compared to PCN.

To determine the market potential for PCN and CT-2, EMCI analyzed the relative positioning of these technologies against existing mobile communications technologies, primarily cellular. Mobile communications technologies can be viewed as existing within a continuum along two axes: mobility and information content (see Figure 6). Mobility refers to the ubiquity of product use. Pagers and cellular have high mobility since they can be used nationwide. Information content relates to the degree of interaction allowed. Pagers have relatively low information content since they allow only a displayed message to be received by the user. Cellular has high information content since it allows users to send and receive voice communications.

CT-2 will compete to some extent with all mobile technologies. It, however, will also be complementary to pagers. CT-2/pager combinations will have relatively higher mobility and information content than standard CT-2 units. Due to similarities in mobility and information content, CT-2 will be most directly competitive with PCN. CT-2 and PCN will likely also be competitive with cellular, particularly if cellular reacts to these products, reducing its pricing while retaining its relatively high mobility and information content.

Through surveys and econometric analysis, EMCI has examined demand for mobile communications products. Within the price levels examined, the demand for CT-2 is roughly half the demand for cellular. The lower demand for CT-2 at a given price is due to the inability to receive calls and the restricted mobility of CT-2.

Given a price differential between CT-2 and cellular, however, CT-2 can generate demand levels similar to cellular. For example, if the average monthly bill of cellular was $70 and the average monthly bill for CT-2 was $40, there would be virtually the same consumer demand for CT-2 as for cellular (see Figure 7). PCN will generate a similar level of demand to cellular if priced at a 25% discount to cellular. The projected demand for CT-2 and PCN is a combination of new mobile communications users not interested in existing products, consumers switching from existing technologies, and consumers using other mobile communications products in conjunction with these products.

These findings, when combined with information on consumer demand for cellular and competing technologies, can be used to derive market potential estimates for CT-2. By tracking penetration trends for recently operational cellular markets, EMCI has determined that a new cellular market can expect to achieve three to five percent pene-

Preliminary Estimate of Consumer Price Sensitivity for Cellular and CT-2.

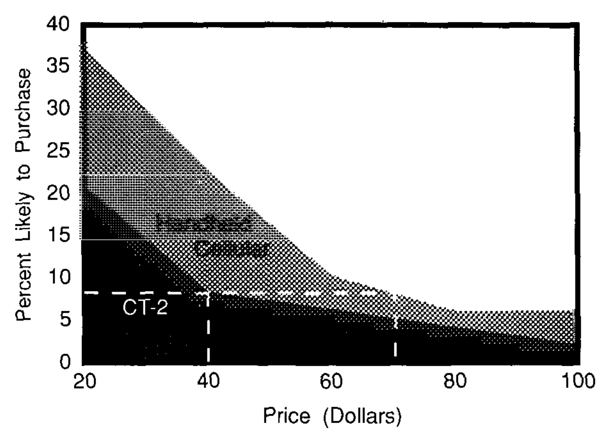

Figure 7

PCN in an Industrial Park

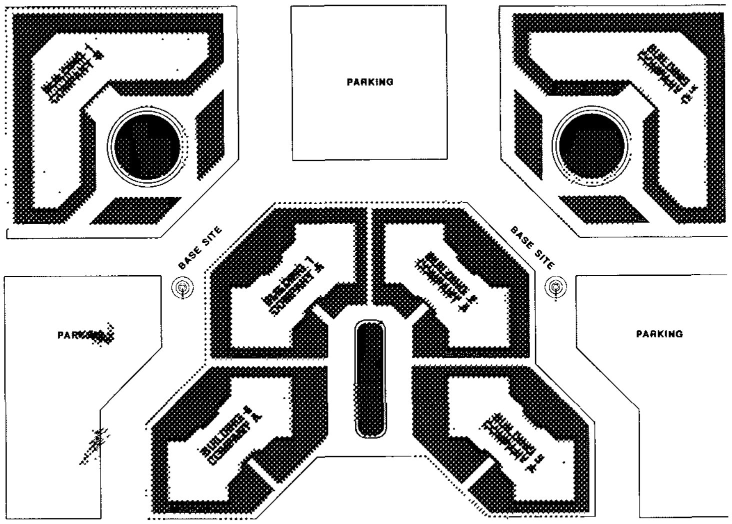

Figure 4

PCN in the Business District

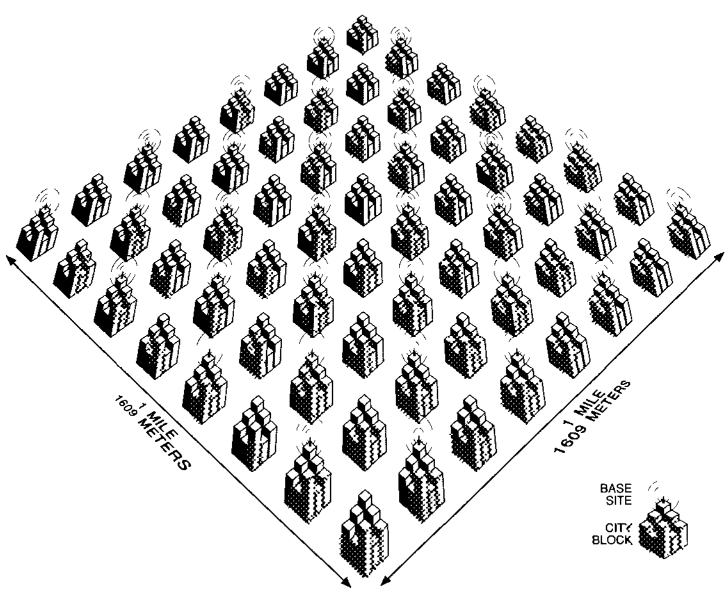

Figure 5

would be permitted to compete with cable TV without also permitting cable TV to compete for PCN subscribers.

Telephone companies, through established tariff procedures, obviously control the charges and terms for access to the Public Switched Telephone Network (PSTN). These are fairly well set for cellular and other interconnected radio communications systems, and not likely to change for PCN. However, telcos also control the charges for pole attachment and duct rental. Already, they have demonstrated apparently predatory tactics by charging as much as $120 per year for fiber optic pole attachments, compared with $5 for the coaxial TV cable on which the fiber is overlashed. This is a matter that must be confronted.

It will take patient investment strategy, lobbying, and litigation by cable TV interests to establish a solid position in PCN. Short of actually operating PCN, cable TV operators could either lease bandwidth on the existing network to unaffiliated PCN operators, or overlay independent networks for lease to PCN.

While the synergies of cable and PCN are clear, operating cost and revenue structures are not yet well-defined. The following analysis by EMCI's President, Andrew Roscoe, examines the market potential, capital requirements, and operating costs for CT-2 and PCN based on anticipated configurations. ■ *AST*

THE MARKET POTENTIAL FOR CT-2 AND PCN

Because the technology and market structure of PCN have yet to be defined, it is difficult to determine financial benchmarks for this service. CT-2, however, has been operational in the U.K. for almost two years. This section examines the market potential for CT-2 service in the U.S., followed by benchmark financial projections using several scenarios. When possible, the implica-

The Market Potential for Telepoint (CT-2) in the U.S. Top 100 Markets

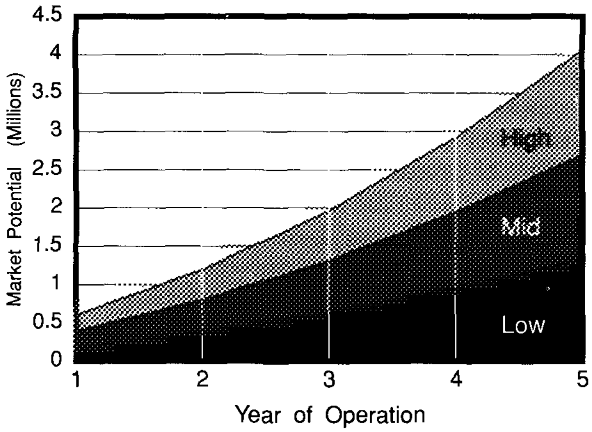

Figure 8

tration of population over a five year period. EMCI's survey research indicates that CT-2 can be feasibly priced at a discount relative to cellular such that total demand for CT-2 is similar to total demand for cellular.

EMCI estimates that the five year market potential for the top 100 U.S. markets is between 1.3 and 4.0 million, with a mid-point of 2.7 million (see Figure 8). The high market penetration estimate is based on little or no competition among other technologies and CT-2 for customers. This forecast implies that CT-2 will generate new subscribers that would not otherwise use mobile communications devices. Since CT-2 does compete with cellular for some of the same market segments, this is considered a maximum market potential. The low market potential estimate assumes that a CT-2 carrier will compete with other carriers, either other CT-2 carriers or cellular and/or PCN carriers and that CT-2 will be competitively priced relative to PCN and cellular.

While survey research indicates that PCN will generate a similar level of demand to cellular if priced at a 25% discount, several industry experts believe that it will be priced at a 50% discount to cellular. If this prediction is realized, the demand for PCN will be several times that of cellular, resulting in a potential PCN market of at least 10 million subscribers after only five years of operation. The U.S. marketplace is highly receptive to mobile communications products. Illustrative is the use of 26 million home cordless phones and over 14 million cellular phones and pagers in 1990 (see Figure 9).

FINANCIAL ANALYSIS OF CT-2 AND PCN OPERATIONS

By applying our market potential estimates to a range of metropolitan markets and combining the results with information on likely pricing levels and capital costs, EMCI has determined the economics of CT-2 operations. This analysis assumes the operation of one CT-2 carrier and omits the impact of roaming. To provide a benchmark for analysis, these results are compared to typical cellular operations. EMCI has examined viability for top 20, top 50, and top 100 metropolitan markets. A net present value (NPV) cash flow per pop (population in the market area) is used as the base for comparison. The NPV for CT-2 is based on a 20% discount rate to account for the relatively greater risk inherent in new technology ventures. Cellular is assessed with a lower level of risk, and a 15% discount rate is used.

In the largest markets, if CT-2 is able to achieve 3.6% penetration by the fifth year of operations (EMCI's high market potential), CT-2 will generate cash flow similar to cellular on a per carrier basis. This is likely an optimistic scenario for several reasons.

The Aggregate Mobile Communications Marketplace Number of Subs by Product

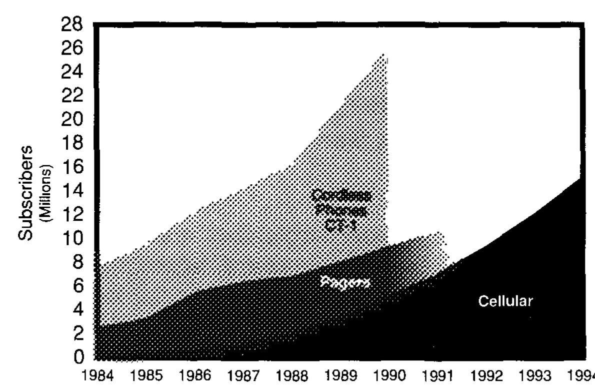

Figure 9

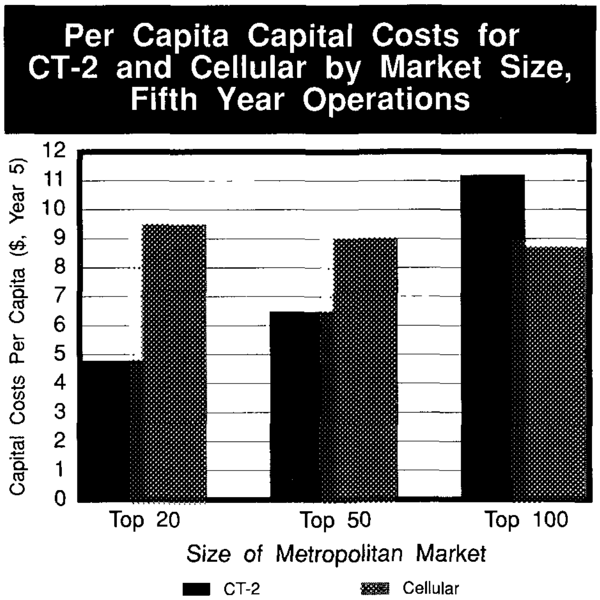

Figure 10

- CT-2 is likely to compete with cellular for some of its subscriber base, reducing its market potential.
- The introduction of PCN services will further erode CT-2's subscriber base.
- The CT-2 system is unlikely to provide coverage convenient for the entire market population. Achieving 3.6% penetration only on covered pops will reduce the subscriber base.

High and mid penetration scenarios present viable business opportunities while the low penetration scenario (approximately 1.2% penetration after five years) represents approximately a break even cash flow business. If CT-2 achieves lower penetration rates, CT-2 is not viable.

EMCI projects both CT-2 and PCN to have lower operating margins than cellular due to the following factors:

- lower revenue per subscriber relative to fixed costs such as management, accounting, and billing;
- significant expenditures to maintain the large number of microcells;
- relatively high interconnection costs which may or may not be directly passed on to the consumer; and
- significant marketing expenses although much lower than cellular on a per subscriber basis.

Under high penetration assumptions, CT-2 will achieve a better than 50% operating margin by the fifth year of operations, somewhat below cellular's average. Because the infrastructure necessary for a viable system cannot be reduced with lower market penetration, operating margins decline as projected penetration declines. This is primarily due to the fact that the number of initial cell sites is high and fixed for coverage purposes unlike cellular where a large area can be covered with a few cells. This affects capital expenditures, technical support and maintenance, and general and administrative expenses. Capital expenditures, in particular, do not decline linearly with market size. While cellular has a relatively constant capital cost per capita over a wide range of market sizes, initial CT-2 capital costs per capita are expected to increase dramatically with smaller markets (see Figure 10).

This analysis indicates that CT-2 may be a viable business even under moderate penetration projections in the largest markets. This result is sensitive to:

- the entry and number of PCN service providers;
- the number of CT-2 service providers; and
- the reaction of cellular service providers to the introduction of CT-2 service.

The variance around these projections increases as the market size decreases. Low population densities present in many metropolitan markets 50-100 will prevent viable CT-2 operations due to excessive capital costs.

PCN services will be differentiated from CT-2 by the ability to offer cell handoffs and full two-way calling capability (send and receive calls). These characteristics require PCN handsets and base stations to have greater technical sophistication than their CT-2 counterparts.

Initial cost estimates for PCN handsets are around $800, but costs are expected to decrease to approximately $300 when mass production becomes possible. This cost would be greater than a CT-2 handset if CT-2 could be mass produced. In the

U.K., CT-2 handsets are available for about $400, but the low demand for this product has precluded its mass production. In comparison to cellular telephones, PCN handsets will be lower cost, weight, and size. EMCI's consumer research indicates that it is important for PCN handsets to be less expensive and smaller than comparable cellular handsets to fulfill its market potential.

PCN and cellular service providers face similar costs. Both must install infrastructure, establish billing systems, and pay interconnect charges. Even though PCN, in theory, can bypass the landline public switched telephone network (PSTN) and switch calls, it must interconnect to the PSTN to access landline telephones and other networks such as cellular and private radio networks. Interconnect fees will be a major variable cost that PCN operators face.

For interconnection, local exchange companies (LECs) charge four to six cents per minute. PCN operators, like cellular operators, will pass this cost through to their customers. But since PCN operators will charge less for service than cellular operators, interconnect fees will represent a higher portion of their costs. A minute of cellular airtime usage costs about 50 cents. For the same minute on a PCN network, a user will only have to pay 25 cents. Thus, as a proportion of revenue, the interconnect charge is 10% for cellular and 20% for PCN. This indicates a lower profit margin for PCN.

Currently available CT-2 base stations cost between $3,000 and $7,000. PCN base stations with their greater complexity will be several times more expensive. They will, however, be closer in price to CT-2 base stations than to cellular cell sites which have a macrocell configuration and cost between $300,000 and $500,000.

As with CT-2, capital expenditures for PCN will not decline linearly with market size. PCN, like CT-2, requires a large number of installed microcells to adequately provide coverage. In the U.K., estimates for PCN network costs range from $1 to $2.5 billion per operator. The midpoint estimate represents an expenditure of $30 per capita per operator. Assuming half the total amount is spent by year 5, capital expenditures still total $15 per capita per operator. This compares to $4 to $12 per capita for CT-2 and $8 to $10 per capita for cellular (see Figure 10).

These prices are not strictly analogous since U.K. PCN costs are compared to U.S. CT-2 and cellular costs. In the U.K., Racal, one of the cellular operators, has spent approximately $14 per capita to date. For CT-2, OFTEL estimated capital expenditures of $100 to $200 million per operator, or about $5 per capita per operator. Ferranti, up through November 1990, had spent approximately $22 million. PCN operators in the U.K., then, have indicated that they will spend more than local cellular and CT-2 operators to implement service. The U.K. implementation of PCN is based on GSM, which is also their digital cellular standard. The U.S. implementation of PCN will likely differ and have substantially different cost characteristics. Whatever the final technology used, clearly, the number of base stations required will result in tremendous costs for the base stations themselves and the connecting network. As discussed above, the availability of existing cable networks, particularly those using the star topology may result in substantial savings for PCN implementation.

CONCLUSION

The concept and structure of PCN continues to evolve. Until the FCC announces a regulatory environment and the industry selects standards, operational characteristics cannot be determined with accuracy. Based on an understanding of existing mobile technology products, the requirements of a microcell technology, and the structure of cable networks, the following points are clear:

- If priced appropriately (at least 25%-50% below cellular for PCN and CT-2), PCN services can realize a large market, possibly several times the size of the existing cellular marketplace.
- The capital expenditures for PCN will be greater than that of cellular.
- Existing cable networks may significantly offset the costs of implementing PCN.

Because of the rapid evolution of the communications marketplace, those MSOs which move quickly to explore the synergies of cable and PCN will be best-positioned to meet the industry's needs when commercial operations begin in the mid-1990s. ■ *ADR.*

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Mr. Taylor has maintained a professional practice as a consulting engineer for the broadcast and cable industries since 1944. He is a Fellow and Life Member of IEEE, a Fellow Member of SCTE (UK), and a Senior Member of SCTE (USA). He is a graduate of Antioch College, with a B.S. in Physics.

He has served on the boards of the National Cable Television Association (NCTA), the Cable Television Advisory Committee (CTAC), and the Broadcast, Cable, and Consumer Electronics Society of the Institute of Electrical and Electronics Engineers (IEEE).

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ANDREW D. ROSCOE, is President and founder of Economic and Management Consultants International, Inc. (EMCI). He has consulted to the communications industry since 1982 on projects including cellular and paging competitive environment analysis, development of financial and business plans, and demand potential projections.

Mr. Roscoe is a leader in research on the potential for PCN and other microcell technologies. Recent projects have included FCC filings for PCN experimental license holders, consumer surveys, and market demand projections. He has interviewed extensively microcell manufacturers and service providers in the U.K.

Mr. Roscoe has published a number of studies on communications technologies and services in the U.S. and abroad. As a leading consultant to the mobile communications industry, he is frequently quoted in the popular, financial, and trade press and often speaks at industry trade shows. Mr. Roscoe holds a Master of Arts Degree in Economics from the George Washington University.

REFERENCES

Cox, Donald C.; *Universal Digital Portable Radio Communications*; Proc IEEE; April 1987; pp 436-477. (Dr. Cox is Division Manager of Radio Research at Bellcore).

Steele, Raymond; *The Cellular Environment of Lightweight Handheld Portables*; IEEE Communications Magazine; July 1989; pp 20-29. (Dr. Steele is Professor of Communications at the University of Southampton, England. In 1986, he formed Multiple Access Communications, a company concerned with digital mobile radio systems).

Cox, Donald C.; *Portable Digital Radio Communications - An Approach to Tetherless Access*; IEEE Communications Magazine; July 1989; pp 30-40.

Matsuo, Naoki; *Personal Telephone Services Using IC-Cards;* IEEE Communications Magazine; July 1989; pp 41-48.

Steele, Raymond; *Deploying Personal Communications Networks'* IEEE Communications Magazine; September 1990; pp 12-15.

Cox, Donald C; *Personal Communications - A Viewpoint*; IEEE Communications Magazine; November, 1990; pp 8-20.

EMCI; *The Market for CT-2*; proprietary report; copyright September, 1990.

EMCI; *European Cellular Market*; Vol I and II; proprietary report; copyright September 1990.

Federal Communications Commission; *Notice of Inquiry;* GEN Docket No. 90-314; June 28, 1990.

Performance of Digital Modulation Methods in Cable Systems

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William Wall

Scientific Atlanta

Abstract

Four digital Modulation Techniques; QPSK, QPR, 4-VSB-AM, and 16-QAM; capable of providing data rates between 10 MBPS and 20 MBPS in a 6 Mhz channel have been studied. A time domain simulation program has been written that allows accurate modeling of the modulation-transmission-demodulation process. Error performance vs. thermal noise has been calculated as a function of amplitude tilt, amplitude ripple, phase noise, quadrature offset, and timing offset. Finally a "typical" cable channel and demodulator have been modeled and error performance calculated.

I. Introduction

Recently introduced new digital audio services and the promise of highly compressed digital video in the near future instigated this study of digital modulation methods for cable applications. Digital modulation offers many potential advantages over analog transmission, such as lower transmission power, more services in a given bandwidth, essentially perfect noise free reproduction of the original source when above some threshold C/N, and unbreakable signal security. In order to utilize these potential advantages a modulation method must be selected that provides reliable reception in a cable environment, a data rate appropriate for the service, and economical demodulation in the subscriber's home. While numerous theoretical analyses of digital modulation performance exist in the literature, a practical comparison under signal conditions typical of cable systems was not found. Further, a method for simulating specific channel conditions and measuring performance for a variety of modulation techniques with those specific channel conditions was desired. To that end a computer simulation has been developed that accurately models digital modulation / demodulation methods in the time domain, adds a variety of controlled impairments, and measures the resultant bit error rate. This simulation technique has been found to be highly accurate and greatly reduces the time to test system design changes as well as test competing modulation methods.

This paper provides a brief overview of digital modulation techniques and the important parameters used in quantifying them, and discusses the major types of channel impairments and their effect on the signal as well as the effects of real (non-ideal) demodulator's performance. The specific modulation methods, QPSK, QPR, 16-QAM, and 4-VSB-AM, modeled in this paper are discussed.

A description is provided of the simulation technique as well as the channel model it simulates.

A model for a "typical" cable channel is discussed along with the performance variation that may be "typical" for a consumer grade demodulator. Simulation results for this model are presented and discussed.

II. Methods of Digital Transmission

As a brief review, digital transmission is accomplished by modulating an RF carrier to a discrete value (or state) that represents one or more bits of information. Each successive discrete state is called a **symbol** and the rate at which they are sent is the **symbol rate**. The modulation may be AM, FM, phase modulation (PM), or a combination of modulation types. If a symbol has only two discrete states, then each symbol represents one bit of data and the symbol rate and data rate are equal. A symbol may represent more than one bit by having multiple values; two bits may be encoded into four discrete values, three bits into eight values, etc. In this manner the symbol rate may be less than the data rate, and the occupied bandwidth of the modulated signal is less. If the transition from one symbol to the next was instantaneous then wide bandwidth modulation sidebands would be created, therefore the modulation must be filtered to limit the bandwidth and control the transitions from one symbol to the next. The key to this filtering is that it must allow the signal to reach the precise discrete value at the proper time independent of the values of the previous symbols. (In one type of modulation discussed, QPR, the current symbol depends on the previous symbol but the current value still must be attained.) A large class of filters, known as **Nyquist** filters meet this requirement. Figure 1 shows the possible trajectories for a four valued signal (QPSK) that is Nyquist filtered. Notice that at the sampling points all trajectories have converged to one of the four discrete values. For a given symbol rate there is a minimum bandwidth known as the **Nyquist bandwidth**, equal to one half the symbol rate at baseband, that a filter can achieve and still not produce any intersymbol interference (**ISI**). This bandwidth cannot be achieved in practice since it requires a perfectly square filter and as one approaches this bandwidth the sensitivity of the signal to distortion and timing errors increases dramatically. The measure of the filter bandwidth is known as **alpha** and varies between zero percent, equal to the Nyquist bandwidth, and one hundred percent, equal to twice the Nyquist bandwidth. Typical values range from 20% for a very spectrum efficient system to 100% where spectrum efficiency is less of a concern than sensitivity to distortion. Figure 2 shows trajectories for a two level signal, one filtered with an alpha of 20% and the other with an alpha of 100%. These diagrams, known as **eye diagrams** for their obvious resemblance, show how the eye becomes narrower due to filter ringing with lower alpha.

The total channel filtering must be Nyquist, that is the product of the transmit filtering plus receive filtering must meet the Nyquist criteria. This objective can be met in a number of ways; one way is to place half of the filter at the transmit side and half at the receive side. The resultant filter, **square root Nyquist**, at each end provides optimum performance for additive gaussian noise but requires a precision filter at each end. A second alternative is to place the entire Nyquist filter at the transmit end and a wider less precise filter at the receive end. Though this has a small noise penalty (approx. 2 dB) associated with it, it may be very cost effective in situations such as cable where there is a single transmitter and many receivers. This approach exhibits less sensitivity to tuning errors and filter imperfections.

Turning now to the receive side, the demodulator must try to make an optimum **estimate** of the state of the transmitted symbol. Depending on the modulation type, the demodulator may be **coherent** or **non-coherent**. Coherent demodulators require a phase reference that tracks the transmitter phase prior to modulation. This phase reference can be derived from a pilot tone or reconstructed from the modulated signal. Since coherent demodulators offer significantly better performance than non-coherent demodulators, only coherent demodulators will be considered here.

Many different forms of modulation are used for digital transmission that vary significantly in their noise performance as well as bandwidth efficiency. These variations have been cataloged elsewhere[1]. For cable applications we have limited the number of techniques studied to those that can provide the relatively high bandwidth efficiency required to deliver between 10 MBPS to 20 MBPS data rate in a 6 MHz channel bandwidth. All of the studied techniques are either AM or PM or combinations of the two. FM techniques do not in general meet these bandwidth efficiencies.

Quadrature Phase Shift Keying (QPSK) is a Phase Modulation technique that has four phase states per symbol located 90 degrees apart. QPSK may be alternately described as a suppressed carrier AM technique where two independently AM modulated signal components in phase quadrature (90 degrees apart) each have two amplitude states. These two components, known as the **in-phase (I)** and **quadrature (Q)** components are orthogonal and may be demodulated separately without mutual interference in the ideal case. Figure 3 shows a phase state diagram for QPSK (alpha 20%) with the I axis horizontal and the Q axis vertical. The four areas of highest density are the four phase states; the remaining lines indicate the trajectories between phase states. With an alpha of 20%, QPSK can achieve a data rate of 10 MBPS in a 6 MHz bandwidth. The noise performance of QPSK is the best of any digital modulation technique.

By adding controlled ISI to QPSK one can effectively generate a third level to each of the quadrature AM components such that the interpretation of that third level depends on the estimate of the previous symbol state, creating a nine state modulation format known as Quadrature Partial Response (QPR). An eye diagram for one component of a QPR signal is shown in Figure 4. QPR provides a higher bandwidth efficiency allowing a data rate of 12 MBPS in a 6 MHz channel bandwidth, with a slight noise performance penalty.

Another modulation method is created by the encoding of two bits in four levels in each of the quadrature AM components which allows the transmission of a total of four bits per symbol. The four by four states gives a total of 16 states per symbol, hence the name 16 state Quadrature AM (16-QAM). Using an alpha of 20%, 16-QAM allows the transmission of 20 MBPS in a 6 MHz channel. Again the addition of the additional states degrades the noise performance.

Taking only the in-phase four level AM component and transmitting it as vestigial sideband AM improves the bandwidth efficiency, allowing a data rate of 12 MBPS to be sent in a 6 MHz channel bandwidth. This technique, unlike the double sideband AM techniques which allow reference carrier regeneration from the transmitted signal, requires the transmission of a carrier or pilot component for a reference carrier regeneration. Referred to as 4-VSB-AM, the noise performance is similar to 16-QAM.

III. Qualitative Effects of Transmission Impairments

There are six main processes that contribute to errors in the demodulated data: thermal noise, phase noise, ISI, timing errors, crosstalk between I and Q, and threshold errors. In most modulation methods thermal noise is gaussian and is additive to each state of a symbol. The demodulator will incorrectly estimate the modulation state if the noise caused the instantaneous received signal to be closer to another state than the actual transmitted state. The "tails" of a gaussian distribution extend far so that even at high signal to noise ratios there is still a measurable error rate. Clearly for a given signal level, the fewer the states in a symbol, the greater the distance between states, and the lower the probability of error; that is the fewer bits per symbol, the better the noise performance. Similarly, the lower the **noise bandwidth** of the receive filter the better the performance assuming that the filter causes no ISI.

Phase noise, either introduced in the transmission path by frequency converters or in the reconstructed reference carrier, degrades performance by introducing apparent thermal noise on the received symbol and by causing crosstalk between in-phase and quadrature components in a given symbol, which reduces the effective distance between states. Inter-symbol interference (ISI) is caused by improper filtering or reflections within the transmission path causing previous symbols to interfere with the current symbol, again reducing effective distance between symbol states (closing the eye).

Nominally the state estimate is made when the eye has the greatest opening, or the distance between states is the greatest. Timing errors in the clock recovery part of the demodulator cause the state estimate to be sampled either before of after the nominal time where the distance between states has been reduced, degrading the demodulator performance. Modulation techniques that use in-phase and quadrature AM components may be degraded by crosstalk between components that reduce effective distance between states. Similarly static bias errors in either phase or amplitude that offset the decision threshold of some symbol states relative to their nominal position will degrade the noise performance of the demodulator.

IV. The Simulation Program and Channel Model

To evaluate the effects of the channel and the filter distortion on various modulation types, a simulation program was written in MATLAB language; MATLAB is an interactive program for numerical linear algebra, matrix computation, and signal processing. The simulation was done using the complex baseband representation of the bandpass modulated signal, meaning a carrier frequency of 0 Hz. The complex baseband representation permits us to sample the time waveform of the studied modulation at a lower rate than what would be necessary if we were using a high carrier frequency. This is done without losing generality and keeping the same properties as a band-limited signal modulating a high frequency carrier. All the parameters of the simulation are normalized to the symbol rate and the filter bandwidths are specified as a ratio of the symbol rate instead of Hz.

The modulator block is an information source which, in the program, is composed of multiple Pseudo Random Binary Sequence (PRBS) generators and two digital to analog converters (D/A), one in-phase (I) and one in quadrature (Q). The simulation being numerical, frames of 4096 complex samples is formed by the sampling of 512 symbols with 8 samples per symbol.

The filters and the channel in the system are modeled as finite impulse response (FIR) digital filters and the coefficients for these filters are real for symmetric filters or complex for asymmetric filters. The filtering is done in the time domain if both the input signal and the filter coefficients are real by convolving the signal with the filter coefficients; if the signal or the filter coefficients are complex the filtering is done in the frequency domain by using a 512 point Fast Fourier Transform (FFT) and the overlap-and-add method for processing long records, instead of doing 2 or 4 convolutions in the time domain. The program also generates all the conventional filter responses plus raised-cosine [2], square-root raised-cosine, partial response and various channel distortions. The group-delay of the designed filters can be specified independently of the magnitude response enabling us to simulate any type of filter technologies like SAW filters, digital filters, or LC filters.

After the signal is passed through the filters and the channel, the signal is demodulated synchronously. In a complex baseband signal, the carrier frequency is zero but the phase of the received signal is unknown and is function of the delay between the modulator and the demodulator. The demodulator extracts the I and Q components along two orthogonal axis offset in phase relative to the modulator phase reference in order to minimize the cross-correlation between I and Q components. The demodulated samples are then resampled by the symbol clock. If the sampling instant falls between demodulated samples, the values at that sampling instant are then estimated by linear interpolation which gives accurate results for a sampling rate to symbol rate of 8 or more.

The Bit Error Rate (BER) measurement is based on the Quasi-Analytical (QA)[3] instead of the direct Monte-Carlo simulation. For linear systems this technique permits accurate measurements of low BER without excessive computation. In the QA technique, also referred to as the hybrid simulation, a simulation is done without the addition of noise and with a source data pattern long enough to obtain all the possible combinations of Inter-Symbol Interference (ISI). A histogram of the clock sampled data is then built, the noise is added analytically to the bins of the histogram, and the average symbol error rate is calculated. To convert the average symbol error rate to BER versus Energy per bits normalized (Eb/No), we need to calibrate our system by measurements of the signal power in the channel, the receiver noise bandwidth, and the number of bits per symbol.

V. Quantitative Effects of Transmission Impairments

We considered the following impairments to a digital signal and divided them in two categories; hardware imperfections and channel distortions.

Hardware imperfections:
- -Quadrature error
- -Phase error
- -Symbol timing error

Channel distortion:
- -Linear slope across passband
- -Sinusoidal ripple across passband

The quadrature error is the deviation from orthogonality of the transmitter or the receiver, the symbol timing error is the deviation from the optimal sampling instant. These two errors are usually due to initial adjustment error or drift caused by components aging or temperature change. The phase error is an indirect measurement of the system sensitivity to phase noise, if the component of the phase noise is a sinusoidal waveform with a frequency less than the symbol rate, the BER degradation due to the untracked RMS phase noise give similar results to the same static phase error in the recovered carrier relative to the optimal carrier phase. This result was verified by simulating both degradations as well as by comparing our data with the results of Tranter, et al.[4]

To simulate the channel distortions encountered in a cable plant we designed two linear phase filters, a linear slope filter and a ripple filter.[5] The slope is defined as the number of dB. variation in a 6 MHz bandwidth. The ripple error is generated by a three tap FIR filter. This filter simulates three path propagation and can simulate the effect of reflection due to mismatch and the triple transit in SAW filters.

The bit rate and the filters for each modulation method in the simulation are selected to have a null-to-null bandwidth of 6 MHz using sharp filters that are today's state of the art. The resultant RF spectrum for each of the modulation methods simulated is shown in Figure 5. The bit rate and filters used are shown in Table 1.

Theoretical performance for each of these modulation methods is shown in Figure 6. Examining these results would suggest that while QPSK is the most robust, there is not much difference among the rest. These results do not take into account the relative sensitivities to distortions and demodulator imperfections.

VI. Results for a "Typical" Cable Channel

Elsewhere in the literature (Ref. 1) results are presented showing the BER degradation of the four modulation types studied in the presence of a single non-optimal condition. In an actual system all forms of degradation will be present in varying degrees simultaneously. The results of each impairment do not linearly contribute to the total system performance degradation, thus it is necessary to model the system with all impairments included. In an attempt to understand the performance of these modulation methods in a "typical" cable environment a set of parameters were chosen to represent a consumer grade tuner and demodulator along with reasonable system performance. These parameters are not intended to be interpreted as worst case conditions. Table 2 lists` these parameters.

The ripple of 0.5 dB peak is typical of triple transit response of a good consumer grade SAW filter. It could also be produced by a reflection 24 dB down on a drop produced by a typical splitter. We included no other frequency response anomalies since most channels are quite flat. The phase noise of 5 degrees rms residual is the total contributed by all sources. The 3 degree quadrature error and 5% timing error are typical of alignment accuracy achieved in consumer products. Drift could increase this number.

The simulation results are shown in Figure 7. These results are presented in Eb/No; to convert to equivalent C/N in a video channel, the ratio of the bit rate to a video channel noise bandwidth of 4.2 MHz converted to dB is added to the Eb/No. This is summarized in Table 3. Also shown is the degradation from ideal performance at a BER of 10^-6.

From these results it is clear that even in this rather benign environment that 16-QAM and 4-VSB-AM require a C/N close to that of video, and that adding margin to account for amplitude tilt and delay distortion may require a higher C/N than existing video services. At slightly higher levels of impairment they may not achieve 10^-6 BER at any C/N. Examining the sensitivity of 16-QAM and 4-VSB-AM to demodulator imperfections indicates that much closer tolerances in manufacturing would be required than for QPSK or QPR. QPR on the other hand appears to be an attractive alternative to QPSK, providing 20% greater data rate with a relatively small penalty in C/N and very little additional complexity in implementation.

VII. Conclusion

Four digital modulation techniques; QPSK, QPR, 4-VSB-AM, and 16-QAM; capable of providing data rates between 10 MBPS and 20 MBPS in a 6 MHz bandwidth, and that are suitable for transmission of digital audio or compressed digital video have been studied. A time domain simulation program has been written that allows accurate simulation of the entire modulation-transmission-demodulation process and calculates error performance. The simulation allows imperfect filters, timing errors, quadrature errors, phase noise, and bias errors to be included explicitly. The results suggest that QPSK and QPR are significantly more rugged than 4-VSB-

AM or 16-QAM. QPR offers 20% greater bandwidth than QPSK with only a minor increase in signal power. Since 4-VSB-AM does not offer any advantages over QPR and is much less rugged, it does not appear to be an attractive alternative. If the higher data rate offered by 16-QAM is essential, a significantly more complex demodulator would be required to provide acceptable performance, and then only with an 11 dB higher signal than QPR.

1. John D. Oetting, "A Comparison of Modulation Techniques for Digital Radio", IEEE Transactions on Communications, Vol. COM-27, No. 12 December 1979.

2. Dr. Kamilo Feher, "Advanced Digital Communications, Systems and Signal Processing Techniques", Prentice-Hall, 1987.

3. Michael J. Jeruchim, "Techniques for Estimating the Bit Error Rate in the Simulation of Digital Communications Systems", IEEE Journal on Selected Areas in Communications, Vol. SAC-2, No. 1, January 1984, pp. 153-170.

4. Walter R. Braun, et al., "CLASS: A Comprehensive Satellite Link Simulation Package", IEEE Journal on Selected Areas in Communications, Vol. SAC-2, No. 1, January 1984, pp. 129-137.

5. William H. Trantor, et al., "Simulation of Communications Systems Using Personal Computers", IEEE Journal on Selected Areas in Communications, Vol. SAC-6, No. 1, January 1988, pp. 13-23.

Table 1: Modulation Parameters

Modulation	Data Rate	Transmit Filter	Receive Filter
QPSK	10 MBPS	Square Root Raised Cosine, alpha 20%,	Square Root Raised Cosine, alpha 20%, Noise BW= 5 MHz
QPR	12 MBPS	Partial Response class 1	Maximally Flat, 6 MHz BW Noise BW= 6 MHz
16-QAM	20 MBPS	Square Root Raised Cosine, alpha 20%,	Square Root Raised Cosine, alpha 20%, Noise BW= 5 MHz
4-VSB-AM	12 MBPS	Square Root Raised Cosine, alpha 40%,	Square Root Raised Cosine, alpha 40%, Noise BW= 3 MHz

Table 2: Typical Cable Performance Parameters

Channel Impairment	Amount
Phase Error	5.0 Deg.
Quadrature Error	3.0 Deg.
Clock Timing	5.0 %
Linear Amplitude Slope	0.0 dB
Amplitude Ripple	0.5 dB Peak

Table 3: Comparison of "Typical" Cable Performance at 10^-6 BER

Modulation Type	Eb/No (dB)	Rate/BW (dB)	C/N* (dB)	Degradation (dB)
QPSK	12.2	3.8	16.0	1.6
QPR	16.2	4.6	20.8	2.7
16-QAM	25.2	6.8	32.0	10.1
4-VSB-AM	26.1	3.8	29.9	10.9

* Note: Equivalent C/N in 4.2 MHz noise bandwidth video channel.

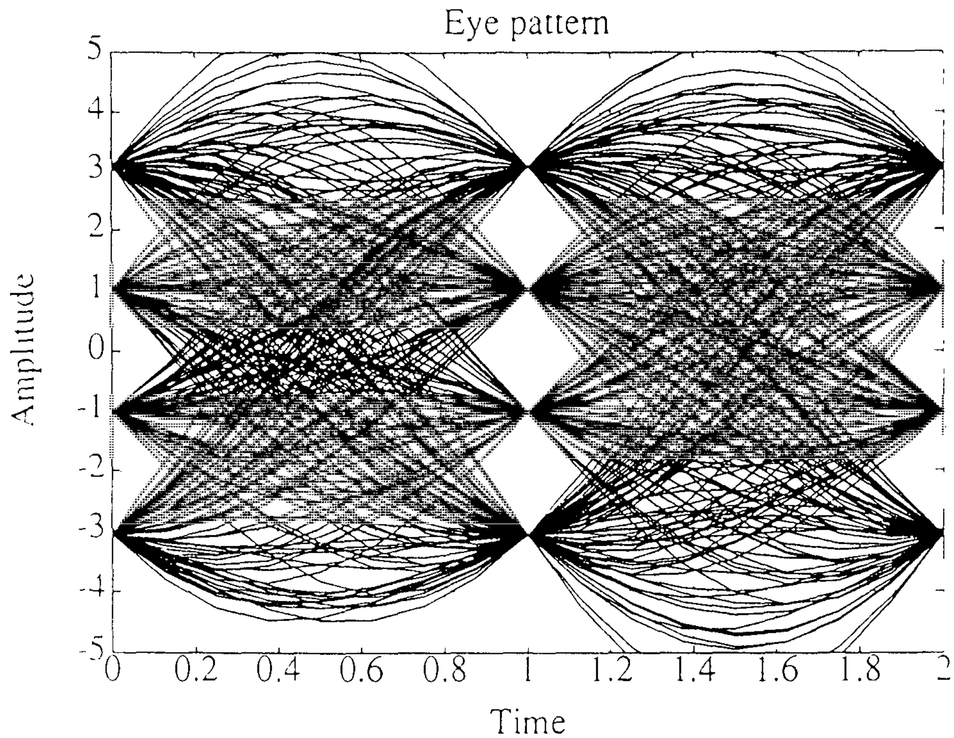

Figure 1. 16-QAM anphase eye pattern, alpha = 20%

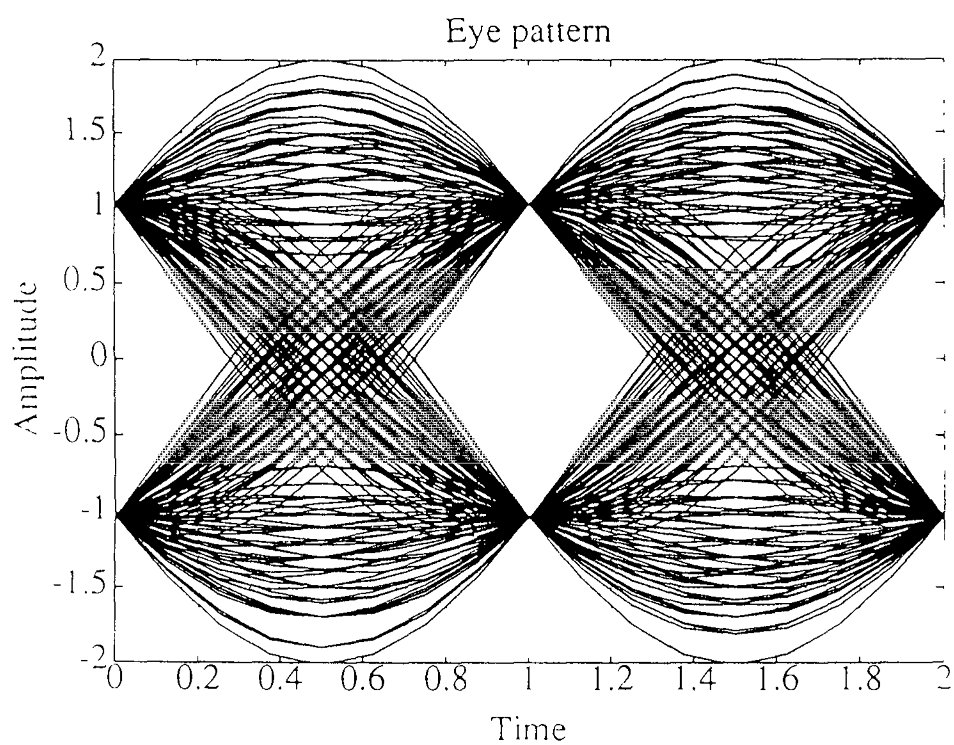

Figure 2a. QPSK inphase eye pattern, alpha = 20%

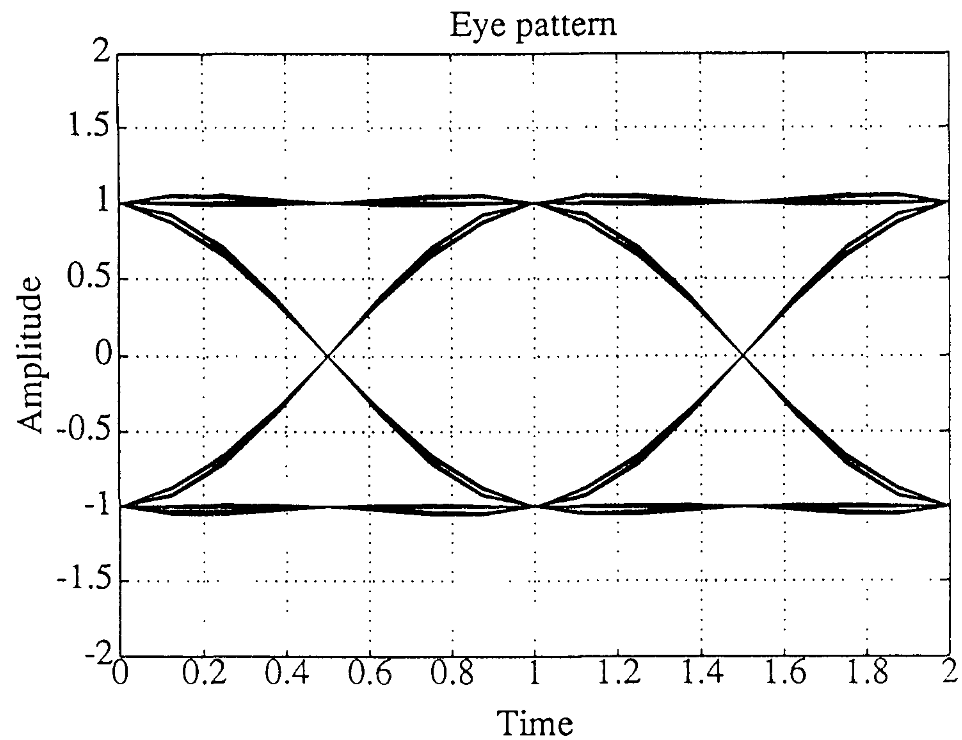

Figure 2b. QPSK Inphase Eye Pattern, alpha = 100%

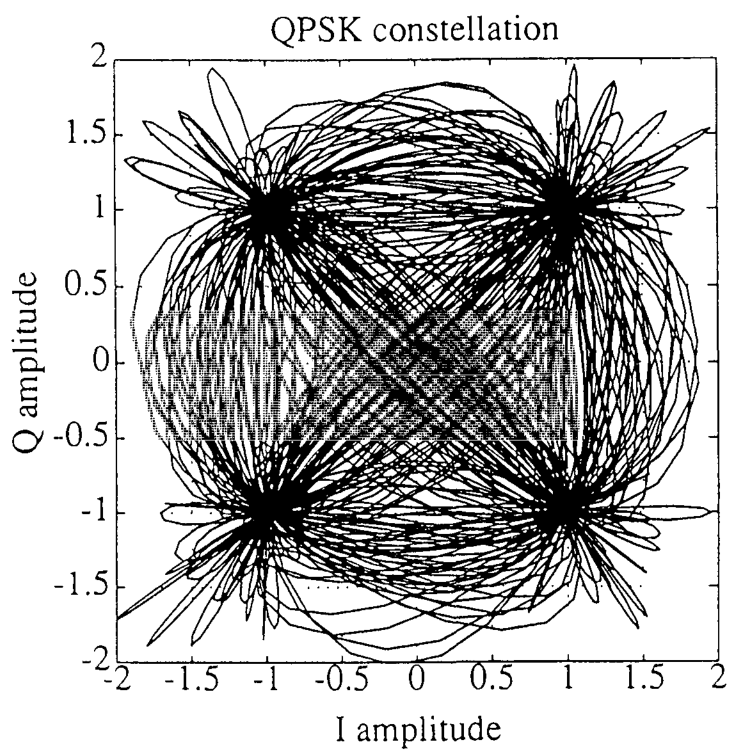

Figure 3. QPSK Phase State Diagram, alpha = 20%

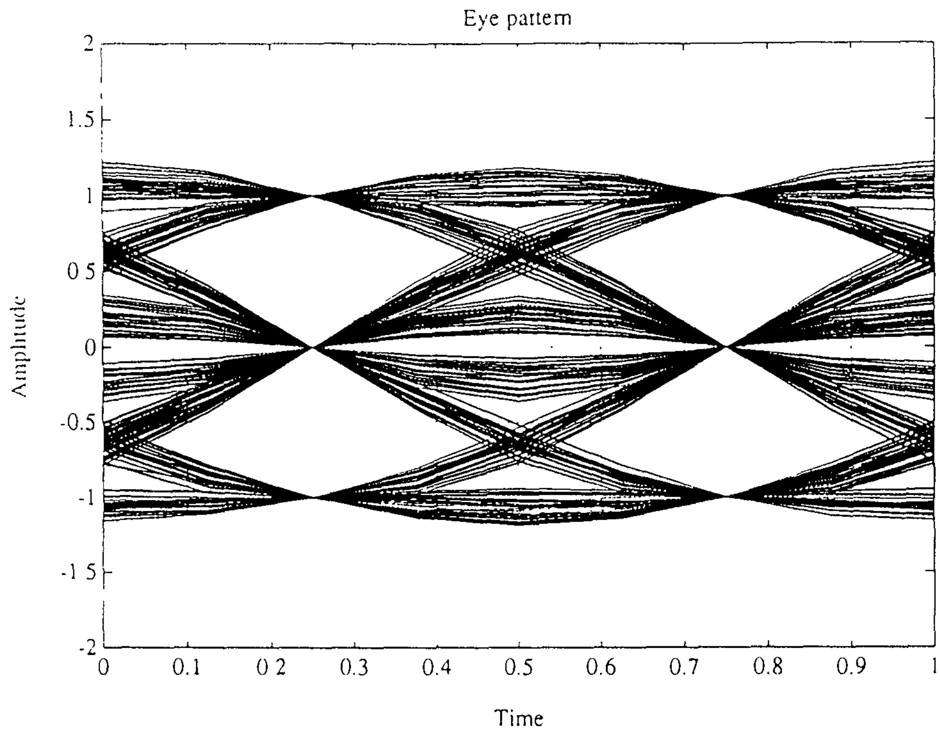

Figure 4. QPR Eye Diagram

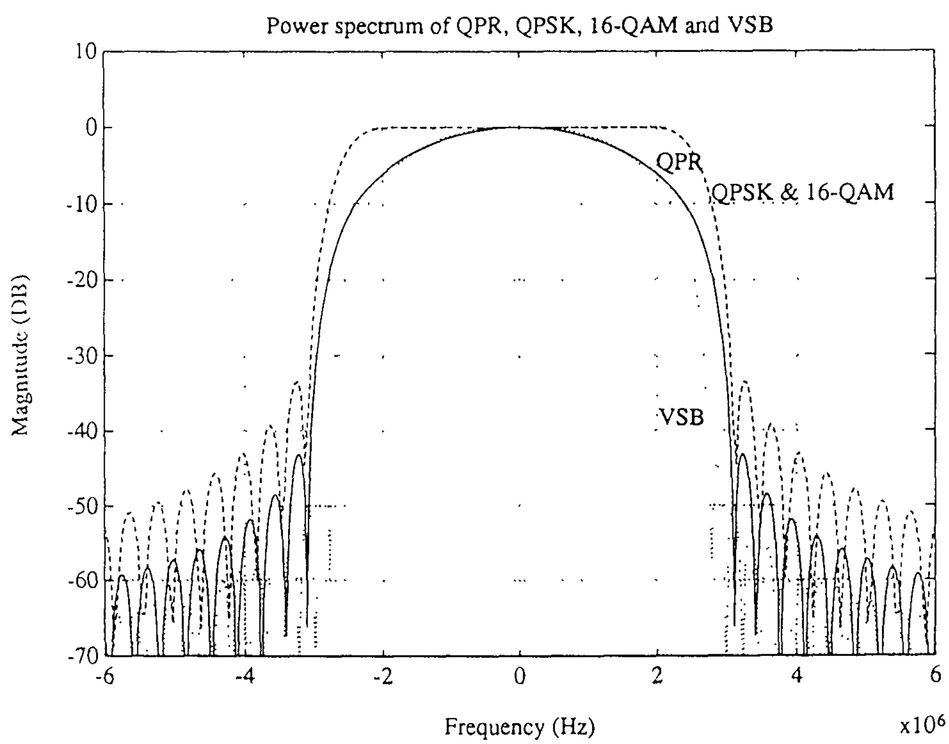

Figure 5. Power Spectrum of the various modulation studied

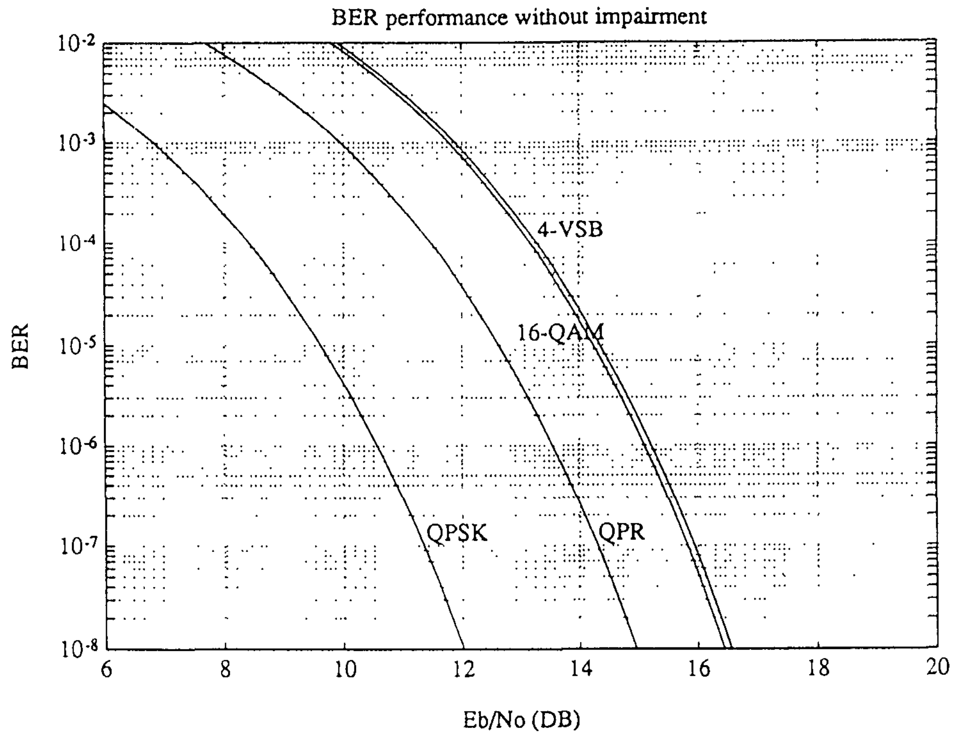

Figure 6. Theoretical Performance of QPSK, QPR, 16 QAM, and 4-VSB-AM

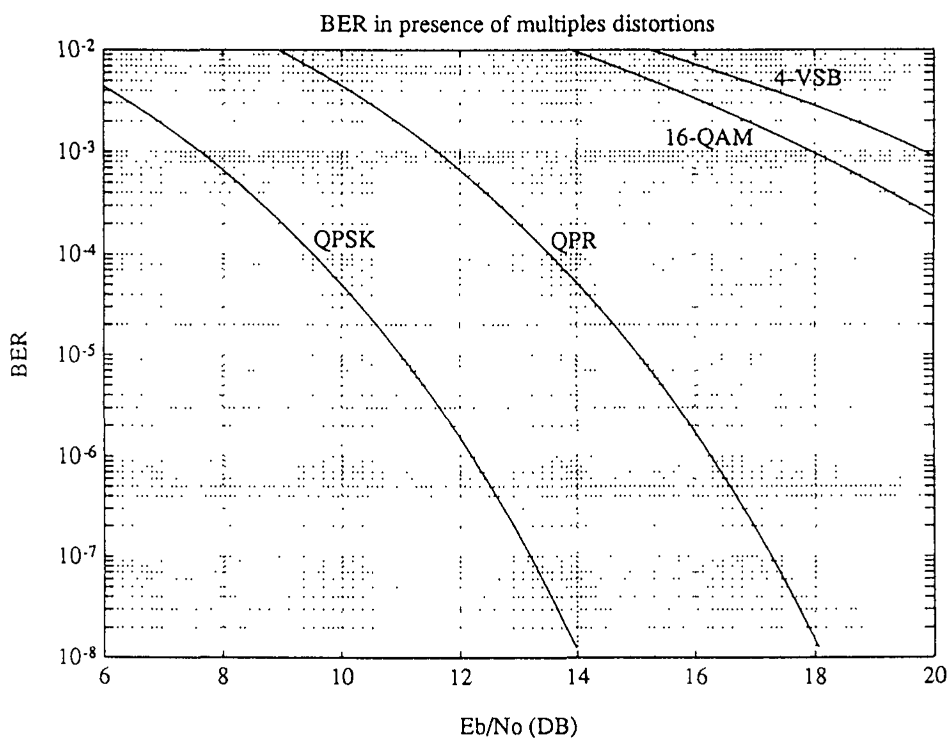

Figure 7. BER performance of QPSK, QPR, 16-QAM and 4-VSB in presence fo the following impairments; Phase error = 5 degrees, Quadrature error = 3 degrees, timing error = 5%, ripple = 0.5 dB peak to peak.

The RF Bypass Converter
An Alternative Broadband Delivery Mechanism

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Abstract

The issue of compatibility between a cable converter and consumer electronics has led to the development of means to concurrently provide all authorized services to the subscriber. Interdiction systems are one example of such means. A significant drawback with interdiction is its use of deny security, which makes it difficult to gradually phase into a cable system.

An alternative approach is presented which provides addressable, concurrent delivery of all authorized services to the home while retaining supply (descrambling) security. This approach is well suited to the subscriber taking one pay channel and desiring access to pay-per-view programming. In some cases an additional, non-addressable trapped pay channel can be offered as well. Because there is a high percentage of single-pay subscribers, this approach might serve the needs of many subscribers while retaining excellent compatibility with consumer electronics.

An overview of a proposed bypass converter is given, and some technical considerations in its design are described.

Compatibility with Consumer Electronics

Consumers and cable operators are sensitized to the potential incompatibility between a cable converter and some consumer electronics devices. For example, the "record one channel while watching another channel" problem requires either a bypass arrangement or two converters, one for the television and one for the VCR. Other examples of incompatibility are with picture in picture, or multiple televisions in the home.

One proposed solution to this interface problem is off-premise control of signals into the home. Cable systems using outside traps are the classic example of broadband delivery of all authorized services to the home, but these, of course, are non-addressable. Other mechanisms for controlling access to signals with outside equipment include interdiction (jamming) and addressable control of trap switching. Both mechanisms allow the full spectrum of authorized signals to enter the subscriber's home, reducing or eliminating any compatibility problem.

Supply versus Deny Security

The interdiction systems generating interest today are typically jamming systems, which interfere with (jam) clear signals before they enter the home. This is an example of deny, or negative, security. The primary advantages of deny security are low distortion of authorized signals and flexibility in jamming non-standard or yet to be standardized signals (for example, HDTV). The clear disadvantage of deny security is the inability to gradually introduce it into a system. With signals in the clear, every subscriber must be equipped with a mechanism to deny unauthorized channels before the service can be turned on. Another disadvantage cited by some operators is the increased risk of piracy with clear channels on the cable plant.

The descrambling converter is an example of supply, or positive, security that allows viewing of an authorized channel by descrambling an otherwise unwatchable picture. As such, the converter has the very significant advantage of being able to be gradually introduced into a system. New services or channels can also be added at any time, since they are secured by scrambling rather than any form of physical security (such as traps). Also, the signals on the cable plant are scrambled and thus less susceptible to piracy. The converter's limitation is its inability to descramble more than one channel at a time, and its single channel output to the television or VCR.

Proposed Bypass Converter

The desirable combination of supply (descrambling) security and broadband delivery of all authorized channels can be at least partially obtained with a bypass converter. This is a converter which can tune and descramble one pay channel, then insert that descrambled channel into the bypassed combination of all other signals. The

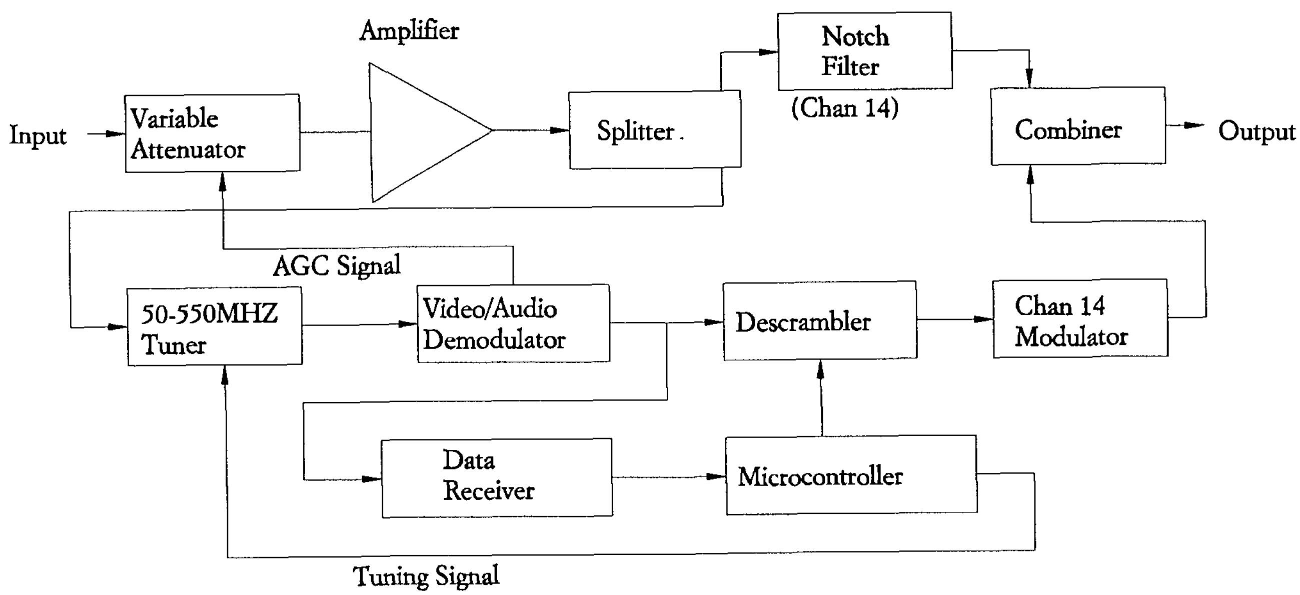

Figure 1: Bypass Converter Block Diagram

block diagram of Figure 1 outlines the basic approach to this type of converter. Such a converter is most applicable in a system with a large non-scrambled basic tier.

A clear understanding of the bypass converter can be gained by study of the block diagram. The input signals, typically in the 50 MHz to 550 MHz range, first pass through a variable attenuator to normalize levels at the input of the amplifier. The amplifier is used to make up for splitting and filter losses, and to set the noise figure of the converter. The output of the amplifier is then split, with one output of the splitter going to a 50 to 550 MHz tuner. This tuner selects the pay channel to be demodulated and descrambled. Tuning is controlled not by a remote control or keyset on the converter, but by a downloaded command from the headend. The descrambled signal is then remodulated to the output channel frequency, for example channel 14. This remodulated signal is then combined with the bypassed signals and presented to the television or VCR.

The bypass path signals are filtered by a band reject, or notch, filter (for example, channel 14), and are then combined with the descrambled and remodulated pay channel signal before being presented to the television or VCR. This filtering on the bypass path removes, for example, channel 14 from the cable plant, making room for the remodulated pay channel to be reinserted. Because this filter cannot have perfectly steep

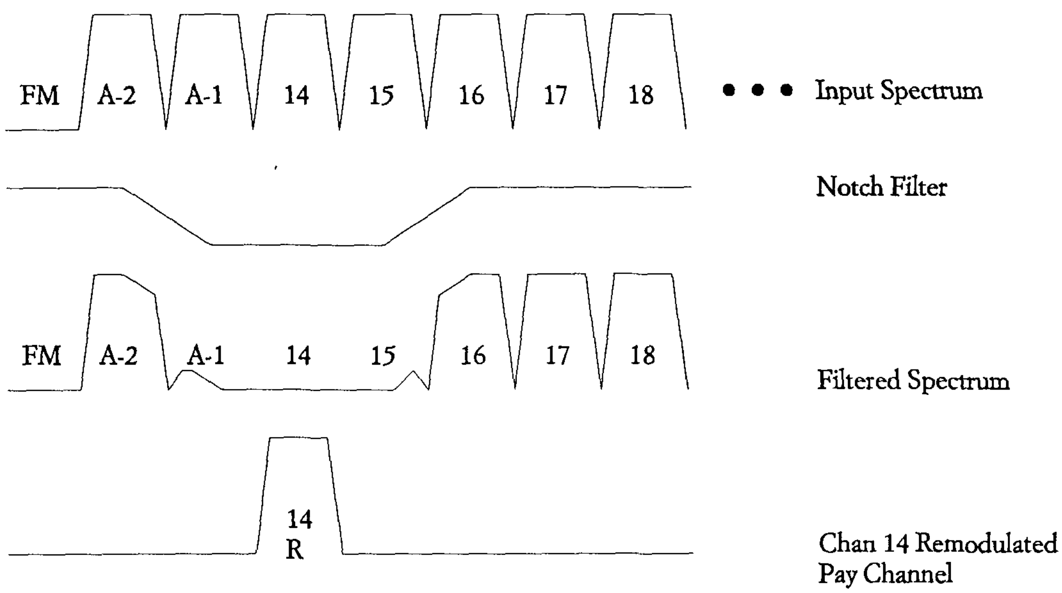

Figure 2: Impact of Notch Filter on Bypass Channels

attenuation skirts, there is some attenuation of channels plus or minus two channels away from the output channel (see Figure 2). The channels on either side of the reinserted channel are thus the ideal location for some of the scrambled pay signals, since they cannot be directly used by the television or VCR.

The result for the subscriber is that the pay channel tuned at any moment is always output on the reinserted channel, while all other channels (many of which are non-scrambled and therefore tuneable by the television or VCR) are bypassed and appear to the television or VCR at their normal channel frequency. As previously mentioned, the tuning of the pay channel is controlled by downloaded data — there is no need for any human interface to the bypass converter. If desired, it can be located behind the set, or even at the cable entrance to the home, allowing service to multiple televisions and VCR's in the home.

The advantages to the subscriber are obvious — the converter can be forgotten and the television, VCR, and remote controls can be used as if in an off-premise system. The VCR can be connected in a straightforward series manner, between the bypass converter and the television, and any authorized channels can be viewed and/or recorded without limitation. The disadvantage is equally obvious — with only one tuner, descrambler and modulator, only one pay channel can be presented to the subscriber at any time. A change of pay service requires a call to the system operator to have the new channel tuning command sent.

While this single pay channel might appear to be quite limiting, many subscribers take only one pay service. An approach to further circumvent this limitation is to use non-addressable, negative trap security outside the home for any service or services which have high and stable penetration. These services don't necessarily need addressability if churn and spin are low, and negative security is appropriate when penetration is very high — only the small percentage of non-subscribers need a trap.

Pay-Per-View Considerations

A bypass converter can be a good choice for those subscribers desiring one (or two, if trapped) pay channels and access to pay-per-view programming. Pay-per-view authorization is accomplished by downloading an event number authorization to the bypass converter, prior to a pay-per-view event. The converter then continuously reviews a list of events on the system and their respective channel numbers. When an event number on the system matches a previously downloaded event number, the bypass converter automatically tunes away from the normal pay channel and to the pay-per-view channel, for the duration of the event. This happens with no input from the subscriber at all, other than a request for the event via CSR, ARU, or ANI. The pay-per-view event automatically appears on the reinserted channel at the proper time.

Technical Issues with a Bypass Converter

There are several technical issues related to a bypass converter that are not issues in the design of a normal converter. These issues include AGC of the bypassed channels, the impact of tilt in the system, control of the relative level of the bypassed channels and the remodulated channel, and the transmission of tagging data to allow preauthorized pay-per-view.

AGC of the bypassed channels as a group is necessary for two reasons — first, to meet the FCC requirement that the output level of a cable system terminal device (CSTD) be limited, and second, to insure a reasonably good level match between the remodulated channel and its adjacent (6 MHz away) and alternate (12 MHz away) bypassed channels. The first requirement must be complied with for FCC acceptance of the device. The level match is required because of the imperfect selectivity of televisions and VCR's. If the channels adjacent or alternate (plus, minus, or plus and minus 6 MHz or 12 MHz from the tuned channel) are much higher than the channel tuned, interference in the picture will result from distortion in the tuner preamplifier or mixer. This interference appears as a beat on the screen.

Televisions are much more tolerant of alternate channel levels higher than the tuned channel, since the broadcast environment uses an alternate channel plan for any geographic area (12 MHz spacing between channels, minimum, except 10 MHz between channels 4 and 5). In this broadcast environment, the level difference between alternate channels can be very significant, as when tuned to a distant channel in the presence of a much stronger local alternate channel. Tolerance of strong adjacents, on the other hand, is not as good.

The adjacent channels can be intentionally reduced in level by the band reject filter at the remodulated channel frequency, easing the accuracy requirement on the AGC. For example, if it is determined that a typical television can tolerate an alternate channel at +20 dB and an adjacent at only +10 dB, the band reject filter can provide the additional 10 dB required rejection of the adjacent. The AGC would then need to keep the bypass channels only within 20 dB of the remodulated channel.

Any AGC works by driving some measured signal to a reference level. The

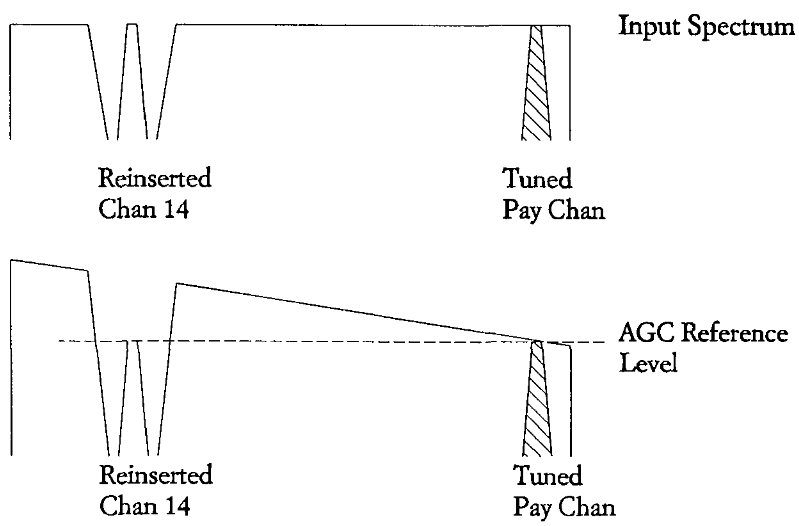

Figure 3: Impact of Tilt on AGC Operation

choice of measured signal to be used for AGC in the bypass converter depends on several factors. CATV amplifiers have used pilot frequencies as the measured signal, sometimes at more than one frequency in the band. The bypass converter cannot rely on the availability of pilot signals, however, so another choice must be made. If the bypass converter is a baseband design, the IF amplifier prior to the video demodulator has an integral AGC detector. This AGC signal provides an accurate measure of the level of the pay channel tuned at any given time. This AGC signal can be conditioned and used to control the wideband attenuator in the bypass path, driving all signals in the bypass path to a known and controlled level equal to the fixed level of the remodulated channel (see Figure 3).

In the absence of tilt on the system, this AGC method has no significant problems. However, if there is frequency response tilt on the system at the input to the bypass converter, the AGC accuracy will be degraded. AGC accuracy can be defined as how closely the channels adjacent and alternate to the remodulated channel are matched to the level of the remodulated channel. Consider the example, shown in Figure 3, of 5 dB of tilt between channel 14 (reinserted pay channel) and a tuned pay channel 61, near the top of the input spectrum. The

AGC drives channel 61 level at the output to match the remodulated channel 14 level. Because of the tilt, however, the channels adjacent to 14 are actually 5 dB higher than channel 14. A positive tilt on the system would have the opposite effect, with channel 14 output 5 dB higher than its adjacents.

Measurement of the alternate channel selectivity characteristics of a number of televisions indicates that tilt is not a major factor in the AGC design. Assuming that the AGC can match the tuned channel output level and the remodulated channel output level with reasonable accuracy, the channels adjacent to the remodulated channel are attenuated enough by the band reject filter to prevent visible interference, even if the adjacents are up to approximately 10 dB higher than the remodulated channel due to tilt on the system. The opposite case, where the remodulated channel ends up 10 dB higher than the adjacents, is of no concern because the television cannot directly tune the adjacent or alternate channels.

Tagging data must be handled differently with a bypass converter than with a regular converter. With a regular converter, tagging data specific to the tuned channel is picked up whenever that channel is tuned. The most important part of this data is the service code or event number. It is this data that tells the converter what service or event is on the particular channel. A check for match with previously downloaded authorized service codes and event numbers is made each time a channel is tuned, and if there is a match the descrambling is allowed. In the case of a pay-per-view channel, the event number is changing with each event.

The bypass converter is not tuned by the subscriber, however, so it must have some other means of knowing what events are on channels it is not tuned to, to allow the comparison to its previously authorized event numbers. This is accomplished by repeatedly sending a list of event numbers and their corresponding channels, on every pay channel. Therefore, regardless of which pay channel the bypass converter is tuned to, it will know what events are "playing" on all other channels. A constant comparison is then made between events on the system and authorized event numbers in the converter, and if a match is found, the corresponding channel is automatically tuned and descrambled.

Summary

The issue of compatibility between a cable converter and consumer electronics continues, and various approaches to improve this compatibility have been proposed or offered to the marketplace. One of these alternatives is interdiction, using jamming carriers. A significant drawback to interdiction is its deny security, which requires fully populating a system with interdiction devices before service can begin.

One primary benefit of a converter is its use of supply, or descrambling, security. A bypass converter has been proposed which combines the advantage of the converter's descrambling security with the consumer electronics compatibility of off-premise systems. The bypass converter is best used in those systems with a significant number of non-scrambled basic channels, and for those subscribers who take one (or possibly two) pay services. It is also an effective method of offering pay-per-view programming to those subscribers.

The Role of A Linearized External Modulator in The Video Distribution Network

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Presentation Summary

The past two years has seen the introduction of high quality linear fiberoptic transmitter into the CATV trunk plant. These lasers to date have been limited to 1.3 um. With the rapid maturity of optical fiber amplifiers (EDFA) which operates at 1.5 um window, linear source operation at 1.5 um are becoming increasingly important. This presentation will focus primarily on the use of external modulator as a 1.5 um signal source. The key parameters influencing the signal source selection will be outlined. Linearization techniques of intensity external modulators will be reviewed. Then, results of an electronic predistortion linearization approach are detailed. The attractive features of using this type of signal source in a video multicast system employing EDFA are also described.

Many research groups have conducted studies on different techniques of linearizing the external modulator. Basically there are two different approaches. One is more involved at the system level. With this approach, both optical [1] and electronic domain techniques [2] [3] have been used. The other technique concentrated more on the device aspects. There are many activities utilizing the latter approach [4] to [6], however, they lack experimental results to date. Regardless of the approach taken, there are fundamentally two distinct problems; one is the optical device domain problems, the other is the RF domain. A good design must take into account the interrelation between these two aspects of the problems. Depending on the type of devices on hand, one has to carefully separate the tasks to be worked on. By properly separating the tasks into the optical and electrical domain, a robust linear external modulator (LEM) can be realized.

The best modulator linearization approach depends on the device under consideration. For CATV applications, the main concern is the performance of the linearized modulator under multi carrier modulation. In the case where electronic signal processing is involved, the EM device response must be made as ideal as possible. The residual non-ideality will likely limit the overall performance.

In the situation where the optical circuit level is involved, the type of optical components making up the device should be carefully selected. Additionally, the stability of the optical circuit component must be taken into account so that the linearized device will maintain a given level of performance over the entire operating range.

In this presentation, an experimental electronic predistortion approach are taken. The standard lumped electrode Mach Zehnder (MZ) modulator operated at 1.5 um is employed. As a result, the well known device non-ideality must be taken into account. This includes the acoustical resonance and the frequency response peaking. The light source used is an Erbium doped glass laser. The spectral content has three distinct longitudinal modes. The center mode is at 1534 nm. The available output power is + 14 dBm.

Linearity measurements of the LEM are conducted using the arrangement shown in the Figure 1. Due to the large frequency response peaking, the experiment is limited to a 20 channel carrier modulation. The frequency range selected is from 330 MHz to 450 MHz. The measurement results show that a CTB improvement of 17 to 21 dB across the band is achieved. The actual operating conditions will be determined by the system specification requirements. Specifically, the CNR and CTB must be given in order to optimize the system.

Evaluations of the transmitter system performance are conducted under the following test condition. The average optical received power is at 0 dBm and +3 dBm respectively. To obtain these levels, either optical attenuators or actual fiber links are used. The modulation index, m, used are 4% and 5.7% per channel, respectively. The CNR are measured to be >56 dBc and >58 dBc across the band for the corresponding modulation indices. The CTB are better than 63 dBc and 67 dBc for the two index of modulation measured. The CSO are measured at selected frequencies outside the band of interest. The purpose is to have a complete system evaluation. They are measured to be better than 67 dBc for both case of measurements.

To achieve higher CNR, one clearly must operate the system closer to the shot noise limit. This means that, both higher modulation index and higher detected power are playing the key roles. The external modulator should have sufficiently high optical output power. However, the maximum modulation index strongly depends on the achievable linearization improvement and the number of modulating carriers for a given CTB specification. For given CTB, m, and N (number of carriers) specifications, the highest CNR will be achieved at the maximum allowable incident optical power onto the detector. The limit will be determined by the non-linearity degradation incurred by the detector used. On the other hand, in a system where an EDFA is employed, a different system parameter will lightly to limit the system performance. Therefore, a different system design approach must be evaluated.

Next, the system is evaluated using different fiber links. One measurement conducted used the 45 km 1.5 um fiber link. All the CNR, CTB, and CSO did not suffer discernible degradation compared to the measured value given in the previous section. The detection are performed at the same incident optical received power. Another measurement is conducted using a 13 km lenght of 1.3 um fiber link. After the 1.5 um signal propagating through the link, there is no detectable degradation in the system dynamic range. These results provide evidence that the transmitter has low source phase noise. This is expected since the Erbium laser has much longer coherent-length compared to a semiconductor laser diode. Another important point is that the system can be used in a conventional 1.3 um fiber video multicast system. Measurement of this transmitter with a high power EDFA will also be conducted. It is expected that this signal source will be ideally matched to this type of multicast system.

In summary, LEM signal sources is an important alternative source for the EDFA AM-FDM video distribution system. A solid state laser and an external modulator can be teamed to produce a high performance optical AM-FDM video link. One important advantage of the solid state laser is that it has very low optical phase noise. Another key advantage is that a standard 1.3 um fiber can be used with this 1.5 um signal without suffering significant degradation on the system dynamic range, especially the CSO.

Reference

1. L.M. Johnson and H.V. Roussell, "Linearization of an interferometric modulator at microwave frequencies by polarization mixing," Conference Digest, LEOS Summer Topical meetings, July 23-25 1990, Monterey, CA, paper BAM14.

2. R.B. Childs and V.A. O'Byrne, "Multichannel AM video transmission Using a high power Nd:YAG laser and linearized external modulator," IEEE JSAC, Vol. 8, No. 7, pp 1369-76, Sept. 1990.

3. Y. Trisno, D. Huber, L. Chen, "A linearized external modulator for analog applications," presented at the SPIE OE/Fiber Conf., Sept 17-19 1990, San Jose, CA.

4. Z. Lin and W. Chang, "Waveguide modulators with extended linear dynamic range, a theoretical prediction," IEEE LEOS Summer Topical Meeting, Broadband analog optoelectronic, PDP-2.

5. G. S. Maurer, P. W. Cornish, R. A. Becker, "New integrated optic modulator design for AM video transmission," Tech. digest of the OFC 91, San Diego, CA., Feb 18-22 1991, paper ThI5.

6. P. Liu, B. Li, and Y. Trisno, "In search of linear amplitude modulator," to appear in IEEE PTL, Feb. 1991.

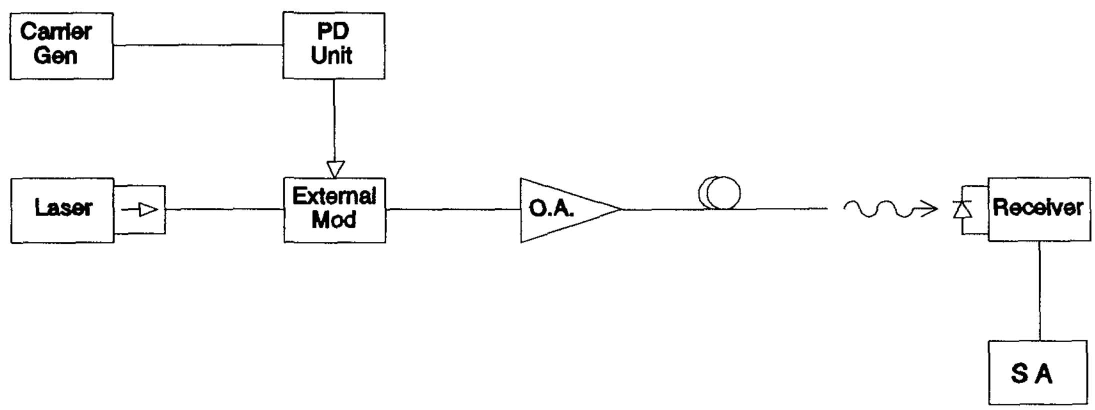

Figure 1. The experimental set-up for measuring both the linearity improvement of the linearized external modulator and the transmitter system performances.

UPGRADING COAXIAL DISTRIBUTION NETWORKS WITH AMPLIFIED TAPS

Exploring a Reliable, Cost-Effective Approach to GigaHertz CATV Plant

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ABSTRACT

Substantial progress has been made in the last few years in improving approaches to the trunking portion of CATV plant, largely through innovations in broadband analog optical fiber transmission technology. While this provides a trunking system with essentially unlimited potential bandwidth and excellent performance specifications, it leaves the remaining coaxial distribution plant as the weak point in these networks. This paper presents an approach to distribution architecture, and to tap design, which addresses this issue. This approach greatly reduces or eliminates the use of in-line amplification in distribution plant, and introduces the use of "active taps." This means that the reach of the distribution plant is determined primarily by cable loss, as splitting loss is largely eliminated. Active devices are used to provide isolation and output levels sufficient to drive subscriber drops. The failure of any active device in such a system would affect only the very few subscribers fed by that individual tap.

In addition to an architectural and strategic overview, specific tap design possibilities are outlined, and the capital and operating economics of such a plant are reviewed. This paper is intended to contribute to a dialogue in the cable television industry which may lead to the development of a new family of coaxial distribution hardware.

INTRODUCTION

The cable industry today is in the process of making dramatic changes to its network architecture. Traditionally, cable plant has been designed with two primary elements, the first of which is trunk plant, providing branched coaxial distribution of high quality signals from the central headend (or regional hubs in very large communities) deep into the system, within one or two miles of every subscriber. The second element is the distribution plant, which consists of a branched and tapped coaxial network passing every possible subscriber in the service area, in sufficient proximity to provide for a service drop distance of no more than 200 to 300 feet.

Because of coaxial cable and branching losses, traditional trunk plant requires broadband amplifiers every 2,000 feet or less. The resulting cascades, or series, of trunk amplifiers, with their additive noise and intermodulation distortion, provide practical limits to the achievable bandwidth, reliability, and signal quality of today's CATV systems.[1] The replacement or reinforcement of the trunk plant with low loss optical fiber can dramatically improve the channel capacity, transmission quality, and reliability in this portion of the system. Advancements over the last several years in low noise, high bandwidth, highly linear lasers and detectors have made this replacement of coaxial trunk plant with fiber trunking cost-effective, and much of the new construction and system upgrades now underway take advantage of this technology.[2,3] It is quite feasible to construct high quality trunking plant with a useable bandwidth well in excess of 1 GHz today, using off-the-shelf lasers and detectors.

The evolution of coaxial distribution plant architecture as bandwidths increase has proven to be more challenging. As channel capacities increase, carrier loading and noise bandwidth increase, and coaxial cable loss increases as well. This means that the distribution amplifiers required for conventional distribution architectures in high capacity systems must have significantly greater performance capabilities than those available today. In addition, losses introduced by splitting and power-passing devices become more critical at higher frequencies.

On top of these challenges lies an opportunity. The cable industry as structured today is remarkably labor-intensive. The drop connection to each subscriber must be physically connected and then disconnected when that subscriber decides to receive or terminate service.

One interesting solution to the formidable problems of bandwidth expansion may be offered by the replacement of today's passive coaxial tapping devices with active devices. This could be realized by the provision of an amplifier for each subscriber or small group of subscribers, coupled to the distribution transmission cables either passively or actively. As will be seen, this may provide an opportunity to substantially extend the reach of coaxial cable without the use of distribution amplifiers, or, alternatively, allow the minimization of amplifier cascades. The introduction of active electronics at the tap means coming to grips with difficult issues of powering and reliability in an electrically and physically hostile environment. It also carries with it an opportunity to significantly improve operating efficiencies, however. Once there are active electronics at the tap, there should be little additional cost in providing on/off switching for each subscriber, eliminating a major source of cable industry labor.

In addition to these advantages, active taps, through the replacement of "lumped" gain blocks within the distribution plant by "distributed gain" in the subscriber leg of each tap-off device, should provide an opportunity to improve perceived plant reliability significantly. Even though a much larger number of active devices would exist in the plant, only one would exist between each subscriber and the fiber trunking system. This means that widespread outages would become much less frequent than in today's system architectures, since device failures would generally affect only one or a very small number of subscribers.

There have been several attempts in the past to realize active tap electronics. Each has met with frustration, but the advent of new types of electronics and new techniques to protect semiconductor devices from voltage transients and current surges, coupled with challenges facing the cable industry regarding channel capacity and customer service, may mean that the time has come to revisit this idea.

THE ACTIVE TAP CONCEPT

Current Tap Technology

The taps used in cable television systems today have one primary function, which is to tap off a percentage of the broadband RF signal power on the distribution line to distribute to the subscribers' homes. An additional requirement is that they allow 60 volt, 60 Hz powering to flow along the coaxial distribution line, while blocking voltage from the tap output ports which feed subscriber drops.

The conventional tap configuration, shown in Figure 1, is a simple transformer-wound directional coupler, feeding a four-output RF power splitter. In order to achieve AC power-passing capability, an RF choke is added in parallel to blocking capacitors which isolate the RF coupler. The "tap value" or coupling ratio of the transformer is selected based on the desired percentage of signal power to be tapped off.

Each tap installed in the distribution line attenuates the signal power passing along the line as it taps off signal. The amount of insertion loss varies with tap value. The total insertion loss caused by a tap can be characterized as the sum of: i) the reduction in signal power resulting from the power split of the directional coupler; ii) the power lost to inefficiencies of the directional coupler's ferrite transformer; and iii) the power lost as a result of the 60 Hz line power bypass, blocking network, and associated matching networks. The excess insertion loss, that is, loss in excess of the theoretical value for the power split, can equal or exceed 1 dB and is frequency dependent.

Tap insertion loss, when added to the signal attenuation due to the distribution cable, dictates the maximum distance between distribution amplifier locations. The cumulative insertion loss caused by the taps in the feeder line is proportional to the required signal level at the output of the taps which feed subscribers. As the bandwidths of cable television systems increase, the tap output level must also increase due to the added attenuation of the drop cable.

Figure 2 shows the effects of tap insertion loss when combined with cable attenuation in a sample feeder line. The diagram also shows how different cable sizes at different frequencies,

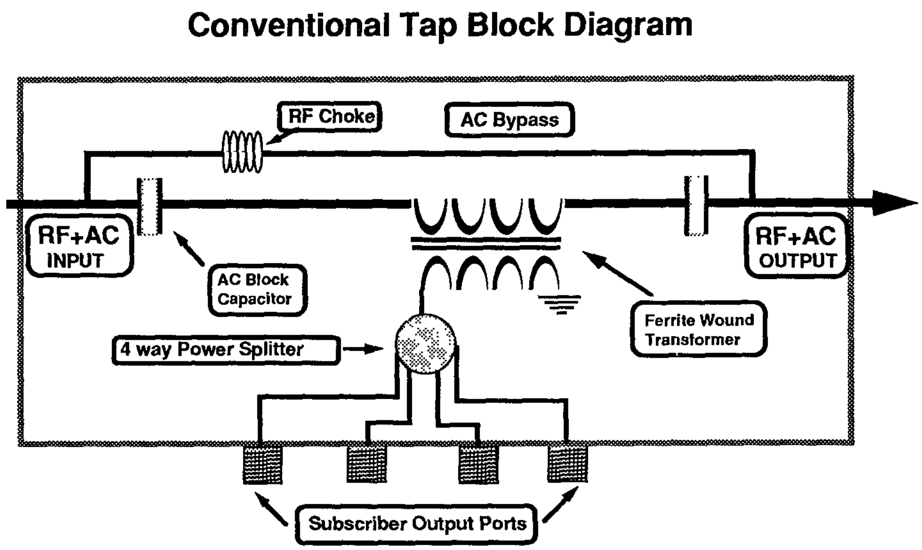

FIGURE 1

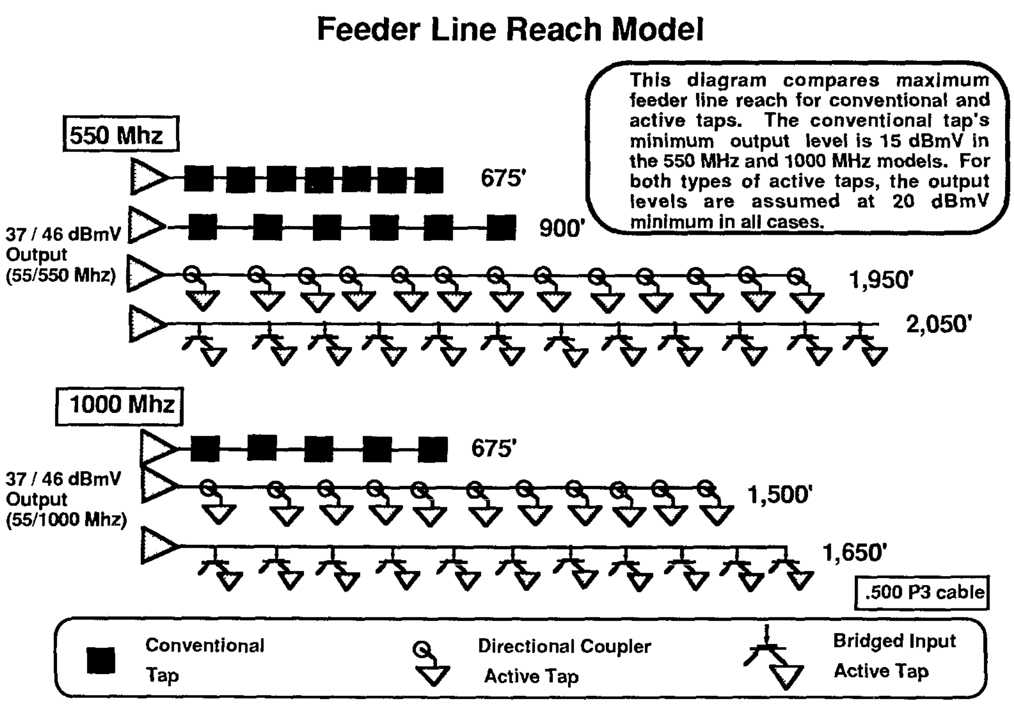

FIGURE 2

along with tap insertion loss, affect the maximum reach after an amplifier, for a given minimum tap output level. Also compared on this diagram are examples of feeder line reach when equipped with the two types of active taps that are discussed in the following sections of this document.

Directional Coupler Active Tap

A "directional coupler active tap" in a distribution line serves the same basic purpose as a conventional tap. Since the tap, as shown in Figure 3, represents distributed amplification, it allows distribution plant architecture which eliminates the use of amplifiers after the fiber optic node. Taps need not be capable of passing 60 Hz, 60 volt line power, assuming that the active tap device is powered directly from the subscribers' homes.

The interest in this type of active tap stems from its amplification ability. By having internal amplification, and by locating the RF power splitters after the gain stage, the active tap reduces the amount of signal power that must be tapped from the feeder line. Since less RF power is tapped, the tap's insertion loss is reduced. By reducing the tap insertion loss, the maximum reach of the feeder line is extended.

The directional coupler active tap shown in Figure 3 begins with a high value (low insertion loss) directional coupler feeding a plug-in attenuator pad. An optional diplex filter for low frequency return signals could be installed between the directional coupler and the plug-in pad if two-way operation were required. The purpose of the pad is to reduce the number of different values of directional couplers that would be required. Due to the discrete nature of the wound ferrite transformer, directional couplers are usually only available in three to four dB steps. In the reach model (Figure 2), ten of the thirteen directional coupled active taps used directional couplers with values of 16 dB or greater. The insertion loss of these couplers was assumed to be 0.8 dB. Since the theoretical insertion loss of a non-power passing 16 dB directional coupler is 0.1 dB, it would appear that improvements in efficiency could be expected. It is important to note that with a 0.3 dB improvement per directional coupler, there would be an additional three dB of signal after the tenth tap. This extra signal might allow an increase of 150 feet in the feeder line reach.

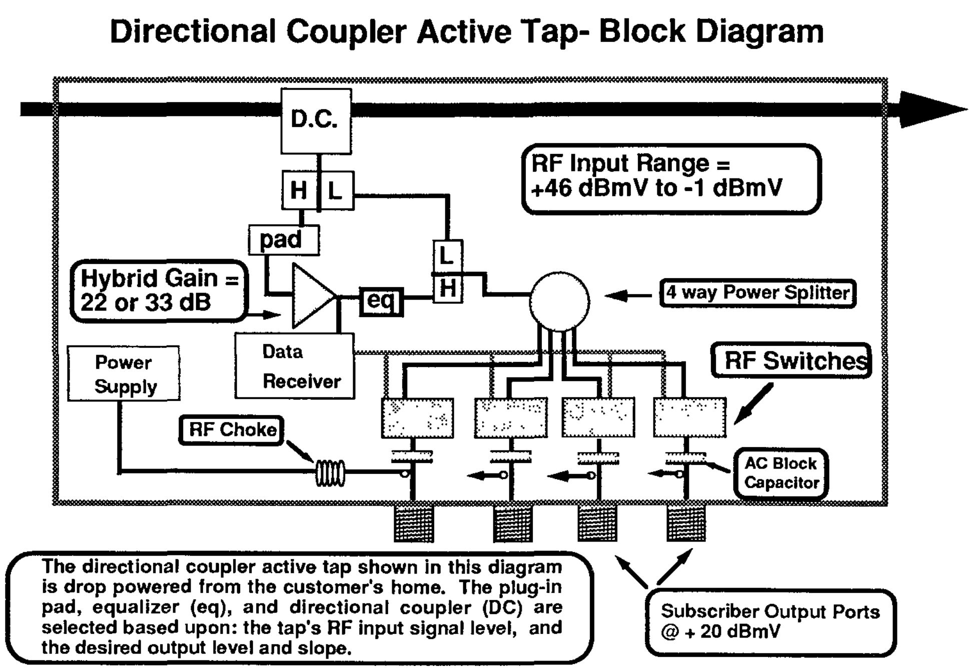

FIGURE 3

The plug-in pad is followed by the hybrid amplifier gain stage. This amplifier chip, for the 550 MHz version of the active tap, would be a push-pull type hybrid with a 5-6 dB noise figure and with 30-33 dB of gain or less depending upon carrier to noise ratio ("C/N"), number of tap output ports, and output level requirements. (The hybrid gain for seven of the thirteen directional coupler active taps shown in the 550 MHz feeder line in Figure 2 could have been 22 dB, while still providing a 20 dBmV output level on four ports.)

The nominal operating output level of the hybrid amplifier would be 30-36 dBmV. It would appear that the operating level could increase to 42 dBmV (with no tilt) before its contribution to overall system distortions would cause the end-of-line performance to degrade below established goals. Following the hybrid amplifier is a plug-in equalizer. The equalizer is selected based upon the active tap's location in the feeder line (or the degree to which the RF signals are tilted) and the amount of tilt that will be added by the average drop fed by that specific tap. Post-hybrid equalization was selected in order to protect the C/N ratio of the low band channels. Reduced C/N ratio for these channels is a common problem in single stage, high gain amplifiers with front-end equalization or slope control operating with significant output tilt.

Following the equalizer is a directional coupler or resistive tap that would feed signals to a data receiver. Past the directional coupler is a second (optional) diplex filter, which completes the upstream signal path around the amplifier. According to the number of tap output ports needed, a two, four, or eight way splitter would be installed after the diplex filter. Connected to each of the output ports would be a PIN diode switch, which is driven by an output of the data receiver, allowing the on/off switching of the signals at each tap port. The last component in the chain is a voltage blocking capacitor/powering extractor circuit that allows the tap to be powered from a subscriber's home.

The Bridged Input Active Tap

This device is similar to the previously described active tap except that the directional coupler is replaced with a field-effect transistor ("FET") with a high input impedance as a tap-off device. (See Figure 4.) As a result of its high impedance, the tap appears to have a zero dB insertion loss across the 75 ohm distribution line. This allows the placement of active taps, in unlimited quantity, along the distribution line until the distance is reached where the cable has attenuated the signals below the required threshold for the active tap.

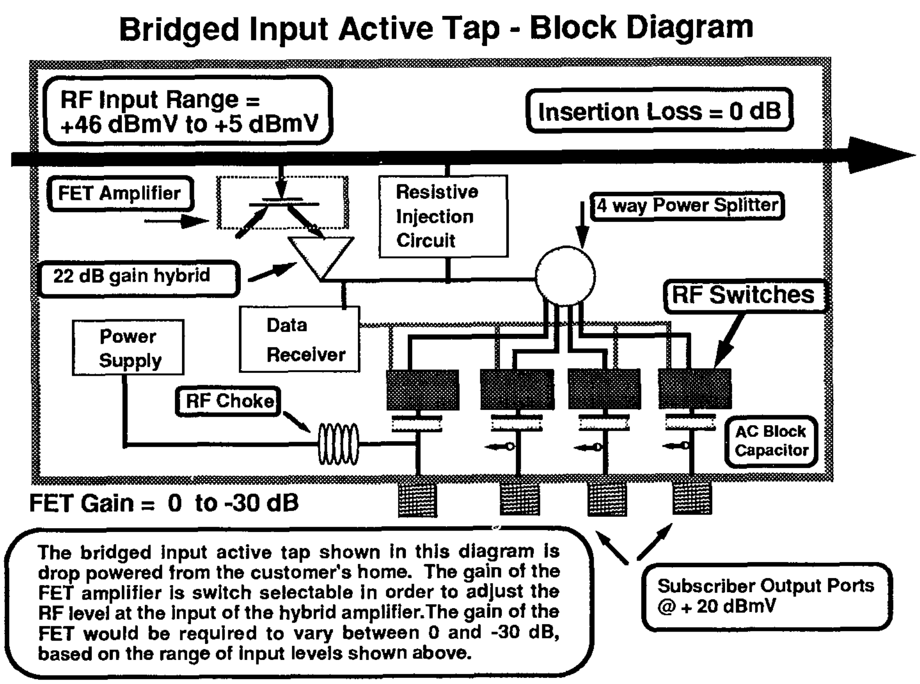

FIGURE 4

The fact that this technology allows any quantity of taps to be placed on a feeder line makes it density insensitive. The maximum feeder line reach from the optical node is therefore dependent only upon the cable type used and the cable system's bandwidth when assuming constant node output levels. The effect of density on the maximum reach of a conventionally tapped feeder line can be seen in Figure 2 (550 MHz case). By maximizing the feeder line reach, one assures that the fiber optic node will serve the largest number of homes possible.

The differences between this tap and the directional coupler active tap are found between the feeder cable center conductor and the input to the hybrid gain stage. As previously mentioned, the directional coupler in the preceding active tap is replaced with a high impedance, voltage sensitive, FET amplifier. This amplifier serves to isolate the 75 ohm hybrid from the feeder line. The gain of the FET could vary from 0 dB (unity gain) to -40 dB. The window of gain variation could be minimized, if necessary, by adding a plug-in pad located between the output of the FET and the input of the hybrid amplifier.

As compared with a directional coupler active tap, the hybrid's gain requirement for the bridged input active tap is significantly reduced. It would appear that a hybrid with 22 dB of gain would suffice in all cases. In fact, hybrid amplifier chips would not be required in the bridged input active taps that were installed within 900 feet of the node, assuming the unity gain FET could directly feed the power splitter.

The amount of gain might be controlled by selectable dip switches in the device. If the gain versus frequency response of the FET amplifier could also be controlled by switches, this might eliminate the requirement for post-hybrid equalization.

The diplex filter located between the directional coupler and the hybrid input in the earlier active tap would be replaced with a passive or active/passive "injection circuit" for return signals. This injection circuit would greatly resemble a 30-40 dB resistive tap. Some amount of gain, although less than the loss of the injector circuit, could be added in series. Since there would be minimal flat loss at the low frequencies used for upstream signals, between the active tap and the input to the fiber node, this would seem to represent a workable approach. As long as the negative gain of this circuit (when added to the gain (loss) of the FET amplifier) exceeds the positive gain of the hybrid, instability as a result of positive loop gain would be avoided.

The circuitry following the amplifier hybrid would be the same as that described for the directional coupled active tap.

Performance Requirements

The performance requirements for both types of active taps are listed in Figure 5. The C/N ratio specification for the directional coupled active tap assumes a distribution line level of +46 dBmV maximum and -1 dBmV minimum. In the case of the bridged active tap, the C/N ratio should be met with feeder line levels between +46 dBmV and +5 dBmV. These levels were selected given the expected noise performance of the bridged active tap product. Both types of active taps should provide a tap port output level of +20 dBmV minimum for two-port and four-port models. For an eight-output port active tap, the output level should be at least +18 dBmV. These output levels are specified at the active tap's maximum rated frequency. The tap should be able to introduce the range of positive slopes as specified in

RF Performance Requirements

	Optical Trunk	Optical Bridger	Active Tap	Total
C/N	50	67	51	47.5
CTB	65	64	59	53
CSO	65	70	55	53

The table indicates the assumed specifications for the optical trunk and the optical bridger. The "total" column indicates the desired minimum end of line system specifications. The column labeled "Active Tap" indicates the performance required of the active tap in order to meet total system specifications.

FIGURE 5

Drop Slope Table

Drop Length	550 MHz	1000 MHz
100'	4.1 dB	6.2 dB
125'	4.9	7.4
150'	5.8	8.6
175'	6.6	10.0
200'	7.4	11.1

The table indicates the amount of negative tilt that will be introduced by the length of RG-6 cable shown in the left column. The negative tilt, in dBs, is specified between 55 MHz and the frequency shown in the top row. Also included in the total negative tilt is the contribution of a two way splitter.

FIGURE 6

Figure 6. The CTB and CSO performance should be at least that shown in Figure 5, at the minimum tap output levels shown above, when operated with the output tilts shown in Figure 6.

On/Off Capability

In order to dramatically reduce connect/disconnect operating labor costs, the active tap should be capable of switching the downstream signal flow on and off at each of the tap output ports. This switching capability would be addressably controlled through the billing system via a data transmitter located at the headend. The data receiver, shown in the active tap diagrams, would command the switches (pin diode attenuators) to open or close depending on the instructions received from the billing system. This data receiver would be similar to that currently used in addressable converters. The frequency of its discrete data carrier could match that of the addressable converters used in the system.

TECHNICAL CHALLENGES

Powering

There are two logical means to power the active tap: from subscribers' homes and via the distribution line. There are advantages and drawbacks to each method. Distribution line powering, used today to power line amplifiers, is straightforward and relatively simple. Given the number of added active devices in relationship to the number of line amplifiers removed, additional power supply locations would be needed. The assumption has been made that, if line powered, this system would require 30% more power supplies. This is based on a power consumption for each active tap approximately equal to consumption of commercially available off-premise interdiction taps. Since the goal of the active tap is to have almost no distribution line insertion loss, the challenge would be to add the necessary AC power passing circuits without noticeably increasing insertion loss.

One solution to line powering active taps without incurring additional losses might be the use of DC powering, particularly for bridged input active taps. In this scenario, current would not be required to pass through a coupling transformer, since it would simply be carried on the transmission line through the tap. Since the gate of the bridging FET would be directly connected to the transmission line, the gate potential would be that of the DC voltage on the center conductor of the coaxial cable. Biasing would be accomplished through networks attached to the source and drain of the transistor. The amplifier hybrid and data receiver's power could be extracted from the transmission line through a carefully designed RF blocking network. With this direct connection to the coaxial cable, protection from power surges and spikes would be particularly critical.

By powering the unit via the drop from the subscriber's homes, the required number of line AC power supplies in the system would be greatly reduced. This would also eliminate the challenge of designing low insertion loss, AC power-passing circuitry. Powering from the home does, however, create its own problems. For example, if two customers are connected to the active tap, which drop (i.e., home) would actually power the tap? If only one drop actually powered the tap, and that customer disconnected his cable service, an outage for the other customer(s) fed from the tap would occur. When powering from the home, it would be useful to have power fed to the active tap on each drop. In that case, the active tap would automatically sense and use the drop with sufficient supply voltage to obtain its power. Two associated costs are the installation of the miniature power supply in the customer's home, and the long-term effects of electrolysis on the drop cable if direct current (DC) powering is used.

The RF signal level provided to the home using an active tap would insure adequate levels on more television sets than is currently provided with conventional architectures. Many cable television systems must use a drop amplifier in the home to provide adequate signal levels for more than two television sets. The active tap, both in terms of functionality and power consumption, is essentially a high quality drop amplifier mounted on the pole, followed by a splitter. The only added circuitry would be that of the data receiver and the drop on/off switches. Since the drop amplifier is often located close to the ground block (i.e., before any splitter) and may already be remotely powered from some other location in the home, one might view the drop powering issue in terms of relocating the drop amp further towards the tap.

Installation

In order to minimize installation costs, it would be useful if the active tap were packaged in an enclosure that could be mounted directly to the existing conventional tap base (for models of taps where the power-passing circuitry is part of the tap face plate). This laborsaving approach would eliminate the need to change the tap housing and associated connectors. The other primary aspect of the installation process would be to confirm, or install, the correct value of pad and equalizer (and directional coupler in the DC active tap). If drop powering is used, it would be necessary to install the small transformer and power inserter in the customer's home.

Maintenance and Reliability

An active tap must be essentially maintenance free. This implies that there should be no potentiometers to adjust output levels, etc. Long-term stability of gain, distortion, and frequency response should be engineered into the product. In the same vein, the reliability of the data receiver and its command of the on/off switches must be flawless over time and exposure to the elements. The product should be able to withstand significant electrical surges and transients as a result of lightning, power utility switching, sheath currents, etc., without damage to the hybrid amplifier, the data receiver, or the FET.

Overall reliability of an active tap product is critical. In a typical 100,000 customer cable system one would find 40,000 to 50,000 active taps. Unlike the addressable converters used today, it would not be possible for the customer to bring in a failed active tap for an over-the-counter exchange.

Dynamic Range

As previously mentioned in the section on performance requirements, the active tap must function over a wide range of input level conditions. The bridged input active tap should be able to accept at least +46 dBmV with up to 9 dB of slope. This tap should also accept input signal levels as low as +5 dBmV and 9 dB of reverse slope without degradation to the C/N performance. This requires a dynamic range of at least 41 dB. The dynamic range of the directional coupled active tap is somewhat less critical as a result of the directional coupler and selectable input pad. Nevertheless, this tap should meet target specifications with slopes from 9 dB positive to 9 dB negative, and with signal level variations of -2 to +5 dB.

ECONOMICS

Modeling Issues

There are many ways to imagine deploying active taps. The most likely would be as part of a system upgrade or rebuild. Another way might be as a result of a plant extension project. Plant modification projects, such as serving a new, unexpected apartment building, may be the case where the use of few active taps can save many thousands of capital dollars by eliminating trunk extensions that would otherwise be required.

Previous work analyzing the economics of off-premises addressable interdiction systems has provided examples of ways to deal with the economic analysis of the kinds of costs and savings represented by active tap technology.[4]

In the analysis that follows, the following factors were taken into account:

- Operational savings from the reduction of disconnect and reconnect labor
- Capital savings from the elimination of line extenders in an upgrade or rebuild
- Capital savings from the elimination of drop amplifiers

Pertinent issues that were not taken into account in the analysis include:

- Added costs to power the plant if line-powered active taps were used

- Added maintenance costs from having more active devices in the field

- Added installation costs when using drop powering from the home, which would require the installation of a power supply

- Cost savings from not having to power the plant, other than the optical trunk nodes, if drop powering were used

- Reduced service calls as a result of increased drop longevity through reduced physical disconnects and reconnects

- Marketing "lift" or increased revenue from "instant on/off" capabilities, e.g., weekend service, timely non-pay disconnects

- Reduction in future converter costs by eliminating the need for front-end pre-amps, since active taps would provide an additional 6 dB of signal at the set in most cases

- Capital cost savings by avoiding the need to replace the subscriber's drop or internal wiring as a part of system upgrade plans as a result of the high tap output level of an active tap

Economic Analysis

The starting base assumptions were as follows:

- Annual churn rate	30%
- Disconnect truck roll	$16
- Reconnect truck roll	$30
- Cost of capital/yr	10%
- Active tap unit cost	$100

Other relevant assumptions:

- 1000 subscribers
- 100 homes per mile density
- 33 taps per mile
- 3 homes per tap

Case 1

Pay-back of capital (see example, Figure 7), only as a result of truck roll savings for reconnects and disconnects:

- w/60% penetration: 5-1/3 yrs

- w/80% penetration: 3-3/4 yrs

Sample Payback Calculation

-For 60 % penetration & 30% churn
-Case 1 analysis (no capital savings)

INVESTED CAPITAL = ACTIVE TAP COST LESS CAPITAL SAVINGS

(Capital savings present only in Cases 2 & 3)
(12 Month Interest Expense calculated on prior year end invested capital less prior year's cash flow savings)

CASH FLOW SAVINGS = REDUCTION IN ANNUAL TRUCK ROLLS

(In this case, 300 disconnects @ $ 16 & 300 reconnects @ $ 30)

For Case 1, 60% penetration:
Capital Investment = $ 55,000
Annual Cash Flow Savings = $ 13,800

Year	YE Capital Balance	Cash Flow Savings
Year 0	$ 55,000	$ 0
Year 1	$ 46,700	$ 13,800
Year 2	$ 37,750	$ 13,800
Year 3	$ 27,527	$ 13,800
Year 4	$ 16,480	$ 13,800
Year 5	$ 4,328	$ 13,800
Year 6	$ -9,039	$ 13,800

FIGURE 7

Case 2

Deployment of active taps as part of a Fiber-to-the-Feeder, or Fiber Backbone system upgrade, taking into account the elimination of capital costs for line extenders:

- Saved line extenders offset 34% of capital cost

  w/60% penetration: 2-1/2 yrs

Case 3

Expanding the previous scenario with the assumption that 20% of all subscribers would require a $67 drop amp to be installed if active taps were not used:

- Saved drop amps offset 24% of capital cost (collectively with LE's - 58%)

  w/60% penetration: 1-1/2 yrs

  w/80% penetration: < 1 yr

  w/80% penetration, but only 15% churn: 1-1/2 yrs

Price Goals

The price used for an active tap in the above analysis was $100. This price was derived by starting with a commercially available high quality drop amplifier. This drop amp features a 550 MHz push-pull hybrid, passive return capability (diplex filters), signal equalization, remote power supply, and power inserter. The circuitry missing for a DC active tap would be a data receiver, the pin diode switches, an output splitter, and an input directional coupler. Packaging the product for a pole-mounted environment would also add to the total cost.

The Market Potential

As the requirements to increase channel capacity cause more systems to be upgraded or rebuilt, system operators will find it necessary in most cases to replace their existing taps. With concurrent needs to increase signal level in the home as a result of higher extra outlet penetrations, increased drop cable attenuation at higher bandwidths, or desired improvements in terminal carrier to noise ratio, the active tap offers a powerful solution to respond to these challenges.

A complementary, lower gain active tap, with power-passing capabilities, would allow the device to be used in plant modifications, as well as in upgrade or rebuild scenarios with conventional trunk and feeder architectures. The lower gain would cause the active tap to have distortion performance similar to a single trunk amplifier.

The combined offering of these products may permit the active tap to be the tap of choice for tap replacement in all cases except routine plant maintenance. In this scenario, the market potential for an active tap is significant.

With 84,000,000 homes passed in the United States, and assuming three homes per tap, one would estimate that there are 28,000,000 taps. If one assumes that all U.S. cable systems will be upgraded using active taps by the year 2000, and that the cost of an active tap is $100, the potential market would be $2.8 billion over a nine year period.

Conclusion

As we have seen, there are a number of approaches to designing active taps which may be of interest. The simplest is an active device fed with a directional coupler. The addition of an active bridging element may provide additional benefit. It is hoped that this discussion will spark additional thinking and work in these areas. There is significant market potential available for vendors who are successful in developing reliable devices with these capabilities. In addition, this technology promises substantial benefit to the cable operator, as it has the potential to dramatically reduce operating costs, improve perceived customer service levels, and facilitate the development of expanded capacity systems.

REFERENCES

1. J. A. Chiddix, D. M. Pangrac, "Fiber Backbone: A Proposal for an Evolutionary CATV Network Architecture," NCTA '88 Technical Papers, 1988.

2. J. A. Chiddix, "Fiber Backbone Trunking in Cable Television Networks: An Evolutionary Adoption of New Technology," IEEE Magazine of Lightwave Communication Systems, Vol. 1, No. 1, February 1990.

3. J. A. Chiddix, "Fibre Backbone for Cable TV Using Multi-Channel AM Video Trunking," International Journal of Digital and Analog Cabled Systems, Vol. 2, 87-93 (1989).

4. J. A. Chiddix, D. M. Pangrac, "Off-Premises Broadband Addressability: A CATV Industry Challenge," NCTA '89 Technical Papers, 1989.

BIOGRAPHIES

James A. Chiddix is Senior Vice President, Engineering & Technology, with American Television and Communications. Mr. Chiddix is responsible for corporate engineering activities as well as research and development. ATC serves 4.5 million subscribers in 33 states and is 82% owned by Time Warner Inc.

Mr. Chiddix is a Senior Member and former Director of the Society of Cable Television Engineers and is a Senior Member of the IEEE. In 1983 he received the National Cable Television Association's Engineering Award for Outstanding Achievement in Operations, reflecting, in part, his role in introducing addressable converter technology.

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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