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A DIGITAL CODING SYSTEM FOR VIDEO SIGNAL TO DISTRIBUE VIDEO MONITORING IMAGES ON COAXIAL NETWORKS

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ABSTRACT

This article describe a digital coding system for video signals called redundancy estimation ADPCM which provide a good quality for transmission of video in 6.5 Mbits or 1.6 Mbits channel.

It uses a differential pulse-code modulation (ADPCM), adaptative prediction and adaptative quantization based on redundancy estimation.

This system could be use for video monitoring on return way of cable networks, so that it could be possible to manage several video monitoring sources in the same numeric channel.

1. INTRODUCTION

Usually analog coding and frequency multiplex system are used in a full duplex coaxial networks to carry return video signals from subscribers taps to the headend.

As video monitoring signal needs 5 MHz bandwith and while return way bandwith is no more than 25 MHz, it is then difficult to transmit multiple number of large bandwith signals on a network. Management of these signals for monitoring or video conference is then complex and not useful in coaxial network.

On the other hand, it is easier to use digital signals, as it is necessary in monitoring, when one needs a large adressing capacity in the system. Nevertheless, digital techniques applied to video need commonly a high rate for transmission, which is incompatible with the small bandwith capacity of return way in coaxial cable.

In fact monitoring signals generally contained a lot of redundancy and it is possible to use highly efficient reduction techniques to decrease the transmission rate needed for those type of signals, in such a way that it could be possible to use simple and low cost receivers. ADPCM techniques seems to be very appropriate for such a purpose and that is the reason why TDF has developped a digital processing system adapted for video monitoring signals transmission.

2. THE PROCEDURE OF COMPRESSION BASED ON DPCM

Generally in the DPCM function, the actual (digital) value of a pixel is compared with a predicted value, and the difference is transmitted. Si ce typical picture has much redundancy, the uncertainty of the difference signal is much less than that of the actual signal. This can result in a low channel capacity requirement for transmitting the video signal information. However, it is difficult to obtain satisfactory picture quality at a lower transmission rate by using a conventional DPCM technique. Therefore, the development of a more efficient bit rate reduction technique is necessary.

To meet this requirement, TDF has recently developped a new coding technique called redundancy estimation ADPCM coding (RE-ADPCM). RE-ADPCM employs both adaptive prediction and adptive quantization based on redundancy estimation. Using RE-ADPCM, a digital TV coding system developed at CERLOR laboratories can transmit a TV program at a rate lower than 6.5 Mbits/sec.

ADAPTATIVE PREDICTION AND ADAPTATIVE QUANTIZATION

Adaptative prediction : In adaptative prediction coding system, several different predictors are operated simultaneously.

However, only one of them will be selected as an actual predictor by a special selection algorithm. The optimum selection method is to choose a predictor which gives the minimum prediction error for each input sample.

Adaptive quantization : Similarly, the adaptive quantization coding system employs several different quantizers simultaneously. Each quantizer has a quantizing level and a quantizing step size which are different from those of the other quantizers. One of them is adaptively selected to fit the prediction error.

It is very important to find an effective selection algorithm which can choose the predictor and quantizer to obtain low bit rate. The redundancy estimation is such an algorithm.

Redundancy estimation : The redundancy estimation algorithm is as follows :
Two successive frame, denoted by frame 1 and frame 2, are stored in memory. Each is grouped into blocks of size 8 x 8 pixels denoted by Bi. The block Bi is divided into two sub-blocks of size 8 x 4 pixels denoted respectively by Bi1 an Bi2 as shown in Fig.2.1. The group of the circled element is Bi1 and the rest of the element is Bi2. Here, i is the ordinal number of the blocks which varies from 0 to 2047.

The redundancy characteristic of block i is described by the correlation coefficient Ci,

Ci = f (Ci1(l1), Ci2(l1), Ci(l2))
where Ci1(l1) and Ci2(l1) are separatly referred to as the intraframe correlation coefficient of block Bi1 and Bi2, Ci(l2) is referred to as the interframe correlation coefficient of block Bi.

Ci1(l1)s are obtained by calculating the dispersion of the value of each pixel of block Bi1 and Bi2. Ci(l2)s are computed by comparing the value of each pixel of block Bi in frame 1 with the one of corresponding pixel in frame 2. According to the values of Ci, the predictors and quantizers are selected. The details will be described in 3.2. The correlation coefficient Ci is represented by a 4-bit word.

3. THE CODING SYSTEM AND ITS ENCODING ALGORITHM

3.1. Configuration

This system is designed on the basis of the redundancy estimation ADPCM coding described above.

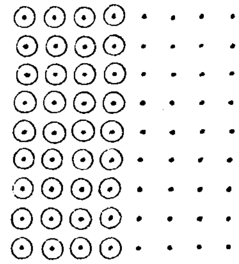

Fig. 2.1. Pixel arrangement in a block
Bi = Bi1 + Bi2

A block diagram of the system is shown in Fig. 3.1.

3.2. Preprocessing

The preprocessing circuit is shown in Fig.3.4. it is consisted of an analog low-pass video filter, a comb filter, an amplifier clamping circuit, a clock generator, an analog-to-digital (A/D) converter, and a subsampling circuit. The amplifier and clamping circuit provide isolation to the input from any following circuits. The amplifier must have a wide dynamic range without loss of linearity to handle a widely varying unclamped input video signal. The clamp circuit is used to restore the DC level to the video signal.

The amplifier drives a clock generator circuit which produces clock signal synchronized to the incoming video signal. These signals include a line clock at the horizontal line rate, a frame clock, a composite blanking, a composite sync and a sampling pulse. The exact nature of these clocks signals must be tailored to the needs of the specific realization of the circuits they drive.
The 2:1 subsampling pattern shown in fig. 3.2. is used, only the circled samples are transmitted. For decreasing the degradation caused by the subsampling, for example zigzag of oblique edge, we employ two comb filters to increase the bandwith up to theorical Nyquist limit.

So, while the sampling frequency used is close to 10 MHz, it is possible to transmit a bandwith of 4.5 MHz without aliasing distorsion due to subsampling procedure.

The comb filter responses and its position at the transmitter and receiver are illustrated in the Fig. 3.3. The subsampling frequency is chosen to be an odd multiple of one-half the horizontal line frequency so that the baseband spectrum and the replicated spectrum interleave. This is made possible by the well-known property of video spectrum of clustering the energy at multiples of the horizontal line frequency. Consequently, the video baseband spectrum and the replicated spectrum tend to interleave and can be separated by a comb filter.

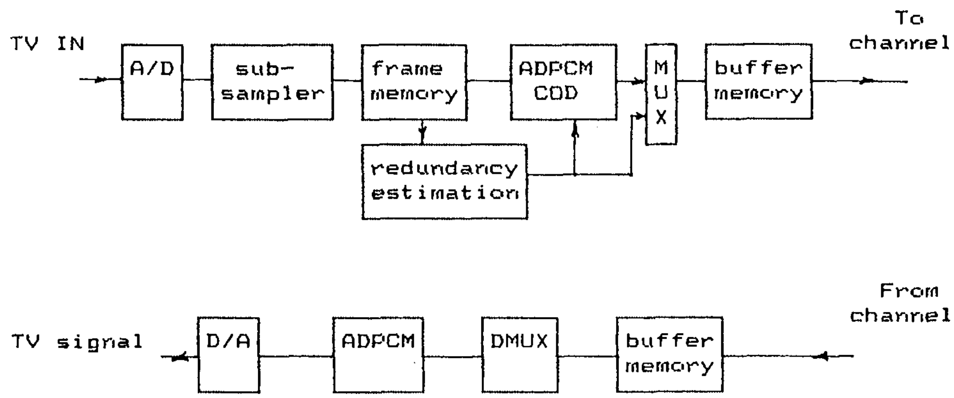

FIG. 3.1.
The configuration of the RE-ADPCM coding system

The comb filter responses and its position at the transmitter and receiver are illustrated in the Fig. 3.3. The subsampling frequency is chosen to be an odd multiple of one-half the horizontal line frequency so that the baseband spectrum and the replicated spectrum interleave. This is made possible by the well-known property of video spectrum of clustering the energy at multiples of the horizontal line frequency. Consequently, the video baseband spectrum and the replicated spectrum tend to interleave and can be separated by a comb filter.

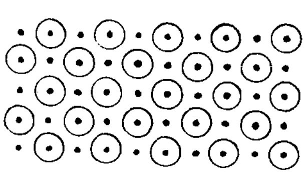

Fig. 3.2.
The subsampling pattern

The full 8-bit output of A/D converter is used. The 8 bit output provides 256 possible levels from o to 255 in decimal.

3.3. Application of the result of redundancy estimation to adaptive prediction and adaptive quantization.

As mentionned before, Ci is the correlation coefficient of block Bi, and Ci = f(Ci1(11), Ci2(11), Ci(12)). Ci1(11) and Ci2(11) are the intraframe correlation coefficient of block Bi1 and Bi2 respectively. Ci(12) is the interframe correlation coefficient of block Bi.

The information rate of block Bi can be estimated as follows :

H= (Si+M1i+M2i+I1ix32+I2ix32)/64

Si is a 4-bit word representing the correlation coefficient Ci.

Ci1(11) Ci2(12)	Quantizer	Bit/pixel	Ci(12)	Quantizer	Bit/pixel
1	0 bit	0.1875	1	0 bit	0.0625
0.5	2 bit	2.1875	0.5	2 bit	2.1875
0	4 bit	4.1875	0	4 bit	4.1875

TABLE 1
Value of Ci and corresponding information rate

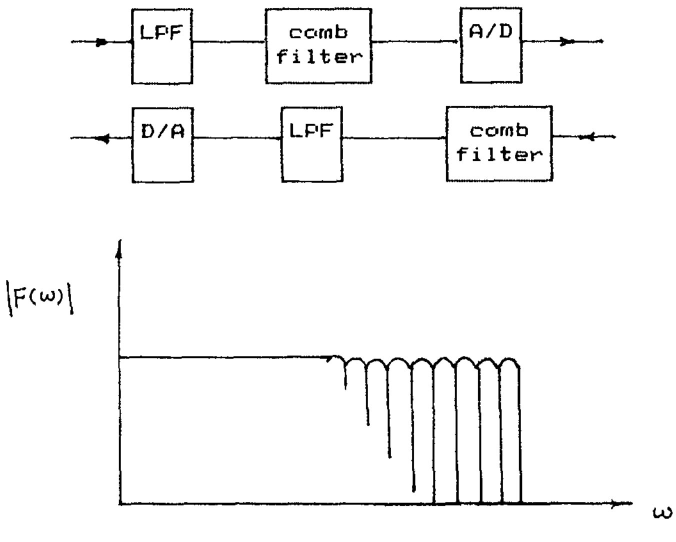

Fig.3.3.
Comb filter's position and its response characteristic

M1i and M2i are the averages of block Bi1 and block Bi2 respectively, coded on a 4 bit word lenght. I1 and I2 are the bit rates of each pixel of block Bi1 and Bi2, which vary from 0 bit to 4 bit. In the following, the selection of the quantizers is described first.

Suppose that Ci1(11), Ci2(11), and Ci(12) take on one of the values 1, 0.5 and 0. The selection of the quantizers and the corresponding bit rate is shown in table 1.

Ci1(11)=1 denotes the case when the difference between the value of each pixel of block Bi1 and the average of this block is quite low. The dispersion of this block is considered to be zero, and only the average of this block is transmitted. At the receiver, the element in the block is reproduced with the average alone. The information rate in this case is 0.1815 bits/pixel.

Ci(12)=1 means that the interframe prediction error of block Bi is so small and therefore it is assumed to be zero. Hence no bit is transmitted for this block. At the receiver, the picture element concerned is reproduced with the interframe prediction values, and the transmission rate is reduced considerably. The information rate in this case is 0.0625 bits/pixel.

The other value of Ci1, are used to quantized the intraframe prediction error on a 2-bits or 4-bits word as shown in table 1.

According to the values of Ci1(11), Ci2(11), and Ci(12), the predictors are selected for blocks Bi1 and Bi2 to obtain the low prediction error. For block Bi1, if Ci(12) > Ci1(11) the intraframe predictor is chosen, else the interframe predictor is chosen instead. For block Bi2, if Ci(12) > Ci2(11) the intraframe predictor is selected, else the interframe predictor is selected. In the case when Ci(12) = Ci1(11) and/or Ci(12) = Ci2(11), the intraframe predictor is selected for block Bi1 and/or block Bi2 to decrease the "dirty window" effect, for a moving image.

3.3. PREDICTORS AND QUANTIZER :

Predictors :

In this study, two predictors that have been proven effective are used. The first simplest predictor is the same pixel in space in the previous frame to the present pixel and is P0. The second predictor is a combination of two pixels from the previous field and present line and is (½ P1 + ½ P2) shown in fig. 3.6. x is the pixel predicted.

```
      P1
. . . . . . .
. . x . . . .
. . . . . . .
      P2
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FIG. 3.6

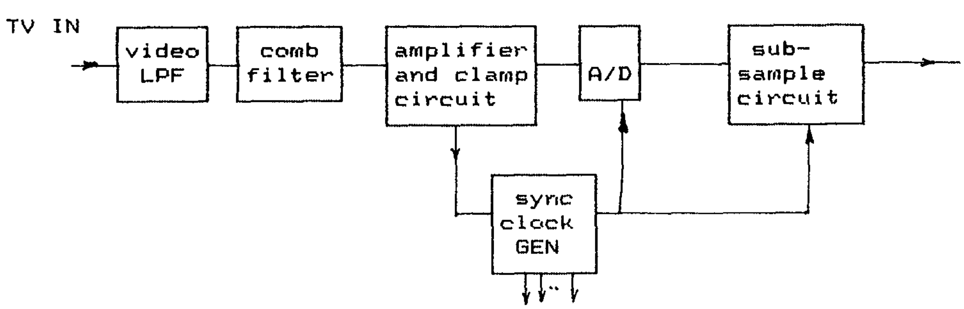

Fig.3.4.

The preprocessing synoptic

These predictors are adaptively selected on the basis of the Ci. Ci is the correlation coefficient of block Bi. The same predictor is used for all element of a block.

Quantization :

The 8 bit/pixel prediction error output data of the ADPCM function is quantized using two tapered non-linear quantizers. The transfer function of the quantizer used in this study is specified in fig. 3.7. The characteristic was chosen according to the subjective experimentation made by CCETT.

The quantizer significantly reduces the number of levels to be coded from 256 to 16 for 4-bit quantizer and from 256 to 4 for 2-bit quantizer. The same quantizer is used for all pixels of a block. The two quantizers are selected by the value of Ci, where Ci is the correlation coefficient of Bi. When the buffer occupancy variable = 0, the 4-bit quantizer is forced to be used.

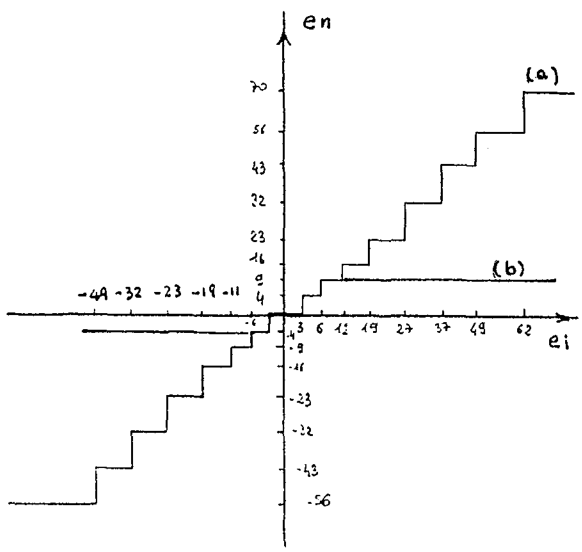

FIG. 3.7

The characteristic of the quantizer

(a) 4-bit quantizer (b) 2-bit quantizer

3.5 BUFFER :

The widely varying output data rate is smoothed in the buffer to a constant bit rate for transmission. Information is read out of the buffer at a constant rate of 1.6 - 6.5 M bits/sec. The size of the buffer is 3 Mbits.

Buffer-occupancy is determined, that is, how many bits are stored or how full the buffer is. The value of the buffer occupancy (VBO) varies linearly with the actual buffer occupancy, with 0 corresponding to an almost empty buffer and 15 corresponding to almost full. When VBO = 15, the input to the buffer stop. This condition prevents buffer overflow. The prediction error data and the value of Ci in this case are not transmitted. The reconstructed pixel values at the receiver are set equal to the predicted values. The interframe predictor is used throughout in this case. When VBO = 0 the 4-bit quantizer is used, and the actual value of each pixel is quantized using 4-bit quantizer.

3.7 TRANSMISSION FORMAT :

The transmission format is shown in fig. 3.8

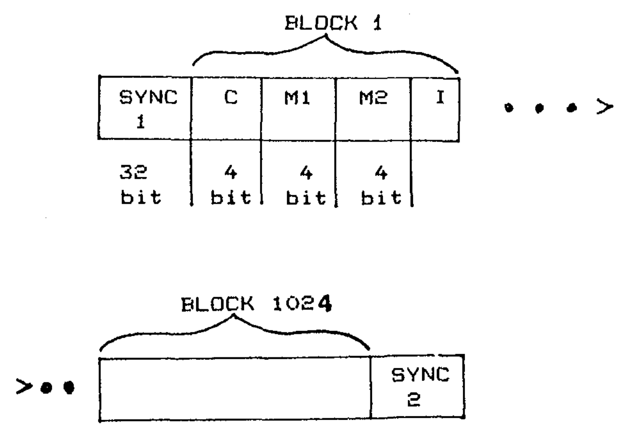

FIG. 3.8 transmission format

The first element consisting of 32 bits in length is the field synchronization signal. This synchronisation signal must be a unique pattern that does not occur in any other part. The receiver continuously looks for this pattern and, upon recognition, signals the start of another field. The sync is followed by 1024 blocks. The arrangement of information in each block is identical.

The first element in a block is the correlation coefficient of the block, which is 4 bits in length. M1 is the average of block Bi1 and the M2 is the average of the block Bi2, which have separately 4 bits word length. It contains the information of each pixel of the block, and its length varies from 0 to 255 bits.

4. THE CODING PERFORMANCE OF THE SYSTEM :

The resolution of picture in this system is 512 lines x 512 pixels.

We have chosen three typical test pictures to evaluate the coding algorithm. These pictures are shown in fig. 4.1 - 4.3. Figures 4.4 - 4.6 show the coded version of the three test pictures. The degradation caused by the encoding will be discussed below.

At 6.5 M bits/sec, i.e 1 bit / pixel

Picture impairment is not severe under this condition . Nevertheless, some slight degradation which can be observed are as follows :

- for the region in which the gray levels changes gradually, the block structure can be perceived, for example the mountain in fig. 4.1. This is due to the usage of the block average. In the flat parts and fine parts of a scene, this deterioration is not observable at all.

- some smear noise is observed when the objects move in the flat blackgroud. This is due to temporal prediction of interframe. But random noise is relatively slight.

At 1.6 M bits/sec i.e (0,25) bit / pixel

For the still picture, the coded picture quality is the same as coded at 6.5 Mbits/sec. If there is a movement associated with the image, two cases must be considered separately. In the first case, all objects in the image move. As a result, the coded image has the same movement but not as fluid. However, if only a certain small object move, for example in a wide scene with several car moving, then the degradation is not observable.

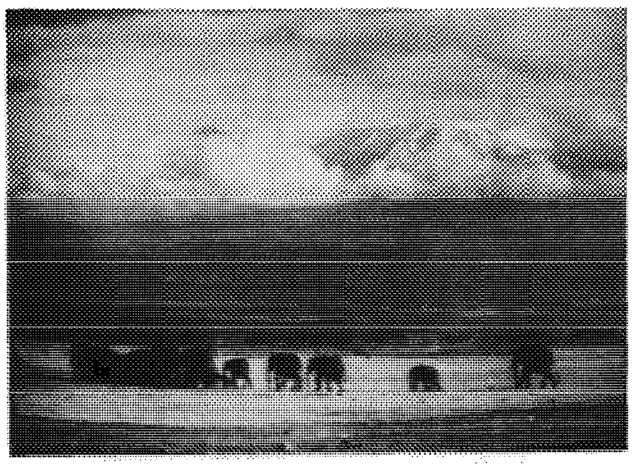

FIG. 4.1

FIG. 4.4

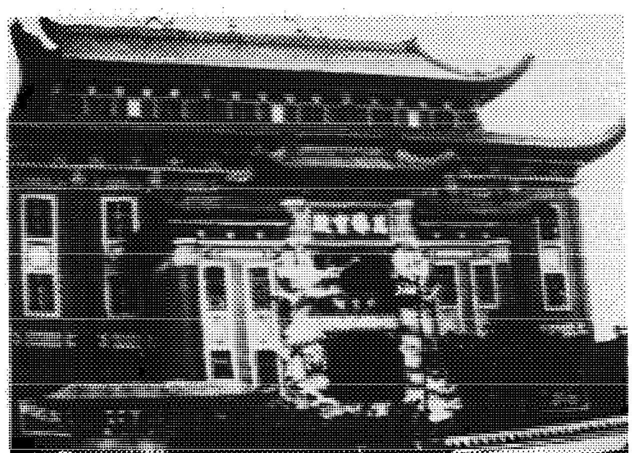

FIG. 4.2

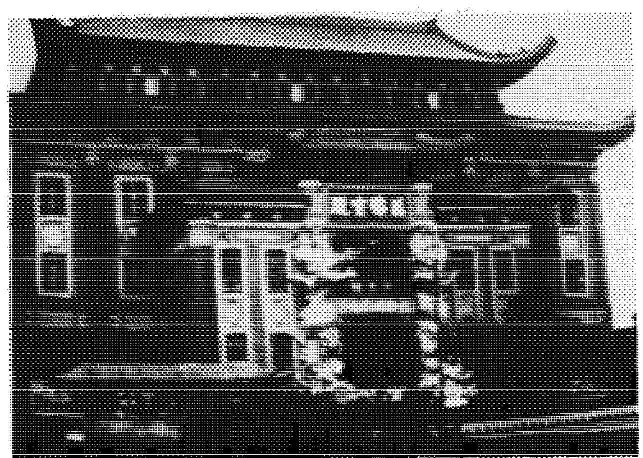

FIG. 4.5

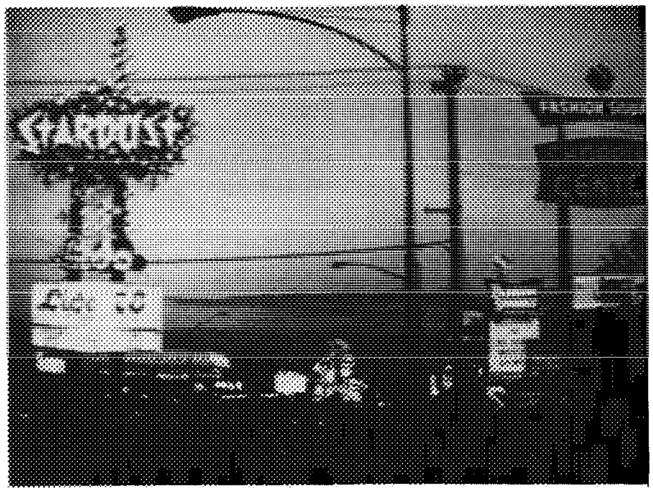

FIG. 4.3

FIG. 4.6

5. USING THE SYSTEM IN COAXIAL NETWORK

The RE.ADPCM has been developped for coaxial network to transmit monitoring video signals with a high resolution and good quality, with the following attempt.

Return way of coaxial network has a small bandwith capacity compared with the services that should be provided. In fact analogic video monitoring signals, either on splitting networks or star coaxial systems, would need the main capacity of the return way so that it could not be possible to experiment other services. In another way video monitoring only concerns a few subscribers. So, such application would not be really economically balanced.

Such discussion exists in France on new private coaxial networks, and most of cable operators think more on interactive system for suscribers in a pay per view or low data rate purpose than for monitoring.

Nevertheless the need for monitoring signals is real and no satisfactory system exists excepted with optical fibers.

It is then reasonnable to consider that the only way monitoring video signals should be implemented in a coaxial system is in the digital domain.

Another reason why analogic video signals should not be transmitted in coaxial network is the bad noise performance of coaxial networks return way. It is common to consider that noise figure of return way in coaxial networks result from an additive operation of the noise figures of each of the amplifiers installed on the network.

That is why taps should be used on the main splitting point of the networks. However the result is that it would be difficult to get a good quality for video signals transmitted in the return way on large coaxial networks. Hence a digital transmission with a high efficiency modulation could offer a better (C/N) ratio.

As it was previously explained, a 1.6 Mbits data rate transmission should be highly sufficient for implantation of digital video monitoring signals.

When a higher rate could be used, it would be then possible using the RE-ADPCM techniques to realize a flexible management system where the instantaneous rate of the channel would be adapted depending whether information has changed or not.

In this case, instantaneous rate for a digital source would be variable form 0.8 Mbits/s for still picture, to 3.2 or 6.5 Mbits/s depending on picture quality required.

On another point we have to consider whether a 4.5 MHz bandwith resolution is really necessary for video monitoring. When a lower bandwith should be sufficient, the same RE-ADPCM techniques would provide a lower rate of transmission only by decreasing the sampling frequency.

Meanwhile we did not experiment a high efficiency multiplex system. We think that an 8 Mbits channel would be sufficient to manage more than 15 monitoring video camera. The reason is that in case of video monitoring, only a small part of the picture is moving, so the RE-ADPCM transmission rate needed is very low and shouldn't be more than 1.6 Mbits/s.

6. CONCLUSION

It has been proved that a RE-ADPCM system permit a considerable bit rate reduction, without severe degradation, and that using this technique it would be possible to realize low cost receivers and encoders.

Those system seems so to be well accurate for video monitoring, but also should be employ for other applications like for example tele-teaching.

This could be a good opportunity for cable networks operator in France to put on cable networks large bandwith services as well as other data transmission services.

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A FIBER OPTIC DESIGN STUDY

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ABSTRACT

A design study was conducted to test the AM fiber backbone concept. A "typical" cable system was chosen in order to compare a fiber design to a conventional tree-and-branch architecture. Fiber's impact on end-of-line performance was investigated, and it was found that while it was possible to improve signal quality using fiber, a number of tradeoffs were involved, and the improvement was limited by distortions. A number of design and installation issues were considered, including amplifier reversal, design procedures and the location of splice points. While each system is unique, fiber designs were generally found to be more complex, and required more extensive planning for future growth. The effect of fiber systems on reliability was also studied. Fiber was determined to have little positive impact on system reliability, but it did equalize subscriber reliability across the system.

INTRODUCTION

The potential market for AM fiber optic systems in CATV applications appears to be almost limitless. Fiber's ultimate success in penetrating the cable plant will depend on several factors, including laser performance improvements and component cost reductions. However, the degree to which fiber can function as an integral part of the existing CATV distribution system will be of equal or greater importance. In order to evaluate the impact of fiber optics on the distribution plant, a detailed design study was conducted by Scientific Atlanta's Distribution Systems applications engineering group.

STUDY OBJECTIVE

The objective of the fiber design study was to test the AM fiber backbone concept, which has been proposed by James Chiddix of ATC. The concept under study involved the use of a number of nodes, each connected by fiber to the headend (or hub), and each in turn feeding cascades of four amplifiers inboth directions. The fiber backbone approach was analyzed in terms of three primary criteria: the impact on end-of-line signal quality, practical considerations for design and installation, and the effect on both system and subscriber reliability.

SCOPE OF STUDY

For the purpose of this study, a "typical" cable system was chosen in a small U.S. city, with approximately 40,000 homes passed. The system was comprised of one headend serving 530 miles of plant, which had recently been rebuilt from 300 MHz push-pull to 450 MHz using feedforward trunk and parallel hybrid feeder electronics. The majority of the plant consisted of a conventional trunk-and-feeder architecture using .750" and .875" coaxial cable. There were some coaxial supertrunks using 1.00" cable, with the longest containing 32 amplifiers in cascade.

FIBER SYSTEM DESIGN

Utilizing four amplifier maximum cascades, the fiber design was completed using a total of 51 nodes. Fiber optic cable could be overlashed to the existing aerial plant, with four main cable runs emanating from the headend. Each cable run fed a cluster of nodes, which were divided into four sectors. The first sector contained 6 nodes, with the most distant one located at 28,500 cable feet, or 8.7 kilometers from the headend. Sector 2 consisted of 5 nodes extending out to 37,310 feet (11.4 km);

sector 3, 24 nodes and 49,233 feet (15.0 km); and sector 4, 16 nodes and 70,337 feet (21.4 km). In terms of node density, 90% of the nodes fell within 16 kilometers of the headend, with 59% between 4 and 12 kilometers distant. A breakdown of node density is presented in Figure 1.

FIGURE 1

NODE DENSITY

Distance From Headend	Node Quantity	Node Percentage	Cumulative Percentage
0 - 4 km	6	11.8%	11.8%
4 - 8 km	11	21.6%	33.4%
8 - 12 km	19	37.3%	70.7%
12 - 16 km	10	19.6%	90.3%
16 - 20 km	4	7.8%	98.1%
20 - 24 km	1	1.9%	100.0%

STUDY RESULTS

1. Signal Quality

The existing coaxial system afforded good NTSC-format signal quality. The distribution plant provided a carrier-to-noise ratio (CNR) of 45.3 dB, with a 54.7 dB composite triple beat (CTB). The existing earth station provided a CNR of 50 dB, resulting in an end-of-line CNR of 44.1 dB. The desired minimum performance standards, for NTSC quality, were 43.9 dB CNR and 53.0 dB CTB. To provide picture quality comparable to Super VHS, a CNR of 46.7 dB and CTB of 54.0 dB would be needed. Achieving studio quality pictures would require minimum performance levels of 51.0 dB CNR and 56.0 dB CTB.

In the fiber design, the maximum amplifier cascade was reduced to four, which raised the coax plant performance to 58.7 dB CNR and 58.2 dB CTB. The required fiber performance for NTSC quality signals was 45.3 dB CNR and 59.9 dB CTB. In order to improve the end-of-line performance, the earth station would have to be upgraded. By increasing the earth station carrier-to-noise ratio to 52 dB, Super VHS equivalent performance could be achieved with fiber at 48.6 dB CNR and 62.3 dB CTB. To provide studio quality signals, the earth station would have to be upgraded to 55 dB CNR, with a minimum fiber performance of 54.6 dB CNR and 69.0 dB CTB. Thus, while today's AM fiber performance has brought Super VHS quality signals within reach, studio quality pictures were not yet attainable in the fiber backbone study system.

2. System Design and Installation

The design objective was to site each node to feed eight amplifiers, four in each direction. This was to be accomplished by using the existing trunk locations, and reversing four amplifiers at each node. However, reversing amplifiers was workable only in certain areas, because specific RF levels were required by sub-trunk passives and terminating bridgers. In most instances, reversed amplifiers would not provide the proper levels. Also, in some cases it was not possible to implement four amplifier cascades because of physical constraints, such as the number of amplifiers in a node cluster not being evenly divisible by four. It should be noted that in the entire system, only 15 amplifiers were eliminated out of 62 supertrunk amplifiers, and all of the 224 conventional trunk amplifiers were retained.

In a traditional coaxial installation, construction begins before the final design is complete. This is possible because once the basic parameters are established, such as cable diameter, trunk and feeder spacing, and trunk cable routing, different sections of feeder plant can be designed and built independently. This means that operators have the flexibility to essentially have construction crews follow in the footsteps of the system designers, with today's strand maps being used for tomorrow's installation.

With a fiber backbone architecture, this procedure can no longer be used. Before construction of a cable run can begin, the total number of fibers leaving the headend must be known. To establish an accurate fiber count, the total number of nodes in a given sector must be known, as well as the number of fibers per node. Future expansion must be planned as well, both in terms of new nodes and additional fibers per node. A node might require additional fibers to accomodate a return path, to increase channel capacity, or even to provide services other than traditional CATV. If system expansions or increased housing densities are anticipated, allowances must be made for additional nodes. Thus, in a fiber backbone system, construction cannot begin until the final design of a node sector is complete.

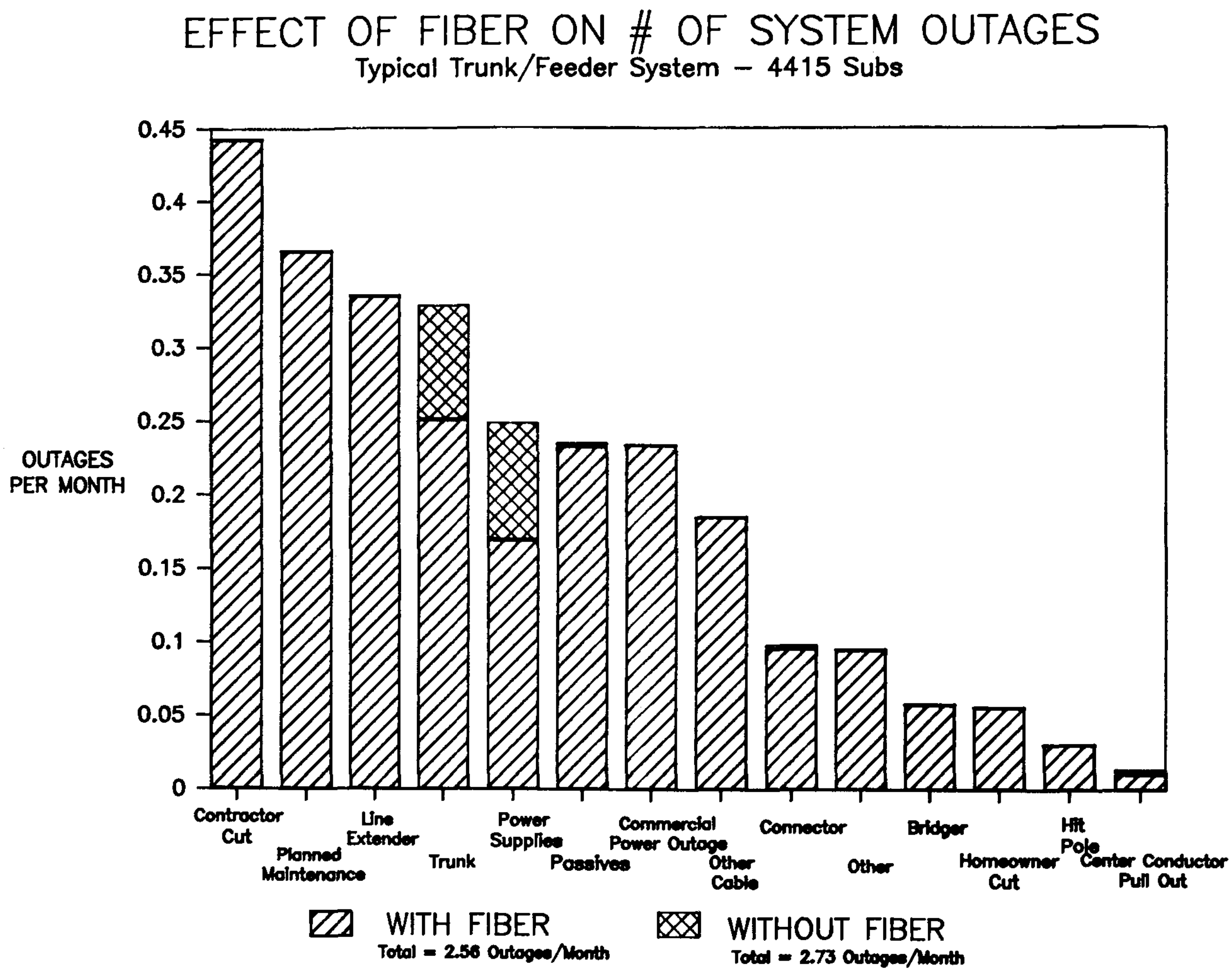
An additional level of complexity is introduced by the necessity to carefully plan the location of splice points. Since most AM fiber system designs allow only a small optical margin, splices must be kept to a minimum to reduce insertion losses and reflections. (This requirement also dictates the exclusive use of fusion splicing for installation of optical cables in CATV systems, which is beyond the scope of this study.) To minimize the number of splices required, cable span lengths must be planned carefully during the design process. Generally, the maximum reel size that can be accomodated in the construction process determines the maximum usable cable length. Up to that limit, which is usually around four kilometers, it is advisable to specify individual cable lengths as required by the design to minimize the number of splice points.

3. Reliability

The reliability impact of fiber was examined in two dimensions, system reliability and subscriber reliability. System reliability was defined as any outage that required the operator to institute repairs. Outage data was collected and sorted according to the cause of the outage and the number of outages per month. Subscriber reliability was defined as any outage that affected a particular group of subscribers. Subscribers were grouped according to their proximity to the headend, and the number of outages per month was predicted based on the cause of the outage.

To determine the effect of a fiber backbone on system reliability, the system outage data was recalculated based on the fiber design. Since no reliability data was available on AM fiber systems, it was assumed for the purpose of this study that the fiber system did not cause incremental system outages. Because the fiber design only eliminated 15 of the 286 amplifiers, there was very little positive impact on system reliability, as shown in Figure 2.

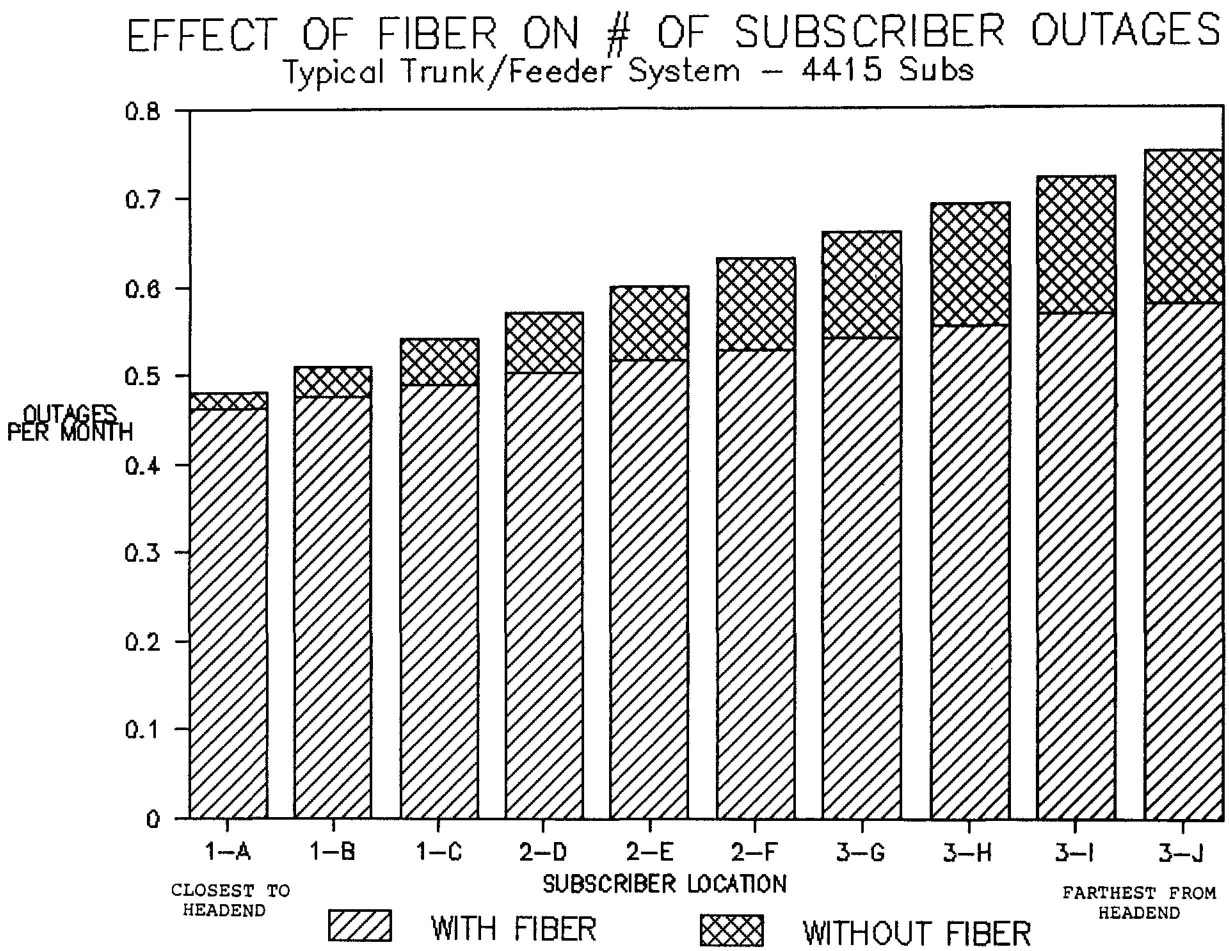
FIGURE 2



Turning to subscriber reliability, the same outage data was then applied to individual clusters of subscribers. In the conventional tree-and-branch architecture, the number of outages per subscriber increased significantly as the distance from the headend increased. This was intuitive since any system outage would be felt by all subscribers from that point on. In the fiber design, however, each subscriber was served by a maximum of four amplifiers. Subscribers located farthest from the headend would be less exposed to amplifier failure. Thus, as indicated in Figure 3, use of a fiber backbone architecture evened out reliability experience across the subscriber population.

There are two elements missing from the preceding analysis, which will require further study and additional data. First, fiber is assumed to cause no incremental system outages. This assumption is obviously not valid, and once accurate data is available, it may in fact show that fiber actually decreases overall system reliability. Second, insufficient information is available on the mean time to repair outages in a fiber system relative to a coax system. With this data and fiber system reliability information, it will be possible to determine the total system and subscriber outage time for a fiber backbone as opposed to a pure tree-and-branch architecture.

FIGURE 3



STUDY CONCLUSIONS

The fiber design study provided useful data regarding the implementation of an AM fiber backbone system in a CATV plant. In particular, the results of the study demonstrated that fiber can be used to improve end-of-line signal quality. In doing so, however, the operator has to trade off the performance of the fiber system against those of the earth station and the coaxial distribution plant. The required performance of the fiber system can be reduced by increasing the antenna size, by upgrading the distribution electronics, or by reducing the number of amplifiers in cascade from the node. These options involve obvious cost implications, which must be addressed in future studies.

There will be a number of design and installation issues in any fiber backbone implementation, many of which will be unique to that particular system. However, it appears that some general conclusions can be drawn. It will be difficult to reverse amplifiers without extensive reconstruction of feeder plant, and few amplifiers will be completely eliminated. The design of a fiber backbone will be more complicated than that of a conventional coax system. Planning for future growth must be done at the outset, and construction must wait until the final design is complete. A variety of cable lengths will be required, since care must be taken to minimize the number of splice points.

A critical factor in determining how successful fiber will be in penetrating the CATV plant will be its impact on overall reliability. In terms of system reliability, since few amplifiers are eliminated, it is apparent that fiber will not have a positive impact. Thus, from the operator's standpoint, it is important to have extremely reliable fiber electronics, and to protect fiber optic cables as much as possible. Close to the headend, where total fiber counts are relatively high, it may be advisable to bury optical cables to protect them from problems with downed lines or pole rearrangements.

In terms of subscriber reliability, fiber does tend to even out performance across the system, so that the most distant subscribers no longer experience every system outage. However, it is not yet clear what impact fiber will have on overall subscriber reliability. The two major unknowns are the reliability of the fiber system and the time required to repair fiber cable and electronics as opposed to coax. If amplifiers are not reversed, operators may choose to retain existing coaxial trunk runs in addition to a fiber backbone. This would provide them the capability to utilize the coax as a backup to the fiber system.

This study has shed light on a number of issues concerning the impact of fiber optics in the CATV distribution plant. Future studies will undoubtedly uncover other issues to be considered in implementing a fiber backbone in a CATV system. Once fiber optic technology becomes an integral part of the CATV distribution system, fiber will have a key role to play in the future of cable television.

A FLEXIBLE SPECTRUM EFFICIENT TRANSMISSION SYSTEM FOR DIGITAL CABLE AUDIO

ABSTRACT ONLY

The ICT digital music system will deliver "CD" quality stereo music to TVROs and cable subscribers using an extremely robust and bandwidth efficient transmission scheme. Nine channels of stereo audio programming will be produced at the origination facility. All programming material will be generated from compact disks or digital audio tapes. The digital outputs of the players will be fed to a digital mixing board. All fading and mixing of the audio programs will be performed in the digital domain. This avoids introducing any distortion through a digital to analog to digital conversion process.

The nine stereo audio channels will be combined with sixty-four data channels to form a single 12 Mbs stream. This bit stream is encoded onto a 4.2 Mhz baseband PAM signal which closely resembles video. This allows the DM signal to be transmitted using standard satellite, cable TV, and microwave equipment.

The sound quality of consumer "CD" digital audio was adopted as the goal for audio reproduction fidelity. Transmitting the required audio and control data in the allocated channel while preserving the integrity of both the music and the data requires the implementation of sophisticated error checking and compression schemes. Satisfying the bandwidth constraints of the overall data channel requires that the audio data be compressed by roughly 30%. The fundamental technical challenge of the system was to achieve this compression without reducing the perceived sound quality of the music delivered.

The data is compressed using a variation of block floating point compression. The variant makes use of the non-random nature of the audio source material to minimize the quantization error, which can result in an increased noise level across the audio spectrum. The expanded signal is identical to the input audio data (16 bits per left and right channel, sampled at 44.1 KHz) with the exception of the minimal quantization error. Noise shaping techniques are used to reduce the perceptibility of this induced noise. Digital filters concentrate the noise at frequencies to which the human ear is least sensitive. These measures provide a signal to noise ratio of "CD" quality for CD or DAT music sources.

The signal format will be used to transmit the DM programming to cable system headends and TVROs. TVRO owners who subscribe to the service will have a tuner which can extract any one of the audio programs and convert it to analog audio for connection to a stereo amplifier. Cable systems will have a choice of transmitting the entire DM signal over a single television channel, or to demultiplex the signal and apply the data for each of the stereo channels on a 600 Khz bandwidth channel of the cable. The narrow band approach will allow DM to be marketed on systems which do not have full channel available.

Jeff Frederiksen-Frederiksen & Shu Laboratories
Joseph L. Stern - Stern Telecommunications Corp.

A REVIEW OF THE CABLE LABS CONSUMER ELECTRONICS INTERFACE SUB-COMMITTEE

THOMAS R. JOKERST
DIRECTOR OF ENGINEERING

CONTINENTAL CABLEVISION - ILLINOIS/IOWA/MISSOURI REGION

The Consumer Electronics Interface Sub-Committee is comprised of a group of engineers from the Cable Labs Technical Advisory Committee. These individuals share the common interest of trying to develop and promote ways to solve the "cable imposed obstacles" which diminish the features and usefulness of consumer electronics hardware when connected to certain types of cable television systems. The sub-committee will temporarily function as staff to Cable Labs until such time as Cable Labs personnel are on board to manage the projects associated with consumer electronics interfacing. After staffing is complete, the sub-committee will maintain a close involvement with consultants retained by Cable Labs as well as the staff of Cable Labs involved with consumer electronics interface projects.

The Consumer Electronics Interface Sub-Committee consists of seven engineers, each of whom are the designated representatives for their company on the Cable Labs Technical Advisory Committee. The companies represented on the Consumer Electronics Interface Sub-Committee are: Cencom Cable Associates, Colony Communications, Continental Cablevision, Douglas Communications, Monmouth Cablevision Associates, Multi-Media Cablevision and TKR Cable.

The goal of this sub-committee is to create and evaluate technical options for resolving the consumer electronics interface dilemma, not to set policy on their use. Cable Labs members need as many alternatives and options as possible for the delivery of their programming products in the most consumer friendly manner. This sub-committee of Cable Labs does not intend to represent or imply that a particular method or system which is developed is to be imposed on the industry, but rather that all participating operators in Cable Labs have the option to use a particular method if it fits their individual needs both from a technical and economic viewpoint. The Consumer Electronics Interface Sub-Committee will strive to study and communicate the economic impact of its recommendations as well as the technical implications of the options that are developed.

Over the years much has been said about consumer friendliness. The issue of consumer friendly interfaces is important for the industry to resolve. If we are successful at doing this we will have happier, more satisfied customers who will not have a reason to decline a premium service or a Pay-Per-View event for example, because it interferes with their ability to use the advanced features of their TV or VCR. This is an important issue from a competitive and strategic planning perspective. Consumer friendliness may someday make the difference between retaining a customer or losing one to a competitor who has solved the interface problems (or who didn't have them to begin with!)

In an effort to further the work and build on the progress which has been made over the past several years with the Joint EIA and NCTA Engineering Committee, Cable Labs has undertaken a project known as the EIA Multiport Field Trial.

The IS-15/EIA Multiport was adopted as an official standard at the most recent EIA R-4 Decoder Sub-Committee meeting in Orlando on February 1, 1989. The EIA is in the process of finalizing the procedural paperwork for approval as a recommended standard. This is anticipated to be completed by the '89 NCTA convention.

The purpose of this project is to demonstrate and analyze the effectiveness of the EIA multiport and IS-15 decoder in solving the consumer electronics interface problems created by the use of addressable converters.

The multiport is an option which may be desirable for the cable operator to utilize to make a more "cable friendly" interface with their customers while still offering new services such as Impulse Pay-Per-View. Many operators currently utilize traps to allow for a consumer friendly interface, however, traps are not designed for use with Pay-Per-View. The multiport decoder and a multiport equipped TV receiver makes it possible to offer "transparent" security while still taking advantage of the added benefits of addressability such as Impulse Pay-Per-View, etc.

The Multiport Field Trial project consists of placing sample decoders in customers homes in carefully selected systems. In addition to gaging the customers satisfaction of the multiport decoders, the project will attempt to reveal operational efficiencies and savings for the cable operator who opts to use the multiport decoder. Additionally, the role of the TV/VCR retailer in selling and servicing multiport equipped TV sets will be evaluated.

This group will work with the EIA to finalize the IS-23 Cable/Consumer Electronics interface standard which, among other things, addresses the issue of improved TV/VCR tuner shielding to resolve direct pick up interference.

Cable Labs intends to continue the demonstration of the multiport decoder concept at industry trade shows such as NCTA, CCTA, CES, SCTE, etc. We hope to create more awareness of the concept and enlist the industry's support for the multiport decoder.

Future projects for the Cable Labs Consumer Electronics Interface Sub-Committee are currently under consideration. Areas for discussion include a technical and economic review of on-premise and off-premise security systems. This would include the interdiction-based security systems, the switched trap type systems, or any other newly announced consumer friendly security system. The technical analysis could study the issues of scrambling effectiveness, hardness of security, potential channel limitations, reliability, longevity, RFI susceptibility and IPPV operational and implementation considerations, etc. The economic analysis could study the costs involved in purchasing and implementing these systems from an initial investment standpoint while reviewing the long term economic consequences associated with system powering costs, replacement costs, etc.

We should anticipate that Cable Labs will be striving to foster even better relationships with consumer electronics manufacturers to aid in their understanding of our industry and our mutual customer's problems. These efforts will hopefully allow both industries to anticipate each other's technical trends and plan appropriately to keep the consumer supplied with products and services which are mutually complementary. Maybe both industries can adopt a somewhat familiar old slogan: "Where the quality (Consumer Electronics Interface) goes in before the name goes on! Let's all work together to make it happen.

ADVANCED SYSTEM UPGRADE REQUIREMENTS AND DESIGN

ABSTRACT ONLY

The demand for increased channel capacity within distribution systems has grown tremendously during the last two to three years. The majority of cable operators, while needing to increase their system capacity, are looking for economical ways to accomplish this task. The immediate answer to this major problem lies in salvaging the cable and amplifier locations while replacing only the active units such as amplifiers. In small channel upgrades such as going from 36 and 54 channels, this task has been fairly simple, but in making larger leaps such as from 36 to 60 and even 77 channels, operators have found that major pitfalls have to be overcome to economically salvage the investment already committed in their systems.

This paper will investigate, from a system level, how operators can optimize their systems to accomplish the larger channel upgrades. Areas that will be looked into are amplifier technologies such as the tradeoffs between push-pull, parallel hybrid, feedforward and quadrapower. Other aspects such as optimization of system tilts and interstage equalization and their effects on system upgrades will be investigated. Also minor consideration will be given to how amplifiers will interface with fiber optic systems.

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ADVANCES IN CLI FLYOVER MEASUREMENTS, THE HELICOPTER APPROACH.

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During 1988 many engineers and field technicians experienced irregularities when trying to correlate the fixed-wing airplane-type CLI FLYOVER test data with the results of their ground surveys. The inconsistencies seemed to be related to the fact that the test dipole was not operating in open space, the high cruising speed of the aircraft and the cockpit's limited visibility. During a helicopter flyover the test antenna is front-mounted, in open space. The helicopter can cruise at moderate speed and the cockpit's vertical visibility is unlimited.

The presentation compares and analyzes the radiation pattern characteristics of the test dipole under the fuselage and in front of the aircraft. This is followed by a discussion of the triangulation method to determine exact leakage source locations. Finally, a few observations are presented about digital leakage recordings and aircraft position identification methods.

RADIATION PATTERN ANALYSIS OF THE HALF-WAVELENGTH DIPOLE, MOUNTED UNDER THE FUSELAGE OF THE AIRPLANE

The "eye" of the CLI FLYOVER signal leakage detection system is the test antenna. If the "eye" is out of focus, the vision is blurred, then the test results become questionable. A radiation detecting antenna, which exhibits poor impedance match, low antenna-gain, or scattered and bi-directional radiation patterns, may produce extremely high or very low system leakage readings. It can also allocate high leakage intensities to areas which may prove to be clean later at the confirmation CLI FLYOVER.

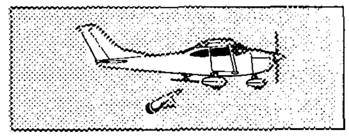

FIGURE 1

Fixed-wing airplane flyovers in the past used almost exclusively a half-wavelength GAMMA-MATCHED dipole mounted coaxially under the airplane's fuselage.(Figure 1). An open space half-wavelength dipole exhibits the well-known bi-directional 8-shape radiation pattern in the horizontal plane. (Figure 2).

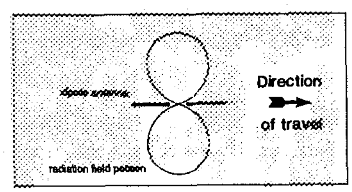

FIGURE 2

The coaxially installed gamma-matched dipole has a deep radiation pattern null in the travel direction while exhibiting two asymmetrical maxima perpendicular to the main axis of the airplane. The critical issue is not how the dipole was installed under the airplane. Rather, it is the fact that the dipole is **not operating** in open space. The fuselage of the airplane becomes an essential part of the antenna system, affecting the vertical radiation pattern of the dipole.

In the *first approximation* this is the classical example of a dipole, located parallel above the ground. In the case of the airplane, it is a dipole, mounted below a large and flat metal surface at a distance *d.*

Radiation pattern conditions can be analyzed by the application of a basic electromagnetic principal, the IMAGE THEORY. Figure 3 illustrates the position of the real antenna and its image on both sides of the metal plate. The current of the "image" dipole is equal in amplitude but 180° out-of phase, under ideal conditions.

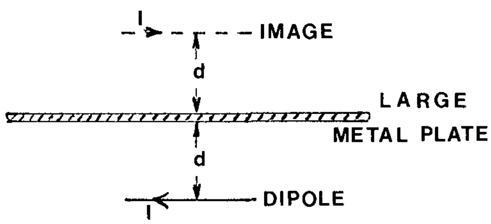

FIGURE 3

At a distant point in the far-field zone, the vertical component of the field intensity will be composed of the direct and *reflected signals*, as described by the following equation:

$$\mathcal{E}_V = 2E \sin(d_r \sin\alpha)$$

E = Maximum field intensity of the half-wavelength dipole in open space, in the = 90° direction.

$d_r = \frac{2\pi}{\lambda} d$, distance in electrical angles.

2E stipulates total reflection from the perfectly conducting metal surface.

Figure 4 illustrates vertical plane radiation patterns of a half-wavelength dipole, located at various distances under a perfectly conducting large metal plate.

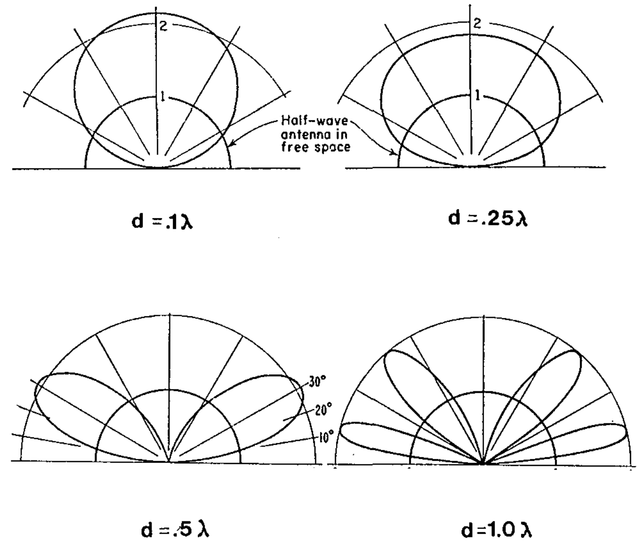

FIGURE 4

Below a distance of 0.25λ the pattern remains a single wide lobe. At 0.33λ it becomes three-directional. Then, between d = 0.5λ to d = 1.0λ the pattern breaks up rapidly into multi-lobes.

Thus conditions under the airplane are a function of the dipole's distance from the fuselage. If the distance is less than 0.33λ, two important antenna parameters, the impedance match and the antenna gain suffer. At 0.33λ or greater distances the radiation pattern breaks up into bi-directional multi-lobe modes, a highly undesirable situation.

Actually, radiation pattern conditions are not even close to those ideal assumptions of the simplistic image theory.

* The conductivity of the fuselage is less than extremely high.
* The fuselage is not flat, nor infinitely large.
* The surface of the fuselage cannot be considered perfectly smooth.
* The space above the dipole is not a single metal sheet. It is loaded with additional material of metal and non-metal substance.

Therefore, the current of the "image" dipole will not be equal with the amplitude of the real dipole, nor will be the "image" current exactly 180° out-of-phase. Also, the leakage signals emanating from the system below may not be perfectly horizontally polarized, or horizontally polarized at all, resulting in extremely complicated and uncontrolled phase cancellations. The real world radiation patterns of the dipole will be transformed into configurations similar to those of Figure 5.

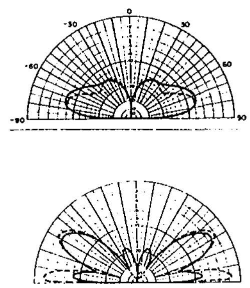

FIGURE 5

THE MULTI-ELEMENT TEST ANTENNA APPROACH, MAST-MOUNTED IN FRONT OF THE HELICOPTER

If the test dipole mounted below the fixed-wing airplane's fuselage develops impedance match, antenna gain, and radiation pattern difficulties why not avoid these problems by moving the test antenna away from the aircraft? Figure 6 shows the solution. The test antenna is mounted in *front* of the helicopter, operating in OPEN SPACE. Neither the impedance match and antenna gain, nor the radiation pattern characteristics will be affected by the cockpit and fuselage of the helicopter. They are located way behind the antenna.

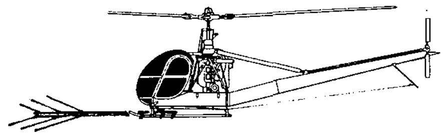

FIGURE 6

The application of a multi-element log-periodic antenna (Figure 7) eliminates an additional difficulty observed in the case of the test dipole: the bi-directional characteristics of the radiation pattern.

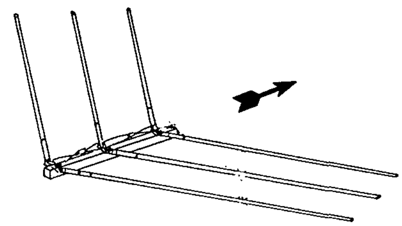

FIGURE 7

A three-element log-periodic antenna, tuned to the relatively narrow 108 to 136 MHz frequency range, has 3 dB gain over a half-wavelength dipole, a *single* 60° wide main-beam toward the front, (70° wide in the vertical plane), and a very favorable 20 dB front/back ratio. (Figure 8).

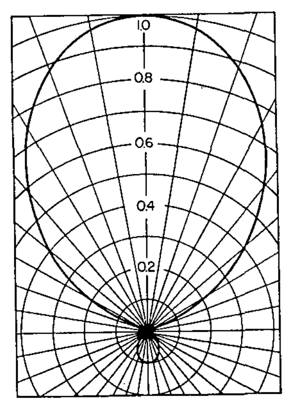

FIGURE 8

The high front/back ratio of the log-periodic multi-element antenna mast-mounted *in front* of the helicopter reduces the coupling between the main-frame and the antenna to a minimum. Consequently, the radiation pattern of the multi-element test antenna remains practically intact.

The helicopter mounted test antenna will "sweep" the cable plant with a single beam, about 60° wide. Also, due to the 3 dB antenna gain, (-3 dB points at 30°) the antenna will deliver the same signal levels obtained from the half-wavelength dipole in open space.

LEAKAGE SOURCE IDENTIFICATION BY THE TRIANGULATION METHOD

The helicopter-type CLI FLYOVER testing has another tremendous advantage: its superior mobility and ability to pinpoint the exact location of the leakage source.

The helicopter can fly at a reduced speed and turn around quickly in a small circle. Then, with the aid of the front-mounted, highly directive antenna the source of leakage can be found by the triangulation method. Follow the flight of the helicopter of Figure 9. The helicopter approaches a strong leakage area (#1). The pilot is directed to make a right turn (#2). Finally, a new right turn verifies the leakage source (#3). This is in sharp contrast to the high cruising speed and limited turn-around capability of the fixed-wing airplane, compounded with the symmetrical and bi-directional pattern of the dipole. Leakage source identification, which is impractical with the fixed-wing airplane, is a high value benefit of the helicopter-type CLI FLYOVER, assisting CATV operators and field engineers in the fast and efficient identification of major system leakages.

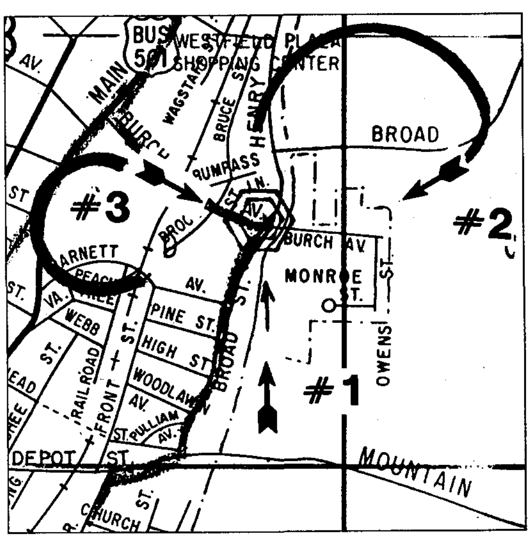

FIGURE 9

VERTICAL VISIBILITY & CRUISING SPEED CONSIDERATIONS

From the fixed-wing airplane's cockpit the view is limited to one side and to the front of the airplane. (Figure 10). Neither the pilot nor the testing engineer is in a position to follow and observe the terrain below. Therefore, the fixed-wing airplane must fly parallel patterns to cover the entire area, using LORAN for position identification and recording purposes.

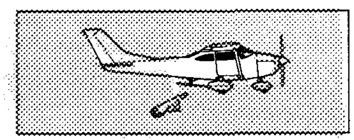

FIGURE 10

Then, due to the high speed of the aircraft which prevents the use of analog instrumentation, the signals from the receiver must be digitally processed, feeding the results into a chart recorder.

This enthusiastic reliance on the LORAN-C and a chart recorder has two serious flaws.

The first one, discussed briefly before, is the fact that the LORAN indicates the aircraft's vertical position over the ground, while the test dipole may receive the strongest leakage from completely different directions. The discrepancy, due to the bi-directional and scattered nature of the radiation pattern can be as high as 2500'.

The second problem area: the chart recorder's inability to differentiate between desired and undesired signals. For a chart recorder every received transmission is a potential leakage problem. The most frequently received spurious signals include harmonics of broadcast stations, two-way radios or CB transmitters, and interference from high voltage transmission lines, welding shops, lightning or other statics.

Observe the excellent vertical visibility from the helicopter's cockpit (Fig. 11). Helped by the low cruising speed of the rotary wing airplane, the test engineer can follow the trunk and distribution lines and mark the exigencies of the cable plant on the system map *directly* in front of him.

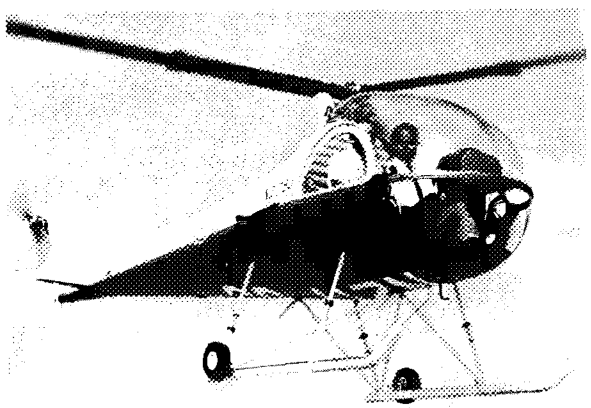

FIGURE 11

The system will be surveyed exactly the same way as conducted during the ground survey. The test results will be marked on a *real system map*, easily understood and appreciated by the system engineer in charge of system leakage integrity.

Major potential aeronautical leakage sources such as high-rise buildings, hotels, and skyscrapers which cannot be checked from the street level are pinpointed from the low cruising speed helicopter. The leakage concentrations can be then identified by the triangulation method.

Last but not least, the helicopter CLI FLYOVER engineer can reliably read the meter of his ANALOG COMMUNICATIONS RECEIVER in the slow moving aircraft. Only those meter movements will be recorded which were accompanied by the 1000 Hz tone modulation of the test signal. All other meter movements will be discounted because they did not emanate from the cable plant.

A receiving antenna, 1500 feet above ground, can pick-up an enormous variety of spurious transmissions which, if taken by their face value, can skew the test results.

ADVANCES IN TECHNOLOGY AND NEW INSTALLATION CONFIGURATIONS SUGGEST MECHANICAL MODIFICATIONS IN CATV AMPLIFIER HOUSINGS

Scot Milne

Magnavox CATV Systems Company

ABSTRACT

While advances in amplifier technology (Power Doubling, Feedforward) have greatly altered the components a housing must contain, and new installation configurations (pedestal, vault) have altered the environment in which a housing must function, the housing itself has changed very little. Results from computer modeling suggest alternatives for housing design to meet these more demanding requirements.

INTRODUCTION

An amplifier housing performs several functions for the circuitry inside the mainstation or line extender. Four of the most important of these functions are:

- Establishing mechanical and electrical connections for the trunk and/or distribution cables.
- Acting as a shield from electromagnetic interference (EMI).
- Offering protection from the weather.
- Providing a heat sink for active devices.

The magnitude of these housing tasks has expanded over several years, primarily from changes in two areas: the amplifier circuitry technology and the housing environment. System bandwidth has spread in measured steps from 5-270MHz to 5-600MHz, placing new demands on the cable connection, the EMI shield, and the heat sink elements of the housing. The environment that the amplifier is exposed to has also changed. Installation in enclosures is now common for domestic systems. International business has increased significantly, with a commensurate broadening of weather extremes and installation configurations. These new environments create challenges to the cable connection and heat sink elements of the housing.

Responding to these changes by modifying the die-cast amplifier housing can become an expensive proposition in terms of engineering effort and tooling investment. A totally new design represents a major commitment of resources. A development program of either scope must start with a set of clearly defined objectives for housing performance for all four of the functions mentioned above. Recommendations for these objectives, or design goals, are introduced in the next few sections of this paper. The process of how the goals should be met follows each goal.

PERFORMANCE BASELINE

The housings currently being produced in the cable industry provide readily available vehicles for measuring the present level of performance, or baseline, of the four functions. These levels should be compared with the design goals to reveal which areas require modification and the magnitude of the modification from the baseline.

CABLE CONNECTIONS

The goal for this function is to provide maintenance-friendly trunk and distribution ports that are transparent to the forward and return signals at frequencies up to 1GHz. It is important to recognize that the historical trend toward greater bandwidth has not lost any momentum in the past few years. HDTV, on-demand services, and data will sustain this expansion. Due to the investment required by the manufacturer (in production tooling) and the system operator (in physical plant), the housing design must be prepared to embrace circuitry delivering these extended bandwidths for product longevity. Initial designs for circuitry are exercised on computers using models that simulate the response of the port as well as the seizure mechanism.

However, the dimensional parameters and material characteristics of the port components should be selected only after exhaustive bench testing to be certain that the difficult-to-characterize anomalies have been captured.

Varying the diameter of the center conductor of the cable from the optimum value introduces impedance changes that reduce return-loss performance. Designing for a pin-type connector permits the housing to achieve a consistently high level of performance in both trunk and distribution ports regardless of cable size. The standard length pin connector (1.60") should fit, without modification, into the housing. Ideally, the "nose" of the port should protrude from the wall of the housing far enough to allow heat-shrink tubing to be attached from the cable, past the insert, to the casting for maximum weather protection. The insert must provide stronger threads than the die-cast housing without creating galvanic incompatibility. A hexagonal or square cross-section would facilitate assembly at the factory as well as an opportunity to counteract, with a second wrench, the torque required to install or remove connectors. Options for cable access to the seizure mechanisms could be more flexible, reducing the contortions necessary for installation of housing and cables in the cramped quarters of enclosures.

EMI/RFI SHIELD

The goal for this function is to offer protection from interference or radiation from higher frequencies (1GHz) at existing signal levels. Even in rebuilds, signal levels have not increased significantly in the past few years. An approximate level of the seal's shielding performance can be established by testing prototype gaskets in simulated housing grooves. However, this limited evaluation does not confirm that the gasket construction can resist permanent deformation and possible loss of performance after the stresses of temperature and compression cycling. Long-term reliability of the seal's shielding effect is crucial for compliance with the cumulative leakage index (CLI) and radiation limits set by the FCC.

When the mainstation base and cover are designed to conform to American Die Casting Institute guidelines, the space between flanges could vary from 0.000" to 0.109". Machining the surfaces of these flanges would significantly improve the flatness. Although widely used in producing military equipment, this approach is too expensive for cable TV products. The gasket must be designed to accomodate the flatness and surface finish variations created by the die-casting process.

WEATHER PROTECTION

The goal for this function is to provide a seal that protects the circuitry from water ingress over repeated temperature, atmospheric, and compression cycling, yet requires a minimal clamping force for easy access. The extremes of temperature and atmospheric pressure can create positive (9.5 p.s.i.g.) or negative pressure (-5 p.s.i.g.) within the housing. A soft rubber gasket would initially provide an adequate seal under these conditions. Unfortunately, a low durometer rubber will typically take a "set" when subjected to a clamping force over time. The "set" is partly a densification and partly a deformation of the gasket material. The net result is a reduction in clamping pressure, possibly leading to a leaky housing during the next summer rainstorm. Increasing the torque on the clamping bolts offsets the reduction in pressure, but may lead to further deformation.

The gasket material must also withstand temperature and humidity extremes as well as chemical attack by ozone and industrial pollutants. Extensive evaluation, starting with prototype gaskets in simulated housing grooves, is essential to establishing the long-term reliability of the seal.

HEATSINK

The goal for this function is to maximize heat-sinking of active components regardless of horizontal or vertical orientation of the housing. This performance will be limited by various restrictions on the weight, size, appearance, and cost of the die casting, as well as upward and downward product compatibility.

Advances in amplifier technology have reduced the distortions required for expanded frequency response; however, these same advances have increased the heat, or thermal, load on the housing. A mainstation loaded with a Feedforward trunk module, a Power Doubling bridger module, a return module, a complete control module with status monitoring, and a switcher-type power supply must dissipate two times more heat (36.5 more watts) than a station with push-pull technology.

Amplifier housings installed at ground level in enclosures shield the housing from the direct rays of the sun, but restrict the free flow of cooling air. Use of a pedestal enclosure forces the casting into a vertical orientation. The fins on the housing are perpendicular to the air flow, which reduces their ability to dissipate heat. This situation was simulated by attaching heaters to the back side of a 6" x 6" piece of heat sink extrusion with 3/4" tall fins. At a surface temperature of 25° Centigrade (C) over ambient, the material dissipated 13.5 watts with the fins in a vertical position. Rotated by 90 degrees, it could only handle 7.6 watts at 25° C over ambient. Another piece of extrusion was cut to the same dimensions with the fins at a 45 degree angle. This piece dissipated 13.5 watts at 25° C over ambient in <u>both</u> 0 and 90 degree positions.

Heavily instrumented housings were installed in a simulated field situation in aerial, pedestal, and vault configurations. A data recorder provided temperature information for the construction of a mathematical model of the housing and modules. The model permitted in-depth study of various design options in any environment without the expense or time required to build and evaluate prototypes. This model quantified the thermal impact of various alternatives for housing design. A thicker housing wall reduces the conduction resistance between the heat source, amplifier modules, and the fins. Deeper fins, optimally spaced, assist convective heat transfer from the fin surface to the surrounding air by increasing the effective surface area of the housing.

Applying a finish to the housing surface will alter the radiation coefficient. This may improve or impede heat flow -- depending on the color and type of coating. As mentioned earlier, angling the fins allows free air flow in either vertical or horizontal configurations and reduces the temperature rise of the circuitry and housing in pedestal applications.

CONCLUSION

A housing designed for the 1990's must be ready for expanded bandwidth, extreme weather conditions, EMI at higher frequencies, and various installation configurations. These demands can be met through innovative design concepts, computer modeling, and exhaustive evaluation.

AM FIBER OPTIC TRUNKS - A NOISE AND DISTORTION ANALYSIS

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ABSTRACT

Nonlinear optical effects in single mode fiber can limit the amount of power that can be coupled into a fiber. The threshold powers for these effects are calculated. The maximum practical optical modulation depth is limited by clipping. This clipping occurs when the peaks of the RF signal drive the laser below threshold. The maximum modulation depth is determined for both standard and HRC systems. The results are used to project system carrier to noise performance.

INTRODUCTION

There are two major issues in distributed feedback laser performance (DFB) for AM-VSB systems. These are output power and linearity. The use of DFB lasers for analog applications is a relatively new idea, having been taken seriously by optoelectronic device manufacturers for only the last year or so. Development of DFB lasers for analog applications is continuing rapidly and there is no indication that the pace is slowing. The question which this paper will attempt to answer is: what are the limits on link carrier to noise performance if highly linear DFB lasers can be developed? In order to answer this question this paper will address the following issues:

1) Output power - Assuming that the output power of laser diode chips will continue to increase over time, what then are the limitations on laser coupled output power due to nonlinear optical effects in single mode fiber.

2) Optical modulation depth - Assuming a perfectly linear laser could be developed, what would be the limitations on optical modulation depth as a function of channel loading. The limitation is due to distortion introduced when the peaks of the signal drive the laser below threshold. This will be examined for both standard and HRC systems.

The power and modulation depth limitations will be used to project realizable system performance as a function of channel loading.

OUTPUT POWER LIMITATIONS

Stimulated Brillouin Scattering

Stimulated Brillouin Scattering (SBS) is a nonlinear optical phenomenon which can limit the amount of power coupled into an optical fiber. When the SBS threshold is reached the energy in the forward wave (signal) couples to a wave at a slightly longer wavelength traveling in the opposite direction in the fiber. The result is that the forward wave is severely attenuated. The threshold depends strongly on the source linewidth because the spontaneous Brillouin bandwidth is less than 100 MHz.[1] As the source linewidth increases beyond 100 MHz, the SBS threshold will increase. For narrow linewidth (<100 MHz) sources the SBS threshold can be calculated using the following equations.[2]

$$P_{TH} = \frac{21 \cdot A_e \cdot K}{g_B \cdot L_e} \quad (1)$$

$$L_e = \frac{1 - e^{-\alpha L}}{\alpha} \quad (2)$$

A_e = effective core area of the fiber
K = polarization factor $(1 \le K \le 2)$
g_B = peak Brillouin gain coefficient (4.6E-11 m/W)
L_e = effective interaction length (m)
α = fiber loss (m^{-1})
L = fiber length (m)

Figure 1 shows the SBS threshold for single mode fiber as a function of fiber length for two different attenuation rates corresponding to 1310 nm and 1550 nm operation. It is assumed that the effective core diameter is 11.5 μm and that K = 2 (complete polarization scrambling). This shows that for long links the maximum input power at 1310 nm is about 10 mW.

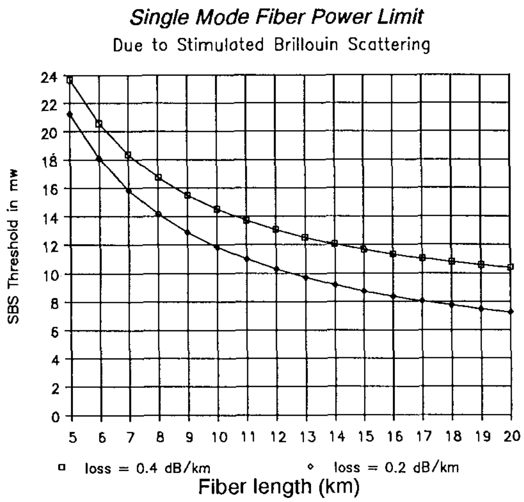


Figure 1

The actual SBS threshold for an AM-VSB system depends on the type of laser that is used and on the method of generating the optical signal. There are two methods commonly used to generate optical signals in AM-VSB systems. They are direct modulation and external modulation (see Figure 2). In direct modulation systems the source laser is usually a distributed feedback (DFB) laser. The linewidth of DFB lasers is typically less than 100 MHz with no modulation applied.[3] However, when the laser is intensity modulated the carrier densities in the active region are modulated. This causes the refractive index of the material to vary which in turn causes the laser output frequency (or wavelength) to vary. This phenomena is referred to as chirp. The amount of chirp a laser exhibits is a function of chip design but it is typically more than 5 GHz.[4,5] This effectively increases the source linewidth to many GHz when the laser is modulated. This means that for systems employing direct modulation the SBS threshold will be considerably higher than that shown in Figure 1.

In systems that use external modulators light is coupled from the laser source to an external electro-optic modulator. The modulating signal is applied to the external modulator. The laser is operated in a CW mode which means the source linewidth can be narrower than the spontaneous Brillouin bandwidth depending on the type of source chosen.

This means that the SBS threshold could be as low as the levels shown in Figure 1.

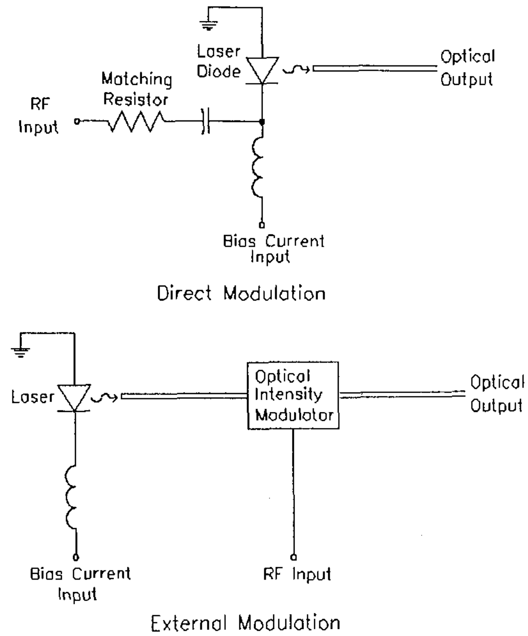


Figure 2

Stimulated Raman Scattering

Raman scattering can be thought of as the modulation of light by molecular vibration. The process generates Stokes light at wavelengths both shorter and longer than the pump wavelength. The Stokes light travels in both the forward and reverse directions. In long optical fibers the signal wave will act as the pump source for the Raman gain. In glass fibers the Raman gain-bandwidth is very wide so laser chirp should not affect the threshold. As with stimulated Brillouin scattering, when the stimulated Raman scattering threshold is reached it causes the fiber attenuation to become nonlinear. The SRS threshold can be computed using the following equation.[6,7]

$$P_{Th} = \frac{16 \cdot A_e \cdot K}{G_R \cdot L_e} \qquad (3)$$

A_e, K and L_e are as defined above

G_R = peak Raman gain coefficient (1.38E-13 m/W at 1.3μm)

Using this equation the SRS threshold is on the order of 8 Watts at 1310 nm for long fibers. From this result it is apparent that stimulated Raman scattering will not be a practical limitation on AM fiber systems.

OPTICAL MODULATION DEPTH LIMITATIONS

Optical modulation depth directly affects both carrier to noise and distortion performance. Generally, higher modulation depths will improve C/N and degrade distortion performance. The exact relationship between modulation depth and distortion depends on the characteristics of the particular laser. There is, however, an upper limit on optical modulation depth for even perfectly linear lasers. This upper limit on modulation depth per carrier depends on the number of carriers and the frequency plan that is being used. In order to mathematically determine the maximum useful modulation depth it is important to understand the characteristics of the composite RF signal. The most important characteristic of the signal is whether the individual carriers are correlated with each other. In a standard system with free running carriers, the carriers are uncorrelated. In an HRC system the carriers are all phase locked to a common reference, which means they are correlated.

In the case where the carriers are uncorrelated the characteristics of the composite CATV signal can be determined using statistical methods.

Let each carrier be represented by:

$$x_i(t) = m_i(t)\cdot\cos(w_i t + \theta_i) \qquad (4)$$

where:
w_i is the carrier frequency
θ_i is the carrier phase
$m_i(t)$ is the modulating signal

The composite signal which modulates the laser can be represented by:

$$y(t) = \sum_i x_i(t) \qquad (5)$$

The laser drive current is given by:

$$I(T) = I_{th} + I_{bias}\cdot[1+y(t)] \qquad (6)$$

Where I_{th} is the laser threshold current and I_{bias} is the nominal bias current above threshold. When the signal, $y(t)$, exceeds the value -1 the laser is driven below threshold. The result of this is that the signal is clipped, causing distortion.

It is useful to examine the statistical distribution of the composite signal. In order to simplify the analysis, let $m_i(t)$ = constant (unmodulated carriers). Define a random variable X_i, formed by sampling the signal $x_i(t)$ at time T as follows:

$$X_i = x_i(T) \qquad (7)$$

$$X_i = m_i\cdot\cos(w_i T+\theta_i) \qquad (8)$$

where θ_i is uniformly distributed over the interval $-\pi$ to π.

The probability density function (pdf) of X_i is:[8]

$$p_{Xi}(x) = \begin{cases} \dfrac{1}{m_i\cdot\pi\cdot\sqrt{(1-(x/m_i)^2)}} & ,|x|<m_i \\ 0 & ,\text{otherwise} \end{cases} \qquad (9)$$

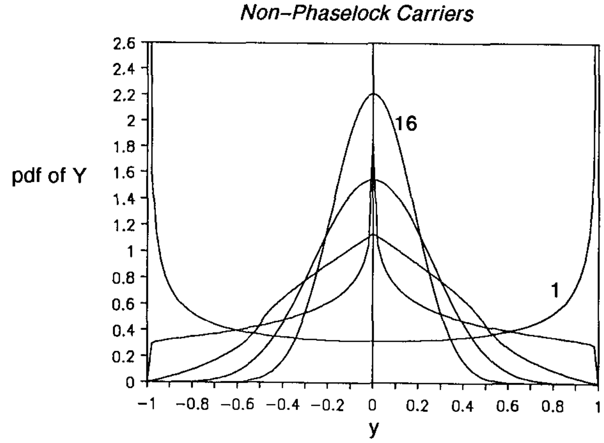
Define a random variable Y, formed by sampling the signal y(t) at time T. Assuming the random variables θ_i are statistically independent, the pdf of Y can be determined by convolving the pdf's of each individual signal.[9] So for N channels the pdf of Y can be determined as follows:

$$p_Y(y) = p_{X1}(y) * p_{X2}(y) * \ldots * p_{XN}(y) \qquad (10)$$

Figure 3 shows the function $p_Y(y)$ for different numbers of channels. The channel amplitudes, m_i, are normalized to $m_i=1/N$ in each case. The function approaches a Gaussian distribution as the number of channels increases. For N>8, $p_Y(y)$ can be approxiamated by a zero mean Gaussian distribution variance given by equation 11.

$$\sigma = m_i\cdot\sqrt{(N/2)} \qquad (11)$$

Statistical Distribution of Sums of Non-Phaselock Carriers



1, 2, 4, 8, and 16 carriers

Figure 3

This result can be used to estimate practical values of modulation depth as a function of the number of channels. Assuming a perfectly linear laser, the only source of distortion in the laser's optical output is due to clipping when the signal drives the laser below threshold. For values of m_i less than 1/N this never occurs. As the modulation index is increased past some threshold, distortion due to clipping increases rapidly. Based on experimental results this threshold seems to be between 5% and 6% per carrier for a 40 channel system and between 7% and 8% per carrier for a 20 channel system. Assuming $m_i = 0.055$, then for 40 channels $\sigma=0.055\sqrt{20}=0.246$. The probability of the signal driving the laser below threshold at any given time is equal to the probability of a zero mean Gaussian random variable with $\sigma=0.246$ exceeding -1. This probability can be evaluated using the Q function[10] and it is roughly 2.5E-5. This probability is quite small but it seems to be significant in terms of measured distortion performance. This result can be used to project the practical modulation depth as a function of channel loading based on the assumption that the maximum practical value of σ is 0.246 regardless of the number of channels.

$$\sigma = m_i \cdot \sqrt{(N/2)} = 0.246 \tag{12}$$

$$m_i = \frac{0.246}{\sqrt{(N/2)}} = \frac{0.348}{\sqrt{N}} \tag{13}$$

In HRC systems the statistical distribution of the signal depends on the phase relationships between the channels. It is possible to choose the channel phases in such a way as to minimize the peak amplitude of the composite signal. It has been suggested by other authors that this could have an effect on system performance[11] but the extent of the possible improvement was not explored for AM fiber systems. In a 40 channel system with $m_i=1$ the peak signal level could be as high as 40. With proper selection of carrier phasing it is possible to reduce the peak signal level to less than 9.1. Figure 4 shows the pdf's of a 40 channel standard signal and a 40 channel optimally phased HRC signal. In both cases the carriers are unmodulated with $m_i=1/40$. The pdf of the HRC signal was determined by computer simulation of the signal. In general, we have determined that the peak signal level with optimally phased unmodulated HRC carriers can be determined using the following equation.

$$\text{Peak} \leq 1.5 \cdot m_i \cdot \sqrt{N} \tag{14}$$

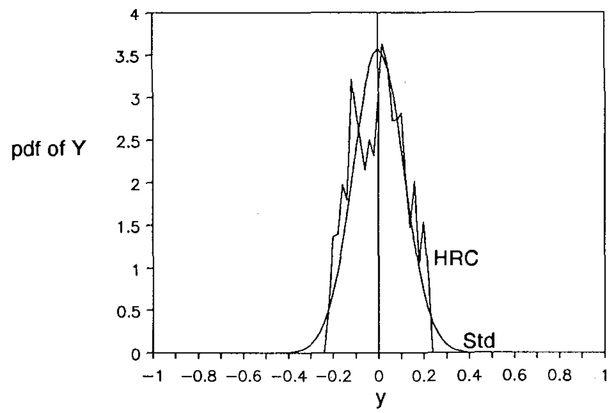


Figure 4

When the carriers are modulated with video information the peak signal level may increase somewhat. The extent of this increase has to be determined experimentally.

This expression for peak signal level can be used to determine the theoretical maximum useable modulation depth for HRC systems. The theoretical maximum is a modulation depth which results in a peak signal level just equal to one.

$$\text{Peak} = 1$$

$$1.5 \cdot m_i \cdot \sqrt{N} = 1$$

$$m_i = \frac{1}{1.5 \cdot \sqrt{N}} = \frac{0.67}{\sqrt{N}} \tag{15}$$

Table 1 shows the calculated modulation index as a function of channel loading using equations 13 and 15.

Table 1

N	m_i (per carrier) Std	HRC
10	0.11	0.21
20	0.078	0.15
40	0.055	0.11
60	0.045	0.086
80	0.039	0.075

This result shows that it is at least theoretically possible to use modulation depths in HRC systems that are twice as large as the modulation depths that are practical in non-phaselock systems. This translates to a possible improvement in link carrier to noise ratio of up to 6 dB. The extent to which this improvement is realizable has to be determined by extensive laboratory testing.

SYSTEM PERFORMANCE

The carrier to noise ratio at the output of the link can be calculated using equation 16.[12]

The C/N is calculated in terms of mean squared currents at the input of the receiver amplifier. The numerator represents the mean squared signal current of each video carrier. The terms in the denominator are due to quantum noise, laser noise and receiver thermal noise (transimpedance receiver) respectively.

Figure 5 shows the AM link C/N versus laser coupled power for a 40 channel system with a 10 dB loss budget. The three curves were calculated based on the assumptions summarized in Table 2.

The quantum limit represents the upper limit on C/N performance due to quantum noise. The modulation depth per carrier was calculated using equation 15 because this represents the maximum achievable optical modulation depth. The two practical limits curves assume a state of the art optical receiver employing a high responsivity pin diode (0.9 A/W at 1310 nm). In order to simplify the analysis the laser RIN was assumed to be -160 dB/Hz. In reality the laser RIN is a function of the laser's output power, with the best DFB lasers consistently exhibiting RIN levels in the low -150's today.

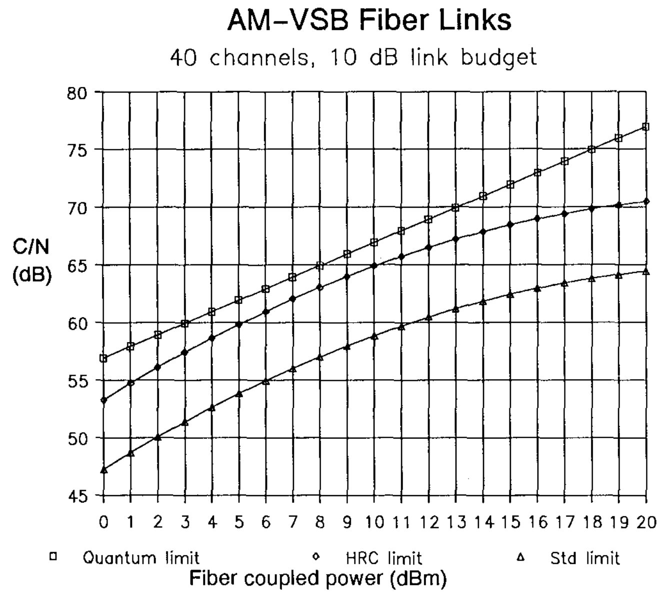


Figure 5

$$C/N = \frac{(m_i \cdot R \cdot Pr)^2}{(2 \cdot e \cdot Bv \cdot R \cdot Pr) + RIN \cdot Bv \cdot (R \cdot Pr)^2 + \dfrac{4 \cdot k \cdot T \cdot F \cdot Bv}{Rz}} \qquad (16)$$

Where:

RIN = laser noise
m_i = optical modulation index
Pr = received optical power
Bv = video bandwidth
e = charge of an electron
k = Boltzmann's constant
R = photodiode responsivity
Rz = receiver preamp transimpedance
F = noise factor of preamp
T = receiver temperature

Table 2

	Quantum limit	Practical limit (HRC)	Practical limit (Std)
m_i	0.11	0.11	0.05
Photodiode quantum efficiency	100%	86%	86%
Laser RIN (dB/Hz)	No laser noise	-160	-160
Amplifier transimpedance (Ω)	No thermal noise	1200	1200
Amplifier noise factor	NA	2	2

CONCLUSIONS

These results show that for standard systems 40 channel link C/N performance of 55 dB is achievable with laser output power of 6 dBm (4 mW). This power level is well below the stimulated Brillouin scattering threshold of 10 mW for narrow linewidth sources. In systems using direct modulation or external modulation with broad source linewidths considerably higher power levels are theorectically possible. This could lead to higher C/N performance or longer link budgets. In optimally phased HRC systems the link C/N performance could be as much as 6 dB better than in standard systems. The actual amount of realizable performance improvement has to be determined by extensive lab testing.

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AN ECONOMIC ASSESSMENT OF AM FIBER OPTIC APPLICATIONS

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ABSTRACT

Recent technological advancements in fiber optic hardware have moved this technology to the forefront of attention for utilization in Cable TV distribution plants. FM fiber optics is well on the way to replacing AML (microwave) sites and/or RF distribution supertrunks. AM fiber optics is under consideration for replacement or supplementing the standard RF distribution trunks. The application of fiber optic technology is viewed as being consistent with future bandwidth expansion requirements of high definition television and increased system reliability.

The purpose of this paper is to review only the economic aspects of fiber optic applications. Based on present AM fiber link performance versus typical RF CATV equipment, three practical distribution system scenarios are examined. In each scenario an economic assessment of an AM fiber node approach is assessed by comparison to a typical RF distribution plant.

The scenarios presented in this document were selected from a lengthy list of actual system upgrade/rebuild analyses which Jerrold's System Design and Proposals Department has performed over the last 18 months. These specific examples were chosen as being representative of the type of system expansions being considered most prevalent.

It is not the intent of this paper to conclude whether AM fiber or typical RF distribution plant is an appropriate economic decision. Rather, it is intended to highlight those areas where each has its advantages and to stress the importance of giving appropriate consideration to the application of a hybrid architecture for CATV systems.

2-1/2 MILES ADDITIONAL TRUNK REACH FOR $20K

The major benefit of an AM fiber optic link, in a CATV distribution network, is the trunk reach advantage AM fiber optics technology has over standard RF amplifier/coaxial cable technology. The issue, however, is whether the benefit is worth the extra cost.

In a straight 550 MHz trunk run analogy, a current generation AM fiber link of approximately 7.3 miles provides a carrier-to-noise performance of 52 dB and a composite-triple-beat performance of 65 dB. Utilizing eight 26 dB gain, 550 MHz feedforward amplifiers and one-inch cable, the same performance specifications are possible at a maximum distance of 4.8 miles. The AM fiber link (equipment and cable) would cost approximately $60-65K compared to $40-45K for the feedforward supertrunk (equipment and cable). In both cases, the cost per mile equates to approximately $8.6K; however the economics of the $20K for 2-1/2 miles of additional trunk reach will vary from system to system.

Equipment Specifications

In order to minimize the variety of equipment used in the following scenarios, all systems will be 550 MHz (77 Channels) new builds, rebuilds or upgrades. Table Number 1 provides the critical operating specifications for the RF and fiber optic active devices utilized throughout the paper. It is also important to emphasize that all economic assessments are made with the assumption of 40 channels/fiber transmitter - receiver links.

TABLE NUMBER 1
EQUIPMENT SPECIFICATIONS

I. STANDARD RF ACTIVE DEVICES

AMPLIFIER DESCRIPTION	GAIN (dB)	NOISE FIGURE (dB)	OUTPUT LEVEL (dBmV)	CTB @ OUTPUT (dB)
550 MHz, 26 dB Feedforward Trunk	26	12	38	85
550 MHz, 30 dB Feedforward Trunk	30	11	41	79
550 MHz, Quadrapower™ Bridger	25	17	48	65
550 MHz, Power Doubling Line Extender	30	13	45	68

II. AM FIBER LINE

Transmitter Input Level (dBmV)	30
Receiver RF Output Level (dBmV)	32
Loss Budget (dB)	6
Channels/Link	40
Carrier-to-Noise (dB)	52
Composite-Triple-Beat (dB)	65

SCENARIO I - 450-550 MHz UPGRADE

System Information

The existing distribution system is carrying 60 Channels (450 MHz) with several trunk runs having 27 amplifier cascades. The active equipment in the system utilizes conventional technology Trunk cable is 3/4 inch foam dielectric and has been tested successfully beyond 600 MHz.

The goal of the upgrade is to expand channel capacity to 77 Channels (550 MHz). If possible, it would be most desirable to save existing trunk locations and, therefore, trunk/feeder tie points. In addition, franchise documents specify the system (at the tap) must meet the following specifications:

Carrier-to-Noise	47 dB
Composite-Triple-Beat	53 dB

Standard RF Upgrade

The equipment selected for the standard upgrade was 26 dB gain, feedforward trunks with Quadrapower™ bridgers and power doubling line extenders. The trunk stations provided for maintaining the existing locations and cable. The desired system performance specifications, however, could not be accomplished with reasonable bridger and line extender levels. Even with bridger and line extender output levels at 43 dBmV and 40 dBmV respectively, the 27 amplifier cascade produced inadequate specifications (C/N: 46.3 dB and CTB: 51.0 dB).

AM Fiber Upgrade

Using an AM fiber link with a 6 dB loss budget (approximately 12 Km) allowed for a maximum cascade of 9 trunk amplifiers. Originating from the headend, the fiber link would terminate at the eighteenth amplifier. At that point, the RF output would be split, amplifiers 10 through 18 reversed and both nine amp cascades (18-10 and 19-27) fed by the AM fiber link.

Using the same RF equipment selected for the standard RF upgrade, the desired system performance specifications are achievable. In addition, reasonable feeder levels (bridger: 47 dBmV, line extender: 44 dBmV) can be maintained.

Scenario I Economics

The massive amount of feeder system changes required for the standard RF upgrade make a detailed economic comparison unnecessary. Based on preliminary calculations, the standard RF upgrade would cost approximately $300-$350/Mi. more (RF distribution equipment costs only) than the AM fiber upgrade. Based on the system mileage of 600 miles, the RF approach would require approximately $195K additional expenditures for RF distribution equipment. The $120K for the three AM fiber nodes would represent a $75K savings for equipment alone. Realistically, the only economic approach would be the AM fiber link. In addition, it represents the only approach to satisfy the system performance requirements.

SCENARIO I - 450 MHz TO 550 MHz UPGRADE

A. EXISTING SYSTEM
- o Channel Loading: 60 Channel, 450 MHz.
- o Longest Cascade: 27 Amps.
- o Equipment Type: Conventional Trunk, Bridger and Line Extenders.
- o Trunk Cable: 3/4 Inch.

B. DESIRED SYSTEM
- o Channel Loading: 77 Channel, 550 MH
- o System Specification Targets: CTB: 53 dB; C/N: 47 dB.
- o Other: Maintain Trunk Amplifier Locations and Cable.

C. CONVENTIONAL RF UPGRADE
- o Equipment Type: Feedforward Trunk, Quadrapower™ Bridgers and Power Doubling Line Extenders.
- o Cascade and Level Analysis:

Amplifier	Cascade	Input Level	Output Level
Trunk	27	14 dBmV	40 dBmV
Bridger	1	--	43 dBmV
Line Extenders (2)	2	10 dBmV	40 dBmV

- o System Distortion Analysis:
  - Composite-Triple-Beat: 46.3 dB
  - Carrier-to-Noise: 51.0 dB

D. AM FIBER UPGRADE
- o Equipment Type: 12 Km AM Fiber Optic Link, Feedforward Trunk, Quadrapower™ Bridger and Power Doubling Line Extenders.
- o RF Cascade and Level Analysis:

Amplifier	Cascade	Input Level	Output Level
Trunk	9	12 dBmV	38 dBmV
Bridger	1	--	47 dBmV
Line Extenders (2)	2	14 dBmV	44 dBmV

- o System Distortion Performance:

	Fiber	RF	System
Composite-Triple-Beat:	65 dB	55.7 dB	53.1 dB
Carrier-to-Noise:	52 dB	48.9 dB	47.1 dB

SCENARIO II - TOTAL 550 MHz REBUILD

System Information

Primarily due to the condition of the existing plant, the system operator had decided to do a complete system rebuild to obtain a 550 MHz, 77 Channel distribution plant. The entire project, with the exception of a few long cascade runs, meets desired specifications with 7/8 inch cable and 26 dB gain feedforward trunks, Quadrapower™ bridgers and power doubling line extenders.

It is desired to reduce the longer cascades to a maximum of 25 amplifiers deep, in order to obtain system specifications of 46 dB carrier-to-noise and 51 dB composite-triple-beat. Reducing the cascade lengths is also advisable for ongoing maintenance purposes.

Alternative Analysis

Three options were selected for consideration in reducing the cascade lengths. The first option was to reduce cable losses by installing 1-1/4 inch trunk cable. This would allow the required distance to be covered with 24 amplifiers. The other two options involved the use of an AM fiber optic link with a 4 dB loss budget covering approximately 7.7 Km. One of the optics options would use 7/8 inch trunk cable, while the other would incorporate 1-1/4 inch cable. These two fiber options reduced the cascades to 20 amps and 14 amps respectively.

Since the majority of the system was in compliance with the desired system specifications, it was decided that the feeder levels would not be altered to make any of the options under

SCENARIO II - 550 MHz TOTAL REBUILD

- SYSTEM INFORMATION
  - Initial design, utilizing feedforward trunk and 7/8 inch cable required several 31 amp cascades. It was desired to reduce all cascades to 25 amplifiers or less.

- ALTERNATIVE ANALYSIS
  - Option 1 - Use the same RF amplifiers, but upgrade cable to 1.125 inch cable.
  - Option 2 - Use the same RF amplifiers and cable, but add a 7.7 Km AM fiber link (4 dB loss budget, C/N: 53 dB, CTB: 65 dB).
  - Option 3 - Use the same RF amplifiers, but upgrade cable to 1.125 inch cable and add 7.7 Km AM fiber link.

		Initial Design	Option 1	Option 2	Option 3
RF Amplifier Cascade		31	24	20	14
Cable:	Quantaity (1000 Ft.)	118.6	118.6	118.6	118.6
	Cost ($K)	$ 56.9	$ 97.2	$ 56.9	$ 97.2
Amps:	Quantity	61	46	61	46
	Cost ($K)	$ 81.8	$ 61.7	$ 81.8	$ 61.7
P.S.:	Quantity	31	23	31	23
	Cost ($K)	$ 37.2	$ 27.6	$ 37.2	$ 27.6
Conn:	Quantity	220	174	220	174
	Cost ($K)	$ 2.2	$ 3.8	$ 2.2	$ 3.8
AM Fiber Link:					
(2) Transmitter	($K)	--	--	$ 30.0	$ 30.0
(2) Receivers	($K)	--	--	$ 10.0	$ 10.0
7.7 Km F.O. Cable	($K)	--	--	$ 13.9	$ 13.9
	TOTAL COST	$178.1	$190.3	$231.9	$244.2
System Performance:					
Carrier-to-Noise (dB)		45.5	46.4	46.1	47.1
Composite-Triple (dB)		49.7	51.2	50.3	51.6

consideration meet the performance parameters. Based on this criteria, option 3 was eliminated (CTB: 50.3).

Scenario II - Economics

Option 1 met the desired criteria (cascade length and system performance) at a cost increase of only $12,200 compared with the existing design. Option 3 also met the criteria; however, the cost increase is estimated at $66,100. Option 1 was selected by the operator. Since option 3 offered approximately the same performance as option 1, the $53,900 incremental cost increase was not economically justifiable. Attempting to justify the extra expenditures, based on the cost savings realized through cascade reductions (Option 1: 24 amps, Option 3: 14 amps) would have been unsuccessful, since the majority of the remaining cascades in the system were between 19 and 24 amplifiers deep.

SCENARIO III - 300 MHz TO 550 MHz UPGRADE/REBUILD

System Information

The existing 355 mile, 300 MHz system is operating with conventional active devices in cascades less than or equal to 20 amps. The feeder line levels are 44 dBmV for bridgers, 43.5 dBmV for line extenders and 8 dBmV at the tap. Trunk and feeder cable (3/4 inch and 1/2 inch) has been tested to 600 MHz and is reusable.

The system, by franchise agreement, is now required to expand channel capacity to a minimum of 70 Channels. The option of constructing a "B" cable system of 300 MHz to obtain the additional 35 Channels was discussed, but eliminated from consideration due to the ongoing maintenance problems it would cause. It was decided to upgrade, if possible, to 550 MHz. Cable would be saved as much as possible; however, system specifications of C/N: 47 dB, CTB: 52 dB and 13 dBmV tap levels would dictate how much of the existing plant could be saved.

Due to design limitations, all taps and system passives would have to be replaced as well as all active devices. In addition, based on preliminary calculations and the vast amount of feeder line construction activity already required, it was decided to upgrade the feeder cable to 5/8 inch cable.

AM Fiber Optic Upgrade

By utilizing an AM fiber link, backbone trunk architecture to reduce RF amplifier cascades to 4, trunk locations could be maintained, in addition to 75% of the existing trunk cable (25% had to be replaced with 1.0 inch cable to reduce losses). 30 dB gain feedforward trunk/Quadrapower™ bridger mainstations were required for trunk spacing and maximum bridger levels (49 dBmV). The additional loss of 550 MHz vs. 300 MHz, in addition to a tap level increase of 5 dBmV, required that most areas needed line extenders to be cascaded three deep.

Following the analysis that proved the compliance of the above hybrid fiber optic/RF distribution plant to system specifications (C/N: 47.7 dB, CTB: 52.1 dB and 13 dB tap levels) a review was conducted to reduce the number of fiber nodes required. A cascade analysis revealed that the end of line performance of a nine trunk amplifier cascade was approximately equal to the AM link, followed by a four amp cascade. Therefore, the AM backboning was modified so that cascades emanating from the headend would be nine amps deep. All other cascades would be limited to four amplifiers, one bridger and three line extenders. Using this approach, the number of fiber nodes was reduced from 14 to 10.

SCENARIO III - 300 MHz TO 550 MHz UPGRADE/REBUILD

A. EXISTING SYSTEM
- 300 MHz.
- Longest Cascade: 20 Amps.
- Equipment Type: Conventional Trunk, Bridger and Line Extenders.
- Cable Type: 3/4 inch trunk, 1/2 inch feeder.
- Levels: Bridger - 44 dBmV.
  Line Extenders: (2) - 43.5 dBmV.
- Tap Port Level: 8 dBmV at 300 MHz.
- 355 miles of plant.

B. DESIRED SYSTEM
- 550 MHz, 77 Channels.
- Reuse Trunk Locations and Cable (where possible).
- Desired System Specifications: CTB = 52 dB, C/N = 47 dB
- Tap Port Level (minimum): 13 dBmV at 550 MHz.

C. UPGRADE/REBUILD INDEPENDENT
- All Connectors will be replaced.
- Cable is reusable if design losses are acceptable.
- All Taps and System Passives will be replaced.

D. UPGRADE ANALYSIS
- Equipment Type: AM fiber optic links, 30 dB gain, feedforward trunk, Quadrapower™ bridger and power doubling line extenders.
- System Changes:
  - 25% of the trunk cable would require change-out to 1.0 inch cable.
  - 100% of the feeder cable would require change-out to 5/8 inch cable.
  - Line Extenders would be required to cascade 3 deep in some cases.
  - Trunk amp cascade limited to 9 deep from headend and 4 off any fiber node.

SCENARIO III

D. UPGRADE ANALYSIS (CONT'D.)

- System Performance: F.O. link + 4 trunks + 1 bridger + 3 line extenders or 9 trunks + 1 bridger and 3 line extenders. C/N = 47.7 dB. CTB = 52.1 dB.
- Tap Port Levels: 13 dBmV at 550 MHz.
- Operating Levels:

Amplifier	Cascade	Input Level (dBmV)	Output Level (dBmV)
30 dB Feedforward Trunk	4	9.0	39.0
Quadrapower™ Bridger	1	25.0	49.0
Line Extenders	3	13.0	43.0

- 73 Km of F.O. cable and 10 AM fiber nodes required.

E. TOTAL REBUILD ANALYSIS

- Equipment Type: 26 dB gain, feedforward trunk, Quadrapower™ bridgers and power doubling line extenders.
- System Changes: All trunk and feeder replaced with 3/4 inch and 5/8 inch cable respectively.
- System Performance: 20 trunks + 1 bridger + 2 line extenders. C/N = 47.0. CTB = 52.6.
- Tap Port Levels: 13 dBMv.
- Operating Levels:

Amplifiers	Cascade	Input Level (dBmV)	Output Level (dBmV)
26 dB Feedforward Trunk	20	14	40
Quadrapower™ Bridger	1	25	45
Power Doubling Line Extender	2	3	42

Standard RF Total Rebuild

A total rebuild approach was analyzed, using 3/4 inch trunk and 5/8 inch feeder cable. 26 dB gain feedforward trunk amps with Quadrapower™ bridgers and power doubling bridgers were required. With bridger, line extender and tap levels of 45 dBmV, 42 dBmV and 13 dBmV respectively, system performance of 47.0 dB carrier-to-noise and 52.6 dB composite-triple-beat were demonstrated.

Scenario III - Economics

More trunk amps are required in the rebuild than the upgrade because each has 4 dB less gain, but also because more trunk is required in the rebuild. At the equipment line, the total rebuild would seem to offer a $250K advantage. This advantage is offset by the cable, strand, hardware and installation labor to replace the entire trunk network, compared to less extensive requirements for the upgrade. Taps passives and feeder cable prices and installation were not part of the analysis, since they would be required in both the rebuild and upgrade.

The economic analysis indicates no clear advantage to either approach. At present, the operator is reviewing the option from an ongoing maintenance viewpoint. The rebuild approach offers a completely new plant but a number of 20 amp cascades. The upgrade offers cascades of 4 to 9 amps (maximum) but, many areas with three cascaded line extenders.

SCENARIO III - COST COMPARISON

	AM FIBER UPGRADE		TOTAL REBUILD	
	Quantity	Total $	Quantity	Total $
A. EQUIPMENT COSTS				
1. Standard RF Equipment				
o Amplifiers	163	$ 233,900	255	$ 357,000
o Line Extenders	1705	530,250	1666	518,126
o System Passives	1737	66,000	1890	71,800
o Power Supplies	102	122,400	133	159,600
2. Fiber Optic				
o AM Transmitter	20	$ 300,000		--
o AM Receiver	10	100,000		--
SUBTOTAL EQUIPMENT		$1,352,550		$1,106,526
B. CABLE, HARDWARE, STRAND AND INSTALLATION				
1. Coaxial Cable				
o 1.0 In. Trunk (1000 Ft)	104	$ 204,200	--	--
o 3/4 In. Trunk (1000 Ft)	--	--	470	$ 651,400
2. Fiber Optic Cable				
o Four-Fiber Bundle (Km)	73	$ 168,000	--	--
SUBTOTAL CABLE AND INSTALLATION		$ 372,200		$ 651,400
TOTAL		$1,724,750		$1,757,926

CONCLUSION

The cost trade-off for the additional trunk reach provided by AM fiber optic technology was examined in three specific, real-life system scenarios. In each case, the focus was solely on the economics of providing the required system performance parameters. On review of the three scenarios we have considered, plus all of our previous experience, there are situations where a hybrid coax-fiber design make economic sense. Especially when one considers the advantages inherent in such a hybrid system in terms of quality, etc., AM fiber optic products need to be given serious consideration on a system by system basis. It has been our experience that there are situations where fiber optics pays for itself or adds only moderate cost without considering the incremental benefits to system performance.

Consideration should also be given to the fact that AM fiber optic products are still in the infancy of their development and are rapidly advancing. Product performance improvements which could dramatically improve the cost vs. performance ratio may happen at anytime. Such changes will alter the economic analyses presented in this document.

It is the opinion of these authors that the new AM fiber optic generation of products should be viewed as another option to be considered for use in system upgrades/rebuilds. Therefore, fiber optics should be added to the list of technology considerations, along with power doubling, Quadrapower™ and feedforward, when considering the economics of system expansions.

ANTENNA CONSIDERATIONS FOR CONTROLLING CABLE SYSTEM LEAKAGE

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ABSTRACT

The success of a cable leakage maintenance program is keyed to the ability of the cable operator to monitor, categorize and locate leakage within the system. The antenna plays an important part of controlling cable leakage since system leakage exists outside the cable system. This paper reviews various antenna alternatives and suggests how they may be used in a leakage maintenance program. Several references are given so that the cable system technical personnel may further understand the considerations of antenna selection.

INTRODUCTION

The required methodology for making field strength measurements in determining the Cumulative Leakage Index or CLI is well described in section 76.605 of the FCC rules. It states the following:

> " The resonant half wave dipole antenna shall be placed 3 meters from and positioned directly below the system components and at 3 meters from the ground."

While this measurement technique will ensure a consistent standard in terms of the law it does present certain logistical problems if the cable operator is to perform the measurement process on a routine basis. What is important for any alternative measurement method is that cable system is controlled by cable leakage operator. Any measurement alternative should have traceable performance to the legal standard.

An antenna will provide a terminal voltage when placed in an electric field according to the following relationship:

Equation 1 $$dB\mu V = dB\mu V/m - K$$

Equation 2 $$K = 20\,\mathrm{Log}(f) - G_{dB} - 31.54\ \mathrm{dB}$$

K = antenna factor in dB
f = frequency in MHz
G_{dB} = gain of the antenna over an isotropic

Any antenna should be able to be used for field strength measurements if its antenna factorcan be established. The antenna factors can relate the measured terminal voltage to that obtained with a dipole. The user can then be assured that the field strength measurements made with the alternative antenna are representative of those he would have obtained using a dipole. More suitable measurement techniques will encourage routine quantitative characterization of leaks resulting in better control of cable leakage.

Currently there are two measurement alternatives which are accepted for the CLI process. First is the use of the inverse distance law which relates field strength to the distance from the RF source. By using this relationship measurements can be made from a more practical distance and the measured results extrapolated to the actual distance. The assumption is more sensitive to parasitic effects such as reflections from conductive elements such as power and phone lines as well as any other reflective elements such as buildings.

The second alternative is the placement of the measurement dipole on a vehicle roof at a height of 1 meter. An antenna height of less than several wavelengths above ground or a reflective element acting as ground causes a distortion of the antenna pattern and the resulting gain at various radiation angles. Knowledge of how the pattern is affected will ensure that the proper interpretation is made of the field strength readings.

The purpose of this paper is to present information about alternative antennas which may be used for cable leakage measurements. Antennas can be classified by their polarization: horizontal, vertical, or circular. This paper will discuss antennas which have vertical and horizontal polarization. Additionally, direction finding (DF) and near field antennas will be discussed.

HALF WAVE DIPOLE

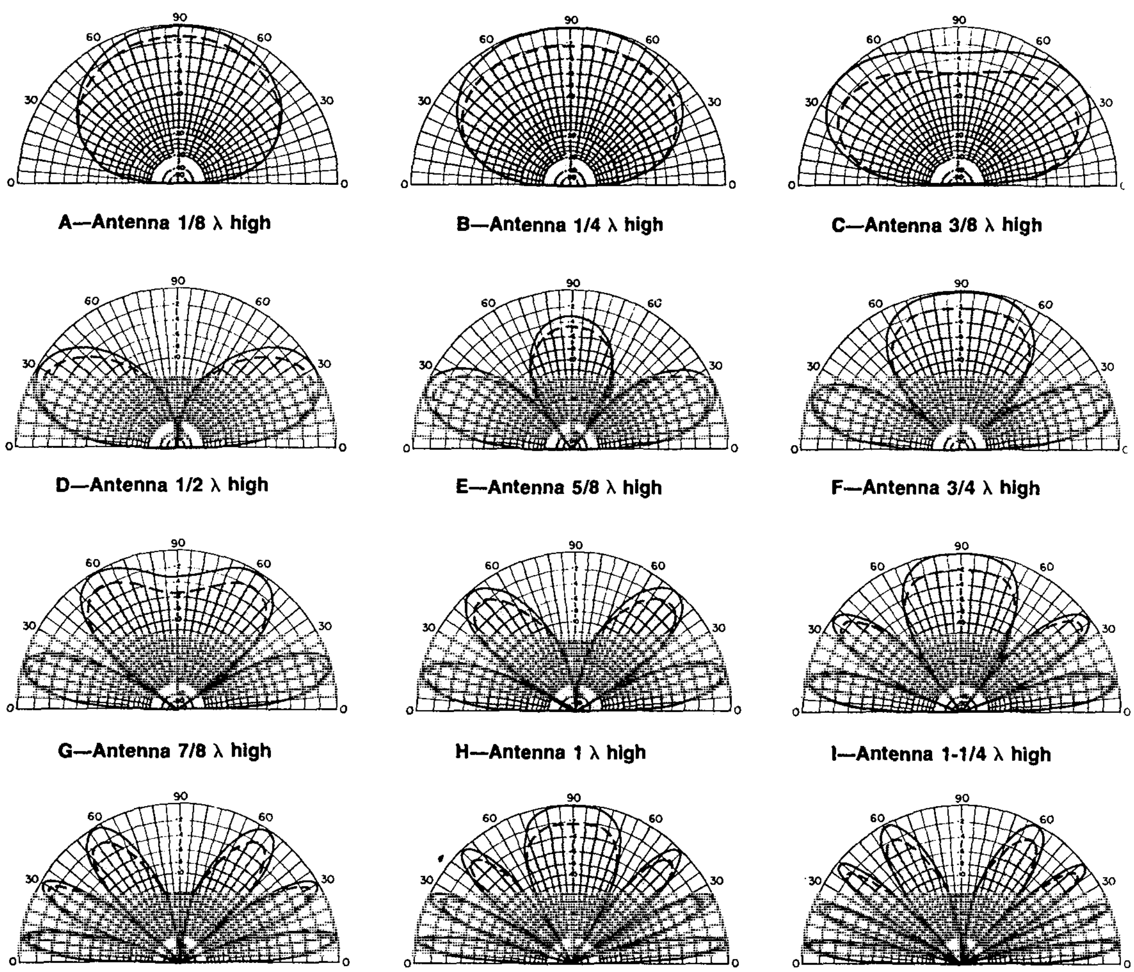
The half wave dipole is a well characterized radiating element which exhibits a gain of 2.15 dB over an isotropic. It is the practical standard that is used for most antenna work. Most other antennas are related to it. However, it can not be used blindly. The radiation pattern and overall gain are easily affected by parasitic reflective elements. The strategic placement of parasitic elements and the resultant affect on the overall antenna pattern is of course the basis of the Yagi-Uda design.

Using Equation 2 the antenna factor for a half wave dipole is given as K = 20 Log(f x 0.021).

Figure 1 illustrates the pattern distortion which can occur for various antenna heights. What results is that the gain of the dipole varies at different angles of radiation. A certain antenna height has advantages over other heights when measuring cable emissions. Cable leakage measurements in an easement made from the street will have a low angle of radiation. Cable leakage measurements made on a strand directly above the CLI vehicle will have a high angle of radiation. These pattern distortions should be taken into account if CLI measurements are to be directly correlated to those made with a dipole outlined in 76.605.

One way to minimize pattern distortion is to select a measurement frequency which is compatible with the desired antenna height. There are two measurement scenarios. First is with the dipole three feet above the roof of a vehicle and second that outlined in 76.605.

FIGURE 1.



A—Antenna 1/8 λ high

B—Antenna 1/4 λ high

C—Antenna 3/8 λ high

D—Antenna 1/2 λ high

E—Antenna 5/8 λ high

F—Antenna 3/4 λ high

G—Antenna 7/8 λ high

H—Antenna 1 λ high

I—Antenna 1-1/4 λ high

J—Antenna 1-1/2 λ high

K—Antenna 1-3/4 λ high

L—Antenna 2 λ high

Table 1 has the antenna heights in wavelength for each of these scenarios at different frequencies.

TABLE 1
ANTENNA HEIGHTS
RELATED TO FREQUENCY OF MEASUREMENT

f	L	10/L	3/L	0.666/L
72 MHz	13.0 ft	0.76	0.231	0.051
108 MHz	8.6 ft	1.20	0.349	0.078
118 MHz	7.9 ft	1.30	0.378	0.084
137 MHz	6.8 ft	1.50	0.439	0.098
225 MHz	4.2 ft	2.40	0.721	0.159
400 MHz	2.3 ft	4.30	1.280	0.290

Note the patten variation between heights at 118, 225 and 400 MHz (3/8, 3/4 and 1 1/4 wavelengths). The pattern at 118 MHz gives a good overall coverage except at low radiation angles. Low radiation angles will be experienced when the leak is in the easement. Leaks in an easement will be covered better with higher antenna heights.

In a similar manner, leaks directly above a vehicle will be covered well with an antenna height which has a predominant response at high angles of radiation.

A popular vehicle configuration is to mount a dipole a very short distance (eight inches) above the roof. Although somewhat difficult to characterize, the resultant antenna pattern will be similar to that of the eighth wave pattern for the frequencies in Table 1.

As the dipole is raised above the ground reflections become less predominant and the free-space radiation pattern emerges. The nulls will be less distinct since there is rarely prefect reflection from the ground surface. The patterns for low antenna height will be typical for vehicle configurations while the patterns for high antenna height will be typical for walk arounds.

YAGI-UDA

The Yagi-Uda or Yagi antenna has two characteristics which can aid in cable leakage measurement and detection. The antenna gain can be used to overcome problems that a dipole will have with increasing frequency as well as extend the measurement range. An improved front-to-back ratio as well as reduced gain on the sides of the antenna will aid in locating cable leaks by minimizing the effects of interference from reflections and other leak sources.

A Yagi's multi-element configuration and resulting size dictates that a high frequency of operation be used. However, at 225 MHz the total boom length of a four element Yagi is less than three feet. A four element Yagi will exhibit approximately 8 dB of gain, 20 dB of front-to-back ratio and good side lobe performance. A four percent bandwidth can be expected at the frequency of interest.

One might consider modifying a commercial off-air channel 13 antenna for operation in the upper aeronautical band. The antenna can then be trimmed for the operating frequency of interest. A return loss bridge and a bench top sweep can be used to retune the antenna. Overall guidelines are listed in Table 2 below for a four element Yagi.

TABLE 2
FOUR ELEMENT YAGI DIMENSIONS

Driven element to reflector	0.20 wavelength
Driven element to 1st director	0.20 wavelength
1st director to 2nd director	0.25 wavelength
Reflector length	0.51 wavelength
Driven element length	0.47 wavelength
1st director length	0.45 wavelength
2nd director length	0.44 wavelength

VERTICAL ANTENNAS

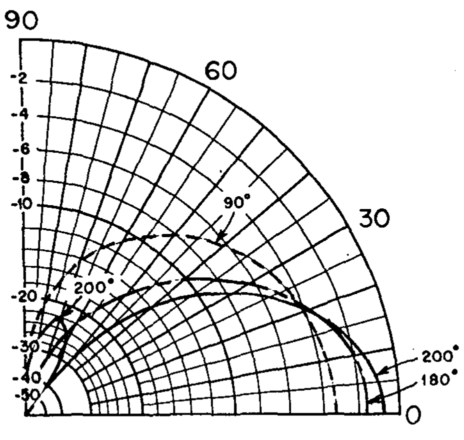
The use of vertical antennas for use in all leakage activities will always be in doubt. The predominant polarization of a leak has been argued for some time. According to the law the use of a vertical antenna is not acceptable. However, it has been shown that most leaks will exhibit both polarizations when measurements are made at a distance. In terms of controlling leakage a vertical antenna can be used to make a field strength measurement. Some determination of the leak severity can then be made. However, decisions should be made on the pessimistic side.

The use of a vertical antenna should be done with the same caution as that for horizontal antennas. Parasitic reflectors on a vehicle will cause a distortion of the antenna pattern which could result in nulls in the response. These parasitic reflectors can come from other antennas or metal objects such as booms and ladders. The objects should either be moved or the overall pattern of the vehicle characterized. Figure 2 shows the antenna patterns for vertical antennas of different lengths. The lengths are given in electrical degrees.

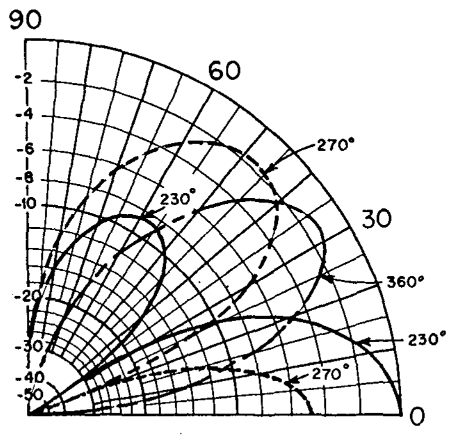
QUARTER WAVE VERTICAL

The use of a quarter wave vertical is a popular choice. It exhibits 3 dB gain over a dipole and is easily configured on a vehicle. It does, however, exhibit a null at high angles of radiation as shown in Figure 2.

FIGURE 2.



(A)



(B)

This antenna can be configured for multiple frequencies to give coverage in all three aeronautical bands.

5/8 WAVE VERTICAL

The 5/8 wave vertical exhibits 3 dB of gain over a quarter wave but has a significant lobe at high angles of radiation. This can be a benefit since leaks directly overhead will be covered better.

A matching network is built into the antenna base since the 5/8 wave is not resonant at the desired operating frequency. The antenna will also resonate at frequencies at which the physical element is 1/4 and 1/2 wavelength. The result is that offair signals will be received and may show up as intermodulation components in the receiver.

The multiple response of a 5/8 wave antenna can also be used to an advantage. Monitoring frequencies can be picked so that several frequencies are scanned. This can give more coverage with one antenna.

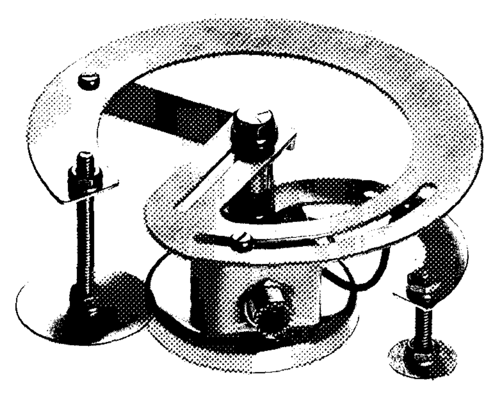
DIRECTIONAL DISCONTINUITY RING RADIATOR

An interesting variation of a vertical antenna is the DDRR. This antenna shown in Figure 3 has an overall antenna height of 2 1/2 inches and a diameter of 8 inches for a unit tuned to the aeronautical band (121.25 MHz). The DDRR is intended for use on a vehicle. A radome is available to protect it from the weather.

This antenna has a similar radiation pattern to a 1/4 wave whip. It exhibits unity gain, however, its high Q gives it an narrow bandwidth. The narrow bandwidth makes it ideal for areas with high intermodulation from offair signals. For a 2:1 SWR the DDRR has a 3 MHz bandwidth versus the 10 MHz bandwidth of a 1/4 wave vertical.

The DDRR has the advantage that it requires very little ground plane area to achieve its characteristics. The ground plane requirements suggest that this antenna could be mounted on the front of a vehicle. This may be a consideration if roof space is a premium.

FIGURE 3.



RUBBER DUCKIES

A popular antenna for use in portable detection is the "rubber duckie" antenna. These are resonant antennas which have a somewhat broad band characteristic. Their gain is not well characterized however, they do display characteristics which make it suitable for detection and location of leaks.

The most important characteristic is the null on the end of the antenna. This antenna will exhibit a 15 to 20 dB null on the end of the antenna and a maximum response perpendicular to antenna. The null can be used to locate a leak. The antenna will be pointing in the direction of the leak source when a null is found after detection is noted.

DIRECTION FINDING ANTENNAS

Several direction finding antennas are possible. The references contain many examples of DF antennas and their application.

One of the simplest vehicle based DF techniques is to use two Yagis at a frequency in the high aeronautical band. The gain of the antenna will makeup for any anticipated free-space losses and the directional characteristics will allow isolation of the leak. The initial direction of the leak can be determined with each Yagi oriented toward each side of the vehicle. Switching between each antenna will tell the technician where to start his search.

A more sophisticated approach is to use multiple vertical antennas and doppler techniques. Several antennas can be switched electronically and the direction of the incoming signal determined from the relative phase of the signal at each antenna. The relative bearing can be read out on an indicator device giving the initial direction of the search. These devices seem to be very sensitive due to multipath and require a lot of patience to use. False indications while the vehicle is in motion as well as at rest are very common.

These systems usually use an audio tone as the antenna commutating signal. The recovered audio is then used to determine the bearing of the signal. This audio tone is usually placed at the lower corner frequency of the audio response of most receivers. Its placement causes some problems if the receiver does not have adequate response at the commutating frequency.

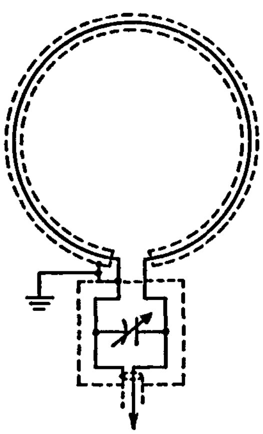
Future possibilities include real-time processing of the recovered doppler signal to determine the validity of the reading. Another possibility is to place the commutating tone within the audio passband of the receiver to ensure as reliable bearing indication.

NEAR-FIELD LOOPS

Once a leak is detected it must be located, isolated and finally repaired. Isolating a leak is usually the most difficult part of the task. Near-field probes can aid in this process by using the magnetic field of the leak instead of the electric field. The magnetic field has a more pronounced attenuation effect with distance and does not seem to exhibit the same extreme standing wave effects that electric fields demonstrate.

The classic near-field loop is shown in Figure 4. It is easily constructed and is commercially available. It is usually tuned to one frequency. There is n specific calibration requirement except that it be tuned for maximum response.

FIGURE 4.



Once connected to a sensitive RF voltmeter it is moved along the strand until a maximum response is achieved.

Another type of near-field probe is a very short vertical antenna. These are usually used with portable detection equipment to locate a leak when the receiver is very near to the source. They take the form of short single element attached to the input of the receiver. The effect is to desensitize the receiver and allow variations in field-strength to be noted.

FURTHER READING

Several references are given at the end of this paper. The best reference for a practical and theoretical understanding of antenna behavior is the ARRL Antenna Book. This reference is revised on a regular basis so that some of the material may not be repeated every issue. However, much of the basic material has not changed since the early 50's and is repeated unchanged. Much of what is changed is related to current work in the amateur arena. It is this authors view that the 1988 and 1970 issues represent good overall references for any technical personnel who has to deal with antennas.

ACKNOWLEDGEMENTS

Several of the illustrations used in this paper are used with permission of the ARRL.

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AUDIO 101 - TELEVISION AUDIO - A SYSTEMS ISSUE FOR CABLE

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ABSTRACT

Audio - how boring - how basic! Audio is really basic stuff, right?? After all, why worry about it when you have your precious video to coddle and perfect. I'll bet you get just as many, if not more, subscriber complaints about sound than about the pictures. I know, the pictures are real important, always have been, always will be, but the sound is a source of subscriber content or irritation.

LISTEN UP YOU BROADCASTERS!

I will call this paper Audio 101 because that's what it's about - the basics, and why if basic good common sense is applied your audio based service calls can be effectively reduced or eliminated. For the sake of this paper, we need to consider ourselves as broadcasters for three very simple reasons:

1. Subscribers are receiving our signals with equipment specifically designed to broadcast standards. (TV SETS).
2. Our signals are generated by cost effective versions of little broadcast transmitters. (MODULATORS).
3. Subscribers compare cable signals to broadcast signals all the time. (SWITCHING CHANNELS).

Try a little experiment when you get back home. Disconnect the cable and connect the antenna. Tune back and forth between the off-air broadcasters and compare the sound. (Turn the brightness all the way down!) Chances are, the sound will be remarkably similar between all stations. Repeat this little test several times during the day and the results will probably be the same.

Chances are that even if the picture is very snowy, (you may cheat and turn the brightness back up) the audio performance could be quantified to be at least as good as:

Frequency Response:	+/- 1dB 50Hz-15kHz
Distortion:	< 1%
Signal to Noise:	> 60dB Unweighted
Stereo Separation:	> 20dB

Granted, these numbers do not represent the pinnacle of audio achievement, but if done properly, will satisfy both "Joe 6 Pack" and "Fred Golden Ears".

GARBAGE IN - GARBAGE OUT

Since the audio signal is FM, as a general rule, whatever you send from the head-end is what will be received at a subscriber's TV set. Even BTSC stereo is a pretty rugged format and remarkably tolerant to "bruises" on the way from head-end to home. So...if you want a good indication of overall audio quality, the tests, measurements, and listening, you in the head-end will go a long way toward insuring subscriber happiness. The secret here is consistency - more on this later.

SATELLITE DELIVERED AUDIO

We receive the bulk of nonbroadcast signals via satellite. Most is digital via the scrambling system. This system is capable of delivering excellent audio quality. Several analog systems are in use including conventional high level wideband as well as narrowband companded subcarriers. A few simple rules here will keep you out of trouble.

1. When dealing with high level subcarriers (6.8MHZ mono audio), be sure the audio section of the receiver has an IF bandwidth wide enough for the signal involved. A majority of programmers using this method have peak deviation of 237 kHz. This requires an IF bandwidth of about 500kHz. Receivers that "cheat" in this area will be subject to annoying "sibilance" distortion (nasty sounding "S" on some program material).

2. When dealing with low level companded subcarriers, be sure the subcarrier demodulator has an expander that matches the compressor in use at the uplink. Many "all in one" receivers have serious problems in this area. Improper demodulation will generally result in signal with excessive noise and/or high frequency compression artifacts.

In general, satellite signals are handed to us ready for broadcast. With day-to-day and program-to-program consistency. As I said, more on this consistency stuff later on.

HOW WE BROADCAST

We broadcast to subscriber via cable using the exact same technical standards as the TV stations. Disregarding stereo for the moment, the basic transmission media is FM with 75 microsecond pre-emphasis. Our old friend, the 75 microsecond pre-emphasis curve was a fine tool in its day; very useful in reducing noise based upon statistic characteristics of audio. Well, things have changed. Audio is getting "hotter" with respect to high frequency content. Digital sources are no longer limited to the old roll-off curves of the LP and tape decks. There are two ways to cope with the situation.

1. Lower the overall program level so that all material peaks will fit within the bounds of the curves or;
2. Use modern audio processing techniques to dynamically "fit" the program material within the curve.

Figure 1 is a basic block diagram of a TV station audio chain. The box labeled "Audio Processor" is worth some comment. First of all, a TV station can afford to spend a few bucks for audio. After all, they only have to worry about one channel, not 35 or 50. The television audio processor is typically a multi-band limiter/compressor specifically designed for TV sound. It is also expensive. The audio processor is used to provide consistency to a station's sound and make it sound similar to other stations in the market.

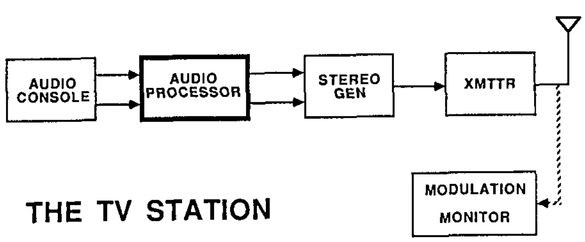

FIGURE 1

A CURIOUS PHENOMENON

Audio processing is relatively new to television. In the "old days" a typical TV station did little audio processing other than protective peak limiting to prevent overmodulation.

When the big fuss started over stereo, broadcasters began to pay attention to audio. When one station in a market converted, it usually meant new audio processing coupled with the new stereo generator. In fact, 2 of the 3 major vendors of broadcast BTSC stereo encoders include sophisticated multi-band processing as part of the package. All it took was one station in the market to process and the rest had to follow just to maintain consistency. Thus, over the past 5 years or so, the "sound" of American TV has changed - for better or worse, it is fairly consistent. In all fairness to the processor manufacturer, today's processing can be set up to sound very clean and natural while maintaining high modulation levels.

CONSISTENCY

Our objective as cable programmers and operators should be to maintain this consistency from channel to channel. It is impractical to expect each head-end to have expensive processing on each and every channel. The task of maintaining consistency belongs to the programmer. A few simple steps if taken by programmers would go a long way toward solving level problems:

1. Insure that the transmission chain from audio board through downlink monitor meets or exceeds specifications.
2. Use broadcast grade audio processing to control loudness, high frequency peaks and ride gain.
3. Have your downlink monitor feed a typical CATV modulator and adjust it for 100% modulation using program material, just like a head-end tech would do. (Using the flashing light.)
4. Compare the signal received via the cable modulator to the major market off-air channels. If there is a significant difference, you have created a problem for your affiliates!
5. Repeat step 4 often.

For the cable operators, if level inconsistencies are a chronic problem on a given channel, consider the use of AGC products on that particular channel. On channels with local commercial inserts, the use of AGC products on the local commercial audio is highly recommended. Again, keep local spot audio consistent in level to network audio.

THE FUTURE

TV sound today is a constant battle to achieve consistency and parity with the broadcasters. In the future as HDTV develops, this may change. It is just too early to tell whether or not TV sound in the future will be natural or highly processed. In the meantime, if Fred Golden Ears wants the highest quality audio he will rent a movie and listen to it in VHS HI-FI. Great sound - but not the same type of sound as broadcast entertainment.

I know this subject is confusing and frustrating, but if programmers and operators strive for consistency today while keeping an eye toward future development, we will have done all we can as an industry to ensure subscriber satisfaction.

BTSC STEREO AND CABLE SYSTEMS
Measurement Techniques and Operating Practices

Alex Best

Cox Cable Communications

ABSTRACT

In August of 1982 the NCTA Engineering Advisory Committee, concerned about the capability of cable systems to provide quality transmissions of stereo audio, formed an ad hoc subcommittee to investigate multichannel television sound. In September of 1982 the subcommittee sent a report to the Chairman of the EIA BTSC committee outlining the areas of technical concern. The cable industry also went on record as being opposed to any of the stereo formats being proposed. The opposition centered on the selection of a subcarrier scheme (very similar to the 30+ year old FM broadcast system) at a time when digital audio systems were becoming widespread in the consumer marketplace. More important, there were cable carriage problems which were explained in detail in a report to the EIA BTSC committee.

In March of 1983 the NCTA subcommittee wrote a comprehensive test plan and hired a test engineer to measure the impact of cable equipment on the proposed stereo systems. The testing was completed in September of 1983 and the results documented in a report titled "Multichannel Television Sound Report." As a result of the efforts of the cable industry, plus others, the FCC granted a "non must-carry" status to the newly selected stereo system in February of 1985.

After the conclusion of the test report in 1983, the BTSC subcommittee entered a phase of providing a clearinghouse for inputs on both successful and unsuccessful attempts by the cable industry in providing quality stereo reception to our customers. As a result of this effort, one fact became evident. The cable industry needed a comprehensive set of measurement procedures and operating practices to verify optimum performance for stereo encoding equipment plus insure quality delivery through the remainder of the system. As a result the subcommittee reconvened its efforts in 1987. In November of 1988 a draft report on BTSC measurements and operating practices was presented to the NCTA Engineering Committee. What follows is a summary of that document.

INTRODUCTION

Listed below is the Table of Contents for the complete report.

I. Measurement Techniques
  a. Signal-to-Noise Ratio
  b. Signal-to-Buzz Ratio
  c. Frequency Response of BTSC Stereo Transmission System
  d. Separation Measurements of the BTSC Stereo System
  e. Total Harmonic Distortion of the BTSC Stereo Transmission System
  f. Incidental Carrier Phase Modulation (ICPM)

II. Operating Practices
  a. Interconnecting with CATV Modulators at Baseband Audio
  b. Interconnecting with CATV Modulators at 4.5 MHz
  c. Interconnecting with CATV Modulators at 41.25 MHz
  d. Online Checks
  e. BTSC Operational Guidelines
  f. Transportation of Stereo Signals Over FM Links

III. Cable Error Budget (Separation)

The complete report consists of 58 pages. Anyone interested in making stereo measurements should obtain a copy of the full document.

During the development of this report, much discussion took place on the quality of test equipment required to obtain meaningful results. The dilemma revolved around whether the subcommittee should recommend precision decoders and demodulators, knowing that many operators don't presently have access to this type of equipment. The alternative was to suggest the use of consumer type test equipment severely limiting the accuracy of the resulting measurements, and in many cases rendering the results useless. After all, in most instances what we are attempting to determine is the performance level of a high quality stereo encoder. Attempting to measure its performance through a device whose specifications are worse than the device(s) being tested will only lead to erroneous conclusions. Because of this, you will find that only precision test equipment is recommended in the "required test equipment" lists in each of the measurement descriptions.

BTSC MODE VS. EQUIVALENT MODE

A second dilemma involved the issue of whether the measurements should be made with the equipment operating in the BTSC Mode or the Equivalent Mode. To help us understand equivalent mode, look at a block diagram of the BTSC system in Figures 1 and 2. What sets Multichannel Television Sound (MTS) apart from conventional FM broadcast stereo is the BTSC noise reduction that makes possible buzz-free stereo operation with intercarrier sound detection. Unfortunately, the presence of this compressor and expander in the L-R channel that is not duplicated in the L+R channel creates a whole host of problems that do not exist in conventional FM stereo. FM stereo broadcasting is much simpler than MTS because the FM system is linear throughout.

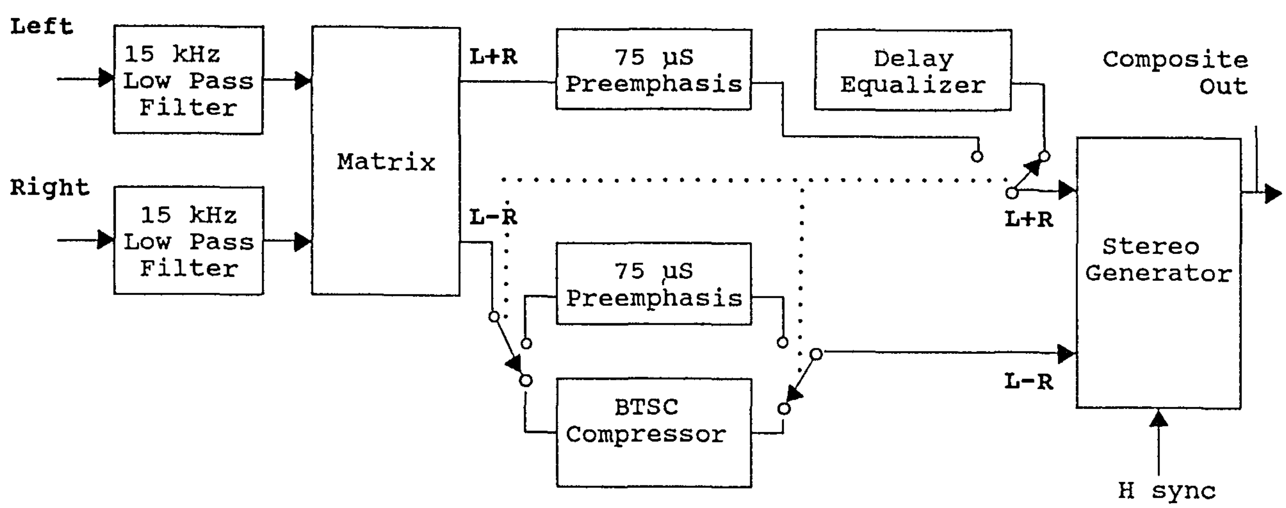

Figure 1 - MTS Stereo Encoder

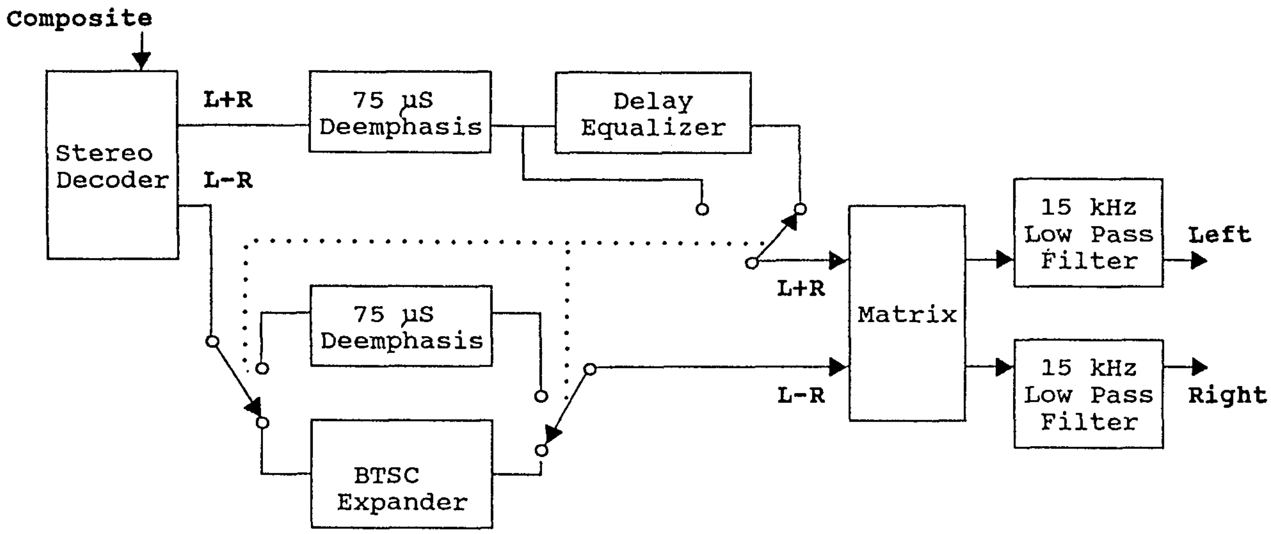

Figure 2 - Stereo MTS Decoder

Measurements made in the BTSC mode (Expander/Compressor active in circuit), while more difficult to make accurately, gives us an indication of how the system actually operates in the real world. One of the major areas of concern is the level setting accuracy. For the expander and compressor to track one another correctly they must have a common reference. For this to occur the absolute FM deviation of CATV modulator must be accurate. The total allowable error budget, input to output, on the L-R channel is ±.1744 db for 40 db of separation and ± .550 db for 30 db of separation. Although many knowledgeable engineers strongly urge that stereo measurements (especially separation) be made with the encoder/decoder operating in the equivalent mode (expander/compressor replaced with 75 usec pre-emphasis/de-emphasis network), we have chosen not to do that for two reasons. First, not all BTSC encoders designed for cable applications have the capability of being operated in the equivalent mode. Second, interpreting the results of measurements made in this mode of operation is confusing when attempting to relate the numbers to real world operation.

MEASUREMENT TECHNIQUES

At this point, I would like to provide a synopsis of each of the measurements defined in the table of contents in the Introduction section of this paper. Because of space limitations, I will not go into the actual measurement procedure but instead will provide the definition, performance objective, and a limited discussion for each of the measurements.

a. Signal-to-Noise Ratio

Definition: Audio signal-to-noise ratio is defined to be the ratio of the audio signal power output to the noise power in the entire audio pass band.

Performance Objective: The BTSC signal delivered to the subscriber's receiver shall be capable of providing a sum channel signal-to-noise ratio, for sum channel peak modulation levels of +/- 25 Khz, of no less than 55 db, when demodulated and decoded by precision BTSC stereo equipment.

When a scrambling system that puts amplitude modulation energy on the sound carrier is employed, this measurement should show no less than 53 db sum channel signal-to-noise ratio.

Discussion: The measurement defined here is referred to as the sum channel signal-to-noise ratio. Since in this measurement L = R, no difference energy should be present and the companding system should be operating at minimum gain. Two measurements may be used to characterize the noise performance of the BTSC system. The sum channel signal-to-noise procedure is used to evaluate output noise effects that are not a function of the companding process. The signal-to-buzz measurement described next evaluates output noise including effects of the receiver's expander.

b. Signal-to-Buzz Ratio

Definition: Signal-to-buzz ratio (S/B) is the ratio in dB of the peak-to-peak signal voltage divided by the peak-to-peak buzz voltage as seen using an oscilloscope at the output of a BTSC stereo decoder. The buzz measurement is made while the signal is present.

Performance Objective: More than 35 db unweighted should be measured using a precision demodulator and stereo decoder at the output of a complete system. In systems with scrambling a minimum of 27 db should be measured. Measurements made with consumer equipment may produce lower numbers.

Discussion: The results of the measurement depend almost completely on the ability of the test demodulator and decoder to reject video and AM on the sound carrier. A consumer demodulator/decoder may be more sensitive to this than precision equipment, and may be more advantageous in terms of finding system buzz problems. For this reason numbers obtained by this measurement should be regarded as comparative and not absolute. The performance objective listed here should be interpreted accordingly.

c. Frequency Response of BTSC Stereo Transmission System

Definition: Frequency response of the MTS stereo system is the variation in system gain versus frequency from input of the stereo encoder to the output of the test decoder. The frequency response measurement may include part or all of the cable distribution system.

Performance Objective: The frequency response, peak-to-valley, should be within three dB from 50 Hz to 14 Khz.

Discussion: Frequency response is a common audio measurement used to determine the transparency of the system to signals within the passband of the system. If the frequency response is greater than the three dB recommended then the subscriber may observe less than a satisfactory performance with the system.

d. Separation Measurements of the BTSC Stereo System

Definition: Stereo separation is the difference in output level between the demodulated left audio channel and the right audio channel exclusive of noise when only a left or right audio input channel is supplied to the stereo system including the encoder, distribution system and decoder. Separation is expressed as a voltage ratio in decibels.

Performance Objective: Separation through the total system, to the subscriber's stereo television or decoder of 20 dB from 100 Hz to 8 KHz with a taper of 4 dB to 14 dB at 40 Hz and 14 Khz.

Discussion: The measurement of separation in the BTSC television sound system is by far the most complex. The BTSC system uses the dbx companding system in the (L-R) difference channel to reduce noise. This channel is very sensitive to transmission error. For the system to deliver a high degree of separation, the dbx compressor in the encoder must track the expander in the decoder with a high degree of accuracy. If you wish to measure an encoder with a 35 dB specification, a decoder whose performance has been verified must be used. If you wish to measure an encoder with a 20 dB specification, a high quality consumer decoder may be used.

e. Total Harmonic Distortion of the BTSC Stereo Transmission System

Definition: Total Harmonic Distortion is defined as undesirable harmonic content of a modulating signal that is detected and presented at the output of a detector. With respect to stereo audio, it is the amount of unwanted input related signal.

Performance Objective: At all levels and all frequencies, the measured total harmonic distortion should be 1.0% or less.

Discussion: Total harmonic distortion is a common audio measurement that is one indication of overall audio quality. Distortion levels on audio equipment should easily meet the 1% objective. The test method outlined exercises all aspects of the BTSC transmission system including the noise reduction system.

f. Incidental Carrier Phase Modulation (ICPM)

Definition: Incidental Carrier Phase Modulation (expressed in degrees) is defined as Phase Modulation of the video carrier with changes in video input signal level, as the signal level varies from blanking to reference white (0 to 100 IRE).

Performance Objective: Incidental Carrier Phase Modulation should be kept below three degrees or audio buzz may become unacceptable.

Discussion: ICPM is a phenomenon that can create both audio and video related problems. The most common malady exhibited by ICPM is audio buzz which is caused in a home TV receiver when the ICPM riding on the video is transferred to the audio carrier in the TV's intercarrier detector.

OPERATING PRACTICES

In addition to the actual hardware measurements outlined above, there are operational considerations which must be addressed when installing a BTSC encoder. Once again, the following sections attempt to provide a summary of the extensive discussions included in the complete report. The reader is urged to obtain a copy of the entire document to get the complete story.

a. Interconnecting at Baseband Audio

Discussion: The BTSC composite signal is created by the stereo encoder. This signal contains energy in the band between 50 Hz and about 47 Khz. When SAP is used, energy components are present to about 90 Khz. The aural carrier is frequency modulated by the BTSC composite signal. The deviation sensitivity of the aural carrier modulator must be precisely set to maintain the performance (separation) of the system. The procedure for setting Audio Modulator Deviation describes how to use the Bessel null technique to precisely set levels from the BTSC encoder to the aural carrier modulator.

Performance Objective: The deviation sensitivity of the aural carrier modulator must be matched to the operating output level of the BTSC encoder to within +/- .1 dB. An error of only .28 dB will limit the best achievable separation to 30 dB.

b. Interconnect with CATV Modulators at 4.5 MHz

Discussion: As discussed in the previous section, the generation of BTSC stereo signals requires precise calibration of baseband audio levels. For this reason, it is very common to interface a BTSC encoder in such a manner that only non-critical adjustments must be made at the time of installation and on an on-going routine basis. Such is the case when the encoder contains a built-in 4.5 Mhz subcarrier modulator. In this case the encoder manufacture is performing all critical baseband level adjustments as an internal part of the encoder.

Summary: When interconnecting a BTSC encoder to a cable modulator using a 4.5 Mhz interface, only two adjustments are necessary. One is the 4.5 Mhz interface level which is normally set at 100 mv p-p and the second is the aural carrier level which is typically 15 dB below the video carrier level.

c. Interconnect with CATV Modulators at 41.25 MHz.

Discussion: This interface scheme is used primarily when the modulator being used has no provisions for accepting an external 4.5 MHz input. In many cases modifications are required to the modulator for interfacing with an external 41.25 MHz source. It should also be realized that when interfacing at a frequency of 41.25 MHz, the visual/sound intercarrier frequency spacing will be determined by two different frequency sources. One is the frequency accuracy of the 45.75 MHz video source and the other is the frequency accuracy of the 41.25 MHz audio source. It is important that the spacing between these two carriers be maintained at 4.5 MHz ± 1 KHz.

Summary: This method of interface is preferred over the baseband audio interface however when this scheme is used one must take care to insure that the proper visual/aural amplitude ratio at the output of the modulator is achieved.

d. On-line Checks: Critical parameters of BTSC stereo performance include signal-to-noise ratio, signal-to-buzz ratio, frequency response, stereo separation, and relative phase between the left and right signals. Presented in this section in the complete report are methods for on-line checks of stereo performance without interruption of service. These methods provide qualitative indications of performance or allow subjective evaluation of various parameters. They provide a continuous method of monitoring stereo signals to obtain a reasonable amount of confidence in signal quality.

Summary: Numerical standards for stereo performance are given elsewhere in the NCTA operating practices. The procedures given in this section in the complete report are sufficient only to indicate the occurrence of catastrophic system failures or major changes in stereo performance. They are supplied as an aid to the recognition and troubleshooting of system failures.

e. BTSC Operational Guidelines

Discussion: In most installations, the operation of BTSC stereo encoding may encounter difficulties caused not specifically by the encoder, but rather due to operating conditions. In most cases, the difficulties can be avoided through adherence to a few basic operating guidelines and practices.

This section in the complete report discusses such issues as: satellite video spectral content, encryption systems, commercial insertion, modulator interfaces, ICPM, character generators, phasing of left and right signals,and signal level variations.

Summary: While it is not possible to cover all potential problems, the adherence to several basic recommendations will alleviate the majority of difficulties.

f. Transportation of Stereo Signals Over FM Links

Discussion: Field experience as well as laboratory tests have shown that the delivery of quality stereo television over a CATV system is feasible. The effect of the distribution system itself is slight. The interface between stereo encoder and modulator, though critical, has received much attention and is well under control. Scrambling is fairly well understood, and at least some systems cause little degradation. The transportation of stereo between hubs, satellite receiving stations, and headends, however, has received very little attention.

AML's and FM links are routinely used in CATV systems. When used properly this equipment is nearly transparent to video and audio signals. Care should be taken however, when adapting existing systems to carry BTSC stereo. FM links in particular can cause significant degradation of the stereo signal if improperly used.

Summary: It cannot be denied that it is possible to transmit BTSC stereo as a 4.5 MHz subcarrier added to video over an FM link. By disabling active clamps, checking frequency response, filtering any spectral overflow and reducing deviation the BTSC signal should survive, though bruised with a marginal S/N. The problem is that the FM link is only a small part of a complex distribution system. No reasonable degradation budget could allow for any one part of the system to be so marginal.

By contrast, transmission of separate left and right audio information over individual channels of an FM link is an excellent way to transport stereo and entirely avoid budgeting any degradation to the FM link.

CABLE ERROR BUDGET (Separation)

Definition: A system separation budget is a calculation of the expected stereo separation through the entire cascade of headend, transportation link, distribution equipment and cable. The budget calculation is based on the required performance of the individual pieces of equipment. Measurements can be performed on the individual pieces of equipment to evaluate suitability, or to initially decide on numbers for the budget. The complete report provides a detailed description of a technique employing "Generalized Error Coefficients" to determine the separation of a cascade of devices once the separation is known for each individual piece.

Performance Objective: A minimum of 24 dB of separation (worse case) should be measured using a laboratory quality receiver and stereo decoder at the output of a complete system. This system performance, combined with a typical consumer decoder separation of 22 dB will provide a worse case separation of 17 dB (typical separation of 21 dB) to the subscriber.

CONCLUSION

BTSC Stereo, having been "protected" by the FCC in 1984, is now the de-facto standard in the United States. To date, more than 20% of the television receivers in use in the U.S. have the capability of receiving stereo audio through built-in decoders. This number will rise to 50% in the next few years.

Cable operators, slow to react initially because of low stereo receiver penetrations, are now providing BTSC stereo on satellite delivered channels in ever increasing quantities. It is not uncommon to see as many as 12-15 channels of BTSC stereo in many cable headends.

Obviously, the quality of this audio service becomes critical to the overall customer satisfaction when viewing channels providing BTSC stereo. The intent of this paper is to provide a description of both measurement techniques and operating practices which could provide the basis for equipment evaluations and installation and operating practices to help insure the required level of audio quality. In addition, a concept is introduced utilizing "Generalized Error Coefficients" to allow the calculation of overall separation for a cascade of devices once the separation for each piece is known.

The credit for authorship of the individual sections of the complete report goes to the following individuals: Jim Farmer (Scientific Atlanta), Brian James (NCTA), Ned Mountain (Wegener Communications), Karl Poirier (Triple Crown Electronics), Louis Rovira (Scientific Atlanta), Dave Sedacca (Scientific Atlanta), and Russ Skinner (United Artist Cable Corp.).

CABLE HEADEND POLARIZATION ALIGNMENT CONCERNS
DURING PEAK SUNSPOT CYCLE

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ABSTRACT

The polarization angles of satellite dishes used for cable headends may have more error during the next few years than ever before. This is due to the diurnal variation of incident polarization angle, which is on an upswing recently due to heavy solar activity which should persist and worsen before abating.

A combination of night and day time polarization alignment of cable headend dishes may be preferred by cable operators in some parts of the world where this effect is strongest. This paper describes the phenomena of Faraday Rotation which causes polarization mis-alignment. It identifies regions of the earth where cable headends would find the strongest effect and recommends methods to minimize degradation due to this effect. Prediction data from NOAA (National Oceanic and Atmospheric Administration) is reported, showing the anticipated peak period of polarization rotation to be early 1990.

BACKGROUND

Some satellite users have experienced interference recently on C-band satellites due to a propagation phenomena called Faraday Rotation. Under current conditions of above normal solar activity, uplink and downlink signals are skewed in polarization as they encounter the ionosphere (see Figure 1.). As a result, a signal in one polarization may bleed into the opposite polariza- tion enough to cause interference (cross-pol interference).

Since cross-pol interference is usually caused by incorrect earth station antenna alignment, many satellite uplinkers have recently received close scrutiny in order to determine the source of interference. During this period of high solar activity, it is more difficult to maintain proper antenna polarization alignment for all users because the amount of rotation depends on frequency and time of day.

The present high level of Faraday rotation is anticipated to abate during the summer before reaching a maximum next winter (see Table 1.). The last maximum was reached in winter 1957/58 and it should be a similar period in the future before the next maximum occurs. What follows is a technical description of Faraday rotation with advice regarding ways to prevent or mitigate interference.

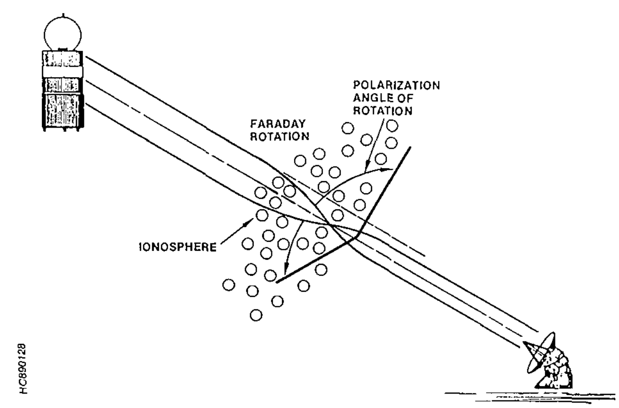

FIGURE 1. POLARIZATION ROTATION

INTRODUCTION

Faraday rotation, a natural occurrence arising from the introduction of ions into the ionosphere via solar radiation, has recently caused interference to some users of C-band satellites. Faraday rotation is the rotation of the electric field vector of a signal as it passes through the ionosphere.

SMOOTHED (OBSERVED AND PREDICTED) SUNSPOT NUMBERS: CYCLES 21 AND 22 (VALUES AFTER APRIL 1988 ARE PREDICTED)												
YEAR	JAN	FEB	MAR	APR	MAY	JUN	JUL	AUG	SEP	OCT	NOV	DEC
1980	164	163	161	159	156	155	153	150	150	150	148	143
1981	140	142	143	143	143	142	140	141	143	142	139	138
1982	137	133	129	124	120	117	115	109	101	96	95	95
1983	93	90	86	82	77	70	66	66	68	68	67	64
1984	60	56	53	50	48	46	44	40	34	29	25	22
1985	20	20	19	18	18	18	17	17	17	17	17	15
1986	14	13	13	14	14	14	14	13	12*	13	15	16
1987	18	20	22	24	26	28	31	35	39	44	47	51
1988	58	64	71	77	84	90	99	108	115	122	127	132
1989	135	139	147	154	161	166	169	172	179	184	185	186
1990	186	186	184	178	172	168	167	164	157	149	142	138

*SEPTEMBER 1986 MARKS THE ONSET OF SUNSPOT CYCLE 22.

HC890124

TABLE 1. (FROM THE JOURNAL OF SOLAR-GEOPHYSICAL DATA, OCT. 1988

The degree of Faraday rotation is now approaching a maximum linked to the long term solar cycle, and interference is arising in cases where it never before existed. The present prediction is that the peak will occur near February 1990, but this date is uncertain. Seasonal and diurnal(daily) variations also occur.

If interference occurs strongest in the middle of the day, and not at all at night, then Faraday rotation may have contributed to the interference.

There is no procedure known which eliminates Faraday rotation. What follows are some recommendations to satellite users of ways to minimize Faraday-induced interference.

RECOMMENDED PROCEDURES TO MITIGATE EXCESS INTERFERENCE

If you have not yet created or experienced any interference due to Faraday rotation, it is possible that you will neither create nor experience any such interference in the future. In this case there is no need to adjust any antennas.

If you have created or experienced interference due to Faraday rotation, or if you wish to take precautionary measures, we recommend following these guidelines.

a. For Fixed Transmit-Only Antennas

The polarization orientation of a transmit-only antenna should be adjusted to be midway between the extreme positions of proper uplink polarization angle (midway should be reached by rotating the top of the polarizer to the East from its nominal position). (See section 4.a for a step-by-step procedure.)

b. Fixed Receive-Only Antennas

The polarization orientation of a receive-only antenna should be adjusted to be midway between the extreme positions of proper downlink polarization angle (midway should be reached by rotating the top of the polarizer to the West from its nominal position). (See section 4.b for a step-by-step procedure.)

c. For Fixed Transmit-Receive Antennas

The polarization orientation of a transmit-receive antenna should be adjusted to be midway between the extreme positions of proper uplink polarization angle (midway should be reached by rotating the top of the polarizer to the East from its nominal position). (See section 4.a for a step-by-step procedure.)

d. For Transportable Uplinks

If the uplink is for a short period (hours vs. days), then present day practices are unchanged. If the uplink is for a period of a few days, then follow the guidelines for a fixed uplink (3.a).

STEP-BY-STEP PROCEDURES

a. For Fixed Transmit-Only Antennas

Step 1. Conduct Night and Noon X-pol Checks. Contact your satellite operator to schedule a nighttime cross-pol (X-pol). Check for a time when it will be dark at the uplink site and a subsequent daytime X-pol check near local noon.

Step 2. Conduct Night X-pol Check. It should be dark during the first X-pol check so that the nominal polarization orientation of the dish may be determined. Mark the nominal polarization angle thus determined.

Step 3. Conduct Noon X-pol Check. Repeat the X-pol check near midday to determine proper noon polarization. Mark the noon polarization angle thus determined. Before ending noon X-pol check, adjust antenna polarization to the point midway between the nominal and noon polarization marks, checking to see if this yields a satisfactory value of X-pol.

On some days the amount of rotation will be higher than other days. Peak rotation for a given day is difficult to impossible to predict. Therefore, if interference persists, it may be required to perform the foregoing procedure again.

b. For Fixed Receive-Only Antennas

Step 1. During dark hours adjust polarization to nominal and mark. (Two port systems should adjust by nulling the opposite polarization.)

Step 2. Near noon readjust polarization to noon position and again mark.

Step 3. Final antenna polarization adjustment should be midway between the nominal and noon polarization marks.

On some days the amount of rotation will be higher than other days. Peak rotation for a given day is difficult to impossible to predict. Therefore, if interference persists, it may be required to perform the foregoing procedure again.

c. For Fixed Transmit- Receive Antennas

Step 1. Follow procedure outlined in 4.a.

Step 2. Observe receive signals for cross-pol interfer- ence. If there is none, the antenna is adjusted properly. If there is cross-pol interference, again adjust cross-pol near local noon and rotate antenna polariza- tion angle from noon mark towards nominal mark (top of polarizer from West to East) until acceptable cross-pol is achieved. Mark as minimum noon position.

The unfortunate fact that the adjustments for transmit and for receive during extreme rotation are opposite, forces the above compromise, which insures adequate uplink cross-pol discrimination while achieving the best possible simultaneous downlink cross-pol discrimination.

TEMPORAL VARIATION OF ROTATION

Diurnal, seasonal and long term variations are such that the night-time rotation is approxi- mately the minimum rotation (nominal) - see Figure 2. Maximum rotation varies a great amount each day during concurrent seasonal and long term maxima. Considering this, one would like to align one's antenna's polarization using the step-by- step procedures previously outlined on a very active solar day, in order to not have to repeat the procedure.

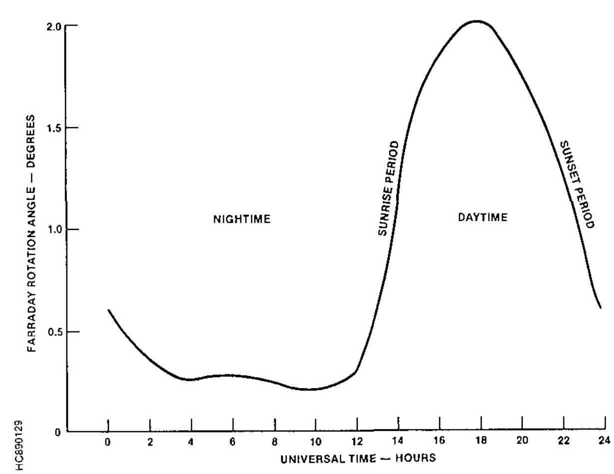

FIGURE 2. TYPICAL DAILY VARIATION OF POLARIZATION ANGLE (1982, NEW YORK @ 4 GHz)

Seasonal and diurnal variations are shown in Figure 3.

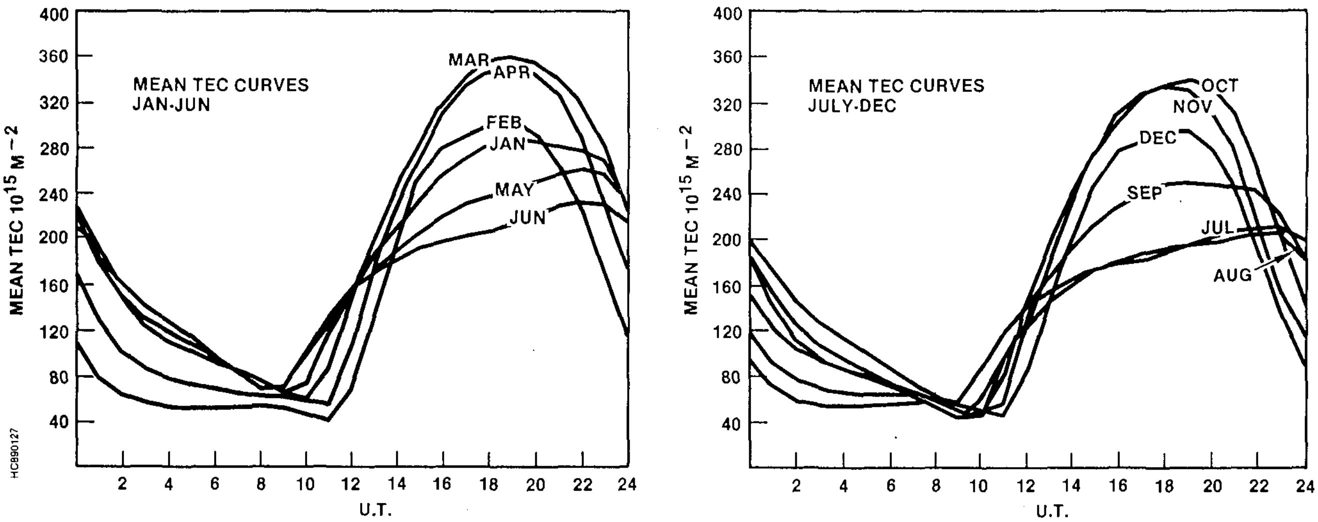

FIGURE 3. DIURNAL VARIATIONS IN TEC, MEAN MONTHLY CURVES FOR 1967 TO 1973 AS OBTAINED AT SAGAMORE HILL, MA USING 136 MHz SIGNALS FROM ATS-3 (AFTER HAWKINS AND KLOBUCHAR, 1974)

DIRECTION OF ROTATION

The downlink rotation experienced in the northern hemisphere is counterclockwise when looking towards the satellite (see Figure 4.). The uplink rotation experienced in the northern hemisphere is also counterclockwise when looking towards the satellite. In the downlink case a counterclockwise correction (top of the feedhorn to the East) is required. In the uplink case a clockwise correction (top of the feedhorn to the West) is required.

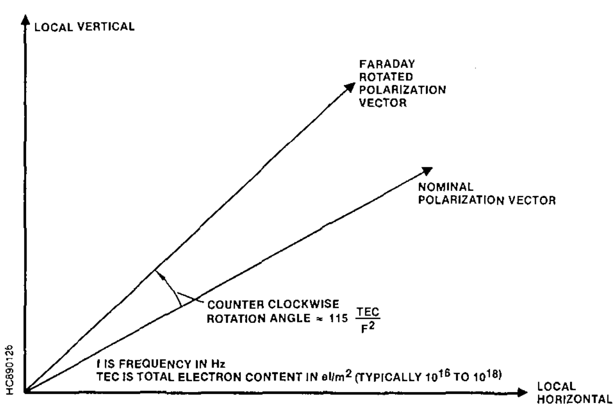

FIGURE 4. FARADAY ROTATION (SATELLITE DIRECTION IS INTO THE PAPER)

FACTORS DETERMINING DEGREE OF ROTATION

The degree of Faraday Rotation primarily depends on: the frequency of signal transmission, the ionization level of the ionosphere, the angle between the direction of propagation and Earth's magnetic field, Earth's magnetic field strength where the signal passes through the ionosphere, and the signal path length through the ionosphere.

The resultant crosspolarization discrimination as a function of rotation angle is shown in Figure 5.

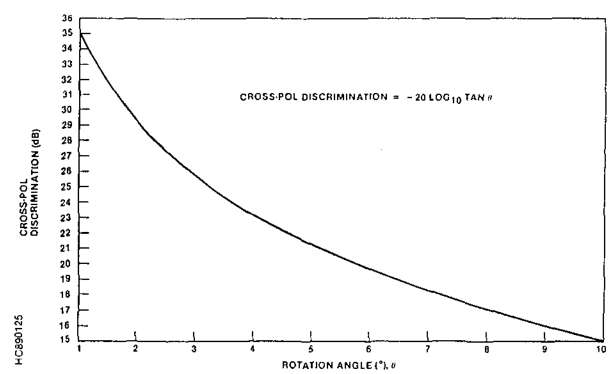

FIGURE 5. CROSS POLARIZATION DISCRIMINATION VS. MIS-ALIGNMENT ANGLE (i.e., FARADAY ROTATION ANGLE)

FREQUENCY DEPENDENCE

Polarization rotation is inversely proportional to the square of transmitted signal frequency (e.g. the relative rotation at twice a certain frequency is one-fourth of the amount found at the lower frequency). If an antenna is aligned at one end of the frequency spectrum, it may be misaligned at the other. Therefore, care should be taken to adjust for the proper polarization orientation at the frequency which is to be used, or at a middle-frequency of the signal frequencies used by an antenna. The polarization orientation disparity for transmit-receive antennas (due to differing transmit and receive frequencies) requires a slight sacrifice of receive polarization discrimination in order to insure (by adjusting for optimal transmit polarization orientation) that interference is not caused by the transmit operation of such an antenna.

TELEMETRY CARRIER AS RECEIVE POLARIZATION REFERENCE

A particularly pure polarization reference for receive-only antenna polarization alignment is the satellite telemetry carrier located between 4,198 MHz and 4,200 MHz (in the horizontal polarization for the Galaxy Satellites). A signal may be present at the same frequency as the telemetry carrier but in the opposite polarization (trans ponder 24), necessitating careful use of a spectrum analyzer in order to null properly. Frequency correction for proper polarization orientation is important when using the telemetry carrier for initial alignment if the antenna is to be used to receive signals at the lower end of the receive band (near 3,700 MHz) as the telemetry carrier is located at the uppermost extreme end of the receive frequency band.

CONCLUSION

Faraday rotation has become a concern to satellite users. Its impact can be minimized by the method described. The best predictions indicate that this rotation will be at its worst near February 1990. The physics of Faraday rotation is well understood (see references) and the variables which determine the degree of rotation are predictable but there is significant uncer- tainty in these predictions and the variables themselves are impossible to control.

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CATV SYSTEM POWERING CONSIDERATIONS

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ABSTRACT

The input AC Power Factor of a DC power supply has a significant effect on input AC current, and therefore affects system design. Since system efficiency and Power Factor can vary significantly as a function of system configuration, analysis of system powering and overall efficiency requires a realistic system model which accommodates the wide variety of modern system designs, re-designs, and upgrades.

This paper will discuss the effects of system configuration, Power Factor, and AC Line Power Supply performance on overall system efficiency. A unique system simulator which permits accurate modeling of all known system configurations will be described, along with test data obtained. A simplified method of system design and performance analysis which was developed using the system simulator will be presented.

INTRODUCTION

Power supply bench-testing has traditionally been done by connecting the unit under test either directly to an otherwise unloaded AC Line Power Supply (LPS), or through an autotransformer to simulate lower voltages commonly encountered in CATV systems. These testing methods do not place a normal load on the LPS, nor do they account for cable resistance normally encountered in CATV systems. In an attempt to more accurately model conditions encountered in real systems, series resistance was added to the output of the LPS to simulate cable resistance: A significant decrease in AC input current was observed over the entire operating voltage range of the DC power supply.

Actual cable resistance and LPS loading can vary widely depending on the type of cable and amplifier technology, subscriber density, and general system architecture. I realized that the most practical way to establish accurate performance guidelines for system design would be to simulate an entire powered segment of a CATV system, and measure actual performance of a variety of system configurations. Instead of using real cable as in cascade measurements, resistors would be used to simulate cable resistance. By using variable resistors, a wide variety of span lengths and cable characteristics could be easily simulated. Any number of trunk stations and line amplifiers could be used. Provisions would be made for the connection of accurate current and power metering equipment anywhere within the segment. The most useful metering locations would prove to be at the input and output of the LPS and at the input of each trunk station and line extender. Metering the input and output of the LPS would also provide a test of its performance under a variety of realistic conditions. The data gathered using the system simulator could then be applied to a general rule for system power design using switched-mode power supplies.

CHARACTERISTICS OF SWITCHED-MODE POWER SUPPLIES

Compared to the traditional series-regulated DC power supply, the most distinguishing characteristic of a modern switched-mode power supply (SMPS) is its ability to operate at relatively constant efficiency across a wide input voltage range. The efficiency and input characteristics of any AC-input DC power supply are defined by the following equations:

$$\text{Eff} = \frac{V(out) \times I(out)}{P(in)} \qquad (1)$$

$$P(in) = V(in) \times I(in) \times PF \qquad (2)$$

Where:

Eff = power supply efficiency (%/100)

V(out) = DC voltage in Volts

I(out) = DC current in Amps

P(in) = AC power in Watts

V(in) = AC voltage in Volts RMS

I(in) = AC current in Amps RMS

PF = input AC Power Factor

Note that unless the AC Power Factor is known, it is not possible to solve for Efficiency unless accurate power measuring equipment is available. Certain "standard" Power Factor values have been used for years in conventional power supply design, but as will be shown, Power Factor can be a significant variable.

Using specialized equipment, it is possible to simultaneously measure AC power, RMS voltage, RMS current, and Power Factor at the input of a power supply. Figure 1 is an example of an equipment set-up which has been used to measure the characteristics of DC power supplies used in CATV applications.

Using the equipment set-up shown in Figure 1, characteristics of the Magnavox model 8PS-60HE are shown in Figures 2, 3, and 4. The 8PS-60HE is a ruggedized high-efficiency transformerless switched-mode power supply.

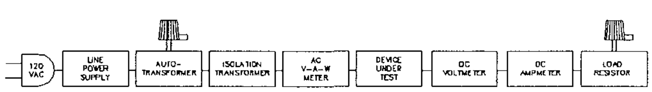


Figure 1. Test Set-Up

As a comparison, the characteristics of the conventional series-regulated ("linear") Magnavox model 7PS-60 are shown in Figures 5, 6, and 7.

Note that compared to the linear power supply, the switched-mode unit maintains excellent efficiency across a wide input voltage and output current range. SMPS Power Factor at 2 Amps output is virtually identical to the linear unit at 1 Amp output.

As shown in equation (2), Power Factor is an important variable which has a considerable effect on AC input current. Equations (1) and (2) can be combined to solve for input current:

$$I(in) = \frac{V(out) \times I(out)}{\text{Eff} \times PF \times V(in)} \qquad (3)$$

Since V(out) and Eff are nearly constant for the 8PS-60HE, equation (3) may be further simplified to:

$$I(in) = \frac{26.7 \times I(out)}{PF \times V(in)} \qquad (4)$$

Use of equation (4) has been suggested for CATV system powering design applications, but its use is difficult for the following reasons:

1. Power Factor varies significantly, not only as a function of V(in), but also as a function of system configuration.

2. AC voltage at a distance from the LPS is a function of current flow in the cable. Since current flow must be estimated until it can be calculated from AC voltage, (and then repeatedly re-calculated based on the new current value) an iterative approach is required for solution.

By using an average value for Power Factor, use of equation (4) may be simplified, but an iterative approach is still necessary for solution.

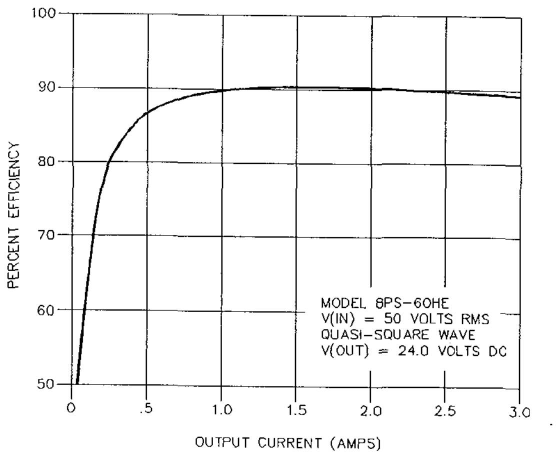


Figure 2.

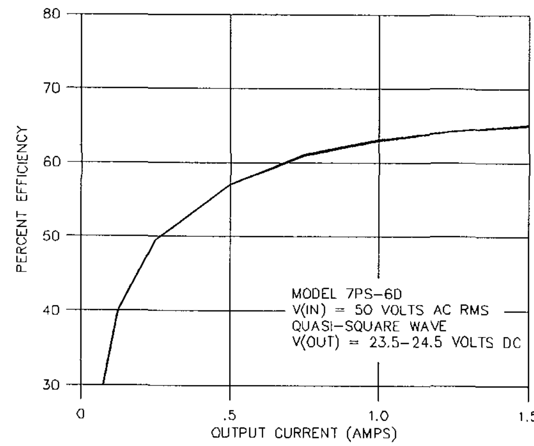


Figure 5.

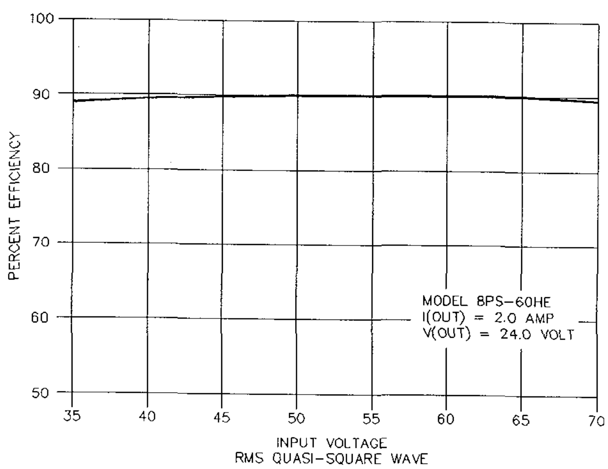


Figure 3.

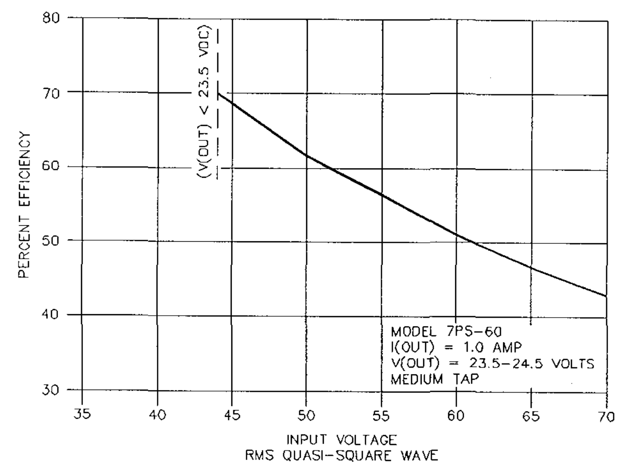


Figure 6.

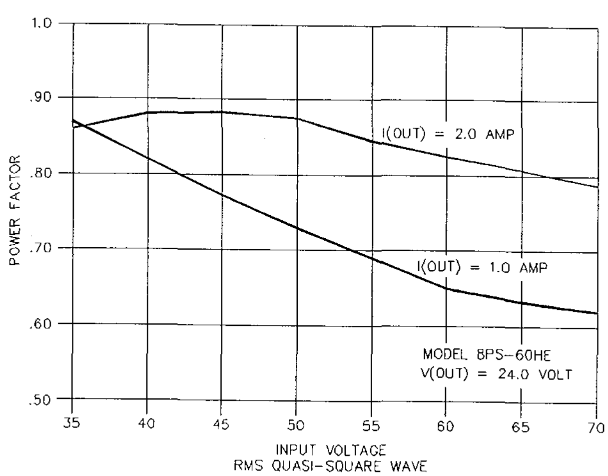


Figure 4.

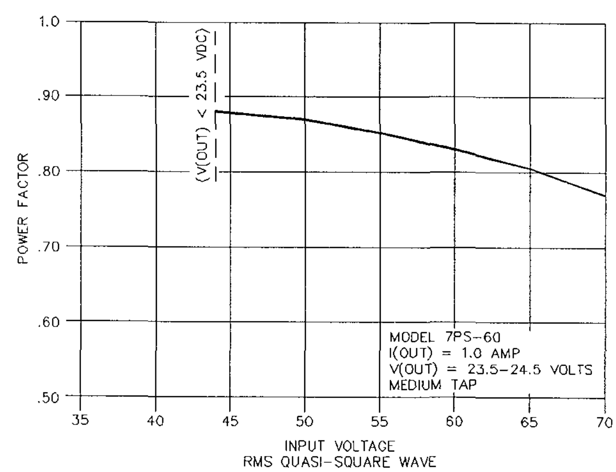


Figure 7.

EFFECT OF SOURCE IMPEDANCE ON SMPS PERFORMANCE

In a CATV system, normal cable resistance would be expected to have an effect on Power Factor. The effect of cable resistance was simulated by adding resistance to the test set-up as shown in Figure 8.

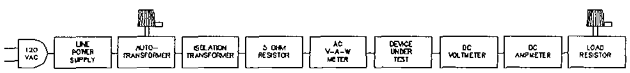


Figure 8. Test Set-Up with 5 Ohms Resistance Added

Using the test set-up in Figure 8 with a simulated cable resistance of 5 ohms, the 8PS-60HE measurements were repeated. A major change in Power Factor was noted, as shown in Figure 9. Small improvements in Efficiency (typically less than .5%) were also noted; these are due to lower AC current.

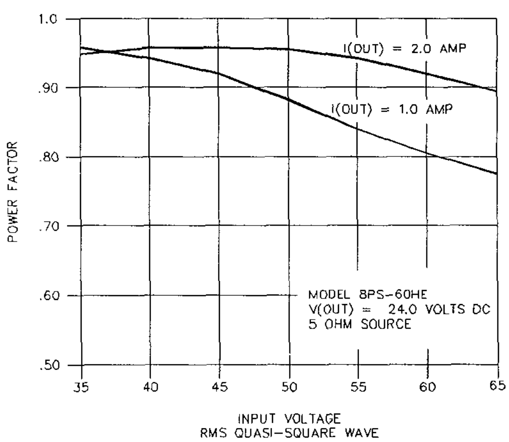


Figure 9.

The selection of a 5 ohm resistor to obtain the data depicted in Figure 9 was based on the typical loop resistance of several spans of cable. The effect of LPS loading was not addressed at all. In a real system configuration, the Line Power Supplies are typically loaded to between 50% and 85% of their current rating, and cable loop resistance is distributed in a network with complex loads applied at many points within the network.

SYSTEM SIMULATOR

To be able to evaluate a variety of system configurations using different types of cable, variable power resistors were used to simulate each cable span loop resistance. An array of 32 line amplifiers was mounted on a wall with the power resistors to permit connection in any configuration. Due to the wide variety of possible trunk amplifier DC load configurations, trunk station loads were simulated using variable load resistors. A standard Magnavox 5-LPS60-14 line power supply delivers up to 14 Amps at 60 VAC to the simulator board. Provisions were made for the insertion of a digital Volt-Amp-Watt meter at the input and output of the LPS, and at the input of each line extender and mainstation DC power supply. The final configuration is shown in Figure 10.

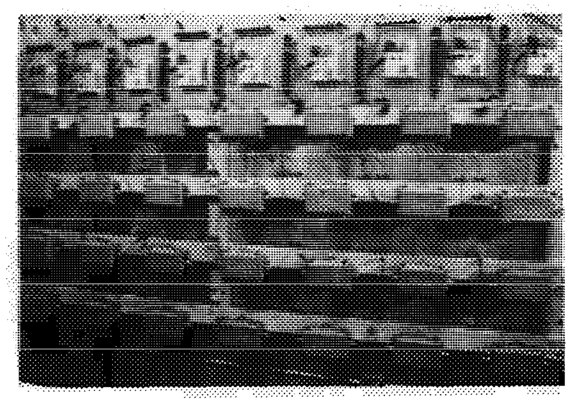
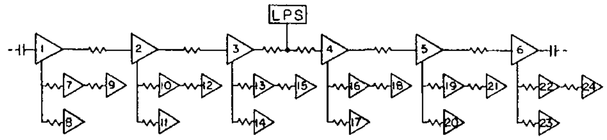


Figure 10. The System Simulator

330 MHz: SERIES-REGULATED vs SMPS

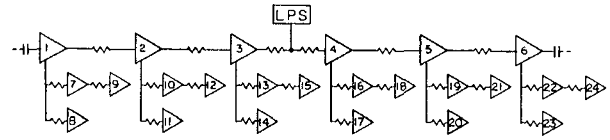
For the first simulation, a typical 330 MHz system segment with conventional power supplies and push-pull amplifiers was built, as shown in Figure 11. This configuration models the use of .750" T4+ trunk cable at 22 dB per span, or 2340 feet; trunk span resistors were set at 1.76 ohms, except for the power insertion location, which was split into two resistors of .88 ohms each. The trunk station 7PS-60 power supplies were loaded at 1.11 Amps DC to simulate the load of a standard trunk amplifier, bridger, and AGC/ASC control module. Standard-gain 5-LEX330-60 line extenders were used. Use of .500" T4+ distribution cable was simulated, assuming a 900 foot span (1.53 ohm) to the first line extender and a 1000 foot span (1.70 ohm) to the second line extender.



LPS Input: 118.0 VAC, 7.11 Amps, 760 Watts, PF = .91
LPS Output: 57.6 VAC, 11.90 Amps, 645 Watts, PF = .94
LPS Efficiency: 85%

Station	VAC(in)	IAC(in)	Power	PF	TAP
1	42.0	1.124	44.0	.93	LO
2	48.5	.970	39.7	.89	MED
3	52.2	.919	40.5	.84	HI
4	52.4	.923	40.7	.84	HI
5	45.5	.963	39.5	.90	MED
6	42.1	1.120	43.6	.92	LO
7	41.1	.350	13.9	.97	2
8	41.7	.338	13.5	.96	2
9	40.5	.331	13.0	.97	2
10	44.9	.295	12.4	.94	3
11	45.2	.300	12.6	.93	3
12	44.4	.304	12.7	.94	3
13	51.3	.337	15.4	.89	3
14	51.8	.328	15.0	.88	3
15	50.7	.334	15.2	.90	3
16	51.5	.350	15.6	.86	3
17	52.0	.340	15.2	.86	3
18	51.0	.335	14.9	.87	3
19	44.8	.304	12.5	.92	3
20	45.2	.298	12.5	.93	3
21	44.3	.294	12.2	.94	3
22	41.0	.345	13.5	.95	2
23	41.5	.350	13.9	.96	2
24	40.6	.347	13.3	.94	2

Figure 11. 330 MHz System with Linear Power Supply



LPS Input: 117.7 VAC, 5.69 Amps, 603 Watts, PF = .90
LPS Output: 58.2 VAC, 9.25 Amps, 498 Watts, PF = .92
LPS Efficiency: 83%

Station	VAC(in)	IAC(in)	Power	PF	TAP
1	46.2	.740	29.8	.87	-----
2	48.9	.782	29.8	.78	-----
3	53.9	.896	29.9	.62	-----
4	54.1	.889	29.6	.61	-----
5	48.8	.780	29.6	.78	-----
6	46.3	.737	29.6	.87	-----
7	45.4	.290	12.8	.97	3
8	45.9	.280	12.5	.97	3
9	45.0	.278	12.1	.97	3
10	48.0	.290	13.3	.95	3
11	48.4	.295	13.5	.94	3
12	47.6	.300	13.6	.95	3
13	53.2	.244	12.3	.95	4
14	53.7	.240	11.9	.92	4
15	53.0	.242	12.2	.95	4
16	53.4	.245	12.4	.95	4
17	53.8	.245	12.0	.91	4
18	53.0	.242	12.0	.94	4
19	48.0	.292	13.5	.96	3
20	48.4	.290	13.5	.96	3
21	47.5	.286	13.1	.96	3
22	45.6	.289	12.4	.94	3
23	45.9	.290	12.8	.96	3
24	45.1	.290	12.3	.94	3

Figure 12. 330 MHz System with Switched-Mode Power Supply

Note that while the output Power Factor of the LPS is .94, its input Power Factor is .91. Power Factors at the DC power supply inputs vary from .84 to .97, with the highest being the farthest from the LPS. Power input to all stations varies considerably due to the nature of the series-regulated power supplies used. The difference between the LPS output power and the total power consumed by all stations (194 Watts) represents power lost in the cable resistance; this total agrees with the total calculated dissipation of the cable resistors. In this case, the total current consumed by the stations agrees exactly with the current output from the LPS. (As will be seen, this is generally not typical.)

For the simulation shown in Figure 12, the configuration set up for the first trial shown in Figure 11 remained unchanged, except that the 7PS-60 power supplies were replaced by 8PS-60HE units. The series-regulated line extenders remained unchanged, except that their transformer tap selectors were reset to reflect the new operating conditions.

Note that while the Power Factors at the 8PS-60HE inputs of stations 1 through 6 have fallen to between .61 and .87, in each case the current input to these stations has also dropped, and power consumption is uniformly low. As a consequence, cable voltage drop is lower, permitting the selection of the next higher voltage tap in many of the line extenders. Current and power consumption of all the line extenders is significantly lower as a result.

In contrast to the first example, note that the total station current consumption is .5 Amp higher than the output of the LPS. This difference is due to the wide range of Power Factors seen in this example. As will be seen, this difference is typical when switched-mode power supplies are used in CATV systems. (Due to current phase differences which are a consequence of Power Factor, the vector sum of the currents is lower than the arithmetic sum.) Total power dissipation in the cable resistance has dropped to 91 Watts, less than half that of the example in Figure 11.

Although the Power Factors in stations 1 through 6 have fallen significantly, the overall Power Factor at the LPS input and output have changed only a small amount. While the Efficiency of the LPS has fallen from 85% to 83% (due primarily to reduced loading), the overall current and power input to the LPS is 20% lower than the original configuration in Figure 11.

450 MHz: SERIES-REGULATED vs SMPS

In the next two simulations, a 450 MHz transportation trunk application will be evaluated: First with 7PS-60 power supplies, then with 8PS-60HE units. Feed-forward trunk amplifiers will be used at 29 dB operational gain, and 1.00" MC[2] cable will be modeled, resulting in a span of 3920 feet and 1.61 ohms loop resistance per span. AGC/ASC control modules are used, resulting in a total DC load of 1.26 Amps per station.

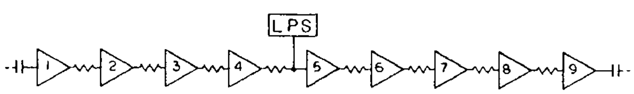
Due to the long spans and lack of distribution, the full output current capability of the LPS unit in Figure 13 cannot be used because of the considerable voltage drop in the cable. This is known as a "Voltage Limited" configuration, and is common in transportation runs and configurations with very low density distribution.

Note that while the Power Factors at the station inputs vary between .84 and .96, the overall Power Factor at the LPS is .95 at the output and .93 at the input. LPS efficiency is excellent at 84%, considering that it is not fully loaded. Total cable dissipation is 141 Watts, and the total station current consumption agrees within 80 mA of the LPS output current.

Figure 14 shows data for 8PS-60HE switched-mode power supplies with the same DC load and physical configuration as Figure 13.

In contrast to the 330 MHz example, Power Factors for Figure 14 station inputs distant from the LPS are equal to those of the conventional power supplies in Figure 13. This is likely due to the considerably higher cable resistance in this configuration. Since the input to station 5 is connected directly to the LPS, its Power Factor is considerably lower. However, the overall .90 Power Factor at the LPS output (and input) is still quite reasonable.

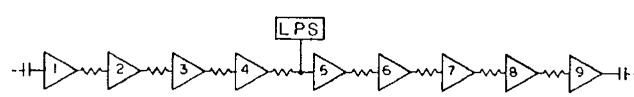
Total power dissipation in the cable has dropped to 63 Watts. Overall LPS power consumption is 30% lower than in Figure 13, and current consumption has dropped 28%. As a result of the low 6.84 Amp loading, LPS efficiency has fallen to 79%; while more savings would be possible if the LPS efficiency was higher, repowering of this configuration for more complete LPS utilization would not be advisable due to considerable cable voltage drop.



LPS Input: 114.8 VAC, 6.19 Amps, 659 Watts, PF = .93
LPS Output: 58.1 VAC, 10.04 Amps, 553 Watts, PF = .95
LPS Efficiency: 84%

Station	VAC(in)	IAC(in)	Power	PF	TAP
1	39.7	1.201	45.6	.96	LO
2	41.8	1.233	48.0	.93	LO
3	45.2	1.047	43.4	.92	MED
4	50.8	.967	43.2	.88	HI
5	57.5	1.066	51.8	.84	HI
6	50.8	.968	43.2	.88	HI
7	45.3	1.056	43.7	.91	MED
8	41.8	1.237	48.2	.93	LO
9	39.8	1.189	45.2	.95	LO

Figure 13. 450 MHz System with Linear Power Supply



LPS Input: 115.9 VAC, 4.43 Amps, 464 Watts, PF = .90
LPS Output: 59.0 VAC, 6.84 Amps, 365 Watts, PF = .90
LPS Efficiency: 79%

Station	VAC(in)	IAC(in)	Power	PF
1	46.9	.753	33.6	.95
2	48.1	.750	33.2	.92
3	50.2	.760	33.1	.87
4	53.7	.828	34.1	.77
5	58.0	.986	33.6	.59
6	53.8	.820	33.7	.76
7	50.2	.770	33.5	.87
8	48.1	.760	33.6	.92
9	46.8	.752	33.6	.96

Figure 14. 450 MHz System with Switched-Mode Power Supply

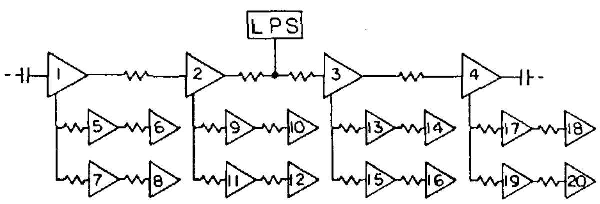
550 MHz APPLICATIONS USING SMPS

The above examples were provided to demonstrate the considerable differences between SMPS and conventional power supplies, and to document the savings possible by upgrading to SMPS units. The remaining examples will show how SMPS use applies to different configurations of 550 MHz systems. All examples will model .750" MC^2 trunk cable and .500" MC^2 distribution cable. A 6-VLE550-SWA Power Doubling SMPS line extender will be used. While the trunk spans and loop resistances will be defined by the type of trunk amplifiers used, for sake of simplicity, all distribution cable will span 600 feet to the first line extender, and 700 feet to the second. Loop resistances of .94 ohm and 1.10 ohm will be used, respectively.

Figure 15 shows a typical system using high-gain Power Doubling trunk amplifiers, for a span length of 2500 feet and loop resistance of 1.83 ohms. Trunk DC load is 1.47 Amps, including trunk, bridger, and AGC/ASC control.

Power Factors range between .68 and .92, and average .84 overall. However, the overall .89 power factor at the output of the LPS is reasonable, and the value of .92 at the LPS input is typical. Note the uniform input power and current for all stations: they appear to be nearly independent of input voltage. Total cable dissipation is 85 Watts, and total station current is .59 Amp higher than the LPS output.

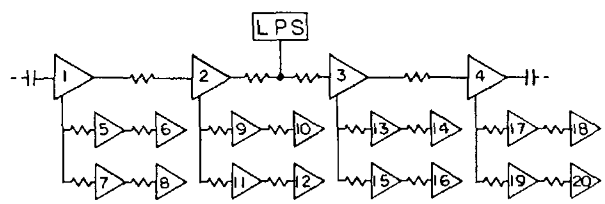
Figure 16 shows another version of the Figure 15 system, except that Feedforward amplifiers are used for maximum span of 2685 feet and loop resistance of 1.96 ohms. Trunk station DC load is increased to 1.96 Amps, and LPS loading is near optimum. Cable dissipation is 126 Watts, and total station current is .4 Amp higher than LPS output. Power Factors are nearly identical to those in Figure 15.



LPS Input: 114.4 VAC, 6.27 Amps, 660 Watts, PF = .92
LPS Output: 58.5 VAC, 10.52 Amps, 548 Watts, PF = .89
LPS Efficiency: 83%

Station	VAC(in)	IAC(in)	Power	PF
1	49.5	.977	38.7	.80
2	53.9	1.093	39.8	.68
3	53.7	1.082	39.5	.68
4	49.5	.987	39.2	.80
5	48.9	.431	18.9	.90
6	48.4	.436	19.3	.92
7	48.8	.435	19.2	.90
8	48.5	.436	19.3	.91
9	53.2	.445	19.3	.82
10	52.8	.437	19.3	.84
11	53.2	.450	19.4	.81
12	52.8	.434	19.2	.84
13	53.2	.433	18.7	.81
14	52.8	.430	19.1	.84
15	53.2	.444	19.1	.81
16	52.8	.429	19.0	.84
17	48.7	.433	19.0	.90
18	48.3	.432	19.2	.92
19	48.6	.430	19.0	.91
20	48.2	.431	19.1	.92

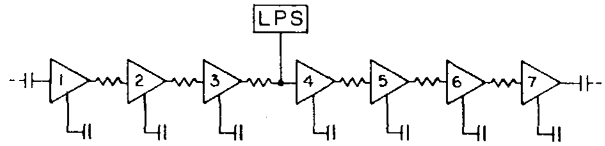
Figure 15. Typical 550 MHz System with Power Doubling Trunk



LPS Input: 114.5 VAC, 7.24 Amps, 763 Watts, PF = .91
LPS Output: 58.2 VAC, 12.4 Amps, 643 Watts, PF = .89
LPS Efficiency: 84%

Station	VAC(in)	IAC(in)	Power	PF
1	46.7	1.307	51.7	.85
2	52.4	1.465	53.2	.69
3	52.3	1.450	52.6	.69
4	46.6	1.313	52.1	.85
5	46.0	.450	18.9	.91
6	45.6	.460	19.4	.93
7	46.0	.462	19.3	.91
8	45.6	.460	19.3	.92
9	51.6	.462	19.3	.81
10	51.3	.454	19.3	.83
11	51.6	.466	19.5	.81
12	51.3	.450	19.2	.83
13	51.6	.448	18.7	.81
14	51.2	.447	19.1	.84
15	51.6	.455	19.2	.82
16	51.2	.444	19.1	.84
17	46.0	.451	19.1	.92
18	46.6	.454	19.2	.91
19	45.8	.451	19.1	.93
20	45.4	.455	19.2	.93

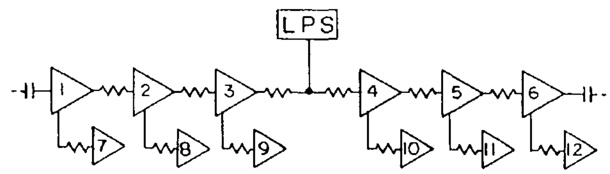
Figure 16. Typical 550 MHz System with Feedforward Trunk



LPS Input: 116.1 VAC, 5.36 Amps, 565 Watts, PF = .91
LPS Output: 58.8 VAC, 8.63 Amps, 460 Watts, PF = .91
LPS Efficiency: 81%

Station	VAC(in)	IAC(in)	Power	PF
1	44.8	1.200	51.3	.95
2	47.3	1.207	51.6	.90
3	51.5	1.262	52.9	.81
4	58.3	1.485	52.0	.60
5	51.5	1.252	52.3	.81
6	47.2	1.221	52.0	.90
7	44.8	1.224	52.0	.95

Figure 17. Rural 550 MHz System with Feedforward Trunk

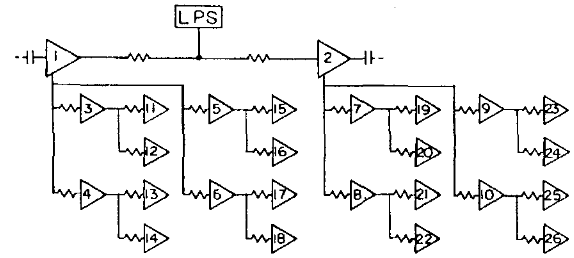


LPS Input: 114.8 VAC, 6.25 Amps, 660 Watts, PF = .92
LPS Output: 58.4 VAC, 10.37 Amps, 551 Watts, PF = .91
LPS Efficiency: 83%

Station	VAC(in)	IAC(in)	Power	PF
1	44.3	1.276	51.2	.91
2	47.0	1.272	51.5	.86
3	53.5	1.423	51.3	.70
4	53.2	1.409	52.6	.70
5	47.1	1.291	52.0	.86
6	43.7	1.296	52.0	.92
7	43.6	.490	19.0	.89
8	46.8	.467	19.4	.89
9	53.1	.466	19.6	.79
10	53.0	.446	18.7	.79
11	46.7	.465	19.1	.88
12	43.4	.488	19.1	.90

Figure 18. Low Density 550 MHz System with Feedforward Trunk

Figure 17 shows a Feedforward system with the very low distribution loading typical of a rural area. This system configuration is similar to a transportation trunk, except that trunk station power consumption is higher. Trunk span and DC load are the same as in Figure 16, but the LPS is not fully loaded due to the lack of line extenders. Station Power Factors cover a wide range from .60 to .95, but LPS input/output Power Factors remain typical. Cable dissipation is 96 Watts, and total station current is .22 Amp higher than LPS output current.



LPS Input: 116.7 VAC, 7.03 Amps, 743 Watts, PF = .91
LPS Output: 58.2 VAC, 12.21 Amps, 619 Watts, PF = .87
LPS Efficiency: 83%

Station	VAC(in)	IAC(in)	Power	PF
1	53.1	1.052	39.0	.70
2	53.4	1.051	39.0	.69
3	52.1	.430	19.5	.87
4	52.2	.437	19.6	.86
5	52.2	.431	19.3	.86
6	52.3	.432	19.5	.86
7	52.3	.432	19.4	.86
8	52.4	.430	19.3	.86
9	52.3	.427	19.2	.86
10	52.4	.427	19.4	.85
11	51.8	.422	19.1	.87
12	51.9	.414	18.9	.88
13	51.9	.424	19.3	.88
14	51.8	.424	19.4	.88
15	51.9	.421	19.1	.87
16	52.0	.419	19.0	.87
17	51.8	.420	19.2	.88
18	51.8	.431	19.7	.88
19	51.8	.430	19.8	.89
20	52.0	.425	19.3	.87
21	51.9	.419	19.0	.87
22	52.0	.426	19.4	.88
23	52.0	.424	19.3	.87
24	52.0	.422	19.2	.87
25	51.9	.419	19.1	.88
26	51.9	.421	19.2	.88

Figure 19. High Density 550 MHz System with Power Doubling Trunk

Figure 18 shows a low density Feedforward system with one line amplifier per trunk station. While station Power Factors are significantly different than in Figure 17, Power Factors at the LPS input/output are typical. Cable dissipation is 124 Watts, and total station current is .42 Amp higher than LPS output current.

Figure 19 shows a high-density system with Power Doubling trunk amplifiers, Power Doubling bridgers, and AGC/ASC control. Trunk span is 2222 feet, loop resistance is 1.62 ohms per span, and trunk station DC load is 1.47 Amps. Overall, line extender Power Factors are very consistent; since input voltages are similar, this is expected. LPS output Power Factor is a bit on the low side of typical, but does not appear to affect LPS input Power Factor significantly. Cable dissipation is 78 Watts, and total station current is .10 Amp higher than LPS output current.

A NEW APPROACH TO SYSTEM POWERING

Initially, it was expected that since the input current of a SMPS unit (at constant load and efficiency) is a function of input voltage and Power Factor as predicted by equation (4), accurate calculation of input current for system powering design could become a complex task. However, since input power is nearly constant for SMPS units with equal loads, equation (2) may be restated to:

$$I(in) = \frac{P(in)}{V(in) \times PF} \quad (5)$$

Where:

$$P(in) = \frac{24 \text{ VDC} \times I(out)}{Eff} \quad (6)$$

For trunk stations, P(in) is calculated from known DC load current and SMPS efficiency specifications; for line extenders, P(in) is specified for each model. For SMPS units, most variation in P(in) is actually due to variations in I(out).

System powering design using equation (5) still requires knowledge of actual Power Factor and an iterative approach. However, system simulator data shows that station AC current is nearly independent of AC voltage. If AC input current is truly constant, system powering design would be greatly simplified.

Referring back to the system simulator data, notice that in each case, Power Factor increases at nearly the same rate as AC input voltage decreases. The increase in Power Factor due to the effects of cable resistance demonstrated by the test configuration in Figure 8 is complemented by a decrease in voltage due to the same cable resistance. It follows that equation (5) may therefore be further simplified to:

$$I(in) \approx \frac{P(in)}{K} \quad (7)$$

Where K = SMPS system power constant

While a value for "K" could be derived using equation (7) and measured data for I(in) and P(in) for each station, errors due to Power Factor related current phase errors can be resolved by using measured data for LPS output current and total station power consumption (excluding cable loss) according to the following:

$$K = \frac{P(in) \text{ total}}{I(out) \text{LPS}} \quad (8)$$

Using equation (8) and data from SMPS trial runs shown in Figures 12 through 19, the following "K" values have been calculated:

FIGURE	CONFIGURATION	"K"
12	330 MHz upgrade to SMPS (trunk-only)	44
14	450 MHz transport upgrade to SMPS	44
15	550 MHz typical Power Doubling	44
16	550 MHz typical Feedforward	42
17	550 MHz ultra-low-density Feedforward	42
18	550 MHz low-density Feedforward	41
19	550 MHz high-density Power Doubling	44

Notice that most "K" values are within a few percent of each other. If an average "K" value of 43 is used for 60 VAC system powering design using equation (7), for typical cases calculated LPS output current agrees within 3% of measurements.

SUMMARY

The "system simulator" addresses two issues which affect all systems using switched-mode power supplies: the magnitude and effect of Power Factor on system performance, and the simplification of system powering design.

System Power Factor as measured in the cable and at the LPS output is only slightly lower than that for a conventional system, even though SMPS Power Factor differs significantly from that of a conventional series-regulated unit. Furthermore, there is no significant difference in Power Factor at the 120 VAC LPS input.

The existence of significant Power Factors in CATV systems using switched-mode power supplies complicates the already complex task of system design using these devices. The system powering approach proposed in this paper takes advantage of the "constant input power" characteristic of the switched-mode power supply, and avoids dealing directly with Power Factor. The new design approach is simpler and more accurate than previous methods.

In a simple upgrade from linear to SMPS units, savings in overall power consumption are substantially higher than the improvement in power supply efficiency alone, due to the sizable reduction in cable losses. Overall, SMPS units are far better suited to CATV use than linear types due to their high efficiency, constant input power characteristic, and their capability for operation across a wide input voltage range.

ACKNOWLEDGEMENTS

I would like to acknowledge the guidance provided to me by Dieter Brauer in the preparation of this paper, to the Magnavox Systems Design Group, and to Jerry Monroe for his assistance in the construction of the System Simulator.

CLI MEASUREMENTS FOR LARGE SYSTEMS

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ABSTRACT

The following paper reviews Cumulative Leakage Index requirements and considers leakage program organizational methods and procedures as they relate to large systems.

Both ground-based Cumulative Leakage Index formulas were analyzed revealing the lack of any system size allowances. Flyover advantages and limitations were compared with system leakage strategy. Its intent is to emphasize the need for immediate planning which will ease the ordeal of passing the first FCC filing due July 1, 1990. Ultimate failure may result in a severe loss of channels.

LARGE SYSTEM CONSIDERATIONS

Large cable systems have an inherent disadvantage when attempting to submit a passing annual Cumulative Leakage Index to the FCC.

There is no mileage adjustment factor included in any of the three methods for collecting and computing a CLI. That means, for example, that a 3000-mile system cannot have any more leaks, which equal or exceed 50 microvolts per meter at a distance of ten feet, than a 10-mile system can if any channels between 108 mHz and 136 (soon to be 137) mHz or 225 mHz to 400 mHz are used.

A review of the following considerations and their possible implementation into your leakage program may make the task of filing a passing annual CLI to the FCC more pleasant, and may reduce the anxiety of the July 1, 1990 deadline.

If you do not presently have an aggressive, routine quarterly monitoring and repair program in place, DON'T EXPECT TO PASS YOUR FIRST ANNUAL CLI. The following list is an example of allowable leaks at various field strengths producing a CLI at the passing threshold:

# of leaks	level	CLI
1000	50 µv/m	63.9
250	100 µv/m	63.98
100	150 µv/m	63.52
1	1600 µv/m	64.08 (failure)

The CLI calculations were derived from this formula:

$$I_\infty = \frac{1}{\emptyset} \sum_{i=1}^{n} E_i^2$$

where:

$\emptyset$ is the fraction of the system cable length actually examined for leakage sources and is equal to the strand miles in the plant;

E_i is the electric field strength in microvolts per meter (µV/m) measured pursuant to Section 76.609(h) 3 meters from the leak i; and

n is the number of leaks found of field strength equal to or greater than 50 µV/m pursuant to Section 76.609(h).

where: $10 \log_{10} I_\infty$ must equal or be less than 64.

More simply stated:

$$CLI_\infty = 10 \log_{10} \left[\left(\frac{\text{plant miles}}{\text{miles monitored}} \right) \text{sum of (leaks}^2) \right]$$

As you can see from this formula, large systems have the same burden of a 1000 leak maximum limit if all leaks discovered emit a field strength of 50 microvolts per meter. As the leak levels increase, the number of allowable leaks decreases exponentially. Therefore, the larger the system, the more advanced planning is required to avoid the last minute panic of how to deal with a system that can not produce a passing CLI.

ROUTINE MONITORING

What is an effective routine quarterly monitoring and repair program? Docket 21006, as adopted in Part 76 of the FCC Rules, states that a sufficient number of vehicles must be equipped with leakage receivers sensitive enough to detect leaks at a field strength of 20 microvolts per meter at a distance of three meters (ten feet) to ensure 100% system coverage every three months. Repairs must be made at all locations which meet or exceed the 20 µv/m threshold and all objectionable leaks (an incident where complaints have been made no matter what the level) even though a minimum level of 50 µv/m is used for CLI computation.

Assess the resources you presently have. The monitoring can and should be included in routine daily work activities. Simply outfit a sufficient number of existing vehicles to provide system wide coverage. Large systems should never have a problem devising monitoring strategy, however, dealing with the initial repair backlog quickly develops into a serious demand on labor and material resources.

REPAIR DILEMMA

PREVENTION is the only path to a passing CLI and an eventual "Closed System" in my opinion. Until all appropriate staff are trained to understand the significance of all the tasks they perform on your coaxial cable plant and develop the necessary skills and professional ethics to deliver high quality work standards every time, your leakage repair backlog will remain an unmanageable problem. Properly preparing and tightening every trunk and feeder connector, as well as drop "F" fittings, will show results quickly. Developing cable configurations at poles which avoid frictional damage and fatigue is another beneficial example.

Multiple Dwelling Units are the single most common cause of excessive signal leakage in the majority of cable systems I have examined. Feeder cables usually present the greatest potential for high level leakage since the highest signal levels in a typical cable plant are present. Therefore, feeder cable in apartment houses, hotels, motels, etc. should be constructed with 5/8"x 24 thread ported amplifiers, aluminum cable and directional taps not 50dB gain amplifiers with an "F" fitting at the output port feeding residential splitters via a piece of single shielded RG 59U as an example. Typically, these bulk billed accounts are simply a result of attaching to an existing MATV system. The rules are clear. IF YOUR SIGNAL IS PRESENT ON A CABLE OR PIECE OF EQUIPMENT, IT'S YOUR RESPONSIBILITY. I recommend adopting the rule "aeronautical channel signal levels may not exceed 20 dBmv on any drop type cable", including short jumpers.

EFFECTIVE RECEIVERS

Leakage receivers should be equipped with some type of meter which can be calibrated. The Rules do not require level measurements during quarterly monitoring, however, the benefits justify the cost (especially to larger systems). Time spent during the locating process is shortened since the direction-finding properties of the receiver are enhanced. Repair time is more productive toward CLI reduction because leaks can be sorted by level for priority.

S meters, LCD bar meters, and LCD light bars can be calibrated just as effectively as a meter which reads microvolts per meter directly. The development of a graph similar to the one in fig. 1 will establish the relationship between various signals in dBmv and the appropriate field strengths in microvolts per meter for the monitoring frequency of your receivers. The graph was produced from calculations derived from the formula:

$$\text{dBmv} = 20\ \log_{10}\left(\frac{E\ (\mu v/m)}{\frac{.021\ X\ f\ (mHz)}{1000}}\right)$$

where: f = frequency being measured

Example:

$$\text{dBmv} = 20\ \log^{10}\ \frac{20}{\frac{.021\ X\ f\ (mHz)}{1000}} = -43.95$$

The example above demonstrates that a signal of -44 dBmv will represent a field strength of 20 microvolts per meter when the frequency measured is 150 mHz. The graph illustrated in fig. 1 shows similar relationships for a 150 mHz receiver including field strengths from 20 µv/m to 340 µv/m.

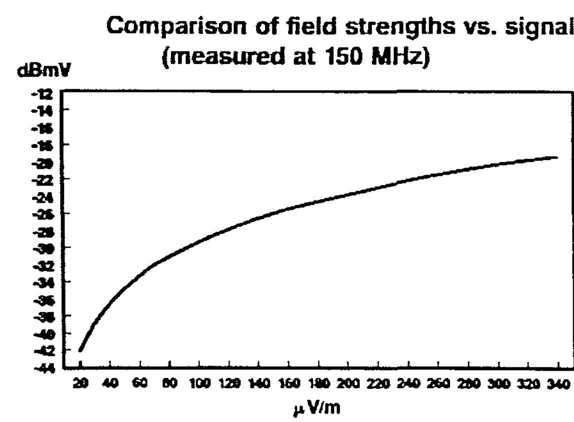
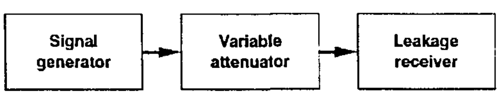


Fig. 1

A test arrangement similar to the one presented in fig. 2 can be used for calibration purposes. A lab quality signal generator is not necessary. Calibration signals from a Sam 1 field strength meter were used for this example along with a 2 meter amateur receiver since both will operate at 150 mHz. We will then assume that channel F will be used for monitoring the hypothetical system. The signal from the generator is fed into a variable attenuator. I recommend that the output of the signal source be adjusted or attenuated to produce 0 dBmv. This will facilitate the readings since a direct relationship will exist between the variable attenuator settings and the negative signals being exposed to the receiver under test.

Since the formula is based on measurements collected from a half-wave dipole, a correlation must be established if an alternative antenna is used. Place an appropriate in-line pad in the test lead to compensate for any antenna loss exhibited when compared to a half-wave dipole. The signal generator output must be adjusted above 0 dBmv an amount equal to the gain factor of any antenna used to maintain the attenuator's direct relationship. Various receiver meter readings are recorded against the field strengths indicated from the graph to serve as a calibration table. Employee participation during this procedure and future routine confirmations will improve their development in learning the relationship between dBmv and µv/m. A sense of confidence with the new skills required develops when various leaks are quantified in µv/m rather than pass/fail. The system's leakage program efficiency will improve parallel to the learning curve of its participants.



Test setup to determine signal level

Fig. 2

To confirm the proper operation of leakage receivers, build a "test zone" at some frequently visited location such as the company gas pump or parking lot entrance. A typical example is the use of a half-wave dipole mounted on a pole fed by signals from your system and adjusted to produce a 20 µv/m leak at a specific location painted on the parking lot pavement. Regular visits to that spot will develop operator confidence in the correct operation of their leakage receivers and save valuable monitoring and repair time avoiding the use faulty equipment.

LOGGING

Prior to the development of logging procedures, careful consideration to some aspects of these records may mean the difference between passing or failing your first CLI, even though you may have collected identical data.

It is advantageous to divide a system for leakage calculations whenever legally possible. John Wong has stated at recent NCTA seminars on leakage that sections of cable systems fed by a separate headend, microwave signals, or fiberoptic cables that are not connected by coaxial cable in any way may be considered as separate systems for the purpose of CLI computation. By no means am I giving permission to make divisions of your system based on any of these criteria, however division is worth consideration with interpretation confirmation from the Cable Branch of the FCC if necessary.

Computation based on the:

$$I_{3000} = \frac{1}{\emptyset} \sum_{i=1}^{n} \frac{E_i^2}{R_i^2}$$

where:

$$R_i^2 = r_i^2 + (3000)^2$$

r_1 *is the distance (in meters) between the leakage source and the center of the cable television system;*

$\emptyset$ *is the fraction of the system cable length actually examined for leakage sources and is equal to the strand miles of plant tested divided by the total strand miles in the plant;*

R_1 *is the slant height distance (in meters) from leakage source i to a point 3000 meters above the center of the cable television system;*

E_1 *is the electric field strength in microvolts per meter (µV/m) measured pursuant to Section 76.609(h) 3 meters from the leak i; and*

n *is the number of leaks found of field strength equal to or greater than 50 µV/m pursuant to Section 76.609(h).*

may provide some relief since the slant height is used to determine the effect of individual leaks upon aircraft. Larger systems potentially offer greater advantages since the distance adjustment may be significant. In order to use this formula, distances from each leak to the theoretical center of the system must be added to the list of information collected while on leakage patrol. Distance averaging may be used as long as an advantage is not derived from its use. The advance development of distance contour lines about the system center and the use of a common distance for all leaks within each contour band may provide a useful tool to ease collection of this information.

Without proper logging of leakage monitoring and repair activities, all your efforts will have been wasted when faced with an FCC inspection. The responsibility of proper documentation is to ensure a "Good Faith Effort" can be demonstrated. Large systems may have the advantages of various computer aided service programs. Inclusion of leakage logging and CLI computation tasks into these programs may streamline the process.

FLYOVERS?

Finally, you must deal with the question, "Do I use aircraft to determine system cumulative leakage?". Consideration of the following information may make that decision easier.

The FCC considers a CLI as a snapshot of your system to determine its total cumulative contribution to potential aircraft interference. Recent question and answer sessions have indicated that they would like the annual CLI pass of the system to take about two weeks with a four week maximum. System size coupled with available staff may make a flyover the only practical means to accomplish this task within that time frame.

Additionally, a ground based CLI must cover a minimum of 75% of your system. Both the I_∞ and the I_{3000} formulas contain correction for partial rideouts, therefore it may be a disadvantage to miss up to 25% of the plant. Inaccessible easements, rough terrain, fenced property, etc. may make sufficient coverage impractical.

The simple estimate of cost to conduct a quick pass of the system and compute the data when compared with a flyover estimate may make the decision for you. Distributing flyover costs among neighboring systems should also be factored within these estimates.

It is very important to remember that a system that cannot pass a ground-based CLI calculation, due to unrepaired quarterly monitoring backlogs, is unlikely to pass a flyover. Be sure the system is tight before spending time and money in aircraft.

LEAKAGE IS A MANAGEMENT PROBLEM

You must gain the support of management to provide the labor and material resources necessary to accomplish this task. Without a cooperative effort, the expectation of failing your first CLI is almost certain. Managers take note. The Cable Branch of the FCC has clearly stated that submission of a failing CLI on or before the July 1, 1990 deadline requires the immediate voluntary shut down of all channels within the aeronautical bands until a passing CLI submission can be produced. Upon receipt of the failing report, the FCC will dispatch an inspector to confirm the discontinuance of their use and report any noncompliance conditions which may result in the possible assessment of fines.

In conclusion, plan the implementation of an aggressive, routine leakage monitoring and repair program including an annual CLI computation strategy or a plan for the operation of a 20-CHANNEL SYSTEM. The choice is yours. If you have not already started, it may be too late.

REFERENCES

FCC Rules, Part 76.

COST AND PERFORMANCE COMPARISON OF FIBER OPTIC CATV SUPERTRUNKS UTILIZING FM AND DIGITAL TRANSMISSION TECHNIQUES

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ABSTRACT

This paper discusses practical implementation of cable TV supertrunks utilizing FM and digital transmission techniques carried on single mode fiber optic links. Architectures and link budgets are discussed, along with cost and performance comparisons.

Noise and distortions inherent in FM and in digital techniques are analyzed. Those factors necessary for good FM performance over a fiber link are considered. In the digital domain, the performance factors of the digital converters are presented. The problems encountered in utilizing video distortion test equipment to evaluate digital systems are reviewed. Finally, some projections for future developments are described.

I INTRODUCTION

A CATV trunk system made up from today's fiber optic components enjoys a number of important advantages over conventional coaxial or microwave links. Chief among these are long transmission paths without amplifiers, no leakage from or ingress into the cable, very wide bandwidth, increased security, and low maintenance costs (all active electronics indoors). Fiber optic cable is exceptionally reliable and impervious to environmental effects like rain fade. The available bandwidth is presently limited by the electronics at either end of the fiber, not the fiber itself. Future advances in transmission techniques will utilize this bandwidth to carry more channels. The cable is suitable for aerial or buried installation, and is not limited by licensing or line-of-site restrictions. CATV super-trunk systems utilizing both FM frequency division multiplexing (FM-FDM) and digital time division multiplexing (TDM) are now coming on the market.

II SUPERTRUNK OPTICAL COMPONENTS

In recent years, considerable progress has been made in theoretical and practical development of single mode (SM) glass fiber, semiconductor lasers, silicon photodiode (PIN) receivers, and avalanche photodiodes (APD) receivers. The characteristics of the components must be understood so that they may be used effectively. For example, when modulated, a laser chirps (changes wavelength slightly). The velocity of propagation through the fiber varies with wavelength (chromatic dispersion). It is desirable to operate a SM fiber at the point that has the most constant velocity vs. wavelength. This minimum dispersion generally occurs at about 1310nm. Proper selection of the laser-fiber-receiver combination will result in nearly zero chromatic dispersion, and in signal attenuation of 0.3 to 0.5 dB per kilometer. Operating at or near the zero dispersion wavelength of SM fiber results in pulse rise times on the order of 0.5ns, and bandwidths in excess of 1GHz. This combination of components and operating parameters yields adequate performance for both FM-FDM and digital TDM supertrunks.

Figures 1a and 1b, laser transfer characteristic, illustrate how the device is light intensity modulated by varying the drive current. This is the case for FM, digital, and for AM modulation. Optical output is proportional to drive current; for FM-FDM modulation, the laser diode must be operated in the linear portion of figure 1b. The DC current bias point, I_b, and the variation of drive current, ΔI, must be controlled to minimize intermodulation products.[1]

III FM SUPERTRUNKS

FM supertrunk equipment is on the market today that can transmit 16 or more channels of video and associated audio on one single mode fiber. With proper design, RS-250B video specifications can be achieved with cable lengths up to 40km. Proper design means selection of FM deviation and channel spacing to minimize second and third order intermodulation products, and to achieve acceptable signal

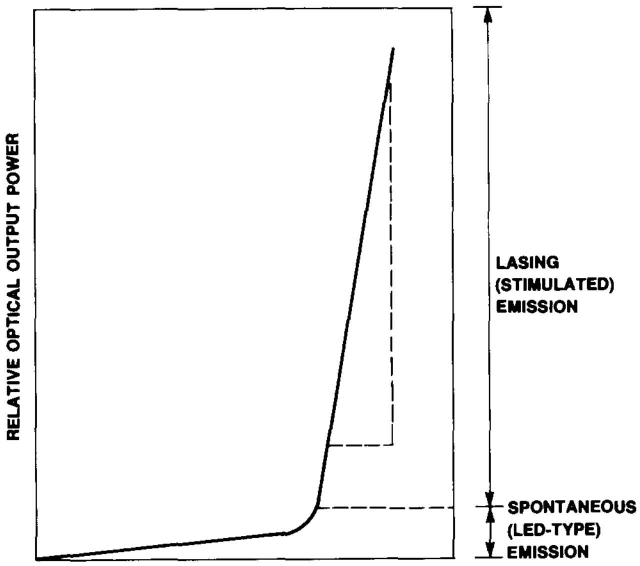


FIGURE 1A

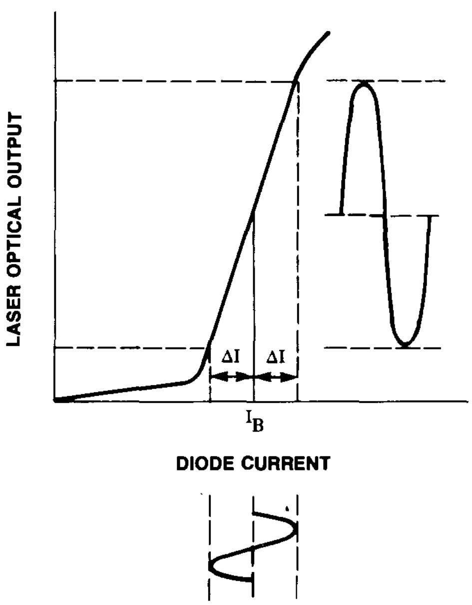


FIGURE 1B

FIGURE 1 LASER TRANSFER CHARACTERISTIC

to noise ratio.

IIIa Intermodulation Distortions

The second order distortions predominate in laser diodes. These may be minimized by a frequency plan proposed by Simons[2]; in this plan the channel frequencies are described by:

$$F_{ch} = f_s/2 + (n \times f_s)$$

where F_{ch} = channel frequency [MHz]

f_s = frequency spacing [MHz]

n = channel number (integer)

The second order products have the form

$$f_{im2} = m \times f_s$$

where m = integer

As a result, the second order products fall between channels; with proper selection and application of the laser diode and receiver, third order products are far enough below the desired signal to be acceptable.

It can now be seen that the wide bandwidth available in a properly designed fiber optic FM system can be used to advantage. This is illustrated in Figure 2, FM supertrunk frequency plan. This plan utilizes over 700MHz of bandwidth to carry 16 wide deviation FM video channels. The channel spacing in governed by the equations above so that second order products fall between channels. This frequency plan also affords adequate adjacent channel protection ratio as described by Gysel.[3]

IIIb Signal to Noise Ratio

Figure 3 illustrates the triangular spectrum of random noise present in an FM system. Due to this characteristic of FM , widening the transmission bandwidth improves the signal to noise ratio. Further improvement can be gained by utilizing CCIR pre- and deemphasis.[4]

By proper selection of the single mode fiber, laser diode source, and diode receiver, a CNR on unmodulated carriers of 34 to 36db is practical. This presumes a reasonable optical loss budget, which will be discussed later. CNR can be measured in the lab with a spectrum analyser. Video S/N can be calculated from C/N from:[3]

$$SNR = CNR + 12db + 20 \log D_{stpw}$$

where SNR = CCIR weighted video SNR, referenced to 100 IRE

D_{stpw} = sync tip to peak white deviation

note: 12Db is gained by pre- and deemphasis

With a measured CNR of 34DB and a deviation of 8MHZ, the calculated SNR is 64DB. This agrees with lab measurements of video SNR with a Rohde & Schwarz noise meter of 63 to 65DB.

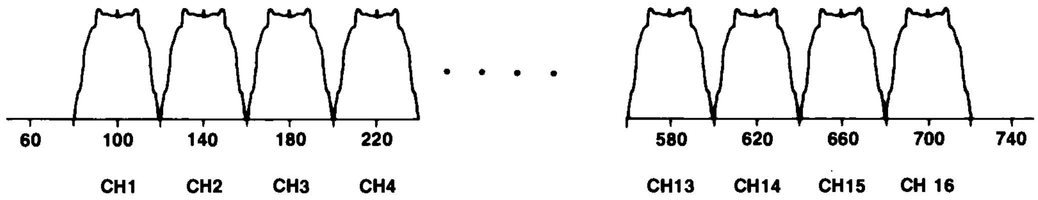


FIGURE 2

FM SUPERTRUNK FREQUENCY PLAN

Schwartz discusses modulation index, FM noise spectrum, and noise improvement in wideband FM in chapter 6.[4]

IIIC Audio Carriage

The stereo (BTSC format) audio program is carried along with the video by wideband FM subcarrier techniques in today's FM supertrunk. The required bandwidth is about 300khz for each BTSC encoded stereo pair. The audio subcarriers may be carried with the associated video carriers, or grouped together in their own portion of the available spectrum. Audio dynamic range better than 60Db is practical, with channel separation of approximately 30db. The audio frequency response is 50hz to 15khz.

IIID Architectures and Link Budget

Figure 4 illustrates one FM supertrunk architecture that carries 16 TV channels with stereo audio from point-to-point. The transmission equipment consists of 16 FM video/audio modulators, an RF combiner, and a laser transmitter. The modulators are driven by baseband audio and video signals, which could come

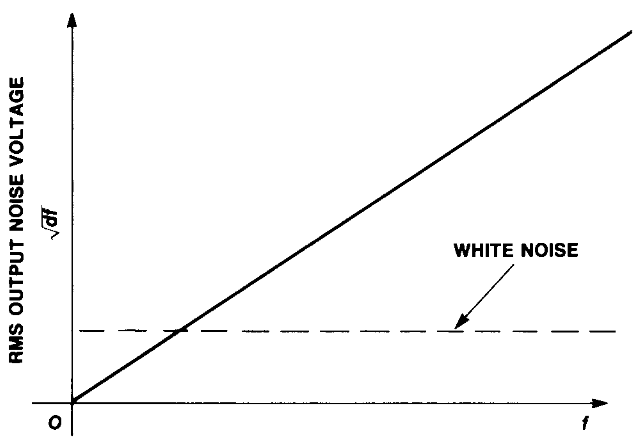


FIGURE 3

SPECTRUM OF NOISE IN FM SYSTEM

from satellite receivers in a typical head end. The SM fiber is likely one of several carried using a loose-tube buffer surrounded by a protective cover.(Figure 5)[10] The receiving equipment consists of the PIN diode or APD receiver with transimpedance amplifier, an RF splitter, and 16 video/audio demodulators. The outputs are at baseband suitable for driving standard AM modulators and BTSC encoders. Thus an 80 channel system could be carried on 5 fibers, with spare fibers, in one cable.

A supertrunk architecture designed to feed two receive sites is illustrated in Figure 6. The optical splitter is a small passive device that typically exhibits a 3DB power split and less than 1db additional insertion loss. At 0.35Db/km, a 4Db loss represents 10km less reach in each leg. An 80 channel system could be carried with additional electronics and one splitter per fiber.

A link budget is based on the following assumptions:

1. The optical power launched by the laser transmitter is -3Dbm, at 1310nm.
2. The semiconductor receiver sensitivity is -25db.
3. The fiber loss is 0.35db/Km at 1310nm.
4. A system margin of 3DB is assumed to allow for component aging and temperature effects.
5. Fusion splice losses are less than 0.1DB and can be ignored for the purpose of a comparative analysis.
6. Connector losses are assumed to be 0.5DB per connector

Figure 7 is an optical power loss model that illustrates the cumulative losses between the laser and the receiver. If P_t is the transmit power in DBm and P_r is the receive power in dbm, then P_l is the total loss budget for the link:[1]

$$P_l = P_t - P_r$$
$$= -3 -(-25) = 22\text{dbm}$$
and
$$P_l = 3xl_c + a_f \times L + \text{system margin}$$

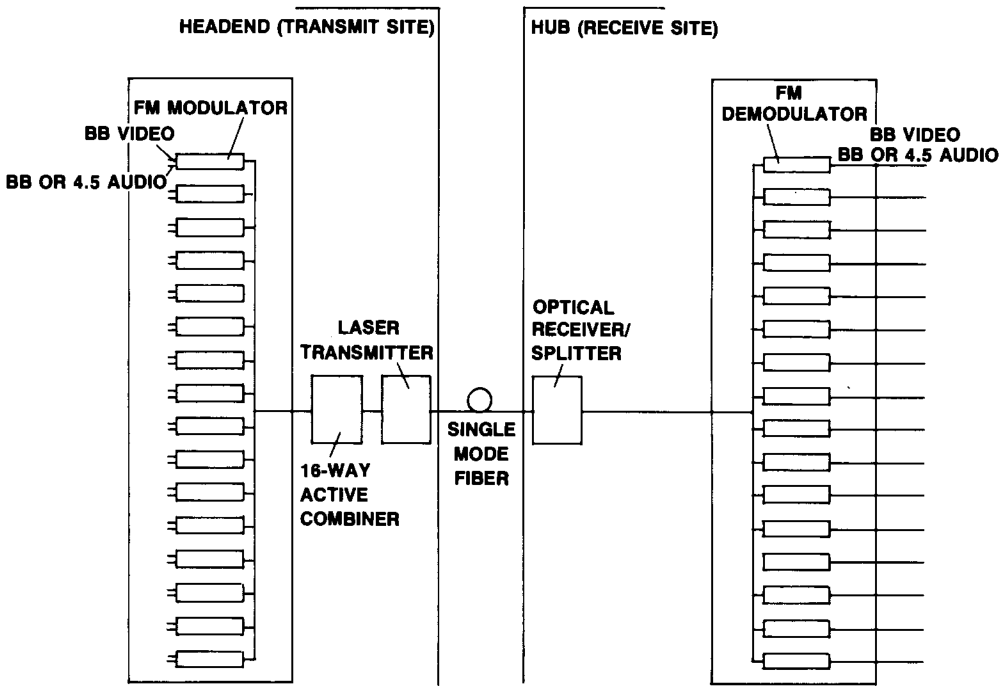


FIGURE 4

16 CHANNEL FM-FDM SUPERTRUNK

where: L = transmission distance (km)
l_c = connector loss
a_f = fiber attenuation in db/km

then:

$$22db = 3x0.5db + 0.35db/km \times L + 3db$$

and

$$L = (22 - 3x0.5 - 3)/0.35db/km$$

we find $L = 50km$

Therefore, the link may be more than 40km long with adequate margin.

By the same method, the two receive sites in figure 6 could be over 30 km from the transmit site.

IV DIGITAL SUPERTRUNKS

Figure 8a illustrates how one baseband channel of video can be quantized, delivered to a D/A converter, and converted back to the analog domain. The distortions and noise inherent in this

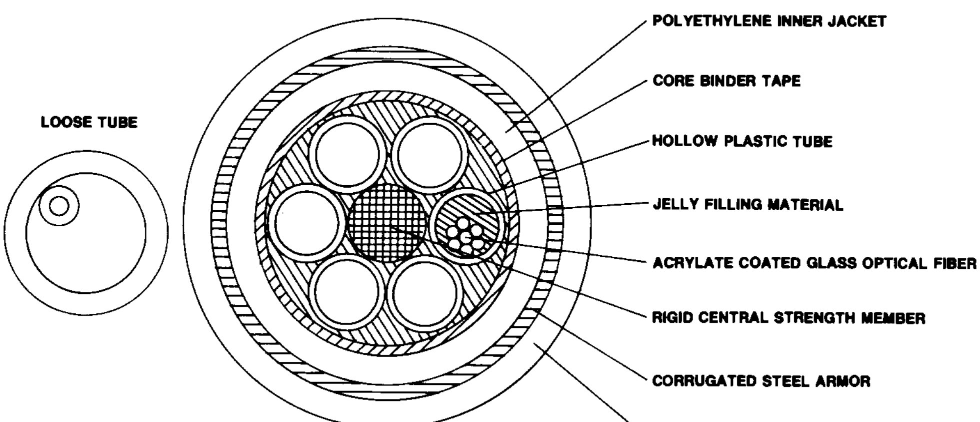


FIGURE 5 LOOSE-TUBE CABLE CONSTRUCTION

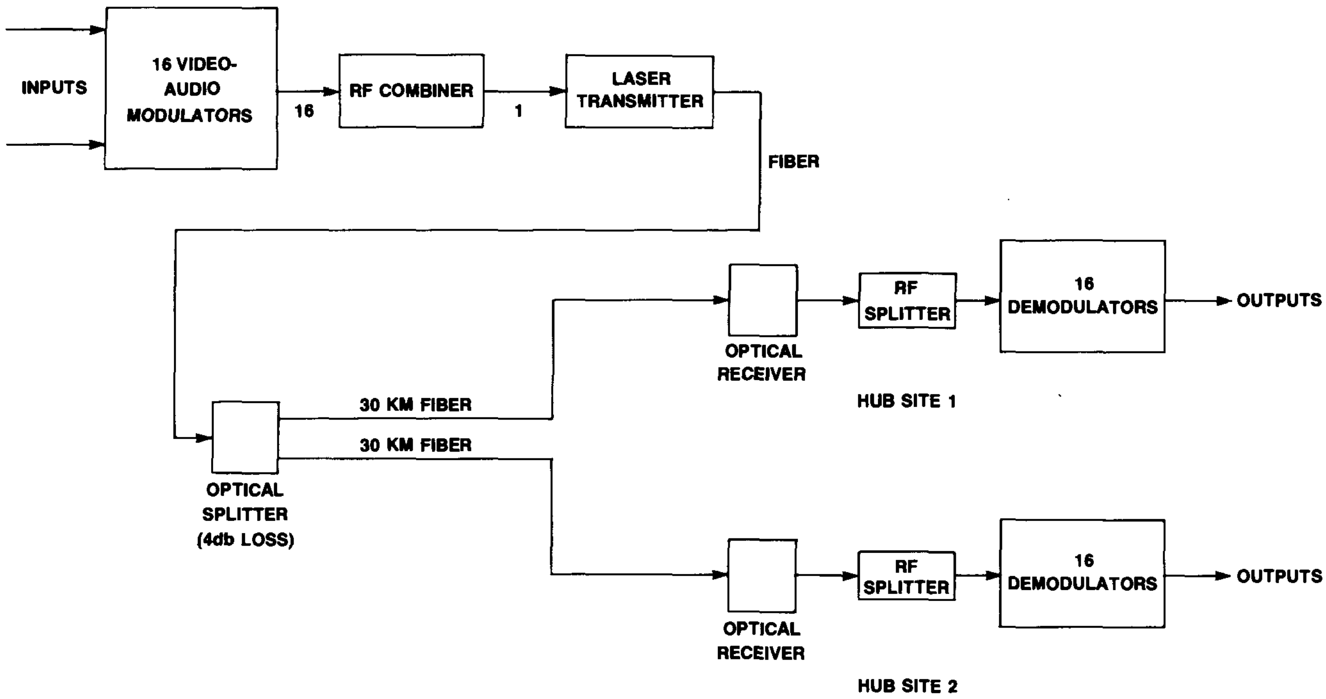


FIGURE 6

FM SUPERTRUNK TO TWO HUBS

process are far different than in FM transmission. Figure 8b shows how several digitized channels of video (or audio) can be time division multiplexed for carriage over a link. In this TDM technique, there are no intermodulation distortions as are present in FM. In a properly designed digital trunk system, all significant noise and distortions occur in the A/D and D/A converters. The presence of channel N in the system has no effect on channel 1.

Other advantages of the digital system are uniform performance over long fiber links (assuming acceptable bit error rates), and the ability to digitally regenerate signals using a receiver and laser while introducing no additional distortion. In a digital supertrunk, the audio is also digitized and time division multiplexed into its own subchannel. There is no interaction with the video, and the digitizing process may be optimally designed for audio.

IVa Noise and Distortion in The Digital Process

In the digitizing process, the baseband signal (video or audio) is converted into a series of quantum values which serve to represent the original

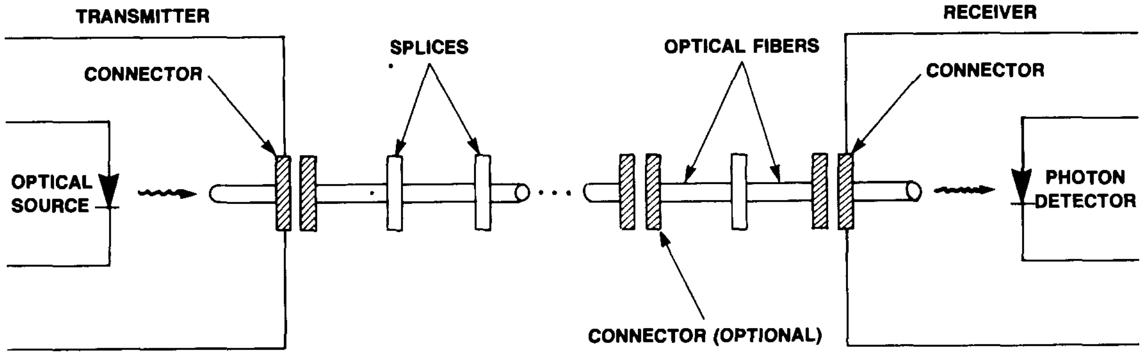


FIGURE 7

OPTICAL POWER LOSS MODEL

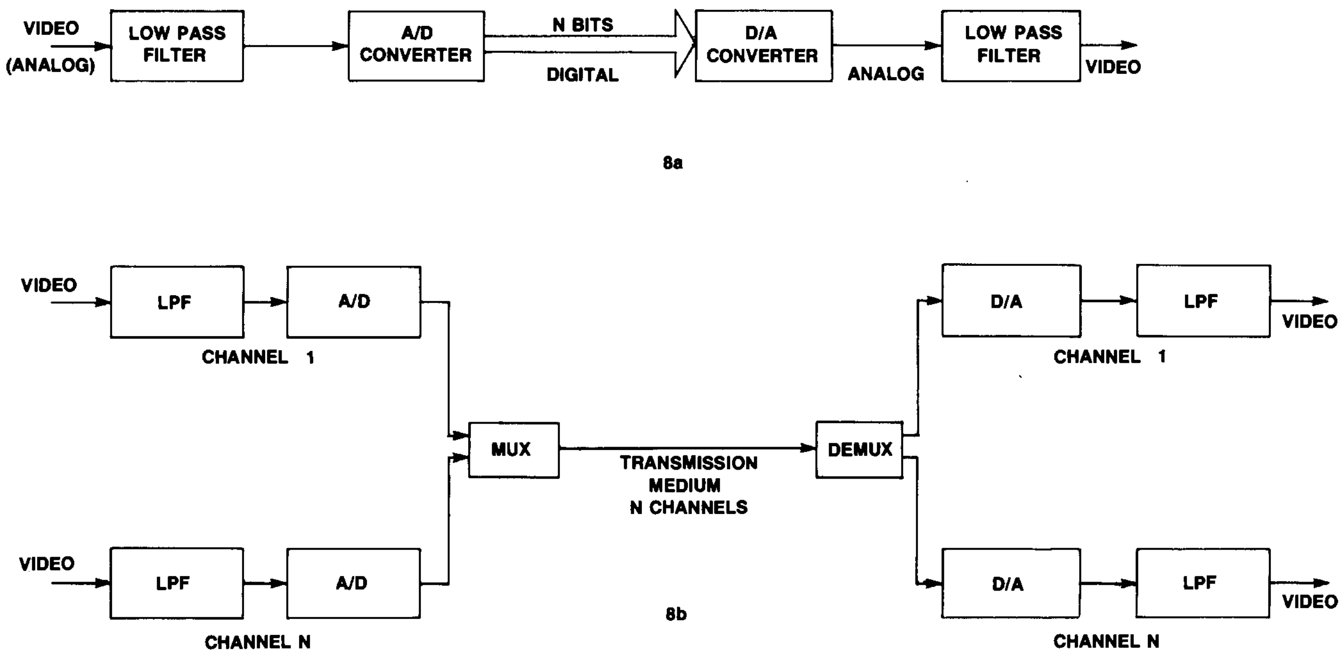


FIGURE 8

DIGITAL TRANSMISSION

signal. Consider that the digital signal contains no information describing the analog signal between samples. The most important factor in converter performance is the number of bits, or resolution. Another important parameter is the linearity when processing low frequency signals, as much important information is contained in these components. This error is best measured using an unmodulated ramp video signal. Test methods to measure this parameter are described in IEEE standard 746-1984.[6] Bellanger discusses sampling theory in chapter 1.[5]

The transfer characteristics of ideal A/D and D/A converters are shown in figure 9. The quantizing error of +/-1/2 LSB is also illustrated in this figure. The nonlinearities that occur in real converters will be discrete, as opposed to the continuous nonlinearities that can occur in analog circuits. If the sampling of the video ramp described above is coherent (synchronous) with the video, these errors can appear as vertical lines on a monitor. Proper choice of converters, resolution, selection of low pass filters, and good circuit design practice will reduce this type of distortion to acceptable levels (not visible on monitor). In today's monolithic converters,+/- 1/2 LSB linearity is practical; this represents +/-0.2% of full scale in an 8-bit system. It can be seen that linearity improves with resolution.

Sampling theory and the Nyquist criterion tell us that we must sample with a clock frequency at least two times the highest component in the transmitted signal to preclude aliasing. If we wish to carry 4.2MHz video, we must digitize at 8.4MHz or higher. It is much easier to design realizable low pass filters if we digitize at a higher rate. A common practice is to digitize at 4 times the NTSC color subcarrier rate of 3.58MHZ, or 14.318180 MHZ.

The signal to quantizing ratio may be calculated as follows:[6]

$$SNR(db) = 6.02N + 10.8db + 10 \log F_s/2F_{vmax}$$

where N = number of bits

F_s = sampling frequency

F_{vmax} = max freq content of video

For 4.2MHZ video sampled at 14.3MHz, the calculated SNR is 61.3db for an 8 bit system. Each additional bit of resolution represents 6.2db of SNR. Lab measurements using a Rohde and Swartz noise meter of an 8 bit system yield 62 to 63db. The care required in measuring SNR will be discussed below.

IVb Video Distortion and Noise Measurements in Digital Systems

A standard measurement technique for weighted signal to noise measurement using a Rohde and Swartz noise meter is to select a blank line in the vertical

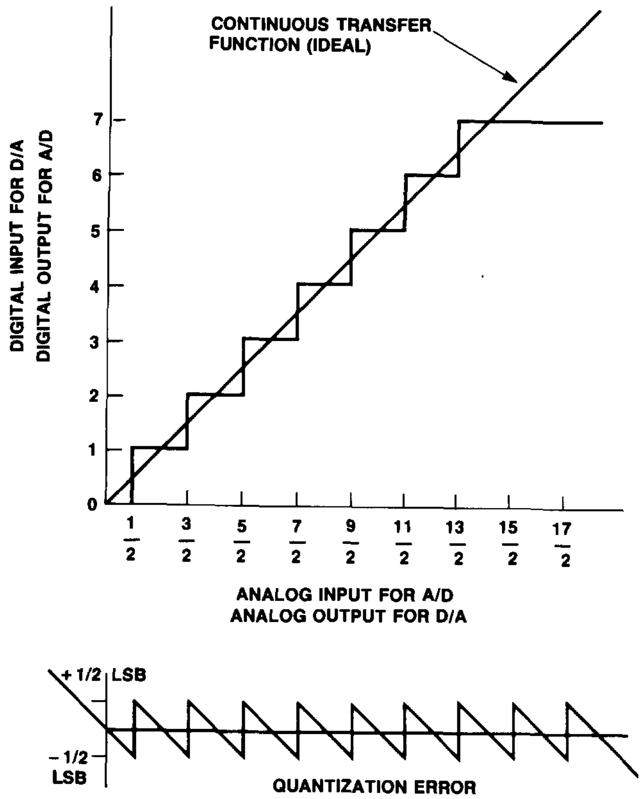


FIGURE 9

A/D AND D/A TRANSFER FUNCTIONS

interval. This line will contain a zero IRE flat field. This is a legitimate test in an analog system; in a digital system, it may yield a misleading result. If the flat field at 0 IRE happens to land halfway between A/D slicing levels, the A/D may output a fixed digital value; there will be no quantizing error and the resulting SNR will be artificially high.

A far more meaningful SNR measurement may be made as described in IEEE standard 746-1984. The test signal is a highly saturated chroma signal with constant luminance. It causes all the bits to change and quantizing errors to occur. Measurement using this technique confirms the calculation of video SNR given above.

Another phenomenon unique to digital processing is the occurrence of gliches in the output of the D/A converter. A glitch is an unwanted excursion that occurs at a D/A converter code change. It is due to unequal switching times within the DAC. In binary coded converters, the largest glitch is likely to occur at the half-scale transition when all the bits change simultaneously. Please refer to figure 10.

In an unmodulated ramp test signal, glitches at the same point in each line will cause a sharp vertical line on a monitor. In the modulated ramp signal used for differential gain and phase measurements, severe glitches will cause a crankcase effect on the differential gain and phase displays of a vectorscope. A severe case is shown in figure 11. This causes noticeable chroma saturation and hue changes in the picture.

In today's monolithic D/A converters, careful design has reduced glitch energy to 50 pV-sec or less. At this level, the peak differential gain can be approximately 2%, and differential phase as low as 1 deg. These levels are not discernible on a television monitor.

There is an in-depth discussion on these measurements in IEEE standard 746-1984.

IVc Audio Carriage

The digital supertrunk enjoys an advantage over the FM supertrunk for audio carriage because digital processing of audio is a mature technology. This technology offers the highest performance of any transmission technique. Low THD, well below 0.1%, 60db separation, and dynamic range approaching 90db are easily achievable. There are 12, 14, and 16 bit linear PCM converters designed for high-performance audio applications. Data used by these converters can easily be time division multiplexed with digitized video. Compact disk quality PCM audio requires approximately 1.4MHz serial data rate per stereo pair carried in the trunk TDM data stream. Dolby Laboratories has developed an adaptive delta modulation technique that requires 650KHz to achieve compact disk quality stereo audio. This technology is now field proven.[7] Either format will give audio performance

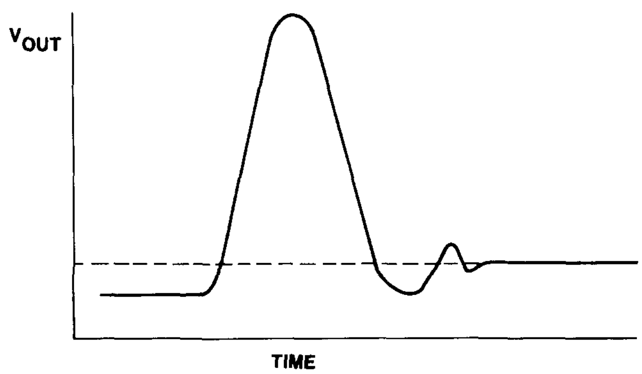
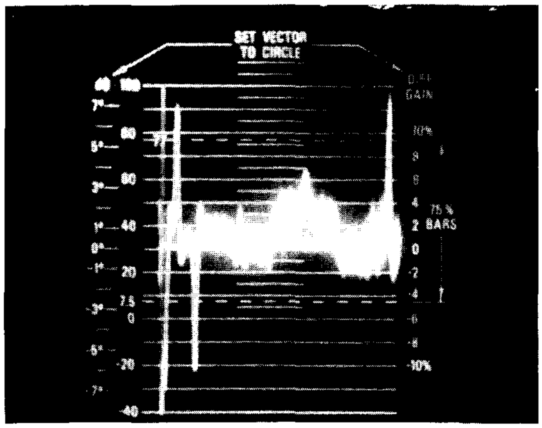
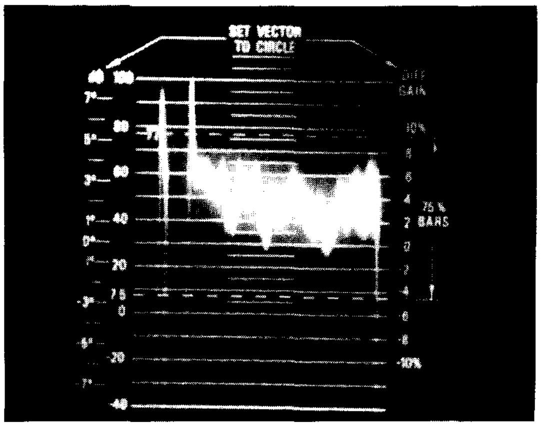


FIGURE 10

D/A OUTPUT GLITCH



DIFFERENTIAL PHASE



DIFFERENTIAL GAIN

FIGURE 11
SEVERE DISTORTION IN DIFFERENTIAL PHASE AND GAIN

superior to that delivered by FM.

IVd Architecture and Link Budget

The architecture of a digital supertrunk employing equipment on the market today is illustrated in figure 12. This equipment carries 8 channels of video and associated stereo audio over a SM fiber at 560Mbit. This equipment employs 7-bit codecs and achieves video SNR of 57 to 58db. Audio dynamic range is better than 65db with channel separation of 60db. The audio encoding is 12 bit PCM.

Since it is relatively easy to detect a logic one or zero with an optical receiver, digital supertrunks will have a somewhat longer reach than there FM counterparts. The parameter analoqous to CNR in FM is bit error rate (B.E.R.) in a digital link. A B.E.R. of 10^{-9} is the criterion for acceptable performance.

A link budget is based on the following assumptions:

1. The optical power launched by the laser transmitter is -3Dbm, at 1310nm.
2. The semiconductor receiver sensitivity is -35DB for a B.E.R. of 10^{-9} or better.
3. The fiber loss is 0.35db/Km at 1310nm.
4. A system margin of 3DB is assumed to allow for component aging and temperature effects.
5. Fusion splice losses are less than 0.1DB and can be ignored for the purpose of a comparative analysis.
6. Connector losses are assumed to be 0.5DB per connector

Figure 7 is an optical power loss model that illustrates the cumulative losses between the laser and the receiver. If P_t is the transmit power in DBm and P_r is the receive power in dbm, then P_l is the total loss budget for the link:[1]

$$P_l = P_t - P_r$$
$$= -3 - (-35) = 32\text{dbm}$$
and
$$P_l = 3xl_c + a_f \times L + \text{system margin}$$

where: L = transmission distance (km)
l_c = connector loss
a_f = fiber attenuation in db/km

then:
$$32\text{db} = 3\text{x}0.5\text{db} + 0.35\text{db/km} \times L + 3\text{db}$$
and
$$L = (32 - 3\text{x}0.5 - 3)/0.35\text{db/km}$$

we find $L = 78$km

Therefore, the link may be 60km long with adequate margin.

It can be seen that an optical splitter may be employed to deliver signals to more than one hub site.

Figure 13 illustrates a hypothetical trunk system to carry 12 channels of video and audio per fiber at 1.2Gbit/sec. This system employs 8 bit converters, as an 8 bit system most closely matches an FM-FDM supertrunk in video SNR (mid 60db range). The 100Mbit/sec data rate output of the transmitters assumes the the audio data is time division multiplexed into the data stream during the horizontal blanking interval of the video. This technique of carrying digitized audio embedded within the video is employed today in the Videocipher equipment used for satellite signal encryption. In this type of system, it is necessary to reconstruct the video composite sync at the receiving

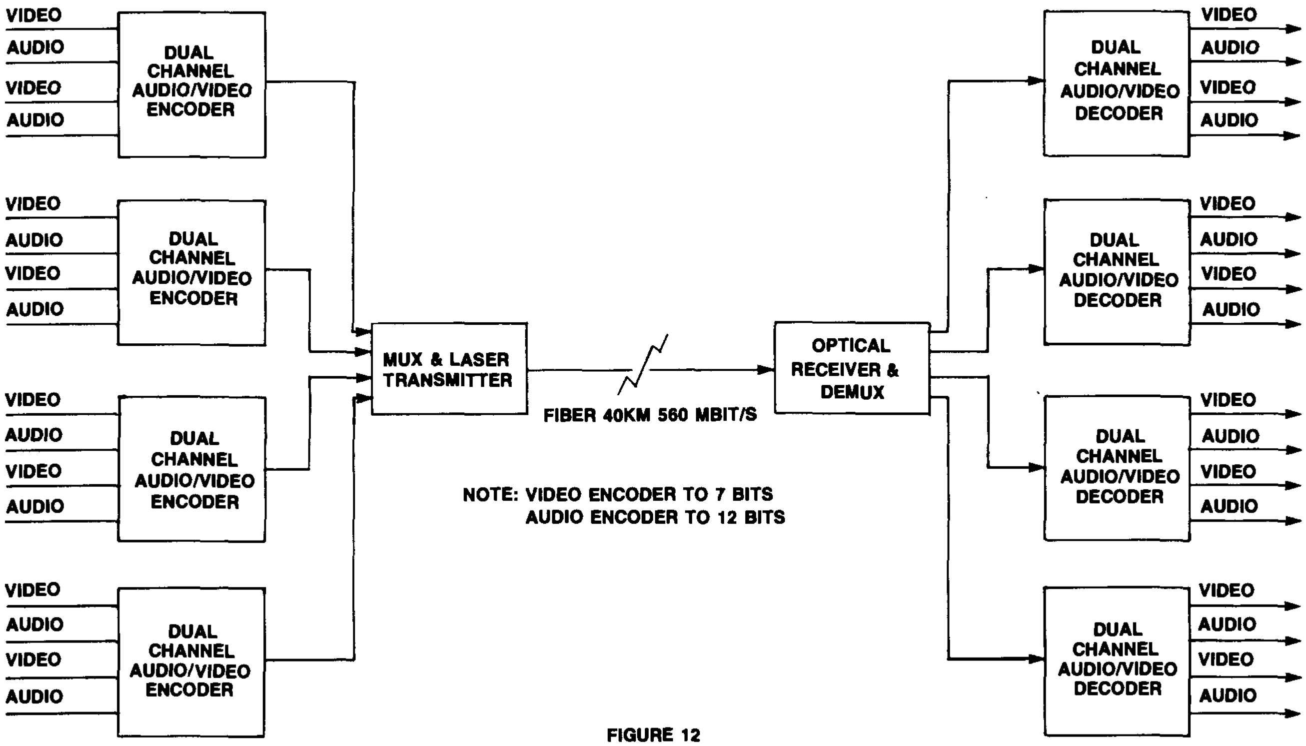


FIGURE 12
8-CHANNEL DIGITAL SUPERTRUNK

decoders, adding some complexity. The technique does make efficient use of the available bit stream.

The bandwidth required to carry 12 channels as described here is greater than on the 16 channel FM-FDM trunk. Figure 14 illustrates the spectrum of non-return to zero (NRZ) data modulated onto a carrier. F_b is the bit rate; the overall bandwidth is approximately equal to the NRZ bit rate, or 1.2Ghz.[8] This compares to 700Mhz for the 16 channel FM system.

A trunk system based on 9 bit converters would require a serial bit rate of approximately 109Mbit/sec per channel. The video SNR would improve to better than 67db, and the differential phase and gain could meet RS-250B short haul specifications. However, a 1.2Gbit link would only carry 10 channels per fiber. To exceed the video performance of the FM-FDM supertrunk (video SNR) requires 9 bit resolution, at higher cost.

V COST ANALYSIS

Fiber optic cable, as illustrated in figure 5, is available with various fiber counts. For the purpose of cost analysis, the following price per foot, when purchasing 40 km of cable, will be assumed:

Number of fibers in cable	price/ft (armored)
4	$0.60
6	0.73
8	0.82
10	0.91
12	1.02
16	1.25

The cost of installing the cable has several constituents, which are assumed as follows:

Description	cost
Route make ready work	$0.45/ft
Hang cable	$0.70/ft
Cable Installation	$1.15/ft
Fusion splicing	$45/splice

(Assume a splice required every 4km and at each end of fiber)

Optical patch panel	$1000
Test & Document completed cable	$2000

The architecture in figure 4 will carry 16 channels per fiber. An 80 channel system would require 5 fibers. A cable with 6 fibers is selected to provide

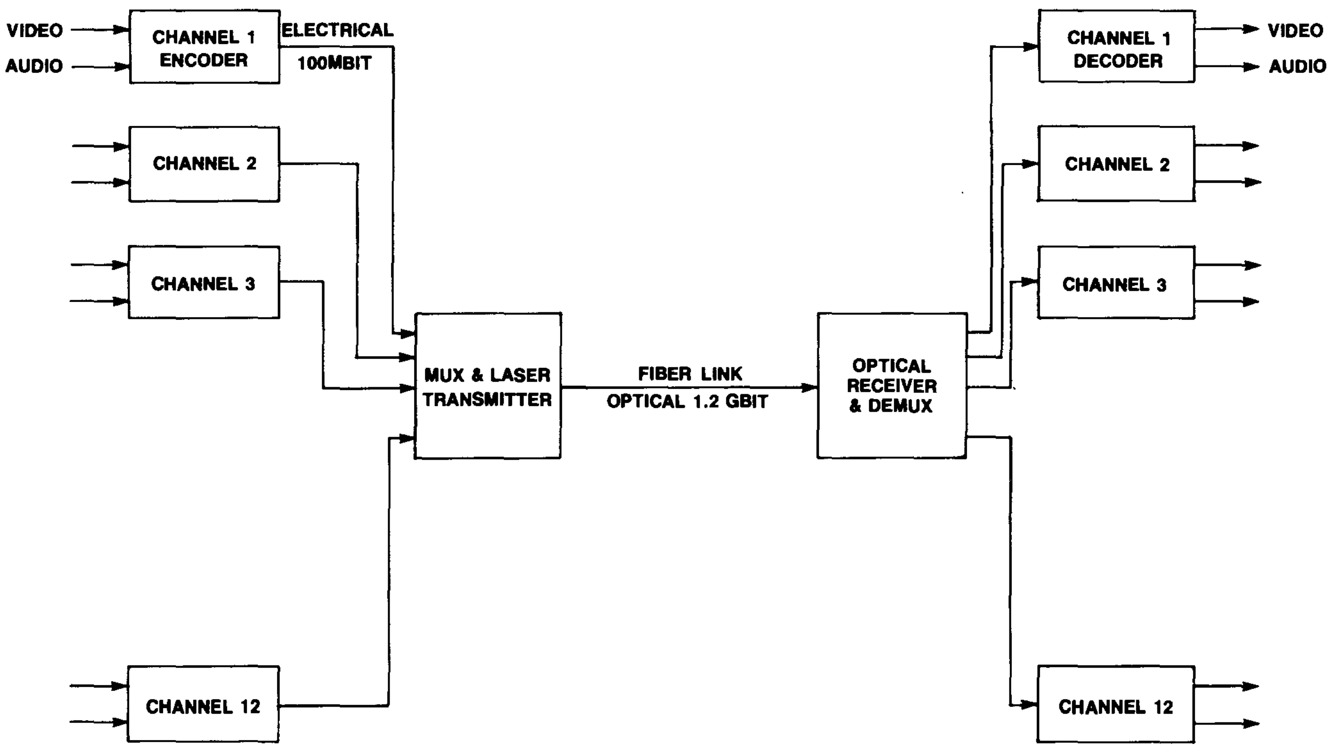


FIGURE 13

PROPOSED 12 CHANNEL 8-BIT SYSTEM

a spare fiber. The cost of pre-wired equipment racks is included in the unit cost of all rack mounted equipment. The cost of a complete 80 channel FM supertrunk is presented in table 1.

Therefore the average cost per channel of the 80 channel FM supertrunk is $7,843.

The digital supertrunk in figure 12 will carry 8 channels per fiber. Note that the encoders and decoders each carry two channels. An 80 channel system based on this architecture would require a cable with 10 fibers; a 12-fiber cable is selected to provide spares. Again the cost of pre-wired equipment racks is included in the unit cost of the equipment. The cost of a complete 80 channel digital supertrunk is presented in table 2.

The cost per channel of the 80 channel digital supertrunk is therefore $8,091. Consider that the video SNR of a 7 bit digital system is approximately 57db, as compared to 65db for the FM supertrunk.

The hypothetical system in figure 13 (8 bit) would carry 12 channels per fiber. An 80 channel trunk would require 7 fibers. A cable with 8 fibers is selected to provide 1 spare. An 8 bit system most closely matches the video performance of the FM-FDM trunk.

A conservative cost breakdown of this 80 channel supertrunk is given in table 3. The length of the trunk is assumed to be 40km.

In this case, the cost per channel is $13,823. It is expected that the cost of the encoders and decoders could be reduced

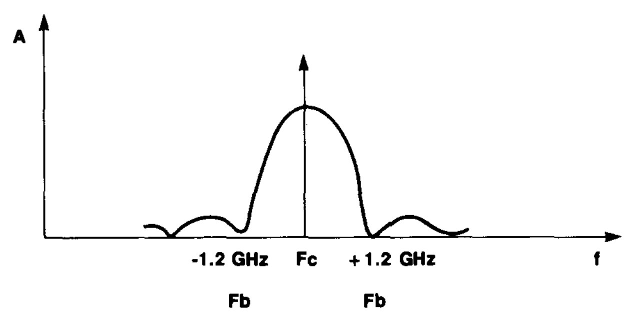


FIGURE 14

NRZ DATA SPECTRUM

TABLE 1
FM TRUNK IN FIGURE 4

ITEM	UNIT COST	QUAN	EXTENDED COST
FM Modulators	$2000	80	$160,000
16 Combiner	$1000	5	Chan $5000
Laser Transmitter	$6200	5	$31,000
Optical Receiver	$3200	5	$16,000
RF Splitter	$300	5	$1500
FM Demods.	$2000	80	$160,000
Cable (6 fibers)	$0.73/ft	40km	$95,805
Cable Installation	$1.15/ft	40km	$150,926
Optical Patch Panel	$1000	2	$2000
Splices (12 per fiber)	$45 ea	72	$3240
Test/Document Installation	$2000	1	$2000
total cost			$627,471

by the application of large scale integration of the digital circuitry. This is important since these equipments are a major cost component of this proposed system.

VI SCRAMBLING

There is no practical need to secure signals while on a supertrunk, especially on a point-to-point fiber link that is difficult to tap. It is desirable, however, to carry signals that are already encoded to a hub site. This precludes the need for additional scramblers at the hub.

There are two basic type of video scrambling in common use today, baseband and RF scrambling. The common techniques are sync suppression and video inversion. Baseband scrambling may be carried over an FM link if a sync driven clamp is used at the receive site. The clamp is required to restore the DC offset of the video, which can not be carried over the FM link, before being AM modulated. RF scrambling has proven to be impractical due to the difficulty of carrying already modulated vestigial sideband (VSB) signals.[3,11]

TABLE 2
DIGITAL TRUNK IN FIGURE 12
(7 BIT)

ITEM	UNIT COST	QUAN	EXTENDED COST
2-Channel Encoder	$3900	40	$156,000
Laser Trans.	$4400	10	$44,000
Optical Receiver	$2400	10	$24,000
2-Channel Decoders	$3200	40	$128,000
Cable (12 fibers)	$1.02/ft	40km	$133,865
Cable Installation	$1.15/ft	40km	$150,926
Optical Patch Panel	$1000	2	$2000
Splices (12 per fiber)	$45 ea	144	$6480
Test/Document Installation	$2000	1	$2000
Total Cost			$647,271

Since a digital system can encode and decode the DC component of a video signal, no clamp is required at the receive hub site. Thus the digital supertrunk can carry baseband (sync suppression/video inversion) scrambling. However, the digital link would encounter the same problems with RF VSB scrambled signals as the FM trunk.

If at some time in the future digital signals are carried directly to the subscriber, the level of security achieved in a properly designed system could be orders of magnitude higher than that in an analog system. A good digital encryption system will introduce no distortion (residual effect) when the desired signal is decoded. The encryption may be time varying. Wechselberger has written an excellent paper on the subject of encryption as applied to CATV.[9]

VII FUTURE TRENDS

FM supertrunking is a relatively low volume, mature technology, with equipment having been in the field for several years. No dramatic breakthroughs in cost reduction can be expected in the modulators, demodulators, combiners, or splitters over that suggested in the

TABLE 3
DIGITAL TRUNK IN FIGURE 13
(8 BIT)

ITEM	UNIT COST	QUAN	EXTENDED COST
Video/audio encoders	$5000	80	$400,000
1.2Gbit Laser trans.	$10,000	7	Mux~ $70,000
1.2Gbit Receiver	$7000	7	Optical $49,000
Video/audio decoders	$4000	80	$320,000
Cable (8 fibers)	$0.82/ft	40km	$107,617
Cable Installation	$1.15/ft	40km	$150,926
Optical Patch panel	$1000	2	$2000
Splices (12 per fiber)	$45	96	$4320
Test/Document Installation	$2000	1	$2000

Total Cost			$1,105,863

previous cost analysis. Additional channels per fiber will require improvements in the lasers and receivers to utilize more of the available bandwidth. This is likely to happen. As the optical components are improved in performance and built in volume, substantial cost reductions are expected. The telecommunications industry is driving the cost of the optical components down. The FM supertrunk will take advantage of this.

Digital technologies as applied to consumer electronics have repeatedly shown dramatic cost reductions with volume production. Digital watches and calculators are good examples. The designer may employ gate arrays, standard cells, or full custom technology to effect dramatic cost reductions. Although the digital supertrunk achitectures in figures 12 and 13 will not be built in the volumes of consumer products, judicious use of integration techniques can dramatically reduce size, power consumption, and cost.

The very high data rates required at the laser and optical diode interfaces can only be achieved using gallium arsenide logic. Today this logic is only available from a few vendors and exhibits relatively low yields; it is therefore expensive. However, only a small portion of the required logic need run at these very high speeds. Thus more cost effective logic families, such as ECL or 74HC, may be used to implement the lower data rate portions of the circuits.

In the last five years, dramatic cost reductions have occurred in monolithic A/D and D/A converters. Digital audio technology is coming into use in high volume consumer products. As lasers and high speed optical receivers come into high volume application, their cost will also come down. Use of these components by the telephone industry for digital transmission is pushing the technology forward. The digital supertrunk will take advantage of all these trends.

VIII CONCLUSION

Table 4 summarizes the factors to be considered when comparing FM and digital supertrunks:

TABLE 4

FACTOR	FM	DIGITAL		
		7 bit	8 bit	9 bit
channels per fiber	16	8	12	10
bandwidth	700Mhz	600Mhz	1.2Ghz	1.2Ghz
video SNR	65db	57db	63db	67db
diff phase	1deg	2deg	1deg	<1deg
Diff gain	1%	2%	1%	<1%
audio dynamic range	65db	65db	85db	85db
audio freq resp	50hz-15khz	- 20hz to 20khz-		
Audio Chan Separation	30db	>60db	>65db	>65db
System cost/channel	$7,843	$8,091	$13,823	?
ease of carrying scrambling	fair	--good--		
cost reduction potential	fair	high		

It can been seen from the table that a 7 bit digital system is cost competitive with an FM-FDM system, but does not provide the same level of video performance. An 8 bit digital system most closely matches the FM-FDM system in performance, at increased cost. This cost differential is likely to decrease with time. A properly designed 9 bit digital system can provide performance superior to the FM-FDM system at higher cost, with fewer channels per fiber.

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DATA COLLECTION FOR STATUS MONITORING SYSTEMS

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ABSTRACT

Status monitoring is a useful tool in maintaining the high quality of service demanded by today's subscriber. Unless a two way cable system is available, it is not always possible to implement status monitoring with the available techniques. This paper examines several techniques for collecting the data from status monitoring devices. Two approaches to using the RF return path are discussed. Four techniques for gathering data when no conventional RF return is available are then discussed.

INTRODUCTION

Cable subscribers are becoming increasingly demanding of cable service providers. With the proliferation of VCR's and satellite receivers, as well as the looming presence of High Definition Television, there is more pressure than ever before to maintain the highest possible quality in the delivery of video to the home. This pressure creates a demand for a viable status monitoring system to aid the operator in maintaining the system at its peak capability.

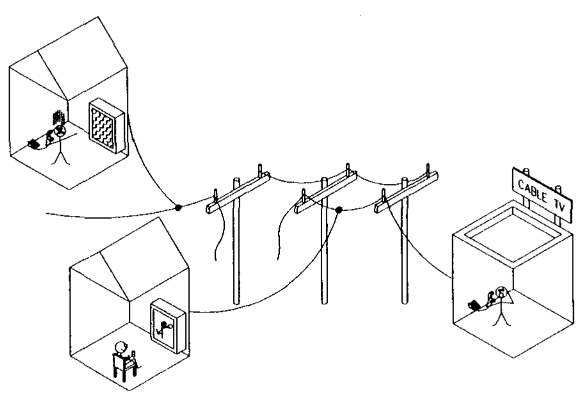

FIGURE 1
SUBSCRIBER-BASED STATUS MONITORING

Every cable operator has in place an extensive status monitoring system. This system covers every tap outlet in the system, providing feedback when picture quality degrades to unacceptable levels. There are many problems with this universal monitor system. For example, feedback is slow, often providing the first indication of problems several minutes after the problem occurs. Collecting data is also very expensive, since someone must answer the phones when all of those angry "status monitors" call in to complain about their picture quality (Figure 1).

Obviously, there is great benefit in a system that can find and report system changes before they result in subscribers becoming upset. These systems are available from all major CATV equipment suppliers. The typical status monitor product resides in the trunk amplifier. It measures the performance of one or more frequencies in the system and reports back to the headend using an RF carrier in the return band. More recently, stand alone monitor products have been introduced that offer more complete measurement capability. For the most part, these also use the RF return system for communications.

These devices work well when a return system is available, but what do you do when this luxury is not available? The advantages of status monitoring are only available when you are able to communicate in some way with the monitoring devices. Talking to the devices is easy. Any cable system can find space to squeeze in one data carrier somewhere in the forward frequency band. The trick is getting back to the operator from the devices in the field.

The easy approach may be to build an RF return system. Many systems have been built with two-way capability, even if the return amplifiers were never installed. A full return system can be expensive to install, however. This is especially painful if the only application of the return system is for end of line monitoring. The cost of all of the return amps must be divided over a very small number of monitors. Furthermore, many systems currently in operation have no facility to be upgraded for operation of a return system. How can we communicate with status monitor devices in these systems?

LOW COST RF RETURN SYSTEMS

First, let's examine the options available for using the RF return system, when an upgrade is possible. While status monitoring will not necessarily allow a system to be maintained with fewer technicians, additional resources should not be required just to maintain the return system. Any return system that is installed solely for status monitoring, therefore, should require little or no maintenance of its own. Set up should be straight forward. The cost of the return system should be small compared to the cost of the monitoring equipment itself. We have examined two alternative return configurations that meet these requirements: the Return Data Relay system and a low cost return amplifier.

Return Data Relay

The first RF return system we studied is what we call the Return Data Relay system. In this approach, the return data pilot is converted back to baseband, timing corrected, and retransmitted back toward the headend by intermediate relay stations in the system. These relay devices can take the place of the return amplifier in selected trunk stations (Figure 2). The distance between relay stations will vary, depending on the spacing of the amplifier stations and the amount of passive loss between stations. In general, we would expect to install a relay in every fourth or fifth station. In between the relay stations, a simple jumper arrangement allows the RF carrier to pass through the station.

The Return Data Relay system offers two significant advantages over traditional return systems. The first is the lack of setup and maintenance required to operate the system. Noise buildup is limited to the span between relay stations. Even with trunk branching, the noise floor will be very low at the input to the receiver. The low noise floor allows the receiver to accept a wide range of input levels without the noise overpowering the signal. Variations in level due to thermal changes are minimal over the short cascades as well. Adjustments for transmit and receive levels can be simplified or even eliminated.

A second advantage of the Return Data Relay is cost. It is not difficult to make a wide dynamic range receiver circuit that will operate in a low noise environment. There are integrated circuit receivers that will perform the job very nicely for either AM or FM data systems. In a system using traditional trunk based status monitors, the transmitter of the transponder can be used as the transmitter of the relay, thus the cost of the transmitter can be eliminated from the repeater. The cost of a repeater station will be considerably less than the cost of a standard return amplifier, and we do not even need one in every amplifier location.

The primary argument against the Return Data Relay system as the RF product of choice is that it is strictly a single use device. It does not allow the return path to accommodate any service other than a single status monitor data pilot. This is not a serious drawback, since our only application for the return path is the monitoring system itself. We have, however, also considered the possibility of a more general purpose system that will provide for services beyond the status monitoring system.

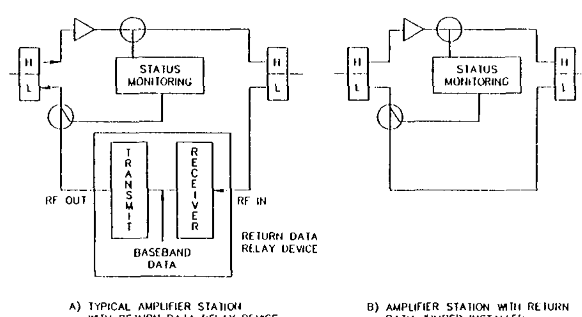

FIGURE 2
TRUNK AMPLIFIER STATION CONFIGURATION FOR RETURN DATA RELAY SYSTEM

Low Cost Return Amplifier

Most return amplifier modules assume that video signals will be sent to the headend on a return channel and retransmitted to subscribers on the forward system. To operate properly in the system, this application requires a fairly expensive amplifier module and extensive setup. If we limit our application to narrow band data communications, the performance requirements of the amplifier module can be relaxed considerably. While a typical sub-split return system operates from 5 to 30 MHz, only a small portion of that bandwidth is required for data. Several data pilots can operate in a single 6 MHz video channel assignment. It is easy to implement a return amplifier to accommodate a narrow-band, data only, return system using any of a number of integrated circuit RF amplifiers. A single IC can provide up to 20 dB of gain at a very low cost. No slope control is necessary for the narrow bandwidth required. All that is needed is a gain control for the purists who insist on setting levels and possibly a simple thermal compensation network to limit the level variations due to temperature changes.

Here is an example of how this system might work: Assume a 300 MHz system spaced at 22 dB at the highest carrier. If we use a return data frequency of 11 MHz, there is about 4 dB of loss per span for the return data. This will increase when passive devices such as trunk couplers are used, but will be less for short spans. For a cascade length of 30 amplifiers, the total loss at 11 MHz will be about 120 dB. For operation over a temperature range of -20 to +120 degrees Fahrenheit, the change in attenuation is estimated by the following equation:

$$A_c = ((t_1 - t_2)/10) \times .01 \times A_{nom} =$$

$$((120+20)/10) \times .01 \times 120 = 12 \text{ dB}$$

where Ac is the change in attenuation due to temperature change, t1 and t2 are the temperature extremes, and Anom is the nominal attenuation of the cable.

The receiver in the hub must accommodate input signals that vary as much as 12 dB over temperature variations. Any variations in the nominal transmit levels must also be accommodated. If we provide a level control for the transmitters, it is feasible to provide a receiver that will track the level variations expected. The system will, therefore, operate reliably with no thermal compensation in the return amplifiers. The lack of compensation simplifies the setup and maintenance of the system substantially.

This does not necessarily provide a system that will be usable by any other data services, however. In order to provide a reliable data path for other services, thermal compensation is required. This still presents an appreciable cost savings over traditional return amplifiers, but loses the advantages of simplified set-up.

Having looked into two different approaches to a RF return system, what conclusions have we made? Each system has some advantages. The relay system eliminates any need for complicated set-up procedures. It also prevents any build-up of noise, which makes the monitoring of the distribution system much easier. The low-cost amplifier approach, on the other hand, provides a less expensive approach to a trunk-only monitor system. It also provides an easier upgrade path to accommodate other data services. Both systems also suffer from one other problem. The devices need to fit into the trunk station in the return amplifier. This means that each unique amplifier product line needs to have a device designed to fit it. A large engineering effort would be required to package the system for the many different types of trunk equipment installed in the field. The low cost amplifier is the preferred solution due to its more universal nature.

THE ONE WAY TICKET

Two alternatives have been identified for implementing a return data system when the return path is usable. In many systems, however, there is no provision at all for using the return band. How can we retrieve status data in these cases?

Visual Reading

One concept we have considered for extracting data from the monitor modules is the use of a visual indicator on the outside of an amplifier station. This could be as simple as a pair of colored lights, a green light to indicate normal operating status and a red light for a trouble alert. This system eliminates the need to climb the pole and open a test port to determine if signals are present in the system. A system of this type may work to speed system diagnostic time when a major outage occurs, but will not do much to allow us to find problems before they become major. Only go/no-go data is available, and collecting even that data is a labor-intensive process. Due to the limitations of this type of device, we have focused our studies elsewhere.

Power Band Communications

Every cable system passes AC voltage between stations to power the amplifiers. To pass the power required, stations contain bypass circuitry that blocks RF signals above 5 MHz, but passes lower frequencies, from DC to as high as 1 MHz (Figure 3). Using this band for data transmission is not a new idea, but is intriguing. More and more devices are becoming available to perform similar functions over shorter distances, in home and office applications. It seems that this technique can be applied successfully in a CATV environment.

At least one vendor in the CATV arena has a product that communicates in the power band. Carriers from 100 KHz to 150 KHz are used. Data repeaters are used to extend the reach of the system so that a full cascade can be covered. Test systems are installed and operating, showing the viability of the technique. Repeaters are installed at the locations of the 60 volt power supply stations. No extra passives are required. The trunk stations that do not pass power must be modified to pass the communications frequencies while preventing the 60 Hz supply voltage from passing through them.

A power band communications system, then, has a large appeal due to its minimum impact on the system architecture. It is a generic system, in that it requires no significant modification to the existing equipment.

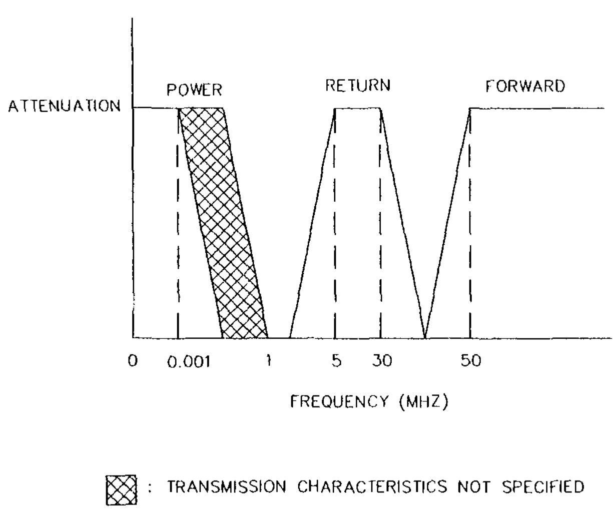

FIGURE 3
CATV FREQUENCY UTILIZATION

Installation of a power band communication system in a CATV network can be quite complicated, however. Although the frequencies around 100 KHz are generally passed through the power system, the characteristics vary greatly within a system. Manufacturers do not specify the operation of their equipment in this region. Not only do products from different manufacturers differ greatly in their response in this range, but also supposedly identical equipment from a single vendor can vary greatly. Setting up a system demands that the characteristics of the particular section of the system be analyzed to determine the optimum frequency to use for the communications channel. This analysis is currently being performed manually by sweeping the frequencies in the 100 KHz range for each repeater span to find the best frequency to use. Equipment is currently being developed to automate the setup process.

The noise generated by switching power supplies, coupled with the low impedance to ground presented by the power system, requires a significant amount of power from the transmitters in this system. It is still not clear whether power band communications is a completely universal system. There may be some installations that will not pass the frequencies desired,making this system unusable.

Phone Line Communications

A conventional means of communicating with remote devices uses the telephone network. If full time communications is desired, leased lines are required. This is an expensive proposition. A voice-grade dedicated line that runs between telephone switching centers can cost over $100 per month, even for a short connection, and over $300 per month for a longer run. This cost would prove difficult to justify.

An easier sell is the installation of a standard dial-up line. This would eliminate the surcharge for connecting between local switching centers. Base rates may run as low as $30 per month, with an additional charge based on the number of calls made from each location.

In a dial-up telephone system the control computer dials into the monitor devices, one at a time, to determine the status of a particular point in the system. The speed of the system is limited by the time it takes to place a call to each station in the system. At best, it takes about 5 seconds to make a dial-up connection. Only 12 stations per minute can be polled using this scheme. Even the slowest conventional status monitoring system can poll over 200 stations in a minute. Thus, the dial-up system is less useful than a conventional system as an aid in rapidly finding major system problems. It still provides the most important feature of status monitoring, however. Data is accumulated over time that will track the performance of the system and indicate areas where performance is beginning to degrade and maintenance is required.

Ride The Air Waves

The FCC has recently allocated a frequency band at about 900 MHz for over-the-air data communications. The hardware required to take advantage of this space is now becoming available. This system is capable of communicating over the distances required by most CATV systems. Since it is a line-of-sight system, the reach that can be obtained depends on the height and location of the master antenna as well as the physical terrain of the installation (Figure 4). This system is considered a microwave product by the FCC due to the high frequencies involved.

This over-the-air system accommodates full-time communications. The monitoring stations can be polled just as if data was running through a two way cable system. If a loss of power takes down a monitoring node, the control system will "know" almost immediately, since it is continually polling the devices in the field. A cable break or a failed amplifier will not impede communications at all, unlike a traditional cable-based system.

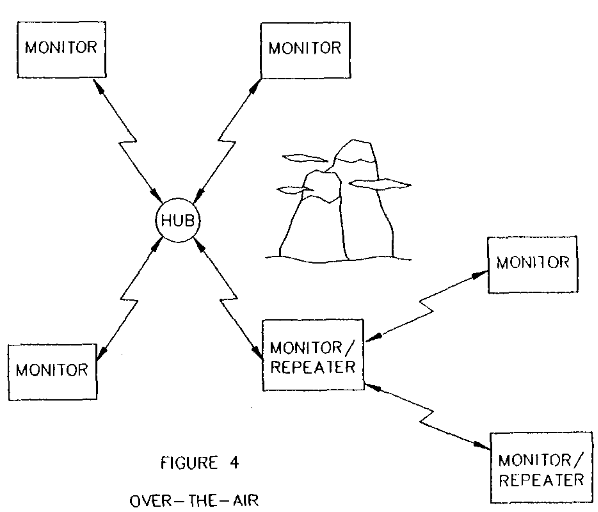

FIGURE 4

OVER-THE-AIR
MICROWAVE COMMUNICATIONS

00159

The hardware to implement an over-the-air data system is not cheap by any means. The transmit/receive stations alone cost about $1500, for a unit that is not rugged, and around $3000 for a unit set in a rugged NEMA type enclosure. That level of expense makes this over-the-air approach unreasonable, except for a few key points in a system. Lease arrangements will allow the operator to obtain this over-the-air service without a large initial expense. The cost of a lease will make this technology cost competitive with a dial-up system, while providing a dedicated, full-time, data link.

The 900 MHz over-the-air concept has not been shown to be appropriate for a CATV status monitoring system, but it holds promise. The spectrum is available. Equipment is becoming more readily available. Magnavox is looking seriously into this technology to determine the problems and advantages that it brings us. Stay tuned.

SUMMARY

We have looked briefly at six techniques for collecting data from status monitor systems. The low cost return amplifier and Return Data Relay systems use the traditional RF return path of the cable network. The low cost amplifier appears to provide a more complete and general solution to the problem at hand. Visual indicators provide too little communication too late to be acceptable for most applications. Using the power band for communications holds promise as a technique that applies in almost any system, but the problems of setting it up for individual systems must be resolved. Telephone communications are certainly viable and can be implemented today. A new high-frequency, over-the-air, data system allows full time, rapid communications at a cost that is comparable to monthly charges for phone service.

The CATV community has not yet determined which of these systems will be accepted for status monitoring qpplications. With the diversity of the applications and personalities involved, it is likely that most of them will be used to some extent.

ACKNOWLEDGMENTS

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DELIVERING DIGITAL AUDIO

JAMES GREEN AND CLYDE ROBBINS

JERROLD - APPLIED MEDIA LAB

ABSTRACT

Cable television has grown to service 60 percent of United States television households. Additional growth will require additional services. Audio is the second largest consumer communications market following video. Cable television has the only pipe presently capable of capitalizing on the audio market. Digital audio is necessary to control and provide premium quality programming. This paper presents an integrated system for digital audio delivery which is compatible with existing cable television methods of operation. Room for future growth in audio programming services is designed into the system.

CABLE INDUSTRY GROWTH

Cable television has grown over the past 40 years by providing television programing not available via local broadcast reception. The first stage of cable television growth came from importing distant broadcast signals. Cable developed during this period primarily in small communities. The second stage of growth was driven by satellite delivered signals which provided programing not available from terrestrial broadcasters. This additional programing allowed growth in both small and large population centers. As cable operators now look toward deriving more revenue from existing systems, it is useful to consider that a cable system is really a communication system capable of delivering other services in addition to television.

THE AUDIO MARKET

After television, the next largest available market is audio.

To put audio into perspective, U.S. television sales in 1988 were 6 Billion dollars. U.S. Consumer Audio Equipment Sales were 2 Billion dollars. The audio recording industry sales in 1988 were 6.3 billion dollars according to the RIAA.

Since the advent of compact disc players, the audio hardware industries' sales on an annual basis have doubled. The recording industry has seen a rebirth. Compact discs offer convenience and quality and are achieving widespread acceptance (over 40 million CD players are in use worldwide). Cable systems can be used to deliver compact disc quality audio directly to the home.

DIGITAL AUDIO

Cable systems can deliver digital audio to the home free of commercial interruptions. Broadcast radio can not deliver commercial free digital audio because of transmission regulations, backward compatibility problems and the inability to derive revenue other than by commercial messages. Cable system operators have an opportunity to provide the first direct digital audio link into the home. Digital transmission is necessary for control and consistent quality. Although some analog methods can yield high quality, encryption is difficult and expensive. Digital encryption is inexpensive and does not affect quality.

Digital audio is audio represented as a series of numbers. There are numerous methods of converting audio from a continuous analog form to a discreet digital form. 16 Bit linear pulse code modulation (PCM)is used in compact disc recordings. This system provides excellent audio quality, but uses excessive amounts of information space or bandwidth. For transmission, PCM is usually companded whereby inaudible information is discarded. The resulting audio quality is dependent on the particular companding implementation used. The interested reader is referred to references 1 and 2 for additional information on digital techniques.

Digital (or analog) companding does have measureable affects on audio performance parameters, the claims of some system proponents not withstanding. The parameters affected are typically harmonic distortion and noise in the presence of large signals (instantaneous S/N). The change in these parameters is not necessarily the important issue but whether the change is audible or not.

The method which must be used to determine transparency of a companding system is to have independent scientifically controlled blind listening testing performed. The source material must be of excellent quality and varied in content. The listening system and environment must also be controlled and of excellent quality. If blind testing done in this fashion shows no statistical preference on any program material for the disk source or the compact disc source processed through the companding system, then the system is proven to be transparent.

Another method of digital audio sampling is Dolby TM Adaptive Delta Modulation (ADM). This system developed by the leaders in both professional and consumer recording technology offers certain advantages over PCM systems. It offers audio transparency at a low bit rate (narrow bandwidth) and low cost decoding.

Another advantage of ADM is the ability to withstand a low level of bit errors (i.e. 10-6 BER) without a significant reduction in audio quality. All PCM systems must be error corrected throughout all transmissions, as severe cracks and pops will result when the most significant bits are in error. The need for error correction translates into increased receiver complexity and cost as well as additional transmission bandwidth.

The goal of a transmitted digital audio system is to be audibly transparent while minimizing receiver cost and transmission bandwidth (data rate). Once a signal is in a digital form, its quality is determined. The sampling and companding system determines the quality, not the transmission process (providing that bit errors are corrected). This is the key difference between analog and digital signals; analog signal quality is primarily determined by the transmission or storage medium. Digital signal transmission and storage (both audio and video) are certain to play major roles in the future of cable.

DIGITAL TRANSMISSION

There are many options available for transmitting digital signals. The choice of modulation format is based on maximizing signal robustness (least received bit errors) while minimizing occupied bandwidth and receiver cost. The more complex the modulation (i.e. number of data levels) the less bandwidth used, but the more complex the receiver is and the more susceptible the signal is to noise, reflections and interference.

A popular modulation choice for digital transmission is quadrature phase shift keying (QPSK) which uses two data levels on each of two carrier phases. Two data levels minimizes the sensitivity of the signal to interferences, as well as the receiver complexity. Two phases doubles the bit rate in a given bandwidth. Using more data levels (4, 8, 16, 32, etc.)will reduce occupied bandwidth by the power of two used, but a 6dB increase in susceptibility to interferences is taken with each power of two increase in data levels. Figure 1 shows C/N vs BER for QPSK signals. An informative text on digital modulation techniques is Reference 3.

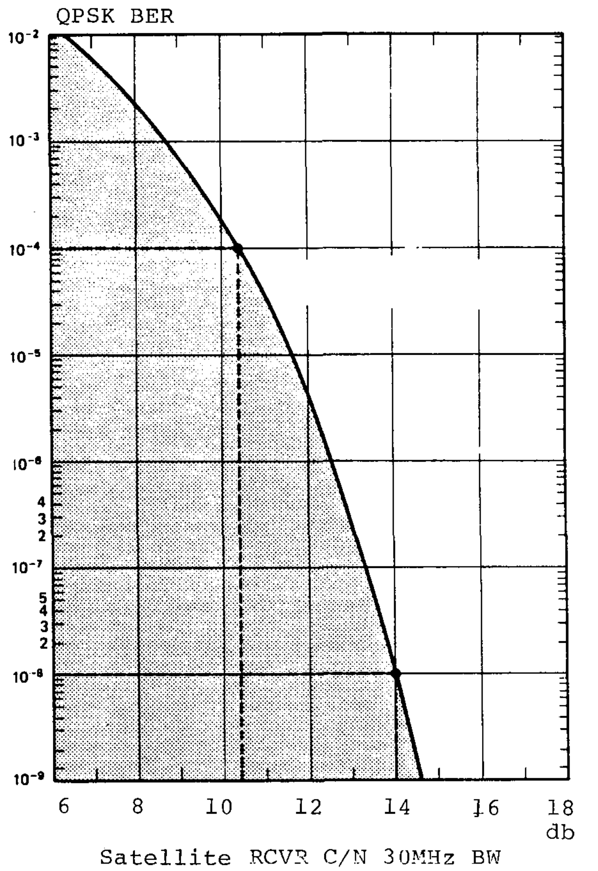

FIGURE 1

DIGITAL AUDIO SERVICE

Commercial free digital audio in numerous musical formats (i.e. rock, country, classical, jazz, etc.) has been proven to be a viable pay service, both in market surveys and market tests. It does not cannibalize video services and has achieved a high level of customer satisfaction due to convenience and quality.

DIGITAL AUDIO SYSTEM

In order to achieve success on a national level, a digital audio system should include the following:

1. A proven addressable control system with major billing system interfaces.
2. A satellite transmission system with no more stringent requirements than those which are required for video reception including immunity to terrestrial interference.
3. A cable transmission system which does not displace present or future video services. The transmission system should also operate to the point of unacceptable video services and not contribute to video distortions.
4. Enough channel capacity to accommodate growth in future audio services. A guideline for the number of audio channels might be comparable to the number of video channels expected in the future.
5. Equipment payback time should be favorable in comparison with other possible service investments.

A system block diagram for nation-wide delivery of pay digital audio services to the home is shown in Figure 2. Key points to consider for each of the blocks follow.

DIGITAL AUDIO DELIVERY SYSTEM

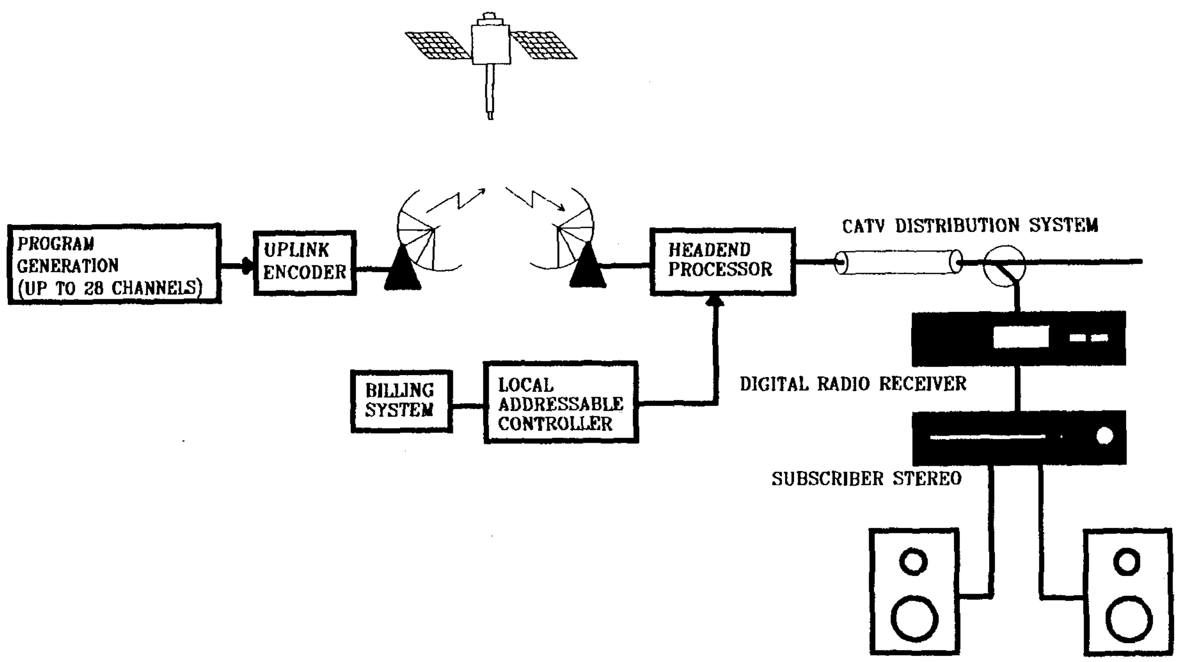

Figure 2.

Program Origination

Each channel of the digital audio service programming is sourced by a computer controlled playback system. The playback system plays from a selection of format dedicated compact discs according to a playlist programmed into the control computer. The playlists for each music format are created by programming experts within that format. The system is capable of continual play, and the playlist and disc stock can be updated as required. See Figure 3.

Satellite Link

Up to 28 stereo audio pairs are converted from analog to digital form by Dolby DP-85 encoders. The uplink encoder encrypts each channel, adds forward error correction and control data before interleaving the channels and framing the multiplexed data. Rather than using a pseudo video signal format and applying it to a standard satellite exciter, the multiplexed data is QPSK modulated at a 70MHz IF frequency and then upconverted to the transponder uplink frequency.

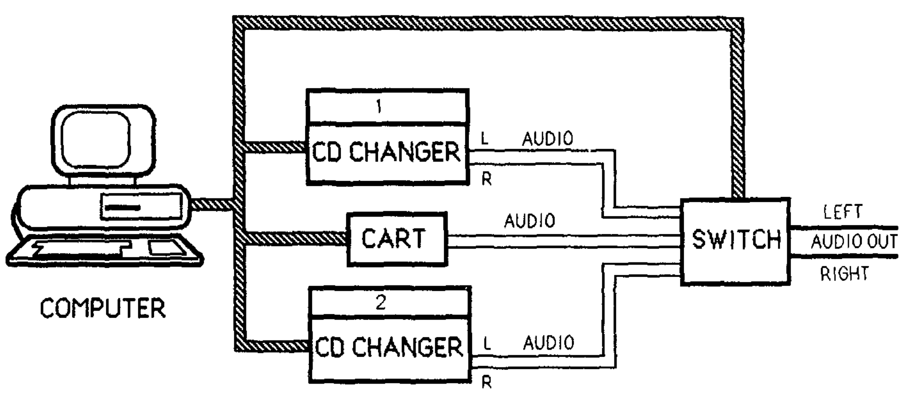

FIGURE 3

This has two key advantages: an improved bit error rate (BER) at the receiver, and a bandwidth narrow enough to allow filtering to reduce terrestrial interference. This system will operate with as little as 10dB C/N. It is desirable to carry the digital audio service on a transponder of a satellite which carries other common cable services. In this case the existing 950 to 1450MHz feed from the existing antenna and LNB is split and fed to the tuner and QPSK demodulator for the digital audio service, as well as to the video service satellite receivers.

Transportation

As shown, the cable system operator will receive the digital audio service via earth station downlink from a cable programming satellite. In those cases where the earth station downlink and CATV headend are not co-located, provision must be made to transport the signals. The CATV community presently utilizes two classifications of equipment to transport video services: FM systems and AM systems. Each of these can be compatible with digital audio.

The FM systems currently employed consist of FM supertrunk, FM microwave (FML) and FM fiber optics. In any of these systems, the digital audio signal will be transported by processing the received satellite IF signal (70MHz) as appropriate to the application. In FM supertrunk applications, the IF signal can be converted to a selected output channel, transported via coaxial plant, and converted back to IF at the CATV headend. FM microwave (FML) applications require that the IF signal be up-converted and amplified for microwave broadcast, then received and down-converted at the CATV headend. FM fiber optic systems will make use of available lightwave processing equipment to deliver the IF signal to the CATV headend.

AM systems in use include standard coaxial plant, AM microwave (AML), and AM fiber optic systems. Transportation of digital audio through these will be accomplished by processing the received satellite signal into discreet QPSK channels (as discussed in the following section), and distributing these in a method similar to processed FM analog audio, or modulated video. Interconnection to the addressable controller will be by established data communications products.

Headend Signal Processing

In order to provide a pay digital audio service that does not take up spectrum suitable for video services, the FM band (88 - 108 MHz) is an ideal location. Ingress generally makes picture quality unacceptable in the FM band. This ingress also makes transmission of wideband multiplexed high speed digital signals unreliable in this band. The solution is to demultiplex the high speed satellite delivered signal and generate individual data carriers which can be spaced 600kHz apart.

Not only are the narrow band signals less sensitive to noise and reflections, but the carriers can be placed to avoid frequencies of maximum ingress or desired existing carriers. QPSK modulation is again a desirable modulation form for transmission over the cable system. Operation with only 15dB C/N or C/I is fully acceptable. A cable system with 35dB C/N video carrying digital audio with QPSK carriers 15dB below video still has a 15dB safety margin for the digital audio service. This is due to the noise bandwidth of the QPSK receiver being about 400kHz instead of 4MHz which reduces noise by 10db relative to video.

The headend signal processing includes transponder tuning, QPSK demodulation, data demultiplexing, error detection and correction, control data multiplexing and data framing, digital filtering and QPSK modulation. Controls are provided for transponder selection, audio service selection, output frequency and output level.

Since the QPSK carriers are more robust than AM video, any transmission component suitable for AM video use will be suitable for digital audio use. This includes AML and AM fiber.

Addressable Control System

Addressable authorization control is an important consideration to the success of the digital audio service. Authorization control is accomplished by first encoding each digital service with a "tag" or identification and then authorizing each tuner to play a package of services. This control is equivalent to and compatible with present addressable control methodology. Additionally, the tuner is configured for operating parameters and channel allocation via the addressable control system. This control can be an upgrade to the existing addressable controller or a standalone system.

The distribution of the addressing data requires a family of communication products that can accommodate a wide range of cable system interconnect architectures. These products have been developed for present addressable control systems and a pool of trained resources is in place to assist in their installation. A wide variety of local and remote (both via RF and telephone) applications can be met.

The business system integration of the digital audio services will make use of existing billing system functions; installing a new converter type, and building a new service package (or packages). The billing system will interface to the addressable control system by using the standard wirelink protocol.

DIGITAL RADIO RECEIVER

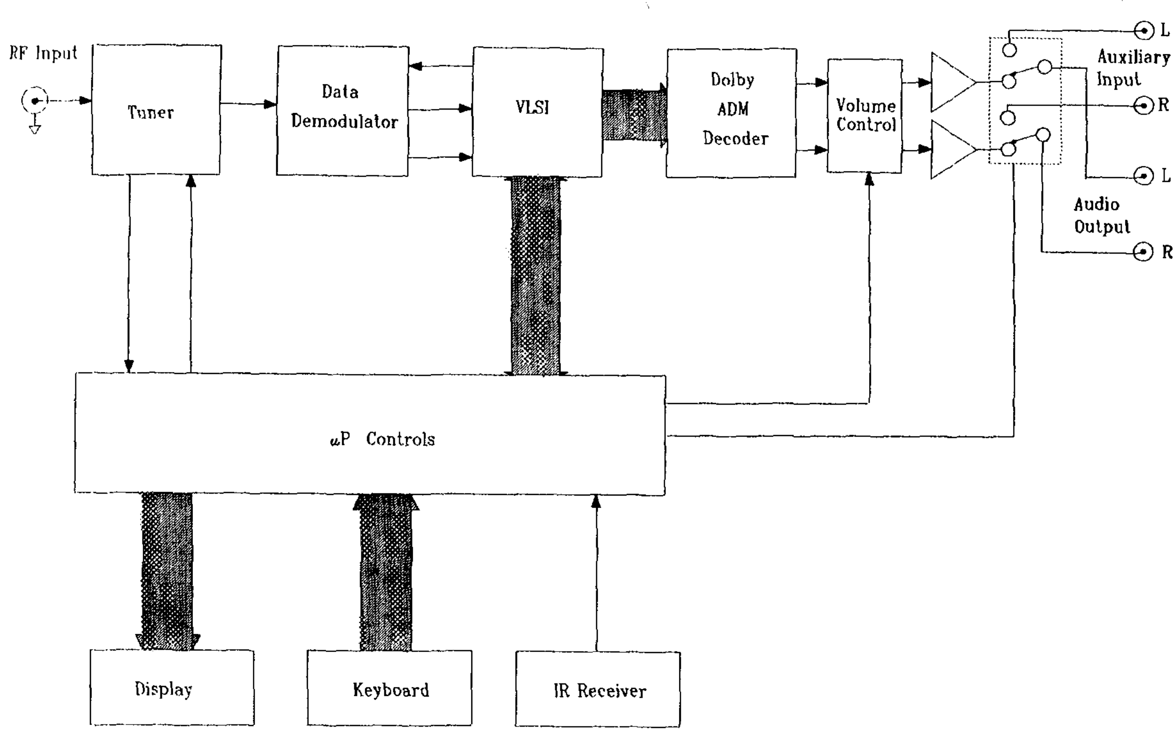

FIGURE 4

Receiving Terminal

The digital radio receiver block diagram is shown in Figure 4. The front end is similar to an FM radio tuner. It is a single conversion tracking tuned RF system. The oscillator is synthesized and operates from a downloaded tuning map. The IF frequency is 10.7MHz allowing the use of wide type ceramic filters for channel selectivity. The 10.7MHz IF QPSK signal is demodulated and fed to a logic LSI which reads and passes control information to a microprocessor.

The microprocessor decrypts control information and in conjunction with the LSI it decrypts the digital audio when authorized. The digital audio data is formatted and provided to the Dolby TM ADM decoder which converts the digital data to analog audio. This audio is then passed through a digitally controlled attenuator to provide the remote volume control function and passed out to the left and right audio output jacks.

Subscriber Installation

The subscriber installation of the digital audio service will be similar to that of an additional outlet. A coaxial feed from the subscriber drop will be routed to the location of the subscriber's home audio system where the tuner will be installed. A wide dynamic range input will facilitate installation on any drop which provides satisfactory video signals using a directional coupler. The tuner will connect to the AUX or CD inputs of the stereo system and connections will be made using standard hardware. The tuner does not require any additional inputs on the subscriber's existing stero amplifier or receiver. Audio inputs are provided on the digital tuner to allow an additional input to the audio system when the tuner is powered off.

CONCLUSION

Delivering digital audio presents the cable system operator with an opportunity to capitalize on the second largest consumer communications market. An integrated system for delivering digital audio to the home can ease the launch of digital audio services into the CATV business environment.

This system can operate on existing cable plant and take advantage of existing business system operations. Finally, the system possesses the flexibility to allow for expansion as the market requires.

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DESIGN CONSIDERATIONS FOR AN ADVANCED TV TEST FACILITY

DAVE WACHOB

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ABSTRACT

Advanced Television (ATV) represents an enormous potential, not only for CATV, but also for the Television and Electronics industry as a whole. Presently, the FCC and others are analyzing the various proposed ATV systems, in an effort to determine the optimum system(s) for delivery. Accurate, thorough and impartial evaluation of the proposed ATV systems is an essential element in the selection process to ensure that the best system is selected.

This paper will describe design considerations for establishing a test facility for evaluating the proposed ATV systems. While there are still many unresolved issues relating to source material, equipment availability and test procedure finalization, facility design and construction can be initiated. Specific areas of the facility design to be discussed include the overall testing strategy, facility layout, video and audio aspects, ventilation and powering requirements, and access and security concerns. The items highlighted will allow for both objective and subjective testing on a wide variety of media.

OVERALL TESTING STRATEGY

In addition to ensuring compatibility with existing NTSC cable systems, ATV testing is critical for a variety of reasons, the most important of which is consumer acceptance. Ultimately, the consumer must decide whether the perceived benefits of Advanced TV justify the cost premium associated with owning an ATV receiver and viewing ATV services. While few will question that ATV can and will technically deliver a better picture, consumer acceptance is by no means assured, particularly given the current cost and availability estimates for ATV receivers. Complicating the consumer decision is competition from pre-recorded tapes and discs, which offer some of the same benefits of ATV at a reduced cost.

Consumer acceptance can therefore be broken down into two parts: 1) Ensuring that the best picture technically possible is presented to the home and, 2) Determining what the consumer is willing to pay to have Advanced Television. Both these elements of ATV testing must be accommodated through objective, subjective, and consumer preference testing in order to undertake a thorough evaluation of ATV. Advance knowledge of this information before actual ATV market introduction is essential to ensure that not only the best system is put in place, but also that once it is in place it will be accepted by the marketplace.

The term "Advanced Television" includes not only high definition televisions systems, but also enhancements and improvements to present day NTSC based systems. Distinction between the two is largely a matter of bandwidth requirements, performance improvements, NTSC compatibility, cost and marketplace availability. Since all forms of ATV may be potential contenders in the marketplace of the 1990's and beyond, they must be carefully analyzed both objectively and subjectively to determine their viability.

When NTSC was introduced, delivery to the home was largely determined by terrestrial over the air transmission, directly to the consumer's television. The advent and growth of cable, DBS, fiber in the future, and other alternative media has considerably changed the outlook for the introduction and delivery of Advanced Television. This new media environment of the 90's requires performance considerations and, therefore, testing on all the possible media responsible for ultimate delivery of the ATV signal in the home.

FACILITY LAYOUT

The design and construction of the physical structure for performing ATV testing requires attention to key details, to assure that a flexible and neutral environment is attained for ATV testing. Adequate provisions for lighting, ventilation, AC powering, interconnections, isolation, access, security, storage, fire protection, and people flow must all be considered in the design. Since audio as well as video characterization of the ATV system is essential, special considerations must also be given to audio aspects of the design.

Ideally, the test facility should be divided into separate areas for subjective and objective testing. This allows for subjective and consumer preference testing in a more controlled environment than a conventional lab would support. Signaling and control between the areas becomes important, particularly during consumer preference testing, where control of the test material, impairments, signal routing or video and audio settings may be desired.

Linear "signal flow" should also be maintained in the facility (from source to presentation) to minimize cable routing and potential feedback problems. Access points to the ATV signal for routing, monitoring, and measurement must also be easily accessible without incurring signal degradation or modification. Signal flow should also allow for convenient interconnection with the desired media, impairments, or simulators to be used during the ATV testing.

VIDEO CONSIDERATIONS

Subjective viewing and consumer preference testing of Advanced Television dictates an environment that is neutral and free from distractions, so that accurate, impartial, and "focused" solo or comparison testing can be accomplished. Critical design parameters to be considered to ensure a neutral viewing environment and minimize distractions include correct ambient lighting, viewing distance and offset angle, wall treatments and environmental control. Psycho-visual research and good engineering practices influence the majority of these key parameters.

Specific ambient lighting and wall preparation requirements for the subjective viewing area are highlighted in CCIR Guideline 500-3. The key emphasis here is that the ATV system under evaluation is to be the "center of attraction" so that the eye (and mind) is not distracted by the surrounding environment. Neutral walls, ceilings, and floors, in both textures and color are optimally recommended to retain subjective attention. Ambient lighting set to 10% of the viewing screen illumination is appropriate. Additionally, a computer controlled lighting system is recommended to "remember" preset lighting conditions to remove ambiguity.

Viewing distance and angle have an immediate impact on the physical constraints of the test facility, in that they dictate room size and shape. Conventional wisdom supports the premise that ATV subjective testing should be done at a distance 3-4x's picture height. While the true benefits of Advanced Television may ultimately be viewed on large screen (>35") television, present display technology still has a way to go in terms of brightness, resolution, viewing angle and cost to realize these benefits. Even with technical advances in this area, it is unlikely that the shape and size of the conventional consumer's living room will change dramatically in the next 10-20 years to take full advantage of Advanced Television receivers.

Complicating the problem of determining a viewing room size is that motion artifact processing will be a key evaluation parameter when the various ATV systems are compared. Supported by psycho-visual research, most ATV systems improve resolution of a static scene at the expense of moving scenes, resulting in reduced resolution during motion, and potential motion artifacts. At close distances (1-2 picture heights), motion artifacts and reduced resolution become more apparent. Further away (beyond about 4 picture heights), they are not as apparent, but unfortunately, neither are the benefits of ATV when compared to NTSC!

Putting all this together yields some minimum and maximum viewing distance requirements for the controlled viewing area. Accommodating a 20" ATV receiver at 1x's picture height yields a 1 foot minimum distance requirement, whereas a 70" ATV receiver viewable at 5x's picture height yields a 17.5 foot maximum distance requirement. Nominal viewing testing would probably be done at 3-4x's picture height, which, assuming a 35" entry level ATV set, would be 5-7 ft, consistent with a typical living room. The viewing angle should be kept within a +/- 30 degree offset from the center line to ensure that adequate detail is available to all the subjects taking part in the viewing testing.

AUDIO ASPECTS

In some respects, the audio aspects of the facility design are more demanding than video, and require even greater attention to detail to minimize the background noise during subjective evaluations. Potential sources of background noise include ventilation, heating and cooling systems, aircraft and environmental noise, power system hum, test equipment residual noise, and good old "people noise". Minimum background noise requirements are highlighted in IEC standard, NC-20.

Beyond background noise, it is also important to minimize distraction to other adjacent test areas during actual subjective audio testing of ATV systems. Fortunately, minimizing background noise <u>into</u> the subjective area also generally minimizes test noise <u>out</u> of the subjective area into adjacent test areas.

Noise is contained through a variety of design techniques, such as floating floor construction, minimizing heating/cooling airhandlers noise, soundproofing of walls, ceilings, and ductwork, isolating access ways into the facility, and reducing transformer AC noise. These concepts are based on sound acoustical practices, and are supported by qualified architectural design.

An optimum reverberation time of .3-.4 sec, 500 HZ octave band is also preferred in the IEC specification. Although specific characteristics can be custom tailored to particular preferences and "listening environments," a more preferred approach would be that which simulates the home environment.

VENTILATION/POWERING REQUIREMENTS

Given the powering (and therefore heat generating) requirements of the ATV hardware to be used during evaluation, obtaining adequate powering and ventilation cannot be overlooked.

While final ATV hardware may someday be "C-MOS custom IC based", the available test and evaluation hardware for the next several years will be anything but! With limited ATV hardware information presently available, an approximate rule of thumb would be to allow provisions for powering and ventilation requirements consistent with 10-15 Kwatt of power dissipation for proponent hardware, plus evaluation and test equipment hardware. Wherever possible, uninterruptable surge protected power supplies should also be considered to reduce down time and equipment damage.

Adequate ventilation in the viewing area is also important for several reasons, primarily to prevent distraction of the viewing subjects during testing. A "comfortable" level of temperature and humidity must be maintained during testing in a smoke free environment. Ventilation must also conform to minimum noise requirements, as discussed in the audio section.

ADEQUATE ISOLATION

In addition to sound isolation mentioned previously, provisions for RF, cable and AC power isolation must be maintained. RF isolation is important to prevent off air interference from either local broadcast channels or spurious interference. Cable shielding becomes essential when routing control and signal cables in and out of the facility, particularly the subjective area. AC power isolation is important to reduce 60 cycle hum and ground loops. All can be minimized with prudent engineering shielding and grounding techniques during the design and construction phase.

ACCESS AND SECURITY PROVISIONS

People access to the test facility must be carefully controlled to prevent unauthorized equipment removal and traffic into the facility. This is particularly important given the high capital expense of the equipment required to support ATV testing. This includes not only equipment provided by the test facility, but also on loan by the proponents.

Doorway access is also important, given the large size of the decoder and test hardware presently available. A minimum door width of 42 inches between critical areas and outside doors is necessary to transport the required equipment in to the test facility. This width should also accommodate shipping containers necessary for transporting the required proponent and test hardware.

Security of the facility is also critical from the standpoint of access to proprietary information, some of which may not be intended for public distribution. "Off hours" tampering could also be a concern, and must be controlled. Given the high stakes involved in the future of Advanced Television, this item must not be treated lightly in the design.

SUMMARY

The design of an Advanced Television Test Facility can be achieved, that will support objective, subjective, and consumer preference testing. Careful attention to detail is, however, required to insure that key aspects of the design are considered beforehand. Successful completion of the facility will provide the foundation for thorough, impartial evaluation of Advanced Television Systems.

REFERENCES

1. CCIR Recommendation 500-3

2. IEC Standard NC-20

ESTABLISHMENT OF BATTERY STANDARDS FOR CATV STANDBY POWERING

Larry Lindner
Bob Bridge
Alpha Technologies

Charles Marks
Johnson Controls

INTRODUCTION

The reliability of cable television signal distribution systems is increasingly dependent on the reliability of 60 Volt A.C. power back-up. The decreasing quality of utility power, and increasingly frequent interruptions in delivery of power add to the list of challenges to the cable system engineer in his efforts to improve service to subscribers.

Manufacturers of standby power supplies have traditionally left the decisions on selection, installation, and maintenance of the battery sub-system to the cable engineer. Without standards of quality, a clear definition of the requirement, and with potentially unrealistic expectations respecting battery service life, the industry has experienced wide variations in performance and reliability of its standby powering equipment.

It is the object of this paper to lay groundwork for establishment of a process leading to the creation of standards for battery usage in the CATV standby power application. The paper reviews the nature of the CATV standby power battery service requirement, and compares this requirement to available classes of battery service. The special needs relating to temperature, environment, price/performance, and maintenance are discussed, and comparisons made within two main battery types.

The paper concludes with recommendations affecting the use of AGM (absorbed glass mat) batteries and Gelled Electrolyte batteries. Recommendations are also made respecting the establishment of a Battery Standards Committee within the SCTE.

BACKGROUND

CATV Standby Power Battery Service Requirements

Size, Voltage Rating and Discharge Capacity

Typical standby power supplies in CATV use a 36 Volt D.C. inverter as the back-up source. Some designs remain that use a 24 Volt source, although these have largely passed from common usage. Dual parallel battery strings are in use where higher capacities are required to support longer run times. The "standard" 3-battery series-connected 36 Volt hook-up will be the example used throughout this paper. At typical run times of 1.5 to 3 and 4 hours, at 60 VAC loadings of between 7 and 12 Amps, battery sizes in the "group 31" case size offering capacities of greater than 150 minute reserve capacities are commonly used. Batteries commonly produced for the automotive, traction, and marine markets suggest themselves for this application, and are widely available in the capacity range indicated.

Charging Requirements

1) Float Service

The standby power supply battery must be continually maintained at a state of full charge, in anticipation of a discharge cycle at unpredictable intervals. Float service is commonly used by power supply designers. A constant-voltage float is impressed permanently across the battery string. Current flow into the battery string is determined by the total battery series impedance, a function of charge level. At the correct float voltage, a fully charged string of 3 batteries will continuously draw only a small amount of replenishment current to compensate for ongoing self-discharge. At any level of discharge, a demand current proportionate to the required charge will be drawn from the float charger.

2) Temperature Variation

The standby power battery will be used in outdoor enclosures and is therefore subject to most environmental conditions. The most important of these will be temperature variation. The battery will be required to charge effectively, to retain significant discharge capacity, and to provide reasonable service life over a wide range of temperatures. Temperatures will range from extremes well below 0°F to high values in the 120°F to 135°F range. Float charge systems must be temperature compensated to avoid overcharging at high temperatures, and to avoid inadequate charge levels during periods of extreme cold.

3) String Charging

The 36 Volt inverter battery comprises three 12 Volt batteries in series. Use of a single charger to float the three batteries at equal voltages requires that the single float voltage value be divided across three equal-impedance batteries. This requires that the batteries be of identical production, identical age, and in identical condition. Variations between batteries, or early failure of a cell within a battery, will unbalance the charge voltage distribution, usually resulting in overcharge of the remaining batteries. Batteries produced with low-cost materials, and/or in a poor quality-control environment, may have inherent imbalance conditions even when fresh from production. Carefully matched batteries, known as "UPS Grade", have been considered outside the price range of CATV budgets.

Discharge Service

Cable back-up requirements come in unpredictable spurts, depending on local conditions of weather and utility reliability. Outages may occur typically for very short periods on average, but on occasion may be lengthy. National averages have been reported to be in the 20 minute range, but outage classifications within CATV have not been standardized. Outages of seconds, or minutes in duration, are common. During periods of storms, extreme lightning activity, or severe winter weather, outages may last for days.

Battery back-up is normally expected to provide inverter power for up to two hours on average at typical loads. Many CATV franchise agreements today call for four-hour run times, but these are largely unrealistic and in any case are seldom referred to a required amount of load power.

Discharge events can be said to be at random intervals, of widely varying duration. For events which last longer than available capacity, the battery may remain without charge for the duration of the outage until restoration of power by the utility or generator. These events may be at extremes of temperature.

Discharge Current Levels

Standby power supply rated load values are currently in the 10 Amp to 28 Amp A.C. range. This translates to power requirements in the 700 Watt to 1800 Watt range, at 36 Volts D.C., when the inverter is called on to support cable loads and operation of the power supply circuitry. For a 36 VDC string, discharge currents will therefore be in the range of 20 to 50 Amps. By far the majority of standby power supplies will operate in the 12 Amp A.C. output range, (at 60 VAC), calling for a 23 Amp average discharge current level from inverter batteries. "Cranking" ratings, given as "C.C.A.", or "cold cranking amps", describe short-term discharge ratings for batteries used in automotive starting applications. For standby service applications, most 12 Volt batteries in use today are rated for "reserve capacity", in minutes, at 25 Amps discharge. The reserve capacity rating then, is almost directly equivalent to "standby time" attainable at average CATV loads. Batteries used in cable applications should be compared on the basis of reserve capacity, not "amp-hour" ratings or C.C.A. figures.

Vibration Environment

Batteries in most CATV applications are mounted on poles in enclosures at 10 to 20 ft. off the ground, or on concrete pad-mounted enclosures. These are all in direct contact with traffic-generated vibration on cable rights of way, and this vibration is relatively constant. Occasional "whipping" or jolts due to falling ice, high winds, or adjacent use of heavy equipment may be experienced. Mechanically, the environment is not exceptionally stable.

Maintenance Environment

Maintenance of line equipment and subscriber connections tends to dominate the time of the outside plant maintenance staff. Power supply familiarity and battery test procedures are generally low on the priority list of many cable technicians. This can lead to low levels of understanding of the role of the battery in the standby equation. On the plus side, training programs from vendors and a growing sensitivity to power protection is rapidly improving maintenance levels focussed on standby power supplies. Progressive cable operators, using optimum maintenance schedules, may direct field technicians to visit power supplies quarterly or even more often, but correct procedures relative to checking battery health remain somewhat hit-and-miss. Virtually no directives exist which mandate removal of batteries from service until after the power supply ceases to function. It appears that battery *failure* is the only symptom that is universally recognized by the industry as indicative of the end of service life.

Battery Chemistry Review

The lead-acid battery, in common use for over a century, in size, price, and capacity ranges, suited to the CATV application, are considered in this paper for their suitability. A review of lead acid chemistry follows.

The chemical reaction between lead dioxide (PbO_2), lead (Pb), and sulfuric acid (H_2SO_4) provides the lead-acid battery's energy. The electrolyte, which is a 1 to 3 solution of acid in water (under full charge) at 77°F. may be in free liquid (flooded cell), finely distributed into a fibrous absorbent carrier (absorbed glass mat), or suspended in a mineral gelling material (gelled electrolyte). The "active" material of the positive electrode is the lead dioxide, and the active material of the negative electrode is metallic lead (Pb). Both active materials are present in the electrode structure as paste, mounted in a metallic lead frame or "grid" structure. The pastes are porous to facilitate required amounts of surface area needed to support anticipated discharge currents.

On discharge, the active materials are converted to lead sulfate ($PbSO_4$) which is deposited on the grids. The sulfuric acid solution turns to water (H_2O) as the sulfate is consumed and water is produced. During recharge, particles of lead sulfate are converted back to sponge lead at the negative electrode and to lead dioxide at the positive, by the charging source driving current through the battery. The well-known reaction is:

$$\underset{\text{Lead + Lead Dioxide + Sulfuric Acid}}{Pb + PbO_2 + 2H_2SO_4} \;\underset{\longleftarrow}{\xrightarrow{\text{discharge/charge}}}\; \underset{\text{Lead Sulfate + Water}}{2PbSO_4 + 2H_2O}$$

Battery Construction Review

Grid Alloys

Lead alloy grids are used to mechanically support the active electrode material. Grids are produced from cast lead, or lead sheet. As pure lead is too soft for grid material, alloying materials such as calcium or antimony are added. Antimony, in 4 to 6% amounts, tends to dissolve from the positive plate and redeposit on the negative plate. This affects operating voltage, increases water consumption, and reduces life as the antimonial battery ages.

Calcium was first used in telephone battery applications as a hardener. Service life was lengthened, and less frequent watering was required. Amounts vary from .03 to 0.10%.

Grid Pastes

Lead oxide pastes applied to grids by hand or machine contain measured amounts of a mixture of lead oxide, sulfuric acid, and water. Fibrous additives may be present for increased strength and binding ability. Expanders may be used to increase porosity, and subsequent current-supporting surface area.

Pastes with low density (high porosity) are used in high-rate shallow discharge service. Pastes where a deep discharge is required at relatively low rates need to be higher in density (less porous). Deep discharge stresses the paste material as it grows during discharge on the grid structure. These stresses will work to loosen the paste material from the grid. High density paste will hold onto the grid through the stress of deep discharge. Each deep discharge cycle of the battery will obviously work to weaken the paste-grid bond, and batteries in deep-discharge service will be cycle-life limited.

The paste curing process during production helps to determine its service characteristic. Flash drying after initial pasting, and subsequent lengthy cures at specific temperatures and humidity levels will determine paste-grid bonding and paste densities.

Battery Assembly

Positive and negative plates are electrically insulated from each other by separators. Separators include PVC, phenolic/celluloid, microporous rubber, or polyolefin. Some sealed batteries use glass fiber mats. Separator materials are porous or permeable to allow ionic conductivity. Current flow through separators will be a function of pore size in the separator: internal impedance then, is partly a function of separator material. Separator types can have a dramatic effect on service life.

"Formation" of the battery is the initial charging of the assembled unit. Electrolyte is added to the battery and charging current applied. During this process, the sulfuric acid in the paste material transfers to the electrolyte, resulting in a final specific gravity higher than that of the electrolyte added. The battery is now ready for service.

<u>Battery Service Categories</u>

In producing batteries for different types of service, the production parameters such as grid size, alloys used, paste mix, paste density, paste cure, separator material, and the internal physical dimensions of the case may all be varied and optimized. The service requirement will be dictated by parameters of capacity, cycle life, "float service" life, cost, and discharge depth. The majority of 12 Volt lead-acid storage battery service requirements have been divided into three classes of service: "SLI" or "starting-lighting-ignition", "Deep Cycle", or "Traction" batteries, and "Stationary" batteries.

SLI Batteries

An ability to deliver maximum amounts of current in minimum amounts of time characterizes this category. The automotive starting application is the most familiar use for this type of battery. The battery design and construction is therefore optimized for low impedance by maximizing plate exposure to electrolyte. This calls for thin plates (to maximize number of plates) and low density (porous) active material. Discharge depth is typically shallow, as perhaps only 1% of the batteries total capacity is taken off during a typical 'short burst' discharge in a starting application. If deep discharges are taken off of an SLI type battery, very few cycles will be available, as the paste-grid bond is severely stressed by the electrode growth during the discharge. The thin plate structure and porous paste will be seriously weakened by deep cycling. SLI batteries are also characterized by low-cost materials and economic construction.

Deep Cycle Batteries

The deep cycle application implies discharge cycles to as much as 80% of the capacity of the battery. The use of thick grids with high paste density is required to support the stresses of discharge to these depths. Premium separator material is usually required, with possible additional use of fiberglass mats, to assist in support of the paste material on the grids. Reduced surface area of plate to electrolyte contact will increase internal impedance and reduce values of discharge currents. Typical applications are in running small electric trolling motors, and in supporting inverters or other DC equipment in motor homes and boats. ("RV" and "Marine").

Stationary Batteries

Batteries which are required to support lengthy discharges (up to 8 hours) and long life in service, particularly in the 48VDC battery plants used by telephone company central office equipment, have come to be known as "stationary batteries." These units are normally racked in place, and are typically high-quality flooded designs using premium materials and high-quality construction methods. These batteries are assembled with much more care than automotive or deep-cycle products. Stationary units typically are rated for 20 year life on float service.

Sealed Batteries

All three of the main service categories may be produced in sealed designs. Sealed batteries are designed to eliminate the need to add water, increasing the convenience and safety to the user.

Flooded batteries, typical of the SLI category, may be sealed if extra electrolyte is used to provide the additional reserve of water. Lead calcium grids should also be used in a sealed flooded model to minimize gas production. These units are normally supplied as "maintenance free" batteries in the SLI market.

AGM batteries, or "absorbed glass mat" products are also known as "starved electrolyte", "acid limited", or "recombinant" batteries. In this sealed lead-acid product, a mat of fine glass fiber material functions as a plate separator with up to 95% void volume in the mats. The mat acts as a "sponge" to hold electrolyte between the plates. About 95% of the void volume is filled with acid, leaving residual volume for gas transfer during recombination. H_2 and O_2 gases, freed from the electrolyte during discharge will normally vent to the atmosphere in flooded batteries. In the AGM product, these gases will recombine with the negative plate and the electrolyte during recharge, with virtually no gas production released to the atmosphere. An internal overpressure is maintained in AGM products to aid in recombination by retaining the gases long enough to recombine.

Gel Batteries

Gelled electrolyte batteries function as starved electrolyte batteries, but do not use a glass mat. The high-density gel material takes on a porosity through development of "fissures" in the gel during initial formation and charging of the battery. Gas transfer during recombination then is facilitated by way of the fissures. Gel materials in common use are fumed silica, sodium silicate, boron phosphate and polymer microspheres. Gel products have exceptional shelf life. Self-discharge rates in storage may average 3% per month at room temperatures.

Battery Charging

Charging methods for lead-acid batteries may be divided into:

1. constant current
2. taper
3. pulse
4. trickle
5. constant potential

Constant current chargers force a fixed value of current through batteries on recharge irrespective of the state of charge of the battery. This can lead to overcharge. Trickle chargers are simply low-rate constant current chargers. Pulse chargers use current pulses which are periodically disconnected to measure open circuit voltage. Constant potential chargers (float chargers) deliver current in proportion to the charge needs of the battery. Float chargers need to be temperature compensated. Taper chargers are low-cost constant potential chargers.

As batteries are only about 80% efficient, more energy is delivered to the battery during recharge than is taken off during discharge. During the last 5 to 10% of recharge, a normal "overcharge" condition exists. Gassing occurs during this portion of recharge. In a flooded battery these gases are released to the atmosphere; in AGM or Gel products, gases are recombined.

<u>Lead-Acid Battery Failure Modes</u>

Loss of Capacity

Normal end-of-life will be corrosion of the positive grid. Corrosion of the positive grid is the slow conversion of the lead alloy material to lead dioxide, resulting in a loss of support for the active material. As the grid material corrodes, its resistance increases. This increasing internal impedance is equivalent to decreasing capacity. As the grid corrodes, the lead dioxide material takes up more room than the lead alloy did: this is "positive plate growth". Without design for this expansion, internal shorts result. In designs where low impedances and high-rate discharges require thin, closely spaced plates, this type of failure mode is more likely than in low-rate, thicker-plate designs.

Thicker grids will obviously take longer to corrode away. Long-life batteries therefore are characterized by thick grids. Stationary batteries may have grids up to 0.30". For similar capacity ratings, (say in the 160 minute reserve capacity category) an SLI flooded grid will be .030" to .040" in thickness, whereas Gel and AGM products may be up to .115". The AGM product, with the mat acting as electrolyte support, needs less room between plates for separator and electrolyte, allowing greater thickness to be used, resulting in closer plate spacing.

Paste Failure

During charge and discharge cycles, lead sulfate builds up on the lead dioxide plate material. Lead dioxide is a brittle, ceramic-like material, and is more dense than lead sulfate. This repeated expansion and contraction of the composite plate stresses the active material, causing it to break loose from the grids. Deep cycling obviously causes great stress. Deep cycle batteries may incorporate several layers of material or envelopes over the plate to retain the active material in place. Again, high density materials will tend to remain in place in the grid under the stresses of deep cycling.

Shorts due to Dendrite Production

Cell shorting can occur in several ways. Most commonly it occurs when the design has not allowed for positive grid growth. Cell shorting can also occur, due to "dendrite" formation, when batteries are very deeply discharged. Dendrites are conductive lead paths that penetrate the separator. Dendrites occur as the battery is discharged and the electrolyte specific gravity (S.G.) is lowered. Lead sulfate from the plates becomes more soluble in the low S.G. electrolyte. Upon recharge, this lead sulfate turns into metallic lead forming dendrites. AGM batteries seem to be more susceptible to this deep discharge problem as they are designed to be an "acid starved" system and to use up most of the sulfuric acid. The gelled product maintains a higher specific gravity which tends to inhibit lead sulfate solubility and resultant dendrite formation.

Accelerated Grid Corrosion

Positive grid corrosion occurs under all conditions in a lead-acid battery. The rate of corrosion is dependent on grid alloy charging voltage, temperature, and electrolyte specific gravity. Improper float voltage, high temperature, and elevated electrolyte S.G. will all increase the rate of positive grid corrosion. In general, AGM batteries operate at a slightly higher S.G. than gelled batteries to compensate for the limited electrolyte volume. If this type of battery is not designed with the slightly higher grid corrosion rate in mind, it will fail prematurely, particulary at higher temperatures. High float voltages that increase grid corrosion point out the need for proper temperature compensation.

Sulfation

Normal production of lead sulfate during discharge is converted to the usual reaction products during recharge. If batteries are left to stand on self-discharge and experience lengthy periods of stand-time at partial charge levels, a hardening of the lead sulfate sets in, which is much more difficult to break down on recharge. This hard sulfate acts to close the "pores" of the active material. Obviously batteries should be charged as soon as possible after discharge. Sulfation is accelerated at elevated temperatures. Sulfation can be reversed somewhat by sustained overcharge, to break down the sulfate material. This will produce excess gassing and is not recommended for sealed batteries. Excess gases in sealed batteries will be vented by pressure-relief valves, resulting in reduced battery life.

Sulfation is also enhanced by low electrolyte level in flooded batteries (at the exposed tops of the plates), by higher specific gravities, and by higher operating temperatures. Continuous undercharging due to inadequate float voltages or inadequate temperature compensation will increase sulfation rate.

Operation at Elevated Temperatures

The chemical processes at work in all lead acid batteries will proceed more rapidly as a function of increasing temperature. The rule of thumb is "..for every 15°F. of continuous operating temperature over 77°F., battery life expectancy will be halved.." This implies that a battery with an estimated 60 month life at room temperature will have a 30 month life if operated continuously at 92°F. At 107°F., the anticipated life is reduced to 15 months. Battery life can be maximized through operation at reduced temperatures, and with correct temperature compensation of charging current.

CONCLUSIONS

It is evident to CATV engineers that a sealed-construction battery is first choice for the standby power supply application on grounds of ease of handling, safety, UL approvals, and local electrical codes. Flooded batteries are to be avoided due to hazards in shipping, transporting to power supply locations, warehousing, disposal, and lack of approval by UL or local jurisdictions for use on poles or in ground-mounted installations on public rights-of-way.

The choices are reduced to Gelled electrolyte products or AGM products, which have been made available in the capacity and price ranges compatible with customary cable industry budgets. This reduces the choice to sealed, non-flooded products aimed at deep-cycle ("traction") applications and stationary applications.

AGM or Gel products with thick plates, designed for potentially lengthy discharges without cycle damage will be suitable candidates for use in the CATV application. The greater susceptibility of the AGM product to accelerated corrosion, dendrite growth, plate growth, and capacity reduction, with all of these increased at higher temperatures, will result in shorter service life for this choice.

AGM units will be economic choices in cooler, or relatively constant room temperature environments. The Gel product should offer lengthier service life, all things being equal, as it appears less susceptible to the processes leading to failure. "Dry-out" of the Gel product due to sustained overcharge at elevated temperatures has been reported as a common failure mode of this battery type in power supplies without temperature compensation circuitry.

Imbalances in internal impedances due to production process variations can be expected to have equal effects on both types. The lower production cost of the typical AGM product relative to the Gel product suggests more proneness on the part of the AGM to variations between individual batteries in a given production run. Gel products now made for long life stationary battery applications can be expected to have tight quality control content and higher quality materials used in construction than AGM's manufactured for trolling motor markets and subsequently marketed as CATV standby power supply batteries.

Prices of AGM products made for mass markets have been a significant attractant for tight CATV budgets. Typical purchases have been driven by "price per unit", with little regard for service life. Simple arithmetic points out that a battery product with a typical two-year service life supplied to the market at (say) $75.00, will have an annual amortization of $37.50. Another product with a potentially longer service life of (say) three years, at pricing of $85.00, amortizes at $28.33 per year. When price/performance comparisons of service life are made in this way, the "more expensive" product may be up to 25% less costly in the long run.

RECOMMENDATIONS

A committee to serve the needs of the CATV community, specialized in battery expertise, should be established to provide guidelines on battery evaluation, battery maintenance, and battery selection for use in CATV standby power applications. The committee could be established under the auspices of the SCTE, and should be made up of engineers from the battery industry, cable engineers, and power supply designers. With the help of such a committee, and the guidelines it could supply, users of standby power supplies will be able to plan battery purchase and usage guidelines for the most efficient use of their powering budgets. Maintenance planning and battery replacement criteria will also benefit immensely from the establishment of clear standards.

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EVOLUTION OF THE BRITISH TELECOM SWITCHED-STAR CABLE TV SYSTEM

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ABSTRACT

After briefly reviewing cable TV progress in the UK, the more advanced developments will be described, focussing on the British Telecom switched-star network, deployed already in Westminster and now with a second generation design coming to fruition. The evolution of the switched-star concept towards a multi-service network is discussed, and an implementation aimed at a small field trial in 1990 is described.

INTRODUCTION

Cable TV has never been widespread in the UK. Nevertheless a significant industry started in the 1950s, using mostly HF provision on twisted pairs, to cover areas of poor off-air reception. This peaked at around 3.5 million homes in the early 1970s, but was dampened as the comprehensive programme of installing UHF transmitters round the country came to completion.

By the late 1970s cable TV was in need of a major boost, and attention turned to the twin thrusts of service improvement (including more channels) and modern technology. British Telecom (BT), who had always been allowed to compete for cable TV franchises, had at that time a few conventional coaxial cable systems. The largest was at the new town of Milton Keynes, which was chosen as the site of a small all-fibre trial to the home using the switched-star concept [1]. Implemented in 1982, it coincided with an upsurge of interest in the potential of cable TV and the advent of subscription film channels.

From its experience with this trial, BT moved to a design for a switched-star network (SSN) that was more practical for large-scale deployment; this was chosen for installation in Westminster, one of 11 large franchises awarded in November 1983 as part of a major initiative on cable TV in the UK. This network and its capabilities are described in the next section in more detail.

The combination used in Westminster of fibre to a street-located switch point and coaxial cable to the home is still seen as the most cost-effective solution, and remains the basis of a second generation design that is nearing completion. The experience gained allied to improved technology has enabled significant improvements to be made, which are described later in the paper.

There is worldwide interest in future broadband multi-service networks, and so effort has been applied to extending the switched-star concept towards this aim, essentially by marrying up the cable TV and remote telecomms multiplexer ideas. These evolutionary developments will be described to indicate the potential that the switched-star concept has to meet future service demands.

EXPERIENCE WITH THE WESTMINSTER SSN

The Westminster system started service in mid-1985, and will shortly reach its initial target of 200 Wideband Switch Points (WSPs) (each WSP being capable of serving up to 300 customers). The network has been described in detail in several papers [2], [3], but Figure 1 shows the topology used, and a brief outline of its operation follows.

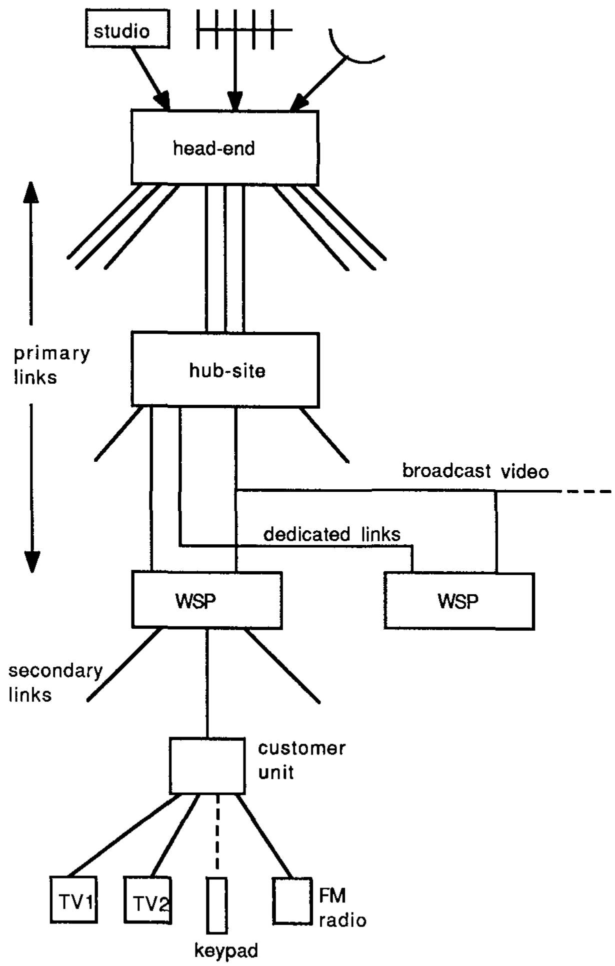

FIGURE 1 - Westminster switched-star network topology

Multimode fibre optics is used to carry multiplexes of 4 frequency modulated (FM) video channels per fibre from head-end to WSP via an intermediate repeatering and fan-out point, the hub-site. This is the primary network which saw one of the earliest uses of large numbers of optical splitters, taking advantage of the spare optical budget to distribute the broadcast TV channels more economically.

At the WSP the incoming channels were demodulated, switched at baseband (0 to 6 MHz for PAL) by DMOS-FET semiconductor devices, and then amplitude modulated (AM) onto a coaxial cable to go over the secondary network into the home. This secondary link carried 2 TV channels on 40 and 56 MHz carriers plus the FM radio band (88 to 108 MHz); signalling placed at the bottom of the spectrum carried customer requests back to the WSP where they controlled the switch directly or, for the more sophisticated services, were relayed back to the head-end. The low total bandwidth on the link allowed a reach of up to 500 metres.

The Services

Broadcast TV was of course the core service, but the network was aimed at providing much more. Text generators housed at the WSP feed into the switch. They provide standard captions and user guidance, but are also the vehicle for information services using data fed down the primary link. A user is allocated a text generator, whilst his requests are relayed back to the head-end to access information either stored in databases there or in external databases accessed via a gateway.

Video library service was also built in from the start, with a limited number of dedicated video links to each WSP to carry channels from the library at the head-end via the switch down to customers. Signals from the customer are relayed up the network to fully control an allocated video disc player, thus enabling flexible access to the contents of a disc.

A carefully designed control structure was built into the network with a hierarchy of processors in the customer equipment, the WSP, the hub-site, and the head-end linked together by a messaging system. At the head-end there is a particularly comprehensive network management and systems administration software package, as well as packages for the information and video library services.

Experience

Technically there have been few problems, the worst being early failures of the 0.85µ lasers, now fortunately overcome. Concern over how the WSP equipment would fare housed in a street-sited cabinet proved unnecessary; the environmental control provided by fans and a heat exchanger has proved very successful. The only real disappointment has been the slow build up of cable TV in the UK; there are signs that this may change and that the switched-star scheme will then get a chance to be exploited to the full.

MARK 2 SSN

The system for Westminster had been designed within a tight timescale to meet an urgent operational demand. It had performed remarkably well, but experience and improving technology were showing where gains could be made; it is often in the move from the first to the second generation design where the greatest degree of benefit can be realised. Hence virtually from the moment that Westminster became operational there was work proceeding on a Mark 2 SSN design.

The Switch Point

The heart of the system is the WSP and this became the first focus of attention for improvement. To reduce size and cost a new more integrated switching chip was developed based again on the successful DMOS-FET crosspoint. At the same time it was decided to change the switch unit construction from a special shelf module to a conventional card based system, allowing standard shelving and racking to be used. The new switch card is shown in schematic form in Figure 2 . It has 48 inputs buffered such that each can go to one or all of 16 outputs. The transmitter units sit as daughter boards on it, taking as standard 2 outputs to each customer. Thus this compact card of 220 x 230 mm serves 8 customers. It can be controlled from customer signalling incoming from the coaxial link or centrally by parallel or serial lines. The opportunity was taken to incorporate some additional return switching capability for video services and monitoring purposes.

The smaller size of the switching circuitry allows the WSP cabinet to reduce to two-thirds of its previous size; there is just one switching bay instead of two, with still the common services bay. With the new switch card, standard 19 inch racking and eurocard shelving could be used throughout; Figure 3 shows the full layout. Since cabinets were often against a wall or fence, it was decided to avoid the necessity for rear access by having only shelf backplanes (no rack backplanes)

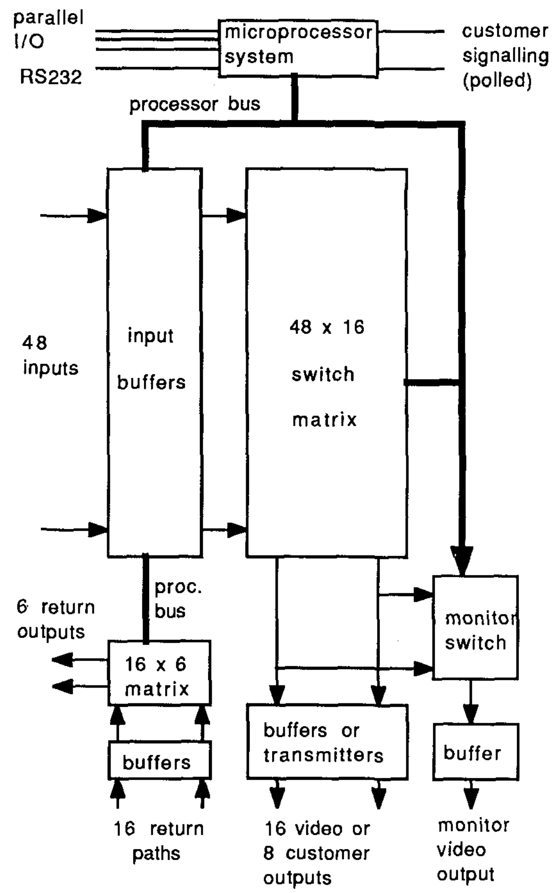

FIGURE 2 - Schematic of the second generation switch card

and all shelf input/output ports brought to the front by interconnect cards.

With the circuitry now more compact an improved heat exchanger arrangement was devised, with all environmental aspects (temperature, humidity etc.) alarmed back to the head-end. Both glass reinforced plastic and stainless steel cabinets can be used, as in Westminster, with the latter having the edge in terms of ease of handling and RF screening.

Customer Service

A new, but compatible customer link was designed so that existing customer equipment could still be used. Plans have been made for an improved set-top unit, somewhat cheaper and with additional display facilities (e.g. message waiting). Greater integration of the coaxial transmitter unit was possible using

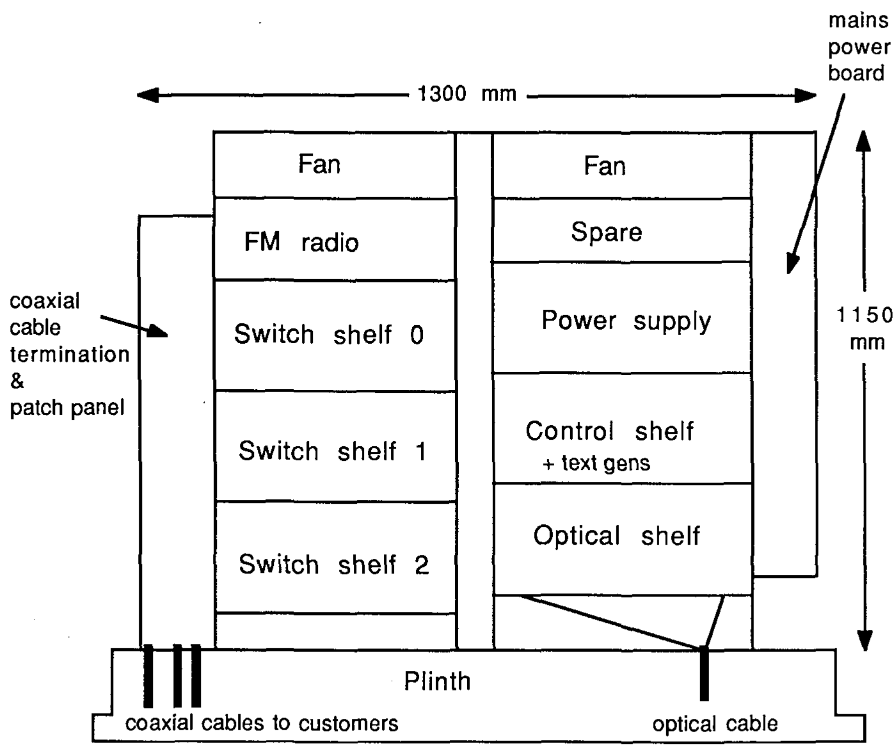

FIGURE 3 - Mark 2 WSP Layout

surface mount technology; one of the above mentioned switch unit daughter boards served 2 customers where the standard option was two separate feeds of 2 TV channels plus the FM radio band. There is flexibility over this arrangement however.

Firstly a version of the board could supply one customer with 4 channels, since there is plenty of spectrum available. A second option is again to use 4 channels on a feed but to share it between two customers, splitting and filtering just outside the home. This gives small savings on the daughter board (shared FM radio and surge protection circuitry) and significant cable and duct savings. This flexibility in customer options needs to be matched by capability in the local control and central administrative software to cater for the alternatives.

Optical Links

One of the clearest changes to consider was to move from multimode fibre working at 0.85μ to single mode fibre at 1.3μ for the primary links. It extends reach without repeatering to 15 km or more and allows greater splitting. The latter means that even with higher laser cost, greater sharing can improve the final link cost. The price differential has in fact now decreased; further 1.3μ devices offer improved performance (noise, linearity) and better reliability.

Frequency modulation is still the chosen transmission technique though this needs constant reviewing. On the one hand digital operation is being examined; PCM devices are getting cheaper and one-bit coding offers a very simple decoder [4]. On the other hand AM operation on fibre is receiving great attention[5]. If this proves viable and the signal can be passed through the switch in its AM form (so it is ready to go straight onto the coaxial cable), then it should lead to lower cost.

In considering the modulation technique to use, it is as well to bear in mind the different TV formats [6] that may have to be carried. NTSC, PAL, and SECAM are still dominant worldwide, but the move to higher definition has started. In Europe the newest format is MAC (Multiplexed Analogue

Components), where each TV picture line is sent as a sequence of :- a burst of data (digital sound), the compressed chrominance signal, and the compressed luminance signal. It offers higher quality by separating the components in this way (so avoiding mutual interaction), and has the facility also to allow a wider screen format. There are several versions of MAC including one termed HD-MAC which is compatible with true high definition TV (HDTV). A 30 MHz bandwidth FM slot or a 140 Mbit/s digital channel will both act as a "universal channel" carrying any of the standard formats or MAC. AM is less robust but requires lower bandwidths; a 16 MHz channel is probably needed for HD-MAC whereas the reduced form D2-MAC can fit in the same 8 MHz channel as the standard formats.

It may seem attractive to put as many channels in a multiplex as practical to share the optical device costs. Dedicated links for video library can sensibly take advantage of recent developments [7] which promise 50 channels or more within a wavelength, using existing GHz microwave technology with broadband lasers. However the higher bandwidths mean lower receiver sensitivity. For a broadcast link it may be just as well to retain optical budget for splitting, and so a multiplex of 16 FM channels per fibre seems appropriate at present, avoiding "putting too many eggs in one basket".

EVOLUTION OF THE SWITCHED-STAR CONCEPT

The Mark 2 SSN represents a good viable cable TV system for deployment now, but it is important to keep in mind potential improvements. Alternative transmission options have already been mentioned above, and there are similarly alternatives in other areas, most obviously for switching. The present card switches baseband signals, the most flexible format if the transmission means is going to be different on either side of the switch. There is however benefit in eliminating signal processing (modulation, demodulation) within the network and having transparent operation, with the one transmission format for primary link, switch, and secondary link (e.g. one of the "universal channels" mentioned earlier).

The FET switching devices have a wide bandwidth, which is however reduced by crosstalk on the card itself. By increasing the bandwidth of the input buffering circuitry the present design has been shown as also able to pass FM signals centred on 25 MHz with good quality. The FM format into the home may become a viable option, since this is used by the Direct Broadcast Satellites and so cheap set-top FM decoders are now on sale.

A different switching technique worth revisiting for fibre based systems is that based on the frequency agile tuner. Now that large multiplexes of channels can be carried on a fibre down to the cabinet, it may be sensible, at least in part, to use frequency translation to put a channel into the required slot on the customer link.

Text generators have been a key element in the SSN. They have made it "user friendly" and enabled extensive information services to be provided at marginal cost. Greater integration with 8 units per card has been achieved for the Mark 2 design, and no doubt this may increase. The question is whether they are now cheap enough to provide on a per customer basis either within the WSP or the set-top unit. There may still remain a role for some centralised text generators feeding into the switch to provide special displays.

Multi-service Networks

The SSN has some restricted ability to provide data, telephony, and even videophone. As standard this is limited to 12 customers per cabinet. However the modularity of the system is such that a switch shelf could be removed (reducing cable TV capacity by one third) and replaced by a "Special Services" shelf. There is spectrum for these to be multiplexed in with the TV channels or of course sent on their own.

The full multi-service provision is being addressed by the BIDS (Broadband Integrated Distributed-Star) development. This is a combination of the switched-star design with a third bay to house a remote telecomms multiplexer. The whole is now termed a Broadband Access Point (BAP), the layout of which is shown in Figure 4. As the network diagram of Figure 5 shows separate fibre links carry the telecomms multiplex and the video channels on the primary side, there being little value in integrating the two. There is though definite benefit in a single feed to the customer. For the present the cost-effective solution is certainly a coaxial link as used for the Mark 2 SSN. However there is a desire at

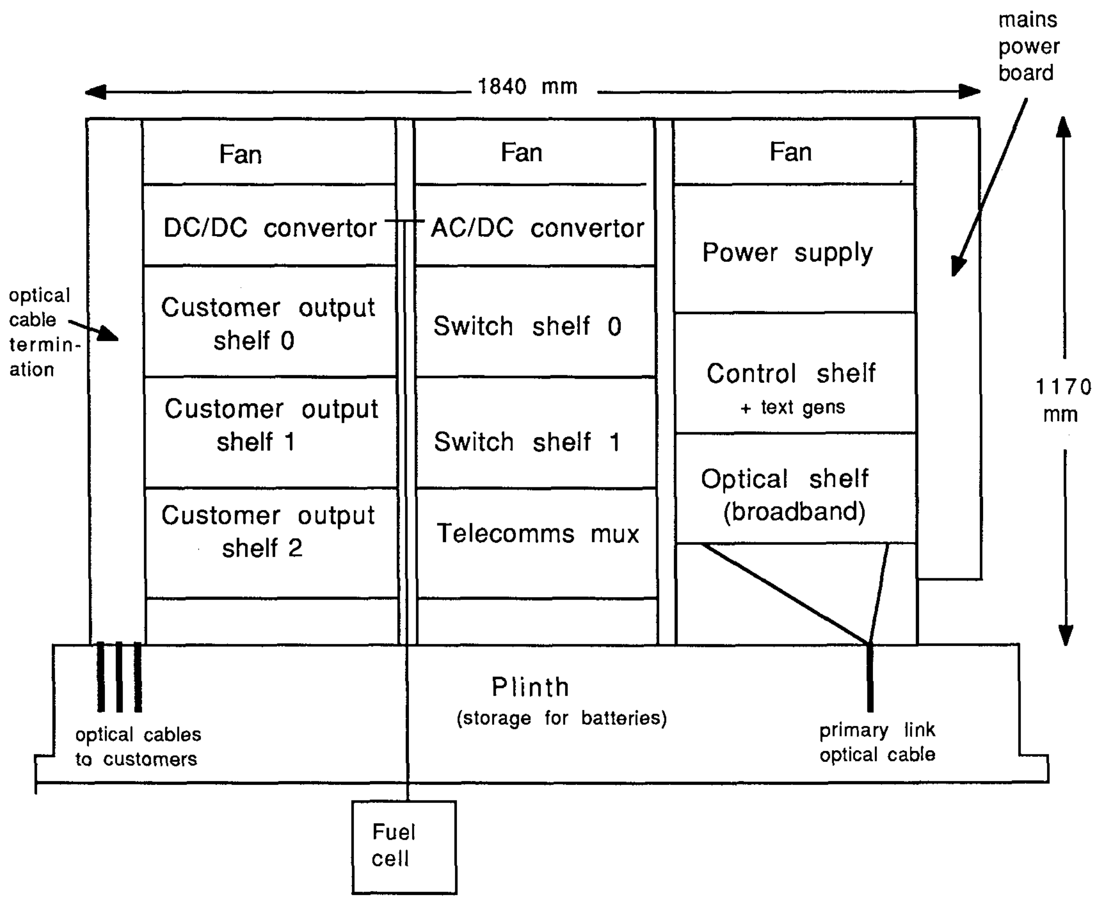

FIGURE 4 - Layout of the Broadband Access Point (BAP)

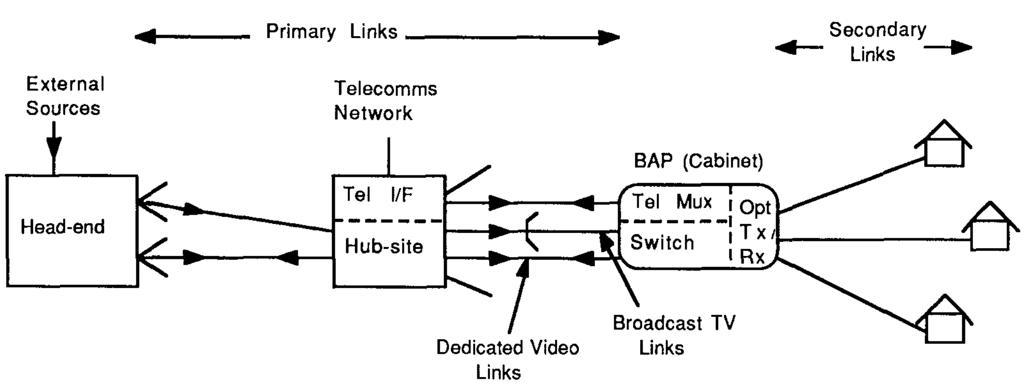

FIGURE 5 - The BIDS Network

some point to reach through to a fibre customer link to achieve a future-proofed network.

This all-fibre option has been chosen for field trial deployment in 1990 to around 100 customers in the town of Bishops Stortford, where the aim is to get practical experience with fibre methods. It will run alongside another local loop trial looking at the passive optical network (PON) technique [8], which is an alternative approach that BT is investigating in parallel.

The fibre based customer link is the subject of substantial development work in order to achieve a low cost solution. This is particularly pertinent to the telephony-only customer, who at least for the present would be in the majority in the UK. The link is short and the bandwidth or bit-rate low, so the task is in many ways very undemanding. However, since the circuitry involved in the copper alternative is minimal, the cost target is very tight. One option is to use a cheap LED/PIN combination packaged with a coupler to match onto a single fibre (bi-directional operation is desirable to keep fibre handling to a minimum). An alternative novel approach to keep costs low is to use a single device as both transmitter and receiver. A laser with the bias removed is a poor, but quite adequate receiver for this particular task. To time share the device in these two roles one can use a burst-mode technique, as developed for copper ISDN (Integrated Services Digital Network), i.e. a sequence of :- a burst of compressed transmitted data, a guard band to allow for transmission delays and devices to change roles, and a burst of compressed received data.

In a switched-star design the broadband services are also not that demanding on the customer link. Two video channels into the home are being provided for the trial, though there is potential for more. To give a simple upgrade to the above telephony link the broadband channels are being optically coupled in; they are transmitted at the same nominal wavelength but are effectively separated within the electrical spectrum (see Figure 6).

Once in the home the fibre terminates at a wall unit close to its entry point. This unit contains all the telephony circuitry which, in addition to local powering, requires battery back-up. The telephony signal is potentially available after the burst-mode circuitry in ISDN form, but for the present is converted to the standard analogue interface. The optical customer link means that some previously exchange-based functions (ringing, line current, etc.) must be provided by the unit in the home.

The incoming broadband signal to the home is decoded by its own optical receiver

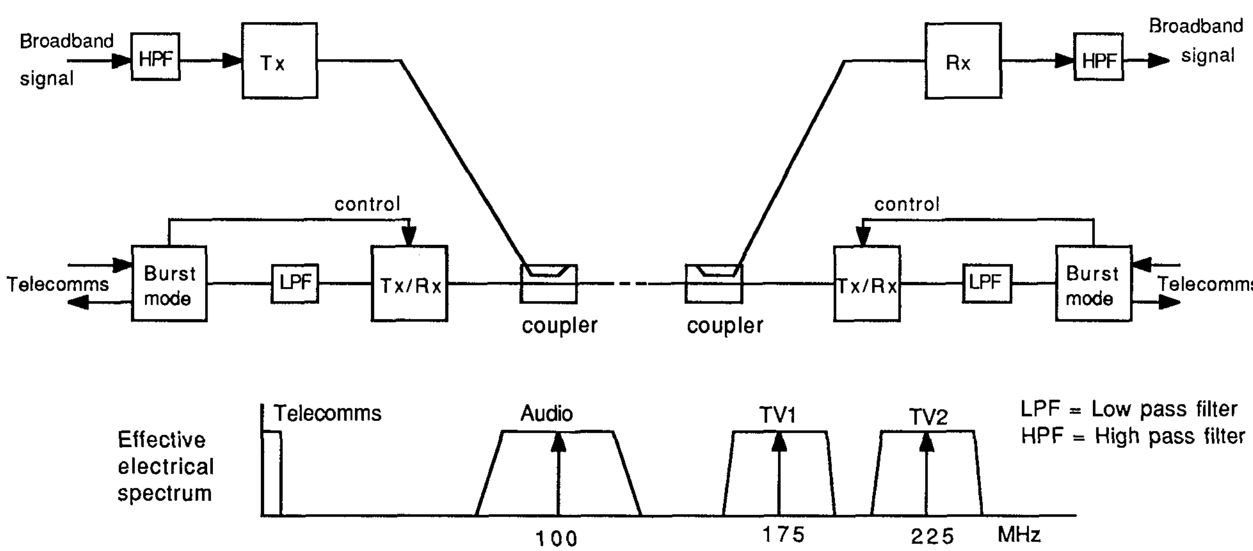

FIGURE 6 - The BIDS fibre customer link and spectrum

co-located with the telephony circuitry in the wall unit, and is then transmitted on by coaxial cable link to a set-top unit. This translates the link signals to the standard TV and audio input formats and receives the customer selections. The signalling associated with channel selection is sent back up the network via the telephony link, utilising the return path it already provides to the BAP.

Video transmission is in FM format all the way through, with the switch card modified to take FM signals. This gives the transparency described earlier, and hence some future-proofing by catering for MAC and HD-MAC. Full high definition TV will require additional measures. The customer link could be extended to add such a channel into its multiplex, but it may be an easier upgrade to add this in on as a separate optical wavelength. While demand is small and HDTV channels few, simple optical splitting in the BAP to broadcast it to a few customers is the simplest option; a switched system may be of value in years to come by which time technology advances will have eased handling the higher bandwidth/bit-rate of HDTV.

CONCLUSIONS

Switched-star has started to make an impact on a cable TV scene dominated worldwide by tree-and-branch. However the provision of additional services beyond entertainment TV by cable TV networks has been far smaller than anticipated in the early 1980s, and this in particular is where switched-star was expected to show benefit. There are signs that this area is taking off now, and it could be that the 1990s will usher in the era of the multi-service network. Certainly if cable TV penetration improves towards the 50% level, as is already the case in parts of the USA and some other countries, then the switched-star approach becomes increasingly attractive because of its simple customer equipment and its centralized resources becoming more effectively shared.

Competing directly with tree-and-branch on TV distribution has forced the switched-star concept to look for low-cost implementation rather than just a technical ideal. This then leaves it in good shape to move into a new era, whether it is as a predominantly cable TV network with strong capabilities for additional services, like the Mark 2 SSN described earlier, or whether it is a total multi-service scheme along the lines of the BIDS scheme.

ACKNOWLEDGEMENT

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FIBER BACKBONE - MULTI-CHANNEL AM VIDEO TRUNKING

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Abstract - This presentation updates a scenario for the evolutionary integration of optical fiber transmission technology into existing cable television networks first presented at the NCTA transactions in 1988. The resulting "fiber backbone" yields a hybrid fiber/coaxial network with significantly better reliability and transmission quality than present systems. System and electro-optical component advances in the last year are reviewed, and the merits of various modulation techniques are examined. The fiber backbone approach emphasizes continuing the broadband delivery of a large number of video signals to the consumer.

INTRODUCTION

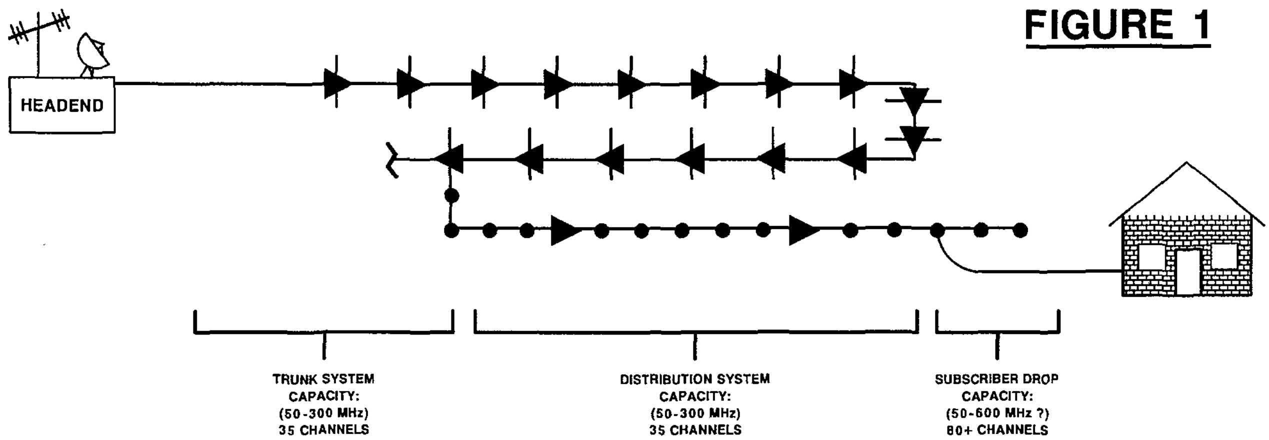
Optical Fiber Transmission Technology has achieved rapidly increasing acceptance by the cable television industry. While 87% of the homes in the United States are already passed by a broadband coaxial CATV network, coaxial technology, as it is presently being used, is beginning to approach its performance limits. Optical fiber offers a high bandwidth, low loss transmission medium which holds out the potential to allow significant performance improvement in today's cable television networks.

For some years, there has been increasing use of fiber for "supertrunking": high quality point-to-point video interconnections between major system hubs, connections earth station/headend connections, and links between headends to allow simultaneous insertion of local commercials. These fiber supertrunks have proven themselves to be highly reliable and cost effective, offering, in many situations, a viable alternative to microwave interconnection.[1,2]

It is, therefore, natural that the CATV industry has sought additional ways to use this new technology to improve its networks. The authors previously described an approach to such a use which they termed "fiber backbone".[3] In examining current CATV system architecture, it was noted that most of the performance limitations, including those reliability, transmission quality, and useable bandwidth, stemmed from the long amplifier cascades required by typical CATV tree-and-branch architecture when used in medium to large communities. This, in turn, is a product of the relatively high loss of coaxial cables (on the order of 1 dB per 100 feet: a loss of half the signal voltage every 600 feet). This loss, and the large number of amplifiers required in series to counteract it, requires that CATV system bandwidth be limited far below the potential of the coaxial transmission medium in order to achieve acceptable signal transmission performance. Current system architecture is illustrated in figure 1.

The optical fiber transmission medium exhibits extremely low signal loss (on the order of 1dB of signal power per mile). The fiber backbone approach to CATV system architecture is designed to replace long runs of coaxial plant, which often contain twenty or thirty amplifiers in series, with completely passive low loss fiber trunks as illustrated in figure 2. With a fiber system within one or two miles of all subscribers, CATV signals can be handed off to an existing RF broadband coaxial network for delivery. By limiting the total number of amplifiers between the CATV system headend and any subscriber to a small number, significant improvements in reliability and signal quality can be achieved, and there is an opportunity to upgrade the remaining coaxial portion of the network to achieve substantially greater bandwidth then possible in traditional CATV systems.[4,5]

FIGURE 1



While conceptually simple, optical terminal technology capable of delivering broadband multi-channel signals to the coaxial portion of a system is technically very challenging. Nevertheless, substantial progress has been made on this front by a number of system developers, and implementation of both demonstration and operational fiber backbone systems has begun by a number of cable operating companies during the last year. There is growing acceptance of the idea that a hybrid fiber/coaxial CATV system has the potential to provide significant improvements in performance and channel capacity at relatively modest cost as the cable industry faces the challenges of the next decade.

DESIGNING A HYBRID SYSTEM

There are many different network topologies which could be adopted using a hybrid fiber/coaxial transmission medium. Questions naturally arise as to what forms of modulation should be used in which portions of the system, and how close to the home the fiber portion of the hybrid plant should extend.

AM MODULATION

There are several types of modulation available for the transmission of video information. The most obvious is amplitude modulation with a vestigial sideband (AM-VSB). This is perhaps the most bandwidth-efficient practical modulation system available for video transmission, and is used for over-the-air television broadcasting as well as in current CATV systems. With it, NTSC video can be transmitted within a 6 MHz channel. In addition to bandwidth efficiency, AM/VSB enjoys tremendous ubiquity. It is estimated that there are over one hundred and sixty million television sets in use in the U.S. today. All of these sets are designed to accept AM/VSB modulated video as their input. It follows that, regardless of the transmission modulation system adopted, AM/VSB modulation must be the final product of a CATV system at the point of hand-off to the customer's television set. Today's cable television systems use AM/VSB modulation throughout, with a simple broadband coaxial transmission medium carrying signals all the way to the television set. Some televisions require a channel converter, a hetrodyne frequency conversion device, if they cannot tune all channels provided bv the cable system directly. Our researcn indicates, however, that 52% of cable homes currently own at least one cable-ready television set capable of tuning non-standard cable channels directly. While converters with built-in descramblers are also sometimes used for signal security, particularly for premium services, the cable television industry's approach of delivering signal all the way to the home in directly tunable multi-channel AM/VSB form is clearly highly attractive.

FM MODULATION

Frequency modulation of video signals in the RF domain allows high quality multi-channel video transmission. FM video requires substantially more bandwidth than AM, unusally from 10 to 40 MHz per channel, depending upon the performance improvement sought. FM video is widely used for satellite transmission, as well as for cable supertrunking. At some point in any distribution system using FM, however, there must be demodulation of the FM signal and AM/VSB

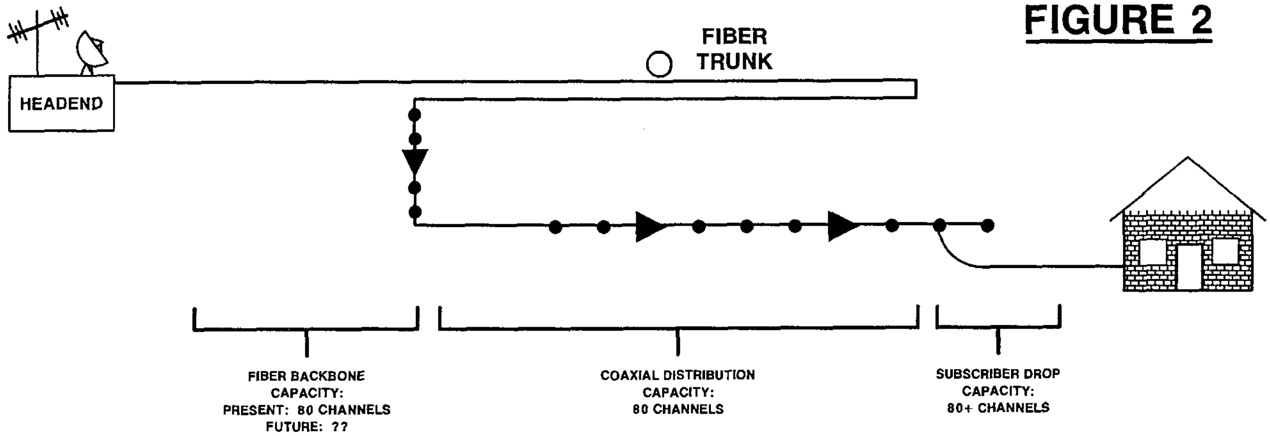


FIGURE 2

remodulation of the resulting baseband video in order that it may be received by today's television sets.

Carrying FM video all the way to the home would require an FM receiver to demodulate the selected channel, and an AM modulator to remodulate it for viewing or recording. One of these receivers would be required for each TV or VCR in the home and many of the built-in features of televisions and VCRs, such as remote control tuning, would be rendered useless. FM is used today for high quality supertrunks. Upon delivery of FM signals to a system hub, each is demodulated and remodulated using AM/VSB modulation onto the correct RF channel for coaxial transmission to the home. FM has transmission quality advantages over AM, but the costs of modulation conversion limit how deep into a CATV system it is economically practical to use it.

DIGITAL TRANSMISSION

Digital modulation is an obvious approach to video distribution in systems which use optical fiber transmission. Although digital modulation is highly bandwidth inefficient, this will matter less in the future as fiber systems realize greater bandwidth. Digital modulation has the advantages of offering high transmission quality and almost infinite repeatability as its binary codes can be recovered and regenerated as needed.

Digital modulation is likely to become wide-spread in CATV supertrunking as large urban systems interconnect hubs with redundant routing to improve reliability. Costs for the electronic components required for digital video transmission will continue to drop. Nevertheless, the cost of converting to AM/VSB modulation will limit the depth into the CATV network to which digital modulation will be economically practical.

TRADE-OFFS

It is apparent that each modulation scheme has its advantages and potential points of application in a CATV system. In a hybrid fiber/coaxial system, it can be assumed that there will be a significant cost for the conversion interface between the optical and RF portions of the system. It can also be assumed that there will be a significant cost for conversions in the type of modulation used. It is assumed that there is a potential role for different modulation schemes and for both optical and coaxial RF transmission. The cost/benefit trade-offs will determine how far into the network both non-AM/VSB modulation and optical transmission should extend, since there is a strong economic motivation to limit the number of conversion points for both the transmission medium and the modulation, and since AM/VSB signals within a broadband RF spectrum is assumed to be the required final product.

Figure 3 shows a plot of relative system improvements as the coaxial portion of a CATV system is shortened and the number of amplifiers in cascade is reduced. These relative benefits include an improvement in both system reliability and transmission quality arising from the use of fewer active components in series, as well as the ability to deliver more channels. There is a direct relationship between amplifier cascade reduction and relative performance improvement.

Also in figure 3 is an estimation of the cost per home involved in the reduction of amplifier cascade through the extension of passive fiber plant closer to the home. This curve rises exponentially, as tree-and-branch architecture dramatically increases the number of conversion points required as the system approaches the home. The point of cavity. This structure acts as a reflector, where light is partially reflected from each corrugation. If the wavelength of the light matches the structure wavelength, all of the reflections from the structure are summed coherently and the light continues to travel in the cavity and to be amplified. If the wavelength does not match the grating, the reflections cancel out. Single mode operation is achievable in DFB lasers.

The best noise and intermodulation performance observed to date in ATC's testing have been in DFB lasers. An important element in maintaining low-noise operation of DFB lasers appears to be limiting the amount of reflected light re-entering the laser from the transmission medium. External reflections which have less than -40dB of attenuation appeared to cause a rapid increase in the laser's Relative Intensity Noise (RIN). In order to achieve useful performance in a multi-channel CATV system, where carrier-to-noise ratios of 53 to 55 dB or better are desirable, RIN's approaching -160 dB/Hz appear to be required. It is, therefore, critical that external reflections be controlled. Success has been observed with both external Faraday rotation isolators, and with systems which utilize careful splicing of geometrically matched fibers and incline-ground connectors and inclined detector faces to minimize reflections.

DFB lasers are relatively expensive, as there are two diffusion steps in their fabrication, and yields are relatively low in each step. Nevertheless, DFB structures hold out great promise for AM fiber backbone systems for the CATV industry.

QUANTUM WELL LASERS

This is a new technology, and quantum well lasers are not commercially available now. In quantum well lasers, the area where electrons combine with holes is made very thin. There are several ways to build a laser having quantum wells, and usually a single mode of operation is achieved, with a line width narrower than that of current DFB lasers. Another benefit of quantum well structure is in having very low threshhold current, enabling operation of the laser with less external electronics.

Quantum well structures may be constructed in a Fabry Perot cavity laser, with the promise of better performance and lower prices than existing DFB lasers. Quantum well structures may also be combined with a DFB cavity to create still better performance, but at a premium price.

Quantum well semiconductor laser technology is expected to develop over the next several years, and holds out the promise of dramatic improvements in noise performance for optical links used in AM fiber backbone systems.

EXTERNAL CAVITY LASERS

In external cavity lasers, the semiconductor laser cavity is optically coupled to a second, external, cavity. There are many different structures possible, but the main idea is to fabricate a device which enables the creation of an extremely narrow line width, and a corresponding decrease in modal noise. External cavities may be combined with DFB and quantum well laser structures, as well as basic Fabry Perot structures.

External cavity lasers exist today only in the laboratory. The main development problems to be overcome relate to physical stability with respect to vibration and temperature change. The possibility of an external cavity integrated on the same substrate with the laser is being explored and holds great promise. The commercialization of these lasers, with the potential for very high performance at low cost, may be 5 to 10 years away.

EXTERNAL MODULATORS

Another promising line of development work involves the generation of low noise, high power light using a constant-output Continuous Wave (CW) laser, feeding an external modulator. This is illustrated in figure 5. This allows the generation of substantially more optical power than is possible in practical, directly modulated lasers. The external modulators available today are of the Mach-Zehnder inferometer type. These devices split the optical input and allow it to follow two paths through the device. One leg is entirely passive, but the other allows variable delay through the application of an electrical field. If this field is varied, the delay will vary, and the output of the device, where the legs are

FIGURE 3

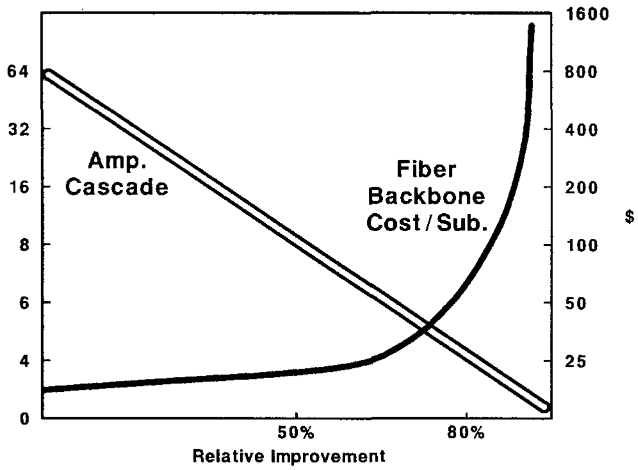
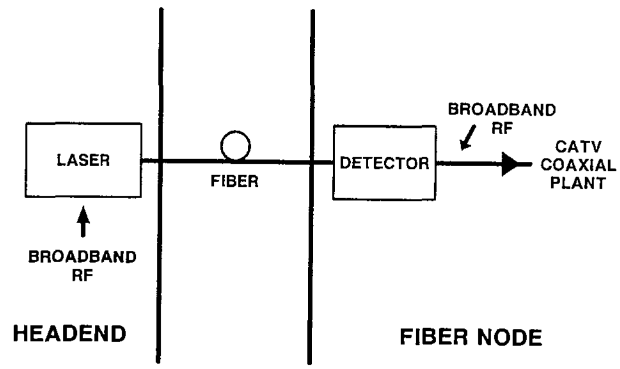


FIGURE 4



diminishing returns is difficult to pin-point precisely, but it appears that the optimum balance between fiber and coaxial plant in a hybrid system comes with a maximum amplifier cascade between two and five trunks amplifiers.

OPTICAL COMPONENTS FOR USE IN AM VIDEO FIBER SYSTEMS

It is apparent that in a hybrid fiber/coaxial CATV system, it would be highly desirable to maintain AM/VSB modulation throughout. If this could be accomplished, the only signal conversion required outside of the headend would be that from optical to RF at the end of each optical trunk. This approach greatly simplifies the electronics needed at each conversion point, since it should be possible to directly detect the intensity modulation of the light on the fiber, with the resulting detected output being the broadband RF spectrum, a complex waveform complete with all the original channel information, scrambling, data carriers, etc. Such a conversion point could be contained in a small weather-proof housing, directly powered off the coaxial portion of the CATV system. Because AM/VSB modulation is relatively fragile, however, this approach is technically quite challenging.

Figure 4 shows a simple block diagram of an ideal system. At the headend, the broadband AM/VSB signal, containing all of the cable channels, is used to directly modulate a laser. This information is transmitted optically through the fiber to a conversion point deep in the cable system, where it is reconverted using a simple detector.

In the last year, substantial progress has been made on the necessary components to effect such a system. It is necessary that the laser used has a high degree of linearity, and adds very little noise to the signal. While there is room for improvement in detectors to insure the lowest possible noise contribution and highest sensitivity, it is the semiconductor lasers used in these systems which dominate system performance. Because this technology is critical to the implementation of practical AM fiber backbone systems, it is worth examining these components more closely.

FABRY PEROT LASERS

In semiconductor lasers, the distance between the laser mirrors defines the possible wavelengths of light amplified in the cavity. Only a integer number of half-waves can oscillate. We call this list of possible wavelengths the "Fabry Perot modes". There are an infinite number of Fabry Perot modes, but within the cavity, only a limited bandwidth is amplified. Normally, 10 to 15 Fabry Perot modes are within the amplification band, and they create the spectrum we see in Fabry Perot lasers.

Laser noise is directly related to the bandwidth of the emitted light. It is of interest, then, to create a laser with just one Fabry Perot mode, and to make this mode as narrow as possible. There are several ways to reduce the number of modes developed in the laser.

DISTRIBUTED FEEDBACK LASERS

Distributed feedback (DFB) is the most common means of achieving single mode operation. The laser is fabricated with a corrugated grating structure along the

FIGURE 5

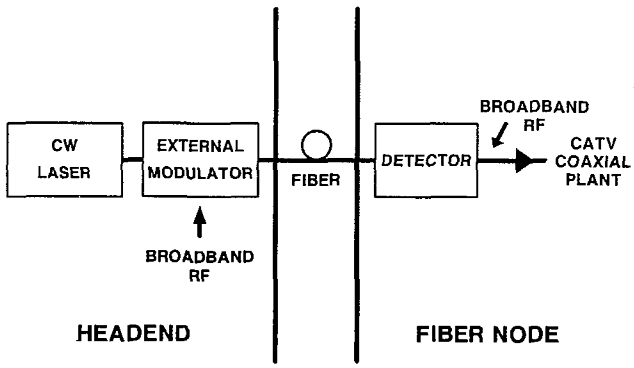


FIGURE 6

DESIRED PERFORMANCE

CHANNELS	60 - 80		
C/N	55	dB	
CTB	65	dB	
CSO	65	dB	
CROSS MOD.	65	dB	
POWER BUDGET	10	dB	(20 Km/12 Mi)

recombined, will vary through signal addition and subtraction caused by the relative phasing of the light through the two paths.

The primary drawback to Mach-Zehnder devices are that the modulation process is inherently non-linear. The change in intensity at the output of the device is related to the change in input voltage by a sine function. Non-linearity in a broadband, multi-channel device creates severe problems in the form of intermodulation products, but the fact that the Mach-Zehnder modulator has a precisely predictable characteristic opens the possibility that either preemphasis, feedback, or feedforward techniques can be used to produce overall system linearity.

It is possible that practical externally modulated optical transmitters will be realized with high enough output levels that they can be used in relatively long-haul applications, or that their outputs can be split to feed a number of conversion points in an AM backbone system, allowing the relatively high cost of the transmitter to be shared over several links.

OVERVIEW

While significant progress has been made in optical links for use in an AM backbone system, it is the authors' opinion that price/performance that the point which has been achieved is still somewhat short of that required for widespread proliferation of AM fiber backbone technology in CATV system rebuilds and upgrades. We continue to believe that the achievement of the goals illustrated in figure 6 will spark massive adoption of this technology by the CATV industry when combined with a cost between $5,000 and $10,000 per link. Nevertheless, developments by optical component manufacturers are the key element in achieving these goals, and it is clear that component technology has many promising avenues to explore.

CURRENT SYSTEMS

Figure 7 illustrates the results of ATC's most current system tests. These results are conservative but repeatable, and are made with CW carriers from a Matrix multi-channel signal generator. Multi-channel systems with asynchronous video modulating signals will yield somewhat better results. It can be seen that the goals set by ATC are being approached relatively closely by some of these systems. While the pricing of these systems ranges from $15,000 dollars to $30,000 per link, we expect that during 1989 there will be delivery to the CATV industry of a significant number of links, that the goals we have outlined will be met, and that prices will begin to drop toward the target we have set.

It should be noted that decreasing the channel loading of AM optical links has a relatively dramatic impact on performance. First, noise performance improves by 3 dB each time the number of channels is cut in half. Secondly, the available power budget also increases dramatically as the number of channels is decreased and the modulation index per channel increases correspondingly. This has lead to the commercial development of AM transmission systems of the type shown in figure 8. These multi-fiber/multi-laser systems are capable of carrier-to-noise ratio performance in the high 50's, with power budgets of 10 dB and more.

FIGURE 7

AM FIBER OPTIC LINK RESULTS

SYSTEM	CHANNELS	C/N	CTB	CSO	LOSS BUDGET
A	40	48	60	58	4
B	40	52	65	66	5.2
C	40	53	69	64	6
D	40	54	70	69	5.7

While their cost is high and they make relatively inefficient use of optical fiber, they are clearly useable in some applications. ATC has constructed and tested multi-fiber AM super trunks in several of its systems using this technology, and Jones Intercable has announced the construction of a fiber backbone hybrid system in an upgrade currently underway in Broward County, Florida.[6]

SUMMARY

1988 was the year when the true potential for integrating optical fiber into its systems began to dawn on the CATV industry. 1989 will be the year when the commercialization of practical systems begins to hit its stride, and field deployment will begin. By 1990, it is expected that the system architecture and economics originally predicted by ATC will be realized, that subsequent years will bring further improvements in both performance and price, and that applications will increase dramatically. We believe that the typical CATV system of the mid-'90s will be a hybrid fiber/coaxial network achieving levels of reliability, signal quality, and channel capacity once though unattainable.

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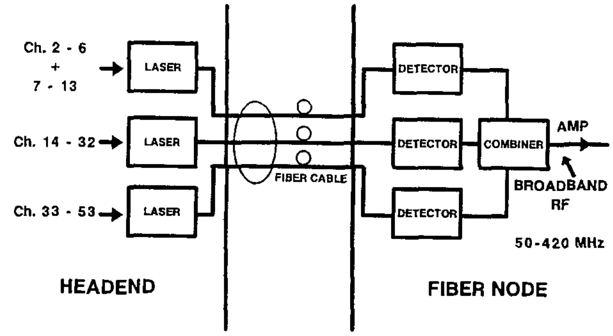
FIGURE 8



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Jim Chiddix - Senior Vice President, Engineering and Technology, for American Television and Communications Corporation, the country's second largest cable television operator, headquartered in Stamford, Connecticut. Mr. Chiddix is responsible for corporate engineering activities as well as research and development. ATC serves 3.9 million subscribers in 32 states and is 82% owned by Time, Inc. Upon completion of its pending merger with Warner Communications, which also owns cable TV operations, the combined companies will serve 5.6 million homes.

ATC leads the cable industry in exploring the use of optical fiber technology in cable television systems. Their "fiber backbone" concept for optical trunking has gained wide acceptance as an evolutionary approach, offering the prospect of improved performance and increased channel capacity from existing cable systems. In recognition of his pioneering role in exploring this use of fiber, Mr. Chiddix was named Man Of The Year by Communication Engineering and Design Magazine in January, 1989.

Mr. Chiddix, 43, has been in the cable television business for 17 years. He spent seven years as General Manager at Cablevision, Inc. in Waianae, Hawaii, and eight years as Engineering Vice President at Oceanic Cablevision in Honolulu, an ATC division. In September, 1986, he joined ATC's corporate office.

Mr. Chiddix is a Senior Member and former Director of the Society of Cable Television Engineers (SCTE). 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.

Dave Pangrac is the director of engineering and technology for American Television and Communications Corporation (ATC), the country's second-largest cable television operator.

Pangrac has been in the cable television business for 22 years. He joined ATC in 1982 as Vice President and chief engineer for American Television of Kansas City and in 1987 joined the ATC corporate staff as director of engineering and technology.

Pangrac is a member of the Society of Cable Television Engineering and past president of the Hart of America Chapter.

Pangrac is currently involved in ATC's effort to develop the use of fiber optic technology in cable television plants.

FIBER OPTIC
RF LINK ANALYSIS

J. Koscinski

Laser Diode, Inc.

ABSTRACT

The key system design parameters for analog fiber optic systems are their noise and distortion specifications. Both the optical transmitter and optical receiver have the potential to significantly limit the link's noise and distortion performance. A thorough system analysis first requires establishing the individual noise and distortion parameters for the laser transmitter and optical receiver, and secondly, examining the complete link's (noise and distortion) performance over the required range of optical loss budgets.

Noise performance of an RF optical link may be limited by either the transmitter (laser noise) or the optical receiver noise depending on the optical loss budget. Additionally, there are several independent noise sources contributing to the optical receiver noise. These are referred to as the "quantum" and "circuit" noise.

Distortion is introduced primarily in the optical source (laser). However, when high optical powers are delivered into the optical receiver, the receiver may also generate significant distortion.

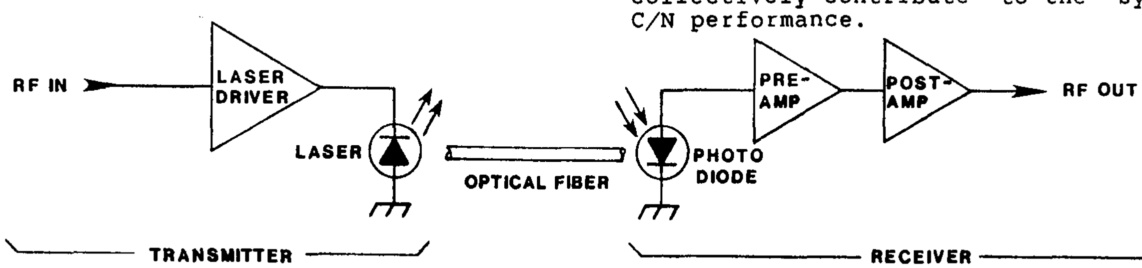
The focus of this paper will be to 1) quantify these noise and distortion parameters for the laser transmitter and optical receiver; and then, 2) apply these system performance parameters by analyzing an optical link over a range of optical loss budgets.

LINK PERFORMANCE PARAMETERS

The key performance parameters used to characterize any multi-channel RF optical link are its carrier/noise ratio (C/N) and intermodulation distortion (C/IMD). Although there are currently many types of optical sources and fiber available, this analysis will deal exclusively with 1300 nm single mode lasers and optical fiber. Single-mode technology provides the essential performance requirements with manageable costs; thus, it is becoming the defacto standard for high performance RF optical link applications.

A block diagram of a typical RF optical link is illustrated in Fig. 1. The main system components are: 1) a laser transmitter, 2) the fiber-optic cable and, 3) the optical receiver. As in conventional RF link analysis, one analyzes the link by first establishing the contribution for each of the individual (transmitter and receiver) components. Then, a complete system analysis can be performed using optical loss as a system variable.

The optical link's C/N performance is highly dependent upon the actual optical loss budget. For low loss budgets (distances < 10 km), the inherent noise of the laser transmitter will generally establish the upper limit on C/N performance. However, as the optical loss increases (high loss budgets), the optical receiver will takeover and limit the optical link's C/N performance. At intermediate loss budgets, both the laser transmitter and optical receiver will collectively contribute to the system's C/N performance.



RF OPTICAL LINK

FIG. 1

Similarly, the distortion performance of the RF optical link may vary as a function of the optical-loss budget. The laser's inherent linearity (or nonlinearity), in conjunction with system reflections, normally determines the intermodulation distortion for the optical link. However, at very low optical-loss budgets, the optical receiver may overload; and thus, the receiver may actually limit the link's distortion performance.

Since both of these key system performance parameters are highly dependent on optical loss budgets, it is important to perform a system analysis over an application's "relevant" loss-budget range. Spreadsheets have proven to be a powerful tool in supporting this type of analysis. DEC's 20/20 spreadsheet software has been utilized to perform the system analyses that are illustrated in this paper.

The required performance for any RF optical link is dependent on the form of modulation (AM or FM) used to modulate the individual carriers. For example, AM modulation is less expensive to utilize; however, it is the most demanding on the link's noise and distortion performance. FM modulation is more costly; however, bandwidth is traded off for a substantial "signal-to-noise" enhancement (typically > 25dB). Additionally, an FM-modulated carrier has a much greater tolerance for "in channel" distortions (typically also > 25dB).

LASER TRANSMITTER NOISE

LASER "RELATIVE INTENSITY NOISE" (RIN)

RIN defines the inherent noise power of the semiconductor laser diode. Minute fluctuations in optical emission are exhibited when the laser is biased above threshold. This intensity noise is neither thermal nor strictly shotnoise in nature. The noise is related to the response of the laser to modulation caused by the intrinsic shotnoise which is present in the laser. This phenomena results from the granular nature of light and electricity.

Analog lasers are currently available (specified) with RIN performance of (-135 dB/Hz) to (-155 dB/Hz). Carrier/noise is calculated from the RIN specification by subtracting 3 dB (for a 100% modulation index and a single carrier). This factor of 3 dB results from RIN being defined as the ratio of average (dc) light power [rather than (rms) signal power] to the (rms) noise power. For other modulation indexes and multiple channels, the carrier/noise (per channel) may be determined using the following equation.

C/N(1) =-RIN-10 LOG (2n*Bw/m^2) (1)

where;

C/N(1) =transmitter C/N (per ch.) (dB)
m = composite modulation index (%)
n = number of carriers
Bw = noise (channel) bandwidth (Hz)
RIN = laser noise (dB/Hz)

It should be emphasized that it is essential to control the fiber and/or connector related back reflections (interactions) to the laser (40 dB typical) in order to preserve consistent, low noise results from the laser transmitter. Laser diode noise significantly increases when there are reflected waves created from the discontinuities of the refractive index in the optical transmission path. Optical isolators are now being introduced, either integral to the laser package or externally, to provide this important isolation.

LASER MODULATION

Referring to equation (1), one can see that the laser's noise performance is highly dependent on its modulation index. This modulation index (m) is defined as the ratio of the peak laser current excursion about the laser's normalized (dc) bias current (Fig. 2). The equation which defines the laser's modulation index is:

m = (delta-I)/Ib' (2)

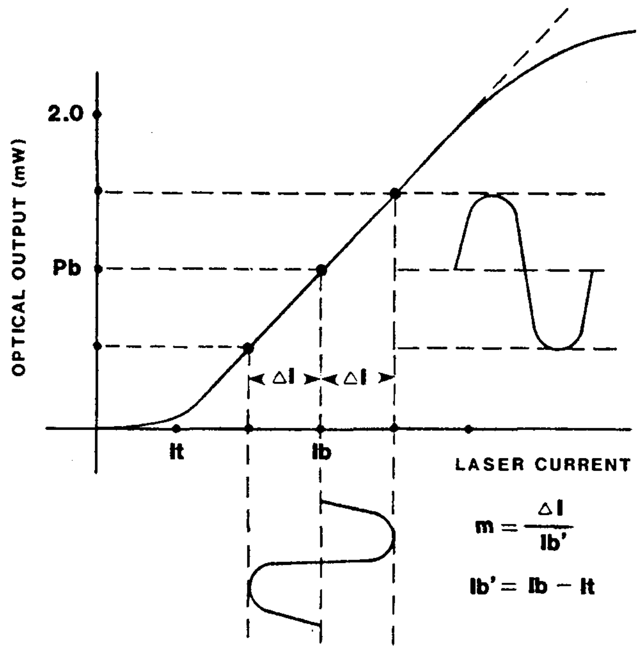
where;

Ib = (bias current)
Ib' = Ib-It (normalized bias)
It = (threshold current)
delta-I = (RF peak current)

Using a higher modulation index (m) provides a greater C/N ratio since the signal detected in the optical receiver is directly proportional to (m) squared. However, a higher modulation index requires operating the laser over a wider range of its light-current (LI) characteristic curve which also results in higher distortion levels. Typically, a laser's modulation index is set at 50 to 70% as the best tradeoff for transmitter noise versus distortion performance.

LASER TRANSMITTER DISTORTION

The laser transmitter is generally the



LASER LI CURVE

FIG. 2

distortion limiting component in high-performance RF optical links. Lasers generate harmonic and intermodulation distortion as a result of their nonlinear (LI) transfer characteristics and sensitivity to system reflections.

As can be seen from Fig. 2, the laser's (LI) curve exhibits nonlinear characteristics (especially at the extremes). Laser threshold determines the lower limit, while the instantaneous rise in laser temperature (as current is increased) causes the upper portion of the (LI) curve to saturate. Thus, for linear operation, it is important to bias the laser in the center of its linear region and to limit the amplitude of the modulation consistent with achieving the required distortion performance. High performance lasers are currently available which provide (specify) third-order intermodulation distortion levels of (> 60 dB) when operating at a 1 mW bias (50% modulation index).

An approach which is currently used to characterize the transmitter's distortion performance is to define its input "Intercept Point" (IP). The expression which defines the three-tone, 3rd-order (3-o) input (IP) is:

$$IP(3\text{-}o) = Pt + (IMD/2);(dBm) \qquad (3)$$

where;
Pt = composite RF input (dBm)
IMD = three-tone, (3-o) laser intermodulation ratio (dBl)

Second-order (2-o) distortion performance of a laser transmitter is generally (10-15 dB) poorer than its third-order performance. The equation which defines the transmitter's two-tone, (2-o) input (IP) is:

$$IP(2\text{-}o) = Pt + (IMD); (dBm) \qquad (4)$$

where;
Pt = composite RF input (dBm)
IMD = two-tone, (2-o) laser intermodulation ratio (dBl)

When possible, the RF optical links should attempt to limit the transmission bandwidth to a single octave. This avoids having the second-order distortion products fall within the desired transmission bandwidth.

When transmitting more than (3) carriers, the carrier-to-intermodulation ratio (C/IMD) will not necessarily degrade as rapidly as one might first expect. The reason for this is due to the laser requiring a constant (average) modulation index (independent of the number of channels transmitted). Otherwise, the transmitter's RF modulation peaks would overload the laser and perhaps might subject the laser to destructive current levels.

Thus, as more channels are loaded onto the transmitter, the individual carrier levels are lowered in amplitude (as required to maintain constant (composite) input power). For example, as the number of channels are increased to 10 channels, the amplitude for each of the individual carriers is lowered by 10 dB (constant RF power loading).

In the following analysis, it will be assumed that the C/IMD ratio is independent of the number of channels transmitted, (contingent on the composite modulation index also being maintained at a constant value).

OPTICAL RECEIVER NOISE

The optical receiver configuration reviewed in this analysis is an avalanche (APD) photodiode driving a low-noise, 50-ohm preamplifier. Although there are alternative front end designs available, this is a competitive (and representative) high performance front end receiver configuration.

There are two major sources of noise present in optical receivers. They are referred to as "quantum" noise and the receiver "circuit" noise.

QUANTUM NOISE

Quantum noise results from the statistical nature of the production and collection of photo-electrons when an optical signal (photon) is incident on a photodetector. Since fluctuations in the number of photocarriers created from the photoelectric effect are a fundamental property of the photodetection process, they set the ultimate "quantum limit" on receiver sensitivity when all other conditions are optimized. The quantum noise current has a mean square value, in a bandwidth (B), which is proportional to the average value of photocurrent.

Although quantum noise theoretically limits the link's ultimate C/N performance; in practice, laser noise usually limits the link's C/N performance first.

The C/N expression for "quantum" limited performance is:

$$C/N(q) = \frac{[(m*M*Re*Pb)\hat{}2/(2*n)]}{[2q(Re*Pb + Id)(M\hat{}(2+x))Bw]} \qquad (5)$$

where;
- m = modulation index (%)
- M = gain value (APD)
- Re = responsivity (A/W)
- Pb = average optical power (W)
- n = number of channels
- q = electron charge
- Id = dark current (A)
- M^x = excess noise factor (APD)
- x = excess noise exponent
- Bw = noise bandwidth (Hz)

CIRCUIT NOISE

There are two main constituents of the optical receiver's "circuit" noise. The first is due to the total equivalent resistance (Req) which is reflected back to the input node of the optical preamplifier (Fig. 3). This resistance constitutes a thermal noise source at the receiver's input. The second component is related to the "noise factor" (Ft) of the preamplifier.

The C/N expression for an optical link limited by the receiver's "circuit" noise is:

$$C/N(c) = \frac{[(m*M*Re*Pb)\hat{}2/(2*n)]}{(4*K*T*Bw/Req)Ft} \qquad (6)$$

where;
- K = Boltzmann's constant
- T = temperature (degrees K)
- Ft = preamplifier noise factor
- Req = equivalent resistance reflected at receiver's input node (ohm)

Combining the "quantum" and "circuit" noise components, the equation for the complete C/N performance of the optical receiver is:

$$C/N(r) = \frac{(m*M*Re*Pb)\hat{}2/(2*n)}{[N(q) + N(c)]} \qquad (7)$$

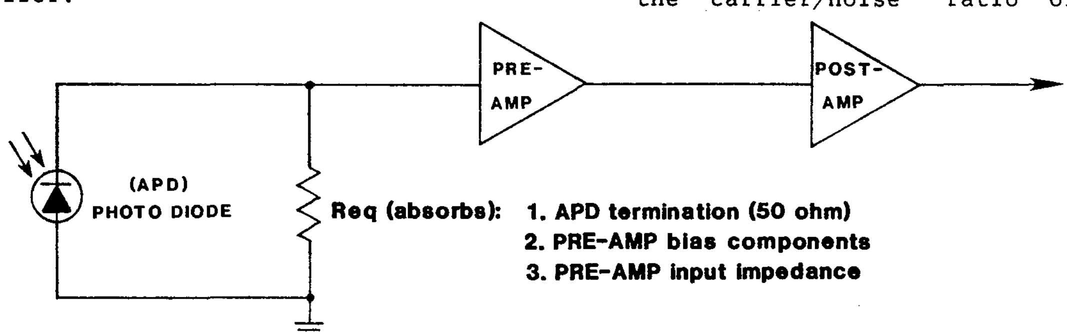
where;

$$N(q) = 2q(Re*Pb + Id)(M\hat{}(2+x))Bw$$

$$N(c) = (4K*T*Bw/Req)Ft$$

Referring to this equation, one may draw the following conclusions:

- At low signal levels and with low-gain (M) values, the circuit noise term dominates.
- At a fixed low signal level, as the gain (M) is increased from a low value, the carrier/noise ratio increases with gain until the quantum noise term becomes comparable to the circuit noise term.
- As the gain is increased further beyond this point, the carrier/noise ratio decreases (due to the APD's excess noise (M^x).
- Thus, for a given set of operating conditions, there exists an optimum gain value (M) of the APD for which the carrier/noise ratio of the



OPTICAL RECEIVER

FIG. 3

receiver is a maximum.

The expression which determines this optimum (M) value for optimum C/N of the link is:

$$Mopt = [A/(q*Req(B)x]^{(1/(2+x))} \quad (8)$$

where;
A = 4K*T*Ft
B = (Re*Pb)+Id

The C/N performance for the complete RF optical link may then be obtained by summing the C/N ratios of the transmitter and optical receiver on a power basis.

$$C/N(s)=-10\log(10^{(-Tx)}+10^{(-Rx)}) \quad (9)$$

where;
Tx=C/N(l)/10
Rx=C/N(r)/10
C/N(l)=C/N of transmitter (dB)
C/N(r)=C/N of receiver(dB)
C/N(s)=C/N of the link (dB)

OPTICAL RECEIVER DISTORTION

The optical receiver's distortion performance is generally negligible (relative to the laser transmitter). However, at high optical input powers, the optical receiver's output amplifier may overload and begin to introduce substantial distortion. Even at high input powers, photodetectors normally contribute a negligible portion of the total optical receiver's output distortion. Thus, in this analysis, their contribution is ignored.

The preamplifier and post-amplifier distortions are determined from their respective output "Intercept Point" specifications. Cumulative receiver distortion levels are then determined by combining "voltage summing" these resultant independent C/IMD ratios of the preamplifier and post-amplifier. Similarly, the composite distortion of the complete RF optical link may be determined by combining "voltage summing" the C/IMD ratios of the laser transmitter and optical receiver. The equations which define these C/IMD ratios for both second and third-order distortions of the optical receiver and complete link are summarized below.

$$C/IMD(3\text{-}o) = 2(IP(3\text{-}o) - RF\ out) \quad (10)$$

$$C/IMD(2\text{-}o) = (IP(2\text{-}o) - RF\ out) \quad (11)$$

$$C/IMD(Rx) = -20*\log(A + B) \quad (12)$$

$$A = 10^{-(G1/20)} \quad (13)$$
$$B = 10^{-(G2/20)} \quad (14)$$

$$C/IMD(Link) = -20*\log(C + D) \quad (15)$$

$$C = 10^{-(Tx/20)} \quad (16)$$
$$D = 10^{-(Rx/20)} \quad (17)$$

where;
IP(3-o)=(3-o) output IP (dBm)
IP(2-o)=(2-o) output IP (dBm)
G1 = IMD ratio of preamp (dB)
G2 = IMD ratio of post-amp (dB)
Tx = IMD ratio of Tx (dB)
Rx = IMD ratio of Rx (dB)

Equations (10 & 11) are used to determine the C/IMD ratios out of the receiver pre-amp and post-amplifier, given their respective output (IP) specifications.

Equations (12 & 15) are "voltage summing" the C/IMD ratios for two independent C/IMD ratios in order to obtain a combined C/IMD ratio.

OPTICAL LINK ANALYSIS

Having defined the noise and distortion parameters for the optical transmitter and receiver, one may now analyze the link's total performance over a range of optical loss budgets. The analysis which follows is modeled onto a DEC 20/20 spreadsheet which provides one with a very powerful tool for evaluating "what if" results when alternative system variables are selected.

The optical link's performance model, illustrated in Tables (1 and 2), actually consists of two sub-models. Table 1 illustrates the C/N performance model exercised over a range of optical loss budgets; whereas, Table 2 illustrates the link's distortion performance model for the same range of optical loss budgets.

Each of these sub-models consists of three subsections. In the upper portion of each model is a list of the system parameters which must be quantified. These parameters are either the fixed or variable parameters used in the model's analysis equations. The model's equations are written to the right of the parameter list to provide a convenient reference for the user.

The lower portion of the model is where the system analysis is actually performed using optical loss budget as an additional variable parameter. Basically, the system analysis is performed by evaluating the optical link's noise and distortion performance at various optical losses using the analysis equations which have been referred to in this paper.

An illustrative RF optical link performance analysis is shown in Tables

(1 & 2). This example analyzes the performance of a 20 channel RF optical link over an optical loss budget range of 12 dB. In this example, the laser RIN noise is specified at -145 dB/Hz with an optical output power of 1mW (50% modulation). The transmitter and receiver distortion components are determined from their respective "Intercept Point" specifications.
Plots of the system's C/N and IMD performance are conveniently displayed with this spreadsheet software. This is especially useful when evaluating which of the system's components are the most dominant at specific loss budgets. For example, Fig. 4 is a plot of the system's C/N performance used in this example. Individual C/N components of the transmitter and receiver are also plotted enabling one to examine which component is most dominant in determining the link's C/N performance. As seen from the curves (Fig. 4), the C/N performance is mainly laser limited at very low optical losses to a limit of 55 dB. At 6 dB of optical loss, the optical receiver is the dominant C/N contributor limiting the link's performance to 51 dB.

Distortion performance for this example is displayed in Fig. 5. These curves reveal the contribution of the laser transmitter versus the optical receiver in determining the link's total distortion performance. In this example, the laser limits the third-order distortion performance of the link to -65 dBc. However, the receiver will degrade this performance to -61 dBc at the maximum available optical input power.

RF OPTICAL LINK PERFORMANCE MODEL

(LASER/APD-PHOTODIODE)

Parameters			Model Equations
n=	20	ch	C/N(1-laser)=-RIN-10Log(2*n*Bw/m^2)
m=	0.5		
M=	var		C/N(q)=10Log(m*M*RePb)^2/(4nq(RePb+Id)(M^2+x)Bw)
Re=	0.7	A/W	
Pb=	var		C/N(circuit)=10Log(m*M*RePb)^2/(8nK*T*Bw*Ft/Req)
q=	1.6e-19		
Id=	1.0e-06	A	C/N(rec'r)=10Log[((m*M*RePb)^2/(2*n))/
M^x	var		(2q(RePb+Id)BwM^(2+x))+(4KTBwFt/Req)]
K=	1.4e-23		C/N(s)=-10Log(10^(-C/N(1)/10)+10^(-C/N(r)/10))
			= System Carrier/Noise Performance
T=	300	K	
Ft=	4.0		M(opt)=((4K*T*Ft)/(q*Req(Re*Pb+Id)x))^(1/(2+x))
Req=	50	ohm	
Bw=	4.0e+06	Hz	
x=	0.95		
RIN=	-145	dB/Hz	

(Carrier / Noise)

(Watts)	Ptx out	1.0e-03	6.3e-04	4.0e-04	2.5e-04	1.6e-04	1.0e-04	6.3e-05
(dB/div)	2							
(dB)	Opt Loss	0	-2	-4	-6	-8	-10	-12
	C/N(c)=	65	62	60	57	54	52	49
	C/N(q)=	62	59	57	54	51	49	46
	C/N(r)=	60	57	55	52	50	47	44
	C/N(1)=	57	57	57	57	57	57	57
	C/N(s)=	55	54	53	51	49	46	44
	M(opt)=	2	3	3	4	4	5	6
	M^x =	2	3	3	4	4	5	5

TABLE 1

(Distortion Performance)

(Laser Transmitter)

Parameters		Model Equations
Pn=	-5 dBm	Pn=per/channel input power
Pt=	8.5 dBm	Pt=total input power
i(l)p/p=	33 mA	i(l)=(((10^(Pt/10))*10^-3/50)^0.5)(2.8)
n%=	3 W/A	n=laser quantum efficiency
Pb=	1 mW	Pb=optical bias power
deltaPo=	0.50 mW	Po=(1/2)(i(l))*10^3*n%/100; (delta Popt)
m=	0.50	m=Po/Pb (modulation index)
IP(3o)=	41 dBm	IP=input intercept point(3rd order)
IP(2o)=	64 dBm	IP=input intercept point(2nd order)
(dBl)=	65(3o)	(dBl laser-3o IMD)=2*(IP(3o)-Pt)
(dBl)=	55(2o)	(dBl laser-2o IMD)=(IP(2o)-Pt)

(Optical Receiver)

Parameters		Model Equations
m=	0.50	i(pd-RMS)=((m*M*Re*Pb)/(2^(1/2)))
M=	var	AMP-3o=2*(3oIP-(Amp Out)):3oG1;3oG2
Re=	0.7 A/W	AMP-2o=(2oIP-(Amp Out)):2oG1;2oG2
Pb=	var	Rec'r-3o=-20Log(10^-(3oG1)/20)+10^-(3oG2/20))
Ft=	4.0	Rec'r-2o=-20Log(10^-(2oG1/20)+10^-(2oG2/20))
Req=	50 ohm	Link-3o=-20Log(10^-(3oTx/20)+10^-(3oRx/20))
BW=	4.0e+06 Hz	Link-20=-20Log(10^-(2oTx/20)+10^-(2oRx/20))
G1-db=	8(pre-amp)	
IP-dBm=	31(output 3o-IP(G1))	
IP-dBm=	55(output 2o-IP(G1))	
G2-db=	8(post-amp	
IP-dBm=	31(output 3o-IP(G2))	
IP-dBm=	55(output 2o-IP(G2))	

(watts)Opt Pwr	1.0e-03	6.3e-04	4.0e-04	2.5e-04	1.6e-04	1.0e-04	6.3e-05
(dB) Opt Loss	0	-2	-4	-6	-8	-10	-12
(RMS) i(pd)=	5.9e-04	4.3e-04	3.2e-04	2.4e-04	1.7e-04	1.3e-04	9.4e-05
(dBm) 50ohms=	-18	-20	-23	-26	-28	-31	-34
Pre-Amp(G1)							
(dB) Gain=	8	8	8	8	8	8	8
(dBm) AMP-OUT=	-10	-12	-15	-18	-20	-23	-26
(dB) AMP-3o=	87	93	98	103	108	114	119
(dB) AMP-2o=	68	70	73	76	78	81	84
Post-Amp(G2)							
(dB) Gain=	8	8	8	8	8	8	8
(dBm) AMP-OUT=	-2	-4	-7	-10	-12	-15	-18
(dB) AMP-3o=	71	77	82	87	92	98	103
(dB) AMP-2o=	60	62	65	68	70	73	76
Optical Receiver	(Composite Receiver Distortion)						
(dBr) Rec'r-3o	70	75	81	86	91	96	102
(dBr) Rec'r-2o	57	59	62	65	67	70	73
Optical Link	(Total Link Distortion)						
(dBc) Link-3o=	61	63	64	64	65	65	65
(dBc) Link-2o=	50	51	52	53	53	54	54

TABLE 2

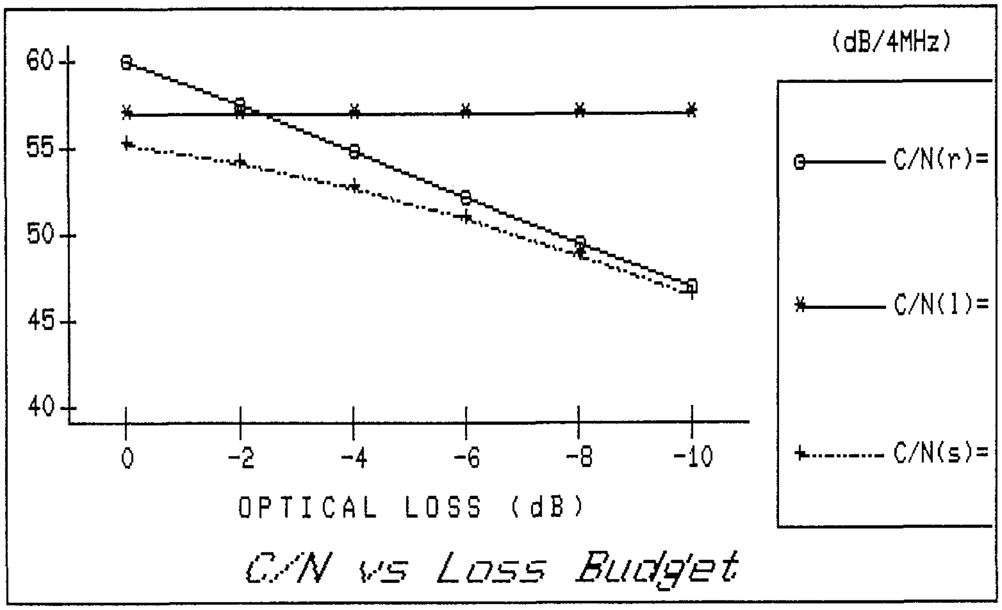


FIG. 4

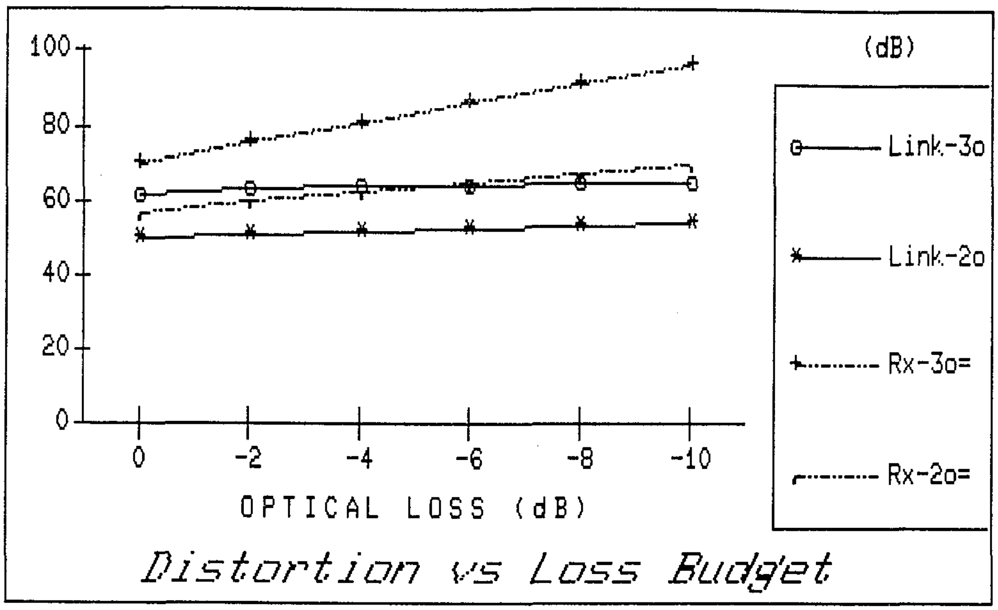


FIG. 5

SUMMARY

A comprehensive RF optical link analysis requires analyzing the C/N and C/IMD performance of the optical link over a range of optical loss budgets. The equations which characterize the optical link's noise and distortion performance have been reviewed and modeled onto a spreadsheet.

In addition to analyzing the optical link's system performance, this analysis technique also enables one to specify the essential performance parameters of the optical transmitter and receiver in order to satisfy specific system design requirements.

HDTV CABLE TESTS: METHOD OF MEASUREMENT

by
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ABSTRACT

A method of measurement was recently developed at CRC to evaluate amplitude and group delay responses of TV channels, broadcast or cable.

This method makes use of conventional test instruments or modified instruments when the channel being investigated has a bandwidth exceeding 4.2 MHz. It was adopted and used by the CABSC (Canadian Advanced Broadcast Systems Committee) during its cable TV field test program in November 1988, which was aimed at characterizing 12 MHz wide cable channels for transmission of HDTV signals. The test method has also been adopted by the ATTC (Advanced Television Test Center) for the characterization of wideband off-air channels in the VHF, 2.5 GHz and 12 GHz bands. Actual tests are underway in Washington DC. This paper describes in detail the concept of the measurement method and its implementation.

INTRODUCTION

The test method described in this paper was developed at the Communications Research Center (CRC) in the context of its contribution to the CABSC's field tests on Cable TV systems. In May 1988, the CABSC Working Group on Channel Characterization undertook to prepare a test plan to characterize 12 MHz cable TV channels. This knowledge would enable the W.G. members to establish the performance specifications of typical wideband cable TV channels. This in turn would allow a realistic assessment of the robustness of the various HDTV signal formats proposed to the FCC, when carried over modern cable TV systems. Although there are some indications[1] that the utilization of wideband channels for off-air HDTV broadcasting is unlikely, due to the major impact on the current frequency plan, there is the possibility that cable TV systems may elect to carry a signal format that is different from the one adopted by broadcast. Also it is believed that the results can be applied to bandwidths smaller than 12 MHz if necessary. Consequently, the test plan[2] was finalized and the tests carried out on three cable systems during Fall 1988.

THE CONCEPT

The objective of the test method is to characterize 12 MHz wide channels by measuring their response to an impulse and then calculating their amplitude, phase and group delay responses. This technique is referred to as the "Complex Impulse Response"[3].

As an example, a low-pass filter can be characterized by its impulse response h(t), (see Figure 1). The Fourier Transform of h(t) provides the mathematical expression of the filter's "transfer function" H(w) which characterizes its amplitude and phase response. The group delay response is simply the derivative of the phase response.

$$F\,[\underbrace{h(t)}_{\text{time domain}}] = \underbrace{H(w)}_{\text{frequency domain}} = \text{transfer function}$$

In the case of an unsymmetrical system[4] such as an rf channel amplitude modulated, with unsymmetrical sidebands (see Figure 2), the response to an impulse includes a carrier frequency w_o which is amplitude modulated by an in-phase term and a quadrature term (i.e. 90° phase from the first term). These two terms brought to baseband through synchronous demodulation constitute the "complex impulse response" of the channel:

$$h(t) = h_I(t) \cos w_o t + h_Q(t) \sin w_o t \quad (1)$$

where w_o = carrier frequency
$h_I(t)$ = in-phase component
$h_Q(t)$ = quadrature component

When an rf channel is perfectly symmetrical about its centre frequency (eg. ideal AM-DSB), $h_Q(t)$ is equal to zero. Therefore, the magnitude of the quadrature term can be seen as a measure of the degree

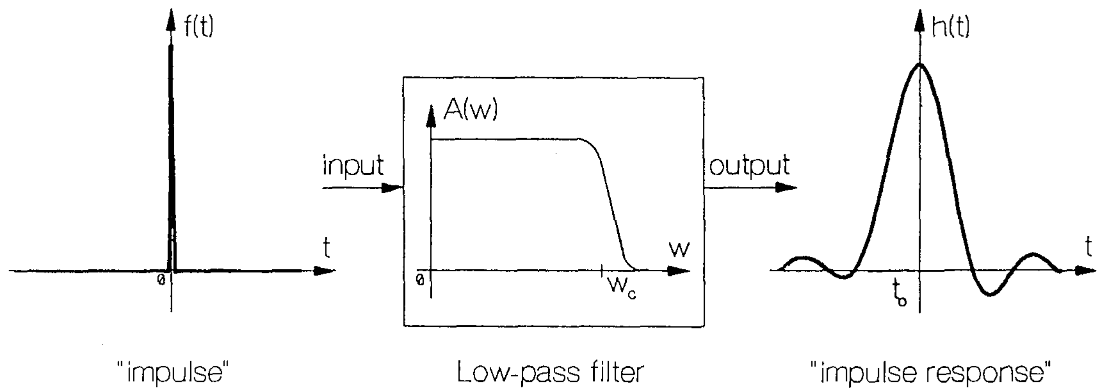

Figure 1a Impulse response

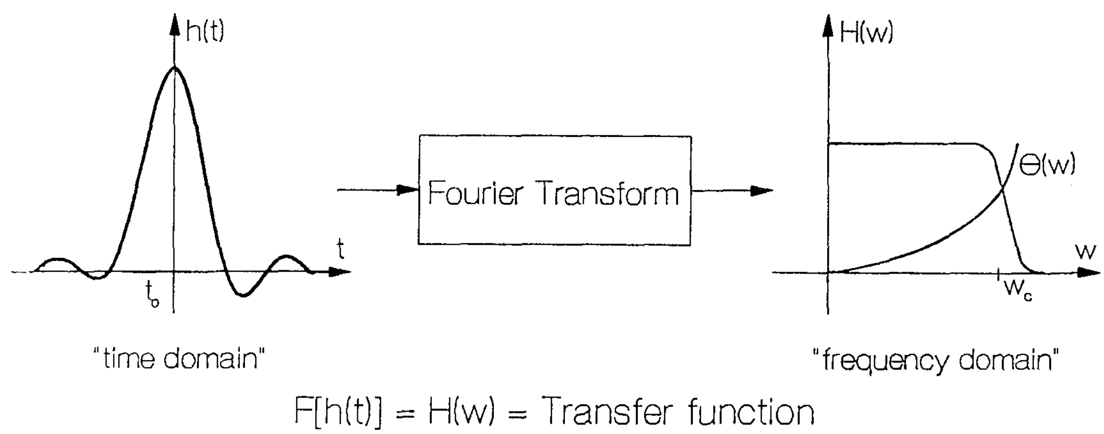

Figure 1b Transfer function

of asymmetry of the channel frequency response.

A synchronous detection process allows the recuperation of $h_I(t)$ and $h_Q(t)$. Then the transfer function H(w) defined over the whole channel bandwidth (12 MHz) can be derived from the expression:

$$H(w) = F[h(t)] = F[h_I(t) + jh_Q(t)] \tag{2}$$

PHYSICAL IMPLEMENTATION

In a real application several simplifications to the theoretical concept are necessary to accommodate the various constraints and limitations of distribution systems and test equipment.

Sin x/x pulse

The impulse had to be traded for a truncated sin x/x pulse[5] whose spectrum has the interesting property of resembling that of the "ideal low-pass filter", which features flat amplitude response and linear phase response up to w_c, the cut-off frequency determined by the width of the pulse (see Figure 3).

The response to a sin x/x pulse is considered equivalent to the response of an ideal impulse since the basic requirement is that the spectrum of the probe signal be constant in the bandpass to be characterized[4].

DSB modulation

A sin x/x pulse with a cut-off frequency of 6 MHz in conjunction with Double-Sideband amplitude modulation was selected to uniformly spread energy in a 12 MHz wide frequency window (see Figure 3d).

By truncating the pulse duration to 24 usec, the pulse could be inserted twice (one positive and one inverted) in a VBI line of a regular NTSC signal (see Figure 6a). The truncation process induced a negligeable amount of ripple on the frequency spectrum of the test signal.

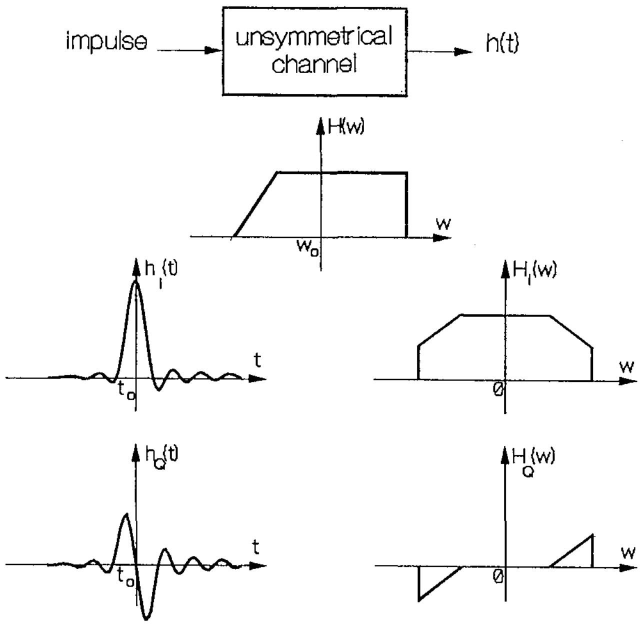

Figure 2 Response to an unsymmetrical channel

The NTSC video signal (Colour Bars) with the 6 MHz sin x/x pulse in its VBI is then fed to an AM-DSB TV modulator whose bandwidth exceeds 12 MHz, and tuned to the visual carrier frequency of the channel under investigation. At the receiving end, a synchronous AM-DSB demodulator also flat on the entire 12 MHz band, locks its phase to that of the incoming carrier and detects the modulating signals on both the I and Q components. The two baseband signals are then digitized to facilitate their processing and storage. This function is performed by a digital oscilloscope with two input channels. The two signals I & Q are tested separately by two 8 bit A/D converters operating at a sampling frequency of 100 MHz (see Figure 4).

<u>Time averaging</u>

The small amount of energy in a sin x/x pulse makes the technique proposed extremely sensitive to noise. Time averaging of video waveforms is a process by which random and time varying components (noise, transients, etc) can be discarded from time invariant components such as the transmitted test signal. In theory the signal-to-noise ratio can be improved by as much as 30 dB when the received test signal is averaged over 1024 transmissions. The theoretical improvement in SNR is expressed by:

$$\Delta\, \mathrm{SNR} = 3 \log_2 n \ (\mathrm{dB}) \qquad (3)$$

where n is the number of times the signal is averaged. Repeated additions of the incoming sin x/x lines take place in the oscilloscope according to a formula that takes into consideration the relative weight of each line. These calculations are done in 16 bit registers, using the 8 bit input samples, (5,000 samples per line) to minimize quantization noise generated by the analog-to-digital conversion process.

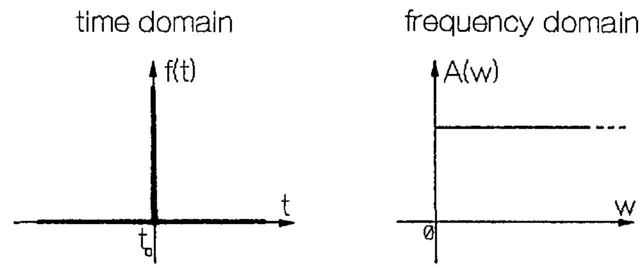

Figure 3a Impulse

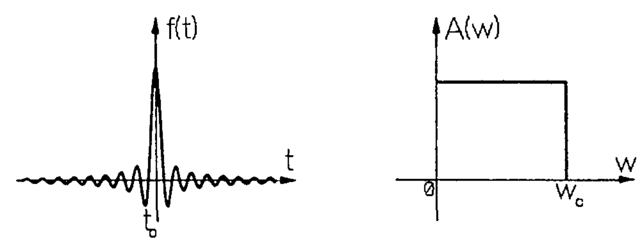

Figure 3b Sin x/x pulse

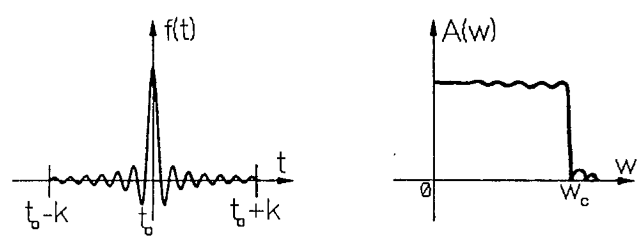

Figure 3c Truncated sin x/x pulse

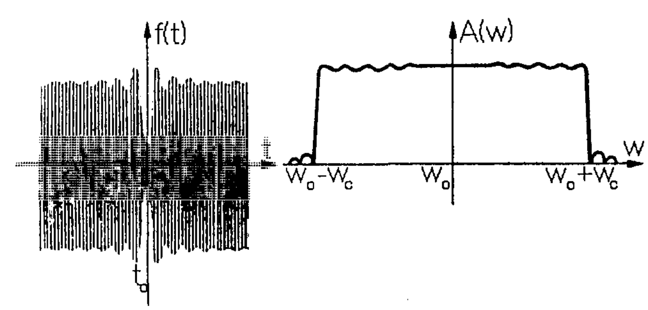

Figure 3d DSB-AM sin x/x pulse

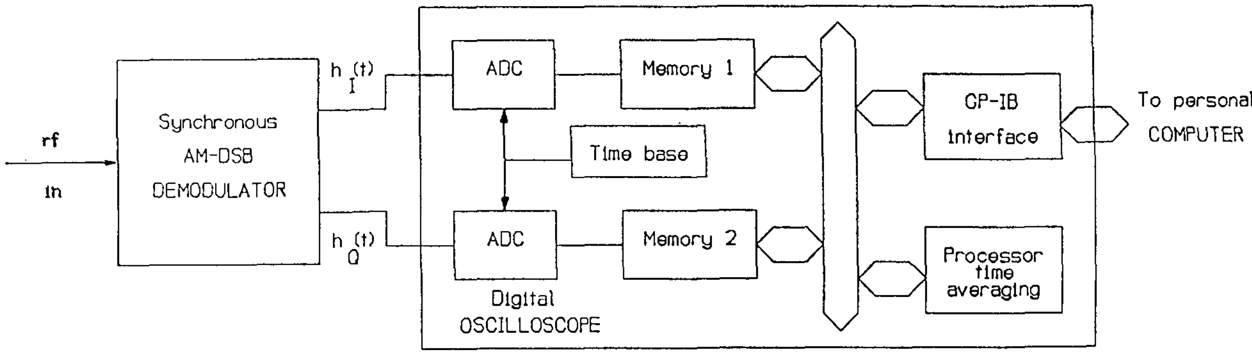

Figure 4 Digitization of impulse response

Triggering

The time averaging process requires that a given sample of the digitized sin x/x signal be averaged with samples corresponding to the same time slot in the test line. This raises the need for a very stable and reliable clock recovery system. The triggering of the oscilloscope must be tightly synchronized to the incoming signal as well as the sampling clock, otherwise a low-pass filtering effect would be observed on the averaged sin x/x pulse due to jitter on the recovered sampling clock. The presence of noise compounds the problem. A digital frame synchronizer which proved to be very reliable even in a 25 dB SNR environment was modified to provide a triggering pulse at the beginning of each 4th field. The oscilloscope could then acquire the sin x/x line after having waited a preset time period. The cadence of one field out of four was dictated by the need to conserve the colour burst of the test line, and the time taken by the processor to do the averaging. It follows that for 2000 averaging periods, each one taking 1/15 sec, the averaging process for one test signal requires a little over 2 minutes. When two signals (I on channel 1 and Q on channel 2) are processed, the time taken by the oscilloscope is doubled and the overall process lasts for 4 minutes.

Instruments Control

In order to automate the process it was decided to control the operation of the oscilloscope with a personal computer via a GPIB bus. A software program was developed to activate the appropriate functions during the acquisition process and to transfer the data from the oscilloscope to data files on the computer hard disk.

Data Processing

Once stored on the computer hard disk, the data files can be retrieved to be processed by a digital signal processing software. The digitized I and Q responses are combined in a Fast Fourier Transform algorithm which produces graphs showing the amplitude, the phase and the group delay responses of the channel measured (see Figure 5).

Later when the signature of the modulator-demodulator is known, it is subtracted from all the impulse responses collected to isolate the contribution of the cable system.

Additional Test Signals

Because a distorted sin x/x pulse does not lend itself to direct and easy interpretation, more "user friendly" tests signals were included in the video signal.

The first one is a special COMPOSITE test signal (see Figure 6b) which includes two $\sin^2$ pulses, one with a half amplitude duration(H.A.D.) of 250 nsec (4 MHz) and one with a H.A.D. of 167 nsec (6 MHz). The second test signal is the "6 MHz line sweep" (see Figure 6c). It consists of a sine wave with a period that decreases linearly from 1.67 usec (600 kHz) to 0.167 usec (6 MHz). This signal allows the field personnel to quickly assess the flatness of the channel amplitude response.

These two additional test signals were to be acquired, time averaged and stored for later comparison with the calculated response of the channel. By adding an X-Y plotter to the set-up, good quality plots of the received test signals can be obtained right at the test site.

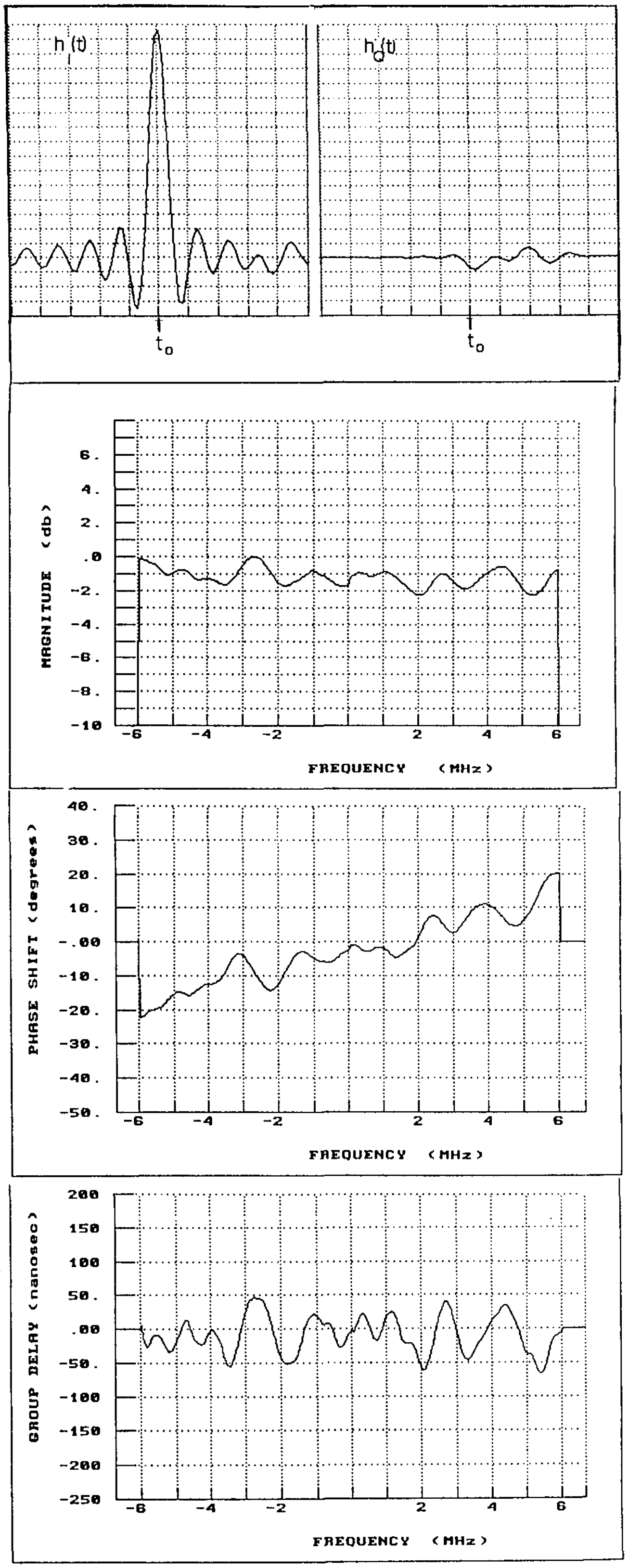

Figure 5 Typical results

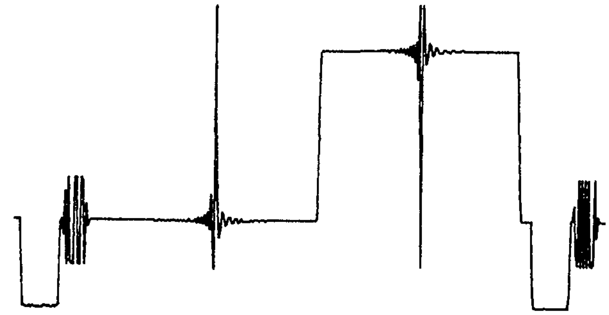

Figure 6a "6 MHz sin x/x" test line

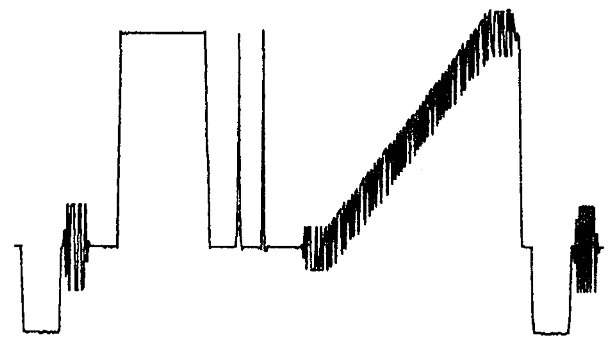

Figure 6b "6 MHz composite" test line

Figure 6c "6 MHz line sweep" test line

CHANNEL TESTS

For the actual tests conducted on cable TV systems, the signal generation equipment was installed at the system headend according to a set-up illustrated in Figure 7. The receive equipment was mounted in a test vehicle which could be driven to any test point of the distribution system (see Figure 8). The "channel" consisted of all the cable equipment and hardware (amplifiers, cable, AML system, taps, etc) comprised between the headend combiner and the subscriber's

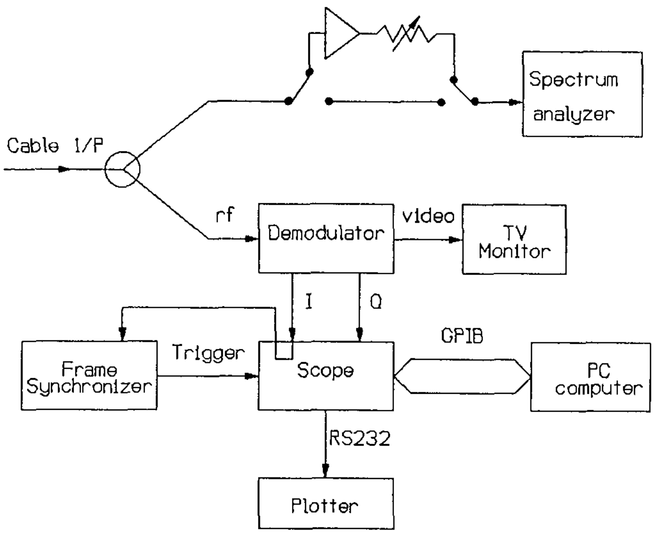

Figure 7 Equipment set-up in the test mobile

drop to which the receive set-up was connected. The carrier frequency was above 400 MHz. Approximately 25 points per system were visited providing results on a good variety of configurations. Also for each system, a "system sweep" was performed by repeating several times the test, at a fixed site, each time using a new carrier frequency to investigate the full CATV spectrum.

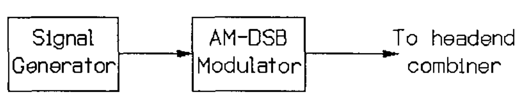

Figure 8 Headend set-up

TESTS RESULTS

At the time of writing this paper, the data collected was being processed and analysed at CRC. Actual results should be available and be presented at the NCTA convention.

CONCLUSION

The concept of the Complex Impulse Response was implemented and used for the characterization of 12 MHz wide cable TV channels. The analysis of the data collected should provide accurate information on certain aspects of modern cable TV systems' performance such as in-band amplitude and group-delay responses, micro-reflections, system group-delay and other parameters that will have an impact on the quality of an HDTV service.

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HDTV PICTURE QUALITY TESTS: METHODS OF MEASUREMENT
An Update from the CCIR Extraordinary Meeting in Geneva

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ABSTRACT

The International Radio Consultative Committee (CCIR) is the branch of the International Telecommunications Union (ITU) responsible for proposing and recommending standards to the world broadcast community. Study Group 10 is specifically charged with broadcast sound services and Study Group 11 is responsible for broadcast television services. On May 10 - 16, 1989, SG 11 held an "Extraordinary Meeting on HDTV" in Geneva, at which discussions continued regarding parameters for a single, world-wide HDTV standard for the studio and for international program exchange.

Subjective assessments of HDTV picture quality provide data which are invaluable in the decision making process. Therefore, the methods used in making such subjective judgments should be the same or equivalent, at least in part, among laboratories worldwide. Single and double-stimulus methods have been discussed and a new recommendation for subjective assessment of television picture quality in an HDTV environment has been drafted.

IMPROVED METHOD FOR VIDEO INVERSION SCRAMBLING SYSTEMS

Blair Schodowski

Scientific Atlanta

ABSTRACT

This paper presents an improved method for inverting and re-inverting video signals. The new technology uses a concept which divides the sync into two levels, which permit calculation of the axis of inversion. The divided sync signal consist of splitting the horizontal sync into a -40 IRE and +100 IRE pulses before baseband inversion. The transmission of the split sync signals eliminates luminances errors when re-inverting the video signal. Splitting the sync greatly improves overall system dynamics, thus increasing security. Deficiencies relating to current video inversion technology, along with a description of the scrambling method will be discussed.

INTRODUCTION

Video inversion involves a process which reverses the light and dark levels in a video signal. The light and dark levels are swapped by rotating the video signal around a reference located between white and sync tip. Ideally in the re-inversion process the video signal is restored to its original integrity by rotating the inverted signal around a reference equal to the axis of inversion. The reference establishes an axis of inversion, which is an essential component in the process of re-inversion. The importance of an accurate axis of re-inversion is best shown through an example. Consider the example waveform shown in figure 1 rotated between points (a) and (b). Figure 2 illustrates the inversion being performed around a 30 IRE axis of inversion. The 30 IRE axis of inversion was chosen primarily for ease of calculating an axis of re-inversion, which will be discussed later. However, it is possible to invert the video signal about any axis, providing the re-inversion axis is identical to the inversion axis. Figure 3 illustrates the consequence of re-inverting around a 40 IRE axis instead of the desired 30 IRE axis. Note that the re-inverted waveform

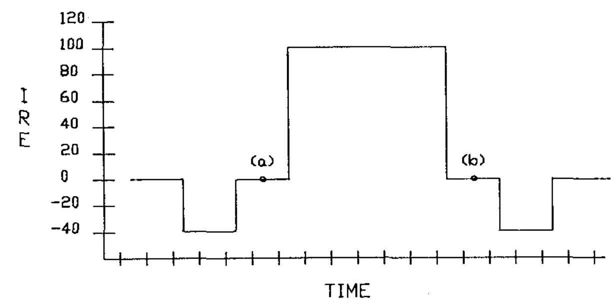

Figure 1. Example video waveform.

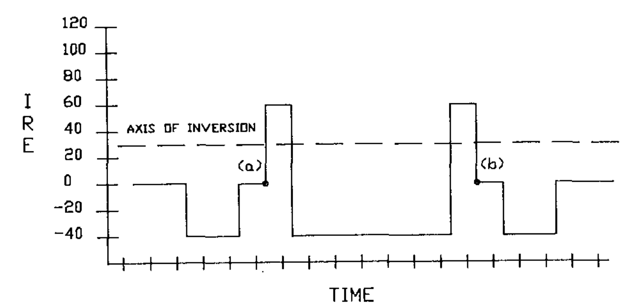

Figure 2. Example video waveform inverted around a 30 IRE axis.

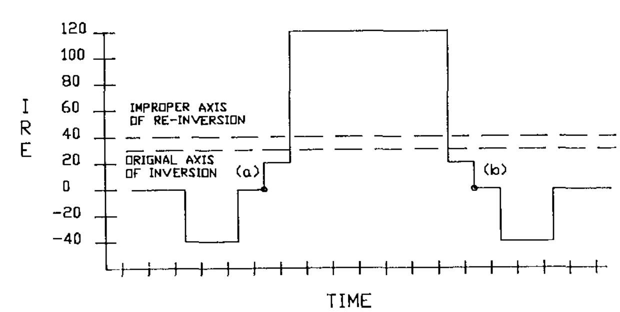

Figure 3. Inverted waveform re-inverted around a 40 IRE axis.

has been expanded with respect to the non-inverted waveform. Note the portion of the signal that initially was at blanking and is now at 20 IRE. Also, observe that the white level is now at 120 IRE. If the axis of inversion was lower than the desired axis the re-inverted waveform would have been compressed with respect to the original signal.

Mathematically the basic equation for inverting a signal can be written as

INV SIG = 2 X AXIS OF INVERSION - SIGNAL TO BE INVERTED.

Similarly the equation for re-inverting a signal can be shown as

RE-INV SIG = 2 X AXIS OF RE-INVERSION - INVERTED SIGNAL,

where the axis of re-inversion is equal to the axis of inversion plus any error between the two axes.

An important relationship in the inversion process shows that the difference between the re-inverted signal and the signal before inversion is equal to twice the axis error. This is illustrated in figure 3 where a 10 IRE axis error expanded the re-inverted signal by 20 IRE compared to the non-inverted signal. In addition it should be pointed out that a precise axis of inversion is equally important as an accurate axis of re-inversion.

There are basically three inversion modes possible for inverting a video signal. The first mode, shown in figure 5, depicts active video inversion only, with normal horizontal blanking. Figure 6 represents sync inversion, which technically is inverted horizontal blanking with normal active video. The third mode, all inversion, shown in figure 7, inverts both the horizontal blanking and active video.

Examination of active video inversion with normal horizontal blanking shows that inverted active video could become the most negative level of the video signal. Since sync recovery circuits in television receivers rely on the most negative level of video for synchronization, reliable synchronization is virtually impossible. Without horizontal synchronization the television picture will tear and become unviewable. In the event synchronization is established the recovered video will be the negative of the original signal. Also, as a result of the inversion process the color information will be rendered incorrect because the phase of the color subcarrier is shifted by 180 degrees. Since the horizontal sync is inverted in both sync inversion scrambling and all inversion scrambling there is no possibility of horizontal synchronization.

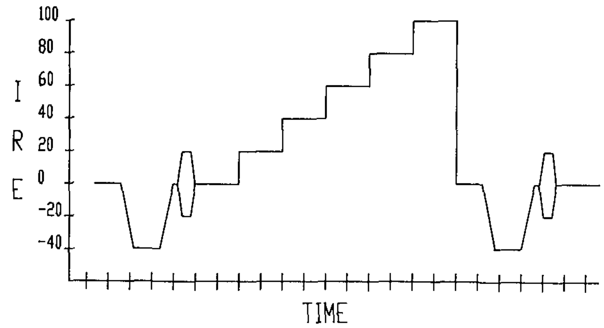

Figure 4. Example waveform.

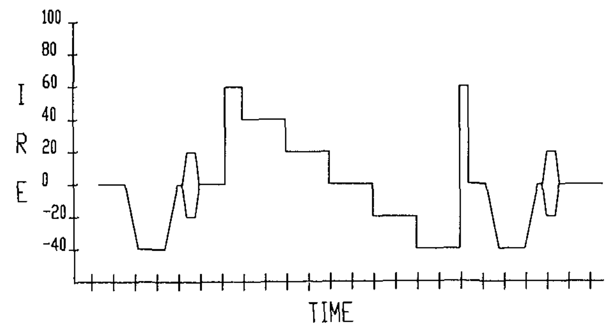

Figure 5. Active video inversion with normal horizontal blanking.

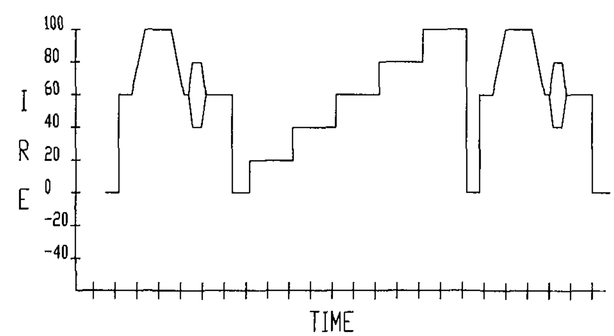

Figure 6. Inverted horizontal blanking with normal active video.

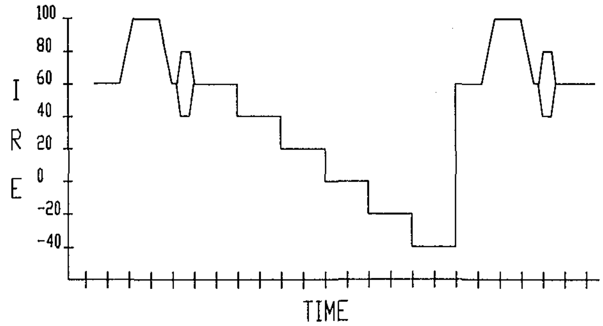

Figure 7. Inverted active video and horizontal sync (all inversion).

DEFICIENCIES OF PAST SYSTEMS

At first glance it would appear that inverting video or horizontal sync would be the ideal method for scrambling a video signal. This would be partially true if the transmission and receiving media was baseband to baseband. However, the baseband signal is transmitted at RF and therefore needs to conform with RF transmission limitations, particularly the modulation and demodulation process. This limitation is the source of one of the two major deficiencies associated with past systems. The deficiency arises when horizontal blanking is inverted and modulated as indicated in figure 8. After modulation, sync tips are no longer the highest amplitude of the RF envelope. Video demodulators rely on sync tips to be the highest RF amplitude for automatic gain control (AGC). Proper AGC is necessary in order to establish correct video level proportions. Systems have been developed that compensate for this deficiency by incorporating very slow time constant AGC circuits that respond to the non-inverted sync pulses in the vertical interval. However, very objectionable artifacts are generated in the recovered signal when utilizing such a measure. The most obvious artifact is a luminance shift when changing channels or scrambling modes. Due to the nature of scrambling systems in which non-inverted sync pulses and inverted sync pulses are used, a video demodulator that AGC's to the most negative amplitude level will not work. Dual video demodulators can be used but basic economics and the inability to match gains and clamping level preclude this approach.

In addition to the demodulator not responding to the inverted horizontal blanking, video modulators react in a similar fashion. Video modulators rely on the sync tip amplitude to clamp the video signal to assure constant peak modulation with varying input levels. Essentially, sync tips are not present when the horizontal blanking interval is inverted. Without costly modifications to dedicated modulators this limitation precludes the utilization of this type of scrambling.

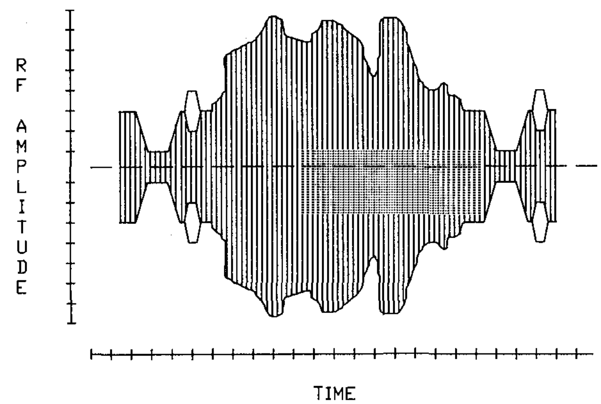

Figure 8. RF envelope of an inverted horizontal blanking waveform.

The second deficiency manifests when the descrambled axis of inversion fails to equal the scrambled axis of inversion. This scenario becomes a reality when headend or descrambling conditions change. Older systems do not have the dynamic capability of maintaining identical axis between scrambling and descrambling with varying conditions. Since any error in axis between scrambling and descrambling will cause a luminance shift of twice the axis error, axis integrity is very essential. In some systems axis integrity relied on factory calibration settings to match inversion and re-inversion axes. Figure 9 illustrates a descrambler configuration which is used in past systems. This type of configuration, which relies on fixed factory calibration, poses a serious axis problem. Consider a calibration signal from the factory's headend which is inverted about a 30 IRE axis and transmitted at a set depth of modulation (DOM). The descrambler can then be calibrated to measure from sync tip a set axis of re-inversion which equals

SYNC TIP LEVEL + (70/140) X (VIDEO LEVEL),

where the video level is from reference peak white to sync tip. If the video level remains equal to the level of video when calibrated, the descrambled signal will match the original signal before inversion. Now consider the consequence when the field modulator DOM differs from the calibration modulator DOM. A change

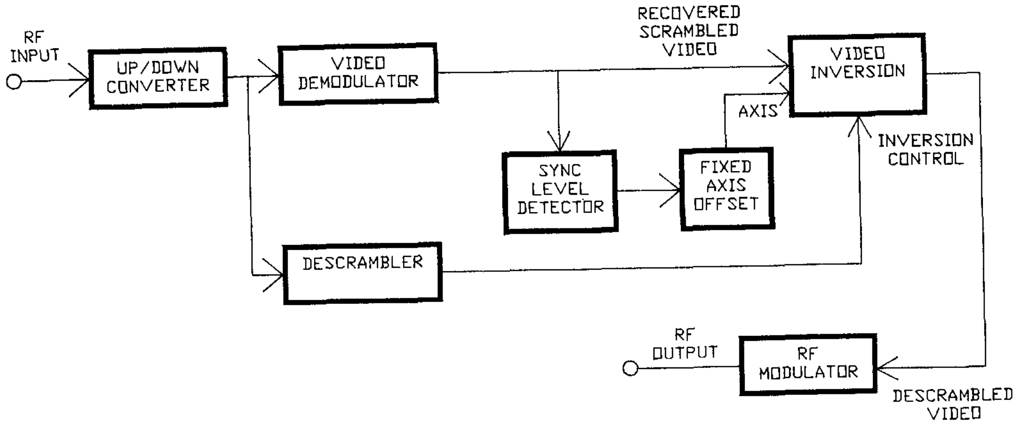

Figure 9. Descrambler configuration which determines axis of re-inversion by fixed offset.

in headend DOM causes the recovered video level to be different from the level of video when calibrated. As would be expected the re-inversion axis is in error by

(70/140) X [(NEW VIDEO LEVEL) - (CAL VIDEO LEVEL)]

For example, if the calibration video level is 1 volt and the video level in the field is .9 volts the axis error will equal 7.78 IRE. By virtue of inversion the 7.78 IRE axis error will cause a 15.56 IRE luminance shift, thus resulting in brightness variations in the television picture.

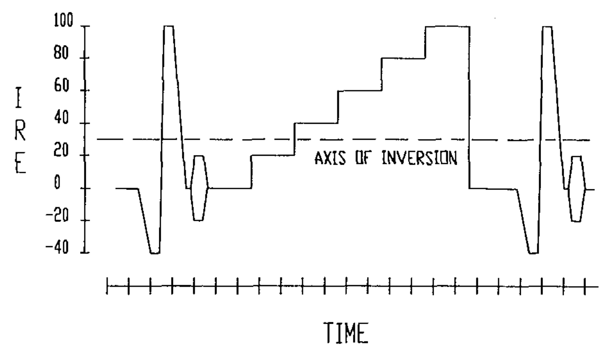

Figure 10. Example waveform with split sync signal.

SPLIT SYNC INVERSION

Fortunately, there is an economical method for overcoming the deficiencies of past video inversion systems. The new method involves utilizing a technology which transmits a modified sync signal during the conventional sync time. The signal is transmitted by splitting the sync interval into two components. Figure 10 illustrates an example waveform which incorporates the split sync signal. The first component of the signal is the conventionally transmitted -40 IRE sync. After a time equal to about 2.0 microseconds the -40 IRE signal level increases to the peak amplitude of 100 IRE. The 100 IRE signal pulse remains at 100 IRE for about 2.0 microseconds before returning to 0 IRE.

The split sync signal is then used in an analog computation by the descrambler to calculate an axis of re-inversion. As previously mentioned a 30 IRE axis of inversion is used in order to facilitate a simple axis of re-inversion calculation.The axis is calculated by simply computing the difference between the peak sampled 100 IRE and -40 IRE signal. The absolute axis which equals

SYNC TIP LEVEL + (.5) X [(100 IRE LEVEL) - (-40 IRE LEVEL)],

is then applied to the inversion amplifiers, which precisely re-inverts the inverted signal to its original integrity. The utilization of a technology that dynamically calculates a re-inversion axis, instead of a fixed axis offset, makes the re-inversion process immune to recovered signal level. For example, assume if the waveform in figure 10 was inverted around a 30 IRE axis and transmitted at a set DOM. After recovering the signal in the descrambler the example waveform would resemble that in figure 11. With the -40 IRE and 100 IRE signal levels equal to 4.0 and 6.0

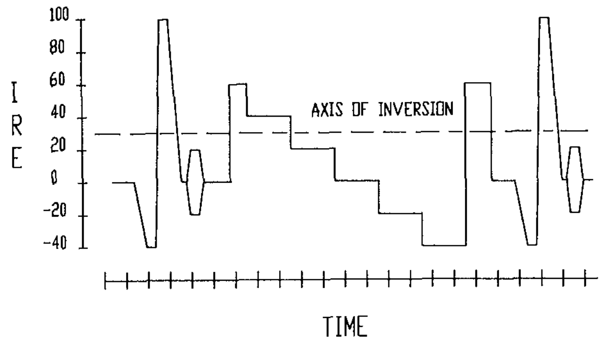

Figure 11. Recovered (inverted) signal.

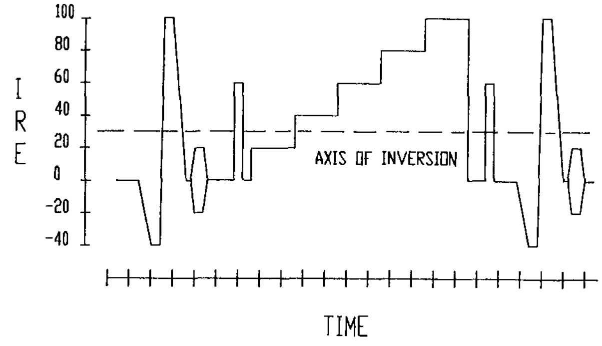

Figure 12. Re-inverted signal.

volts respectively, the calculated absolute axis of re-inversion would equal 5.0 volts. The calculated axis of 5.0 volts lies exactly centered between the peaks of the split sync signal, which correspond to a 30 IRE axis. Re-inverting the inverted signal around this calculated 30 IRE axis will restore the signal to its original integrity, as shown in figure 12. Now consider that the headend DOM changes, resulting in recovered signal split sync peaks of 4.0 volts and 5.5 volts. The calculated absolute axis of re-inversion would equal 4.75 volts, again placing the axis centered between reference white and sync tip. As demonstrated by varying the headend DOM one can see that the axis of re-inversion is independent of DOM. The ability to dynamically calculate an axis of re-inversion not only removes the dependency on headend DOM adjustments, but also improves the systems ability to dynamically change scrambling modes.

An advantage of scrambling a signal by splitting the sync is that many television receivers require the entire 4.7 microsecond sync to establish synchronization. However, in the descrambling process the signal transmitted to the suscriber's television must be capable of working with all television receivers. Therefore, the descrambler must restore the split sync signal so that the horizontal blanking interval complies with the NTSC video signal standard. This is accomplished by replacing the split sync signal in the descrambler with the proper sync level prior to re-modulation. The diagram in figure 13 illustrates a portion of a descrambling system which dynamically calculates the axis of re-inversion along with restoring the split sync signal.

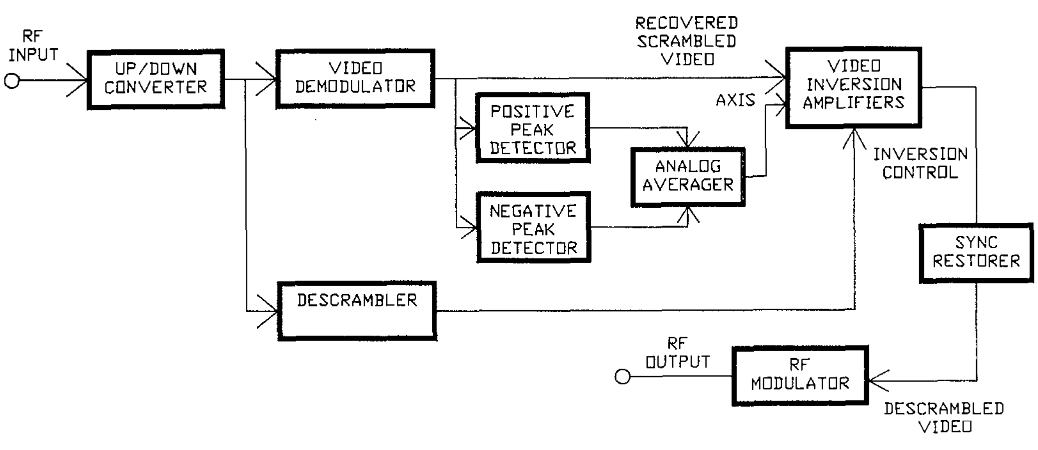

Figure 13. Descrambler configuration which calculates axis of reinversion.

An inherent advantage to splitting the conventional sync with a 100 IRE and -40 IRE signal is that headend modulators do not require any modifications. Modifications are not required because adequate sync tip always remains to permit proper clamping. Traditionally, the inversion modes that a modulator would have trouble with are sync inversion and invert all. Inspection of the example waveforms shown in figures 14 and 15, which represent sync inversion and invert all respectively, shows that the modulator will always have a -40 IRE signal required for proper operation. This is possible due to inverting the 100 IRE split sync signal around a 30 IRE axis. As illustrated the 100 IRE signal becomes -40 IRE and the -40 IRE signal becomes 100 IRE.

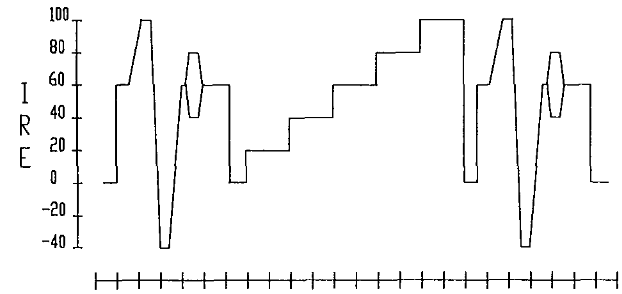

Figure 14. Sync inversion with split sync.

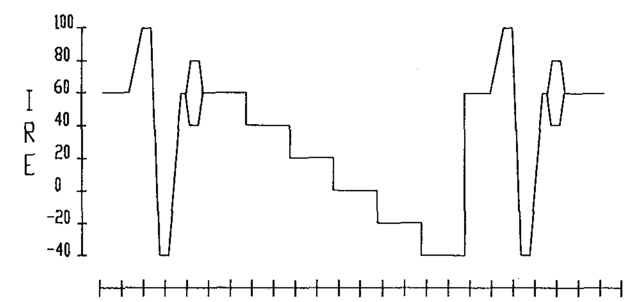

Figure 15. Invert all with split sync.

In addition to furnishing the headend modulator an adequate signal for modulation, the split sync signal also allows the descrambler's demodulator to function properly. The inverted 100 IRE signal establishes the required signal necessary for proper demodulator AGC operation. Figure 16 shows an example of an RF envelope that has been modulated with an all inverted split sync video signal. Note that when modulating the inverted 100 IRE signal it becomes the most positive amplitude of the RF envelope. This positive signal is what is required by the demodulator for reliable operation. By always having a peak signal level present reliable demodulator performance is maintained in all modes of scrambling.

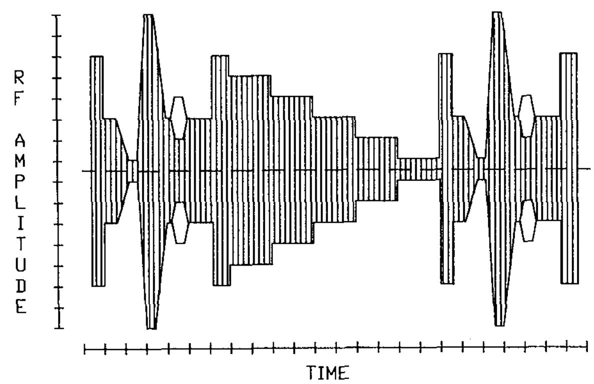

Figure 16. RF envelope of inverted sync & active video with split sync signal.

CONCLUSION

This improved method of video inversion eliminates many of the limitations affecting some previous video inversion systems. Besides eliminating deficiencies the improved inversion system increases overall system performance and capability. Dynamic generation of the re-inversion axis eliminates the need for precise factory modulator adjustments. Axis integrity in all modes of inversion allows for dynamic changing of scrambling modes without any chrominance or luminance artifacts. The ability to dynamically scramble increases system security, which is becoming a major concern throughout the industry. The new technology increases security further when integrated with older scrambling techniques such as sync suppression.

INTEGRATED AML/FIBER BACKBONE SYSTEMS

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ABSTRACT

A fiber optic backbone system fed by AML is a cable system architecture that provides both performance and cost advantages. Although both AML and a fiber backbone have been separately proposed as means of improving the overall cable system carrier-to-noise ratio, the attributes of AML and AM-fiber are in this case complimentary rather than competitive. By combining the two technologies, one can overcome the drawbacks of each. Line-of-sight and zoning restrictions sometimes limit the location of AML receive sites. Shot and thermal noise sharply limit the carrier-to-noise ratios achievable with multiple-carrier AM fiber on long paths. When the latest AML technology is used to reduce the average length of the fiber backbone, the overall system C/N can be improved. At the same time, the savings in the cost of the glass can more than offset the cost of the microwave. This paper reviews AM fiber and recent AML system performance. Examples of integrated AML/fiber backbone architecture are analyzed for both cost and performance. It is shown that an overall C/N in large cable systems of 50 dB or better at the last subscriber terminal can be obtained with today's technology.

INTRODUCTION

The fiber backbone system concept was described in a series of papers presented at the 1988 NCTA convention.[1-3] The performance goals of this system were stated to be a 10-dB optical loss budget, 42 channels, and 55 dB C/N with 65 dB C/CTB and C/CSO. By cutting the trunk amplifier cascade length to two to four amplifiers, the fiber backbone concept should provide the advantages of improved reliability, quality, and maintainability for the overall cable system. Back in 1976, similar advantages were found to apply when AML microwave was used to cut trunk cascades to a maximum of ten amplifiers.[4] However, it is not always feasible to use microwaves for these purposes. A clear line of sight with adequate path clearance is required. Zoning restrictions may ban the installation of receive sites, particularly in residential neighborhoods. In addition, if the trunk cascade is to be cut to two to four amplifiers, the number of receive sites in major cable systems would imply a broadcast type of transmit antenna. With existing power limitations, the microwave system would be restricted to very short range even if such a broadcast antenna pattern were permissible under CARS band rules. Currently, the largest point-to-point AML system utilizes only 32 receive sites.

On the other hand, it must also be acknowledged that today's AM fiber systems still fall short of the above-stipulated performance goals, particularly at larger distances. Moreover, with a large number of fiber hubs, and multiple glass fibers to each hub, the overall cost of glass is not an insignificant item. For these reasons, it us useful to consider a system architecture using AML microwave to sharply reduce the length of the fiber runs. Each microwave receive site, aside from taking the place of one fiber hub, then becomes the source for feeding a dozen or more fiber backbone hubs. With modern AML equipment, it is possible to achieve high-quality performance at distances in excess of 20 miles. This reach should not be confused with 32 kilometers of fiber. Whereas microwave is "as the crow flies" distance, fiber must follow routings dictated by local conditions. Even when there are no natural barriers, such as river crossings, involved, a reasonable expectation might be that the required fiber distance exceeds the microwave distance by 30 percent. Thus, the equivalent reach is 41 km of fiber. To this, one can add up to 10 km of AM fiber backbone for a total equivalent reach of over 50 km.

AM FIBER SYSTEM PARAMETERS

The general characteristics of the C/N performance of an AM fiber system have been clearly described.[5] The three contributions to overall C/N are

$$C/N_{SOURCE} = \frac{m^2/2}{RIN \cdot B} \tag{1}$$

$$C/N_{QUANTUM} = \frac{m^2 R P_R/2}{2qB} = \frac{m^2 \eta P_R/2}{2h\nu B} \tag{2}$$

$$C/N_{RECEIVER} = \frac{m^2 R^2 P_R^2 R_{eq}/2}{4kTB \cdot F} = \frac{m^2 R^2 P_R^2/2}{< i_N >^2 B} \tag{3}$$

where m is here taken as the modulation index for each individual TV channel, which is often assumed to relate to a total modulation index $M = m\sqrt{N}$, with N being the number of channels. RIN stands for "relative intensity noise" and normally describes the intensity noise of the laser. However, multiple reflections on the fiber system, aside from possibly directly degrading laser RIN, can also give rise to additional RIN through conversion of phase noise to intensity noise.[6] A typical linewidth for a DFB laser is 50 MHz. With this linewidth, a better than 40 dB return loss must be required of all fiber system components to keep the additional RIN at channel 2 (54 MHz) under -160 dBc/Hz. This is important when, with the use of optical isolators, the laser RIN is maintained at -152 dBc/Hz or better.

In equation (2), η is the quantum efficiency, a measure of the probability that an incoming photon of energy, $h\nu$ (h = Planck's constant and ν = optical frequency) will generate a hole-electron pair that is collected across the junction of a p-i-n photodetector. Although quantum noise is identified with receiver shot noise, it is based on a fundamental limit intrinsic to the electromagnetic field, wherein the background noise radiation at optical frequencies is approximated by $h\nu B$, rather than kTB as in microwave satellite receive terminals. A factor of two arises because direct detection is less sensitive than heterodyne detection. Since η is already quite high (a 1.3 μ detector responsivity, R of 0.7 amps/watt implies a 67 percent quantum efficiency) the only available means of significantly increasing the C/N when quantum noise is dominant is to raise either m or the average optical received power, P_R. Note that with electron charge, $q = 1.6 \times 10^{-19}$, P_R is in watts. With the NCTA definition of C/N, $B = 4 \times 10^6$.

A great deal of effort has been expended within the last decade in optimizing optical receiver sensitivity. This continuing effort[7] has focused on transimpedance amplifier designs suitable for high speed data communications. Standard receivers of this type can respond out to 550 MHz with an equivalent transimpedance, R_{eq}, of 2 kΩ and beyond 330 MHz with $R_{eq} = 5$ kΩ. Unfortunately, at the high P_R required by equation (2), standard receiver designs suitable for data communications are not sufficiently linear for 40-channel CATV applications. In particular, second-order distortion limits the transimpedance to on the order of 500 ohms for high-level input. Equivalently, one can ascribe an equivalent input noise current density, i_N, whose square is proportional to a noise factor, F, divided by R_{eq}. In either case, noise can be expected to increase somewhat with frequency so that the worst case $C/N_{RECEIVER}$ occurs at the high frequency channels.

Table I summarizes the assumed contributions to C/N for a hypothetical 42-channel link. It is obvious that all three contributions to system C/N must be improved to meet the original fiber backbone requirements. A 3 dB increase in laser power output would result in a 6 dB improvement of $C/N_{RECEIVER}$ but the receiver distortion limit must be raised with higher P_R. Raising transmitter output by 3 dB also increases $C/N_{QUANTUM}$ by 3 dB. At this point, C/N_{SOURCE} would become the dominant term and RIN would have to improve.

The only factor that enters into all three terms is the modulation index, m. Improved laser linearity would be required but "crash point" saturation limit cannot be very far removed since even with 4-percent per channel modulation, the 42-channel instantaneous current can, however briefly, drive the laser to below its threshold current. It has been pointed out[8] that phase fiddling in HRC systems could be useful in this regard.

The optical loss is normally assumed to be 0.5 dB/km at 1.3 μ. This includes an allowance for splice loss, but connector losses at transmitter and receiver ends and residual link margin are not included. The CATV operator will have to decide whether the planned fiber link distance can be based directly on the optical loss required for given C/N or whether 1 or 2 dB should first be subtracted before applying the 2 km/dB formula. Figure 1 plots the Table I C/N versus distance assuming a 1 dB loss holdback for connectors.

RECENT AML DEVELOPMENTS

Figure 2 summarizes the relative output capability of AML transmitters. The point to be made is not only the wide range in output capability but also the wide diversity of choice. The day when AML transmitters were available in only two varieties is long since gone.

Two transmitters are of particular recent significance. The SSTX-145 is a solid-state high-power channelized transmitter[9] that is almost comparable in power with traditional

TABLE I
ASSUMED FIBER OPTIC LINK PARAMETERS

Optical Loss (dB)	C/N_{SOURCE}	$C/N_{QUANTUM}$	$C/N^*_{RECEIVER}$	C/N_{LINK}
2	55.0	60.6	68.0	53.8
4	55.0	58.6	64.0	53.1
6	55.0	56.6	60.0	52.0
8	55.0	54.6	56.0	50.4
10	55.0	52.6	52.0	48.2

m = 4% (N = 42)
RIN = −152 dBc/Hz
P_{LASER} = 2 mW (into fiber after isolator)
R = 0.7 A/W
i_N = 5 pA/$\sqrt{\text{Hz}}$

*Distortion, particularly at higher input levels, may be excessive.

high-power AML but uses half the floor space and one fifth of the primary power. At the recent Western Cable Show, this transmitter was teamed with a new Compact Outdoor Receiver[10] for a live demonstration of a simulated eight-output 40-channel 32-km microwave link with 60 dB

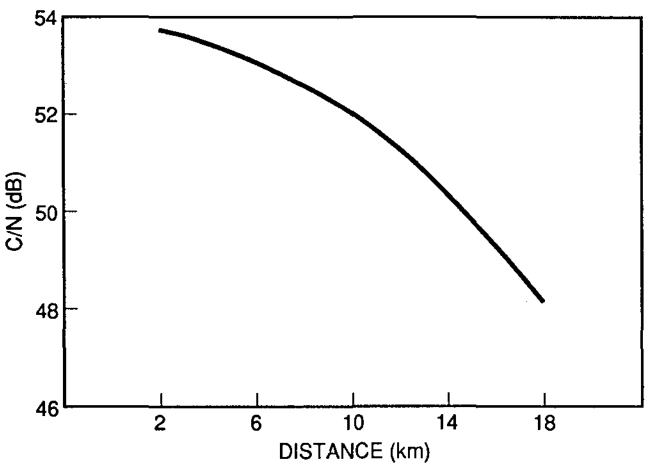


Figure 1 Calculated AM fiber-optic link C/N versus distance.

S/N. The AML demonstration equipment is depicted in Figure 3. By measuring baseband characteristics including differential gain and phase, it was shown that the signal was indeed of a high quality. The S/N was largely determined by the higher than normal receiver microwave AGC threshold setting. This level setting trades off C/N against C/CTB and C/CSO. At the normal factory setting of -46 dBm for the COR-299 6-dB noise-figure receiver, C/N is 56 dB, C/CTB is 75 dB, and C/CSO is 70 dB for 40-channel loading.

Figure 4 shows another recent AML development, the block upconverting IBBT-116 transmitter.[11] Table II summarizes its performance capabilities. This transmitter with a two-tone 3-IM intercept point of +57 dBm has 8 dB greater output capability than any previous CARS-band block-conversion type of transmitter. It is capable of full 80-channel loading, but when loaded with only 42 channels, its output is +9 dBm with 60 dB C/N, 65 dB C/CTB, and 65 dB C/CSO. Including a four-way split to 16-km microwave paths, the received signal level would be -42 dBm.

It is clear from the above that for supertrunk applications, AML microwave performance far outpaces what AM fiber systems can deliver. Moreover, for the two examples given, overall system costs for AML microwave will be far less than for the corresponding fiber system (disregarding for now the performance differences). Cost will of course vary greatly depending upon site availability, type of fiber construction,

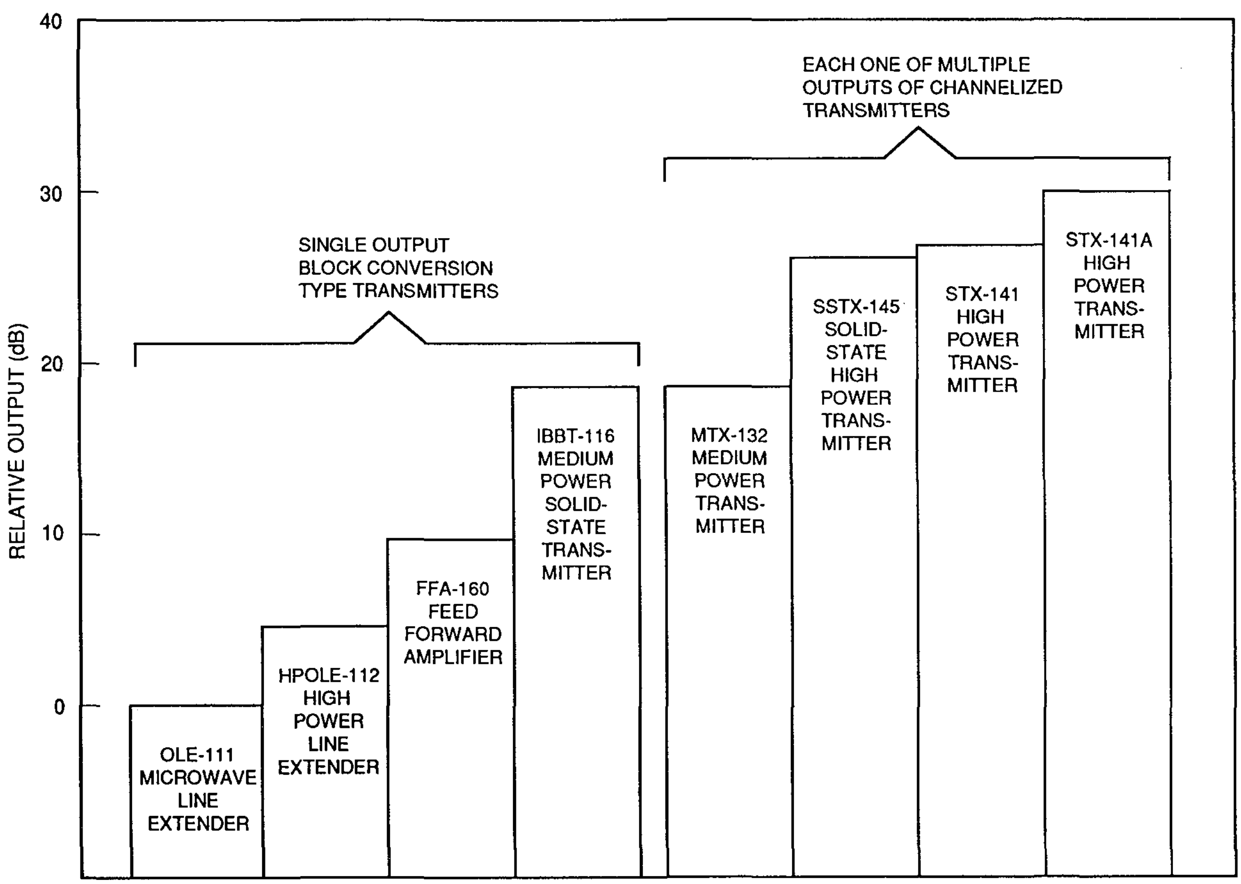


Figure 2 Relative output capability of AML transmitters.

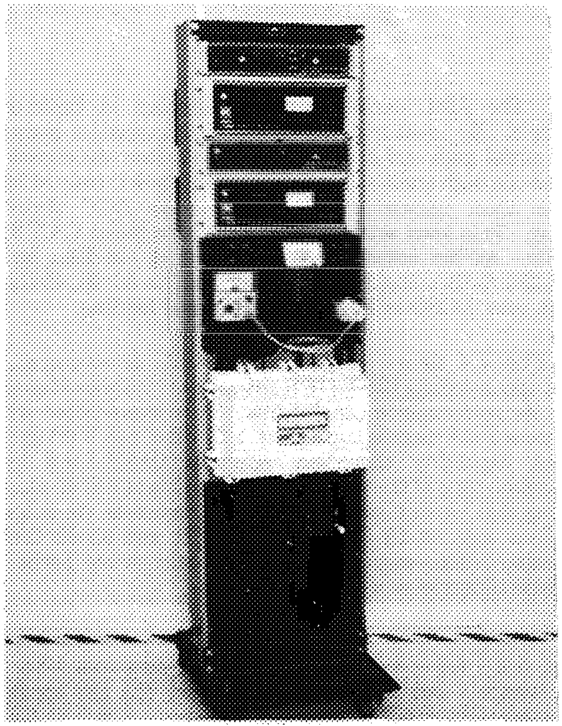


Figure 3 AML equipment used in 60-dB S/N demonstration.

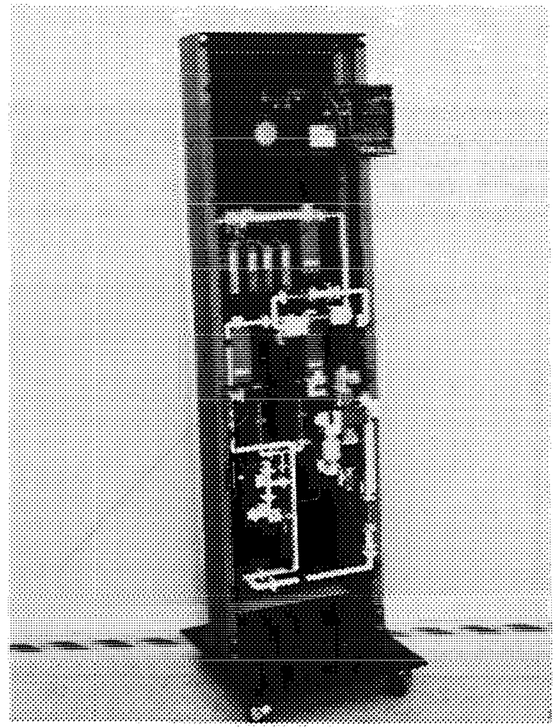


Figure 4 AML IBBT-116 transmitter.

TABLE II

IBBT POWER OUTPUT AND C/N FOR 65 dB C/CTB AND 65 dB C/CSO

No. of Channels	P_0 (dBm)	C/N (dB)
12	15	66
21	13	64
35	10	61
60	7	58
80	5	56

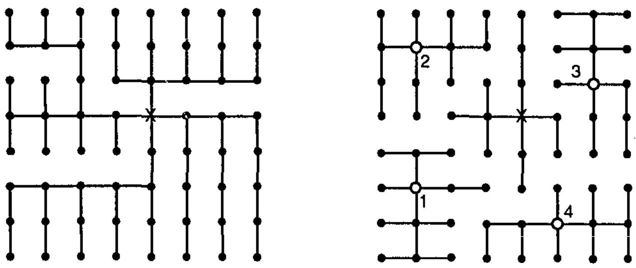
etc., but in general, microwave will be more economical except for applications involving multiple paths under two to three miles in length or where the total of all path lengths add up to less than ten miles. Thus, if cost and performance are the criteria, AML microwave will be preferred in most supertrunk applications. However, in the fiber backbone application, the one technology complements the other.

COMBINED AML AND FIBER BACKBONE

Consider a rather idealized fiber backbone system in which the fiber nodes are uniformly spaced on an 8 by 8 grid. Assume further that the central head-end is located at the point "X" shown in Figure 5a. If the streets run north- south and east-west, the fiber routes might exit the head-end as shown. In total, there are 63 fiber hubs with the four directions connecting respectively to 17, 16, 15, and 15 hubs. If the spacing between hubs is conceived to be unity distance, the maximum length fiber run is eight units long, and the average distance is four units.

Contrast this with the situation in Figure 5b, in which four AML receive sites, indicated by the circles, have been added. The maximum length of fiber run is now reduced to three units, and the average length is 1.85 units. The number of fiber hubs has also been reduced down to 59, because the AML receivers replace the fiber hubs at their locations. The total cable distance is likewise reduced from 63 to 59 units. The central head-end services 11 fiber sites, while each of the AML receivers connects to 12 fiber hubs. Table III summarizes the situation. The cost savings that can be realized in the fiber plant will, of course, depend critically on the actual unit distance. Typically, the "unit" will be in the range of one to two miles or even greater if the length of the trunk cascade is allowed to grow above 4.

A second critical parameter is the number of fibers that will be dedicated to each hub. An estimate of four (including spares) may not be unreasonable, but in some cases there may be even more. One reason for using multiple fibers is to reduce the channel loading on the individual fiber link. In particular, if the loading is reduced to 18 channels, a frequency plan that avoids in-band second-order distortions can be constructed. Aside from being able to increase the per-channel modulation index, m, roughly in proportion to the inverse square root of the



(a) WITHOUT AML (b) WITH AML

Figure 5 Idealized fiber backbone systems.

number of channels, a further increase in m may be possible if filtering is applied to remove the out-of-band second-order products at the photo receiver output prior to recombining the channels. In all, the C/N shown in Figure 1 might then be increased by about 4 dB, assuming all other DSB laser and receiver parameters were held the same. The exception would be the cross-over channels since the broadband noise would leak through and degrade C/N at the filter band edge. When the signal source is also broadband, as is the case with the AML receiver, it is probable that a guard-band channel would have, in any case, to be set aside to prevent undesired signal phasing effects due to inadequate overall filtering at the source and fiber receiver ends. In any case, multiple fiber links to each fiber hub, although increasing complexity and cost of the electronics (assuming the same quality laser and receiver) is another option which presents itself to the CATV system designer.

To make a numerical comparison between the fiber backbone systems with and without AML, it is necessary to assign a definite length to the unit distance in Table III. With 2-1/4 km, the maximum fiber run length without AML is 18 km. It is assumed that increased C/N can be traded 1:2 for C/CTB without "crashing" the fiber system. Adding 1 dB to Figure 1, one then achieves a more respectable 49.2 dB. However, for the shorter 6-3/4 km maximum fiber distance with AML, normal 65-dB CTB operation is assumed. The AML system consists of an IBBT-116 transmitter backed off to +7 dBm/channel output to improve C/CTB. This can be done, since the maximum AML path length here is only five miles long. The calculation assumes that C/CTB from a chain of dissimilar devices will add randomly, i.e., on a power-addition basis. The AML system cost includes the transmitter, four receivers, antennas, waveguide, typical installation costs, and a $30 K allowance for a transmit tower. The advantage in both cost and performance is evident even at these small distances. As the unit distance increases, the advantages of incorporating AML will tend to increase further.

It is of interest to compare this idealized system with a real CATV system layout. For this purpose, an enlarged cable system trunk route map corresponding to the fiber backbone system described in references 2 and 3 was obtained. Figure 6 shows the originally proposed 61-node fiber plant with four AML receive locations (circles) superimposed. With fiber rerouting, the receive sites service 7, 11, 12, and 14 fiber hubs, respectively, while the central point is connected to only 13 hubs. Although the fiber maximum distance was, without AML, only 9 miles (14.4 km), the ratio of average fiber route distance with and without AML worked out to be 0.50, which compares fairly well with the 0.46 ratio in Table III. The ratio of maximum fiber length correlated less well: 0.44 in the real system versus 0.37 in the idealized case.

TABLE III
COMPARISON OF IDEALIZED FIBER BACKBONE SYSTEMS

System Parameters	Without AML	With AML IBBT-116
Number of fiber hubs	63	59
Maximum fiber-run distance (unit)	8	3
Average fiber-run distance (unit)	4	1.85
Total fiber distance for one fiber/hub (unit)	252	109
Total fiber distance for four fiber/hub (unit)	1008	436
Total fiber cable distance (unit)	63	59
Max. distance from head-end to AML receive site (unit)		3.6
<u>If unit distance = 2–1/4 km</u>		
C/N of longest 42-channel fiber link (dB)	49.2	52.8
C/CTB of fiber link (dB)	63	65
Combined C/N with AML (dB)	49.2	50.3
Combined C/CTB with AML (dB)	63	63.2
Installed cable cost saving @ $6.8K/mile		$38K
Glass cost savings @ 7¢/foot and four fibers/hub		$297K
Fiber hub savings @ $20K/Tx–Rc pair		$80K
AML IBBT-116 System Cost		<u><$267K></u>
NET SAVINGS		$148K

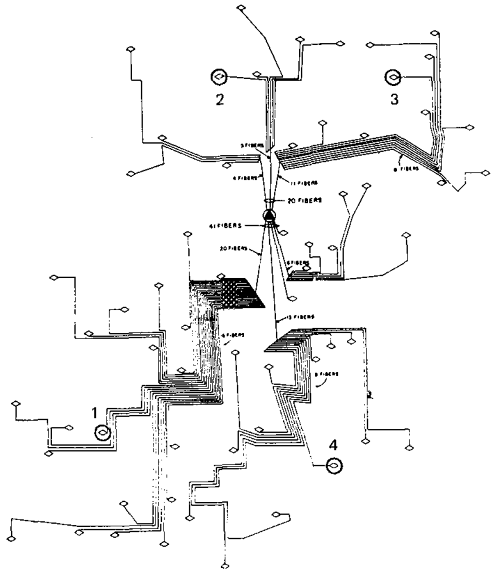


Figure 6 Fiber backbone system[2] with overlay of AML receive sites (fiber layout corresponding to Figure 5b, not shown).

Although there are many similarities between the idealized and real systems, two factors diminish the AML advantage. One is the aforementioned smaller distance. The second factor stems from the Florida location where the rainfall environment is particularly severe. Nevertheless, another possible option in this case serves to illustrate a general point. The central hub site is itself fed from an existing channelized 7.6-mile-distant AML transmitter with parallel 47 dB C/N AM fiber being used to provide a fail soft type of route redundancy to protect against rain fades. With presently unused AML transmitter outputs, additional paths could potentially be implemented to provide signals to one or more of the AML receive sites indicated in Figure 6. Although the cost of possibly upgrading the transmitter must be considered, in many cases the only real cost would be the addition of the receive path(s). In such a case, the economic advantage with AML would be overwhelming.

To achieve the goals[1] of the fiber backbone system with present-day systems, one could construct a system based upon either the AML MTX-132 transmitter or the SSTX-145 transmitter and the above-described reduced channel loading fiber plant. The channelized AML transmitters lend themselves to fiber backbone systems with many more fiber hubs than considered in Figures 5 and 6. The geographic coverage of such systems would extend over large urban and suburban areas. The principal drawback to such systems would be the complexity and cost associated with filtering and multiple laser sources to service each fiber hub.

One could, however, achieve the 50-dB distribution-system goal required by HDTV carriage[12] without reducing the per-fiber channel loading to below 40 channels. For instance, by assigning 58 dB C/N to AML, 52.3 dB to the fiber, and 56 dB to the remaining trunk and distribution, one calculates an overall 50 dB C/N. Table IV summarizes the reach of such a system in a mid-Atlantic (average) rain zone region.

TABLE IV
42-CHANNEL LARGE-AREA FIBER BACKBONE SYSTEM

	MTX-132	SSTX-145
AML transmitter output (dBm)	+9	+16
Number of outputs before splitting	8	8
Assumed antenna diameters (feet)	10	10
Assumed waveguide loss (total transmit and receive) (dB)	4	4
Maximum microwave path length (km)	21.6	30.4
AML system C/N with AGC disabled (dB)	63.3	67.3
AML system C/N (dB)	58	58
Path availability[1] for 54 dB C/N (%) (49 dB total cable distribution system C/N)	99.5	99.6
Path unavailability[1] for 35 dB C/N (hrs/yr)	1	1
Maximum fiber reach for 52.3 dB C/N (km)	9	9
Total equivalent[2] path reach (km)	37	48

[1] Combined rainfall and multipath for CCIR region D2 and 0.25 multipath factor.

[2] Microwave distance multiplied by 1.3 for equivalent fiber distance.

CONCLUSION

A system architecture in which AML drives a fiber backbone system can result in both performance and cost advantages. Generally speaking, the larger the system, the greater the advantage in utilizing AML. However, utilization of recently developed block-conversion type AML equipment can even lead to advantages in modestly sized systems. The tradeoffs are sufficiently complex and employ such a widely ranging set of parameters that each case must be analyzed on its own.

ACKNOWLEDGEMENTS

Thanks go to S. C. Johnson, Senior CATV Project Engineer at ATC, for useful discussions and for providing a detailed map of the proposed fiber backbone system at the Pine Hills, Florida, hub site. The valuable inputs provided by J. Lipson of AT&T are also gratefully acknowledged. Finally, thanks go to L. Kaufman for assistance in calculating AML path reliability.

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INTERPRETATION OF AIRBORNE LEAKAGE DATA

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ABSTRACT

Qualification of cable systems for leakage integrity is being performed by both ground and airborne procedures. A correlation of the airspace and ground measurements was established by the Advisory Committee on Cable Signal Leakage in the late 1970s. This data was taken on a relatively few cable systems and is, therefore, subject to refinement as more data is collected. Some initial observations of recent data provide certain insights into the leakage patterns observed in airborne observations and their sources on the ground. The effects of large single leaks on these patterns, some probable causes, and the implications on ground monitoring procedures are treated in this paper.

You would think that enough had been said about cable signal leakage to last a lifetime. Unfortunately, leakage control probably will last a lifetime and the discussion may never be done. When the FCC rules for qualification of cable systems to the leakage standards become effective in July, 1990 we will only be at the "first hurdle in the race" since qualification must be done yearly and perhaps forever. Leakage is a relatively simple subject on the surface, however, there are many nuances some of which we have yet to learn.

The cable industry has gathered considerable data taken both from ground and airborne measurements. There is a pressing need to investigate correlation between ground and air results. To date, little work has been done toward investigating correlations because of a lack of concurrent ground/airborne data plus the complexity of the situation. Analysis of data taken on a few systems has shown major disparities between ground and airborne results. In these cases the airborne data usually indicates more leakage signal in the airspace than predicted by the groundbased CLI. As a matter of fact, flyover measurements of some systems look very bleak indeed, with large sections of the system showing leakage in excess of the limit of ten microvolts per meter (10uV/m) at 1500 feet above the cable system.

THE FCC REQUIREMENTS AND THEIR IMPACT

In order to better understand the governing factors let us review some of the Part 76 rules and their implications.

Sections 76.605 and 76.611 of the FCC rules require limitation of leakage from any leak to 20 uV/m at a distance of 3 meters (10 feet), while 76.613 prohibits harmful interference regardless of the magnitude of the leak. Calculation of CLI per 76.611(a)(1) requires only that leaks of 50 uV/m or greater be included. Section 76.611(a)(2) perscribes a 10 uV/m total leakage limit at 450 meters (1500 feet) above the cable system.

The implications of these sections relevant to this discussion are:

1. Leaks greater than 20 uV/m at 3 meters are in violation. Smaller leaks are also in violation if they cause harmful interference.

2. In the calculation of CLI, leaks smaller than 50 uV/m need not be included. The reason for this is that these smaller leaks are of minimal significance in the total field.

3. In a flyover measurement the limit of 10 uV/m at 450 meters (1500 feet) above the cable system, could be caused by a single leak. Such a leak measured on the ground at 10 feet, would have to be 150 times larger, (the ratio of 450 meters to 3 meters) or 1500 uV/m in order to equal this threshold.

It is interesting to note that the 1500 uV/m field strength at 10 feet is 75 times the permissible value of 20 uV/m. This is equivalent to 5625 times the power or 37.5 dB excess. You must admit that this is a very wide and generous margin, courtesy of the FCC.

FLYOVER EXPERIENCE

Getting back to actual flyover results, the data from most systems shows at least one area where the 10 uV/m threshold is exceeded. This is in contrast to the ground monitoring and CLI data which usually indicates that the system complies. Although this may seem strange, several factors can contribute to the effect. Remember that the ground measurements for CLI probably required substantial elapsed time, so that it is likely that during the ride-out new leaks developed in the areas which were first measured. In fact, if the CLI was measured and computed and no further monitoring was done prior to a later flyover, the opportunity existed for many new leaks to develop in the interim.

Addressing the fact that most flyovers do show some areas where the leakage exceeds 10 uV/m, the FCC rules have made still another provision in the cable operators favor. This is known as the "90th percentile" and requires that only 90% of the points taken where digital recording is used, show values equal to or less than 10 uV/m. As a result of this provision a cable system showing a few areas of excessive leakage can still qualify. In review, it is fair to say that the FCC regulations are generous and allow for compounded problems without unduly penalizing the cable operator. Nevertheless, a flyover report showing substantial areas of excessive leakage can be very discomforting.

In some flyover reports areas of excessive leakage are seen as circular or eliptical patterns. Figure 1 illustrates this effect. The regularly spaced lines depict the path of the overflight to scale on a latitude/longitude plot. Areas of signal strengths greater than 10uV/m are shown by heavier lines. This plot is simulated because of the inability to reproduce the colors normally used to portray different signal strengths on the flight path.

ANALYSIS OF THE EFFECT

To investigate this effect we will consider the area of excessive leakage generated at 1500 feet above the cable system by a single large leak. Assume that the leak radiates equally in all directions producing a hemispherical pattern (this assumption is unlikely but may be used for this simple example). Figure 2 illustrates this model. An airborne detector directly over the leak (point "A") would receive the maximum energy while in other locations the energy from the leak would have to travel further thereby reducing its effect. For instance, a single leak measuring 15 uV/m at 1500 feet directly above (point "A"), would measure only about 71% of that (about 10 uV/m) if the observer were 1500 feet away from the center (point "B"). This reduction in field strength is governed by the length of the hypotenuse of the triangle formed by the altitude and the radius. The hypotenuse is, in this case, 1.414 times the length of either leg. Signal traveling 1.414 times the distance will produce a received amplitude proportional to 1/1.414 or about 71%. Table 1 has been developed to illustrate the extent of this effect upon the signal strength at 1500 feet from a single large leak. The table records the radii and areas of the circles bounding the region of excessive signal at 1500 feet altitude.

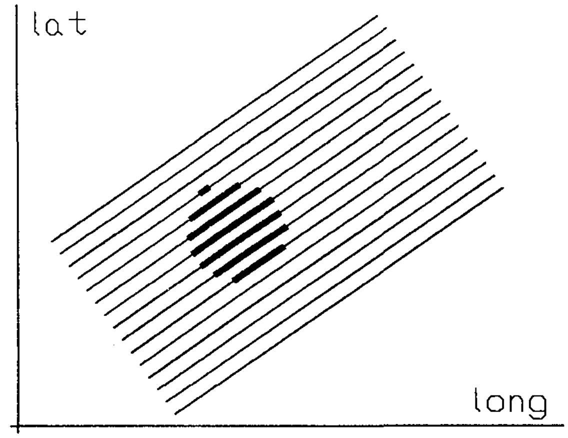

Figure 1

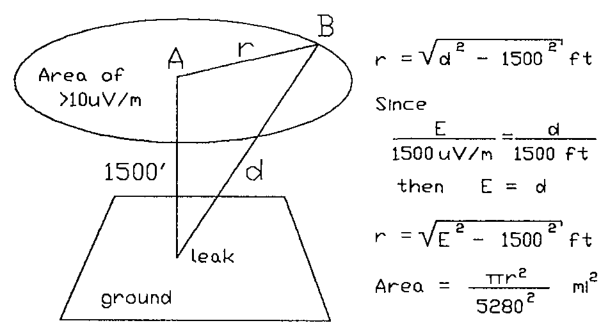

Figure 2

TABLE 1

Magnitude of leak uV/m	Circle of excess signal at 1500' radius (ft.)	area (sq.mi.)
1,500	0	0
2,000	1,323	.197
3,000	2,598	.761
4,000	3,708	1.55
5,000	4,770	2.56
10,000	9,887	11.0
20,000	19,944	44.8

As an example, a single leak of 1500 uV/m will cause only a single point of threshold level directly over the leak while a leak of 3000 uV/m generates a circle of excess single strength with a radius of 2598 feet and encompasses an area of about 0.76 square miles. The dimensions for larger leaks increase rapidly. From the table it can be seen that a single large leak can have a devastating effect on the overall survey results and, as a matter of fact, can be a serious threat to the aircraft navigation and communication circuits which we are trying to protect.

Experience has shown that hot spots observed from the air can often lead directly to the locations of large leaks when they are the sole cause. In very leaky systems where there are many intermediate size leaks the areas of excess signal shown in Figure 3, generally follow the areas of the plant which are in worse shape rather than circular or eliptical shapes illustrated in figure 1.

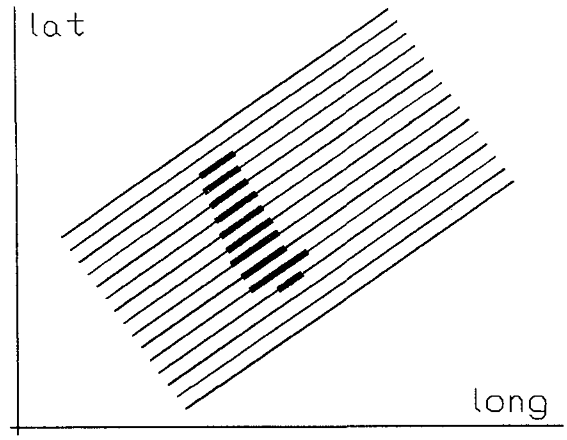

Figure 3

IMPLICATIONS

The implications of all this are a little more subtle. You may ask, "but, where is this large leak which I did not see in my monitoring?" It is possible that it developed since you rode out that area, however, it is also possible that it existed at a fair distance off of the right-of-way so that it was overlooked as you surveyed that area of the plant. Or, perhaps, it is in a high-rise building where you probably could not get close enough to observe it.

The good news is that if you have a relatively clean plant and one of these "blockbusters" has made your aerial survey look far worse than expected, you probably have a simple job to locate and repair. The bad news is that you may have to develop some other monitoring techniques in order to avoid missing these big ones in difficult to access locations.

Is Fiber Optic Cable Fragile?

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ABSTRACT

Communications grade optical fibers are very sensitive to bending, impact and tension. These forces can adversely affect fibers optical performance immediately and can also reduce the expected lifetime through the mechanisms of macro and micro bending. The objective of the cable manufacturer is to "package" those fragile glass fibers in such a way that they will survive the rigors of installation and the installed environment without suffering performance or life expectancy loss. This objective has been far exceeded by today's cable designs and in fact fiber optic cables are "NOT FRAGILE" but are more rugged than coaxial cables used in CATV systems.

TENSILE FORCE

Figure 1

A cable when put under a tensile load will have a tendency to stretch. The plastic material having a elasticity greater then that of glass will increase in length as compared to the glass fibers. The cable design must allow for that stretch without putting stress on the fibers which could, depending on the magnitude of the stress, cause reduced performance or premature failure. A laboratory test which is generally used to prove that a cable design will withstand a given load without damaging the cable is the Electronic Industries Association 455-33A (EIA 455-33A).

The test set-up is shown in Figure 2.

Tensile Testing
EIA 455-33A

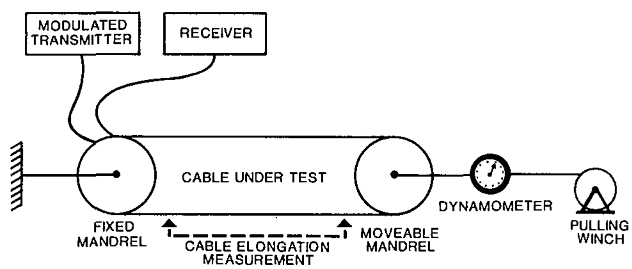


Figure 2

Typical limits for fiber optic cables are 600 pounds force during installation and 250 pounds force installed. Coaxial cables are typically specified at 200 pounds force for .500 inch size and around 400 pounds force for .750 inch size.

Although fiber cables are obviously specified significantly higher than coaxial cables, there is one difference in their installation which may reduce the significance of the difference. That is, fiber cable installed lengths are generally on the order of magnitude of 2-6 km (6,000 to 20,000 ft.) rather than the typical coaxial length of 2000 ft. or less. Obviously then, the weight from the long lengths of cable and the frictional forces developed during installation (whether aerial or duct) can generate greater actual installation tension for fiber cables than for coaxial cables. Under some aerial installations of fiber cable, can generate 600 lbs/f after 1 km of pull. Therefore, when pulling long lengths of fiber cable, tension monitoring is a must.

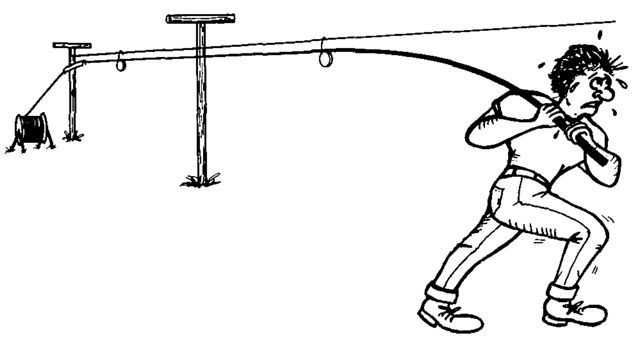


Figure 3

IMPACT AND CRUSH

Between the factory floor and finished installation many unforeseen accidents can happen to the cable. Sufficient ruggedness is built-in to protect the fibers under most conditions.

Figure 4

The Electronic Industries Association 455-25 (EIA 455-25) test apparatus for crush resistance is shown in figure 5. The hydraulic apparatus puts a significant crushing load on a small section of cable under test with no measurable performance degradation.

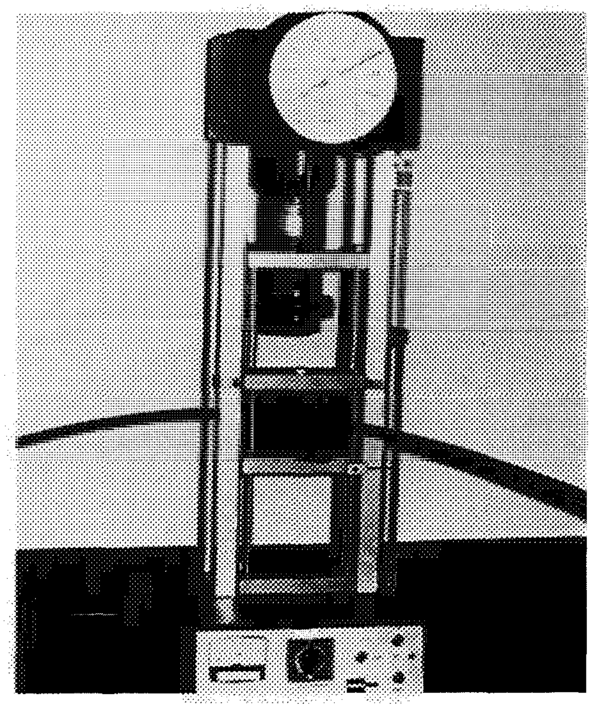


Figure 5

Obviously to people in the CATV installation business impact and crush resistance is coaxial cables weakest link. We all are familiar with dented cable. Fiber cable is very rugged in this respect and with reasonable care during shipping and installation no damage to the fibers should occur.

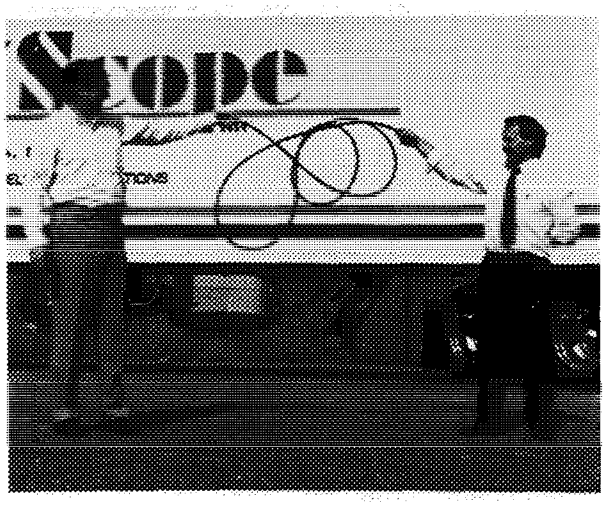


Figure 6

BENDING

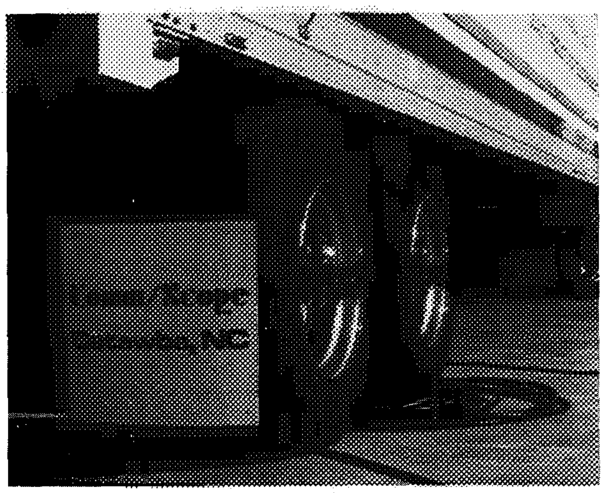


Figure 7

Fiber cables generally act more elastically in the bending mode than do coaxial cable. Therefore they are less susceptible to buckling and kinking. However because of macro and micro bending characteristics the performance and life of the fibers inside the cable can be affected even though no observable deformation to the cable is done. Strict adherence to the minimum bend radius specifications is important

But again fiber cable demonstrates a considerable edge over coax in specified minimum bend radius.

Fiber Optic Cable	7"
1/2" Coaxial	8"
3/4" Coaxial	10"

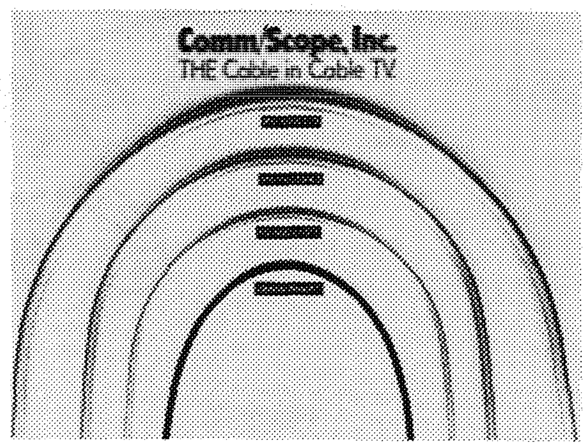
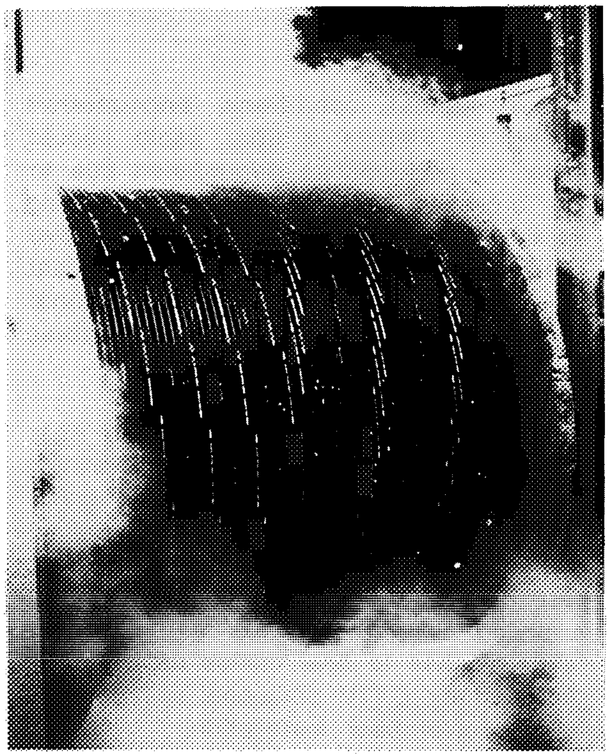


Figure 8

TEMPERATURE

Like all CATV outside plant, fiber optic cable will see extreme environmental changes during its life.



Environmental Test Chamber

Figure 9

Environments include the tropics to the arctic; the mountains to the deserts; swamps; industrial pollution, acid raid, seacoast salt and beneath the streets of major cities. The materials and cable designs chosen must be carefully and extensively tested to assure adequate performance.

The Electronic Industries Association tests that evaluate how the fiber optic cables will react under various environmental conditions can be found as follows:

a) Fiber optic cable bend test at high and low temperature (EIA 455-37).
b) Fiber optic cable twist test (EIA 455-85).
c) Fiber optic cable jacket elongation and tensile strength (EIA 455-89).
d) Fiber optic cable external freezing test (EIA 455-98).
e) Fiber optic cable cyclic testing (EIA 455-104).

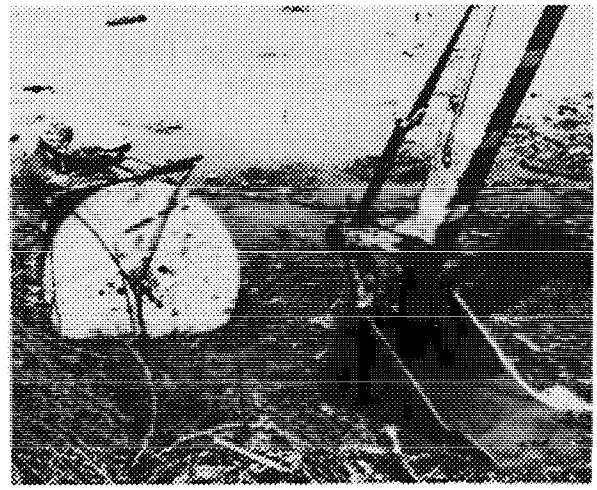


Figure 10

WATER

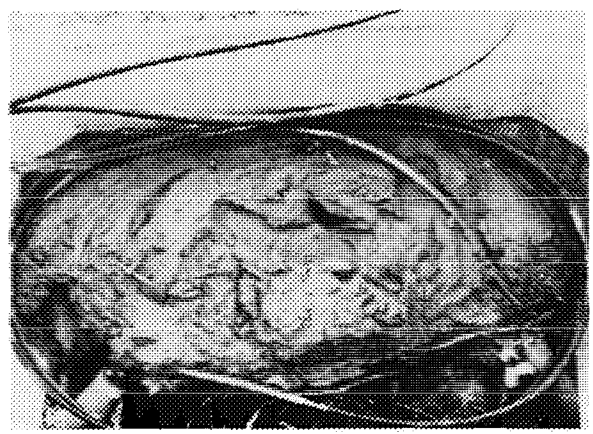


Figure 11

As with any cable water ingress will be detrimental to performance. Optical cables are filled with grease like materials to prevent water entry and there is an Electronic Industries Association 455-82A (EIA 455-82A) test which is performed on samples of finished cable to prove the performance of the filling material.

Fluid Penetration Test

EIA 455-82A

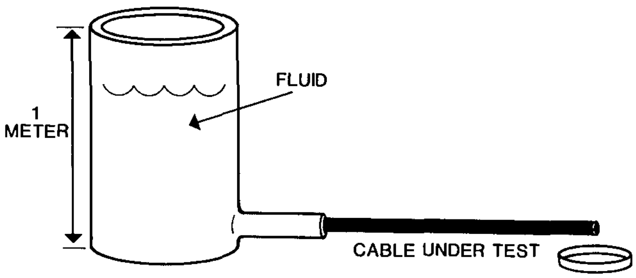


Figure 12

THE BOTTOM LINE

The end result is that cable manufacturers have been very successful at developing cable packages that protect the fragile fibers during manufacturing, shipping, installation and service life. Under typical conditions encountered during its lifetime a fiber cable should show no performance degradation due to the forces we have talked about. As long as the fiber remains inside the protective cable, it is as resistant or more resistant to harmful forces than the traditional coaxial trunk and distribution cables widely used in CATV.

We have intended to demystify fiber optic cable by effectively demonstrating that CATV construction and service crews who have customarily dealt with coaxial cables should unequivocally have no reservations about handling fiber cables. Applying the same rules and common "street" sense to fiber cables as is widely practiced by CATV personnel will produce a successful fiber installation.

LINEARITY CHARACTERISTICS OF DFB LASER DIODES AT HIGH OUTPUT OPTICAL POWERS

Ernest M. Kim, S. Lee Cummings, and Mark E. Tucker

TACAN Corporation

Thomas J. Gibbs

Raynet Corporation

ABSTRACT

Long wavelength Distributed Feedback (DFB) laser diode linearity characteristics under modulation at high output optical powers are investigated. Chirp, emission wavelength, and thermal characteristics have been measured.

It was found that the laser diodes tested had "sweet-spots" of average power which yielded the highest linearity. Second order distortion was reduced dramatically for low temperatures.

INTRODUCTION

State-of-the-art CATV fiber optic systems are, in many cases, limited by the performance of the semiconductor laser diode source. One of the major limitations is the low average launched optical power for the linearity required for high quality multichannel CATV transmission.

In this paper, we report on a study in which the linearity of a 1300 nm distributed feedback (DFB) laser diode was examined as a function of average optical power and temperature. The optical spectra of the DFB laser diode, both with and without modulation, at various operating temperatures and average optical powers were determined.

EXPERIMENTAL SET-UP

The linearity of the laser diode was determined by the two-tone measurement method. In this method, two frequencies of equal radio frequency (RF) power is used to drive the laser diode. The RF power of each frequency is adjusted such that, at the optical output of the laser diode, the optical modulation index per channel (or frequency), OMI/ch, is 0.4. For two tones, the overall OMI is 0.8. This OMI/ch was chosen so that the laser diode could be characterized as having approximately a quasi-linear transfer function. If the total OMI exceeds 0.9, there is increasing deviation from the quasi-linear approximation.

The experimental set-up is shown in figure 1. Two RF frequency sources are used to excite the laser diode. The output of each RF source is passed through a 10 dB attenuator and then amplified using a standard hybrid CATV amplifier with a gain of approximately 17 dB. The outputs of the amplifiers are combined and two frequencies are used to excite the laser diode. The attenuators and amplifiers are used to isolate the source from any reflected signal from the combining process. Without this isolation the combined signal could distort the source frequencies, yielding false laser diode linearity data. The output signal from the combiner was characterized by an RF spectrum analyzer. The second and third order intermodulation distortion was found to be greater than 80 dB below carriers.

The laser diode was driven through a wideband bias-T with a constant current drive. The optical signal from the laser diode was measured with a Hewlett-Packard lightwave signal analyzer. The analyzer determined was used to determine the OMI/ch and the linearity of the laser diode.

The temperature was controlled by driving the thermo-electric cooler packaged in the laser diode module to the desired operating temperature. Measurement of the operating temperature was performed by the in-package thermistor.

Second and third order intermodulation distortion ratios (IMD2 and IMD3, respectively in absolute dB from carrier) were measured using the two-tone method for the following two pairs of frequencies:

199.25 MHz & 205.25 MHz
535.25 MHz & 541.25 MHz.

The measurements were taken at 0, +2.7, and +5 dBm average optical power. Each measurement was made at both 0 and 25 degrees Celsius.

After each distortion test, the optical spectrum of the laser diode was recorded. The experimental setup is identical to that of figure one with the exception that an optical spectrum analyzer is used instead of the lightwave signal analyzer.

EXPERIMENTAL RESULTS

The results of the two-tone tests for 0 and 25 degrees Celsius are presented for the low and high frequency pairs in figures 2 and 3, respectively. At 25 degrees, the best performance in terms of IMD3 was for and average optical power of +2.7 dBm. At +25 degrees and +2.7 dBm optical the IMD3 for the lower frequency pairs are 73 dB, and 68 dB for the upper frequency pairs. IMD2 was 38 dB for the lower tones and 38 for the upper tones. At 0 dBm optical, the IMD2 was better by 10 dB than that for the +2.7 dBm case. However, at 0 dBm optical the IMD3 was degraded by 14 dB. Curiously, at +5 dBm optical the distortion, as a function of frequency, was similar to that at +2.7 dBm.

At 0 degrees Celsius and 2.7 dBm optical, we find an increase in IMD2 of 10 dB and a slight decrease in IMD3. However, the distortion is still relatively constant over frequency. Significant change in either IMD2 or IMD3 is not evident at 0 and +5 dBm from the 25 degree Celsius operation.

The optical spectra for 0, +2.7, and +5 dBm operation with modulation with two tones (at 0.4 OMI/ch) at 25 degrees Celsius are shown in figures 4a, b, and c respectively. Note that there is some spectral broadening and better definition of the chirp. The amount of spectral broadening at the high optical powers will not significantly affect the propagation characteristics through the optical fiber. The amount of chirp exhibited was approximately 0.4 nm at 35 dB from the peak emission wavelength. As such the amount of chirp is not of any significance. In each case, there was a broadening of the optical spectrum and a slight increase in chirp amplitude when the laser diode was modulated.

What is interesting is that when the laser diode is operating at 25 degrees Celsius and +2.7 dBm optical and is driven with 40 channels (figure 4d), its optical spectrum is nearly identical to that of the two channel case at +5 dBm optical. This indicates that under large channel count modulation, there is increased spectral broadening and chirp.

Similar results were evident at 0 degrees Celsius. As expected, the spectral widths were slightly narrower than for the 25 degree condition.

SUMMARY

Distortion measurements using the two-tone method was performed on a state-of-the-art long 1300 nm DFB laser diode. Because of the high linearity and low intrinsic noise characteristic of this class of laser diodes, they have become an attractive source for high quality CATV fiber optic transmission systems.

One of the limiting characteristics of the CATV fiber optic systems is the low launched optical powers required for the high linearity desired. In this paper, we have reported on the performance of a DFB laser diode at varying average optical powers and temperatures. The results indicate that there is a "sweet-spot" for maximum linearity, regardless of temperature. At that operational optical power, the distortion was relatively flat over frequency. Additionally, second order distortion was dramatically reduced when operated at 0 degrees Celsius at the "sweet-spot." The average optical power for optimum performance was +2.7 dBm.

The optical spectrum did not vary significantly for varying average optical powers or temperature. However, we observed that there was slightly increased broadening for higher operational temperatures, higher average optical powers and large number of channels.

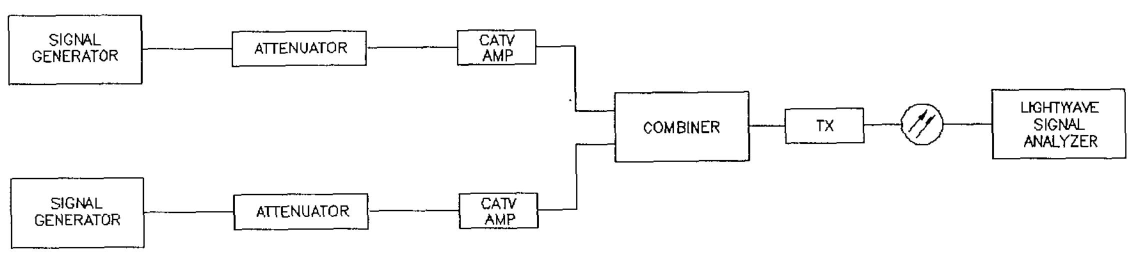


FIGURE 1:
TWO-TONE LASER DIODE MEASUREMENT SETUP

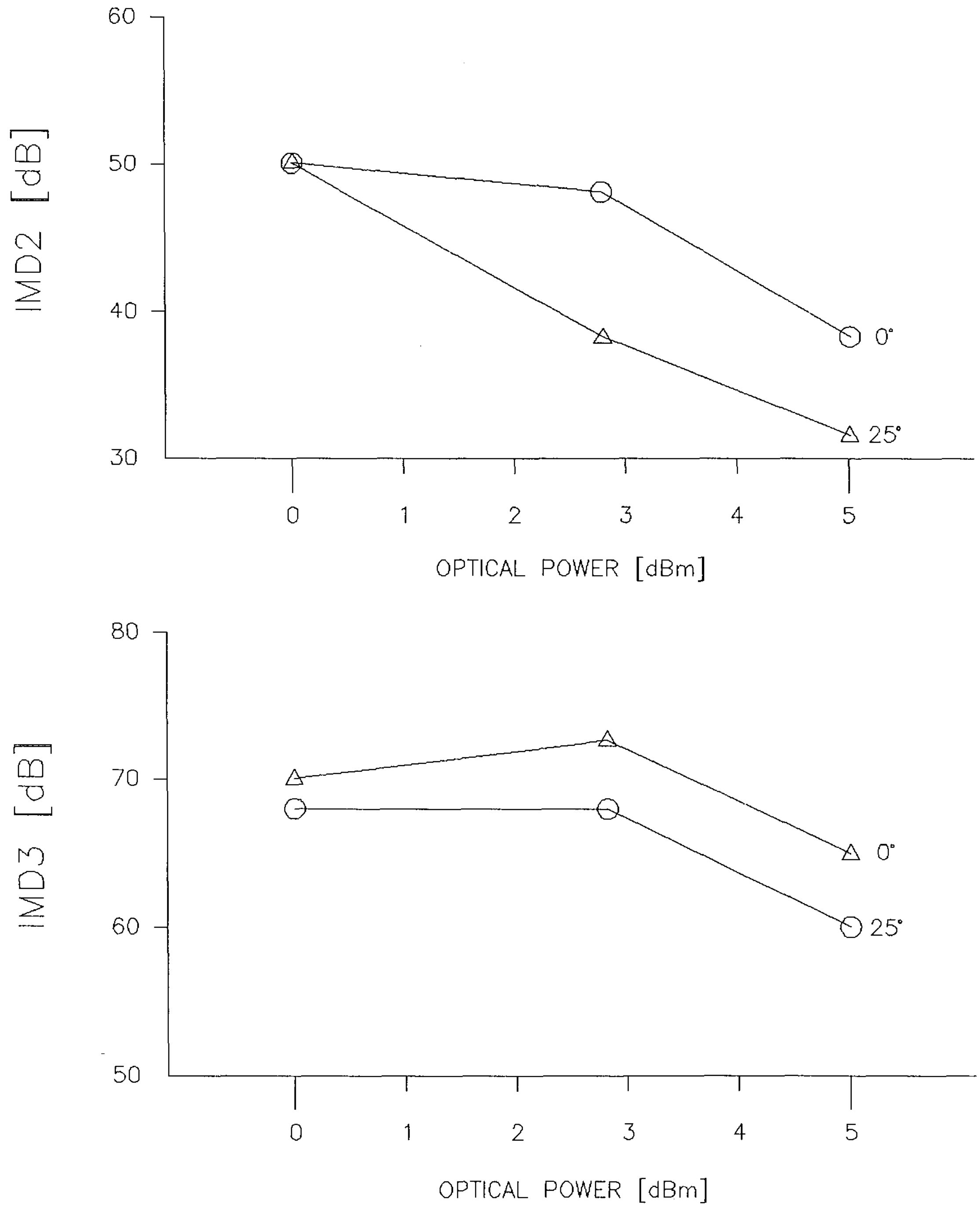


FIG 2:
IMD2 AND IMD3 FOR THE LOWER FREQUENCY PAIR

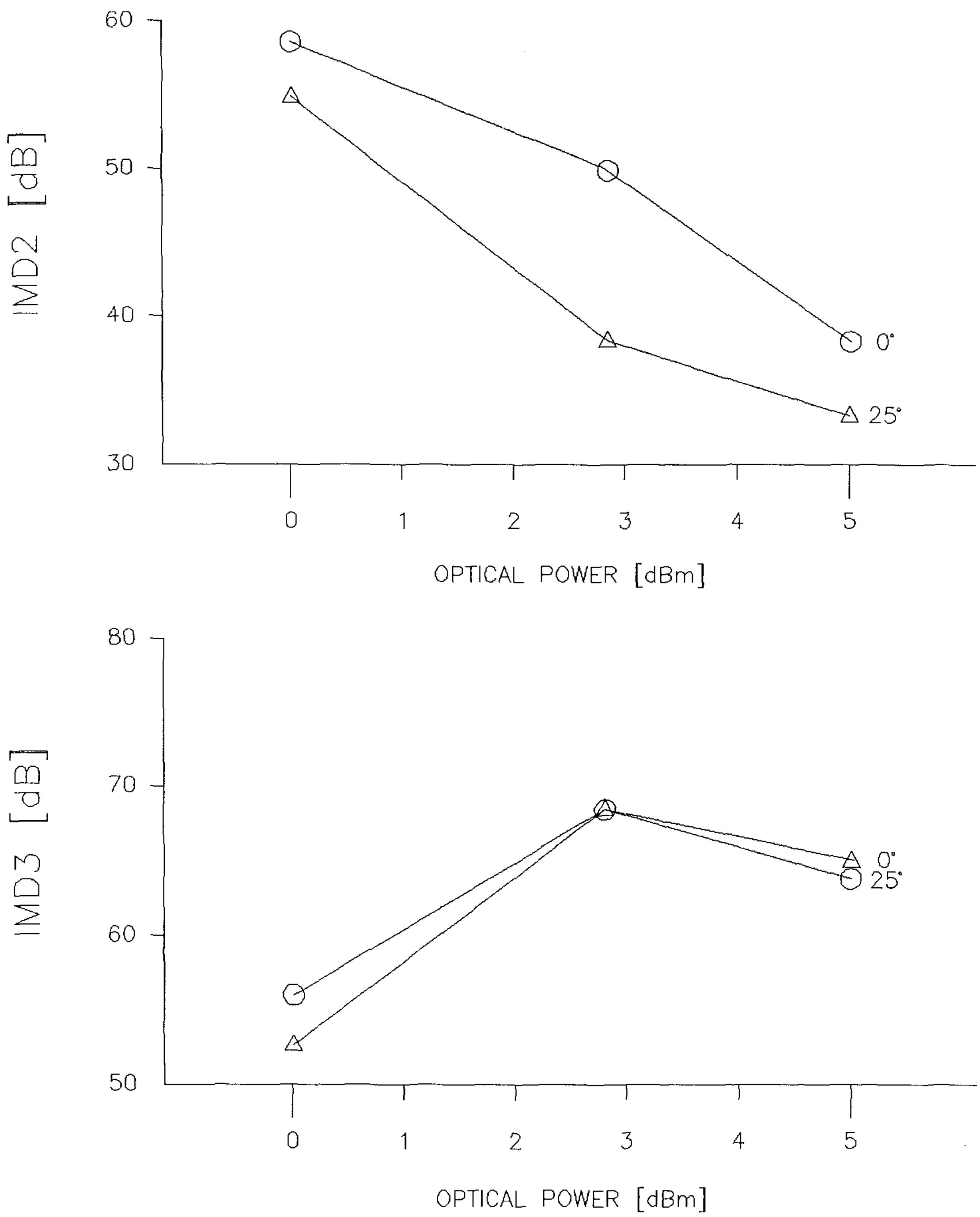
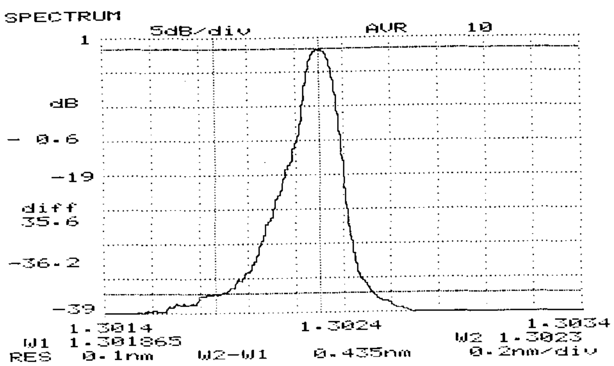
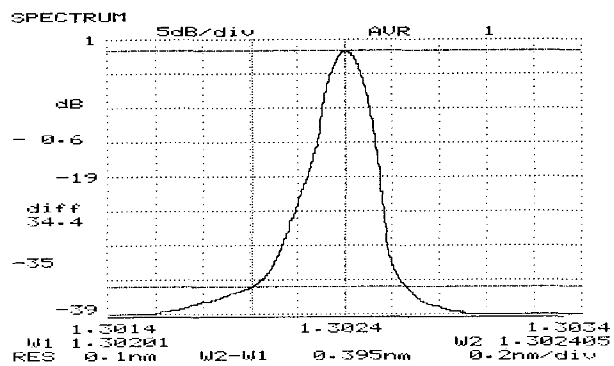


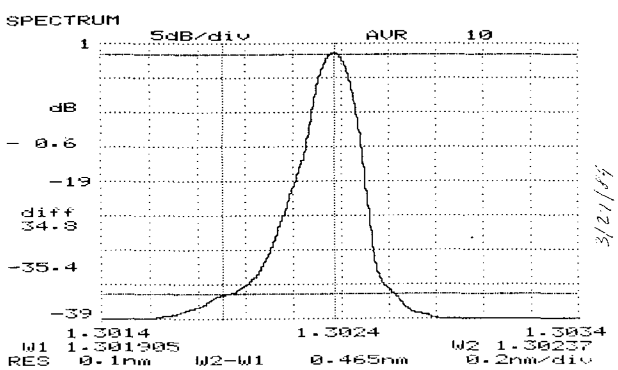
FIG 3:
IMD2 AND IMD3 FOR THE UPPER FREQUENCY PAIR



(a)

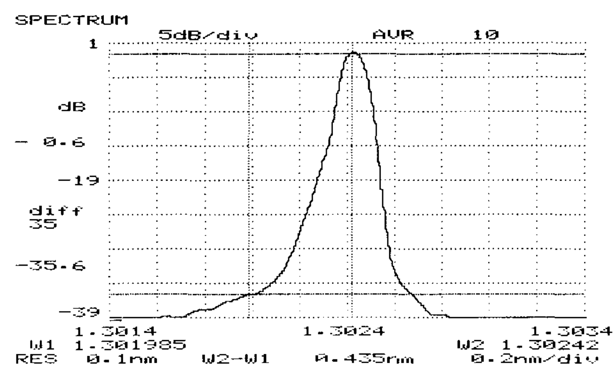


(b)



3/21/89

(c)



(d)

FIGURE 4
Optical Spectra For A DFB Laser Diode Operating At 20 Degrees C Under Two-tone modulation at: (a) 0dBm, (b), +2.7 dBm, (c) +5dBm, and (d) 40 channel modulation at +2.7 dBm.

METHOD FOR INCLUDING CTBR, CSO, AND CHANNEL ADDITION COEFFICIENT IN MULTICHANNEL AM FIBER OPTIC SYSTEM MODELS

Ernest M. Kim, Mark E. Tucker, and S. Lee Cummings

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ABSTRACT

A model of multichannel fiber optic AM CATV links is presented. The analysis yields the worst case Carrier-to-Noise Ratios (CNRs) as a function of average received optical power for desired Composite Triple-Beat Ratios (CTBR) and Composite Second Order (CSO) ratios. A method for determining the required Optical Modulation Index (OMI) per channel for desired CTBRs and CSOs, and number of transmitted channels is included.

The overall OMI is related to the per-channel OMI for given numbers of channels.

INTRODUCTION

Publications to date on multichannel AM fiber optic system models have concentrated on yielding the CNRs as a function of average received optical power without explicitly including OMI related distortion or the appropriate channel power addition coefficients [1, 2, 3]. In this paper, we propose a model in which the laser diode OMI is specifically related to the desired CSO and CTBR.

The per-channel OMI, (OMI/ch), to distortion relationship is derived from two figures of merit for laser diodes. The new parameters are the Optical Second Order Intercept Point (OIP2) and Optical Third Order Intercept Point (OIP3). Similar to radio frequency (RF) amplifier intercept points [4], multichannel composite distortions for a given channel input power at a specific average launched optical power, may be estimated using the simple two-tone measurement method of characterizing the laser diode optical output distortion. The relationship between the two-tone measurements and desired CTBRs and CSOs for given numbers of channels is presented.

Using OIP2 and OIP3, OMI/ch can be determined. Then, the CNR as a function of average received optical powers can be computed by using the receiver noise, shot noise, photodetector responsivity, and channel bandwidth for desired CTBRs and CSOs. The overall (total) statistical OMI, OMIt, can be calculated to confirm that the quasi-linear assumptions of this model are not violated. Comparisons of the projected performance through the use of the model to experimental results at four different channel loadings are included. The comparison showed excellent correlation.

For multichannel AM CATV systems OMIt is related to OMI/ch by [1]

$$\text{OMI/ch} = \text{OMIt}/N^{\zeta}, \qquad (1)$$

where N is the number of channels and ζ is the multichannel addition coefficient used to combine the multiple carriers. A simple graph and chart is presented which allows AM CATV fiber optic designers to find the appropriate ζ factor.

MODEL

There are several steps included in the model. They are:

1. Determine the number and type of distortion products falling into the frequency band of interest. Determine "penalty" numbers.

2. Determine the fiber optic receiver noise sources. The required parameters are: amplifier noise figure in dB (Ft), effective receiver load impedance in (R_L), photodiode dark current in amperes (Id), operating temperature in Kelvins (T), and photodetector responsivity in amperes/watt (η).

3. Using the two-tone test, determine the third order intermodulation distortion IMD3 and second order intermodulation distortion IMD2 of the laser diode at the desired average launched optical power.

Using data from the first three steps, the projected performance can be calculated.

Distortion Products

The determination of the number of intermodulation distortion products falling into the channels of interest is a critical step in the model. All calculations are performed with unmodulated channels, consistent with the measurement procedure used when MATRIX type systems are utilized.

To see the effect of a multichannel load, it is instructive to consider an input of three sinusoids of frequencies , and α,β and γ because the form of all modulation products can be found from a three-frequency input. If we assume a quasi-linear transfer function, the output eout is a function of the input ein,

$$\text{eout} = a_1\text{ein} + a_2\text{ein}^2 + a_3\text{ein}^3 \qquad (2)$$

where ein is made up of the three sinusoids. The resulting eout is presented in Table 1 for the three frequencies at three phases [5]. The output then contains modulation products at all possible sums and differences of all multiples of the input frequencies, up to order three. The products of interest are;

$$\begin{array}{l} \alpha - \beta, \\ \alpha + \beta, \\ \alpha - 2\beta, \\ 2\alpha - \beta, \\ 2\alpha + \beta, \\ \alpha - \beta - \gamma, \\ \alpha + \beta - \gamma, \\ \text{and } \alpha + \beta + \gamma. \end{array} \quad (3)$$

CSO is primarily caused by the second order sum beats [6]. The reason being that the permissible limits for interfering signals in relation to visual carriers indicate that the sum beats (which falls at fc + 1.25 MHz, where fc is a visual carrier frequency), the carrier-to-beat ratio is approximately 52 dB. For the second order difference beats (which falls at fc - 1.25 MHz), the visually limiting carrier-to-beat ratio is approximately 30 dB. Therefore, we calculate the CSO from the sum beats since the difference beats will not have a significant effect on the picture quality.

Although the calculation of the beats is complex, a simple graphical method can be employed using figure 1. The graphs show normalized numbers of products of a given type as a function of normalized channels. The number of products outside the limits of the curves is zero.

In Figure 1,

U =
Total possible products of a given type.

N =
Number of channels transmitted with carriers n_1f to n_2f inclusive, where f is the base frequency in Hz (6 MHz in CATV) and n_1 and n_2 are integers $n_2 > n_1$.

k =
Channel of interest associated with carrier kf_o within the fundamental band $n_1 < k < n_2$.

M =
Channel of interest associated with carrier Mf_o where;
$M = k - n_1 + 1$.

To clarify the above, a 40 channel example is given. For simplicity, we are assuming consecutive channel loading (i.e. no FM radio channels).

$f = 6$ [MHz]
$f_1 = 55.25$ [MHz]
$f_2 = 289.25$ [MHz]

$n_1 = \text{integer\_truncate}(f_1/f_o)$
$= 9$

$n_2 = \text{integer\_round up}(f_2/f_o)$
$= 49$

$N = n_2 - n_1$
$= 40$

$k = f_1/f_o, f_1/f_o + f_o, \dots, f_2/f_o$
$= 55.25/289.25, 55.25/289.25 + 6, \dots, 289.25/6$

Therefore, Channel 2 at 55.25 MHz is designated as

$M = 55.25/6 - 9 + 1$
$= 1.208$

Referring to figure 1 at M = 1.208 (Channel 2), for the $2\alpha - \beta$ distortion products we see that at M/N (1.208/40) is 0.03. Using the graph, we find that U/N is 0.5. Knowing that the number of channels, N, is 40, U is found to be 20; corresponding to 20 intermodulation distortion of this type falling into Channel 2.

Table 2 provides the maximum number of beats of each kind observed over the total channel capacity of interest. The calculations are for consecutive channels without dead bands. If there is an FM band, the beats are worse than experimentally observed for low channel counts (up to about 20) and approximately correct for 30 channels and up.

The intermodulation "penalties" are correction factors to desired CTBRs or CSOs used to determine OMI/ch after finding the laser diode OIP3 and OIP2. They are dependent only on the channel loading and products found from calculations using figure 1. The second order penalty correction factor uses the number of sum beats,

$$P2 = 10\log[U(\alpha + \beta)]. \quad (4)$$

The third order penalties are more complex. The penalties must be made in terms of the triple beat components. From Table 1, we observe that the triple beat products are twice the amplitude of the other non-harmonic third order intermodulation distortion. Knowing that the triple beats are twice the amplitude of the other third order intermodulation products, the penalty P3 is,

$$P3 = 10\log\{1^2[U(2\alpha + \beta) + U(\alpha - 2\beta)] + 2^2U(\alpha + \beta + \alpha)\}. \quad (5)$$

Laser Two-Tone Test

The distortion characteristics of the optical source must then be quantified. The laser diode of interest is tested with the two-tone method at the desired average launched optical power. The measurement set-up is shown in figure 2. Each tone is set at 0.4 OMI/ch, where Pmod is the optical modulation of each carrier, Pav is the average optical power, and

$$OMI/ch = Pmod/Pav. \quad (6)$$

The measurement is made over the entire frequency range of interest. A typical laser diode two-tone measurement result at 2.73 dBm optical is shown in Table 3.

The second order (a) and third order distortion (b) ratios, in dB, measured from the two-tone test are used to determine OIP2 and OIP3. Since the laser diode at total modulation indices less than 0.8 follow approximately (less than 10% deviation) the polynomial rule for quasi-linear systems, RF intercept point concepts can be used. The intercept points are,

$$OIP2 = rms(OMI/ch) + a,$$

$$\text{and } OIP3 = rms(OMI/ch) + b/2. \quad (7)$$

The relationship between CTBR and CSO to OIP2 and OIP3 for given channel loading are defined as,

$$CTBR = b - P3, \quad (8)$$

$$\text{and } CSO = a - P2.$$

Using equations 7 and 8, the required OMI/ch for given numbers of channels can be calculated for desired CTBRs and CSOs. The expressions are,

$$\text{rms(OMI/ch)} = \text{OIP2} - (\text{CSO} + \text{P2}), \qquad (9)$$

$$\text{and rms(OMI/ch)} = \text{OIP3} - (\text{CTBR} + \text{P3})/2.$$

As an example, for the characteristics of the Distributed Feedback (DFB) laser diode, with integrated optical isolator shown in Table 3, with 40 consecutive channel loading,

OIP2 = 39 dBm,

OIP3 = 19 dBm.

From table 1 the penalties for CSO = 60 dB and CTBR = 65 dB found by computing the number and types of distortion products, using equations 4 and 5 are;

P2 = 10 dB,

P3 = 33.9 dB.

The peak OMI/ch can then be calculated from equation 9,

OMI/CH = 0.04.

Channel Addition Coefficient

Knowing the OMI/ch, it is sometimes useful to determine the overall OMI (OMIt) to make certain that we do not exceed 100% modulation. If we approach 100% OMIt, the quasi-linear assumptions do not hold.

The relationship of OMI/ch and OMIt is dependent on the number of channels and the channel addition coefficient ζ. ζ can be determined either experimentally or through statistical analysis. We chose to experimentally determine ζ.

The measurements indicated that, as expected, for low channel counts (e.g. 2), ζ is 1. For large channel counts, ζ approaches 0.5. The later condition is approached as a result of the averaging effect produced by a large number of subcarriers with random phases.

The results are shown in figure 3 and presented in tabular form in Table 4. Using Table 4 (or figure 3) and equation 1, a solution to OMIt can be found. To maintain the integrity of the model, OMIt must be less than 1 (preferably less than 0.9). If OMIt is greater than or equal to 1, the assumption of a quasi-linear system is violated, and the model is invalid. If the computation yields an OMIt of greater than 1, the OMI/ch must be reduced such that OMIt is within the bounds required for the quasi-linear assumption of the model.

Model Development

Many authors have developed equations for analog fiber optic systems to determine CNR. A concise equation for PIN photodetector receiver systems is given by Koscinski [2] in linear CNR:

$$\text{CNR} = 1/2\, \frac{(\text{OMI/CH})^2\, \eta^2 .\, P_{AV}^2}{[(\text{RIN})\eta^2 . P_{AV}^2 B) + 2q(\eta P_{AV} + I_d)B + (4kTB/R_L)Ft} \qquad (10)$$

$$\text{LET Ns} = \text{RIN}\eta^2 P_{AV}^2 B + 2q(\eta + P_{AV} I_d)B + (4kTB/R_L)Ft$$

where
q = electron charge[C]
k = Boltzmann's Constant [J/K]
B = bandwidth of channel [4 MHz]

RIN = laser relative intensity noise [dB/Hz]
= 148 dB/Hz for the laser diode of Table 2
RL = 470 Ω, I_d = 0.5 nA, η = 0.75 A/W, T = 290K, Ft = 4dB

Substituting the expressions for rms(OMI/ch) we arrive at;

$$\text{CNR [dB]} = \text{OIP3} - \frac{(\text{CTBR} + \text{P3})}{2} + 20\log[\eta P_{AV}] - 10\log\text{Ns} \qquad (11)$$

for the desired CSO for N channels, and;

$$\text{CNR[dB]} = \text{OIP2} - (\text{CSO} + \text{P2}) + 20\log[\eta P_{AV}] - 10\log\text{Ns} \qquad (12)$$

for the desired CTBR for N channels.

The CNR as a function of average received optical power is shown in figure 4a for desired CTBRs and figure 4b for desired CSOs for 40 channel loading. For a CTBR of 65 and CSO of 60, the required OMI/ch is approximately 0.04.

Note that the linearity of the receiver is not included. The receiver in question was tested with the two-laser, two-tone measurement and exhibited acceptably high linearity and wide bandwidth.

COMPARISON TO EXPERIMENTAL DATA

A comparison between the model and experimental data was made with the laser diode and receiver exhibiting the behavior above. The measurements were taken with a MATRIX Multiple Frequency Signal Generator and R-75 Signal Analyzer. The output from the receiver/AGC was set at +30 dBmV +/-1 dB over the channels of interest.

Comparisons were made for four channel loadings: 10, 20, 30, and 40 channels. In the first case, a comparison was made for consecutive channel loading from Channel 14. There were no second order products for the 10 and 20 channel case. The second case was that in which the model was run for consecutive channel loading from channel 2 in 6 MHz increments without a dead band for the FM channels. The experiment, however, did include an unused FM band. The results are shown in Table 5 exhibiting the differences between the model and experimental CNR, CTBR, and CSO. The + in the CTB and CSO columns indicate higher ratios found in experimental results than that of the model.

CNR difference between the model and the experiment agreed to within 2 dB. The CNR difference was equal to the expected CNR calculated by the model for 65 dB CTBR and 60 dB CSO to that of the experimental results. For the most part, the CTBR differences were within 2 dB. Notable exceptions were for the 30 channel consecutive from Channel 14 case. At -2 dBm average received optical power, the CTBR differed by as much as 3.8 dB and CSO by as much as 3.0 even though the CNR results were excellent. Another notable deviation were the CSO differences for the 20 and 30 channel cases for consecutive loading from Channel 2. The large deviations are indicative of beat stacking at the higher frequencies not accounted for in the model beyond the calculated maximum carrier.

CONCLUSIONS

Fiber optic multichannel AM CATV links are being developed and deployed in increasing numbers. Enhancements of the analytical tools which will aid in the design of fiber optic AM CATV systems is becoming ever more important.

In this paper, a model was presented which yields the CNR as a function of average received optical power, and desired CTBR and CSO. By characterizing a semiconductor laser diode using the two-tone method, projected multichannel distortion performance can be calculated. Those calculations involve the determination of the numbers and types of intermodulation distortion products, and the computation of intermodulation "penalties". The "penalties", P2 and P3, are used in conjunction with the optical intercept points, OIP2 and OIP3, found from two-tone measurements of the laser diode and the desired CTBR and CSO to find the appropriate OMI/ch. The resulting OMI/ch is used in conjunction with the intrinsic noise of the laser diode (RIN) and receiver parameters to calculate the expected CNR as a function of average received power at desired CTBR and CSO under specific channel loading.

The channel addition coefficient as a function of the number of channels has been tabulated. OMIt can then be determined from the OMI/ch. By limiting the OMIt to 0.9 or less (and the corresponding OMI/ch), the quasi-linear assumptions of this model can be maintained.

The results of the model was compared to experimental data under various channel loading conditions. The comparison produced favorable correspondence.

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TABLE 1

Terms in $e_{out} = a_1 e_{in} + a_2 e_{in}^2 + a_3 e_{in}^3$ for $e_{in} = A\cos(\alpha t + \phi_1) + B\cos(\beta t + \phi_2) + C\cos(\gamma t + \phi_3)$

dc

$$\frac{1}{2}a_2(A^2 + B^2 + C^2)$$

First order

$$\begin{aligned}
&a_1 A\cos(\alpha t + \phi_1) + a_1 B\cos(\beta t + \phi_2) + a_1 C\cos(\gamma t + \phi_3)\\
&+ \frac{3}{4}a_3 A(A^2 + 2B^2 + 2C^2)\cos(\alpha t + \phi_1)\\
&+ \frac{3}{4}a_3 B(B^2 + 2C^2 + 2A^2)\cos(\beta t + \phi_2)\\
&+ \frac{3}{4}a_3 C(C^2 + 2A^2 + 2B^2)\cos(\gamma t + \phi_3)
\end{aligned}$$

Second order

$$\begin{aligned}
&\frac{1}{2}a_2\left[A^2\cos(2\alpha t + 2\phi_1) + B^2\cos(2\beta t + 2\phi_2) + C^2\cos(2\gamma t + 2\phi_3\right]\\
&+ a_2 AB\left\{\cos\left[(\alpha + \beta)t + \phi_1 + \phi_2\right] + \cos\left[(\alpha - \beta)t + \phi_1 - \phi_2\right]\right\}\\
&+ a_2 BC\left\{\cos\left[(\beta + \gamma)t + \phi_2 + \phi_3\right] + \cos\left[(\beta - \gamma)t + \phi_2 - \phi_3\right]\right\}\\
&+ a_2 AC\left\{\cos\left[(\alpha + \gamma)t + \phi_1 + \phi_3\right] + \cos\left[(\alpha - \gamma)t + \phi_1 - \phi_3\right]\right\}
\end{aligned}$$

Third order

$$\frac{1}{4}a_3\left[A^3\cos(3\alpha t + 3\phi_1) + B^3\cos(3\beta t + 3\phi_2) + C^3\cos(3\gamma t + 3\phi_3)\right]$$

$$+ \frac{3}{4}a_3\left\{\begin{aligned}
&A^2B\left\{\cos\left[(2\alpha + \beta)t + 2\phi_1 + \phi_2\right] + \cos\left[(2\alpha - \beta)t + 2\phi_1 - \phi_2\right]\right\}\\
&+ A^2C\left\{\cos\left[(2\alpha + \gamma)t + 2\phi_1 + \phi_3\right] + \cos\left[(2\alpha - \gamma)t + 2\phi_1 - \phi_3\right]\right\}\\
&+ B^2A\left\{\cos\left[(2\beta + \alpha)t + 2\phi_2 + \phi_1\right] + \cos\left[(2\beta - \alpha)t + 2\phi_2 - \phi_1\right]\right\}\\
&+ B^2C\left\{\cos\left[(2\beta + \gamma)t + 2\phi_2 + \phi_3\right] + \cos\left[(2\beta - \gamma)t + 2\phi_2 - \phi_3\right]\right\}\\
&+ C^2A\left\{\cos\left[(2\gamma + \alpha)t + 2\phi_3 + \phi_1\right] + \cos\left[(2\gamma - \alpha)t + 2\phi_3 - \phi_1\right]\right\}\\
&+ C^2B\left\{\cos\left[(2\gamma + \beta)t + 2\phi_3 + \phi_2\right] + \cos\left[(2\gamma - \beta)t + 2\phi_3 - \phi_2\right]\right\}
\end{aligned}\right\}$$

$$\begin{aligned}
+ \frac{3}{2}a_3 ABC\Big\{&\cos\left[(\alpha + \beta + \gamma)t + \phi_1 + \phi_2 + \phi_3)\right] + \cos\left[(\alpha + \beta - \gamma)t + \phi_1 + \phi_2 - \phi_3\right]\\
&+ \cos\left[(\alpha - \beta + \gamma)t + \phi_1 - \phi_2 + \phi_3\right]\\
&+ \cos\left[(\alpha - \beta - \gamma)t + \phi_1 - \phi_2 - \phi_3)\right]\Big\}
\end{aligned}$$

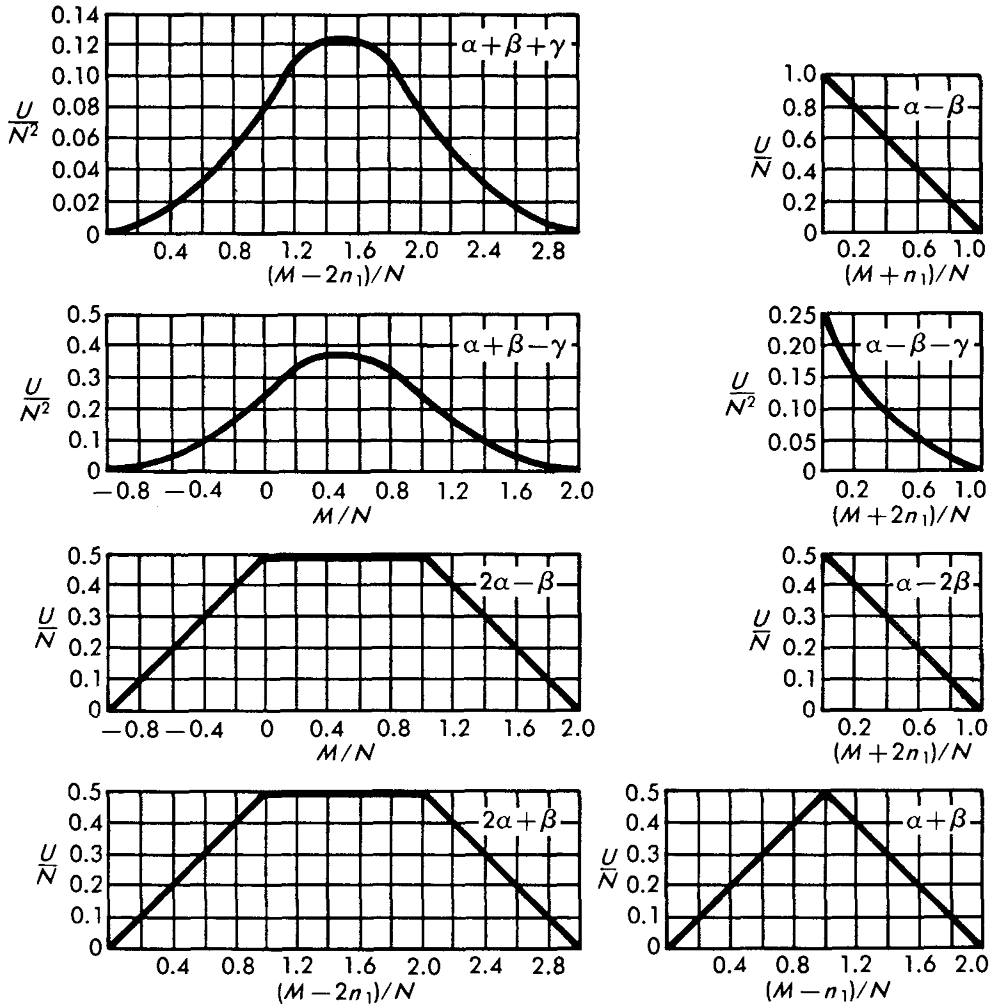


Figure 1 Spectral Distribution of Various Types of Imtermodulation Products

INTERMOD. TYPE	No. CONSECUTIVE CHANNELS FROM Ch. 2: 10	20	30	40
$\alpha-\beta$	0	9.8	19.7	29.8
$\alpha+\beta$	0.6	5.6	10.6	15.6
$\alpha-2\beta$	0.2	5.2	10.2	15.2
$2\alpha-\beta$	5	10	15	20
$2\alpha+\beta$	5	10	15	20
$\alpha-\beta-\gamma$	0	2.3	56.6	158.4
$\alpha-\beta-\gamma$	37.5	150	337.5	600
$\alpha+\beta+\gamma$	0	1.2	11.3	35.8

TABLE 2 Number of Intermodulation Distortion Products

Fundamental Freq's in MHZ	IMD3 [dB] $2f_1$-f_2	IMD3 [dB] $2f_2$-f_1	IMD2 [dB] f_2-f_1	IMD2 [dB] f_2+f_1
55.25 61.25	58	58	49	49
199.25 205.25	60	60	50	45
301.25 307.25	62	62	50	39
445.25 451.25	63	63	49	31

TABLE 3 Typical Laser-Diode Two-tone Measurement Result At +2.73 dBm Optical Power And 0.4 OMI/ch

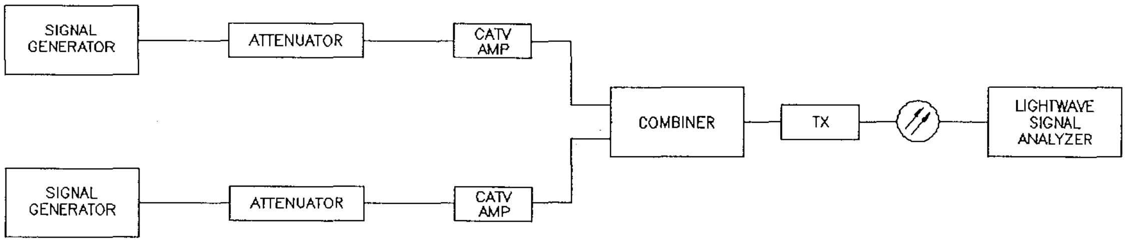


FIGURE 2:
TWO-TONE LASER DIODE MEASUREMENT SETUP

NUMBER OF CHANNELS	ζ
10	0.7
20	0.67
30	0.62
40	0.59
50	0.57
60	0.55
70	0.54
80	0.53

TABLE 4 Channel Addition Coefficient, ζ, for Given Channel Loading

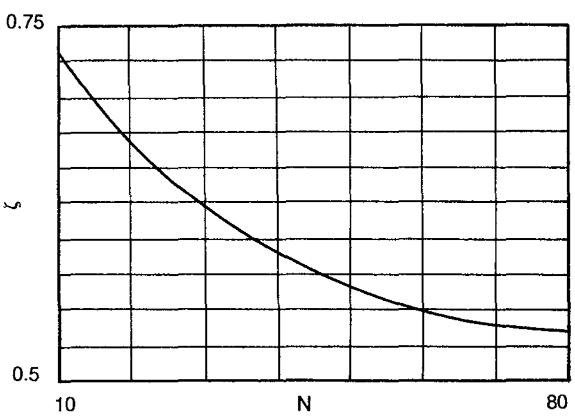
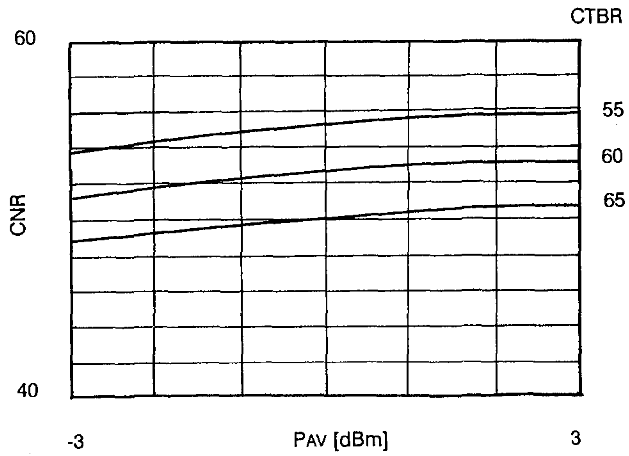


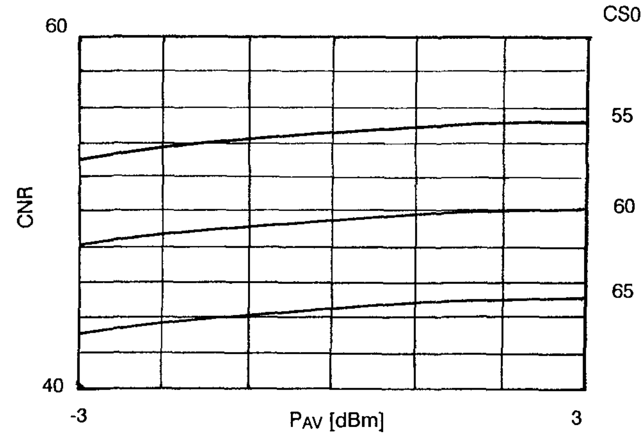
Figure 3 Channel Addition Coefficient ζ for 10 to 80 Channels

No. of Channels	Avg. Opt. Pwr [dBm]	ΔCNR	ΔCSO[dB]	ΔCTBR[dB]
10	-1	1	N/A	+1.6
From Ch14	-2	<1	N/A	+1.4
20	-1	1	N/A	-1.1
From Ch14	-2	1	N/A	+1.1
30	-1	1	+1.1	-1.4
From Ch14	-2	<1	+3.0	-3.8
40	-1	1	+1.1	-0.4
From Ch14	-2	1	+1.1	-1.7
10	-1	1	N/A	+1.6
From Ch 2	-2	<1	+4.7	+2.4
20	-1	2	-1.6	-0.2
From Ch2	-2	<1	+4.7	+2.4
30	-1	2	+2.7	-0.8
From Ch2	-2	1	-4.7	+2.4
40	-1	1	-1	-2
From Ch2	-2	1	+2.3	-0.3

TABLE 5 Comparison of the Model Results to Experimental Data



(a)



(b)

Figure 4 Results of the Model for a 40 Channel System
CNR as a function of average optical power for desired (a) CTBR and (b) CSO

Multi-Channel AM Fiber Optic CATV Trunks - From Lab to Reality

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ABSTRACT

In the past 12 months practical multi-channel AM fiber optic links have moved from the R&D lab into real world applications. In this paper, we focus on the design, characterization and performance capabilities of systems intended for use with signal spectra covering 20-80 CATV channels using distributed feedback (DFB) laser diodes and single mode optical fiber.

We discuss first the fundamental concepts used in a direct modulated AM laser based communications link. The noise and other degradation sources are identified and techniques used in mitigating their affects on performance are presented. Measurement techniques and practical results are also discussed.

We then discuss results on several laboratory demonstrations and field installations using the broad band AM link technology, with attention to the implementation issues faced by operators in the real-world environment of CATV networks.

1. INTRODUCTION

A year ago in Los Angeles, we heard several papers [1] on the architectural and technical aspects of fiber optic transmission for the CATV industry. Digital transmission, a combination of sophisticated Encoders, Decoders (CODECs), and off the shelf, mature, telephony-oriented transmission equipment, had been with us for many years. Frequency Modulation (FM) based systems were also available and being deployed in several markets. It was proposed, however, that neither Digital nor FM were on an appropriate cost-performance track to meet the most critical needs of the CATV operator - the trunking and distribution portions of the network. The solution? AM! (Amplitude Modulation).

Why AM? What was really being said was the following:

> "We know how to build high quality stacked VSB/AM signals in our head ends. The equipment is mature, cost effective, familiar and exists everywhere. We have set top converters and TV front ends in everybody's house, and they expect that stacked VSB/AM spectrum. And we can't afford to change everything at once, so whatever we add must be compatible on an incremental basis if we're to *evolve* to a fiber based network over a number of years."

Cost and available technology make AM an obvious choice for CATV fiber optic trunking. We have already observed that per-channel Digital and FM systems were applicable only to the high end part of the network (i.e., super trunking) and broadband interfaces for these techniques are not yet available. The challenge then for technologists is to solve the signal processing problem in the most direct manner - minimize the processing and maximize the performance of the transmission channel.

We at AT&T Bell Labs summarized these demands in the following set of design objectives:

1. The system must be cost effective.
2. The system must fit into existing architecture, yet be flexible enough to incorporate evolution.
3. The system must be compatible with the physical and environmental constraints of the typical CATV network.
4. The system must be installable and maintainable by the typical CATV technician.
5. The system must perform, now, and the technology must be capable of moving ahead with the advances in channel capacity, network size, and demands on performance anticipated for the future.

During 1988, several labs worked the issues that surfaced in LA and by year end, AM products were announced, delivered, installed, and put into service by several MSOs. Two basic architectures have emerged, one which recognizes the present limitations of off-the-shelf laser devices and uses several lasers in parallel to handle the spectrum, and a "home-run" single laser broadband architecture which demands premium performance from its components but delivers the simplest implementation.

In the following sections, we discuss the latter - a "home-run" architecture delivering 40 to 80 high quality CATV signals. Our focus is first on the technology issues, fundamental limitations and device characteristics, and finally on achieved performance.

2. A SIMPLE SYSTEM MODEL

A simple model of a fiber optic CATV trunk system is shown in Figure 1. The head end electronics here are modeled as N (N = # of channels) video modulators, converting a baseband video + audio signal to a VSB/AM signal at frequency f_n. The individual outputs are passively combined in several stages to form the composite spectrum.

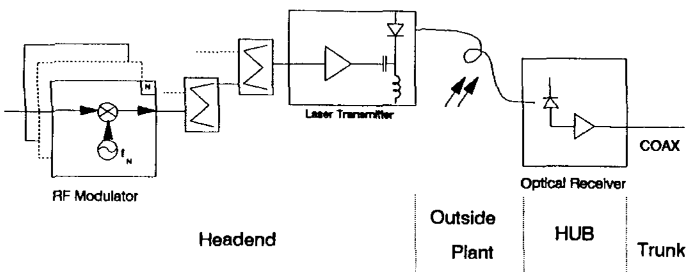


Figure 1

For our initial analysis, we will consider the performance with unmodulated carriers (CW case), resulting in frequency and time domain characteristics shown in Figures 2a and 2b, an analytical view of 42 cosinewaves summed together.

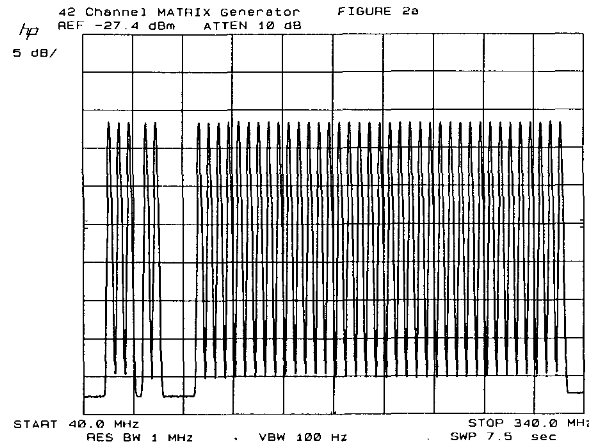
The laser transmitter is assumed to consist of an amplifier/driver device and a laser diode, at this level viewed as a current to light (optical) power converter. The laser launches this power into a single mode optical fiber, characterized by loss and dispersion (bandwidth), which delivers the power to a photo-detector diode at a remote location. The photo detector converts the incoming optical power to current, which is amplified and delivered to a load, here assumed to be a COAX cable distribution network.

3. A LOOK AT THE COMPONENT PARTS

3.1 Laser

The laser diode converts input current (modulation) to output light, a relationship often shown diagrammatically as in Figure 3a. This "L-I" characteristic shows several important parameters often considered when specifying lasers:

1. Threshold - The current level at which lasing (stimulated emission) begins.
2. Efficiency - The slope of the L-I characteristic, often referred to as dL/dI, in mw(opt)/ma.



Composite CATV VSB AM Signal

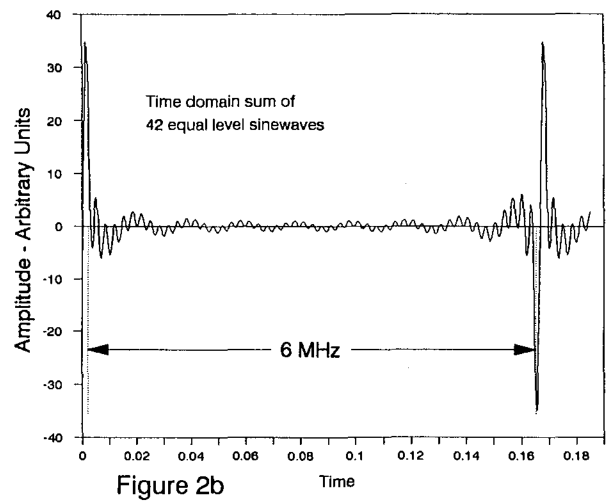


Figure 2b

3. Linearity - In general, you can only detect poor linearity from an L-I plot, not good linearity. A perfectly linear device follows a straight line over the region of operation, yet the deviation from "ideal" permissible for CATV applications is generally not measurable using L-I techniques.
4. Maximum Output - There is no simple definition of the maximum optical output from a laser device, rather it is a complex and device specific set of rules ultimately limiting the current density in the semiconductor junction. Most lasers exhibit a noticeable "rollover" or "current saturation" effect as shown in Figure 3a, where the non-linear L-I relationship becomes noticeable. For CATV applications, the maximum power is somewhat below this "observable" point on the L-I curve.

Not addressed on Figure 3 is the noise performance of the laser, normally specified as the *Relative Intensity Noise* (RIN). RIN is a significant contributor to overall AM link performance and will be discussed further below. As a device parameter, it is highly dependent on device structure and packaging.

Typical LASER L - I Curves

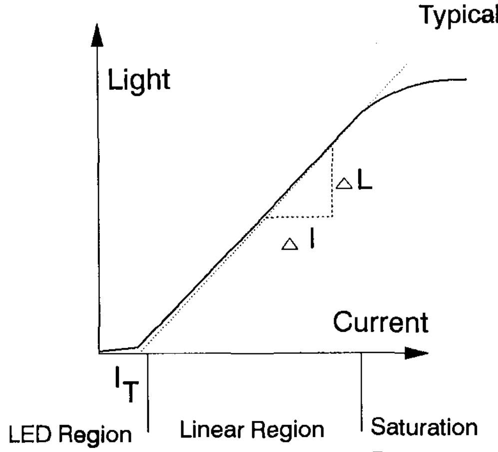


Figure 3a

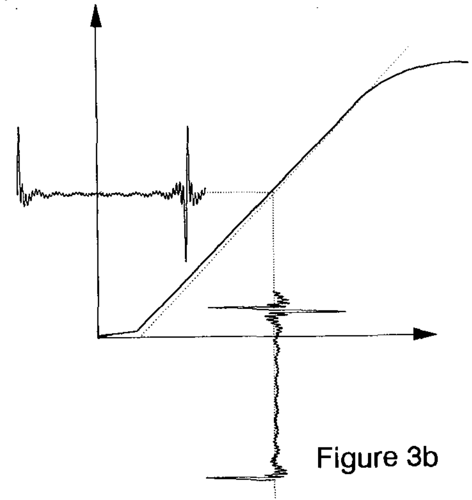


Figure 3b

Intensity modulation, or modulating the amplitude of the optical oscillator (laser), is achieved by changing the current level in the device. Since we know our signal (time domain, Figure 2b), is symmetrical about zero mean (it is a sum of zero mean sine waves), a DC operating point for the laser will need to be established if the RF modulation is to see a uniform I → L conversion over its amplitude range, as depicted in Figure 3b.

3.2 Optical Isolator

A laser may be viewed as an oscillator whose amplitude and stability characteristics are strong functions of cavity (semiconductor material) purity, current stability, thermal stability, and input energy from intended and unintended sources. A significant source of unintended energy is a reflection somewhere in the output circuit which, due to (optical) impedance mismatch, couples energy from the load back into the oscillator at a random time, a function of the propagation time from the laser to the point of mismatch. As we will discuss later, we have determined that certain limits must be placed on the amount of reflected power that may return to the laser.

An isolator is a device that has very low insertion loss in one direction, high insertion loss in the other. These devices, mounted in or near the packaged laser, provide the necessary limiting of reflected power.

3.3 Fiber and Connectors

A detailed discussion of fiber and connector systems is beyond the scope of this paper. For our purposes, we need consider only the loss of the fiber and installed connectors (in dB) and, to some extent, the reflection performance of the complete optical circuit. In the context of this work, with lasers at $\lambda = 1.3\mu$ wavelength, the fiber dispersion is low enough to be insignificant, or in essence, the transmission medium is assumed to have infinite bandwidth.

3.4 Optical Detectors

Optical power transmitted through the fiber must be converted back to an electrical signal for input to the COAX cable network. Semiconductor diodes, typically *InGaAsP* or *Ge* at $\lambda = 1.3\mu$ wavelengths, are ideal for this application due to their small size, high bandwidth, high reliability, and low voltage operation. Two types of diodes are candidates; PINs and avalanche photo-diodes (APDs).

As we will see below, APDs are not applicable for high channel load applications since a significant portion of the noise in the system is present at the input to the detector in the form of laser noise and shot noise, both of which would be amplified by the APD along with the signal.

The PIN diode is characterized by an efficiency, η, in units of ma (detected) per mw (optical) input. Typically η is defined and measured to include the loss of the connector and fiber pigtail. A PIN diode is typically modeled as a current source, shunted by a parasitic capacitance. The bandwidth of this current source is much larger then the CATV spectrum and is not of concern, although the parasitics in the package will combine with other receiver components to limit the overall system performance.

3.5 Amplifiers

Amplifiers and drivers are used at various points in the overall system to match the typical CATV RF levels to those appropriate for laser based systems. These amplifiers are conceptually no different from units used in COAX amplifiers, and are likewise characterized for noise figure, linearity and bandwidth performance. The required performance will be discussed as part of the actual analysis of a laser based trunk, to follow.

4. CATV TRUNK APPLICATIONS - PERFORMANCE CRITERIA

The key performance criteria [2] for CATV trunk applications are:

C/N - Carrier to Noise. The dB ratio of the peak carrier power for a given channel to the noise floor near that carrier, assuming a noise bandwidth of 4 MHz.

CTB - Composite Triple Beat. The dB ratio of the peak carrier to the peak power in the composite third order intermodulation tone which for CATV signals appears at the carrier frequency.

CSO - Composite Second Order. The dB ratio of the peak carrier to the peak power in the composite second order intermodulation tone. For standard and IRC frequency plans, the CSO appears at the carrier ± 1.25 MHz. For the HRC frequency plan, the CSO beats appear at the same frequency with the CTB beats.

We will look at C/N and intermodulation performance separately, since the noise performance of most components is well understood and may be accurately modeled. Intermodulation performance, on the other hand, must be measured and characterized on each individual unit.

5. C/N - NOISE SOURCES IN A FIBER OPTIC AM LINK

We will use the model [3] shown in Figure 4 to discuss noise sources. Regardless of source, we are ultimately interested in the total noise present at the input to the front end amplifier at the receiver. This approach also makes comparisons among these sources simpler. There are three dominant noise sources, modeled here as current sources since the receiver diode (PIN) is modeled as an ideal current source. We review these noise sources in detail below.

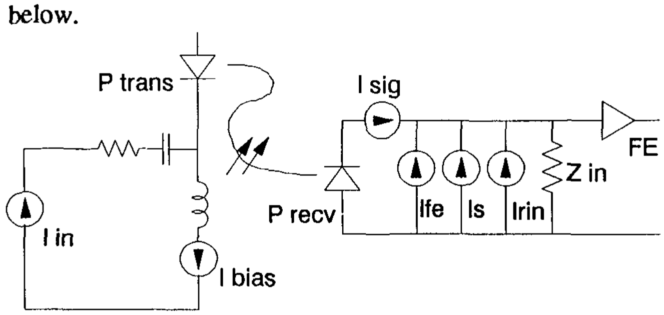


Figure 4

5.1 Front End Noise

All electronic amplifiers add noise to the input signal when delivering the output to a load. In RF systems, we typically deal with an amplifier in terms of its Noise Figure, a measure of the equivalent noise power that appears at the input. For this analysis, that noise power is converted to an equivalent current, commonly expressed in picoamperes per square root of Hertz ($pa/\sqrt{Hz}$).

The equivalent input noise is a function of many circuit and device parameters. It is generally not flat across the frequency spectrum of interest, may vary with temperature and load conditions. It must be characterized for each device or family of devices considered for use.

Low noise digital fiber optic system receivers have achieved equivalent input noise currents in the 3-5 $pa/\sqrt{Hz}$ range, although these receivers are typically limited in their RF output capability and are not yet useful in broadband CATV applications. Amplifiers useful for these broadband applications are more typically in the 12-16 $pa/\sqrt{Hz}$ range. Further improvements in this performance can be expected as CATV applications expand the need.

5.2 Shot Noise

The conversion of light energy, which arrives at the PIN junction in photon "packets", each at a particular energy level, to an electrical current flow involves the generation of hole-election pairs in the Intrinsic junction region as the discrete photon energy "packets" are absorbed. The effectiveness of this conversion is a statistical function and the deviation from perfect conversion is referred to as quantum or shot noise on the detected signal. It is given by:

$$\overline{i_s^2} = 2e\, Ip \quad A^2/Hz$$

where e = charge on electron

Ip = detected current

This noise is assumed to be spectrally flat over the CATV region of interest. Shot noise represents a fundamental limit on overall noise performance, to be asymptotically approached as other noise sources are reduced through improved device performance.

5.3 Relative Intensity Noise - RIN

The final dominant noise source, in the AM system, is laser intensity noise. When observed at the receiver, RIN is a function of many electro-optic and optical mechanisms, including

- Quantum effects in electron to photon conversion in the laser
- Reflection effects on the laser cavity
- Mode partitioning and modal dispersion
- Phase noise to intensity noise conversion in external reflective cavities.

RIN is a device performance parameter and must be specified and measured for each device. Because it is critically dependent on the optical circuit configuration, it is important to carefully specify and characterize the test setup when measuring laser RIN. In addition, intensity noise has a potential for significant spectral shaping, depending on the dominant source of intensity variation. Reflection induced intensity noise can be particularly frequency dependent due to transit times between the reflectors and the source.

Typical multimode (Fabry-Perot) digital system lasers have RIN performance in the -110 to -140 dB/Hz range, and laser noise is of little concern with respect to error rate performance. For AM CATV applications, RIN must be better than -145 dB/Hz for typical system applications. We have routinely achieved RIN performance from distributed feedback (DFB) lasers, with optical isolators, that span the range of performance from -148 to -152 dB/Hz in system level applications.

5.4 Noise Measurement and C/N

When viewed as equivalent unity bandwidth current sources, it is relatively easy to separate the total noise power into its component parts. First we assume:

$$\overline{i_{tot}^2} = \overline{i_{FE}^2} + \overline{i_S^2} + \overline{i_{RIN}^2}$$

The frontend noise power, $\overline{i_{FE}^2}$, is independent of the presence of an input optical signal and hence may be measured with the laser shut off. Secondly, the shot noise is a function of the DC detector current and may be calculated under those conditions. The RIN component then is derived by subtracting the $\overline{i_{FE}^2}$ and $\overline{i_S^2}$ components from the total. Device data sheets typically specify laser RIN as a dB ratio, so:

$$RIN = 10 \log \left[\frac{\overline{i_{RIN}^2}}{\eta P_o^2} \right] \qquad dB/Hz$$

To obtain the C/N ratio, we now must look at the achievable per channel carrier amplitude. Referring back to Figure 2a, our signal is modeled as a sum of N equal amplitude sine waves.

$$P_{TRANS}(t) = P_{cw} \left[1 + \sum_{i=1}^{N} m_i \cos(\omega_i t + \phi_i) \right]$$

Each channel i has a unique frequency defined by the frequency plan in use (STD, HRC or IRC), and even if phase locking HRC and IRC are used some random phase ϕ_i will be introduced by electronics and combiner cabling. For large N, N > 40 or so, we will therefore assume that the resulting amplitude distribution for P_{TRANS} is Gaussian. If we further assume that we do not wish to exceed the laser threshold with probability > .1%, the L-I characteristic shown in Figure 3a limits the achievable index of modulation, m_i, to about 4.4% for N=42 channels. In general, given m_i for a channel, the rms carrier power is given by:

$$C = \frac{m_i}{\sqrt{2}} P_{RECV} \cdot \eta \qquad ma,rms \text{ for } \eta \textit{ in } ma/mw$$

and

$$C/N = 10 log \left[\frac{C^2}{\left[\overline{i_{tot}} \cdot 10^{-9} ma/pa \right]^2} \right]$$

We will defer a detailed look at this equation until intermodulation is discussed, since it directly impacts the achievable index of modulation, m_i.

6. INTERMODULATION NOISE - CTB and CSO

The theory and mathematics of intermodulation noise were well developed [4] in the early days of broadband (relative) linear telephony and further analyzed in the early days of CATV[5], when channel loads on COAX began to exceed the original 13 off-air channels.

Basically, if we model the transfer characteristic of any transducer (amplifier, laser, detector, etc...) as a third order polynomial $e_{out} = a_1 e_{in} + a_2 e_{in}^2 + a_3 e_{in}^3$ and apply our

$$P_{TRANS}(t) = \sum_{i=1}^{N} m_i \ cos(\omega_i t)$$

signal spectrum, the resultant e_{out} is shown to consist of linear terms plus countable intermodulation products at frequencies related to $\omega_1 \pm \omega_2$ due to $a_2 e_{in}^2$ expansion (second order non linearity) and $\omega_1 \pm \omega_2 \pm \omega_3$ due to $a_3 e_{in}^3$ expansion (third order non-linearity).

In CATV, unlike telephony, the energy in each channel is highly concentrated at the carrier frequency, resulting in intermodulation products which fall in very narrow frequency ranges.

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The composite power in these frequency bands, a power based summation of the intermodulation products, is measured as the CSO (Composite Second Order) and CTB (Composite Third Order), interference power. It is normally measured relative to the carrier peak and reported in dBc, dB relative to the carrier.

Each composite second or third order beat is theoretically made up of a countable number of equal amplitude beats, assuming that the generating spectrum is flat, or in other words, $m_i = m_j$ for all i, j. If we assume that these beats are uncorrelated in frequency and phase, then the expected channel to channel relative differences in intermodulation performance will follow 10 log (N), where N is the number of second or third order products. In Figure 5a, we show a plot of the predicted second order intermodulation performance, for 42 CATV channels. In Figure 5b, we show a similar plot for third order beats.

In a real laser system, the achievable index of modulation, m_i, will be governed by the second and third order distortion coefficients a_2 and a_3 above, rather than by the simple Gaussian-threshold relationship reviewed in the idealized look at achievable carrier to noise performance above. We have achieved system level performance with indices, m_i, in the typical range of 2.5% to 5% .

7. TYPICAL RESULTS

In Figures 6a, b, c and d, we summarize the results of measurements on a 42 channel laser trunk link. Figure 6a is a spectrum analyser plot for a typical channel under test, showing the carrier , noise floor and second and third order composite intermodulation tones and the measurement results. Figure 6b is a derivation of the specific noise and device performance characteristics from those measurements. Figures 6c and 6d plot the broadband performance of the device on the theoretical $10\log(N)$ plots presented in Figures 5a and 5b. Measured parameters and broadband results can be compared with the theory and models above.

During the presentation, we will look at more statistical data from the Laser Link ™ units delivered to CATV MSOs during the first Quarter of 1989.

8. SUMMARY

We have reviewed many of the performance degradations and system considerations which are key to the application of AM modulated lasers in the CATV trunk networks. While the overall application is still in its infancy, these performance models will provide a foundation for unit to unit comparisons as well as evolutionary trends.

9. ACKNOWLEDGEMENTS

Credit is due to many members of staff in AT&T Bell Labs for their support and ideas in attacking these issues. I would like to particularly mention G. L. Fenderson and M. S. Schaefer who helped on the refinement of these models and the development of real hardware to test them.

I would also like to thank those members of the CATV engineering community who have openly shared their ideas, needs, techniques and time in helping our efforts. In no particular order, special mention to Louis Williamson, Dave Pangrac, Jim Chiddix, "Shorty" Coreylle, Tom Elliott, Richard "Rex" Rexroate, Dick Kreeger, Jack Ramsayer, John Walsh, Jim Hayworth, Bob Luff and Hugh Bramble.

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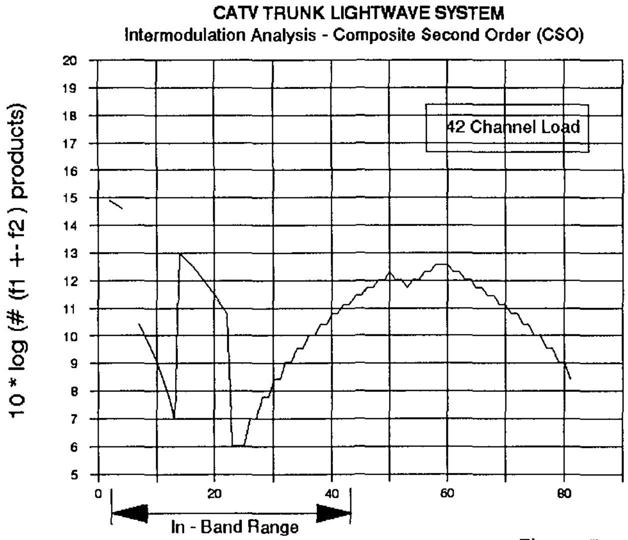

Figure 5a

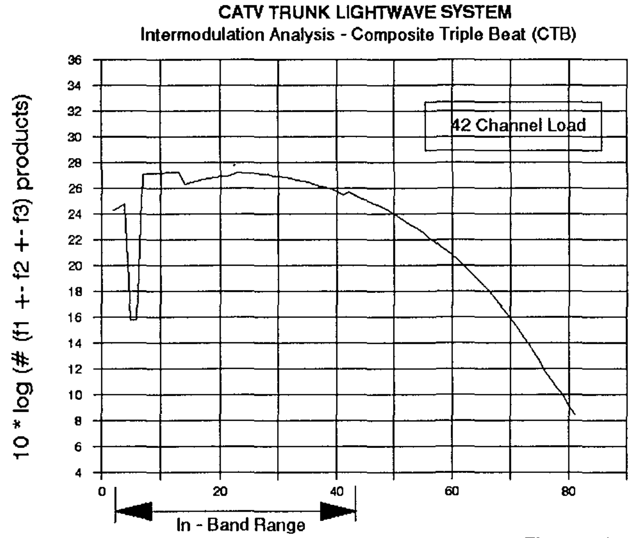

Figure 5b

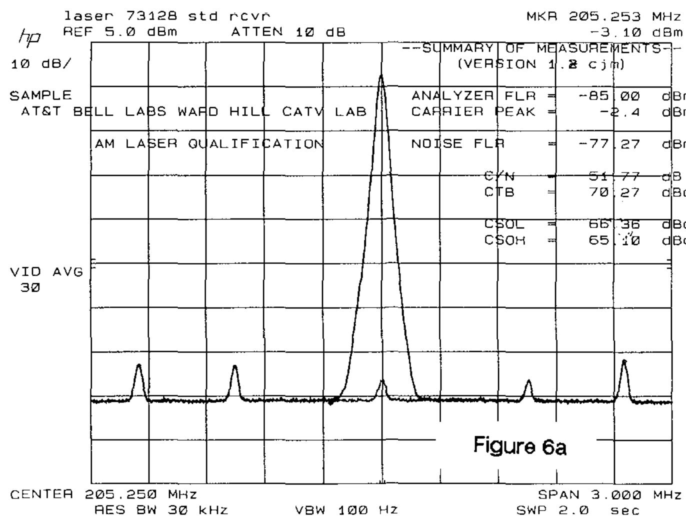

Figure 6a

AT&T Bell Laboratories CATV Trunk Design

Fiber Loss - dB =	5.15		12:06 PM
Received Opt Power - dBm =	-4.05		
PIN (standard)		Calculations -->	
Detected curent - ma =	0.345	Responsivity ma/mw =	0.877
Signal power out-dBm =	-48.2	Index of Mod =	0.071
at Impedance - ohms =	50		
SYSTEM (Under Test)		Calculations -->	
Analyzer floor dBm/Hz =	-131.90		
Analyzer Impedance ohms	75.00		
PIN detected current-ma =	0.316	GB2 Effective Gain dB =	15.20
Tone Out GB2 - dBm =	-4.2	GB1 Effective Gain dB =	9.70
Tone Out GB1 - dBm =	-19.4	NF GB1 + GB2 dB =	12.63
Tone Out FE - dBm =	-29.1	FE Transimpedance ohms=	602.76
Noise GB1+GB2 - dBm/Hz =	-130.60	FE Noise pa/sqrtHz =	12.37
Noise Nolite GB2 dBm/Hz =	-125.00	LASER RIN dB/Hz =	-150.92
Noise FE+GB1+GB2 dBm/Hz =	-122.30	C/N FE OUT dB =	52.77

Figure 6b

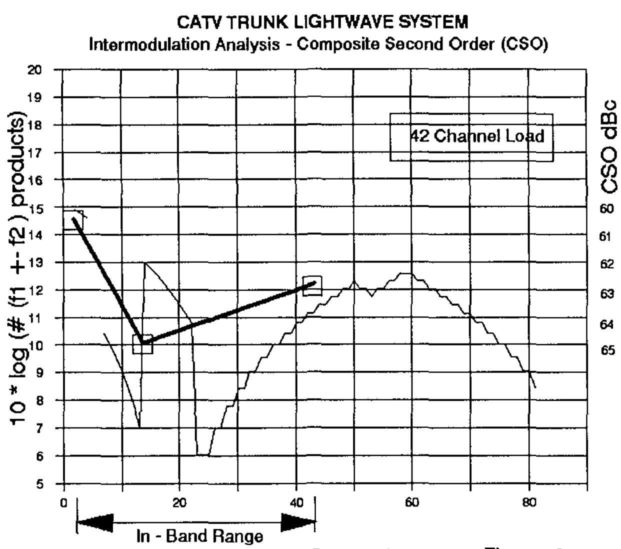

Figure 6c

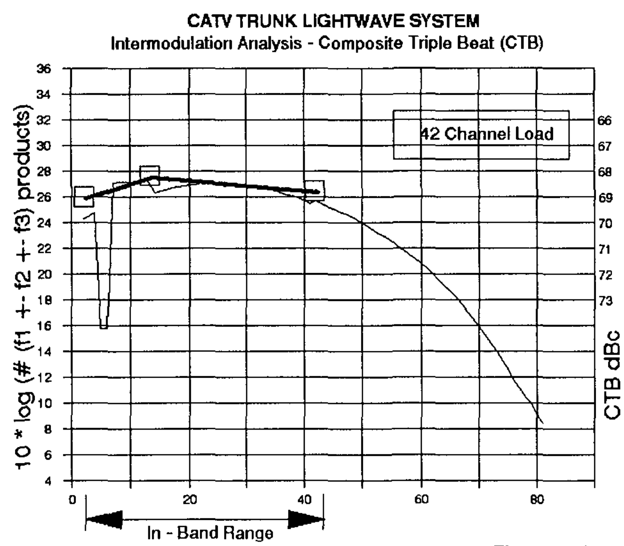

Figure 6d

NEW APPROACHES TO CATV SYSTEM POWERING

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Joseph L. Stern - STERN TELECOMMUNICATIONS CORP.

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ABSTRACT

As CATV grows and becomes more complex, the need to provide additional power for extensions, upgrades and special equipment can pose some interesting challenges, especially if the system wishes to avoid new powering locations or system power sector redesign.

INTRODUCTION

This paper will explore two options: (1) increasing the efficiency of 60 Hz power transmissions, and (2) adding high frequency AC power to the existing 60 Hz power. An overview of research and development is included to demonstrate the feasibility of equipping new amplifiers or off-premise devices with current controllers, and of retrofitting existing AC power supplies with high frequency add-on power sources.

REQUIREMENTS FOR TELACTION PROJECT

In early 1988, Telaction authorized Stern Telecommunications Corporation (STC) to begin a feasibility study to determine if existing CATV powering schemes would accommodate an auxiliary power delivery system to provide power for Telaction Frame Store Units (FSU's). Telaction FSU's are installed at trunk amplifier - bridger stations. The number of FSU's is determined by the number of subscribers connected to the distribution legs of the system as well as the activity of these subscribers. Each FSU requires approximately 50 watts of power and is designed to operate from the standard 60 VAC - 60 Hz CATV system power supplies.

Most CATV systems do not have the excess power capacity available to provide for operation of the FSU's. Any excess capacity that does exist is usually reserved for future expansion of the CATV system. Most system powering designs limit power supply loading to approximately 75% of the output capacity. Some systems are already using the full capacity of the existing power supplies.

Several FSU powering options have been investigated. These include:

1) Direct 120 volt AC powering.

2) 60 volt power using new power supplies at each FSU location or by use of an overlashed powering cable.

3) Redesign and rebuild of the CATV power system to provide for existing system needs as well as FSU power.

4) Alternative power supply designs to provide power in unconventional ways.

120 volt powering requires special construction to comply with local electrical codes. Each installation must be scheduled with the local utility company, and permits must be obtained from the local electrical inspector. This option is considered costly and would require long lead times for implementation.

60 volt powering also requires long lead times for utility installations, plus the additional cost of new AC power supplies and/or the cost of overlashing a new cable to the existing plant. This option may not be acceptable to some of the CATV system operators.

Redesigning and rebuilding the powering system to enhance power availability for the existing plant and to provide for the FSU power requirements would involve considerable expense, long lead times, and would cause disruption to the operation of the CATV system during the rebuild.

Alternate powering systems would require devising an auxiliary power source that could provide useable power to the FSU's using the existing cable as a conductor but with minimal impact upon the existing plant.

APPROACH TO THE POWERING PROBLEM

Modern CATV systems are powered by applying a 60 volt, 60 cycle AC potential between the coaxial cable center conductor and the sheath. The power source is typically a square-wave output ferroresonant transformer, which in reality produces a waveform similar to a trapezoid.

The power supply can be either a pole mounted unit or a pedestal mount for underground systems. Most systems use a standby-type power supply which contains batteries for back-up operation in case of utility power failure. The local utility company provides electrical service to the power supply enclosure. Typically a service disconnect as well as a watthour meter is installed at the power supply location.

Most CATV AC power supplies provide up to 16 amps at 60 volts (960 VA). The 60 volt AC power is rectified, filtered, and regulated to provide the one or more DC voltage(s) required to operate the active components within the amplifier. Recent amplifier power supply designs utilize high frequency switchmode technology to provide high conversion efficiency as well as small size and tight voltage regulation. In some cases an isolation transformer will be located at the input of the power supply to step-down the 60 volts AC to a lower voltage prior to rectification. See Fig. 1. These transformers typically have taps for adjusting for different AC input voltages to accommodate IR drop. All amplifier power supplies utilize electrolytic capacitors to filter the rectified (pulsating) DC and to provide energy storage for the switchmode regulator circuit. These capacitors do not draw energy from the AC input for the full duration of the 60 Hz AC cycle. Due to the nature of "peak charging" of the capacitors, the current input pulse occurs 10 to 20 degrees after the "zero voltage cross," and during the ramp-up of the cycle to the maximum peak voltage. The current pulse usually lasts for less than 60% of the half cycle (100 - 120 degrees). See Fig. 2.

It was imperative to characterize existing CATV systems to understand operation of the power supplies interacting with the coax. cable, the parasitic inductive, capacitive, and resistive elements, and the amplifier loads. Current and voltage waveforms were measured on a number of CATV systems and on laboratory test cascades. The systems tested include: B-Q Cable (Warner) in Queens; Post Newsweek in Deerfield and Highland Park; Alpha Technologies Research and Development Labs in Bellingham, WA, and Burnaby, B.C.; the test laboratory at Cableshare in London; and the burn-in room system at Telaction, Schaumburg.

The measurements show that there is a "time window" during which there is very little or no current flowing to the load. It is the existence of this window that provides the opportunity to deliver power to the FSU's without disrupting the existing power supplies or amplifiers.

Fig. 2 shows the typical 60 volt - 60 Hz voltage waveform as well as the current pulse waveform drawn by several amplifiers in a cascade. It is evident that there is a portion of the half cycle where very little energy is delivered to the amplifiers. Fig. 3 is the product of the voltage and the current (Volts x Amps = VA). This is a waveform of the actual power delivered to the load during the half cycle. Fig. 4 defines the "time window" available where additional power could be inserted and delivered via the coax. cable to the FSU's.

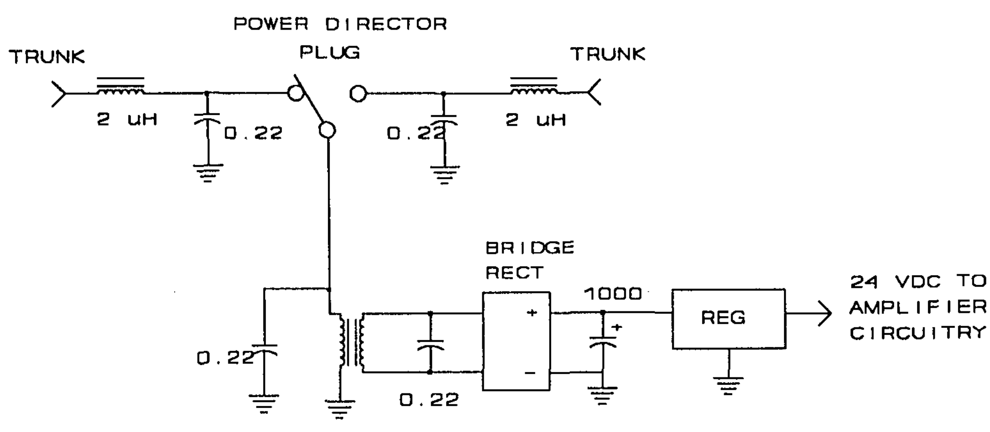

FIG. 1
TYPICAL INTERNAL AC/DC
POWER PATH, TRUNK AMP

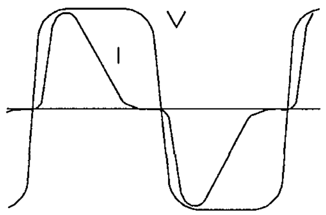

FIG. 2

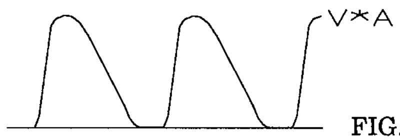

FIG. 3

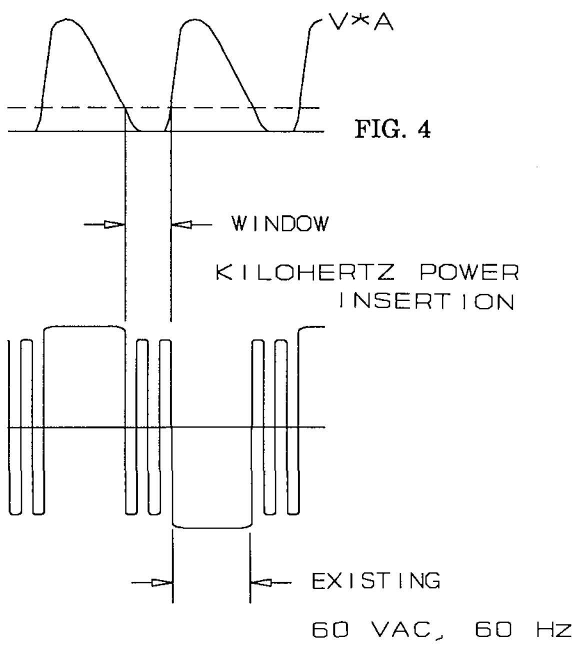

FIG. 4

It is important to note that the "window" will vary in duration depending upon the actual voltage at each amplifier location and the current draw, i.e. the AC voltage at the end of a powering sector is usually lower due to the draw of current through the resistance of the coax. cable and the resultant voltage drop. An amplifier at this location will draw current for a longer period of the cycle in order to supply constant power to its circuitry. The window of minimal power consumption can also be altered due to power factor of a particular location. A load with a capacitive component (such as CATV amplifier power supplies and high frequency by-pass capacitors in the amplifier AC power bus) will exhibit a phase shift of the current and voltage. The current waveform will tend to "lead" the voltage by a few degrees (see Fig. 2). The same is true if there is an inductive component in the load. The current waveform will tend to "lag" the voltage waveform by a few degrees. In the majority of the systems tested, the waveforms indicated either a slight "leading" power factor or the voltage and current were in phase (unity power factor).

UNIQUE APPROACHES

Based on the understanding of the CATV current and voltage relationships, two possible approaches to supplemental powering became evident:

1) "Shared time control"

2) "Kilohertz power"

"Shared time control" of existing 60 volt loads to optimize power draw over the voltage cycle. Theoretically, existing power transfer would be more efficient, thus making extra power available for powering FSU's.

A prototype "shared time controller" (STC) was tested. This approach used solid state switches to intelligently control the AC input to the FSU's. A simple synchronization circuit detected the zero cross of the AC waveform and waited for a preset time delay before allowing the FSU to draw power. Controlling the FSU current draw in relation to the existing CATV current draw provided an improvement in power delivery by spreading the total CATV and FSU current draw over the entire voltage cycle. The major effect was a reduction in IR drop in the cable. This approach also supports the use of existing power supplies (if there is enough excess capacity) or the addition of a supplementary 60 Hz power supply. Although promising for some applications, it was determined that this approach could not provide the extra power required.

"Kilohertz power" inserted during the previously described "window." A high frequency energy burst with an amplitude less than the 60 volt waveform would be coupled onto the coax. cable via an impedance matching network. See Fig. 4. This "Kilohertz power" would be recovered by a special "power receiver" in the front end of the FSU power supply. The FSU would ignore 60 VAC - 60 Hz power and only accept the "Kilohertz power" bursts. This energy would be converted to the appropriate DC voltages required by internal circuitry of the FSU.

IMPLEMENTATION OF THE CHOSEN APPROACH

In mid-1988, Alpha Technologies began a cooperative effort with Stern Telecommunications to research and develop a "Kilohertz Power Unit" (KPU). The preliminary design specifications included an "intelligent" controller to measure the existing 60 Hz voltage and current. This would allow determination of the "window" in which to insert a high frequency power burst with a power amplifier and coupling network. The most efficient frequency of this power burst had to be determined. A "Kilohertz Power Receiver" (KPR) was required to mount in the FSU. This unit would ignore 60 volt - 60 Hz power and only receive the Kilohertz power. The power would be regulated to provide the appropriate DC voltages for FSU operation.

Other important requirements included:

1) Small physical size in order to allow retrofit in existing AC power supply enclosures. This would provide the advantage of deriving input power from the existing AC receptacle, thus eliminating additional utility installations.

2) Minimum impact on the existing CATV plant. The KPU must not introduce interference with the existing AC power supplies or with the video signals.

3) High reliability.

4) Provisions for status monitoring.

Work began at Alpha Technologies to construct and test a prototype KPU system. The first requirement was to determine the appropriate frequency for the Kilohertz power. A test cascade was constructed with Alpha AC power supplies and Jerrold trunk stations and passives to provide a convenient system simulator for testing. A variable frequency / variable amplitude power amplifier was constructed to "sweep" the cascade to determine the losses at various frequencies. A simple controller was added to couple the power bursts during an adjustable window time in the 60 Hz voltage cycle. Frequencies from 400 Hz to 25 kHz were tried and the losses through various components in the cascade were recorded (see Fig. 5). At low frequencies (less than 700 Hz), the amplifier power supplies loaded down the Kilohertz power, which was unacceptable. At higher frequencies, the AC bypass capacitors in the passives and in the trunk amplifiers also loaded down the Kilohertz power. It was determined that 1200 Hz was the optimum frequency for power transmission with relatively low losses.

After testing and analysis, it was determined that it would be necessary to actually disconnect the existing 60 Hz AC power supply from the load during the "window" when Kilohertz power was coupled onto the coax. By doing this, the window became much more defined and consistent in width. A rugged diode-switching circuit was designed to accomplish this function. A "fail-safe" relay was included to bypass the existing AC power around the KPU in the event of a malfunction (see Fig. 6). A 400 watt power amplifier stage was designed to provide the Kilohertz power. The amplifier was designed to drive a low impedance load without damage. A current limit circuit was included to protect the amplifier from short circuit faults.

The main control and coordination of the system was provided by the microprocessor control

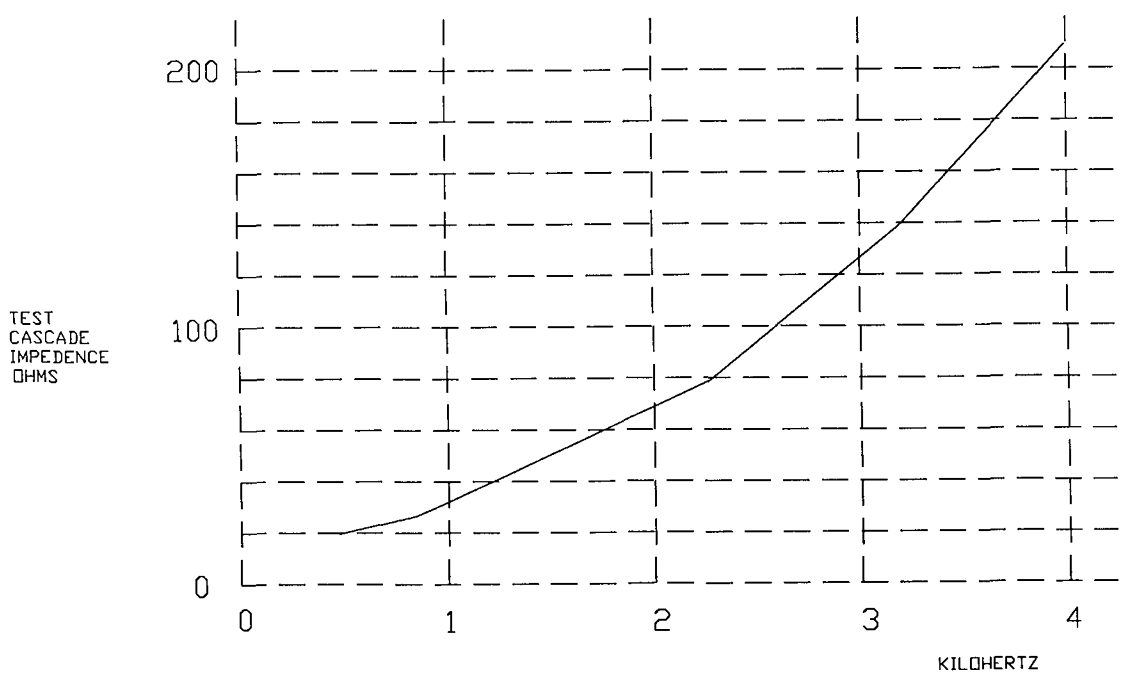

FIG. 5
TEST CASCADE
IMPEDENCE VS. FREQUENCY

unit (MCU). The MCU uses a high speed analog to digital converter to measure currents and voltages in real time. This enables the MCU to determine the 60 Hz AC and Kilohertz power being delivered to the loads in order to adjust the duration of the Kilohertz insertion window. The MCU decided when to drive the diode switches to disconnect 60 Hz power and connect the Kilohertz power to the system. It also determined the duration of the Kilohertz insertion, then re-connected the 60 Hz power for the next half cycle. The MCU continuously adjusted its operation to accommodate voltage or load fluctuations.

The KPR (Kilohertz Power Receiver) was installed in the FSU's and included a coupling circuit to acquire the Kilohertz power and convert it to DC. The KPR was synchronized to the zero cross of the 60 Hz power. It featured an adjustable window duration to enable input of the Kilohertz power (as well as some of the 60 Hz power, if desired). This is advantageous if a particular CATV system has some excess 60 volt capacity.

In lab testing at Alpha, the system was able to provide power to 4 KPR's (50 watts each) in the cascade. No degradation of the cable signals was observed, and no adverse effects on the existing 60 volt - 60 Hz power supplies or amplifiers were noted.

CONCLUSION

It has been demonstrated that there are viable alternatives to standard 60 volt - 60 Hz power for specific applications with limited performance criteria. Assembly of several pre-production Kilohertz power units will be completed by Alpha Technologies and installed for actual system tests. Although experimental in nature, this approach shows promise for implementation by Telaction in systems where other options for powering are not available.

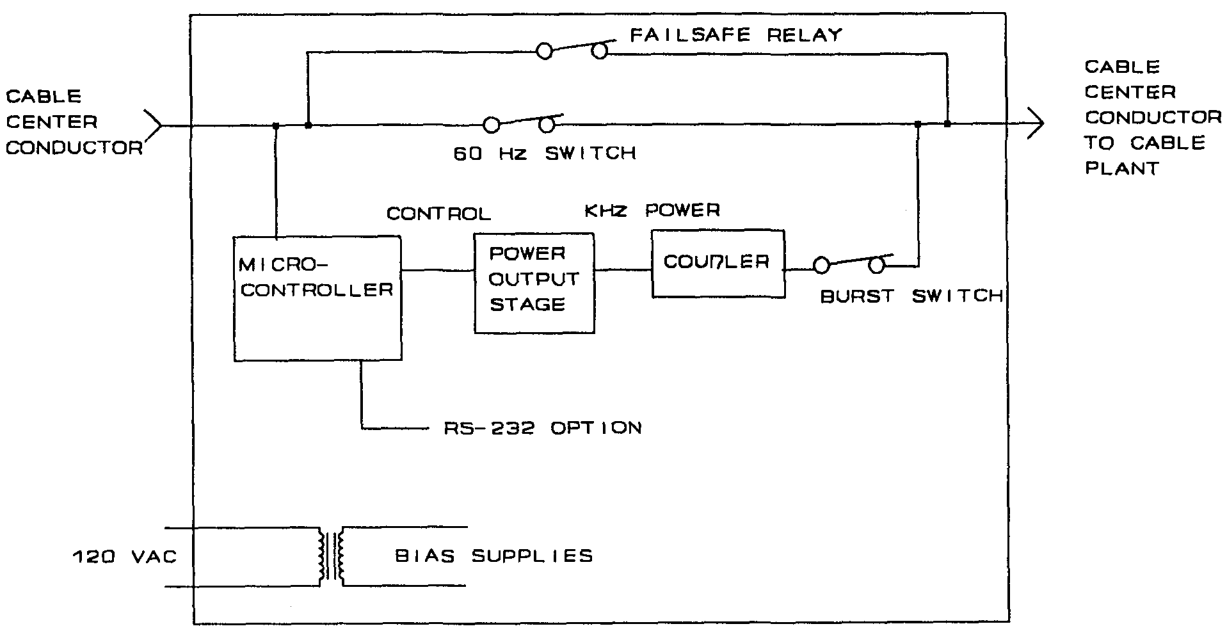

FIG. 6
KPU FUNCTIONAL BLOCK DIAGRAM

NOISE MEASUREMENTS FOR A CATV SYSTEM-C/N,S/N,Phase Noise

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INTRODUCTION

A study of the theoretical calculations of noise in a CATV system and the methods to measure the noises is presented.

This paper is focused on the effect of phase noise on the signal-to-noise ratio (S/N). It is well known that excessive phase noise will degrade the S/N but lack of understanding of the exact relationship between phase noise and S/N causes unneccessary confusion about the importance of phase noise. It is the purpose of this paper to clarify the significance of phase noise and provide a detailed calculation and measurement method for phase noise and its conversion to S/N.

I. CARRIER-TO-NOISE-RATIO

Carrier-to-noise ratio (C/N) is defined as the power ratio of the unmodulated carrier to the noise in the communication channel. In the CATV system, the noise is mainly white noise (thermal noise). The white noise of each component of the CATV system will accumulate when they are cascaded. To specify the degree of noise containmination of a piece of equipment, the term Noise Factor (f) is used. The f of a piece of equipment represents the increase in C/N of a signal which passes through that equipment. The f can be written as follows:

$$f = \frac{C/N\ (output)}{C/N\ (input)} \qquad [eq.1]$$

Note that the f and C/N's in eq. 1 are numerical ratios. If a system is noiseless, then the f will be 1. Noise factor f should not be confused with the noise figure F which is defined as the logrithm of noise factor F = 10* log (f) and is specified in db. Noise figure is more familiar than noise factor but for most noise power calculations noise factor is more practical.

It is often more convenient to use the term "equivalent input noise" to describe the noise contribution. The noise at the output of a piece of equipment is contributed by the noise accompaning the input signal plus the noise generated by the equipment itself. The self-generated noise can be treated as a kind of noise that occurs at the input of the equipment and the equipment is treated as noiseless. In this way, the output C/N can be calculated easily once the equivalent input noise is known. The equivalent input noise (Pn) is given by : (note 1)

$$Pn = Pn(in) + (f-1)kTB \qquad [eq.2]$$

where Pn (in) is the power of the noise that accompanies the signal, f is the noise factor of the equipment, K is the Boltzmann constant (1.38E-23), T is the absolute temperature (290 at room temperature) and B is the bandwidth (4 Mhz for standard baseband video).

Note that eq. 2 is a formula for noise power addition so that both the input noise power and the noise power generated by the noise factor should have the dimension of watts instead of dbmv or dbm.

Table 1 provides a conversion for dBmV (dB relative to one millivolt at 75 ohms) and watts. Table 2 is the conversion for noise factor and watts.

The idea of equivalent input noise is very helpful to calculate the C/N. For example, suppose the input signal level is +30dBmV and C/N is 70dB and this signal is going through an equipment with 10 db Noise Figure and we want to know the C/N at the output.

First the signal power of +30dbmv is calculated as 13.33 uW, the noise power of -40dbmv is calculated as 1.333 pW. The noise generated by the 10 db noise figure is equal to 0.144 pW so that the total input noise is 1.477 pW and C/N at output is 69.6dB.

If the noise figure is increased to 30dB, the equivalent noise power will be 16 pW, the total noise becomes 17.33 pW and C/N becomes 59 dB.

If the input C/N is 50dB, in other words, input noise power is 133 pW; the output C/N will be also 50 dB when the noise figure is 10 dB and 49.5 dB when the noise figure is 30dB.

Compared with the other noise parameters, C/N is relatively easy to measure. A spectrum analyzer or signal level meter can do the job with good accuracy.

A sophisticated spectrum analyzer such as the HP8568B is especially convenient for C/N measurement. The noise floor level can be automatically measured in 1Hz resolution with no need analyzer or meter as the resulting compression can produce an artificially low reading.

The noise power measured in a 1 Hz bandwidth should be converted to 4 Mhz video bandwidth. The conversion is 10* log (4000000) which is equal to 66 dB. The C/N, in dB, can be calculated as follows:

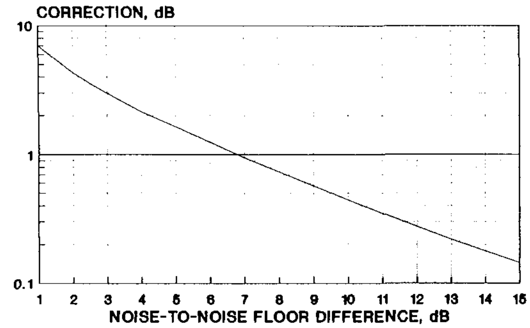
C/N = carrier level - noise level (dbmv/Hz) - 66 + correlation [eq. 3]

for IF filter, detector response and LOG amp correction. However, the spectrum analyzer noise floor correction is needed when the noise floor measured is less than 10 dB above the noise floor of the spectrum analyzer. This correction is required not just for spectrum analyzers, but for all RF level measuring equipment when used to measure equipments with low noise floor levels like headend modulators.

To determine the correction factor, the difference of noise floor level with and without the equipment under test should be measured. The correction factor can be read from the noise floor correction chart in figure 1. If the noise floor difference is less than 3 dB, in other words, the signal being tested has lower noise than the spectrum analyzer, a low noise pre-amplifier should be used to boost the signal level or the input attenuation of the spectrum analyzer should be reduced. However, care must be taken not to overdrive the

FIGURE 1

CORRECTION CHART FOR NOISE FLOOR



For example, a typical Jerrold C5M modulator output at CH 2 is to be measured for C/N. As shown in figure 2, the carrier level is 58 dBmv and the noise floor level is -75.6 dbmv. The noise floor level difference is only 7 db so that a correction factor of 0.9 dB is needed. The C/N is calculated as follows:

C/N = 58 - (-75.6) - 66 + 0.9 = 68.5 dB

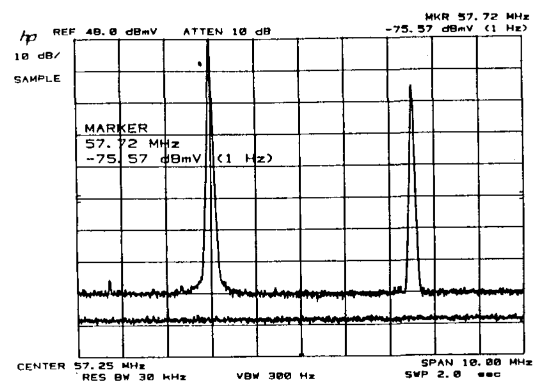


FIGURE 2

C/N MEASUREMENT FOR C5M

The C/N measurement is a very straightforward and accurate way to determines the degree of containmination of the carrier after passing through a system or a piece of equipment. But for the person who receives the signal at the end of the communication system, it is the quality of the video signal, instead of the carrier, that is of concern. In other words, video signal-to-noise ratio (S/N), instead of C/N, should be used to judge the performance of the system. Fortunately, since the carrier is AM modulated by the video signal, the noise floor is directly converted to baseband video noise so that we are able to calculate the S/N from the C/N. The conversion of C/N to S/N needs to be corrected , however, if phase noise prevails. This will be discussed in the following section.

II. SIGNAL-TO-NOISE-RATIO

A carrier modulated to 87.5% by a 100 IRE flat field video signal is shown in figure 3. It is clear that the full power of the carrier only occurs during the sync period while the video signal is carried at a lower power level. For a 100 IRE flat field signal the carrier power for the video is only 12.5% of the full power carrier . This is the case because the synchronization signal is more important than the video signal. If a picture can not be synchronized, it can not be watched no matter how clean the video signal is. It is also the reason that positive sync instead of negative sync is adopted.

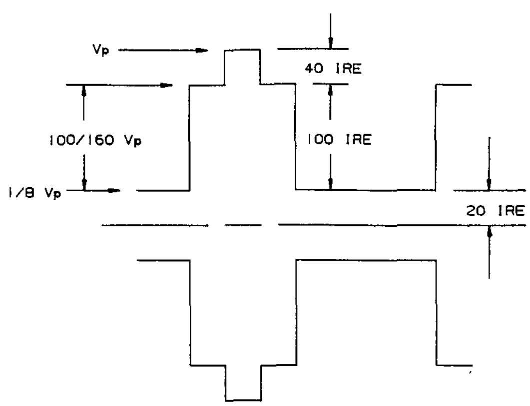


FIGURE 3
100 IRE FLAT FIELD MODULATION 87.5%

It is evident that the C/N measurement, which measures the full power carrier vs. noise, will yield a higher value than the S/N measurement which measures a fraction of the full carrier power vs. the same noise.

In order to calculate the equivalent signal-to-noise ratio, it is assumed that the signal level to the TV demodulator is high enough to overcome the noise figure of the TV tuner so that the noise contributed by the demodulator is negligible. It is further assumed that the Nyquist slope of the demodulator is accurate so that the 6 Mhz RF channel is equivalent to a 4 Mhz baseband video channel. These assumptions are valid in most cases. Assume also that the noise in the RF channel is white noise only and the phase noise of the carrier is low enough not to affect the measurement.

Suppose the carrier peak level is Vp and the noise power is Pn, the Carrier to RMS noise ratio (C/N), by definition, can be written as:

$$C/N = 10 * \log\left(\frac{Vp*Vp/2}{Vn*Vn}\right) \qquad \text{[eq.4]}$$

The video S/N is defined as the ratio of the 100 IRE video level to the rms noise level. In a 75 ohm system, it is equivalent to the ratio of the peak-to-peak power of 100 IRE video level to the rms noise power. As in figure 3, the peak carrier level is 160 IRE which is equal to 8/7 * (100 IRE + 40 IRE). So the video level is equal to Vp*100/160 and S/N is

$$S/N=10 * \log\left(\frac{Vp^{2} * (100/160)^{2}}{Vn^{2}}\right) \qquad \text{[eq.5]}$$

so that

$$C/N=S/N + 10* \log((160^{2} / 100^{2})*2)=S/N + 7.1 \text{ dB} \qquad \text{[eq. 6]}$$

According to the NTC-7 video test standard (note 4), three filters are needed for video noise measurement. Two filters, the 4.2 Mhz low-pass filter and the 10 Khz high-pass filter, must be in use at all times. The noise power measured with only these two filters is called the unweighted noise. A weighting filter, which simulates the visual response of human eyes to white noise, may be used and the noise measured with the weighting filter is called the luminance weighted noise. Noise below 10 Khz, like hums, is usually periodic and should be measured as peak-to-peak. This type of low frequency periodic noise is treated seperately from the random white noise.

It should be noted that the S/N calculated in eq. 6 is unweighted. With a luminance weighting filter, the S/N is improved by 6.8 dB from the unweighted S/N (note 2). As a result, C/N is almost equal to S/N with a weighting filter.

$$C/N = S/N \text{ (luminance weighted)} + 0.3\text{dB} \qquad \text{[eq.7]}$$

Eq. 7 is based on the assumption that the RF noise sidebands above and below the picture carrier are coherent so that they add directly. For a CATV modulator, the RF noise floor is normally much lower than that of typical video sources so that the modulated RF noise sidebands are coherent. In such a case, equation 7 accurately relates C/N and S/N. In cases where the noise phase is not coherent above and below the picture carrier, as with broadband distribution system noise, the equation is modified to

C/N = S/N (luminance weighted) [eq.7a]

Refer to the paper described in Note 2 for the derivation.

To measure the S/N, a demodulator is required. The Tektronix demodulator 1450 is a typical precision TV demodulator for accurate S/N measurement. The S/N of that demodulator is specified as 60dB min and the equivalent S/N with luminance weighting filter is 67dB. Typically, the S/N of the Tek 1450 with luminance weighting filter is about 70 dB.

The equipment setup for S/N measurement of a headend modulator is shown in figure 4. The video noise can be measured by using either an AC RMS meter and a set of filters or the automatic video test equipment, such as Tektronix 1980 Answer system or the Tektronix VM700 video measurement set. When the automatic video measurement equipment is used, it is important to make sure that the modulation depth on the test modulator is exactly 87.5 %. The modulation depth is not a concern when an AC RMS meter is used to measure the noise because the video signal has to be removed from the modulator anyway.

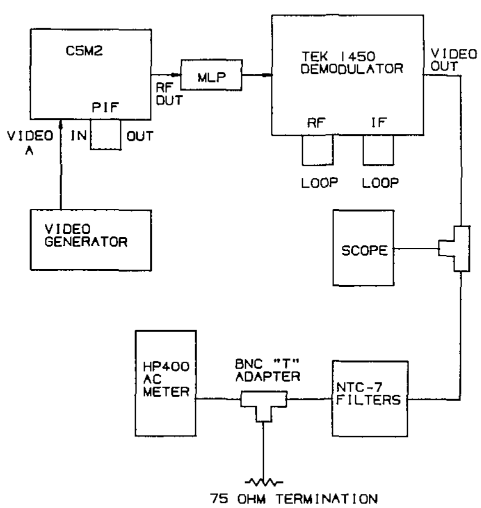


FIGURE 4
TEST EQUIPMENT SETUP FOR S/N MEASUREMENT

If the Answer system is used, the S/N measurement can be done quickly by typing a simple command. The Answer shows the S/N in four different ways --- unweighted, luminance weighted, chromance-weighted and low frequency peak-to-peak noise.

The Answer system, however, can not measure the low frequency peak-to-peak noise accurately. Such a system cannot separate the tilt and random low frequency noises from the hums and other periodic low frequency noises and measures the total low frequency peak-to-peak disturbance. A 1% tilt will cause the Answer to show -40 db low frequency noise. Some automatic test systems such as the VM700 have a function to correct the error due to tilt and the result is more accurate.

If an RMS meter and filters are used for S/N, it is wise to make the high-pass filter switchable (as we will see later). The HP3400A AC meter is good for the video noise measurement. The S/N can be read directly from the HP3400A AC meter by adding 0.7 dB to the noise level measured in db. This correction is needed because the meter 0 db reference is equal to 1 mw/600 ohm (775 mv) while the signal level 0 db reference is 714 mv.

The conversion of C/N to S/N by eq. 6 is found to be very accurate. If there is a discrepancy between C/N and S/N measurements, it is caused by either spurious signals or the phase noise of the carrier. The effects of spurs depends upon the location of the spurious signal. But how does the phase noise affect the S/N measurement?

III. Phase Noise

Phase noise is a general term used to describe the phase perturbation of a signal. Depending on the rate of the perturbation, the phase can have several characteristics, such as white phase, flicker phase, white FM and flicker FM. Phase noise is inherent to any oscillator and it is strongly affected by the Q of the oscillator. For example, a crystal oscillator has lower phase noise than an L-C oscillator due to the high Q of the crystal. Phase noise can be measured and specified in several ways, such as time variance, single-sideband power density or residual FM. In CATV systems, residual FM is probably the most easily and commonly used approach to the phase noise.

The residual FM of a carrier indicates the total power of the phase noise. In other words, the total phase noise power can be represented by a noisy FM modulation with deviation equal to the residual FM. The higher the residual FM, the more the carrier phase jitters. Residual FM can be measured directly by using a modulation analyzer such as the HP8901A, which characterizes the residual FM by measuring with different baseband filters. The alternative is to measure the single-side-band (SSB) noise power density with a spectrum analyzer and calculate the residual FM from the noise power density. As we will see later, the SSB noise power density is the better way for residaul FM measurement due to limitations of the modulation analyzer.

In a television receiver, FM noise of the picture carrier is converted to amplitude noise by the Nyquist slope of the demodulator. Refer to figure 5 and suppose the FM noise causes the carrier to jitter by f(KHz). After the Nyquist slope, the noise becomes amplitude jitter of f/750 %. Since the phase noise is random, it must be measured in rms.

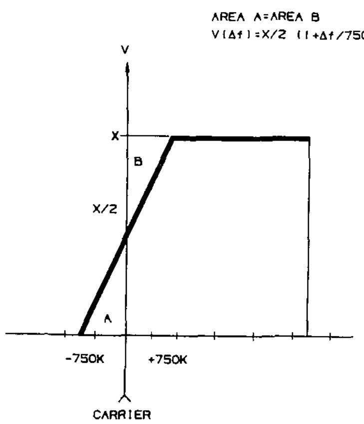


FIGURE 5
TV RECIEVER NYQUIST RESPONSE

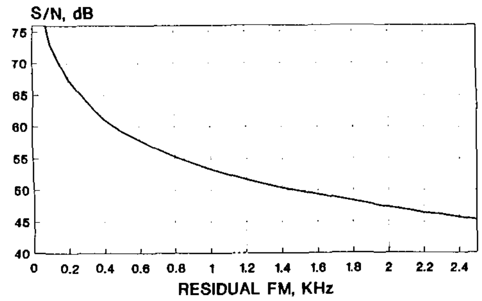
The S/N caused by the phase noise can be calculated as:

$$S/N = 20 * \log \left(\frac{Vp*100/160}{Vp*\ f/750} \right) \qquad [eq.8\text{-}a]$$

$$S/N = 53.4\ dB - 20 * \log f(KHz) \qquad [eq.8\text{-}b]$$

Eq. 8-b, plotted in figure 6, provides a conversion chart of residual FM to S/N.

FIGURE 6
RESIDUAL FM TO S/N CONVERSION CHART



Eq. 8-b and the chart show that a phase noise with residual FM of 1 KHz is equivalent to an AM noise that produces a S/N of 53.4 dB, and that 100 Hz residual FM produces a S/N of 73.4 dB. Since the amplitude noise caused by the phase noise will add to the noise floor, the total system noise is the sum of the two noises. For example, if the S/N requirement of the headend is 60 dB and the C/N is 68 dB, then the phase noise must be less than 400 Hz. The 400 Hz value can be used as a rule of thumb to judge the phase noise performance of CATV headend frequency agile equipment.

The basic process for measuring the residual FM with the HP8901 is simple. Connect the test signal to the HP8901 and choose the FM function in AVE mode. Since video S/N is always measured with a 10 Khz HPF, the residual FM also has to be measured with a 10 Khz HPF. The HP8901, unfortunately, only has 50 and 300 Hz HPF's and 3KHz, 15KHz and 20 KHz LPF's. So the residual FM can only be measured approximately and indirectly by this method.

First measure the residual FM with all the filters OUT and the residual FM measured is noted as f1. Then measure the residual FM with the 15KHz LPF IN and note the residual FM as f2. The residual FM with a 15 KHz HPF can be calculated as:

$$f\ (res) = \mathrm{sqrt}(\ f1^{2} - f2^{2}\ ) \qquad [eq.9]$$

f1: residual FM measured without filter
f2: residual FM measured with 15 Khz LPF
f(res): residual FM within the bandwidth 15 Khz to 200 Khz

Due to the bandwidth limitation of the filters of the HP8901, we can only measure residual FM from 15 Khz to 200 KHz. Since the Nyquist slope covers +/- 750 KHz about the carrier, the phase noise between 200 KHz and 750 KHz can not be measured by using HP8901. The phase noise in this region, depending on the SSB noise power slope, may or may not contribute to the final residual FM. In general, the total residual FM is higher than the residual FM calculated from eq.9. As a result, eq. 9 can show excessive residual FM but cannot obtain the actual residual FM.

The only way to find out the real residual FM is to measure the phase noise SSB power density and calculate the residual FM from it. The process, unfortunately, is very time consuming and error prone.

The residual FM, by definition, can be written as (note 4) :

$$f_{res}^2 = 2 * \int_{fa}^{fb} f_m^2 * L(f_m)\, df_m \qquad \text{[eq. 10]}$$

$$L(f_m) = K * f_m^x \qquad \text{[eq. 11]}$$

where f_{res} is the residual FM,

f_a, f_b are the bandwidth corners of residual FM,

$L(f)$ is the phase noise power density in dbc/hz,

f_m is the deviation frequency,

x is the slope of the sideband and usually is a negative integer like 0,-1,-2, etc., and

K is a constant.

The process for obtaining the total residual FM is illustrated by the following example: For a typical Jerrold model C5M modulator , the residual FM measured with the modulation analyzer is 202 Hz without filters and 160 Hz with a 15 Khz LPF so that the residual FM between 15 Khz and 200 Khz is calculated from eq.9 as 123 Hz. The residual FM between 200 Khz and 750 Khz is calculated as follows:

Refering to Figure 7, the carrier level is recorded as 4.5 dBm. It is clear from the figure that between 200 KHz and 750 KHz the SSB noise power follows the slope of f**-2. (noise power level reduces by 6db/octave). The level of noise at 200 KHz is -113.9 dBm/Hz. Since the noise level at 200 KHz is only 11db above the analyzer noise floor, a correction factor of 0.3 dB is needed. The resulting noise power density L(200Khz) is -113.9 -0.3 -4.5 = -118.7 dBc/Hz which is equal to 1.34E-12.

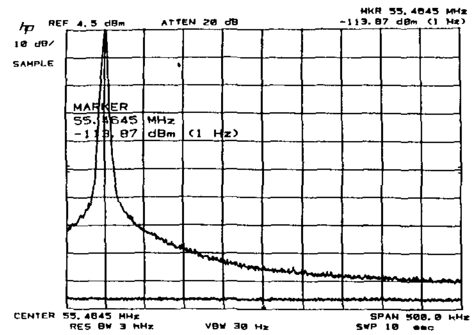


FIGURE 7
NOISE SIDEBAND MEASUREMENT

The constant K can be calculated from eq. 11 as:

$$K = L(200KHz) * 200KHz^2 = 1.34E-12 * 200000^2 = 0.0536$$

The residual FM can then be calculated from eq. 10,

$$f_{res}^2 = 2 * \int_{200E3}^{750E3} f_m^2 * \frac{K}{f_m^2}\, df_m$$

$$f_{res} = \text{sqrt}\ (\ 2 * 0.0536 * (750000 - 200000)\ ) = 243 \text{ Hz}$$

Therefore, the total residual FM from 15 Khz to 750 Khz is equal to

$$f_{(total)} = \text{sqrt}\ (\ 243^2 + 123^2\ ) = 272 \text{ Hz}$$

The total residual FM calculated above can be used to calculate the equivalent unweighted S/N. The luminance weighted S/N can be calculated from the weighted residual FM as follows:

$$f_{(total,\ weighted)} = \text{sqrt}\ (243^2 * 0.7 + 123^2 * 0.9\ ) = 228 \text{ Hz}$$

A weighting filter correction factor of 0.7 is used for 200 to 750 KHz and a factor of 0.9 is used for 15 to 200 KHz. These two factors are needed to predict the effect of the luminance weighting filter which is a kind of low pass filter. The numbers 0.9 and 0.7 are obtained experimentally and by calculating the approximate response of the filter.

The equivalent S/N due to the residual FM can be obtained by using the conversion chart in figure 6 and it is found to be 66.2 dB.

Continuing the above example, the Jerrold C5M2 frequency agile modulator has 228 Hz weighted residual FM and 68.4 dB C/N from the example of section 2. The S/N is the sum of 66.2dB (due to phase noise) and 68.1 dB (due to noise floor). Since the difference between 66.2 dB and 68.1 dB is 1.9 dB, referring to figure 1, the sum will be given by subtracting 2.2 dB from the smaller number. So the result is 64 dB. When the S/N of the modulator was measured by using Tektronix 1450, the result was 63 dB. The discrepancy can be ascribed to the contribution of the demodulator's noise floor which is 70 dB and is only 6 dB above the noise floor of the modulator S/N, causing a 1 dB degradation.

If a modulation analyzer is not available, the residual FM between 15 KHz and 200 KHz has to be measured and calculated from the phase noise SSB power density spectrum. The procedure is similar to the residual FM mesurement between 200 KHz and 750 KHz as shown in the above example.

First, the corners, or break points of the phase noise sideband spectrum slope should be identified and the noise level at each corner should be measured in dBc/Hz. Then the slope should be measured and used to calculate the constant K in eq. 11 and the residual FM in eq. 10. The constant K will not be the same for different slopes so that the residual FM associated with each segment must be calculated seperately and added by :

$$FM_{(res)} = \mathrm{sqrt}\,(f1^2 + f2^2 + f3^2 + \ldots.) \qquad [eq.\ 12]$$

f1,f2,f3 are the residual FM calculated for each different slope.

The calculation process for the exact residual FM is very tedious. To simplify the calculation, the conversion charts in figure 8 and 9 can be used to find the approximate residual FM from the SSB noise power density directly. Figure 8 and 9 are for weighted residual FM only.

To use these charts, the noise power level at 200 Khz offset should first be measured in dBc/Hz. Then the slope below and above 200 Khz should be decided. Figure 8 is the conversion chart for noises between 15 Khz and 200 Khz and figure 9 is the chart for noise between 200 Khz and 750 Khz. There are four curves in each chart for slopes 0, -1, -2 and -3. The residual FM can be read directly from these charts. For example, the noise power level at 200 Khz offset of a typical C5M2 is measured -119 dBc/Hz and the slopes are -2 both above and below 200 Khz. The residual FM, according to the chart, is 132 Hz for the noise between 15 Khz and 200 Khz, and 195 Hz for the noise between 200 Khz and 750 Khz. So the total residual FM, by eq. 11, is 235 Hz. This result is close to what we calculated before (228 Hz). The accuracy of the approximation depends on the judgment of the noise side-band slope. Usually, the slope is -3 (for -9dB/oct), -2 (for -6dB/oct) or -1 (for -3 dB/oct). For a few cases the slope is measured as -7dB/oct or -8 dB/oct). For these cases, the slopes can be approximated as -2.3 and -2.6 and the residual FM can be read between the lines -2 and -3.

FIGURE 8
WEIGHTING FACTOR = 0.9
(for F = 15 KHz to 200 KHz)

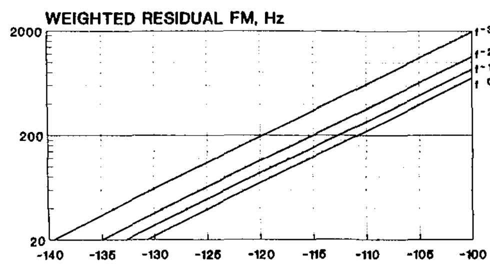
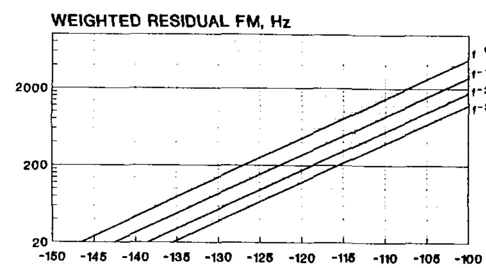


FIGURE 9
WEIGHTING FACTOR = 0.7
(for F = 200 KHz to 750 KHz)



IV. <u>LOW FREQUENCY PHASE NOISE</u>

The residual FM measured with 10 KHz LPF shows the amount of low frequency phase noise which is not included in the S/N measurement. The rms low frequency phase noises should be added to the low frequency peak-to-peak S/N. Subjectly, the low frequency noise is not like white noise which appears as grains in the picture. The low frequency phase noise looks like flickering horizontal bars due to its long period.

The low frequency video S/N can be measured very easily using the rms meter and filters. By toggling the ON/OFF switch of the 10 KHz HPF, the S/N difference between filter ON and OFF is the low frequency noise. For our sample C5M the residual FM measured with the 10 KHz LPF was 160 Hz, which from Fig. 6 corresponds to 69.8 dB low frequency S/N. Alternately, the S/N measurement with 10 KHz HPF is 64 dB and without the 10 KHz HPF is 63.3 dB. The difference of 0.7 dB results from the low frequency S/N which can be found from Fig. 1 to be 63.3 + 8.2 = 71.5 dB. The results of the phase noise calculation and the S/N measurement are reasonably close.

Many subscriber converters and low cost CATV/MATV frequency agile modulators have low frequency S/N worse than 50 db due to phase noise. Theoretically, the low frequency noise should be suppressed by the AGC system of the TV/demodulator if the noise is within the bandwidth of the AGC. But for some noisy equipments, the noise jitters so rapidly that the phase noise appears in the TV picture. To look for the low frequency noise, a flat field 10 IRE test pattern can be used. This pattern gives a very low intensity picture so that the contrast and brightness of the TV monitor should be turned to maximum. The noise can be readily seen if the ambient light is lowered and appears as flickering horizontal bars.

The phase noise can be seen more easily if a waveform monitor is available. Set the monitor to display 2 video fields (2V) and look at the sync region of the video waveform where the phase noise can be seen as low freqeumcy ripples.

The low frequency phase noise will be accumulated when equipment such as satellite receivers, modulators, FM links and subscriber converters are cascaded. In a practical CATV system, the subscriber converter is the major contributor to low frequency phase noise while the others are negligible.

V. COMPARISON

To verify the phase noise theory and calculations, four CATV/MATV frequency agile modulators were used to compare the theory with actual measurements. The noise side band spectrums of these modulators, brands "A" through "D", are shown in figures 10 through 13.

FIGURE 10 BRAND A

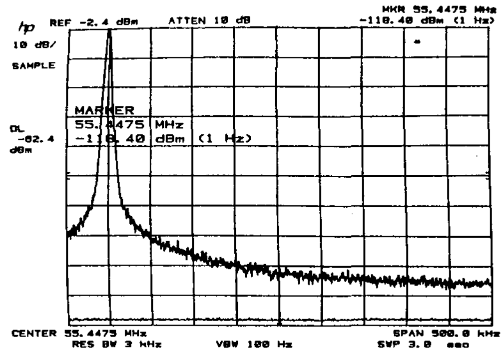


FIGURE 11 BRAND B

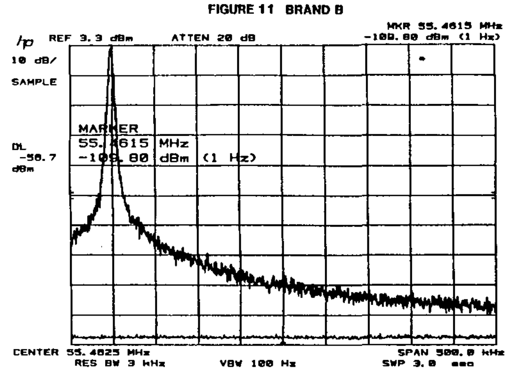


FIGURE 12 BRAND C

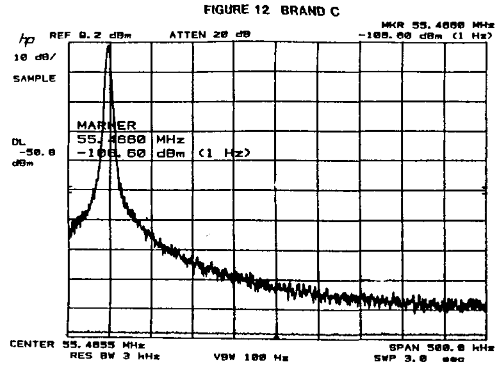
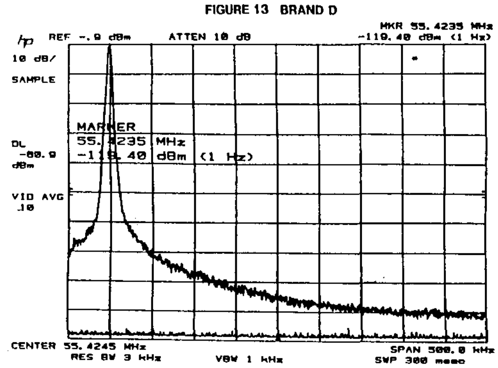


FIGURE 13 BRAND D



For these modulators, the S/N are measured first by using the rms meter and demodulator. Then the measured S/N is compared with the S/N calculated from the C/N and residual FM. Figures 8 and 9 were used to calculate the residual FM from the SSB noise power at 200 Khz offset.

The spectrum analyzer can be set up in such a way that the slopes and the SSB noise power at 200 Khz offset can be measured easily. As in figure 10, the span is set at 500 Khz so that each division is 50 Khz. The carrier is set one division from the left edge so that 200 Khz is located at the center of the screen. The SSB power at 200 Khz offset and the carrier level should be measured, converting the noise power to dbc/Hz.

The slope above 200 Khz can be found by measuring the level difference between the center and the point one division from the right edge (400 Khz offset). The slope below 200 Khz can be found by measuring the level difference between the center and two divisions to the left (100 Khz offset). This slope can be double checked by measuring the level difference between 100 Khz offset and 50 Khz offset which is one division to the left of 100 Khz offset.

The results of the calculations and measurements for the modulators are listed in the following table.

Brand	C/N	L(200Khz)	slope above	slope below	S/N (res. FM)	Total S/N	Measured S/N
A	54.7 dB	-116.2 dBc	-1	-2	60.0 dB	53.3 dB	53.8 dB
B	58.2 dB	-112.0 dBc	-2	-2	59.2 dB	55.7 dB	56.1 dB
C	61.7 dB	-115.8 dBc	-2	-3	60.8 dB	58.1 dB	58.0 dB
D	58.4 dB	-118.1 dBc	-2	-2	65.1 dB	57.3 dB	57.5 dB
C5M	68.6 dB	-118.9 dBc	-2	-2	66.2 dB	64.0 dB	63.0 dB

From these measurements we know that the calculated S/N values are very close to the measured S/N. The accuracy of the calculation is better than the authors expected. As mentioned earlier, the accuracy is decided by the judgment of slopes. If the slope is not an exact integer value then an error might occur due to approximation. It may also be necessary to compensate for the spectrum analyzer's phase noise. The phase noise power of HP8568 at 200 Khz offset is specified as 125 dbc/Hz but typically it is about 128 dBc/Hz. If the measured phase noise is below 118dbc/Hz, a correction similar to the noise floor correction in the C/N measurement (see section II) should be used.

The above measurements illustrate how phase noise can significantly degrade the S/N of a system and therefore, the subjective noise performance. As mentioned earlier, the phase noise is inherent to any oscillator and greatly affected by the Q of the oscillator. The block diagram of a frequency agile modulator system is shown in Figure 14. The modulator consists of two parts, the baseband to IF conversion and the IF to RF conversion. In the IF to RF conversion, two high frequency oscillators are needed to convert the IF to the desired channel output. The first oscillator is a fixed frequency oscillator and the second oscillator is a variable frequency oscillator (VCO) with its frequency controlled by a PLL. Since the second oscillator must cover a wide frequency range, the Q is usually low and contributes most of the phase noise. The resonant circuit of the VCO usually consists of a fixed inductor and a variable capacitor that results in the -2 phase noise slope. So if the oscillator is clean, the phase noise should follow a -2 slope. But in some cases the oscillator transistor has extra noises, especially in the low frequency region. This noise is called flicker noise and causes the close-in noise to have a -3 slope as with brand "C"

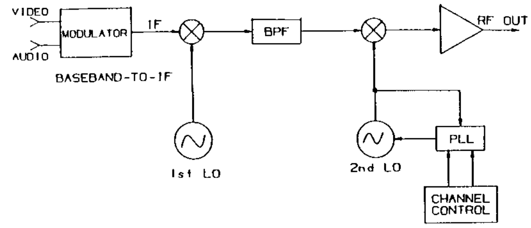


FIGURE 14 BLOCK DIAGRAM FOR FREQUENCY AGILE MODULATOR

The noise floor of the oscillator buffer and amplifiers following the second mixer should be as low as possible. This noise floor will be added to the phase noise spectrum of the oscillator. In brand "A", the phase noise slope became -1 above 200 KHz offset because of the noise floor of the RF output amplifiers.

The noise figure of the oscillator should be as low as possible. The phase noise level is directly affected by the noise figure of the oscillator. Every dB improvement of the oscillator noise figure will improve phase noise by a dB. Brand "B" is an example wherein the phase noise slope is good but the oscillator itself is too noisy.

The phase noise spectrum of brand "D" was the best among the frequency agile modulators to be tested so far. The noise slopes are -2 both below and above the 200 KHz offset and the level is as low -118 dBc/Hz.

CONCLUSION:

The C/N measurement does not completely characterize the noise performance of a CATV system. A S/N measurement is absolutely necessary for this purpose. With subscriber converters and new generation frequency agile headend modulators , the S/N can be degraded by phase noise. A high quality modulator will reduce this degradation to the minimun. For Jerrold's model C5M, with phase noise included, the typical S/N is about 64 db as required for negligible system degradation. The phase noise is not necessarily a problem for CATV systems but must be carefully considered in the design and selection of frequency agile equipment.

Note 1: MODERN COMMUNICATIONS AND SPREAD SPECTRUM, by George R.Cooper and Clare D. McGillerm, published by McGraw-Hill Company.

Note 2: THE RELATIONSHIP OF S/N IN A VHF CABLE SYSTEM TO THE S/N IN THE VIDEO SYSTEM, by Mike Jeffers, 7/85.

Note 3: TEKTRONIX 1450-1 TELEVISION DEMODULATOR INSTRUCTION MANUAL.

Note 4: NTC-7 REPORT

Note 5: DIGITAL PLL FREQUENCY SYNTHESIZERS, by Ulrich L. Rohde, published by Prentice Hall Co.

CONVERSION TABLE FOR dBmV AND POWER

dBmV	Power (W)	dBmV	Power (W)	dBmV	Power (W)
60.0	0.133E-01	20.0	0.133E-05	-20.0	0.133E-09
59.0	0.106E-01	19.0	0.106E-05	-21.0	0.106E-09
58.0	0.841E-02	18.0	0.841E-06	-22.0	0.841E-10
57.0	0.668E-02	17.0	0.668E-06	-23.0	0.668E-10
56.0	0.531E-02	16.0	0.531E-06	-24.0	0.531E-10
55.0	0.422E-02	15.0	0.422E-06	-25.0	0.422E-10
54.0	0.335E-02	14.0	0.335E-06	-26.0	0.335E-10
53.0	0.266E-02	13.0	0.266E-06	-27.0	0.266E-10
52.0	0.211E-02	12.0	0.211E-06	-28.0	0.211E-10
51.0	0.168E-02	11.0	0.168E-06	-29.0	0.168E-10
50.0	0.133E-02	10.0	0.133E-06	-30.0	0.133E-10
49.0	0.106E-02	9.0	0.106E-06	-31.0	0.106E-10
48.0	0.841E-03	8.0	0.841E-07	-32.0	0.841E-11
47.0	0.668E-03	7.0	0.668E-07	-33.0	0.668E-11
46.0	0.531E-03	6.0	0.531E-07	-34.0	0.531E-11
45.0	0.422E-03	5.0	0.422E-07	-35.0	0.422E-11
44.0	0.335E-03	4.0	0.335E-07	-36.0	0.335E-11
43.0	0.266E-03	3.0	0.266E-07	-37.0	0.266E-11
42.0	0.211E-03	2.0	0.211E-07	-38.0	0.211E-11
41.0	0.168E-03	1.0	0.168E-07	-39.0	0.168E-11
40.0	0.133E-03	0.0	0.133E-07	-40.0	0.133E-11
39.0	0.106E-03	-1.0	0.106E-07	-41.0	0.106E-11
38.0	0.841E-04	-2.0	0.841E-08	-42.0	0.841E-12
37.0	0.668E-04	-3.0	0.668E-08	-43.0	0.668E-12
36.0	0.531E-04	-4.0	0.531E-08	-44.0	0.531E-12
35.0	0.422E-04	-5.0	0.422E-08	-45.0	0.422E-12
34.0	0.335E-04	-6.0	0.335E-08	-46.0	0.335E-12
33.0	0.266E-04	-7.0	0.266E-08	-47.0	0.266E-12
32.0	0.211E-04	-8.0	0.211E-08	-48.0	0.211E-12
31.0	0.168E-04	-9.0	0.168E-08	-49.0	0.168E-12
30.0	0.133E-04	-10.0	0.133E-08	-50.0	0.133E-12
29.0	0.106E-04	-11.0	0.106E-08	-51.0	0.106E-12
28.0	0.841E-05	-12.0	0.841E-09	-52.0	0.841E-13
27.0	0.668E-05	-13.0	0.668E-09	-53.0	0.668E-13
26.0	0.531E-05	-14.0	0.531E-09	-54.0	0.531E-13
25.0	0.422E-05	-15.0	0.422E-09	-55.0	0.422E-13
24.0	0.335E-05	-16.0	0.335E-09	-56.0	0.335E-13
23.0	0.266E-05	-17.0	0.266E-09	-57.0	0.266E-13
22.0	0.211E-05	-18.0	0.211E-09	-58.0	0.211E-13
21.0	0.168E-05	-19.0	0.168E-09	-59.0	0.168E-13

dBmV = decibels relative to 1 millivolt in a 75 ohm system

TABLE 1

CONVERSION TABLE FOR NOISE FIGURE, NOISE FACTOR, AND NOISE POWER

Noise Figure	Noise Factor	Noise Power (W)
49	79432.8	0.127E-08
48	63095.8	0.101E-08
47	50118.7	0.802E-09
46	39810.7	0.637E-09
45	31622.8	0.506E-09
44	25118.9	0.402E-09
43	19952.6	0.319E-09
42	15848.9	0.254E-09
41	12589.3	0.202E-09
40	10000.0	0.160E-09
39	7943.3	0.127E-09
38	6309.6	0.101E-09
37	5011.9	0.802E-09
36	3981.1	0.637E-10
35	3162.3	0.506E-10
34	2511.9	0.802E-10
33	1995.3	0.319E-10
32	1584.9	0.254E-10
31	1258.9	0.201E-10
30	1000.0	0.160E-10
29	794.3	0.127E-10
28	631.0	0.101E-10
27	501.2	0.801E-11
26	398.1	0.636E-11
25	316.2	0.505E-11
24	251.2	0.401E-11
23	199.5	0.318E-11
22	158.5	0.252E-11
21	125.9	0.200E-11
20	100.0	0.158E-11
19	79.4	0.126E-11
18	63.1	0.994E-12
17	50.1	0.786E-12
16	39.8	0.621E-12
15	31.6	0.490E-12
14	25.1	0.386E-12
13	20.0	0.303E-12
12	15.8	0.238E-12
11	12.6	0.186E-12
10	10.0	0.144E-12
9	7.9	0.111E-12
8	6.3	0.850E-13
7	5.0	0.642E-13
6	4.0	0.477E-13
5	3.2	0.346E-13
4	2.5	0.242E-13
3	2.0	0.159E-13
2	1.6	0.936E-14
1	1.3	0.414E-14
0	1.0	0.000

TABLE 2

OFF PREMISES TECHNOLOGY COMPARISONS

Lamar West, Himanshu Parikh, Neil Robertson, Allen Childers, Mark Doremus

Scientific-Atlanta

INTRODUCTION

The last decade has witnessed a significant change in the direction of the CATV industry. The advent of satellite delivery of premium programming has broadened the appeal of CATV as a method to supply home entertainment. This, in turn, has resulted in a huge increase in the CATV subscriber base. However, in addition to this new base, it has also resulted in the need for viable denial technologies to control the dissemination of this premium programming.

A large segment of the CATV industry accomplishes the task of programming control by means of electronics that are physically located inside the subscriber's home. These electronics may consist of a set-top converter (addressable or non-addressable). The converter or other electronics inside the subscriber's home result in many problems for the system operator, such as theft, access, and subscriber education. In addition, these electronics typically aggravate the subscriber interface problem.

A number of methods have been devised to remove the electronics from the subscriber's home. In the following paper we examine several of the various methods that have been proposed to accomplish this task.

I. NEGATIVE TRAPS

The use of negative traps is one of the first methods employed for the purpose of controlling subscriber access to premium programming. A negative trap consists of a high-Q band-reject (notch) filter with a stopband frequency centered on the picture carrier of the channel(s) to be controlled. The filters are typically located at the tap port for a particular subscriber and "notch-out" the un-authorized channels for that subscriber.

One important figure of merit for negative traps is their ability to remove energy at and around the frequency of the picture carrier of the channel(s) to be controlled. Empirically it has been shown that it is necessary to attenuate the picture carrier of a CATV television channel by at least 60 dB in order to render it unusable. A second figure of merit for a negative trap is its effect on adjacent non-secured channels. It is desirable to minimize any attenuation of the lower and upper adjacent channels.

These concerns result in a requirement for excellent shape factor of the band-reject filter response. However, conventional negative traps are built using L-C filter technology. The achievable shape factors are limited by the Q available from the L's and C's used in the filter. Temperature considerations are critical, as any drift in notch frequency may result in reduced carrier attenuation. Thus conventional negative traps generally are built at frequencies below 216 MHz. Additionally, attenuation of the adjacent channels of up to several dB may occur, especially in the upper part of the usable range of channels.

An additional concern for negative trap performance is passband return loss. High return loss is desirable in order to allow the cascade of multiple traps to secure multiple premium services. The use of conventional L-C technology has limited this cascadability to three or four traps before significant passband deterioration occurs in many cases.

It is important to note that, despite these technical challenges, negative traps have, and are, being used successfully in the industry today.

II. POSITIVE TRAPS

Positive traps are electrically very similar to negative traps. However their application is significantly different. In a positive trap system, an interfering signal is inserted in the channels to be controlled at the CATV headend. This interfering signal renders the channel unusable by the unauthorized subscriber. In the case of an authorized subscriber, a positive trap is installed at the CATV tap. The positive trap is a band-reject (notch) filter designed to remove the interfering carrier while removing a minimum of the television signal information.

The technical requirements for a positive trap are very similar to the technical requirements for a negative trap. The positive trap must attenuate the interfering signal by at least 60 dB in order to render its effects invisible in the recovered picture. Typically the interfering signal is inserted at a frequency that is midway between the picture carrier and the sound carrier of the channel to be controlled. This placement gives maximum distortion of both audio and video signals (as a result of the way inter-carrier detection is used in virtually all television receivers) and occupies an area of the channel where the energy density is very low. The low energy density is important to ensure that any television signal that is removed from the channel by the positive trap will have a minimal effect on the recovered picture quality.

However, despite this placement the shape factor limitations of L-C based positive traps result in undesirable energy removal from the television channel. With the conventional placement of the interfering carrier this energy comprises the high frequency components of the luminance signal (picture detail). This is one reason why positive traps are often called "soft traps" in the industry, as they tend to soften the picture of the channels they secure. Pre-compensation for the amplitude and delay (non-linear phase shift) distortion at the head end can minimize the perceptibility of this phenomenon. Positive trap approaches are likewise limited to frequencies below 216 MHz.

As with negative traps these limitations have not prevented positive traps from being used successfully in the industry today. The economic benefits of requiring hardware at only those subscriber locations where a premium service is being received are seen to far outweigh the technical limitations.

III. ACTIVE (DYNAMIC) NEGATIVE TRAPS

A modified negative trap has been developed that uses active circuitry to solve some of the limitations of the conventional negative trap. This technique uses a notch that is too narrow to ensure frequency stable picture carrier attenuation in the conventional negative trap sense. However, this notch is slightly frequency variable by electronic means. This is accomplished by embedding a varactor diode in the notch filter topology. During operation the notch center frequency is modulated at an audio frequency rate. If such a notch is placed at the frequency of the picture carrier of a television channel to be secured, the resulting amplitude and phase modulation of the picture carrier will render that channel unusable.

The advantages of this technique lie in the requirements for the notch. As stated earlier the conventional or static negative trap requires excellent shape factor and at least 60 dB of stop-band attenuation. However it is the parasitic modulation of the picture carrier and not the attenuation of that carrier that results in the "scrambling" of the picture. A dynamic notch can obscure a channel with only 30 dB to 40 dB of ultimate rejection. Additionally this attenuation may be very narrow in frequency. The requirement of the notch frequency stability versus temperature is also relaxed with respect to static negative traps, as the notch is intentionally moved in frequency.

Conceptually the idea solves several problems that have plagued static negative traps. However, the physical embodiments of this technique have thus far proven to have serious deficiencies. The reduced shape factor requirement allows for significantly lower attenuation of the adjacent channels. Unfortunately there still remains a small amount of attenuation and this attenuation is modulated along with the notch frequency. The parasitic modulation imparted on the adjacent channels will result in visible artifacts even if the amount of modulation is only a small fraction of a dB. In addition, the introduction of any active device into the CATV system brings with it the requirement for powering with all of its associated problems.

This technique has met with extremely limited acceptance in the CATV industry.

IV. OFF-PREMISES CONVERTERS

A great deal of study has gone into the development of an off-premises (or "pole-mounted") converter. The technical challenges of this task are substantial. Attempts at developing and using an outdoor converter have met with only limited technical success.

There is, however, a more fundamental reason for why systems using off-premises converters have been unsuccessful. The main problem with this approach is that it does not solve any of the problems that lead to the examination of off premises technologies in the first place. For example:

1. It does not solve the consumer interface problem. It delivers a narrowband signal to the home.

2. It does not remove hardware from the home. Some sort of channel selection console is required in the home.

Even if a technically sound approach to this technique could be found, it is unlikely that it would offer anything of interest to the system operator.

V. ADDRESSABLE BASES

Addressable bases have been proposed to be used in conjunction with the filter techniques that were mentioned previously. The simplest form of an addressable base is an addressable service disconnect. The addressable service disconnect is useful but does not solve the problem of control of individual premium services.

More sophisticated addressable bases do permit the addressable control of individual services. However they do not solve the other technical problems associated with these techniques. Many operators see addressable bases for filter techniques as stepping stones to more sophisticated programming denial techniques.

VI. INTERDICTION SYSTEMS

Interdiction is defined in Webster's Ninth New Collegiate Dictionary as "to destroy, cut or damage". In CATV, interdiction means injecting an interfering carrier in one or multiple TV channels to deny the viewer these TV channels. The TV channels may be in any format of i.e. NTSC, PAL I, HDTV, etc.

The interdiction system is basically a negative technology. This means that some electronic device must be installed at the users locations to deny the TV channels which have not been subscribed. There are several methods to implement such electronics. The first method is to design one fixed oscillator per TV channel denied in any given system. Thus, if a system is designed to deny 16 pay channels, the number of oscillators required is 16.

The second method is to design one or more electronically tunable oscillators which will hop among all denied TV channels. So if a system is designed to deny 16 pay channels, the number of oscillators required will be less than 16. One can do a trade off between the number of pay TV channels and the number of oscillators to hop.

There are practical limitations to both methods. It is difficult to combine a large number of oscillators with a broadband signal path containing television channels in the first method. In the second method, it is difficult to control jamming oscillator side bands generated by hopping.

The most important consideration in the interdiction system is how well the system obscures the TV channel which is denied. The parameters associated with the interfering carrier are very important (i.e. dwell time, hopping rate, interfering carrier strength with respect to video carrier, placement of the interfering carrier with respect to video carrier, AM/FM side bands etc.)

The interdiction system delivers a broadband signal which is compatible with all home electronics equipment. Interconnection of the in home electronics becomes less confusing. The TV channel that is received by the authorized subscriber has not been subject to any scrambling techniques. Thus the signal contains no descrambling artifacts. The electronics associated with the interdiction system will always be off premises. So the system operator will always have direct access to and control over his investment. It will also be less intruding to the end user.

Certain technical requirements exist for the interdiction approach. Interdiction systems prevent reception of a TV channel by placing an interfering carrier (jammer) near the video carrier frequency of the channel. The jammer is injected at each subscriber's tap location, and may be turned on or off addressably to control access to the channel. Jammer level and frequency must fall within certain bounds. Level must be high enough to mask the picture, but not so high as to cause distortion in the television set or interference to adjacent channels. Jammer frequency must be located accurately within a tight window with respect to the video carrier to assure masking of the video and to prevent adjacent channel interference. There are two further benefits of locating the jammer close to the video carrier. First, severe interference to audio occurs because the jammer disrupts the intercarrier detection process within the television set. Second, it is impossible to pirate the signal by trapping-out a jammer located close to the video carrier.

As stated earlier, there are several ways to implement an interdiction system. For example, a very simple method would use several fixed-frequency oscillators to generate jammers, the oscillator frequency being controlled by crystals or perhaps SAW resonators. This approach has the disadvantage that the choice of controlled channels is fixed and limited to the number of oscillators. The cost of such a system is high if more than a very few channels must be controlled, since one oscillator is required for each channel.

Using several tunable oscillators would add flexibility in channels controlled, but one oscillator per jammed channel would still be required.

These disadvantages can be overcome by using a voltage-controlled oscillator (VCO) which hops in frequency among the channels in its tuning range. Several techniques have been proposed to do this. Conventional oscillator techniques result in significant energy at harmonics of the oscillator fundamental frequency. If not dealt with properly, these harmonics can result in objectionable artifacts in unsecured channels. One approach to an interdiction system uses a single variable oscillator to jam all channels to be secured. The frequency range of this oscillator is the upper octave of the CATV band. Placement of the frequency in this range results in harmonic energy falling outside of the CATV band where it does minimal damage. However, this approach makes the transition to interdiction technology from conventional denial technology very difficult as programming to be secured is traditionally located in the lower part of the CATV spectrum due to the reasons mentioned earlier. It can also be shown that as a single oscillator is used to secure more and more channels, the amount of scrambling of these channels is reduced.

A wide range of channels can be controlled by using several hopping oscillators. For example, four oscillators, each having a tuning range of 1.3:1 can cover all channels from 120 MHz to 325 MHz. The relative frequency range is determined by the ability of practical filters to attenuate harmonics of the oscillators to an acceptable level. The outputs of four VCOs are combined and then added to the broadband signal. An isolation amplifier prevents jammers from leaking onto the feeder. VCO frequency is controlled by a staircase tuning waveform which is generated from digital words stored in memory for each jammed frequency. Frequency accuracy is assured by calibrating these stored values using a frequency-locked loop. Finally, there must be a means of preventing the jammer from interfering with authorized channels while hopping in frequency. This function is performed by RF switches located at the output of each VCO.

A basic question regarding the hopping-jammer system is: how many channels can be securely controlled by each oscillator? (Here, secure control means that the jammer will cause any television set to lose horizontal synchronization.) As the jamming carrier hops to an increasing number of channels, it dwells a shorter period on each channel. Not surprisingly, this reduces its jamming effectiveness. Eventually, a point is reached where loss of horizontal synchronization cannot be guaranteed.

However, dwell time is not the only determinant of jamming effectiveness. Jammer amplitude and hopping rate are also important. Higher jammer level can to some extent compensate for short dwell times, but jammer level is limited by distortion and adjacent channel considerations. And there is a range of hopping rates which produces maximum jamming. Combining the factors of dwell time, hopping rate, and jammer amplitude, it has been determined that four channels can be securely controlled by each oscillator using the definition of "secure control" given above (this result has been verified on a large number of television receivers of various makes and models). Eight or more channels can be controlled per oscillator if loss of horizontal synchronization is not a requirement. Eight channels jammed per oscillator produces severe degradation of the picture and sound.

An important aspect of the interdiction system is control of jammer level relative to video carrier level. Low jammer level results in reduced security, while high jammer level can cause distortion and adjacent channel interference. AGC is thus a necessity. Two methods of AGC are possible: AGC of the video carriers and AGC of the jammer with respect to the video carrier being jammed.

AGC of the video carriers is fairly simple to implement and has the advantage of providing a stable video carrier level to the subscriber. One approach controls the average level of several carriers located near the center of the bandwidth of the VCOs. Average level variations due to temperature changes or other causes are removed. AGC of the video carriers does not remove level variation due to non-flatness of the video carriers or seasonal tilt changes. These variations are not so large, however, as to cause jammer level to fall outside the window defined by loss of secure jamming at low levels, and distortion or adjacent channel interference at high levels. Level error due to cable tilt at a fixed temperature can be avoided by the use of a cable equalizer in the interdiction unit.

AGC of the jammer with respect to the video carrier being jammed has the advantage of eliminating level error due to non-flatness and seasonal tilt changes. It is most appropriate when only one VCO is used, due to circuit complexities which result when multiple VCO levels must be controlled.

VII. POWERING

Since the use of active Off-Premises technology requires the replacement of passive tap devices with active Off-Premise devices, the system power consumption will be increased. As the larger number of active devices require an increase in the number of power supplies, "Who pays for the increase?", becomes more of an issue with cable operators.

There are basically three powering schemes to address the problem.

1. Feeder Cable Powering
2. Subscriber Drop Cable Powering
3. Hybrid Powering

As with any design, there are certain pros and cons that must be evaluated before decisions can be made. This section hopes to address some of these issues.

Conventional CATV plants are powered by 60 volt AC ferroresonant power supplies situated in the trunk and feeder lines. Additional power supplies may be required to cable power new active devices in the plant. Since most existing system ferroresonant power supplies have high current ratings (15 amps) and are loaded to near maximum, one cost effective solution is the use of smaller ferroresonant power supplies inserted in the feeder system itself. This approach minimizes system redesign requirements as well as the cost impact of the larger 15 amp ferroresonant supplies.

The Off-Premise unit itself must make every effort to reduce its power consumption and thus limit its impact. The use of DC switching power supplies (with efficiencies of 70 to 80 percent) instead of linear power supplies (with efficiencies of 40 to 50 percent) is one way of reducing power consumption. With emphasis on such parameters as power factor and component losses, much can be done in ways of optimization.

As with any aerial mounted active device, reliability (especially the ability to withstand lightning strikes and other transients) are extremely important.

The use of subscriber drop cable powering may resolve the problem of plant redesign for power. In addition it reduces power cost impacts by having the subscriber pay to power the off-premises electronics. There are several considerations associated with the use of home powering:

1. The payback of the cost of the transformer in the home

2. The possibility of local zoning requiring the house transformer be hardwired

3. Service calls for unplugged transformers

4. Access to the drop (i.e. lock boxes over ground blocks)

5. UL restrictions

6. Home circuitry needing to be hardened for transient and lightning survivability

7. Customers may try to split the signal between the home transformer and the Off-Premise unit

8. The possible uneven sharing of powering burden depending on cable drop lengths.

9. The liability from electrical shock.

Additionally, one must decide on a format for the power to be delivered over the drop cable. Two major approaches have been proposed: AC powering and DC powering.

Some of the advantages associated with AC powered drop cable are: low cost, simplicity and ruggedness to lightning. The disadvantages are: sheath currents, potential problems with power summing in Off-Premise unit, shock hazards, U.L. violations (?), and corrosion. With ferroresonant AC powering, there are additional disadvantages of the home transformer cost, size and weight.

DC powered drop cable advantages are: minimal sheath current effects, ease of power summing, maximum power transfer in long drops, low cost, fairly rugged and potentially lower shock hazard. The main disadvantage with DC powering is corrosion unless special precautions are observed.

The hybrid powering approach involves using feeder cable powering for part of the electronics while using drop cable powering for the remainder. Unfortunately, this method has the advantages and disadvantages of both powering schemes. While it is possible to argue in favor of this approach, it is not a "cure-all".

VIII. DISTRIBUTION ISSUES

As one of the main requirements of an Off-Premise is to act as a functional tap replacement, there are certain distribution system concerns that must be addressed.

One issue of immediate concern is the insertion loss of the Off-Premise unit compared with conventional taps. If the Off-Premise units were to have more through loss, additional line extender amplifiers would be needed. It is therefore desirable to make any off-premise device insertion loss comparable with conventional taps.

The power passing capability of an Off-Premise approach is a significant concern. It may become necessary to pass more current than the typical 6 amp limit of conventional taps. A 10 amp limit is suggested as a reasonable goal. It is important that hum modulation is monitored to ensure that the additional current does not create objectionable artifacts in the television pictures.

IX. MECHANICAL CONCERNS

As described elsewhere in this paper an off premises unit combines functions performed in the headend (scrambling/ interdiction), taps and set top terminals. The headend and set top terminal functions are not normally performed out-of-doors, but in this unit they are. Therefore we add many of the mechanical concerns normally associated with distribution equipment to a product that performs few distribution functions.

Although the housing required to enclose all of these functions is much larger than a tap, it should resemble a tap for ease of installation and replacement. The installation of this type of product should be no more difficult than installing a tap. To ease the design of the system a unit that has the same number of drops as a tap is also required. In addition, if the active part of the electronics are not used at a location, then provisions should be made for the unit to function exactly like a tap.

Since the unit is normally out-of-doors, a major concern is to make the housing weather resistant. Especially important is to provide a water proof enclosure for the electronics. Since temperature changes can change the air pressure inside the enclosure, the connections and the housing seal should be pressure tight to approximately 10 psi.

Being outside also subjects the unit to corrosion problems. Selecting a corrosion resistant alloy for the housing will reduce the problem in most cases. Salt air and constant immersion in water require that the unit be protected with an epoxy resin and that stainless steel hardware be used to retard the corrosion. The aluminum alloys typically used in housings may show good corrosion performance, but they do not show good electrical performance and are not used in connectors. So even when corrosion resisting alloys are used, problems associated with dissimilar metals can occur. The use of compatible materials and platings will minimize this problem.

The addressable nature of an off premises product will allow service to be electronically disconnected from the headend. Because this eliminates removing the drop as the primary way to remove service from the home, damage to the port from frequent connections is also eliminated. The permanent nature of the drop, similar to a telephone connection, will support the development of a more rugged F connector.

When fully populated, the unit will serve multiple customers. But, if the unit is designed for modular functions, it is not necessary to fully populate the unit to bring only one customer on line. A unit with field replaceable modules will reduce the time that a subscriber would be without service in case of an electronic failure. If a major failure were to occur it is also important to be able to remove and replace the entire unit very quickly to restore service with a minimum of interruption.

As the unit is larger than a standard tap, how to mount it is an important question. The design of the housing should follow the guidelines used for most strand mounted equipment. Properly designed, this equipment should not cause any problems in an aerial mount. For pedestal mounts the design should be such to allow use of a standard pedestal enclosure. As the unit incorporates functions not usually found in a tap, it may be necessary to use a pedestal enclosure slightly larger than that used for conventional taps. An eight inch enclosure, typically used in amplifier applications, should be the largest enclosure used to house this unit. This pedestal imposes a maximum length constraint on the unit of 16 1/2 inches.

Since use of interdiction technology allows the transmission of the original signal in the clear, use of a secure housing is extremely important. In an unsecured housing the clear signal is accessible to a pirate. The housing developed for an off premises product should have provisions for a mechanical lock or tamper evident seal. Electronic protection that shuts off service when the housing is opened without authorization should also be included.

The heat generated by the various power supplies used to support the electronic functions should be considered in developing the outside housing. Adequate mountings for transferring heat from the modules to the housing should be included. The exterior of the housing should be designed to insure sufficient heat transfer to the surrounding air to keep the unit with in its specified operating temperature range.

Besides the purely mechanical functions previously described, the housing has to provide several electronic functions. Attenuation of RFI/EMI signals to acceptable levels and grounding of power and signal voltages are important functions of the enclosure. In addition to the water seal, a conductive gasket should be used to assure that a conductive path exists between the halves of the housing.

CONCLUSIONS

Several methods have been described that will accomplish the task of programming denial. Each of the methods has its own technical strengths and weaknesses. It is important for the system operator to weigh these technical concerns against the economic benefits associated with each technique.

As work and development continues in our industry, there will most certainly be new additions to the methods described herein. As responsible members of the CATV industry we should all review these methods to determine which will provide the most satisfying service to the ultimate customer, the subscriber.

OFF-PREMISES BROADBAND ADDRESSABILITY: A CATV INDUSTRY CHALLENGE

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Abstract - A marketplace need for automating control of broadband CATV signal delivery is described. Past and current efforts to produce equipment meeting this need are outlined, and some concepts for future approaches are suggested. The economic forces at play in the implementation of such a system are described, along with an approach for modeling the operating cash flow needed to offset the required capital investment. The conclusion is drawn that a need for such a delivery technology does exist, and is likely to grow as competitive forces increase the cable industry's need to improve compatibility with consumer electronic equipment, deliver an increasing number of switched video (pay-per-view) services, and control operating expenses. Meeting this need is seen to involve meeting significant technical and economic challenges.

INTRODUCTION

Over the last several years the cable industry has been undergoing an agonizing reappraisal of the role which addressability should play in its operating systems. While there is not yet industry consensus, the outcome of this debate will be a major factor in determining cable's future. On the one hand, some operators are moving aggressively away from addressability, finding refuge in the simple negative and positive trap technology which initially built the pay TV business. Other operators are moving more aggressively into addressability because of their belief in the future of pay-per-view services.

The original dream of addressability encompassed automated delivery of multi-pay and pay-per-view, operating savings from reduced truck rolls, reduced converter losses, and the ability to market more flexibly. In retrospect, we see a number of unanticipated problems. Addressability introduced additional layers of complexity to virtually all operational aspects of our systems, and there were varying degrees of success in coping with this. Some vintage addressable converters were unreliable, wiping out potential operating savings and angering subscribers to boot. While most addressable set-top units being delivered today have achieved acceptable reliability, these problems will be with us for some years in our universe of older converters.

Additionally, the multi-pay environment did not require the number of channels once expected. Three or four services appear to meet the needs of most markets and trapping is often a viable delivery option. Problems with consumer friendliness, which resulted from the introduction of scrambled signals at the same time that "cable-ready" consumer equipment was being introduced in volume, were largely unforeseen, but have growing significance. According to research done by ATC over a large sample, over 52 percent of cable subscribers own cable-ready equipment and over 68 percent have VCR's. As an industry, we have not been particularly successful in addressing the resulting issues.

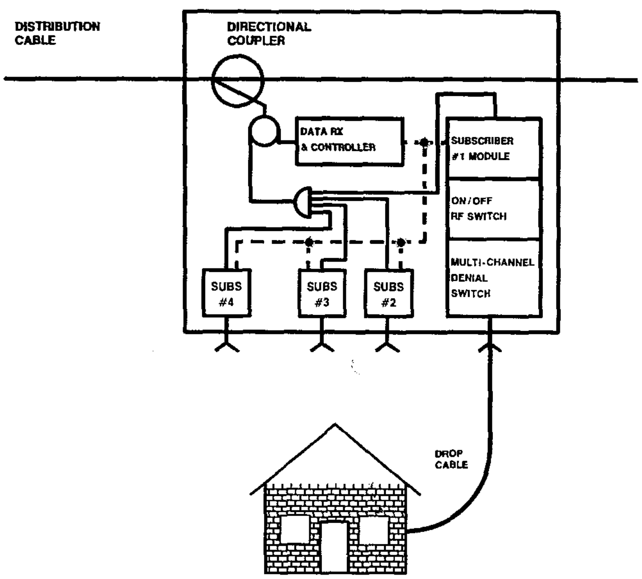
The experience of a number of operators indicates that there is additional revenue available from pay-per-view, although the magnitude remains unclear. In addition, our most likely long-term competitors, who will employ direct broadcast satellites and switched telco delivery systems, may well be capable of pay-per-view delivery to all of their subscribers. Thus, to the extent that pay-per-view offers things that consumers want, our moving away from addressable technology may put us at a competitive disadvantage.

The operating economies which are an unrealized part of addressability's potential are more important than ever. This is true in improving present-day margins, as well as in positioning for future price competition. In addition, skilled labor will continue to become more expensive and increasingly scarce in years to come.

We also need to capitalize on the proliferation of cable-ready equipment, with its potential to decrease the need for capital investment inside the home ATC's research indicates that 52% of cable subscribers have cable-ready TV sets currently. Further, the consumer expects us to be compatible with the equipment he purchases. While traps can satisfy the need for broadband, unscrambled delivery to the home, in the long term it is important that we explore ways to combine this feature with addressability.

The heart of the challenge is the separation of scrambling from addressability, and the provision of unscrambled, broadband signal delivery under addressable control. A generalized approach which would meet these goals is shown in figure 1. This represents a device located outside a subscriber's premises which would allow broadband unscrambled delivery of all services ordered by that subscriber. The device would have the ability to turn "off" or "on", and to intercept premium services not ordered by the subscriber. This would allow a subscriber to use any cable-ready equipment he might own, and to receive all services to which he subscribed at all outlets within the home. Any TV or VCR not having the channel tuning capabilities necessary to receive this service would, of course, need an RF converter. However, subscribers with cable-ready equipment would not need any additional equipment inside their homes. The cable operator would have full control over each subscriber's reception remotely.

FIGURE 1



The system outlined would behave very much like a current CATV system with individual channel traps except that customer connection, disconnection, and changes in authorized services would be fully automated. This would have a number of implications. First, there would be an opportunity to substantially reduce operating costs through the elimination of physical visits to the subscriber's home in order to change to status of his service. This would be further enhanced by an expected increase in drop reliability due to a dramatically reduced need to physically handle drops. Drop cables, once installed, could be permanently secured and waterproofed, removing a major cause of future service calls.

The system would have the positive consumer equipment interface aspects outlined above, avoiding a significant cause of subscriber dissatisfaction in systems that currently use scrambling as a means of signal security. The cable company would reduce the amount of equipment necessary inside the home, which would result in a decrease in related capital and operating expenses as the universe of cable-ready equipment continues to increase.

In addition to reducing operating costs and improving customer satisfaction, the system outlined would also be capable of providing pay-per-view services to any subscriber. Marketing flexibility would be increased with the ability to demonstrate cable's products for any period of time desired.

Finally, such a system begins to set the stage for the future. The ability to authorize "slices" of spectrum leaves open the door to controlling potentially non-standard HDTV signals. In addition, this form of addressability would, in essence, provide distributed video switching, which could, if combined with switching elsewhere in the network, ultimately result in selective delivery of video to individual homes.

TECHNICAL CHALLENGES

While there are a number of conceptual approaches to realizing off-premises broadband addressability, implementing such technology in a practical way poses a number of challenges. Clearly, an outdoor device can be built with the capability of turning a drop off and on under remote control. The control system would, in fact, be very much like that used for addressable descrambling systems today, and PIN diodes or relays could serve to disconnect an unauthorized drop with sufficient signal isolation. Additionally, there are a variety of approaches available to selective delivery of individual channels to the subscriber drop cable. These include fixed frequency and frequency-agile positive and negative traps, as well as various types of fixed and frequency-agile jamming signals to be summed with individual unauthorized channels.

The challenges in realizing a practical off-premises broadband addressable system arise from the need to deliver unimpaired signals on authorized channels, to remove or disrupt video information from unauthorized channels sufficiently to prevent practical use, and to prevent defeat scenarios which would involve signal processing inside the home. In addition, powering a large number of active devices in the CATV system is not a trivial matter. If they were to be powered from the CATV plant, it is likely that a substantial increase in system power supplies would be required, necessitating a significant capital investment in power supplies, in adding to increased on-going power expenses.

An additional concern is the maintenance and reliability implications of adding a large number of active devices to the network in a hostile physical and electrical environment. While this is partially true of addressable set-top converters as well, it should be remembered that a number of years passed before satisfactory reliability was achieved with those devices. Prior to that time, significant expense and subscriber disruption was caused by converter malfunction and failure. If off-premises broadband addressable devices cannot be produced with very high long-term reliability, it is clear that any operating cost reductions will be more than offset by maintenance costs, and subscriber satisfaction gains created by compatibility with cable-ready equipment will be destroyed by dissatisfaction due to service disruptions.

Thus, the goal of mass-producing an affordable, practical device for selective broadband signal delivery located outside the home, with a high degree of reliability, is a major challenge. This is further exacerbated by the hostile environment in which such a device must be placed, with the hazards of moisture, wide temperature variations, and electrical discontinuities caused by power utility fluctuations and surges. This challenge has, in fact, defeated several attempts in past years to produce such equipment.

PAST APPROACHES

The attraction of off-premises addressability is not new. A system was developed by AMECO in the late 70's which utilized relays along with a data receiver in a line extender housing to produce an off-premises addressable tap. Latching reed relays were used to turn off and on individual subscriber drops, and to switch in and out a single negative trap on each output. The system was field tested, but was never implemented on a large scale, possibly due to the advent of multi-pay services at about that time as well as, it is surmised, concerns about cost-effectiveness.

In the early 80's, an addressable tap was marketed and was installed in a few cable systems by Delta-Benco-Cascade. The system was sold in both an outdoor, four-port addressable tap configuration, and an addressable wall-plate configuration for loop-wired multiple dwelling units. The DBC system used phase modulation of the AC powering waveform to transmit data from each power supply location to each tap or wall plate. This allowed the construction of an exceedingly simple data receiver within each tap, with a more complex RF data receiver located at each power supply receiving addressable instructions from a computer at the headend.

The DBC addressable taps could turn signals on and off, using PIN diode RF switching, as well as control two pay channels, using a negative and a positive trap. The product was ultimately discontinued and all known installations were dismantled, due to reliability problems with both the tap units and the data modulated power supplies. This is a clear illustration of the lack of reliability destroying any possible operating cost savings.

During the early to mid-80's, a variety of off-premises converter systems were developed and tested. These included the DST system developed by ATC and Toshiba, Texscan's TRACS converter system, C-COR's SCAT system, and Times Fiber's Mini-Hub I and II (Mini-Hub I used multi-mode optical fiber for the connection from the addressable converter to the home). While these approaches differed in specifics of powering, design and construction, the essentials of an addressable set-top converter were located outside the home, with only a control head at the television set. Up-stream signals from the control head instructed the external converter as to which channel to tune, and a single channel was delivered downstream to the television. Sometimes provision was made for several control heads and converters to share a single drop, using several channels. The external converter electronics contained a data receiver which received authorization information from the headend. All of these systems were field tested, and some were installed in some quantity in operating cable systems.

The introduction of these systems coincided with an increasing proliferation of cable-ready consumer equipment. These systems shared all the consumer interface drawbacks of addressable descrambling converters, and most lacked any ability to deliver broadband signals to the home for use by cable-ready TV sets and VCR's. In addition, the electronics moved outside the home were the inner-workings of a highly complex RF hetrodyne converter, and most systems had a variety of reliability problems. Consumer interface problems and the failure to realize operating economies proved fatal to these approaches, and all have been, or are being, discontinued from production and removed from service.

Thus, attempts, to date, to accomplish practical off-premises addressability have been defeated by failure to achieve cost-effective operation on a scale which justifies the capital expenditures involved, and, in the case of off-premises converters, to provide sufficient subscriber utility. The lessons which appear to have been learned are that broadband signal delivery from an off-premises device is important, both in terms of consumer interface issues and achieving a practical level of the simplicity, and that reliability is an absolutely critical factor in implementing this technology. Experiences with powering such devices from the cable system clearly involved high costs for additional power supplies, and for the substantial number of kilowatt-hours required. Approaches which used powering of the drop from the home avoided those problems but necessitated accessing the home, an additional source of trouble calls due to subscribers' inadvertently disconnecting power.

CURRENT APPROACHES

There are three basic approaches to off-premises broadband addressability currently available commercially. The first involves "signal interdiction" in an addressable tap at the pole. Variations of this are offered by AM Communications and Scientific Atlanta. In both cases, the pole mounted tap includes a data receiver and a jamming oscillator or oscillators which are frequency agile, and can be selectively switched onto a subscriber output port. In the case of the AM Communications product, a single oscillator can frequency hop to as many as sixteen channels, while Scientific Atlanta employees four frequency agile oscillators which can cover a larger number of channels. In both systems there is a clear tradeoff between the number of channels which share an oscillator, and the level of security and "signal masking" on unauthorized channels. Both allow a decrease in the number of channels sharing an oscillator to allow better masking of particularly controversial programming. While there are differences in features and costs of between the two systems, both are currently being installed, or will be installed in the near future, in working systems. This will hopefully result in the capture of meaningful data about their reliability and the actual operating savings realized.

A second approach to broadband addressability which is currently being offered has been termed "on-premises addressability" This approach essentially automates positive and negative traps at a location outside each home, as opposed to being located at the pole or equipment pedestal. In these approaches, a data receiver controlling PIN diodes, turns the drop off and on and switches positive and negative traps in and out the circuit. They receive their power from inside the home, and can be located in an environment less hostile than that of pole-mounted equipment. Advantages of this approach include an incremental investment which can be selectively deployed against subscribers most likely to order pay-per-view services, or against some other rational. Drawbacks include the inability to share system costs across more than one subscriber and concerns about physical security.

A third category of addressable broadband addressability is being offered for the multiple dwelling unit environment. These generally are capable only of turning individual drops on and off remotely, and do not address the control of pay services. This technology is relatively simple, with the cost of the data receiver being shared by many subscribers. These units seem to be finding utility in multiple dwelling units with a high degree of subscriber turn-over, especially in resort and university environments.

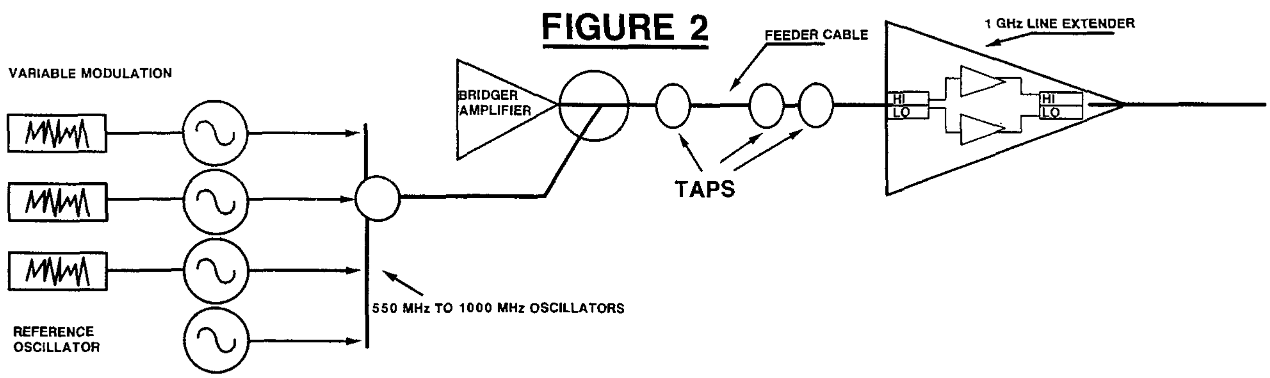
ALTERNATE APPROACHES

In examining other possible approaches to off-premises broadband addressability, the goals are to shed complexity and to share costs, while maintaining the ability to turn off and on individual subscriber drops and to control a reasonable number of individual channels. Figures 2 and 3 show such an approach. In this approach, a number of jamming oscillators, at frequencies well above those of the channels delivered by the CATV system, are located at the bridger amplifier. These are modulated to provide a high degree of video and audio masking to channels with which they are ultimately mixed. Also located at the bridger amplifier location is an unmodulated master oscillator, also well above the frequencies of interest in the system. These signals are combined with the bridger output, and are transported through distribution at high frequency. This requires tap electronics capable of passing frequencies perhaps as high as 1 GHz. It also requires that line extender amplifiers make provision for amplifying these frequencies. Because noise and distortion are not of great concern with regard to these signals, a separate amplifier stage for the high frequency jamming signals could be used within line extenders, in addition to a high-performance broadband amplifier for the CATV signal spectrum.

Figure 3 shows the inner workings of the tap. Switched notch filters are used to turn off and on individual jamming oscillators. The master oscillator frequency is recovered and applied to a mixer, hetrodyning the jamming oscillators down to their final frequencies within the CATV band. The summing of the switched jamming frequencies with the CATV spectrum results in a broadband signal to the subscriber with unauthorized channels obliterated. Notch filtering of jamming signals could also be performed after down conversion. This approach would allow one oscillator per channel, since the cost of oscillators would be shared across many subscribers.

Figure 4 shows another possible arrangement, using a jamming oscillator at 74 MHz, between channel 4 and channel 5, located at the headend. Within the tap, this jamming frequency would be divided by two and applied to a comb generator which would generate multiples of 37 MHz. This would result in interfering carrier frequencies

FIGURE 2



at 111 MHZ, 148 MHz, 185 MHz, 222 MHz, etc. These jamming frequencies could then be selectively filtered before being combined with the CATV signals to each subscriber. Thus, a degree of system simplification could be achieved at the cost some inflexibility regarding the channels used for premium services.

FIGURE 3

TAP DIAGRAM

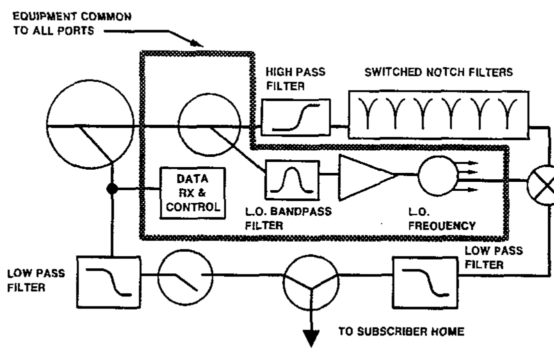
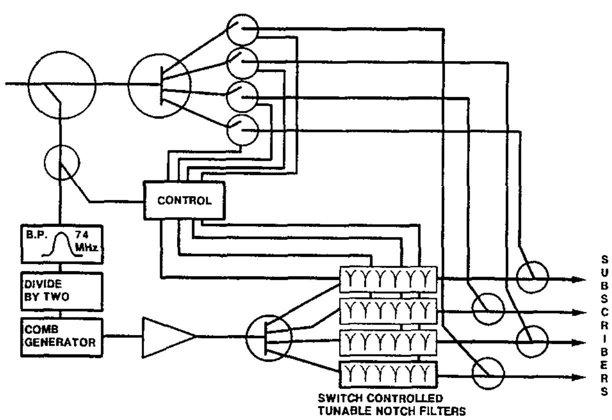


FIGURE 4



These are but a few of the possible architectures for use in an off-premises broadband addressable signal delivery system. Since such systems incur significant penalty for both initial capital expense and complexity, there is a premium to be obtained in simplifying the system and spreading the cost of expensive components or subsystems across a number of subscribers.

ECONOMICS

There are a number of positive and negative forces at work when we examine the cost-effectiveness of off-premises broadband addressability. Economic modeling of the equilibrium between these forces can become highly complex, as there are many variables. While only field experience will resolve some of these issues, it is worth examining key factors in building an economic model.

Reduction Of Subscriber Visits

This item has potential to be a major justification for the installation of off-premises addressability. It is assumed that an off-premises broadband addressable system would eliminate the need for most visits to the home. Once an installation had been performed, future disconnections, reconnections, and changes in level of service would be automated. The cost of rolling a truck to a subscriber's home is estimated to be between 20 and 30 dollars per visit. Basic churn in most cable systems is between 1 and 3 per cent of subscribers per month. It is may be assumed that installation of outlets in additional or different rooms in a subscriber's home would be billed at cost and would therefore, be cash flow neutral.

Universal Pay-Per-View

One obstacle to the growth of pay-per-view has been the limited number of homes in most systems which have addressable converter/descramblers. In systems which use set-top addressability, it can be argued that most potential pay-per-view subscribers also subscribe to scrambled pay services, but that hypothesis is untested. Additionally, it is clear that major pay-per-view events, such as boxing and wrestling matches, with their substantial revenue potential, could sell to a wider audience if a delivery mechanism were in place. When compared with a trapped system, off-premises broadband addressability has significant revenue potential in terms of pay-per-view.

There is no consensus in the cable industry about the size of this potential revenue, but it is an important factor to be examined in modeling off-premises addressability.

Consumer-Friendly Broadband Delivery

Systems which employ addressable scrambling, as opposed to trapping, in order to control selective delivery of pay television or pay-per-view product provide a fair degree of frustration to that majority of their subscribers which have cable-ready consumer electronics equipment. If there is any benefit to be gained from improved subscriber satisfaction, off-premises broadband addressability should capture it. Such a benefit should take two forms. First is an economic advantage, in the form of improved retention and, therefore, penetration. This is a difficult effect to isolate from other factors in subscriber penetration, and is a potentially large but difficult factor to use in economic modeling. The second advantage of improved utility of consumer electronics is strategic. With a variety of alternative video delivery systems on the horizon, cable's strategic ends are not well served by providing a source of subscriber frustration.

Reduced Set-Top Converter Capital Investment

When an off-premises broadband addressable system is compared with a set-top addressable descrambling system, the off-premises system has a clear advantage in its ability to benefit from cable-ready consumer equipment in the reduction of the set-top converters needed in the system. Since set-top addressability requires a device in the home regardless of the kind of television set the subscriber owns, and since the number of cable-ready TV's and VCR's is steadily increasing, a system using off-premises addressability should show a decreased need in future years for set-top converters. In addition to gradually reduced set-top capital requirements, elimination of converters from an increasing number of homes decreases the need for service call, and converter delivery and pick up. Additionally, this would result in fewer unretrieved converters.

High Capital Cost

Off-premises broadband addressable signal delivery systems currently available have an installed capital cost between $75 and $125 per subscriber. This represents a very significant incremental investment, and we can reasonably expect to make it only if offset by sufficient benefit.

Powering

Powering from the home involves no incremental additional power cost, but does involve accessing in the home for the installation and maintenance of a low voltage power supply and power inserter. This is somewhat at odds with the goal of using off-premises addressability to reduce operating costs and subscriber contact. Such a scheme also increases the capitalized investment necessary to implement an addressable system. Powering from the plant has the potential of requiring many additional power supplies. This item is highly dependent upon power consumption of the addressable devices, and provides a powerful incentive for developers to minimize power requirements.

Maintenance

Even though off-premises broadband addressable taps are conceptually quite simple, the fact that they would be deployed in very large numbers has potential to have enormous impact on system maintenance economics. In a sample design of a 3,000 mile plant, 105,000 active addressable taps were found to be required. Thus, there is substantial economic impact from anything but exceedingly high device reliability. In addition, any lack of reliability will result in a loss of subscriber satisfaction.

Economic Modeling - An Approach

A practical means of developing a feel for the economic trade offs involved in installing an off-premises broadband addressable system can be derived from examining the annual incremental cash flow requirements necessary to provide a reasonable internal rate of return (IRR) against incremental capital required for the installation of the system. In the following example, the assumption was made that the existing system used traps for signal security, and was in need of a major plant upgrade, involving splicing in new system taps throughout. Thus, no incremental labor was included for the installation of addressable taps. Figure 5 shows basic statistics regarding the system sampled.

FIGURE 5

SAMPLE CATV SYSTEM
($ X 1000)

280 MILES OF PLANT (TRAPPED FOR SECURITY)
21,550 PASSINGS
16,500 BASIC SUBSCRIBERS
10,500 PAY UNITS

REVENUE	
TOTAL BASIC	**$3,183**
TOTAL PAY	**1,197**
MISCELLANEOUS REVENUE	**227**
TOTAL REVENUE	**4,607**
COST OF SALES (PROGRAM COST)	**735**
OPERATING EXPENSES	**1,615**
TOTAL CABLE CASH FLOW	**2,257**

The following assumptions were used for the modeled off-premises addressable system:

The subscriber unit would be made up of two pieces. The housing and back plane would have the potential to serve for subscribers, and would cost $150. Additionally, one subscriber module would need to be added for each active customer served. These modules would cost $50 each. It was also assumed that the unit could be driven by standard tap input levels, so a system would require the same number of active as a non-addressable system. It is further assumed that this system would be powered from the home, at an installed cost of $10 per home.

It is also assumed that the increased maintenance cost from the installation of over 7,000 additional, but highly reliable, active devices is offset by the service call savings resulting from decreased drop handling and the ability to permanently weatherproof drops.

Over time, with churn, it is assumed that 88% of homes passed would be installed, requiring capital investment in off-premises modules.

Although each tap is capable of serving four homes, it is assumed that the design is 70% efficient; that is that 30% of the tap outputs (4 per device) will be unused, on average. This means that the 21,550 passings will require 7,395 devices to be installed.

It can be seen in the highly simplified example in Figure 6 that the installation of off-premises broadband addressable taps in this previously trapped system, during its normally scheduled upgrade, results in a reasonably favorable economic scenario driven primarily by operating economies and contributed to by pay-per-view revenues. If such an installation were contrasted with addressable set-top converter/descramblers, subscriber satisfaction and converter capital reduction elements would be introduced, but the pay-per-view benefit would be reduced, since that capability exists with set-top addressables as well.

FIGURE 6

SAMPLE SYSTEM ECONOMICS

CAPITAL INVESTMENT REQUIRED

HOUSINGS & BACK PLANES	
7,395 LOCATIONS X $150	**$1,109,000**
SUBSCRIBER MODULES	
88% X 21,550 PASSINGS X $50	**948,000**
POWER SUPPLYS	
88% X 21,550 X $10	**190,000**
TOTAL	**$2,247,000**

ANNUAL INCREMENTAL

CASH FLOW REQUIRED FOR 10YR IRR OF 10%

$365,000/YR

CASH FLOW GENERATORS

TRUCK ROLLS SAVED FOR DISCON & RECON	
30% CHURN X 16,500 SUBS	
X $20/TK. ROLL X 2 TK. ROLLS	
(DISCON & RECON)	**= $198,000/YR**
TRUCK ROLLS SAVED FOR SPIN	
10,500 PAY SUBS X	
20% SPIN X $20/TK. ROLL	**= $42,000/YR**
20% PPV REVENUE/MONTH X 12 MO X 16,500	
SUBS X $2 NET/TAKE	**= $79,200/YR**
	$319,200/YR
ACTUAL IRR =	**7 %**

The example illustrates the difficulty of viewing off-premises broadband addressability as a highly attractive investment in terms of direct pay back. However, when viewed in the context of a more competitive environment the argument for its installation becomes far more compelling. Clearly, reduction in the hardware cost, or a more aggressive view of Pay-Per-View revenue potential would have a major favorable impact.

SUMMARY

We've seen that the quest for an improved cable television signal delivery system leads us to seriously examine off-premises broadband addressable delivery of our services. We have also seen that there are significant technical and economic challenges in our path as we seek to realize hardware which would meet these goals. There is clearly a substantial reward to the cable industry in finding such a solution. It is hoped that in working with potential manufacturers of such hardware, the industry as a whole can realize the goal of reduced operating costs, increased subscriber satisfaction with our service, enhanced pay-per-view revenues, and a network which is better positioned for a more competitive future.

ACKNOWLEDGEMENTS

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Louis Williamson - ATC

Jim Chiddix - Senior Vice President, Engineering and Technology, for American Television and Communications Corporation, the country's second largest cable television operator, headquartered in Stamford, Connecticut. Mr. Chiddix is responsible for corporate engineering activities as well as research and development. ATC serves 3.9 million subscribers in 32 states and is 82% owned by Time, Inc. Upon completion of its pending merger with Warner Communications, which also owns cable TV operations, the combined companies will serve 5.6 million homes.

ATC leads the cable industry in exploring the use of optical fiber technology in cable television systems. Their "fiber backbone" concept for optical trunking has gained wide acceptance as an evolutionary approach, offering the prospect of improved performance and increased channel capacity from existing cable systems. In recognition of his pioneering role in exploring this use of fiber, Mr. Chiddix was named Man Of The Year by Communication Engineering and Design Magazine in January, 1989.

Mr. Chiddix, 43, has been in the cable television business for 17 years. He spent seven years as General Manager at Cablevision, Inc. in Waianae, Hawaii, and eight years as Engineering Vice President at Oceanic Cablevision in Honolulu, an ATC division. In September, 1986, he joined ATC's corporate office.

Mr. Chiddix is a Senior Member and former Director of the Society of Cable Television Engineers (SCTE). 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.

Dave Pangrac is the director of engineering and technology for American Television and Communications Corporation (ATC), the country's second-largest cable television operator.

Pangrac has been in the cable television business for 22 years. He joined ATC in 1982 as Vice President and chief engineer for American Television of Kansas City and in 1987 joined the ATC corporate staff as director of engineering and technology.

Pangrac is a member of the Society of Cable Television Engineering and past president of the Hart of America Chapter.

Pangrac is currently involved in ATC's effort to develop the use of fiber optic technology in cable television plants.

OPERATION ISSUES

Thomas G. Elliot

Tele-Communications, Inc.

These are exciting times for the CATV industry. We are able to offer our subscribers more and more original programming. Fiber optics and HDTV are hot topics. Cable Labs is off and running.

Behind all of this excitement, we must not forget what got us here - **OUR CUSTOMERS.** There are numerous areas that we can address that will provide better service to our subscribers and lower our operating costs. Several items to think about are:

1. Interfaces
2. Installation procedures & practices.
3. Equipment standards
4. Maintenance procedures
5. Performance budgets

Interfaces continue to be the root cause of most of our problems. RF leakage, intermittents, interruptions, outages, water migrating into our plant, corrosion problems, and on and on are caused by bad design, bad installation, or bad maintenance of interfaces. This problem is expensive for us and very irritating to our subscribers.

Proper construction and installation procedures and practices, coupled with properly designed materials will go a long way toward eliminating many of our problems. This is an area too often taken for granted, where our field people are forced to "live with" what they have. We must focus more attention in this area.

Equipment standards are important for many reasons, not the least of which is to avoid confusion. Why do we still have some equipment with 20 dB test points, some with 30 dB test points, some with resistive probes, some with direct probes, etc.? This leads to major confusion and difficult training problems.

Maintenance procedures are slowly getting better, but not much. How many of our people really understand cable slope, proper equalization, block versus linear versus cable tilt, how to properly set up an amplifier for slope and gain over a wide temperature range, etc.? We have not worked with the different suppliers of actives and test equipment to normalize tilts, levels, and set up procedures so that we can simplify our maintenance people's lives and properly take advantage of microprocessor-based test equipment.

We must start thinking of the image transfer process in total. In other words, what is the end-to-end performance required from the point the image is generated to the point it is displayed? An overall performance budget must be established and then broken down into budgets for each part of the system. (A sample noise budget is attached.) This budget process is essential if we intend to improve the quality of the images we deliver to our customers.

Fortunately, all of the things we need to do to improve our systems for today's use are well understood. We just need to do it. The picture is not as clear as we move toward the future, as our consumers acquire better TV sets and VCR's, with BTSC stereo, with HDTV on the horizon, with consumers demanding better quality over time. However, the best way to get there is to do the best possible job now, while keeping an educated eye on where we are going.

THEORETICAL C/N AND S/N

Calculations are based on:

Headend C/N	**60**	**dB**
AML System		
Output Power	7	dBm
TX and RX Antenna	48.8	dB
Path Length	12.02	miles
AML C/N	53	dB
Distribution		
Trunk Amps		
Cascade	30	
Noise Figure	9	dB
Input level	10	dBmV
Trunk C/N	45.23	dB
Bridger Amp		
Noise Figure	8	dB
Input Level	20	dBmV
Bridger C/N	71	dB
Line Extenders		
Cascade	2	
Noise Figure	9	dB
Input Level	20	dBmV
Line Extender C/N	66.99	dB
Distribution C/N	**45.2**	**dB**
Converter C/N	**53**	**dB**
Television C/N	**55**	**dB**

Theoretical subscriber C/N is:

$$C/N = 10 \log \left(\frac{1}{\frac{1}{\left(\frac{HE\ C/N}{10}\right)^{-1}} + \frac{1}{\left(\frac{AML\ C/N}{10}\right)^{-1}} + \frac{1}{\left(\frac{DIST\ C/N}{10}\right)^{-1}} + \frac{1}{\left(\frac{CONV\ C/N}{10}\right)^{-1}} + \frac{1}{\left(\frac{TV\ C/N}{10}\right)^{-1}}} \right)$$

System C/N = 43.5 dB and System S/N = 43.2 dB
To get the actual Signal to Noise (S/N), the consumer views the camera, tape, and satellite S/N's must be added.

Camera S/N	55 dB
Tape S/N	55 dB
Satellite S/N	52 dB
System S/N	43.2 dB

$$\text{Subscriber } S/N = 10 \log \left(\frac{1}{\frac{1}{\left(\frac{CAM\ S/N}{10}\right)^{-1}} + \frac{1}{\left(\frac{TAPE\ S/N}{10}\right)^{-1}} + \frac{1}{\left(\frac{SAT\ S/N}{10}\right)^{-1}} + \frac{1}{\left(\frac{SYS\ S/N}{10}\right)^{-1}}} \right)$$

Subscriber S/N = 42.2 dB

1/25/87 TGE

PAY-PER-VIEW (PPV) COPY PROTECTION IN THE CABLE SYSTEM ENVIRONMENT

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Eidak Corporation
Cambridge, MA

ABSTRACT

Video copy protection has become an essential component of cable's PPV technology infrastructure necessary for PPV to attract the competitive programming required for healthy growth. A specially designed modification of the PPV video signal assures non-recordability of copy protected programming, but still permits normal viewing operation of television receivers. The signal modification is optimized for compatible operation with cable system plant, with particular emphasis on the addressable descramblers used to control viewing of authorized PPV programs. Particular emphasis is placed on copy protection throughout cable plant, including distribution to multiple hubs.

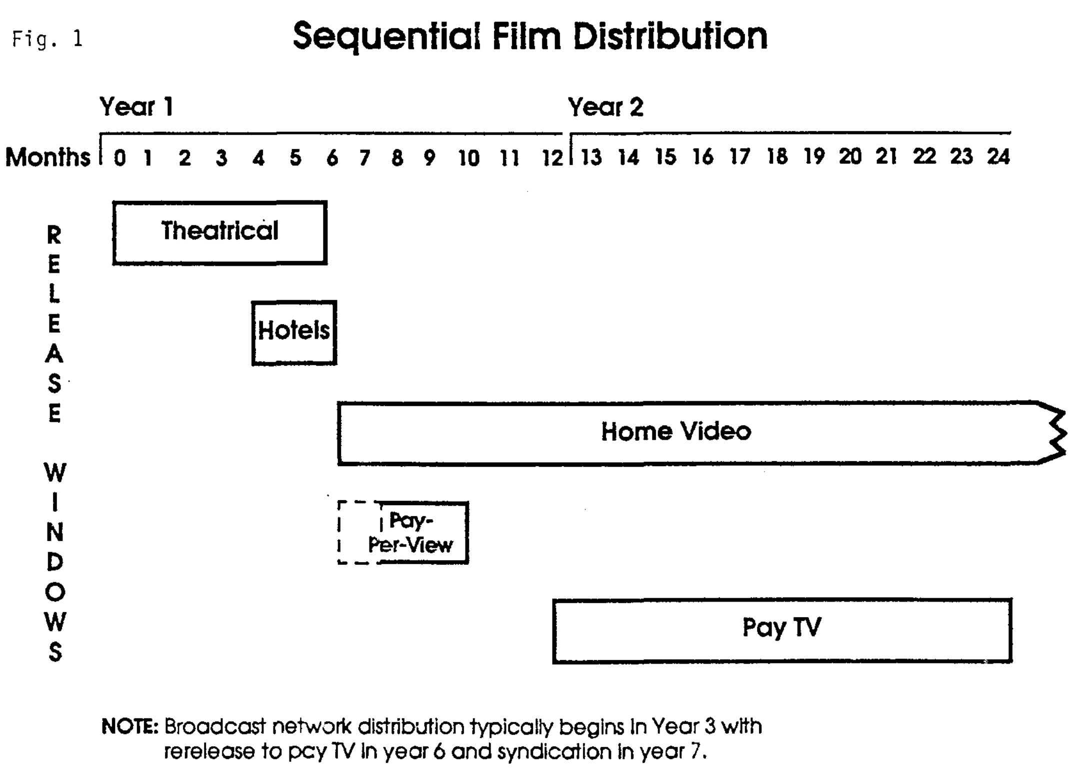
INTRODUCTION

The primary economic driving force of the PPV segment of the cable industry is the appetite of the consumer for recently released movies on television. Important though movies are to PPV, income from PPV presently represents only a very small percentage of the total revenues to Hollywood from movies. The principal revenue sources are:

- Home video tape rental;
- Theatrical;
- subscription pay TV;
- Cable PPV;
- Hotels.

Of these, home video rental is the largest revenue source.

Fig. 1



Motion pictures are usually released to these various forms of distribution in a distinct sequence (see Fig. 1). The timing within the sequence is designed to maximize total revenues and to protect each distribution medium in turn. At the present time, home video rental movies are released to approximately 6 months after theatrical release. The release to pay-per-view was, in the past, close to the rental release, but has now slipped to 4-6 weeks later than home video. A major concern of the studios is the threat of unauthorized copying of PPV movies and the potential for such copies to erode the rental business. This concern has resulted in the steady trend of delay in PPV release dates relative to home video rental.

Compared with video rental, cable PPV has major advantages for the subscriber. Ordering and delivery of programming is extremely convenient, and PPV does not suffer from the "depth-of-copy" problem of home video rental. (The "depth-of-copy" problem refers to the availability at the video rental store of only a limited number of copies of any new release and the resulting wait.) However, because of the disparity in release dates, cable PPV is at a distinct disadvantage relative to video rental when it comes to newly released movie titles.

In order to capitalize on its inherent advantages, cable PPV requires competitive release windows. One of the keys to advancing release dates to PPV is the assurance to the movie studios that PPV programming cannot be copied, and that the technology employed for copy protection cannot readily be circumvented.

During the past year, technology for effective video copy-protection has been demonstrated and tested in cable systems, and is now entering commercial use on a market test basis. The technical feasibility of copy-protection is no longer questioned. This paper describes the specific requirements and implementation of a copy protection system for use in cable.

REQUIREMENTS FOR COPY PROTECTION FOR CABLE PPV

The unique historical development of the cable industry, with the very substantial investment of equipment already in place, places some specific requirements on the design of a copy protection system for cable PPV.

Security

The industry's experience with signal piracy dictates a high level of security. Experience has shown that if a security scheme of any kind can readily be defeated or circumvented, then it's likely to be so! Copy protection must be inherently secure, costly to attempt to defeat, and leave no unprotected signals anywhere in the system.

Subscriber Hardware

Cable systems providing PPV have already invested in subscriber equipment (either in the home or outside). There must be no new hardware for the subscriber.

Compatibility

Whatever the modifications which are made to the video signal to achieve non-recordability, the signal must still be compatible with existing distribution systems, especially with the addressable descramblers already in place. Obviously, the copy protected signals must also be compatible with the large population of television receivers, of all makes and models, in subscribers' homes.

Ease of Operation

The addition of copy protection should not require additional operational steps in distributing PPV programs. Operation of the copy protection system should be automatic, i.e., protecting those PPV programs which require protection and leaving unmodified program material which does not require it.

Effectiveness

In order to achieve the goal of attracting more timely programming to cable PPV, a copy protection system must be effective in preventing VCR's from making useful tape copies of PPV programs. As compared with existing methods of protecting video tapes to deter duplication, the requirements for a PPV copy protection are more stringent. Absent some form of copy protection, copies are more easily made from PPV than from rental tapes. Copying of a tape requires two VCR's, presently found in less than 13% of U.S. homes. Copying of a pay-per-view movie, on the other hand, requires just one VCR, owned by more than 60% of cable subscribers. An additional highly desirable feature of a PPV copy protection system is prevention of recording by a CAM-CORDER directly from the TV screen.

SELECTION OF A METHOD

Because of the very different purposes of video cassette recorders and television receivers, there are just three fundamental differences in the way television signals are processed. Compared with a TV receiver, signal processing is optimized in VCR's in the following ways:

- Precise control of video signal levels is required for high quality recording and is usually achieved by measuring and controlling the amplitude of synchronizing pulses.
- For bandwidth-efficient recording, the VCR separates out the color component of video and records it in a different manner than the luminance. On playback, the color component is processed in a unique fashon to restore its frequency and phase.
- For efficient use of tape, and to minimize tape velocity, a mechanical scanning system is used to record in diagonal stripes across the tape. The mechanical scanning system in synchronized to the vertical sync component of the video waveform.

Various techniques have been devised (Ref. 2) to exploit each of these three differences between television receivers and VCR's.

1. The "PSEUDO-SYNC" AGC method exploits the VCR's unique ALC system by adding Pseudo-Sync pulses during several lines of the vertical blanking interval. The effect on an attempted recording is a weak looking video signal on replay. This technique has been used to deter duplication of video tapes. Its effectiveness is, however, quite dependent upon the specific VCR's used to record and playback.

2. The "COLOR STRIPE" method exploits the VCR's color restoration system by altering or selectively removing the back porch color burst signal. The effect on an attempted recording is loss of color or horizontal bands of color across the picture. It can, however, be negated by turning off the color on a TV receiver.

3. The "TIME BASE VARIATION" method exploits the synchronized diagonal scanning of the tape by a recording head mounted on a rotating drum. The vertical time base of the copy protected signal is time-varied by adding or deleting lines from frames. A momentary change of vertical frame rate disrupts the drum synchronization and servo systems of the VCR, causing it to miswrite the video recording and disturb control track information. The effect on an attempted recording is intermittent break-up of the picture, rolling artifacts caused by head switching and distortion of audio.

The TIME BASE VARIATION method has the following advantages by comparison with the other two:

- Effectiveness: a wide range of time-varying patterns can be used to confuse VCR's.
- Difficulty to Remove: there's no simple method to remove the treatment. The treatment is not removed by signal processing within a TV set, thus, the program material cannot be recorded from signals within a TV receiver.

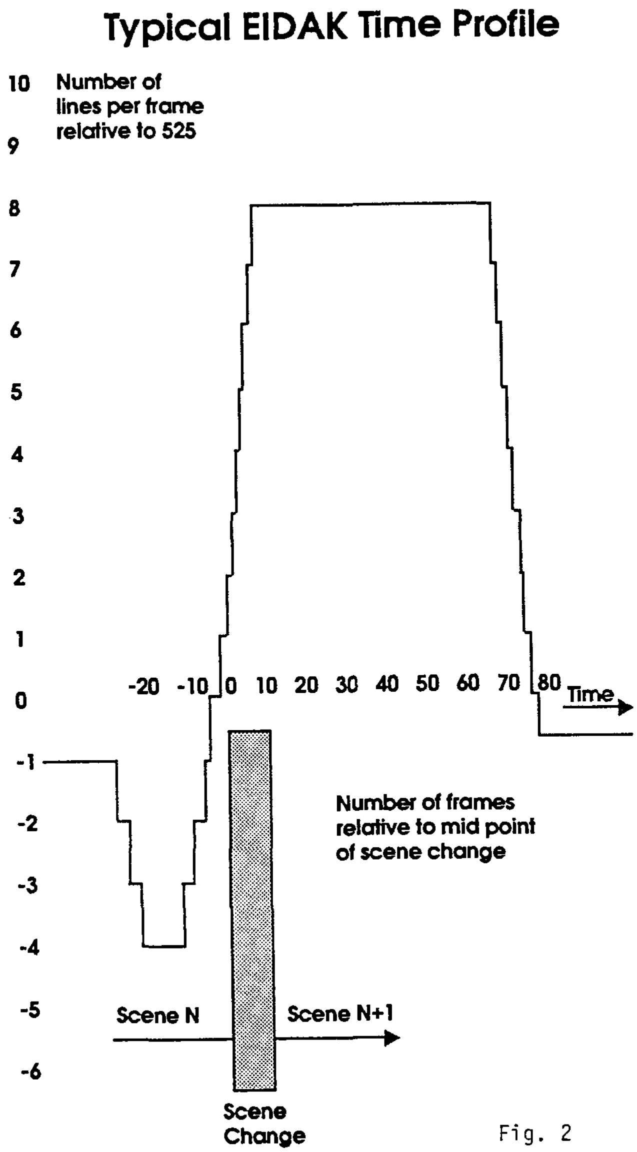


Fig. 2

- Works Against Camera Recording: an attempted recording from a TV screen is left with an objectionable moving pattern.

The TIME BASE VARIATION method is also the only one of the three methods which is not readily defeated with inexpensive components.

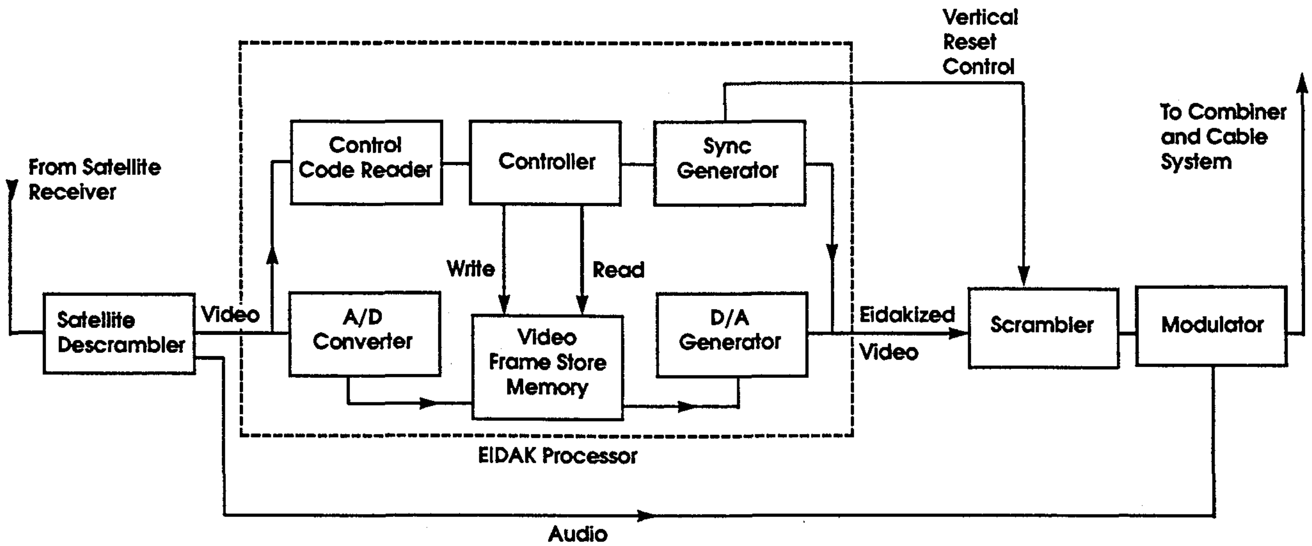
THE EIDAK CABLE PPV COPY PROTECTION SYSTEM

The Eidak Copy Protection System employs the TIME BASE VARIATION method, optimized for compatibility with TV receivers and cable distribution equipment. The vertical time base (frame rate)is varied by adding and deleting lines from video fields in a careful and systematic way. The specific time variation of lines-per-frame is called a "profile", an example is given in Figure 2. A variety of these "profiles" can be used to confuse the widest variety of VCR servo/synchronization systems. In the example shown, the number of lines per frame starts at 524 (one less than the standard 525) goes to 521 lines for a few frames, ramps rapidly to 533 lines (8 more than the standard 525) and then returns to 524 lines. It remains at 524 for several seconds before undergoing another similar "profile". This sudden variation in line count is sufficient to throw out the synchronization of the rotating recording head in a VCR and cause intermittent break-up of an attempted recording. At the same time that the number of lines is being changed, the position of the "real" active video within the field is varied in order to compensate for potential vertical movement of the picture on the TV screen. Without the vertical compensation, the TV picture would tend to move up or down with each change of field line count.

Variation of the frame length is accomplished digitally by changing the rates at which frames of digitized video are written into and out of a multiple frame store buffer memory. Up to $\pm 3\%$ variation in frame rate is achieved by adding or deleting up to 8 lines per frame. Lines are added or deleted from frames in pairs in order to maintain interlace. Although care is taken to keep the displayed picture centered on the TV screen to within about ± 1 line, even this small variation may be noticeable. In order to mask the movement, the time varying pattern is applied usually at scene changes. For movies, identification of the exact timing of the profiles is achieved by analysis of the movie prior to transmission. Data identifying the profile timing is then keyed to the SMPTE time code track. (For live events, this process is performed in real time.)

Fig. 3

EIDAK Processor Block Diagram



A block diagram of the processor is shown in Fig. 3. The analog portion consists of an A/D converter, operating at 4x color subcarrier frequency, and a corresponding D/A on the output side. The memory section consists of eight video field buffers, configured as a FIFO. The control code reader extracts profile command data from the vertical blanking interval and passes it to the Controller (an Intel 88 Wildcard) which controls the television line read/write rates of the memory. A vertical reset sync generator is used to interface with cable scramblers.

System Characteristics

- Non-recordability for both movies and live events;
- high degree of copy protection;
- secure throughout the program distribution system;
- no new or modified hardware in subscribers' homes;
- compatible with cable scramblers/descramblers;
- transparent operation in wide variety of cable plant configurations;
- compatible with the wide range of TV receivers in subscribers' homes.

SIGNAL DISTRIBUTION (Fig. 4)

The steps of signal distribution are:

a. For PPV signals delivered to cable systems by satellite:
- signal analysis, profile generation and control data insertion;
- scrambling overlay for satellite transmission;
- copy protection processing at the cable headend;
- cable distribution.

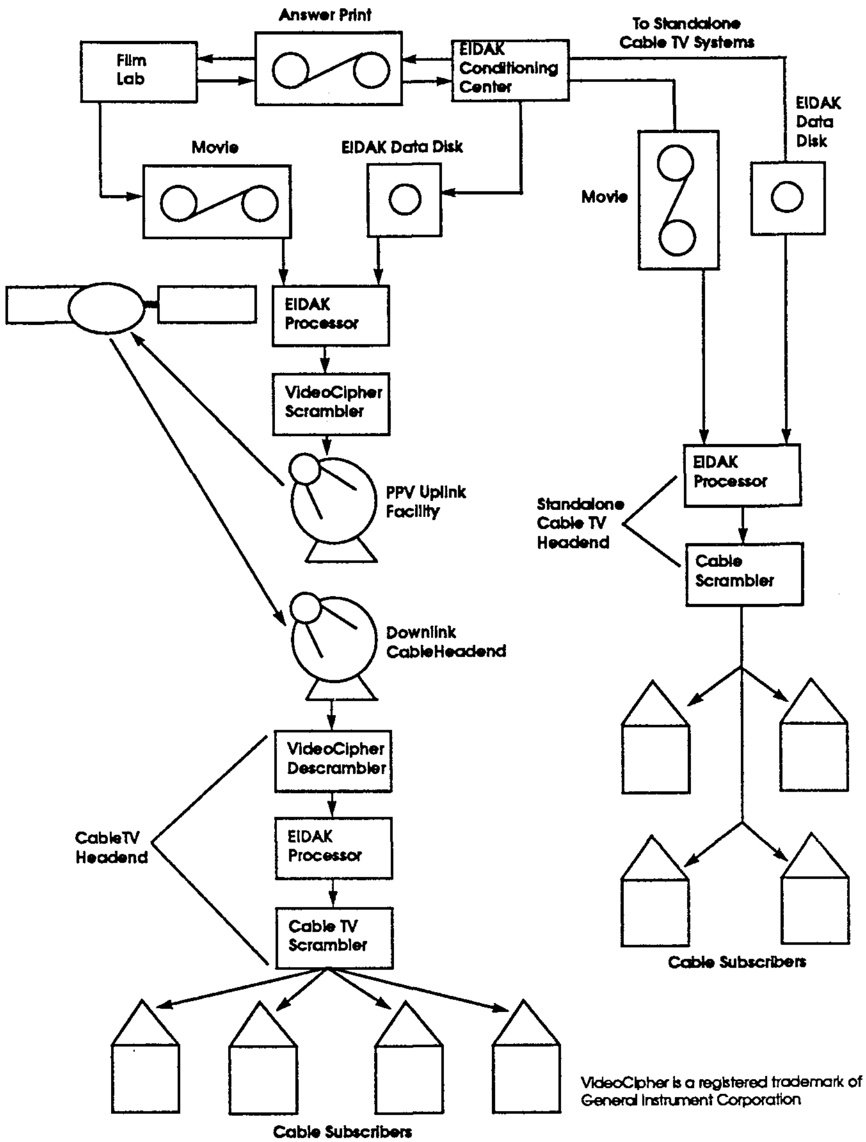
b. For standalone cable systems:
- signal analysis, profile generation, and control data diskette;
- copy protection processing at the cable headend;
- cable distribution.

Signal Analysis

Movies (and other pre-recorded program material) are analyzed prior to transmission to determine the optimum timing of "profiles". This analysis is performed at a conditioning center which generates data defining the location of profiles throughout a program keyed to time code. At transmission time, this data is sent simultaneously with the movie over the satellite link. (For live events, the data is generated by real time analysis of the video signal.) Upon receipt at the cable head end, the data is used in the copy protection processor to generate the profiles which define the patterns of varying line count.

Fig. 4

The EIDAK System



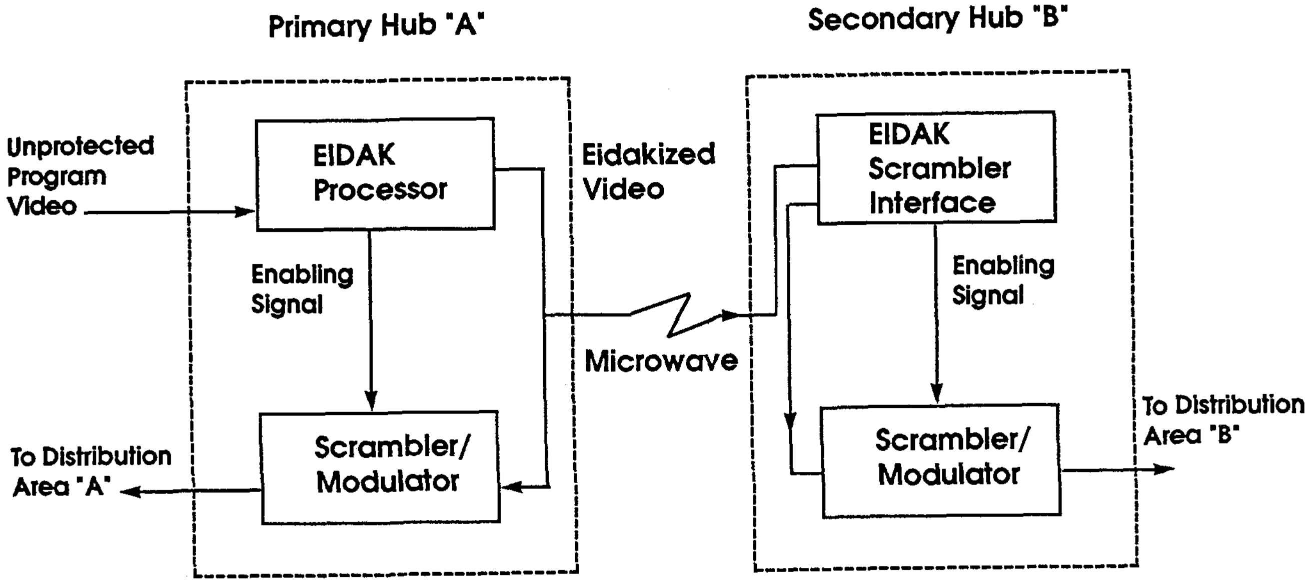
Satellite Distribution and the Scrambling Overlay

Satellite distribution of PPV programs presents a specific challenge regarding copy protection. Because the copy protection is applied at cable headends, there exists in the satellite link signals which have the potential to be copied, either at cable headends (prior to copy protection treatment), or through signal piracy. In order to guard against these possibilities, a "scrambling overlay" is applied prior to the uplink satellite scrambling.

This overlay can only be removed by passing the signal through an authorized copy protection processor. Thus, received signals have either the overlay scrambling or are copy protected. A useful signal is thus available for distribution only when it has been copy protected. The use of the overlay (and associated control signals) makes system operation completely automatic at the cable headend.

Fig. 5

EIDAK Processing in Hub Systems



Standalone Systems

Distribution of PPV programming to standalone cable systems is by videotape. The tape and a diskette with profile data are delivered to the system. The diskette loads the copy protection processor with date necessary to generate the time varying frame length variation as the tape is played.

Cable Scramblers

Most addressable cable scramblers generate and use field rate signals, either for scrambling/descrambling or for control signalling. In most cases, the circuit implementation of the scrambler is designed around standard NTSC (525 line) video. The use of a non-standard line count requires a timing signal to re-set the circuits which determine the vertical interval timing. This re-set timing signal is generated by the copy protection processor (or by the scrambler interface in the case of multiple hub systems).

Multi-hub System Operation

As cable systems are consolidated, and as they are built to cover ever expanding areas, a strong trend in system architecture has been the use of hubs. Distribution of signals to hubs is usually by one of two means:

(a) AM modulated RF signals (e.g., RF supertrunk and AML;
(b) baseband-video fed transmission links (e.g., FM microwave).

The copy protected RF signals can be transmitted transparently through the RF links. However, in the case of baseband-fed links, it is usual to employ a scrambler at each hub location. In this case, it is necessary to provide each scrambler with an enabling signal (see Fig. 5). At the primary hub, the enabling signal is generated by the processor. The copy protected baseband video signal is fed by microwave to one or more hubs. At each hub, a scrambling interface device generates the enabling signal for the cable scrambler. The enabling reset signal from the received copy protected video. Thus, PPV program video is copy protected throughout the hub distribution system.

CONCLUSION

The need for copy protection as part of the technology infrastructure of PPV is well established. In the cable environment, a copy protection system must be secure, and compatible with the wide range of equipment used in the construction of cable systems. The timebase variation method is effective and secure and has been optimized for compatible operation with addressable descramblers. A system configuration has been described which leaves <u>no</u> unprotected signals, even in a multi-hub distribution environment.

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PERFORMANCE OF AM MULTI-CHANNEL FIBER OPTIC LINKS

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Frank Little
Lee Thompson

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ABSTRACT

This paper is a presentation of how the characteristics and specifications of the basic components of an AM fiber optic link are interrelated to determine system performance. A simple theory and basic equations for calculating link performance is developed. Of particular emphasis is the calculation of carrier-to-noise ratio for a fiber optic link. Factors determining system distortions are discussed, and trade-offs indicated. Performance data for a current state-of-the-art AM system is presented.

INTRODUCTION

Since the beginning of lightwave communications, fiber optic systems have been designed for digital transmission. For digital communications, the intensity of the optical source is modulated on and off (referred to as on-off keying, or OOK) in response to logic levels "zero" and "one". Because the modulation is digital, modulation linearity is not an issue. For analog modulation, however, modulation linearity is a basic system parameter and a key factor in some systems. Also, laser noise requirements are much more stringent with AM modulation. Subcarrier FM has been used to advantage in multi-channel CATV systems since it is less affected by modulation nonlinearities than AM systems and requires a carrier-to-noise ratio of only 16 dB or so. However, since the cable distribution system to the home must carry signals in the AM format, AM technology is the preferred technology in cases where system objectives can be met with AM.

This paper first discusses direct laser modulation, laser noise, and the effects of optical reflections on noise. Expressions for CNR (carrier-to-noise ratio) due to laser noise are given and CNR expressions are derived for the optical receiver. Laser distortion is discussed and relationships between distortions and channel loading are given. Data for a CATV prototype system are presented.

INTENSITY MODULATION OF OPTICAL SOURCE

Light is generated in a semiconductor laser by forward biasing the semiconductor junction with a dc current. The relationship of light intensity to input current is given by the L versus I curve. An example of a distributed feedback (DFB) laser L-I curve is given in Fig. 1. As indicated in Fig. 1, lasing begins at a bias current referred to as the threshold current, I_{th}, and increases nearly linearly for bias currents greater than threshold. The light intensity, L, is commonly given in milliwatts or dBm. The efficiency of the electrical-to-optical conversion is given by the slope efficiency SE of the laser, which is defined as the slope of the L-I curve at the operating point I_b:

Eq. 1 $$SE = \frac{\Delta L}{\Delta I} \qquad \text{mW/mA}$$

Slope efficiency is also referred to as differential quantum efficiency. For the laser of Fig. 1, the slope efficiency is 8.5 percent. Note that efficiency is not dimensionless since the laser produces watts of output in response to amperes of input current.

To amplitude modulate a laser by a multi-channel AM CATV source, the broadband RF signal is added to the laser dc bias current. The amount of intensity modulation produced by the broadband RF signal is given by the modulation index. Modulation index is normally defined on a per-channel basis and is equal to the peak change in optical intensity divided by the average optical intensity. In this paper, it is assumed that all carriers are of equal amplitude. Modulation index m is defined as

Eq. 2 $$m = \frac{\Delta L_p}{L_o}$$

where ΔL_p is the peak change in optical power caused by a single RF carrier and L_o is the average optical power. The term optical modulation depth, OMD, is also used to define the amount of modulation and is identical to modulation index m.

Typically, for a 40 channel AM system, m ranges from .035 to .05.

Linearity of laser modulation is an important parameter in analog fiber optic systems. Laser linearity is measured by some manufacturers as the percent change in slope efficiency over the operating range normalized to the slope efficiency at the bias point, i.e.,

$$\text{Eq. 3} \qquad \text{LINEARITY} = \frac{\Delta SE}{SE_o}$$

where ΔSE is the change in slope efficiency and SE_o is the slope efficiency at the bias point. A plot of normalized slope efficiency as a function of current is given in Fig. 2. In this example, laser linearity is 8.5 percent for an optical modulation depth of 1.0.

Laser linearity is also specified by the amount of harmonic distortion generated by the laser, and by two-tone second- and third-order distortion. Distortion is discussed in later sections in this paper.

LASER OPTICAL NOISE

In analog lightwave systems, noise from the optical source contributes to the optical link CNR and is an extremely important factor in practical system applications. Laser diodes produce fluctuations in light output, or intensity noise. This intrinsic intensity noise is caused by the statistical nature of the carrier re-combination process. Laser noise is defined by RIN (relative intensity noise) as

$$\text{Eq. 4} \qquad \text{RIN} = \frac{\langle L_n^2 \rangle}{L_o^2}$$

where $\langle L_n^2 \rangle$ is the mean-square spectral intensity of light output noise. Noise power is referred to a 1 Hz bandwidth and RIN is dimensionless. RIN is normally expressed in dB/Hz and is equal to 10log(RIN).

Theoretical analyses [1] show that intrinsic laser noise is maximum for laser threshold current and decreases as the bias current increases as follows:

$$\text{Eq. 5} \qquad \text{RIN} \propto \left(\frac{I_b}{I_{th}} - 1 \right)^{-3}$$

Generally, for commercially available lasers, laser types with a lower threshold attain better noise performance than those with a higher threshold for the same output power [1]. Intrinsic noise is essentially independent of modulation frequency at low frequencies and increases to a resonance peak corresponding to the relaxation-oscillation frequency of the laser. The overall shape of the RIN response curve has the same general characteristic shape as the modulation frequency response [2].

Knowing RIN and the modulation index, one can calculate the carrier/noise ratio in a 1 Hz bandwidth (C/No) and CNR for a 4 MHz bandwidth according to CATV practices. The equivalent input noise current $\langle I_n^2 \rangle$ that would produce optical noise equal to that produced by the laser is, from (1) and (4), given by

$$\text{Eq. 6} \qquad \langle I_n^2 \rangle = \text{RIN}\left(\frac{L_o}{SEo} \right)^2 \quad A^2/\text{Hz}$$

Likewise, for optical modulation depth of m, the peak input signal current is mL_o/SE_o. The mean-square signal current I_s^2 is

$$\text{Eq. 7} \qquad I_s^2 = \frac{1}{2}\left(\frac{mL_o}{SE_o} \right)^2 \quad A^2$$

From (6) and (7), the carrier-to-noise ratio for a 1 Hz bandwidth is

$$\text{Eq. 8} \qquad \text{C/No} = \frac{m^2}{2\text{RIN}}$$

Of particular interest is the carrier-to-noise ratio in a 4 MHz bandwidth due to laser noise:

$$\text{Eq. 9} \qquad \text{CNRrin} = \frac{m^2}{8 \cdot 10^6 \text{RIN}}$$

Expressed in dB,

$$\text{Eq. 10} \qquad \text{CNRrin(dB)} = -69 + 20\log(m) - \text{RIN(dB)}$$

Note that CNRrin due to laser noise is independent of laser power. The effect of link loss and the contribution of optical receiver noise on the link CNR is given in later sections.

As an example, if RIN = -153 dB and m = .04 (typical for 40 AM channels), CNRrin due to laser noise is 56 dB. Commercial DFB lasers with integral optical isolators are available with RIN better than -150 dB/Hz. These lasers are capable of meeting current objectives for AM applications.

Intrinsic laser noise can be altered considerably due to the interaction of the laser and optical fiber. Laser diode noise increases significantly when light is reflected into the laser by discontinuities in the optical path [1]. Near-end reflections, less than ~10 cm, interact with the laser cavity and cause mainly low-frequency noise in the kilohertz range. Reflections from ~10 cm to ~100 m cause periodic noise peaks in the RF spectrum, and reflections from greater than ~100 m cause noise with an almost flat noise spectrum in the HF and VHF range [1]. In [1], the quantitative evaluation of reflection effects on laser noise characteristics was reported. It was found for the three types of lasers investigated that the maximum laser-coupled reflected power should be -65 to -73 dB to limit the increase in induced noise to within a few dB of the intrinsic laser noise level.

To prevent excess reflection-induced laser noise in practical AM systems, lasers with internal optical isolators should be used. These devices are commercially available with 30 dB of optical isolation. Furthermore, because of the high isolation required, fusion splices are recommended to ensure optimum system performance.

PHOTODETECTION OF OPTICAL SIGNAL

An optical receiver must be employed to convert the intensity-modulated optical signal to an RF signal for distribution in the CATV feeder network. For AM CATV systems, PIN photodiode detectors are usually employed. FM and digitally modulated systems operate at lower signal-to-noise ratios than AM systems and thus the received optical power is usually lower. For those systems, an avalanche photodiode (APD) is often used. An APD functions similarly to a PIN photodiode except that the APD can provide current gain whereas the PIN is limited to unity gain. However, the APD generates more noise in the optical/electrical conversion and is therefore at a disadvantage where the received optical power is large. Therefore, this discussion will be limited to PIN photodiode detectors only.

A photodiode emits electrons in response to incident photons. Quantum efficiency η is defined as

Eq. 11 $$\eta = \frac{\textit{number of photoelectrons}}{\textit{number of photons}}$$

and is equal to (reflection loss) $\cdot$(absorption loss)$\cdot$(absorption efficiency). Typically, quantum efficiency for a PIN photodiode is approximately 80 percent at 1.3-1.5 μm, but an efficiency of approximately 95 percent can be realized.

Responsitivity R is the measure of detected current due to incident optical power. Responsitivity is given by

Eq. 12 $$R = \frac{\textit{detected photocurrent}}{\textit{incident optical power}} \quad \text{A/W}$$
$$= \eta\frac{q\lambda}{hc}$$

where q is electron charge ($1.6\cdot10^{-19}$), h is Planck's constant ($6.63\cdot10^{-34}$), c is light velocity, and λ is optical wavelength. In the ideal case, η = 100 percent and responsitivity is

R = 0.684 A/W at = .85 μm
= 1.046 A/W at = 1.3 μm
= 1.248 A/W at = 1.5 μm

PHOTODIODE SHOT NOISE

A photodiode detector also generates a noise current called shot noise. Shot noise is caused by the discrete nature of electrons. In a photodiode, discrete charge carriers are generated by the incident optical signal and each contributes a pulse of current to the total dc current. These pulses are emitted randomly in time and thus produce a noise current referred to as shot noise.

For a PIN photodiode, shot noise $\langle I_{sn}^2\rangle$ is equal to

Eq. 13 $$\langle I_{sn}^2\rangle = 2qI_o \quad \text{A}^2/\text{Hz}$$
$$= 2qRP$$

where I_o is the dc current that flows in response to the incident optical power P. This shot noise limits the signal-to-noise ratio that can be achieved by the photodiode detector for a given optical input signal power. This limit is referred to as the quantum limit. If shot noise dominates in system operation, the system is said to be quantum limited.

Consider the signal current that flows in response to an incident optical signal. If the power incident on the photodiode is P watts, then from (2), the peak signal power is mP. The resulting peak signal current is mRP, and the mean-square signal current I_s^2 is

Eq. 14 $$I_s^2 = \frac{1}{2}(mRP)^2 \quad \text{A}^2$$

From (13) and (14), the signal-to-noise ratio in a 1 Hz bandwidth is

Eq. 15 $$\text{C/No} = \frac{m^2RP}{4q}$$

Thus, the quantum-limited CNR for 4 MHz bandwidth is

Eq. 16 $$\text{CNRsn} = 3.906\cdot10^{11}m^2RP$$

Expressed in dB,

Eq. 17 $$\text{CNRsn} = 85.9 + 20\log(m) + 10\log(R) + P(\text{dBm})$$

For example, if m = .04 (typical for 40 channels) and R = .85 A/W, the quantum limited CNRsn is 57.2 dB for P = 0 dBm. CNR decreases 1 dB per dB decrease in received optical power.

DETECTOR AMPLIFIER NOISE

Consider now the noise added by the amplifier that amplifies the output current of the photodiode. Even in an ideal case in which the amplifier contributes no excess noise, thermal noise is added by the load resistor that terminates the photodiode. Thermal noise current $\langle I_n^2 \rangle$ in resistor R_l at temperature T is

Eq. 18 $$\langle I_n^2 \rangle = \frac{4kT}{R_l} \quad A^2/Hz$$

where k is Boltzmann's constant ($1.38 \cdot 10^{-23}$) and T is °Kelvin.
If the amplifier noise factor is F, the equivalent input-current spectral density $\langle I_{an}^2 \rangle$ is

Eq. 19 $$\langle I_{an}^2 \rangle = \frac{4kTF}{R_l} \quad A^2/Hz$$

Note that the noise factor of an amplifier is a function of the source impedance, which, in the case of interest herein, is a current source shunted by a small capacitance in the range of 1 pF. Thus, the amplifier noise figure in situ is likely quite different from that measured in a characteristic impedance of 50-75 ohms, as is generally the practice. Equivalent input noise current is better suited for the transimpedance amplifier concept than the more common noise figure specification.

As indicated in (19), it is desirable to increase the photodetector load resistance in order to decrease the amount of amplifier noise. FET amplifiers are designed for that purpose. However, the impedance level that can be achieved practically is limited by the inherent circuit capacitance and the bandwidth required. Practical values for AM CATV applications range from approximately 500 to 2000 ohms. The design value is, in general, a function of the received signal power and sensitivity required.

The signal-to-noise ratio due to amplifier noise only can be determined in a manner similar to that for the quantum limited case. The signal current is given by (15). The signal-to-noise ratio for a 1 Hz bandwidth due to amplifier noise is

Eq. 20 $$C/No = \frac{(mRP)^2 R_l}{8kTF}$$

The amplifier-limited CNRan (for 4 MHz bandwidth) is

Eq. 21 $$CNRan = 7.81 \cdot 10^{12} (mRP)^2 \frac{R_l}{F}$$

Expressed in dB, CNRan is

Eq. 22 $$CNRan(dB) = 68.9 + 20\log(m) + 20\log(R) + 10\log(R_l) - F(dB) + 2P(dBm)$$

For example, if the received power P = 0 dBm, and if m = .04, R = .85 A/W, R_l = 1000 ohms, and F = 3 dB, then CNRan due to amplifier noise is 66.5 dB. Note that CNRan due to amplifier noise decreases 2 dB per dB decrease in received optical power.

LINK CNR

The CNR for the fiber optic link can be obtained from the individual CNRs defined above in a manner similar to that used in computing the cascade CNR in a CATV system. Specifically,

Eq. 23 $$CNR = \frac{1}{\frac{1}{CNRrin} + \frac{1}{CNRsn} + \frac{1}{CNRan}}$$

If the CNR's are expressed in dB,

Eq. 24 $$CNR(dB) = -10\log\left[10^{\frac{-CNRrin}{10}} + 10^{\frac{-CNRsn}{10}} + 10^{\frac{-CNRan}{10}}\right]$$

For the preceding examples, CNR_{rin} = 56 dB, CNR_{sn} =57.2 dB, CNR_{an} = 66.5 dB, and, from (24), the total CNR is 53.3 dB. If the received power is decreased to -5 dBm, the system CNR is 49.9 dB.

Fig. 3 is an example of a plot of link CNR and CNR due to RIN, photodiode shot noise, and receiver amplifier noise. The laser output power is 2 mW and other parameters are the same as in the previous examples. Also, link distance is shown assuming the link loss budget is 0.5 dB/km.

INTERMODULATION DISTORTION

The main source of nonlinear distortion in a well designed fiber optic system is the laser itself. Other sources of distortion include interaction of the fiber with the laser and reflections and discontinuities in the fiber system. Laser linearity can be degraded by the reflection of light into the laser cavity [3], but with the laser optically isolated, as it should be to prevent reflection-induced excess noise, this effect should not be a problem. In addition to nonlinear distortions from reflected light, connectors and splices can generate additional distortion because the loss of connectors and splices is a function of optical frequency [4]. Nonlinear distortions occur since direct modulation of a semiconductor laser not only modulates the light intensity but also the wavelength. The photodiode and receiver should not add significant distortion. In [5], the nonlinearity of photodiodes was measured and it was concluded that photodiode distortion is negligible.

Intermodulation distortion studies have provided a theoretical basis for determining distortion in a laser as a function of physical parameters of the device [6][7]. In [7], expressions for second- and third- harmonic distortions and two-tone third order distortion are given. It was also concluded that those expressions are valid for a variety of lasers, including DFB and Fabry-Perot devices at wavelengths of 1.3 and 1.5 μm. In theory, only the small-signal response characteristics of the laser are required to predict distortion levels. In [8], experimental tests are reported which show that measured data at microwave frequencies agree well with theoretical calculations, including triple-beat distortion of the form F1 + F2 - F3.

In CATV and other systems, distortion is often calculated assuming the nonlinear device is without memory (nonlinearity is independent of frequency) and the transfer function of the device can be expressed by a power series. Although this is not a rigorous approach, the results can be reasonably valid and a meaningful relationship between system variables can be derived. This method has been used [9] to accurately describe laser nonlinearity and predict intermodulation products. Also, since in CATV applications the maximum modulating frequency is low compared to the resonant frequency of the laser, the simple model should be useful [10].

The development that follows is patterned after [9]. First, neglecting distortion, for a single carrier of modulation index m, the optical output $L_{(t)}$ of a laser is given by

Eq. 25 $$L_{(t)} = L_o(\ 1 + m\cos\omega_m t\ )$$

A laser with nonlinearity is represented by the series

Eq. 26 $$L_{(t)} = L_o(\ 1 + m\cos\omega_m t + C_2(m\cos\omega_m t)^2 + C_3(m\cos\omega_m t)^3\ )$$

where C_2 and C_3 are second-order and third-order distortion coefficients. The ratio of the second harmonic to the fundamental is $mC_2/2$, and the ratio of the third harmonic to the fundamental is $m^2C_3/4$. From this, it is evident that second-harmonic distortion, relative to the fundamental, increases in proportion to the per-channel modulation index. Third-harmonic distortion, relative to the fundamental, is proportional to m^2.

By applying two or more carriers, each with modulation index m, the results can be extended to the other second-order and third-order beats. Table 1 gives the relationship of the various beats and crossmodulation. It also shows the familiar principle that all second-order distortions, relative to the fundamental, increase in proportion to m, or at a 1 dB/dB rate. Likewise, the relative change in third-order distortion, including crossmodulation, is proportional to m^2 and changes at a 2 dB/dB rate. Note that the ratios in Table 1 are amplitude ratios; the factor 20log is used to convert to dB.

TABLE 1

ORDER	FREQ. TERMS	DISTORTION RELATIVE TO FUNDAMENTAL	RELATIVE VALUE (dB)
2	2F1	$\frac{mC_2}{2}$	0
2	F1 + F2	mC_2	6
3	3F1	$\frac{m^2C_3}{4}$	0
3	2F1 + F2	$\frac{3m^2C_3}{4}$	9.5
3	2F1 + F2 + F3	$\frac{3m^2C_3}{2}$	15.6
3	F1 (XMOD)	$\frac{3m^2C_3}{2}$	15.6

Composite triple beat (CTB) distortion and composite second order (CSO) distortion are the results of power addition of all second-order or third-order beats at the nominal frequency of interest. In systems not harmonically related and phase locked, frequency and phase uncertainties cause each beat to be distinct. The composite distortion is, therefore, given by the power addition of all beats at the nominal frequency. Distortion is calculated by counting the number of beats of a given type that fall at specific frequencies, and dividing the carrier/distortion ratio for a single beat of that type by the number of beats.

Crossmodulation is a third-order distortion and can be calculated based on parameters in Table 1 and the number of TV channels. Crossmodulation is measured according to CATV practices with all interfering carriers synchronously modulated. Therefore, as measured, crossmodulation distortion adds on a voltage basis. For N channels there are N-1 interfering channels to produce crossmodulation. The composite crossmodulation ratio is the ratio given in Table 1 (a power ratio) for one interfering channel multiplied by $(N-1)^2$. However, it has been the authors' experience that laser crossmodulation is not always predictable, due perhaps to the nature of synchronously modulating the laser at 15 kHz with a high modulation index. In addition, the laser semiconductor is thermally modulated causing the emission to be wavelength modulated. But, based on other perceptibility tests [11], crossmodulation is not expected to be a major factor with laser video modulation.

Figs. 4 and 5 present the distribution of beat counts as a function of channel loading. This data can be used to calculate CTB and CSO from knowledge of harmonic, two-frequency, or three-frequency distortion. For these figures, beat counts are calculated for the standard frequency plan (excluding channels A-2 and A-1). Fig. 4 presents beat counts for determining CSO distortion. Curve (a) is the beat count (in dB) for the top channel ($F_1 + F_2$ beats plus second harmonics); channel 2 is the bottom channel. Curve (b) is the beat count (in dB) for channel 2 ($F_1 - F_2$ beats).

Fig. 5 presents beat count data for determing CTB. Curve (a) is the equivalent triple-beat count for the worst channel in N channels. All channels start with channel 2. In some systems, it is advantageous to split the total number of channels into two or more bands on one fiber, with each band modulating a laser, in order to reduce CSO and achieve better performance. For those applications, curve (b) shows the beat count data for a contiguous band of N channels starting at any channel above A-2. These beats are triple beats of the form $F_1 + F_2 - F_3$ and two-frequency beats of the form $2F_1 - F_2$. The relative value of the latter is 6 dB less than that of the triple-beat and is weighted accordingly (1/4 the power) when determining the equivalent triple-beat count.

For the simple model of the static L-I characteristic described by Eq. 26, linearity as given by Eq. 3 can be related to the distortion coefficients C_2 and C_3 and, by means of Table 1, the various distortions. For a single carrier with optical modulation depth = 1, second-harmonic distortion is $C_2/2$, and linearity due to parabolic curvature of the L-I characteristic is $4C_2$. Thus, C_2 = (linearity)/4, and the relative amplitude of the second-harmonic component = (linearity)/8. On this basis, for 40 channels with m = .04 and linearity = 4 percent, calculated CSO at channel 2 is 53.7 dB.

An example will illustrate how CTB and CSO can be predicted from knowledge of harmonic, two-frequency, or three-frequency distortion. Assume that the specified second harmonic distortion is -55 dBc for a modulation depth of 0.25. Calculate CSO for 20 channels assuming the per-channel modulation index is .06.

First, the harmonic distortion is calculated for the change in modulation index. Table 1 shows that the relative amplitude of second-order distortion is proportional to m. Therefore, the improvement in second-order distortion for a modulation index of .06 is 20log(.25/.06), or 12.4 dB. Thus, the carrier/second-harmonic distortion ratio at m = .06 is 55 dB + 12.4 dB = 67.4 dB.

Next, the difference in a two-frequency beat and a second harmonic is accounted for. From Table 1, $F_1 \pm F_2$ distortion is 6 dB greater than second-harmonic distortion, so the carrier/($F_1 \pm F_2$) distortion is 67.4 dB - 6 dB = 61.4 dB. (The preponderance of second order beats are of the type $F_1 \pm F_2$; only one harmonic component at most can be included in CSO beats).

Finally, a correction is made to account for the number of beats on a particular channel. From Fig. 3, a factor of 8.5 dB is added to account for 7 beats at channel 2 for 20 channel loading. Thus, the calculated CSO is 61.4 dB - 8.5 dB = 52.9 dB.

SYSTEM PERFORMANCE

Laser technology for CATV applications is currently progressing rapidly as more effort is expended in laser development for this market. With this changing technology, there is presently much variance in performance and yields from laser sources, particularly with regard to distortion specifications. For awhile it may be desirable for manufacturers to select and grade lasers to meet specific system requirements. Lasers that do not meet CSO objectives but are satisfactory otherwise could be used where the bandwidth is less than an octave or so. As the technology improves, yields and variances are expected to improve.

The system data in Table 2 was taken with one of the better lasers of those available at the time from different manufacturers. This system performance cannot be guaranred at this time in a standard product. Processes and specifications for this laser are being improved by the manufacturer, which should make this device suitable for production systems. This laser exhibits good linearity which enables a high modulation index to be used and still achieve very low distortion. Data was taken on a production prototype developed for the CATV market. Some of the system parameters are:

laser type- DFB
laser wavelength- 1330 nm
output output power- 2.6 mW
link distance- 15 km
link loss- 5.6 dB
channel loading- 40
bandwidth- 330 mHz (channels 2-EE)

TABLE 2
SYSTEM PERFORMANCE

FREQ. (MHz)	CNR (dB)	CTB (dB)	CSO (dB)	XMOD (dB)
55.25	54	69.7	69	57
83.25	53.7	74	71	57
121.25	54.1	67.7	>	56
145.25	53.8	66.9	69.8	
175.25	54.3	67.6	70	58
205.25	54.3	66.9	70	58
241.25	53.8	66.6	69	58
265.25	54.3	66.9	66.5	57
295.25	54.3	67.5	65	59
325.25	54.1	67.5	62.0	60

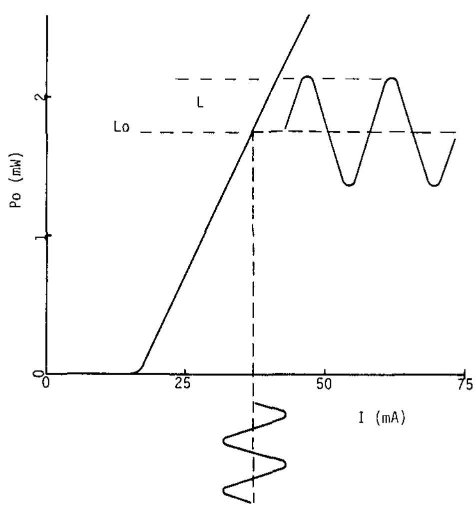

Fig. 1. Laser light intensity vs bias current. Laser bias is in mA and optical output is in mW. The response to a sinusoidal modulation current is shown.

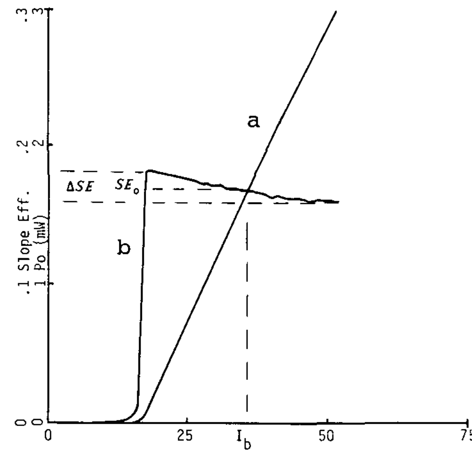

Fig. 2. Laser L-I curve (a), and slope efficiency SE (b).

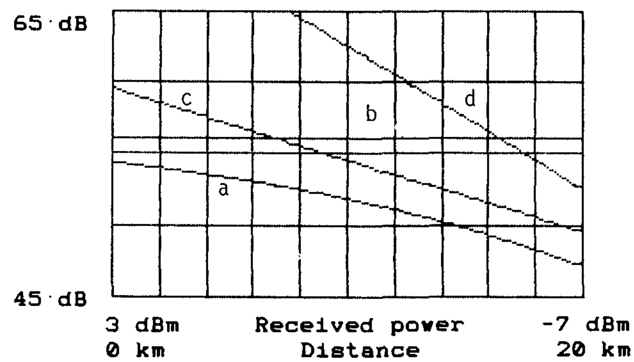

Fig. 3. (a): Link CNR. (b) through (d) are CNR due to: (b); laser intrinsic noise, RIN = -153 dB/Hz, (c); photodiode quantum noise, responsitivity = .85 A/W, (d); amplifier noise, R_1 = 1000 ohms, F = 3dB. Laser output is 2 mW and modulation index is .04/channel. Fiber loss budget is assumed to be .5 dB/km.

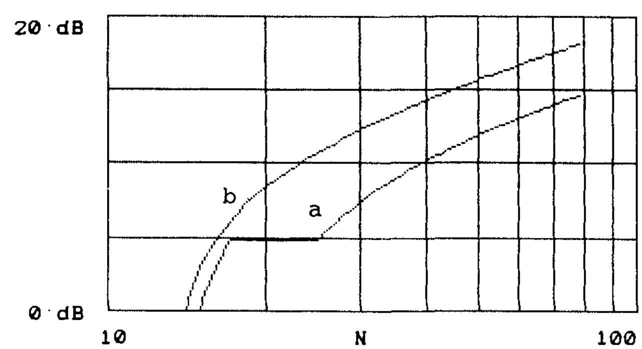

Fig. 4. Number of discrete second-order beats (in dB) that comprise CSO as a function of the number of channels, N, in the standard frequency plan. (a) is for highest channel; (b) is for channel 2.

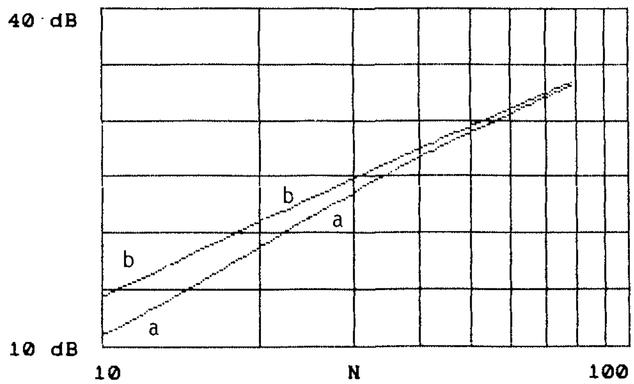

Fig. 5. Maximum number of discrete third-order beats that comprise CTB as a function of the number of channels. (a) is for channel assignments starting at channel 2, and (b) is for the contigious channels starting at A-2 and above. N is the number of channels in the standard frequency plan. Ordinate is 10log(number of channels).

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TIMING CONSIDERATIONS IN RF TWO WAY DATA COLLECTION AND POLLING

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ABSTRACT

The transport of control data streams for Addressable Two Way CATV equipment can take several forms, across several different media, with a variety of data formats. The coordination and sequencing of the various data streams to ensure accurate and complete responses becomes complicated when several technologies are combined. Two parameters are of particular concern: delay through a given channel, and the jitter (or delay uncertainty) incurred for each format translation.

This paper presents several scenarios for data transport serving an out-of-band data carrier Addressable system utilizing combinations of RF band, microwave, and telephone line transmission technologies for data signal delivery. Definitions are given for both continuous and discontinuous data streams. Since discontinuous data formats are most sensitive to delay and phase distortion, special solutions are developed for compensating these parameters.

INTRODUCTION

The maturation of Addressability as a technology, coupled with the advent of Pay-Per-View (PPV) and Impulse-Pay-Per-View (IPPV) services has effectively turned the CATV system into a two way data communications network. Regardless of the format of data transport over the cable, some basic principles of data communications govern the behavior of the data streams.

The CATV system can be represented as a tree-and-branch network. In the forward direction (controller to terminals), there is a single source, multiple destination data signal. Since various components in a CATV Addressable communication network are located at geographically distinct sites, and interconnection between the sites may be done with various media, depending on the particular geography, and availability of resources, a means to transport the addressable control data (one and two way) must be provided.

In most implementations, the forward data stream runs continuously, and is transported either on an out-of-band carrier, or in-band in the video itself from the headend or hub to the converter. If the Addressable Controller is at a business office that is located remotely from the headend or hub, data must be transported from the Controller to the headend or hub. This may be done through various media (e.g., RF cable, telephone lines, microwave, etc).

In the return direction, data is transmitted from the converter to the hub or headend site, where it is received and routed back to the Addressable Controller, over whatever media has been selected. In the case of multiple hubs or headends, return data is routed from each of the hubs back to the addressable controller, and combined for reception.

In the case of one way data, propagation and processing delays in the one way path are of no significant consequence. However, in a two way system, delays in both the forward path and the return path are critical to insure collision free high speed polling operation.

TIME DIVISION MULTIPLE ACCESS

Time Division Multiplexing (TDM) is a technique used in communication networks to allow multiple, unrelated lower speed data streams to be transported on a single higher speed data stream. The technique provides time slots in a given, predefined sequence for segments of each component data stream to be inserted at the transmit site, and extracted at the receive site. An advantage of TDM is the ability to carry multiple signals on a single wire or channel, with the associated saving in equipment over what would have been required if each signal was carried over its own wire or channel. In a TDM system, the multiplexing operation occurs at a single site, thus delays for each component data stream are identical. A typical application of TDM is the transport of telephone signals from one central office to another.

Time Division Multiple Access (TDMA) is an extension to the TDM concept that allows each component data stream to be inserted from different geographical locations. The complication involved in designing and operating a TDMA system is the difference in propagation delays from each source to a single destination.

Since collisions would be detrimental to operation, each source must be time compensated so as to insure that the arrival of its transmission at the destination site occurs during its allocated time window. The time compensation must account for any processing or propagation delays in the path from a particular source to the destination. TDMA systems are typically used in digital telephony over wire, terrestrial, or satellite channels. The advantage is that multiple lower speed users, located at multiple origination sites, can share a single higher speed channel, with the associated saving in equipment and channel space as compared to individual channels from each source to the single destination.

CATV APPLICATION

The typical tree and branch CATV system provides, in the forward direction, a single source, with multiple destinations. In the return direction, multiple sources provide transmission to a single destination. This architecture is analogous to the TDMA system described above. One major difference between the typical TDMA system and a two way CATV system is the number of operating nodes. A large CATV system may have on the order of several hundred thousand operating nodes in the network. Time compensating each of the nodes in a network of this magnitude would be an extremely time consuming task. Another major difference is the low cost nature of the subscriber terminal.

For these reasons, an approach requiring time compensation down to the hub level, but not to the subscriber node level is more appropriate. In addition, because the base of subscribers defined on the network is in a constant state of flux (box swaps, churn, etc.), fixed time slot assignment would slow the network down. Thus, a more efficient system allows assignment of time slots to subscriber terminals on the fly. The following sections of this paper introduce several system scenarios and describe the impacts of each technology.

When data passes through a communications channel, they are impacted by the physical properties of the channel. Of particular importance to this discussion is the propagation time of the transmission through the cable (fractions of the speed of light).

If the channel is actually comprised of several separate media with electronic translation on the endpoints of each segment, the propagation parameters are no longer the prime impact. The delay and delay parameters of each electronic translation is orders of magnitude more significant than the propagation time thru the channel.

Delays are generally incurred due to baseband data rate translations or format translations. Digital Signal Processing (DSP) operations and other digital data manipulations also contribute to delay. RF modulation and demodulation do not of themselves contribute significant delay to the channel relative to other delays.

Jitter, or delay uncertainty, is introduced whenever the clock phases of an asynchronous baseband transmitter and receiver at the endpoints of a channel differ. This occurs whenever data are reclocked (for processing by data path devices), or when data are sourced from numerous different devices running from their own, non-coherent clock sources, or located various distances from the receive point, with corresponding propagation delay differences (see Poll Format Data below).

Transaction Formats

The most straightforward format for data transmission has been shown to be a command/response transaction. In this format each transmission of information is framed by synchronization and error-checking data to assist in interpretation. This packet format is used extensively in CATV control and other applications. The transaction format provides a coherent query-response sequence for communicating with Addressable devices. It is constrained to communication with one device at a time in that the return channel can only accommodate one packet at a time.

Figure 1 shows a transaction system. The Addressable Control system provides a query command requiring a response from the addressed device. It then waits a prescribed period to allow that device to answer before continuing with more transmission. In this fashion, a controller can maintain coherence between queries and responses. The first chart shows the timeout period for a simple RF- only system, that is, one with minimal propagation delays. The second chart includes provision for an arbitrary delay inserted at some point in the channel.

The optimum time out period is based on worst case delay time through the system. Too large a timeout valve causes inefficiency in a system where many network nodes will not respond during various activities. Too small a valve may cause some valid responses to be ignored because the reached the receiving node after the timeout had expired.

Poll Formats (TDMA)

Poll formats can be described as transmission schemes which allow for maximum communication throughput at the cost of error-checking and framing. This technique borrows from the TDMA technique discussed above, in that several responding devices share a single response channel by synchronizing their transmission to some marker in the request data stream after their address is recognized. In this fashion, the responses are queued in the same order as requested, and each responding device is given a time slot for its response to arrive, if all delays are equal. A single command can initiate a response sequence from a range of Addressable Devices. the primary advantage in using this type of system is to achieve a very high poll rate for a given data rate.

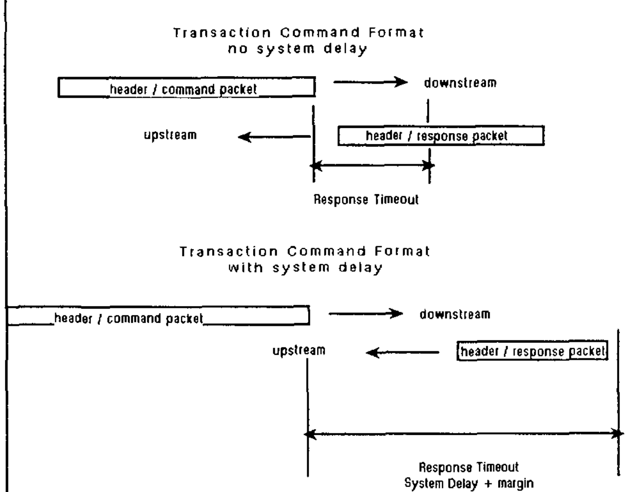

Figure 1 Transaction Command Timing Diagrams

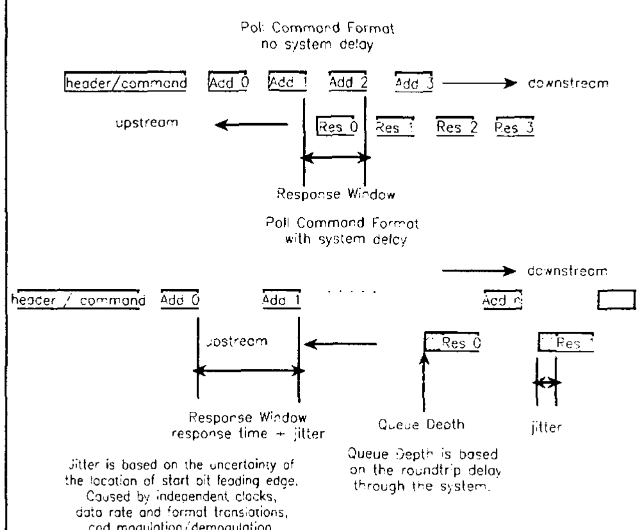

Figure 2 Poll Command Timing Diagrams

Figure 2 shows a poll format protocol in which a command is presented, followed by the addresses of all devices expected to respond. Each device formats and transmits a response on encountering its address in the data stream following a command. As can be seen, there is then an expected order, and time slot, in which the response will arrive at the Addressable Controller. The upper chart shows the sequencing in a non-delayed system. The lower chart makes allowances for both round-trip channel delay and delay uncertainty from data translations.

In this manner, responses are pipelined, with the depth of the pipeline determined by the absolute delay in the system. If the time delay for responses from each of multiple hubs differ, they must be compensated so they are equal. This is accomplished by inserting additional delay in each hub interconnect that has an inherent delay less than the hub with the maximum delay. The goal of this process is to make the delay from each hub equal to the delay in the hub with the largest delay.

Differences in delay from each responding node on a given hub are accommodated by allowing a large enough response window to receive a response with the shortest and longest expected delay within that hub.

SYSTEM DESIGNS

The RF plant of the CATV system is a known quantity. Any transmission delay in the signal is directly attributable to propagation delay in the amplifiers, passive devices, such as, combiners, splitters, and directional couplers, or cable. These quantities are easily calculated or measured, and are relatively small.

The opportunity for time distortion of the data streams occurs wherever there is a translation from one media to another. The most common place for this is the link between the Addressable Controller and the RF headend or hub sites. Often the Controller site is geographically separated from the Headend site. If there is a cable link, the delay will be minimal. This cable link can be a direct connect baseband connection between the Controller and headend, or an RF modem link to the headend. If there is no cable link, alternate technologies must be employed. When they are used, there is an impact on the communications system timing. Most sensitive are systems with multiple headends connected to the same Addressable Controller through different media and at different distances.

The following sections describe some of the more common interconnect options available for the link between the Addressable Control system and the RF CATV plant.

RF systems

In systems where the Addressable Controller is located at or near the RF headend, it is possible to make direct baseband connection to the RF modulation/demodulation equipment. This is the most efficient means for transport of data. This system exhibits only cable propagation delays in the channel. This basic configuration is shown in Figure 3. RF system delays are calculated from the physical parameters of cables, and distribution equipment, like amplifiers. In addition, an Addressable Controller must compensate for worst-case response set-up time in the subscriber terminal. These

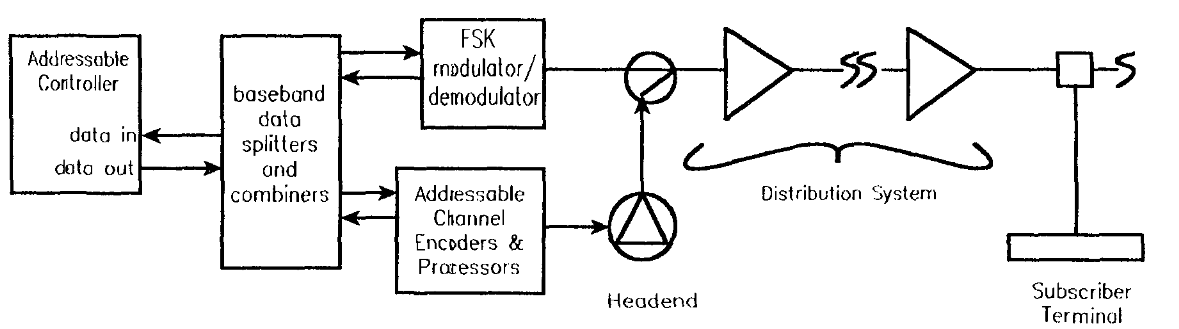

Figure 3. Colocated Single Hub Addressable System

values, once specified, become the baseline timing for the system. That is, a simple RF direct connect system defines the minimum timing compensation for any CATV data system.

More complex is the situation where the Addressable Controller is located on the cable plant downstream from the headend site. This is shown in Figure 4. To accomplish this, two additional modems must be installed in the system. Forward data intended for the terminals are modulated onto a sub-band carrier for transmission to the headend. There, the stream is demodulated and remodulated onto a carrier in the FM band for transmission back downstream to the subscriber terminals. In addition, the demodulated forward baseband stream is distributed to addressable video processing equipment within the headend. Terminal responses are received in the sub-band at the headend (on a unique frequency from that used to carry the forward stream), and demodulated. Responses from the addressable baseband equipment at the headend, along with the demodulated responses of the subscriber terminals, are modulated onto a unique FM frequency for transmission back to the Addressable Controller downstream.

RF modulation and demodulation do not add significantly to the delay found in a minimal configuration system. However, if the demodulation process is coupled to any form of error detection and/or recovery device which manipulates data at baseband, there is some effect on the overall timing of the system. In the RF non-colocated system described above, there are 3 opportunities for delay and uncertainty changes, one at each modem. These will add to the original baseline timing values. Depending on the Addressable Controller, additional delay and timing compensation may be necessary.

Telephone interconnect systems

When there is no cable interconnection between the RF headend, and the Addressable Controller site, alternative technologies must be used to transport the data streams between the two sites. Several different systems are available for this purpose, including telephone lines and microwave.

Telephone line communications can take two forms: one within the normal telephone network using dial-up modems, or the other using dedicated point-to-point lines which are always connected. Due to the heavy data traffic and the time sensitivity of the communications, dedicated lines are used for this type of activity.

High speed modems encode data into a trellis format which allows very high bit rates to be transferred via a low

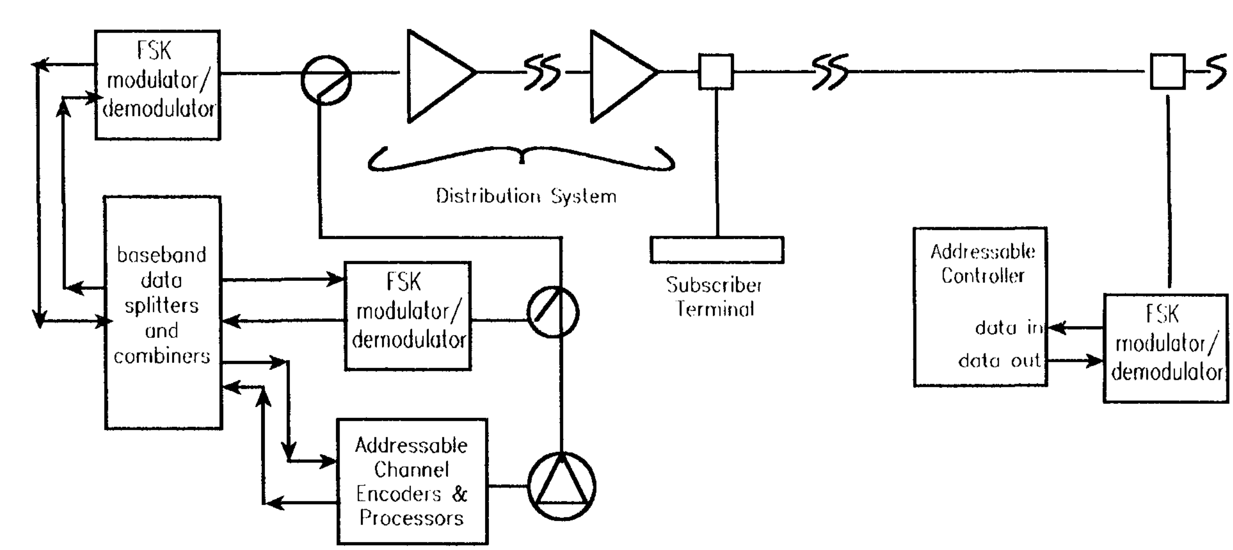

Figure 4 RF-linked Noncolocated Addressable Control System

bandwidth channel (3 Khz.). These data format and rate translations are usually the work of one or more microprocessors within the modem block itself. The delay and clock phase variations encountered are the result of not only propagation delay, but different clock rates, and phases, and nonlinear delays due to runtime variations in the formatting software of the signal processing microprocessors. In fact, these run run delays can be sizeable. For example, a 14.4 kbps. V.33-compatible trellis code modem may induce a 23-25 msec. delay in each direction. This amounts to an approximately 50 msec. roundtrip delay.

If the clock rate of the telephone modem is not identical to that of the incoming baseband stream, there are bit slippages and bit insertions which occur during encoding and decoding of the trellis-coded stream. This contributes to the jitter or delay uncertainty.

There is a final factor in the timing of a telephone linked system which needs consideration. The point-to-point telephone line connecting the Addressable Controller with the headend may be longer than the geographical separation of the two sites (see Figure 5). In fact, it is possible that a relatively short separation (<30 miles) can be connected by a very long telephone line (> 100 miles). Although propagation delays through a 4 wire telephone line are small, they are no longer insignificant in relation to the data rate when distances start to increase. This is why direct measurement of the roundtrip delay of the channel is desirable. Most modems can be placed into loopback modes. This allows the roundtrip delays in a given channel to be measured directly. If this is the last link before the RF interconnection to the distribution plant, the total delay can be calculated for that network leg.

Microwave systems

Another alternative for non-cabled data path is via microwave point-to-point transmission. A full duplex system capable of supporting two way RF terminals must incorporate transmission and receiving equipment for both directions between the Addressable Controller site and the RF headend or hub. (Two way satellite systems are generally not feasible due to large uplink costs at each remote site. One way systems are in use in several locations.)

The system design is shown schematically in Figure 6. The baseband data streams are first modulated by FSK modems and then presented to a microwave upconverter for the appropriate frequency translation. At the receiving end, a downconverter translates the stream back to its original carrier

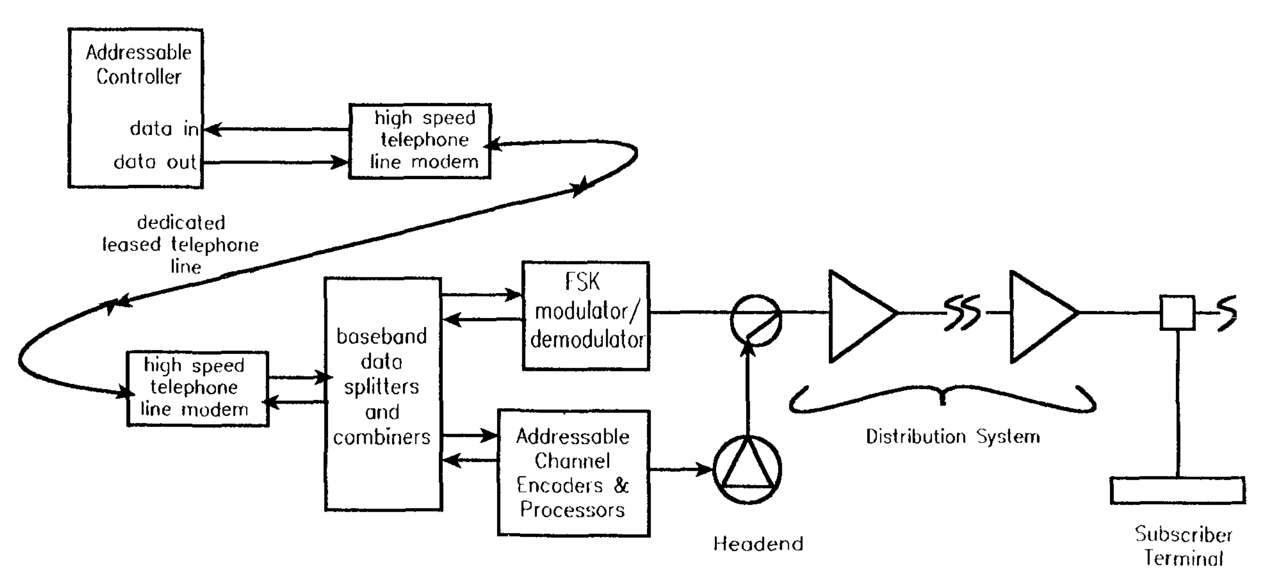

Figure 5 Telephone line-linked NonColocated Addressable Control System

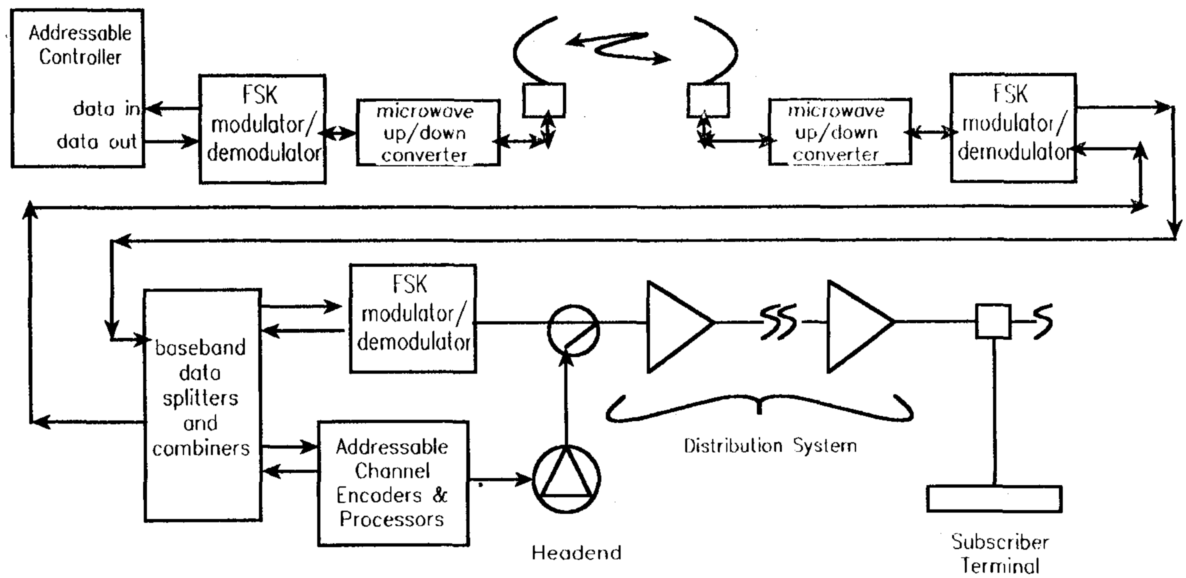

Figure 6 Microwave-linked NonColocated Addressable Control System

frequency. In most AML microwave systems, there is little discernible time delay for a continuous data stream. Propagation delays through the channel (upconverter, transmit, downconverter) are not significantly different from those in an FM band and sub-band RF system.

Combination systems (Multihub)

A system where several headends or hubs are serviced by one Addressable Controller is the most sensitive to differences in timing from hub to hub. When there are several different transport technologies implemented, the timing becomes more complex. Figure 7 shows a multiple hub system which utilizes all of the above described transport scenarios.

If timing between terminal and hub is considered constant, then the remaining areas for timing differences are in the data path equipment. The effects can be considered a combination of each single technology impact described in the preceding paragraphs.

In a multiple hub system, if delays are not accounted for, the command data in the forward stream does not arrive at the subscriber terminals simultaneously. In many applications it is not desirable to control addresses of terminals by geographic location, thus it is possible that consecutive addresses will be on different hubs. In fact, depending on the delay in a given channel, a command can arrive at consecutive address devices at very different times. This is significant in polling command formats where the expected response position identifies the responding device. If there is disparity in the arrival time of the command, the response cannot return in the proper order and may collide with responses from other devices.

The solution in a multiple hub system is to equalize the delay in all hubs so that responses from devices on each hub arrive at the destination at the same time regardless of the delay incurred on that leg of the network. The compensation is described in detail in the following section.

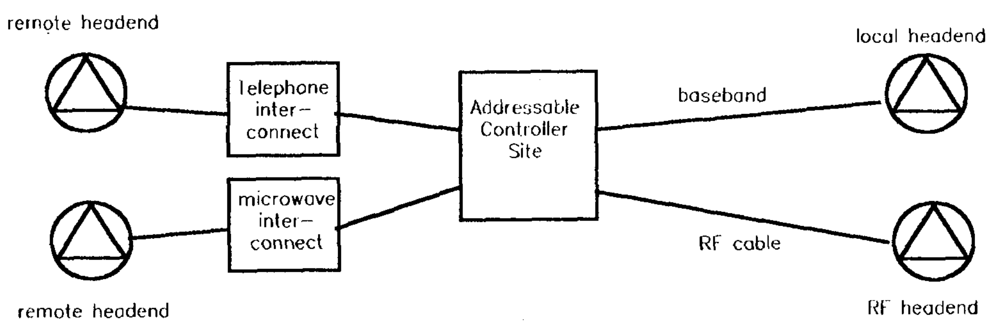

Figure 7 Multiple Hub, Multiple Data Path Addressable System

TIMING COMPENSATION

There are two places where timing compensation is required to accommodate the various data transport technologies. The first is in the data path itself using additional electronics to provide for the delay values. The second place is within the Addressable Controller software. The data communications parameters and protocols should be adjustable for the full range of delay and jitter which can be encountered in a CATV data path and distribution system. Ideally, in the Controller, this adjustment should be automatic. That is, the Addressable Controller can determine the type of network it is using, make timing measurements, and provide automatic compensation.

In single hub systems, the compensation required is minimal. Adjustment of timing parameters within the Addressable Controller is usually sufficient to assure adequate system performance. However, in multiple hub systems, timing must be equalized between hubs. The concepts of aggregate delay and delta delay become important as one deals with several differing length network leg timings.

In the Data Path

Aggregate Delay is calculated separately for each leg of the data path network. It is the combination of data path equipment delays, delays in translation to RF equipment, the propagation delay of the distribution system, and the turn around time of the subscriber terminal.

If there is more than one telephone-linked leg in the network, do not assume that the delays are the same. Each telephone line may have different delays.

Delay uncertainty is based on the accumulated uncertainty for all data path devices in the chain. Each device must be carefully characterized, both through device specification and empirical measurement. With the resulting base of information, the delay uncertainty can be calculated for any chain of devices.

The term "Delta Delay" is used to describe the difference in delay between a response from a given leg (hub) of the network, and a response from the leg (hub) with the longest delay. This is the amount of additional delay that must be inserted in that leg to equalize it (make it equal to the delay in the longest leg).

Delay insertion can be implemented using a programmable device under addressable control. It should be programmable on a channel by channel basis for a wide variation of delay values. The controller can then program the device to insert the appropriate Delta Delay for each channel in the network. When unequal delays are not compensated within the data path, there is a risk of collision between responses returning from different hubs.

In the Addressable Controller

An Addressable Controller operating in the environment presented can be thought of as a half duplex system connected to a full duplex line during transactions, and a full duplex system during polls. The controller sends a command and waits for a response for a specified length of time. This is true in either the Transaction or Poll modes of operation. The treatment of the two modes of operation is different however, and bears discussion.

In Transaction mode, the Addressable Controller addresses a single device with a fully framed request or command. The response is also fully framed. The controller can wait for a prescribed period of time for the response to arrive, or declare it failed. This wait period is the Transaction Response Timeout. The value is the maximum aggregate delay through the network. This is shown graphically in Figure 1.

Poll format commands are structured to permit one command to provoke responses from a group of terminals on the system. As delay becomes larger, there is more elapsed between the time when the Addressable Controller has sent the address of a given terminal and the return of that terminal's response. In order to keep up throughput, the controller will keep sending addresses to the remainder of the group. The delay incurred has the effect of forcing the Addressable Controller to allow more addresses to be transmitted before expecting an answer from an earlier address. This is referred to as Queueing or pipeline responses. That is, there is a set of addresses transmitted before the first response returns. The size of that set is a parameter called Queue Depth.

Delay uncertainty, or jitter in the system causes the response to move around within its expected window of response. The window is defined as a period of time in the response stream sized to the response plus some margin. The larger the jitter, the more margin is required to assure that the response will be received. This window size should be an adjustable parameter within the Addressable Controller.

CONCLUSIONS

The data communications functions of the CATV system have been described with respect to timing variations in the network. It is possible to compensate even the most complicated networks for timing to ensure efficient, reliable data communications performance.

The various data transport technologies may cohabit a system if their delay and jitter parameters are understood and accounted for within the system. This "fine tuning" is necessary in systems where high traffic polling and RF- based data collections are necessary functions.

If the time delays are understood, and the delay uncertainty can be measured and/or calculated for each leg of the CATV data network, the integrity of data responses from subscriber terminals will be reliable regardless of the complexity of the network. This results in more accurate, and complete data collections from the terminals.

Since timing compensation is done only down to the hub level, a typical system will have a relatively small number of nodes requiring compensation.

Since compensation is based on a mathematical model, and all measurements may be done under computer control. The entire compensation process can be automated, relieving the cable system operator from the laborious calculations.

Once timed, a system should only require retiming when data path devices or configurations are changed.

While the concepts have been presented in reference to an out-of-band data path system, it is possible to extend them to in-band systems as well. If there is a means for measuring or calculating the delay in the data channel between the Addressable Controller generation of a command and the arrival of the response, the delay values can be ascertained. If real time polling or other time sensitive communication is used, it is possible to measure the delay uncertainty of the return channel. These two parameters can be incorporated to fine tune the system for maximum reliable throughput.

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