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# 1 GHZ FIBER-TO-THE-PEDESTAL SYSTEM FOR CATV

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*Abstract*

*The design of a one gigahertz fiber-to-the-pedestal CATV system is described and discussed. The system consists of numerous unrepeatered fiberoptic links from the headend to receiver-amplifier units in pedestals. There is no cascading of amplifiers. The two types of amplifier units for this system are described and the system performance is discussed. Typical equipment costs for the deployment and design trade-offs are outlined. Other operational features and efficiencies are also described.*

*At the present time the architecture of this fiber-to-the-pedestal system is particularly well-suited for dense concentrations of dwelling units. However, the paper will indicate the likely path of progression from today's fiber-to-the-feeder systems to this fiber-to-the-pedestal architecture for more moderate subscriber densities.*

## BACKGROUND

During the past 18 months the cable television industry has made a dramatic move toward fiber-to-the-feeder (FTF) system architectures. In a paper[1] at last year's NCTA national meeting we analyzed the reasons for this shift in detail, but the key advantages of FTF can be summarized as:

- High signal quality
- Cost-effectiveness
- Reduced outage rates
- Subscriber-base segmentation

The one new feature underlying FTF that has made it such a successful architecture for cable TV, however, is its use of laser transmitters and optical fiber to free the system design from the requirement for a long amplifier cascade with low distortions. The fiber link takes the signal all the way into a neighborhood with reasonable fidelity. The concomitant high cost of the optoelectronics, however, is offset by lengthening the distribution strings from the conventional 2-3 amplifiers up to 4-6 amplifiers, thus serving greater numbers of subscribers from each fiber node. Additional reach is achieved by allowing operating levels to increase -- at the expense of distortions. The higher level of distortions is not a problem in FTF because there are so few amplifiers in the cascade.

We report here on a system under design that takes this distortion-vs-reach trade-off to the limit. In this system we have reduced the "cascade" to a single amplifier that is located in a pedestal within drop-length reach of a substantial number of subscribers (appropriately called "fiber to the pedestal" or FTP).

## FIBER-TO-THE-PEDESTAL SYSTEM

Increasingly during the past year cable TV operators and equipment vendors have been faced with a dilemma. There has been a clear need to increase bandwidth for additional programming services in the United States and for UHF transmission compatibility in Europe, but the key electronic components -- the amplifier hybrids -- have not generally been available in the bandwidths, gains and technologies required. Recently this has begun to change, however, as push-pull hybrids with good distortion performance up to 1GHz have come onto the market, but in limited gain versions. The challenge for the equipment vendors, then, has been to find ways to apply this limited selection of hybrids to solve the requirements of system operators without developing whole new product lines for each application.

Recalling that the fiber-to-the-feeder architecture was based on the idea of running a small number of amplifiers at relatively high levels -- thus enduring high distortion levels in individual amplifiers but compounding these distortions over only a few units -- we were led to explore what would happen if the amplifier cascade was reduced down to its minimum, one amplifier. It was proposed that this might permit the use of push-pull hybrids in applications where only power doubling and feedforward would normally have been considered. What we found was that one could, in fact, have considerable reach and reasonable costs in cases where one was able to reach each of 20-30 homes with 100-meter-long drop cables from that single amplifier. The amplifier would be a combination fiberoptic node and post-amp, with the optical and RF signals split as many times as possible.

The system architecture is shown in Figure 1. In the following sections the individual components are described, the system performance is reported and the system economics are discussed.

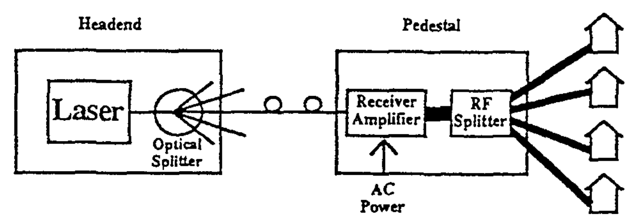


Fig. 1. Fiber-to-the-pedestal system architecture

## SYSTEM COMPONENTS

### Laser Transmitters

Clearly one of the keys to any cost-effective application of fiberoptics in cable television is the availability of high-power, low-noise and low-distortion laser transmitters. In this application the requirements are the same as in others, so the laser options are, as well: single-output DFB lasers and multiple-output YAG lasers. In particular we have designed for DFB's with output powers of

4-6mW (6-8dBm) and for YAG lasers with four 10mW (10dBm) outputs.

Commercially available DFB lasers have RF input drivers capable of 860MHz operation. These can be upgraded to 1GHz using the new push-pull hybrids. The modulators in the YAG laser systems, on the other hand, require high enough drive levels that feedforward amplifiers appear to be required. Since the commercial availability of such high bandwidth feedforward gain blocks is unclear at the present, this paper deals only with tests of the DFB's.

For European applications, where UHF frequencies are needed but the channel count is modest (i.e., 30 to 40, approximately), single-fiber transmission appears to provide quite adequate performance. For US applications, where a full 150-channel line-up is intended, dual-fiber systems appear to be required.

Optical splitting is called for and is generally done in the headend. We have designed for splits up to 16-ways, which adds 13dB to the optical link loss.

Receiver/amplifiers

Two different fiberoptic receiver/amplifier stations have been developed for this application. One is a single-output "GlasPAL" device in a line-extender housing, with plug-in single or dual receiver and transmitter cards. The unit utilizes two 18dB hybrids in the forward path and provides a 40dBmV output level for typical received optical powers.

The second unit is a four-output "Flamethrower" station with similar plug-in optical receive and transmit capability. This unit uses up to seven of these hybrids and can provide high-level outputs from each of its four ports.

Tap array and drop cable

The system design objective is to be able to reach 24-36 homes located within a 100m radius of the amplifier, for each of the amplifier outputs. This is done by providing an assembly of taps immediately at the amplifier output ports inside the pedestal enclosure. As can be seen from Figure 1, there is no need for passing AC power through the taps, thus they can be designed for minimum loss. Typical loss values at 1GHz would thus be 3.6dB for a 2-way split and 19dB for a 24-way.

In order to reduce losses at these high frequencies the drop cable selected is 1.6/7.3 (Cordaillod) cable. A 100m length of this cable has attenuations of 15dB and 4dB at 1GHz and 100MHz, respectively.

## SYSTEM CONSIDERATIONS

Powering

AC power must be provided at each pedestal since there is no other way to bring power to the receiver/amplifier. In these units the power enters the station via a separate AC feed and is transformed down to 48VAC for direct application to the power pack inside the unit.

Return signals

Laser transmitters for data return can be provided in each station at reasonable cost (see economic discussion below). The need for video return at specific locations can also be accommodated (at an increased cost, however). Perhaps of greater interest is the ability to readily plug-in one of the higher-cost video laser cards temporarily at any station for "on-the-spot" remote transmissions (such as electronic newsgathering), since one is never far from an optical node station in the FTP architecture.

## END-OF-LINE PERFORMANCE

### Design objectives

The intriguing aspect of FTP systems is that the end-of-line performance is determined, for the most part, by the characteristics of only one fiberoptic link with post-amp.

In the most immediate application of this system architecture -- a European system transmitting forty 8MHz channels plus data spread over the spectrum from 87-860MHz -- the performance required at the set-top is:

| | |
|---|---|
| Carrier-to-Noise: | 49dB, min |
| Composite Triple Beat: | -55dB |
| Composite Second Orders: | -53dB |

The level to the subscriber is 6-18dBmV, but the maximum difference in signal levels at any particular set-top can be no more than 8dB.

Furthermore, the system economics require that all of these specifications apply to lengths of drop cable from 15 to 100m emanating from each receiver/amplifier.

The average fiber cable length in this system is 6km, which translates to an optical loss of 2.4dB, plus splitting loss.

### Performance test results

In order to deliver the required levels to 24 homes from one receiver/amplifier, the output of the amplifier has to be 40dBmV. This results in the system configuration shown in Figure 2.

At a -4dBm input to the optical receiver and a 4% optical modulation depth, the CNR and CSO performance requirements can be achieved and the CTB requirement well exceeded. Interestingly, there is essentially no distortion contribution from the receiver/amplifier hybrids at this output level and loading.

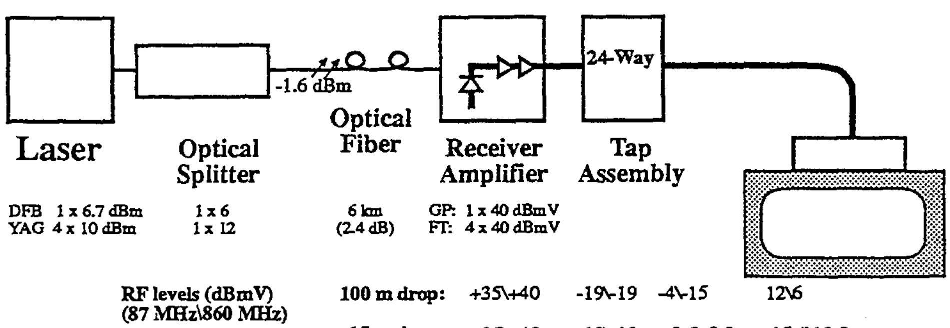


Fig. 2. Specific architecture to be deployed

In order to supply -4dBm to the receiver with a 6km fiber length, the output of a 4.7mW laser can be split six ways. Each output of a 10mW YAG laser can be split twelve ways.

## SYSTEM ECONOMICS

Using published laser transmitter cost estimates[2], we can analyze the costs of the opto-RF equipment for FTP in various service scenarios. (See Table 1 for the assumed prices for the key components.) It must be kept in mind that each of these scenarios will assume that all of the calculated number of homes can, in fact, be reached by drop cables from the receiver/amplifier node. This works out to a required subscriber density of 2000 per square mile, which is certainly not typical either in the US or Europe. There are numerous urban locales where such densities are found, however. For reference, the average household density in Bern (Switzerland), Washington, DC (US), and Paris (France) is approximately 2800, 5000 and 27,000 per square mile, respectively.

In the remainder of this section, we will examine the cost per subscriber and the group size in a few different deployment scenarios for this architecture. The costs of the opto-RF equipment and the sizes of the service groups are summarized in Table 2.

### Scenario 1: High density, low channel load

We first consider a scenario similar to the one discussed above in the performance testing section. It applies to many situations in Europe, where there is sufficient residential density and the programming is limited to about 40 channels that are widely spaced across the VHF and UHF bands. As has been shown above, a single DFB laser can provide transport of these signals with acceptable CNR and distortion performance. In Table 2, the columns labelled 1a and 1b show the results of a 4-way and a 6-way split of this optical output, followed by a single-fiber GlasPAL whose RF output is split 24-ways. The additional optical split in case 1b limits the fiber loss budget. On the other hand Column 1c shows that (for the particular equipment prices given in Table 1) a YAG laser can deliver as much optical power to each pedestal as the initial case (1a) but at the same cost as case 1b, because the YAG laser allows much more splitting. This has unfavorable failure group size implications, however. Note that case 1b is the same as that shown in Figure 2, except for the optical fiber loss.

Table 1. Estimated costs of opto-RF components

| | |
|---|---|
| DFB laser transmitter (DFB) | $12500[2] |
| YAG laser transmitter (YAG) | 65000[2] |
| Single-fiber receiver/amplifier (1GP) | 1750 |
| Dual-fiber receiver/amplifier (2GP) | 2250 |
| Single-fiber Flamethrower rcvr/amp (1FT) | 3250 |
| Dual-fiber Flamethrower rcvr/amp (2FT) | 4000 |

Table 2. Fiber-to-the-pedestal scenarios

| Scenario: | 1a | 1b | 1c | 2a | 2b | 3a | 3b | 3c | |
|---|---|---|---|---|---|---|---|---|---|
| Transmitter type | DFB | DFB | YAG | DFB | YAG | DFB | DFB | YAG | DFB |
| No. of transmitters | 1 | 1 | 1 | 1 | 1 | 2 | 2 | 1 | 1 |
| No. of outputs | 1 | 1 | 4 | 1 | 4 | 1 | 1 | 4 | 1 |
| Opt pwr/output (mW) | 5 | 5 | 10 | 5 | 10 | 5 | 5 | 10 | 5 |
| Optical splits | 4 | 6 | 8 | 4 | 8 | 4 | 5 | 8 | 4 |
| Opt fiber loss (dB) | 4 | 4 | 4 | 4 | 4 | 4 | 4 | 4 | 4 |
| Opt pwr rcd (dBm) | -3.5 | -5.4 | -3.8 | -3.5 | -3.8 | -3.5 | -4.6 | -3.8 | -3.5 |
| Receiver/amp type | 1GP | 1GP | 1GP | 1FT | 1FT | 2GP | 2GP | 2GP | |
| No. of outputs | 1 | 1 | 1 | 4 | 4 | 1 | 1 | 1 | |
| RF splits | 24 | 24 | 24 | 24 | 24 | 24 | 24 | 24 | |
| No. homes served | 96 | 144 | 768 | 384 | 3072 | 96 | 120 | 768 | 96 |
| Opto-RF $/home | 203 | 160 | 158 | 66 | 55 | 354 | 302 | 309 | |

## Scenario 2: Very high density, single fiber

In an area of high subscriber concentration, where a four-output receiver/amplifier may be used to advantage, we can see dramatic dilutions of the laser transmitter cost. Case 2a shows that a DFB laser system delivers signals at a cost of only $66/sub. In comparison a YAG-based sytem delivering the same picture quality can reduce that cost by 17%, but at a eight-fold increase in failure group size.

## Scenario 3: High density, dual fiber

In a scenario more relevant to the US, where one fiber could carry the lowest 77-channels and a second fiber could carry almost 500MHz in a single octave, the costs are, of course, higher, as shown in case 3a. Case 3b shows that an extra optical split reduces the cost/sub, but delivers what may be marginal optical signals for this loss budget. An interesting possibility is the use of a YAG laser for the lower band, where distortion performance will be critical, along with a DFB laser of lower linearity spec for the upper band. If the high output power of the YAG laser can be shared efficiently, as outlined in the final two columns of Table 2, this scheme would lead to a cost of just over $300/sub. Furthermore fewer than 100 homes would be served by the high-band laser, so considerable market segmentation would still be retained.

## Discussion

There are a few additional observations that we have made in the course of this investigation. First the models for single-fiber systems should be applicable to 550MHz transport in the US today. Future upgrades could be obtained by substituting dual-fiber receiver cards for the single-receiver plug-ins (provided, of course, that enough extra fibers have been included in the cable provisioning).

As with all of the new fiber-based architectures, the system operator will have to make sometimes difficult trade-offs between service group size and capital cost. It should be noted that the high-power YAG laser may be most useful for long optical loss budgets, where its power is needed. In the scenarios we have discussed above, which had relatively short fiber lengths, the YAG unit tended to generate excessively large group sizes.

Finally the additional cost for 2-way systems must be considered. The return of conventional data (in this case by laser, of course) adds approximately $2000 to the cost of a fiber node. Since the node serves 24 homes, this would add about $85/sub. In the future, one might envision that return of video might be required, which could also be done with a plug-in card, but at a significantly higher cost.

## SYSTEM EVOLUTION

Fiber-to-the-feeder designs being installed today provide highly cost-effective systems that serve on the order of 2000 subscribers from each fiber node. In addition there are ways to readily adapt some of these installed systems so that the fiberoptic signals can be brought further into the network and the service group decreased by a factor of four, to about 500 subs.

The fiber-to-the-pedestal architecture can be made to supplement such systems. As has been mentioned, FTP applies today to high concentrations of subscribers, as in urban areas. It can also be used to bring new services to specific areas within a larger system.

As costs decrease (particularly for the lasers and photodiodes) the cost-effective group size will also decrease, which means that FTP systems will apply to less dense population areas. Specifically, it should be noted that when the cost can be divided among as few as four subscribers, then this equipment would be capable of full 2-way video service, since four NTSC channels can be accommodated on one video laser return.

## CONCLUSIONS

In recent years many people have endeavored mightily to bring the advantages of fiberoptic transport to users of AM video signals. The technical successes on the component level has engendered great changes in the way we think about our cable TV system architectures. In this paper we have explored yet another step in this conceptual evolution. Our observation is that a single-amplifier FTP system can be deployed effectively today in certain cases. We also see it as a logical outgrowth of the fiberoptic technology and systems already deployed and as further evidence of the strong position of the CATV industry as the provider of broadband services in the future.

## REFERENCES

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# A COMPARISON OF LEADING EDGE COMPRESSION TECHNOLOGIES

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*Abstract*

*Digital Television is fast becoming a reality after decades of research. Broadcast, Cable and Satellite distribution of entertainment and sports television presents a unique set of requirements. In this paper, these requirements are reviewed and are followed by a brief description of the compression techniques. A compression system is then presented that combines several leading techniques to satisfy these special requirements. It is clear from this that the compression technology is real, cost-effective and can offer consumers and service providers significant benefits.*

## INTRODUCTION

Digital television is fast becoming a reality after decades of research.[1] It is rapidly proliferating variety of applications such as video-conferencing, multimedia computing systems and entertainment television over broadcast, cable, satellite and tape media. Many of the technical impediments which previously prevented the dream of digital television from becoming a reality are going away. This is due to significant progress in compression technology, integrated circuits for signal processing and memory, and reduced cost of transmission and storage. Moreover, digital compression provides the vehicle for a step function increase in the capacity of video delivery systems which can enable new services that were not previously possible. Thus we have a simultaneous "push" from the new services side and "pull" from the technologies of algorithms and hardware. The challenge now is to "engineer" the systems in a cost-effective way to provide an array of services to the customers.

Television signal when digitized produces an enormous bit rate, too high for economical transmission or storage. For example, the CCIR-601 television signal, when sampled and digitized requires over 200 Mbits/sec. Digital compression is therefore, essential in reducing this to a bit-rate dictated by the application. Obviously, performance of the compression hardware has to be judged by the quality of the picture at the affordable bit rate, but in addition, compressed digital television has to allow all the other functionality that analog television currently has. This functionality will vary from application to application. For example, the quality of compressed digital television should not be lower than analog television for the type of entertainment and sports material that we are used to watching over the cable, but, a lower quality may be sufficient for videoconferencing. Therefore, the choice of compression algorithm will depend on the application. Already different algorithms are in the process of getting standardized for different applications such as video teleconferencing ($p \times 64$ kbit/sec-algorithm[2]), multimedia (MPEG-1 algorithm[3]), and digital cable TV and HDTV, being led by the Cable Labs and the FCC, respectively.

In this paper, we start with a description of the desirable characteristics of the video compression algorithm for the cable, terrestrial broadcast and satellite applications. We then discuss some of the basic techniques used by most of the compression algorithms. We compare these basic techniques on the basis of compression performance as well as complexity of implementation. Following this, we describe an algorithm that appears ideally suited for cable and broadcast applications. It is clear from this, that the compression technology is ripe and will revolutionize the nature of television in the coming years.

## COMPRESSION TECHNOLOGY REQUIREMENTS

The basic digital video compressor and decompressor are shown in Figure 1. As mentioned earlier, the function of the compressor is to reduce the bit rate as much as possible without sacrificing the picture quality required for the service. In addition, a number of other requirements have to be satisfied, particularly in the cable and broadcast environment. In this section we outline these requirements so that they will serve as a guide in evaluating the different compression algorithms described in the next section.

The first requirement, of course, is to achieve transparent coding for the class of pictures that are typically used in the particular service. In entertainment and sports applications, the picture material may contain high detail, complex (not necessarily predictable) motion and a large number of frequent scene cuts or changes. The picture material may be created by a television camera or synthesized by a computer or an arbitrary mixture of the two. In a number of situations, the source material may contain noise (e.g., old film or electronic news gathering cameras). The ability to compress such diverse material is indeed very challenging and very different from either video teleconferencing or the multimedia applications. Moreover, the picture quality standard accepted and practiced in the cable and the broadcast industry is significantly more demanding than what is practical in videoconferencing or multimedia applications.

The second requirement is that the compressed video bit stream must not be very fragile. Of course, compressed bit streams are always more fragile than raw uncompressed digital video. However, their robustness to transmission impairments can be significantly improved by error correction and ghost cancellation techniques. In addition, the compression algorithm must be such that a large fraction of the errors that escape the error correction mechanism can be concealed and their effects localized both in terms of space (i.e., horizontal and vertical dimensions of a picture) and time (i.e., the number of frames).

In a cable or broadcast environment, a viewer may change from channel to channel with no opportunity for the transmitter to adapt itself to such channel changes. It is important that the buildup of resolution following the channel change takes place quite rapidly so that the viewer can make a decision to either stay on the channel or change further depending upon the content that he wishes to watch.

A cable, satellite or broadcast environment has only a few transmitters which do compression but a large number of receivers which have to do decompression. Therefore, the economics is dictated to a large extent by the cost of decompression. The choice of the compression algorithm ought to make the decompression extremely simple by transferring much of the cost to the transmitter, thereby creating an asymmetrical algorithm. The existing video standards such as px64 kbit/sec for videoconference and MPEG for multimedia do not

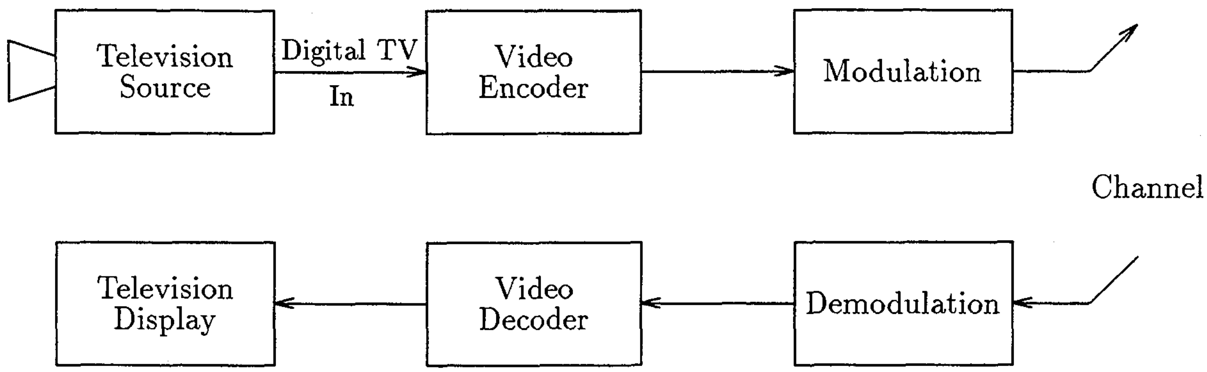


**Figure 1: Video Compression and Decompression**

explicitly incorporate this requirement. In a number of situations, cost of the encoder is also important (e.g., camcorder). Therefore, a modularly designed encoder which is able to trade off performance with complexity but which creates the data decodable by a simple decompressor may be the appropriate solution. Obviously, this tradeoff will change as a function of the integrated circuits technology.

In a number of instances, the original source material may have to be compressed and decompressed several times. In studios it may be necessary to store the compressed data and then decompress it for editing. Such multiple encodings of the signal is bound to increase the visibility of the coding artifacts; however, a choice of the coding algorithm should minimize the loss of quality associated with multiple encodings.

It is commonly believed that much of the material for the services based on digital compressed television will be from films. The conversion from film to video using 3:2 pull down creates a unique type of correlation in the signal. It is desirable for the compression algorithm to automatically detect such correlation and adapt itself to achieve a high degree of efficiency without increasing the complexity of the receiver.

The technology for storage of digital signals on a variety of media has made significant strides over the last few years. If the compressed digital signal is stored on a digital tape recorder, then some of the functions that we have become accustomed to should be easy to provide from the compressed digital signal. These include fast forward and backward searches, still frames, etc. This was an important consideration in the development of the MPEG-1 algorithm.

As we move toward digital television, we see a large number of possible picture resolutions (e.g., NTSC, CCIR-601, HDTV). It is desirable for the compression scheme to be compatible over these different resolutions. This will allow, for example, an HDTV decompressor to decode the compressed NTSC as well as CCIR-601 signals without much duplication of the hardware. Also, a compatibility between the transmission formats chosen for the NTSC, CCIR-601 and HDTV signals would be desirable. Moreover, such a common transmission format should allow easy interconnection or transmission over different media and telecommunications networks.

In a number of situations, particularly in the cable head end, one may wish to add an insert into a compressed digital bit stream.

It is desirable to add the insert without having to fully decode the signal. If only a small part of the coded bit stream can be affected and these effects can be localized on the picture signal, then the adding of inserts can become less damaging to the original signal.

It is clear by looking at these requirements that the broadcasting, cable, and satellite applications have requirements that are different from teleconferencing and multimedia applications. Therefore, it is not surprising that a different class of algorithms would suit these applications. In the next section we describe the basic compression techniques, and compare them in preparation for the description of an algorithm that handles the above requirements rather well.

## BASIC COMPRESSION TECHNIQUES

A number of compression techniques have been developed for coding of video signals.[4] A compression system typically consists of a combination of these techniques to satisfy the type of requirements that we listed in the previous section. The first step in compression usually consists of **Decorrelation** i.e. reducing the spatial or temporal redundancy in the signal. The candidates for doing this are:

1. Making a prediction[1] of the next sample of the picture signal using some of the past and subtracting it from that sample. This converts the original signal into its unpredictable part (usually called prediction error).

2. Taking a transform[2] of a block of samples of the picture signal so that the energy would be compacted in only a few transform coefficients.

The second step is **Selection** and **Quantization** to reduce the number of possible signal values. Here, for DPCM, the prediction error may be quantized sample at a time or a vector of prediction error of many samples may be quantized all at once. Alternatively, for transform coding, only important coefficients may be selected and quantized. The final step is **Entropy Coding** which recognizes that different values of the quantized signal occur with different frequencies and, therefore, representing them with unequal length binary codes reduces the average bit rate. We give below more details of the following techniques since they have formed the basis of most of the compression systems:

a) Predictive Coding (DPCM)

b) Transform Coding

c) Motion Compensation

d) Vector Quantization

e) Entropy Coding

f) Incorporation of Perceptual Factors

### Predictive Coding (DPCM)

In predictive coding, the strong correlation between adjacent pels (spatially as well as temporally) is exploited. As shown in Figure 2, an approximate prediction of the sample to be encoded is made from previously coded information that has already been transmitted. The error (or differential signal) resulting from the subtraction of the prediction from the actual value of the pel is quantized into a set of discrete amplitude levels. These levels are then represented as binary words of fixed or variable lengths and

[1]This is also the first step in a family of compression algorithms known as Differential Pulse Code Modulation (DPCM).

[2]This is called Transform Coding. Transform is simply a linear combination of all the pels in the block.

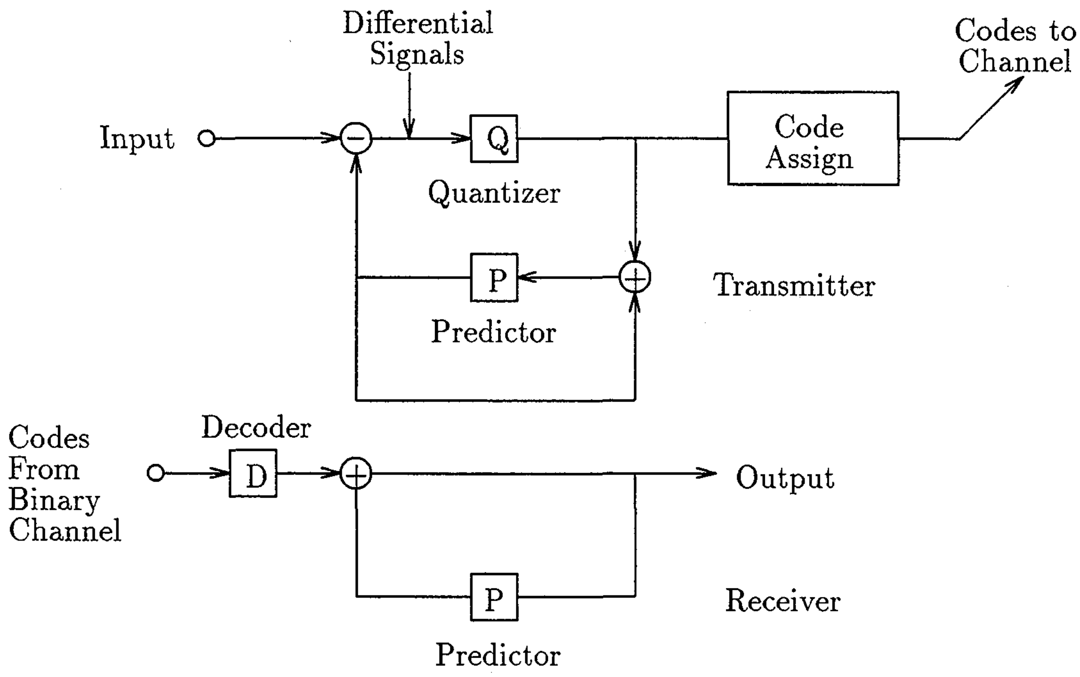


**Figure 2: Block Diagram of a Predictive Encoder and Decoder**

sent to the channel for transmission. The predictions may make use of the correlation in the same scanning line or adjacent scanning lines or previous fields. A particularly important prediction is the **motion compensated** prediction. If a television scene contains moving objects and an estimate of frame-to-frame translation of each moving object is made, then more efficient prediction can be performed using elements in the previous frame that are appropriately spatially displaced. Such prediction is called motion compensated prediction. The translation is usually estimated by matching a block of pels in the current frame to a block of pels in the previous frames at various displaced locations. This is shown in Figure 3. Various criteria for matching and algorithms to search for the best match have been developed.[4] Typically, such motion estimation is done only at the transmitter and the resulting motion vectors are used in the encoding process and also separately transmitted for use in the decompression process.

Transform Coding

In transform coding (Figure 4) a block of pels are transformed by transform **T** into another domain called the transform domain, and some of the resulting coefficients are quantized and coded for transmission. The blocks may contain pels from one, two or three dimensions. The most common technique is to use a block of two dimensions. Using one dimension does not exploit vertical correlation and using three dimensions requires several frame stores. It has been generally agreed that **Discrete Cosine** transform is best matched to the statistics of the picture signal and moreover, since it has a fast implementation, it has become the transform of choice.[3] The advantage

[3]Also, recent advances have dramatically decreased the cost of implementing transforms using LSI[5].

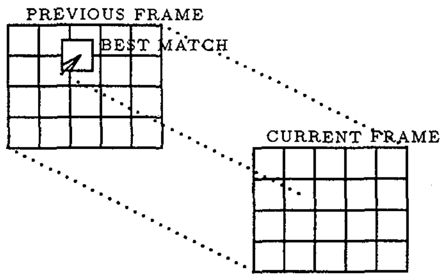


**Figure 3: Block Matching for Motion Estimation**

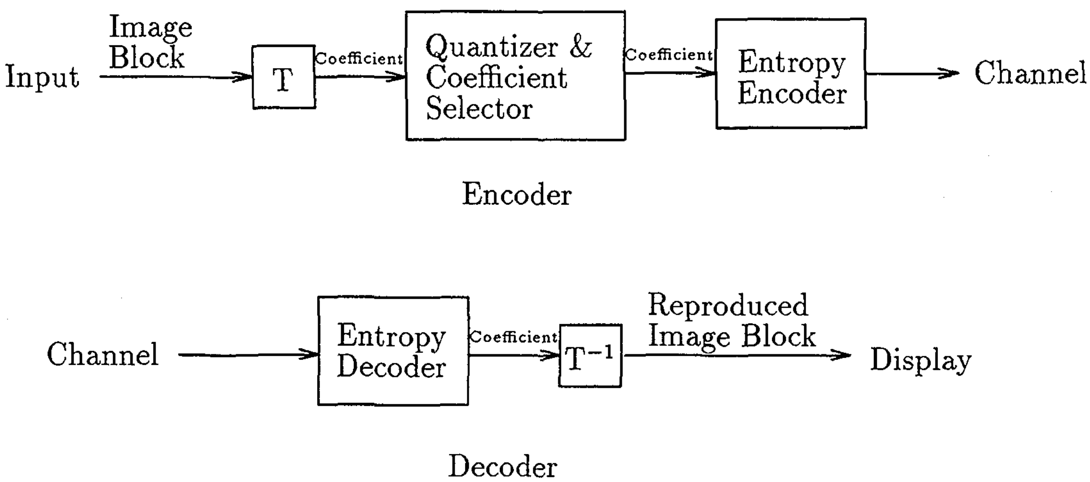


**Figure 4: Transform Coding**

of transform coding comes about mainly from two mechanisms. First, not all of the transform coefficients need to be transmitted in order to maintain good image quality, and second, the coefficients that are selected need not be represented with full accuracy. Loosely speaking, transform coding is preferable to predictive coding for lower compression rates and where cost and complexity are not extremely serious issues. Most modern compression systems have used a combination of predictive and transform coding. In fact, motion compensated prediction is performed first to remove the temporal redundancy, and then the resulting prediction error is compressed by two-dimensional transform coding using Discrete Cosine transform as the dominant choice.

### Vector Quantization

In predictive coding, described in the previous section, each pixel was quantized separately using a scalar quantizer. The concept of scalar quantization can be generalized to vector quantization in which a group of pixels are quantized at the same time by representing them as a code vector. Such a vector quantization can be applied to a vector of prediction errors, original pels, or transform coefficients. As in Figure 5, a group of 9 pixels from a $3 \times 3$ block is represented to be one

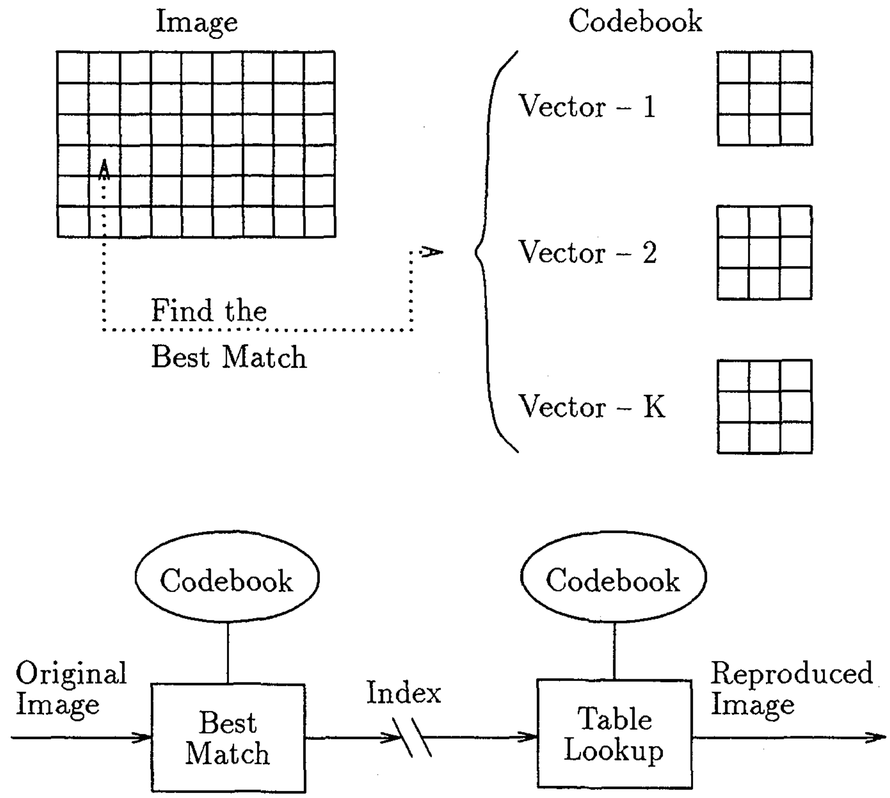


**Figure 5: Vector Quantization**

of the $k$ vectors from a codebook of vectors. The problem of vector quantization is then to design the codebook and an algorithm to determine the vector from the codebook that offers the best match to the input data. The design of codebook usually requires a set of training pictures and can grow to a large size for a large block of pixels. Thus, for an $8 \times 8$ block compressed to two bits per pel, one would need $2^{128}$ size codebook. Matching the original image with each vector of such a large size codebook requires a lot of ingenuity. However, such matching is only done at the transmitter, and the receiver is considerably simple since it does a simple table lookup.

Entropy Coding

If the quantized output values of either a predictive or a transform coder are not all equally likely, then the average bit rate can be reduced by giving each one of the values a different word length. In particular, those values that occur more frequently are represented by a smaller length code word. If a code with variable length is used, and the resulting code words are concatenated to form a stream of bits, then correct decoding by a receiver requires that every combination of concatenated code words be uniquely decipherable. A variable word length code that achieves this and at the same time gives the minimum average bit rate is called **Huffman Code**. Variable word length codes are more sensitive to the effect of transmission errors since synchronization would be lost in the event of an error. This can result in several code words getting decoded incorrectly. A strategy is required to limit the propagation of errors in the presence of Huffman Codes.

## Incorporation of Perceptual Factors

The perception based coding attempts to match the coding algorithm to the characteristics of human vision. We know, for example, that the accuracy with which the human eye can see the coding artifacts depends upon a variety of factors such as the spatial and temporal frequency, masking due to the presence of spatial or temporal detail, etc. A measure of the ability to perceive the coding artifact can be calculated based on the picture signal.[6] This is used, for example, in transform coding to determine the precision needed for quantization of each coefficient. Perceptual factors control the information that is discarded on the basis of its visibility to the human eye. It can, therefore, be incorporated in any of the above basic compression schemes.

## Comparison of Techniques

Figure 6 represents an approximate comparison of different techniques using compression efficiency vs. complexity as a criterion under the condition that the picture quality is held constant at a "8-bit PCM level". The compression efficiency is in terms of compressed bits per Nyquist sample, and therefore, different resolution and bandwidth pictures can be simply scaled by proper multiplication to get the relevant bit rates. The complexity allocated to each codec should not be taken too literally; rather, it is an approximate estimate relative to the cost of a PCM codec which is given a value of five. Furthermore, it is the complexity of only the decoder portion of the codec, since that is the most important part of the digital television. The relation of cost to complexity is controlled by an evolving technology, and codecs with high complexity are fast becoming inexpensive through the use of application specific video DSP's and submicron device technology. Also, most of the proposed systems are a combination of several different techniques of Figure 6, making such comparisons difficult. As we remarked before, the real challenge is to combine the different techniques to engineer a cost-effective solution for a given service. The next section describes one example of such a codec.

## A COMPRESSION SCHEME

In this section we describe a compression scheme that combines the above basic techniques to satisfy the requirements of Section 2. We have used this compression scheme successfully for both the noninterlaced HDTV signals and interlaced NTSC and CCIR-601 signals.[7,8]

Three basic types of redundancy are exploited in the video compression process. Motion compensation removes temporal redundancy, two-dimensional DCT removes spatial redundancy, and perceptual weighting removes amplitude redundancy by putting quantization noise in less visible areas.

Temporal processing occurs in two stages. The motion of objects from frame to frame is estimated using hierarchical block matching. Using the motion vectors, a displaced frame difference (DFD) is computed which generally contains a small fraction of the information in the original frame. The DFD is transformed using DCT to remove the spatial redundancy. Each new frame of DFD is analyzed prior to coding to determine its rate versus perceptual distortion character-

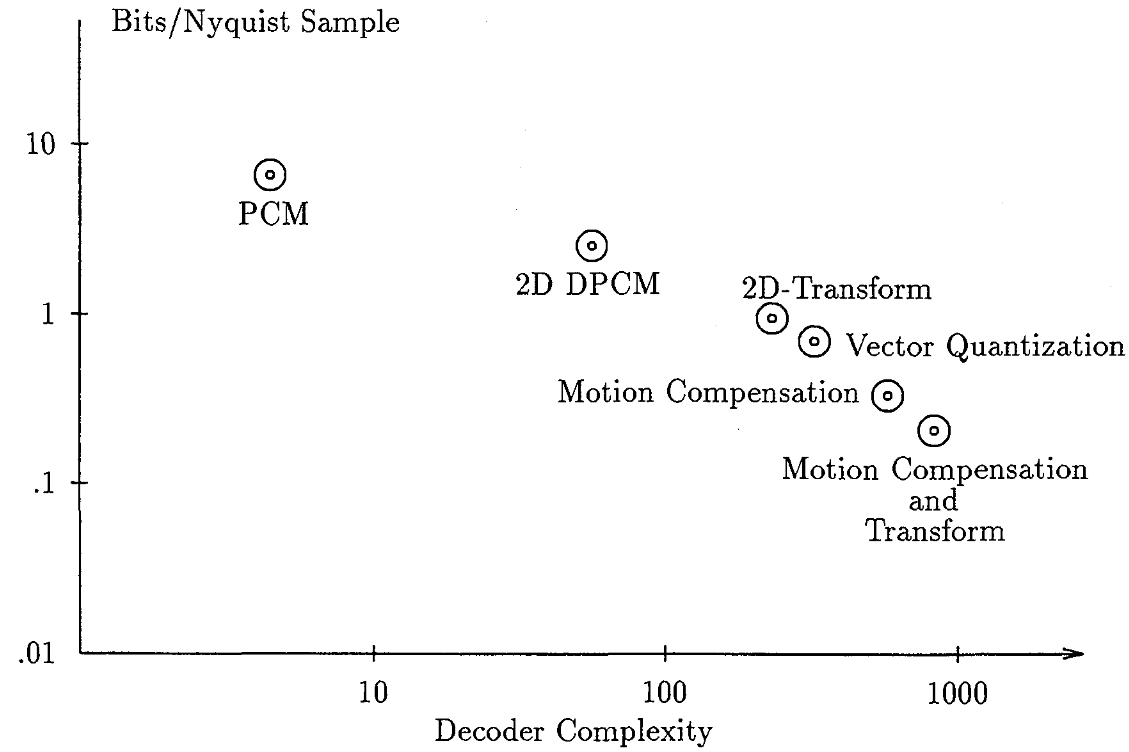


**Figure 6: Compression vs. Complexity**

istics and the dynamic range of each coefficient (forward analysis).[4] Quantization of the transform coefficients is performed based on the perceptual importance of each coefficient, the precomputed dynamic range of the coefficients, and the rate versus distortion characteristics. The perceptual criterion uses a model of the human visual system to determine a human observer's sensitivity to color, brightness, spatial frequency and spatial-temporal masking. This information is used to minimize the perception of coding artifacts throughout the picture. Parameters of the coder are optimized to handle the scene changes that occur frequently in entertainment/sports events, and channel changes made by the viewer. The motion vectors, compressed transform coefficients and other coding overhead bits are packed into a format which is highly immune to transmission errors.

[4]This also helps us to automatically detect the 3:2 pull down signals and adapt to them.

The encoder is shown in Figure 7. Each frame is analyzed before being processed in the encoder loop. The motion vectors and control parameters resulting from the forward analysis are input to the encoder loop which outputs the compressed prediction error to the channel buffer. The encoder loop control parameters are weighed by the buffer state which is fed back from the channel buffer.

In the predictive encoding loop, the generally sparse differences between the new image data and the motion-compensated predicted image data are encoded using adaptive DCT coding. The parameters of the encoding are controlled in part by forward analysis. The data output from the encoder

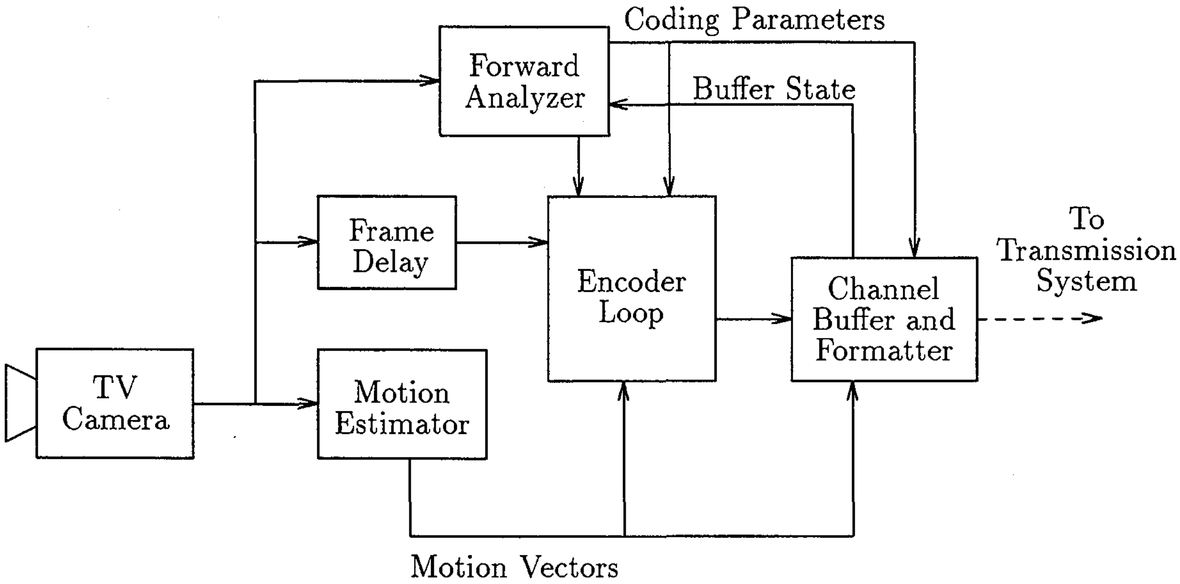


**Figure 7: Encoder**

consists of some global parameters of the video frame computed by the forward analyzer and transform coefficients that have been selected and quantized according to a perceptual criterion.

Each frame is composed of a luminance frame and two chrominance difference frames which are half the resolution of the luminance frame horizontally. The compression algorithm produces a chrominance bit-rate which is generally a small fraction of the total bit-rate, without perceptible chrominance distortion.

The output buffer has an output rate of between 2 to 7 Mbits/sec and has a varying input rate that depends on the image content. The buffer history is used to control the parameters of the coding algorithm so that the average input rate equals the average output rate. The feedback mechanism involves adjustment of the allowable distortion level, since increasing the distortion level (for a given image or image sequence) causes the encoder to produce a lower output bit rate.

The encoded video is packed into a special format before transmission which maximizes immunity to transmission errors by masking the loss of data in the Decoder. The duration and extent of picture degradation due to any one error or group of errors is limited. The Decoder is shown in Figure 8. The compressed video data enters the buffer which is complementary to the compressed video buffer at the encoder. The decoding loop uses the motion vectors, transform coefficient data, and other side information to reconstruct the NTSC images. Channel changes and severe transmission errors are detected in the decoder causing a fast picture recovery process to be initiated. Less severe transmission errors are handled gracefully by several algorithms depending on the type of error.

Processing and memory in the decoder are minimized. Processing consists of one inverse spatial transform and a variable length decoder which are realizable in a few VLSI

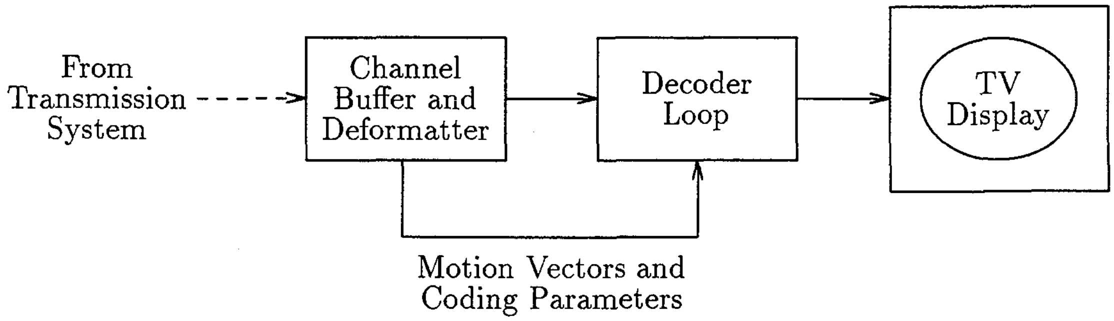


**Figure 8: Decoder**

chips. Memory in the decoder consists of one full frame and a few compressed frames.

## CONCLUSIONS

We have presented in this paper a set of requirements specific to digital compression of television signals for broadcast, cable and satellite distribution. These requirements are considerably different than those for videoconferencing or multimedia computing. We have outlined a collection of state of the art basic compression techniques and synthesized a compression scheme that meets the special requirements of the broadcast, cable and satellite industries. Rapid advances in video compression, and semiconductor technology (processing as well as memory) have made digital television realizable.

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[8] A. N. Netravali, et al., "A High Performance Digital NTSC/CCIR-601 Coder," SCTE Fiber Optics Conference, 1992.

# A DIGITAL COMPRESSED VIDEO TRANSMISSION SYSTEM - WITH SIMULATION RESULTS OF ECHOES IN 64-QAM TRANSMISSION

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Jerrold Communications

Abstract

Digital compressed transmission is developing as an attractive method of video delivery to both headends and subscribers. This paper will review compression and transmission, and then focus on some key performance parameters for digital transmission. Finally we present the results of computer simulations of the effect of echoes on 64-QAM digital transmission.

## Introduction

The US space program, through a multitude of outstanding examples, made the creation of spin-off technologies famous. Digital compression for Cable was born as just such a spin off, not from NASA, but from the desire to deliver high definition TV to consumers. The need to squeeze a high bandwidth HDTV signal into existing 6 MHz TV channel bands created the means to deliver several standard NTSC programs in the spectrum currently occupied by just one.

Various methods of transmitting HDTV signals have been proposed to the FCC. Of the six advanced TV systems under review four propose digital transmission. Each of these digital HDTV proponents has also proposed a multichannel NTSC delivery system in response to the CableLabs compression request for proposal.

## Compression

Since the landmark announcement of the DigiCipher system in 1990, several other digital video compression systems have been proposed. Most of these, including all of the digital HDTV proponent systems and the work of the International Standards Organization's MPEG[1] committee, have been similar in function, based on the discrete cosine transform (DCT) and motion compensated inter-frame processing. This has proved to be an efficient technique for removing the many redundancies and, as needed, lower value components, from the video signal, with a minimum perceived effect on the reconstructed picture.

The use of digital compression offers operators and subscribers much more than just an increased number of programs. Digital transmission means every subscriber will get the same very high quality picture. It will be free of the noise and distortion common in analog systems. Most digital compression systems use component, not composite, color. With these systems there will be, for the first time a means of delivering a true component signal to the S-Video jack on high end consumer TV's and VCR's. Sound quality will also be uniformly excellent, indistinguishable from compact disc. True digital encryption will provide a level of security never before possible for video on Cable. Programming these new digital channels will become easier and more reliable with the introduction of digital switchers and digital compressed storage at the headend.

With satellite delivered digital compressed programs passing through to digital transmission on the Cable plant, the operator will no longer need to worry about picture quality anywhere in his operation. Headend processing to realize this, ranges from changing only the modulation, which offers very limited local features, to full video decompression and recompression, which allows all the control and programming options the operator now has with analog video.

The advantages of digital media have been realized by the telephone industry. They are arriving in TV network studios now, and they will soon offer major benefits to Cable operators system-wide.

Transmission

For analog video transmission, different forms of modulation are required on satellite and Cable. The satellite channel has a bandwidth of at least 24 MHz, but a reliable carrier to noise ratio (C/N) of only about 8dB. The Cable channel has only a 6 MHz bandwidth but typically 40 dB or more C/N. To make effective use of these very different transmission paths, we use a unique form of modulation in each. Frequency modulation (FM), a wide band noise insensitive technique, is carried on satellite. Amplitude modulation (AM), a noise sensitive but bandwidth efficient approach is appropriate for Cable.

For digital transmission of video the same logic applies. Satellite systems use quadrature phase shift keying (QPSK) which is an FM-like modulation for digital carriers. QPSK requires a wide bandwidth channel but offers high immunity to noise. For Cable, two AM techniques are proposed for digital carriers: double side band quadrature amplitude modulation (QAM), and the vestigial sideband multi-amplitude technique often called 4-VSB. These two approaches to amplitude modulation are very similar in performance. Both trade off noise immunity for spectral efficiency relative to QPSK. The following work addresses QAM with 6 bits per symbol.

To compare the data capacity of the disparate satellite and Cable channels we turn to Claude Shannon's pioneering work in information theory[2]. He developed a formula for calculating the maximum theoretical capacity of a channel based on bandwidth and S/N alone. For a signal of power S transmitted over a channel with noise power N and bandwidth W, the channel capacity C in bits per second is:

$$C = W \log_2(1 + S/N).$$

A satellite channel with 24 MHz bandwidth and 8 dB optimum signalling S/N would have a Shannon limit of 69 million bits per second (Mbps). Of course, this is for an ideal modem, operating in a channel free of secondary impairments. Commercial satellite modems commonly approach about one half this theoretical maximum rate.

A Cable channel with 6 MHz bandwidth and 35 dB optimum signalling S/N has almost the same Shannon limit as the above satellite channel: 70 Mbps. The Cable channel may have more secondary impairments and certainly the tolerable modem cost is much lower, but it is clear the Cable channel has a data transmission capacity similar to that of satellite.

Transmission Analysis

In designing a new digital communications system many factors must be considered. Among these are a wide variety of channel impairments, many of which will occur simultaneously in a working system. It is the combined effects of these impairments that define the transmission system operating parameters.

For the traditional Cable environment there are three principle impairments to the transmission of digital QAM carriers: channel noise, echoes, and modem implementation loss. Although there are many other impairments they are expected to be either avoidable, like not running high-level sweeps through active digital spectrum, or of lesser significance to digital transmission, like CTB and CSO.

Many prior papers have defined the effect of the primarily white channel noise on QAM. The following analysis will briefly discuss the specification of digital C/N in the Cable plant. We will then detail the results of our simulations of echoes on QAM transmission in the presence of white noise. Modem implementation loss, the third major factor, will be significant due to the severe cost constraints on the subscriber terminal for Cable applications. It includes such factors as filter imprecision, phase noise, receiver noise figure, clock jitter, and computation error. These factors are hardware specific and beyond the scope of this paper.

The block diagram for the following simulations is shown in Figure 1. It consists of a random data source, 64-QAM modulator, 5 MHz Nyquist bandwidth (BW) transmit filter, white noise and recursive echo sources, a matched receive filter, 64-QAM demod, and an error counter. All blocks are ideal floating point simulation models. Figures 2 and 3 show the eye-diagram and constellation with the noise and echo sources set to zero amplitude.

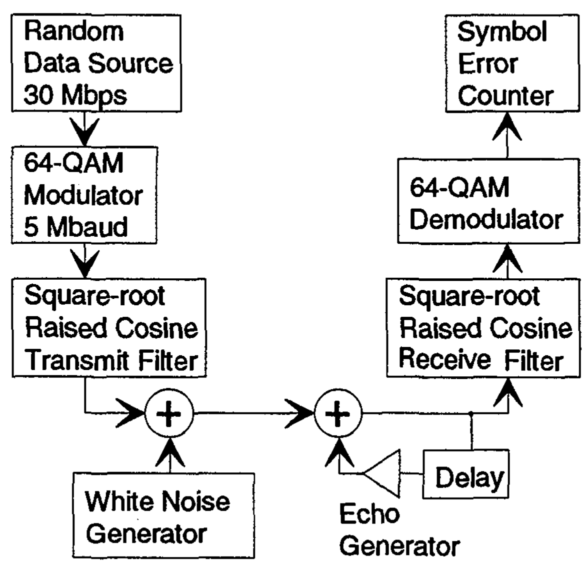


Figure 1: Simulation Block Diagram

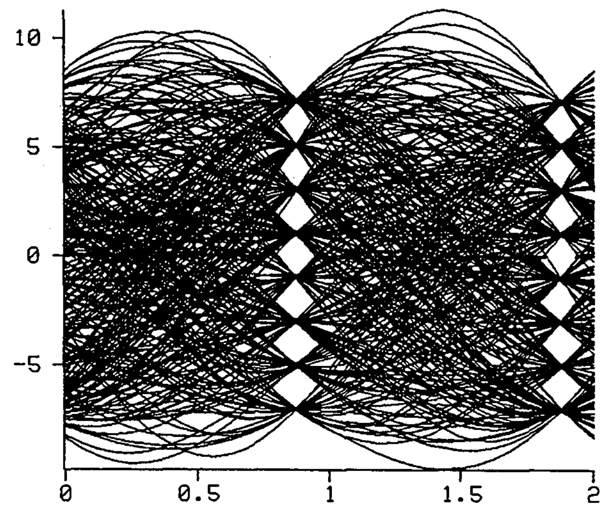

Figure 2: Eye Diagram - Noiseless

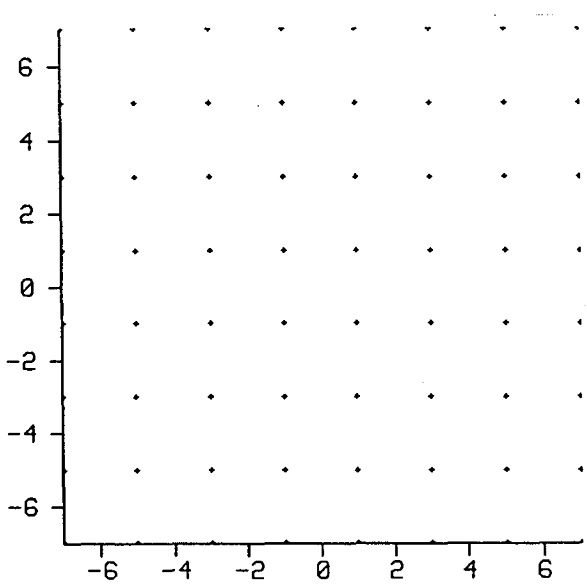

Figure 3: 64-QAM Constellation

Noise
Channel noise is the familiar broadband white noise floor we now measure as C/N on the Cable distribution plant. Easily seen with a spectrum analyzer, it comes primarily from the active devices in the distribution plant: trunk amps, line extenders, AML, and fiber links. Digital carrier to noise as referred to in the literature and this paper, is average modulated digital carrier power divided by rms noise power in the modulation bandwidth. Digital carrier modulation bandwidth will be about 5 MHz. This contrasts with our standard analog C/N spec. We measure analog carrier power during sync tip peak level and divide by noise power in the 4 MHz video bandwidth.

To relate these different C/N measures, consider that average modulated analog carrier power depends on picture content. Indeed, this is why it's impractical to measure average power on a modulated analog carrier. The peak to average ratio ranges from 7 dB to 2 dB for 0 to 100 IRE pictures respectively. A long time average of all possible picture contents will eventually yield a peak to average ratio near that for a 50 IRE flat field: 5dB. From this we can say that a digital carrier with the same average power as an analog carrier will have a 6 dB lower measured C/N. Five dB for the peak vs. average power measurement method and 1 dB for the 4 vs 5 MHz noise bandwidth.

Figure 4 shows symbol error rate as a function of (average modulated) digital carrier to noise. Simulation results are shown to be close to calculated theoretical values. Symbol error rate (SER) is very close to bit error rate (BER) when error rates are low ($<10^{-2}$). Figures 5 and 6 show the effects of -24 dB white noise on the eye-diagram and constellation.

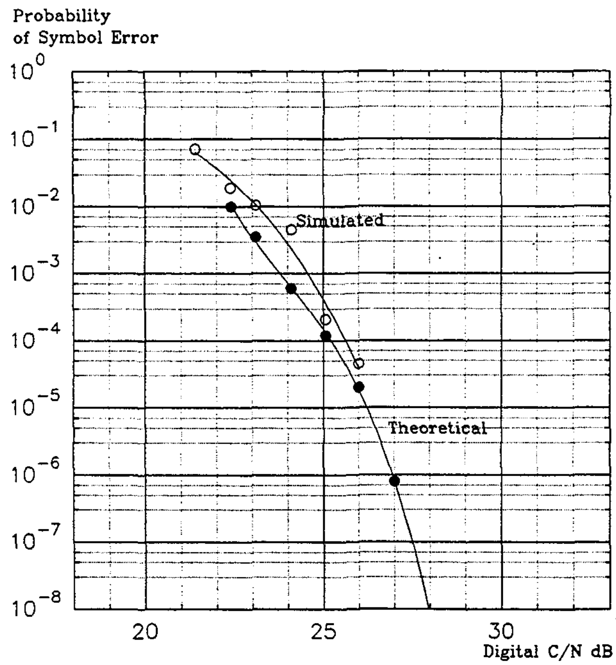


Fig.4: Error Performance of 64 QAM with White Noise

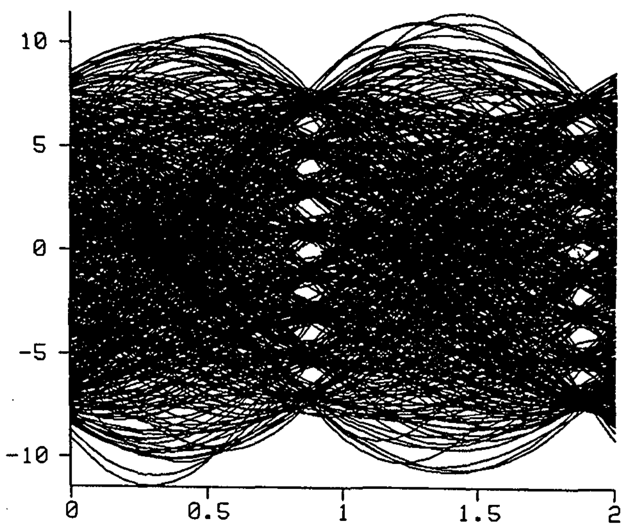


Figure 5: 24 dB Digital C/N

## Echoes

There are many sources of echoes in the traditional Cable plant. Most of these are the result of imperfect impedance matching at the myriad connection points along the RF distribution path. At each of these points a

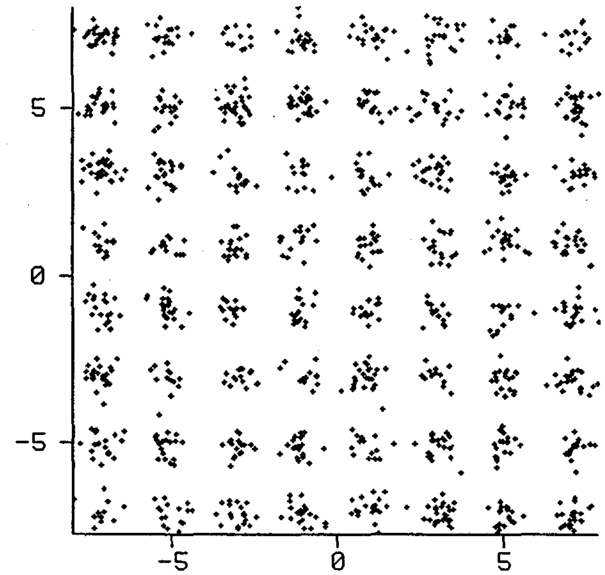


Figure 6: 24 dB digital C/N

small amount of the downstream RF power reflects back to its source. The example in figure 7 shows the reflection through the 8 dB return loss of a digital receiver input. Reflected power returns to the next upstream device, delayed and attenuated by its double pass through the connecting coax. A fraction of this returning power then reflects again through a second return loss. The twice reflected signal is now a downstream echo with a delay and power relative to the primary signal.

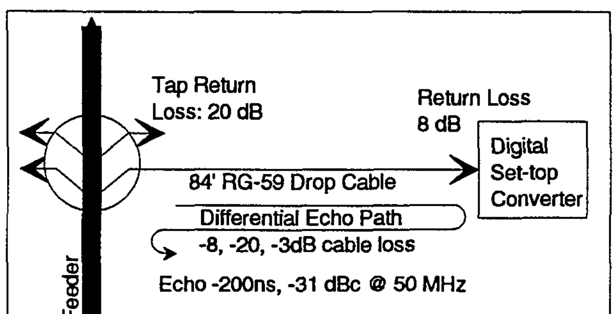


Figure 7. Echo Example

The simulation model for echo generation is a recursive structure as shown in

Figure 1. This is appropriate as the echo is by nature a repeating effect. A -20 dB 200ns delayed impulse will be followed by impulses at -40 dB 400ns, -60 dB 600ns, ..., as the twice reflected signal continues to be reflected four, six, ..., times with diminishing amplitude.

Echoes will have a specific phase relative to the primary signal carrier, determined by their exact delay in carrier cycles. A 200ns echo on a 50 MHz carrier will have a 0 degree phase because the delay is exactly 10 cycles of carrier. The constellation of a signal with this echo is shown in Figure 8. A 205ns echo on a 50 MHz carrier will have a 45 degree phase as shown in Figure 9. In this example, the difference would be an 86 versus 88 foot run of coax between the reflecting devices.

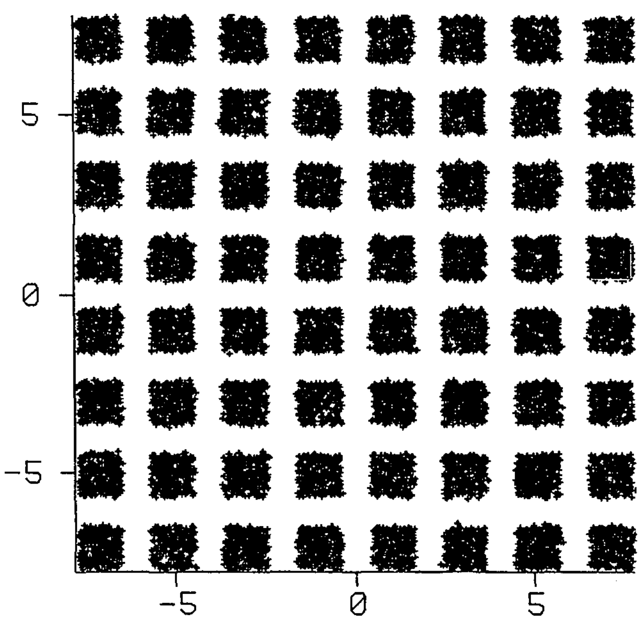

Figure 8: 0° Echo Phase

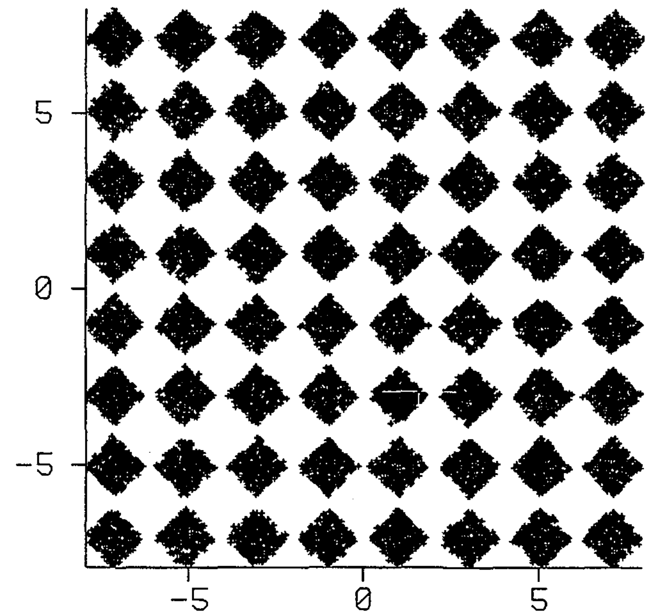

Figure 9: 45° Echo Phase

The effect of phase on QAM transmission in the presence of other impairments is shown in figure 10. We used white noise at -30dBc to model the combined effect of all non-echo impairments for this example. The phase of echo has a 1 to 2 dB effect on sensitivity, with the maximum effect at 45 degrees. All following simulations assume a 45 degree echo phase.

Figure 11 shows the effects of a range of echo powers on QAM with three different levels of white noise. From Figure 4, the SER for -25 dB noise alone is $10^{-4}$. At this noise level even a -37 dB echo increases the SER substantially, to $10^{-3}$. With white noise in the range of -27 dB echoes of -30 dB are still significant. As used here 'white noise' includes all of the non-echo transmission impairments including: channel noise, implementation loss, and secondary transmission impairments. For the sum of these in the range of -25 to -30 dB, even very low level echoes will be significant.

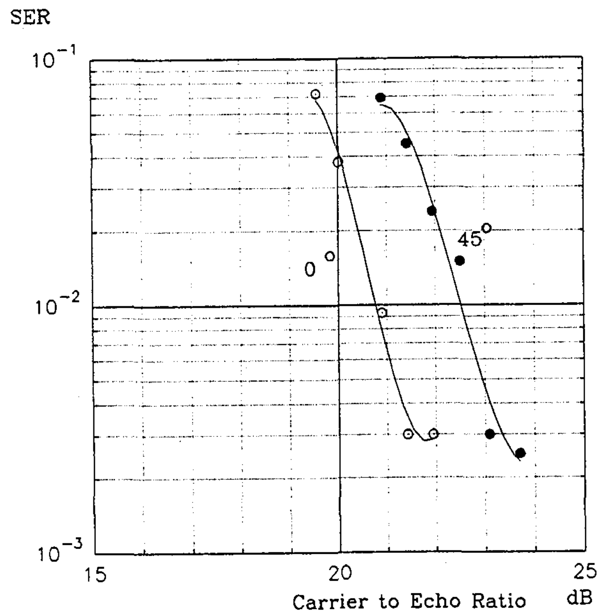


Fig. 10: Effect of Echo Phase

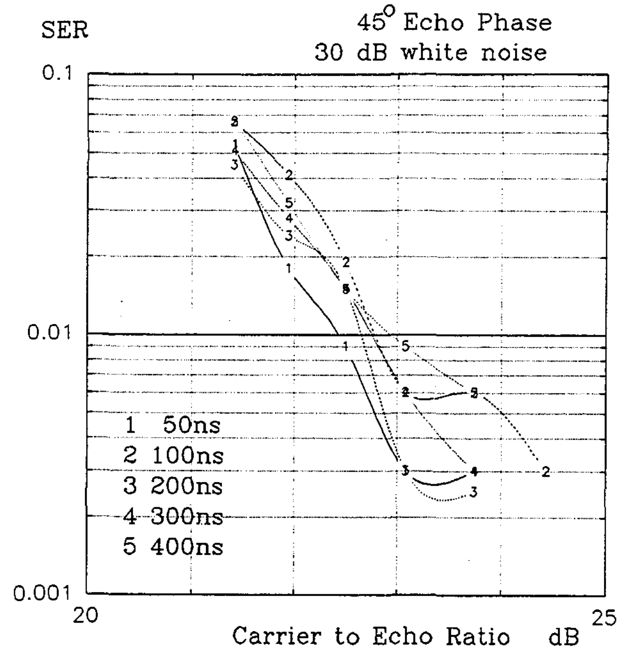


Fig. 12: Effect of Delay on Echo Sensitivety

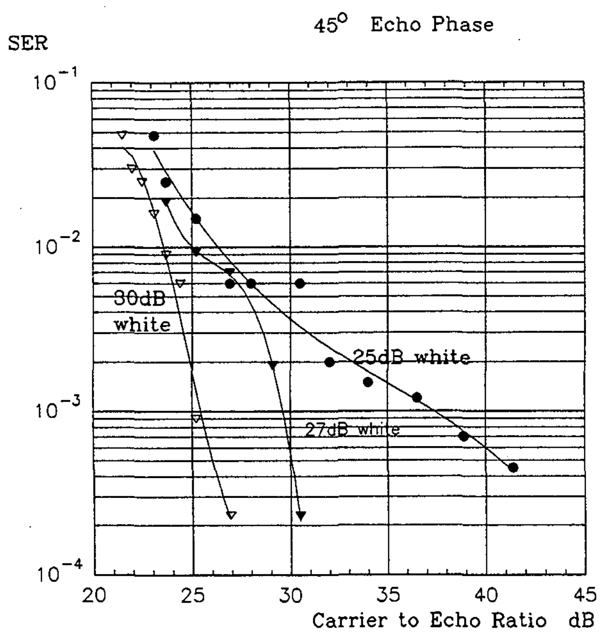


Fig.11: Echo Sensativity For Three Level of Added White Noise

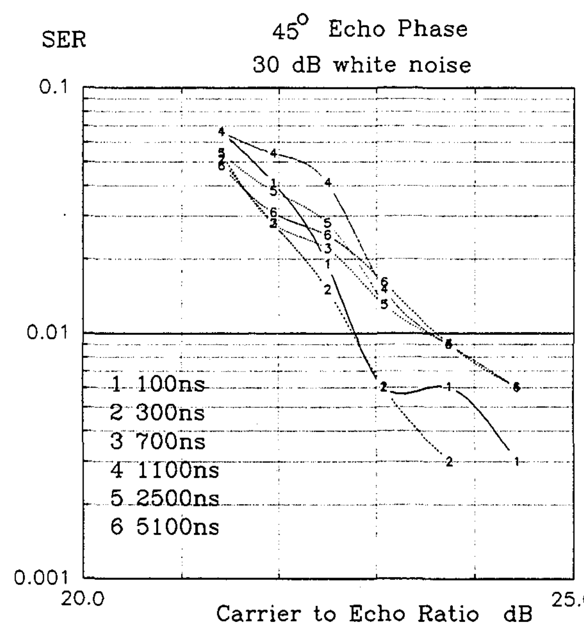


Fig.13: Effect of Delay on Echo Sensitivity

Figures 12 and 13 show the results of simulations over a range of echo powers for several different delays. All include white noise of -30 dB. The echo power is far more significant to SER than the delay. The effect on SER is almost unchanged over the range of delays. Even the short delays of 50ns and 100ns, one quarter and one half the 200ns symbol period, have the same general level of impairment as the longer echoes.

## Conclusion

Digital compressed transmission will offer substantial benefits to the Cable operator and subscriber. Data rates similar to those of satellite systems are feasible using unique modulation, and demodulation, optimized for the Cable environment. These new digital carriers will require a new C/N measurement specification.

Digital transmission of 64-QAM will be sensitive to echoes at very low power. For a sum of other impairments equivalent to white noise of -30dBc or above, even echoes below -30dBc, will be significant.

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[1] Motion Picture Expert Group (MPEG)

[2]Shannon, Claude E., a series of papers including "A Mathmatical Theory of Communications," Bell System Technical Journal, volume 27, pages 379-423, July 1948.

# A NEW TECHNIQUE FOR MEASURING BROADBAND DISTORTION IN SYSTEMS WITH MIXED ANALOG AND DIGITAL VIDEO

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## I. ABSTRACT

Several proposals have been made to augment the channel capacity of existing or proposed CATV systems by adding compressed digital video signals to conventional analog VSB signals. Such a proposal brings with it the difficulties of quantifying the degradation of broadband distortion performance (CTB, XMOD, DSO and CSO) associated with the additional loading caused by the new digital signals. Conventional techniques that measure device and system distortion performance by modeling the digital signals with CW carriers at the picture carrier frequencies of conventional analog VSB channels will give extremely disappointing results.

A new technique is described that models the digital signals in a manner that accurately represents the digital channel energy spectral density. Measurements using this technique are given that accurately predict distortion performance that is far better than that predicted by conventional CW carrier techniques.

A simple means of generating this signal is described. Test results are given that compare the performance of test devices when loaded with only conventional analog VSB signals to the performance of the test devices when loaded with mixed analog and digital signals.

## II. CONVENTIONAL DISTORTION TECHNIQUES

It is a primary goal in the electronic processing of CATV signals to minimize any corruption of these signals. This design goal, as it interplays with economic considerations, has set the practical limits for performance of the system components (whether they be program supplier studio electronics, satellite delivery systems, headend electronics, distribution plant or subscriber electronics). In the case of broadband electronics, the principle technical trade-offs have been between broadband distortion and noise limitations.

Well established techniques have been used throughout the industry for several years in order to characterize the broadband distortion performance of conventional CATV electronics. Composite triple beat, discrete second order, composite second order and crossmodulation have historically been used as figures of merit for system performance. These techniques are closely linked with the types of signals used in conventional CATV applications.

In the United States, Canada, Mexico, Japan and several other countries, NTSC-M is the standard for color television transmission. As we are all aware, this system utilizes vestigial sideband modulation with a 6 MHz channel bandwidth. The picture carrier contains the most significant amount of channel energy, from a distortion standpoint, and is located 1.25 MHz from the lower channel edge. Another significant characteristic is the horizontal line frequency of 15734.264 Hz. The distortion techniques mentioned above utilize this characteristic in order to quantify system performance.

Knowing that the majority of the energy in an NTSC-M television signal is located at or very near the picture carrier leads to the understanding that the majority of the distortion in a CATV system results from that energy. In many laboratory or test applications the modulated video signal is replaced with a CW carrier at the picture carrier frequency. In a CATV system with a standard frequency line-up (not HRC or IRC) the frequency characteristics of these carriers have the useful property that third order distortion products fall at or very near to the picture carrier frequencies of affected channels. A simple examination of the possible linear combinations of an odd number of picture carrier frequencies will demonstrate this fact (for example: Channel 7 + Channel 8 - Channel 37 = 175.25 MHz + 181.25 MHz - 301.25 MHz = 55.25 MHz = Channel 2). This is the driving factor in the definition of CTB (composite triple beat). Note that the name CTB is somewhat of a misnomer as this distortion product is composed of not only third order beats, but also many higher odd-order distortion products.

A similar examination of the linear combination of the frequencies of any two picture carriers will result in a frequency that falls at 1.25 MHz above or below the picture carrier frequency of an affected channel in a standard system. The result is of course discrete second order distortion. A simple extension of this phenomena to any combination of an even number of carrier frequencies results in an explanation of composite second order, a phenomena that grows significantly as the broadband system bandwidth exceeds 450 MHz.

Crossmodulation is closely linked to the frequency components contained within the modulating signal of the RF spectrum. In the case of NTSC-M, there is a significant spectral content at 15.734 kHz. In a test environment, the modulating waveform is simplified to a 15.734 kHz square wave. Under these test conditions, it is relatively easy to quantify the parasitic modulation impressed upon an unmodulated carrier when included in a broadband spectrum where all other carriers are modulated and then passed through a device that can create nonlinear distortion. The ratio of the amplitude of that parasitic modulation to the modulation generated by 100% modulating that carrier with a 15.734 kHz square wave is defined as the crossmodulation.

In all of the cases mentioned above, the distortion measurements, by definition, are linked closely to the baseband video format and RF modulation techniques. An easy extension of these concepts may be used to quantify broadband distortion in the case of HRC or IRC frequency line-ups. It is also possible to extend these concepts to apply to other video formats, such as PAL and SECAM as well as their various RF spectrum formats, such as I, B, G, N, etc.

## III. INITIAL STUDIES OF LOADING OF CATV DEVICES FOR DIGITAL VIDEO

A popular scenario for expanding the capabilities of a CATV plant is based on the inclusion of digital video in the upper part of the CATV broadband spectrum. Digital video promises to greatly increase the capability of this spectrum by permitting digital compression of the analog video signal. This compression will allow the carriage of up several times the number of channels in a given block of spectrum as would be possible using conventional VSB techniques. For example, one scenario might be to include conventional analog signals in the spectrum from 54 MHz to 550 MHz in order to serve the existing base of CATV subscribers. Additional digital video services could be included from 550 MHZ to 750 MHz. However, it is not clear how the inclusion of this potentially large number of video signals would affect the overall system distortion performance.

Initial testing of typical equipment was performed by Scientific-Atlanta in order to quantify the performance of 550 MHz CATV electronics when loaded with video carriers from a Matrix signal generator from 54 MHz to 750 MHz. This is the natural extension of the existing art of distortion measurements. Under these test conditions it was shown that distortion deteriorates significantly when conventional video carriers between 550 MHz and 750 MHz are included in the spectrum. These results were initially somewhat discouraging.

## IV. AN EXAMINATION OF DIGITAL VIDEO SIGNAL CHARACTERISTICS

As one might expect, the RF characteristics of the compressed digital video spectrum are substantially different from those of the conventional VSB analog signal. For the purposes of this discussion, we will examine the characteristics of the 4-level VSB modulation technique employed by Scientific-Atlanta. However, the significant characteristics as described herein may be applied to most other compressed digital video RF modulation systems. Most of the conclusions drawn are easily extended to all systems.

Upon examining the RF spectrum of the digital video, the most striking characteristic is the uniform distribution of the energy as a function of frequency. There are no discrete high level carriers that result in the concentration of the RF energy at particular frequencies. The distribution of the energy is further homogenized by various multiplex techniques that allow the inclusion of several video programs within one 6 MHz slice of spectrum. A final smoothing of the energy is accomplished by the various encryption techniques that allow conditional program access. In the case of the S-A 4-level VSB technique, there is a suppressed pilot carrier at the very low end of the spectrum, but its impact upon the overall energy distribution is negligible.

Another significant difference in the signal characteristics is level. The average level of the signal is reduced with respect to the level of a conventional analog signal. The average level of the power in the channel is 10 dB below the peak envelope power (sync tip power) of a conventional analog channel. There are instantaneous peak powers that are higher than the average power but because of the multiplex and encryption techniques mentioned above, it is virtually impossible to relate the frequency components of these peaks to any characteristics of the original video waveforms.

Finally, a performance advantage is obtained with the compressed digital video that allows it to be more robust with respect to undesirable interference. Excellent bit error rates (BER) may be obtained with average channel power to interference ratios of only 20 dB. This results in a required channel dynamic range of 30 dB (resulting from operation at a level of -10 dB with respect to the peak carrier power of the analog video and the the requirement for a 20 dB signal to interference ratio), as compared with the dynamic range required by conventional analog video of 57 dB.

These characteristics result in a channel energy distribution that very closely resembles Gaussian noise at an average channel power that is -10 dB with respect to a peak power of a conventional analog signal.

## V. DISTORTION TESTING TECHNIQUES

A new distortion measurement technique was developed that allowed the simulation of a mixed system containing both conventional analog channels and compressed digital channels. For the purposes of this test it was proposed that the device under test (DUT) would be loaded with conventional analog carriers from 54 MHz to 550 MHz and loaded with simulated digital channels from 550 MHz to 750 MHz. The device chosen to be tested in this experiment was the Scientific-Atlanta Model number 9504 four-port interdiction unit.

The test set-up for this experiment is shown in figure 1. The conventional analog carriers were supplied from a matrix generator. A conventional distribution equalizer was cascaded with the matrix generator in order simulate the typical tilt encountered in a conventional feeder leg of a distribution plant. The optional video generator, optional modulator, and optional directional coupler will be discussed in detail later. The conventional analog signals are combined with the output of the digital video signal simulator.

For the purposes of this test, the digital video is simulated with band limited gaussian noise. The noise was generated by amplifying thermal noise in a cascade of two high gain indoor distribution amplifiers. The noise must be band limited to the 550 MHz to 750 MHz band in order to match the desired test conditions. It was determined that a very complex and high order band-pass filter would be required to meet the shape factor requirements of limiting the noise to this band.

A simpler approach was identified that utilized a series of low-pass filters that produce a noise spectrum occupying the band from approximately 3 MHz to 100 MHz. This spectrum is then mixed with a 650 MHz LO to produce band limited noise spectrum from 550 MHz to 750 MHz. The SBL-1X type mixer used gave adequate LO rejection at the output to ensure that the LO content would not significantly affect the distortion measurements. Finally a 550 MHz high-pass filter is included to ensure that any inadequacies in the mixer isolation would not degrade the distortion measurements by masking distortion products in "leaked" noise. An added benefit of this technique is the ability to "turn-off" the digital channels by simply removing the LO from the mixer. Comparisons could therefore be easily made between performance with and without the additional loading of the digital video signals.

The resulting total RF spectrum is shown in figure 2. Note that in the simulated digital portion of this spectrum, the average level is -12.5 dB with respect to the peak analog carrier level. This apparent discrepancy with respect to the normal operating conditions described above is due to the resolution bandwidth of the spectrum analyzer. The widest available resolution bandwidth was 3 MHz. Correcting this for 6 MHz and compensating for the equivalent noise bandwidth of the analyzer filter, the resulting observed average power level is -10 dB with respect to the analog carriers, as described above.

Also note the relatively low level of the mixer LO leakage and the narrow gap of noise introduced around 650 MHz by the inability of the indoor distribution amplifiers to amplify the noise all the way to DC. It was decided that these characteristics would cause negligible effects on the measured

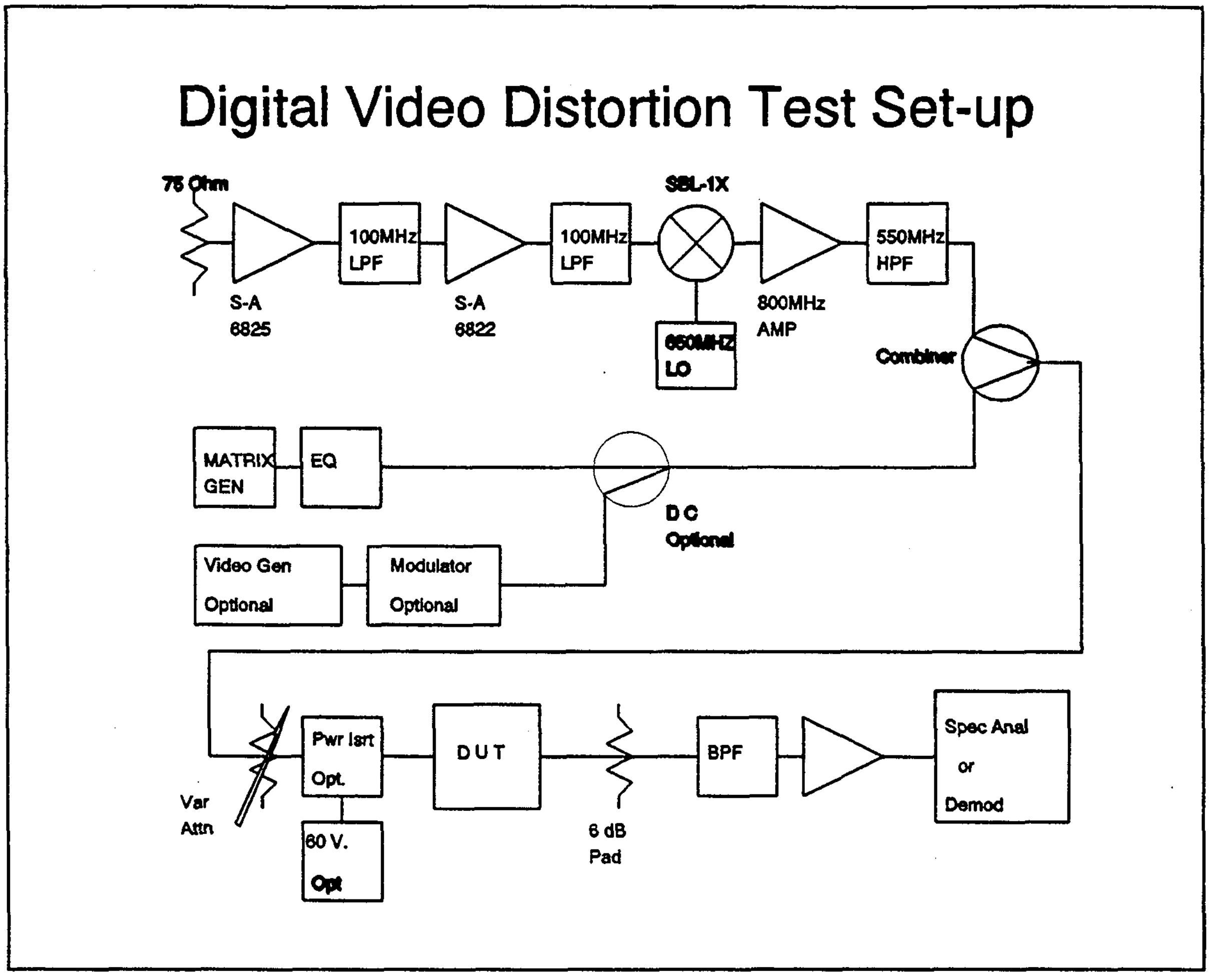


Figure 1

distortion performance.

## VI. MEASURED DISTORTION RESULTS

Conventional distortion techniques were utilized to quantify the distortion performance on the analog channels with and without the loading of the digital video. The distortion results were initially measured on a total of 4 different subscriber ports. Distortion was measured at five different frequencies spaced throughout the 54 MHz to 550 MHz band. CTB, CSO, and XMOD were measured. The data from this initial experiment is given in figure 3.

In general, the additional noise loading of the simulated digital video signals causes a deterioration of the distortion performance of only a few tenths of a dB. In no instance does the addition of the noise degrade the signal by more than 1 dB (note that in a few cases the signal appears to actually get better with the addition of the noise, but these measurements are limited by the measurement system noise floor).

To further confirm the results, an additional 20 ports were measured for CTB at elevated output levels to ensure that system noise would not mask the results. A summary of this data is shown in figure 4. Note again, that the performance, though clearly deteriorated by the higher levels and now clearly above the system noise floor, shows distortion performance impacts of only tenths of a dB when the

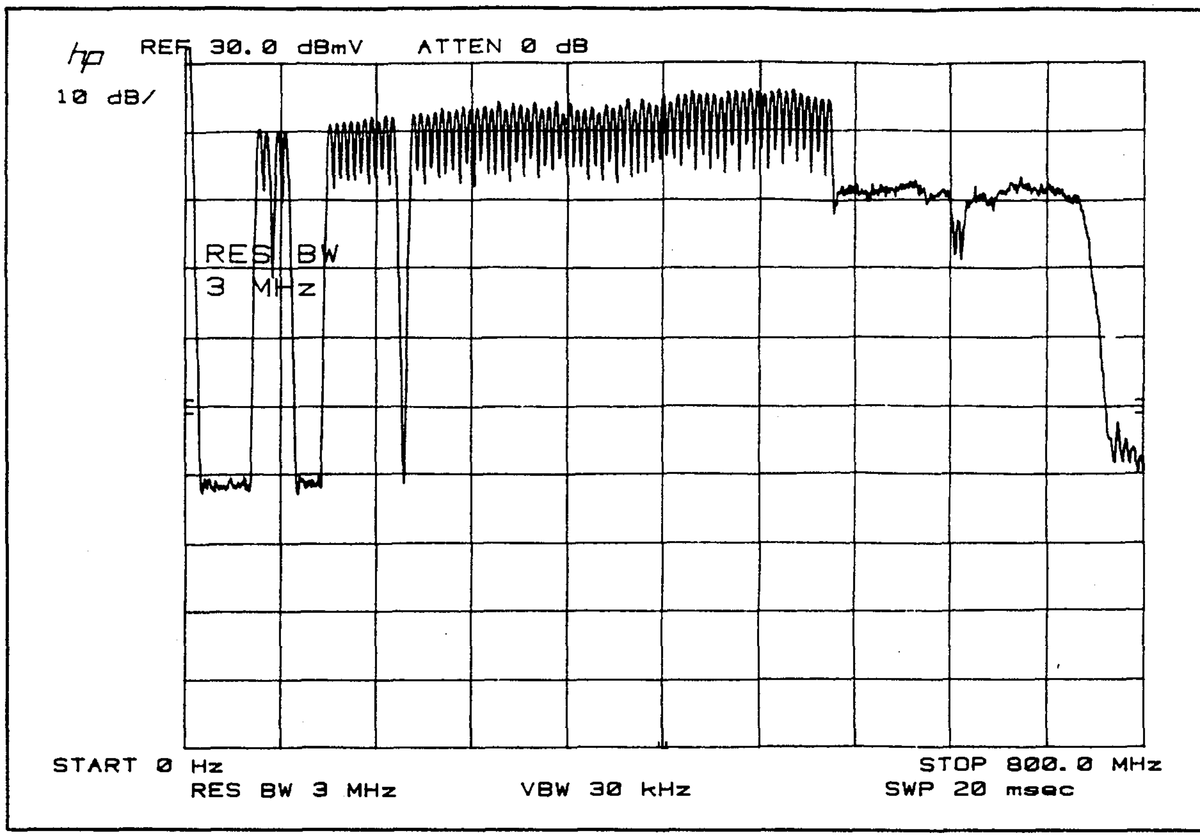


Figure 2

additional noise loading of digital video is added. This encouraging result implies that the additional loading of compressed digital video at frequencies above the conventional analog channels will have a minimal impact on system performance.

## VII. EXPLANATION OF RESULTS

As described above, the energy in a conventional NTSC-M VSB channel is primarily concentrated around the frequencies of discrete RF carriers. Therefore the loading of a DUT with these types of signals results in discrete distortion products or beats that fall at easily predicted discrete frequencies (CTB, DSO, CSO, XMOD). In the case of compressed digital video signals, however, the energy is at a much lower average level and evenly distributed across the entire 6 MHz channel.

The distortion products from such a signal do not lie on discrete frequencies, but rather spread themselves across the entire 6 MHz of an affected channel. Fortunately, as the energy is spread, its impact on perceived picture quality is minimized. It was theorized that the result would appear as a slight degradation of the video signal to noise on the affected channel.

## VIII. VIDEO S/N EFFECTS

The S/N effects were quantified using the optional video generator, optional modulator and optional directional coupler mentioned above. To use these devices, a single carrier of the matrix generator was turned off. The modulator was then used to replace that carrier with modulated video. An S-A agile modulator was used to allow the modulated carrier to be moved throughout the 54 MHz to 550 MHz band, and specifically to the frequencies where the distortion had been measured earlier.

| Distortion Data | | 55.25 MHz | | 199.25 MHz | | 295.25 MHz | | 379.25 MHz | | 499.25 MHz | |
|---|---|---|---|---|---|---|---|---|---|---|---|
| | | w/o dig | w/ dig | w/o dig | w/ dig | w/o dig | w/ dig | w/o dig | w/ dig | w/o dig | w/ dig |
| CTB (dB) | Port 1 | 73.6 | 73.6 | 72.7 | 72.8 | 74.3 | 74.1 | 72.7 | 72.5 | 75.5 | 75.2 |
| | Port 2 | 74.4 | 74.0 | 72.9 | 72.6 | 74.5 | 74.5 | 72.5 | 72.4 | 75.5 | 75.3 |
| | Port 3 | 73.5 | 73.4 | 72.3 | 72.6 | 73.7 | 73.8 | 70.6 | 70.3 | 74.2 | 74.4 |
| | Port 4 | 73.4 | 73.4 | 72.3 | 72.3 | 72.5 | 72.7 | 70.7 | 70.6 | 75.3 | 75.2 |
| CSO- (dB) | Port 1 | 64.8 | 65.0 | 64.0 | 64.0 | 62.9 | 63.3 | 63.5 | 63.5 | 77.1 | 77.3 |
| | Port 2 | 64.0 | 64.0 | 63.0 | 62.8 | 62.8 | 62.8 | 62.8 | 62.9 | 77.5 | 77.6 |
| | Port 3 | 64.5 | 64.6 | 64.2 | 64.0 | 63.8 | 63.8 | 63.5 | 63.4 | 75.7 | 75.5 |
| | Port 4 | 64.0 | 64.6 | 64.0 | 63.8 | 63.6 | 63.2 | 63.2 | 63.1 | 76.5 | 76.4 |
| CSO+ (dB) | Port 1 | 74.8 | 75.0 | 76.8 | 76.6 | 71.7 | 71.5 | 66.7 | 66.6 | 64.2 | 65.0 |
| | Port 2 | 75.5 | 75.6 | 76.7 | 76.9 | 71.5 | 71.3 | 66.2 | 66.3 | 64.3 | 64.2 |
| | Port 3 | 74.6 | 74.8 | 74.7 | 74.6 | 71.1 | 70.9 | 66.8 | 66.8 | 66.7 | 66.6 |
| | Port 4 | 74.6 | 74.7 | 74.7 | 74.7 | 71.1 | 71.1 | 66.1 | 65.9 | 64.0 | 63.8 |
| XMOD (dB) | Port 1 | 65.1 | 65.1 | 66.1 | 66.4 | 66.8 | 66.4 | 65.3 | 65.6 | 66.2 | 68.9 |
| | Port 2 | 67.1 | 66.4 | 64.9 | 65.5 | 66.6 | 67.5 | 65.7 | 65.8 | 67.0 | 67.0 |
| | Port 3 | 69.8 | 69.3 | 65.1 | 67.6 | 67.1 | 67.6 | 65.0 | 65.2 | 66.7 | 66.2 |
| | Port 4 | 68.5 | 68.6 | 66.6 | 66.5 | 66.8 | 66.0 | 64.2 | 64.0 | 65.5 | 67.4 |

Figure 3

COMPOSITE TRIPLE BEAT (dB)
Measured at elevated operating levels

| Analog Video Only | | | | | |
|---|---|---|---|---|---|
| | 55.25 MHz | 199.25 MHz | 295.25 MHz | 379.25 MHz | 499.25 MHz |
| Minimum | 53.5 | 55.0 | 51.1 | 49.8 | 48.0 |
| Mean | 54.5 | 55.2 | 51.7 | 50.8 | 49.0 |
| Maximum | 55.5 | 55.4 | 52.0 | 51.8 | 49.8 |
| Analog And Digital Video | | | | | |
| | 55.25 MHz | 199.25 MHz | 295.25 MHz | 379.25 MHz | 499.25 MHz |
| Minimum | 53.3 | 54.4 | 50.4 | 49.4 | 47.4 |
| Mean | 54.3 | 54.6 | 51.1 | 50.2 | 48.7 |
| Maximum | 55.3 | 54.8 | 51.4 | 51.1 | 49.6 |

Figure 4

Reduction of Unweighted Video S/N
With the Addition Of Digital Video Signals
In the 550 MHz to 750 MHz Band

| Frequency | 55.25 MHz | 199.25 MHz | 295.25 MHz | 379.25 MHz | 499.25 MHz |
|---|---|---|---|---|---|
| Average Change of S/N | 0.3 dB | 0.4 dB | 0.5 dB | 0.5 dB | 0.3 dB |

Figure 5

During this test, the spectrum analyzer was replaced with a video demod and a Tektronix VM-700 video analyzer. The VM-700 was then used to measure the video S/N on that channel. Care was taken to ensure that adequate pre-selection filtering was placed in front of the demod so that additional distortion was not generated by that demod. The results of these measurements are summarized in figure 5.

The Video S/N degradation resulting from the loading of the simulated compressed digital video was typically less than 0.5 dB.

## IX. CONCLUSIONS

It has been analytically shown that band-limited Gaussian noise is a good simulation for compressed digital video. Testing with noise of this type may be used for quantification of distortion performance of broadband devices. This method easily allows the simulation of mixed systems with partial analog video and partial compressed digital video.

Additional study of this subject is required. However, based on the preliminary data presented, the distortion performance impact of system or device loading with compressed digital video should be substantially less than the effect of loading the same amount of spectrum with conventional analog signals. In particular, the Scientific-Atlanta model 9504 4-port interdiction device shows very little distortion deterioration (less than 1 dB) with the additional loading of simulated digital video from 550 MHz to 750 MHz. The deterioration of the video signal to noise ratio resulting from this loading is typically less than 0.5 dB.

## ACKNOWLEDGEMENTS

The author wishes to gratefully acknowledge the invaluable assistance of the following people: Charles Bramhall, John Canning, Scott Cheek, Dean Fredriksen and Steve Webb. This paper would not have been possible without their help.

# A Novel Approach to Improving Picture Quality of Cable Television Converters

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*Abstract*

*The quality of pictures sold to cable subscribers is of paramount importance. However, methods commonly used in cable to provide conditional access may worsen picture distortions caused by the cable plant. One such distortion, which we will refer to as scratch noise, is a result of an interaction between sync suppression scrambling and the automatic gain control of the television receiver.*

*This paper describes an innovative circuit design (patent application filed) to eliminate scratch noise. This circuit design produces a trade-off due to the effects of plant power supply hum on the cable signals. The paper concludes by presenting and evaluating one possible approach for dealing with hum.*

## INTRODUCTION

Cable television signals are prone to picture degradations from a variety of causes, both internal and external to the cable plant.[1] One example of interference resulting from external causes is co-channel interference. Co-channel may result from leakage into the cable plant from a broadcast channel sharing the same frequency as a channel on the cable. Internal to the plant, nonlinearities in amplifiers, for example, can cause intermodulation. Previous investigations have helped in many ways to minimize many of these types of interferences.

To secure cable television signals and prevent unauthorized reception, many cable systems scramble premium channels at the head end. They then restore the signal in the homes of paying subscribers using a de-scrambler. Although designed to be as benign to the de-scrambled signal as possible, these signal processing steps may, themselves, introduce additional types of interference.

## RABBIT EARS AND SCRATCH NOISE

In particular, consider the effects of signal security methods which scramble the signal by suppressing the horizontal synchronization pulses. Such horizontal sync suppression is an integral part of most RF and baseband types of scrambling. Figure 1 shows the three stages of this scrambling and de-scrambling process. These stages are the original signal, the sync suppressed signal, and the final restored signal.

First, notice that the suppression window at the encoder is wider than the expansion window at the decoder. Because of manufacturing tolerances, it is not wise to try to make these two windows exactly the same width. Doing so would result in narrow and unpredictable spikes. Some of these spikes might extend in the direction of the sync tip causing erratic behavior of the television receiver. As a result, it is common to use a fixed overlap. This fixed overlap creates the characteristic sync suppression artifact commonly called rabbit ears.

Now, consider the effect of passing this sync suppressed television signal through the cable plant. Each amplifier through which the signal passes, will add some noise to the signal. This noise adds evenly to all parts of the video signal. However, restoration of the suppressed sync pulses in the decoder effectively amplifies the noise only in the

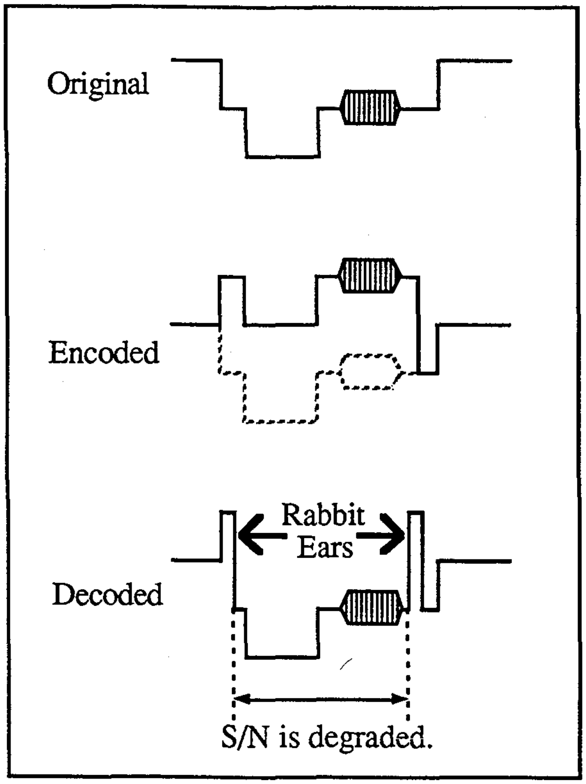


Figure 1: Rabbit Ears Resulting from Sync Suppression

sync portion of the signal. Because this noise affects primarily the sync and blanking area, and not the active video information, it is usually not considered to be a concern. However, careful investigation into the television receiver operation clarifies the problems caused by this noise.

Television receivers use Automatic Gain Control (AGC) circuitry to allow consistent reception of signals with varying strengths. The sync tip is the highest amplitude portion of the video signal. Therefore, additional noise there will affect an AGC level derived from signal amplitude. Figure 2 illustrates this.

Virtually all modern television receivers use a form of AGC called keyed or gated AGC.[2] This type of AGC samples the level of the received signal only once per horizontal line of video. Sampling occurs in the area of the color burst. The sampling is done on the luminance signal after separating the chrominance and luminance components. This location for AGC sampling is ideal for normal broadcast television because the sync and blanking area is the only predictable portion of the composite video signal.

Once sampled, this level determines the appropriate gain for the remainder of the horizontal line. However, for a de-scrambled channel, the noise riding on this area of the blanking pulse provides an incorrect gain for the entire line of video. Figure 3 illustrates the effect of such noise on the luminance signal. The result is a darkened horizontal line on the viewed picture. Each such darkened line looks like a momentary scratch. The rapid changes in these scratches have prompted the term scratch noise.

In addition, some television receivers may drift with age and sample part of the rabbit ear as well. This, too, can result in scratch noise effects.

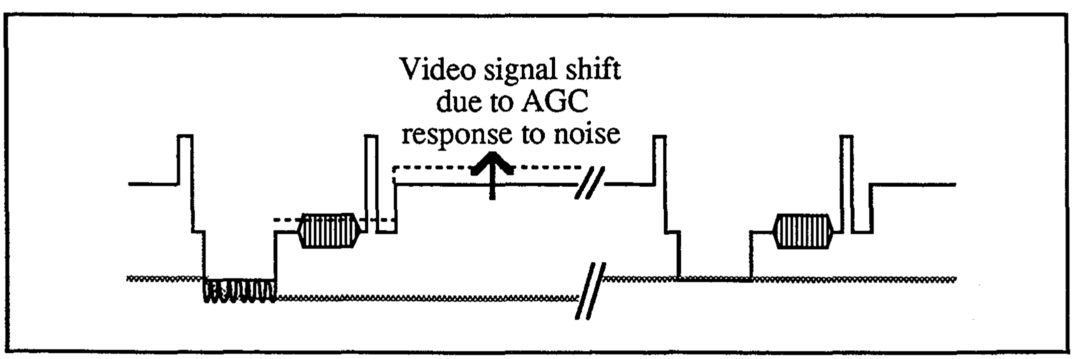


Figure 2: Sync Tip Noise Effects on AGC

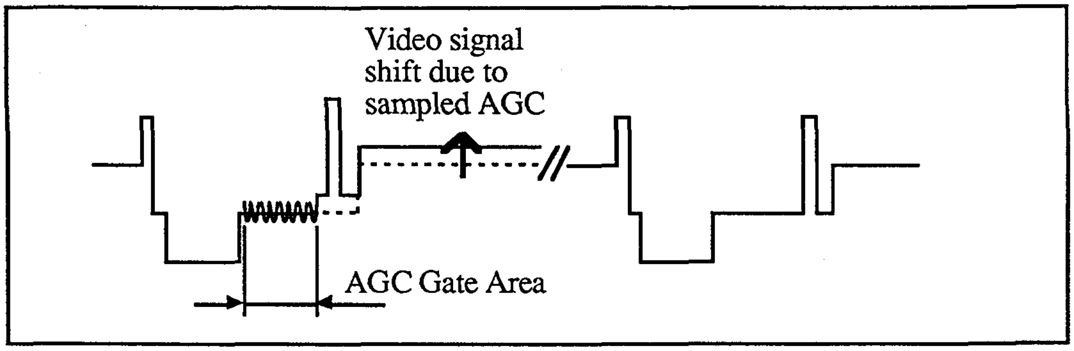


Figure 3: Noise Effects on Keyed AGC

EPQ CIRCUIT DESCRIPTION

Understanding this phenomenon has allowed development of an Enhanced Picture Quality (EPQ) circuit (patent pending). This EPQ circuit reduces scratch noise significantly and eliminates problems resulting from rabbit ears. To accomplish these two functions, the EPQ circuit consists of two sections for:
1) the removal of artifacts (rabbit ears)
and 2) improvement of signal to noise ratio on sync and blanking.

Rabbit ear removal

The first section of the EPQ circuitry uses sample and hold switching to replace rabbit ears with the immediately preceding signal level. Figure 4 shows the basic circuit. The associated timing diagram appears in Figure 5.

During active video, switch 2 is in the up, or bypass, position. Time T1 marks the beginning of the rabbit ear on the front porch of the horizontal sync. At time T1, switch 2 moves to the lower position selecting the pedestal level sampled at the previous horizontal sync's back porch. After the rabbit ear, at time T2, switch 2 moves back to the bypass position.

At the beginning of the back porch, time T3, switch 1 closes. This causes a sample and hold of the level just prior to the next rabbit ear. At time T4, switch 1 opens and switch 2 changes to the lower position. In this way, the pedestal value just sampled replaces the rabbit ear. Finally, at T5, switch 2 moves back to the bypass position. As a result, the sampled pedestal level substitutes for both rabbit ears.

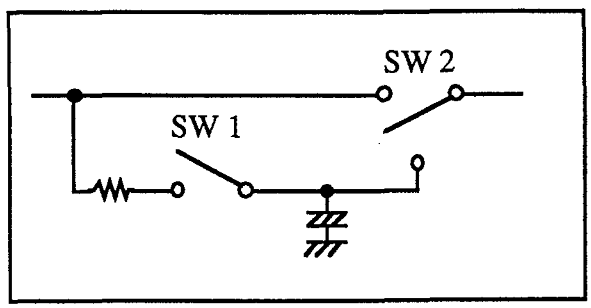


Figure 4: Rabbit Ear Removal Circuit

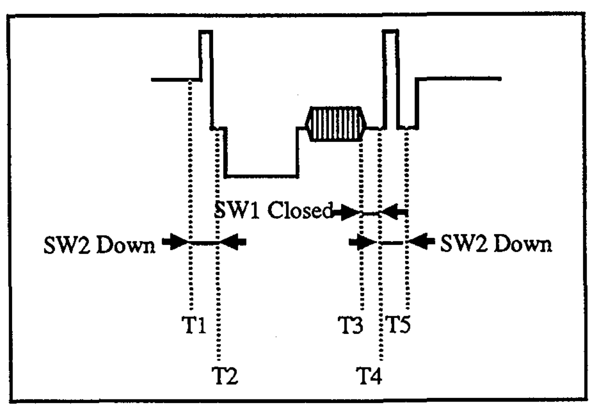


Figure 5: Rabbit Ear Removal Timing

S/N improvement on sync and blanking

The second section of the EPQ circuitry switches in filtering circuits to remove noise on the sync tip and blanking. Figure 6 shows the basic circuit. The associated timing diagram appears in Figure 7.

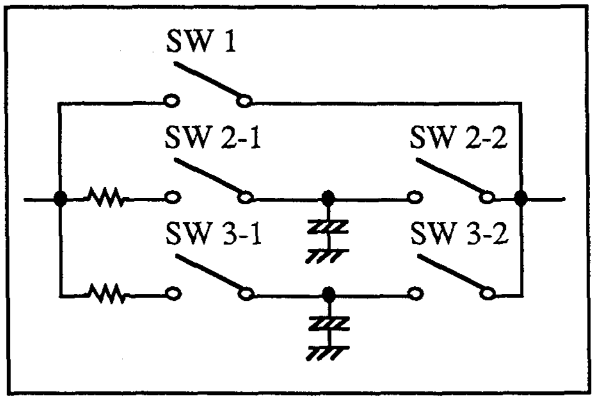


Figure 6: Noise Reduction Circuit

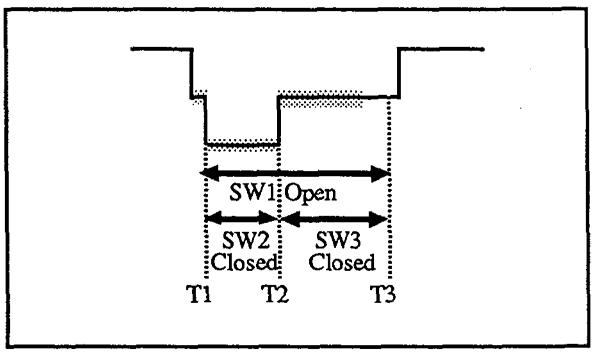


Figure 7: Noise Reduction Circuit Timing

Because of the presence of the color burst, we must position this circuit after the chrominance/luminance (Y/C) separation. The luminance signal then passes through this switch and filter circuit. During active video, switch 1 closes to bypass the signal. At the beginning of the horizontal sync tip, time T1, switch 1 opens and both sections of switch 2 close. This switches in a filter circuit of the appropriate value for the sync tip.

At T2, the transition from sync tip to breeze way, both sections of switch 2 open and both sections of switch 3 close. This switches in a filter circuit of the appropriate value for the breeze way and back porch. Finally, at T3, both sections of switch 3 open and switch 1 closes to pass the active video portions of the signal unfiltered. Thus, appropriate circuitry separately filters the sync tip and breeze way/back porch portions of the signal.

## TRADE-OFFS IN DESIGN

Of course, the television receiver is not the only device involved which uses AGC. The cable converter also has it's own AGC. Selection of a slower than normal AGC further improves the immunity of the converter to fast, random noise.

System Power Supply Hum

However, there is one additional consideration. Cable systems employ long cascades of amplifiers. Power supply failures in any one of these active devices may cause power supply hum modulation to appear on the television signal. Excessive current drain due to overloaded system power supplies can also have a similar result.

If the hum includes sharp transitions like a square wave or spike, it will become more noticeable on the television receiver. In such cases, the NCTA recommends that systems limit hum modulation to 2 percent or less; otherwise 4 percent or less is acceptable with smooth sinusoidal hum.[3] The Federal Communications Commission proposed a limit of 3% in their technical requirements docket of last year.[4] With modern switching power supplies, well maintained systems may achieve hum levels below 0.3 %.

What effect does this hum modulation have on the subscriber's picture ? Normally, television receivers cancel any hum. The keyed AGC in the receiver samples the signal at a frequency much higher than that of the hum. Thus, the AGC rises and falls with the hum detected on the horizontal sync.

With the EPQ circuit, however, the converter removes the power supply hum only from the sync, not from the active video. The keyed AGC in the television receiver never detects the hum remaining on the active video. As a result, the television receiver displays the hum directly in the picture, with no cancellation.

Viewers can tolerate moderate amounts of hum in the viewed picture, while still enjoying the noise reduction benefits of the EPQ circuitry. However, as hum increases, the benefits of EPQ become completely masked by the objectionable hum bars in the picture.

Hum detector and AGC Control

One way around this problem is to use the converter's AGC as a surrogate for the television receiver's. However, the slow AGC chosen to complement the EPQ circuitry now becomes a disadvantage. Ideally, we would like to combine the benefits of EPQ with AGC speeds appropriate to the changing conditions of hum. Figure 8 shows the circuit proposed to do just this.

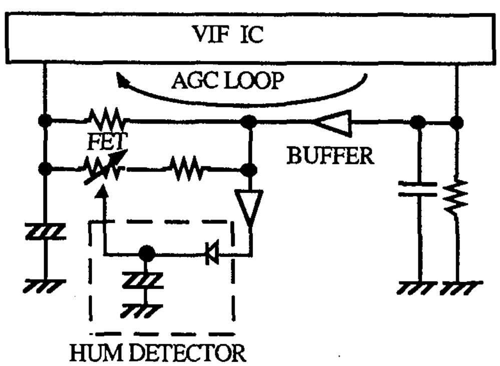


Figure 8: Hum Detection and AGC Speed Control

This circuit senses the amount of hum on the signal. It then adjusts between the slow AGC needed for EPQ, and the fast AGC to deal with the hum distortion. Based upon subjective tests to determine the optimum points, the control points chosen for AGC speed are as follows:

Slowest AGC 0.3% and below
Fastest AGC 1.2% and above

Between these two values, the speed of the AGC increases linearly.

## CONCLUSIONS

Conditional access to cable programming requires a method of denying programming to non-subscribers. Signal scrambling involving sync suppression has become a common means of providing this control. However, the added noise introduced by sync suppression has adverse affects on television receiver AGC. The EPQ circuit provides a method to avoid the resulting scratch noise, while still handling hum modulation on the cable signals.

## ACKNOWLEDGEMENTS

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# A PROGRESS REPORT ON THE WORK OF THE ATSC SPECIALIST GROUP ON INTEROPERABILITY AND CONSUMER PRODUCT INTERFACE

by

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Abstract

*Since August of 1989, a group of dedicated television engineers has been meeting approximately every six weeks to help ensure that the Advanced Television standard chosen by the FCC for terrestrial broadcast will be friendly to Cable TV and other alternate media as well as the American consumer. The work of this group and the results it has achieved in the last 2 1/2 years is described. It is concluded that a single baseband ATV format should be adopted by all television delivery media. It is also concluded that a single standard for encryption of conditional-access programming is technically feasible.*

## INTRODUCTION

In August of 1989 the Advanced Television Systems Committee (ATSC) created a Specialist Group on Interoperability and Consumer Product Interface to study issues relating to interoperability among the various media that will be employed to deliver Advanced Television (ATV) service to U.S. consumers and to study the resulting impact on the interface between consumer products and the various media. An objective of the Specialist Group (known as T3/S2) is to develop a body of technical information relating to the requirements of the alternate media that the ATSC can provide to the FCC Advisory Committee on Advanced Television Service (ACATS) to assist in the choice of an ATV standard for terrestrial broadcast. A further objective is to encourage the alternate media to adopt voluntary standards for transmission of ATV signals that will maximize the interoperability among media and especially between the alternate media and terrestrial broadcast. A final objective is to encourage the adoption of a voluntary interface standard by television receiver manufacturers that will accommodate the needs of the alternate media. The goal is to achieve a set of harmonized standards that are friendly to the various delivery media and the consumer.

The membership of T3/S2 totals approximately 30 and includes a broad cross-section of the television industry with representation from broadcasters, receiver manufacturers, programmers, the telephone industry, Cable TV operators and equipment manufacturers, ATV system proponents, satellite operators and equipment manufacturers, etc. The Group has met over 20 times with a typical attendance of 12 to 15 members at each meeting.

In carrying out its work, T3/S2 has concentrated on issues relating to the delivery of ATV by Cable TV and DBS. A major part of our effort has been focussed on conditional-access systems for ATV. The Group has sought to identify which, if any, aspects of conditional-access systems are suitable for voluntary industry standardization. Working with our sister Group, the ATSC Specialist Group on Digital Services (T3/S3), we developed a list of attributes for scrambling (The term "scrambling" is used here and throughout this paper in the broad contest of rendering the signal unintelligble.) and conditional access to be applied to the proposed ATV systems.

To assist T3/S2 in its work, written inputs were obtained from the various ATV system proponents and other interested parties. We have also had presentations at our meetings by system proponents and others. The Group is now in the process of distilling these many inputs to organize them into a final report by mid-1992. Toward this end we recently developed a strawman list of nearly 30 possible points of agreement relating to interoperability and consumer product interface as a basis to achieve a consensus.

## BACKGROUND

As the U. S. moves towards the adoption of standards for a terrestrial broadcast ATV service, it is important to recognize that ATV services will also be provided by Cable TV and the other alternate media. Since Cable TV and the other media have differing needs as well as differing technical and regulatory constraints, it is important to insure coordination and cooperation among all media in the development of standards so that program material delivered by any one medium can also be easily delivered by all other media and so that consumer receivers can be easily interfaced to all possible media. If this is not done, expensive conversion equipment might be required to exchange programming between media and, worse yet, consumer television receivers might require complex, and potentially user-unfriendly, interface boxes to receive programs from the various alternate media.

This need has been recognized by the FCC and has been addressed in part by the Advisory Committee through the Planning Subcommittee Working Party 4 (PS-WP4) on Alternate Media and the Systems Subcommittee Working Party 4 (SSWP4) on System Standards. Specifically, SSWP4 recognized the importance of Cable TV in delivering terrestrial broadcast signals to consumers and has explicitly stated that any standard(s) adopted for terrestrial broadcast must be capable of being transmitted over Cable TV systems. The HDTV Subcommittee of the NCTA Engineering Committee, PSWP4 and Cable Television Laboratories (Cable Labs) developed a basic test plan to evaluate the performance of proposed ATV systems when transmitted through Cable TV systems and over fiber optic links. Cable Labs converted this basic test plan into a specific series of tests and installed the necessary Cable TV and fiber optic equipment at the Advanced Television Test Center (ATTC) where Cable Labs is now conducting tests of the proposed ATV systems.

In late 1988, PSWP4 developed a strawman proposal for an ATV Multiport receiver interface that would make it possible for ATV receivers to interface to alternate media sources. Subsequently the Electronics Industries Association (EIA) ATV Committee created an ATV Multiport Receiver Subcommittee. This Subcommittee, which is now a part of the EIA R-4 Engineering Committee (designated as the EIA Receiver Interface Subcommittee - R4.1), developed a detailed generic model of an ATV receiver multiport interface.

It is in the context of these various related activities that T3/S2 undertook its work in August of 1989. To insure that there would be neither competition nor unnecessary duplication of effort, T3/S2 established and has maintained liaison with EIA, PSWP4, SSWP4 and the NCTA.

## ACTIVITIES OF T3/S2

Early in its work, T3/S2 realized that the alternate media were free to choose ATV standards totally unrelated to those developed for terrestrial broadcast. We concluded that such a scenario was both unwise and unlikely, and in any event, unless and until some medium chose such a standard, there was little if anything we could do to deal with its interoperability with other media. Recognizing that the various media will employ different modulation methods and may format and condition the signals for transmission differently, we concluded that the requirements for interoperability and consumer product interface would be most easily met if all media were to adopt substantially the same baseband video signal format. We made this statement intentionally vague to allow for the possibility that variations among media will allow exploitation of extra capability by a given medium or fitting within a constraint by another medium without unduly compromising interoperability or complicating the consumer product interface. As an example, VHS tape has less luminance bandwidth than NTSC and S-VHS has more luminance bandwidth.

Therefore T3/S2 has focussed on examining the standards proposed for terrestrial broadcast of ATV, assuming that the terrestrial standard chosen by the FCC will form the basis for the standards employed by the alternate media. We have attempted to evaluate how well each of the proposed terrestrial standards meet the unique needs of Cable TV and the other alternate media with regard to interoperability and consumer friendliness.

### Cable Television

T3/S2 decided initially to concentrate on Cable TV and to characterize the needs of Cable TV in an ATV environment. We agreed that most of the issues of concern to the Cable TV industry would be covered by the test program now being conducted by Cable Labs at the ATTC. The major issue that is not covered by the Cable TV test plan is scrambling and the need for data transmission to control conditional access to scrambled programming. Since the data transmission for controlling access falls within the charter of the ATSC Specialist Group on Digital Services (T3/S3), starting with our January 18, 1990 meeting, T3/S2 has met jointly with T3/S3. At that meeting we agreed to develop an attributes list for scrambling and conditional access to be applied to the proposed ATV systems.

At our March 7, 1990 meeting, we reviewed the criteria for scrambling and conditional access used by the Direct Broadcast Satellite Association (DBSA) in 1986 as well as a list of attributes generated by Graham S. Stubbs, the Chairman of T3/S3. We reached a consensus on four basic desireable attributes for a scrambling and conditional-access system.

- The images displayed on receivers not authorized to receive the scrambled programming must be unrecognizable.

- It must not be possible to recover the image by inspecting the transmitted signal and performing any reasonable real-time processing on the information contained therein.

- The details of how the system operates must be general public knowledge.

- The security of the system is entirely contained in the delivery and processing of the key.

During subsequent joint meetings of T3/S2 and T3/S3, this list was modified and expanded and it was augmented by a second list of desireable features and other considerations. A first draft was mailed by T3/S3 to the ATV proponents and other interested parties in August of 1990. The comments received were incorporated into a second draft which was mailed by T3/S3 to a wide cross-section of the television industry in December of 1990. The final result is ATSC document T3/180 "ATV Encryption System Characteristics" dated 16 May 1991 (see Appendix), which has been widely distributed to the ATV proponents and others for use as a guideline in developing and evaluating ATV systems.

During 1990 T3/S2 solicited and received information from the ATV system proponents concerning how they planned to meet the original four basic desireable attributes for a scrambling and conditional-access system. This was at a point in time when most of the proposed systems were based on analog transmission. By the end of 1990, that had all changed. Four of the five full high-definition ATV simulcast systems are now based on digital transmission, and it is clear that they can all meet the six general requirements for System Security listed under Desireable Attributes in the Appendix by encrypting the transmitted digital signal bit-by-bit and providing an appropriate replaceable security module.

The attention of T3/S2 then turned to the question of which, if any, aspects of such a conditional-access system might be subject to voluntary industry standardization. To address this issue, during 1991 we invited ATV system proponents and others to present their views at our meetings. Those presentations and the ensuing discussion have led us to the following tentative conclusions, assuming an all-digital simulcast ATV system:

- For conditional-access programming, the source-coded digital video, audio and data will be scrambled by encrypting the transmitted digital data stream using a DES-like algorithm.

- The process of scrambling and descrambling the transmitted digital data stream will be implemented by performing a mathematical operation on the source-coded digital signal and a pseudo-random number generated from a seed provided by the encryption algorithm.

- With respect to the equipment in the consumer's home, the security of the system shall reside entirely in the hardware/firmware used to implement the encryption algorithm, i.e., process the delivered keys to produce the seed for the pseudo-random number generator.

- In order to provide for recovery in the event that the security module, i.e., the hardware/firmware used to implement the encryption algorithm, is cloned by a pirate, this portion of the in-home system must be easily replaceable.

- It is highly desireable that the security module be replaceable by the consumer himself, not requiring a service call.

Given the above five tentative conclusions, all of the system security attributes listed in the Appendix can be met, and if ATV receivers have a multiport interface which passes the source-coded digital signal, a decoder box can be connected to the interface to perform the descrambling operation. Different service providers can use different encryption algorithms if desired. However, this may lead to multiple boxes connected to the multiport interface with each box having its own replaceable security module.

T3/S2 also concluded that it is possible to design and standardize a conditional-access system that meets all of the requirements for the desireable attributes listed in the Appendix. The hardware for such a standardized conditional-access system could be contained in a back-of-the-set box connected to the ATV receiver multiport interface, or alternatively, except for the replaceable security module, it could be built into the ATV receiver. If the conditional-access hardware is built into the ATV receiver, a separate interface (not the multiport interface) will be required to accommodate the replaceable security module and still another interface will be needed to provide signals to an upstream data link to order pay-per-view programs. The EIA R4.1 Receiver Interface Subcommittee is studying both of these interfaces as well as the multiport interface.

Although a standardized conditional-access system built into the ATV receiver, rather than external, is technically feasible and would provide the most cost-effective and user-friendly system for both service providers and their customers, the design of such a system will require a major cooperative effort by all segments of the television industry. However, since the various U.S. television industry segments traditionally compete with one another and each has vested short-term business interests that argue against a single conditional-access standard, T3/S2 has concluded that agreement on a single voluntary standard for conditional access is not likely, despite its technical feasibility and obvious long-range advantages of cost and convenience for both service providers and the American public.

DBS

Beginning at our March 7, 1990 meeting, T3/S2 began to study satellite (DBS) delivery of ATV programming to consumers. We understood that the Satellite Broadcasting and Communications Association (SBCA) was planning to distribute a questionnaire to present and prospective satellite programmers concerning ATV-related issues. We established a liaison with SBCA to initiate a cooperative effort and avoid duplication of work. It quickly became clear that the conditional-access issues for DBS and Cable TV were substantially the same, except possibly for data capacity requirements to address a larger universe of subscribers from a single transmission point in the case of DBS.

The major difference between DBS and Cable TV is the different modulation method likely to be used by DBS. T3/S2 concluded that the proponents of ATV systems should be asked how they proposed to deliver ATV signals over satellites, and that it would be desireable to conduct comparative satellite delivery tests of the proposed ATV systems. PSWP4 had developed a skeleton test plan for satellite transmission and T3/S2 volunteered to convert this into a specific test plan, if the SBCA could put a test program in place. During the period from December, 1990 to July, 1991, there was an ongoing dialog between T3/S2, SBCA and ATTC. In July, 1991 SBCA created a Working Group on Satellite Testing of ATV Systems, which first met on July 17, 1991. By early December of 1991, the Working Group developed a Conceptual Test Plan for Satellite Delivery of ATV. This plan differed significantly from the earlier PSWP4 plan because of the change from analog systems to all-digital systems.

Later in December of 1991, the SBCA Working Group initiated discussions with PSWP4 and in January, 1992, the Working Group became a sub-group of PSWP4. The present plan is to conduct a theoretical evaluation of the proposed ATV systems based on information supplied by the proponents about how they propose to deliver ATV by satellite.

Consumer Product Interface

With respect to the television receiver interface issues, T3/S2 believed from the outset that the EIA Multiport Subcommittee (now R4.1) had this well in hand. During 1989 and 1990, we reviewed drafts of their reports at various stages and provided our comments. We held a joint meeting with them in April of 1990 and the Chairman of T3/S2 has been attending their meetings as an active participant since early 1991. Initially, R4.1 developed a generic model of an ATV receiver multiport interface. Recently the Committee has focussed on being more specific. Although precise details of the interface must await the choice of a terrestrial broadcast standard by the FCC, R4.1 has identified four possible interfaces: An analog interface which probably will be luminance and two color difference signals; a digital interface which will most likely be source-coded compressed digital video, audio and data; a conditional-access interface to accommodate a replaceable security module; and a control interface to pass receiver status to other devices, receive status from other devices and to allow user inputs (either directly to the TV or via remote control) to be passed to other devices, e.g., to order pay-per-view programs via a modem or other up-stream communications link.

In early February of 1992 the EIA R4.1 Committee distributed a questionnaire concerning these four possible interfaces to a wide cross-section of the television industry. The responses to the questionnaire will be analyzed this spring and, along with inputs from T3/S2, T3/S3, PSWP4 and other industry groups, will help to define appropriate interface ports that can lead to EIA recommendations for voluntary standards for television receiver manufacturers.

## CONCLUSIONS

Since its inception in August of 1989, T3/S2 has been working to achieve harmonization of standards among the various media that will deliver ATV to the American consumer. We have focussed on Cable TV and DBS and it appears at this point that both Cable TV and DBS will adopt standards that embody substantially the same baseband format chosen for terrestrial broadcast. It also appears that other media (telephone companies, pre-recorded video, etc.) will do likewise. Working with our sister group T3/S3 we have, with the help of many in the television industry, developed a list of ATV Encryption System Characteristics to serve as guidelines for the industry. We have also, through SBCA and PSWP4, initiated an evaluation of the proposed ATV systems for DBS transmission. With the EIA R4.1 Committee, the definition of TV receiver interfaces that will provide user-friendliness in an alternate media environment is proceeding apace. Finally, T3/S2 has achieved agreement that a single standard for conditional-access programming is technically feasible and can, with an interface for a replaceable security module, be implemented within the TV receiver. The challenge to achieve a single conditional-access standard is not technical; it is a business challenge that will require cooperative efforts within the industry to achieve win-win rather than win-lose scenarios.

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## APPENDIX
ATSC DOCUMENT T3/180
Dated 16 May 1991

**ATV Encryption System Characteristics**

### I. DESIRABLE ATTRIBUTES

1. System Security

- General -- The system security design should provide:
  - System security entirely contained in the delivery and processing of encryption keys; it should permit recovery from any security compromise.
  - Secure operation even when threatened by a party with total system information; i.e., the system must be secure even if all details of how the system operates should become public.
  - To the extent that withstanding piracy threats over a long period of time may require the periodic exchanging of at least some key components of consumer decoders, this capability for changes should be provided in a way so as to minimize the costs of such exchanges.

    - Images and audio of transmitted signals should be unrecognizable when received and displayed by receivers not authorized to receive the scrambled programming.
    - Non-feasibility of decryption by inspection of the encoded signal and performance of any reasonable processing on the encoded information.
    - Secure transmission of ancillary services.

- Physical -- Physical security measures should:
    - Preclude a typical homeowner with household tools from defeating any security function.
    - Preclude a commercial enterprise from making cloned units or modifying legitimate units such that security measures are defeated. The cost of cloned devices should be much greater than the value of the deferred service cost.

2. <u>Signal Quality</u>

- Under perfect signal conditions:
    - There should be no perceptible decoding artifacts in video.
    - There should be no perceptible decoding artifacts in audio.

3. <u>Signal Robustness</u>

- The effect of noise or signal degradation on reception and decoding of encrypted signals should be no greater than the effect of such signal-path imperfections on non-encrypted signals.

4. <u>Universality</u>

- Applicability, where possible, to alternate media intended for program delivery to the consumer.
    - Terrestrial broadcast.
    - Wired networks (cable, fiber, etc.)
    - Satellite broadcast.
    - Pre-recorded.
- It is desirable that an encrypted format be useable in all of the alternate media without completely decoding and re-encoding as the signal passes from one medium to another.

## II. DESIRABLE FEATURES AND OTHER CONSIDERATIONS

1. <u>Addressability and Tiering</u>

- Pre-authorized tiers.
    - Number available.

    - Independent control of audio, data, and text.
- Addressability
    - Size of the universe (per operator/system).
    - Data space required per addressable consumer terminal.
    - Universal key and fail-safe modes.
    - Interface to automatic ordering systems.

2. Subscriber and Pay-Per-View

- Lock-out functions.
    - By rating (parental control).
    - By channel or time of day.
    - By pay-per-view.
- Program tracking.
- Transactions per month.
- Impulse pay-per view: increment/decrement function.
- Return loop provision.
- Multiple operator pay-per-view.
- Consumer-friendly operation.

3. Multiple Operator Use and Control

- Feasibility of simultaneous multiple operator use.
- When encrypted signals are delivered to cable headends, provision for local cable system intervention to manage conditional-access control and key distribution within the cable system.
- Separate billing systems for multiple operator use.
- System capacity: how many simultaneous billing systems.

4. Ancillary Services

- Multiple sound channels.
- Teletext/captioning capability.
- Allocation flexibility for ancillary services.

# Advances in Videodisc Technology

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*Abstract*

*Cable systems are becoming increasingly active in the local origination of both programming and commercials. Meanwhile, the technology available for video storage and playback is developing rapidly. This paper describes some of these advances, especially erasable video discs. A description of replicated and recordable video discs provides a baseline for comparison of the new technology of erasability. An understanding of all of these approaches should help the cable system operator to select the best technology for his applications.*

## INTRODUCTION

Since the early days of cable television, cable operators have concentrated on the distribution of programming that was originated elsewhere. Recognizing the economic potential of both commercial insertion and pay per view, cable operators are now considering new methods for sourcing programming right in their systems. This local recording and playback requires that the cable operator understand new technologies and apply them wisely.

Introduced over ten years ago, replicated video discs remain a high quality, reliable source of playback video. However, video discs are now available in other formats, the newest of which is the erasable video disc. To best understand the operation and advantages of this new format, it is helpful to review the earlier approaches.

## REPLICATED VIDEO DISCS

Audio and video information on the disc takes the form of tiny "pits" measuring 0.4 micron wide and 0.1 micron deep. The lengths of these pits represent the analog video information, and the associated audio. Rows of these pits form a single track spiraling from the inside of the disc to the outside. The distance between adjacent tracks is approximately 1.6 microns.

### Formats

There are two common formats for replicated video discs. They are Constant Angular Velocity (CAV) and Constant Linear Velocity (CLV). The CAV format of a 12 inch disc has a storage capacity of 54,000 frames of video. At a constant rotational speed of 1800 RPM, each revolution of the disc corresponds to one frame of video. The playtime of a CAV disc is 30 minutes per side. Refer to Figure 1 for a detailed illustration of the CAV format.

Generally referred to as standard play, some of the beneficial capabilities of the CAV mode are freeze frame, slow motion, and step frame.

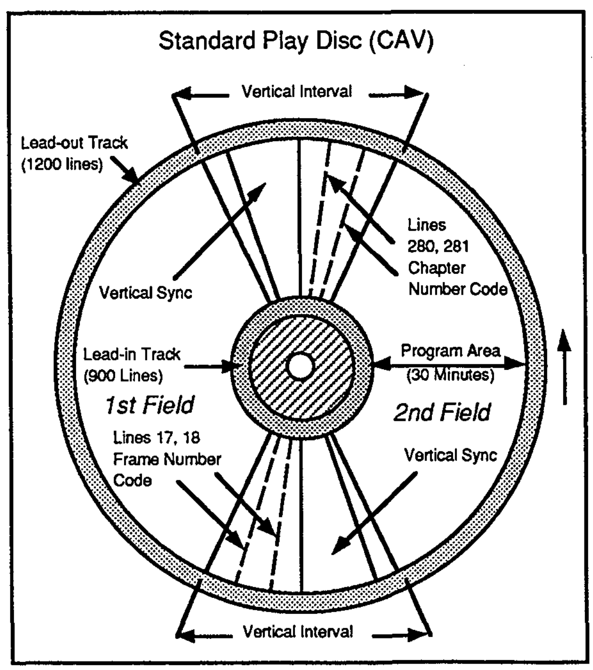


Figure 1: Standard play (CAV) format

When recording in the CLV format (extended play), one frame of video occupies the innermost track revolution. This linear distance is constant for all video frames on the disc. As a result, three frames of video occupy the outermost track revolution. Refer to Figure 2 for the CLV format. The rotational speed of the player varies from 1800 RPM at the center to 600 RPM at the outside edge. This compensates for the amount of information recorded per revolution. As a result, this format stores 60 minutes of play time per side of the disc. Without the use of additional electronics, such as frame store memories, the CLV format cannot do freeze frame, slow motion, or still step.

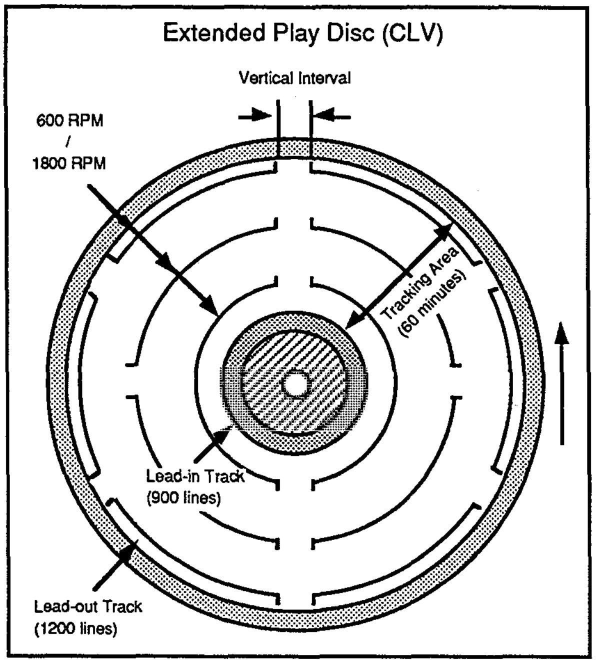


Figure 2: Extended play (CLV) format

Playback Process

The surface of a video disc contains billions of tiny pits of uniform width and depth. The length of each pit, and spacing between each pit, varies according to the modulated video and audio signals represented on the disc. As a laser beam shines on the surface of a disc, 80% of the incident light normally reflects into a detection lens. When the beam passes over a pit, however, only about 20% of the incident light reflects into the detection lens. These detected changes in reflected light modulate the photo diode current of the pickup, regenerating video and audio signals from the playback circuits.

Finished discs have two sides, of course. Manufacturing of each side, however, occurs independently. Only the final stage of production adhesively bonds the two sides together. The clear polymethyl-methacrylate (PMMA) protective coating used in the molding process is susceptible to shrinking and warping in humid conditions. Therefore, the process of bonding two discs together is necessary to insure that the video disc will not warp, causing the disc's vertical acceleration to change during playback.

Generation of Master

There are various methods of mastering and reproducing video discs. A cross sectional diagram of a replicated disc is shown in Figure 3.

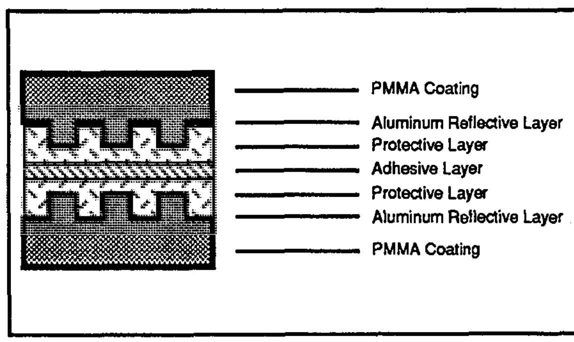


Figure 3: Cross section of the replicated video disc

Because it can replicate video discs in mass quantities, the photo resistive replication technique is used on most discs on the market today. This process, shown in Figure 4, starts with cleaning and polishing a piece of glass to a flat smooth finish. Application of a photo resistive (PR) material forms a thin layer on the glass. The thickness of the PR material is crucial in determining the C/N of the signals during playback. For optimum C/N performance, the material thickness needs to be 1/4 wavelength of the laser used to read the disc.

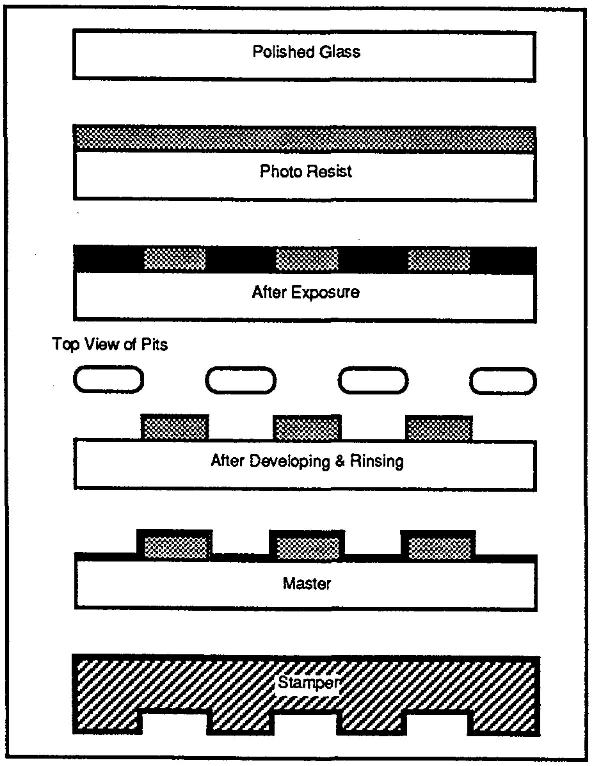


Figure 4: Production of stamper for replicated disc

## Recording Process

The recording process begins by using an FM modulated composite audio/video signal to modulate the recording laser beam. The laser beam modulations soften the exposed photo-sensitive surface areas of the master disc. A chemical developer solution washes the disc to remove the softened photo-sensitive material. This step actually forms the pits. In a vacuum deposition chamber, a vaporized finish of nickel or silver covers the glass master, making the surface conductive.

In an electro-forming bath, a thick coating of electroplated nickel covers the surface of the master, forming the stamper. After additional steps to protectively seal, cut, and punch the stamper, it is ready for mounting in an injection molding machine, and the start of the mass production process.

## Mass Production

Figure 5 shows the steps for mass production. The first step consists of an injection molding or hot press operation. A hot plastic compound (PMMA) enters the mold cavity under high pressure. Once the plastic solidifies and cools, the disc is ready for removal and metalization with a layer of aluminum. Finally, the addition of a layer of protective overcoating to the aluminum completes one side of the disc.

The adhesive bonding of the protective layers of two sides of a disc forms a complete replicated disc. A quality assurance inspector performs a final check on all discs to insure they meet all standards.

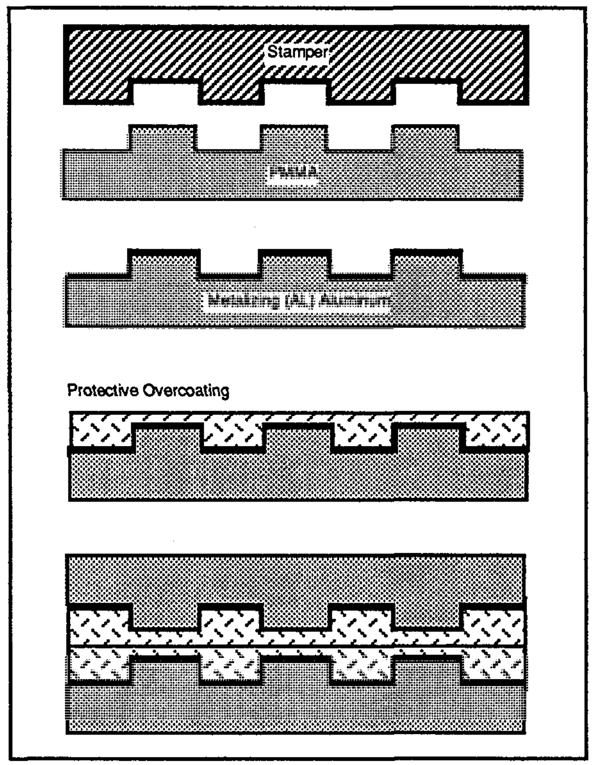


Figure 5: Mass production process for replicated disc

## REAL TIME VIDEO DISC REPLICATION

The real time recording process uses a Direct Read After Write (DRAW) optical system to record onto a blank video disc. Immediately after recording, a disc is ready for playback on any standard laser disc player and produces the same image quality as a mass produced video disc. Real time disc replication is a relatively fast and inexpensive alternative to replication of single copy and low volume disc orders.

The real time videodisc recording system records all source material onto a blank disc in one process. Blank discs are available in either plastic (PMMA) or glass substrates. The recording medium consists of a dye-polymer/metal bi-layer. The dye-polymer layer is thermally reactive to the exposure of laser light. Upon contact of laser light, the surface reacts by converting solids to gas-forming pits with no debris.

## ERASABLE RE-WRITEABLE VIDEO DISC

The most recent entry into the video market is the magneto-optical disc, an erasable / re-writeable technology. As the term magneto-optical implies, magnetic and optical principles combine to achieve recordability.

One of the major benefits of this new technology is that it offers the user the ability to erase and rewrite over one million times to a single disc, with no degradation to the re-writeable media. Other advantages associated with magneto-optical technology are high quality video, frame by frame recording capability, and independent digital audio recording.

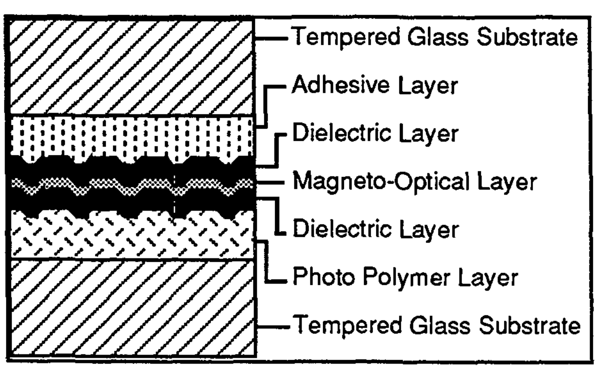


Figure 6: Magneto-optical disc composition

### Disc Composition

Layers of magneto-optical, dielectric, adhesive, and photo polymer materials make up the disc. The addition of a chemically strengthened glass substrate provides durability against warping and bending. Figure 6 illustrates a cross section of the disc.

Use of a constant angular velocity (CAV) disc format for the magneto-optical technology fixes the speed of rotation at 1800 RPM. Each rotation of the disc contains one frame of video. A thirty-two minute disc capacity, then, equates to 57,600 individual frames of video. The CAV format permits frame by frame recording, as well as slow motion, freeze frame and random search features.

All discs are pre-grooved with precisely formed microscopic "valleys" in the surface. This allows the disc recorder to record and play information accurately in a spiral track.

### Features

The constant rotation speed also allows for the use of two separate heads. The second head is one half rotation away from the first. The first head can erase or play. The second can record or play. Together, the two heads make possible seamless playback cuts from one area of the disc to another. Recording can occur at the same time as erasing on a used disc. The two heads can even copy from one portion of the disc to a, previously erased, section of the disc.

The player and disc together provide several approaches to accurate timing. First, each disc contains permanently pre-etched frame numbers. The drive uses these frame numbers to derive elapsed time for precise time searches to any section of the disc. Secondly, a portion of the VBI may store the SMPTE time code. Playback restores the original time code, whether linear or VITC.

### Component Video

Video is recorded on the disc in an analog time compressed component form. Refer to Figure 7 for details on the recording process.

process. The recorder first separates luminance and color components of the video. It then time compresses these and separately records them within each line of video.

Recording and playback in component form eliminates cross color and cross luminance problems. These normally result from interference between luminance information and color subcarrier frequency. The component approach also avoids chroma dot crawl and 2-field jitter problems associated with phase relationship between color subcarriers. As a result, recording in the time compressed component form produces image quality equivalent to that of professional broadcast video.

PCM Audio

The recorder converts two channels of analog audio to a PCM format. It then time base compresses these audio channels and records them within a portion of the vertical blanking interval, independent of the video signal. Refer to Figure 7.

During the playback mode, the reverse process occurs. The quality of this recorded audio is equivalent to Hi-8 digital audio, with a dynamic range of 85 dB at a bandwidth of 15 kHz.

Comparison to Replicated Discs

The technology associated with magneto-optical is not compatible with, nor similar to, replicated discs, other than disc size and format (CAV). The MO disc contains billions of individual segments, where each can be magnetically polarized.

Playback Process

Even the operation of the laser pickup differs in the MO approach. When reading information from a MO disc, all laser light equally reflects off the surface of the disc. As shown in figure 8, playback circuits detect the spatial coherence of the reflected light between segments, producing an FM signal. When demodulated and decompressed, this FM signal reveals the video signal in its original component form.

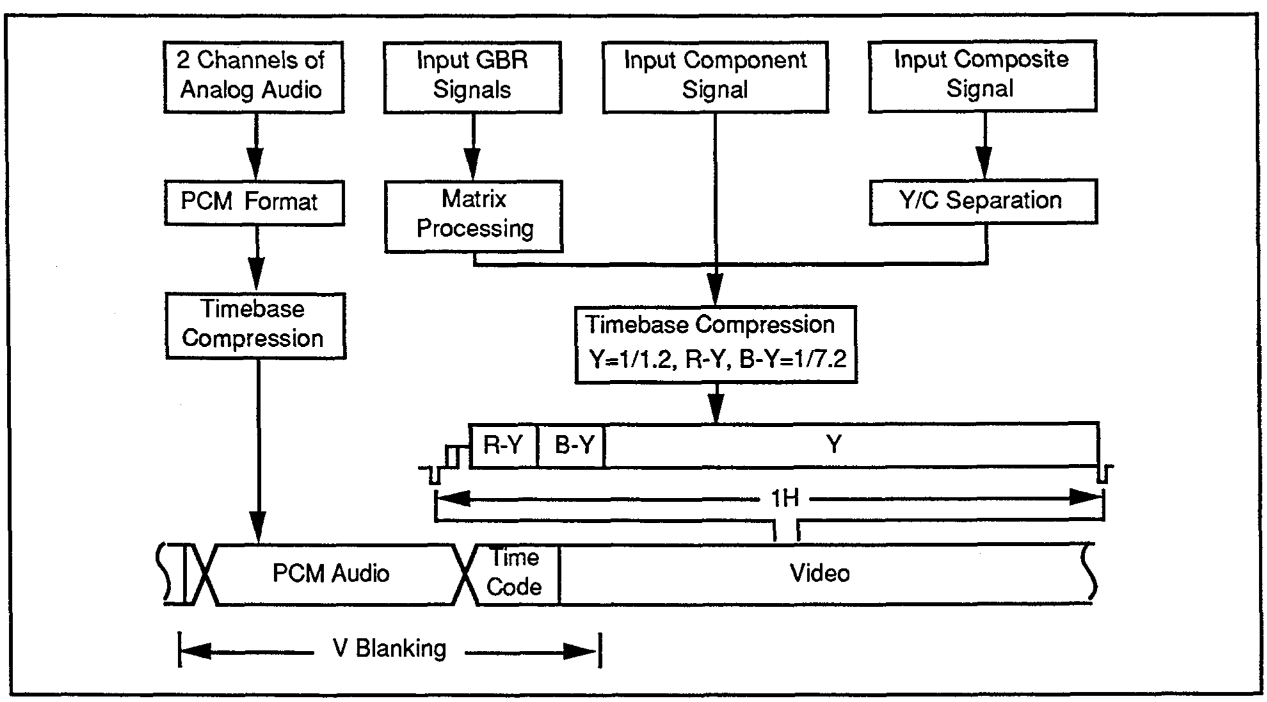


Figure 7: Magneto-optical disc recording process

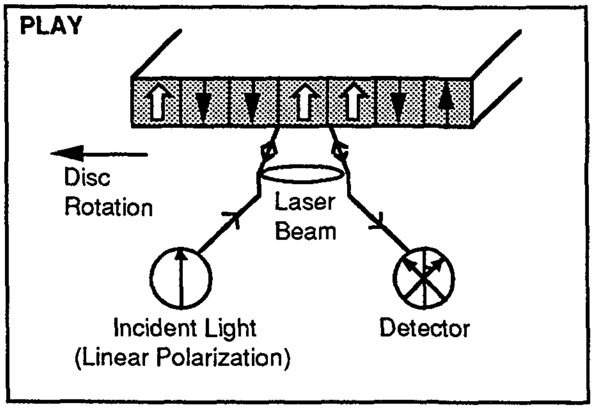


Figure 8: Playback of magneto-optical disc

Recording Process

Figure 9 shows the record method. While heating a segment of the disc with a laser beam, application of a magnetic field changes the orientation of the north and south poles of that segment. As the segment cools, the changed orientation remains, permanently storing the information.

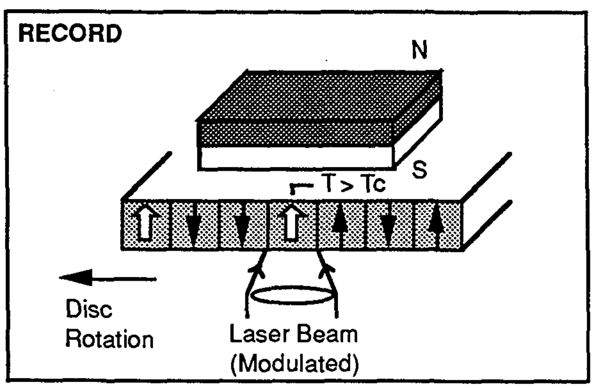


Figure 9: Recording of magneto-optical disc

Erasure Process

Laser and magnetic fields again combine to erase a segment as Figure 10 illustrates. As a laser beam heats the segment, the applied magnetic field reorients the polarity of the segment to its north pole position. On blank or erased discs, all segments have the same, north pole, polarization.

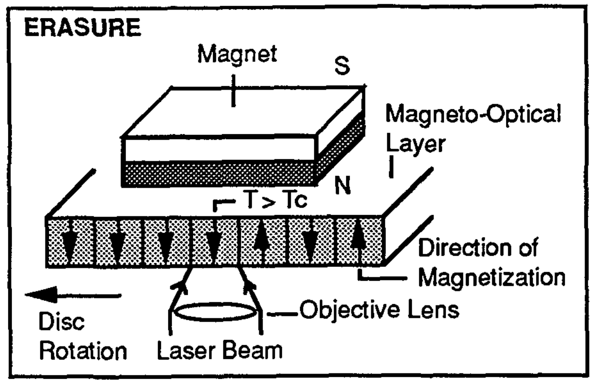


Figure 10: Erasing magneto-optical disc

This recording approach allows one million erase/rewrite operations with no degradation to the media. Thus, the use of laser optics and magnetic polarization can provide significant video recording and playback capabilities.

## CONCLUSIONS

The storage and playback of video continues to grow in importance as commercial insertion and local origination spread. Advances in this area now offer significant alternatives to cable operators. The erasable, re-writeable video disc using magneto-optical technology provides many benefits which deserve consideration.

## ACKNOWLEDGEMENTS

The authors wish to acknowledge the varied contributions of, among others, Andy Parsons, Mike Johns, and John C. Mitchell

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# An MPEG Standard Based Video Compression System and Applications

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## ABSTRACT

An MPEG based video compression system for delivering multiple channels of digitally compressed video over a single satellite transponder or a single cable channel is described. The video compression scheme is MPEG compliant and is based on the Discrete Cosine Transform and motion estimation followed by Huffman coding, and is described in detail. The system is capable of handling both interlaced video as well as progressive sources such as movie material.

## INTRODUCTION

Recent years have seen significant developments in the area of digital compression as applied to video. These compression techniques have been applied to a variety of media including storage devices, satellite and terrestrial broadcast channels, and in cable and telecommunication networks. There has also been significant activity to establish image and video compression standards. The JPEG (Joint Photographic Experts Group) standard for compression of still images is currently in the final stages of standarization[1]. MPEG (Moving Picture Experts Group) is another body whose activities occur within the framework of the ISO (International Organization for Standardization)[2,3]. MPEG examines the issues involved not just in video compression, but also that of compressing associated audio and the system issues of audio-video synchronization. Differing industries such as the cable and telecommunications industry, the consumer electronics industry and the computer industry are looking at using the same technology in their applications. The use of a common standard will go a long way towards resolving the problem of interoperability of these systems. Furthermore, large volume production of standards based IC's can lead to substantial cost reductions, thus enabling even wider applications.

In this paper, we will give an overview of a video compression system based on the evolving MPEG (Moving Pictures Expert Group) standard, and describe its application to a cable network to provide multiple channels of compressed video. In recent months the cable industry has been examining applying digital compression techniques to satellite and cable, to increase the number of programs that can be carried over a single cable channel or a satellite transponder. The block diagram of such a system is shown in Figure 1. A program encoder compresses the video and audio associated with a particular program and multiplexes the output with the auxiliary data after separately encrypting them. Outputs of multiple program encoders are then multiplexed and transmitted over a single channel, after appropriate forward error protection and digital modulation. In an SCPC mode, the output of a single program encoder would be modulated on to one carrier. A cable head-end receives the compressed video, either over a satellite link as shown in Figure 1, or over a fiber-optic link. In either case, the received signal is demodulated, and then re-modulated for transmission over cable to the consumer's home.

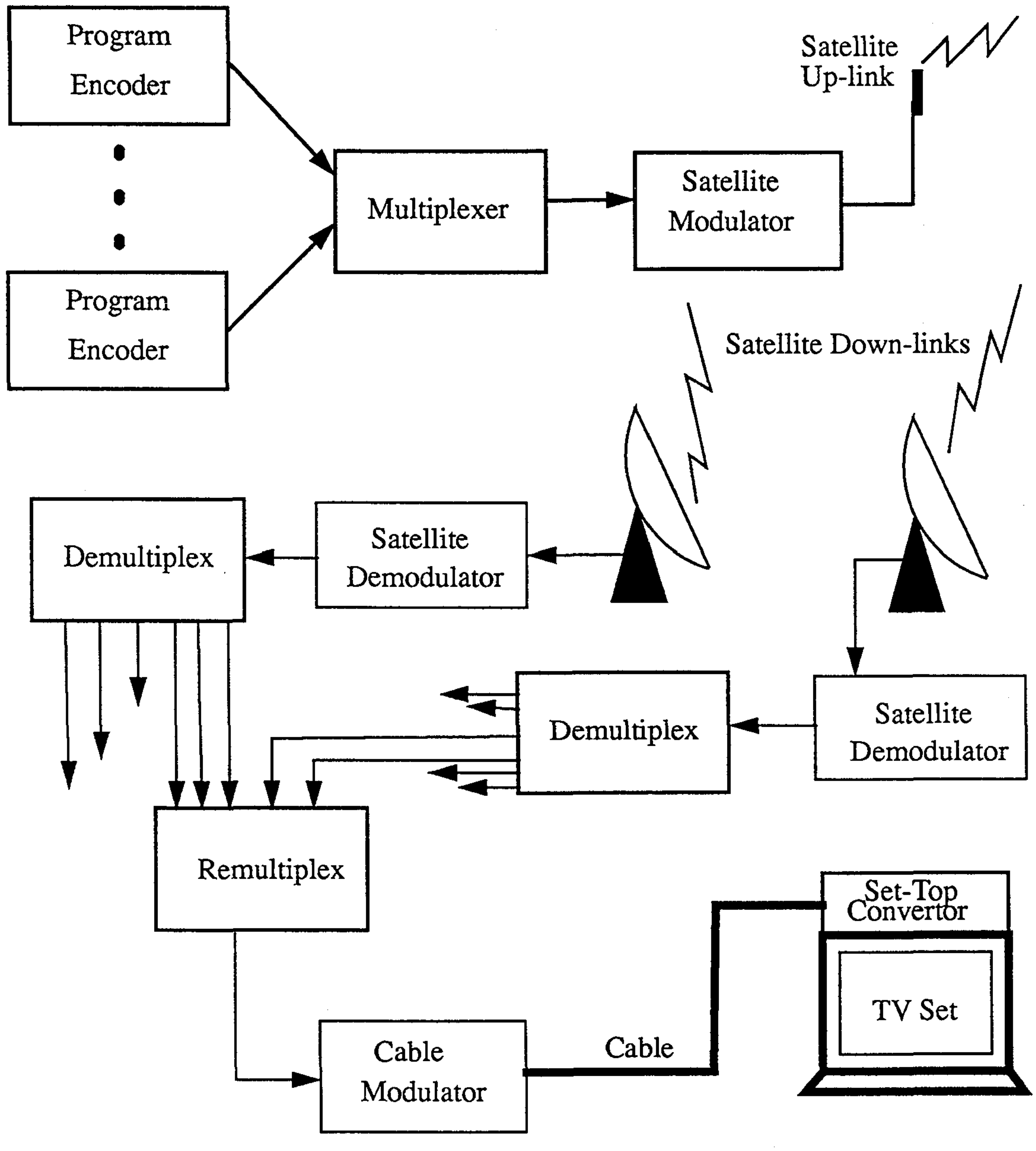


**Figure1: Transmission link from satellite uplink through the head-end to the consumer's home.**

The cable operator can also demultiplex the incoming signals and re-mix them in any appropriate combination, if it is so desired. The digital transmission techniques used for the cable and the satellite channels are very different and are so indicated in Figure 1. In the early stages of deployment of such a system, the cable operator can also decompress the compressed video and audio, at the headend, and transmit uncompressed analog video over cable, as is done today.

## SYSTEM OVERVIEW

Figure 2 shows the function of each program encoder in greater detail. An individual program encoder compresses the video and audio associated with one program, separately encrypts them and the auxiliary data, and outputs a multiplexed stream. In the system we describe in this paper, the video and audio compression algorithms used are MPEG (Moving Pictures Expert Group) compliant. The MPEG video compression algorithm is currently undergoing standardization by the ISO (International Standards Organization), and is based on the Discrete Cosine Transform and motion compensation. The audio compression is provided by MPEG audio.

**TABLE 1. System Parameters**

| | |
|---|---|
| **Video** | |
| Raster Format | 525 lines 2:1 interlaced |
| Frame Rate | 29.97 Hz |
| Aspect Ratio | 4:3 or 16:9 |
| Pixels per line | 720 |
| Lines per frame | 480 |
| **Audio** | |
| Number of channels | 4 or more |
| Bandwidth | 20 kHz |
| Sampling frequency | 32, 44.1 or 48 kHz |

The different encrypted bit streams are then multiplexed, as shown in Figure 2, to form the output of a single program encoder. The outputs of different program encoders can possibly have different bit rates, due to the variability in the program information, and are statistically multiplexed together to form a single bit stream. The multiplexer shown in Figure 1 therefore, also includes a transport layer protocol to allow for effective synchronization, and identification of the bit streams emanating from the different program encoders. At the cable headend, the multiplexed data is demultiplexed into different streams corresponding to the data output of each program encoder. These can now be remultiplexed in any combination desired. The different data streams multiplexed together can also originate from different satellite transponders as shown in Figure 1. This provides enormous flexibility to the headend programmer to selectively show certain programs in a particular viewing area.

Table 1 provides a listing of the video and audio related system parameters for the system.The video can be at either 4:3 or 16:9 aspect ratio. Four mono audio channels, each at 128 kbps are available for every video channel. These can be used to provide four separate mono audio tracks or two CD quality stereo pairs. The system transmits 30 Mbps of total data in the 6 Mhz cable channel, as well as the 24, 27 or 36 Mhz satellite channels.

**TABLE 2. Total bit rates required for different levels of video quality.**

| Quality | 60 Hz Video Source | 24 Hz Movie Source |
|---|---|---|
| VHS | 2.0 - 2.5 Mbps | 1.5 - 2.0 Mbps |
| NTSC | 2.5 - 6.5 Mbps | 2.5 - 5.0 Mbps |
| CCIR 601 | 3.5 - 9.5 Mbps | 3.0 - 7.5 Mbps |

Table 2 lists the total bit rate that is required for 24 Hz movie and 60 Hz interlaced video, for three different levels of quality. VHS quality is equivalent to that which can be obtained on video played off a VHS tape, while NTSC quality is similar to that seen on today's TV sets. CCIR 601 quality refers to video at virtu-

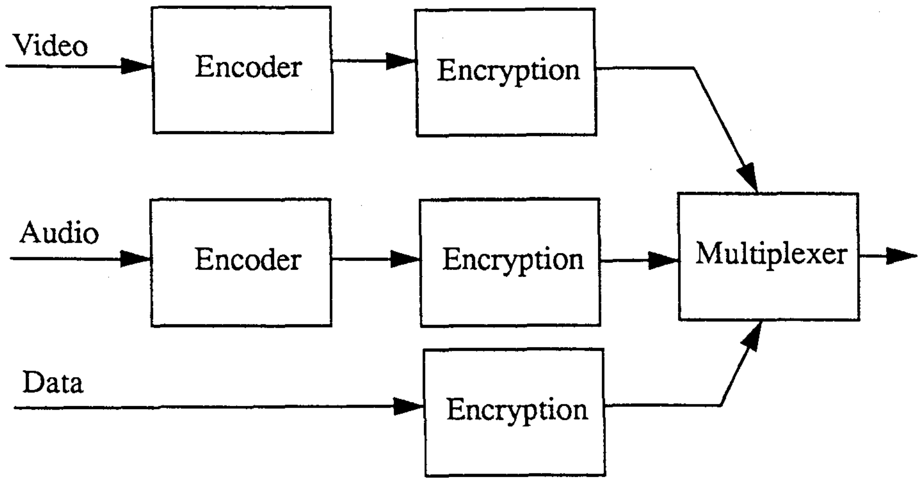


**Figure 2: Block Diagram of the Program Encoder.**

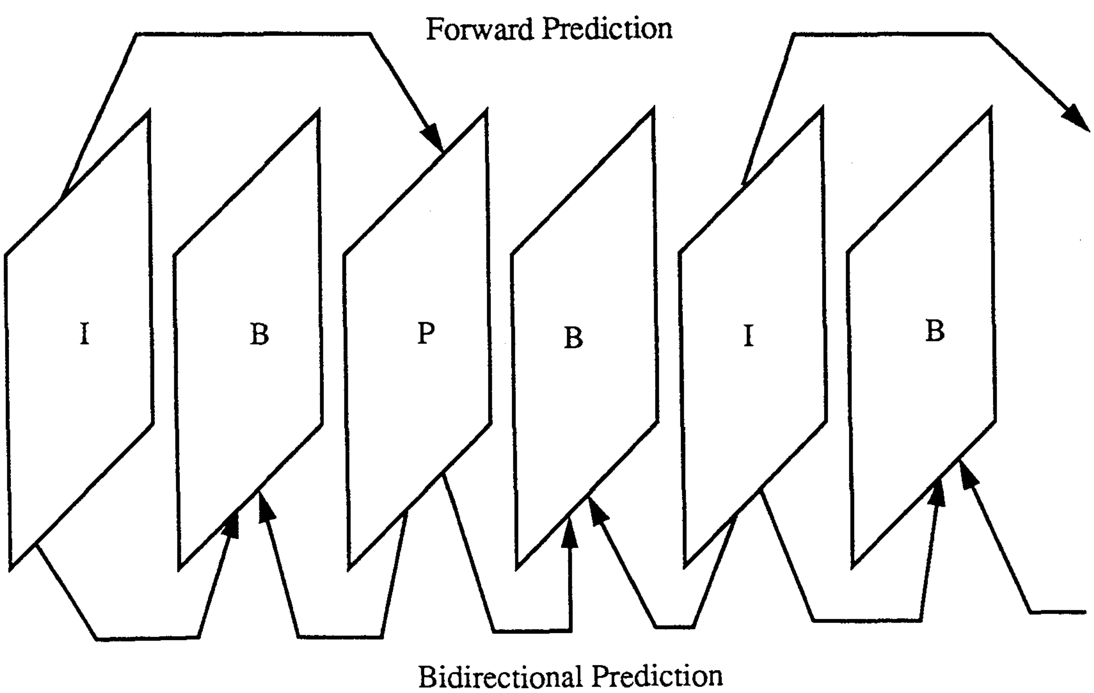


**Figure 3: Frame structure in MPEG**

ally studio quality. These rates include over 700 kbps of audio and data per channel.

## VIDEO COMPRESSION AND PROCESSING

The input video to the video compression module of the system is of CCIR 601 resolution, since CCIR 601 is the standard for studio quality digital television. It also provides for easier exchange of material between the NTSC, PAL and SECAM television formats. For 525-line signals such as NTSC, each video frame has 720 samples per line, and 480 lines per frame. The input is also 2:1 interlaced. The video is converted to the YUV format, and the chrominance components subsampled to 240 lines per frame, and 360 samples per line. The YUV components are then compressed using an MPEG compliant algorithm.

The MPEG standard has been designed after extensive computer simulations, and provides a high degree of compression while maintaining good video quality. The proposed MPEG standard is based primarily on two compression techniques: the Discrete Cosine Transform (DCT), and motion compensation. Motion compensation attempts to exploit the redundancy between frames, by estimating the direction of motion of blocks from one frame to another, while the DCT reduces the spatial redundancy that exists between samples.

MPEG differs from other video compression schemes in the way it implements motion compensation. Unlike many other video compression schemes that use frames in the past to predict frames that occur later, MPEG uses frames from the past as well as from the future (these frames are collectively termed reference frames) to perform motion compensation. The use of non-causal and bidirectional motion compensation is one of the strengths of the proposed MPEG standard, and sets it apart from other video compression techniques.

The use of motion compensation implies that the decoder requires the reference frames in order to completely reconstruct the motion compensated frame. To allow a decoder to start decoding from the middle of a sequence (as in the case of a channel change), it is necessary to have frames that use no motion compensation. To support this and to restrict the propagation of any bit errors that occur during transmission, MPEG compresses frames in one of three different ways. Intra-coded or I-frames use no motion compensation at all and use only the DCT to perform data compression. As such, they can be decoded independent of other frames. Predicted or P-frames use motion compensation, but use only those frames that occur in the past as reference frames. Typically, this is the closest I or P frame. Finally, the interpolated or B-frames (B for bidirectional prediction), use frames from both the past as well as the future as reference frames in its motion compensation. This enables the encoder to predict uncovered areas, as well as objects that have recently moved into the frame. The reference frames used by the B-frame are the nearest I and P frames.Since B frames require far fewer bits than I or P frames, it is possible to get increased data compression while maintaining excellent video quality. Figure 3 illustrates one possible sequence of I, B and P frames in MPEG.

Figure 4 shows a block diagram of a typical MPEG encoder. The input video is converted to YUV component video. The chrominance components are then decimated to be a quarter of the resolution of the luminance. All three components are then grouped into blocks of 8x8. The four adjacent 8x8 luminance blocks comprising a larger 16x16 block, and the two associated chrominance blocks comprise what is termed a macro-block. Motion estimation and compensation is done for each macroblock. As mentioned earlier, I, B and P frames are coded differently. In the case of I frames, the DCT is applied directly on a block by

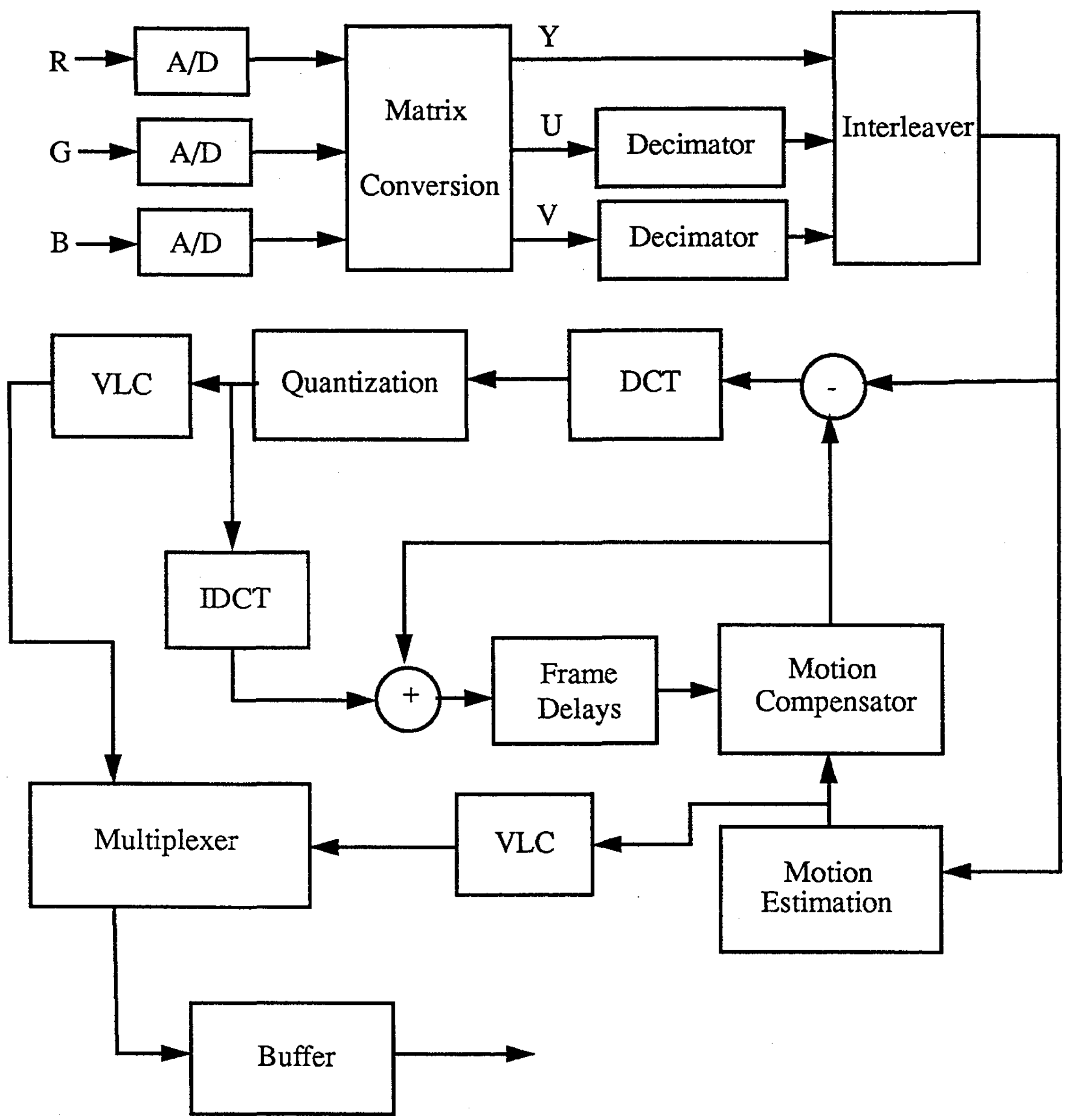


**Figure 4: Digital Video Encoder Block Diagram**

block basis. The DCT coefficients are then quantized, and coded using a variable length coder.

For B and P frames the DCT is applied on blocks obtained after motion compensation. Thus, the DCT is in this case applied on a difference signal. Motion compensation is done with respect to the reference frames (two reference frames for B-frames, and one reference frame for P-frames), and a "mode-selection" technique used to determine the best way to handle each macroblock. For instance, if the motion estimation is not successful, the algorithm intra-codes the macroblock using just the DCT. Furthermore, for B frames, motion-compensation could be done in a multitude of ways. The two reference frames can be used together to predict the macroblock. Or, the prediction could be done by using just one of the two frames. This is typically done when a scene change occurs between the frame being coded, and one of the reference frames. The quantized DCT coefficients and the motion vectors are then coded using a variable length coder, and transmitted.

## Motion Estimation

Motion estimation is typically done on a macroblock basis, and involves determining for each macroblock in the current frame, the corresponding macroblocks in the reference frames. P-frames use one reference frame (the previous I or P frame) while B-frames have two reference frames (previous and future I or P frames). Motion estimation is one of the most computationally intensive operations that is required by MPEG. For each macroblock, it involves locating within a search window in the reference frames, the macroblock with the best match.An MPEG encoder can do this in many ways - full search, telescopic search and hierarchical being some of these.

## Discrete Cosine Transform

On the B and P frames, the DCT is used to further reduce the spatial redundancy, subsequent to motion compensation. For I-frames, the DCT is applied directly on the frame. The DCT is an orthogonal transform, is filter-bank oriented, and thus has a frequency domain interpretation. This allows the system to exploit the properties of the Human Visual System (HVS). The HVS is less sensitive to high diagonal frequencies than it is to high horizontal and vertical frequencies. Thus, the DCT coefficients corresponding to diagonal frequencies are quantized more coarsely than are those corresponding to the horizontal and vertical frequencies. Furthermore, most naturally occurring images have higher energy distribution in the low frequency coefficients and less in the higher frequencies. To utilize this, the DCT coefficients are ordered according to the zig-zag scan shown in Figure 5. This one-dimensional ordering of the two-dimensional DCT coefficients increases the possibility of a large number of the quantized coefficients being zero. This is exploited by run-length coding the DCT coefficients as a "run of zeroes - amplitude" pair.

## Movie Material

Movie material originates at 24 frames per second, and is displayed on television sets at 60 fields/second using a process termed three-two pulldown (Figure 6). The video encoder recognizes video material that has gone through a three-two pulldown process, and converts it back to 24 Hz. The material is then compressed and transmitted and the three-two pulldown done at the decoder, prior to display. This enables the encoder to take advantage of the lower frame rate as well as avoid an intraframe scene change.

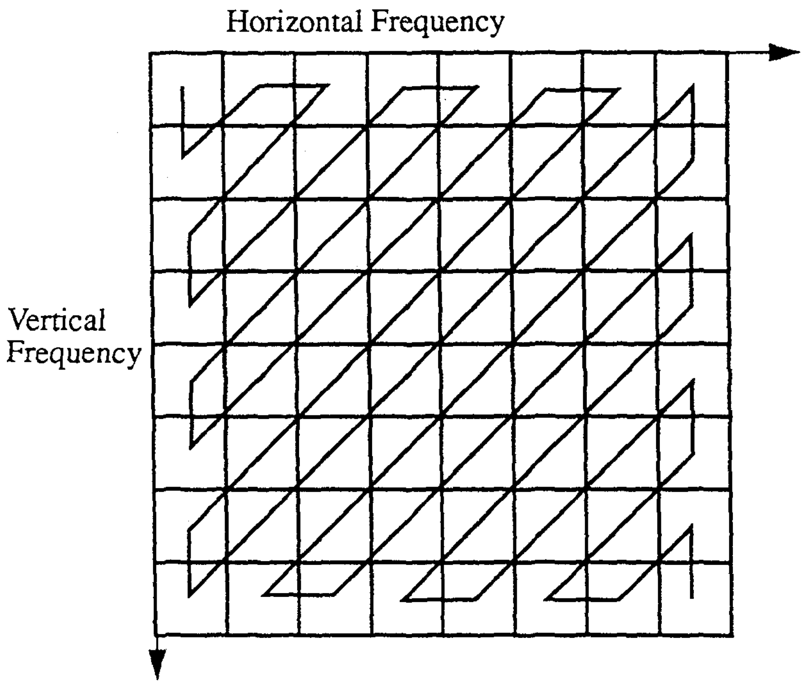


**Figure 5: Zig-zag scan for the Discrete Cosine Transform**

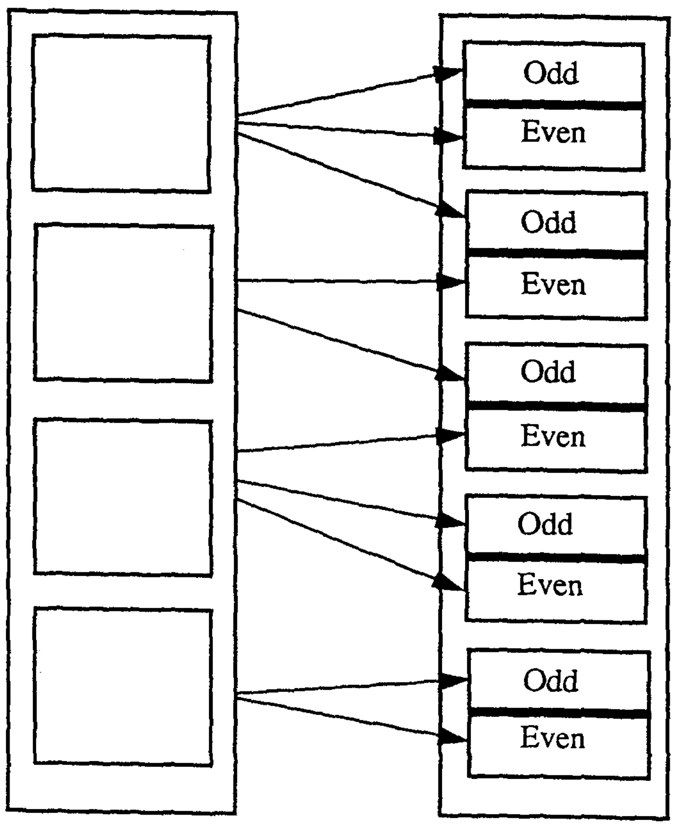


**Figure 6: Three-two pulldown for movie material.**

### Error Concealment

The error rate for the system is restricted to less than one error event per day. However, even when the error does occur, it is concealed by the use of appropriate error concealment techniques. Since, the video compression algorithm uses frames from the past and the future to perform data compression, these reference frames are buffered at the decoder, in order to assist in the decoding process. Information from these frames is used to predict the pixel values in the current frame that have been lost due to transmission errors. Thus, the displayed video suffers minimal visible degradation.

## REFERENCES

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# APPLICATION OF ERROR CONTROL TECHNIQUES TO DIGITAL TRANSMISSION VIA CATV NETWORKS

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## ABSTRACT

Digital carriage of data via CATV networks is becoming more prevalent with each passing year. Digital radio services are presently being offered by numerous cable operators. Digital compression of video is just over the horizon, and other data services will likely follow. Although these digital delivery systems do not exhibit the impairments of their analog counterparts, bit errors may cause catastrophic distortions. Forward error correction is one aspect of a solution to this problem.

This paper introduces the CATV system engineer to the concepts of forward error correction, and discusses its benefits, complexity, and limitations. It also touches on the interdependence of forward error correction with channel equalization and efficient modulation. Several important concepts, such as coding gain, are discussed in detail.

## INTRODUCTION

A CATV system meets the definition of a communication system because it connects multiple information sources to users of this information. A general communication system is illustrated in Figure 1. For purposes of this paper, the source is any source of television programming; the **source encoder** might be some form of video and audio compression. The **forward error encoder** and decoder in Figure 1 are the subject of this paper.

Error control techniques can be very effective against random noise impairments, but are not a panacea for microreflections on digital transmission in a CATV network. Error control can be teamed with channel equalization, as shown in Figure 1, to develop a very robust and cost effective communication channel. Both the rate at which the errors occur and their distribution must be known before the optimum error correction scheme may be designed. This may be accomplished by a combination of simulation, laboratory tests, and field tests.

Error control applied to future CATV networks using video and audio compression (the source encoder and source decoder in Figure 1) is essential because compression eliminates the redundancy from the original analog signals. Errors occurring during transmission may cause severe impairments to the reconstituted analog signals. In general, errors will propagate through the decompression process.

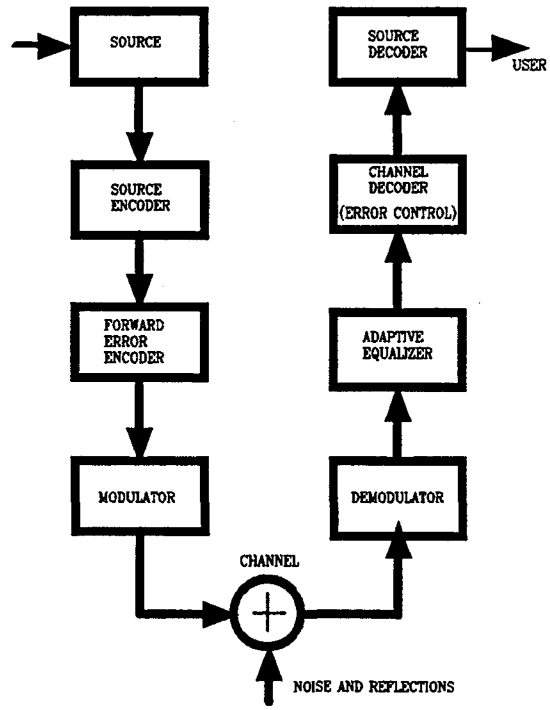
## HISTORY

The history of error control began in 1948 with Claude Shannon's famous paper on channel capacity[1]. His channel capacity theorem says the following:

$$C = W \log_2(1 + S/N) \text{ bits/sec} \quad (1)$$

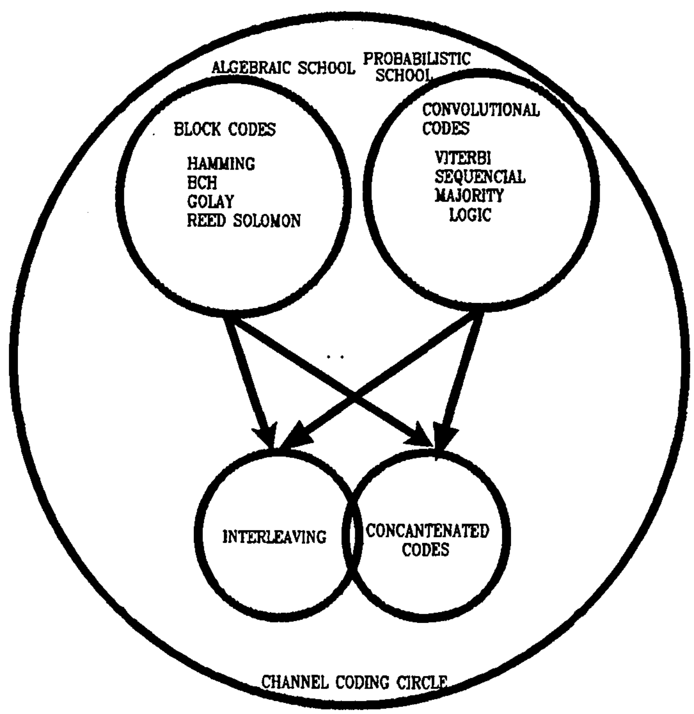
where:

$C$ = capacity in bits/sec
$W$ = bandwidth in Hz
$S$ = signal power
$N$ = noise power




DIGITAL COMMUNICATIONS SYSTEM

FIGURE 1




CODING FOR ERROR CONTROL
FIGURE 2

What Shannon said was that the noise limits the rate at which we can send information, but not the accuracy. Today designers are moving ever closer to this limit with a combination of error control and efficient modulation. Shannon's work also tells us it is more cost effective to employ error coding than to try to build an error-free channel.

During the 1950's and 1960's the search for good codes continued. It was during this period that two mathematical bases developed to solve the error coding problem. This concept is illustrated in Figure 2. The two bases are the **algebraic** and **probabilistic** approaches. The algebraic codes are most commonly known as "block codes". The first of these were introduced in 1950 by Hamming; his are a class of single error correcting codes. Another major milestone occurred in 1960 when Bose, Ray-Chaudhuri, and Hocquenghem found a class of multiple-error-correcting codes now known as BCH codes . Reed and Solomon also developed their codes in 1960; these codes are related to the BCH, but for non-binary channels.[2,3]

The second mathematical approach to coding, the probabilistic approach, led to the development of "convolutional" or "tree" codes. In the late 1950's, studies led to the notion of sequential decoding and to the introduction of non-block codes of indefinite length. However, the most well known algorithm, the Viterbi algorithm, did not appear until 1967. Such techniques have allowed reception of digital data from deep space probes. The steady improvement in the performance of telephone modems has also resulted from advances in error coding and sophisticated modulation techniques.

## ERROR CONTROL AND EQUALIZATION

Error control is very effective at mitigating the impairments caused by additive noise. A CATV channel presents other phenomena that limit channel performance. Chief among these are microreflections due to impedance mismatches at the television receiver or unterminated taps. This results in **intersymbol interference** (ISI), which is the tendency of received symbols to flow into one another.[2] This can not be overcome by increasing signal power; there is an ISI noise floor that increases with signal power. ISI may be overcome by **adaptive equalization**, which is outside the scope of this paper. Error control and adaptive equalization may be combined to result in a very robust communication system (refer to Figure 1).

## DEFINITIONS

Certain terms appear throughout the literature of coding theory. These are defined here for the convenience of the reader:[2,3]

Symbol A symbol is a group of bits within an error control block. It is also defined as a signal representing a group of "k" bits in some analog manner, such as amplitude or phase. Thus, there are error control symbols and modulation symbols.

Weight The weight of a symbol, codeword, or "vector" is the number of non-zero elements.

Hamming distance The Hamming distance between two vectors having the same number of elements is defined as the number of positions in which the elements differ. This is a key concept in error control and will be discussed in more detail later in this paper.

Minimum distance The minimum distance "d" of a linear block code is the smallest distance between pairs of different codewords in the code.

Codeword A codeword or "code block" is a group of bits or symbols made up of information elements and parity (error control) elements.

Code rate Assume that a block encoder accepts information in successive "k"-bit blocks and for each k bits generates a block of "n" bits, where n > k. The code rate R = k/n is a dimensionless ratio that indicates the portion of an encoded block that carries information.

Overhead This is the percentage of parity bits that must be appended to the information bits in constructing a code.

Hard decision A hard decision demodulator makes an absolute 1/0 choice on each received bit (or symbol). The symbol is quantized to two levels.

Soft decision In making a soft decision, the demodulator makes a bit-quality measurement on each bit or symbol. The symbol is quantized to more than two levels.

Erasure This is the process of flagging a bit or symbol as unreliable. It is the result of a soft decision. This flag is passed along to the error control circuitry.

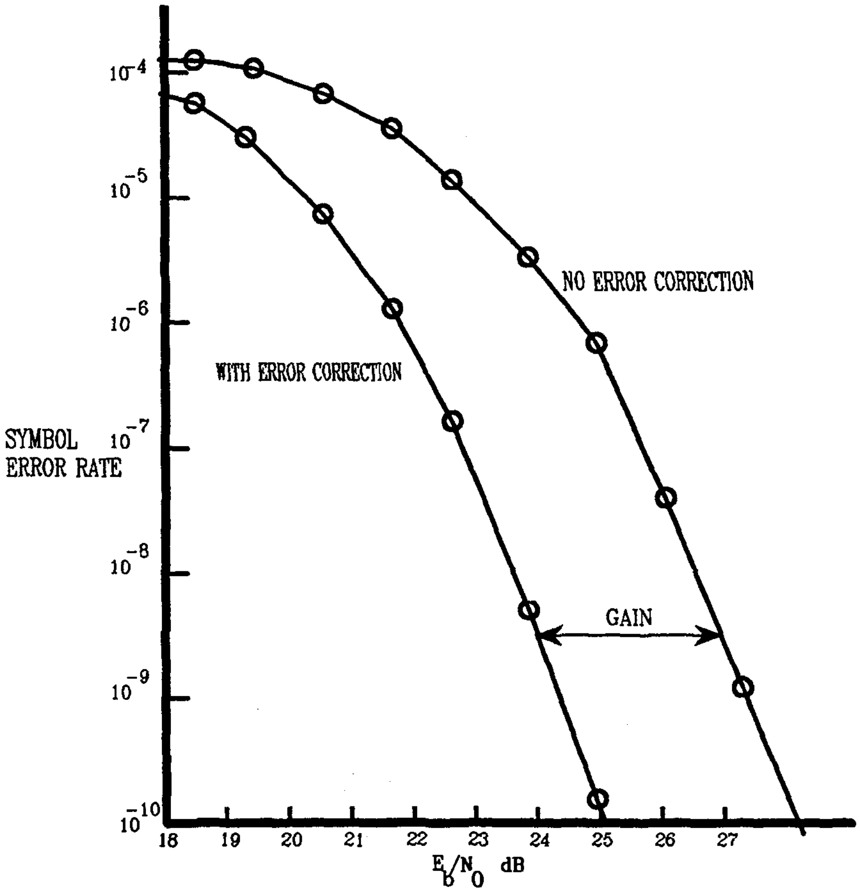
Coding gain This term describes the amount of improvement that is achieved when a particular coding scheme is used. Figure 3 illustrates coding gain on a logarithmic plot of bit error rate vs $E_b/N_0$ (energy/bit divided by spectral noise density). At low signal to noise ratios, the gain will become negative.

Vector This term is based in linear algebra and is familiar to us from physics. In coding theory vector space is one of the most important algebraic concepts. The vector provides a convenient representation of field elements that may be implemented with simple digital functions. The term is also used in matrix notation, where the vector consists of the coefficients of a polynomial. Refer to section 3.3 of reference 2.

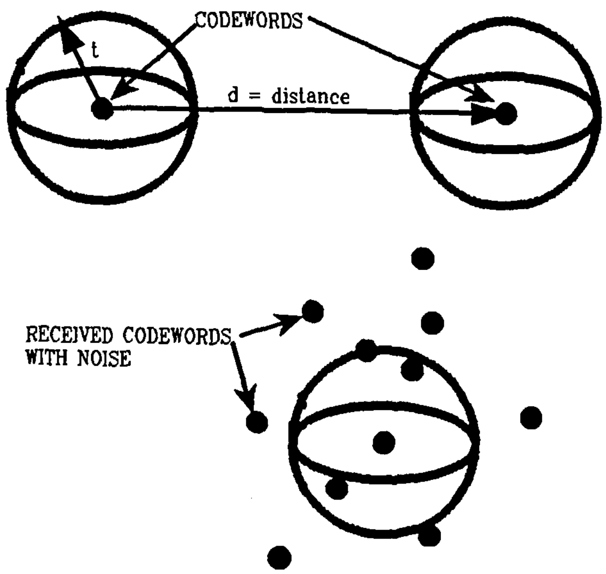
The syndrome The syndrome is defined in the dictionary as "a number of symptoms occurring together and characterizing a specific disease".[6] In coding theory, a syndrome is a sequence of discrepancies which occur when received parity bits are compared with calculated parity bits. The syndrome may take on the form of a "vector" in a matrix. Calculations of syndromes are used in many decoding algorithms to locate errors in received data.

Constraint length In a convolutional code, the constraint length is the number of data frames used in the generation of the encoded data. Each input frame may consist of one or more bits. The process occurs on a continuous basis. In terms of the actual circuit elements, the constraint length is the length of the input data shift register in the encoder.

Galois field A field having a finite number of elements is called a finite or Galois (pronounced gall-wa) field. It is denoted by GF(q), where q is the number of elements in the field. These fields are named after Evariste Galois (1811-1832), a French mathematical prodigy who established group theory mathematics by age 17.[2] Chapter 4 in reference 3 treats this theory in detail.




CODING GAIN
FIGURE 3




DECODING SPHERES
FIGURE 4

## THE DISTANCE CONCEPT

This concept is crucial in visualizing the operation of error control circuitry. Figure 4 illustrates "decoding spheres" in a geometric fashion. Recall the definition of minimum distance $d_m$. We will define t as the number of errors that a particular code can correct. If more than t errors occur in transmission, the decoder may incorrectly decode the data or it may indicate with a flag that it can not decode the message. [3]

In Figure 4, the code is designed so that the minimum distance between codewords is defined as:

$$d \geq 2t + 1 \qquad (2)$$

where t is the number of errors that can be corrected. A codeword received error free will land at the center of a sphere. If t errors occur, the codeword will be on the surface of the sphere, and the decoder will correct the error(s). Received codewords with more than t errors may fall between spheres or within another sphere; those falling in another sphere will be incorrectly decoded. Those falling between spheres may or may not be correctly decoded, but would be erased in a soft decision decoder. One can see that the minimum distance is a critical property of a code.

## BLOCK CODES

In a block (algebraic) code the encoder accepts k information bits and appends r parity-check bits to form a block of n bits, such that:

$$n = k + r \qquad (3)$$

where n = block length
r = number of parity bits

The code is referred to as an (n,k) code. The code rate R is k divided by n. Each block is independent of all others; the check bits are completely determined by the information bits within the same codeword. Also, there are $2^k$ codewords in the code set. The code is designed to make the codewords very different from each other to resist channel errors.

Arithmetic operations in the Galois field GF(2) are simple because no overflow or round-off error is permitted. The operations of addition and multiplication are mod-2. This is illustrated in the following tables:

| + | 0 | 1 |
|---|---|---|
| 0 | 0 | 1 |
| 1 | 1 | 0 |

ADDITION

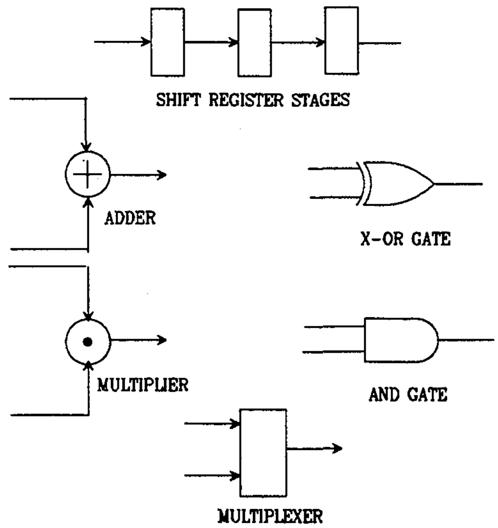
| * | 0 | 1 |
|---|---|---|
| 0 | 0 | 0 |
| 1 | 0 | 1 |

MULTIPLICATION

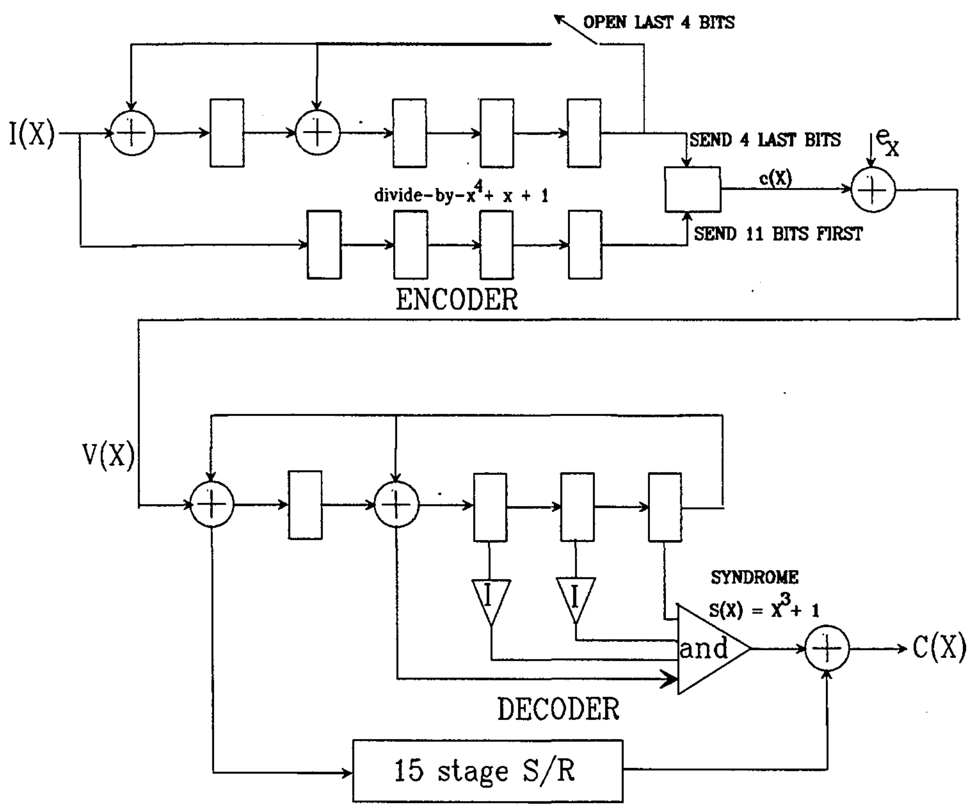
Addition bit-by-bit is accomplished with an "X-OR" gate. Multiplication is done with an "AND" gate.

Polynomial arithmetic in a Galois field (in this case GF(2)) can be used in the description of block codes. Fortunately, digital logic circuits may be constructed to mimic this special polynomial arithmetic. These circuits take the form of digital filters, and are constructed of shift register elements, X-OR gates, AND gates, and multiplexers (Figure 5). The form of the encoders and decoders are similar.

We choose for this paper "binary cyclic block codes" to illustrate the relationship of the GF(2) polynomial arithmetic to the actual circuits. We do so because these codes have proven useful and efficient in practice. Binary cyclic block codes are a subset of




CIRCUIT ELEMENTS
FIGURE 5




CYCLIC BLOCK ENCODER/DECODER

FIGURE 6

linear block codes, and fall in the "algebraic school" circle of Figure 2. A binary code must meet two criteria to be cyclic:

a. The code is linear; bit-by-bit addition of two codewords in GF(2) is again a codeword.

b. Any cyclic (end around) shift of a codeword is also a codeword. [2]

Chapters 4, 5 and 6 of reference 3 give the reader a clear understanding of the mathematical basis and implementation of cyclic block codes. The polynomial description of a codeword is also found in chapter 4 of reference 2 as follows (in general form):

$$\text{let } c(x) = c_0 + c_1x + c_2x^2 + \dots + c_{n-1}x^{n-1} \quad (4)$$

where n = block length, and the polynomial is of degree n-1. Now we will develop an example, as shown in Figure 6. If the information polynomial is:

$$i(x) = i_0 + i_1x + i_2x^2 + \dots + i_{k-1}x^{k-1} \quad (5)$$

and the generator polynomial is:

$$g(x) = x^4 + x + 1 \quad (6)$$

(derivation of generator polynomials is given in references 2 and 3)

then the codeword takes the form:

$$c(x) = x^{n-k}i(x) + t(x) \quad (7)$$

where $t(x)$ is the remainder, and is equal to:

$$t(x) = -R_{g(x)}[x^{n-k}i(x)] \quad (8)$$

this reads "t(x) is the remainder after dividing by g(x)".

and thus

$$R_{g(x)}[c(x)] = 0 \quad (9)$$

The encoder in Figure 6 is a systematic encoder that implements a divide-by-g(x) using shift registers and X-OR gates; it produces a (15,11) Hamming code. Assume that the register stages are first cleared to zero. Eleven information bits are shifted into the circuit; division begins after four clock shifts. The circuit produces eleven information bits followed by four parity bits, to produce a fifteen bit codeword. The four parity bits are the result of the division.

Refer again to Figure 6. As the codeword passes through the channel, noise may cause bit errors. This noise is represented as the error polynomial e(x), which has degree n-1. The sum of the codeword c(x) and noise e(x) is v(x), the received codeword:

$$v(x) = c(x) + e(x) \quad (10)$$

The decoder in the figure implements a divide by g(x), where g(x) is the same generator polynomial used in the encoder. If no error has occurred, the remainder is zero. If the remainder is non-zero, it is calculated as:

$$s(x) = x^3 + 1 \quad (11)$$

s(x) is the **syndrome** defined earlier! The decoder circuit in the figure calculates s(x) by dividing by g(x); if s(x) is non-zero, the appropriate information bit is inverted, yielding the original information codeword c(x). The encoder and decoder of Figure 6 constitute a single-error-correcting system. Note the simplicity of the circuit, but remember it is limited to correcting single errors.

The example just presented is of a **binary** block code; the coefficients of all the polynomials are either binary 0 or 1. As you may recall from our brief history lesson, Reed and Solomon developed multiple error correcting codes in a 1960 paper. These codes (and there are many) are are very effective in the presence of burst errors. The **overhead** of these codes is typically 10% or less, making them very efficient. However the decoding hardware is far more complex than described above for the binary code. Algorithms for decoding of R/S codes must calculate two syndromes, one for error location, and one for error magnitude. This is because the mathematics is over a Galois field $GF(2^m)$, where m is a small integer on the order of 7 or 8. In hardware, a parallel bus m bits wide is required. The data bits are arranged into "symbols" of m bits, and the arithmetic calculations are done on these symbols. A number of sophisticated decoding algorithms have been developed for the many Reed Solomon codes. They have found many practical applications, such as compact discs.

## CONVOLUTIONAL CODES

These codes are based on a probabilistic approach to the problem of error control. They were originally called **recurrent** codes, and are also referred to as **tree** codes, from the use of a tree or trellis diagram used to visualize the sequence of events. A convolutional code does not have a simple block structure, with each codeword independent from all others. Rather, the codewords are generated using a **sliding window** over the information symbols. A continuous stream of encoded symbols is produced, where successive codeword frame are coupled together by the encoder.

Figure 7 illustrates a generic convolutional encoder, and will be used to define terms common in the literature. The input information is broken into **information frames** of $\mathbf{k_0}$ symbols; **m** is the number of these frames stored in the encoder shift register. The length of the shift register is **m** X $\mathbf{k_0}$, which is the **constraint length**, denoted by **v.** The output **codeword frame** is made up of $n_0$ symbols. The code is referred to as an **$(n_0,k_0)$ code.** **K** is the **wordlength** of the code and is equal to $(m + 1)k_0$. **Blocklength N** is equal to $(m + 1)n_0$, and is the length of the output code that may be influenced by an input frame $\mathbf{k_0}$. The rate **R** of the code is $k_0/n_0$. The input to the encoder is data at a rate of $\mathbf{k_0}$ symbols per second, and the output is data at a rate of $n_0$ symbols per second.[3]

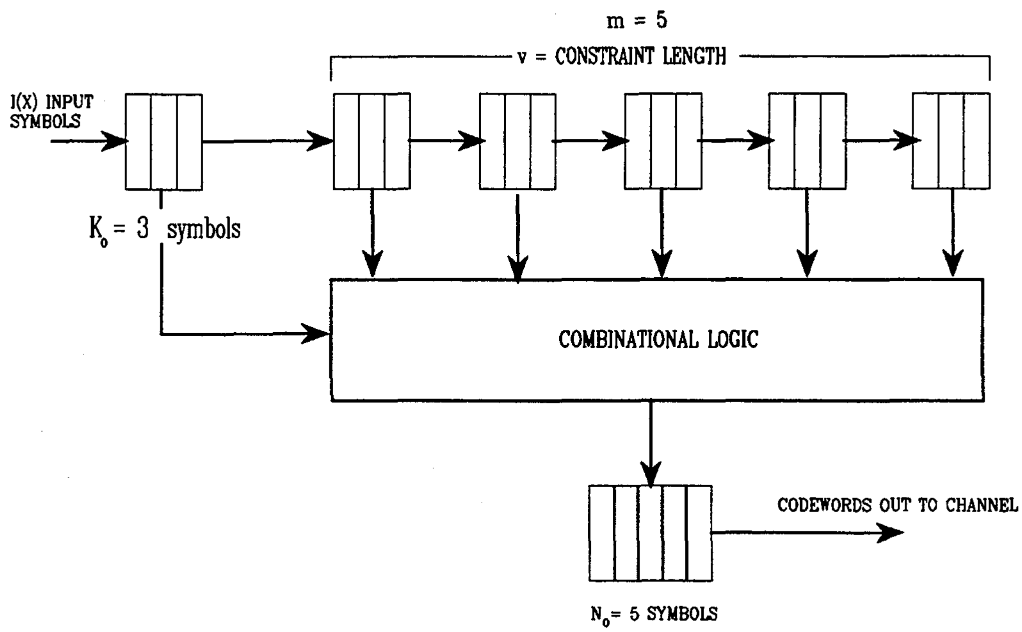
Next we will consider the mathematical basis for these codes. We used a **generator polynomial** in constructing a block code. Convolutional codes require a set of multiple polynomials to describe them; these are best described by a mathematical **matrix.** Matrix notation provides a means of writing a number of simultaneous equations (polynomials) in compact form. Appendix A of reference 2 presents a summary of matrix definitions and manipulations.

A matrix is made up of row and column **vectors**, whose elements are the coefficients of the polynomials.

The generator-polynomial matrix is given by:

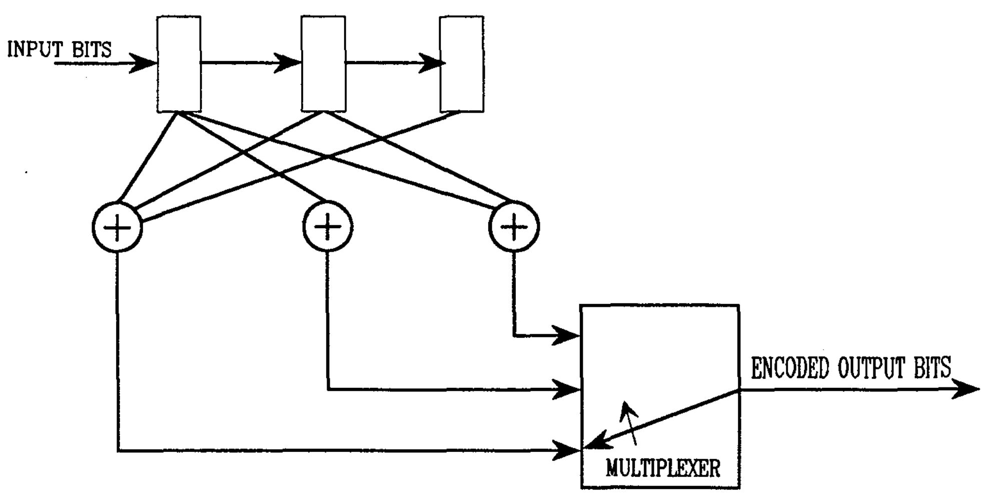
$$G(x) = [g_{ij}(x)] \qquad (12)$$

This is a $k_0$ by $n_0$ matrix of polynomials. Further, if d(x) is a




GENERIC CONVOLUTIONAL ENCODER

FIGURE 7




(3,1) CONVOLUTIONAL ENCODER

FIGURE 8

set of $k_0$ information polynomials and c(x) is a matrix of $n_0$ codeword ploynomials, then:

$$c(x) = d(x)g(x) \qquad (13)$$

Also, there is a parity check matrix H(x) that satisfies

$$G(x)H(x)^T = 0 \qquad (14)$$

and there is a **syndrome-polynomial vector** given as

$$s(x) = v(x)H(x)^T \quad (15)$$

where v(x) represents the received codewords.

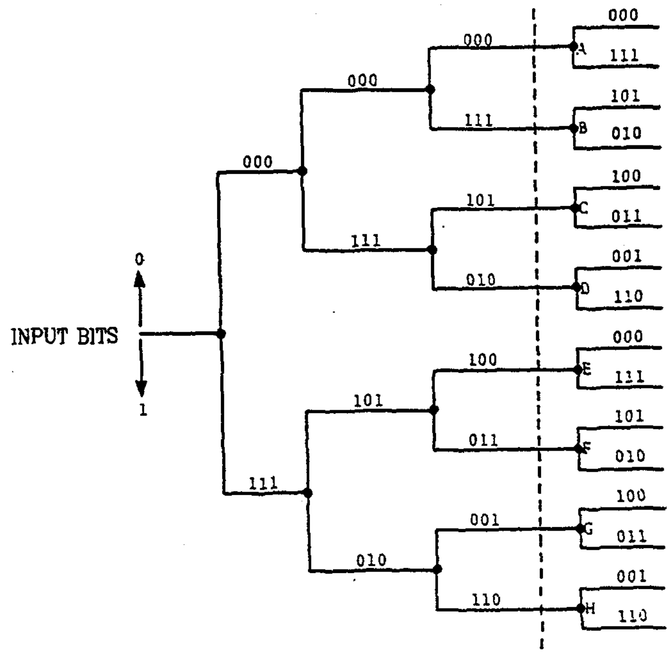
We will now go on to describe the **tree** and **trellis** structures, as these are very useful in visualizing the generation of the convolutional code. We will use an example taken from a paper by Batson. [4] Figure 8 shows a simple encoder made up of a three stage shift register, three XOR gates, and a multiplexer. This is a (3,1) tree encoder. The coefficients of the generator polynomials specify which stages of the shift register are connected to each modulo 2 adder. In Figure 9, we see the tree that describes the operation of the encoder. Assume the shift register contains all zero's to start the sequence. In the tree, an input of zero causes the circuit to follow the upper branch, a logic one the lower branch. The labels on the branches indicate the resulting output code. An input sequence of 1011 results in an output sequence of 111 101 011 010. It can be seen that this tree would grow very large after a relatively short input stream, and would be unwieldy.

An alternate structure which is much more compact is the **trellis diagram.** This may be seen in Figure 9. The state of the encoder is the most recent contents of the $k_0$-1 stages of the encoder shift register. Time is from left to right. The circuit steps from state to state at clock times. The path is up for a logic zero, and down for a logic 1. The encoded output bits are shown on the branches. The repetitive structure of the trellis is immediately apparent.

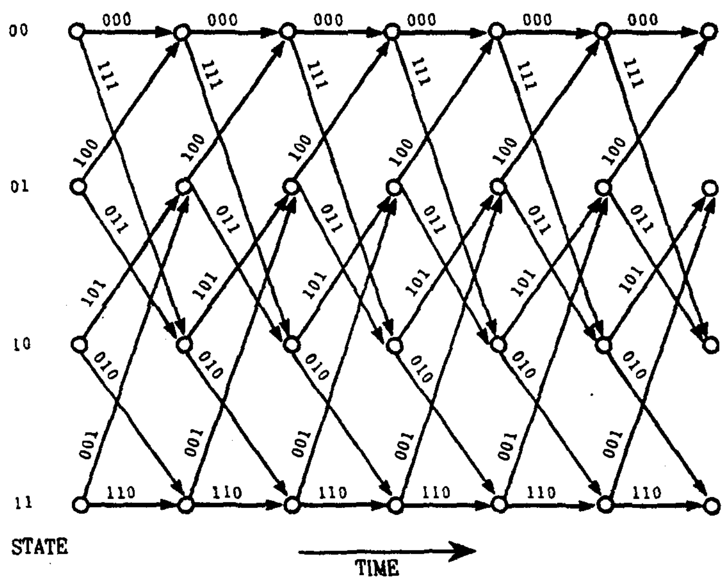
The trellis structure is useful is understanding decoding algorithms for convolutional codes, such as the Viterbi algorithm[4] , which has found wide application. It basically attempts to find a valid path through the trellis that is as close as possible to the received sequence. This method is very effective, but it should be noted that the hardware requirements for the Viterbi decoder grow exponentially with constraint length.

The coding gain of the Viterbi algorithm may be improved by the use of a **soft decision** demodulator. Such a demodulator takes into account the distance of a received symbol from the center of it's decoding sphere. This is accomplished with an analog-to-digital converter (A/D) to quantize the received signal.

We made brief mention of the **syndrome vector** earlier. Next, we will illustrate its use in a syndrome feedback decoder of a convolutional code. Figure 10 illustrates both the encoder and decoder for such a code. The encoder and decoder both calculate the same parity bits if the data is received error-free. However, if a data bit is received in error, the locally calculated parity will differ from the received parity. When this difference occurs, a logic 1 will appear in the syndrome register. It is the function of the decoder decision table to find the **most likely** bit error location. [2]

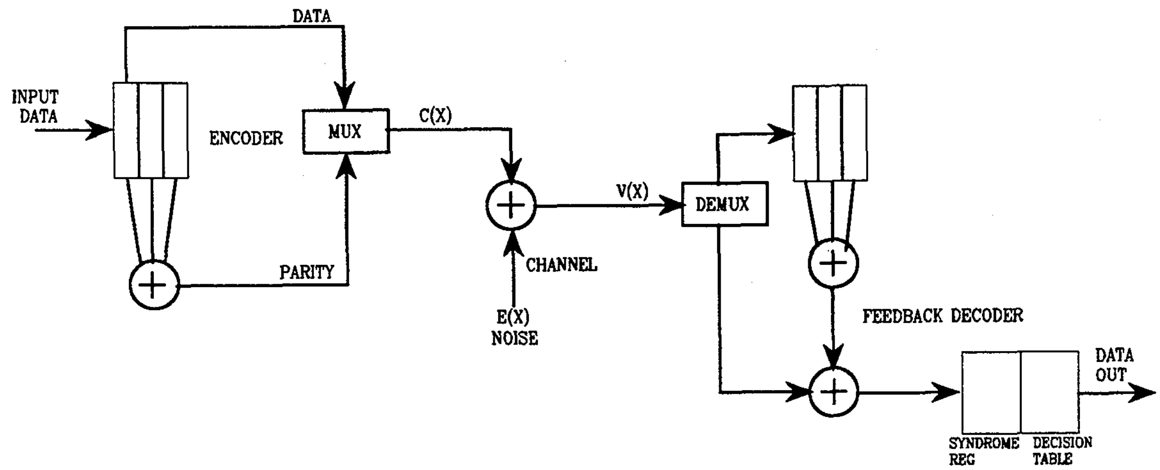



CODE TREE



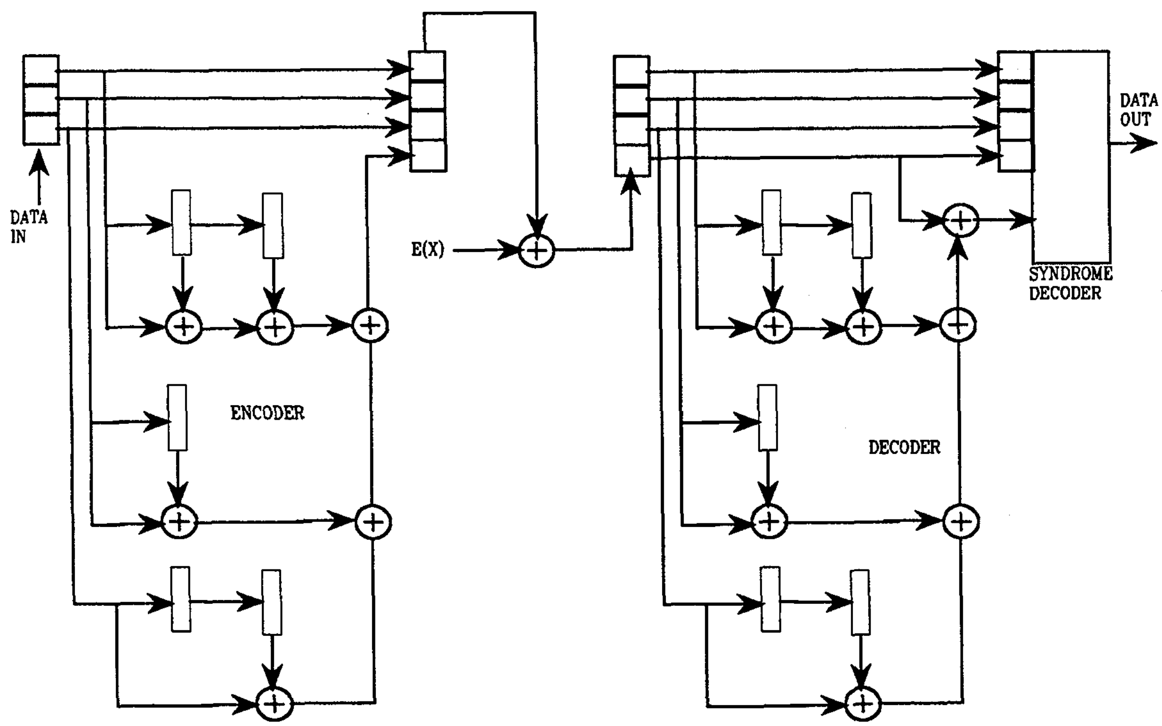

TRELLIS DIAGRAM

FIGURE 9




CONVOLUTIONAL SYNDROME FEEDBACK DECODER

FIGURE 10




WYNER-ASH ENCODER AND DECODER

FIGURE 11

More than one error pattern can result in the same syndrome; the decoder will choose the pattern with the least errors and compensate for that pattern.

One more example of a convolutional encoder/decoder combination is illustrated in Figure 11; the Wyner-Ash code is used here.[3,5] The decoder uses the syndrome concept. The Viterbi decoder described earlier is a better method of improving coding gain.

## INTERLEAVING AND CONCANTENATED CODES

Figure 2 illustrates how these two techniques relate to block codes and convolutional codes. Figure 12 depicts a hardware block diagram of a system employing these techniques.

Interleaving is used to transform a bursty channel into an independent error channel by scrambling the encoded symbols before transmission. An interleaver structure is built from semiconductor memory in a rectangular array. Encoded data is written into the array by rows and out by columns before transmission. After reception and decoding by the decoder, the process is reversed. This techniques has proven effective in satellite links that are subject to long bursts of errors.

Concantenated codes are used to increase coding gain. A Reed-Solomon code is used with the Viterbi decoder in Figure 12 due to the bursty nature of uncorrected errors out of the Viterbi decoder.[2,3]

## CONCLUSIONS

This paper is intended as an introduction to error control theory for the CATV system engineer. In order to determine the optimum strategy to develop a practical digital delivery system for CATV, a number of factors should be considered:

1. The worst-case allowable symbol-error-rate. This determines the required coding gain.

2. The environment in which the system will operate, to include channel C/N, expected reflections, and other impairments.

3. The type of digital modulation selected (eg, 16QAM, 64QAM, etc). The digital channel bandwidth, the allowable signal power relative to AM channels, and any effect on those AM channels must be considered.

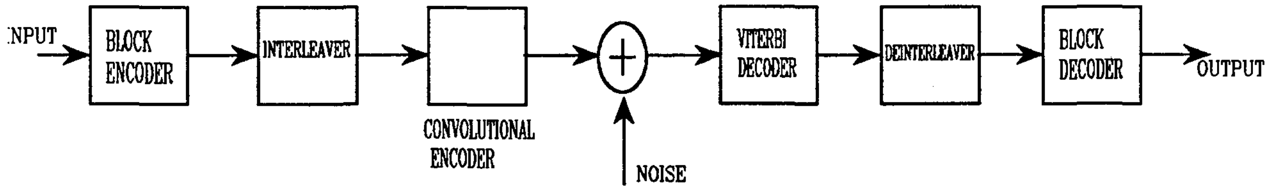
4. The required parity overhead, which increases the symbol rate.

5. The distribution of errors in the channel. A CATV channel is subject to random errors, not burst errors.

6. The behavior of the required adaptive equalizer under various conditions.

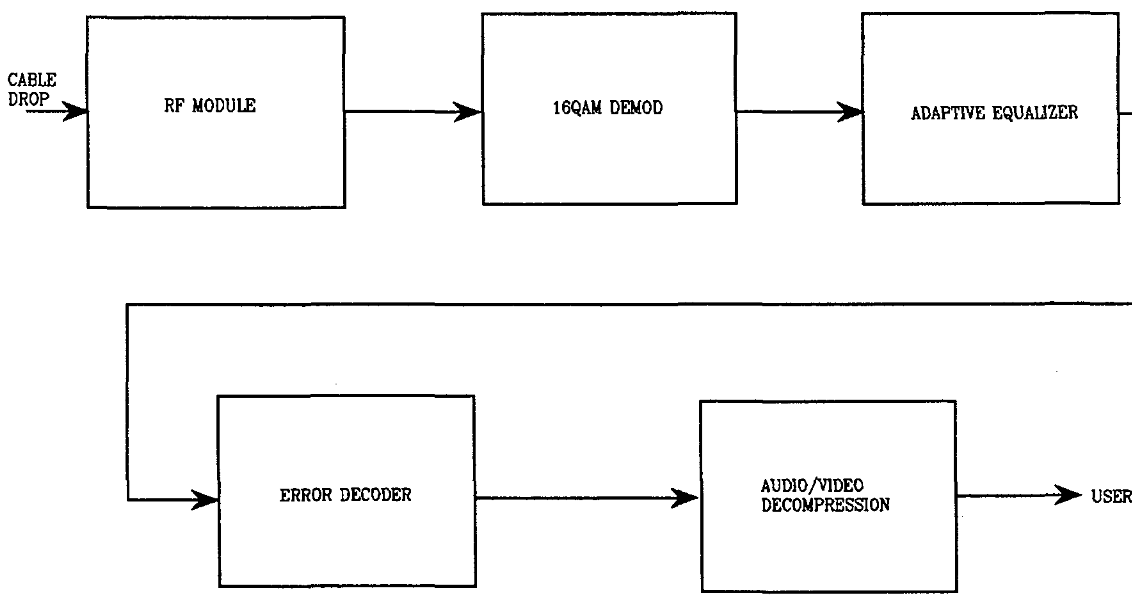
7. The circuit complexity and cost of the hardware, especially in the subscriber terminal.

Figure 13 illustrates the likely functional blocks in a CATV subscriber terminal employing digital data delivery. The demodulator, adaptive equalizer, error decoder, and decompression hardware must be designed to operate in concert. A properly designed system promises to deliver consistent high quality video and audio to all subscribers.




INTERLEAVING AND CONCANTENATED CODES

FIGURE 12




SUBSCRIBER TERMINAL

FIGURE 13

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# BENEFITS OF COMPRESSED VIDEO DATA TRANSMISSION THROUGH AML

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## ABSTRACT

*The advantages of digitized compressed video are particularly relevant to microwave transmission in that such formats permit transmission of many more than the maximum 79 channels which otherwise could be carried in the 500 MHz wide 13 GHz CARS band. A further advantage is the potential for significantly increased microwave fade margin when digitized video is utilized in place of analog VSBAM. Since under normal propagation conditions the noise contribution of AML is only a very small portion of the overall CATV system noise, the focus here must be on the performance of the digital modulation scheme under faded conditions. Transmission of the QPSK Skypix signal through AML systems has already been successfully demonstrated. Considerations relating to the carriage of higher level modulation formats such as 16 QAM through existing AML systems are explored. In this regard, the differences between CATV systems utilizing channelized transmitters and block conversion transmitters are analyzed.*


## DIGITAL VIDEO ADVANTAGES

The principal reason for the intense interest in digitized compressed video transmission through cable systems is the promise of greatly increased video channel capacity. Traditionally, cable systems have increased their channel capacity by extending the bandwidth. Alternatively, dual cable systems offered a way of getting around the limits of the existing technology. Since its introduction in 1971, AML microwave has been able to fulfill the channel requirements of the cable industry through both of these techniques. Starting with only 14 channels (2 to 13 plus J and K) at that time, AML microwave systems today span the full range between 54 and 552 MHz. This, however, fills up the 12.7 to 13.2 GHz CARS band so that continued bandwidth expansion at these frequencies is no longer possible due to the FCC regulatory limit. It is true that the 18.14 to 18.58 GHz band is also open to utilization by CATV operators, but such use would require a separate transmitter and receiver, much as in dual cable systems. At least the 13/18 GHz option is not limited by the site constraints of microwave frequency reuse as in the 120 channel AML system installed in Dallas some years ago.[1]

In fact, considerable development of broadband 18 GHz AML has taken place in the last year in response to the market demands generated by the authorization granted in 1991 for use of 18 GHz by SMATV operators. CATV systems, having had access to 13 GHz where both atmospheric propagation and equipment performance is better, had not, with one exception, utilized 18 GHz AML even though the band was open to them since 1985.[2] If a CATV operator wished to carry 150 VSBAM channels via AML, in theory he could do so by utilizing both 13 and 18 GHz. Nevertheless, it must be recognized that this approach is less attractive than if all 150 channels could be accommodated by a single 13 GHz transmitter/receiver pair with interface up to 1 GHz. Compressed digital video overcomes this problem by squeezing all 150 channels, or more, into the regulatory 500 MHz wide limit.

Another advantage of digital format is that it is not as susceptible to noise and interference as the VSBAM signal. The situation is similar to the so-called "FM improvement factor" which is obtainable at the expense of bandwidth. Indeed, if it were not for the large reduction in the number of bits through the video compression algorithms, the digital signal would have to occupy a much wider bandwidth than the standard VSBAM signal. For a given bit rate, the wider the bandwidth, the greater the immunity to noise and interference. That is why QPSK will be utilized in the satellite downlinks. Higher order digital modulation schemes such as 16 QAM are relatively less robust but should still be satisfactory for transportation through cable systems.

The above characteristic of digitally encoded signals is of particular advantage to AML microwave system operators since path fades are a factor which must be taken into account. With VSBAM, the signal quality degrades linearly with received signal level if the fade depth causes the microwave system's contribution to the overall CATV system C/N to become predominant. The standard design procedure includes a calculation of the probability of the fade causing the C/N to drop below a given value, usually 35 dB. This corresponds to a weighted S/N of 37 dB which, according to a recent survey[3], was judged to be between annoying and very annoying impairment of picture quality. An older survey judged a 38 dB S/N to be "slightly annoying" and the 1958 TASO study described a 28 dB S/N as "somewhat objectionable". This not only illustrates the fact that viewers' expectations have increased over the years, but also that the VSBAM signal is still viewable down to at least 25 dB C/N. By contrast, a digital signal would not degrade in quality as long as the C/N remained above threshold. A threshold of 20 dB has been reported[4] in an experiment involving a 16 QAM signal carried over a test cable system. In the test the VSBAM C/N was 30 dB but the digital signal was carried at a level of −10 dBc relative to the VSBAM pictures.

In an actual cable system application the degree to which the digital signal level would have to be depressed relative to the VSBAM signal, if any, would depend on system loading considerations relative to the quality of the VSBAM signals. As reported, the test illustrates the relative immunity of the digital signal to CTB distortion generated by the VSBAM carriers. Thus, if a system were to carry only digital signals, the microwave transmitter power limiting linearity constraints would be greatly eased and the power output could be increased. This further increases the fade margin. When combined with the improved threshold level, the improvement in microwave path reliability could then in many cases exceed an order of magnitude; i.e., a predicted reliability of 99.9% would become better than 99.99%.

## AML CARRIAGE OF DIGITAL SIGNALS

The fact that AML microwave can successfully carry digital signals is already well established by experiment. A 30 MHz wide Skypix compressed video QPSK digital signal was carried through a cable system which included two distinct AML hops in tandem.[5] It was shown that with the data signal 6 dB below the standard VSBAM video level, the signal could fade to a VSBAM 18 dB C/N before the data signal would start to exhibit tiling. The test illustrates that phase noise in Hughes AML equipment did not degrade the data even with a total of four frequency conversion operations.

Because of the required bandwidth, the above test did not include filters which might conceivably degrade the digital signal threshold. However, a 3 MHz wide digital audio service employing a nine-level QPR modulation is already widely carried in cable systems employing channelized AML transmitters. Two such signals can be accommodated within a standard video channel upconverter and up to six digital audio signals can be accommodated in channels employing wider bandwidth filters such as the standard 88 to 108 MHz FM broadcast channel. The main limitation is imposed by audio signal intermodulation products which fall in the adjacent video channels. For this reason the digital audio signals are carried at least 10 dB below video. The required channel flatness over the 3 MHz increments is readily obtainable with a minimum of care.

## AML CHANNEL CHARACTERISTICS

As one progresses up the ladder of bandwidth efficiency from QPSK to 16 QAM and beyond, the digital system becomes more sensitive to noise and other channel impairments. RF channel characteristics such as AM/PM conversion, frequency response, return loss, and group delay may have to be improved in both the overall cable system and specifically on the AML link when the less robust modulation schemes are considered.

Broadband AML systems employing block upconverter type transmitters as well as the block downconversion receivers, are the least likely to encounter any difficulty in transport of higher order digital modulation signals. Considerations regarding such a system are essentially no different than those pertaining to the trunk portion of the cable plant other than the possibility of a microwave path fade which might temporarily reduce the C/N below 35 dB. This in itself may suggest that modulation schemes less robust than 16 QAM or 4 VSB-AM may not be compatible with existing cable systems.

Channelized AML transmitters employ narrow band filters which introduce group delay. The largest amount of delay can be expected in high power AML systems employing the STX-141 transmitter. This is because there are two such filters plus the klystron, all of which contribute to the delay, in each transmitter channel. Even carriage of 20 Mb/s 16 QAM may be somewhat problematic without filter modification. The STX-141S will be slightly better in this regard since the wide band FET amplifier replaces the narrow band klystron in this unit. The SSTX-145 is better yet with a typical 15-nanosecond delay within the inner 5 MHz of the 6 MHz wide channel. This is a consequence of utilizing a wider band filter in the upconverter stage. Similar performance can be expected of an MTX-132 transmitter channel. Thus it is distinctly possible that even some channelized AML systems may be essentially transparent to passage of digitized compressed video signals without any modification.

## SUMMARY

Compressed video data transmission allows an increase in channel capacity and greater fade margin in AML microwave systems. These advantages can be expected with little or no modification to existing AML systems. It is however important that one recognizes the desirability of achieving a threshold C/N of less than 25 dB so that cable system subscribers will enjoy the full added quality benefits which digitized video is capable of providing.

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# CABLELABS ATV TESTING STATUS REPORT

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**Cable Television Laboratories, Inc.**

*Abstract*

*The Federal Communications Commission (FCC) appointed the Advisory Committee on Advanced Television Service (ACATS) in 1987 to recommend a standard or standards for advanced television in the United States. The committee has developed criteria for the selection of the standard or standards, procedures for the testing of proposed systems to determine whether they will work as proposed and is presently in the middle of the laboratory evaluations of the proposed systems. The laboratory tests are expected to be completed by late summer or early fall of 1992. Follow-up field tests will occur in the winter of 1992/3.*

*This paper reviews the tests that have been utilized by Cable Television Laboratories, Inc. (CableLabs) in performing the tests for the Advisory Committee. At the time this paper was being written the test results of the first systems had not been released by the committee but were expected to be public at the time of the NCTA convention.*

## Background

The FCC appointed Richard Wiley as the Chairman of the Advisory Committee. Three subcommittees were organized: Planning Subcommittee, Systems Subcommittee, and Implementation Subcommittee. The Planning Subcommittee was charged with developing the criteria by which the proponent systems would be evaluated. The criteria included technical parameters, spectrum requirements, economical considerations and alternate media capabilities. The Systems Subcommittee was charged with evaluating and testing the systems against the requirements set down by the Planning Subcommittee and recommending a system or systems for adoption by the FCC. The Implementation Subcommittee was responsible for determining the economic, time, and legal implications of introducing the recommended system or systems. Each of the subcommittees was further divided into working parties and task forces as necessary to complete the work of the committee.

The time frame for determining a standard required that the final report be submitted to the FCC in the third quarter of 1992. This due date has been extended from the original date due to delays in the development of a test facility to conduct the tests and changes to digital transmission techniques by the proponents.

## The Plan for Laboratory Testing

One of the working parties of the Planning Subcommittee developed criteria for the evaluation of proposed systems but did not specify any minimum requirements for the advanced TV (ATV) system. A second working party developed the test plan to evaluate the systems against the requirements for over-the-air transmission while a third working party developed the test plan for cable television transmission. Other working parties developed the procedures and the test material for the subjective tests, considered spectrum requirements and implications, etc. These criteria and test plans were passed on to the Systems Subcommittee working party responsible for overseeing testing of the proponents' systems.

During the time period that the Planning Subcommittee was developing the test plans, the broadcast and cable industries were determining how to fund the tests, since the government had indicated it was unwilling to underwrite the tests. The broadcast industry, through the networks and trade associations, agreed to create and fund the Advanced Television Test Center (ATTC). They were later joined by the Electronics Industries Association. The cable industry considered joining the ATTC but decided instead, through CableLabs, to fund the cable portion of the tests.

After a lot of discussion, CableLabs and the ATTC entered into an agreement whereby CableLabs would perform the cable portion of the tests at the ATTC and pay the ATTC for lab and office space and access to common equipment. This decision helped reduce the costs for the proponents and ensure that the same system was tested for both broadcast and cable performance.

At the time of the agreement there were two sets of test procedures, a set written for over-the-air transmission and a set written for cable transmission. These two documents were reviewed by CableLabs and the ATTC to eliminate duplicate tests. The final decision called for the ATTC to perform the objective, "broadcast-only" and the joint tests, while CableLabs would perform the "cable-only" tests.

The objective tests are designed to determine the basic system parameters of the proponent systems. These tests include horizontal, vertical, and diagonal resolution, both static and dynamic, and in both the luminance and chrominance channels.

The remaining tests considered the proponent system's ability to operate in the terrestrial and cable environments with the various impairments that are present. Some of the impairments considered for terrestrial transmission were co-channel, adjacent channel and UHF taboo situations, discrete carrier interference, interference to and from NTSC signals, carrier-to-noise performance and multipath considerations.

These are subjective tests which require input from observers looking for the presence of interference in the picture. They are performed in two parts. The first part utilizes expert observers to determine the point at which impairments just become visible. That point is determined by varying the impairment level above and below the threshold point while observers "vote" on whether or not the impairment is visible. The observers then determine the level of impairment that makes the picture unwatchable plus a number of intermediate impairment levels. Video tapes are produced to show the impairment at randomly selected levels in comparison with unimpaired pictures. Non-expert viewers observe the pictures and indicate how annoying they considered a given impairment level to be.

While all "over-the-air" transmission impairments are of interest to the cable industry some are of greater importance and were proposed by the Advisory Committee as cable tests. These include carrier-to-noise performance, discrete carrier interference, and multipath. After the test labs reviewed the requirements of the tests it was decided the tests did not need to be performed twice and the one set of tests would be performed by the ATTC with CableLabs observing.

Cable-Only Tests

The remaining tests, to be performed by CableLabs, were designed to determine how well the proponent systems will operate in a typical cable system. The tests are composite second- and composite third-order intermodulation, the effect of multiple micro-reflections, high-level sweep interference, hum and low frequency noise, phase noise and residual FM. In addition, the proponent signal is passed through an AM fibre system and, if an ancillary data channel is present, the bit error rate in the presence of various impairments is determined.

Composite Triple Beat - Composite triple beat (CTB) is considered to be one of the major limiting factors on current cable television systems. The composite third-order test equipment was designed to test CTB alone and not CTB with other impairments. That requirement was satisfied by designing the computer-controlled test bed so that the CTB products falling in channel 12, the test channel, are generated without using a channel 12 carrier. The channel 12 product is selected by a narrow bandpass filter, amplified as necessary, then combined with the channel 12 ATV signal. The level of the impairment, relative to the desired signal, is varied while the expert viewers observe a received ATV picture.

Four sources of CTB can be used for the test. Two types, unmodulated NTSC carriers or carriers modulated with an NTSC signal, are the normal sources of CTB for the tests. Some of the proponents had proposed non-standard carriers for their ATV systems in order to minimize interference to existing services. The effect of these non-standard carriers had to be tested. The test bed was designed with two modes for testing the CTB generated by the ATV signal. The first mode tested the effect when half the signals were NTSC and half were ATV, a situation that will occur some time after the introduction of ATV services. The second mode used all ATV signals to generate the CTB product, representing a future situation when NTSC is no longer broadcast.

<u>Composite Second-Order</u> - Composite second-order (CSO) interference is generated in the same manner as the third-order products and similar tests are performed to determine the threshold levels for various combinations of NTSC and ATV signals. The CSO test was added to the procedures to account for the second-order distortion of current lasers being used in fibre optic links.

<u>Summation Sweep</u> - Two types of high-level sweep signals are tested. The first type is the typical sweep signal operating at a higher than visual carrier level and swept across the band periodically. The second type of sweep tested uses short bursts of carriers located lower in level than the visual carriers and at frequencies selected to produce the minimum amount of interference. The proposed ATV signals make extensive use of digital compression and, in some instances, digital transmission techniques. The impact of a high-level signal sweeping through the band may have a very detrimental effect on some systems and but little effect on other systems. It is desirable to have a transmission system which is not affected by the sweep signal.

<u>Hum and Low Frequency Noise</u> - Long cascades of amplifiers, poorly designed systems or poorly regulated power supplies can amplitude modulate TV signals with power line frequency signals. NTSC sets are designed to be very tolerant of hum modulation and only show the effect when it exceeds a few percent. ATV signals will be subjected to the same environment and must be capable of tolerating this interference at least as well as the NTSC signal. Low frequency noise can be created in the switching power supplies in common use on cable systems. The test equipment amplitude-modulates the ATV signal by passing it through a voltage-controlled attenuator. This attenuator is driven by a 120 Hz source or a low frequency noise source.

<u>Phase Noise</u> - Phase noise is the result of slight instability in oscillators used to heterodyne signals both at the cable headend and in the subscriber's home. Frequency synthesizers have been a significant source of phase modulation and their use has been increasing in recent years. The ATV receiver must be capable of receiving and decoding a signal with phase noise present. In test, the phase noise is introduced by phase modulating the local oscillator used in the ATV signal upconverter.

<u>Residual Frequency Modulation</u> - Residual FM has been be introduced by power supply ripple in a frequency synthesizer. Only a

small amount of ripple was necessary to introduce a significant amount of frequency modulation in the output signal. The residual FM signal is introduced into the ATV signal by frequency modulating the local oscillator with a 120 Hz sine wave signal. The maximum modulation possible is 99 kHz and, if desired phase modulation could be introduced at the same time to observe the combined effect.

Channel Change Time - The length of time necessary to change channels and produce a picture on the next channel has been a major concern to cable operators. The wide availability of remote controls has produced consumers who demand the ability to quickly flip through channels to find a program of interest. If it were necessary to pause at a channel for any length of time to determine the program, the consumer might become very frustrated. Many of the ATV proponents are proposing very complex compression schemes which could possible take an excessive amount of time to tune the channel, decode it, and produce a picture. The test is performed by switching channels and determining the length of time required to produce a picture.

Multiple Micro-reflections - Cable distribution systems by necessity, are composed of many active and passive devices located at fairly close spacings. The input and output matches of the devices plus the return loss of the cable itself are never perfect and produce many low level reflections. The effect of these reflections is sometimes a slight smearing of the NTSC picture. The reflections are normally not long enough to produce an visible ghost. The effect of the micro-reflections could be the production of a visible ghost if a time compression technique is used, or the effect could be the creation of sufficient group delay in the band to upset a digital signal via inter-symbol interference. The test bed includes a sample distribution system which consists of twelve multitaps spaced 60 feet apart and includes two line extenders. The taps have drops attached with lengths varying randomly from 50 to 150 feet and which are terminated in either a short or a three dB pad and a short. The ATV signals are passed through the distribution system and recorded. Non-expert viewers compare the pictures which are passed through the distribution plant with reference pictures and indicate any differences observed.

Fiber Optic - Many cable operators have been introducing fiber optic transmission paths into their plant. It is considered to be imperative that the ATV signal pass through the AM fiber system without degradation. In test, the ATV signal is passed through a 20 km AM fiber system. Expert observers compared the received signals with reference signals to determine if there is any signal degradation. If any degradation is observed the signals are rated by non-experts to determine the amount of degradation the average observer notices.

Bit-Error Rate - Ancillary data channels, proposed by some of the proponents, could be used to transmit decoding information, text, or other data services. The data channel should be designed to be capable of operation to the point that the video channel became unusable. In test, the bit-error rate is observed while impairments such as CTB, CSO, Phase Noise and Residual FM are increased.

## System Selection

At the conclusion of the tests, the results of all the proponents will be compared to determine the best compromise between picture and audio quality and the ability to operate in the presence of the various impairments, both on cable and over the air. The best system will be recommended for field tests, scheduled to take place in Charlotte, NC. The second-best system will be identified as a backup in case of the failure of the first system to succeed in field testing.

Based on the output of the lab and the field tests, the Advisory Committee will recommend an ATV standard to the FCC. After reviewing the recommendation the FCC will begin the rule-making process necessary to adopt an ATV standard. Expected decision date for the standard is late Spring 1993.

# CABLELABS' GHOST CANCELLER TESTING PROJECT

**Tom Williams**
**Cable Television Laboratories, Inc.**

*Abstract*

*Digital technology has progressed to the point where it is possible to remove ghosts from television pictures by using baseband digital filters. However for the digital filters to work, they must be programmed with the correct set of coefficients. The best way to determine coefficients is to put a training signal in the vertical interval and analyze the received training signal. The Advanced Television Systems Committee (ATSC) is sponsoring an effort to select a voluntary standard training signal. The committee work is being carried out by the ATSC T3,S5 specialist group. The effort consists of off-air field tests by the National Association of Broadcasters (NAB), lab tests and computer simulation by the Canadian Research Center (CRC), and cable lab and field tests by CableLabs. The BTA ghost cancelling system is currently in use in Japan. This paper discusses the CableLabs tests and results.*

The most generally agreed upon characteristics that determine a good training signal are highest possible energy within the time and power constraints of NTSC transmission, flat frequency spectrum, an impulsive autocorrelation function with near zero residual correlation for all time displacements, ease of extraction, immunity from non-linear impairments and small VBI usage. It is important to note that ghost cancellers are designed to fix linear distortions only.

The whole idea of testing hardware to pick a training signal is somewhat tenuous because the proponents' equipment was not all equal in terms of development, and it is possible for a better designed training signal to perform worse in the testing process because of hardware implementation. On the other hand, it is impossible for a system to deghost better than the training signal design allows. The CRC has performed computer simulations on three of the five proponents training signals, and their effort provides valuable information to the committee on the theoretical limits of the systems.

## Proponent Description

Five proponents have proposed training signals and built hardware which was evaluated. The waveforms are illustrated in Fig. 1, and are incomplete samples. Typically the waveforms change in a multifield sequence to provide more functions such as DC level reconstruction and noise averaging. The proponents are:

1. ATT/Zenith (ATT) with a pseudo-noise (PN) sequence
2. BTA of Japan(BTA) with an integrated sinx/x pulse
3. Philips (PHI) with a modulated frequency sweep or chirp.
4. Samsung (SAM) with a PN sequence and
5. David Sarnoff Research Center (DSL), also with a PN sequence.

The ATT/Zenith unit was a prototype in the early stages of development. It occupied about of foot of rack space, and had a manual video AGC knob that needed to be adjusted. It produced a number of artifacts, which were presumably related to the development stage it was in. It did incorporate an interesting optional feature: The unit could put on line 12 the solution it had found for the impulse response of the channel. The BTA unit that we were supposed to test failed during the interface week, and a commercial Toshiba deghoster was substituted. This unit was a fully developed commercial product designed to sit on a TV, but had been modified to accept a baseband video input. The unit took a while to deghost, but included a front panel

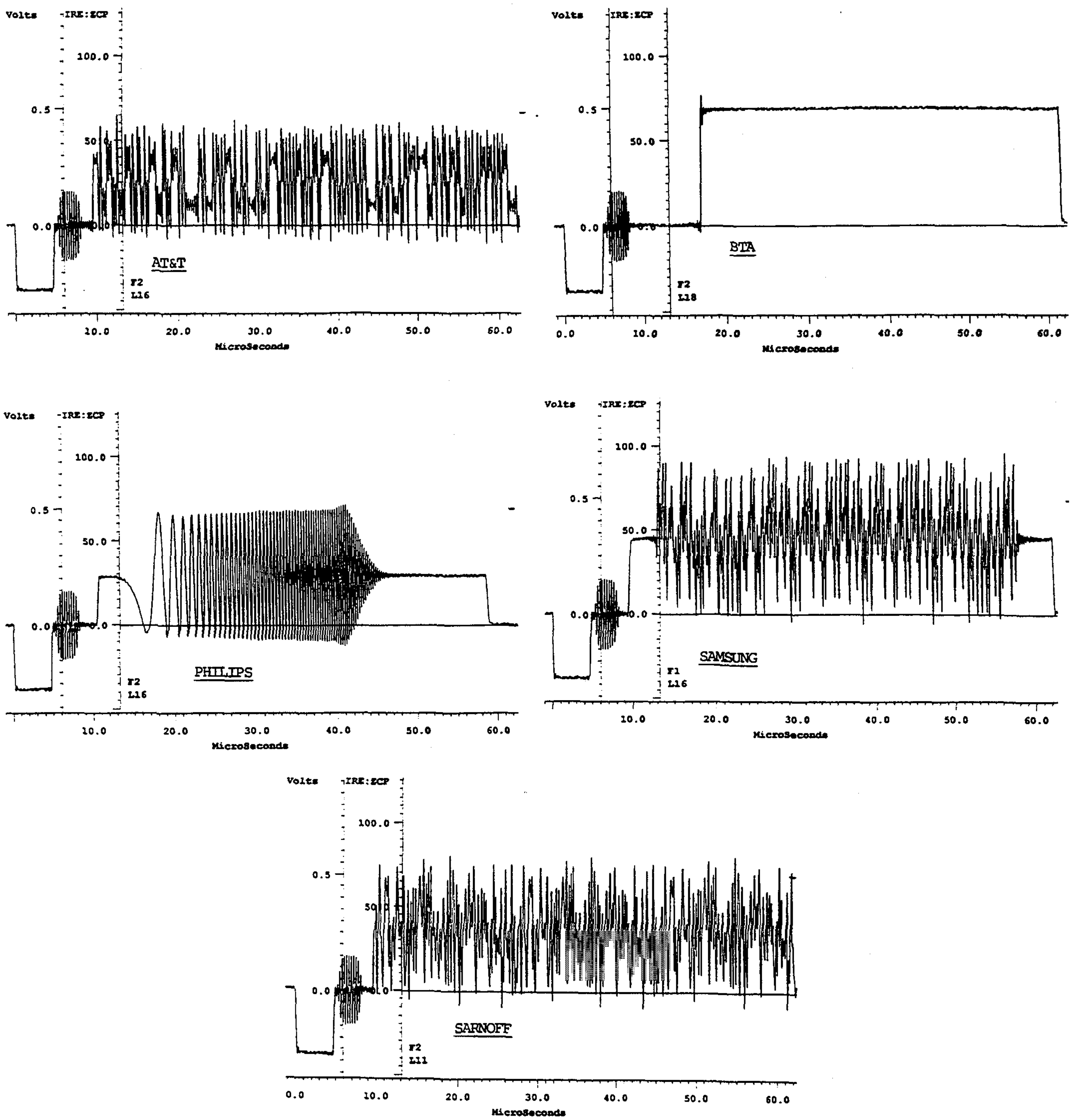



**Figure 1. Proponent Training Signals for Ghost Cancellers**

display to indicate when it was done. The Philips unit was a prototype in an advanced stage of development, and was also of a size and configuration to sit on a TV. In operation, it computed a whole new set of coefficients every few seconds, and did not appear to do a sliding average. The Samsung unit was another unit in the early stages of development. It stood in a rack about 5 feet high and included a DOS computer. It did not deghost the picture continually, but only when commanded by the computer. The David Sarnoff Research Center unit was a relative of their ACTV enhanced definition ATV system and stood in a rack about 3 feet tall. The unit deghosted continually and used a sliding average deghosting solution. The Sarnoff unit was unique in that it was the only one with no ability to freeze the correction (at lease during the CableLabs and NAB test periods), and it required a full bandwidth quadrature channel.

## Testing

The testing that CableLabs did was performed in the Washington DC area by CableLabs personnel during November and December of 1991 in association with the ATSC and National Association of Broadcasters (NAB). We used the same van and some of the same equipment used by the NAB in their field test phase, which took place in September and October. CableLabs testing consisted of two phases, a laboratory phase and a field phase.

For both the lab phase and the field phase, the 5 ghost training signals were all programmed into the vertical intervals of three identical TEK1910 video generators. The video generators were used only for vertical interval insertion. For the cable field tests, two of the training signal

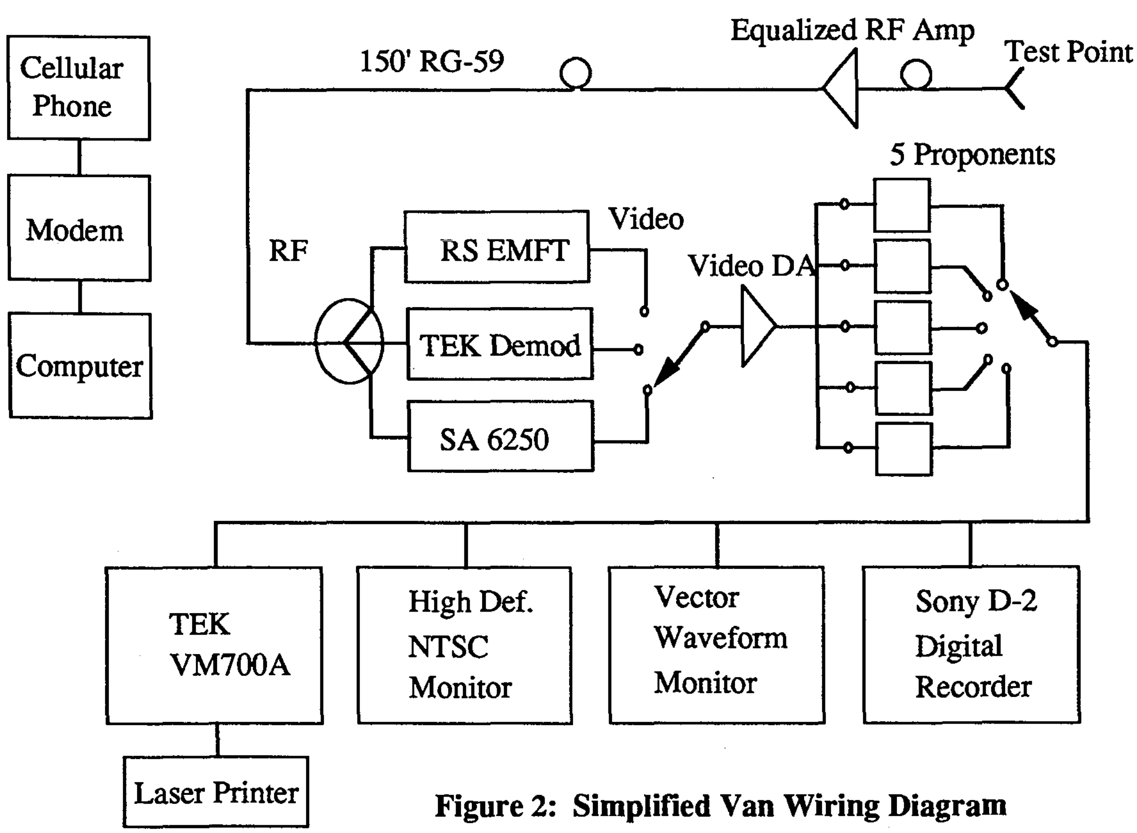



**Figure 2: Simplified Van Wiring Diagram**

generators were used on the off-air channels 4 and 20. The other training signal generator was located in the headend and put on an unused channel with a video pattern that allowed easy identification of ghosts and other impairments (such as loss of resolution).

In the test van, all 5 proponent's deghosting equipment was loaded along with our test gear. Figure 2 shows the wiring diagram of the van. A cellular phone was used with a modem to communicate between the mobile test van and the 1910 generators for purposes of changing proponents signals. The main demod used whenever possible was a TEK1450-1 because Sarnoff Labs needed a full bandwidth quadrature channel. The Rohde & Schwartz EMFT demod was used as a backup unit to the Tek demod for tests where phase instability was encountered, because it is more tolerant of phase noise. The synchronous SA6250 was used for one specific test because it represents cable gear typically used in headends. Various pieces of video analysis and recording gear were used to analyze and capture the data.

Digital recording was done on a Sony portable D2 composite video tape recorder and waveforms were recorded from and analyzed by a Tek VM700A waveform analyzer driving a laser printer.

The lab portion of the tests were performed by backing the test van up to the CableLabs test bed in Alexandria, Va., and running an impaired channel 12 carrier out to the van, as shown in Fig. 3. The lab tests lasted about a week and consisted of tests with ghosts, as well as ghosts with additional impairments such as Gaussian noise, composite triple beat and residual FM plus phase noise. Tests were also done with negative traps and with different demodulators and set top converters to check compatibility.

Figure 4 is a diagram of the Cable field tests. The field tests were performed on four different cable systems, Jones in Alexandria, District Cablevision in Washington DC, Cable TV Montgomery , and Media General of Fairfax. The general procedure was to install our equip-

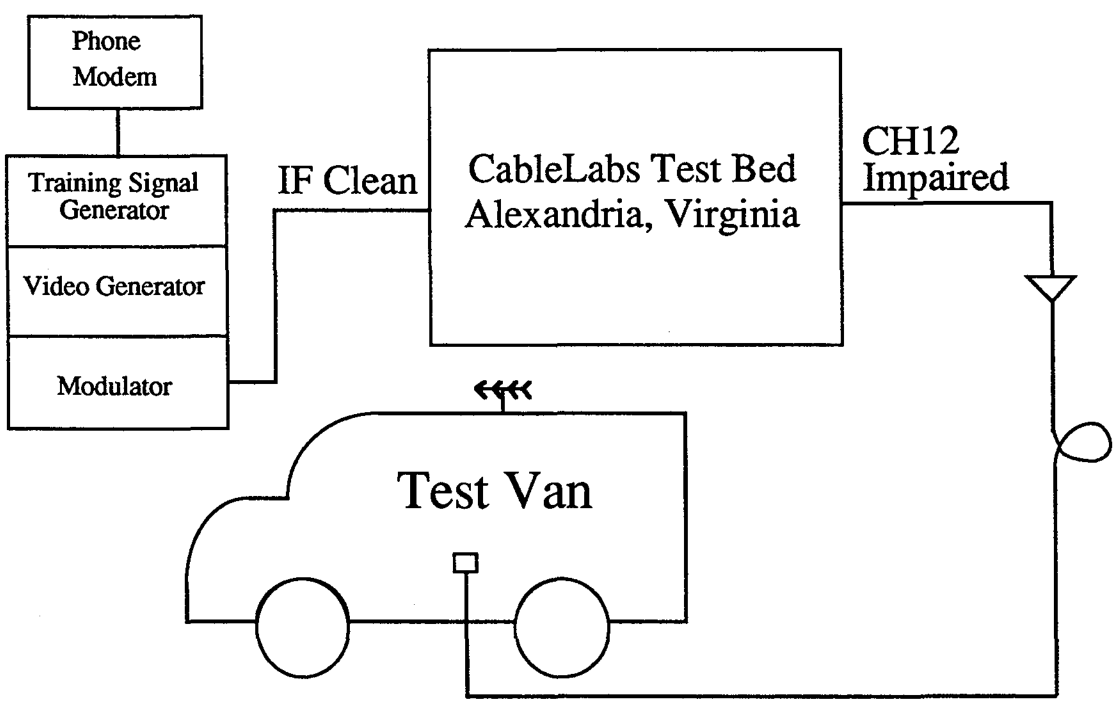



**Figure 3: Lab Test Diagram**

ment into the headend on an unused channel, characterize the location for both off air and the locally originated channel, and then test the proponent's equipment. Then the van was taken out into the field to 1 AML hub site, one fiber hub site, 12 taps, and 12 individual homes. Typically, the channels were characterized, and then each of the ghost cancellers were allowed to correct the channels one at a time. The tests that were actually performed at each location varied, and were chosen to reveal system compatibility problems as well as to identify performance differences between the proponents. The video and waveforms of the corrected video were digitally recorded.

Additionally, tests were done with each of the proponents on selected RF and baseband converters. The concern with the RF set top converters was that the phase noise and residual FM from the up-down converter would cause problems with the true synchronous demods that are necessary to demodulate video for ghost cancelling. The concern with the baseband converters is that they use non-synchronous demodulators, and these may cause poor deghoster performance. Although the handful of set tops tested did perform relatively well, more investigation of representative samples of the overall population converter boxes is merited.

## Results

Figure 5 summarizes the results of the lab test using the magnitude frequency response ripple to 3.58MHz as the vertical axis. The baseband

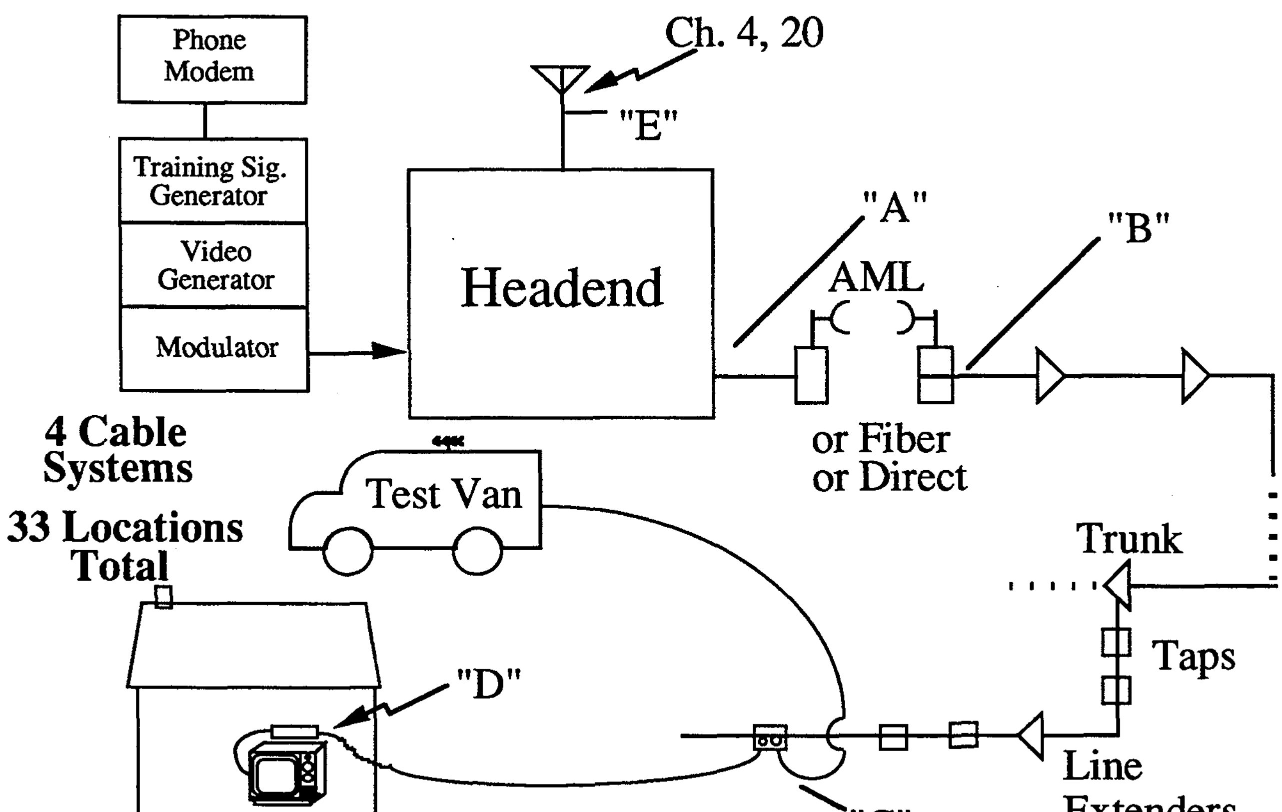
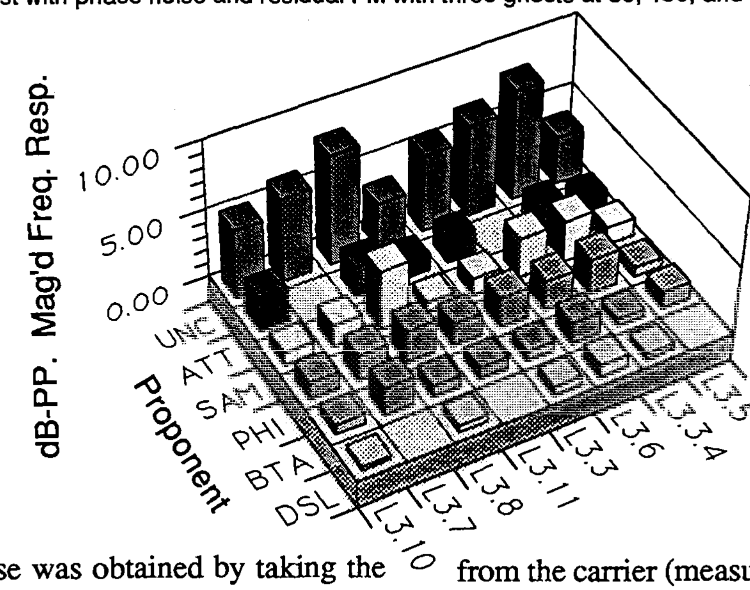



**Figure 4: Field Test Diagram**

Fig. 5 Peak-to-peak magnitude frequency response versus proponent versus lab test number.

| | |
|---|---|
| L3.10 | Test with traps at channels 11 and 13 |
| L3.7 | Test with SA8580 RF converter and ghosts at 150, 600, and 2500ns, each -15dBc. |
| L3.8 | Test with -40dB CTB and three ghosts at 300, 600, and 1250ns. |
| L3.11 | Test with SA6250 synchronous demod and three ghosts at 80, 150, and 1250ns. |
| L3.3 | Test with three ghosts at 40, 150, and 2500ns. |
| L3.6 | Test with SA8590 baseband converter and three ghosts at 150, 600, and 2500ns. |
| L3.3.4 | Test with three ghosts at 300, 600, and 1250ns. |
| L3.5 | Test with phase noise and residual FM with three ghosts at 80, 150, and 2500ns. |




frequency response was obtained by taking the fast fourier transform (FFT) of the sinx/x waveform in the vertical interval. The corrected frequency response ripple does not take into account the fact that a long ghost is more noticeable than a short ghost, but it provides, along with K factor, a good indication of how the ghost canceller is functioning as both a ghost canceller and channel equalizer. The uncorrected data was labeled "UNC" and placed as the last set of bars in the graph. In the case of ATT the missing blocks are due the equipment being out of service. Most of the missing data on DSL is due to the fact that a demodulator with a broadband quadrature output, that would withstand large amounts of phase noise, was not available. Two test were done with ghosts only. One test used ghosts at 300, 600, and 1250ns. and the other used ghosts at 40, 150, and 2500ns. The levels of the ghosts were all 15dB below the carrier at random phases. The residual FM and phase noise tests were done with 120Hz. FM at 8kHz. deviation, combined with a phase noise of -82dB below the carrier at +/- 20kHz from the carrier (measured in a 1Hz bandwidth). Set top converter tests were done with a RF heterodyne converter and with a baseband unit borrowed from Jones Intercable. The Rohde & Schwartz EMFT demod had to be used to demodulate the signal from the RF set-top converter because of phase noise and residual FM. A test was also done with ghosts using the SA6250 demod. Again, the lack of a quadrature channel meant that Sarnoff could not be tested. The test with traps was done to see if the roll-off caused by adjacent traps could be corrected by the ghost cancellers.

Figure 6 shows the frequency response of the received off-air channels along with deghosted response. Most of the data was taken using headends, but location 16 data was taken using the van antenna. Although there were not any long, large prominent ghosts encountered in the headends we visited, we observed short ghosts and equalization problems as a result of ghosts.

Fig. 6 Peak-to-peak magnitude frequency response versus proponent versus location and off-air channel.

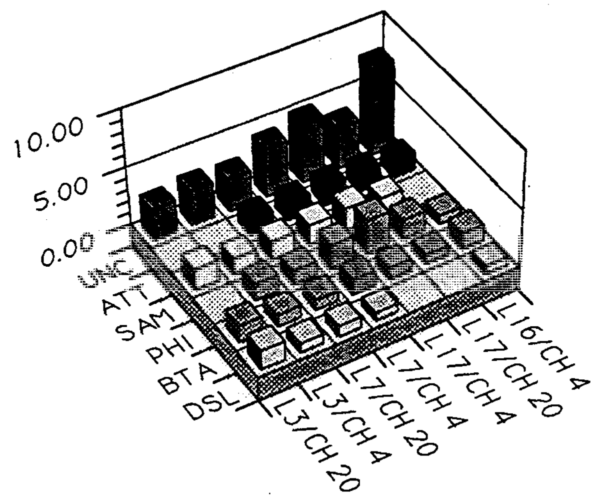



Table 1

| Proponent | ATT | BTA | PHI | SAM | DSL |
|---|---|---|---|---|---|
| Carr. to Noise | 35dB | 40dB | 30dB | 35dB | 25dB |

Table 1 summarizes how the ghost cancellers performed with noise. The test was performed by increasing the noise level in 5dB increments and noting the last place that a proponent's equipment was seen working. The observation that should be made from this data is that the BTA system fails to cancel ghosts at a higher carrier to noise level, and this is expected because their differentiated training signal is a low energy one. The differences between the other proponents is more than can be explained away by a theoretical analysis of the energy in their waveforms, and was probably due to differences in the amount of averaging done to remove noise, and differences in the performance of individual sync separator circuits. The proponents have been nondescript of their system implementation technical details, so we can only surmise the reasons for some of the test results.

The tests indicate that these ghost cancelling devices work well and will produce dramatic improvements in the quality of the off-air broadcast signals. Cable systems should deghost off-air signals for their subscribers and can additionally flatten signals that only need equalization. Additionally, TV sets with built in deghosters should work without artifacts on cable. This means that the cable industry needs to work with the consumer electronics manufacturers to insure that the necessary features such as a phase noise tolerant demod, and an "OFF" switch are built in. Another unresolved issue is the ability to track ghosts that are time varying. In heavy wind loading conditions, transmit and receive towers sway, and this means that the ghost canceller must be very fast to track movement. The higher the UHF channel, the worse the problem will appear to be. Training signal acquisition time and computation time must both be figured evaluated.

## Conclusions

It appears that ghost cancellers can soon be in the headend, and should provide us with a valuable tool to improve the quality of off-air channels. Many thanks to the systems that provided equipment, personnel, and time to our tests, and to the CableLabs technical people who supported the testing and analysis effort.

# CATV POTENTIAL OF EXTERNALLY MODULATED LASERS

Larry C. Brown
Magnavox CATV Systems, Inc.

*Abstract*

*In the last few years, DFB lasers have gained widespread use in CATV as the foundation for AM optical transmitter design. More recently, an important new generation of transmitters has emerged. These devices offer much higher optical launch power and spectral purity than is possible from a DFB design. This paper compares these externally modulated transmitters with their DFB predecessors, and explores potential benefits of the technology to CATV.*

## DIRECT VS. EXTERNAL MODULATION

Direct modulation of a Distributed Feedback laser (DFB) involves summing a bias current with a modulating signal, as shown in Figure 1. In AM CATV applications, the modulating signal is an entire broadband rf spectrum of television channels and other rf carriers spanning a bandwidth of 50-550 MHz or more. The result is an optical "carrier" frequency (wavelength), intensity (amplitude) modulated with the analog composite CATV spectrum.

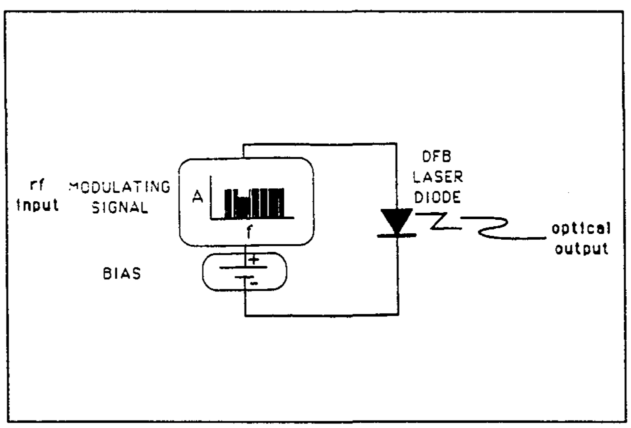



*Figure 1.*
*Direct Modulation Process*

In contrast to the DFB optical transmitter, the External Modulation transmitter employs two separate components to create an intensity modulated optical carrier: a solid state laser and an external modulator. The process is pictured in Figure 2.

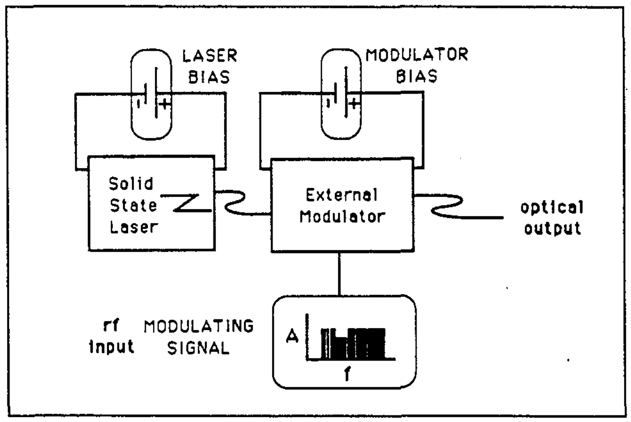



*Figure 2.*
*External Modulation Process*

The laser is d.c. biased, and thus emits a c.w. optical carrier having extremely stable amplitude and wavelength. To date, solid state lasers of Nd:YAG composition have been most common.

The YAG's output is coupled via a polarization-maintaining single mode fiber to the input of an external modulator. To date, interferometer designs, such as the Mach-Zehnder modulator, have been most common. Optical carrier energy is split at the input of the modulator, propagated through two parallel paths, and finally recombined. The modulating signal is added to a modulator bias voltage, then fed to conductive strips parallel to the light paths. The properties of the lithium niobate material employed in the modulator cause complementary changes in propagation time to occur along the two light paths as the modulating signal is applied to the conductive strips. Interferometric recombination of optical signals at the output of the modulator naturally creates two optical outputs, each intensity modulated with the input rf spectrum.

The transfer characteristics of a DFB laser diode vary significantly from piece to piece, and in today's lasers is quite linear. In contrast, the transfer characteristic of a YAG laser feeding an interferometer external modulator is quite nonlinear, but very predictably so. The process of "straightening" the YAG/interferometer transfer characteristic is commonly called "linearization." Several techniques of linearization are possible and proven, but their discussion is beyond the scope of this paper.

Typical DFB laser transmitter optical launch powers available today are in the 4-8 mW range. Meanwhile, typical external modulation transmitter optical launch powers are in the 8 mW - 15 mW range, or as much as four times that of the DFB. The additional power translates directly to a longer reacher for the external modulation transmitter. Thus, the same noise and distortion performance which a DFB can offer on an 11 dB optical link may be achievable on a 16 dB optical link with external modulation. The 5 dB difference translates to more than seven miles of additional reach for the externally modulated transmitter.

The spectral purity of the externally modulated transmitter's output is also beneficial to link performance. Today's DFB lasers are susceptible to Interferometric Intensity Noise (IIN).[1] A portion of the light travelling downstream toward the node is reflected backward, then reflected forward again. IIN is caused by both connectors included in the link and Rayleigh backscattering effects. The effect can degrade DFB AM link CNR performance 1-2 dB or more.

The externally modulated transmitter, however, has a much narrower linewidth than the DFB. This causes links using the external modulation transmitter to be immune to IIN degradation.[2]

Another benefit of the external modulation transmitter is its naturally superior second order distortion characteristics compared with a DFB. The transfer characteristic of the DFB usually makes the DFB link second order distortion limited, i.e., maximum transmitter drive level becomes limited by link CSO performance rather than CTB. This makes it more difficult to employ DFB lasers with high channel loadings such as 550 MHz (77 channels). On the other hand, the transfer characteristic of the externally modulated transmitter is symmetrical about its bias point. The symmetry gives the externally modulated transmitter inherently superior CSO performance under conditions of high channel loading, much as the symmetry in a push-pull rf hybrid amplifier produces the same result.

Inherent noise generation in the YAG laser relative to the DFB is another issue where the YAG excels. The Relative Intensity Noise (RIN) of a YAG laser typically runs -170 dB/Hz, compared with a more typical -155 dB/Hz from a DFB. This can result in substantially superior CNR obtainable on an externally modulated link vs. a DFB.

In summary, higher launch power, IIN immunity, much better CSO and lower RIN are all benefits provided by external modulation.

## EXTERNAL MODULATION COST

External modulation is naturally a more involved process than direct modulation of a DFB. Accordingly, for applications of only a few links of under 10 km each, the DFB may remain a more economical solution for some time. However, for links which are simply too long for a DFB to reach with acceptable performance, or in cases where many links emanate from one optical transmitter site, the external modulation transmitter can provide savings of 20% to 70% per link, depending upon the application.

Without external modulation AM, the only long-link alternative may be FM or digital transmission, inherently expensive because it requires channel by channel processing. The additional 5 dB of reach offered by an external modulation transmitter over its DFB alternative can translate to a user savings of $250,000 or more per optical link in these applications.

Figure 3 shows the economic benefits of splitting each output of a typical externally modulated transmitter. In this application, the transmitter's high launch power is used to reach many nodes with only one transmitter.

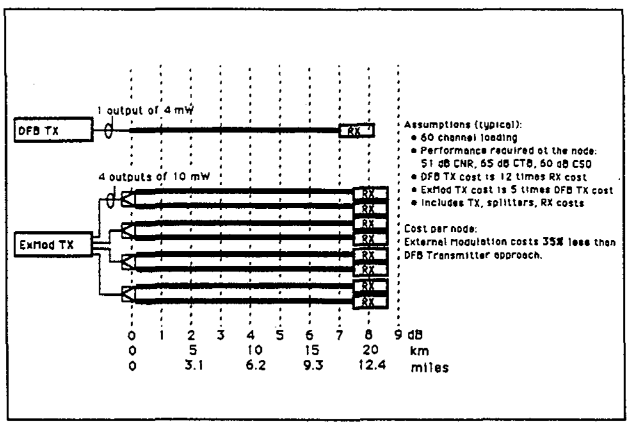



*Figure 3.*
*External Modulation Economic Benefits*

The illustration shows one external modulation transmitter doing the work of 8 DFBs. Noise and distortion performance at the nodes is the same in both external modulation and DFB scenarios.

A heavy weighting of cost in the transmitter (vs. receiver or node) side of the optical link is inherent in all CATV optical transmission. This factor causes the cost PER NODE for the DFB solution to be much greater than its external modulation equivalent ... about 55% more in this example!

## EXTERNAL MODULATION RELIABILITY

Reliability of an opto-electronic transmission network is naturally a function of the total number of components involved in the network and the individual reliability of each component.

From Figure 3 the application where an external modulation transmitter is split to serve many nodes it is obvious that the component count inside one external modulation transmitter will be much smaller than the total component count of 8 DFB transmitters which would be required to provide equivalent performance. The same would obviously be true in the application where external modulation is used as an alternative to FM or digital transmission, inasmuch as the need for individual channel processing is eliminated with the external modulation transmitter. Thus, from a pure component count viewpoint, the external modulation approach naturally enjoys a much smaller probability of random component failure than alternate optical transmission technologies.

All the components used in an external modulation transmitter are similar in nature to those found in the DFB, with the exception of the solid state laser and external modulator. Therefore, given a consistent quality of manufacturing, examination of reliability of these two components should be sufficient to understand any reliability difference which an external modulation transmitter may exhibit compared to a DFB.

The Mach-Zehnder interferometer typically used as an external modulator is a solid state design using a lithium niobate substrate as a foundation. Lithium niobate is the same material used in the manufacture of a myriad of surface acoustic wave (SAW) filters, a technology in widespread use in both industrial and consumer electronics for decades. Substantial work has been done on both material and device reliability assessment of lithium niobate optical guided wave devices, concluding specific package designs can result in a highly reliable product. A solid mechanical design, similar to that used within a DFB laser package, provides a stable coupling between modulator optics and optical fibers at the modulator input and output. Therefore, the external modulator itself can reasonably be expected to have a reliability comparable with today's DFB lasers.

The solid state Nd:YAG laser typically employed as the light source is structured as shown in Figure 4. A typical "pump" laser diode launches a relatively high c.w. optical power at about 808 nm onto the end of a Neodymium doped YAG crystal rod. This causes the rod to lase and emit a c.w. optical carrier at 1319 nm. It is this optical carrier that is output through an optical fiber to the external modulator.

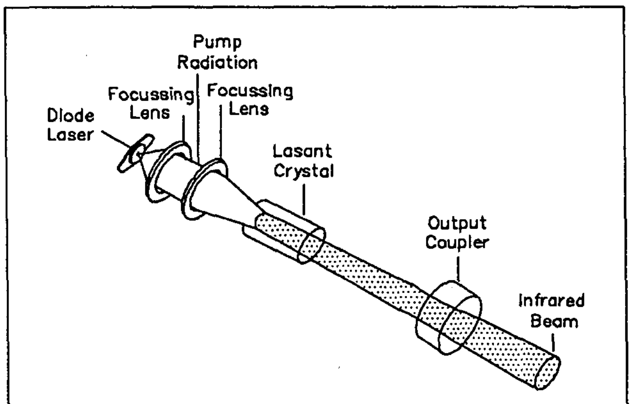



*Figure 4.*
*Nd:YAG Laser Structure*

Commercial results to date indicate the pump diode can continuously emit an optical power of 1 watt and still provide a mean time to failure for the external modulation transmitter which approaches today's DFB transmitters' MTTFs.

Keys to achieving this performance include a multi-stripe array laser structure as shown in Figure 5. It is not the total power at the pump facet that threatens lifetime, but the power DENSITY at the facet. The stripe design keeps this optical power density in an acceptable operating range.

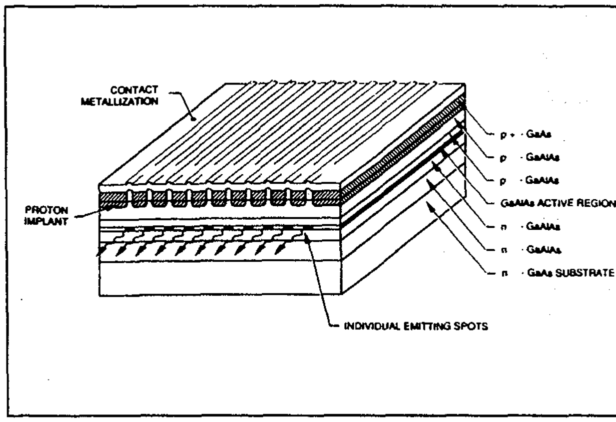



*Figure 5.*
*Pump Laser Diode Structure*

Another important life-extending technique is to lower the temperature of the pump facet by using a thermoelectric cooler. Lowering the facet temperature by 10C can double pump life. Thermal management provided by generous heat sinking and a pump package design optimized for c.w. operation can also extend product life.

YAG lasers have been utilized in medical electronics since the 1970's, and pumped YAGs since the '80's. Thus the technology itself is not new ... only its application to CATV optical transmission.

Accelerated life testing is commonly performed today by laser diode manufacturers by operating a sample lot of diodes at one or more temperatures substantially above their normal operating temperature until failure is observed, then attempting to extrapolate life expectancy at normal operating temperatures. But even the best diode manufacturers acknowledge major uncertainties in this process. The true MTTF of either DFB or external modulation transmitters will only be assuredly known after a quantity of both units have performed under normal operating conditions for many years. Since first commercial deployment of ANY AM CATV laser technology took place less than five years ago, no such history for either will be available for years to come. Nevertheless, no evidence has surfaced to date which would indicate either technology is not capable of producing an optical transmitter with an MTTF in excess of ten years.

## FIELD DEPLOYMENT RESULTS

Commercial deployment of external modulation transmitters employing an Nd:YAG solid state laser combined with a Mach-Zehnder Lithium Niobate modulator in a CATV application began in the fall of 1991. The first year of deployment has seen tens of units installed at locations throughout North America and Europe. Field proofs of performance have confirmed the technical performance superiority of external modulation over the DFB alternative.

Initial installations are seeing a variety of applications. Many cases utilize the external modulation transmitter for the extended link i.e., a link which would alternately require expensive FM or digital transmission. Many of these situations are AML replacements, and hub or headend eliminations. A number of other cases are utilizing the high optical launch power for its splitability to serve many nodes in an FTF situation.

Some technical references have implied that optical launch powers into a single mode fiber in excess of even 9 mW would result in severe power falloff on links only a few tens of kilometers long due to the Stimulated Brillouin Scattering (SBS) effect[3] characteristic of the fiber. A significant field experience has been the absence of any detectable SBS effects, despite the relatively high launch power available from the external modulation transmitter. Further investigation has revealed that in each technical reference the laboratory research involved utilized a test method and/or equipment which was not typical of today's actual product implementation and field situation. As of this writing, optical powers of up to 20 mW have been successfully launched into an optical fiber installed in the field without any evidence of SBS effects, and evidence exists that substantially higher powers may be operable without SBS problems.

## CONCLUSIONS

In any CATV application, the external modulation transmitter's lower RIN, higher launch power, immunity to IIN, and better CSO performance combine to make this technology a superior choice to achieve the best possible technical performance.

From an economics viewpoint, external modulation can save 70% of the cost of the alternative FM or digital link for long reach applications (where DFB reach is simply inadequate). External modulation can reduce cost per node 40% or more below an alternative DFB approach in cases where many optical nodes are involved, such as fiber-to-the-feeder and fiber-to-the-curb system architectures.

There appears to be no technological characteristic which would prevent achievement of external modulation optical transmitter reliability comparable to or in excess of today's DFB transmitters.

The first year of field installations is encouraging and bodes well for the future of external modulation for AM CATV optical links. External modulation may prove to be to the DFB laser what push-pull and power doubling hybrids proved to be to the original single ended amplifiers used by our industry ... a natural successor technology offering improved performance and system economics over its predecessor.

## ACKNOWLEDGEMENTS

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# COEXISTENCE OF ANALOG AND DIGITAL VIDEO

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*Abstract*

*The coming introduction of compressed digital video delivery systems has caused many industry watchers to question the viability of alternate forms of signal security, including converter/descramblers and interdiction products. Some suppliers have backed off from plans to deliver new analog control systems to the industry, choosing instead to focus on digital delivery systems to support their business for the next several years.*

*This presentation will focus on the relationship between digital video and analog services. It will propose a scenario by which the two technologies may coexist on a network, making the new technology available to those subscribers willing to pay for new services, while keeping capital investment low for more cost-sensitive subs. The report will concentrate on the next five to ten years and how the new technology may be phased in over time instead of immediately dominating the market.*

## INTRODUCTION

Digital Video Compression for standard NTSC and High Definition Television (HDTV) has received a great deal of attention over the last few years. Subscriber service control and compatibility with consumer electronics are major issues to be considered for a new technology to be successful. As efforts progress towards making digital compression practical, issues such as compatibility with existing analog channels and consumer friendliness become ever more important. One of the major issues today with the analog converter is its inability to interface with consumer electronics like VCR and TV in a friendly manner. Since video compression is likely to result in some form of converter, we are looking at the possibility of compromising customer friendliness for more channels and more features. This article will propose a method for dealing with these issues.

## AVAILABLE TECHNOLOGIES FOR SUBSCRIBER SERVICE CONTROL

Today subscriber service control exists in several forms. All of them share the same delivery system using analog channels on the cable. Existing service control systems fall into one of two major categories, either IN Premise (mostly cable converters) or OFF PREMISE (mostly traps). In either case, service control is either manual or addressable. Addressable converter-based systems offer flexibility at the cost of consumer (electronics) friendliness; theft of service prevention; and (in)accessibility of converter at customer's residence. Trap-based systems offer friendliness at the cost of flexibility; most trap systems are not addressable and are not easily adaptable to cable system growth in number of channels or channel reassignments. Traps, however, are transparent to analog and digital video channels. Subscribers interested in receiving enhanced services may be provided digital de-compression converters to augment their existing analog service.

## NEW TECHNOLOGIES

Two new technologies that are receiving a great deal of attention in the recent days are the INTERDICTION and DIGITAL COMPRESSION SYSTEMS.

Interdiction systems are at a point where immediate market introduction is possible. Digital compression systems are at final stages of laboratory evaluation. At first sight interdiction and digital compression seem to be mutually exclusive technologies; however, a careful examination will show the merit of the combined technologies in offering the most flexible and consumer-friendly interface.

## INTERDICTION TECHNOLOGY

Interdiction technology is typically employed outside the subscriber's residence - either on the pole or on the side of the house. In either case service control is accomplished by using frequency agile jamming sources in combination with subscriber drop switch under headend control. Interdiction devices, when used on the pole as tap replacement devices, can act as "active taps" resulting in longer reach or better signal quality or both.[1] The "active tap" configuration also makes bandwidth extension of the cable systems easier to realize and improves operational reliability by reducing the number of cascaded amplifiers. Interdiction offers both consumer friendliness and operator friendliness. It is transparent to digital signal delivery. Figure 1 shows an implementation of an Interdiction System.

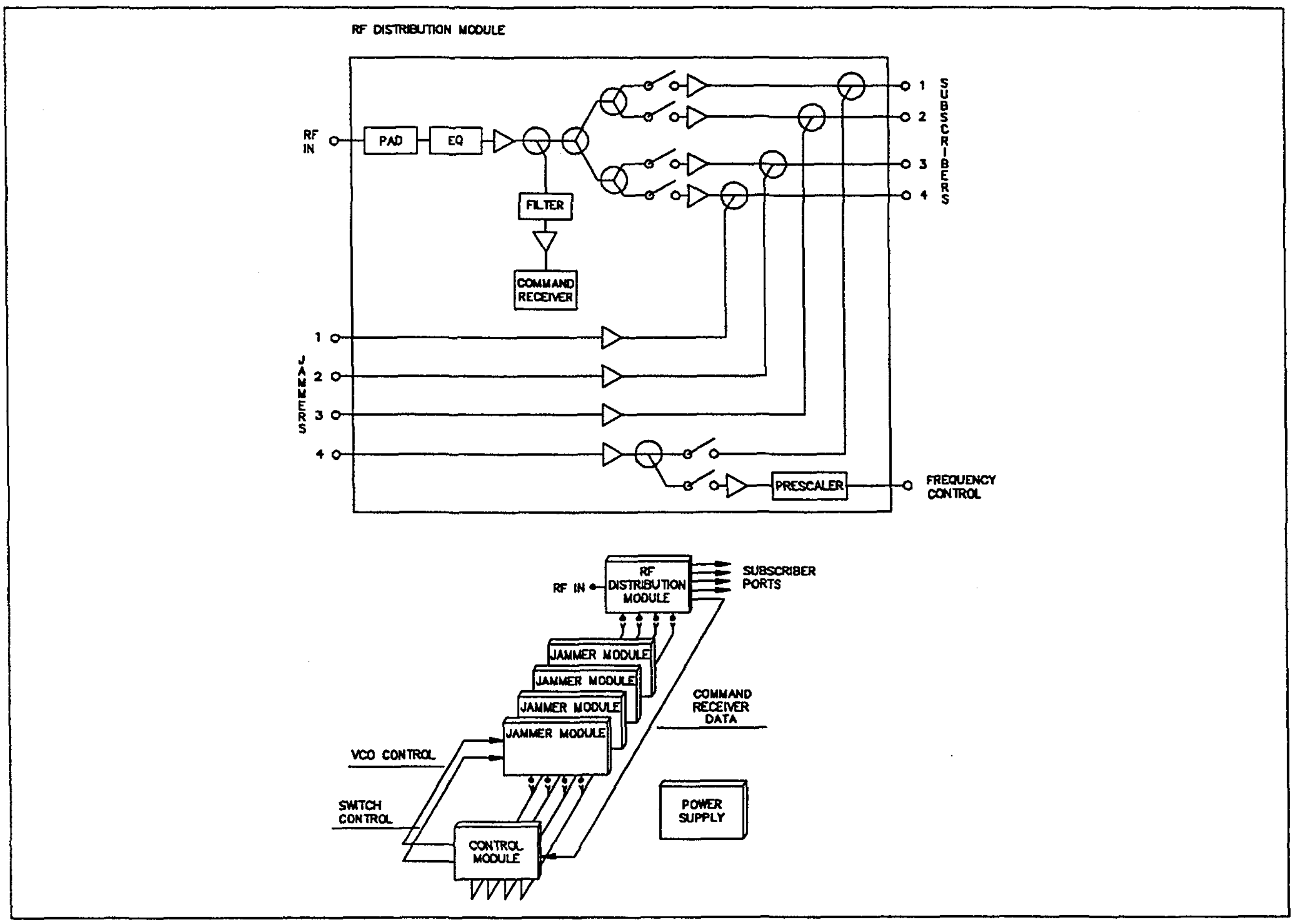


*Figure 1.*
*Interdiction System*

## DIGITAL COMPRESSION TECHNOLOGY

Digital compression is accomplished in two ways. Source coding reducing the amount of information required to describe a video sequence by reducing or even suppressing both temporal (picture-to-picture) and spatial (horizontal sample-to-sample and vertical line-to-line) redundancies. Channel coding involves adding error correcting codes and a digital-modulation subsystem (e.g., QAM).[2] The digital coding and modulation techniques offer, in addition to increased channel capacity, the ability to transmit clean images with a lower carrier to noise than is practical with analog delivery. This allows carrier levels to be lower, stretching the reach of the distribution system.

The spectral efficiency of digital compression will depend on the desired quality of the received signal. In any case, multiple channel delivery at a single NTSC channel space is possible.

## APPLICATION FOR COMPRESSION

For expanding the revenue stream, compression can offer expanded pay-per-view like near-video-on-demand; it will also offer HDTV signals when a commercial market develops for such signals.

Digital compression offers the ability to greatly increase the channel capacity of cable television systems. This provides an opportunity to add revenue-increasing services such as extend pay-per-view and near-video-on-demand. This will give system operators a chance to claim a portion of the revenues that are presently going to videotape rental stores.

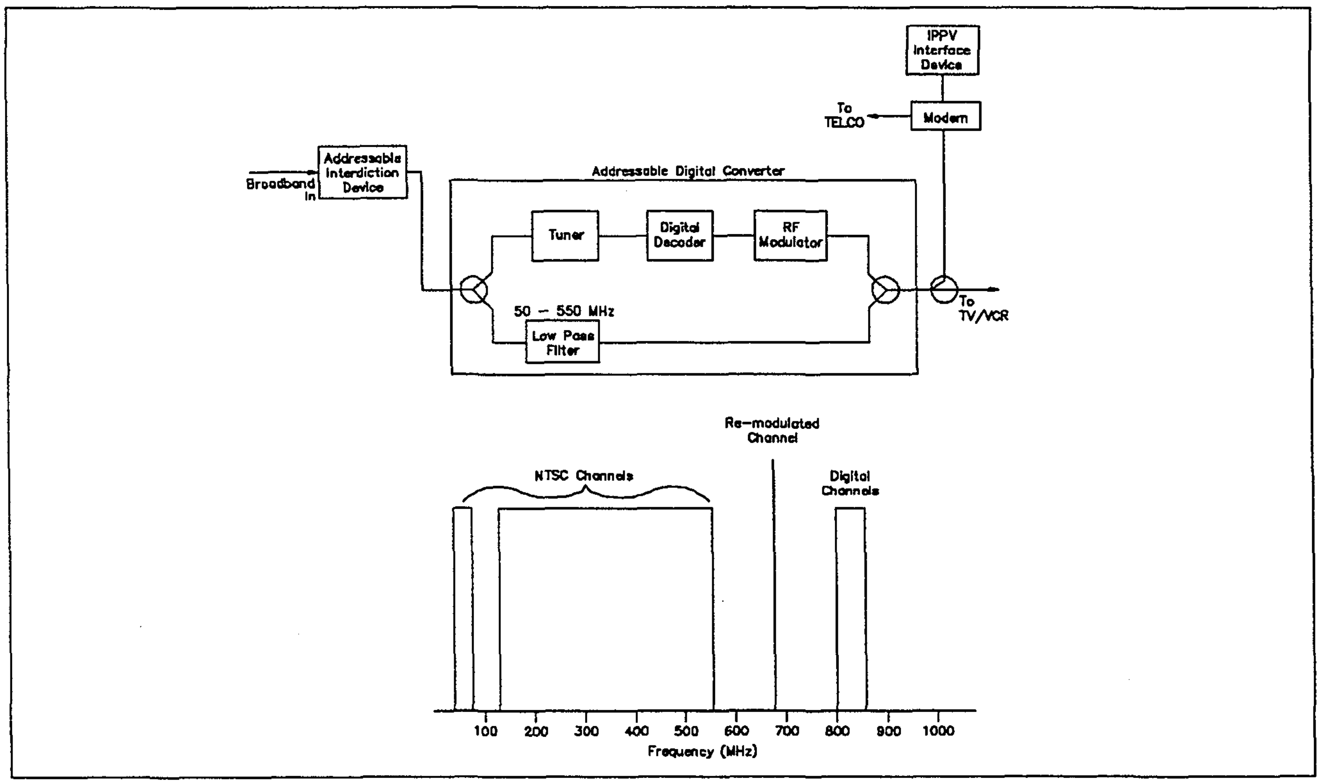


*Figure 2.*
*Concept Digital Converter: Block Diagram*

## WHAT COMPRESSION WON'T DO

In the foreseeable future, compression cannot replace most of the current cable channels that are NTSC compatible. This is due to the fact that the compression technology is not capable of displacing millions of NTSC compatible television receivers and VCRs that have already found their place in households and are continuing to have an excellent market.

## COEXISTING ANALOG AND DIGITAL VIDEO CHANNELS

With a clear picture on what compression can and cannot do, it is easy to see that cable will carry a significant number of NTSC compatible channels along with digitally compressed and HDTV signals. The proportion of NTSC to digital channels will depend on available cable bandwidth and compression yield along with other considerations such as the desired programming scenario. It is likely that in most cases the analog channels will remain between fifty and sixty channels while the number of digitally-compressed channels will keep increasing with increased bandwidth, penetration of near-video-on-demand and higher compression yields possible from technologies that continue to evolve. If we continue on the premise of consumer friendliness and near-video-on-demand using digital compression, then we have a scheme where the consumer continues to buy the analog channels that he is used to and adds to the revenue stream by purchasing programs using near-video-on-demand. If we treat near-video-on-demand as an event or a program and not a channel, we can arrive at the following scheme to provide consumer friendliness without compromising operator revenue and flexibility.

## THE "UN-CONVERTER" CONCEPT

Figure 2 shows a concept digital converter (the "un-converter") working with an interdiction unit to provide the customer friendliness. The off-premise interdiction unit provides control for analog NTSC channels and subscriber drop control. The digital converter consists of tuner-decoder combination to tune and descramble the desire digitally-compressed channel. After decompression, the baseband output of the converter is modulated to a channel at a frequency higher than any NTSC channel on the cable. The output frequency of the converter can overlap the digital channels as a low pass filter can be used to pass analog channels and remove the digital channels. An analog version of this type of converter was previously proposed.[4]

In its base form, the "un-converter" has no user controls. It is controlled completely through the addressable data stream received from the access control system. This allows the decoder to be removed from the set-top and placed out of sight, behind the set, in the basement, or even outside. It is even possible that eventually the decoder and the interdiction device will co-exist in a common enclosure.[3]

As the analog channels are controlled by the interdiction unit, the consumer electronic interface is handled by the subscriber's own cable-ready TV and VCR. This "un-converter" concept does not compromise the cable operator's flexibility and can add to his revenue stream. The subscriber always gets the channels he is authorized for in the clear and whenever he purchases a program, it is added to his channel line-up always at the same frequency.

A typical interface the consumer needs for program control is simply a telephone modem with a remote control or better yet, just a touch-tone telephone. In normal operation, the special services channel displays information about the services that are available (what movies are playing and when they start). This channel could also contain previews of the current features. The subscriber orders a movie by placing a telephone call. The order is logged and verified via ANI and/or ARU (Automated Response Unit) processing. The access control system then authorizes the decoder to receive the next available showing of the requested movie. This prompts the decoder to tune to the appropriate RF signal, decode the appropriate video service, and modulate it onto the output channel. The delay experienced between buying a movie and having it start is minimized by running multiple copies of a movie with staggered start times.

## CONCLUSION

Digital compression technology will open the door to new revenue streams, but will not eliminate the need for distribution and control of analog signals in the near future. There will continue to be a strong demand for user-friendly means of providing secure service to the millions of television sets currently connected to cable. The combination of interdiction for control of the analog signals and the "un-converter" approach for delivery of digital services provides a cost-effective, user-friendly way to tap the potential of digital compression technology.

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# COHERENT OPTICAL TRANSMISSION SYSTEMS

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*Abstract*

*Coherent optical transmission, through its high sensitivity and large capacity, offers a very attractive solution for subscriber loop distribution systems.*

*In this paper the world's first fully-engineered coherent multi-channel (CMC) transmission system will be described. Using established technologies of semiconductor lasers and high-frequency electronics a system demonstrator has been realized, having consumer TV comparable performance. The resulting network flexibility will be discussed.*

## INTRODUCTION

In this paper we will try to sketch the relevant issues in the evolution of broadband networks, both for telco and for CATV applications. From the perspective of service provision various network architectures will be discussed, with an emphasis on network flexibility and evolution towards interactivity. The role of new components and technologies will be addressed, and the options for present and future networks will be outlined.

It will be shown that using different optical wavelengths for different programs or services allows for a very flexible network architecture, at the same time making optimal use of the practically infinite transmission capacity of the optical fiber. The ultimate form of wavelength multiplexing is coherent multi-channel transmission, which in principle allows for thousands of channels to be transmitted over one single optical fiber.

The main part of the paper will be devoted to the description of a coherent multi-channel system demonstrator, which has been implemented in the context of the European RACE (Research for Advanced Communication in Europe) Program. This demonstrator is the first in the world in which fully-engineered transmitter and receiver units have been built according to predetermined specifications by different parties, thus showing the maturity of the techniques employed.

In final part of the paper the expectations about future developments will be addressed, and some pertinent conclusions will be formulated.

## BROADBAND SERVICES

Ever since the advent of low-attenuation glass fibers by the end of the seventies, engineers have been dreaming about fully integrated digital broadband networks. Basically they are right: the fiber, and in particular the single mode fiber, offers a transmission capacity which can be considered as infinite for all practical purposes. By deploying this "optical ether", all information streams and all communication services any individual could ever wish to have at his or her disposal, could easily be taken care of.

Reality turned out to be somewhat different: although the transmission window of the fiber is 30,000 GHz wide, thus allowing for a very low price per unit of bandwidth, the cost of the associated electronic and optoelectronic components was (and still is) prohibiting the realization of that dream. In addition, in almost all countries political issues turned out to play an important role in establishing a broadband integrated services digital network (BISDN). The basic reason for political interference is evident: for several years to come, the only real broadband service is television distribution, and the investments necessary for BISDN can only be recovered if real broadband services are supported by that network. This insight generated a controversy between the public telecom operators (PTO's) and the CATV operators.

However, there are more problems associated with an integrated digital network, such as:

- different tariffs for different services
- lack of digital consumer terminals
- lack of digital video standards
- high bitrate required for a large number of TV channels

The first issue recognizes the fact that a tariff structure simply based on kb/s or Mb/s will not be acceptable: either TV distribution would be too expensive, or telephone would be (nearly) free. Consequently, the operator will have to discriminate between different services, which means that the concept of "integration" is partly lost.

The second and third issues are strongly interrelated. In every country there is a large installed base of TV sets, VCR's, etc. which all expect analog signals at their input. Although this situation is changing (digital ATV in the USA, the acceptance of the MPEG digital coding standard, etc.), analog terminals will be around for many years to come, and future networks will have to take that into account.

Finally, there is the fact that digital transmission requires more bandwidth than its analog counterpart, unless additional measures are taken. Straightforward digitization of one PAL or NTSC channel leads to a bitrate of about 140 Mb/s. Transmission of a multitude of channels in time multiplex would require an unacceptably high bitrate on the subscriber line. Possible solutions to this problem will be discussed in the next section.

## OPTIONS FOR TV DISTRIBUTION

The basic problem in providing television distribution services in an integrated digital network is due to the fact that on the one hand TV requires quite some bits/second/channel, whereas on the other hand TV distribution via the traditional CATV networks has been a cheap consumer-oriented service for decades. Various solutions to this problem are being studied, and the most important options are the following:

- fiber sharing
- analog AM or FM transmission
- optical amplification
- reduction of bitrate/channel
- wavelength multiplexing

All of the above options may be instrumental in reducing the cost per channel, and several of them are relying on sharing of resources in the network.

### Fiber sharing

In many pilot projects in Europe and in the USA fiber-to-the curb or fiber-to-the-home solutions are proposed which rely on sharing the fiber and the related transmission equipment between tens of customers. The pros and cons of these solutions often depend strongly on local situations, so that it is difficult to give an assessment which has a general validity. An important issue however is the question of upgradability of such networks in future: a physical infrastructure requires high investments, and tends to have a long lifetime. Sharing the fiber inherently limits the network capacity, maybe for a long time.

Analog AM or FM transmission

By abandoning the concept of digital transmission in an integrated network, and introducing transmission in analog format, the problem indicated above could be solved in principle [ref.1.]. Several papers in this conference are devoted to this subject, so we will not address this issue further.

Optical amplification

In an optical tree-and-branch network the fiber and the transmission equipment is shared by many customers. However, the sharing factor is limited by the fact that at each branching point the optical power is distributed over the branches. The introduction of optical amplifiers at appropriate locations in the network may considerably increase the sharing factor by boosting the optical power.

There are however some disadvantages related to optical amplifiers located in the network. First of all there is active optical equipment out in the field, which will require surveillance. A second problem to be considered is that the network may lose transparency, in particular for return channels.

Reduction of bitrate per channel

A straightforward way to escape the problem of a (too) high bitrate per TV channel is the application of source coding [ref.2.]. A lot of work in this domain has been carried out worldwide, and the results seem to indicate that a bitrate of 1.5 - 2 Mb/s is the minimum required for an acceptable quality. In this domain the main trade-off is the advantage resulting from the bitrate reduction versus the increase in complexity of the decoder.

Wavelength multiplexing

One of the options for the separation of services in an optical network is to use different wavelengths for different services. Usually, a discrimanation is made between three domains in wavelength multiplexing:

- wavelength division multiplexing (WDM), with a wavelength separation of several tens of nanometers,
- high-density WDM (HDWDM) with a separation of 2-10 nanometers, and
- coherent multi-channel (CMC) multiplexing, with a separation of less than 0.1 nanometer between adjacent carriers.

In the first two cases, the optical carriers are separated by optical frequency selective devices [ref.3], in the case of coherent multiplexing the selectivity is obtained by electronic means, as will be outlined in the next section.

One of the advantages of all wavelength multiplexing schemes is that the various optical carriers are completely independent, so that different modulation schemes can be used, adapted to the particular information to be transmitted.

Furthermore, the bandwidth available in optical fiber between 1300 and 1550 nanometer is about 20,000 GHz, and this capacity can only be exploited fully by applying wavelength multiplexing techniques.

## COHERENT TRANSMISSION

Coherent optical transmission is fully analogous to heterodyne or homodyne techniques applied in radio and microwave systems already for decades. The only difference is that the carrier is now an optical carrier, with a frequency of 200 THz instead of a radiowave in the MHz or GHz domain.

The basic principle of coherent transmission is outlined in fig. 1.

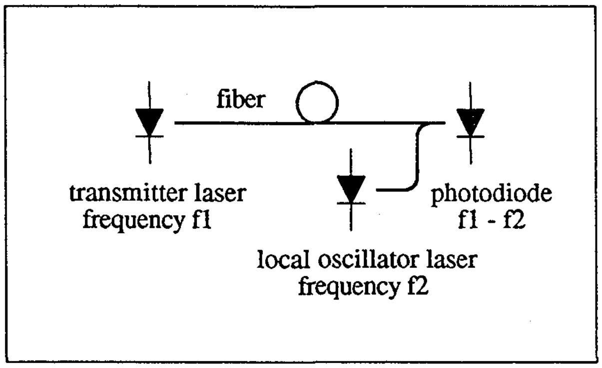


Fig. 1. Principle of coherent transmission.

The transmitter laser, which emits at a central frequency $f_1$, is modulated with a digital signal, either in amplitude (ASK), in frequency (FSK), or in phase (PSK). The modulated light is tranmitted via the fiber to the receiver, where it is mixed with the light of a local oscillator laser, emitting CW at a frequency $f_2$. One of the mixing products which are created in the photodiode is the frequency difference $f_1$-$f_2$. If $f_1$= $f_2$ homodyne detection takes place, if $f_1$ is unequal to $f_2$ heterodyne detection.

The three main advantages of coherent transmission are the following:

- high sensitivity
- high selectivity
- high transmission capacity

The high sensitivity is a result of the presence of the local oscillator laser: the large optical power reaching the photodiode makes that the total receiver noise is not determined by electronics, but by shot noise from the photodiode. Depending on the modulation scheme used, the resulting sensitivity can be 10-20 dB better than that of direct detection receivers.

The high selectivity is a consequence of the fact that selection takes place electronically, not by optical means. As an example, in an FSK system with a modulation amplitude of 1 GHz and an IF frequency $f_1$-$f_2$ of 1 GHz, the signal band is in the region from 0.5 GHz to 1.5 GHz, and can easily be filtered electronically.

Evidently, for being able to detect an FSK modulation with an amplitude of 1 GHz, the inherent stability of the optical carriers must be much better than 1 GHz. Calculation shows that a laser stability of about 20 MHz is required for such systems. With today's DFB or DBR lasers this requirement can easily be fulfilled.

However, if the control over the optical spectrum on the transmission fiber is that precise, then there is no reason why one should transmit only one signal over the fiber. It is then possible to connect several transmitter lasers, with operating frequencies several GHz apart, to one and the same fiber. By changing the frequency of the local oscillator laser ("tuning"), each of the transmitted channels could be selected. If the transmitters were modulated by TV signals, such a receiver would represent an optical TV tuner.

Since adjacent channels can be separated by only a few GHz, the bandwidth of the fiber would allow for the transmission of thousands of channels. The challenge to implement such a system, in a well-engineered form and according to pre-defined specifications, has been taken up in 1988 by the RACE Project R1010 "Subscriber Coherent Multi-Channel System", consisting of a consortium formed by HHI in Berlin, GEC-Marconi in Caswell, Siemens of Munich, and IMEC of Ghent, with Philips as Prime Contractor. The results obtained will be described in the next section.

## RACE DEMONSTRATOR

The fundamentals of the Demonstrator that has been implemented in the course of 1991 is shown in fig.2.

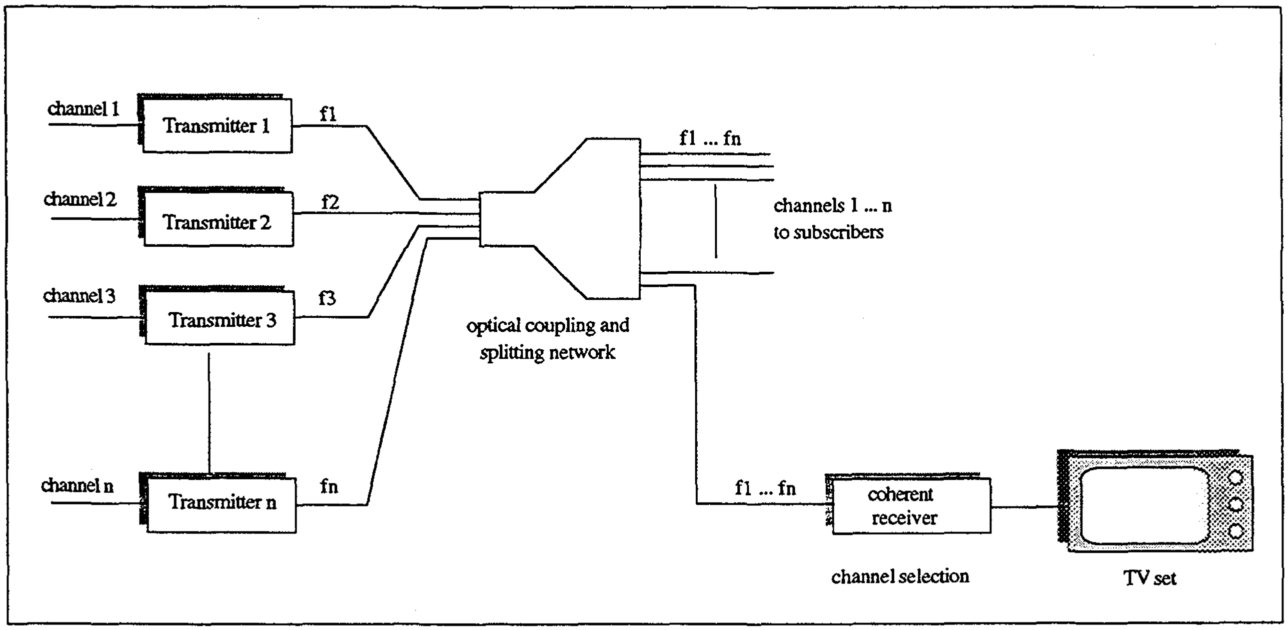


Fig. 2. Architecture of the Demonstrator.

In this Demonstrator, ten coherent FSK transmitters are modulated with a digital TV signal at 140 Mb/s. The optical outputs of the transmitters at frequencies $f_1$ .... $f_n$ are combined in an optical coupling network, and subsequently split to about a thousand subscribers. Each subscriber receives all transmitted TV channels, and by tuning the local oscillator in his coherent receiver he can select any channel he wants. In the present Demonstrator only ten channels are available, but today's technology would already allow for about 50 channels at 140 Mb/s, separated by 5 GHz. Taking into consideration that with a moderate amount of compression 10 TV channels can be accomodated in one 140 Mb/s stream, it is clear that this technique makes it possible to transmit hundreds of channels. In the following sections some details on the Demonstrator will be presented.

### RACE Demonstrator specifications

In the following list the relevant technical parameters of the RACE Demonstrator are presented, as they were fixed in 1989.

- Bitrate/channel 140 Mb/s
- Modulation type FSK
- Modulation amplitude 1200 MHz
- Optical output transmitter > 0.5 mW
- Receiver sensitivity < - 45 dBm
- Optical channel separation 10 GHz
- Number of channels 10
- IF frequency 1250 MHz
- Operating wavelength 1560 nm

As compared to today's state-of-the-art, these specifications are rather conservative, but in 1989 they were not, and they had to be fixed in order to allow the various partners in the project to build the transmitters and receivers according to these specifications.

But even with these conservative specifications, the capabilities of this system are impressive. With an optical output power of 0.5 mW, and a receiver sensitivity of better than -45 dBm, more than 1000 subscribers can be connected to one single transmitter. The network itself is fully transparent, and it would be easy to extend the system with ATV or HDTV transmission, just by adding transmitters for such services [ref.4.].

The fact that this Demonstrator has been built according to predefined specifications by different institutes, also shows that the technology is relatively mature. Since 1990 parts of the equipment have been moved to several locations in Europe, and switching on the mains was sufficient to put the equipment into operation. By using a normal remote control unit, any of the available channels can be selected by tuning the local oscillator laser in the receiver, within 0.2 seconds.

During 1992 the complete Demonstrator will be subjected to various tests, and the transmitters and receivers will be under continuous surveillance in order to gather information about laser stability etc. Finally, in a next phase of the RACE Programme, similar equipment will be used in field tests by European operators.

## CONCLUSIONS

Various problems related to the implementation of broadband networks were addressed, and options to solve these problems were discussed. It was argued that wavelength multiplexing, and in particular coherent multichannel techniques were very attractive for the distribution of a large number of TV programs.

A demonstration system resulting from a European cooperation was described in some detail. This Demonstrator clearly shows the capabilities of coherent multi-channel systems, and the maturity of the technologies required.

In summary, it has been shown that coherent techniques are not only promising, but will be reality in the near future. The main advantages are that the network remains fully transparent and allows for flexible upgrading in the future, that the transmission capacity is practically unlimited, and that the technique can be combined with additional options like fiber sharing, optical amplification and bitrate reduction.

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# "Comparison of Near Video on Demand Methods for CATV"

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## Abstract

This paper will describe a simulation model for evaluating trade-offs between Video On Demand, (VOD), schemes and transmission methods. A statistical approach is used and the concept of a "Demand Function" is introduced. The model is then applied to the specific case of a CATV system which assumes new fiber optic architectures which use "Star" configurations from the Headend and standard coaxial "Tree and Branch" networks to the subscriber. This is similar to the architecture used by ATC in their Queens project [1]. The model is also can be used to compare the capacity of different networks, such as Telephone, CATV, and DBS, to provide NVOD services.

## 1.0 Introduction

There is currently a clear demand by consumers for prerecorded video program material, as evidenced by the success of Video Tape Rental stores, (VTR), with revenues estimated at greater than $15 billion.

Several industries are currently developing strategies for getting a portion of these revenues; such as DBS, Telcos, CATV, and even Terrestrial Broadcasters. Recent advances in digital video compression will allow these industries more channels to compete with the Video Tape Rental stores.

But, how can we compare the efficacy of these different industries for providing NVOD services electronically to the home? And how can different methods for supplying video programs electronically to subscribers in their homes, within the same industry, be compared and evaluated?

Video On Demand, (VOD), is a term that has been used by Telcos [2] to mean a video program delivery system that retains the features of a VCR, (such as Play, Rewind, Fast Forward, Pause, etc.), and provides virtually instant access to large video libraries.

The implication is that :

* any Subscriber can view,
any Library Title, at
any Time.

This scenario could take infinite, or nearly infinite, resources to be able to satisfy all of the demands of all of the subscribers all of the time.

A less ambitious goal has been called Near Video On Demand, (NVOD), where less resources are required, ie. Sources, Titles, or Switching.

Such that:

* not all Subscribers can view,
any Library Title they want , at
any Time they want.

The Cable industry is currently going to 150 channels with 1.0 Ghz systems [1], and with compression, greater than 400 channel systems are realistic. But, is this enough channels or maybe more than enough to compete effectively with VTR stores?

This paper will outline a method for answering these questions by using computer simulations of the various NVOD scenarios.

## 2.0 NVOD Scenarios ( Simulated)

1. "Time Shifted Broadcast" uses several copies of the same program with different start times, similar to ATC's Queens project[1]. This method uses fixed scheduled Start times. The scheduled start times are displayed to the potential subscriber via a barker channel.

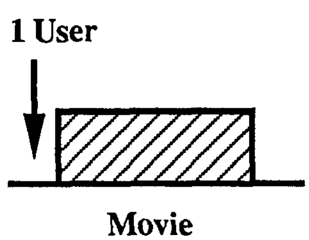
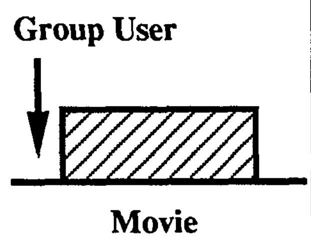
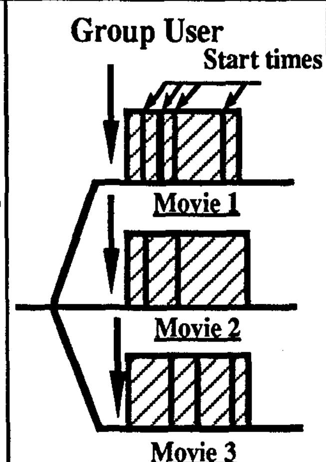
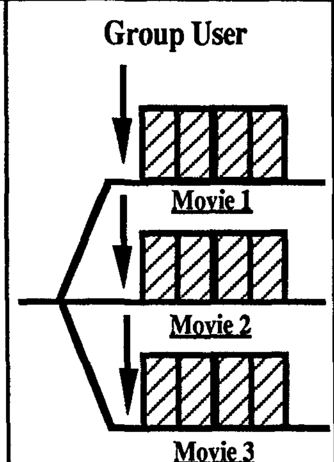
2. “Subscriber Directed Broadcast” is similar to Case 1., but does not exclusively have fixed schedules. This method uses Subscriber requests to modify scheduled start times.

3. “Partial Switch” adds switching to the Subscriber Directed Broadcast system so that more than one Program Title can use the same Channel or Source. This is the Case preferred for CATV.

4. “True VOD” (Telco style)[2] allows a program data base to be switched and dedicated to an individual Subscriber, on Demand. Because each subscriber has a dedicated Channel, the typical functions of VCRs can be implemented, such as Rewind, Fast Forward, Pause, Stop, etc.

In table 1, some features, advantages, and disadvantages of these NVOD scenarios are compared.

Table 1. Various NVOD Scenarios

| **CASE 4**<br>True VOD<br>*Video Rental* | **CASE 3**<br>Partial Switch | **CASE 2**<br>Subscriber Directed Broadcast | **CASE 1**<br>Time Shifted Broadcast |
|---|---|---|---|
|  |  |  |  |
| Interactive 2-Way<br><br>Fully Switched | Interactive 2-Way<br><br>Part Broadcast<br>Part Switched | Indirect 2-Way<br><br>Broadcast<br>(No Switch) | Indirect 2-Way<br><br>Broadcast |
| Easily Saturated<br>Most Titles<br>Least number users | Reduced Saturation<br>More Titles<br>More users | No Saturation<br>Slightly More Titles<br>Unlimited Users | No Saturation<br>Least Titles<br>Unlimited Users |
| High Customer Satisfaction<br>Most Variety<br>Most Convenient | Convenient<br><br>Good Variety | More Convenient during peak loads. | Least Convenient<br><br>Moderate Satisfaction |

## 3.0 Simulation Model Description

In [3], several authors have discussed several simple NVOD scenarios by using analytical means. In practice, however, the Cases can become very complex, taking into account various conditions, such that closed form solutions may not be obtained.

The model, discussed in this paper, uses building blocks in a computer simulation to test the effect of various input conditions and system configurations.

The Subscriber is introduced into the systems in terms of a "*Demand Function*", which is made up of several parts such as: popularity of video program title, national and regional viewing habits of consumers, average number of hours people watch TV, probability that a Subscriber will want to watch a movie more than once, availability of Titles, Start times, etc. The "*Demand Function* " allows the Simulation model to test various Subscriber conditions by entering simple curves, similar to "Load Distribution Curves" listed in [5]. As the Simulation runs, the cumulative probability of a Subscriber making a request is calculated at fixed increments and Outputs are generated for calculating financial data, channel usage, and a quality factor, (which is related to Subscriber satisfaction).

The basic model consist of a data base of "Set-Up" conditions which are loaded into the "Simulation Model", (Inputs 1 to 5). These inputs are then run through the Simulator and pre-defined "Outputs" are generated, (Outputs 8 to 10). While the Simulation is running the Operator can also make minor adjustments to some parameters associated with the Set-Up conditions.

The following is a list of the NVOD Simulation Model's Inputs and Outputs.

**Model Inputs (Figure 1)**

1. Sources.

a. Hardware Configuration

1. Total number of active sources, ,which assumes 1 active source per channel.
2. Number of Titles allocated to a single source, ( includes total storage capacity of source).

b. Type

1. Includes physical parameters of playback device, "Source". ( S-VHS, Optical, digital, etc.).

c. Storage

1. Usable media "Storage time", (how many programs can be stored on a single media- -depends on Program lengths and media size).
2. Access method; Random access, serial tape, etc.

d. Source Selection Method

1. Time to locate a specific Source, Queue Program, and Play Program.
2. Selection method for connecting a Source to a Channel, (Multiplexing method and switch configuration).
3. Media loading method - ( how does Source get access to Library; hand load, carousel , electronic, etc.

2. Program Statistics

a. Statistics from "Variety"[4] , " Top 50 Video Titles" are used to assign a probability index to Programs. This allows weighting of different Program titles.

b. TV ratings from "Nielson", also listed in "Variety" on a weekly basis, allows weighting of the viewing habits of Subscribers in Network and CATV programs against NVOD programs.

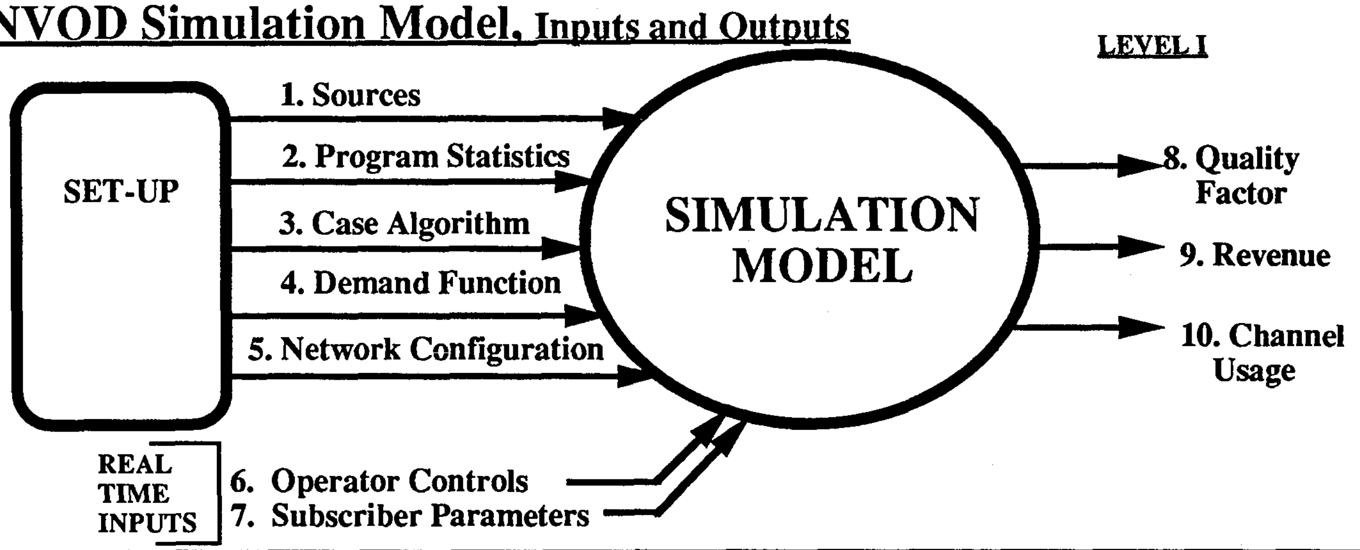
c. Program RUN TIME statistics are derived from movies on the "Top 50 list".

3. Case Algorithm

a. Time Shifted Broadcast, uses fixed schedules. Similar to ATC's Queens NVOD experiment[1].

# NVOD Simulation Figure 1.

## NVOD Simulation Model, Inputs and Outputs




### MODEL INPUTS

**1. Sources**
- Configuration
- Type (Characteristics)
- Storage
- Selection Method

**2. Program Statistics**
- Top 50 Video Rentals
- AM ratings (Nielson)
- Program RUN TIME

**3. Case Algorithm**
- * Broadcast, fixed schedule
- * Broadcast, user influenced schedule
- * Partially Switched
- * True VOD.

**4. Demand Function**
- Viewer statistics
  - * Global, (Nielson, etc.)
  - * Regional
  - * Individual
- Ordering Statistics
  - * Request / Buy Rate
  - * Cancel Rate

**5. CATV Network Configuration**
- Total Subscribers.
- Channel Line Up
- Node size.

### REAL TIME INPUTS

**6. Operator Controls** Changes to Operating Parameters

**7. Subscriber Parameter** Changes to Subscriber Parameters

### MODEL OUTPUTS

**8. Quality Factor**
- Efficiency of channel usage
- Customer satisfaction.
- Growth factor / current profit

**9. Revenue**
- Rev./Sub, Rev./Ch., Rev. / Title
- Profit
- Fixed cost
- Variable cost

**10. Channel Usage**
- Avg. # of Ch. used
- Peak usage
- Unused channel capacity.

b. Broadcast, with Subscriber influenced scheduling. In this algorithm the specific start times of Programs are allowed to vary as a function of both Subscriber "Demand" and a fixed schedule.

For example; The 8:00 P.M. showing of Program A may not start until 8:04 P.M. in order to allow more Subscribers to order the program. Or it may start earlier to reduce waiting time and increase buy rate.

c. Partially Switched uses algorithm b, but also allows simple switching of sources, such that if some channels are not fully utilized, other Programs can be substituted.

d. True VOD is the case of dedicating a source and a channel to a single Subscriber. Although this case is most like the Video Tape Rental scenario, it is expected that Resources are under utilized.

In "Cases" "b" and "c" a "Real Time" Two-Way data transmission system is required. Cases "a" and "d" can use "Store and Forward" technologies as currently used for PPV.

Case "c" is the case which best utilizes modern CATV system resources, with a broadband transmission to the Subscriber for Programs and a narrowband return channel for Subscriber requests.

4. Demand Function

The "Demand" function is a semi-random function which statistically models the viewing habits of Subscribers, by Global, Regional, and Individual statistics.

This function is developed such that it can update the pre-loaded statistical data bases , used for generating Subscriber requests and specific Subscriber's ordering and buying patterns. (ie., a Subscriber who has already watched a Program once, will have a reduced probability of requesting the same program a second and third time.

5. CATV Network Configuration

Specific CATV system parameters are added to the Model. These parameters can be changed during a simulation.

a. Total number of Subscribers
b. Node size
c. Channel Line-Up, including Basic, Extended Basic, Pay, PPV, Audio, Data , NVOD.

## Real Time Inputs

6. Operator Control

Operator Control is a means for the person who runs the Simulation to change certain operating parameters during the Simulation.

7. Subscriber Control,

Subscriber Control is a means to change certain Subscriber parameters, specific to the "Demand" function, during a Simulation .

## Model Outputs

8. Quality Factor

a. A measure of the efficiency of using NVOD channels, ( NVOD channels versus other channels).

b. A measure of Subscriber satisfaction. Measured as a ratio of Requests versus Purchases of NVOD and is dependent on Program availability and Subscriber availability.

c. Growth factor versus current profit is a measure of the revenues and cost of current plant relative to the potential for expansion, such as adding channels, adding switching, adding titles, etc.

9. Revenue

This is the primary measured system performance criteria for the Cable Operator and it consists of :

a. Revenues
b. Costs - fixed
variable
c. And a breakdown of financial data such as Rev./Sub., Rev. / Ch., Rev. / Title. etc.

10. Channel Usage.

This output has two primary functions.

a. To determine an optimal size for a specific system with given specifications.
b. To be used as feedback in the actual operation of the Simulation, (and potentially in real CATV systems), to adjust and control the current configuration of switched programs dependent on previous conditions.

## 4.0 NVOD Simulations (Figure 2)

To begin a Simulation the *Simulation Preset Conditions* are batch loaded into the Simulation program prior to start-up. In this way different "Source Configurations", "Case Algorithms", and CATV "*Network Configurations*" can be simulated.

Once the Simulation has "Started" the *Simulation Timer* provides the primary control of the NVOD Simulations . It provides daily and weekly schedules for the normal Broadcast Channels . It is also used to activate various elements of the "Demand " function, such as scheduling of normal Broadcast TV.

It also provides a *Random Number Generator* to be used for generating Subscriber "Requests" and "Orders" on a statistical basis.

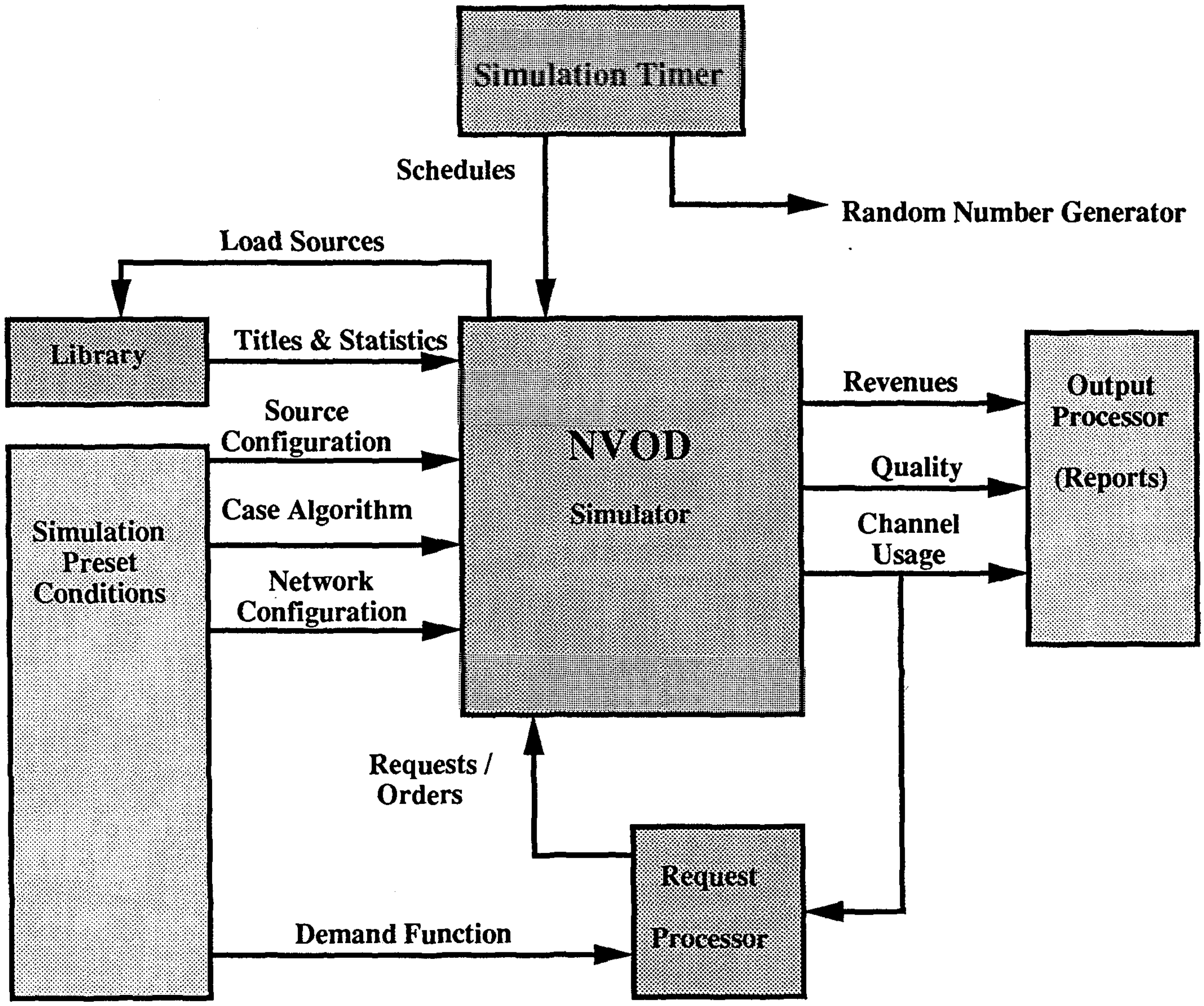
Figure (3) shows a typical "Program Packet". All programs transmitted throughout the CATV system can be modeled as a time ordered Matrix of Program Packets. A complete Simulated system is made up of a matrix of these Program Packets, which represents the traffic on all channels.

Specific *Demand Functions* are loaded into the *Request Processor*. The *Request Processor* then generates random *Requests / Orders* which reflect Subscriber Demand as it relates to the specific *System Configuration* modeled, *Channel Usage*, and pre-loaded statistics. The random nature of the *Request Processor* comes from the *Random Number Generator* output of the *Simulation Schedular*.

The *NVOD Simulator* is a logical block which manipulates various matrices. The logic is fixed, but certain "Real Time" inputs allow changing System and Subscriber parameters such as, Node Size, Total Subs., relative ratios of Demand Function.

As the Simulation program runs NVOD Programs need to be changed, since after a period of time, (which is statistically determined from the "Top 50 Video Titles" listings and the actual viewing by the Subscribers ), the probability that a Subscriber would want to watch a particular Program changes. To simulate this function , the *Library* is accessed to load new Program Titles into the input matrix.

The *Output Processor's* primary function is for producing reports and to give the Simulation operator a view of what is happening during a simulation. The *Channel Usage* output is used to feedback to the NVOD Simulator the actual system activity, ie. what programs are actually being sent to what Subscribers.

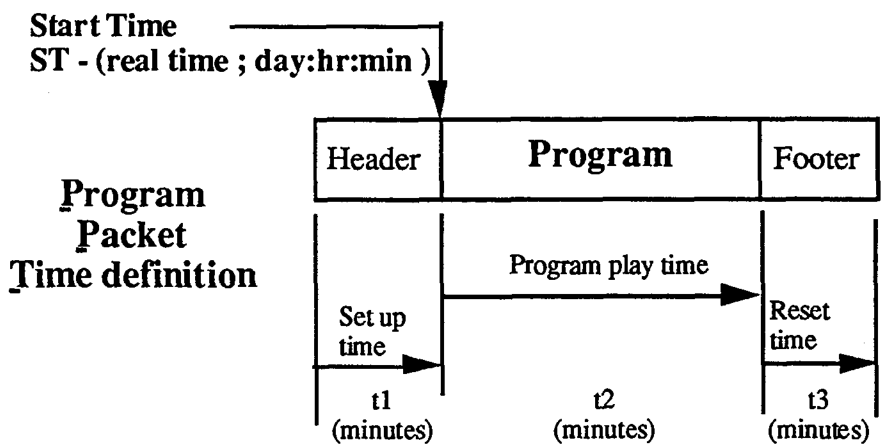



## Simulation System Configuration

Figure 2

PROGRAM PACKETS -

Analysis uses Packet concept for Programs.




t1 - Set up time ; time to queue Program to starting point. Including head-loading, seek-tiem, etc.
t2 - Program play time ; average time to play Program from beginning to end.
t3 - Reset time ; time from end of program to Header of next program. Includes Rewind time.

**Figure 3**

## 5.0 Conclusion

The task of determining how to provide NVOD Services in a cost effective and competitive manner, to Video Tape Rental stores, is quite complex. To launch most NVOD services, such as those envisioned by the Telephone and CATV industries, requires large capital investments. These large investments will require a quick inexpensive method to evaluate potential Return On Investment. The current method chosen , by Telephone and CATV companies, is to do a limited experiment with a small segment of Subscribers and with a limited selection of Program Titles. This is an evolutionary method and would probably be a good approach if it weren't for the competition among industries and the rapidly emerging technologies such as Compression and HDTV.

To further complicate the problem, different industries and communication networks will implement NVOD services in different ways. A method is needed which will inexpensively test the differences for providing NVOD by Telephone , CATV, or DBS networks. And an inexpensive method of comparing different cost effective NVOD schemes within industries is also needed. This paper recommends a viable approach. That is to develop a "robust" Simulation model which will be able to adjust to various NVOD schemes and will be able to include the random variations of consumer's demands.

It is envisioned that the Network which is able to deliver the most cost effective and Subscriber friendly system will enjoy a successful future. After all, there is only so much money the Subscriber has to spend on entertainment; or is there ?

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# COMPRESSED DIGITAL COMMERCIAL INSERTION: NEW TECHNOLOGY ARCHITECTURES FOR THE CABLE ADVERTISING BUSINESS

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**Cable Television Laboratories, Inc.**

*Abstract*

*Advances in compressed digital audio/video technology have paved the way for the introduction of a new technology platform for commercial insertion. These concepts embodied in digital video production, transmission, storage, and playback will redefine cable advertising sales operations in the 1990's. The basic concepts of a compressed digital commercial insertion network (CDCINet) are explored in conjunction with highlighting the various benefits this digital platform should provide.*

## Introduction

The cable industry is uniquely positioned to implement an architecture which enables advertisers the ability to "precision market" its products on the basis of geographical and demographical boundaries. Cable's growing slice of the advertising expenditure pie has demonstrated this unique capability, however there are technological and operational hurdles which impede further significant growth in this important revenue stream [1].

The use of tape-based video cassette player (VCP) technology coupled with the various forms of interconnects (hard, soft, hard interconnect/soft playback) has been the historical technological platform for cable ad sales operations [2]. However, new technology platforms based on integrating compressed digital video mass storage systems and powerful communication network architectures should allow an infrastructure which supports future revenue growth for cable advertising sales. This compressed digital commercial insertion network (CDCINet) will be a critical element of adding value to products and services (e.g., ability to deliver "on the fly" insertions) and supporting cable operators in reducing costs of operations (e.g., significant reductions in cost of tape duplication and tape distribution).

Cable Television Laboratories, Inc. (CableLabs) believes that dramatic improvements in the cost/performance ratios of mass storage and network technologies will allow for the realization of a CDCINet. Recently, it was reported that proponents of compressed digital commercial insertion systems are costing between $2,000 to $5,300 per insertion channel [3] for random-access style systems. Contrast these digital system costs to VCP random-access style systems currently priced at $8,000 to $10,000 per insertion channel.

The majority of the cost for today's random-access style system is centered around the need to allocate up to four VCP's to a single network. Therefore, if VCP's are costing around $1,500 each, an operator spends approximately $6,000 for spot storage for each random-access network. Compressed digital storage for 200 30 second spots is currently costing between $1,600 to $2,000. With the advancements in storage technology, it is conceivable that tens of GBytes of storage will be available, thereby driving spot storage costs to a few cents per spot. This follows "Moore's Law" where the cost of computing is theorized to half every three years.

It can also be argued that the cost of tape duplication (editing, compiling, dubbing, and generating multiple copies of each spot) and the associated cost of physically transporting these tapes to cable headends can be significantly reduced by utilizing the "electronic highway " for spot distribution to a digital storage medium.

Therefore, to analyze the creation of a CDCINet, there are at least three critical elements to explore:

- the network architecture(s) necessary to create an "electronic highway" which transports local, regional, and national spots to cable headends;
- the evolutionary needs of integrating existing VCP-based systems with new compressed digital playback systems, and;
- the need to establish interface guidelines for "front-end sales software", traffic and billing software, and electronic verification software.

Architectural Overview of a CDCINet

In proposing the creation of a CDCINet, which focuses on the general distribution of digitally encoded information, we create the need to examine the network requirements from both an architectural and transport protocol point of view.

Remember, the purpose of creating this network is to promote the interconnection and interoperability of hardware and software which subsequently allows advertisers to buy local, regional, and national spot avails easily and conveniently. Ideally this environment would allow several vendors of hardware and software to coexist; contrast this environment to today's

| | | |
|---|---|---|
| Layer 7 | **Application** | Provides access to the OSI environment for users and also provides distributed information services. |
| Layer 6 | **Presentation** | Provides independence to the application processes from differences in data representation (syntax). |
| Layer 5 | **Session** | Provides the control structure for communication between applications; establishes, manages, and terminates connections (sessions) between cooperating applications. |
| Layer 4 | **Transport** | Provides reliable, tranparent transfer of data between end points; provides end-to-end error recovery and flow control. |
| Layer 3 | **Network** | Provides upper layers with independence from the data transmission and switching technologies used to connect systems; responsible for establishing, maintaining and terminating connections. |
| Layer 2 | **Data Link** | Provides for the reliable transfer of information across the physical link; sends blocks of data (frames) with the necessary synchronization, error control, and flow control. |
| Layer 1 | **Physical** | Concerned with transmission of unstructured bit stream over physical medium; deals with the mechanical, electrical, functional, and procedural characterics of the access medium. |

**Figure 1:** Seven Layer OSI

airline reservation system where several hardware and software systems coexist.

Therefore, it is beneficial to discuss the attributes of a CDCINet in terms of the OSI model. The International Organization for Standardization (ISO) established a subcommittee to develop an architecture for these communications tasks. The result was the development of the Open Systems Interconnection (OSI) reference model, adopted by ISO in 1983 and endorsed by the CCITT in 1984. The OSI model is a framework for defining standards for linking heterogeneous computers. The term "open" denotes the ability of any two systems conforming to the model and the associated standards to connect these systems together [4].

Traditionally the OSI model has used a "layering" concept to promote this interconnection capability. Figure 1 shows a seven layer OSI model protocol suite. From layer 1 (the lowest) to layer 7 (the highest), we can see that interconnection and software interfacing is described.

Realizing that compressed digital commercial insertion is only one application within an architecture that interconnects cable headends, the Advanced Network Development Subcommittee at CableLabs has described a network architecture plan which has interconnectivity and interoperability at its center.

The importance of utilizing an OSI model approach to developing a new "connectivity" infrastructure within the cable industry is illustrated by viewing the Advanced Network Subcommittee's network architecture plan as shown is figure 2.

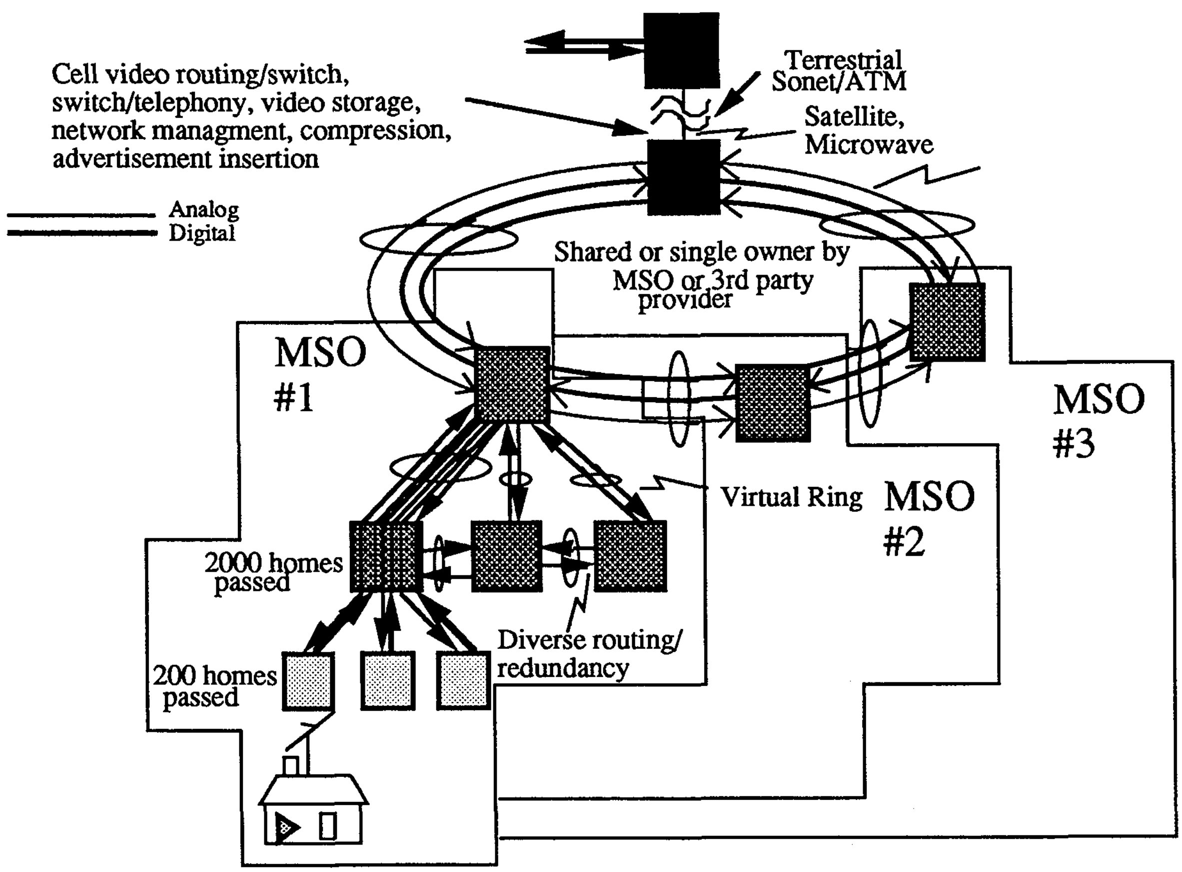


**Figure 2:** Advanced Network Development Subcommittee's Network Architecture Plan

It is clear that current developments in compressed digital video technology will soon be used for program delivery via satellite. This delivery of satellite delivered programming and the creation of some type of CDCINet should be some of the first applications deployed under the Advanced Network Subcommittees network architecture plan. When combined with non-video applications (alternative access, datacomm, PCN, and other non-entertainment services) the need for this network to migrate via an "open architecture" approach is evident.

Architectural Complexities of Local, Regional, and National Spot Delivery Systems

One of the challenges of maximizing revenue from the advertising community is to integrate the insertion needs of the local, regional, and national spot buyer. Each buyer has unique requirements and there are competing forces at play in cable operations when determining what percentage of avail time should be allocated to local, regional, national, and even cross-channel promotional insertions. Obviously each cable market is different and the cable operators in these markets will need to determine the appropriate "mix" of local, regional, and national ad spots which maximize revenue. Therefore it is recommended that any technological platform for commercial insertion should have scheduling capability which is controlled by the local operator.

It is necessary therefore to integrate the distribution technologies for these different insertion situations. Local cable advertising in many markets is targeted to community businesses which have consumers within a defined geographical boundary. For example, a neighborhood hardware store owner realizes it most likely draws customers to this store because of its geographical proximity to consumers. For this hardware store owner it is important that its advertising dollars are spent effectively within its geographical sphere of influence. A broadcast style ad insertion which covers a large metropolitan area and costs several times more than a "localized" insertion would probably be a waste of money.

Now contrast the needs of these geographically defined businesses to other firms which desire entire metropolitan coverage. Many businesses will still want broadcast style coverage and the current insertion systems are typically structured to meet this need. The majority of the $600 million the cable industry received in 1991 was from this type of advertising buy [5].

However, cable advertising executives realize that significant revenue growth opportunities will come from the national spot delivery business. The cable industry received approximately $150 million from this market segment in 1991 against total advertising expenditures of approximately $16 billion [5]. Many operators feel that cable's fair share of this market should be in the $4 billion to $5 billion range.

Therefore the architectural design of a compressed digital commercial insertion network (CDCINet) should look at the needs of local segmentation, regional distribution (via these hard, soft, hard distribution/soft playback techniques), and national spot market interconnection.

CDCINet Concept

A comprehensive end-to-end compressed digital commercial insertion network (CDCINet) which covers the need of local, regional, and national advertisers would likely need the following platforms:

- A national/regional distribution center(s) which utilizes both satellite and fiber optic based communications architectures. Both analog and compressed digital delivery of spots would be delivered to cable headends.

One would expect this platform to migrate to an all digital delivery system within the next five to ten years. These types of distribution centers would facilitate more efficient advertising buys for regional and national accounts.

- Coexistence of current VCP based insertion systems and new all digital systems at the local headend (or playback facility) which have "intelligent switching" integrating both. However, most operators should realize the advantage of storing spots digitally and subsequently begin phasing out VCP systems. Because spots don't need to be delivered to these headend storage devices in "real time", cost effective transmission schemes can be developed for locally produced ads.

- Interfaces and interoperable software packages which allow for efficient buying, scheduling, billing, and verifying of local, regional, and national ads. Sales oriented front-end packages should interface into traffic and billing systems, thereby creating the "paperless" environment so desired by both ad agencies and cable operators. It is important to look at common data exchange formats and integrate these into software systems to maximize advertising revenue.

When integrated, the various components of a CDCINet might resemble the architectural description in figure 3. These components: local ad production/sales facilities, regional interconnection facilities, and national uplink facilities, are diagrammatically shown to coexist.

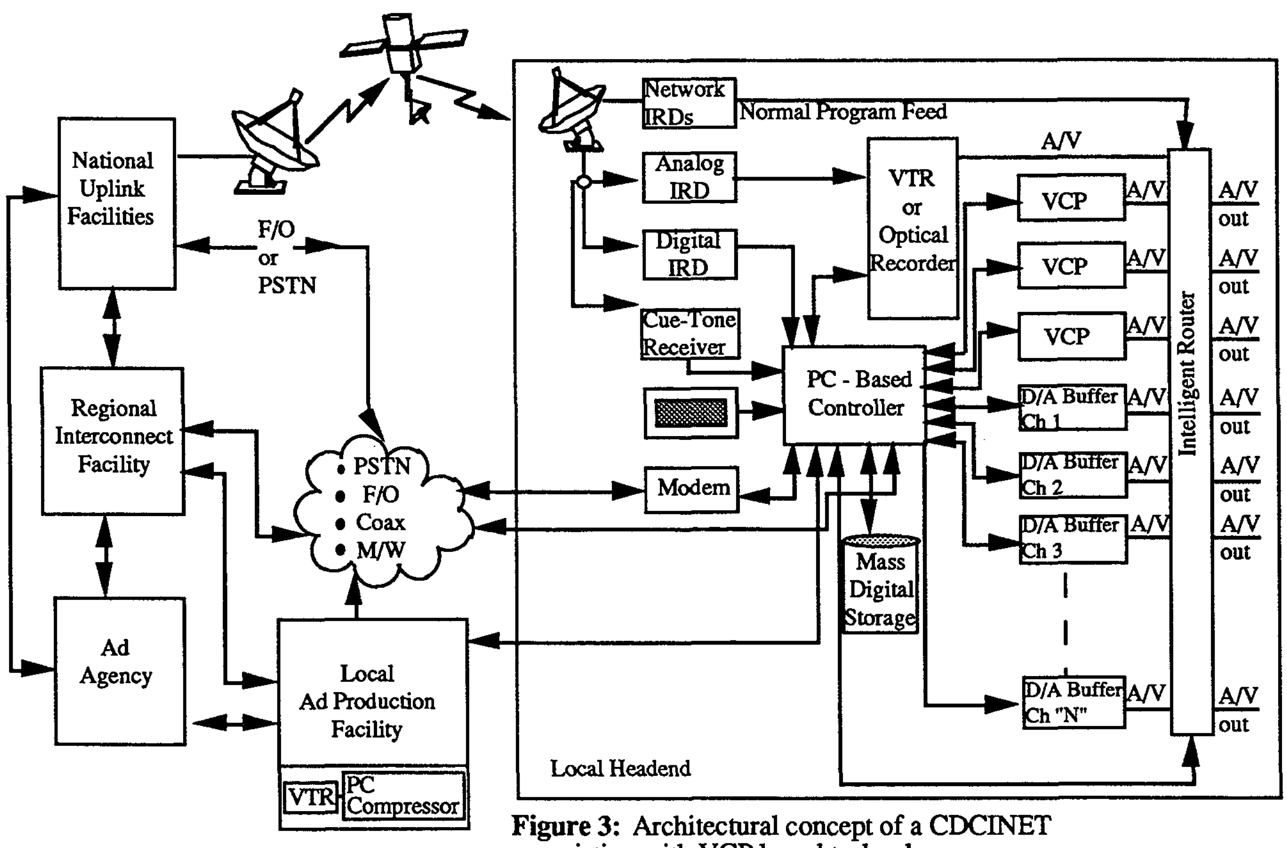


**Figure 3:** Architectural concept of a CDCINET co-existing with VCP based technology

While national and regional spot delivery are accomplished by transmitting analog and digital signals to addressable headend storage and playback systems, the delivery of geographically segmented local advertising will likely be based on AM-VSB fiber optic nodal architectures. Figure 4 diagrammatically shows routing different spots to different areas of the community even though the same program source will be viewed in both parts of the system. The first systems deployed for this type of market segmentation will likely be point-to-point FM or digital fiber optic hubs. Further research and development will be necessary to cost-effectively switch a large number of different commercials into the same program source when using AM-VSB fiber optic transmitters. However a system might find it economically viable to deploy a few channels of geograghically/demographically segmented commercial insertion.

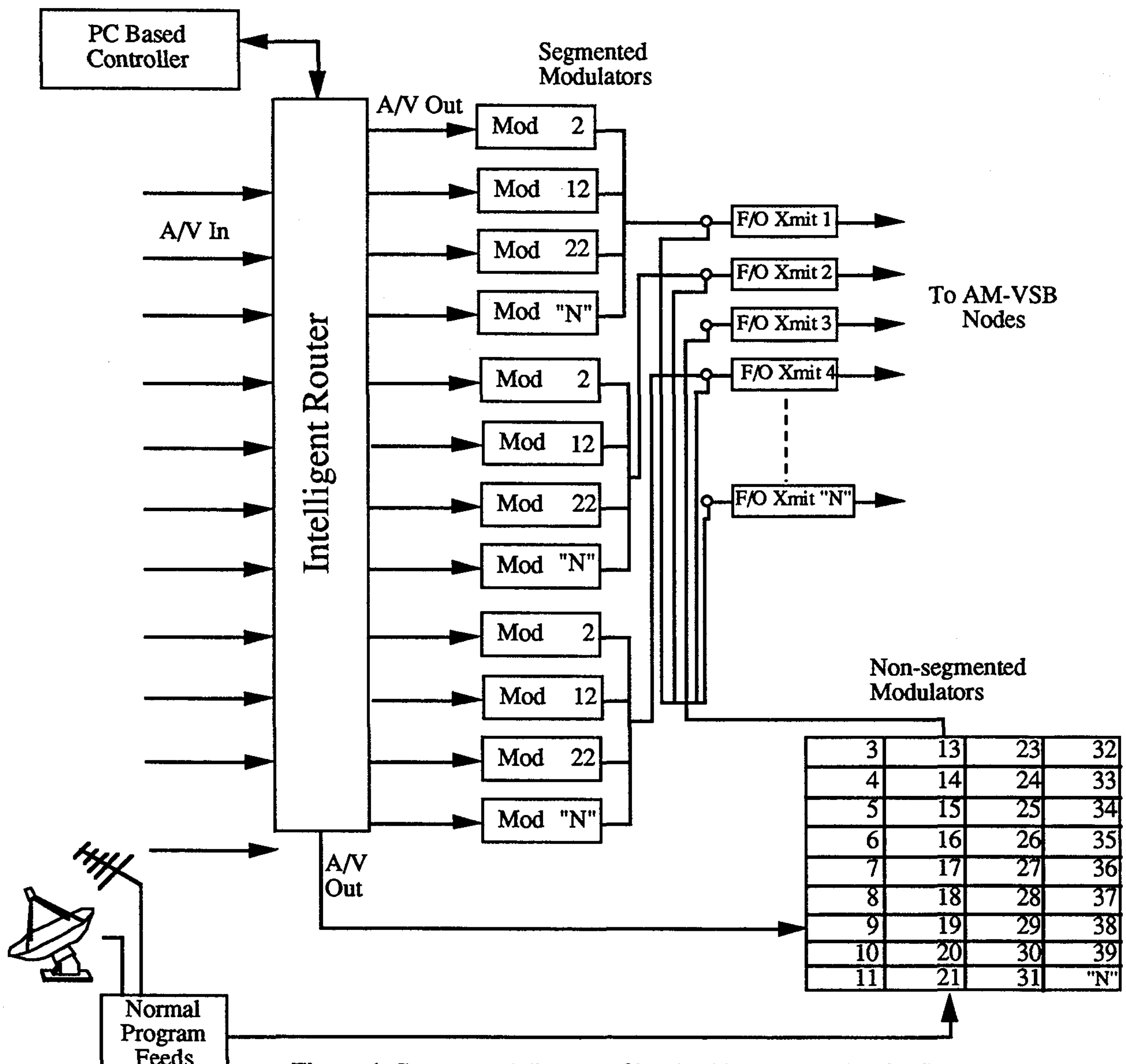


**Figure 4:** Concept and diagram of headend interconnection for fiber optic nodal segmentation

Order Entry, Traffic, Billing and Verification Issues

While the hardware portion of this CDCINet is an integral part of the solution to maximizing advertising revenue, the software requirements of this network is often cited as the greater challenge for the industry to grapple with.

It is recognized within the national spot advertising community that buying and coordinating multiple avails across multiple headends is difficult. Tom Winner, from the ad agency of Campbell, Mithum and Esty's, recently intimated this difficulty in the November/December, 1991 issue of CableAvails magazine. In this article he indicated that,

> *"I see selling spots on cable as terrifically difficult". When told of the electronic wizardry slated to eliminate difficulty, he responded flatly: "We'll believe it when we see it."* [6]

Many MSOs and industry service providers are beginning to address these limitations. In his 1991 white paper on local advertising sales, Larry Zipin, vice president-advertising sales for Warner Cable Communications, Inc., says:

> *"The other half of the Commercial Insertion Technology issue is defined by the Software that drives the Process. This two-part software is usually referred to as the Traffic and Billing System. Traffic is the component that is fundamental to the creation of revenue potential, because it is within this software that the degree of Accessibility to the commercial insertable inventory is determined. In other words: if the combination of ad insertion hardware and Traffic software enables the maximum assessibility to individual spots (i.e., fixed positions), then revenue potential is increased; as compared to random rotations (i.e., ROS) in which revenue potential decreases significantly.."* [1]

To further illuminate this issue, Zipin goes on to say:

> *"At the present time, most of the major ad agencies that control the Spot Market transactions are not willing to deal with Cable's fulfillment limitations. As a result, the Spot Cable opportunity will not be realized until a multi-market, commercial insertion/delivery system coupled with multi-market spot scheduling/traffic software is in place.*
>
> *This Spot Cable opportunity, however also has the added complication of requiring that its fulfillment technology be able to interact and interface with the Local Ad Sales technology so that the Local Ad Sales owner maintains control over its inventory. (the airline industry's reservation system has been cited as being analogous to the ad sales situation)."* [1]

Again we see the convergence of cost-effective computing plus the use of cost-effective communication networks can be the enabling architectures for an integrated software solution to the above challenges.

Therefore, as it was shown to be necessary to promote interfaces between hardware, the "open architecture" approach to software interoperability is also needed to accomplish this "multi-market" capability. Hopefully industry leaders can recommend an "interface platform" that allows multiple developers of front-end sales software, traffic and billing software providers, and electronic billing and verification providers to coexist within this CDCINet.

Conclusion

The cable advertising industry has an opportunity to develop and construct an integrated network system (CDCINet) which allows for the coordination of local, regional, and national advertising avails.

It is critical for the cable industry to develop interoperable hardware and software systems at and between each level of this advertising network: the local level, the regional level, and the national level. This physical network, when coupled with powerful sales, research, traffic, billing, and verification systems, can do what advertisers want it to do: delivery the right message to the right group of consumers cost-effectively.

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# CONDITIONAL ACCESS SYSTEM FOR DIGITAL TV: THE EUROCRYPT STANDARD

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## ABSTRACT

*Digital compression of video signals will provide a breakthrough in TV broadcasting market by enabling the transmission of several TV programmes per channel. Moreover the digital technology for transmission will allow to provide the consumer with various services such as:*

- *TV programmes with several stereo sounds.*
- *Stereo sound radio programmes.*
- *Data transmission...*

*The tremendous increase in number and type of services will require a permanent use of conditional access. In Europe, the EUROCRYPT standard has been defined for the MAC transmission requiring conditional access. The module in charge of the conditional access function is a Smart Card.*

The EUROCRYPT system offers the usual types of access modes. Those are:

- Subscription: Access is given for a certain type of programme for certain period of time.
- Pre-booked Pay per View: The user asks in advance for the access to one or several programmes.
- Impulse Pay per View (IPPV): If a user wants to access this type of programme, he will have to accept to buy the programme. The cost of the programme will then be debited from the credit in his Smart Card. In this case, the user watches what he pays for.

To get a personal control of his Smart Card, the user is given the possibility to define:

- Personal maturity rating
- Secret code to get access to IPPV programmes and programmes under maturity rating

Some facilities are also offered to the broadcaster to control the audience:

- Blackout with replacement under geographical and/or subject basis
- Fingerprinting to ensure that no misuse of video recordings from his programme are made.

The following pages describe the current implementation with the Smart Card and the way it will be extended from MAC (Multiplexed Analogue Components) to digital TV.

## SYSTEM ASPECTS

The system can be split into two parts [see fig. 1]: The receiver side and the transmission side. The receiver side is made out of a receiver which incorporates the descrambler. A Smart Card performs the conditional access function. The transmission side has to manage the following functions:

- The encoding of the signal
- The scrambling function of video, sound, data ...
- Transmission of conditional access information related to scrambling using a master Card.

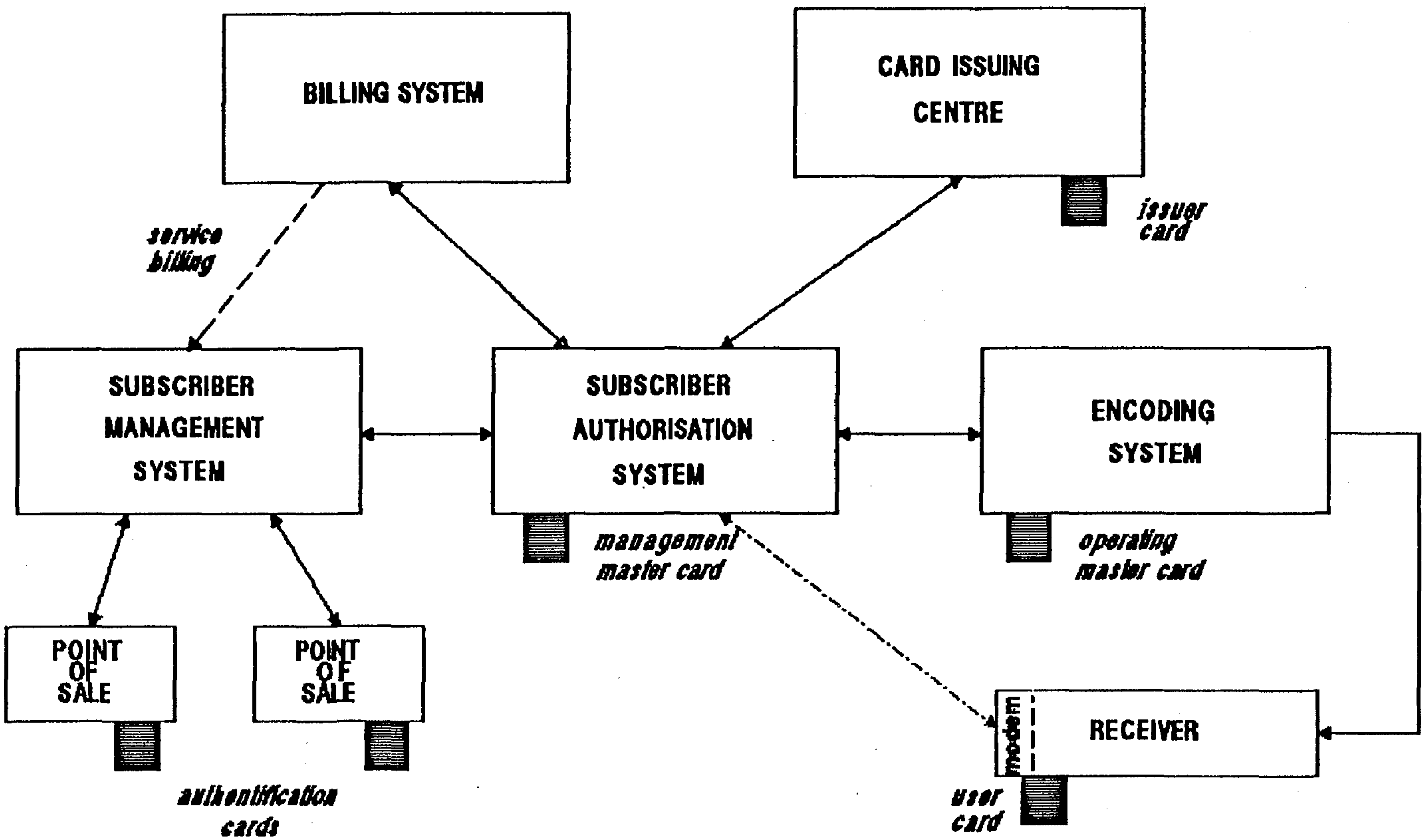


fig. 1

EUROCRYPT SYSTEM

. Transmission of entitlements related information using a master Card.
. Handling of the Commercial, Billing and Subscriber data base.

## SMART CARD

The Smart Card is the part which is in charge of the conditional access and which handles all the entitlements and secrets of the conditional access system. The Smart Card is similar to a credit card with a built in chip. The Smart Card offers high protection against piracy:

. The chip includes protections versus physical piracy.
. The built in component is a monolithic chip which means no easy access to the internal bus.
. The microprocessor of the Smart Card controls the interface with the external world. No sensitive information can be read in the Smart Card through this interface.
. No secret information appear in clear on the Smart Card interface. There is no use in monitoring this interface for secret information.

The Smart Card is built up in two major parts [see fig. 2]:

. The program part which contains the software to manage the Smart Card and also the algorithm to decrypt messages.
. The memory which contains all the information related to the services (e.g. the name), the keys used by the broadcasters to encrypt the messages and also the entitlements the user has for his programmes.

The master Card is a specific card for the broadcaster who has the capability to encrypt messages.

EUROCRYPT allows several broadcasters or "Service Entities" to share the same Smart Card. An authority called "Issuer Entity" supervises the whole Smart Card and attributes to each Service Entity memory to write his data. The Issuer Entity will also supply the Service Entity with his first secret key.

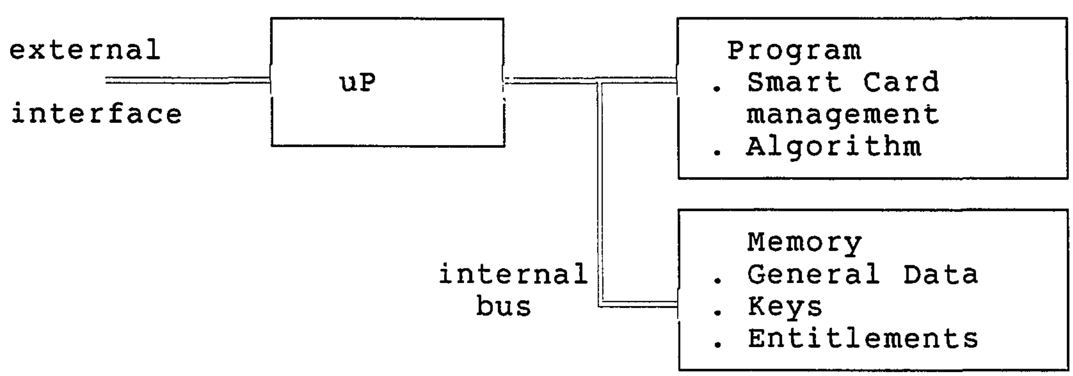


fig. 2

The current version of EUROCRYPT Smart Cards used in Europe can contain around 2000 entitlements of any type. This of course depends on the number of Service Entities in the Smart Card. The new version with EEPROM will be erasable which means that it could theoretically last for ever.

One of the most obvious advantages of a Smart Card is portability. A user who has to travel can take his Smart Card with him and insert it in the receiver available in his room at e.g. a hotel. The receiver is not specific for each user: It contains no secret key. When a problem occurs with the entitlements the Smart Card can be taken to a shop and checked or loaded with some special entitlements. Furthermore, if by any chance someone could pirate the Smart Card, the broadcasters would only have to replace the Smart Cards instead of complete receivers which from a price point of view is a big advantage.

## DIGITAL TV APPLICATION

The data in the digital transmission, like in MAC, are organized in packets and each component can be selected and extracted from the packet multiplex by filtering its packet address.

The scrambling in a digital system consists, as in the digital part of MAC, in adding a binary pseudo random sequence to the bit information extracted from the signal. This scrambling is totally symmetrical (identical operation on transmission and reception side) and does neither give any consistent information about the initial signal nor about the scrambling sequence itself.

The receiver is in charge of the demultiplexing of the data, the recovery of the components (video, sound, data...) and also of the descrambling of the components which are under conditional access.

The scrambling sequence is the output of a complex Pseudo Random Generator initialized by a Control Word and the frame counter of the video. The Control Word is generated (e.g. randomly) on the transmitter side and transmitted in a encrypted form to the receiver side. If the user did subscribe, the conditional access will deliver the decrypted Control Word to the receiver which will load them in the pseudo random generator. The user will then get a clear signal.

Each component part of the transmission (sound, video, data ...), can be scrambled with different Control Words, which means that each can be accessed separately.

Concerning a Control Word:

. It has to be long enough to prevent pirates from trying all possible combinations. In the case of EUROCRYPT this Control Word is 60 bit. This means that trying all the combinations, say one per NTSC frame, would take around 1 billion years.

. It has to be changed rather often to prevent subscribers from distributing the Control Words to e.g. neighbors. In the MAC application the Control Word can be changed every 256 frames (about every 10 seconds in MAC).

Two types of messages are defined for transmission of conditional access data: One for control Words (ECM) and the other type is for entitlements transmission (EMM).

## ECM MESSAGES

The Control Word is sent in the TV signal itself in a special message called Entitlement Checking Message: ECM. This ECM is transmitted every half a second to offer a quick descrambling of the signal. This message does not require significant data capacity of the signal.

An ECM message contains:

. PPID (Programme Provider IDentifier): reference of the broadcaster or Service Entity in the Smart Card. This field gives also the reference of the key to be used for decrypting the following data. This information is of course not encrypted.

. The access modes of the programme: The authorization valid to get access to this programme, the maturity level, etc. This information is not encrypted.

. The Control Words in Encrypted format: The current one to get access immediately to the programme and the one for the next 256 frames period.

. The HASH parameter which is the signature of any sensitive parameter of the message. This is to guarantee the integrity of the message, that is that nobody has tried to modify for example the access modes part.

## EMM MESSAGES

The user also needs to get his entitlements from the broadcaster. The message used for this purpose is the Entitlement Management Message: EMM. This message can contain several types of information:

. Initialization of a Service Entity
. Keys
. Entitlements
. Replacement or fingerprinting data

This message is sent within the transmission. The receiver can filter the messages for a specific user at the user address present in the message. The user addresses are memorized in the Smart Card and passed on to the receiver. There are four types of EMM:

. EMM_U: Uses a Unique address which means that this message is intended for one particular user. This unique address is 36 bits (around 64 billion potential users). A Unique Address EMM is used to send entitlements to one user. Using the capacity of one sound channel at mono medium quality in MAC, around 250 customers can be addressed their entitlements per second.

. EMM_S: Uses a Shared address which means that this type of message can be addressed to several users. This type of EMM can be used when the users having the same address are asking for the same entitlements. This address is 24 bits (about 16 million possibilities). The audience is split into groups of 256 subscribers. The entitlement itself is sent in the EMM_G and this entitlement is registered in the conditional access system only if the information available in the EMM_S for the user indicates to do so. Using the equivalent capacity of one mono medium quality sound in MAC, 1 million users can have their entitlement updates within 20 seconds.

. EMM_C: Uses the Collective address. In the current application of EUROCRYPT, only the value 0 of the address is defined. The Collective EMM is used to send replacement or fingerprinting information. These are not used for addressing entitlements and do not require a big capacity of the channel.

. EMM_G: Uses no address at all. These message are intended for the whole audience. The General EMM is received by all the users. This type of EMM is used in combination with an EMM_S for addressing a large amount of people at the same time. It is also used to send free entitlements for customers who are subscribing to the channel for the first time. The user will be able to watch some programmes during which he will receive the entitlements he paid for.

Only the secret information such as keys are encrypted in an EMM message. The entitlement for example is sent in clear format or at most scrambled to guarantee confidentiality. A HASH field as in the ECM ensures that no fraudulent operation was performed on the message.

## CONCLUSION

EUROCRYPT has proved to be reliable and secure in the current applications in Europe. EUROCRYPT can fit for new applications like Digital TV.

The EUROCRYPT standard offers many advantages for the development of Digital TV. The Digital TV will require very soon a conditional access function which EUROCRYPT could supply today. The market development of Digital TV with conditional access is also dependant on the price the customer will have to pay. With the Smart Card, the user will not have to pay for several specific receivers. The same neutral receiver, produced in large quantities at a low price, will be used by all customers for all conditional access programmes. If the broadcasters have to change the conditional access system, they just have to change the Smart Cards. These are some of the reasons why EUROCRYPT is the solution for conditional access in digital TV.

REFERENCE:

- Système d'accès conditionnel pour la famille MAC/paquet: EUROCRYPT

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# CONSTRUCTING A CONVENTIONAL HARDENED TRUNK

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## ABSTRACT

With higher bandwidths and increased gain, today's generation of active equipment consumes more power than the previous generation. The methodology behind powering this equipment, however, has remained essentially the same. Power supplies are added to the plant to compensate for the expected additional load or for potential future modifications. Along with additional power supplies, however, comes a decrease in system reliability. A power supply is dependent on the integrity of the local utility, and, based on our standard TREE AND BRANCH architecture and resistance to standby power, our reliability can never be better than theirs. To achieve satisfactory reliability, power supply cascades, or the number of power supplies serving a subscriber, must be reduced.

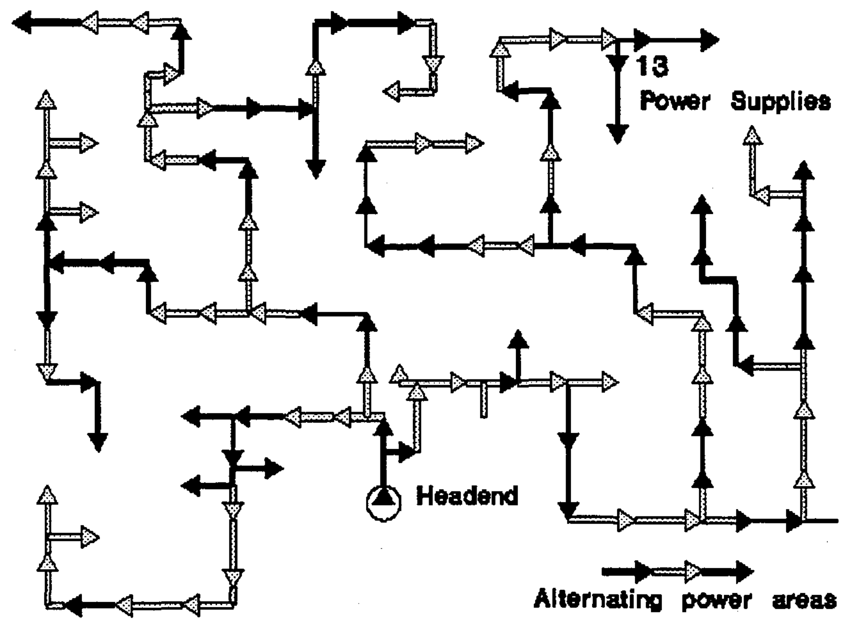
Headend operations are likewise impacted by power outages. In addition to the interruption in the cable service, the down time can contribute to headend equipment failures.

This paper introduces an alternative to the standard method of system powering. Described as a HARDENED TRUNK, it is a highly reliable, low maintenance, method of powering a cable plant. It offers relative ease of implementation, additional equipment protection, and, most of all, fewer power outages.


## INTRODUCTION

The Nashua, N. H., system was recently upgraded from 300 to 450 Mhz. The system has 285 miles of plant and 85 power supplies. Trunk spacings and cascades remained the same as before, however, the heavier loads of the new equipment seemingly dictated that power supplies be added to the plant. The equipment vendor was instructed to design for complete two-way operation, maximum equipment loads, and to load the power supplies to no more than 80% of their maximum capacity. New power supplies were simply inserted into the plant where needed and none of the existing power supplies were moved. This is common practice for an upgrade throughout the industry.

The new system, however, was less reliable. The number of power supplies had increased by 40, or almost 100%. Cascades of 25 amplifiers had accompanying cascades of 13 under - loaded power supplies, none of which were equipped for standby operation. The picture quality in the system was judged by our customers to be very good, however, it did not make up for the fact that any isolated failure of the power grid resulted in catastrophic cable outages. In common with this plant problem was a power related trend that had developed at the headend locations: A loss of power, no matter how brief, would seem to coincide with equipment failures up to two weeks later. Headend equipment failures had previously been overlooked and/or accepted as the norm as long as they were not excessive. The headend was equipped with a diesel generator to provide emergency power during outages, however, the reason for the failures seemed to be the down time while the generator reached transfer speed.



ORIGINAL DESIGN

## SOLUTION

The plant had to be more reliable. To achieve a higher level of reliability, power supplies needed to be eliminated and cascades needed to be reduced. Fewer power supplies mean fewer problem areas and less reliance on the power company. Eliminating power supplies required starting over and re-powering the entire network.

To re-power the network, the contribution of each component had to be known. Cost was a factor, both the cost to implement and the cost to maintain thereafter. Simply changing all of the power supplies to standby would be cost prohibitive and too difficult to maintain. Whatever was done, the end result had to be an actual improvement. It had to be both maintainable and reliable.

## DESIGN PROCESS

**Considerations** - The upgrade to 450 Mhz required the use of three different types of amplifier: Push Pull, High Gain Power Doubling, and Feedforward. Switching regulated power packs had been installed to power the higher gain equipment. The RF portion of the design was similarly inefficient. Maximum cable losses had been used in the design, and because this was a "drop in" upgrade, the high gain Power Doubling amplifiers were used quite often; many times when they were not needed. The high gain version of the Power Doubling amplifier not only consumes more current but also has worse distortion specifications than a push Pull amplifier. To Achieve the higher gain, two Power Doubling stages are used and the advantage of Power Doubling technology is lost.

Switching regulated power packs had been installed in approximately 50% of the trunk stations. A switching power pack is approximately 90% efficient, compared to 50-60% with a linear regulated supply. Efficiency of the power pack is important, it determines the amount of AC current used by the trunk station, which determines the voltage drop, and, in turn, the load on the power supply.

**Testing** - Each type of active equipment was bench tested to measure the **DC LOAD CURRENT**. The DC load current is the current that the module requires from the power pack. The AC power differs from the DC current depending on the efficiency of the power pack.

$$\textbf{AC POWER} = \frac{\textbf{24VDC x DC LOAD CURRENT}}{\textbf{EFFICIENCY}}$$

The cable was then tested. The plant consists of several generations of air dielectric cable. A composite loop resistance was developed using the test data. The composite loop resistance was approximately 10% better than published specifications.

**Design method** - A **CONSTANT POWER** design method was chosen based on the existence of the switching regulated power packs, that is, the equipment consumed a constant wattage regardless of the input voltage, with higher voltages consuming less current than lower voltages. This method would more accurately reflect the operation of the switching power packs in the plant. The existing power areas were then tested using the new specifications and the numbers were adjusted to fit the real world accordingly.

Using the new specifications, the entire system was then redesigned on paper. Power supplies were loaded to 90-95% of maximum, 5 ampere power supplies were added to power the sub-trunk legs that branch off of the main trunk, and existing power supplies were moved to the most efficient location. Several other important steps were taken to improve reliability.

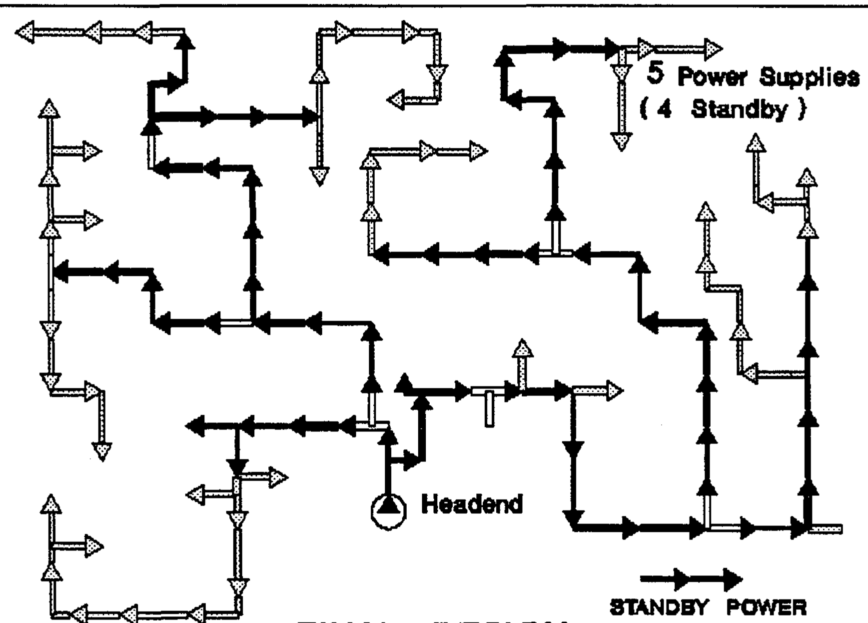
- During the initial redesign, and using actual field measurements, Push Pull amplifiers were moved to the main trunk and combined with a switching regulated power pack where the extra gain was not needed. This combination allowed for the most efficient design, low current consumption combined with a highly efficient power pack.
- A major focus was to stop power at the input of any trunk split. For example, each leg of a split is powered from a separate supply, eliminating the possibility that a power outage on one output leg would affect cable service on another. Each trunk section is therefore isolated from the others and responsible only for its own operation.
- Power supplies were moved to the location where they would have the greatest impact on the cascade and have a balanced load on either side of the power supply. Having a balanced load assures that voltage losses are kept to a minimum, and, as a result, the greatest efficiency is achieved.
- Standby power was then added to the longest trunk cascades. Having reduced the power supply cascade, a limited amount of standby power would greatly improve reliability.

The result of the new design was a much more reliable network. Power supply cascades were cut in half. There were fewer critical power supplies and loads were brought to an efficient level. Despite installing only 4 standby power supplies in the first year of the project, the placement of the units greatly improved reliability.

## IMPLEMENTATION

A three year implementation plan was developed. In the first year, 1991, 5 ampere power supplies were took on a more significant role. The standby power supplies were placed where they would do the most good. Each one powers 6 to 7 trunk stations. The sub-trunk legs are powered by standard, non-standby power supplies which power 1 to 4 trunk stations.

This year, the second phase of the project, effort was placed on deeper penetration of the standby network, making most households no more than one power supply away from a standby. AMP CLAMP technology has been installed for the "hardened" portion of the trunk and the maximum power supply cascade has gone from 13 to 6.



FINAL DESIGN

installed on the sub trunk legs. Four 15 ampere standby power supplies were then installed on the "main" trunk, and 30% of the existing power supplies were moved.

The biggest impact to system reliability occurred during this phase of the project. The main trunk

80% of the households are now only one power supply away from a standby unit. This is reasonable, when the power is out in a neighborhood, there is no need for the cable service to be operable in the same confined area. With fewer than 12 standby units installed, the maintenance crew is able to perform regular maintenance on the standby power supplies.

The trunk network now has two unique identities, one a main trunk system composed of several long cascades of amplifiers powered by a limited amount of mostly standby power supplies, and another sub - trunk network composed of several short amplifier cascades branching off of the main trunk and without standby power.

**Headend** - Headend outages and equipment failures were occurring no more than in any other system, however, it was a problem worth correcting. Two problems existed relating to headend power outages and the delay until transfer to a generator. One, the cable service was interrupted to all subscribers. In the past this was acceptable, however, today we are more critically judged and we should not allow even the briefest outage if it can be prevented.

The second problem was related to equipment failures. The headend equipment is of various types and ages. The problems were varied and some were as common as modules becoming unseated in their housings and needing to be re-seated. This problem has been with us for years. The key factor is that the occurrences were traced to within two weeks of a power event. Adequate transient protection was in place and failures were not thought to be caused by line surges.

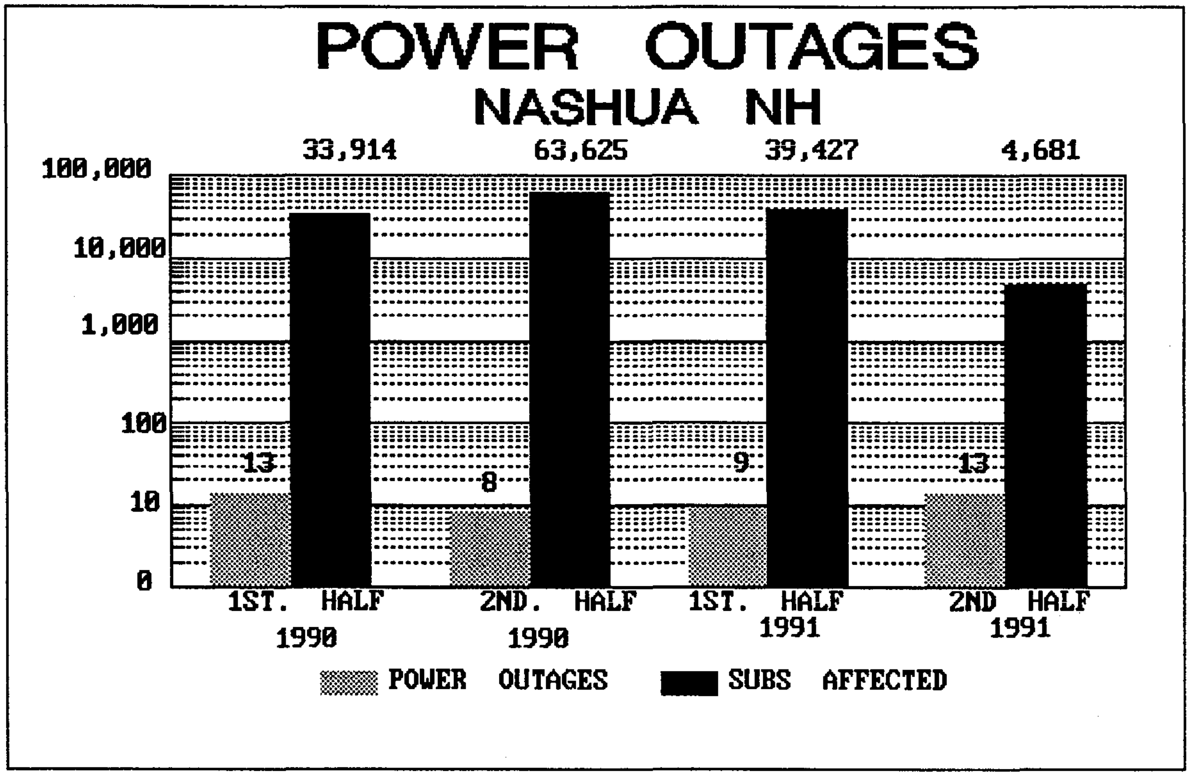
The cause of the failures was thought to be the equipment's inability to withstand brief losses of power. Equipment that had been working year in and year out developed a trend towards failure only after power was removed. The generator will transfer the load only after it has reached transfer speed, 5 - 10 seconds. If the lack of power was the cause, then eliminating the power outages in between would eliminate the corresponding equipment failures.

To solve this dilemma, uninterruptable power supplies were installed at one of the headend sites. An uninterruptable power supply, or UPS, is an instantaneous, battery powered AC source. It provides backup power when needed and is a source of additional over-voltage protection. The equipment would never notice a loss of power, nor would our customers ever see an outage. This was an enhancement to the existing backup generator. The UPS network only carries the load for the brief period between the time of utility power failure and the transfer to the diesel generator.

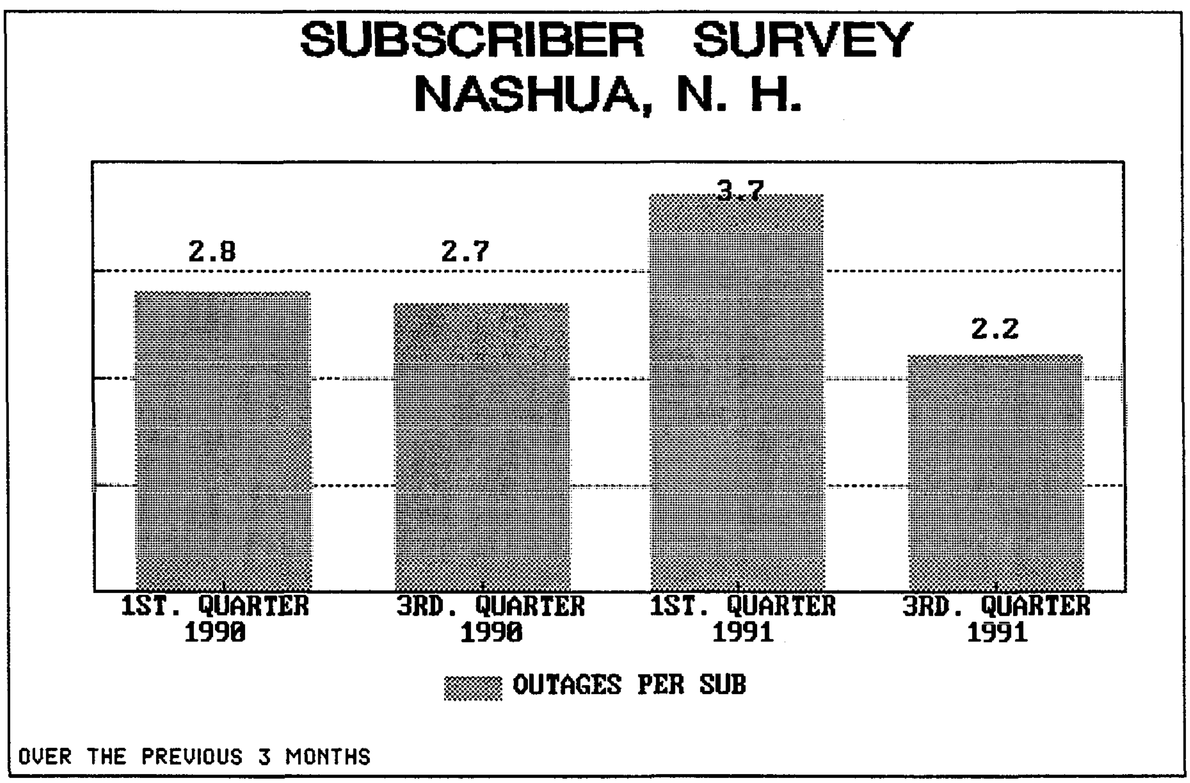
## RESULTS

Two years after the concept was developed, the system is more reliable. Some power outages have been eliminated and the impact of each remaining occurrence is now limited in scope. No longer does a power outage result in catastrophe, and headend equipment failures seemingly have been eliminated. Because there are so few standby power supplies, each one is maintained properly and with far less effort.

One year after the UPS installation, March 1991 - March 1992, headend failures have been reduced to practically zero at the selected location. Despite power outages, numerous lightening storms, and a major hurricane, the headend has not gone off and the equipment has not failed. This is a difficult concept to visualize. The results are encouraging, but the theory would need to be tested in a controlled environment to prove a direct correlation.



POWER OUTAGES
NASHUA NH
33,914
63,625
39,427
4,681
100,000
10,000
1,000
100
10
0
13
8
9
13
1ST. HALF 1990
2ND. HALF 1990
1ST. HALF 1991
2ND HALF 1991
POWER OUTAGES
SUBS AFFECTED

SUBSCRIBER SURVEY
NASHUA, N. H.
2.8
2.7
3.7
2.2
1ST. QUARTER 1990
3RD. QUARTER 1990
1ST. QUARTER 1991
3RD. QUARTER 1991
OUTAGES PER SUB
OVER THE PREVIOUS 3 MONTHS

The first graph tracks the number of power outages versus the number of customers affected. This project was begun midway through the first half of 1991 and was completed late in the second half. During the second half of 1991, far fewer customers were affected by power outages than during any other time. This reduction occurred despite a hurricane in August that seriously damaged the power grid.

The second graph shows the results of a bi - quarterly survey of our customers. Again, despite the hurricane, outages were perceived to have been fewer.

## SUMMARY

This paper outlines an approach towards system powering that can greatly increase plant reliability without consuming large sums of capital.

This trunk network is now **HARDENED** because it is better protected against outages, both power and transient related. It is **CONVENTIONAL** because it was constructed with off the shelf material and required no new technology. More than anything, it is an improved application of existing technology. It would interest those systems that are experiencing similar problems and will not be installing fiber in the near future. It can also be used as a first step towards a fiber optic installation. Regardless of the impetus, this technique is a simple solution to a problem that many system operators face.

Cost - It is important to mention that the switching power supplies had already been purchased and installed as part of the earlier upgrade project. For this project to be successful, it required that they be moved to the selected locations.

## COST

| INVESTMENT | TRUNK SYSTEM | HEADEND | PER PLANT MILE |
|---|---|---|---|
| 1991 INVESTMENT | $14,000 | $14,000 | $ 98.00 |
| 1992 INVESTMENT | $10,000 | $ 5,000 | $ 52.50 |
| PROJECTED 1993 | $10,000 | $ 5,000 | $ 52.50 |
| TOTAL INVESTMENT | $34,000 | $24,000 | $203.00 |

The basic premise is to construct a network which reflects the realities of the plant. Budgeting for the needs of tomorrow, which may never occur, requires that power supplies be added today. Standby power is an important asset, however, its use is limited due to the intense maintenance requirement. Many systems have standby power supplies that do not work because they could not be maintained.

The efficiency of the switching power supply is required to completely duplicate the results of this project. Without them, power supplies could still be moved to better locations, however, fewer of them would be eliminated because of the heavier loads. Purchasing new power packs for the main trunk will increase the cost by $100 - $200 per plant mile, effectively doubling the investment.

# CONSUMER ELECTRONICS FRIENDLY CABLE

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## ABSTRACT

This paper reviews the issue of the interface between Cable Television systems and Consumer Electronics hardware. The current status is summarized along with some possible solutions to the problems. The impact of new technologies is projected.

## INTRODUCTION

The Consumer Electronics Interface problem has been with us for a long time. It is now more serious than ever for a number of reasons including: high penetration of "cable ready" consumer hardware, increased interest in Impulse Pay Per View, IPPV, multi-channel IPPV, and Near Video On Demand, NVOD, and an emphasis in the political arena.

The promise of new technologies such as Digital Video Compression, DVC, GHz cable bandwidths, and High Definition Television, HDTV bring increased concerns about the Consumer Electronics interface. There may be opportunities to make progress if work is done during the formative stages of these new technologies.

Some progress has been made. Solutions and partial solutions have been found; but some have not been implemented.

In issues as complex as these, there are a variety of opinions and priorities. There are even misunderstandings. This paper presents my views on this complex subject.

## WHAT ARE THE PROBLEMS

The fundamental problem is one of failed expectations and a resulting frustration. Some of the subscribers' expectations come from the belief that technology should make things simple. When it doesn't, it's only human to want someone to blame. That is often the last entity to come onto the scene. In most cases, that is the cable company. A major factor in failed expectations is the use of the term "Cable Ready" by the Consumer Electronics Industry. This is discussed below.

Placing ourselves into the shoes of the subscriber, several problems with the usage of Consumer Electronics products connected to cable become readily apparent:

a) Connecting up a collection of hardware to cable is confusing and difficult.

b) The subscriber would like to use the remote controls that came with his TV and VCR—and not need to use any others.

c) Programming the VCR to tape shows is difficult. There are a number of constraints added by a cable connection that are bothersome.

d) It is difficult to know what is on cable. As more channels are added, the confusion grows.

e) Certain features of Consumer Electronics products, like Picture-In-Picture, PIP, become impaired when cable is added.

Upon initial connection to cable, the cable installer helps the consumer. Occasionally, the installer will arrive at the subscriber's residence to find a new VCR still in its box. The subscriber expects the installer to connect it and provide instructions on its use! Imagine buying a telephone answering machine from the corner drug store, bringing it home, calling the phone company and asking them to install it and demonstrate its use. While this sounds preposterous, subscribers have these kinds of expectations when it comes to cable. The cable company gets no credit or recognition for the assistance and education it provides on Consumer Electronics product installation and usage. The Consumer Electronics retailers have managed to escape responsibility for this instruction.

Remote Controls

Remote controls are probably the biggest subscriber headache. A decade ago, when a remote controlled TV was a high end purchase, the cable box with remote control rental provided a welcome service for a dollar or so a month. Two things happened to change that. First the availability of remote controls on Consumer Electronics products became pervasive. Today, only the very cheapest products come without a remote control. Second, the relatively low monthly rental for cable remote control has also grown. The subscriber is confused by the usage of the cable remote and irritated by the rental fee.

Most likely, the subscriber must use both the remote control that came with his TV or VCR and the one that came with his cable box. The TV and/or VCR must be turned on. Older, simpler products could be plugged into a switched convenience outlet on the back of the cable box. Turning on the cable box also turned on the TV. Now, most TV's and VCR's have a single "on/off" light-touch button. If the TV or VCR is unpowered at its line cord, it will not come on when repowered. The switched convenience outlet on the cable box becomes useless. The latest TV's and VCR's go through an automatic set-up procedure when powered up at the line cord and first turned on. This takes several minutes. If the cable box's convenience outlet is switched, its use with such products presents major inconveniences.

Even if the cable box has a volume control feature, the TV's volume must be properly set to avoid noisy audio or a loss of stereo separation. This is a complication most subscribers don't understand or quickly forget.

Universal Remote Controls

The advent of the universal remote control has caused more problems. The subscriber expects to be able to buy one and use it with his cable box. If the cable company disables the remote feature on the cable box, the subscriber becomes angry. Another problem with universal remote controls is that they aren't always "universal". They don't accommodate all brands of cable boxes, TV's, and VCR's. If the subscriber's cable box is not accommodated, the subscriber is likely to blame the cable operator. Sometimes the subscriber expects the cable technician to set up and provide instructions on the usage of the universal remote control. If the universal remote control doesn't have non-volatile memory, the set up codes will be lost when the battery dies. The subscriber may expect help with this, especially if he has lost the instructions that came with the device.

VCR Usage

The use of a set-top box ahead of the VCR presents further complications and aggravations and failed expectations.

The first problem subscribers face is attempting to tape shows from different channels at different times. Consecutive taping from different channels is accommodated by a large fraction of cable TV set-top descramblers, but not all. Most popular brands of these devices have models with built-in clocks and timers that allow them to be programmed to change channels. These units are similar to VCR's in their complexity of operation. Older models are like older VCR's before On Screen Displays. They rely on cryptic symbols on the channel indicator. The limiting factor is the subscribers' ability to program his VCR. If the subscriber can program his VCR, he can program the set-top descrambler using the very same skills to allow the consecutive taping from different channels. This, of course, is easier said than done.

In addition, a number of "universal remote controls" from third party vendors have timers built into them. These units will change channels at different times even if the set-top descrambler does not have this feature. This is only the case if the subscriber can master the programming of both his VCR and the extra device.

Subscribers often want to be able to tape one show while watching another. There are some limitations in scrambled cable systems. If the subscriber has only one descrambler, he may:

1) Watch a scrambled channel while at the same time taping that same channel.

2) Watch a scrambled channel while at the same time taping a non scrambled channel.

3) Watch a non scrambled channel while at the same time taping a scrambled channel.

If the subscriber has two descramblers, he can watch one scrambled channel while taping another. However, he has a remote control problem. Likely, his remote control changes both settops' channels at the same time.

The "VCR Plus" product has been very successful. It uses specially coded numbers printed in the TV guide. These numbers tell it when to issue commands to the VCR to turn on and what channel to tape. It sends the channel number commands to the cable box.

While the "VCR Plus" product works very well, there are some constraints. The subscriber must leave the VCR "off" but tuned to the output channel of the cable box and the cable box "on". Most importantly, he must put in a blank tape with adequate remaining capacity.

The most important limitation to VCR Plus usage is that the VCR Plus numbers must be available to the subscriber. If the cable operator provides them in his guide, he may be viewed by the subscriber as endorsing VCR Plus. This then invites service calls when the subscriber has problems with the device. The second constraint is that the device must be initialized. This is not a trivial procedure. In fact, it is difficult to believe that someone who can initialize the VCR Plus device can't operate the VCR! Perhaps the answer is that only one family member needs to be able to initialize the VCR Plus so all can use it. The most difficult part of initializing the VCR Plus is entering the cable channel line up. The cable operator could facilitate this by providing this information in an easy to use form. Once again, this invites inquiries if the subscriber has troubles.

Other Feature Difficulties

TV's and some VCR's come with a Picture-In-Picture, PIP, feature that operates in one of at least two modes. In the first mode, two channels can be watched simultaneously. The big picture and the little picture can be interchanged with the push of a button. The sound comes from only the big picture. The second mode fills the screen with still pictures taken from successive channels. This provides a way of scanning channels to determine which to watch. The scanning mode of usage for PIP is frustrated in scrambled cable systems by the fact that the TV does not control the cable box's tuner.

In fairness it must be noted that only the most expensive TV receivers have two tuners built into them for use with PIP. Most intend for the consumer to use the base-band outputs of his VCR or the output of a video camera for the second input. This results in the need to use the VCR's remote control for the PIP display and the TV's remote control for the normal display. When the unit is used with cable, the same options remain. Most subscribers do not understand this and blame cable for the complications.

In the case where a viewer wishes both the main picture and the PIP picture to come from scrambled channels, two descramblers are required. MultiPort technology can facilitate this function.

It cannot scan channels and provide still pictures in a matrix filling the screen.

## THE POLITICAL PROBLEM

Senator Leahy has proposed amendments to cable bills and has proposed stand alone bills involving the Consumer Electronics interface. The following quotes from the Congressional Record give his opinions:

"...they show little regard for their customers when they choose a means of protection that will sabotage the customer's television and VCR."

"...you will not be able to use any of those features you paid for. But as far as the cable company is concerned, that is your hard luck."

"My amendment is designed to create more user-friendly connections between cable systems on the one hand and televisions and VCR's on the other so that consumers will actually get to use the TV and VCR features they paid for."

"...require cable operators to allow customers to buy their own remote control units from any source rather than having to pay $3 or $4 a month —month after month, year after year—for a remote control that probably does not cost more than $30..."

"...to create a user-friendly connection between cable systems and consumer electronics is more important now than ever before."

A lot of what is in the amendment and Bill is emotional and an oversimplification of the situation. Yet these comments are representative of what subscribers feel. While we can argue with the details, the big picture of discontent and frustration is undeniable.

## "CABLE READY"

The terms "cable ready" and "cable compatible" are frequently applied to TV receivers and VCR's. Major cable subscriber frustrations are due to this terminology being used too loosely by the Consumer Electronics industry. Purchasers of these products are often led to believe that these products can be directly connected to the cable system without loss of functionality of either the cable service or the features included in the TV or VCR. This is generally not the case. There frequently are technical deficiencies in these products that impair performance when they are directly connected to the cable. Consumer Electronics manufacturers should not unilaterally declare their products to be "cable ready" without the consent and concurrence of the cable system for which they are supposedly "ready"! This is only logical and fair. The Consumer Electronics industry shouldn't be allowed to call products "cable ready" if they don't work properly when directly connected to cable!

The concept of "cable ready" is really very straight forward. If a product is truly "cable ready", it can be connected directly to the cable system and a) not interfere with the reception of others, b) provide all the services the subscriber has paid for without the need of additional hardware installed between the cable system and the product, and c) comply with the FCC rules concerning radiation.

The goal is simple: avoid the need for a set-top cable box. If a set-top box is needed to enjoy a paid-for cable service, the TV or VCR is not "cable ready".

For a Consumer Electronics product to be truly "cable ready", there are a number of basic technical requirements:

1) The TV or VCR must **conveniently** tune all channels offered on the cable system. Otherwise a set-top converter is required for access to channels the subscriber has paid for.

2) The tuner must be of sufficient quality to function with **all the channels simultaneously available at its input terminals without introducing distortions** or noticeable noise. The tuners used in cable converters are generally of higher quality than those used in TV receivers or VCR's in order to accommodate these needs.

   a) If the tuner is not of adequate quality, it will combine signals from several channels in a manner that produces disturbing moving background bars and patterns in the picture.

   b) Cable converters use a more expensive "double conversion" tuner that eliminates "image response". Without this added expense, an unwanted channel's signal may be mixed with the desired channel, distorting the picture.

   c) Cable converter tuners typically have a lower "noise figure" which introduces less "snow" into the picture than a TV or VCR tuner.

   d) The tuner must not feed back extraneous interfering signals into the cable system to cause reception problems on other receivers.

3) The internal circuits of the TV or VCR must be adequately shielded so as not to pick up signals off-air directly. When this shielding is inadequate, signals directly picked up off-air are mixed inside the TV (or VCR) with signals from the cable producing an unpleasant (and sometimes unwatchable) mess. In

many of these cases, the only solution is to add a set-top converter with its superior (and more expensive) shielding.

4) If the cable system uses scrambled signals, the TV or VCR must accommodate a descrambler that can be plugged into the rear of the TV or VCR to allow descrambling after the TV or VCR's tuner and remote controls. This approach has been defined by the Electronic Industries Association's (EIA) and the National Cable Television Association's (NCTA) Joint Engineering Committee and has been endorsed by the American National Standards Institute, ANSI, as the ANSI/ EIA 563 MultiPort. Without the MultiPort plug, a set-top descrambler is required to give access to scrambled signals.

5) If the cable system uses two-way cable technology for conveniently ordering Impulse Pay Per View, IPPV, services, the MultiPort implementation must include the pass through of remote control signals to the descrambler module. In the case of multiple start times for movies, the descrambler module must be able to send remote control signals to the TV or VCR's tuner. Otherwise a set-top box is required to enjoy IPPV services.

It will be appreciated that this is a situational definition. For example, in the new 150 channel system in Queens New York, no existing TV or VCR can satisfy the definition of "cable ready" since none can tune the 1 GHz spectrum containing 150 cable channels. For subscribers to be able to enjoy all the signals they have paid for, a set-top converter must be supplied by the cable operator. As another example, the same TV receiver that gave acceptable performance in a suburb far away from any broadcast television transmission towers may require a set-top converter to reject these signals if the subscriber moves near a broadcast tower. If the TV receiver's internal shielding is inadequate, direct pick-up interference may be experienced which spoils the picture reception. The only solution available to the cable operator is to install a set-top converter with superior internal shielding.

Another source of confusion generated on the TV sales floor is the manner of specification of the number of channels a "cable ready" television can tune. The number is given as the sum of the broadcast VHF (12) plus VHF (69) plus cable channels that can be tuned. A purchaser who is told that the TV will tune 116 channels can be forgiven for being impressed and thinking that it is adequate. Yet this only accommodates 35 cable channels. More and more, this is inadequate. Under this counting scheme, a "cable ready" TV for use in the Queens New York 150 channel system needs 12+69+150 = 231 channels!

Some VCR's sold as "cable ready" have sixteen little wheels behind a door. Each wheel can be tuned to one channel. The remote control selects which wheel controls the tuner. While the VCR may be able to tune all cable channels, it can only do sixteen conveniently. Will this make for a satisfied customer?

It is not well recognized that the invention of the cable converter was not to tune more channels. The first converters tuned less than twelve channels. The cable converter was invented to overcome deficiencies in TV receiver tuners. They were meant first to combat the direct pick-up problem described above. Then, as more channels were added to cable service, the cable converter took on added technical burdens. The cable converter was required to counter the effects of non-linear performance and "image response" of less expensive tuners. Improved noise performance is also important. Lastly, the tuner must not back feed interfering signals into the cable system.

Without more discipline in the use of terms such as "cable ready" and "cable compatible", there is the potential for continued consumer anger, confusion, and losses as money is spent on product features erroneously thought to be usable when directly connected to cable.

## CONSEQUENCES

The obvious consequences of Consumer Electronics Interface problems are frustrated and unhappy subscribers and the resulting political pressures. More subtle consequences also exist. Pay penetration is reduced because subscribers simply don't want the box. This is also true of IPPV sales.

Another consequence of the poor Consumer Electronics Interface with cable is that fewer channels are scrambled than might otherwise be the case. Consequently, there is more signal theft.

Some cable operators have undertaken a "de-addressing" campaign in response to the subscriber backlash over interface problems.

## POTENTIAL SHORT TERM SOLUTIONS

There are a number of potential short term solutions and partial solutions that should be considered for implementation as soon as possible.

Perhaps the two most important are MultiPort and a Digital Program Guide. MultiPort is a solution waiting to be implemented. It requires no further development, only cooperation between industries and commitment. This doesn't mean MultiPort cannot evolve to even more useful versions. Yet this potential for upgrade should not get in the way of a speedy implementation to help those presently living in scrambled cable systems. "Backwards compatible" upgrades that accommodate new technologies while continuing to serve older implementations are a feature of the Electronic Industries Association's standards process.

A second short term solution is the Digital Program Guide. This is an urgent matter because a number of agencies are working on conflicting versions. This has the potential of becoming another lost opportunity for cooperation between the Cable and the Consumer Electronics industries.

A third short term solution is the cable set-top box with two descramblers in it. One descrambler is intended for the TV receiver while the other serves the VCR.

Yet one more short term solution is the in-line descrambler that descrambles one channel at a time and places it on the cable at the point where the cable enters the house. All TV's and VCR's receive the descrambled channel just as if it were a trapped service.

## MULTIPORT

### What is MultiPort?

"MultiPort" is the commonly used name for the technical standard for the connection of descramblers to the backs of TV receivers and VCR's that are equipped with a special plug. The standard was developed and tested over a period of several years by engineers from the Cable and Consumer Electronics industries. The Joint Engineering Committee that did the work was sponsored by the Electronic Industries Association, EIA, and the Engineering Committee of the National Cable Television Association, NCTA. The standard developed by the EIA/NCTA Joint Engineering Committee was submitted to the American National Standards Institute, ANSI, and is now formally know as "The EIA/ANSI 563 Decoder Interface Standard".

The name "MultiPort" comes from the original intention of the Joint Engineering Committee to develop a standard that would have multiple applications in the consumer, cable, computer, and other related video fields. At one time the standard was referred to as "IS-15". This was while it was an "Interim Standard". It is no longer interim. It is a fully negotiated and accepted technical standard between the engineers of the two industries.

The MultiPort plug has 20 pins and looks somewhat like the plugs on the back of computers. It allows access to the internal circuitry of television receivers and VCR's in a manner that facilitates the operation of descrambler circuits. The video and audio signals go into and out of the television receiver or VCR and then into and out of the descrambler. In addition, a very critical signal is conveyed to the descrambler: the Automatic Gain Control, AGC adjustment. Without this signal, descrambling would not be possible for most systems.

An optional enhancement allows the television receiver's or VCR's remote control to communicate with the descrambler and the descrambler to communicate with the tuner in the TV or VCR. This means that Impulse Pay Per View, IPPV, ordering with the TV or VCR's remote control is accommodated. In addition, the descrambler can "force tune" the receiver to the correct channel to make certain functions such as Emergency Alert or pre-ordered Pay Per View programs easy to use.

Because video signals exit and enter the

MultiPort plug, such functions as On-Screen Displays, OSD's, for assisting the subscriber in ordering PPV and for electronic program guides are possible. In fact, because the video signals do not have to go through the tuner of the television receiver, the OSD will be much crisper and easier to read.

It is possible to connect multiple devices to a TV or VCR so that, for example, a Captioning for The Hearing Impaired decoder and a Cable Descrambler can be used.

What Are The Advantages of MultiPort?

The principal advantage of MultiPort is that is allows the descrambler to be connected **after** the TV's or VCR's tuner and remote control circuitry. **Scrambling becomes transparent**. The consumer regains all the functionality of his hardware. It is just as if the TV and/or VCR were connected to a video source which had no scrambling. In particular, the VCR's tuner is once again under the control of the VCR's timer so different channels can be easily recorded at different times.

The MultiPort descrambler avoids duplication of the TV's or VCR's tuner, remote control, and channel indicator. In addition it doesn't require a "remodulator" to condition the signal so it is acceptable to the TV or VCR. Not only do these missing items save money, they reduce the "bruising" the signal undergoes as it is unnecessarily processed in the conventional set-top descrambler. A further advantage of MultiPort is that the descrambler goes in back of the TV or VCR and thus is out of sight. This means its cabinet can be simpler with less attention and expense devoted to esthetics. The device is called a "set-back descrambler" rather than a "set-top". All of these factors will result in lower costs and significantly higher reliability. The latter advantage comes from two factors: 1) there are fewer components to fail, and 2) the amount of heat and power consumption is significantly reduced. Heat is the main killer of electronic components.

Isn't It Obsolete, What about Compression?

Some have said that the time for MultiPort has past. The world has changed and we now have GHz cable, Near Video On Demand, NVOD, and Digital Video Compression. Their claim is that MultiPort cannot accommodate these new needs.

This not true. These arguments are a red herring based on misunderstanding.

According to Paul Kagen, there are currently about twenty million U.S. TV households which have addressable, analog descramblers. That means that **twenty million homes could benefit from the availability of MultiPort on their next purchase of a TV receiver or VCR.** This is a number which grows every year. When scrambling is installed in a cable system, it remains in place for ten to twenty years. This is because the capital investment and the commitment to training, inventory, etc. is huge. Almost certainly, any technological advances which occur in the next several years will not change this. Currently, the only electronic signal protection technology being purchased in volume for use in cable is analog based scrambling which can benefit from MultiPort. This will remain the case for a long time.

What about Digital Video Compression, DVC? This is a promising technology. At present no products exist to be purchased. The most optimistic scenarios promise hardware in three years. Initially, DVC products will be very expensive. The rate of penetration will be slow. It will most probably be a decade or more before the number of DVC units in use exceed the number of analog descramblers which would benefit from MultiPort. The mere promise of DVC is not a reason to shy away from a MultiPort implementation which will help millions of subscribers for more than a decade.

What about HDTV? This technology also is a few years away from implementation. Because of the high cost of HDTV receivers, it is likely that penetration of HDTV receivers and VCR's will take years. It is probable that HDTV penetration will lag the penetration of DVC units.

The EIA has a long history of compatible upgrades of standards. A standard is said to be "backwards compatible" if it retains compatibility with older units when it is upgraded to accommodate new technology. EIA/ANSI 563 should be implemented as soon as possible. Simultaneously, the Joint Engineering Committee should begin work on defining backwards compatible advanced versions. These will probably be called EIA/ANSI 563 A, B, C, etc. just as other standards have suffix versions (such as the computer peripheral EIA standard: RS 232 C). This approach allows consumers to benefit from standards while providing for future upgrades.

Why Not Interdiction?

Some have said that MultiPort should not be implemented because Interdiction should be encouraged instead. This too is a specious argument which would deprive subscribers of an early solution to interface problems while promising a tentative future benefit. In fact, MultiPort does nothing to prevent Interdiction when, and if, it ever becomes practical. (Those who are interested in a Consumer Electronics friendly environment can't help but be enthusiastic about Interdiction, but a dose of reality is also important! Analog scrambling will be with us for a long time. Interdiction may or may not make an impact in that time. Both should be encouraged where they make sense.)

Interdiction is the process of protecting signals by jamming them with interfering signals when the subscriber has not paid. The hardware is usually located outside in the cable plant. The advantage is that the paid-for signals enter the home free of scrambling and can be simultaneously used on all TV receiver and VCR's.

There are several problems with Interdiction.

Most interdiction approaches result in the cable signals being "in the clear", that is unscrambled on much of the cable plant. In the section below, it is explained that there is a technical need to scramble even broadcast channels in some situations. Interdiction is not compatible with that need.

Interdiction is a technology which has yet to become practical. Only a few products have been offered for sale. None have achieved success. This is an expensive approach which adds more electronics and power consumption to the cable plant. This can only impact reliability and maintenance expenses in a negative way.

A cable system which is analog scrambled in a manner that is compatible with MultiPort will not likely change out to an Interdiction approach until the investment in scrambling hardware is depreciated. Any other approach is simply wasteful and will raise subscriber costs.

Technical Need to Scramble Broadcast Channels

In cable systems in large cities with Multiple Dwelling Units the cable system is often installed from roof to roof. This is the only practical method of installation in these situations. Unfortunately, people climb to the roof and tap into the cable to steal the signals. There are technical problems caused by this which overshadow the loss of revenue to the cable system and the loss of franchise fees the city government expects. When these illegal installations are done, they nearly always are done in a manner which:

a) damages the cable, causing expensive repairs and occasional signal outages

b) allows signal leakage which is a hazard to aircraft navigation and communications, the emergency communications of police, fire, ambulance, etc., and commercial uses of radio. The FCC has rules about the maximum allowable leakage. When unauthorized installations are made, these rules are almost impossible to meet.

c) allows interfering signals to enter the cable and disturb the reception of legitimate subscribers, distorting their paid-for pictures

The only way to discourage these illegal and dangerous installations is to ensure that those who make these installations derive no benefit. All signals which would be attractive to the illegal installer must be scrambled on all parts of the cable plant. This includes basic and local broadcast signals as well. This precludes the use of such technologies as interdiction. Unless this is done, hazardous conditions will be caused by those acting illegally and paying-customers will suffer degraded service.

Why Hasn't MultiPort Been Adopted by Now?

To be successful, MultiPort requires the cooperation of Consumer Electronics manufacturers, cable operators, and cable descrambler manufacturers. MultiPort has suffered from a lack of patience. A few years ago a number of TV manufacturers, RCA in the largest volumes, produced receivers with MultiPort plugs. A number of box manufactures, Zenith most notably, produced MultiPort descramblers. Several cable companies stocked and supplied MultiPort decoders to those who requested them. Only Bang & Olufsen produced VCR's with MultiPort.

At that time the numbers were just too small and the patience too thin. The participants lost heart and gave up. The severity of the Consumer Electronics Interface with cable was not as well appreciated as it is now. At this point, each of the participants need assurance that the other two will do their share. This might best be accomplished with a requirement to implement that gives comfort that progress will be made.

The party with the least enthusiasm for MultiPort has been the supplier of descramblers. Several of these suppliers see MultiPort as a lower cost version of what they already sell and are reluctant to give up the incremental revenues. It is our conviction that there are several attractive alternatives for descrambler manufacturers: 1) the lower cost and increased consumer friendliness will stimulate a lagging market which has suffered from subscriber back lash over the inconvenience of it all, 2) subscribers will want one MultiPort device for the TV receiver and another for the VCR, and 3) cable operators will be able to spend more for the program protection circuitry since they no longer will have to purchase tuners, remote controls, channel read-outs and remodulators.

Those who are worried about business disruptions should take comfort in the fact that the sales of TV's and VCR's takes place at a steady and deliberate pace. The requirement to provide MultiPort will not change any business overnight. It will, however, provide an option for the consumer who wishes a cable-friendly installation and is in the market for a new TV or VCR. That is something that is simply unavailable now!

The Spoilers

It is important to recognize that a MultiPort equipped TV or VCR which suffers from the other deficiencies described in the section on the definition of "Cable Ready" will still require a set-top converter. The benefits of MultiPort will be lost if the TV receiver or VCR fails the other requirements for being truly "Cable Ready".

Industry Support

The CableLabs Board of Directors during its December 12, 1991 meeting endorsed the need for a Definition of "Cable Ready" in any Consumer legislation. Support for MultiPort was included in this definition in the case of scrambled cable systems. The NCTA Board supported MultiPort during its January 13, 1992 meeting.

## DIGITAL PROGRAM GUIDE

The Digital Program Guide, DPG, is a concept which would provide a carefully specified data stream down most cable systems. This information could be used to greatly simplify the use of cable. The data stream would provide data about cable programs and their schedules as well as a few details about their content.

In-home hardware would use On Screen Displays, OSD's, to present the information to the subscriber in a convenient format. The subscriber would be in a better position to take maximum advantage of what cable has to offer. This would increase the value of the cable subscription and the satisfaction of the subscriber.

More advanced versions of the DPG would use techniques to control the VCR directly. A subscriber would not have to program his VCR in the conventional, confusing way. It would be significantly simplified.

Still more advanced versions would use memory and microcomputer power to advise the subscriber about program which may be of interest. Viewing would be enhanced.

Free or Pay?

A fundamental question is whether the DPG should be "free" to all subscribers or a separate pay service. Since nothing is really "free", the question boils down to whether its costs are lumped into the basic cable bill or are a separately charged item.

The arguments for a separate charge are based on the fact that this would be a truly useful and valuable service. It would greatly enhance the utility of cable. Some would argue that on that basis it should bring revenue. Still another argument in favor of a fee for this service is that those who don't want it won't pay to subsidize its use by those who do use it. Yet a third argument is that other services can be added to it to create a new "information age" business. The investments made by some cable operators in guide services further clouds the issue.

Estimates on the price of a guide service range from $12 per month to $2 to $4. The high end has been proposed by entrepreneurs outside the cable industry with little first hand knowledge of cable economics.

The arguments for a "free" service are based on the benefits of universal usage and the desire to avoid the hassles of a pay service. If the DPG improves subscriber satisfaction as much as anticipated, it may be in everyone's best interests for it to be included in the basic service. Not only will it help subscribers understand the services they have signed for, but is will help them understand what they are missing. As such it may be a powerful promotional tool.

There is good reason to wish to avoid the hassles of a pay service when the expected fee is just a couple of dollars a month. Making the DPG a pay service requires:

a) addressability

b) encryption or scrambling

c) billion, and

d) a continual fight against piracy.

Those functions may be too expensive for a $2 to $4 per month service. Put another way, these functions would eat up most of the margins of such a service.

Perhaps the strongest argument for a "free" service is that it would encourage the Consumer Electronics industry's participation. This would save the cable industry from yet another capital investment for in-home hardware while yielding significant advantages.

DPG in TV's and VCR's

The best place for the DPG is in the VCR. The second best place is in the TV. Next on the list is a subscriber owned stand alone box. The least desirable location is in the cable set-top box. Here it adds capital costs, heat build up, and reduces reliability.

If the DPG is in the VCR, it is best situated to make VCR operation easy. The subscriber moves the cursor next to the program he wishes to record and presses a "record" button. The VCR has all the information it needs to implement the command. Unlike external devices (such as the VCR Plus described above), a built-in DPG has access to information about the current status of all internal controls. For example, many VCR's have a manually operated switch which determines whether an input from the video terminals will be recorded or an output from the VCR's tuner. A built-in device will have access to this information. An external device has no way of knowing.

A built-in DPG knows if a tape is present. If the VCR records time codes on the tape, the DPG can even know if enough tape remains to accommodate the desired recording.

It is possible to put a "stick on" magnetic stripe on the tape cassette. Loading the cassette into the VCR moves it past a simple, inexpensive magnetic head which reads the magnetic stripe. Ejecting the tape again moves it past the head allowing more information to be added to the magnetic stripe. The information on the magnetic stripe comes from the DPG. An index of what is on the tape and its location is provided. This can be displayed with an OSD. This kind of functionality will not be possible in an external unit.

It is not likely that the Consumer Electronics industry will pursue a built-in DPG if its signal is not (almost) universally available.

Advanced DPG Features

One of the exciting possibilities of the DPG concept is to allow the creativity of cable vendors and Consumer Electronics manufacturers to enhance the utility of the DPG. They can decide how the guide should look on the screen and how its information should be accessed. Different brands will implement different "looks" and "feels" to the operation of the guide. Features can be added in higher priced models. At the bottom of the line might be a one-day guide while the top models may do a month.

Since the DPG contains some content information, a "Mood Guide" may be implemented. The subscriber can indicate the kind of programming he is "in the mood" for and the advanced DPG would inform him of his options. A still more advance version could keep tract of what has been watched and "learn" the viewer's preferences. Based on this analysis, the DPG could advise the subscriber of what is likely to be appealing. This could be done separately for each family member.

Urgency

There are a number of entities working on proprietary versions of a DPG. If these issues are not settled soon, the opportunity for an industry wide universal DPG will be lost. We will be faced with a multitude of incompatible approaches with little hope of having a DPG built into Consumer Electronics products.

## VCR DESCRAMBLER

A number of cable descrambler vendors are working on versions of their products with two descramblers and two tuners. The motivation is to eliminate the constraint which prevents the simultaneous taping of one scrambled channel while viewing another scrambled channel.

As an alternative use of such a device, the PIP feature can be used with two different scrambled channels.

The advantages of this device is that is can simplify VCR usage. It may be slightly cheaper than two set-top descramblers. It solves the remote control problem in that two boxes with the same remote control codes is avoided.

The main operational disadvantage appears to be an inventory problem. A secondary disadvantage involves generating sufficient industry support to encourage manufacturers to design and produce such a unit in a cost effective manner. An industry position expressed in purchase orders is needed.

## IN-LINE DESCRAMBLER

The in-line descrambler occurs in at least two forms. The first removes a scrambled channel from the cable, descrambles it, and then reinserts it back into its original spot. The subscriber enjoys the same kind of benefits as in the case of negative traps. The major difference is that this is an addressable approach. The re-insertion of the signal is difficult to accomplish without harming adjacent channels.

The second in-line descrambler approach places the descrambled channel in another part of the spectrum. A major advantage of this method is that it is easier to accommodate multiple channels. Only one channel is placed on the in-home cable system, but it can be chosen from several scrambled channels. This approach accommodates IPPV. Selection can be made either with a remote control whose signals are relayed to the in-line descrambler or using Automatic Number Identification, ANI, and phone calls. Techniques like this have been proposed for the 1992 Summer Olympics.

## NEW TECHNOLOGY, NEW URGENCY

There are a number of technical developments which make the issue of Consumer Electronics interface urgent.

Near Video On Demand

The initiation of Near Video On Demand, NVOD, in Queens New York along with other forms of multi-channel IPPV has increased the interest in addressable signal protection. If we are to avoid a subscriber back lash, we need to implement these services in a Consumer Electronic friendly way. Otherwise, the growth of these services will be impaired and the results disappointing.

Digital Video Compression

Perhaps the most exciting current topic in cable is Digital Video Compression, DVC. It promises hundreds of channels for a variety of applications. Implementing DVC in a Consumer Electronics friendly manner will be a significant challenge. Once again, the promise is of more set-top boxes to get in the way of the remote control and VCR timer. The cable industry would be well advised to search for solutions which avoid new subscriber hassles.

HDTV

Fortunately, HDTV is building on the Multi-Port experience. A couple of industry committees are working to minimize the problems NTSC has experienced in signal security.

## CONCLUSION

The importance of the Consumer Electronics interface with cable is more important than ever. Old solutions need to be implemented. New solutions for new technologies need to be found and implemented in a backwards compatible manner. If cable services are to continue growing, this important aspect of subscriber satisfaction needs serious attention and commitment.

## THE AUTHOR

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He has nine patents issued. He has presented over a hundred papers and published about fifty, two of which have received awards from the Institute of Electrical and Electronic Engineers.

He is currently chairman of the Technical Advisory Committee of CableLabs. He serves on the Executive Committee of the Montreux Television Symposium. Walt is a Fellow of the IEEE, a Fellow of the Society of Motion Picture and Television Engineers, and a Senior Member and member of the Board of Directors of the Society of Cable Television Engineers. Other memberships include Tau Beta Pi, Eta Kappa Nu, and Beta Gamma Sigma. He has served on several industry standard-setting committees. Current interests center on cable-competitive technology, the consumer electronics interface with cable, video compression, and High Definition Television.

Walt received the 1987 NCTA Vanguard Award for Science and Technology. Communication Engineering and Design magazine named him 1990 Man of the Year.

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Walt has a Ph.D. in Electrical Engineering from Illinois Institute of Technology dated 1969. The BSEE and MSEE are also from IIT. He received an MBA from the University of Chicago in 1979. He has taught Electrical Engineering in the evening division of IIT for seven years.

Hobbies include reading, wood working, photography, skiing, amateur radio (WB9FPW), and helping with his wife's horses.

# CONTROLLING THERMALLY-INDUCED LEVEL CHANGES IN TAPPED-FEEDER DISTRIBUTION SYSTEMS

Peter Deierlein
Magnavox CATV

*Abstract*

*With the introduction of fiber-optic trunk systems and fiber-to-the-bridger system design concepts, tapped-feeder cascade lengths have doubled. As a result, feeder performance has become a more critical factor in overall system performance. Seasonal temperature changes can cause significant variations in distribution signal levels. Considering the presence of taps, tapped-feeder level changes have different characteristics than level changes in conventional trunk lines. These changes are magnified by the longer feeder cascades which are common in modern system designs. Consideration of feeder level changes must therefore become an integral part of modern system performance evaluation.*

## INTRODUCTION

Conventional CATV level control systems are used only in Trunk amplifiers, where precise RF level control is required to achieve maximum performance over long cascades. These control systems compensate level changes that result from thermally-influenced variables: cable attenuation and amplifier gain. Closed-loop control systems are required because minor open-loop errors would become major when multiplied by the cascade length. To achieve the best combination of cost and performance in conventional systems, tapped-feeder distribution cascade lengths are kept short, negating the requirement for active feeder level control.

Modern fiber-optic distribution systems eliminate long trunk cascades by substituting a single low-attenuation optical fiber, with the output of the optic receiver often directed immediately into the tapped-feeder system. Although fiber-optic performance advantages permit additional cost savings by allowing longer feeder cascades, the feeder signal levels may then require active control in order to realize the additional potential. Besides being complex and expensive, trunk-type level control systems are not intended to correct for changes in tap attenuation. This paper documents the effect of taps on thermal level stability, and suggests new approaches to controlling signal levels in tapped feeders.

## EFFECT OF TAPS ON CATV SYSTEM PERFORMANCE

CATV systems are designed for "zero gain," with the gain of each amplifier station replacing the loss of the transmission-line segment which precedes it. The transmission-line segments are subdivided into the basic categories of "Trunk" and "Feeder," each consisting of coaxial cable plus a combination of splitters, taps, and other hardware necessary to meet system requirements.

In Trunks, taps are seldom used, and the sum of all losses other than cable loss is generally low. To maintain a constant amplitude frequency response through long cascades, amplifiers and their control systems compensate for the cable's well-documented characteristics.

In Feeders, taps are numerous, and their combined loss can dominate the total loss of the transmission-line segment of which they are a part. A tap's insertion loss differs significantly in the frequency domain from that of an equivalent length of cable, and the ratio of tap loss to cable loss can vary considerably from segment to segment. This is particularly apparent when comparing a segment in a rural area to one in a densely populated urban center.

While adjustable amplifier frequency response permits compensation of a wide range of initial tap-to-cable loss ratios, the effect of temperature change on frequency response due to the variability of these ratios is not clear. The characteristics of these thermally-induced level changes must clearly be understood before control can be provided. The first step in this process is defining tap performance over the applicable temperature range.

## MEASURING TAPPED-FEEDER PERFORMANCE OVER TEMPERATURE

Theoretically, tap insertion loss should be constant regardless of frequency, and should not be temperature sensitive. However, due to the realities of broadband transformer performance, actual tap performance is loosely defined. Tap insertion loss specifications are generally given as either a maximum limit or a "nominal," with no specific data on thermal stability.

### Tap Measurement Procedure

A high-precision method of tap insertion loss performance was developed. Magnavox 8000 series multi-pads were connected in strings of 5 equal-valued types, using 6 dB pads as interconnects. After making insertion loss measurements over frequency and temperature in an environmental test chamber, the test was repeated with only the 6 dB pads. The process was repeated using various values and types of taps, and the data was captured by a computer. This procedure yielded averaged insertion loss data with over 4 digits of precision, permitting highly accurate computer simulation of long feeder cascades.

Measured tap insertion loss changes over the -30F to 100F temperature range generally amount to well under .2 dB per tap; thus the requirement for a high degree of precision. While this may seem to be an insignificantly small change, the total change over an entire feeder cascade is significant. In Figure 1, tap data from 2-way, 4-way, and 8-way tap measurements are averaged and plotted along with cable characteristics. In both cases, insertion loss change over the -30F to 100F temperature range is plotted as a percentage of maximum insertion loss at 68F.

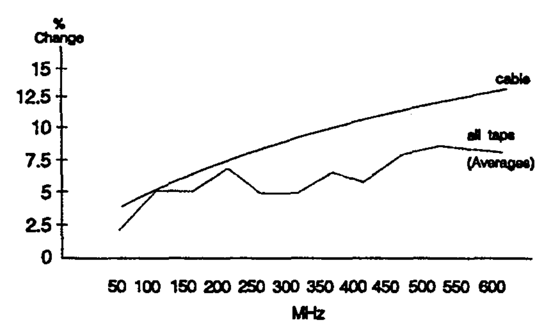



*Figure 1.*
*Insertion Loss Change Over 130F, Relative to 600 MHz Loss*

Figure 1 graphically illustrates two important differences between cable and taps. The tap's thermal change function is complex over frequency (even after averaging), and the average change is lower. Both of these factors have a potential impact on the performance of the thermal compensation system. Figure 1 shows the average thermal change function of all the taps tested; characteristics of individual taps are considerably more erratic and vary between types and lots. The thermal change function is therefore predictable only in a general case.

### Tapped-Feeder Analysis

The type of tap used is generally dictated by subscriber density in the area served, and feeders can range in length from a few hundred feet to over a thousand feet (between amplifiers). The impact of this is illustrated in Figure 2, where the thermal characteristics of three hypothetical feeders have been plotted using the measured tap data. The high-density model has a total of 64 subscriber drops over 333 feet using 8-way taps; the medium-density model has a total of 20 drops over 721 feet using 4-way taps, and the low-density model has a total of 6 drops over 937 feet using 2-way taps. In each case, the cable length was adjusted for 25 dB loss (taps plus cable) at 600 MHz using .412 cable at 68F. Longer distances are possible using low-loss cable, which does not alter the plot. The data is plotted to show thermal change in dB over the -30F to 100F temperature range, with a plot of trunk characteristics for reference.

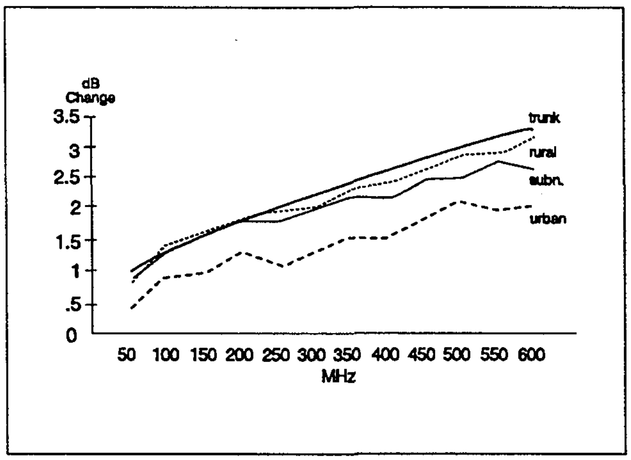



*Figure 2.*
*Change Over 130F (100F -30F)*
*25 dB Span @ 600 MHz (68F)*

In Figure 3, the same data used in Figure 2 has been normalized to the trunk cable characteristics, and shows level change deviation over 4 feeder spans. This models a typical four line extender cascade (not including line extender anomalies), and helps illustrate the thermally-induced frequency response distortion due to taps. The tap effects are shown over a long cascade to model fiber-to-the-bridger applications.

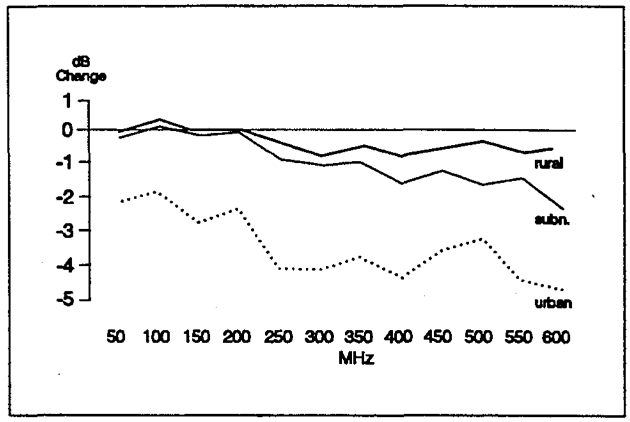



*Figure 3.*
*Cascade of 4, Change Over 130F*
*- Relative to "Ideal" Cable Compensation*
*- Actual Open-Loop System Would Not Be This Good*

Keep in mind that Figure 3 does not show the total change, only the difference between cable and tapped-feeder (The reference line represents the change in 100 dB of trunk cable). Tap loss in the low-density "rural" and medium-density "suburban" models is generally less than 35% of the total and shows little deviation from the cable except at high frequencies. The distortions in the high-density "urban" model are more severe due to the large number of taps, which comprise over 70% of the total feeder loss. While a closed-loop level control system could hold the levels constant at one or two frequencies, the response irregularities would remain in all three cases.

Effect of Amplifiers

Thermally-induced amplifier gain change differs from that of Trunk or Feeder in that the amplifier gain change is relatively constant over the active frequency range. The gain change varies based on the number and type of hybrids used, and is typically between 1 and 2 dB over the 130F range.

## CONTROL SYSTEMS

"Closed-loop" control systems are defined as those where the controlled variable is fed back to be compared with a reference. Hence, the term "feedback control system" is also used to define such systems, and the term "open-loop" defines control systems which do not use feedback. While open-loop control systems have stability, cost and simplicity advantages, their usefulness in CATV level control applications is limited due to relatively poor accuracy and linearity.

In CATV applications, closed-loop systems are generally referred to as "automatic" controls. Actually, CATV amplifier systems have true closed-loop control of only one or two "pilot" channels, while the remainder of the channel level controls are open-loop functions of the closed-loop system. True closed-loop control accuracy is provided only at the pilot channel(s) because only their levels are fed back to the control circuitry. Providing closed-loop control for all channels is neither required nor feasible for the application and would probably be a problem from a maintenance point of view. (The time required for setup and periodic adjustment of every channel level, at every amplifier in a system, would be prohibitive.)

Open-loop control systems can perform quite well if the control function accurately matches system variables. In conventional trunk amplifier cascades, amplifier gain and cable attenuation are the only significant variables, and their performance over frequency is well-characterized. Combined open-loop and closed-loop level errors in automatically-controlled trunk amplifiers typically amount to less than 2 dB over a 16-amplifier cascade.

### Control Functions

Regardless of whether or not they include automatic control systems, CATV amplifiers generally require gain control and frequency response control. Since these functions are often electrically controlled, they form a part of a closed-loop control system; however, they are open-loop functions from a broadband point of view. Of the three commonly used control functions, the Slope and Bode functions are different methods of frequency-response control, while the Gain function affects all frequencies equally.

The Gain function is used in all CATV amplifiers for manual level control, and can be used in either open-loop or closed-loop configurations to correct for thermally-induced gain changes in amplifiers. In closed-loop applications, Gain control can be used with the Slope function to correct for thermally-induced cable attenuation changes.

The Slope function permits the frequency response to be manually or automatically adjusted from a relatively "flat" low-attenuation response to a sloped response where the attenuation varies from over 8 dB at low frequencies to near zero at high frequencies. While Slope is theoretically a linear function, CATV applications of the Slope function are designed to approximate the non-linear characteristics of coaxial cable. CATV Slope functions must also be tailored to the bandwidth of the amplifier in which they are used. They have a "pivot" frequency where attenuation remains constant regardless of the degree of slope. This characteristic reduces interaction with the Gain function, and placement of a pilot channel near the pivot frequency provides increased stability in closed-loop dual-pilot applications.

The Bode function was developed specifically for the compensation of thermally-induced attenuation changes in coaxial cable, where attenuation change increase is non-linear with frequency, as was shown in Figure 2. The Bode function is best used in the closed-loop configuration, since manual adjustment can be tedious due to interaction with the Gain function. The Bode function is never used with the Slope function and is completely independent of amplifier bandwidth, but since it is specifically matched to cable, performance is compromised by the addition of taps.

## APPLICATIONS

Trunk amplifiers have traditionally used combinations of Gain control and either Slope or Bode control in automatic systems with closed-loop feedback using two "pilot" channels. The two pilot channel frequencies are generally separated by at least 50% of the total bandwidth, with the Bode function (if used) controlled by the highest-frequency pilot. If the Slope function is used, it is always controlled by the lowest-frequency pilot, and in either case, the Gain function is controlled by the remaining pilot.

Dual-pilot configurations are ideal for long cascades, as the inevitable open-loop errors are reduced by closing the control loop at two points on the frequency spectrum. Dual pilots are used in most automatic Slope/Gain applications, as the Slope and Gain functions must work together to compensate for cable change. While the Bode function is capable of compensating for cable change with only a single pilot, a second pilot improves the Gain control accuracy and helps correct minor open-loop Bode errors which would multiply in long cascades.

Single-pilot configurations are available in some line extenders, and they offer a significant reduction in cost and complexity (compared to dual-pilot systems). They are considerably better suited to cable applications than purely open-loop systems. Since the highest amplitude carrier levels have a greater effect on overall system distortion, the pilot is usually selected near 3/4 bandwidth. The simplest single-pilot configuration is direct control of the Bode function; the pilot can also be used to control the Gain function directly with the Slope function operating as an open-loop function of the Gain control voltage. While this "coupled" Slope/Gain function would not match pure cable as well as the Bode function, it has proved to be a good approximation.

Open-loop thermal sensors have been used in some trunk amplifiers for level control in short cascades, but this approach has generally been restricted to Return amplifiers and line extenders. Open-loop thermal Gain compensators are available as replacements for plug-in attenuator pads, making it simple to add thermal compensation to existing equipment.

Open-loop thermal compensation works well only when the thermal sensor is located in the same environment as the controlled variable. While this is not a limiting factor in amplifier gain compensation, it makes the open-loop approach a poor choice for cable compensation. In cable, the variable is distributed over the entire length of the cable, and the temperature of the amplifier may differ substantially from that of the cable. Throughout its length, cable often passes through areas of differing ambient temperature conditions, and jacketed cable can be subjected to temperature differentials of over 25F (relative to ambient) due to solar-heating effects.

Since most real systems require both amplifier gain compensation and cable (or tapped-feeder) compensation for adequate control, a combination of open-loop thermal compensation and closed-loop tapped-feeder compensation is a good option for single-pilot applications. The tapped-feeder compensation can be implemented as either a Bode function or a coupled Slope/Gain approach, with advantages to each. The Bode function provides a better match to pure cable or rural tapped-feeder, and it is not bandwidth-dependent. The coupled Slope/Gain approach is simpler overall because the amplifier gain compensation function can be added by simply changing component values (without adding passive insertion loss).

## SYSTEM PERFORMANCE

As shown in Figure 3, the open-loop cable compensation limitations illustrate the inadequacy of open-loop compensation in tapped-feeder systems. However, that example remains an "ideal" case which does not include the effects of temperature differentials or amplifier anomalies. By their very nature, closed-loop systems can correct for such errors, but since such errors are unpredictable by definition, they will not be evaluated further. Suffice it to stipulate that actual system performance will generally be poorer than that shown in Figure 3 and all to follow.

This system performance analysis shows how the closed-loop systems deal with the measured thermal response non-linearities introduced by taps. To that end, the same measured data which was used to generate Figure 2 is factored (as in Figure 3) as appropriate to model the ideal characteristics of three functions (Gain, Slope, and Bode), in four combinations. Each of Figures 4 through 7 is displayed under the same conditions as Figure 3: The cumulative change in frequency-response of four 25 dB spans of cable or tapped-feeder is modeled over a 130F temperature range (from 100F to -30F), neglecting amplifier anomalies.

In the dual-pilot configuration shown in Figure 4, the low-frequency pilot at 100 MHz controls a linear Slope function, while the high-frequency pilot at 450 MHz controls a Gain function. Note that the trunk cable plot is severely curved due to the cumulative difference between the linear Slope function and the non-linear cable characteristics. This effect would not be as pronounced in a real system (since CATV Slope functions are not perfectly linear), and all four plots would therefore be somewhat less curved at the ends. Amplifier Gain compensation is accommodated automatically by the closed-loop Gain function.

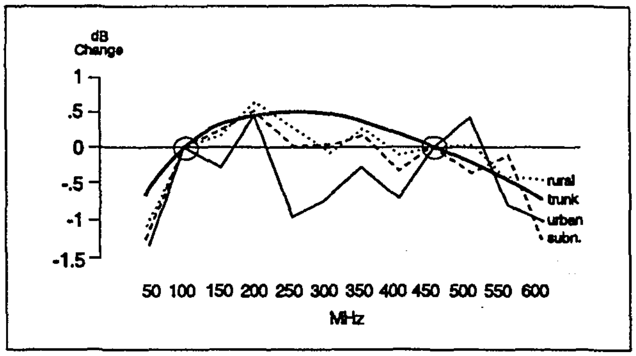



*Figure 4.*
*Cascade of 4 Dual Pilot Slope/Gain*
*Over 130F -*
*Pilots at "O"*

Figure 5 is also a dual-pilot configuration, with the 100 MHz low-frequency pilot controlling a Gain function, and the 450 MHz pilot controlling a Bode function. Since the Bode function compensates exactly for cable, the trunk cable plot is coincidental with the 0 dB reference line. Note that all three tapped-feeder plots now have a slight "belly" in the 250-400 MHz range which is more pronounced in the high-density "urban" example. Amplifier Gain compensation is accommodated automatically by the closed-loop Gain function.

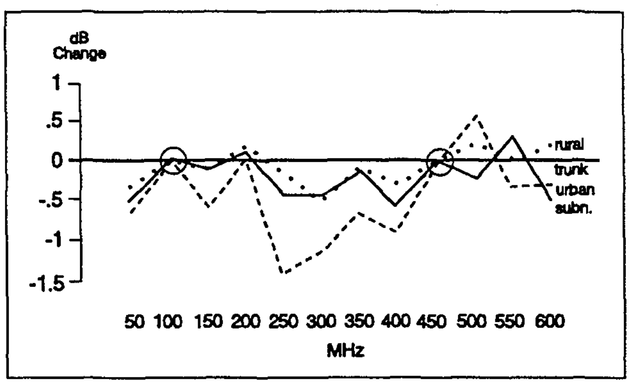



*Figure 5.*
*Cascade of 4 Dual Pilot Gain/Bode*
*Over 130F -*
*Pilots at "O"*

In Figure 6, the Gain function is controlled by a single pilot at 450 MHz. The Slope function is open-loop-coupled to the Gain function, and is designed to compensate for cable at 100 MHz. Comparing Figure 6 to Figure 4, note that while the cable plots match exactly (0 dB at 100 and 450 MHz), the tapped-feeder plots are equal only at 450 MHz. The resulting error is much less than the purely open-loop configuration shown in Figure 3, even at low frequency. Amplifier Gain compensation can be accommodated as necessary by changing component values.

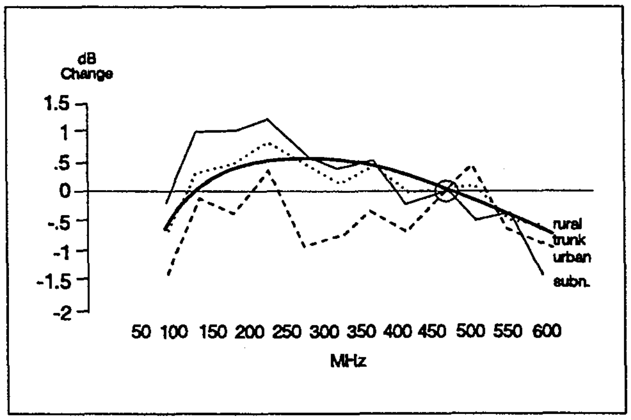



*Figure 6.*
*Cascade of 4 Single Pilot Coupled Slope/Gain Over 130F - Pilot at "O"*

Figure 7 shows a Bode function controlled by a single pilot at 450 MHz. As in Figure 5, the trunk cable plot is coincidental with the 0 dB reference. Amplifier Gain compensation cannot be accommodated by the Bode function, and must be added separately (if necessary).

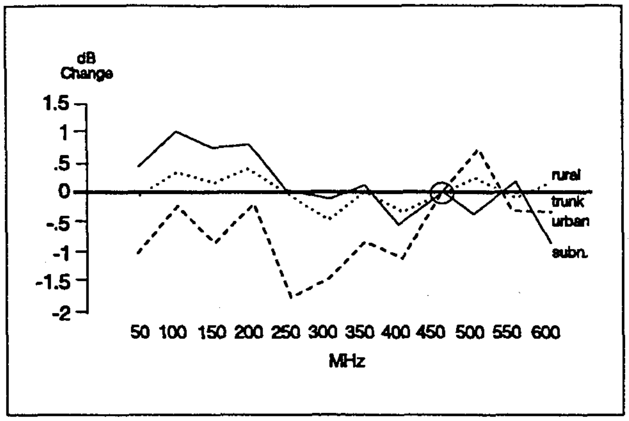



*Figure 7.*
*Cascade of 4 Single Pilot Bode Over 130F - Pilot at "O"*

## CONCLUSIONS

Where wide temperature variations exist, uncontrolled level changes build up rapidly in cascade; this applies equally to trunk and feeder.

Tap thermal change is slightly lower than that of an equivalent section of cable; the difference is particularly significant in high-density areas.

The tap thermal change function is complex and unpredictable over frequency, but is generally similar to cable.

Simple open-loop compensation performs poorly in most CATV applications, but remains preferable to no compensation at all.

Dual-pilot control offers little advantage over single-pilot control in tapped-feeder applications; it is not worth the additional cost and complexity.

The single-pilot Bode function offers slightly better compensation flatness than the coupled-Slope/Gain function, but usually requires a separate Gain compensation function for good performance.

# COST EFFECTIVE CABLE TELEVISION TRANSPORT FOR PCN

George Hart
Rogers Engineering

## ABSTRACT

A novel approach for universal transport of PCN signals on Cable Television facilities is described. Experimental Remote Antenna Driver (RAD) hardware has been tested on Cable Television facilities in a variety of user environments including residential, office and shopping mall. Network architecture evolution based on fiber optic signal transport supports consumer demand paced capital deployment. Bidirectional cable transport, essential for PCN is discussed. A comparison with standalone base-station deployment is made.

## INTRODUCTION

Many believe that Digital Cordless Telephone (DCT) service represents the next major step in the evolution of wireless telephone service in North America, with the potential to make wireless telephone service available to all consumers and to lead the way to the personal communications environment, where a single telephone number could reach users no matter where they are located[1].

Moreover, cable television companies have a potentially important role to play in this challenging vision of the future for personal communications[2].

An ubiquitous wireless telephone service will require a large number of base-stations to be deployed. In order to provide high quality service in an affordable technology package these base-stations will need to be closely spaced and will require a means of connection to the public switched telephone network.

The deployment and interconnection of these many base-stations with the Public Switched Telephone Network (PSTN) requires a substantial telecommunications network. The entities currently in the best position to provide this Personal Communications Network (PCN) include the telephone companies, cable television operators, alternate access carriers, and private communications network owners including the power utilities.

## PERSONAL COMMUNICATIONS SERVICES

Achieving a successful consumer acceptance of Personal Communication Services (PCS) requires that these new communications services have the following characteristics:

- good service quality: a grade of service defined as percentage calls blocked plus 10 times percentage calls dropped totalling less than 0.5 percent of total call attempts;
- extensive network coverage: general availability of service where people are;
- a highly accessible retail network for the marketing and sale of subscriptions to the wireless telephone service;
- efficient business office and customer services, sensitive to local market needs;
- a diversity of services and service features, including virtual private networks for business users; and
- excellent price value, based on exceptionally cost-effective network design and operation.

Users of PCS will expect to have access to basic telephone service features such as operator services, long distance calling, credit card calls, directory assistance and emergency services.

In addition, service providers will consider enhanced telephone service features including call management services and voice store and forward. These service features will be accompanied by network design features such as autonomous registration, which will allow users to make and receive calls no matter where they are.

## NETWORK REQUIREMENTS

The low-cost, lightweight transceivers associated with PCS will necessitate low transmit power, in the vicinity of 10mW[3]. The resulting limited radio path loss budget supports a typical outdoor base station service radius of 150m hence requiring a very large number of radio base-stations to provide the extensive availability necessary for a successful service.

The PCN necessary to support the numerous base-stations must satisfy the following criteria; widespread, reliable, low-cost, flexible, and support sufficient traffic capacity. A brief description of each parameter follows.

### Widespread

Users' reasonable expectations of widespread PCS availability means the network of base-stations needs to extend into most urban and suburban areas. Those existing communications networks already close to meeting this requirement are advantageously positioned to provide base-station signal transport.

PCS availability will be expected in most public areas such as playgrounds, parks, sidewalks, and within buildings such as airports, shopping malls, recreation facilities, hotels, stadiums, ice rinks, and variety stores; in short, wherever a person is likely to spend more than a few moments.

### Reliable

The Grade Of Service (GOS) suggested above (0.5%) for successful PCS depends on a combination of critical elements. Traffic capacity related elements are treated later in this paper. The GOS budget is mainly allocated to the radio link, which is subject to significant shadowing and multipath fading in the low-power, high-clutter environment anticipated for most PCS base-stations[4].

The telecommunications infrastructure portion of the PCN must therefore achieve virtually 100% availability in order not to affect the overall GOS. A single component should not affect a large number of base-stations, implying a need for redundancy and fail-safes inherent in the network architecture. Maintenance activities must not result in frequent service interruptions. Expeditious repair of equipment failures is mandatory.

### Low-Cost

Two elements dominate the total cost of operating a PCN: purchase and installation of base-stations and interconnect network operating costs. The interconnect network between base-stations can easily become the greatest operating cost element in PCS delivery. Achieving an affordable rate structure commensurate with widespread service penetration necessitates a low-cost communications network.

The interconnect network may also contribute indirectly to the base-station costs. Removing as much signal processing as possible from the base-station will in general reduce its complexity, power consumption (another significant operating cost) and manufacturing cost.

### Flexible

A large degree of uncertainty exists among wireless telephone experts as to the details of how PCS will evolve. The desire to easily add or change services is driving the current intelligent network (IN) developments in the telecommunications industry. Unfettered evolution of PCS will require IN capabilities in the support infrastructure of base-stations and interconnect network.

The inexact radio engineering for thousands of base-stations (precision engineering would be prohibitively expensive) will require that individual stations be easily added or moved by relatively unskilled technicians working autonomously in the field. The base-station interconnect network must therefore be easily and cost-effectively extended to practically anywhere in urban and suburban areas.

### Capacity

Together with network reliability discussed above, proper matching of traffic channel capacity with user demand at any particular location determine the GOS. Unsuccessful call attempts resulting from a shortage of voice channels are reduced by the addition of base-station equipment, which the interconnect network must be capable of supporting. PCS demand is expected to be low when the network is built, and subsequently grow with time as service penetration increases. Perspective PCS providers face the challenge of achieving sufficient traffic capacity at low cost when building the base-station network while ensuring expansion capability as the service matures.

## CABLE SYSTEM SUITABILITY

The possibility of cable television systems providing the PCN was suggested earlier in this paper. Cable's ability to meet the requirements listed above will now be reviewed.

### Widespread

Cable television systems currently serve virtually all urban and suburban areas in North America. Areas typically poorly served are commercial and industrial areas. Cable facilities criss-cross residential areas in a tree-and-branch architecture which passes 95% of all TV households and many public buildings.

These extensive coaxial networks are typically one-way, however. Most of the amplifiers must be upgraded to bi-directional operation and properly aligned to support a PCN.

Many cable operators are exploring business communications opportunities in urban centers, leading to network expansion in traditionally poorly served areas. These extensions are frequently fiber-optic based two-way systems.

### Reliable

The network availability required to support PCS is generally not supported by cable television operators. This is one parameter that cable owners will need to examine closely in their deliberations of PCS involvement.

Technical operating and Maintenance practices vary across the industry with extended service outages common in significant portions of cable networks. Many operators who have tried to use the coax networks for supporting business communications have invariably become frustrated and installed fiber systems to deliver the quality of service expected by the customer. A radical change in technical operating philosophy is required by most cable operators before serious exploitation of existing coax facilities for PCS can be realistically achieved.

Network reliability improvements are realized through system architecture and equipment modernization in addition to maintenance practices. The current deployment of fiber to serving areas of approximately 2000 homes reduces significantly the number of components between the headend and the customer with attendant lower probability of failure[5,6]. Status monitoring of system actives enables rapid response to equipment failures. Amplifier bypass reduces the impact of failure, supporting "lifeline" service continuation until repairs are made. Standby powering of components serving a substantial number of customers further reduces service outages. The PCS availability expectations may require all actives be standby powered.

Two-way plant operation introduces additional challenges. Ingress in the traditional upstream band 5 to 33 MHz can significantly impair communications. The objective of meeting Cumulative Leakage Index (CLI) requirements for aeronautical band occupancy on cable should simultaneously give satisfactory ingress immunity. Technician training for proper two-way alignment and maintenance procedures is required. Success among various operators shows two-way operation is possible but only through a comprehensive commitment to incorporate upstream system design and maintenance into the cable system operation.

### Low-Cost

The extensive cable distribution networks which already exist in residential areas can be combined with an innovative signal processing concept to extend high quality, cost-effective wireless telephone service to all public areas in a neighbourhood, including streets, parks, hospitals, plazas and school yards.

Under the authority of an experimental license granted by the Canadian Department of Communications, Rogers Cable is undertaking a field trial of cordless telephone technology in

Vancouver, Canada. This trial, which commenced on September 26 1991, is designed to test the performance of CT2 Plus technology produced by a number of different manufacturers. In addition, this field trial is designed to test the feasibility of a wireless signal processing concept developed in conjunction with Cable Television Laboratories Inc.

This new concept, referred to as remote antenna driver (RAD), uses cable distribution plant, equipped with two-way amplifiers, to connect a number of remotely located antennas to centrally located base-stations. In this way, an extensive coverage area capable of accommodating many users may be established without the need to deploy numerous base-stations. This not only improves the cost-efficiency of the network but also provides for continuous uninterrupted service as the user moves from antenna to antenna within the coverage area.

RAD technology will be particularly effective in providing wireless telephone service in residential areas where there is an installed base of cable distribution plant and where the cost of providing extensive coverage of public areas using conventional base-station technology would be prohibitive.

Flexible

The deployment of RAD on the existing broadband coaxial cable television networks is an ideal means for supporting a growing PCS customer base. Through the sharing of base-stations over an extensive area of low traffic, optimum base-station acquisition for the number of users is achieved.

Intelligent Networks (IN) enable a variety of advanced telecommunication services of interest to wireless telephone customers. Advanced call routing and termination, dynamically controlled by the user and provided through IN gives the customer the choice to accept certain calls while sending other calls to voice-mail, with the selection particulars dependant on location or time of day. The concentrating of base-stations at a limited number of locations eases IN implementation by reducing the quantity of sites requiring separate IN interconnect facilities.

The virtual base-stations that RAD units represent increase installation flexibility through ease of deployment anywhere within a cable television system. Adaptive subsystems within the RAD conform to the cable environment with a minimum of technician support.

Capacity

Additional base-stations are easily added, while frequency division multiplexing on coax and ultimately space division multiplexing on fiber allow unrestricted expansion of traffic capacity. The expected continuation of fiber deployment in cable systems leads to ever smaller areas of shared coax facilities. Supporting PCS via a "last mile" of coax is therefore practical even as user traffic increases.

## THE RAD TECHNOLOGY

Cable Television Laboratories and the engineering staff within the Rogers family of companies, including Rogers Engineering, have developed a method for remoting antennas relative to wireless base-station equipment. This work finds potential application in micro-cellular and PCN deployment. A number of experimental installations of antennas connected by coaxial cable to host base-stations have been tested for micro-cellular with good results.

The concept of remoting antennas from a set of radio base-stations is considerably enhanced if a standard CATV distribution system were able to provide the connection. The evolution in cordless telephone technology towards PCS makes this a very desirable means for serving an extremely large number of small micro-cells. The likely frequencies to be used for PCS (ie. 1850-2200 MHz) are, however, beyond the normal range of CATV systems. Therefore a conversion process is needed at each remote antenna connected to the cable system.

The primary goal of the remote antenna link is transparency for voice modulated wireless radio signals. The "radio" link protocol between base-stations and portable telephones must remain uncorrupted to maintain full capabilities for call hand-off and radio channel assignment. Therefore a unity gain, broadband path for both downstream and upstream signal flow is preferred.

The CATV plant and any signal processing equipment used to transport wireless telephone signals between base-stations and remote antennas should match the operating bandwidth and dynamic range of the base-station. The technical challenge is to develop a device which can interface between the RF requirements of PCS and the limitations of existing CATV plant.

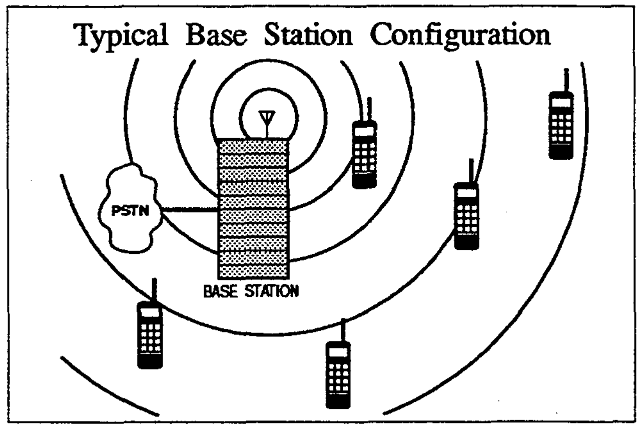
The Remote Antenna Driver (RAD) is an attempt to create such an interface device. It has proven feasible in a variety of RAD field trials to convey wireless telephone signals over standard two-way cable television facilities with no adverse effects. It has also been shown that antenna-to-antenna service continuity is practical. This is extremely useful as it trivializes the hand-off process between micro-cells.

The amount of processing required for CATV transport of PCS signals has a direct impact on both the cost and performance of the service. The minimum equipment required at each micro-cell for wireless communication is a radio transmitter and receiver. If simple amplifiers are sufficient for this task, a minimum-cost base-station is achieved. Any further processing, regardless of production volumes, is at incremental cost. In addition, processing will potentially degrade service quality or feature functionality as a result of efforts to reduce its cost.

## RAD CONCEPT DESCRIPTION

The RAD concept can best be understood through a progression that starts with a single multiple voice-channel cordless telephone base-station as shown on Figure 1. The number of users able to simultaneously communicate with the base-station equipment is determined by its voice channel capacity. All users are free to roam anywhere throughout the coverage range assumed to be path-loss limited. Call setup and maintenance is per the Common Air Interface (CAI) specifications for the wireless radio protocol. A number of alternative CAI proposals have been submitted to the FCC for consideration as a PCS standard. The CAI particulars do not affect this RAD discussion. The base-station equipment connects to the PSTN via any of a variety of access technologies. The introduction of RAD concentrates this PSTN connection.

Figure 1




The next step in the progression to RAD considers an extension of the base-station equipment coverage area through a simple coaxial connection to a remote antenna as shown in Figure 2. All of the voice channels supported by the base-station appear at both antennas. A user may freely roam throughout the overlapping coverage areas provided by the two antennas with no hand-off processing required.

Figure 2

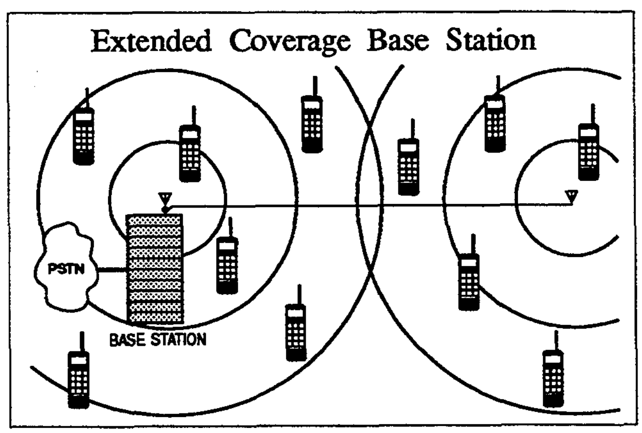
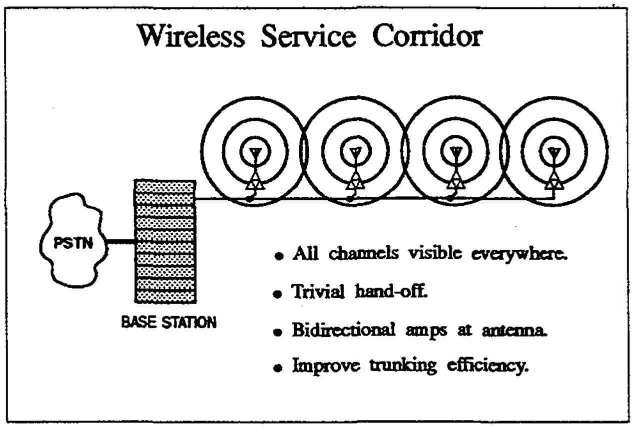


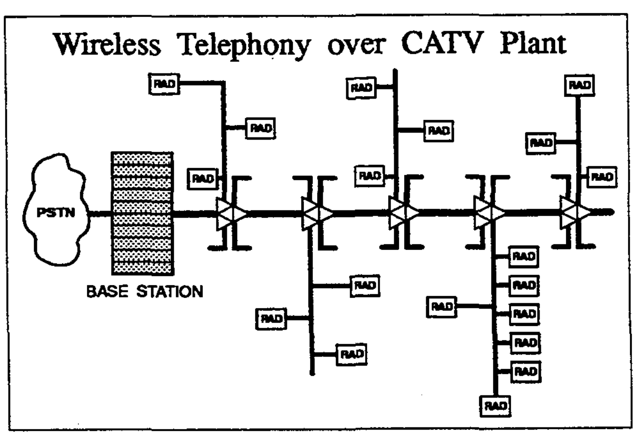

Figure 3 illustrates an extension of the remote antenna concept. A cascade of couplers and antennas served by a coaxial cable connected to the base-station antenna port supports the creation of a corridor of service coverage. Amplifiers at each antenna compensate for the attenuation of the couplers and cable. Bidirectional amplification is made possible with duplexing filters in Frequency Division Duplexed (FDD) CAI or an RF switch in Time Division Duplexed (TDD) CAI. Users have access to the full channel capabilities at any antenna and may roam without service interruption throughout the coverage corridor. The total capacity in the corridor is determined by the base-station capacity, but there is no restriction on how the capacity is shared between antennas.

Figure 3




The RAD replaces the bidirectional amplifier at each antenna. A CATV distribution system, depicted in Figure 4, connects RADs and the remote antennas they serve to a common set of base-station radios. Users at any RAD have access to the full capacity of the base-station equipment and are free to roam between overlapping RAD coverage areas without a hand-off process required. Similarly, the capacity of the base-station is shared among the RADs with no restriction. The ideal RAD is capable of full flexibility in call capacity and coverage range while not impairing the signal transmission between remote antennas and base-station radios.

Figure 4




## VOICE CHANNEL PROCESSING

Implementation of PCN requires intelligence to locate wireless handsets wherever the user may be. A large number of complex radio base-stations must be deployed to provide sufficient radio access to stimulate use by the general public for PCS to succeed. Each radio base-station translates the duplex voice signals from its twisted pair or fiber optic transmission format, into radio frequency (RF) signals suitable for communication with the portable handset.

In most of the proposed PCN standards, the radio spectrum is divided into a number of discrete RF channels using Frequency Division Multiplexing. Additionally, some systems such as CT-3 employ Time Division Multiple Access (TDMA) to multiplex several traffic channels into each RF channel. In all cases, traffic channels are processed individually at the base-station before interconnection to the PSTN occurs.

Each radio channel served by the base-station generally requires a modulator and demodulator for radio communications with handsets, a multiplexer and demultiplexer for handling the transmission of voice and signalling information over the radio link and a control system for error correction, establishment of radio links, synchronization, and other functions related to digital radio transmission. An encoder and decoder for converting the voice messages into digitally-coded streams of data are required for each traffic channel.

When more than one user is served by a base-station, duplication of the circuitry for one duplex conversation or voice channel is needed for each simultaneous user. The number of users who can access a particular radio base-station will therefore be limited to the number of voice channels it is equipped to support.

### PSTN Interconnect Requirements

A PSTN interface to many individual stand-alone radio base-stations requires additional loop facilities throughout the community. Each channel of each radio base-station demands a new telephone line which may be supplied by new or existing telephone company facilities or alternatively by

CATV facilities. In most cases, new transmission equipment will be needed to service this demand and the CATV operator will not have any advantage relative other network providers. Additional intelligence must be provided to locate DCT handsets wherever the user may be. Providing the intelligent network required to support the enhanced service features at all base-stations will be a difficult challenge, particularly if they are widely dispersed. Cable operators may have particular difficulty addressing this challenge using conventional telecommunication approaches.

Individual base-station package deployment requires capacity engineering for each. A determination of the number of voice channels to support at each radio base-station package must be made based on the user traffic anticipated. Most cases require at least two voice channels to adequately support users for call blocking less than 0.01. Generally this is more capacity than needed, particularly at service launch. Excess capacity at each radio base is cumulative and a less than optimum deployment of equipment results in excessive cost of PCS provision.

## An Alternate Approach

In contrast to the dedicated PSTN loop interconnect configuration associated with many stand-alone base-stations, RAD deployment requires minimal consideration for the amount of voice traffic expected. A common set of base-station equipment serving a large number of RADs supports dynamic adaptation to traffic patterns throughout their collective coverage area. Traffic engineering efficiency is greatly improved since base-station capacity is determined for the neighbourhood, rather than each individual base-station.

## Consumer Demand Paced Investment

As subscriber penetration increases the number of base-station channels can be increased on a very cost effective basis. Conventional telephone traffic engineering practices rely on "trunking efficiency" resulting from many users sharing a large pool of voice channels. This is illustrated in Table 1 which is taken from Erlang B load vs. loss curves. For example, a thirty-channel equipped base-station with the assumption of 0.1 Erlangs per user will support 200 users within the area served by those base-stations, with a probability of busy signal of less than one percent. One Erlang is equivalent to a single telephone line busy for one hour. A sub-unit commonly used is hundred (centum) call seconds (CCS), with 36 CCS in one Erlang. A user offering 0.1 Erlang is therefore using one phone line (or voice channel) for a total of 6 minutes every hour. In this case 7 users share each base-station voice channel. A radio base-station with 7 channels only supports 20 users, with only 3 users sharing each channel.

**TABLE 1**
**WIRELESS VOICE TRAFFIC CAPACITY**

| Neighbourhood Population | 2000 | 2000 | 2000 | 2000 |
|---|---|---|---|---|
| DCT Service Penetration | 1% | 5% | 10% | 25% |
| User Population | 20 | 100 | 200 | 500 |
| Voice Traffic (@0.1E) | 2E | 10E | 20E | 50E |
| Voice Channels Required (P(b) = 0.01) | 7 | 18 | 30 | 64 |
| Users per channel | 3 | 6 | 7 | 8 |

The figures in Table 1 indicate the growth in channel capacity required for service penetration from 1% to 25%. It is clear from this table that even at 25% penetration, only 64 channels of voice capacity are required to service the 500 users that this level of penetration represents. The case of very low penetration is not indicated. However, for a guaranteed probability of blocking to be less than 1%, it is essential that a minimum of two voice channels be available for the user population. This rule can be applied to the stand-alone base-stations that might be deployed to offer this service to the general public. In other words, virtually every base-station installed in a public area would need to provide at least two voice channels of capacity in order to offer a level of service comparable to wire line telephone availability.

The 2000 home residential neighbourhood dimension used in the Table is consistent with coax cable plant networks resulting from deployment of fiber to the serving area. It bears clarifying that the 30 channels required for a 10% service penetration implies a 30 voice-channel radio base-station located at the headend for each serving area fiber node. An equivalent level of service requires approximately 80 individual base-stations in order to provide ubiquitous coverage over each 0.8 mile$^2$ area this represents. Instead, a similar number of very simple RADs are deployed at a fraction of the cost of complex radio base-stations.

A comparison of the two alternatives is presented in Table 2. The same 2000 home serving area used in Table 1 covers 0.8 mile$^2$, requiring 80 micro-cells of 25 homes each. The micro-cells are served by two-way coax and use either a standalone base-station or alternatively a RAD with base-station equipment concentrated at the headend.

Voice channel requirements are based on Erlang "B" curves with each user offering 0.1E (3.6 CCS) busy hour traffic load and a 0.01 call blocking probability.

**TABLE 2**
**RAD DERIVED EFFICIENCY**

| | PCS PENETRATION | 1% | 2% | 5% | 10% | 25% |
|---|---|---|---|---|---|---|
| Stand | Users per cell | 0.25 | 0.50 | 1.25 | 2.5 | 6.25 |
| Alone | Traffic per cell | 0.03E | 0.05E | 0.13E | 0.25E | 0.63E |
| Base | Voice Channels /Cell | 1 | 1 | 2 | 3 | 4 |
| Stations | Base-station/ Cell | 5 | 5 | 6 | 7 | 8 |
| (A) | Cost/Area | 400 | 400 | 480 | 560 | 640 |
| (B) | Voice Channels /Area | 80 | 80 | 160 | 240 | 320 |
| | Users per area | 20 | 40 | 100 | 200 | 500 |
| RADs | Traffic per area | 2.0E | 4.0E | 10.0E | 20.0E | 50.0E |
| (C) | Voice Channels /Area | 7 | 10 | 18 | 30 | 64 |
| | Base-Station Cost | 7 | 10 | 18 | 30 | 64 |
| | RAD Cost | 80 | 80 | 80 | 80 | 80 |
| (D) | Total Cost/ Area | 87 | 90 | 98 | 110 | 144 |
| | C/B | 0.09 | 0.13 | 0.11 | 0.13 | 0.20 |
| | D/A | 0.22 | 0.24 | 0.20 | 0.20 | 0.23 |

The cost comparison, D/A, in Table 2 is relativistic, normalized to the complexity of a RAD. Concentrated base-station equipment at the headend is assumed lower cost than standalone due to more favourable packaging and environmental requirements. In the case of CT2 Plus, which uses FDMA, each voice channel requires an incremental radio transceiver and voice processing unit which is assumed equal in cost to a RAD suitable for outdoor mounting.

The significant savings in voice channel trunking required resulting from RAD is indicated in the ratio C/B. Voice traffic is concentrated when using RAD, reducing the network interconnect volume.

In effect the total number of users in the neighbourhood share the total number of base-station channels. On average, this means fewer total channels are required to support a given population of users, as documented in the Erlang traffic tables. Provided the total capacity of the shared base-station equipment is not exceeded, low probability of blocking is efficiently achieved. The expense associated with the construction and operation of a PCN is potentially reduced by the use of RAD technology deployed through existing cable distribution networks.

## RF Hand-off

Connecting RAD access-ports spaced approximately 200 m apart via a cable system as depicted in Figure 4 provides contiguous coverage to form a neighbourhood area of service. A portable handset is free to move about within the cluster of contiguous antenna areas because it remains "connected" to the same base-station and no hand-off processing is required as it moves from one antenna zone to the next. This is contrasted with the conventional approach of individual base-stations at each antenna location, where a coordinated hand-off process is required for roaming. The neighbourhood extension of radio base-station service coverage is efficiently achieved with the RAD system.

## CONCLUSION

In summary, the advantages of the RAD system can be listed as follows:

1. Efficient extension of coverage to areas where user density is much lower than is supported by individual base-station deployment. The cost effective deployment of PCS to the mass-market in particular is achieved via the existing cable infrastructures found in virtually all neighbourhoods.

2. The provision of seamless coverage between micro-cells such that a user may roam from one micro-cell to the next without creating a large switching burden.

3. The seamless hand-off between micro-cells supports focused coverage patterns such as streetscapes and pedestrian malls.

4. Small micro-cell electronics package (RAD) for unobtrusive mounting on buildings, street light standards and utility poles.

5. The concentrating of base-stations optimizes the number of voice channel interconnects required over the total area of the system. This also facilitates advanced network interconnect necessary for the anticipated service features.

6. Consumer demand paced equipment deployment is achieved through the concentrating of radio base-stations made possible by RAD.

7. Elimination of micro-cell capacity engineering, significantly reduces the cost of adding service coverage.

Through innovative techniques, cable operators may succeed in efficiently and cost effectively transporting wireless telephone signals for PCS providers. However, significant improvements in plant reliability and upgrades to two-way operation will need to accompany techniques like RAD and fiber deployment to provide an interconnect facility applicable to these new communication services.

## ACKNOWLEDGEMENTS

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## BIOGRAPHY

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Mr. Hart has most recently been devoting his attention to PCN and managing the development and field testing of RAD. Prior to this he developed the Rogers Fiber Architecture and managed the procurement and deployment of fiber optic equipment required to implement the architecture within Rogers operating divisions.

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# DIGITAL AUDIO AND ANCILLARY DATA SERVICES FOR ATV -- THE WORK OF THE ATSC SPECIALIST GROUP

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## ABSTRACT

*This paper describes the advice and suggestions put forth by the Technology Group on Distribution (T3) of the Advanced Television System Committee (ATSC) regarding digital services for Advanced Television (ATV). These recommendations were based on the background work of the Specialist Group (T3/S3) on Digital Services which conducted technical studies, and surveys, and developed the suggestions and recommendations.*

*Rapid advances in multichannel composite digital audio coding technology now make it possible to plan to provide the consumer (e.g. cable subscriber) with an expanded audio experience to match wide screen high definition TV pictures. This paper summarizes the suggestions adopted by ATSC's T3 Group regarding audio and ancillary data services, including the advice that a standard service for ATV should include capacity for a minimum of five audio channels with composite encoding.*

*Particular emphasis was placed on the need for flexible allocation of data to audio and ancillary data services on an as-needed basis.*

## INTRODUCTION

The Advanced Television Systems Committee and other organizations, including the FCC Advisory Committee on Advanced Television Service, EIA, and NCTA, have recognized that much of the effort spent on ATV development and standardization has been put on the development of vision signal encoding schemes and that it is important that appropriate emphasis should be placed on defining the accompanying sound channels, the features of the ancillary data and control services, and the way in which they may be included in the ATV transmission format.

The charter of the ATSC Specialist Group on Digital Services (T3/S3) has been to conduct industry surveys and technical studies and to develop technical information and recommendations on the following subjects. The Specialist Group commenced its studies in December, 1990.

A. Sound & Ancillary Data Services

Identification of the range of ATV sound channel requirements and how they might be satisfied with recent rapid advances in the state-of-the art digital audio coding. Identification of the range of desirable ancillary data services--including those already in use and in some cases mandated--and estimates of the data capacity required for each.

B. Conditional Access

Establishment of system attributes and features required to allow conditional access to the ATV transmission in the various media in particular Cable Television, with emphasis on the structure and capacity of the required control data channel.

C. Multiplex Structure

Determination of requirements for flexibility in the data multiplex struc ture to permit re-allocation of the available data capacity for the various services.

The goal of this work has been to influence the characteristics of the digital services (other than video) which are to be provided by the ATV emission system selected for use for terrestrial broadcasting in the United States. (However, it is not ATSC's intention to affect the ATV testing program already in progress as part of the FCC selection process.) Specifically, it is suggested that the final ATV system selected for terrestrial broadcasting, cable, and other media follow the guidance detailed in reference 1 (see Appendix for Executive Summary of the ATSC document).

Recognizing that different media (broadcast, cable, DBS, etc.) will use different modulation methods, possibly carry different ancillary signals and have other requirements unique to each medium, T3/S3 has focussed its technical studies to facilitate the adoption of voluntary standards for all media based on the same baseband signal representation for video, audio, data and control to maximize interoperability and to provide a simple common interface to the consumer ATV receiver. Work by T3/S3 on Conditional Access is reported in References 2 and 5.

## WORK ON AUDIO AND ANCILLARY DATA SERVICES

Advances in audio bit rate coding technology have occurred extremely rapidly in the past two years, now making it possible to encode multiple audio channels using data rates previously allocated to a single audio channel. It is now technically possible to create multiple channel audio "images" which can truly complement the high definition pictures in a restricted 6 MHZ wide HDTV digital channel. Audio techniques developed for motion picture theaters may become applicable to the consumers' living room.

As a discussion vehicle, the Specialist Group prepared a "strawman" proposal for TV Audio and Data Services for Simulcast ATV systems. The "strawman" proposal suggested desirable program audio attributes and listed possible ancillary digital services, including some of those already in use with NTSC such as closed captioning. It was suggested that advantage should be taken of the opportunity to design a new generation of ancillary data services into the system from the beginning. Starting in April 1992, the "strawman" proposal was circulated for comments and suggestions to a wide range of parties with interest in ATV development and introduction, including the proponents of systems to be tested in the selection process. By an iterative process, the document was refined to reflect as much industry consensus as possible, with particular emphasis on flexible allocation of data capacity to audio and ancillary data services.

The following summarizes some of the principal suggestions as to the characteristics of the audio and digital services which should be provided by an ATV emission system adopted for use in the United States, both for terrestrial broadcast and by alternate media. Certain services are identified which the final distributor of programming may provide in a standardized manner. Standardization of services is necessary so that receiver manufacturers may produce receivers that can receive these services. A number of capabilities are identified which every receiver should be required to provide.

## MAJOR PRINCIPLES

### Flexible Allocation

The number and type of audio and data services which an ATV system should deliver will vary significantly depending on what services are available to accompany the picture, and the needs of the particular service area for the distribution medium, e.g. broadcast, cable. etc. In order to provide a maximum level of utility, many audio channels and data services might be needed. A fixed provision for many audio channels and data services would require, however, that a significant portion of the available transmission capacity be reserved for these services. This would unnecessarily constrain the video quality, which would suffer if its available data rate were restricted. Since much of the programming may not require many audio channels or data services, a fixed allocation of transmission capacity to these services would be inefficient. Therefore, the ATV transmission system should be capable of a flexible allocation of data to audio and data services on an as-needed basis, with the remaining data available to the video system. At the point of emission or final distribution, a choice may be made based on the available services and the needs of the intended audience as to which services are to be delivered via the final distribution medium.

### Extensibility

It is not possible to envision and provide for all potential audio and data services which might be useful components of the ATV service in all media. It is feasible to allow new digital services to be added in a compatible manner by means of the flexible allocation of data capacity. The key is to provide a way to identify new types of allocated data services, so that new or upgraded receivers may make use of the new data types, while older receivers simply ignore them. The new data types could offer new kinds of audio services (perhaps with more audio channels) and new kinds of information services. Some of the additional data types could be intended for private or commercial reception, with the data capacity being sold by the final program distributor to provide a supplemental revenue source.

### Multi-Channel Audio

The ATV system should be capable of delivering multi-channel sound appropriate to the wider, higher definition picture. Consistent with demonstrated psychoacoustic principles, the preferred channel assignment is: Left, Center, Right, Left Surround, and Right Surround. A significant advantage of three front channels (rather than two as in present TV stereo) is stabilization of the audio dialogue image in the vicinity of the TV screen.
[Note: This is consistent with the CCIR Task Group 10/1 Draft Recommendation on Multi-Channel Sound (Reference 3), and the SMPTE Film Sound Sub-Committee Report](Reference 4). An optional low bandwidth channel for low-frequency enhancement (subwoofer channel) may also be provided. Monophonic and two-channel stereophonic transmission modes of operation should also be provided, and will offer the most efficient method of delivery for mono or two-channel stereo programming.

Even with the use of advanced low bit-rate audio coding technologies, provision of five high quality independently coded discrete channels would require significant data capacity. Current estimates of required bit rates per individual audio channel are:

| | | |
|---|---|---|
| High Quality | 128 | kbits/sec |
| Medium Quality | 96 | kbits/sec |
| Low Quality | 64 | kbits/sec |

However, recent developments in composite multi-channel audio coding technology show that five-channel audio may be delivered with a data rate only slightly larger than that required by two high quality independently coded channels. It is recommended that a five-channel composite coding mode structure be pre-defined as part of the ATV system.

Suggested are three composite coding modes, which offer different numbers of high quality audio channels. The numerical designations below (e.g. 3/2) indicate the number of front channels / rear channels. Some audio coding technologies incorporate a low bandwidth channel (<200 Hz) intended to deliver low-frequency enhancement information (subwoofer channel), the use of which would be optional for the program distributor, receiver, and viewer.

| | | | |
|---|---|---|---|
| Five channels | 3/2 | 300 - 400 | kbits/sec |
| Three channels | 3/0 | 256 | kbits/sec |
| Two channels | 2/0 | 192 | kbits/sec |

With composite coding, all of the audio channels which have been coded into the composite data stream must be reproduced together (similar to the way the R,G,B colors must be reproduced together to form the viewed picture), although not necessarily out of independent loudspeakers. The channels may be mixed together to reduce the number of loudspeakers required for reproduction.

When surround information is relevant and available, the 3/2 composite coding mode is preferred. Three front channels and two surround channels are provided.

When surround information is not available, the 3/0 mode is sufficient. Three front channels are provided; the 3/2 mode may also be used.

Two-channel stereophonic audio may be most efficiently transmitted with the 2/0 mode. The 3/2 or the 3/0 modes may also be used.

It is preferable that audio programs not be simultaneously provided by a monophonic or two-channel stereophonic service intended for low cost receivers, in addition to a separate multi-channel service intended for high end receivers. That would be wasteful in data capacity, and would inhibit the use and growth of multi-channel audio. It would be necessary, though, that all receivers be capable of decoding a multi-channel service into the desired number of reproduction channels. Lower cost receivers would not need to completely decode all five channels, but could decode the five channel service into a conventional monophonic or two-channel stereophonic program with attendant cost savings.

The ATV receiver would be required to produce sound for any of the pre-defined coding modes that the broadcaster chooses to use. This doe not imply that every receiver should be capable of fully decoding a 3/2 service; only that it make sound from it. It would be acceptable for a receiver to decode all modes into monophonic or two-channel stereophonic audio.

### Uniform Loudness

The ATV audio system should provide the means to control loudness in a uniform manner among various programs and delivery channels. The viewer should perceive the same subjective dialogue loudness when a program ends and a new program begins, or when channels are changed. The ATV audio data should inform the receiver of a dialogue reference level so that the receiver can reproduce the normal spoken dialogue at an acoustic level chosen by the viewer.

### Dynamic Range Control

There is a conflict between the needs of many viewers for program audio to have a narrow dynamic range, and the desires of some viewers to reproduce the audio in the full dynamic range intended by the program producer. The ATV audio coding system should incorporate an integrated dynamic range compression method which delivers data to the receiver representing the compression characteristic employed.

Receivers may include circuitry which gives the viewer the ability to control the reproduced dynamic range. Using the data representing the compression characteristic employed, the receiver may partially or completely reverse the dynamic range compression intentionally introduced by the program provider.

### Error Correction and Concealment for Audio Services

It is recognized that error correction and concealment is a more critical issue for an audio service than for the video service, as reproduced audio errors may be more annoying than reproduced video /errors. In addition, some types of delivery (terrestrial, DBS, etc.) will be used by some viewers at the threshold of the service area. Therefore, the ATV audio system should include effective concealment methods to minimize audible disturbances caused by uncorrectable errors.

### Audio Services to the Visually and Hearing Impaired

The ATV audio system should allow programmers to deliver special audio services to the visually impaired (VI) and hearing impaired (HI) using the flexibly allocable channels. The VI service would typically contain a narration describing the picture content, and would be reproduced along with the main audio program in the receiver. The HI service would typically contain only dialogue, and would be processed for improved intelligibility.

## SERVICE IDENTIFICATION DATA (SVID)

Service Identification Data (SVID) should be incorporated into the ATV data stream. The SVID supplies descriptors for all digital services delivered by the ATV signal. The descriptors identify the digital services and indicate their locations within the overall data multiplex. Several methods may be used to deliver this information (such as packets with headers) providing they meet the requirements for this function. This issue has not been studied in depth and no recommendation is made as to the technique to be employed.

The SVID must be very reliable, since errors could cause total loss of ATV service. The SVID should be recognizable by the receiver as soon as possible after a channel change. The information carried by the SVID should be repeated frequently, so that a receiver can quickly recognize the available services after a channel change, and so that the redundancy can be used to improve reliability of the data. The format that is adopted for the SVID must be very flexible and allow new digital services to be defined and delivered in a manner that is compatible with all ATV receivers.

## ANCILLARY DATA SERVICES

Several types of data services should be pre-defined. Others may be added using the flexible allocation capability. Some types of data services, such as conditional access, may only be used by the alternative media.

### Captioning

The ATV emission system should provide a captioning system capable of delivering multiple versions of captions. Different versions of captions may be used to provide service in multiple languages, for the hearing impaired as well as the non-impaired. A minimum requirement on the captioning system should be that it allows three versions of captions: primary language, secondary language, and primary language for slow readers. A data rate sufficient to support this level of service should be allocable to the captioning service when captioning information is available. It should be possible to allocate additional data to the captioning service to support delivery of additional versions of captions (or subtitles). All ATV receivers should be required to decode and display captions.

### Program Guides

The program guide is an optional service. The program guide is intended to inform the viewer about the programming available on the particular ATV channel being viewed. The program guide service should be capable of delivering text accompanied by graphics. For the benefit of the viewer who is quickly scanning channels, lines of text identifying the current and upcoming programs should be provided and the program guide data should the found quickly in order to minimize the delay in displaying the current program information.

If an additional frequency agile tuner is available, either in the ATV receiver or at a cable headend, all available ATV channels may be scanned and the program guide information for all available channels combined to form a multichannel program guide.

### Conditional Access

Some applications of the ATV system will require data to be allocated to conditional access control. The required data capacity will depend on the size of the audience being served and the type of service being offered (i.e. pay-per-view or monthly subscription). There should be no fixed allocation of data to conditional access. Data capacity should be allocable to conditional access control at the option of the particular service provider. The T3/S3 Specialist Group made no recommendation as to the minimum data rate required to support a conditional access system.

(NOTE: Meetings of the ATSC Specialist Group on Digital Services T3/S3 were held concurrently with meetings of the ATSC Specialist Group on Interoperability and

Consumer Product Interface T3/S2. The two Specialist Groups worked closely on Conditional Access issues. This work on desirable attributes and features of encryption systems is reported in a companion paper, Reference 5.)

## CONCLUSIONS

The "strawman" approach to soliciting industry opinion and advice regarding digital audio and ancillary data services for ATV resulted in responses from a broad range of parties with ATV interests. The results of the T3/S3 studies indicate that an acceptable set of audio and ancillary services can only be provided by a flexible, extensible system which allocates data capacity to these services only on an as-needed basis. All remaining data capacity should be used by the video system for improved picture quality.

Advantage should be taken of the opportunity to design into a new ATV standard the ability to distribute multichannel audio with qualities commensurate with wide screen high definition pictures. At the same time, a flexible approach to data allocation will provide new ancillary data services and should satisfy the conditional access needs of Cable and other media.

### Acknowledgements

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## Appendix

Executive Summary of "Digital Audio and Ancillary Data Services for an Advanced Television Network", ATSC, February 3, 1992. (T3/186)

The Specialist Group on Digital Services (T3/S3) of the Technology Group on Distribution (T3) of the Advanced Television Systems Committee (ATSC) has been working for two years to identify the digital audio and data services that should accompany the Advanced Television (ATV) picture. T3/S3 has conducted technical studies and industry surveys, and has circulated a "strawman" proposal to stimulate industry discussion and feedback. Based on this work, T3 offers the guidance contained in this document. On February 3, 1992 the ATSC Executive Committee approved release of this material. It is hoped that the ATV system selected for use in the United States for terrestrial broadcasting, as well as ATV systems for the alternative media, will follow this guidance.

A major finding is that it is not desirable to select a fixed set (number and type) of digital audio and data services for inclusion into the 6 MHz ATV channel. This is because the data rate required for all potential services would negatively impact the picture data rate and affect the picture quality. It is recommended that the ATV system allow data to be allocated to digital audio and data services only on an as-needed basis. A flexible system of data allocation will require the use of only the minimum data capacity necessary for digital audio and data services at any time. Flexible allocation also allows the addition of new types of digital services in the future, with older receivers ignoring the new data types.

The ATV service will offer widescreen pictures, and this feature is expected to increase consumer interest and enjoyment of this new format. The audio corollary to the wider aspect ratio is multi-channel audio. Recognizing the significant limitations of two-channel stereophony for sound accompanied by pictures, it is recommended that the ATV service be capable of delivering five channel audio (left, center, right, left surround, and right surround). This is generally consistent with recent trends in the application of digital audio in the motion picture industry, and the draft recommendation from CCIR Task Group 10/1. Recent advances in multi-channel audio coding technology have reduced the data rate required for five channel audio nearly to that required for two independent audio channels. All ATV receivers would need to decode the provided service (which could vary from one to five channels) into the number of loudspeaker channels to be used (e.g. the five channel audio service may be decoded into mono for the low-cost mono receiver).

The ATV audio system should provide a solution to the problem of loudness uniformity among programs, channels, and delivery media. The average perceived loudness of dialogue should be uniform. The coded audio data should indicate the average dialogue loudness within the dynamic range of the coding system. Different programs may have varying amounts of headroom above this level which is available for dramatic effect.

The ATV audio system should allow audio service to be provided with a wide dynamic range. Since experience indicates much of the audience will prefer a restricted dynamic range, the audio coding system should incorporate an integrated dynamic range compression system. Information about the compression introduced may be incorporated into the audio data stream so that the receiver may optionally reverse the compression to restore the original dynamic range of the program.

The ATV system should allow service to be delivered to the visually and hearing impaired. The visually impaired may be served by an optionally allocated narrative audio channel. The hearing impaired may be served by the captioning system, or by an optionally allocated audio channel containing only dialogue which has been processed for improved intelligibility.

Flexible allocation of data capacity to audio allows audio service to be optionally provided to specialized audiences, including those with other languages. Audio services should be tagged to indicate language and type, so that the receiver may assume the burden of choosing the correct language audio service for the viewer (assuming the receiver has knowledge of the viewers preferred language), and so that the viewer may determine the types of audio available.

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# Digital Transmission Fundamentals for Cable Engineers

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## ABSTRACT

This paper will examine various techniques as they apply to digital transmission on CATV systems. A brief review of analog to digital conversion techniques, as well as an overview of digital modulation methods, will be presented.

CATV system infrastructure will be examined both from a current capabilities viewpoint and from a forward looking perspective considering the evolution of delivery and networking techniques with fiber, wireless and star-bus bidirectional topologies.

Because analog signal transmission will exist on CATV networks well into the future, techniques for integrating digital capabilities into a broadband hybrid analog/digital network will also be discussed. The evolution of these networks, and their interactions with existing and evolving TDM telephony networks will also be considered.

## INTRODUCTION

New cable television system technologies are being introduced at an ever-increasing pace, requiring that technical personnel assimilate the new technologies into their existing operations. Digital signal transmission on cable systems is just such an example. This paper will provide an overview of some of the key technical considerations involved in any kind of digital signal transmission, and more specifically, issues involved in a hybrid analog/digital cable television environment.

Some questions will be raised about specific techniques for cable digital transmission which are the subject of investigations at laboratories both nationally and internationally.

Additional questions posed relate to the deployment of digital technologies in a cable system, especially issues relating to system test and measurement and subscriber location equipment configurations.

## ANALOG TO DIGITAL CONVERSION

The first step in preparing an analog signal for transmission through a digital transmission system is the process of analog to digital conversion, often referred to as A/D conversion. The A/D process determines the maximum level of performance that is achievable in a "perfect" digital system, where perfect implies all coding and transmission processes are implemented with zero error.

The A/D conversion process is shown diagramatically in Figure 1. The first step is sampling of the properly band limited source waveform. The Nyquist Sampling Theorem[1] states that a signal of baseband bandwidth 1/T Hz can be completely represented by evenly spaced instantaneous samples at a rate of 2/T samples per second (sps). In order to eliminate ailiasing, or spectral overlap and interference in the signal reconstruction process, it is necessary to strictly band limit the source spectrum to 1/T Hz. Since perfect filters are impractical, sampling is typically performed at rates that are higher than 2/T, or signals are band limited with realizable filters to less than 1/T, or a combination of both techniques is used. As an example, a standard "4 kHz" voice channel for telephony services[2] is sampled at 8 Ksps after the source is band limited by a filter whose 3 dB bandwidth is 3.4 kHz and provides 80 dB of rejection at 4.0 kHz.

The result of the sampling process is a stream of impulses, evenly spaced at a rate of 2/T sps, with amplitudes that exactly represent the amplitude of the signal waveform. The next step in A/D conversion is to map the infinite set of amplitude values for these impulses into a codeword set that simultaneously meets the performance objectives for the channel and the capacity constraints due to the transmission channel. The mapping process is known as quantization, the impairment introduced is referred to as quantization noise. From a practical standpoint, an additional limitation must be considered, that being the availability of practical A/D conversion circuits which cover the bandwidth of the signal. For example, typical A/D circuits for sample rates around 50 Ksps, such as those used for CD audio, are readily available at 16 bit accuracy, with 18 bit units available for premium applications. A/D circuits for video channels in the 10 to 15 Msps range are typically 8 or 9 bit accuracy, with 10 bit units representing the premium or "best currently achievable" level of performance.

The simplest quantization process is linear pulse code modulation, or linear PCM. As depicted in Figure 1, linear PCM represents each impulse of exact amplitude with the code word for the amplitude which provides the closest match from the $2^N$-1 values available, where N is the accuracy or resolution of the A/D converter. Linear PCM means that the $2^N$-1 values are equally spaced. The total range covered by the available code words must be carefully chosen in order to accurately represent the source at its maximum and minimum instantaneous amplitudes without clipping.

The fundamental degradation inherent in the A/D process is quantization noise, a measure of the average error introduced in mapping the continuous analog amplitude space into the discrete code words available. For a simple sine wave which completely spans the $2^N$-1 code words, the quantization noise is given by[3]:

$$S/N = 6.02 * N + 10.8 \text{ dB } (\textit{pk - pk sig/rms noise}) \quad \text{(Eqn 1)}$$

For complex wave forms such as video, standard code word coverage of the amplitude space must account for the peak factor of the composite signal if distortion due to clipping is to be adequately controlled. Signals with large peak factors (peak to rms ratio) will experience larger quantization error if minimal peak clipping is required. Typical values for 8 bit coders (255 code words) are $V_B = 16$ and $V_W = 235$ ($V_{PP} = 219$) for component systems[4] (YUV), $V_B=64$ and $V_W = 212$ ($V_{PP} = 148$) for NTSC composite system coders, where $V_B = V_{Black}$, $V_W = V_{White}$ and $V_{PP} = V_{Peak\text{-}to\text{-}Peak}$. Coders using 9 and 10 bit code words ($2^9$-1 = 511 or $2^{10}$-1 = 1023 code words) would use comparable ranges.

Referring once again to Figure 1, it should be evident that sampling the wave form at higher rates permits more accurate representation of continuous amplitude wave by the sample stream, albeit at an increase in coder output data

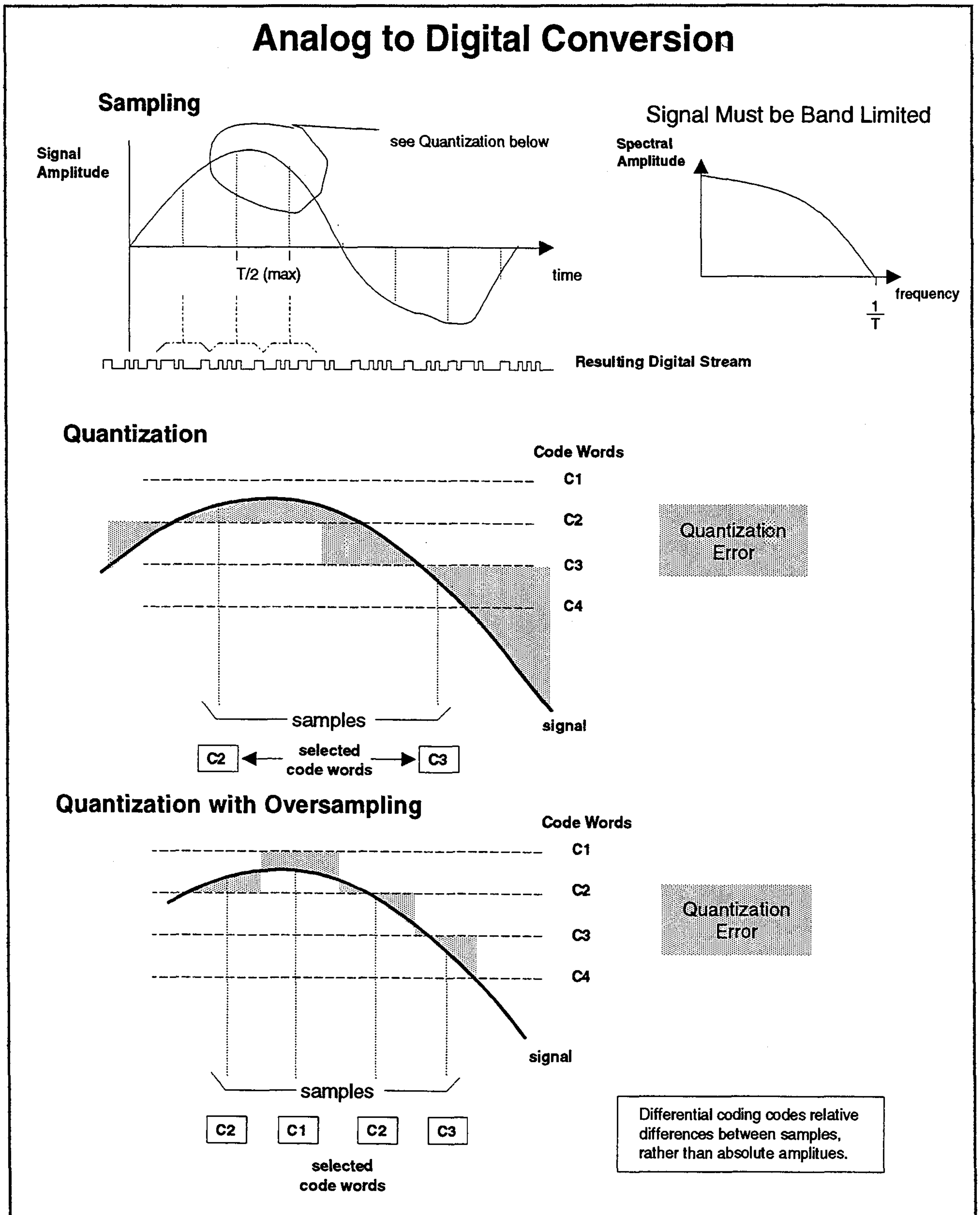

Analog to Digital Conversion
Sampling
Signal Amplitude
see Quantization below
T/2 (max)
time
Signal Must be Band Limited
Spectral Amplitude
frequency
1/T
Resulting Digital Stream
Quantization
Code Words
C1
C2
C3
C4
Quantization Error
samples
signal
C2
selected code words
C3
Quantization with Oversampling
Code Words
C1
C2
C3
C4
Quantization Error
signal
samples
C2
C1
C2
C3
selected code words
Differential coding codes relative differences between samples, rather than absolute amplitues.

**Figure 1**

rate. Taking into account over-sampling and dynamic range, equation 1 becomes:[5]

$$S/N = 6.02*N + 10.8 + 10*\log(f_s / 2f_v) + 20*\log[(V_W - V_B)/S]$$
where
$f_s$ = sampling rate
and $f_v$ = bandwidth of the source
(Eqn 2)

## BIT RATE REDUCTION

As described above, linear PCM generates a digital bit stream whose bandwidth is large relative to the signal being processed. For example, a "4 kHz" voice grade telephony channel becomes a 64 Kbps digital stream (8 bit linear PCM), an NTSC composite signal (6 MHz nominal) becomes an approximately 100 Mbps stream, depending on detailed coding parameters. In many applications, such as networked telephony, the robustness, ease of processing (switching, echo canceling, etc.) and low incremental cost of digital transmission have made this bandwidth increase not only acceptable but the superior choice from both quality and cost of service perspectives. The same can not be said for wide band signals, such as broadcast or cable video, where the available bandwidth of existing delivery system cannot support a simple analog to digital signaling

| | NTSC Composite Signal | | | | CCIR 601 Component YUV |
|---|---|---|---|---|---|
| Sampling Rate | 12 MHz | | 14.36 MHz ($4*f_{sub\text{-}carrier}$) | | 13.5:6.75:6.75 MHz |
| Bits/Sample | 8 | 9 | 8 | 9 | 8 |
| Rate (Mbps) | 96 | 108 | 115 | 129 | 216 |
| S (pk-pk) (digital words) | 148 | 297 | 148 | 297 | 219 |
| S (pk-pk)/N(rms) | 54 dB | 60 dB | 55 dB | 61 dB | 58 dB |

Table 1
S/N performance of Linear PCM for Video

Table 1 summarizes the S/N results, for typical signals and coder parameters that are in use or of potential interest in digital video. It must be noted that these results are ideal and do not address the degradations due to implementation difficulty. A detailed discussion of sampling and conversion degradations such as aperture effects (finite width versus impulse sampling), jitter on the sampling clock, ailiasing due to imperfect filters and digital noise effects is beyond the scope of this paper but covered well in several excellent references[6,7,8].

conversion.

Bit rate reduction techniques for digital signaling can be placed into two general classes—coding techniques such as Adaptive Differential PCM (ADPCM) and advanced compression techniques such as sub-band coding and transform coding. There is wide overlap among the various approaches, with simple coding improvements yielding reduction factors in the 2 to 4 range and advanced context sensitive techniques yielding improvements of 25 to 50 or more, depending on the source.

From a practical viewpoint, the simplest bit rate reduction techniques are similar to data compression techniques used with computer data where sequences of binary data are replaced with coded sequences. Differential coding techniques focus on transmitting the difference between successive samples (Figure 1) rather than the absolute sample amplitudes individually. Such techniques have achieved bit rate reductions in the 4X to 8X range in typical audio applications. The large (25X to 50x) reductions in coded video images, achievable with significant signal processing, are more fully discussed in Reference[9].

## PAYLOAD FORMATTING

After successful conversion from the analog

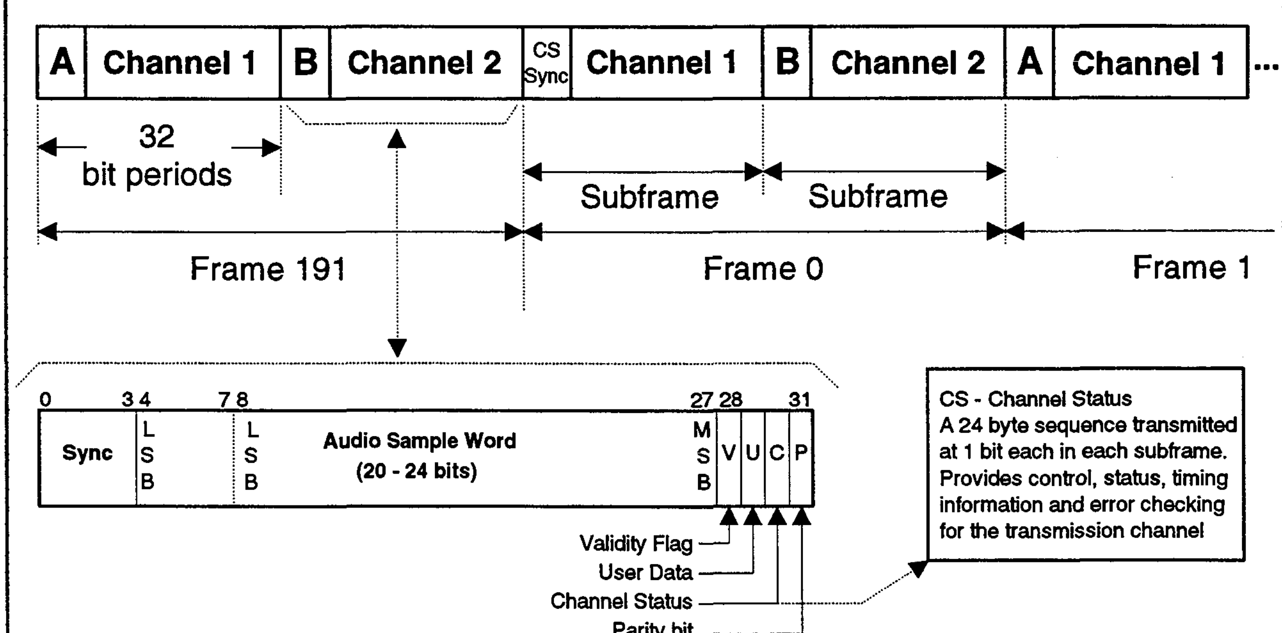



**Figure 2** - Payload Formatting Example -- AES/EBU Digital Audio

domain to the digital domain, the next step in processing for transmission is to add to the encoded stream a series of synchronization and control signals which enable the receiver to properly decode the signal stream. Perhaps the simplest familiar payload is the asynchronous character transmission protocol used in typical asynchronous data interfaces. In this instance, an initial synchronization bit ("start bit") and one or two "stop bits" are added to the signal stream along with optional parity bits for error detection.

A far more complex example of a formatted payload signal is shown in Figure 2, showing the byte (8 bit) level structure of the AES/EBU digital transport format[10] used for high performance digital audio channels. In this case, one or two digitized audio channels are formatted into a fixed bit rate digital stream with predefined functional and padding bits such that a common transport format interfaces with the transmission system. The added control information in the 48 bit header field delivers to the receiver all information needed to decode and interpret the serial stream. Once the receiver has recognized the "unique" 12 bit syncword, it is able to read other control bits for decoding and error detection functions.

Formats such as Figure 2 provide for a universally accepted ("standard") framework for the transfer of payload information and associated control information. Similar standards exist for video streams (e.g. D1 parallel and serial interfaces in CCIR 649), data communications (e.g. HDLC (high-level data link control) data packet protocols) and are used extensively in voice communications systems in international telephony systems (e.g. DS0, DS1, etc. for the North American Digital Hierarchy). Strict adherence to open interface definitions makes it possible for equipment from various vendors to interconnect and transfer information effectively. The increase in apparent complexity of formats such as the AES/EBU format of Figure 2, compared to a DS0 telephony channel at 64 kbps, is a direct result of the availability of economical high performance digital processing LSI. An obvious benefit for the user of this format is its ability to adapt to several different specific payloads utilizing the same common interface. There is a general trend in such standard interfaces towards higher complexity, driven by the need for flexibility and enabled by economical real time processing power.

## COMBINING AND MULTIPLEXING

In many cases, the available capacity of transmission facilities is significantly greater than that required by any individual service or communications system. In these cases, multiple formatted payloads are combined, or multiplexed, into a higher capacity signal for transport over the facility. In typical applications, the multiplexing system uses a small portion of the transport capacity for its own synchronization and control functions, similarly to the discussion of payload formatting above. The remaining capacity is then divided among the digital streams which share the channel capacity.

Several techniques are used to achieve the multiplexing step, synchronous and asynchronous time division multiplexing being the most prevalent at this time.

### Asynchronous Time Division Multiplexing

Each channel operates at a fixed clock rate which is arbitrary (asynchronous) relative to the multiplex system clock. Information is transferred to the multiplex interface at a rate which varies but is always guaranteed (by design) to be less than the "allocated" capacity for that channel. The multiplex interface then bit stuffs or adds additional bits to the channel such that its instantaneous rate matches the allocated capacity. This technique, one of several in use, is commonly called positive bit stuffing. As part of the multiplexing process, control information must be conveyed to the decoder end of the link to enable proper removal of the stuffed bits.

### Synchronous Time Division Multiplexing

In the case of synchronous multiplexing, the amount of data transferred between the digital source and the multiplex system is forced to exactly match the allocated capacity for the channel, typically by providing a clock signal from the multiplex system to the signal source (or, more typically, locking both the source and the multiplex to a common clock source). The synchronous multiplex has higher throughput (utilization of capacity for true information bearing applications) but typically requires a more complex timing distribution system.

### Statistical Multiplexing

As multiplex processing power increases, it becomes possible to efficiently operate a statistical multiplex. In this case, the multiplex capacity is made available to all input channels on an "as-needed" basis and channels randomly send data to the multiplex. The multiplex system allocates available bandwidth based on a current need basis. This implementation depends on the average statistics of the combined signal sources to not exceed the total capacity of the multiplex. While managing and communicating the continuously varying channel capacity information to the demultiplex end of the system, the benefit of statistical multiplexing is that channels see higher capacity available on an as-needed basis. This capability is particularly effective in situations where the signal sources do not generate continuous information streams and are "bursty" in nature, requiring occasional periods of high capacity relative to the long term average.

### Transmission

The transmission of a digital bit stream over a medium is performed by modulating a carrier signal with the digital information in the signal stream. The carrier might be an electrical signal, in the case of metallic media, or an optical signal for fiber optic signals. A simplified view of a digital link or regenerative section is shown in Figure 3. We use simple bipolar (alternate mark inversion) amplitude keying for this example, more extensive analysis of techniques can be found elsewhere[11].

The digital regenerator is comprised of a front-end amplifier, a carrier recovery or timing extraction stage, and a decision circuit which combines the signal and timing information to recreate the digital information. As shown in Figure 3, the transmission media introduces amplitude and edge degradation (in general) in the transmitted stream. Engineering rules specific to the application will specify the allowable loss in the media in order to ensure adequate margin at the decision circuit of the regenerator. Since most real noise sources are statistical in nature, the engineering rules will typically specify a minimum received power in

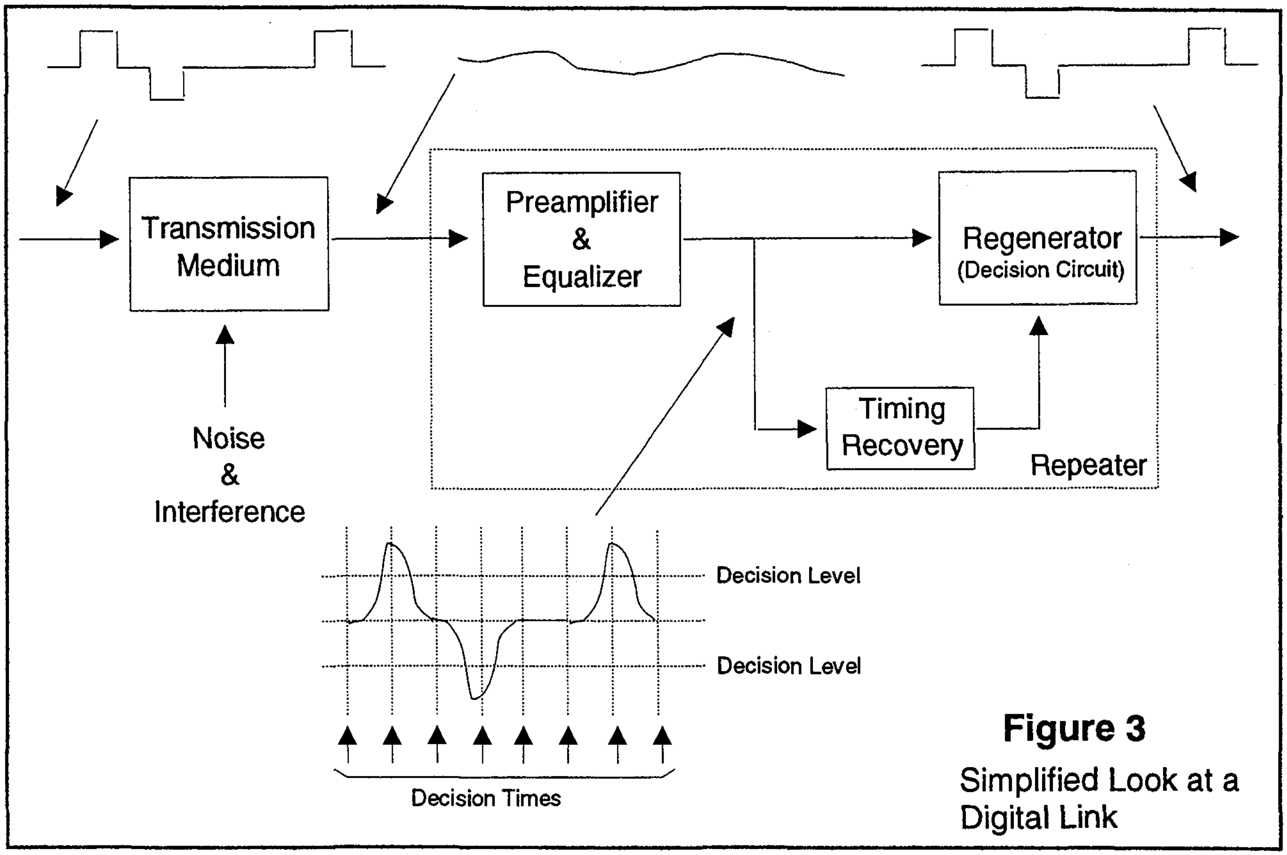



**Figure 3**
Simplified Look at a Digital Link

order to guarantee a particular error rate or probability of error.

Typical digital transmission systems consist of several concatenated regenerative sections, often with different media (wire, fiber, RF carrier) intermixed in the end to end connection. The dominant performance metric is the probability of error or error rate expected for the link. No link is completely error free (statistically), hence signal sources must be capable of properly handling error occurrences.

Binary baseband signaling has been the predominant format for digital transmission, particularly in fiber optic systems, due to its simplicity and robustness. For cable applications, however, multilevel transmission techniques based on passband channels and MODEM technology are popular due to the nature of the delivery channel and the need for compatibility with existing passband analog services. In a simplified view in Figure 3, we look at 4 level signaling. In this case, two bits of transport information are converted to one of 4 levels ($00_2$=0,$01_2$=1,$10_2$=2,$11_2$=3) and the appropriate level transmitted and detected by a four level decision circuit. Such a system requires significantly better noise performance than the binary signaling system but achieves higher throughput per unit of bandwidth. More on cable applications in the following sections.

## DEMULTIPLEXING AND PAYLOAD SEPARATION

The process of recovering individual bit streams at the receive end of the system is performed by continuously scanning the signal stream for the format or synchronization patterns (frame words) inserted in the multiplexing process, using the embedded control information to interpret the bit stream and routing the individual channel streams to their appropriate connections. An important consideration in the demultiplexing process is robustness or immunity to errors in the digital stream.

In systems that use fixed frame word spacing, word detectors typically "fly wheel" on the frame pattern, requiring that several frame words be detected with errors before initiating a new frame search, thus minimizing the effects of random errors. Systems that use variable synchronization spacing generally require a lower (improved) bit error rate performance from the transmission channel in order to achieve equivalent performance.

## DIGITAL TO ANALOG CONVERSION

The final step in the process, assuming an analog interface to the end user (speaker, analog TV) is digital to analog (D/A) conversion. This process consists of reconstructing the sampled impulse stream from the digital code words, converting the code words to impulse amplitudes and low pass filtering to recover the original signal envelope.

## DIGITAL SYSTEM DEPLOYMENT

With the above digital overview as background, the impacts that digital technology will have on CATV system infrastructures will be examined. The traditional tree and branch system architecture has serviced the industry well for almost four decades. In fact, this point to multi-point network is a basic strength of the cable industry. The delivery of entertainment services has required large amounts of bandwidth in the downstream direction from the headend to the subscriber.

This will continue to be the case in the future as cable networks evolve from just entertainment networks to telecommunications networks. What will change as these networks evolve is that lightwave systems will be increasingly deployed deeper and deeper into the system in a fiber to the service area scenario. This will further reinforce cable system bandwidth expansion, setting the stage for expansion into digital technologies discussed above.

The cable television industry has become comfortable over the years with FDM (frequency division multiplexing) technology as a convenient means to allocate spectrum on a cable system. As digital technologies become available to the cable industry, it seems logical to continue using FDM techniques to apportion spectrum for both analog and digital services. This will allow a hybrid analog/digital structure to exist well into the future to support both types of services.

In a paper titled "The Evolution of CATV to Broadband Hybrid Networks", by Carl J. McGrath, AT&T Bell Laboratories, which is being presented at this conference, the author discusses possible methods of implementing a hybrid analog/digital headend architecture.

The cable system forward bandwidth is divided into a lower frequency portion for analog signals and a higher frequency portion for digital signals (compressed digital video/audio and other data signals). Digital "bandwidth" is added to the system as needed to support digital services, allowing for a modular expansion capability by utilizing digital modem technology.

## MODERN MODEM TECHNOLOGY

As discussed earlier under "Transmission", the digital signals must be modulated onto an RF carrier to be carried on a broadband cable television system. The last 10 to 20 years or so have produced an enormous improvement in modem technology in two general classes: power efficient modems and spectrally efficient modems.

Power efficient modems generally are described as those modems which produce less than 2 b/s/Hz (bits per second per Hertz). They are used in power-limited situations such as existing satellite communication systems and digital cellular radio systems.[12]

Of more interest to the cable industry are spectrally efficient modems which produce more than 2 b/s/Hz. In fact, the range is from 2 b/s/Hz all the way to 10 b/s/Hz for the most exotic technology (1024 QAM). (See Table 2.) Likely candidates for cable applications are M-ary QAM, 4-VSB AM, and QPRS (quadrature partial response systems).

| Modulation technique | Nyquist rate theoretical efficiency bits/Hz | Practical efficiency bits/s/Hz | C/N Required for Pe = 10 (theoretical) |
|---|---|---|---|
| QPSK 4-QAM | 2 | 1.2-2.0 | 15.0 |
| 9-QPRS | 2 | 2.0-2.8 | 17.5 |
| 8-PSK | 3 | 2.5-3.0 | 20.5 |
| 16-QAM | 4 | 2.5-3.5 | 22.5 |
| 49-QPRS | 4 | 2.5-4.3 | 24.5 |
| 64-QAM | 6 | 4.5-5.0 | 28.5 |
| 128-QAM | 7 | 4.5-5.5 | 31.5 |
| 225-QPRS | 6 | 5.7-6.3 | 31.0 |
| 256-QAM | 8 | 5.0-7.0 | 34.5 |
| 512-QAM | 9 | 5.5-7.5 | 37.5 |
| 961-QAM | 8 | 6.0-8.0 | 37.0 |
| 1024-QAM | 10 | 7.0-9.0 | 40.5 |

**Table 2**

Spectral Efficiency and C/N Requirement of Wideband Modems
*Adapted from K. Feher, "Advanced Digital Communications"*

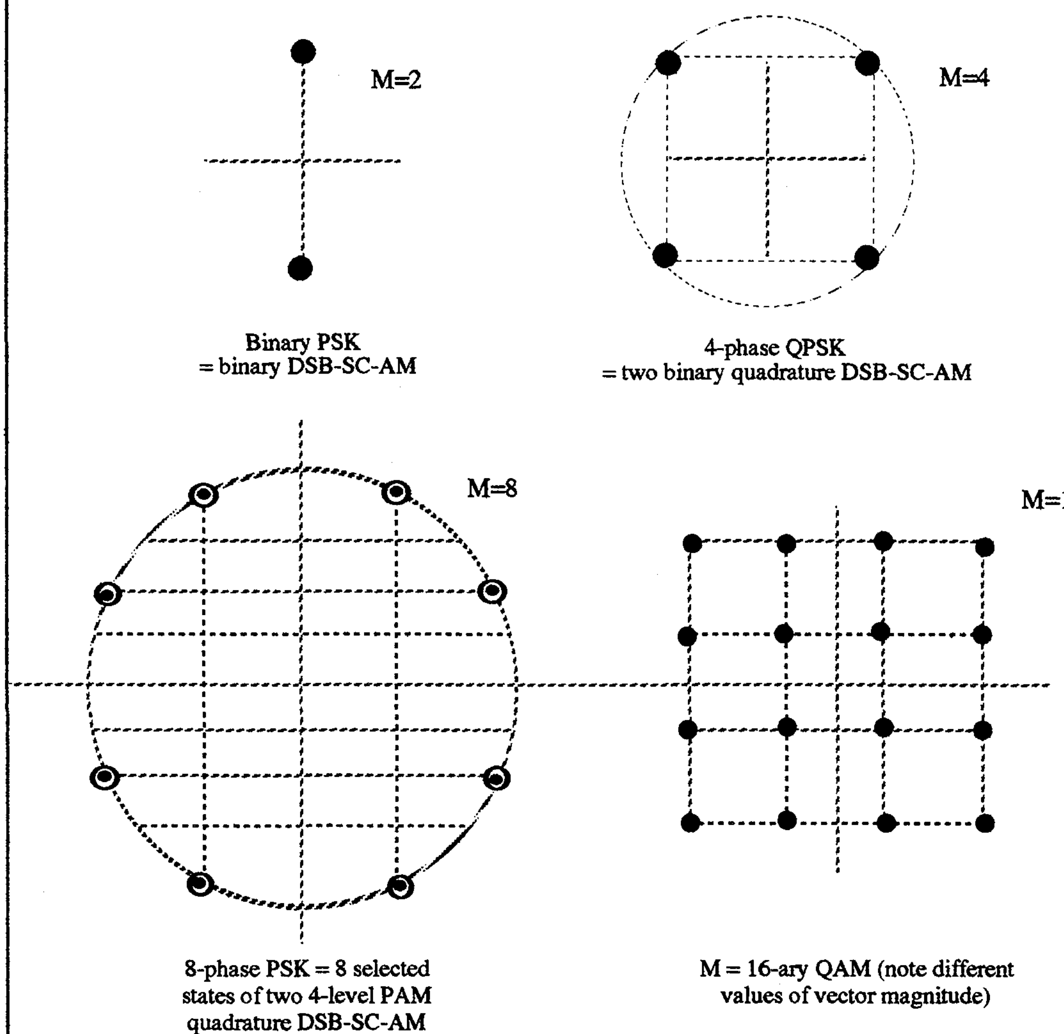


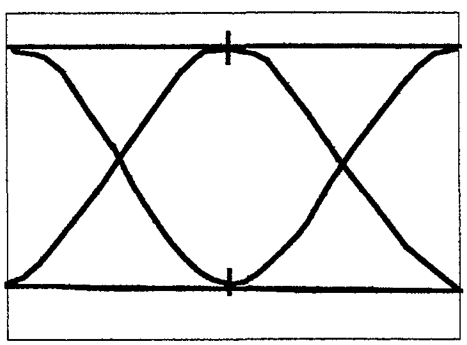

**Figure 4**

Signal-State Space Diagrams of M-ary PSK and QAM Signals

*From K. FEHER "Digital Communications: Microwave Applications"*

When dealing with these modem technologies, two of the concepts involved are vector state space diagrams (Figure 4), also known as constellation diagrams, and eye diagrams.[13] A vector state space diagram is a display of all possible trajectories of the carrier — as it passes through each of the possible phase and amplitude states. For example, the constellation for 16 QAM shows 16 possible states of phase and amplitude of the carrier.

An eye diagram is formed by applying the data signal to the Y axis of an oscilloscope, externally triggering with the system clock, with the sweep time set approximately equal to the symbol time. (See Figure 5.) An eye diagram is a useful tool in analyzing system performance since by measuring the percent of eye height in the demodulated received data, one can determine the likelihood of the system meeting the design goal error probability (Pe).



Idealized Eye Diagram

Figure 5

## COEXISTENCE/PERFORMANCE ISSUES

Now that the digital signals are packaged for transmission, the question of compatibility arises. Will the digital signals interfere with the analog signals? Conversely, will the analog signals interfere with the digital signals? How robust will various forms of digital modulation be on cable systems in the presence of phase noise and reflections? Should digital channels be multipexed into one high-speed digital stream which requires high performance circuitry at the subscriber location? Should digital channels be packaged individually like analog to ease the decoding costs and complexities in the subscriber terminal? These questions are just some of the issues being investigated by various laboratories both at the national and international levels.

## TEST AND MEASUREMENT ISSUES

With the advent of digital transmission on cable systems, there will be some tests required which are not at present in the repertoire of cable system tests. Such tests might include one or more of the following:

1) Phase noise
2) Phase delay
3) BER (Bit Error Rate)
4) Eye height of demodulated data

Some testing will be required at the headend on satellite delivered digital signals as well as on signals delivered by digital headend interconnects. Testing locations should be identified which are representative of subscriber locations which are "deepest" in system cascades whether it is an all coaxial plant or a hybrid fiber/coax system. Some specialized digital test equipment will undoubtedly be required for the cable television analog/digital environment.

The goal here is to maximize the height of the eye diagram for the demodulated data so that the BER is maintained to system design standards. While in analog pictures, phase noise might result in graininess and reflections might result in ghosts, they both can cause the eye diagram to partially close, resulting in difficult digital data detection. As long as the system design goal BER is maintained, the digital transmission process will not degrade the SNR of digital video images as measured at the headend.

It is important to note that as the CNR of an analog picture degrades, it results in a progressively noisier picture. In the digital domain however, once the BER degrades below a minimum threshold value, digital data detection may not be possible. At this point, the subscriber's decoding equipment would most likely result in a freeze frame display of the last digital video frame which was accurately received.

## SUBSCRIBER EQUIPMENT ISSUES

Once the analog and digital signals get to the subscriber's location, several questions must be answered about the topology of the subscriber's distribution "system". Depending on what method of securing analog signals is being used in the cable system (positive or negative trapping, interdiction or scrambling) there will be subscriber equipment located at the tap, on the side of the house, or at the TV set.

Are any of these locations the appropriate place to locate the digital circuitry necessary to receive, demodulate, demultiplex, decrypt, and convert from digital to analog the digital signals appropriate to a particular subscriber?

If the tap location is chosen because of possible cost sharing in a multi-output device, there are considerations of powering and sensitivity to temperature swings and other elements of the rather hostile outdoor environment. In some of the newer feeder designs utilizing "superdistribution" techniques, there is no power in the tapped feeder cable so that drop powering would have to be used.

A location on the side of the house would still be subject to all the conditions of an outdoor environment, would be powered from the home, and would service all outlets in the house. The configuration would provide for modular implementation, allowing "service - specific" cards to be plugged in on an as-needed basis.

A variation of the side of the home approach would be a unit designed for an indoor environment in some out-of-the-way location (such as in a closet or in the basement). Obviously, an indoor location would not work for interdiction of analog delivered services since the signals are in the clear on the drop coming into the unit, but eliminates outdoor environmental issues. Descrambling of analog signals could be also performed for all outlets at this central location.

A top of the set or back of the set location requires a box for each TV set to which any secured services (analog or digital) are to be delivered. A multiport equipped TV or VCR could utilize a multiport decoder for secured analog services but a separate digital receiver would be needed for digital services. The output of the digital receiver would be both RF (channel 3 or 4) to accommodate standard TV sets and baseband audio and video to feed a receiver/monitor.

Wherever the digital receiver is located, it would be helpful from a system troubleshooting viewpoint to have a demodulated data port available so a service technician could plug in a BER meter. By comparing the reading to the system design goal BER, the technician could quickly determine if there was a problem at the particular location with the received data.

## CONCLUSION

An overview of digital transmission fundamentals has been presented, as well as a discussion of some implications of digital deployment in a hybrid analog/digital cable network.

As fiber gets deployed deeper into the network in various star-bus architectures serving smaller numbers of homes (less than 1,000), the industry will be in a good position to offer near-video on demand services and other personalized communications services.

Modem technologies will be further refined for the cable environment, and coupled with the improved carrier to noise ratios provided in the smaller service areas, will allow higher constellation state digital modulation techniques to be applied (e.g., 64 QAM). The result will be higher bandwidth efficiencies allowing for narrowcast, more personalized services to cable subscribers.

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# EVALUATION OF SOLID-STATE CROWBARS AND GAS-DISCHARGE TUBES IN CATV SURGE SUPPRESSION APPLICATIONS

Peter Deierlein
Magnavox CATV

*Abstract*

*Gas-discharge surge arrestor tubes and the recently-introduced solid-state AC crowbars perform identical surge suppression functions in CATV systems, but each operates on very different principles and possesses different performance limitations. CATV equipment must withstand two distinctly different types of surge phenomena, and each surge suppression device is uniquely suited to protect from the effects of a particular type and level of surge. Analysis of both equipment failure and field test data is used as a guide in the selection of appropriate surge protection devices, which are then tested to determine their relative strengths.*

## INTRODUCTION

"Outages" are a major issue for the CATV industry, and the primary cause of service outage is equipment damage (or blown fuses) from exposure to surge voltages and/or currents in excess of design limits. Bonding, grounding, and equipment ruggedness have increased significantly, but no matter how much the design limits are improved, surge suppression devices are required to suppress surges in excess of the limits. With the move towards elimination (or up-sizing) of fuses, the surge suppression devices have become the "weak link" in the system.

Meanwhile, the primary protective devices have improved performance and ruggedness. Solid-state devices have advanced from their secondary role so that they now rival the performance of traditional primary devices. Unfortunately, the devices have fundamentally different characteristics and limitations, and direct comparison of device capabilities has been difficult because the device performance is stated under different conditions. The purpose of this work is to compare the performance of the different types of surge protection devices under conditions which are appropriate to CATV applications.

## SURGE CHARACTERISTICS

While the electrical term "surge" is most often meant to define a potentially damaging temporary increase in circuit voltage, the term "power surge" is more accurate in defining this condition as it relates to CATV equipment.

This is because any surge protection device that does not function by disconnecting the protected equipment causes an increase in circuit current and power as the voltage is clamped to an acceptable level. (While a "disconnecting" type of surge protector would be desirable, such devices are too slow and/or not suitable for use in RF circuits.)

Electrical surges may be divided into three general groups based on duration and amplitude. The most common type of surge is of relatively low amplitude, and in most CATV applications, any surge that does not result in a voltage increase of more than 50% may be disregarded. Surge events that cause voltage increases over 50% will be defined as either long or short duration, with a dividing line of 1 milli-second.

Short-duration surges (also known as "impulses") due to lightning strikes and switching transients are well-known and have been characterized by the IEEE for various applications, which unfortunately do not include CATV. The "IEEE Guide for Surge Voltages in Low-Voltage AC Power Circuits," (ANSI/IEEE C62.41-1980, formerly designated IEEE Std 587-1980) establishes standards for devices connected to 120 VAC power, and their location category "B" (for major indoor feeders and short branch circuits) appears to be a worst-case for CATV applications.

Two impulse waveshapes are defined by the IEEE, a 100-kHz oscillatory wave of .5 micro-second rise time decaying by 60% every 10 micro-seconds, and a uni-directional impulse of 1.2 micro-second rise time with 50 micro-second decay ("1.2 x 50 uS") for high-impedance loads and 8 micro-second rise time with 20 micro-second decay ("8 x 20 uS") for low-impedance discharge current. Amplitudes for these waveshapes are defined as 6000 Volts for 100 kHz and 1.2 x 50 uS high-impedance waves, 3000 Amps for the 8 x 20 uS low-impedance wave, and 500 Amps for the 100 kHz low-impedance wave. While power dissipation in a surge protector can be quite high (up to 900 kW peak), total energy is low (about 10 Joules) due to the short duration of the surge.

Long-duration surges due to imbalances in distribution system powering range upwards from 1 milli-second to many hours. While the 60 VAC cable power system would appear to be isolated from the effects of power distribution fluctuations by the regulating qualities of the line power supply's ferroresonant transformer, Herman and Shekle showed how current sharing between the power company's neutral conductor and the CATV system's cable sheath can cause significant increases in cable voltage. The power company uses primary fuses and circuit breakers to interrupt high-amplitude imbalances and overloads over 200 Amps, but these devices can take up to 11 cycles to activate. Power dissipation in a surge protector can be moderately high (over 60 kW), but total energy (10,000 Joules over 11 cycles) can be tremendous.

## FIELD EXPERIENCE

It is possible to deduce a significant amount of information regarding the type and magnitude of CATV surge phenomena by studying field failure patterns and equipment failure modes. Over a two-year period, the equipment failures were concentrated near the ends of powered segments (normally the lowest voltage points), and even oversized Metal Oxide Varistors (MOVs) experienced catastrophic failures at these points. Since many failures occurred during clear weather, they did not appear to be statistically coincidental with lightning storms (although the study areas were located in high-lightning portions of the country).

When the CATV system's DC power supplies were ruggedized to withstand peak input voltages of 400 and 500 Volts (up from 150 Volts), failure rates were substantially reduced even when other types of surge protection was removed. Often, MOVs would fail in the open condition with no other failures in the ruggedized equipment. As will be shown, MOVs offer good protection from short-duration surges, but do not provide appropriate protection against long-duration surges. All the data pointed away from the short-duration impulses, strongly implicating the long-duration surges as the major cause of equipment damage.

While equipment ruggedization resulted in substantial reductions in failure rates, an opportunity for further study occurred at a site experiencing a unique failure mode (multiple instances of circuit conductor destruction at a single location) along with a higher than normal overall failure rate. An "RMU" (Remote Monitor Unit) commercial surge monitoring device was obtained, and a custom interface was constructed to facilitate its use in CATV systems. The equipment, which was intended to measure short- and long-duration surge voltages on the 120 VAC power line along with voltage differences between line neutral and safety ground, was selected specifically for its small size, low power consumption, and unattended operating capability.

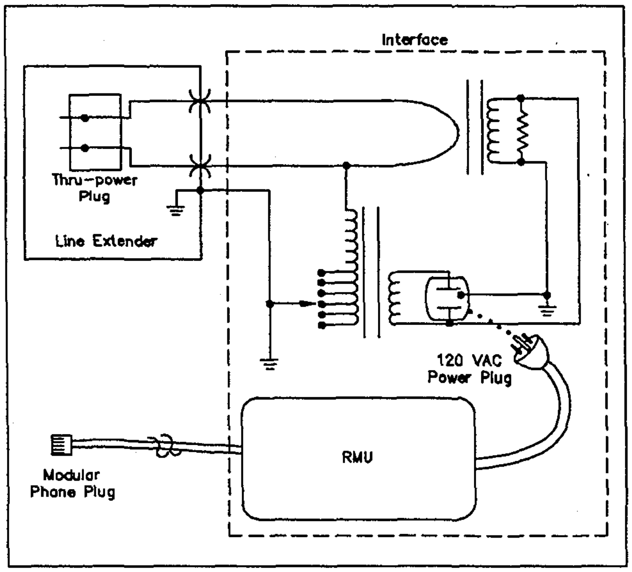


*Figure 1.*
*Remote Monitoring Interface for CATV*

The CATV interface enclosed the RMU, allowed it to be powered by (and to monitor) the cable system's 40-60 VAC power with negligible loading effects, and adapted the neutral-to-ground feature to measure center conductor current surges. A detailed functional diagram of the configuration is shown in Figure 1. The only non-cable connection to the device was a local telephone line, connected to the RMU's internal modem. The RMU contained an internal battery backed-up memory, and was polled several times a week for a period of 7 months from June 1989 through January 1990.

The RMU was installed at the "problem" location, a line extender in a residential section of Cleveland, Mississippi near a cotton processing facility. This location had experienced repeated outages (some of which were not related to severe weather), and had suffered damage to internal AC circuits on two occasions. The nominal AC cable power at this location was 47 VAC with less than 1 Amp through-current to one following line extender on the feeder. The through-current carrying conductors were ruggedized to prevent further damage (no damage occurred during the test). The line extender was equipped with its normal complement of two medium-duty "gas tubes."

The RMU was programmed to log the time, duration and maximum RMS value of each long-duration voltage surge over 50 Volts, the time and maximum peak amplitude of each short-duration impulse over 80 Volts, and the time and peak amplitude of each current surge over 14 Amps. While the current surge measuring subsystem did not possess sufficient bandwidth or resolution to measure short-duration impulse currents, current surges over 1 mS were logged with 16 mS (one-cycle) resolution. The logging thresholds were intentionally set relatively low to avoid losing data associated with other surges, since the RMU's firmware was designed to treat each surge type individually.

Over the seven-month period, 67 "events" were logged. An "event" is herein defined as any surge or series of surges within a 1-second period. In 16 cases, power was lost for periods ranging from 1 second to 30 seconds, and in one case, power was lost for a period of 7 minutes following a single 66 Amp current surge. In all but one event, one or more surges were logged in conjunction with the power loss.

Short-duration voltage surges were relatively rare and of low amplitude, but one event exceeded the 800 Volt measurement capability of the RMU. The >800 Volt measurement was accompanied by the highest amplitude current surge sequence measured throughout the test: 5 cycles ranging from 306 to 408 Amps peak, followed by a loss of power for 2 seconds. According to the system engineer, the weather was clear all day on the day of this event. A total of 14 surge events included short-duration voltage surges, and in all but four, at least one (usually two or more) current surge was logged in conjunction. More than one impulse was logged in a total of five events. During one event, six individual impulses were logged within 1 second (the highest was 382 Volts, which was also the 2nd highest overall). Sixty-four percent of the short-duration impulses were between 100 and 382 Volts.

The long-duration voltage surges were relatively low in amplitude at this location, with a maximum of 88 Volts for 140 mS. While they were often associated with other surge types, long-duration voltage surges were logged alone in 15 out of 31 events. During one large lightning storm, four individual long-duration voltage surge events were logged within 70 minutes, with only one other surge event (a 16-Amp current surge) that day.

Current surges were logged in 48 events, with a total of 113 surges and as many as eight in a single event. Of these, only five events contained a surge over 100 Amps, while 33 events were entirely below 50 Amps. In 25 events, current surges were logged alone (not in conjunction with other surge types), with none of the 25 over 100 Amps and all but four under 50 Amps. Only two events contained surges between 200 and 300 Amps, and the five highest individual surges (in order: 408, 306, 322, 318, and 322 Amps) were all within the single event previously mentioned. This single event is of great interest, since it also contains the highest impulse measured (over 800 V) followed by a 2-second power loss, all under clear skies. This event may have been associated with a major power fault at the nearby processing plant, but it is difficult to explain the high-amplitude impulse under clear skies.

## PROTECTIVE DEVICES

Three basic types of protective devices are commonly used for protection of consumer and telecommunication devices: the Metal Oxide Varistor (MOV), the gas-filled surge arrestor ("gas tube"), and various types of silicon-based devices such as ruggedized zener diodes ("Tranzorb" is a common trademark) and thyristors. (SIDACs, SCRs, and TRIACs are members of the thyristor family.) Of these, the MOV and the zener diode are simple voltage-limiting devices, while gas tubes and thyristors are "crowbars" which clamp the circuit voltage to a low value when activated by a higher "trigger" voltage.

### MOVs

Voltage-limiting devices are fundamentally restricted in that for equal surge current levels, power dissipation is 100 to 1000 times higher than for the various "crowbar" devices. While the MOV makes up for this limitation by spreading the dissipation over the largest area, the zener diodes have a very small active area. By clamping circuit voltage to a low value, crowbar devices reduce dissipation and, therefore, reduce surge energy.

MOVs are available in a wide range of voltage and energy ratings. They are widely used for protection against the impulses defined in the IEEE guide because they are inexpensive and generally perform well. Energy ratings for devices of manageable dimensions vary from 10 to 70 Joules, with maximum current ratings of up to 6500 Amps. MOVs have been well-characterized for their impulse-suppression application, and are known to degrade significantly over time when subjected to events near their maximum ratings. MOVs are not recommended as protection against the long-duration surges due to the

MOV's limited energy ratings. While they are used for "secondary protection" inside CATV power supplies and some modules, MOVs are not suitable as sole protection.

Gas Tubes

More accurately described as gas-filled surge voltage protectors, gas tubes are crowbar devices which have been used for surge suppression in telecommunications applications for many years. They are triggered at relatively low voltages by a gas ionization process similar to that of neon "glow tubes" and cold-cathode displays. When current flow increases beyond about 100 mA, their terminal voltage drops to around 20 Volts as an arc forms. Gas tubes in CATV applications have been ruggedized so that the heavy-duty types have over 100 times the ratings of early telecommunications types. While some have traditionally exhibited limited life due to the formation of internal debris during high-current surge conditions, recent versions have been developed with specially-coated electrodes to eliminate debris formation.

Special gas tubes developed for CATV use exhibit "follow current" ratings over 400 Amps. The "follow current" rating is that current for which the device will immediately return to its normal high-impedance state following a surge. Gas tubes are the only surge protection devices which are specifically characterized for suppression of long-duration surges, and the heavy-duty types now used in some CATV equipment are rated at 20,000 Amps impulse current. This is by far the highest rating of the devices considered here, exceeding the ANSI/IEEE specification by a large factor. Unfortunately, no manufacturer's data relating to service life is available for high-amplitude, long-duration current surge applications.

Zener Diodes

Due to their limitations, all but the largest ruggedized zener diodes are used only in sensitive low-voltage circuits for which the only concern is the short-duration transient, and for which the circuit impedance limits the maximum current. In this application, zener diodes are a very effective "last line of defense" against short-duration pulses due to their ultra-fast response time. Peak power ratings for typical devices are 1200 Watts for 1 milli-second, or 1.2 Joules. Larger versions rated for up to 15 kW are available, but besides being too large and expensive for CATV applications, they fall far short of requirements.

SIDACs

SIDACs are solid-state members of the thyristor family. These small devices emulate many of the electrical characteristics of gas tubes, but the SIDACs have limited energy dissipation capabilities. While SIDACs are recommended by their manufacturers as "line transient clippers" and for other AC line uses, their maximum rating of 20 Amps (for 16 mS) restricts their use in low-impedance circuits. SIDACs are ideal triggers for other thyristors (SCRs and TRIACs) in crowbar circuits. Some manufacturers have combined SIDACs and TRIACs into monolithic components designed specifically for surge suppression, but these devices are equivalent only to the light-duty gas tubes.

SCRs and TRIACs

With their continuous current capacity, low terminal voltage, and seemingly limitless service life, high-power thyristor devices such as SCRs and TRIACs appear to be ideally suited for use in "crowbar" applications. Crowbar circuits containing them have been successfully tested for several years in CATV power inserters, and these circuits have successfully completed testing in accordance with the ANSI/IEEE standard.

However, close scrutiny of the manufacturer's notes and specifications relating to all SCRs and TRIACs raises many questions relating to their suitability for use in surge suppression applications. Thyristors are not characterized for high-current impulses below 1 mS, and the technology specifically limits maximum dI/dt (relating to rapid changes in current) to values well below those required by the ANSI/IEEE standard.

While the published non-repetitive peak surge current specifications for some thyristors are quite high, they are considered to be overloads. One reputable manufacturer notes that "Usually only approximately 100 such current overloads are permitted over the life of the device," and goes on to state that "... neither off-state nor reverse blocking capability is required on the part of the thyristor immediately following the overload current." (This constitutes a limitation similar to the "follow current" gas tube rating.) Thyristor "AC-crowbar" applications in past linear CATV power supplies have not been very reliable. Although thyristors have long been used as DC-crowbars at power supply outputs, such applications are not subject to more than one or two surges over the life of the power supply.

In all fairness, none of the devices considered here has ever been fully tested under the entire range of surge conditions measured, and there has long been questions relating to device service life in CATV applications. Given the limited life of light-duty gas tubes in some severe CATV applications, it has been surmised that the lifetime of a device may be significantly limited by rapid sequences of surges similar to those which were logged in the field measurements described above. However, little if any data was available relating to reliability under rapid-sequence conditions, and a comparative testing program was suggested. Since no equipment is available for the type of testing needed, it was necessary to construct equipment which could generate (under controlled conditions) the repetitive current surge recorded most frequently in the field.

## LAB TESTS

The initial strategy was to test at least five devices of each type to failure using as many combinations of current and duration as possible, under conditions similar to CATV applications (with 60 VAC applied). The equipment had to generate a variable current surge with controllable duration and sufficient power to cause immediate failure of the most rugged devices, and had to be able to record current and voltage waveforms during and immediately following the surge. Generating up to 1000 Amps for 11 cycles (176 mS) with sufficient voltage to trigger the protective device (at least 200 Volts) proved to be the most difficult challenge.

## TEST SETUP

Due to the substantial power requirements, special facilities were required. Power Technologies Inc. of Schenectady, NY, professional consultants to the power distribution industry, suggested the use of a site normally used to test high-voltage transmission lines. The site was powered by a dedicated primary branch line from a major power company substation, and was capable of delivering over 1000 Amps with an open-circuit voltage of 500 Volts.

Figure 2 shows the test set-up in which 208 VAC from the main power panel connects through an adjustable reactor to the secondary of a standard 50 kVA distribution transformer, providing an isolated 4160 VAC source to a step-down transformer delivering 442 VAC (open circuit). A high-speed contactor delivers timed test current pulses adjustable from 1 to 11 cycles to the device under test, which is connected across a conventional 60 VAC ferroresonant CATV line power supply to simulate the CATV environment. The adjustable reactor limits the maximum surge current to values between 140 and 1000 Amps, and the 60 VAC (quasi-squarewave) line power supply is powered by a small generator to keep the surges isolated from the rest of the power system. A digital storage oscilloscope captures the surge event through current and voltage transformers (for isolation), and the oscilloscope display is downloaded to a computer for permanent storage.

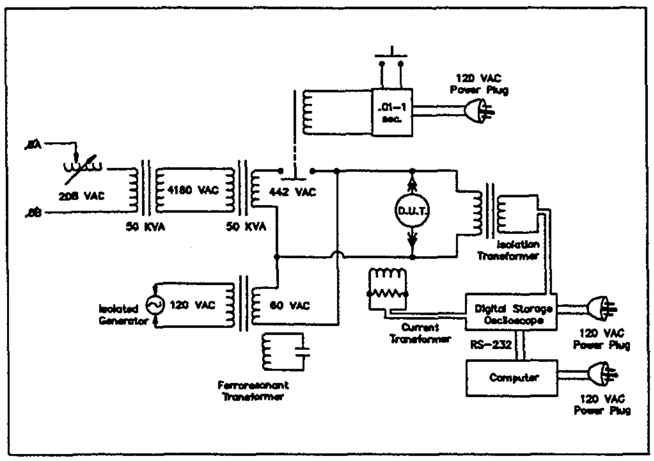


*Figure 2.*
*High-Energy Surge Testing System*

Initially, short-circuit surge currents were recorded for each reactor tap, yielding an available range of 140 Amps to 640 Amps (plus 1200 Amps without the reactor). Test results appear below in tabular form, with gas tubes listed separately from SCR-based crowbar circuits for easy reference. (While the actual testing sequence tested gas tubes and SCRs together for setup consistency, the results are numbered and grouped for clarity.) All devices were tested with surges of 5 cycles duration at 60 Hz (80 mS), and were allowed to cool (typically 15 to 30 seconds, depending on device and current) between surges. All currents are given in peak amperes. Figures 3-5 can be referenced as examples of device behavior.

## TESTING GAS TUBES

The first six gas tubes were tested to determine the maximum level that they would withstand.

| | |
|---|---|
| Tube #1 | Survived 1 surge at 400 Amps<br>Failed shorted during a second surge at 640 Amps |
| Tube #2 | Survived 2 surges at 640 Amps<br>Failed opened during a third surge at 640 Amps |
| Tube #3 | Survived 1 surge at 640 Amps<br>Failed shorted during a second surge at 640 Amps |
| Tube #4 | Survived 1 surge at 600 Amps<br>Failed shorted during a second surge at 600 Amps |
| Tube #5 | Survived 3 surges at 600 Amps<br>Failed shorted during a fourth surge at 600 Amps |
| Tube #6 | Failed shorted during the first surge at 600 Amps |

Figure 3 shows typical gas tube performance at 600 Amps, with current on the top trace and voltage on the bottom: At the beginning, the voltage trace shows the 60 VAC, which drops to 20 VAC as the surge current begins. Following the completion of 5 cycles, the surge current is interrupted, and the 60 VAC resumes with no evidence of "follow current."

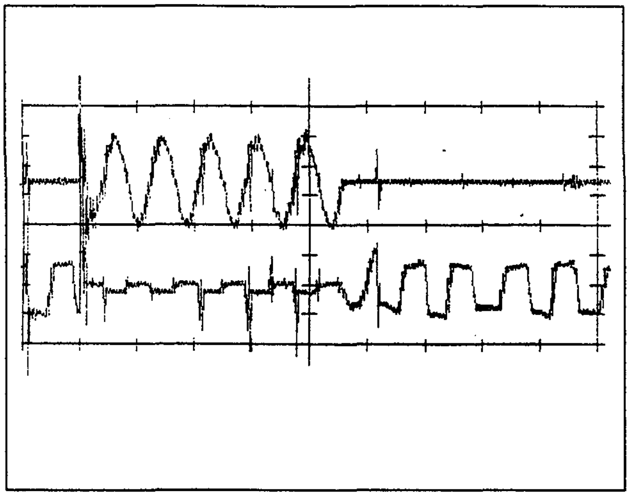

*Figure 3.*
*Typical Gas Tube at 600 Amps*

Of the six failed devices, all but one failed in the shorted condition as shown in Figure 4 (gas tube #1), where continuous "follow current" (barely visible at about 20 Amps) follows the completion of the 640 Amp surge.

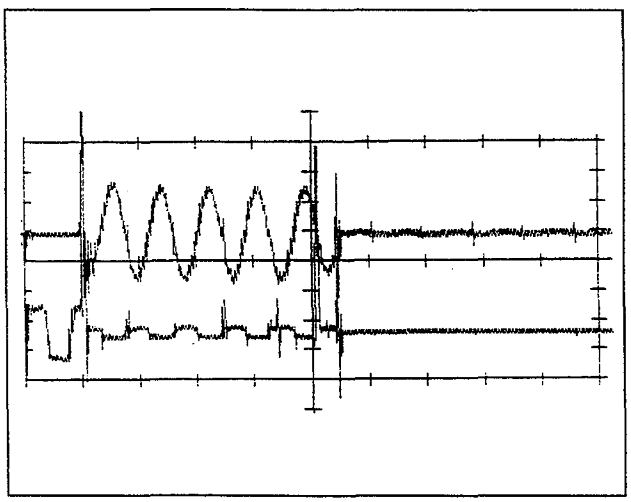

*Figure 4.*
*Gas Tube Failure at 640 Amps*

Gas tubes #7 through #11 were tested for ruggedness at reduced current levels, and although the final five gas tubes could not be tested to failure (due to time limitations), they provided statistical data for comparison with the SCR-based devices.

| | |
|---|---|
| Tube #7 | Survived 18 surges at 320 Amps<br>Failed opened during surge #19 at 320 Amps |
| Tube #8 | Survived 117 surges at 320 Amps<br>Failed opened during surge #118 at 320 Amps |
| Tube #9 | Survived 419 surges at 320 Amps<br>Failed opened during surge #420 at 420 Amps |
| Tube #10 | Survived 55 surges at 320 Amps<br>Failed opened during surge #56 at 320 Amps |
| Tube #11 | Survived 161 surges at 320 Amps<br>Failed opened during surge #162 at 320 Amps |
| Tubes #12-#16 | Functioning normally after 100 surges each at 255 Amps |

Based on the performance of gas tubes #8, #9, and #11, it is reasonable to estimate that one or more of devices #12-#16 could have survived over 1000 surges at the 255-Amp level, if time had allowed.

## TESTING SOLID STATE "CROWBAR" CIRCUITS

These first seven solid-state SCR-based circuits were tested to determine the maximum level that they would withstand. The 340-Amp tests were performed under the same conditions which provided 320 Amps in the gas tube tests, with the difference caused by the difference in terminal voltage (20 Volts for the gas tubes vs 2 Volts for the SCRs). NOTE: In three cases, the surge exceeds the manufacturer's specification: Itsm =

500 Amps (peak 1-cycle), and a table implies a maximum of over 300 Amps (peak) for 5 cycles.

SCR #1 Failed opened during the first surge at 600 Amps

SCR #2 Survived 1 surge at 340 Amps
Failed opened during a second surge at 400 Amps

SCR #3 Failed opened during the first surge at 400 Amps

SCR #4 Survived 6 surges at 340 Amps
Failed opened during a 7th surge at 340 Amps

SCR #5 Survived 3 surges at 340 Amps
Failed opened during a 4th surge at 340 Amps

SCR #6 Survived 2 surges at 340 Amps
Failed opened during a 3rd surge at 340 Amps

SCR #7 Survived 5 surges at 308 Amps
Failed shorted during a 6th surge at 308 Amps

Figure 5 shows SCR-based solid-state crowbar circuit #1 failing at the 600-Amp level, where the SCR increased resistance (note the simultaneous decrease in current and increase in voltage) and then blew up (opened) after only 1.5 cycles of the first 5-cycle surge. The odd-looking waveform in the remainder of the plot is characteristic of the ferroresonant transformer output following a high voltage un-clamped surge. (It settles back to normal after about 5 to 10 cycles.)

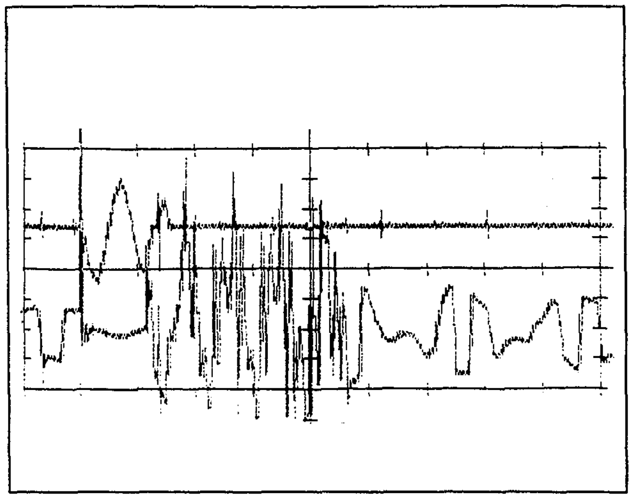

*Figure 5.*
*SCR Failure at 600 Amps*

The next six solid-state devices were tested at a reduced current level for ruggedness comparison to the gas tubes. The tests were performed under the same conditions which provided 275 Amps in the gas tube tests, with differences due to device characteristics as noted above.

SCR #8 Survived 24 surges at 288 Amps
Failed opened during surge #25 at 288 Amps

SCR #9 Survived 36 surges at 288 Amps
Failed shorted during surge #37 at 288 Amps

SCR #10 Survived 25 surges at 288 Amps
Failed shorted during surge #26 at 288 Amps

SCR #11 Survived 5 surges at 288 Amps
Failed opened during surge #6 at 288 Amps
(circuit PCB trace failed; SCR later tested OK)

SCR #12 Survived 29 surges at 288 Amps
Failed opened during surge #30 at 288 Amps

SCR #13 Survived 49 surges at 288 Amps
Failed shorted during surge #50 at 288 Amps

SCRs #9, #10, #12, and #13 all showed the "follow current" effect starting at about the 18th surge. Typically, this "follow current" lasted 3 and 6 cycles following the surge.

## SUMMARY & CONCLUSIONS

Field testing has shown that not only do long-duration current surges dominate the totals, they usually accompany short-duration impulses, and can exceed 100 mS duration. While many of the long-duration events are moderate (under 100 Amps), the extreme amplitudes which were recorded require serious consideration.

Given the amplitudes and durations of the most serious long-duration current surges, most of the devices which are traditionally used for surge protection are not appropriate for most CATV applications. Common devices such as MOVs and "Tranzorbs" are suitable only in secondary applications, where primary protection is provided by clamp-type ("crowbar") devices such as high-power thyristors and gas-filled surge arrestors.

Although high-power thyristor circuits have performed well in limited field and laboratory testing, their manufacturer's specifications and application notes cast considerable doubt on their use as surge protectors.

The long-duration surge testing illustrated the limitations of both gas tubes and thyristors, and proved that the modern heavy-duty gas tubes offer significantly superior performance overall.

## ACKNOWLEDGEMENTS

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# *EXPANDED BANDWIDTH REQUIREMENTS IN CATV APPLICATIONS*

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Converters have played an important role in the cable industry for years. As system bandwidth requirements have grown, so have the tuning capability in converters. During the 1980's, the introduction of "cable-ready" TV's and VCR's raised serious questions as to the benefits of converters. Unfortunately, however, most cable-ready tuners selected by consumer electronics companies are not acceptable in a direct cable environment. This paper presents an analysis of TV versus CATV tuners and the measureable benefit of converters in ensuring high quality, distortion minimized pictures are delivered to the viewing screen. It also touches upon the future challenges in tuner designs to achieve extended bandwidths beyond 550 MHz.

Bandwidth expansion in Cable Television has been a routine occurrence since the early days of cable. In the beginning of CATV, bandwidth requirements were limited to only the re-transmission of local programming in areas which could not access any off-air programming. Since these simple days, bandwidth has continued to double every 12 - 15 years due to the increased availability of diverse programming. Today, state-of-the-art system designs include 550 MHz or 750 MHz spaced plant with the plans to use the added capacity to offer new services such as Cable On Demand or multichannel Pay Per View. Time-Warner Cable has taken these designs even further with the 1 GHz plant in Queens, New York.

As bandwidth requirements have grown, so has the need to provide expanded bandwidth converters to allow subscribers access to these programs. Over the past several years, however, much has been written regarding the availability of "Cable-Ready" televisions and the related desire to remove the converter as a tuning device from the home. This argument has been based upon the perception that cable-ready TV's offer all the tuning capability required and converters simple duplicate this functionality in a less than consumer friendly fashion. As this paper will clearly demonstrate, however, even with only the current standard of 550 MHz bandwidth requirements, converters play a key role to ensure high quality, distortion minimized pictures are delivered to the viewing screen. Beyond 550 MHz, converters will play an even greater role as the tuner performance requirements exceed the capability of any TV tuner design. In moving beyond 550 MHz, a new technology is being investigated (GaAs) to provide these enhanced bandwidth capabilities.

Before analyzing the performance of TV tuners relative to CATV tuners, a brief review of the bandwidth expansion growth is in order. In the early 1960's, channel capacity on the cable plant exceeded the tuning capability of TV's for the first time. To add additional services, operators were required to deploy an 8 channel mid-band tuner. By 1967, the first 20 channel electro-mechanical cable converters were available. TV tuners were still limited to the VHF and UHF bands, so all homes receiving the expanded service needed to take a converter. 1973 brought about the need to expand the cable converter bandwidth to 300 MHz (36 channels) and by 1979 this had expanded to 400 MHz. During the 1980's, bandwidth continued to expand, first to 450 MHz and then to the current standard level of 550 MHz.

It was also during the 1980's, however, when TV manufacturers set out to solve the TV tuner limitation by adding "cable-ready" performance. Basically, this meant a TV was capable of tuning the entire cable band, first to 300 MHz, then to 450 MHz and over the past few years to 550 MHz and 806 MHz. What many TV manufacturers have failed to take into account, though, is the added tuner performance requirements necessary to survive in the cable environment without a cable converter. To fully understand this issue, it is necessary to review the technology employed in TV tuners as compared to that in CATV tuners.

## SINGLE CONVERSION TUNERS

TV tuners utilize an approach commonly known as single-conversion tuning (Figure 1). In this implementation, all signals are first fed through an input tracking filter, and then are received by a mixer and the proper channel selected. A local oscillator (L.O.) is used to select the proper frequency under the control of the tuning system. The resultant single channel intermediate frequency (I.F.) signal is then made available to the TV demodulator for conversion to a baseband signal for viewing by the consumer. This approach to tuning is very cost effective and thus the preferred approach in the very cost competitive consumer market.

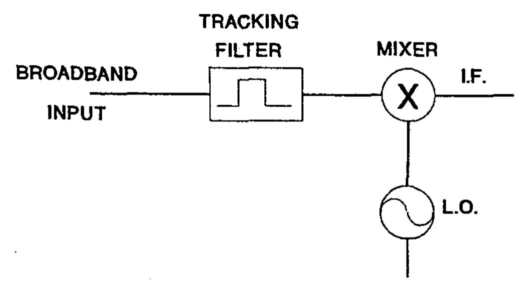


FIGURE 1: SINGLE CONVERSION TUNER

Single-conversion TV tuners were designed to operate in a broadcast environment where the tuning dynamic range is large, but the actual number of channels being tuned is limited and adjacent channel reception is seldom found. In this scenario, the power loading on the mixer due to the input channels is somewhat limited, thereby not overloading its capability which would result in distortion. When these same tuners are utilized in a cable environment, the channel loading is much greater and thus the power loading on the mixer is greater. The result of such loading is a reduction in the overall distortion performance of the tuners above a given bandwidth. In addition, TV's used in the broadcast environment are generally not coupled to one another within a home and therefore are not susceptible to L.O. leakage from another TV's tuner. Even when multiple sets are connected to the same antenna, leakage from the L.O. does not fall into the frequency of another channel. When these devices are linked in the cable system, however, L.O. leakage will inevitably fall within another channel and thus result in distortion.

To determine the actual performance of TV tuners in various channel loading scenarios, 21 TV's and VCR's ranging in date of manufacture from 1981 to 1990 were obtained. These devices were selected at random from the variety of equipment available within Jerrold. The bandwidth on the TV's and VCR's ranged from 300 MHz to 806 MHz, representing a cross section of all tuners manufactured over the past ten years. Measurements were made on the various devices on four basic criteria:
(1) L.O. leakage out the input connector,
(2) composite second order beats (CSO),
(3) cross modulation (adjacent channel),
and (4) image rejection.

L.O. leakage is a measurement worthy of further discussion in the cable market. The basic design of tuners uses an L.O. mixed with the input signal to select the proper channel. In a single conversion tuner, the L.O. frequency is set 45.75 MHz above the tuned channel. With an unbalanced mixer and high input power, it is difficult to prevent the L.O. frequency from leaking out in both directions from the mixer. The result is signal leaking out the input connector and into the input signal of a connected TV of VCR (Figure 2). In a single TV household, at least 25 dB of isolation is provided by the tap. Between sets in the same household, however, commercially available splitters provide less than 20 dB of isolation. Because a minimum of 60 dB of isolation is required to eliminate interference, the frequency of the leakage from the TV tuner L.O. becomes critical.

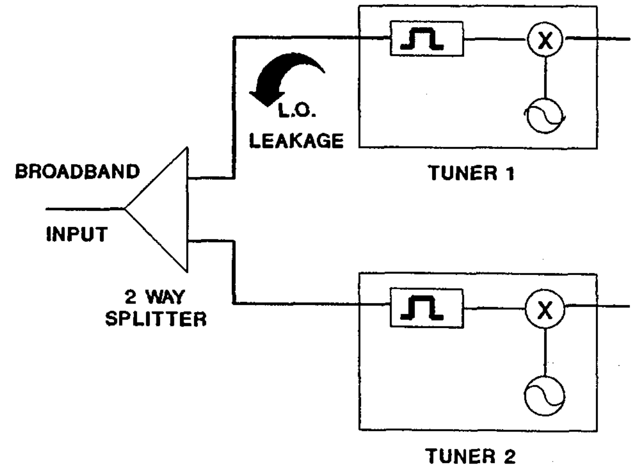


FIGURE 2: L.O. LEAKAGE SITUATION

Image rejection is also more critical in a cable environment than in the broadcast environment. The image frequency is 90 MHz above the channel being tuned. While TV and VCR tuners do employ filtering techniques, 60 dB of rejection is required in a cable environment. Because an image is seldom present in the broadcast environment, TV and VCR tuners are only designed for a maximum of 45 dB of rejection. Therefore, image rejection will cause problems with TV's and VCR's in a direct cable environment.

The summary results of the tests are shown in Tables 1 to 3. Table 1 reflects the number of tuning devices which passed the test in a 300 MHz loaded system, while table 2 assumes 450 MHz loading and table 3 assumes 550 MHz loading. Throughout the testing, live video feeds were used on all channels to simulate a real headend. The criteria used to determine acceptability was visible distortion which from years of experience would result in the following distortion numbers: (1) L.O. leakage > 40 dB with a 0 dBmV input level, (2) CSO >53 dB, (3) cross mod >57 dB, and (4) image rejection < 0 dB.

As can be seen from the results, only 4 of 21 tuners were able to perform within acceptable levels in all four categories with 300 MHz loading, with the number of acceptable sets dropping to 1 at 450 MHz and 550 MHz. While many people have proposed the use of cable-ready TV's along with some form of broadband technology to secure basic and/or premium services in cable systems in the future, these results clearly demonstrate the fallacy of such beliefs, even in 300 MHz analog systems, due to performance considerations. As system channel expansion grows, the use of broadband security approaches with "cable-ready" TV's and VCR's will result in the deterioration of performance. For reference purposes, table 4 provides the actual measured performance of the best and worst device tested in the study at 550MHz loading. The best TV was a 806 MHz model manufactured in the late 1980's which is only indicative of tuners sold into the market over the past few years. As the lifespan of TV's far exceeds ten years, the "cable-ready" sets which cannot truly survive in this environment without degrading picture quality will dominate the market for at least another ten years.

While no testing was performed with digital signals, the impact of L.O. leakage upon QAM modulated signals in sets connected through splitters becomes even more critical. Unlike analog signals, digital signals interfered with by L.O. leakage are not forgiving. Therefore, if one TV in the household is connected to the broadband signal in a cable-ready capacity and a second set is connected via a splitter and a QAM demodulator, the digital signals are likely to be overloaded by the L.O. leakage of the cable-ready TV. The final result may be either the loss of digital signal(s) or a very long time acquiring the channels due to the limitation of the adaptive equalizer. The implication of such a result is conclusive: direct connections to a cable-ready set for analog signals along with the connection to a digital demodulator for another TV or VCR will create problems. Therefore, the need for cable converters to maintain picture quality, in either analog or mixed analog/digital applications, will continue to grow.

## DUAL CONVERSION TUNERS

Dual conversion tuners were available in the satellite industry long before their application in cable television became known. The first recorded use within cable was the introduction of a 260 MHz tuner in the early 1970's. In a dual conversion tuner, broadband signal are first converted to a high IF by mixing the input with the first L.O. and then downconverting it to the proper L.O. by mixing with the second L.O. (Figure 3). The first L.O. frequency is selected such that the leakage frequency falls outside the input video spectrum and thus eliminates the L.O. leakage concern. The first IF is then filtered before being subjected to the second L.O. and converted to either the proper TV frequency in a heterodyne converter or to the proper IF frequency for demodulation in baseband converters.

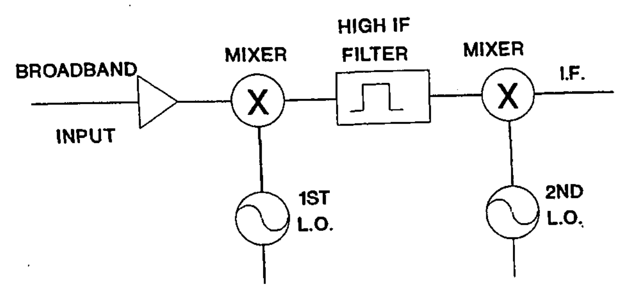


FIGURE 3: DUAL CONVERSION TUNER

| TUNER B.W. | TOTAL TESTED | L.O. LEAKAGE | CSO | X-MOD | IMAGE | OVERALL |
|---|---|---|---|---|---|---|
| 300 MHZ | 4 | 0 | 4 | 1 | 3 | 0 |
| 400 MHZ | 3 | 3 | 3 | 1 | 2 | 0 |
| 450 MHZ | 2 | 0 | 2 | 0 | 2 | 0 |
| 806 MHZ | 10 | 4 | 10 | 8 | 8 | 4 |
| SP-650 | 2 | 0 | 2 | 2 | 0 | 0 |

TABLE 1: 300 MHZ LOADING

| TUNER B.W. | TOTAL TESTED | L.O. LEAKAGE | CSO | X-MOD | IMAGE | OVERALL |
|---|---|---|---|---|---|---|
| 450MHZ | 2 | 0 | 1 | 0 | 2 | 0 |
| 806MHZ | 10 | 4 | 8 | 1 | 5 | 1 |
| SP-650 | 2 | 0 | 2 | 0 | 1 | 0 |

TABLE 2: 450 MHZ LOADING

| TUNER B.W. | TOTAL TESTED | L.O. LEAKAGE | CSO | X-MOD | IMAGE | OVERALL |
|---|---|---|---|---|---|---|
| 806MHZ | 10 | 1 | 5 | 1 | 4 | 1 |
| SP-650 | 2 | 0 | 2 | 0 | 1 | 0 |

TABLE 3: 550MHZ LOADING

| | L.O. LEAKAGE | X-MOD | IMAGE REJECTION |
|---|---|---|---|
| BEST DEVICE | -2dBmV @129MHZ | -7dB CH 86 | -24dB CH 70 |
| WORST DEVICE | -26dBmV @485MHZ | +10dB CH 16 | >+15dB CH 16 |

TABLE 4: PERFORMANCE AT 550MHZ

To determine the performance of cable tuners in various power loading situations, published specifications for converters from each of the major converter manufacturers were compared. The results of this comparison clearly indicate the superiority of the dual tuner approach in all situations, including 550 MHz loading. Therefore, dual conversion tuners must be utilized to provide the best picture quality possible in the cable environment.

## EXPANDED BANDWIDTH CHALLENGES

As the cable industry moves beyond 550 MHz bandwidth, whether it be 750 MHz, 860 MHz, or 1 GHz analog and/or digital signals, the complexity of the dual conversion tuner design will grow. Due to the need to up-convert to such high IF's (up to 2 GHz in a 1 GHz tuner), the basic discrete design utilized to date presents several problems. One solution to this problem is to utilize a tracking filter on the front end of the tuner to limit the power to the first mixer. The downsides to such an approach, however, are the additional cost involved, the precision of the required alignments, and a decrease in noise performance. Another critical concern is the applicability of such an approach to digital signals due to phase noise considerations. With such overwhelming concerns, the need to discover a better approach is clear.

The most promising approach to solve the expanded bandwidth challenges involves the use of compound semiconductors. Silicon devices have been utilized in high precision microwave applications for a number of years. GaAs MESFET devices have demonstrated their superiority in low noise, wideband capability, and high frequency amplification. New generations of devices based on band gap engineering with the use of heterojunction structures have been explored. With improvements in material processing, lithographic techniques, and handling capability, yields have increased significantly with four inch wafers, reducing costs to the point that GaAs is extremely competitive with silicon. As a result, GaAs technology has become available in the high volume, low cost commercial applications such as cellular communications and automotive radar systems.

GaAs MMIC technology will shortly become available and cost effective for such applications as cable tuners as well. When used in this application, GaAs technology allows for the integration of low noise, low distortion amplifiers, low distortion mixers, wide dynamic range voltage controlled oscillators (VCO), and the attenuators required for AGC. Integration of these IC's in tuner designs requires extremely close coordination between the IC design house and the tuner manufacturer due to the total system implications of the IC performance. The long-term benefits of such an approach are high quality and consistency in manufacturing due to the component reduction and the elimination of several manual alignments.

The initial implementations of GaAs IC's at 550 MHz have recently been completed with performance equalling and in some cases exceeding discrete tuner designs. Tuners utilizing such technology are currently incorporated in one cable vendors product. Development is also far along on the use of GaAs IC's in 1 GHz applications. The results of such products will not be known for several months, although initial testing has proven quite successful. Long-term, the use of IC tuners in extended bandwidth applications will enable cable to achieve significantly enhanced performance relative to existing TV tuners, independent of the bandwidth or the composition of analog and digital signals.

## CONCLUSIONS

"Cable-Ready" TV's and VCR's are truly an oxymoron. In today's cable environment, these devices are not capable of handling the system performance requirements. The use of an alternate access control mechanism to enable broadband signals to enter the home will ultimately result in a picture quality which is inferior to that which is provided through cable converters and thus will be judged unacceptable by most subscribers. This phenomena will be exacerbated over the next several years as bandwidth expansion takes most systems to at least 450 MHz of analog channels. With the introduction of digital signals on the plant and the potential expansion to 750 MHz/ 1GHz, the need for high quality extended bandwidth tuners is critical. TV and VCR manufacturers have recently begun to utilize dual conversion tuners in their more expensive products as the result of strong encouragement from the cable industry. Unfortunately, the window of opportunity has all but passed as cable bandwidth expansion will occur significantly faster than the replacement TV/VCR market. Therefore, converters will continue to play an important and beneficial role in the consumer's home for many years to come.

# FAST DIGITAL DATA CHANNELS ENABLE NEW CABLE SERVICES

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*Abstract*

*With the advent of digital delivery systems in CATV, fast data channels of multiple megabytes per second are feasible. Such channels open new possibilities for multimedia interactive applications. They can be used to provide an enhanced TV guide service that includes still pictures or even video clips. They can provide an electronic advertising magazine with pictures. Under appropriate conditions, they can lead to a variety of CD-I like applications. This paper outlines some of the new cable services made possible with fast, one way data channels, and explains the issues to be addressed in developing those services.*

*Keywords: television, cable, digital, teletext, multimedia, database, CD-I, TV guide.*

## INTRODUCTION

For many years broadcasters have included digital data in the video signals they send to TV receivers. Examples are teletext [1,2] and closed caption systems [3]. We consider such data channels "slow", as they carry less than 2 Kbytes/second. For closed captions, the rate is only 60 bytes per second. If in place of video an entire 6 MHz TV channel is used only for data, much higher rates are possible. For example, full channel teletext could carry roughly 0.5 Mbytes/second.

The advent of digital delivery systems in cable TV (CATV) provides the opportunity for even faster data channels of multiple megabytes per second. In the Philips Digital Video system [4], each 6 MHz TV channel can carry roughly 3.5 Mbytes/second. Such fast data channels open new possibilities for multimedia interactive applications.

In the following section we explain the concept of locally interactive applications, based on the repeated, one way transmission of complete databases. This is the primary technique for exploiting fast data channels. We then describe three applications that use this technique: an electronic TV guide, an electronic advertising magazine, and a system help and tutorial facility.

Following these examples we explain how a wide variety of applications, including currently unforeseen applications, can be supported through the transmission and loading of an application's code, along with its database. We then describe a further extension, the use of local storage such as a digital tape or optical disc, to capture all or selected portions of a database for later play back. Finally, we discuss some of the database design issues, including the overall structure of the database and the format of its constituent data items.

## FAST DATA CHANNELS

In this paper we assume a one way transmission channel, as supported by most current CATV systems. With a two way interactive channel, many applications would be handled differently. We do not expect two way fast data channels to be widely available at low cost in the near future. However, the addition of a separate, slow back channel to a one way fast data channel is an extension that will be considered in our future work.

We also assume a single, 6 MHz TV channel is being used. In the future, multiple adjacent channels could be combined to provide even faster digital data channels.

With a one way data channel, it is not possible to send a query to a remote database to obtain a specified item of interest. Instead, the entire database can be transmitted from the remote location and scanned sequentially by a receiver to find and extract the required information. This is the basic technique used in teletext systems.

With a fast data channel of 3.5 Mbytes/second, a database of several megabytes can be transmitted within a few seconds. If the database is transmitted repeatedly, a query about the database can be answered within the time it takes to transmit the database again, in this example within a few seconds.

The underlying query handling system within the receiver is conceptually quite simple. It merely waits until the required data are transmitted again and picks them up at that time. In this way an application can interact with a user and provide responses to the user's input, based on the contents of the database. In practice, the high data rate of the transmission channel necessitates the inclusion of special hardware in the receiver, to scan the data stream for the items of interest and extract them for use in the application.

The scanning hardware can be simple, such as comparing a fixed bit field within each block or packet of the database against a given pattern. The scanning hardware could also be quite complex, such as matching each data item against a Boolean combination of several variable length and variable position fields, or against a complex relational database query [5].

The structure and format of the database, and the complexity of the query handling software within the receiver, are strongly affected by the type of scanning hardware used. In this paper we assume simple hardware, like the fixed bit field matcher described above. Some of the associated database design issues are discussed later, in the section on Database Format.

For databases larger than a few megabytes, the query response time can become unacceptably long, because of how long it takes to transmit the database each time. Depending on the application, this problem can be alleviated to some extent by one of two techniques. First, copies of the most frequently accessed data items can be repeated at regular intervals throughout the database. Second, copies of the most likely items to be accessed next can be cached by the receiver, usually in RAM. Both of these techniques are discussed in more detail in the section on Database Format.

To give a more concrete impression of approximately how much data can be carried in a given time on a fast data channel of 3.5 Mbytes/second, the following list provides a number of examples. These examples assume that state of the art digital compression techniques are employed:

- 5000 full TV screen text pages in 1 second.
- 300 fax pages in 1 second.
- 1200 two page letters in 1 second.
- 15 novels in 1 second.
- 25 one minute voice messages in 1 second.
- 15 high quality video stills in 1 second.
- 300 small color still pictures in 1 second.
- A daily newspaper with black and white pictures in 1 second.
- A weekly newsmagazine with color pictures in 3 seconds.
- The contents of four 800K floppy disks in 1 second.
- One full CD-ROM or CD-I disk (600 Mbytes) in 3 minutes.

## EXAMPLE APPLICATIONS

We now briefly describe three example applications that are based on the repeated, one way transmission of multiple megabyte databases. First we consider an enhanced electronic TV guide service that demonstrates some of the features made possible by fast data channels. There already exist some electronic TV guides. Two examples are SuperGuide [6], built into the UNIDEN UST4800 satellite decoder, and InSight [7], soon to be broadcast via the PBS network. However, because these systems work with relatively low data rates, they only present the user with screens and menus consisting of text.

By exploiting a fast data channel, an electronic TV guide can be augmented with still pictures, audio clips, and even video clips. These can be used to attract consumers to the shows, for instance to advertise pay-per-view offerings. A still picture can be used as an identifier for a show, both to make it more attractive and to make it more immediately recognizable. This is particularly valuable for children who cannot comprehend text. Some shows, such as movies, can have their identifying pictures replaced by video clips. This would be like the clips shown on current "preview" channels, or like the "trailers" shown in movie theaters, and hence be much more effective for attracting viewers.

The size of the text portion of a TV guide database, containing schedule information and brief summaries of all the shows over a one week period, is roughly 1 Mbyte. There are approximately 1000 different shows in one week, each of which could have a separate identifying still picture. Each of those pictures requires roughly 5 Kbytes of storage. Thus, the total size of the TV guide database, including still pictures, is roughly 6 Mbytes. It can be transferred on a fast data channel in under two seconds.

Video clips require much larger amounts of information to be transferred. A fast data channel can simultaneously carry over 15 VHS quality, full screen movies, or over 100 small, picture-in-picture (PIP) size movies. By reserving half the bandwidth of the data channel for video clips, over 50 continuously repeating, PIP size clips could be provided. The text and still pictures part of the TV guide database could then be transferred in about four seconds.

Four seconds is too long for a viewer to wait for information about a TV show, so parts of the TV guide database would have to be cached in the receiver. Only information about the shows that are currently playing or coming up in the near future need be cached, since that is what the viewer will most often be interested in. As the viewer looks further into the future, it will take him/her a few seconds to study each screen of information. During that time the data for successively later shows can be cached, so that the information will be immediately available when needed.

Because of the quantities of data involved, video clips are never cached. Instead, they are played in real time, beginning at whatever point they are currently showing. Since each clip is typically quite short (30 to 60 seconds) and repeated continuously, starting in the midst of a clip should be acceptable to most viewers.

Our second example is an advertising magazine that uses a database of classified ads and optional associated pictures. A mechanism to search through the magazine by category or by keyword would make this more powerful than its printed equivalent. Color pictures, audio, and video could be exploited by advertisers to attract and inform a highly targeted audience of viewers. The viewers of an ad are those who, through their interactions and selections, have already indicated an interest in the product or service.

The size of the advertising magazine database, excluding video clips, would be comparable to a weekly newsmagazine with color pictures. Thus, it would take about 3 seconds to transmit on a fast data channel. Data caching techniques similar to those discussed above for the TV guide can be used to ensure that the system response times are acceptable.

A third example application is a system help and tutorial facility that can explain to users how to operate their receivers, and introduce them to the various TV, audio and data services available. By pressing a HELP button on the remote at any time, users can obtain context sensitive help information. Procedures can be explained and demonstrated through tutorial sequences. Pictures, audio and video can significantly enhance the learning experience.

The text part of the help database, along with simple illustrations, would probably be less than 1 Mbyte. With more elaborate pictures for demonstration sequences, the size might grow to around 2 Mbytes. This can still be transmitted in under one second. Hence it may be possible to provide the help facility without any caching of the database.

## LOADING APPLICATION CODE

The flexibility of the data receiving device to support a wide variety of applications, including currently unforeseen applications, can be supported through a mechanism for loading application code. One obvious approach is to transmit an application's code along with its database. The receiver would first load the application code and begin executing it. The application code would then interact with the user and access information from the database as required.

If the data processing computer inside the receiver conforms to an open standard, third party software developers and service providers will be encouraged to produce applications for use with the system. A wide variety of applications would then be available. For example, one could imagine a "Video Game of the Week" data service channel that transmitted the code and data of multimedia video games, with the selection of games changing on a weekly basis.

For the standard data processing computer, we advocate a Compact Disc Interactive (CD-I) compatible platform [8]. CD-I was designed especially with consumers and TVs in mind. It has been adopted by a number of consumer electronics manufacturers, and an increasing selection of products and applications based on the standard are now becoming available. Although CD-I was originally developed with compact disc technology in mind, much of the standard and many of the existing software development tools can be readily adapted to the fast data channel environment.

## LOCAL STORAGE

A further extension of the data receiving device, to increase its range of capabilities, is the addition of local mass storage facilities. We discussed earlier the use of RAM for caching selected data items from the transmitted database. With local mass storage, such as an optical disc, large portions or the entire contents of a database could be captured. For example, the complete contents of a compact disc would normally take about three minutes to transmit, which is too long to wait for the retrieval of a selected data item. However, if the transmitted data is first stored onto a local disc, it could be "replayed" later with faster interaction.

Local mass storage opens additional application possibilities. For example, music libraries or video game libraries could be made available on data service channels. Subscribers could then create personal collections of their favorite music or games by capturing selected items from the transmitted libraries and storing them on digital tapes or optical discs. The various copyright issues to be addressed by such applications, while very important, will not be discussed in this paper.

Another way to exploit local mass storage is to combine a local database with a transmitted database. For example, the contents of an encyclopedia could be stored on a CD-I disc. The latest updates to the encyclopedia could be transmitted on a fast data channel. By combining the two databases, the user would have interactive access to the latest information in a very large database. Updates to the CD-I disc could then be made at relatively infrequent intervals.

## DATABASE FORMAT

Having looked at some of the potential applications of fast digital data channels, we can now summarize some of the associated database design issues. First it should be noted that although the above multimedia databases are transmitted over a TV channel, the format of their pictures, video and sound need not be identical with the compressed video and audio formats of a digital cable TV system. Sometimes the TV formats are the most convenient to use, such as for the encoding of video clips. But in other cases different formats might be preferred.

The type of data processing computer used, in our case a CD-I compatible platform, has a strong influence on the format of data items. To maintain CD-I compatibility, most items, such as still pictures, graphical objects, and even application code, will follow CD-I conventions. However, some database providers may prefer to supply their information in other formats. This can be accommodated by converting the data into a CD-I compatible form, either before transmission or after it is captured in the receiver. The needs of the database scanning hardware and software could also force additional restructuring of the data items to facilitate retrieval.

One major concern when designing the overall structure of the database is to minimize the number of times the user might have to wait for long periods to receive requested information. Two main techniques are available for dealing with this problem: repetition and caching. Repetition involves copying the most frequently accessed data items and transmitting them more frequently, thereby reducing the expected latency. The down side is that the overall size of the database increases, and thus the time to receive the less frequently accessed data items increases.

The second technique is caching. This involves storing in the receiver the most likely information to be requested next. The main problem is determining what information should be cached. This also has to be traded off against the amount of memory, usually RAM, to be used for the cache. Caching can be especially effective for hypertext structured applications where the user is presented with screens of information that are linked together in a predetermined fashion. While the user is viewing one screen, all of the screens directly linked to it can be retrieved. One of those screens will be what the user wants to see next.

For accessing the database information in different ways, index files can be transmitted along with the database and cached in the receiver. For example, the TV guide database might have one index file containing pointers to all of the shows for each channel, and a second index file with pointers to all of the shows for each time slot. By carefully designing the index files according to the application's needs, it should be possible to answer most queries in a single scan of the database, looking for an item identified by a particular pointer. Simple scanning hardware that matches a fixed bit field in each block or packet of the database against a given pointer/identifier is all that is needed.

## CONCLUSION

In this paper we have explained how the fast digital data channels, made possible by the advent of digital delivery systems in CATV, can enable new cable services. We described a variety of locally interactive, multimedia applications, based on the repeated, one way transmission of complete databases. We also discussed some of the issues involved in the design of those applications.

Work is currently proceeding on the development of an experimental system that can support the types of applications discussed above. For demonstration purposes, an experimental, enhanced, electronic TV guide is being developed as an initial application. Studies by ourselves and others indicate that users like the concept of an electronic TV guide, and are even willing to pay for it. We believe that the TV guide may be the driving application that finally leads to the broad acceptance of interactive services by American consumers.

Throughout the paper, a single 6 MHz TV channel was assumed as the basis for a fast digital data channel. However, nothing prevents multiple adjacent channels from being combined into a single, even faster data channel in the future. This will enable even more elaborate interactive, multimedia applications to be developed.

## ACKNOWLEDGMENTS

The ideas and work presented in this paper are the result of comments and contributions from many people. We especially want to thank Paul Rutter, Jhumkee Iyengar, Bill Lord, Brian Johnson, and everyone associated with the Briarcliff TV Data Services project.

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# FIBER OPTIC RETURN TRANSMITTERS FOR CATV

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*Abstract*

*This paper discusses the use of fiber optic transmitters utilizing Fabret-Perot (FP) lasers for return links in CATV distribution networks. The dynamic range achievable with a link driven by this type of transmitter is dependant on many factors which limit the achievable performance. Knowledge of their effects is important for optimizing the return link. Performance of prototypes operating over a 22 kilometer (7.5 dB) fiber link are reviewed and some of the design details for achieving the performance are explained.*

## I. INTRODUCTION

Fiber optic return links are becoming increasingly important in CATV applications and the use of the inexpensive FP laser as a directly modulated optical source is attractive since the return band channel loading is low. A return cable path often has only a 5 to 50 MHz passband which can accommodate up to 7 NTSC channels. Usually, only a few channels are required and a low cost FP transmitter without an isolator can be used effectively to transmit the video carriers through a link with excellent carrier-to-noise (C/N) and distortion performance.

FP lasers with external optical isolator have been demonstrated for 42 channel forward band CATV transmission over a short (5dB) link [1]; however, for a return fiber link the use of an isolator is an expensive addition and optimization of other system parameters precludes its use.

Return channels sent back to the headend are required to be of high quality as they are to be transmitted back out through the forward cable system. CNRs in the high fifties are generally required for this type of application.

A prototype link capable of delivering 2 channels through a 7.5 dB link with Carrier-to-Noise ratios (CNR) of greater than 59 dB with no distortion products greater than -75 dBc has been demonstrated. This same link can deliver 4 channels with 55 dB CNR and -65 dBc worst case CSO and -70 dBc CTB. When the channel loading is increased to 7, 52 dB CNR with -62 dBc CSO and -67 dBc CTB is obtained and this performance is only slightly degraded as the transmitter is temperature cycled up to +60 celsius.

In order to achieve this performance, a laser with excellent linearity and fiber coupling must be selected. Also important is that the link it is driving must have minimal reflection back into the laser. Correspondingly, the receiving photodetector in the return receiver, as in the forward optical receivers, must have a high optical return loss even though the channel loading is low. In the return link the FP laser, with its characteristic multiple longitudinal mode optical spectrum, is extremely sensitive to reflection.

## II. PROTOTYPE LINK PERFORMANCE

Figure 1 shows a block diagram of a prototype return transmitter using an FP laser with no isolator.

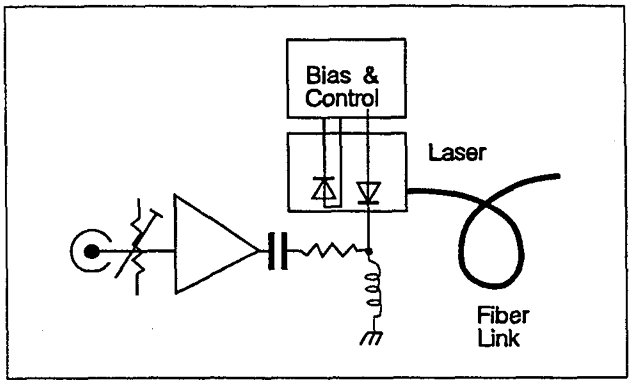


*Figure 1.*
*Transmitter Block Diagram*

The transmitter consists of a Motorola Return Band hybrid CATV amplifier driving a Philips CQF-57D FP laser. Resistive matching is used between the amplifier and the laser, though transformer matching has distinct advantages for return band transmitters. The use of transformer matching will be discussed later in this paper.

Optical power leveling is achieved by means of a differential amplifier which adjusts the bias to the laser to maintain the voltage across a resistor loading the monitor detector at a temperature compensated reference level. Figure 2 shows the 7 channel spectrum viewed at the receiver separated from the return transmitter by 7.5 dB loss of fiber.

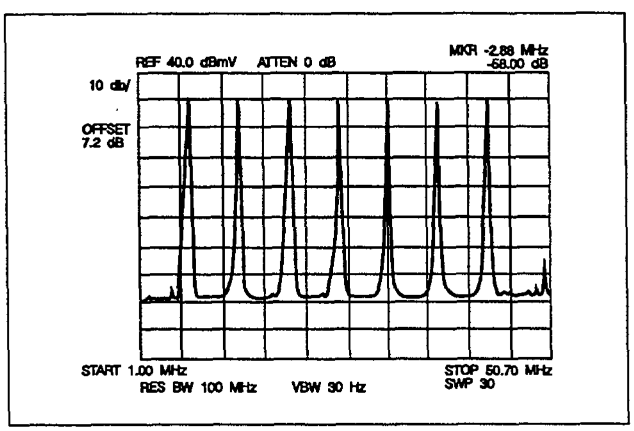


*Figure 2.*
*Seven Channel Spectrum*

The distortion products (second order distortion predominates) visible in this plot are both link and spectrum analyzer generated and calculation of their actual levels or use of a preselector filter reveals their true levels to be -62 dBc. Third order distortion must be characterized using a filter and worst case third order distortion (CTB) was measured to be less than -69 dBc. The return transmitter was fusion-spliced to the link for these measurements.

## II. OPTICAL REFLECTION

With a fiber connector spliced into the link at the transmitter, there was a slope in the noise floor across the frequency band. This slope, when observed over a much larger frequency range, was found to be a periodic peaking in the noise floor of the link over its modulation bandwidth. This effect is caused by optical reflection from the connector and is similar to having a secondary resonant cavity external to the FP resonator. The frequency spacing of the noise peaks is equal to the reciprocal of the round trip travel time of light in the fiber from the laser to the connector interface. In order to improve the noise floor of the link so that it is more flat, one could use an optical isolator, a laser with a narrower optical linewidth, or use a lower reflection fiber link terminated in a better optically matched photodetector. The first two approaches add significant cost to the return fiber link; however, the third is potentially low cost with significant performance enhancement when properly implemented. Using this approach alone, increased channel loading can be accommodated.

## III. CNR CALCULATION

The calculation of CNR for a directly modulated laser diode-driven link has been treated extensively in the literature and is covered in most standard text books on fiber optics so the details will not be reviewed in this paper. The FP laser link performance is effected significantly by optical reflection in the link so an analysis including only laser relative intensity noise (RIN) and receiver thermal and shot noise will predict CNR performance higher than what is observed if an optical reflection such as with a connector is present. For an FP link with no isolator, a more refined analysis is required. It is useful, however, to consider the achievable CNR without considering reflection, as if an ideal isolator were used at the output of the laser. The received CNR vs. link budget for an FP driven link with modulation index typical of 4 channel transmission is shown in Figure 3. It is assumed that; the transmit power of the laser is 2 mW (+3 dBm), the RIN of the FP laser is -135 dB/Hz, the receiving photodetector works into a 300 ohm load (achieved through a 4:1 impedance transformer from the 75 ohm receiving amplifier input) and has a dark current of less than 1 nA, and that the noise figure of the receiving amplifier is 4 dB.

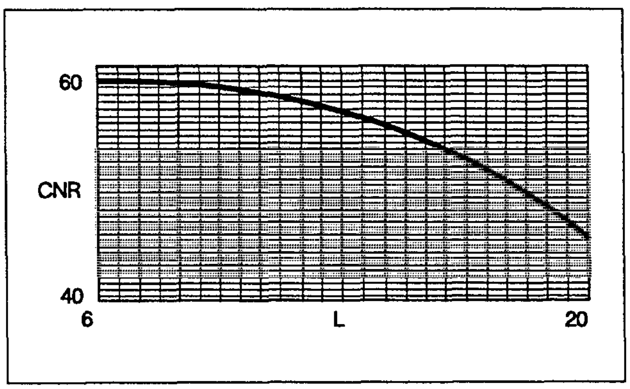


*Figure 3.*
*CNR vs. Link Budget*

This analysis is actually conservative for a return link. With low channel loading and narrow bandwidth, many of the design constraints of a receiver intended for forward band operation are lessened. Special purpose low noise receivers [2] with increased photodetector load impedance and low noise GaAs FET amplification can be applied and have excellent CSO performance, often a problem in return band links.

In order to more accurately predict the received CNR of an FP laser driven link, the effect of optical reflections must be modeled. Unfortunately, this is not particularly simple for an FP laser because the optical spectrum consists of a grouping of closely spaced longitudinal modes with wavelengths that are a half integer sub multiple of the FP laser cavity length and there is a combined effect from these modes (Figure 4). If there is dispersion in the fiber, the calculation of this effect is more than a simple extension the analysis of reflection in a DFB driven link.

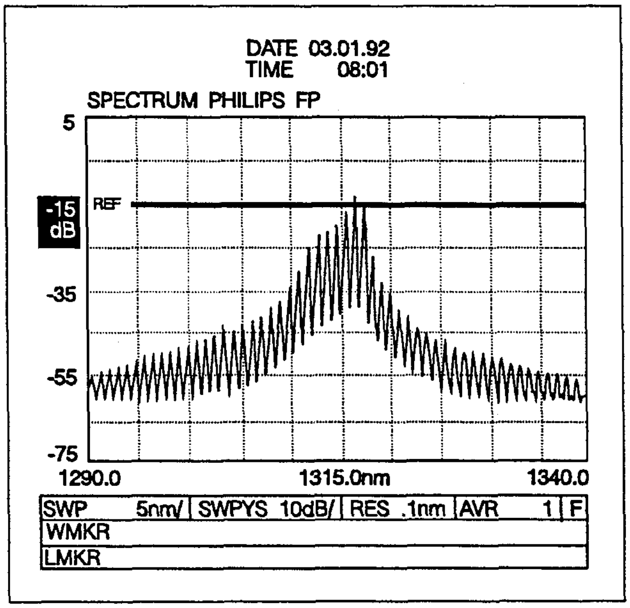


*Figure 4.*
*FP LASER SPECTRUM*

## IV. TEMPERATURE EFFECTS

The optical spectrum of the FP laser is effected significantly by variations in temperature. Not only does the center wavelength of the peak in the mode distribution vary, but the mode distribution can vary significantly depending on whether most of the optical power is in a single, central mode, or shared by several adjacent modes. Often, an optimal operating temperature can be found at which point one mode is dominant and the effective linewidth for the particular laser is minimized. Under these conditions, the CNR and reflection immunity will be improved.

The use of a cooler is not always practical as they consume large ammounts of power (several watts is not unusual if a linear power supply is used). Not using a cooler can limit channel loading and CNR performance as well as the operating temperature range of the transmitter, so the choice of using a cooler or not is an important one.

## V. RF DESIGN OF FIBER OPTIC RETURN BAND TRANSMITTERS

The RF circuits associated with the design of a return fiber optic transmitter include the input attenuator and the driver amplifier. Additionally, directional couplers for monitoring the laser drive or for combining auxiliary inputs are sometimes desired.

The primary requirement for the driver amplifier is that the output compression levels must be well above the input levels that distort the laser so that no significant amplifier distortion is produced and the dynamic range of the link is not limited by RF amplification. Almost any return band CATV hybrid amplifier is capable of driving the laser sufficiently, though improved efficiency through use of a discrete amplifier is favorable for the sparse channel loading required of a return system.

Transformer matching from the output of the amplifier down to the low impedance (5 ohms) of the laser diode is advantageous for narrow band applications because it reduces the required drive level to the laser. For an ideal transformer, the reduction in required drive level to attain a desired optical modulation depth is 6 dB, though for practical transformers with core loss, 4 or 5 dB can be realized. This lessens the output drive level required from the amplifier, thus relaxing the required output compression levels and gain. Less overall power consumption is required by the transmitter if this is properly implemented.

## V. CONCLUSION

FP lasers can be used effectively without an isolator or cooler for return band CATV fiber links provided the channel loading is low and the link provides a low back reflection optical match to the laser.

Use of coolers and/or isolators can significantly enhance system performance of FP driven laser links; however, the cost and power requirements of the improved performance links are large.

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# Hidden Influences on Drop Reliability: Effects of Low Level Currents on F-Interface Corrosion and Performance

*Brian Bauer*
*Raychem Corporation*

## ABSTRACT

The CATV drop, the source of a high percentage of trouble calls, has a lifetime greatly dependent upon corrosion at the F-Interface. This corrosion can be influenced in many ways, including the presence of currents on the drop.

Many emerging technologies, such as those required for signal security and on-premise amplification, will have power requirement schemes which draw from the subscriber premise or mainline via the drop. In addition, many drops currently carry stray sheath currents originating from stray effects of power on the feeder. As a result, the drop can see many waveforms of primarily low level AC and DC currents.

As part of a CableLabs funded project, this paper explores the nature of existing and future current loads on the drop and how these different forms effect corrosion of the F-interface. Practices are suggested which may improve reliability and performance based on findings.

## INTRODUCTION

This is the second phase of a three phase CableLabs project, the three phases being 1) Basic corrosion in the drop and the effects of CASS, Copper Acetic Acid Salt Spray[6] 2) the effects of low currents, simulating ground currents and 3) the comparison of CASS conditioned samples to actual field aged samples.

In this phase, we first look at the basic driving forces of corrosion to understand why externally induced currents can have an effect on corrosion. Then the types of currents that may be seen on the drop are examined as well as engineering practices that may influence these. Finally, results of CASS and low current conditioning of f-interface components are summarized with recommendations.

### Background, Electrochemistry

At a typical corrosion sight, an electrochemical reaction occurs which can be modeled with a simple voltaic cell. Three conditions must be present for wet corrosion, the most common type, to occur, **Figure 1**.

1) Two dissimilar metal surfaces must exist at the site
2) An electrolyte (eg, liquid solution) must be present and in contact with the two surfaces
3) The surfaces must be in electrical contact with each other.

Corrosion occurs when ions in the liquid solution chemically react with one of the metal surfaces, known as the cathode, causing positively charged metal ions to enter the solution, leaving this metal with a negative charge. This is known as the 'oxidation' reaction. The other metal, the anode, develops a positive charge as its electrons free react to the positive ions in the solution. This is the 'reduction' reaction. Due to the difference in charge, a voltage potential occurs between the metals and a current flow results.

In the case of the familiar car battery, as shown, lead is the reactant at one electrode (anode) and lead dioxide is the reactant at the other electrode (cathode).

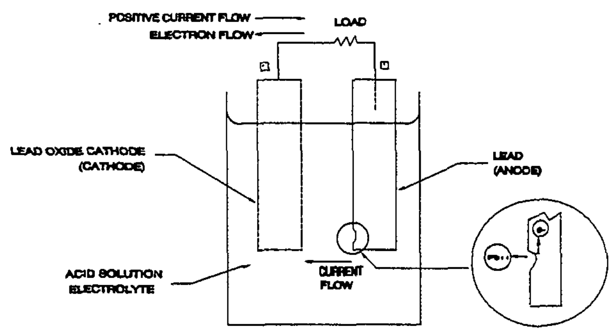



*Figure 1. Battery Cell, analogous to the corrosion dynamic.*

In the battery described above, one electrode is oxidized, one electrode is reduced, and the electrons flow in the outer electrical circuit. In freely corroding systems there is no outer electrical circuit. Therefore, the oxidation (which is the corrosion reaction) must occur in local proximity to the accompany reduction as in **Figure 2.**

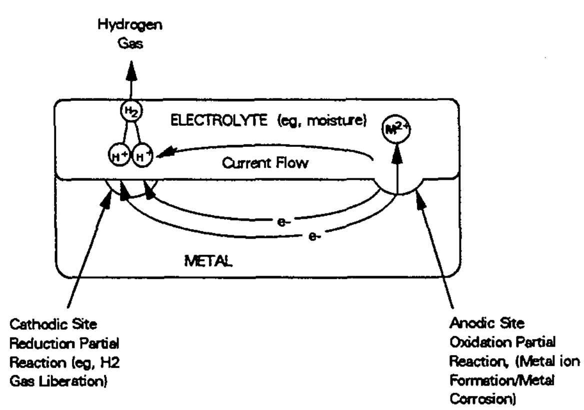



*Figure 2: Corrosion of Single Metal in Presence of Electrolyte.*

In the case of a single, 'freely corroding' metal, the oxidation and reduction can occur on separate surfaces (that are electrically connected). However, it is still very likely that the reactions are in close proximity to each other.

Corrosion is an electrochemical process and as such it is driven by the voltage difference between the oxidation reaction and the reduction reaction, as in our battery above. An example of an oxidation would be the dissolution of iron to form iron ions, and an example of the concurrent reduction is the reduction of protons to form hydrogen. (if you drop iron in acid you will see a high degree of bubbling, which is hydrogen evolution.) Some metals have different driving forces (voltages) so it is expected that some metals will corrode less than others. We also see that the corrosion rate will depend on the reduction reaction. If no electrochemical reduction could occur, no corrosion could occur. Typical reductions that accompany the corrosion reaction are proton (or water) reduction to form hydrogen (as mentioned above), and the reduction of oxygen to form water.

## Galvanic Corrosion

Since voltages are involved in the driving force for corrosion, we can change the voltage of a given metal to effect a change in the corrosion rate of the metal. One way to change the potential of a metal is to connect it to another metal that has a different potential. This is referred to as galvanic coupling.

A typical CATV galvanic couple involves copper of a center conductor (cathode) and tin of a port female contact (anode). When a galvanic cell develops, the sacrificial, or anodic metal typically deteriorates at an accelerated rate relative to its freely corroding state. **Figure 3** suggests that many potential combinations of galvanic corrosion cells may occur in an f-interface.

In summary, we see that the rate of corrosion depends on the type of metal and the environment of the metal. The environment of the metal includes any coupling to one or more metals, the chemical composition, and the temperature.

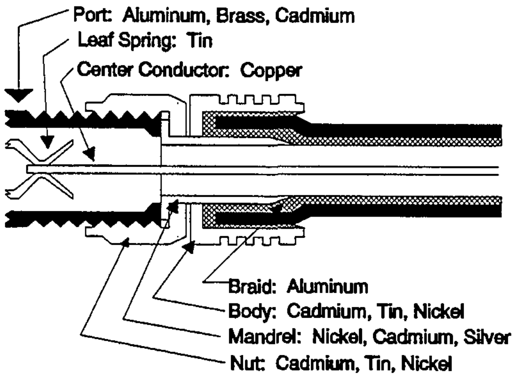



***Figure 3.*** *F-interface. The materials shown are those typically found at the surface of the given parts. Base metal of connectors are usually brass (copper, zinc, lead).*

## Impressed Currents

As corrosion is a current related phenomena, induced external currents can also significantly affect the rate of corrosion. Depending upon the direction of induced current, the reaction can be accelerated or hindered. In most drops, connections are mirrored on either side of a ground block or cable ends. In this case the induced current tends to drive the corrosion rate in one direction on one connector while in the opposite direction in the other. Therefore, inducing currents can have a positive effect on connections in one direction and negative in other.

The use of impressed currents to reduce the corrosion rate is known as cathodic protection. Aluminum, the weakest material in the melange of f-interface materials, is very sensitive to the level of impressed current. 'Overprotection' can occur if cathodic influence is to high. This is due to the quite unique *amphoteric* nature of Aluminum, whereby too much current causes increased corrosion.. It is for this reason that it is difficult to rely on cathodic or other impressed load methods to reduce Aluminum corrosion.

## Currents On the Drop

To address loading these concerns, it is important to understand the types of drop currents that can exist and practices that affect their levels. Low level currents are ubiquitous in the feeder portion as well as other parts of the CATV system. The reasons for low level currents on the drop vary. Two types of currents can exist in a system drop. These are 1) 'sheath' currents', which are low level currents most prevalent between the tap and ground block and 2) power load currents on the drop used for powering active devices outside of the home.

## Sheath Currents

Many drops experience small AC loads on the drop in the form of 'sheath currents'. Typical current levels have been documented for two different systems in **Figure 4a**. Note the wide variation of levels and presence of these currents. Of the homes that had measurable sheath currents, the average was approximately 173 milliamps (**Figure 4b**).

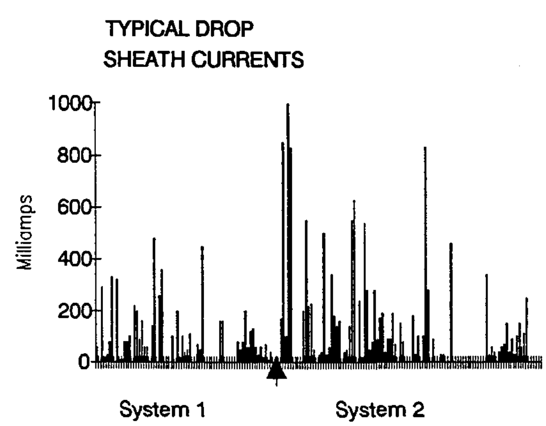



***Figure*** *4a, Typical sheath current values. Measured at two different systems. Courtesy Scientific Atlanta.*

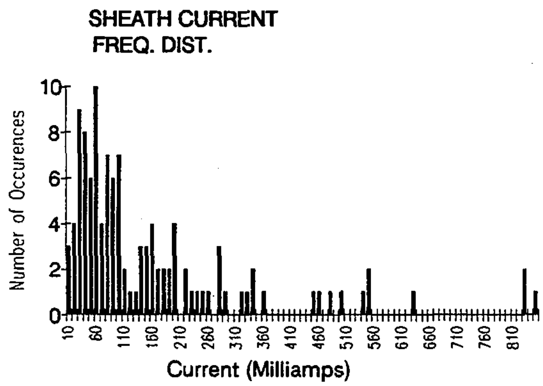



*Figure 4b. Sheath Current Frequency Distribution for the above systems. Shows number of subs at each current level.* Average of all subs=114 mA. Average of subs with measurable sheath current = 173.5 mAmps. Std Deviation = 191.1 Percentage of subs with measurable sheath current = 66%. Courtesy Scientific Atlanta.

These currents are often the result of distribution line Longitudinal Sheath Currents, generated on the feeder line, finding a path down the drop. Current first develops on the outside conductor of the distribution line due to bonding with a non-neutral utilities 'neutral' [1]. If this current is not grounded well or often enough along the distribution plant, much of the sheath current will go to ground via the drop to the ground block.

## Grounding and other Current Influences

Many practices currently exist for grounding along a distribution line that are meant to minimize sheath currents and to avoid damage due to lightning or other surges. Typical practices include bonding the feeder to the utilities 'neutral' at regular intervals, grounding at every active device, and grounding at the 'end of line' device (eg, terminating tap). Many practices are governed by city ordinances as well. Each of these practices, particularly the regularity of these grounds, has varying effects on the degree of resulting sheath currents.

A typical CATV grounding is represented by **Figure 5**. This shows the power line, its so-called 'neutral', and the CATV system plant. The power company typically runs three power lines with a common neutral. With any imbalance in these loads, a current is set up in the 'neutral'. As a result, it is not uncommon for the neutral to carry significant currents[1].

If we look at the schematic of current distribution, we see that the cable system, the power company, and the neutral-to-earth grounds all share longitudinal sheath currents originally generated on the power company lines, **FIGURE 6**. This schematic shows that the distribution plant consists of parallel circuits to three types of major grounds. The current in any one of the potential ground paths will be inversely proportional to their equivalent resistances. If the resistance of the ground is not low, much of the load will be shared by the CATV system. Up to 50 amps[1] can be expected on the CATV system due to the unbalanced load from just one power drop outage.

## Grounding of the CATV System

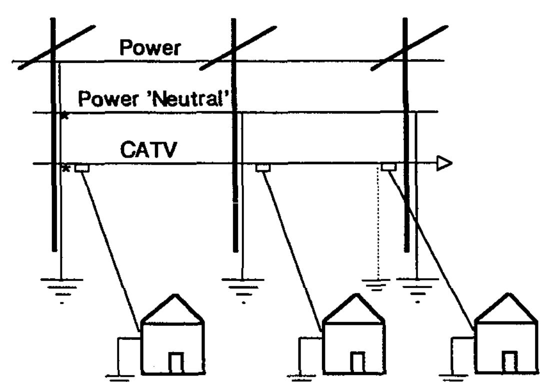



*Figure 5. Typical grounding of the CATV plant. Additional grounds (dotted line), are made at various intervals depending on the system.*

Further detail of the CATV feeder ground resistance distribution, **Figure 7**, shows the effect of drops on the ground paths. If any drops are not grounded, it can be shown that other drops will carry more of the distributed current. Also, if grounds were driven at every pole, the drop current sharing would decrease dramatically.

Clearly, the frequency and quality of many grounds contribute to the level of sheath currents on each drop. Any ground potential, such as the potential between the soil around the ground wire and the nearest feeder strand ground, may also set up a drop sheath current. In summary, the following will minimize sheath current on the drop.

1. Frequent grounding of the feeder (best at every pole).

2. Quality, low resistance feeder grounds.

3. Consistent grounding of drops.

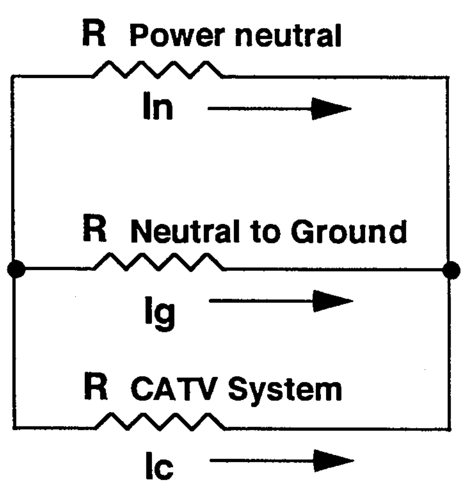



*Figure 6. Ground current distribution in the feeder/drop environment.*

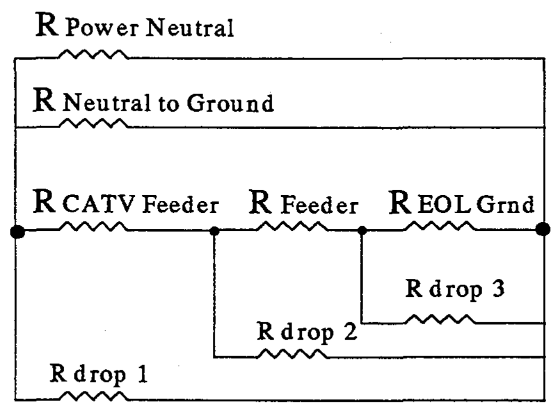



*Figure 7. Expanded current distribution showing the effect of drops.*

**Active device loads**

Some current and many emerging technologies such as interdiction, drop amplifiers, point of entry devices, and active taps are or will utilize AC and DC currents carried by the drop. These follow the standards set by UL 1409 and other manufacturing requirements. In general, these devices operate below 20 volts and around 200 milliamps, 4 watts being the approximate average load. As both the current imposed by active devices and that typically found in the form of sheath currents are roughly 200 milliamps, this was the value chosen to condition samples.

## OBJECTIVES

To see how these currents can actually electrically and physically effect the f-interface and hence the drop system, this study has been set up to explore corrosion of samples at the 200 milliamp level. The objective of this series of tests was to determine the following.

1) Effects of currents on the electrical performance of the CASS exposed f-interface.

2) Effects of currents on the physical degradation of CASS exposed f-interface components.

3) Any correlation between material and electrical deterioration of the f-interface.

With appropriate current levels determined, the goal is to seek information at the typical level. Future tests could explore different waveforms, pulsed currents, and other methods. The methodology for this series of testing is described below followed by results.

## PROCEDURE

### CASS Electrical Performance

In order to determine the effects of various loads on corrosion with CASS salt fog conditioning, performance of samples were compared after being organized into the following categories. The nominal power levels chosen are those which can be expected to be found due to sheath currents (primarily AC) and active drop devices (AC or DC) as mentioned above.

CASS with no load
CASS with AC, 200 mAmps, 15V
CASS with DC, 200 mAmps, 15V

As control categories, the variations of these loads were also measured without CASS conditioning. Effects were sought in two ways. First, F connector interface degradation versus time for these different conditions were characterized from a performance standpoint. Performance measurements were signal transmission, inner and outer conductor contact resistance, and signal egress.

Secondly, analysis of the corresponding physical material degradation was observed by Electron Microscopy.

### Powering Configuration

Ten samples each of DC and AC loaded connectors were subjected to CASS while 6 control samples were left in room temperature conditions throughout the test period. Samples consisted of standard hex crimp connectors and 60% shield PVC jacketed cable. The basic layout, shown in **Figure** 6, consisted of two hex crimp F-connectors, spliced together with an F-81 barrel splice. Each splice assembly was mounted securely to stabilizing boards. The splice was connected to ten feet of quad cable on either side each of which extended outside of the CASS environmental chamber. These non-conditioned cable ends were used to measure the electrical characteristics of the internal sample.

Samples were loaded as shown in **Figure** 7 Two power supplies fed 8 samples each, both in the case of the DC and AC. Currents were monitored daily for proper loading. Preliminary tests showed that the number of connections per supply must be kept low, for as the resistance of samples rose over time, the load necessary could exceed the capabilities of the supply and be difficult to monitor on a daily basis.

Measurements were conducted over a 56 day period, with electrical performance measurements taken at graduated intervals. Data was collected after 1, 3, 7, 14, 21, 28, 35, 42, 49 and 56 days. The line diagram for equipment used for each of the measurements are shown in **Figures** 10, 11, and 12 respectively.

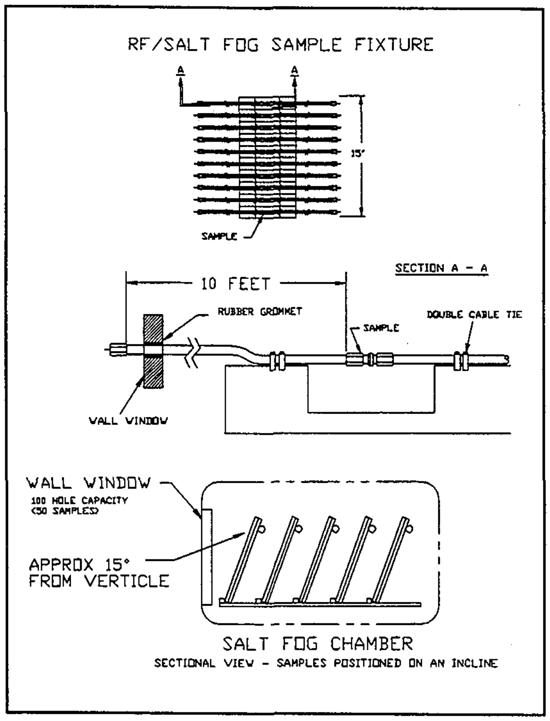



*Figure 8:* *Layout of CASS chamber and configuration for sample stabilization.*

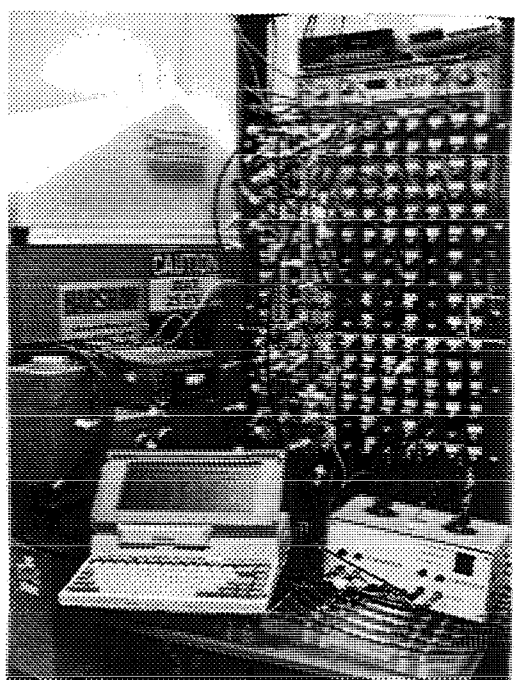


*Figure 9a.* *CASS Chamber with Control Panel/Computer*

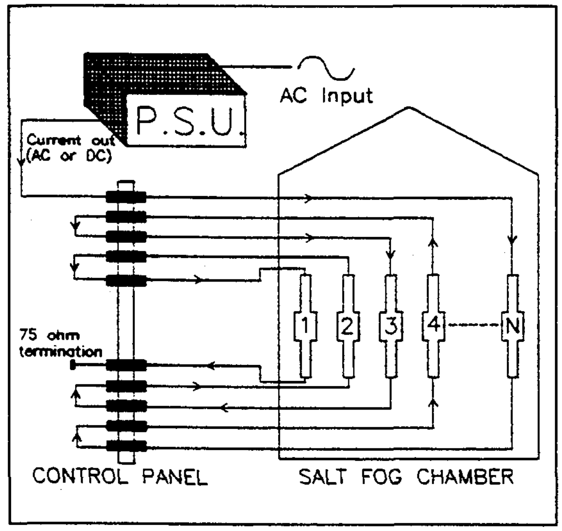



*Figure 9b. Line Diagram of current loading scheme.*

## Signal Egress

Signal egress measurements, **Figure** 10, were taken using an HP 11940A close-field probe. The probe is a balanced magnetic field sensor which provides an output voltage proportional to the strength of the magnetic field at its tip.

Egress measurements were probed at the rear of each of the two connectors, A and B, where the jacket meets the connector. The chamber was temporarily turned off and the lid removed for ventilation while taking egress readings. Samples remained otherwise undisturbed through the duration of tests. Other measurements were conducted outside of the chamber whereby the cable terminates externally.

The voltage values were read into a PC using a program which reads, stores, displays, and charts results. The frequency range taken was between 0 and 1000 Mhz. Egress levels are displayed in dB micro volts. The near field probe is useful as a relative measurement device, with one drawback the difficulty in correlating to shielding effectiveness. Typically, a very corroded sample degrades 10-20 dB. Also, the probe is very sensitive to physical handling. As a result, measurements for this study appear to be out of calibration. However, past studies do show some correlation to outer contact resistance measurements. Therefore, results are based primarily on contact resistance.

## Contact Resistance

Contact resistance measurements, **Figure 11**, were taken on either end of the conductor under test. using a Cambridge Technology Model 510 Micro-Ohmmeter.

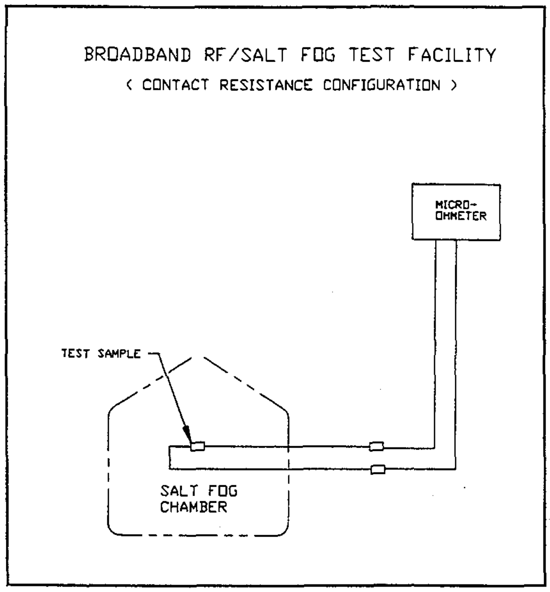



***Figure 11:** Contact Resistance measurement setup.*

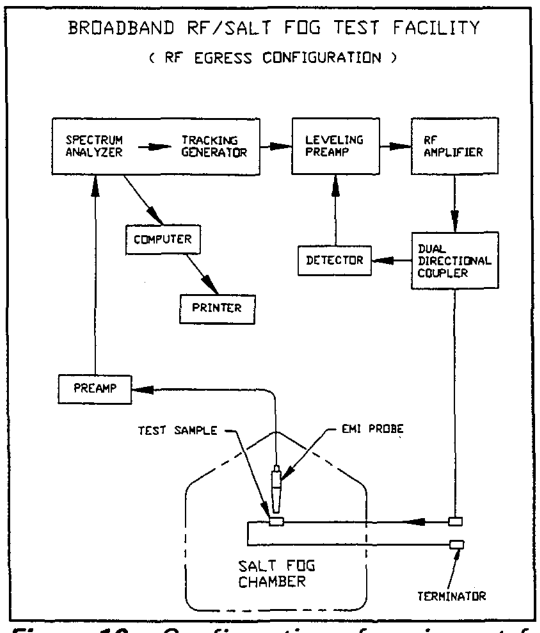



***Figure 10:** Configuration of equipment for signal egress measurements.*

## Signal Transmission

Signal transmission is measured with the equipment shown in **figure** 12. Radio Frequency signal is provided by the internal tracking generator of the HP 8590B Spectrum Analyzer. The signal was subsequently amplified by the Amplifier Research Amplifier and pre-amp. A feedback loop was constructed to keep the swept signal in range and consistent. This signal was transmitted into the drop sample and received directly by the spectrum analyzer. The results were automatically read, stored, and displayed through an IEEE interface. Again the frequency range was from 0 to 1000 Mhz. Signal level is shown in dBmV.

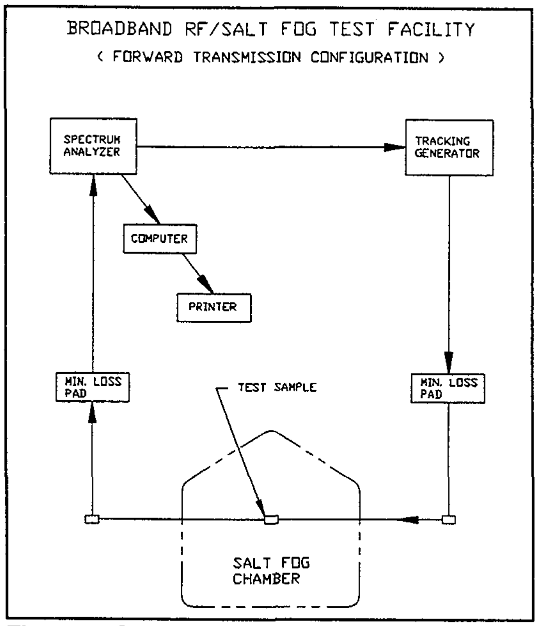



**Figure** 12: Setup for signal transmission measurement.

## Microscopy and Material Characterization Procedure

After samples were tested for electrical performance degradation during CASS exposure, samples were examined for corresponding moisture paths and material change.

## SEM

The scanning electron micro- scope was used to take electron generated micrographs for the purpose of showing paths of moisture salt spray product deposition. Note that the presence of salt spray products doesn't necessarily mean that significant corrosion has occurred. It does, however, show that an electrolyte is present. Actual corrosion is evident when material has been extracted (ie plating goes away exposing the base metal).

Energy Dispersive X-Ray Spectroscopy, allows one to determine the locations of material degradation (actual corrosion), particularly where plating has been removed and base metal is exposed.

## DISCUSSION

1) There will be minimal or no deterioration, with respect to currents on the drop, as long as moisture sealing is used. As shown in resistance data, **Figure 13**, regardless of the presence of load, no degradation occurs without CASS/moisture exposure. The virtually equal, flat profile, dotted lines in this graph represent non CASS samples.

2) When f-interfaces are exposed to CASS, deterioration occurs from most to least rapidly in the following order, DC, AC, and no load. In **Figure 13**, the outer conductor resistance measurements show substantial degradation of CASS samples in that order. It is known that the ratio of resistances, outer to inner, reflects the impedance match and hence the return loss and shielding effectiveness of the interface. As shown, this ratio has substantially changed with current loading as compared to no loading.

3) The inner (center conductor) resistance, when exposed to CASS, was to a high degree variable and intermittent, **Figure 13**. Loaded samples did appear to be slightly more variable. However, again there is no effect when the samples were kept free of the corrosive environment.

4) As variability in electrical performance occurred via center conductor components, the degree of moisture migration into their interfaces varied as well. It appears from micrographs that substantial variability exists between samples regarding the degree of moisture migrating to the center conductor contacts. **Figure 15** shows a typical sample whereby substantial moisture has penetrated one side of the contact while the other remained relatively clean. Due to this occurring in many samples, the degree of moisture contacting the center conductor could not be considered controlled. Micrographs show consistent moisture and corrosion

occurring at the outer conductor interfaces, typically appearing as in **Figure 16**. Compared to the moisture path leading to the outer shield and the connecter outer conductor pieces, the path of moisture to the center conductor has many barriers. This may explain the variance in results relative to the outer resistance results of **Figure 13**.

Other factors which may have caused variance of center conductor related performance are fretting corrosion and the spring effect of the f-81 interface. Fretting corrosion, caused by micro-motions in the contact areas, creates repeated opportunity for oxide layers to develop between contacts. Conversely, contact at the cc f-81 interface may be restored when minimal vibration causes oxide layers and corrosion products to be dislodged and the spring regains contact due to constant compression inward against the cc. Measures were taken to minimize these occurrences by stabilizing samples, however, small yet significant movements may have occurred. These counteracting actions may explain rapid performance losses followed by rapid gains.

5) Signal transmission, unlike center conductor contact resistance, showed almost no change over the time tested, **Figure 14.** The gap caused by corrosion between the center conductor and the f-81 leaf spring only affects the signal transmission slightly. However it greatly affects resistance. As many pressure 'probing' tap manufacturers have shown , RF transmission can occur with a small gap between contact surfaces, whereas resistance increases dramatically or goes infinite with only a slight gap. Due to the transmission less sensitive nature, significant deterioration (10-20 dB down) could be expected to follow the resistance results if the test time were extended.

It is recommendation for further reasearch to extend the test period and/or use lower quality, larger opening F-81's, to advance the corrosion process and assure all contacts see uniform corrosion potential moisture exposure.

## RECOMMENDATIONS

It has been shown that currents can be detrimental to the life of the drop. In order to minimize the effects of low level currents from sheath or active drop devices, the following are recommended.

1) Means of sealing the f-interface should be used. This will avoid any effects that currents may have on deteriorating the drop.

2) Means for powering external to the standard coaxial cable RF conductors should be considered, such as a messengered type of coextruded conductor.

3) If existing drop is to used as a current carrying media, AC current is recommended over DC. The ideal level of current loading from a corrosion standpoint is left for further study.

4) Efforts should be made to minimize the level of sheath currents on the drop. The following practices are recommended.

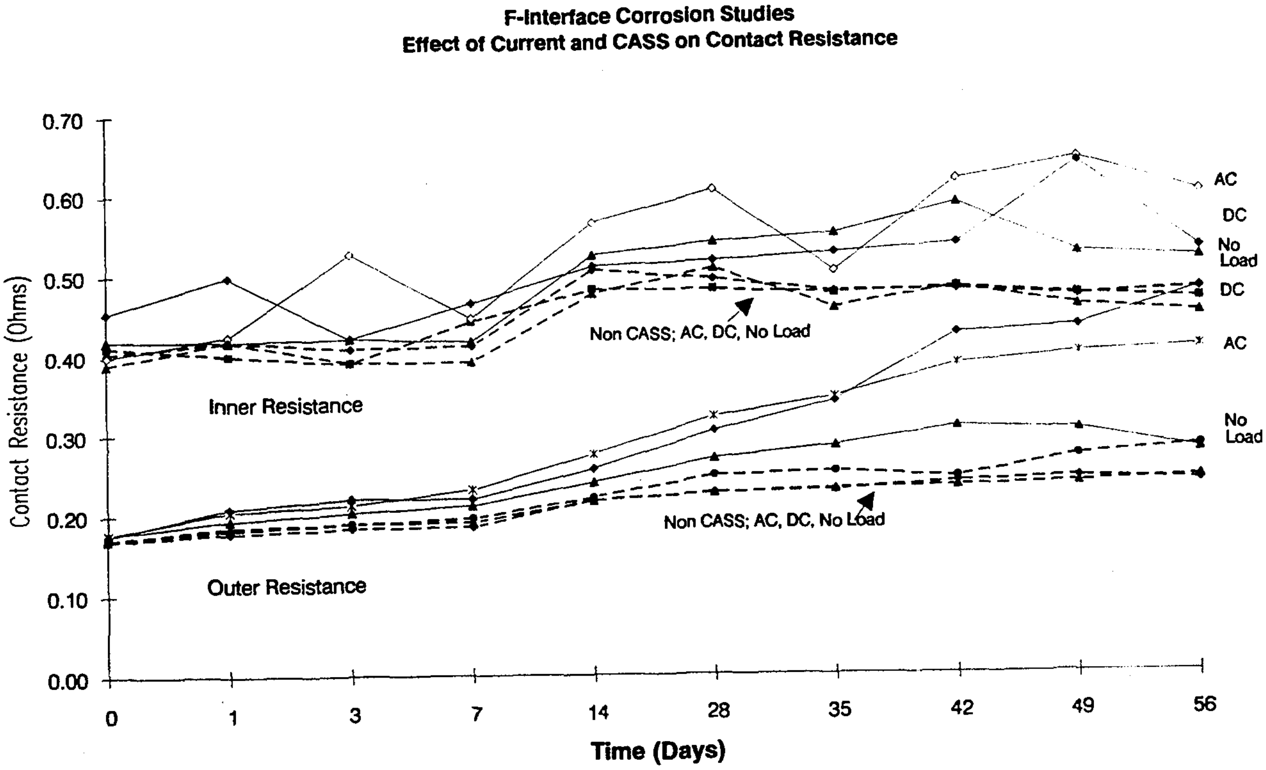
a) Use proper procedures to assure sufficient grounding at the terminating tap.

b) Ground at frequent and regular intervals along the distribution lines.

c) When feasible, assure that no stray currents or other conditions exits which may cause a potential difference between the grounding of the drop and the distribution line. The age and condition of the Utilities grounds are not always trivial and should be considered.

d) Operators are encouraged to get involved in the formation of standards and practices currently being drafted. The SCTE and CableLabs are currently developing sound advice and recommended practices in this area.

## Figure 13




## Figure 14

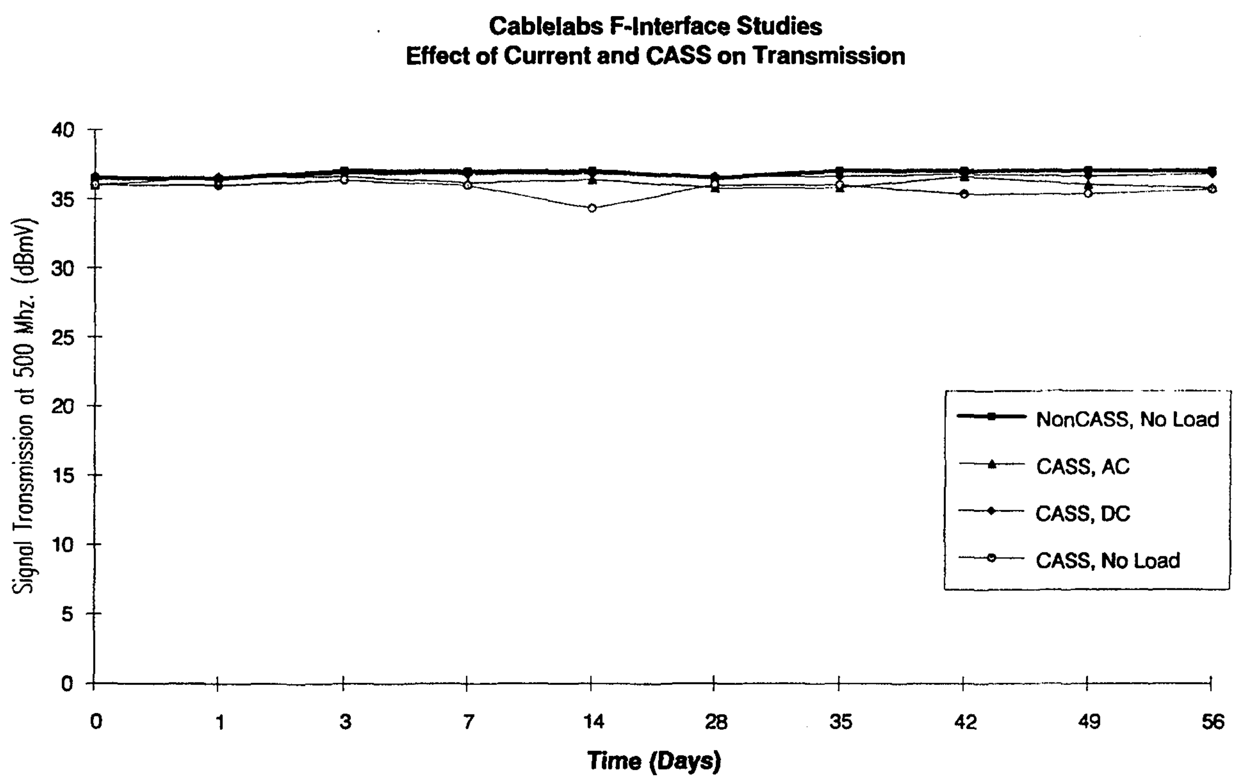
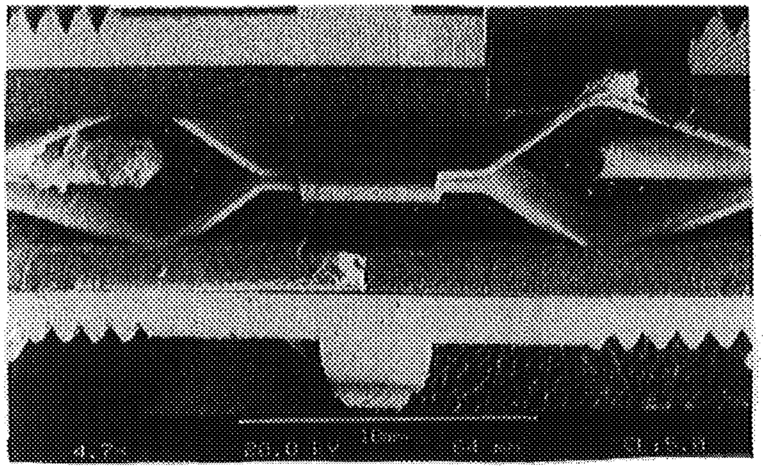


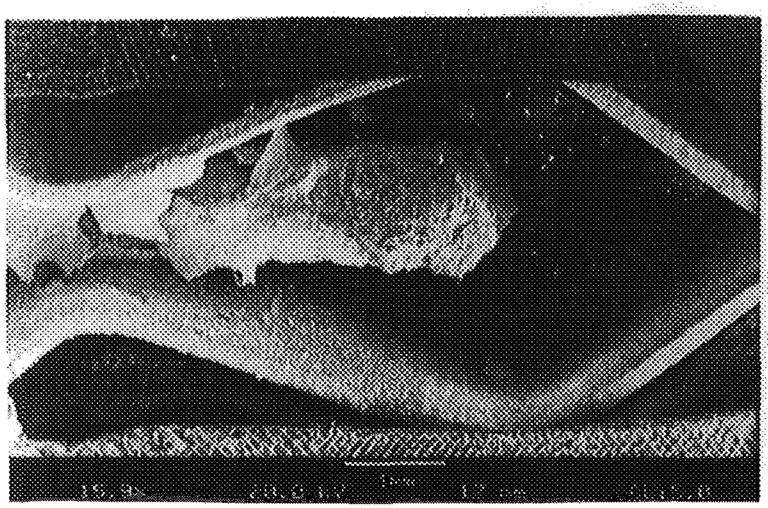
FIGURE 15

Typical variance in Moisture Migration, F-81 Splice Center Conductor Contact Area

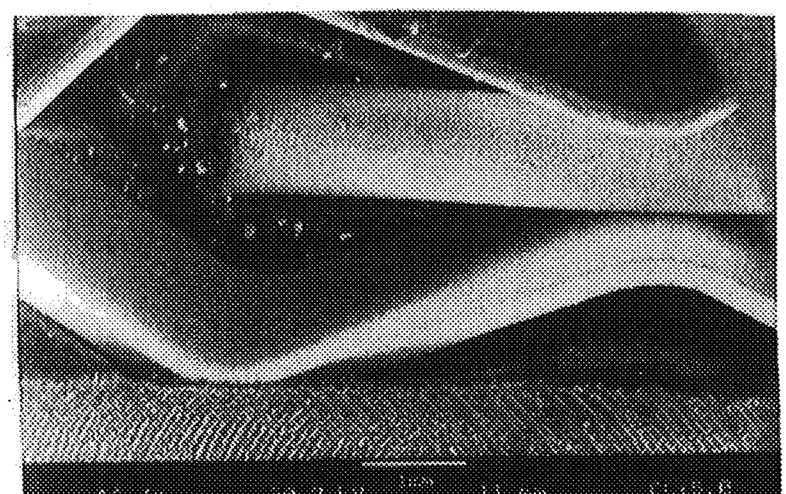


Cass Exposed

Internal Contacts

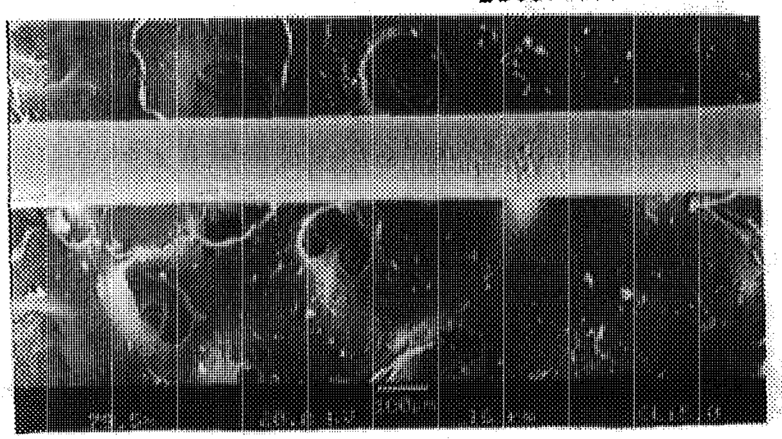


Cass Exposed, Right Contact Area

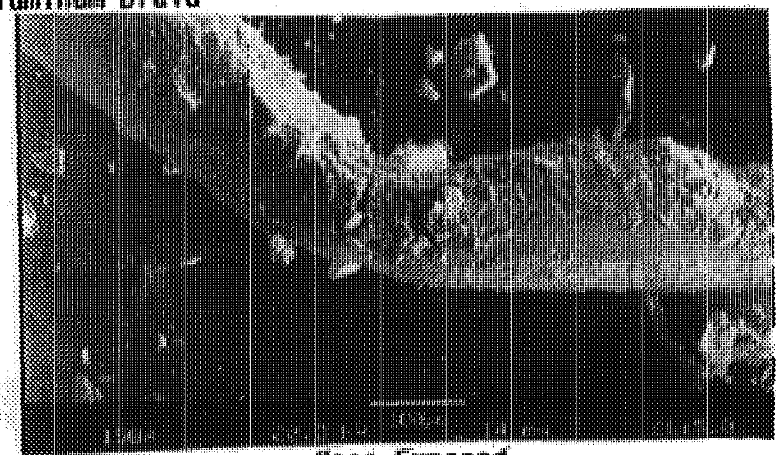


Cass Exposed, Left Contact Area

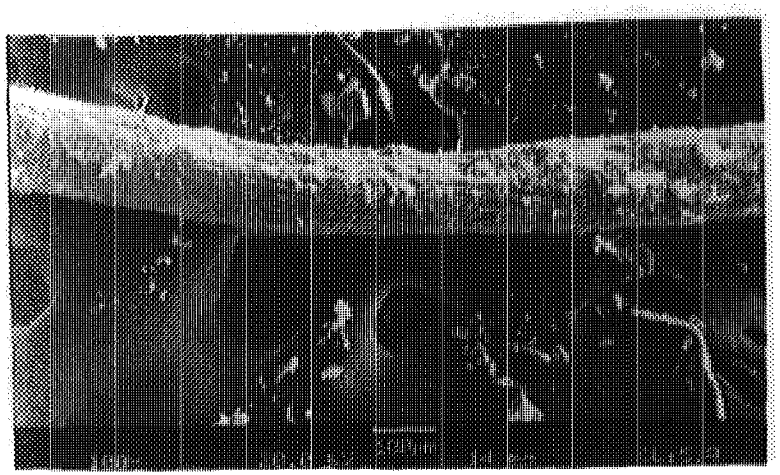
FIGURE 16
Deterioration of the Aluminum Braid



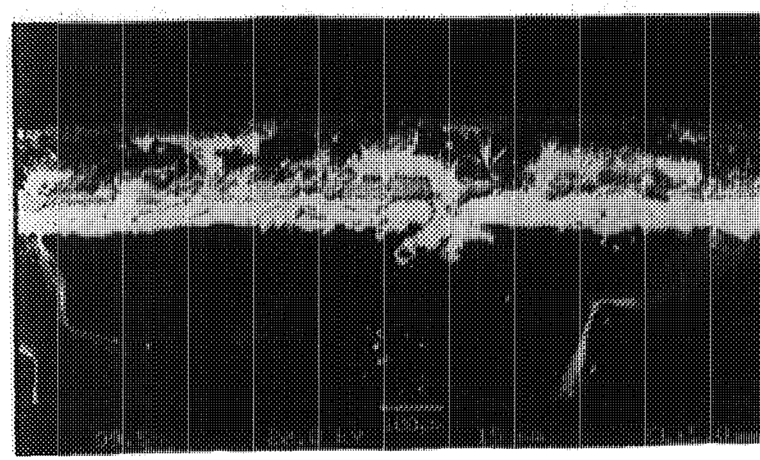
Unconditioned



Cass Exposed



Cass Exposed, Left DC Current



Cass Exposed, Right DC Current

## ACKNOWLEDGEMENTS

The author would like acknowledge the assistance of the following people; Marty Lee, Gregor Sandhaus, Bob Holt, Derek Leong, and Marianne Stephen of The Raychem CATV Team for countless hours of work, the advice of Scott Bachman (CableLabs) and Tom Elliott (CableLabs and TCI), discussions with Tom Williams from CableLabs, coordination of sheath current information and shared experience from Jim Farmer of Scientific Atlanta and Jack Bryant of Antec, and field visits with George Campbell of Viacom.

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# Implications of Ghost Cancelling to the Cable Industry

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**Abstract**

The ghost cancelling effort is steadily migrating from a developmental stage to commercial applications and will eventually target to a wider application - into consumers' television sets and VCRs. The influx of ghost cancelling into the cable industry is imminent. This paper will examine the issues that are important to the cable industry; it is based on earlier experiences on testing ghost cancellers done in: Vancouver - British Columbia (B.C.), Kitchener - Ontario, and the CableLabs' tests in Washington, D.C.

## Introduction

Multipath is one of the major impairments cable companies encounter in receiving over-the-air television signals. Traditionally, RF cancellation technique (sometimes baseband technique) is used. However, it requires trimming long lengths of cable to match the appropriate delays or adjusting sensitive phase-

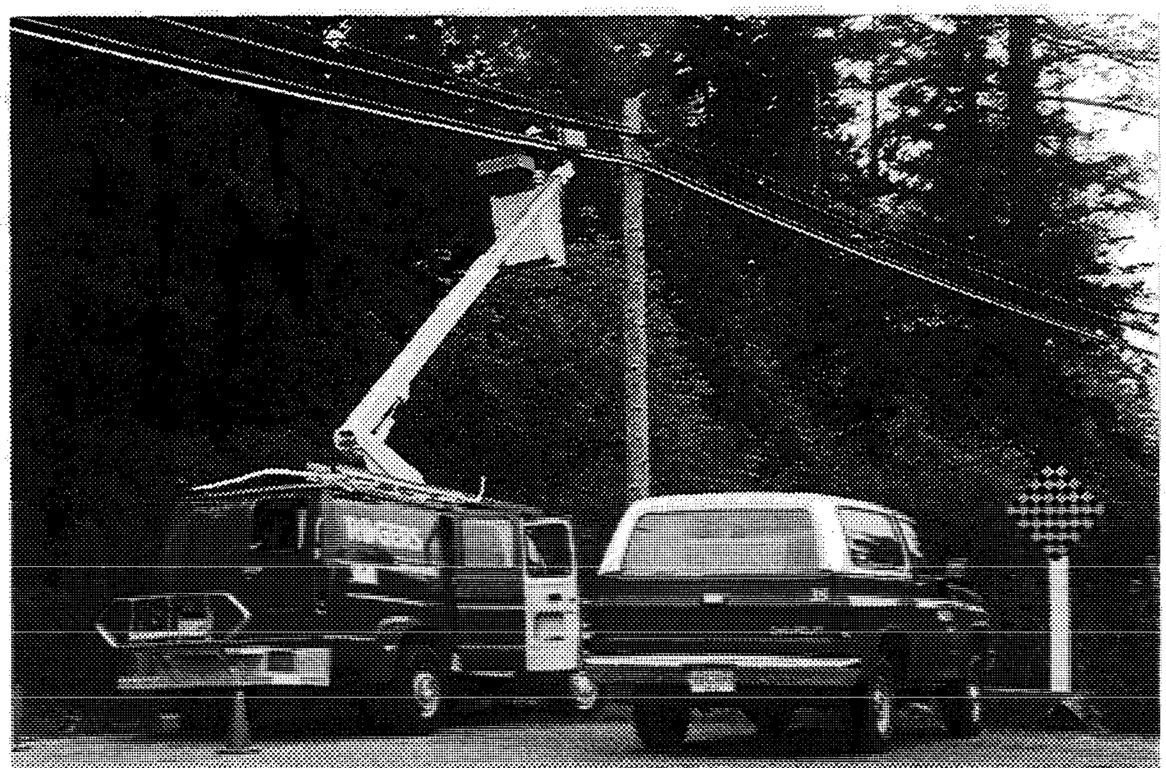

Figure 1 Ghost Cancelling Crew Investigating at a System Location.

shifting networks and matching precisely the amplitude of the reflections. In addition, the resultant signal suffers a signal-to-noise degradation during signal recombination and, most often, 100% cancellation cannot be achieved. Rogers Cablesystems initiated the ghost cancelling project in early January of 1990. A prototype ghost canceller was tried in Salt Spring Island, B.C. on April 6 & 7, 1990 immediately after the NAB convention. A month later, the first ghost canceller was put into full operation serving subscribers in Vancouver and its vicinities. Figure 1 shows a photograph of the Vancouver ghost cancelling crew investigating at a system location in Victoria, B.C.

ATSC's Effort to Establish a GCR Signal Standard

In 1989, the Advanced Television Systems Committee formed a specialist group on Ghost Cancelling called T3S5 to examine the various ghost cancelling techniques. There were five systems - AT&T, BTA, NA Philips, Sarnoff and Samsung - submitted to the ATSC for consideration. It was hoped that with a succession of field test, cable test and laboratory test, a single voluntary standard could be established for the television industry in North America. Six organizations were involved in this effort. The broadcast field test was conducted in mid-September of 1991. It took place in the Washington D.C. area. Three television stations, one in the VHF and two in the UHF bands, were used for the test. Over a hundred and fifty test locations were visited.

The cable test followed immediately after the field test. It consisted of two parts: the cable impairment test and the cable system test. The impairment test encompassed a series of cable impairment simulations which was carried out in the CableLabs HDTV testing facility in Alexandria, Virginia. The cable system test was performed on four cable systems in the vicinity of Washington D.C. and comprised of thirty-three subscriber locations. Measurements were made at headends, AML hubs, distribution taps and within subscribers' homes. The subscriber locations were carefully selected with a balanced mix of AML, trunk and optical link.

The laboratory test was carried out at the Communications Research Center in Ottawa last January. The test consisted of computer simulations of the different GCR signals being impaired by ten different combinations of ghosting and noise conditions; observations were focused on each proponent's GCR signal and its software implementations, such as, its ability in characterizing the transmission channel, the duration required to obtain a good approximation of the channel response, and the number of iterations necessary to achieve convergence. The laboratory test also used segments of the CCIR test tape, originally prepared to assess video codecs, to subjectively evaluate the picture quality improvement. Ghosts and random noise were introduced at IF and RF frequencies. The laboratory test was concluded at the end of January.

The results of the field test, the cable test and the laboratory test will be submitted to the ATSC. ATSC will recommend a standard to its membership for a vote in April of 1992. At the same time, a line in the vertical blanking interval will be determined for the transmission of Ghost Cancel Reference (GCR) signal. The tabulation of the vote will be disclosed in June of 1992.

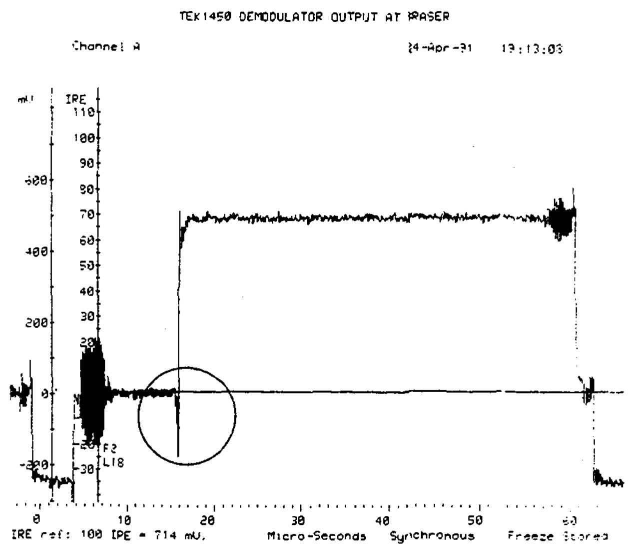


Figure 2 GCR Signal Received at the Fraser Hub.

<u>Vancouver Test</u>

Vancouver was the initial test site chosen for the ghost canceller test. There were a few important findings in the Vancouver test that prompted further investigations into the impact of ghost canceller to a cable system. For example, AM and FM microwave systems are widely used for cable television distribution. However, if they are not well maintained, non-linearity introduced by AM and FM microwave links can adversely affect the ghost cancellation performance and convergence time. Figure 2 shows the received GCR at the Fraser AML receive site. The GCR signal was inserted before the AML transmitter in Burnaby on channel 32. Notice that there was considerable undershoot but no overshoot on the integrated (sin x)/x signal. Under normal circumstances, the undershoot did not create any problems because it was blacker than black. However, when the GCR signal was captured and processed by the ghost canceller, the ghost canceller interpreted the undershoot as a ghost and tried to cancel it. The net result was that a ghost was introduced by the canceller itself.

<u>Kitchener Test</u>

Prior to the commencement of the Kitchener test, there were many uncertainties about ghost cancellers operating in a cable environment:

- Group delay caused by duplex filter and mop-up equalizers often cause chrominance to luminance delay inequality. Could these ghost cancellers be able to improve these distortions and to what extent?
- How would these ghost cancellers perform in a typical cable subscriber's environment, with the presence of a variety of cable distortion products, which satisfied the specifications outlined in

the FCC rules Part 76?

- Microreflection is one of the major concerns for in-house wirings that do not conform to CATV standards. Would these ghost cancellers be as effective as to cancel ghosts which were distinct and far apart than to cancel ghosts which were clustered very close together near the picture carrier?

In August of 1991, a test was launched in Kitchener, Ontario to investigate the effectiveness of a ghost canceller operating in a modern and well maintained cable system with AML distribution. Two test channels were used: one on channel 2 and the other on channel 48. A total of nineteen test points were selected giving thirty-eight sets of data consisting of subjective picture quality evaluations and objective channel waveforms. Since different brands of ghost canceller react differently to the signal that was contaminated by ghosting and other distortions, to simplify the analysis, it was decided that only one ghost canceller would be used.

There were a few specific findings as a result of the Kitchener test. From the subjective data, the results indicated the process through the ghost canceller did not alter the picture quality. However, assisting the judgement with the objective channel waveforms reviewed that 89% of the test points actually showed an improvement and only 11% indicated that the picture quality was degraded after the ghost cancelling process. Figure 3 shows a demodulated cable signal at a distribution tap and Figure 4 illustrates the same signal after it was processed by the ghost canceller. Notice the small improvement on the 2T pulse and

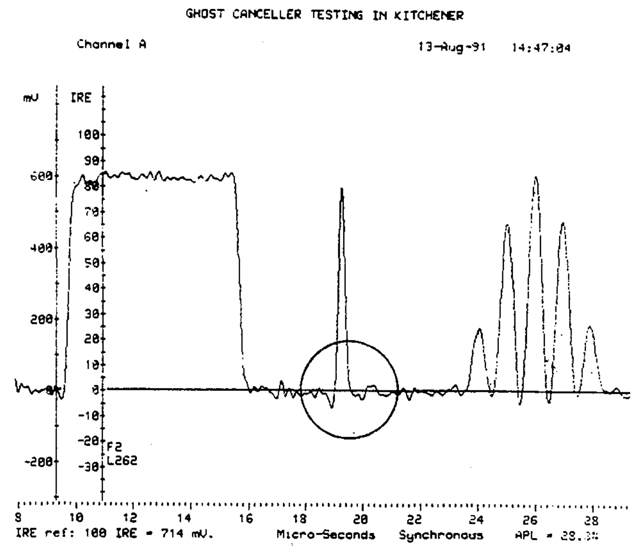


Figure 3 Demodulated Signal at Distribution Tap.

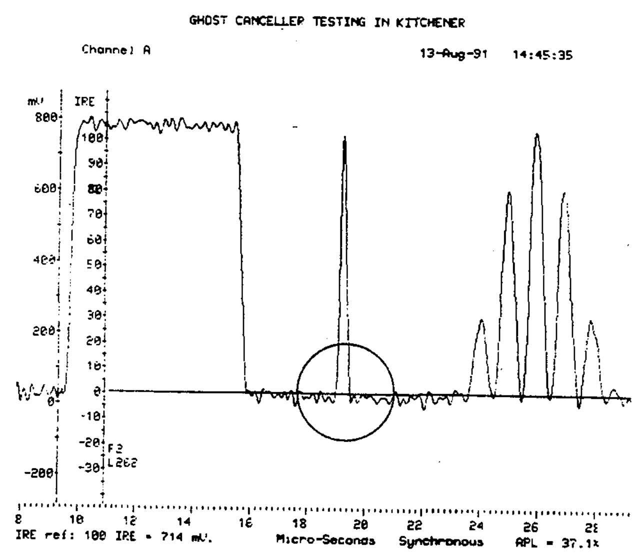


Figure 4 Demodulated Signal Processed by Canceller.

subjectively, it is difficult to perceive the difference by merely observing the pictures shown on the television set.

Contrary to the result of the Fraser hub in Vancouver, the Kitchener data indicated that AML equipment, if well maintained, did not impose any restriction on the ghost cancelling equipment to perform.

<u>Suggestions on purchasing Ghost Canceller</u>

The technology to implement ghost cancelling system clearly exists. Japan introduced the ghost cancelling system as part of the ClearVision implementation in 1988. North America will have a GCR signal standard this year. It is believed that commercial ghost cancellers will be introduced to the broadcast and cable industries even before the 1992 NCTA Convention. Yet, what does a cable company need to consider before buying a ghost canceller? Rogers Engineering would like to share some of our experiences:

- These devices exhibit some peculiar characteristics and, quite often, require a software reset to cure the condition. Therefore, it is highly desirable if a front panel as well as a remote reset (for unmanned antenna sites) functions are available.
- Occasionally, broadcasters may necessitate the removal of the GCR signal for maintenance or trouble-shooting purposes. Normally, a ghost canceller updates the ghosting conditions by continuously captured and processed the GCR signal. However, it may exhibit an instability state due to the loss of GCR signal. Thus, there is a need for ghost cancellers being able to freeze the last set of filter coefficients during which the GCR signal is missing.
- If electrical interference (or sometimes called impulse noise) is present on the received signal, here is another feature to ask for: its robustness to electrical interference. Figure 5 shows the BTA GCR signal corrupted by electrical interference. As a result, there may be momentary loss of the picture for a short duration. If the impulse noise happens often enough, the momentary loss of picture becomes very annoying.

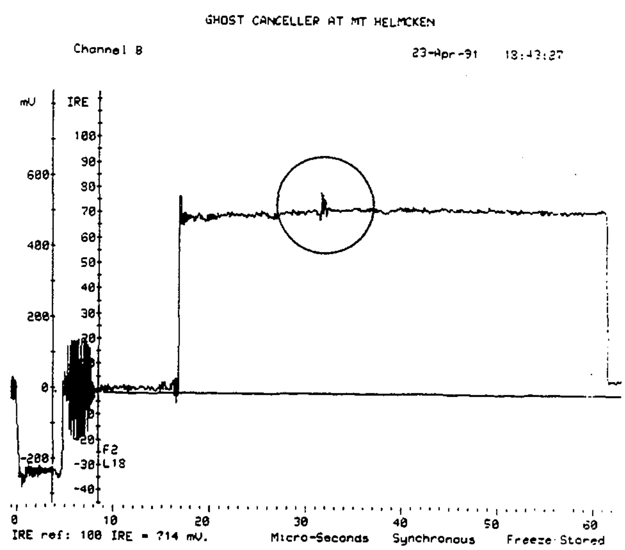


Figure 5 GCR Signal Corrupted by Electrical Interference.

- Under certain circumstances such as a deep signal fade or equipment failure, there may be a need to by-pass the ghost cancelling process. Therefore, front panel as well as remote by-pass switches are useful.
- Ghost cancellers are

baseband devices. If channel processors are presently used to process over-the-air signals, one may want to ask ghost canceller manufacturers whether there is an IF interface option available.

- Lastly, line 19 on the vertical blanking interval is being proposed by the ATSC for the transmission of GCR signal and, as a consequence, the usage of line 18 and line 20 may be restricted to time invariant test signals only. But, line 20 is being used by some cable operators for conditional access and the information on this line is time-varying. Cable operators may want to solicit ghost canceller manufacturers to incorporate GCR signal deletion circuitry once the television signal has gone through the ghost cancelling process. And, maybe, there are other alternatives than deleting the GCR signal.

## Conclusions

The list of suggestions presented on buying a ghost canceller is by no means complete. It is the intention of Rogers Engineering to share the experiences of our ghost cancelling system development effort with other cable companies and, via the sharing of ideas, to stimulate more discussions and generate more ideas on the implementation and deployment of ghost cancellers.

## Acknowledgement

The author would like to thank Rogers Cable TV in Vancouver, British Columbia and in Kitchener, Ontario for their assistance in conducting the field tests. The evaluation of ghost cancellers is sponsored by CableLabs. The cooperation of Tektronix, KOMO, KIRO and KING stations, Communications Research Center, and the support of various ghost canceller manufacturers is vital to this project. Their help and support are most appreciated.

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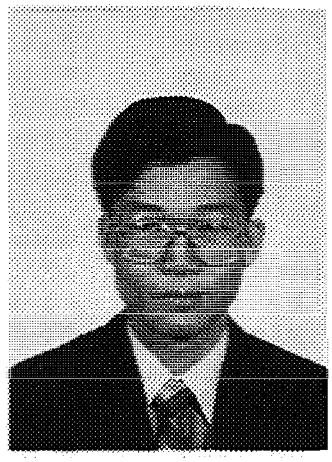

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# Leading Edge Photonic Technologies

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## ABSTRACT

The impact of photonic device research on future lightwave systems, in particular, those that use wavelength-division multiplexing (WDM), is discussed. WDM techniques that maximize the capacity of fiber require increasingly complex transmitters and receivers. Examples of current state-of-the-art prototype devices are presented in which this complexity is built into monolithic photonic integrated circuits.

## 1. Introduction

Recent enthusiasm for lightwave systems within the cable television industry has been driven by rapid advances in laser technology, most notably that of linear high-power distributed feedback (DFB) lasers. Development of linearized external modulators, for use with high-power diode-laser pumped YAG lasers, and improvements in erbium-doped fiber amplifiers (EDFAs) have opened exciting new possibilities. Continued deployment of these technologies over the next few years will provide fiber-based cable networks with unsurpassed capacity and diversity of services.

Prior to this interest in analog systems, the availability of high speed lasers and photodetectors created a similar revolution in digital long-haul, loop feeder, and data networking systems. EDFAs are now enabling non-regenerated trans-oceanic lightwave systems at 5-10 Gbps over 10000 km.[1,2] New photonic device capabilities continue to drive new system concepts.

What new types of devices can be expected and how will these effect system concepts of the future? In attempting to answer these questions it is useful to consider what types of systems might be required.

Already we are seeing the merging of analog and digital services within industries that were once strictly analog or digital. Of course, we always see potential applications for increased capacity. Hence it is essential that future systems have increased capacity, and flexibility in the way in which services can be provided. Wavelength-division multiplexing (WDM) is key to increasing both capacity and flexibility. Yet the photonic devices required for all but the simplest forms of WDM remain confined to research laboratories.

As long as the access to the subsciber is either coaxial cable or twisted pair, then the system capacity must always be constained by this bottleneck in the distribution plant. One must imagine that fiber will, some day, be viable all the way to the home. This will required drastic reductions in the cost of photonic transmitters, receivers and multiplexors. Photonic integration, in which sources, detectors, interconnecting waveguides, electronic devices and passive optical components are fabricated as integrated circuits in one material system, are key to reaching cost targets.

This paper discusses recent milestones resulting from photonic device research, as applicable to these future possibilities.

## 2. Wavelength Division Multiplexing

Three categories of WDM can be defined. The first and simplest uses fiber WDM couplers to combine and separate sources at 1.3 and 1.55 $\mu$m wavelengths. This will be referred to as $\mu$m-WDM, and is feasible with technology that is readily available. It provides a convenient way to reduce fiber requirements by a factor of two, increase channel capacities, or diversify services.[3]

Next, far greater capacities can be realized if multiple wavelengths are used within each of the 1.3/1.55 $\mu$m wavelength windows. Since each of these windows are in the order of 50 nm wide, dramatic increases are possible with channel separations as narrow as 1 nm. (1 nm = 10 Å = 130 GHz = 0.001 $\mu$m)

These nm-WDM techniques require multiplexor properties that cannot generally be obtained from fiber WDM couplers. Integrated planar waveguide WDMs are required to achieve suitable channel densities.[4] Fiber Fabry-Perot filters can be used to select channels. Optical sources must be manufactured to exact specified wavelengths, or must be broadly tunable over the entire wavelength window. The operating wavelength must then be locked to within approximately 1 Å from the center wavelength of the channel. Frequency-locking lasers to within this large margin for error is reasonably straightforward. If these technical issues are overcome, nm-WDM techniques have great potential.

The third, and ultimate WDM technique involves optical channel frequency separations that are in the order of a few GHz, hence this will be referred to as GHz-WDM. Various coherent and direct-detection techniques have been researched that enable this highly efficient use of optical bandwidth.

Direct detection can be used for GHz-WDM in two main ways. The simplest approach is to modulate the intensity of each carrier with an external modulator, then select each modulated channel with a narrow optical filter. An alternative approach is to modulate the frequency of the optical carrier. New tunable laser structures are ideal for this. An optical filter can then be used as an optical frequency discriminator to pass only the desired channel, and simultaneously convert the FM to intensity modulation (IM). One then simply detects the IM on a photodetector.

For a considerable increase in complexity, coherent detection can provide slightly higher spectral efficiency. Carrier (optical) frequencies must be controlled to within fractions of 1 GHz. The polarization of the received signal must be controlled, or polarization diversity techniques must be used. Laser linewidths must be small. Considering the complexity, it is difficult to imagine a system application that will require the capacity for which coherent transmission is essential. Yet, since coherent represents the ultimate in capability, research continues.

Given the potential of nm-WDM, why might GHz-WDM ever be needed? With 10 GHz channel separations, each wavelength window could support 500 channels, each with a bandwidth of a few GHz. Therefore it is unlikely that ultimate bandwidth will be the issue. The choice will eventually be made based on how photonic devices evolve to suit each option.

## 3. New Photonic Devices

The ideal source for nm- or GHz-WDM systems would be tunable over an entire wavelength window (50 nm). This could then be used with an external frequency-lock circuit as a source for any desired channel. Several tunable lasers have been invented. The tunable distributed Bragg reflector (DBR) laser [5] has great potential, but the best tuning range reported to date is less than 10 nm. A new approach [6] has resulted in the demonstration of a tuning range greater than 50 nm.

Broad tunability is achieved with this new structure, shown in Figure 1. The laser consists of a gain section, pumped by current $I_g$, and a tunable filter. Current $I_t$ controls the wavelength allowed to circulate throughout the vertical coupler filter (VCF) structure. The VCF is the key to broad tunability.

The grating on the upper edge of the upper active waveguide allows power to couple to the lower passive waveguide. Varying $I_t$ varies the wavelength that is coupled. Light of the appropriate wavelength couples to the lower waveguide, is reflected from the cleaved reflector, and then returns to the upper waveguide. The VCF structure provides an enhancement in the tuning range of the filter, for a given amount of tuning current.

The tuning characteristics for the VCF laser are shown in Figure 2. Tuning of 57 nm has been demonstrated. The tuning current was pulsed at for extreme values to minimize heating within the chip. Work continues on understanding and optimizing the performance of these devices.

One reason to favor GHz-WDM over nm-WDM has been the limited tuning range of the lasers. With a limited tuning range it is advantageous to maximize the bandwidth accessible to each laser. The high spectral efficiency offered by GHz-WDM then appears worth the added complexity. Devices with extremely broad tunability remove this limitation, adding strength to nm-WDM opportunities.

In addition to broadly tunable sources, simple means to modulate the optical carrier must be developed. Direct laser modulation is adequate for intensity modulation (IM) if chirp (unintentional frequency modulation (FM)) is tolerable. If FM modulation of the optical carrier is desired then new laser structures can provide reasonably pure FM. But nm- or GHz-WDM applications may require pure IM, or phase modulation (PM). This requires external modulation. Unfortunately, coupling a tunable laser made from InP-based semiconductor material to a polarization-dependent external modulator made from $LiNbO_3$ will likely remain expensive.

The solution lies in fully integrated laser-modulators, all made in the InP material system. Figure 3 shows a prototype device that couples a tunable DBR laser to an elecro-absorption modulator[7]. Unlike the more common interferometric intensity modulators, this modulator operates by varying the absorption of the material in the layer that has a bandgap energy corresponding to 1.1 $\mu$m wavelength. The multiple quantum well (MQW) stack provides gain for the DBR laser formed between the grating and the outside edge of the MQW stack. Light from this laser overlaps the absorbing 1.1 $\mu$m layer and is modulated before exiting the anti-reflection coated facet. The DBR laser is tunable over several nm wavelength.

Devices such as those presented so far will open the way for nm-WDM and some forms of GHz-WDM. Coherent systems would require such devices, but if costs are to ever be reasonable, several types of devices must be integrated into low cost modules. Photonic integration has the potential to combine these devices on low cost mass-producible semiconductor chips.

An example of an integrated photonic circuit for a coherent receiver[8] is shown in Figure 4. Here the local-oscillator (LO) laser, combining coupler, waveguide interconnects and balanced receiver are combined onto one chip.

The LO laser is a 3-section broadly-tunable (10 nm) DBR laser. Signals received in the input waveguide are combined with the LO in the directional coupler. Both outputs of the coupler are detected by a pair of photodetectors.

This photonic integrated circuit has been tested for coherent digital transmission at several hundred Mbps. It does not provide polarization control or polarization diversity, an additional complication that must be addressed.

N. Conclusions

Several new photonic devices have been presented. By increasing the complexity of the device structures, effectively combining several devices into one monolithically fabricated photonic circuit, these devices have the potential to make futuristic WDM proposals into reality.

Unfortunately, years of development are required before these research prototypes become components for widespread deployment. The cost of such development cannot be justified without the definition of applications that require the enormous capacities offered. Such requirements may not appear until far in the future, when fiber extends all the way to the subscriber. Meanwhile, new photonic devices will continue to emerge, inspired by the demands of increasingly imaginative system applications.

The devices presented were conceived and produced by T. L. Koch, U. Koren, B. I. Miller, M. G. Young, R. C. Alferness, C. A. Burrus and a large group of collaborators at AT&T Bell Laboratories.

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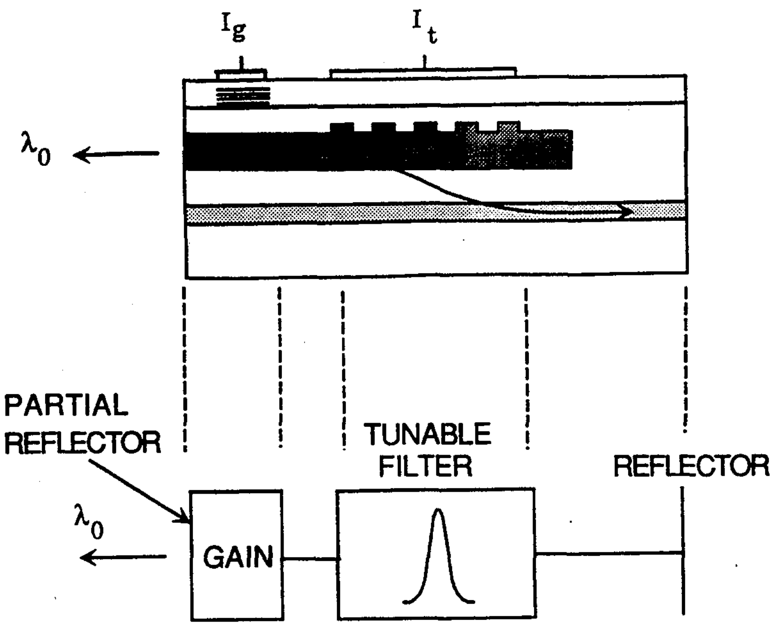
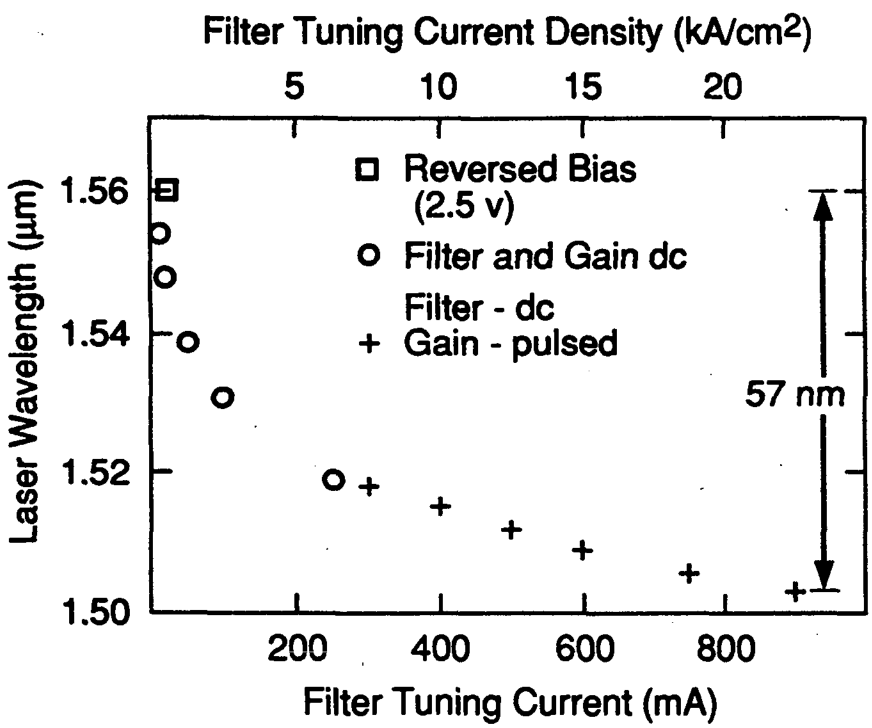
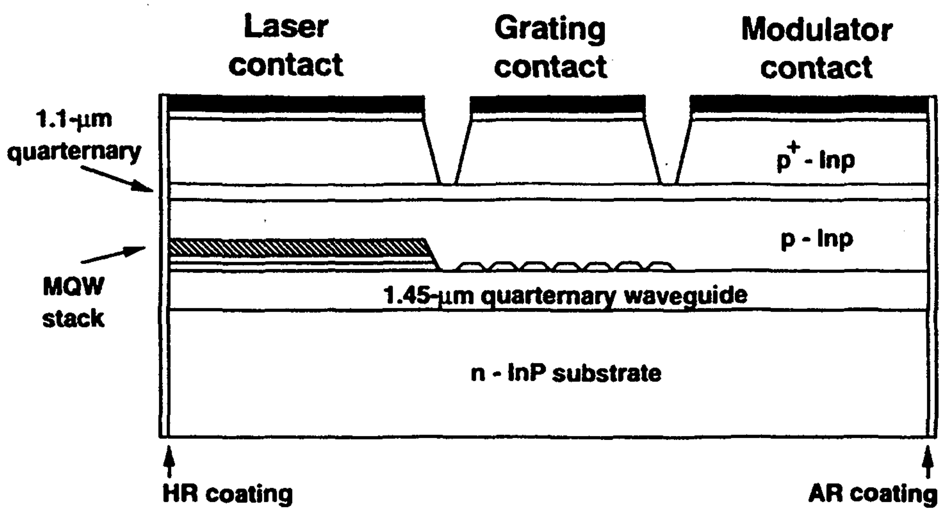


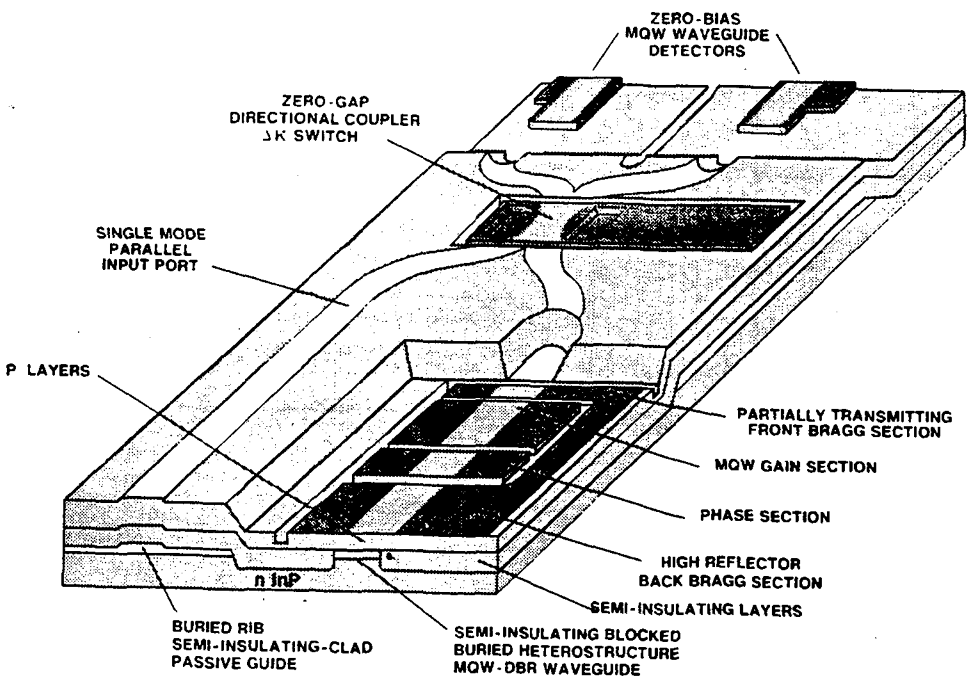

1. Widely tunable vertical coupler-filtered laser. Optical cavity is uses both upper and lower waveguides as indicated by the arrows.




2. Tuning capability of vertical coupler-filtered laser.




3. Integrated tunable DBR laser and electro-absorption modulator.




4. Photonic integrated circuit that combines the components required for a coherent receiver. The local-oscillator laser, which is a tunable DBR laser, a waveguide power combiner and photodetectors are included on the same substrate.

# MEASURING AND EVALUATING VIDEO SIGNALS IN THE HEADEND

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## ABSTRACT

The NCTA and the cities have recently agreed to, among other things, a set of basic measurements to be made on video signals. We'll show you how to make the new measurements, and tell you what they mean. We'll also discuss other basic measurements that should be made in order to provide good signals. The measurement emphasis is on practical things that can be done in the headend, using available equipment augmented with devices that can be built in the system. In looking at various cable systems, we have collected some prime examples of what good video does NOT look like. Baseband video and modulation are reviewed.

## INTRODUCTION

Consumer devices are adding features which emphasize the need for quality in the delivered signals. These added features include larger and sharper pictures viewed from a shorter distance, on-screen displays in VCRs and other equipment, and more recording options. In addition, as this is written, the FCC is about to formalize new rules aimed at providing the subscriber with better service. These new Rules are based on an agreement reached recently between the cities and the cable industry. These rules are good, but you can still do more to improve pictures. Complying with the rules will allow you to put out better pictures, yielding fewer subscriber complaints.

New to cable rules is the requirement to make certain baseband measurements on signals you carry. These measurements may not be familiar to some. We show you ways to make the required measurements, and why they are important. In additional, we will discuss some of the problems we have observed in cable systems.

## THE NEW MEASUREMENTS REQUIRED

The cable industry and the cities have recently reached agreement on, among other things, technical performance standards for cable systems. These are to be incorporated into the FCC Rules in the near future (they may even be in effect by the time this is published). Among the changes are that, for the first time, you will have to make certain basic baseband measurements in your headend. This section seeks to explain the measurements you must make. By no means does this represent a comprehensive set of measurements, and minimum compliance does not necessarily mean that you are supplying impeccable pictures. However, they are a starting point for improving pictures.

A remaining task associated with the agreement is to agree on measurement techniques for the governed parameters. This task is beginning as this is being written. Since we are writing this before the measuring techniques have been agreed, we are taking some risk that we will say something that is not exactly in accord with the final agreement. However, the techniques presented are all standard techniques in use in the broadcast industry, at CATV manufacturers and at some systems. Certainly the principles will remain valid.

## Frequency Response

No measurement on a TV channel is more basic than that of frequency response. Errors can cause picture softness, close ghosting, smearing, weak color and other nasty things that cable TV claims to improve. You are familiar with measuring the RF response of a channel. This is directly related to the baseband frequency response, but the relation is not always simple. In some parts of the spectrum, a big error at RF can translate to a small error at baseband, and at other frequencies a big error at RF can produce a big error at baseband. Since the RF carrier is only something we add to the baseband picture for part of its journey from studio to home, we are primarily interested in the baseband response of a system. Frequency response may be measured in many ways, and this article cannot cover all of them. Rather, we concentrate on one simple means of measuring frequency response. It is accurate enough for most everyday needs and should satisfy the needs of the agreement with the cities.

The technique uses the "multiburst" waveform. This is a simple signal composed of a pedestal (40 IRE in the example - you will see multiburst elements of other amplitudes) on which bursts of different frequencies have been added. By measuring the relative amplitude of each frequency burst, one can get a good idea of the channel response. Figure 1 shows a multiburst waveform. Above it is a spectrum diagram indicating the spectrum occupied by each burst. Intentionally, the spectrum is not shown exactly above the corresponding burst. The X-axis of the waveform is time. The X-axis of the spectrum diagram represents frequency. Each burst in the multiburst occupies a different part of the baseband spectrum, and will be affected by the amplitude response of the channel. By measuring the amplitude of each burst after demodulation, one can ascertain the response of the channel from the point of injection of the multiburst.

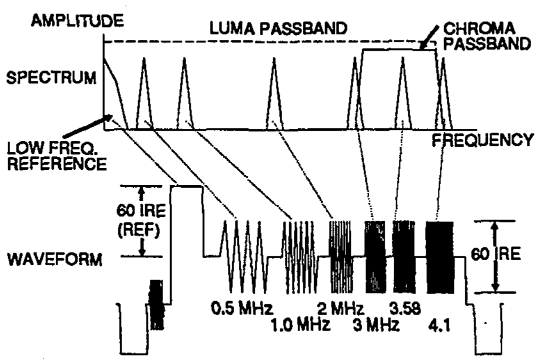



Figure 1. Multiburst Waveform

For example, suppose that we measure the 0.5 MHz burst amplitude to be 60 IRE (the nominal amplitude) and the 1 MHz burst to be 40 IRE. The frequency response of the channel at 1 MHz *compared with 0.5 MHz* is

$$20LOG(40/60) = -3.52 \text{ dB}.$$

We use 20LOG because we are measuring voltage. Frequency response is measured with respect to some reference frequency. Many times that reference frequency is taken to be 0.5 MHz, this being the frequency of the lowest burst. This is not a preferred way to evaluate the channel, however, because it leads us to ignore the channel response below 0.5 MHz. These lower frequencies constitute an important part of the channel, which contains most of the luminance energy. Errors in low frequency response can cause picture streaking, brightness variation from left to right in the picture, and synchronization problems, among others. A preferred measurement technique is to measure response with respect to the reference bar on the left of the figure 1 waveform. This bar also has an amplitude of 60 IRE, the same as that of each burst, measured between peak white and the 40 IRE pedestal on which the bursts are imposed. The energy in the bar occupies the lowest frequency part of the spectrum. We can use this as a reference and get a much more relevant picture of channel response. The easiest way to make the measurement is to set the gain of your oscilloscope to make the bar amplitude 60 IRE as shown, then measure the IRE amplitude of all the bursts, calculating frequency response as shown above.

Notice that we have drawn each burst as occupying a narrow band of frequencies rather than a single point (If we drew this accurately, we would spread out the spectrum even more). This is because each burst can be thought of as a carrier in its own right, 100% amplitude modulated by a rectangular wave that turns it on for a short time every so often. This modulation causes sidebands, as would any other modulation of a carrier. Thus, the spectrum spreads out. Is it difficult to think of the bursts as carriers? The 1 MHz burst is in the middle of the AM broadcast band and 3.58 MHz is near the lower edge of the amateur 80 meter C.W. band.

In practice, you will never see the last burst (between 4.0 and 4.2 MHz, depending upon the generator) look very good after a trip through a band limited channel. This is because the burst is at the very edge of the passband, which theoretically ends at 4.18 MHz in NTSC transmission, but which rarely extends this high (NCTA standards assume a video bandwidth of 4 MHz). The modulation sidebands associated with this burst extend past the channel edge. The result is often a trapezoidal envelope on the last burst, which may not have time to reach its C.W. amplitude. For this reason, frequency response is normally measured only to 3.58 MHz when using the multiburst. Interpretation of the last burst is dangerous and is best left to an expert (most of whom ignore it).

Notice that we show the chroma passband as occupying the spectrum about ±600 KHz from the 3.58 MHz carrier. (Technically, one component of the color extends 1.2 MHz below the color subcarrier, but this extended bandwidth is rarely used in consumer equipment.) The color bandwidth can be smaller than the luminance bandwidth due to the way our eyes perceive color. We perceive sharpness in the luminance signal much more than in the color information. The luminance bandwidth is shown overlapping the chrominance bandwidth. Lower cost TVs cannot recover this bandwidth, but higher cost sets with comb filters can separate the chrominance and luminance, as a result of the way information is interleaved in the spectrum. The details are beyond the scope of this paper.

Figures 2 and 3 show examples of deficient multibursts which we have found on cable systems. (The pictures were taken with a low cost oscilloscope and the clamp/sync separator shown in the appendix.) Figure 2 shows a frequency response roll-off of about 2.9 dB at 3 MHz. This part of the spectrum carries luminance detail information, so the picture will not be quite as sharp as it should be. By the way, we measured the baseband response by using a ruler to measure the heights of the bursts from the photo. We then took 20 LOG the ratio of heights in inches. Measuring in volts or IRE is not necessary: when you are making ratio measurements such as this, it is

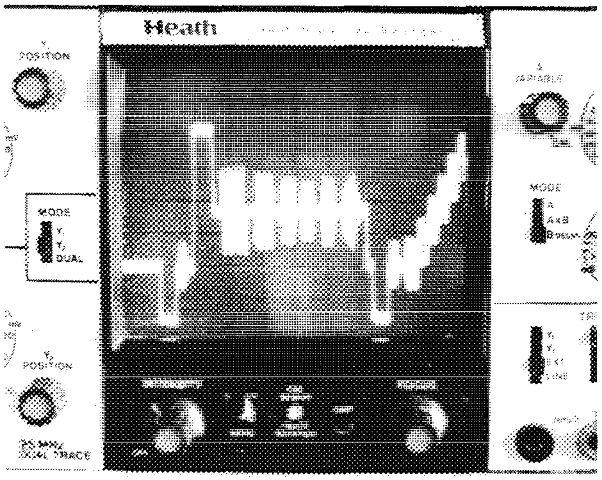


Figure 2. Multiburst With
2.9 dB Mid Frequency Rolloff

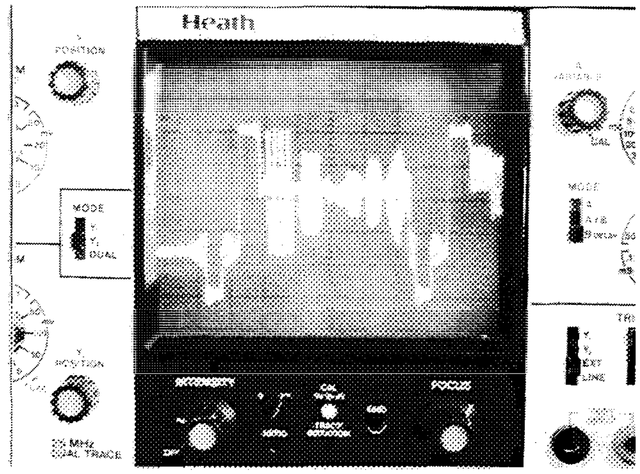


Figure 3. Multiburst on Channel
Protected by Positive Trap

only necessary to make both measurements in the same units.

You can also see from figure 2, the effect of spectrum truncation of the last burst. It has a triangular shape due to the inability of the system to pass the high sideband.

Figure 3 shows the effect of a positive trap. The 2 MHz burst is missing and the higher bursts are grossly reduced in amplitude. This is the reason that channels protected by positive traps have the reputation of being "fuzzy." Also, note that the multiburst in figure 3 is a full amplitude multiburst, meaning that the amplitude of each burst is 100 IRE peak to peak, and so must be measured with respect to the leading edge of the bar rather than with respect to the trailing edge. Transmission of full amplitude bursts is not recommended but is permitted for cable programmers.

## Differential Gain

The amplitude of the color subcarrier determines the saturation, or "purity" of the color on the screen. Adjusting the "COLOR" control of a TV receiver effectively adjusts the amplitude of the chroma signal. One of the important parameters which you are asked to measure in the new FCC Rules, is differential gain. This is a measure of how much the chroma amplitude changes as the luminance level on which it rides changes. To appreciate the importance, consider a picture of a baseball stadium, with green grass on the playing field. Now consider that half of the stadium is in shade and the other half is in sunlight. The "greeness" of the grass is the same in sunlight or shade, but if we have differential gain in the system, the grass in one area will appear greener than in the other.

Figure 4 shows the idea behind measuring differential gain. A test signal is generated on one or more TV lines,

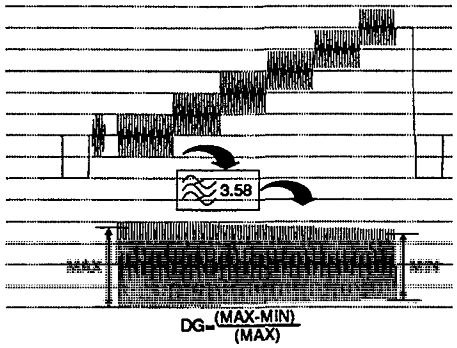


Figure 4. Measurement of Differential Gain

consisting of several (5 or 10) steps of luminance, from 0 to 100 IRE. A gradually rising ramp is also used. Each luminance step has superimposed on it a sample of the color subcarrier, 3.58 MHz. (OK, for those of you who take pleasure in examining things with a micrometer, you caught us: the subcarrier on this and the next figure is drawn as if it were 2 MHz. We just did this so the figure would be a little clearer. Most other figures are literally correct as far as frequencies and durations are concerned.) If the system has differential gain, the amplitude of the subcarrier component will change on different steps. The signal is passed through a 3.58 MHz bandpass filter to eliminate all but the color subcarrier. The change in peak-to-peak amplitude can be measured, and the differential gain computed as shown in the figure. The amount of differential gain shown is 20%, probably equal to the limit negotiated between the cities and the cable industry.[1] By the way, this is a LOT of differential gain: you should be ashamed to have this much in your headend. (Not that it will affect the picture that much, but modern equipment can do much better, and you are not to be forgiven for low quality in your headend. For cost reasons, most of the tolerable distortion must be allocated to the subscriber end.)

Differential gain may be measured on a waveform monitor using the method shown. Vector scopes pass the filtered signal to a detector, and display a line calibrated in percent differential gain.

## Differential Phase

The phase of the chroma subcarrier determines the actual color, or tint. Adjusting the "TINT" control of a TV receiver is analogous to changing the phase of the subcarrier with respect to the burst. The primary operative specification is differential phase. Differential phase is similar to differential gain. Indeed, by sheer coincidence, they often have about the same numerical value. Differential phase is a measure of how much the phase of the signal changes as the luminance changes. In our stadium example, if the system has differential phase, the grass could look green in the shade and blue in the sun! Obviously this is an extreme example, but the idea is that the color of the grass could change between light and dark areas if the system has differential phase.

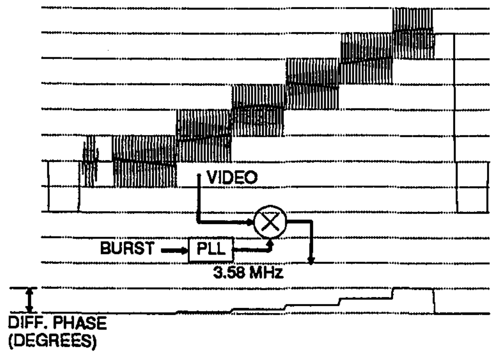



Figure 5. Measurement of Differential Phase

Figure 5 shows the idea behind measuring differential phase. A 3.58 MHz oscillator is phase locked to the burst, and supplies one input to a phase detector. The other input to the phase detector is the video signal itself. Output of the phase detector may be applied to a CRT, and will produce a plot as shown at the bottom of the figure. As the relative phase of the chroma subcarrier changes with different steps, the phase detector output changes. Note that this is not an absolute measurement: one measures the *change* in phase as luminance changes from 0 IRE to 100 IRE. The calibration shown is not absolute, but if one division on the lower part of the scale represents 10 degrees, the differential phase shown is at the agreement limit. Again, you should be able to do better in a headend.

Differential gain must be measured with a vector scope: a waveform monitor cannot measure it.

## Chrominance to Luminance Delay

The final baseband specification to be measured according to the agreement, is chrominance to luminance delay.[2] Refer again to figure 5, which includes the baseband spectrum. Delay is a nasty side effect of filtering which must be done to the RF and baseband signals. Some frequency components in the TV spectrum go through a filter faster than do other components. This can cause what is often termed the "funny paper effect." The name comes from the tendency in funny paper printing, to misalign the three primary colors, each of which requires the paper to pass under a different press. If the paper is not positioned precisely at each pass, the colors are not properly registered. The corresponding situation in a television signal is group delay, a consequence of the picture spectral components at different frequencies getting through at different times. The most obvious problem (and a very practical one) is that the color information will not make it through as quickly as will most of the luminance information. This is often due to the sound trap, which is located only 300 KHz above the theoretical edge of the video passband.

Figure 6 shows a modulated 12.5T

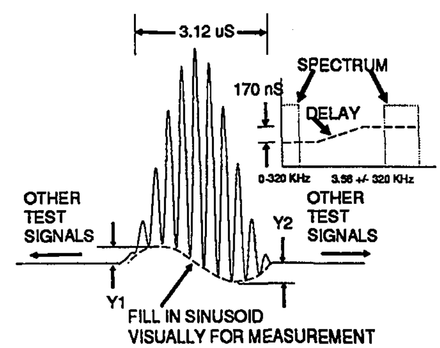



Figure 6. Modulated 12.5T Pulse Showing about 170 nS Delay

pulse used to measure group delay, after the signal has passed through a system having about 170 nS of group delay, the limit specified in the agreement. Later you will compare this with an ideal modulated 12.5T pulse. Of interest in figure 6 is the small spectrum diagram to the right, which shows the situation modelled in developing the figure. The lower frequency video components get through the system faster than do the chrominance components, which are "delayed." By measuring the deviation from a flat baseline, of the bottom of the modulated 12.5T pulse, we can compute both delay and amplitude differences between the luma and chroma components. The modulated 12.5T pulse is on a line of video which also includes a bar, which is set to 100 IRE. Then the positive and negative deviations of the baseline, Y1 and Y2, are measured in IRE units. A nomograph may then be used to read both the amplitude and group delay. The nomograph has been published many places, including the video waveform chart that appeared in Communications Technology magazine in November 1991.

## A Somewhat Intuitive Approach to Group Delay

The other measurements with which we have dealt have been rather more intuitive than is group delay. It will be profitable for us to take a few moments to develop the group delay concept and measurement further.

Chroma to luma delay is usually measured using a modulated 12.5T pulse. "T" is a constant measured in microseconds, related to the bandwidth of the TV system. It is the shortest pulse that can theoretically pass through the system. The NTSC system has a maximum channel frequency response of 4 MHz (more accurately 4.18 MHz, but good luck getting this). Thus, the maximum frequency that can pass is 4 MHz. Now consider the width of the minimum pulse that can produce a white vertical line on the screen. A 4 MHz wave will, in one cycle, produce a white and black line pair. The minimum pulse that can produce a luminance level is thus half of this 4 MHz period.[3] This is the period we call "T." It is equal to one half the reciprocal of the maximum passband frequency of the system. For NTSC, this is

$$T = 1/(2(4\text{MHz})) = 125 \text{ nS}.$$

The *unmodulated* 12.5T pulse has a half amplitude duration (from 50% rising edge to 50% falling edge) of 12.5 times this minimum, and a total duration of twice this. Mathematically, the 12.5T pulse is expressed as

$$v(t) = \sin^2(t/\tau),$$

where:

$v(t)$ = voltage waveform

$t$ = time

$\tau = 2(12.5)T/\pi$, and

$T$ = 125 nS for NTSC.

The upper waveform of figure 7 is

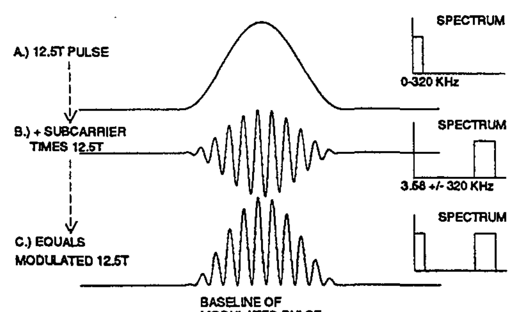


Figure 7. Composition of 12.5T Waveform

this 12.5T pulse, generated from the above formula. It occupies a spectrum from nearly

0 frequency, to 320 KHz. The spectrum occupancy is relatively uniform over this range (though not as uniform as illustrated), which is also where the majority of the luminance energy is located. To make a modulated 12.5T pulse out of this, we use it to amplitude modulate the 3.58 MHz chroma subcarrier. The modulation is done by multiplying the subcarrier by v(*t*) above. This is equivalent to double sideband suppressed carrier amplitude modulation, and produces the waveform and spectrum shown in the middle of figure 7. The spectrum is centered at 3.58 MHz (the "carrier"), and extends 320 KHz either side. This is the spectrum occupied by a majority of the chroma energy (the extremes of the chroma band are further out, but this band includes most of the energy). We can measure the chrominance to luminance delay by combining (adding) the two signal components shown in figure 7a. and b., obtaining the waveform shown in c. This is the ideal modulated 12.5T pulse, having the combined spectrum shown.

We can use this waveform to test for channel amplitude and delay irregularities, as intuitively developed below. We obtain the waveform of figure 7c by adding the waveforms of a and b. The baseline of the modulated pulse is flat because waveform b is of precisely the correct amplitude to fill in waveform a. If the channel response is higher at 3.58 MHz than at low frequencies, the amplitude of b will be greater than it should be to fill in a. It is easy to see that in this case, the baseline of waveform c will "hang down" below the normal flat baseline, with the maximum value in the center of the pulse. The envelope of the baseline describes a cosine function in this case. On the other hand, if the amplitude of waveform b is correct but is delayed with respect to waveform a, then the peak of the 3.58 MHz envelope will arrive too late to properly fill in the center of waveform a. This is the delayed chroma case illustrated in figure 7. The baseline envelope describes a sinusoid. In general, a channel can have both amplitude and delay errors, resulting in a waveform between a sinusoid and a cosinusoid. The response errors can be determined with the aid of the nomograph referenced above.

Figure 6 includes a small spectrum display showing the concept of delay in the frequency domain. The theory behind measuring group delay with a modulated 12.5T pulse is based on the assumption that the delay is flat over the luma bandwidth occupied by the modulated 12.5T pulse. It is also assumed to be flat over the spectrum occupied by the chroma subcarrier, but between the two bands the delay changes. This is a useful approximation, but is hardly real world! In practice, one can often assume that the delay is flat over the luma band. Over the chroma band, the amplitude and delay response are often anything but flat. The effect is to distort the baseline of the 12.5T pulse, so that it is not sinusoidal. Reading the delay in such a case is somewhat questionable, as is the whole concept of group delay. The group delay concept is none-the-less useful, and going further is the subject of future work.

## Getting All of These Test Waveforms

The test waveforms shown in the previous section are all available from any of a number of test signal generators that you might own. However, you don't need (or really want) to use them for most signals, because what you need to measure is the performance of the entire headend signal chain. Fortunately, almost all broadcasters and cable programmers supply these signals free of charge. These and other signals are located in the vertical blanking interval

(VBI) of signals you are carrying now. We call these "Vertical Interval Test Signals," or VITS. To evaluate the overall performance of your system, simply demodulate the output of the channels you are carrying, and look for the VITS in the VBI. You will need a waveform monitor (vector scope required for measuring differential phase) and a good demodulator. Many measurements can be made with an oscilloscope and the circuits described in the appendices.

Your first requirement will be for the best demodulator you can get. Since demodulators will add distortion, it will be worth your while to buy a professional demodulator if possible. Some excellent professional demodulators only tune one channel, having been designed originally for off-air applications not requiring agility. Fixed channel input converters will yield the highest quality input, but make it difficult to test all channels. If your demodulator will accept an IF input at the TV standard IF of 45.75 MHz, you can take the IF output from your modulators and processors, and demodulate it. This will generally allow you to test most of the signal chain, with the exception of the output converter of the modulator or processor. The output converter can affect frequency response but is not likely to have an appreciable effect on delay response. You can sweep the output converter with a conventional sweep set-up to confirm its performance.

Alternatively but not recommended, you can use an RF (never a baseband) set top converter in front of the demodulator, but you must be very careful to confirm the performance of the set top converter. They are not made for measurement applications, and do not have appropriate specifications. If you must use one, you should consult the manufacturer for instructions as to how to sweep it, and pick out the best converter response you can find. After taking into account the response of the converter, it may still be a limiting factor in how well you can measure VBI signals. Caution: the response of a converter will not be the same on every channel. Some of the photographs in this article were taken using S-A RF set top converters and 6250 demodulators.

If you can't get anything better, you will still be able to get some information from the baseband output of a TV or VCR. In this case, unless you have some way to independently test the performance, your ability to make measurements will be very limited, and you may well not be able to measure to the limits prescribed in the agreement between the cable industry and the cities. However, you will be able to get some idea of the performance of your signals in the video blanking interval, and you should be able to check depth of modulation with the calibrator we'll tell you about later.

In order to measure baseband characteristics, you ideally will have a waveform monitor and vector scope. At this time, we don't know of any way to measure differential phase except by using a vector scope. Lacking a waveform monitor, you can do most tests using any good dc coupled oscilloscope having a video response of 10 MHz or more. See appendix B for more information.

Thus, the minimum set-up for observing baseband signals is a TV or VCR with a baseband output, an oscilloscope and the sync separator circuit here-in described. This will be enough to make some important measurements, though not all those required by the agreement. A better set-up is a professional quality demodulator with agile front-end (or using IF interface as described above), with a waveform monitor and vector scope. Automated test indicators are also available, but are very expensive and have their own set of limitations in some situations.

## IMPAIRMENTS WE HAVE SEEN

### Depth of Modulation

While not specified in the agreement, video depth of modulation (DOM) is a most important parameter, and we have found that some systems are not good about maintaining it. Low depth of modulation will cause pictures to look dark, and can sometimes cause sync circuits to work poorly. With some scrambling systems, incorrect depth of modulation will cause even more severe problems. High depth of modulation will cause light areas of the picture to wash out, and will cause an audio buzz in some sets.

The FCC specifies that depth of

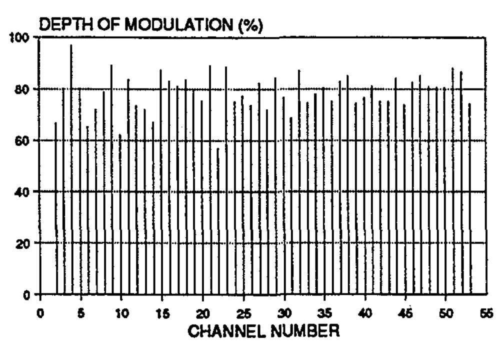



Figure 8. DOM of All Channels Measured on One Cable System

modulation should be set to 87.5% for optimum picture quality. One would expect that the DOM on the majority of channels would be close. However, this turned out not to be true at two cable systems measured. Figure 8 shows the result of DOM measured on one of the cable systems. The system offered 53 channels. The average DOM of all 53 channels measured was 81%. The lowest DOM measured was 65% and the highest DOM measured was 94%. The standard deviation was 6.5%. It was interesting to note that even on several of the premium channels, DOM was below 80%.

Depth of Modulation is a measurement of percent modulation. It is the ratio of an RF carrier's amplitude change during peak white modulation, to the maximum amplitude. Figure 9 illustrates DOM of a video modulated carrier. For a

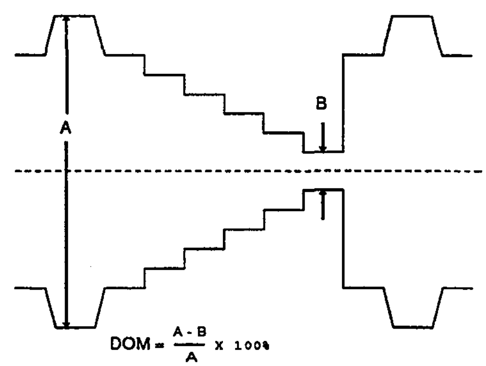



Figure 9. Depth of Modulation of a Video Carrier

discussion of what constitutes a video signal and the definition of various levels, see appendix A.

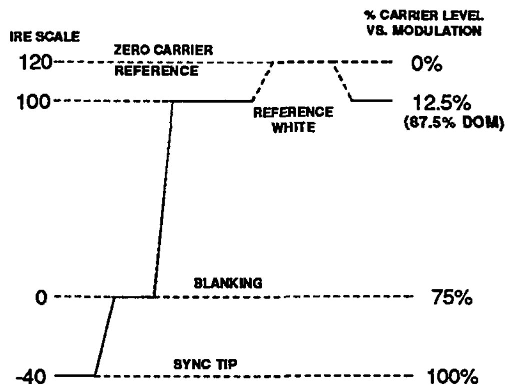


Figure 10. Relationship Between Video Level and Carrier

In an NTSC video system, peak carrier is reached during the sync tips, -40 IRE. During this time 100% of the video carrier is transmitted. Horizontal and vertical blanking, 0 IRE, result in 75% of the RF carrier being transmitted. White level, 100 IRE, produces 12.5% of the available video carrier. Figure 10 illustrates the NTSC depth of modulation.

## DOM Measurement Methods

The three most common methods for measuring DOM are: A) spectrum analyzer, B) demodulator with zero chop, and C) calibrated video demodulator. The method one uses is primarily a function of available equipment, knowledge of measurement technique, or personal preference.

### A. Spectrum Analyzer

Depending on the type of analyzer used, analog or digital, there are several variations and limitations regarding the spectrum analyzer's settings. With an analog spectrum analyzer, DOM can be monitored either at a video field rate or horizontal line rate. With a digital spectrum analyzer DOM becomes very difficult to monitor at video line rates. Regardless of the type of analyzer used, the spectrum analyzer must have an IF bandwidth of at least 300 KHz, zero frequency span capability, and a linear display scale mode or calibrated decibel vertical axis. Video triggering is also required if sweeping at a horizontal line rate.

To measure depth of modulation:

1. Connect the modulated RF or IF output from the modulator under test to the spectrum analyzer's input.

2a. Configure the analog spectrum analyzer as follows for displaying a full video field:

Bandwidth:300 KHz min.
3 MHz preferred
Frequency Span:Zero span
Trigger:A.C. Line
Scan Time:200 ms
Video Filtering:None,
1 MHz preferred

2b. Configure the analog or digital spectrum analyzer as follows for displaying several video lines.

Bandwidth:300 KHz min.
3 MHz preferred
Frequency Span:Zero span
Trigger:Video
Scan Time:200 8 µs
Video Filtering:None,
1 MHz preferred

3. Tune the analyzer to the output frequency of the modulator under test. The displayed waveform represents the peak detected video signal. Fine tune the

analyzer's center frequency control to maximize the amplitude of the detected video signal.

4. Use the linear amplitude display, positioning the sync tips on the top most

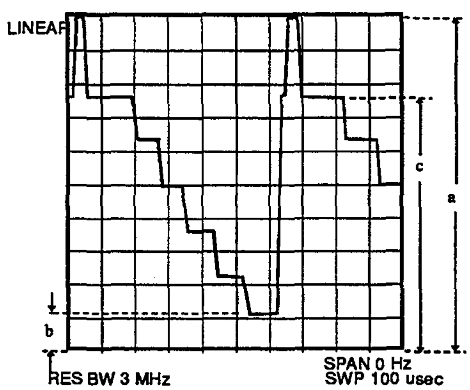



Figure 11. Measurement of Depth of Modulation Using A Spectrum Analyzer

graticule, as shown in figure 10, by adjusting the amplitude sensitivity (often called *reference level*). This calibration sets the top most graticule to 100% modulation and the bottom most graticule to 0% modulation. Percent modulation of the displayed waveform can now be expressed as follows:

$$\%DOM=\frac{a-b}{a}\times 100\%$$

Where:

"a" is equal to the total number of major displayable vertical graticules.

"b" is equal to the number of major graticules the 100 IRE signal is displaced from the bottom most graticule.

For example, if the RF carrier is modulated by a 5 step linearity signal such as that shown in figure 11 and an analyzer display of 10 major graticules, 87.5% DOM would place the 100 IRE step at 1.25 major divisions from the bottom most graticule.

Unfortunately, when observing DOM with active video signals, a 100 IRE reference pulse is not always present. In this situation DOM can be monitored by observing the blanking level with respect to sync tips. The percent of modulation referenced to blanking, assuming no sync compression from the headend, can be expressed as follows:

$$\%DOM=3.5\times\frac{a-c}{a}\times 100\%$$

Where:

"a" is equal to the total number of major displayable vertical graticules.

"c" is equal to the number of major graticules the 0 IRE, blanking, signal is displaced from the bottom most graticule.

3.5 is the ratio of peak to peak video with respect sync tip amplitude.

An even easier way to set up the spectrum analyzer display is to position the sync tips 8 divisions up from the bottom. The peak white should then just reach 1 division from the bottom. Sync tips should be 6 division from the bottom.

B. Zero Chop

In order to measure depth of modulation using the zero chop method, two references must be established. One reference corresponds to the transmitted peak carrier level, 100% of available carrier, and the other corresponds to zero, 0% of available carrier. By establishing these two references and knowing that a standard NTSC video signal has an amplitude of 140 IRE units (40 IRE units from sync tip to blanking level, 100 IRE units from blanking to peak white, and 160 IRE units from peak carrier to zero carrier) DOM can easily be determined by taking the ratio of the 100 IRE signal amplitude (measured from sync tip) with respect to the amplitude of the zero chop signal (measured from sync tip).

A zero carrier reference cannot be

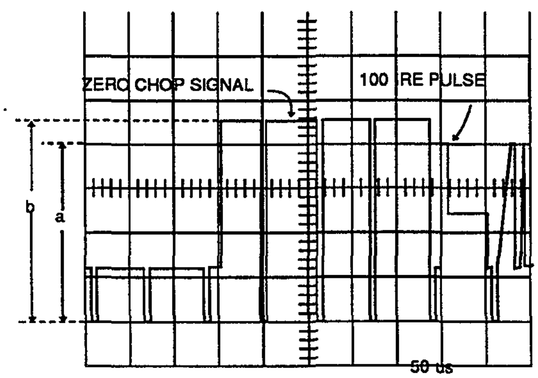



Figure 12. Zero Chop Pulse Produced by a Professional Demodulator

transmitted from the modulator without causing severe system problems. Most professional video demodulators have the ability to switch off the IF signal for a few microseconds during the vertical interval. This simulates a situation in which the transmitted signal is 100% modulated. Figure 12 shows the zero chop signal generated by an S-A 6250 demodulator.

To measure depth of modulation:

1. Tune or set the input converter of the video demodulator to the output frequency of modulator under test.

2. Switch the demodulator's zero chop function on.

3. If incidental carrier phase modulation is present, set the demodulator to envelopment detection. If not, synchronous detection often gives more accurate results.

4. Display the output of the video demodulator on a video waveform monitor or oscilloscope. If using an oscilloscope without TV triggering, use the TV triggering circuit described in this paper.

5. By using a combination of the

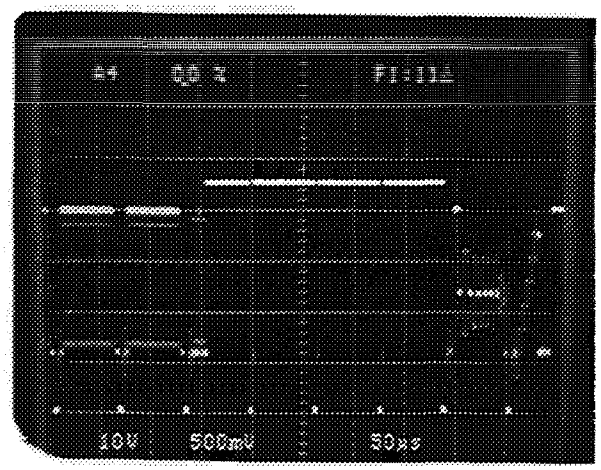


Figure 13. Zero Chop Display

demodulator's signal output level adjustment and the oscilloscope's vertical gain adjustments, set the amplitude of the video signal to 160 arbitrary units from sync tip to peak of zero chop pulse. Figure 13 illustrates this setup by showing a portion of the vertical blanking interval with an inserted zero chop signal. The 100 IRE signal identified in the illustration is a portion of the VITS reference for the mulitiburst signal. DOM can now be determined using the 100 IRE bar reference and zero chop signal as follows:

$$\%DOM=\frac{a}{b}\times100\%$$

Where:
"a" is equal to the amplitude of the 100 IRE signal.

"b" is equal to the amplitude of the zero chop signal.

If VITS or a 100 IRE signal in not available, DOM can be measured using the amplitude of the sync pulse as follows:

$$\%DOM=3.5\times\frac{c}{b}\times100\%$$

Where:
"c" is equal to the amplitude of the sync pulse.

"b" is equal to the amplitude of the zero chop signal.

3.5 is the ratio of peak to peak video (inclusive of sync) with respect to sync amplitude.

C. Calibrated Video Demodulator

Video depth of modulation can also be determined by using the calibrated video demodulator technique. This method is primarily used when a zero chop demodulator is not available. It is also an excellent method for measuring DOM when using a VCR with a baseband video output. An advantage of using a VCR with a baseband video output, is that it can double as an agile demodulator.

As the method's name implies, the demodulator needs to be calibrated. Calibration is accomplished by measuring the output of the demodulator when supplying a known input signal. Once the output of the demodulator is calibrated, DOM can easily be determined by taking the ratio of the amplitudes of the unknown signal with respect to the calibrated signal. The calibrated input signal can either be from a video source and modulator pair or from the demodulator calibrator circuit described in appendix C. If using the video source and modulator pair, the modulator should be set to 87.5% DOM using the spectrum analyzer method or zero chop method. If using the demodulator calibrator circuit, the output signal simulates a signal with 87.5% DOM. This is accomplished by attenuating the input carrier by 18.06 dB at the horizontal line rate. The attenuation level of 18.06 dB was derived from the following equation:

$$attn\ level(dB)=20\log(1-\frac{\%DOM}{100})$$

Where:

"%DOM" is equal to 87.5%

To measure depth of modulation:

1.Configure the test setup as shown in figure 14.

2.If using the demodulator calibrator circuit, verify that there is no video modulation on the output from the channel 3 modulator.

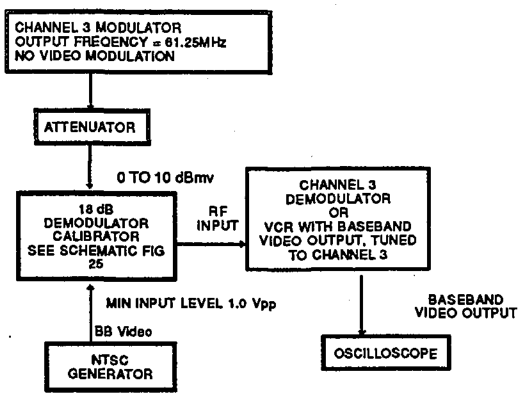



Figure 14. Calibrated Demodulator Set-up

3. If using the video generator and modulator pair, verify that the modulator is calibrated to 87.5% DOM.

4. To minimize the effects of incidental carrier phase modulation distortion, set the demodulator to envelopment detection if necessary (though you may not get a choice).

5. Display the output of the video demodulator on a video waveform monitor or oscilloscope. If using an oscilloscope without TV triggering, use the TV triggering circuit described in this paper.

6. For maximum accuracy, adjust the oscilloscope's gain control for maximum displayable peak to peak signal.

7. Measure and record the peak to peak amplitude of the displayed signal, $V_{\text{cal signal peak to peak}}$. The amplitude of this signal represents this modulator's output amplitude for a signal with 87.5% DOM.

8. Remove the demodulator calibrator and connect the output of the modulator under test to the input of the demodulator.

9. Tune the demodulator to the output frequency of modulator under test.

10. Adjust the oscilloscope's delayed trigger control to view the 100 IRE pulse located in the vertical interval test signal.

11. After measuring the amplitude of the 100 IRE pulse, $V_{\text{100 IRE peak to peak}}$, DOM can be determined as follows:

$$\%DOM=\frac{87.5\times V_{(p-p)}}{V_{(cal\ p-p)}}$$

12. Similar to the spectrum analyzer method and zero chop method, if a 100 IRE pulse is not available DOM can be determined by measuring the amplitude of the sync pulses. The following equation is used:

$$\%DOM=3.5\times 87.5\times\frac{V_{(0IRE\ p-p)}}{V_{(cal\ p-p)}}$$

## Vertical Blanking Interval Problems

The only video signals that are permitted to descend below blanking, 0 IRE, are the color burst and synchronizing pulses. Some dark colors may also cause the color subcarrier to go below 0 IRE during active video. All other signals descending below blanking are improper and may cause trouble. As previously mentioned, if any portion of the video signal were to cause grief it would likely be the vertical interval. Unfortunately, one cable system demonstrated several good examples of

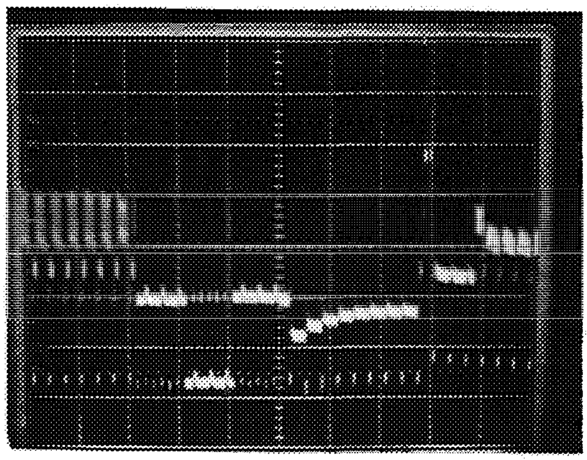


Figure 15. Descending Vertical Blanking

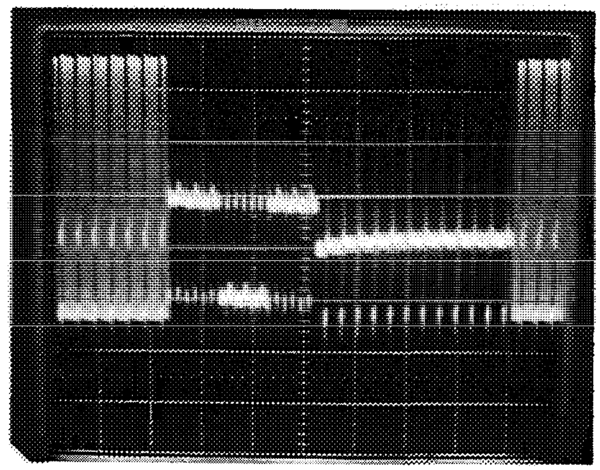


Figure 16. Shift of Sync Tips in Vertical Blanking Interval

illegal video. Figure 15 shows a good example of a problem seen in the vertical interval, in which the vertical interval blanking signals are descending below normal blanking level. Notice the nearly 20 IRE of blanking droop starting immediately after the post equalizing interval. This descent exponentially decreases to 0 IRE after about 10 lines. Notice also that the sync pulses after this period, beginning at about line 17, are raised. Problems of this type may not be noticed during casual television viewing. However, descending vertical blanking could manifest itself in professional and consumer equipment that relies on the integrity of the vertical interval. The most likely cause of the problem is a defect in baseband equipment. In one case it was traced to a defective VITS generator at the uplink, and in another case it was due to a defective character generator at the headend.

Another problem seen in the vertical interval was the shifting of many line of sync. Here the pre equalizing, vertical sync pulse, and post equalizing intervals were shifted above the horizontal sync pulses by 10 IRE. Figure 16 shows an example of sync shifting. A problem of this nature also has the potential of causing problems with vertical sync separators in VCR's, TV's, and closed caption equipment. It was again traced to a defective character generator at the headend. Also notice the extremely high chroma. This channel had lots of problems!

## Sync Compression

This is a common problem with lower cost and consumer equipment, which can manifest itself as unstable sync on some TV sets, poor character generator performance and poor video tape results. The FCC specifies that the amplitude of synchronizing pulses should be 40 IRE. Anything less than 40 IRE is considered to be compressed. Sync compression is measured by taking the ratio of blanking to peak white, divided by blanking to sync tip. This ratio should equal 2.5.

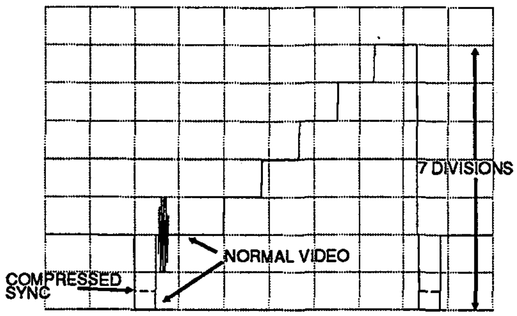



Figure 17. Sync Compression

Figure 17 illustrates sync compression. The easiest way to check this is to adjust the vertical position and gain on your oscilloscope until the blanking interval is 2 divisions from the bottom, and the peak white is 7 divisions from the bottom. If the sync is not compressed, the sync tips will be on the bottom graticule as shown. Compression will result in higher sync tips.

One source of sync compression is the headend modulator. Compression in the modulator could be a result of poor clamping or output amplifier nonlinearity. Another less controllable source of sync compression is from local UHF and VHF transmitters. The high output power from these transmitters has been know to compress sync pulses. We encountered one such example a few months ago, at a UHF PBS affiliate. A call to the station's engineer indicated that he was operating on one Klystron rather than the three normally used. The station was installing a new transmitter, and he couldn't get replacement klystrons for the month until the new transmitter was ready. He was amazed that we weren't measuring even more severe problems! If your headend is processing a signal with sync compression there is nothing to be done short of correcting this problem at the transmitter. If converting to baseband then remodulating, a video processor should do the trick.

## Descending Characters

Most headends today have the capability of inserting on-screen characters on some channels. For example, the Weather Channel allows local cable operators the option of inserting local weather forecasts and current conditions using on-screen characters. The problem with overlaying locally inserted on-screen characters is that the incoming video signal level may not match the inserted on-screen character levels. One manifestation is that the black level boarding the characters will drop below the input video's black level. When this occurs, video equipment that relies on sync pulses becomes confused when it sees additional pulses dropping below the black level. Fortunately, this problem is usually easily corrected by monitoring the video signal at the baseband input to the modulator while making the appropriate adjustments to the on-screen inserting equipment.

## Audio Deviation

Similar to video depth of modulation, headend audio deviation is not one of the proposed operational standards. However, inconsistent deviation is the source of many subscriber frustrations and complaints. Deviation is a measure of the loudness of a signal. Figure 18 shows the relative detected audio levels measured on one cable system. The relative audio deviation was computed by first measuring the detected audio level on all channels using a VCR's vu meter. (By the way, "vu" stands for *volume units*, not "view" or some of the other strange

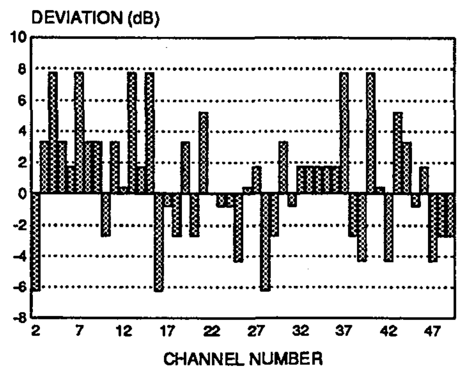



Figure 18. Audio Deviation as a Function of Channel

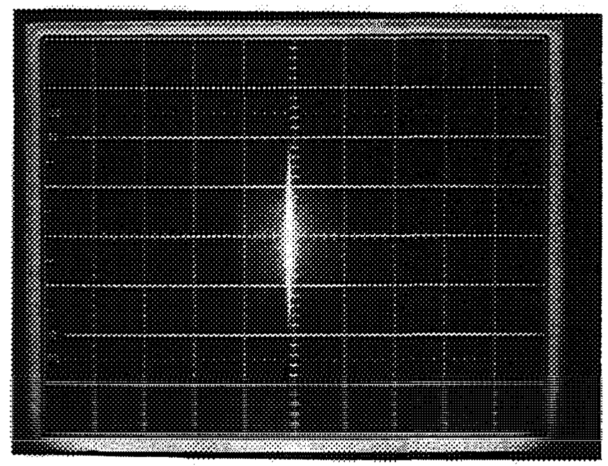


Figure 19. Peak Audio Deviation Displayed on an Oscilloscope Having No X-Deflection

terminology sometimes applied.) Since typical audio programs vary in level the average meter reading of the channel being measured was recorded. After recording all channels, the average reading of all channels was determined. This average was then used as a reference to compute the relative audio deviation on each channel. The horizontal line on the chart indicates the average level of the three network stations.

Notice the wide variation in relative audio deviation across the 50 channels measured. This type of variation can become annoying when scanning the dial. Subscribers are frequently adjusting the TV or converter volume control for the desired audio level. Today's television viewers have become more discriminating to audio quality. CD players, hi-fi VCR's, BTSC stereo, and digital music on cable has sensitized many subscribers to high quality audio. Therefore, it is imperative that cable operators maintain high quality audio through-out the headend.

Audio levels can be evaluated several ways, average, peak, peak factor, and loudness. Average refers to averaging the amplitude of the audio signal over a period of time (RMS). This is usually accomplished with the use of a VU meter.

Peak is an instantaneous measurement of the audio signal's peak amplitude. A peak program meter is usually used to make this measurement. However, an oscilloscope can also be used with adequate accuracy. Figure 19 shows a sample of audio metering using an oscilloscope with the sweep speed set to XY and no deflection horizontally. The meter is first calibrated by connecting it to the output of a demodulator tuned directly to an off-air signal. Broadcasters are usually pretty good about maintaining the correct peak deviation, so you can use a broadcast signal to calibrate your modulation indicator if you have nothing better available.

Peak factor refers to the ratio

between the peak voltage and the RMS voltage in an audio signal. Measuring peak factor requires a specialized piece of equipment called an
audio deviation meter.[4] The final method of monitoring deviation is loudness measurement. The human ear's ability to sense loudness is very similar to measuring the RMS value of an audio signal. This is true because the human ear perceives loudness as a power derived factor. Unfortunately, the human ear's poor loudness memory makes it very difficult to set deviation accurately.

Figure 20. Time Elapsed Oscillograph Tracing of Deviation

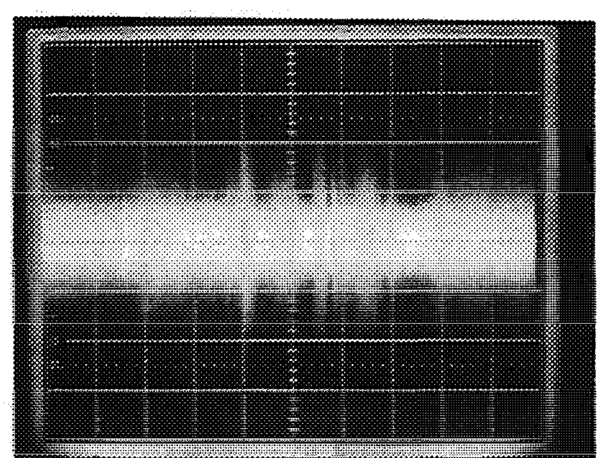
As previously mentioned, an oscilloscope can be used to balance the audio deviation in a headend. Basically, this is done by monitoring the demodulated audio signal using an oscilloscope with its horizontal sweep set to 500 ms or higher. The vertical trace, similar to the one shown in figure 19, sweeping across the oscilloscope is the audio signal. The peak to peak excursions represents peak to peak deviation. Figure 20 shows deviation v.s. time for about 3 seconds of audio. The audio deviation on all channels can now be set to some predetermined reference peak level. Since off-air network channels usually have their volume levels carefully monitored, they should be used to establish this reference.

Using an oscilloscope calibrated against a local broadcaster may not be the perfect method to set deviation, but is a lot better than what some systems are doing now. The peak flasher on many modulators may also be used, but is really intended more as an alarm to indicate that the legal deviation is being exceeded. The NCTA engineering committee is painfully aware of many subscriber complaints of poor audio consistency, and is working with program suppliers to do something about it.[5] Hopefully this fall, a reference tone will be transmitter once a week from one or more program suppliers, which will allow you to set deviation on all modulators supplied from VideoCypher decoders. We urge you to take advantage of this test tone when it becomes available.

## ACKNOWLEDGEMENTS

The authors wish to acknowledge the help of the few cable systems who unwittingly provided a plethora of examples of what not to do, for us to write about.

Brenda Roberts constructed and tested the circuits shown and took many of the photographs.



## APPENDIX A. A REVIEW OF VIDEO BASICS

### The Picture Composition

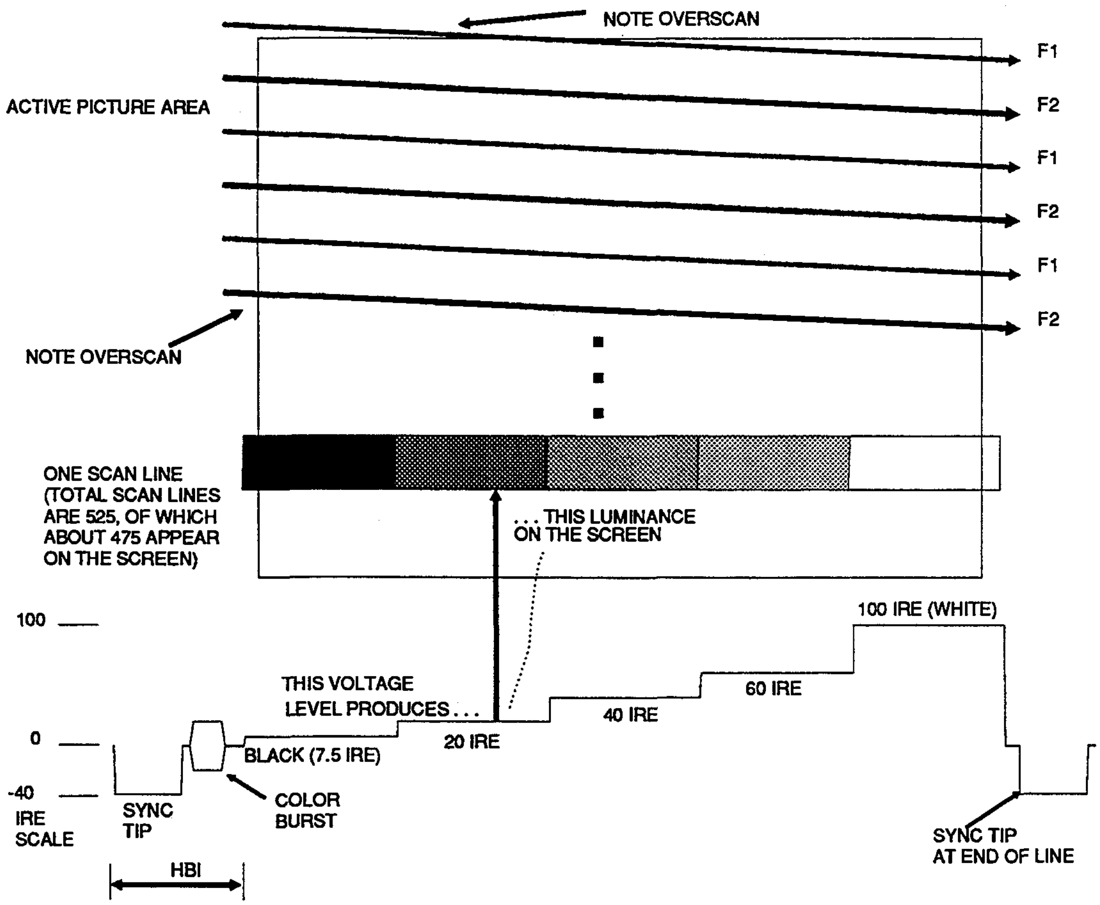



Figure 21. Composition of TV Picture

The picture is painted on a TV screen one line at a time, moving from left to right and top to bottom of the screen - the same trajectory our eyes follow as we read. A complete picture consists of 525 lines, of which, at most 475 actually appear on the TV screen. The complete picture is "painted" about 30 times a second, and consists of two halves, each half "interlaced," or occupying every other line.[6] The complete picture is called a *frame* (similar to the motion picture use of the term), with each half picture being called a *field*. Unimaginatively, each line of the picture, is called a *line*. This is illustrated in figure 21, which shows the active picture area of a TV set in the proper 4:3 ratio (four units wide to three units high). Scanning starts above the active picture area in field 1, and proceeds from left to right as illustrated by the lighter lines labeled "F1." When the electron beam reaches the bottom of the picture tube, it resets to the top and begins painting, or *scanning*, the in-between lines which, form field 2.

Intentionally, the lines reach beyond the left and right edges of the active picture area, as well as above and below. This is done so that under almost all conditions the picture will still take up the entire screen. Most scrambling systems take advantage of this overscan, as do a number of VBI services.

Thus, we have the following composition of the picture.

1 complete picture = 1 frame
1 frame = 2 interlaced fields
each frame = 525 lines

We can compute some important parameters of the picture from this. We said that we painted about 30 frames a second - the actual frame rate is 29.97 frame per second. We have twice as many fields as frames, so the field rate is 59.94 fields per second.[7] Since we have 29.97 frames per second and 525 lines in each, we have a line rate of 29.97 times 525, or 15.734 KHz. The time or period of a horizontal line is 1/15734 = 63.56 µS. This time is often referred to as the H-time, or simply H. You will often see times in video referenced with respect to H. For example, the width of a sync tip is often stated as 0.075H, or 0.075 times 63.56 = 4.767 µS (we frequently round down to 4.7 µS). These are all important numbers which should be committed to memory.[8]

## Horizontal Synchronization

Obviously, if the TV is going to paint the same picture seen by the camera, it must scan in synchronization with the camera. When the camera beam resets left to right or top to bottom, the TV must do the same. How does the TV know when to do this? Synchronization ("sync") information is carried with the TV signal. Horizontal sync tells the TV that its electron beam should be at the right edge and should be resetting to the left. Vertical sync does the same thing top to bottom.

Figure 21 shows a line of TV signal properly positioned with respect to the screen. Sync tips occupy a unique voltage level in the picture. As the baseband TV signal is normally presented, this is the most negative voltage level in the picture. Horizontal sync tips last for 4.7 µS (microseconds). When sync arrives, the TV makes sure that the electron beam is just off the screen to the right, and that the beam resets to the left. Thus, the sync tip shown to the left is really the same as the sync tip on the right, but one line earlier. The time from just before the sync tip begins to just before active video begins, is called the "horizontal blanking interval," or HBI. The other primary feature of the HBI is the color burst, with which we shall deal presently. To an engineer this is one of the most exciting parts of the television picture.

## Luminance

During the active video line, voltage levels represent brightness of the picture. The higher the voltage level the brighter the picture at that spot. We illustrate one line on the screen, which consists of 5 shades of gray, from black to white. The higher the voltage level the brighter the picture at that spot. Note that we are only talking about the black and white picture now - we'll add color later. The black and white picture information is known as the *luminance*, or sometimes, *luma*. It is measured on a scale called the "IRE" scale.[9] The starting point of the IRE scale is 0 IRE, which is the blanking level. Black level is defined to be 7.5 IRE[10], and white is 100 IRE. Sync tips are at -40 IRE. This IRE scale can be related to voltage, in that video is normally at a voltage of 1 volt peak to peak, which comprises the sync tip to peak white, or 140 IRE. Thus, one IRE is normally 1/140=7.14286 millivolts. Note that 0 volts is not generally 0 IRE because video is often interfaced using ac coupling. The actual 0 volt level will be a different IRE level depending on the complete make-up of the signal.

## Vertical Synchronization

You will sometimes (including in this paper) see the signal drawn inverted from that shown. For example, in the FCC Rules, the waveform is drawn upside down from this. Doing so makes reasonably good sense when modulation of the waveform is discussed. You should be able to recognize the signal regardless of the polarity shown.

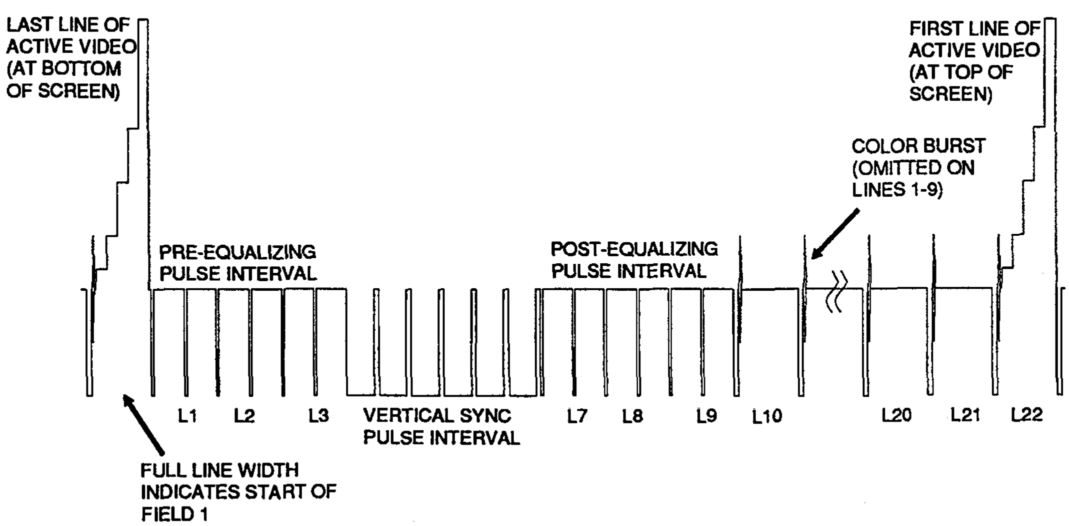



Figure 22. Vertical Blanking Interval

We have seen how the picture is painted on the screen and how horizontal synchronization is achieved. Remaining is to see how vertical synchronization is achieved. Figure 22 shows the vertical blanking interval, or VBI. This corresponds in function to the HBI, in that the vertical sync tells the TV that the beam should be at the bottom of the picture and should now reset to the top. Figure 22 is on a much different time scale than is figure 21. One line of the horizontal signal shown in figure 22 occupies only a small portion of figure 21 (see for example, L22 (line 22), in figure 21, which bears the same video signal as did the illustrated line in figure 21). The vertical blanking interval consists of several features, the most significant of which is the three lines of vertical sync. The vertical sync signal spends almost all of its time at -40 IRE, and this is the way that the TV recovers vertical: it looks for a signal staying at -40 IRE for a long time. The first 9 lines of the field, beginning 3 lines before the vertical sync, are "serrated," or split into two halves separated by a brief transition between sync level (-40 IRE) and blanking (0 IRE).[11]

Notice that the average voltage during the VBI is much lower than during the rest of the signal. This often causes grief for equipment handling the signal. We show an example in this text, of a very distorted VBI on a real cable system. If you have a channel that seems to jitter up and down on some TVs, or if you have complaints about VCR or on screen display problems, look for a problem in the VBI.

The reason for at least 11 (usually more) non video lines after the post equalizing interval, is that the electron beam in the TV, being deflected by magnetic circuits, requires a finite time to retrace from the bottom to the top of the picture tube. The TV signal waits around doing nothing (with video) for a while, waiting for the retrace. This retrace time has recently become one of the most popular parts of a TV signal. It is used to transmit test signals ("VITS," or vertical interval test signals) and VIRS, or vertical interval reference signals. It is also used to transmit closed captioning information (standardized on line 21 of field 1), and is often used to transmit other data. The VBI is at the heart of a great controversy now: does this time belong to the TV broadcaster, the cable operator, or someone else? For example, a PBS subsidiary sells this time to people who want to distribute data. Other broadcasters and cable programmers do the same. Some operators are considering asking for additional money to carry these signals, which generally are not part of the TV program. Some scrambling systems also use the VBI to transmit control and/or addressing information.

## Adding Color

The complete process of adding color to the signal is beyond the scope of this paper. Here we will delve into the subject only far enough to relate to measurements we need to make. Color is transmitted as a subcarrier on the main TV signal. The frequency is about 3.58 MHz. Color information is carried in both the amplitude of the color subcarrier and in its phase.

Figure 23 shows part of a TV line, magnified so we can examine the details related to color. The reference at the top of the picture shows the portion of the line that we are magnifying (we modified the first luma level a bit compared with figure 21, from 7.5 to 20 IRE). The color burst consists of a minimum of 8 cycles of the color subcarrier. Color information in active video rides on this subcarrier. Two characteristics of color apply: color saturation and tint. Saturation is the degree to which a color is "pure," or not tainted with white. As you adjust the "color" control on many TVs you are adjusting saturation. Tint is determined by comparing the phase of the subcarrier with that of the burst. When you adjust the tint control of a TV you are effectively adjusting this relative

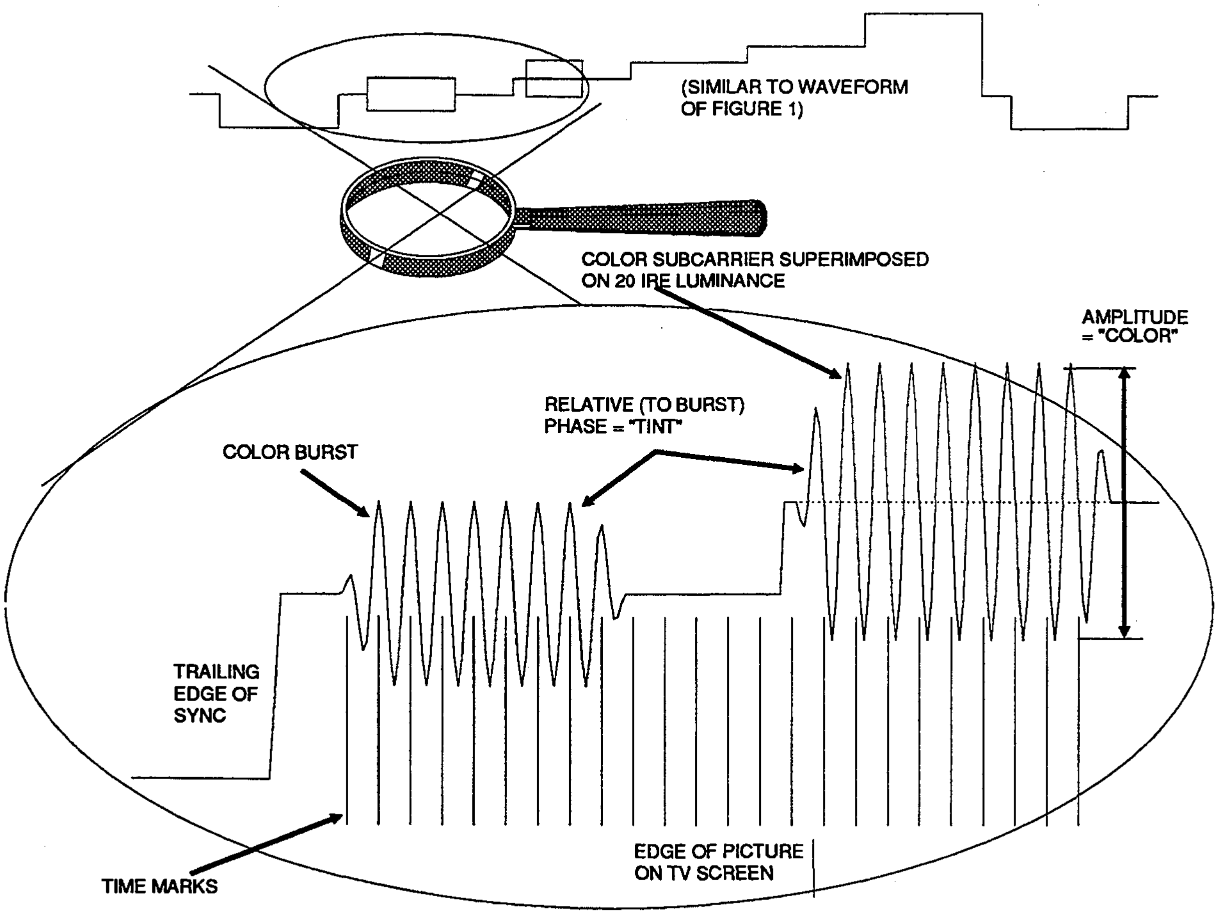


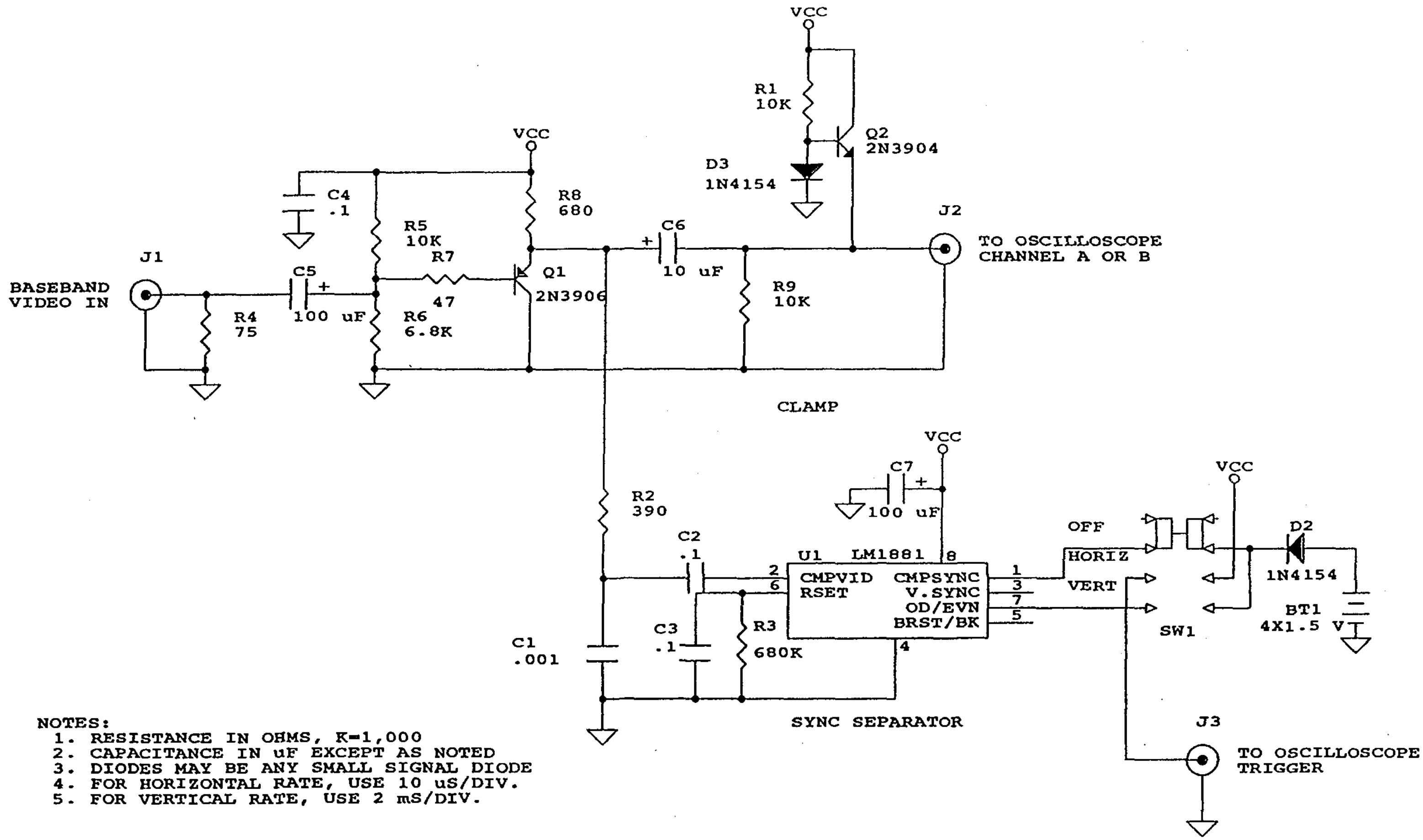

Figure 23. Beginning Part of a TV Line, Showing Color Information

phase. We have added time marks spaced one subcarrier cycle apart, so you can see that the phase of the chroma (color subcarrier) during active video is shifted with respect to the burst - note that the bottom of the color subcarrier aligns differently with the time marks during the burst and during the active video.

## APPENDIX B A SIMPLE CLAMP AND SYNC SEPARATOR

We will now show you a couple of simple circuits that may prove useful in making measurement. We do so at the risk of giving the incorrect impression that this is all you need. *It is not.* These, with an oscilloscope and simple demodulator, will allow you to measure certain important parameters. Rather, they will allow you to get an *indication* of the measurement. These techniques are no substitute for good, professional equipment. If you can get such, please do so. We present these ideas because we know that not every system can get the proper equipment. Intelligent use of this kind of equipment should be better than nothing but, we repeat, it is not nearly as good as getting the proper equipment and doing the job right.

One of the problems you will have making baseband measurements without a waveform monitor, is that an oscilloscope does not have a clamp, or dc restorer, to allow you to see a signal without getting




NOTES:
1. RESISTANCE IN OHMS, K=1,000
2. CAPACITANCE IN uF EXCEPT AS NOTED
3. DIODES MAY BE ANY SMALL SIGNAL DIODE
4. FOR HORIZONTAL RATE, USE 10 uS/DIV.
5. FOR VERTICAL RATE, USE 2 mS/DIV.

Figure 24. A Simple Clamp and Sync Separator for use with Oscilloscopes

vertical bounce as the program material changes. Also, many low cost oscilloscopes have sync separators, but don't have the ability to distinguish between fields one and two, yielding an almost impossible display when you try to look at field rate signals. Figure 24 shows a schematic of a simple clamp and sync separator you can build to use with an available oscilloscope. Your scope will need to be dc coupled and have a bandwidth of at least 10 MHz. In order to observe the VITS you will need a delayed trigger time base. Many scopes in the $500.00 range these days have these characteristics. A convenient way to measure video is to set the signal from sync tip to peak white, to take up 7 divisions vertically. Then the scope is calibrated at 20 IRE per division. When using the zero chop method of measuring DOM, the chop pulse will be on the top graticule.

The circuit in figure 24 consists of three parts. An input amplifier, Q1, buffers the input signal. The clamp, Q2, forces the sync tips to be at ground potential (practically, just below ground - you can make the sync tips closer to ground by lowering R1, at the cost of more battery drain). This provides a signal to the scope that doesn't change level as the video information changes. The third part of the circuit is the sync separator, U1, which uses an integrated circuit make expressly for this purpose.

Resistor R4 provides a proper termination to the video source. If your demodulator doesn't drive 75 ohms (almost all do), you can remove this resistor. Transistor Q1 is an emitter follower to isolate the load from the source. When the input video goes negative at the sync tip, if the output side of C6 is at too low a voltage, clamp transistor Q2 turns on, charging C6 more positive on the output side, raising the output voltage. If the output voltage is higher than zero volts, Q2 never comes on, and the output voltage drops as C6 discharges through R9.

Video signal from the emitter of Q1 is applied to the sync separator through low pass filter R2-C1, which provides a degree of immunity to noise on the input signal. A clamp circuit inside U1 charges C2, to form a clamp circuit similar to that of C6/Q2, but with time constants optimized for sync separation. The IC includes all circuits necessary to recover composite sync, which is switched to the output for horizontal rate triggering. For vertical rate triggering, we use the "OD/EVN," or *odd/even*, output. This signal changes state during every vertical interval, going positive at the beginning of field 1 and negative at the beginning of field 2. This allows you to select one field or the other by changing the trigger polarity on your oscilloscope from + to -. You will need this feature in order to measure VBI signals. One problem you may have with low cost oscilloscopes is that the brightness may be low when observing VBI signals. This is because the signal is repeated only 30 times a second, and the sweep rate during the signal is high. The only solution outside of buying a more expensive scope, is to observe the signals in near darkness (the author must do this with his Heathkit scope).

All parts including the IC are available through DigiKey, (800)344-4539, or through many commercial parts distributors. All parts except the IC are likely available at Radio Shack. You are not too likely to go wrong unless you get hold of a leaky capacitor for C6, or a bad transistor. Diode D2 is included to protect the circuit should you put the batteries in backwards. You could also run the circuit from a 6 volt cube power supply or from a lab supply.

## APPENDIX C. A SIMPLE MODULATION CALIBRATOR

The circuit shown in figure 25 is intended to simulate a video signal modulated 87.5%. This is accomplished by attenuating a 61.25 MHz input signal by 18 dB at the horizontal line rate. Resistors R5, R6, and R12 are configured as a tee attenuator. Since C6 and L5 are in series with R12, the reactance of C6 and L5 at 61.25 MHz was also considered in determining the proper resistor values. In order to improve matching between the signal source driving the switched attenuator and the demodulator's RF input, pads were placed either side of the tee attenuator. Diodes D1, D2, and D3 are PIN diodes. PIN diodes were used because they exhibit very low AC on resistance. In order to supply complementary signals to the diodes for proper switching, Q1 and Q2 are used in a differential switch configuration.

When the input on J3 is at low, D3 is off while D1 and D2 are conducting. This essentially shorts the tee attenuator, thus allowing all of the input signal to pass to the output. When the input on J3 is high, D3 is conducting while D1 and D2 are open circuit. During this time the 18 dB attenuator is switched in line with the input signal and output.

The input on J3 can come from any 15 KHz signal source. The composite sync output available on most NTSC generators makes a convenient source. If a function generator is used, the waveform's period is not too critical, but should roughly match the 4.7 µS period of horizontal sync.

If possible, after constructing the video demodulator output level calibrator it would not hurt to verify the 18 dB attenuation level. This can be accomplished by using the spectrum analyzer method for measuring DOM.

This method of measuring DOM puts you at some risk of error, because not all demodulators have the linearity to make the measurement. However, if you calibrate to 87.5% DOM then adjust sources to 87.5% DOM, you should be OK.

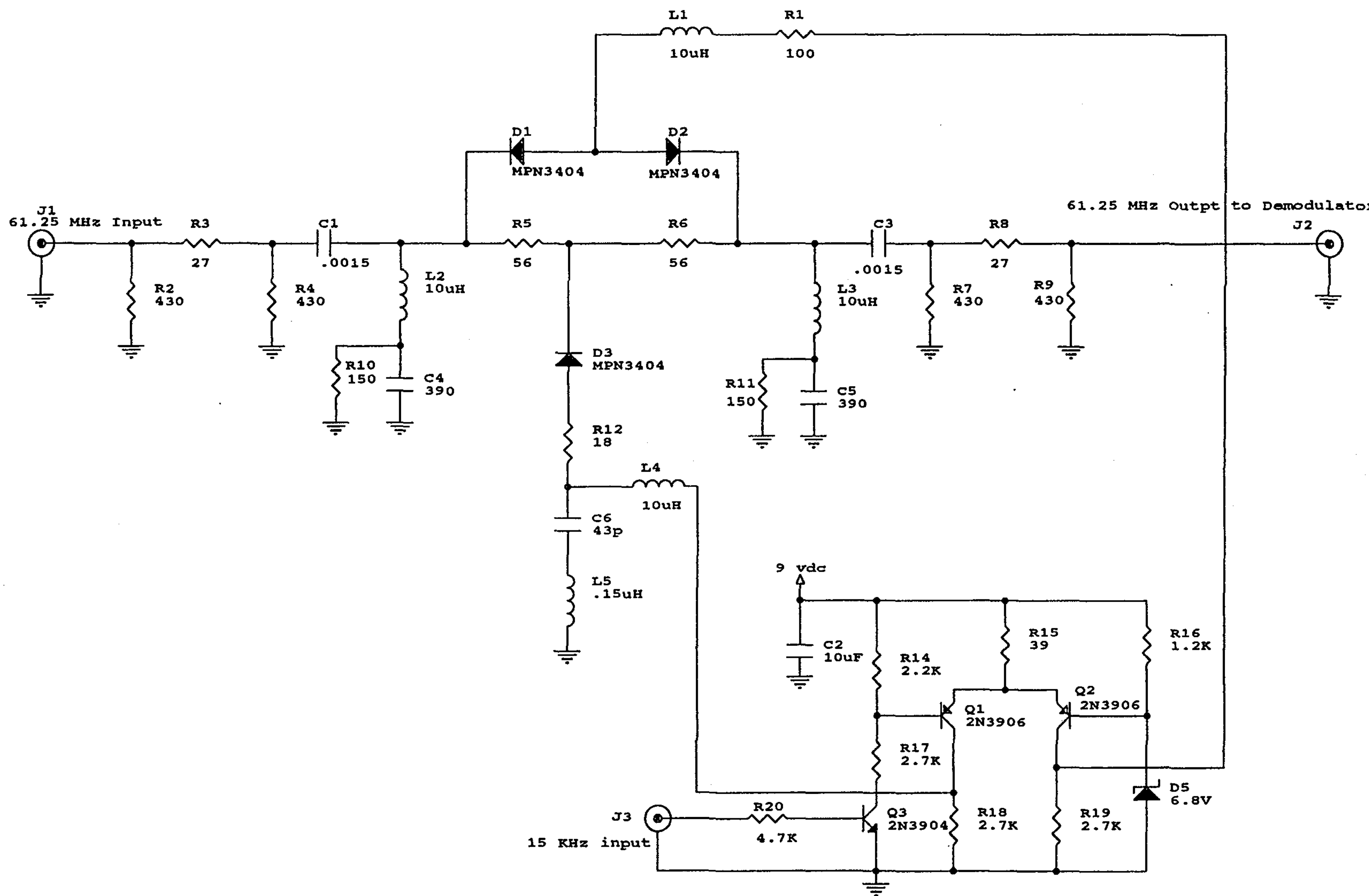



Figure 25. A Simple Depth of Modulation Calibrator

## END NOTES

1. Ambiguity exists in the definition of worst case differential gain and phase in the agreement. This will be resolved shortly by a joint committee which is defining measurement procedures.

2. More generally, the term *group delay* is applied. We often use the terms interchangeably, though technically, they are not quite the same: Group delay is a general term, and chroma to luma delay is a special case.

3. We are aware of the sloppy use of "pulse," v.s. half cycle of a sine wave. The two are not the same, but the effect of a pulse and a half cycle of a sine wave are similar when viewed on a TV screen, and the concept is useful.

4. McClatchie, Frank F. "How To Measure It, Set It Right, And Keep It That Way" NCTA 1988 Technical Proceedings

5. Mountain, Ned, private communication.

6. In the PAL and SECAM systems used in Europe and elsewhere, the picture has 625 lines and is updated 25 times a second. The added number of lines is one reason that European TV is often "sharper" than is North American TV, but the reduced number of pictures per second results in more flicker than is seen in North America.

7. In the old black and white days, the field rate was identically equal to the ac frequency of 60 Hz (then cycles per second). In those early days, it was harder to filter ac to get dc to operate the TV, and the result was hum bars on the TV. To reduce the visible effect of hum, the picture was locked to the ac line, resulting in still hum bars. The change from 60 Hz to 59.94 Hz came about when NTSC color was added, and results from the need to interlace the beat between the sound and color carriers. The details are beyond the scope of this paper.

8. An outstanding reference for baseband video is the <u>Video Reference Data</u> chart published in <u>CT</u> magazine, November 1991. This issue includes several good articles on baseband video.

9. The Institute of Radio Engineers (IRE) was one of the organizations which merged to form the Institute of Electrical and Electronic Engineers (IEEE) about 1960. Much of the early work aimed at standardizing the TV system we have was done under the auspices of this organization.

10. Black is at blanking in PAL. It is offset for historical reasons in NTSC.

11. The reason for the serrations in vertical sync is historical, relating to how early receivers operated. The need today is debatable, but no one is willing to remove something, the removal of which could cause unpredictable reaction on some TV.

# MODULATION AND CODING TECHNIQUES FOR HIGH CAPACITY COAXIAL CABLE AND SCM FIBER DIGITAL TV-HDTV DISTRIBUTION

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## ABSTRACT

New emerging digital modulation-demodulation (modem) and coding techniques suitable for spectrally and power efficient digital TV and HDTV distribution systems over coaxial cable and subcarrier multiplexed (SCM) fiber optics systems are described. Efficient and robust modulators/demodulators (modems) combined with low redundancy forward error correction (FEC) - integrated circuits attain a spectral efficiency in the 2 b/s/Hz to 7 b/s/Hz range and a bit error rate (BER) in the $10^{-7}$ to $10^{-10}$ range. Recently developed VLSI-ASIC chips, having a universal QPSK-QAM-QPRS modem architecture combined with FEC-chips, operating up to 100 Mb/s, with a 3% to 10% redundancy will enable the transmission of 30 Mb/s rate (or even higher bit rate) digitized TV signals in one conventional analog TV band.

Requirements for FDM shared analog and digitized TV distribution systems are highlighted. The performance of a new class of QPRS filtered modems invented by the author is compared to conventional QPRS and QAM-8-PSK cable systems. The performance advantages of our new class of "above Nyquist rate" modems and of an operational SQPRS (staggered) QPRS system developed by Digital Radio Laboratories are highlighted.

## 1. MODEM AND CODED MODEM ARCHITECTURES FOR DIGITAL CABLE TELEVISION AND SCM FIBER OPTICS DISTRIBUTION

Numerous modulation/demodulation (modem) architectures have been implemented and considered for digital cable TV and subcarrier multiplexed (SCM) fiber optics digital TV distribution systems. From simple QPSK, 8-PSK to more advanced coded 64-QAM and even 1024-QAM cable systems have been designed and studied.

In general, the purpose of increasing the number of modulation states is to increase the spectral efficiency expressed in terms of b/s/Hz. For example, a theoretical QPSK system could transmit 12 Mb/s while a 16-QAM system 24 Mb/s in a 6 MHz wide video channel. The theoretical spectral efficiency and BER = f(C/N) performance of ideal coherent a = 0 filtered (raised-cosine-Nyquist filtered with a roll-off parameter of a = 0) uncoded modems is illustrated in Figure 3 and 4. The practical spectral efficiency is typically 5% to 50% below the theoretical values of QAM systems and is 4% to 30% above the "theoretical ISI-free" values for QPRS systems [1; 6; 11; 12; 13]. Steep filters, having an a = 0.1 to a = 0.2 and 60 dB out-of-band rejection have been used in several modem designs [1; 2] in order to approach the theoretical spectral efficiency limit of QAM, within 5%.

For example, in the **DigiCipher** radio broadcast system [J.A. Kraus, 18] trellis coded and Reed-Solomon FEC coded 16-QAM and 32-QAM a 19.51 Mb/s and a 24.39 Mb/s rate modulated signal is transmitted in bandwidth of 6 MHz. This corresponds to a 19.51 Mb/s: 6 MHz = 3.25 b/s/Hz (as compared to a theoretical uncoded of 4 b/s/Hz for a = 0, 16-QAM system.

For the DSC-HDTV (Digital Spectrum-Compatible High-Definition Television) system of **Zenith** [Luplow & Fockens, 8], the transmission of 11.1 Mb/s and of 21.0 Mb/s data (representing the digitized video, audio, ancillary data and error protection bits) 2-level VSB and 4-level VSB pilot aided methods are described [8]. The symbol rate is 10.76 M Symbols/second.

For the **ATVA-Progressive System**, a 19.43 Mb/s rate, 16-QAM system is described for the 6 MHz wide video channel [10].

For relatively lower data rate cable systems (1.544 Mb/s to 2.048 Mb/s) and hybrid FDM, analog video and digital data cable systems, more spectrally efficient modems have been implemented by our design teams including 256-

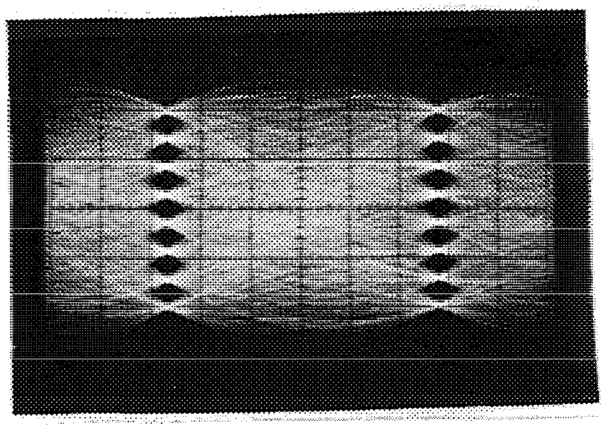

**Fig. 1** Experimental 45M b/s coherently demodulated I-channel of a 64-QAM system with a = 0.1 filtering [1 and 2].

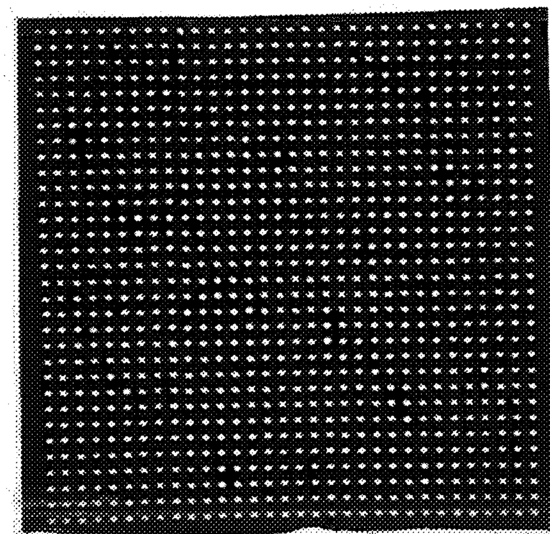

**Fig. 2** Measured constellation of a 1024-QAM subsystem [1 and 2].

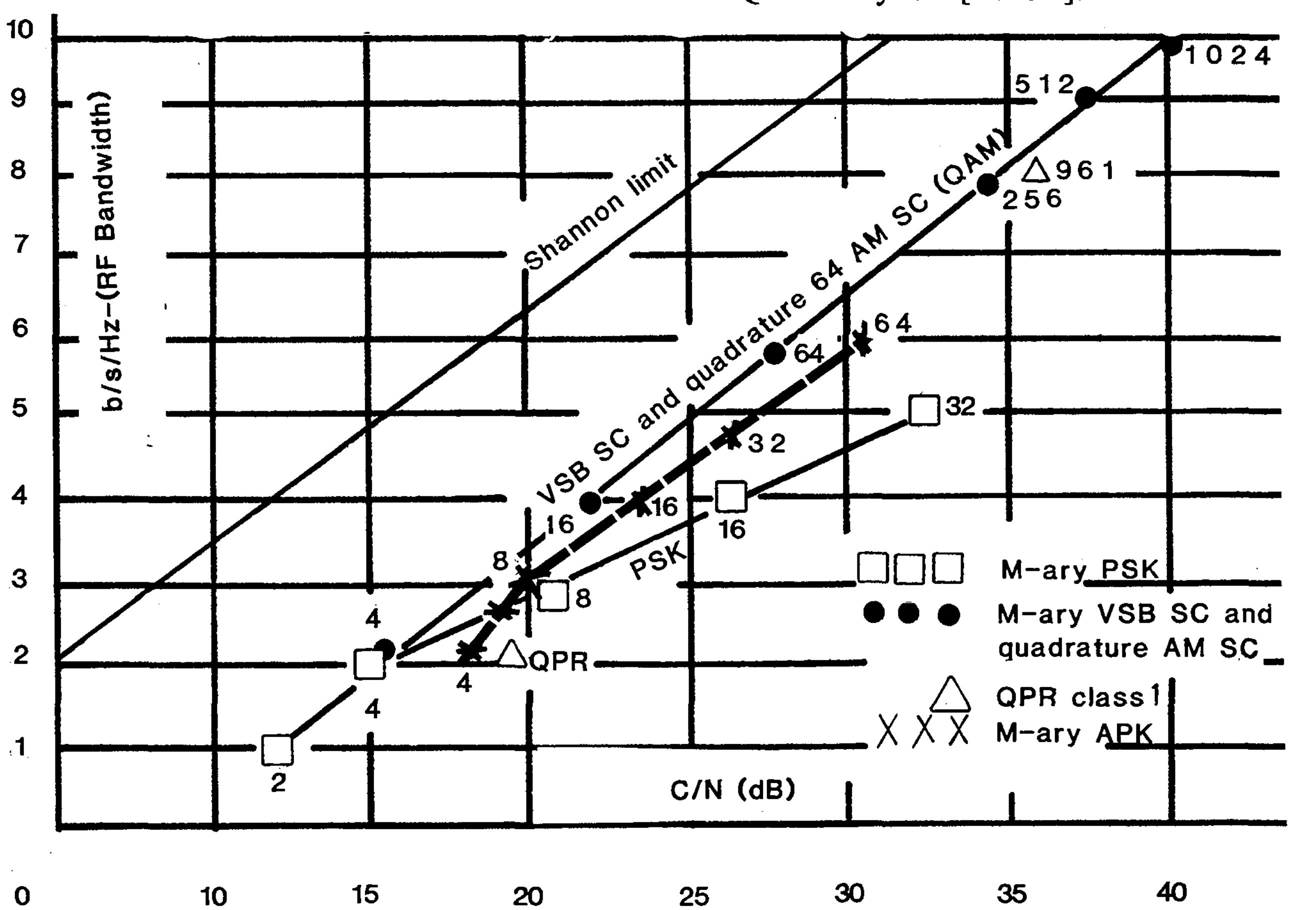


**Fig. 3** Theoretical bit rate efficiency of uncoded-modulated coherent systems as a function of the available C/N at P(e) = $10^{-8}$. The average C/N is specified in the double-sided Nyquist bandwidth which equals the symbol rate. Ideal a = 0 filtering has been assumed. Shannon limit is for coded-modulated systems [1 - 5].

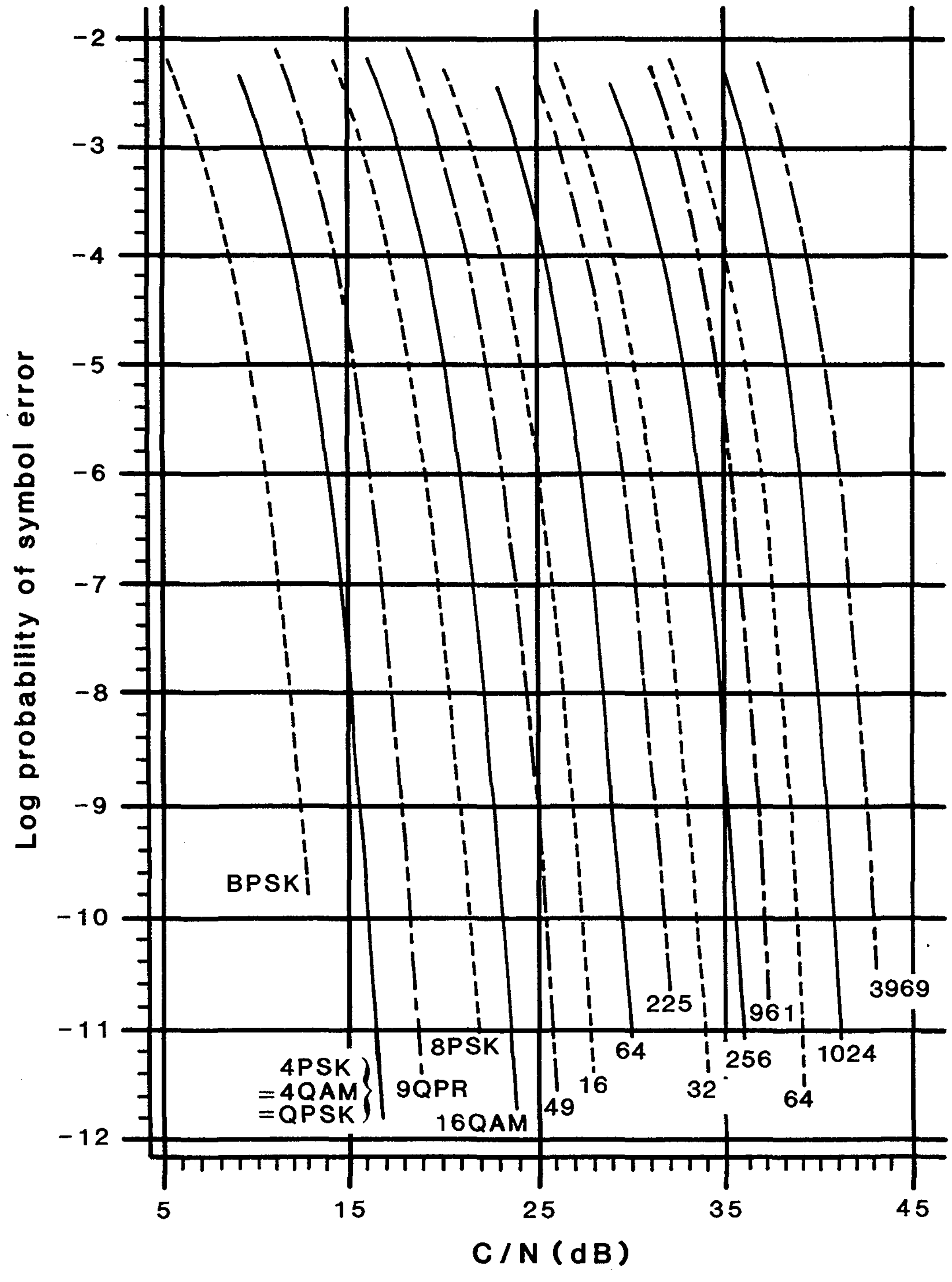


**Fig. 4** Probability of error-theoretical performance curves of M-ary PSK, M-ary QAM and M-ary QPR modulation systems vs. the carrier-to-thermal noise ratio in dB. White Gaussian noise channel only - no phase noise. Double-sided Nyquist RF bandwidth equals the symbol rate bandwidth [1 - 5].

QAM modems which achieve a practical spectral efficiency of 6.66 b/s/Hz specified at the 60 dB out-of-band attenuation point [1; 2; 14]. A practical block and implementation diagram of an FEC-coded 256-QAM modem (could be extended to 1024-QAM) modem is illustrated in Figure 5. Even though our modem designs incorporate advanced baseband adaptive time equalizers (BATE) (fully digital), we found that the residual group delay, i.e., group delay and/or echo not equalized by the BATE and IF adaptive equalizer could significantly degrade the performance, as illustrated in Figure 6. We also found that by "staggering" or "offsetting" the 256-QAM or 1024-QAM systems, for short S-QAM we could reduce the sensitivity of the cable system to residual imperfections, see Figures 5 and 6. Experimental field data of one of our (2196, 2136)-BCH-FEC coded (2.7% redundancy) 256-QAM cable systems is illustrated in Figure 7 - Figure 9. For a BER = $10^{-6}$ our 256-QAM required a C/N = 32 dB and had a spectral efficiency of 6.66 b/s/Hz [2]. Extensive field experiments over LOS hybrid microwave and coaxial cable systems indicated that 99.875 percent or better EFS (error free second) performance could be obtained with these modems over regenerative spans exceeding 1000 km and excellent performance was also obtained on the cable system between Vancouver, Canada and Hawaii.

Probably one of the most spectrally efficient digital modulation cable systems studied so far has been the 1024-QAM system described by Feher in [2]. See the experimental constellation/hardware photograph of Figure 2. Evidently such an efficient system having the theoretical potential of **10 b/s/Hz** and practical potential of more than **9 b/s/Hz** requires an increased C/N requirement and very advanced complex equalization and interference (including echo) cancellation subsystems. For a 6 MHz video channel, a future 1024-QAM system could have the potential of 54 Mb/s. However, it could take some time prior to design completion and implementation of these types of systems.

Simpler, more robust modem architectures have also been considered for digital TV and digital audio cable distribution systems and for **SCM fiber optics** system applications [15-17]. A spectral efficiency in the 2 b/s/Hz to 3 b/s/Hz range has been found to be a good compromise for low C/N requirements, robust performance and low cost ASIC implementations.

Among the simplest and most robust digital modems are the BPSK, G-MSK and QPSK [1]. For a raw BER = $10^{-6}$ a low C/N in the range of 8 dB to 14 dB is sufficient. The spectral efficiency of these modems is less than 2 b/s/Hz, that is, these modem architectures are not suitable for spectrally efficient digital cable TV and digital audio cable distribution applications. Among the robust, i.e., low C/N operation, modems 8-PSK and staggered 9-state QPRS or S-QPRS are of significant interest. For coaxial cable and for SCM fiber optics systems [15], 8-PSK modems have been considered by several corporations. A brief comparative study of these techniques is highlighted in the next section.

## 2. COMPARISON OF 9-QPRS AND 8-PSK SYSTEMS; MULTIPLEXED DIGITAL AUDIO CABLE, TV AND SCM FIBER SYSTEMS

QPRS and S-QPRS systems offer several advantages over 8-PSK systems, including a 3 dB lower C/N requirement, simpler architecture, less sensitive to filter imperfections including group delay or cable roll-off, and more robust performance in a phase noise dominated channel. The 3 dB lower C/N could be potentially even further reduced, i.e., to a lower C/N with non-redundant QPRS error correction and/or Viterbi decoding.

The 8-PSK system has 3 b/s/Hz theoretical spectral efficiency, while the Nyquist rate for 9-QPRS is 2 b/s/Hz. However, as highlighted in our publications and patent [1; 12-14], we note that the practical spectral efficiency of 8-PSK systems is in the 2.5 b/s/Hz range with more complex filters - leading to significant system sensitivities. The practical spectral efficiency of the 9-QPRS system is in the 2.2 b/s/Hz to 2.6 b/s/Hz range. It has been proven by numerous organizations (since Dr. Lender's discovery around 1960) that it is feasible to transmit with simple/robust hardware significantly above the Nyquist binary rate with 3-level partial response (PR) signals - which are the baseband part of 9-QPRS modems. See Lender's chapter in [4].

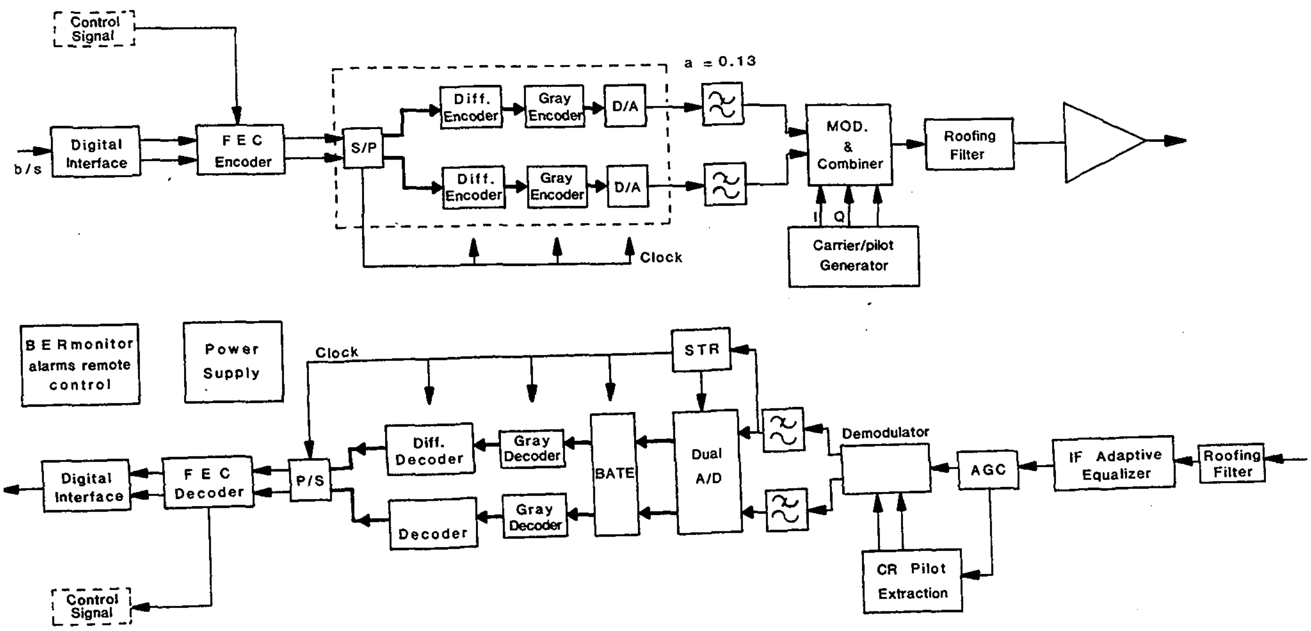


Fig. 5 Block diagram of FEC-Coded QAM and QPRS systems including baseband adaptive time equalizers (BATE).

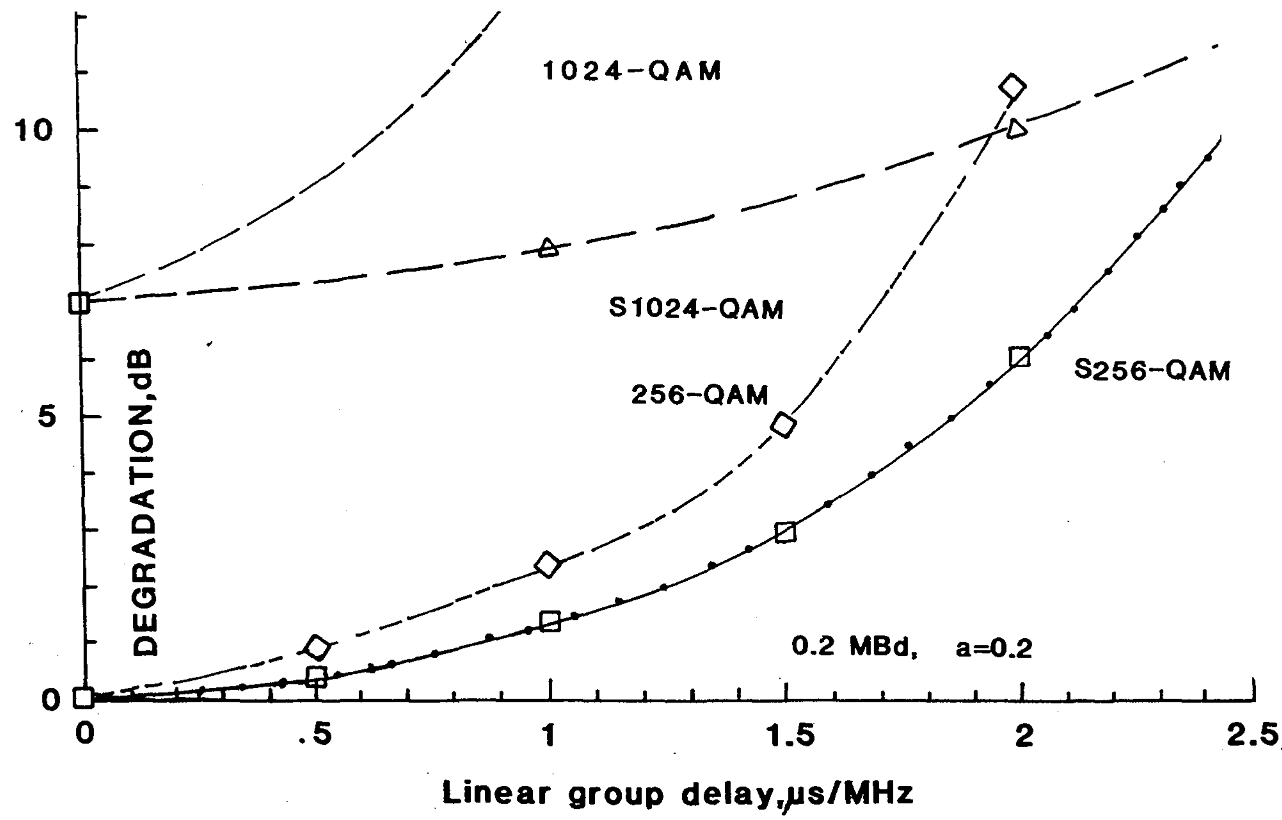


Fig. 6 Performance degradation at BER = $10^{-6}$ of conventional QAM and of staggered QAM, i.e., SQAM due to residual (unequalized) linear group delay distortions of a cable TV system. These normalized curves are for a scaled-down 0.2 M Baud system [1 and 2].

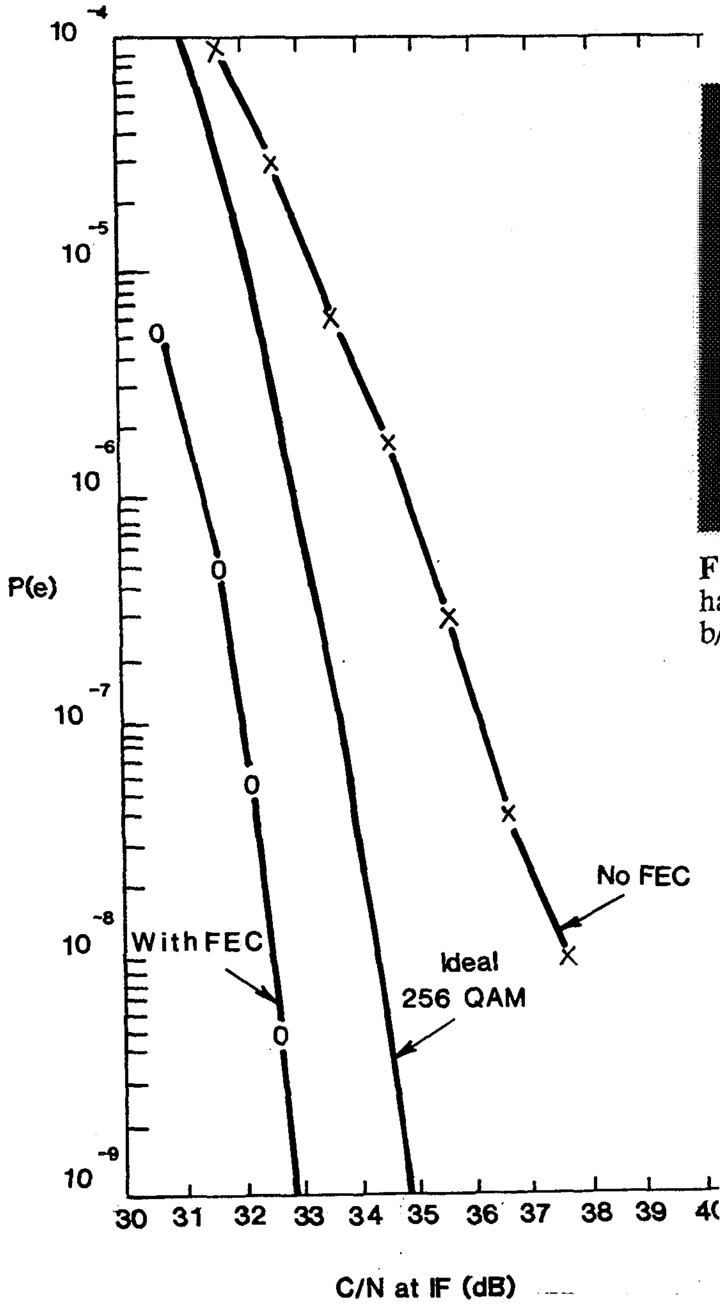


**Fig. 7** Experimental 256-QAM-FEC coded performance in cable systems and LOS microwave systems, measured by the author [1 and 2].

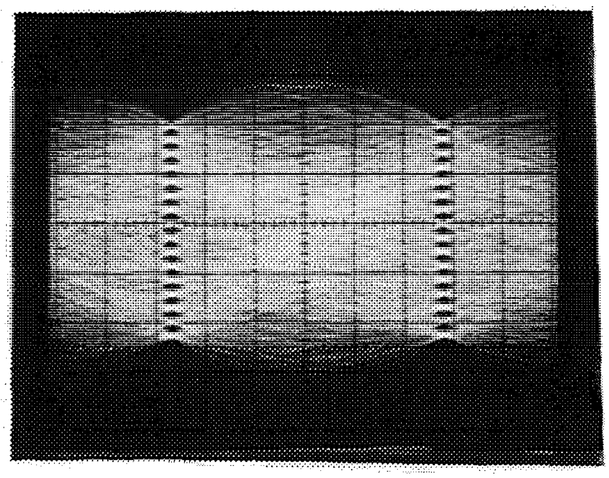

**Fig. 8** 256-QAM experimental eye diagram having a practical spectral efficiency of 6.66 b/s/Hz.

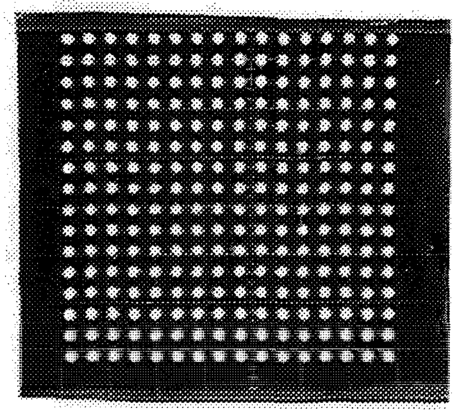

**Fig. 9** Constellation of a 256-QAM digitally equalized cable system tested between Vancouver (Canada) and Hawaii [2].

The theoretical BER = P(e) = f(C/N) advantage of 9-QPRS systems as compared to 8-PSK is illustrated in Figure 4 and Table 1.

**Table 1** Reduced C/N requirement of 9-QPR as compared to 8-PSK in AWGN systems. The 9-QPR system is about 3 dB more robust than the 8-PSK system.

| | 9-QPR | 8-PSK |
|---|---|---|
| $10^{-4}$ | 13.5 dB | 16.5dB |
| $10^{-6}$ | 16 dB | 19 dB |
| $10^{-8}$ | 17.5 dB | 20.5 dB |

The practical efficiency of 8-PSK is in the 2.5 b/s/Hz to 3 b/s/Hz range if an out-of-band rejection of 20 dB to 30 dB is sufficient. For cable TV applications, we assume that an integrated our-of-band spectral rejection of about 60 dB is required, thus the practical efficiency of 8-PSK is assumed to be less than 2.5 b/s/Hz.

An additional advantage of 9-QPRS systems (as compared to 8-PSK) is that they are approximately 3 dB more robust to **phase noise** [1; 14]. For example, at BER = $10^{-6}$ the integrated phase noise requirement is $(C/N_p)$ = 23 dB for a 2 dB degradation in case of 8-PSK, while it is only 20 dB in case of 9-QPRS. thus the 9-QPRS has an additional 3 dB phase noise advantage.

Increased capacity-improved performance QPRS systems, i.e., $3 \times 3 = 9$ state; $7 \times 7 = 49$-QPRS and even $15 \times 15 = 225$-QPRS systems can be designed by using a new invention (see [13], patent by Feher et al.). The advantage of the patented QPRS are illustrated in Figs. 14 and 15. A capacity increase of 8% without increasing the number of states and thus the C/N requirement is attained in the experimental results of Fig. 15.

## 3. DRL'S NEW SQPRS MULTIPLEXED DIGITAL AUDIO CABLE BROADCAST SYSTEM

Staggered 9-QPRS or S-QPRS systems have additional advantages as compared to conventional QPRS. In references [1; 11; 12], it is demonstrated that staggering reduces the peak-factor and the potential intermodulation problems and leads to simpler, better performance coherent demodulator design. In Figure 10, the constellation diagram of a 9-state SQPR digital cable system is presented, courtesy of Digital Radio Laboratories (DRL), Carson, CA. The DRL modem, 9-SQPR is the **'industry first'** operational staggered QPRS cable system implemented in ASIC, see Figure 11. The 9-QPR modem is currently in operation, see Figure 12. The DRL digital audio system carries 18.56 Mb/s data, i.e., two 9.28 Mb/s in two 4 MHz SAW filtered channels. Exceptionally high fidelity MUX (multiplexed) digital audio channels and commercial free music and audio is distributed to cable broadcast systems and directly into homes.

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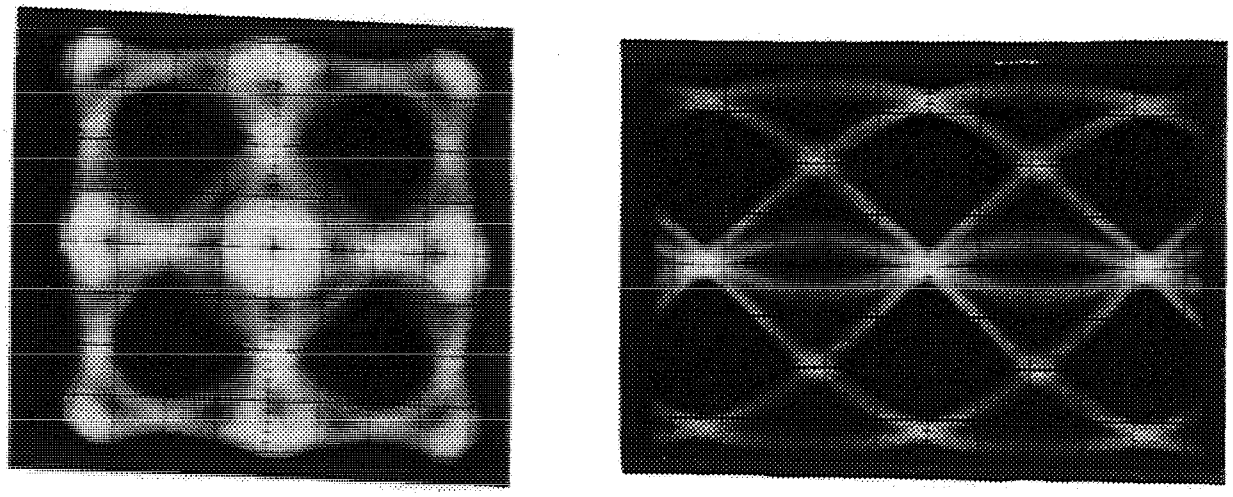

**Fig. 10** Constellation and eye diagram of a SQPR digital cable system operated at an 9.28 Mb/s rate. Courtesy of Digital Radio Laboratories, Carson, CA [6].

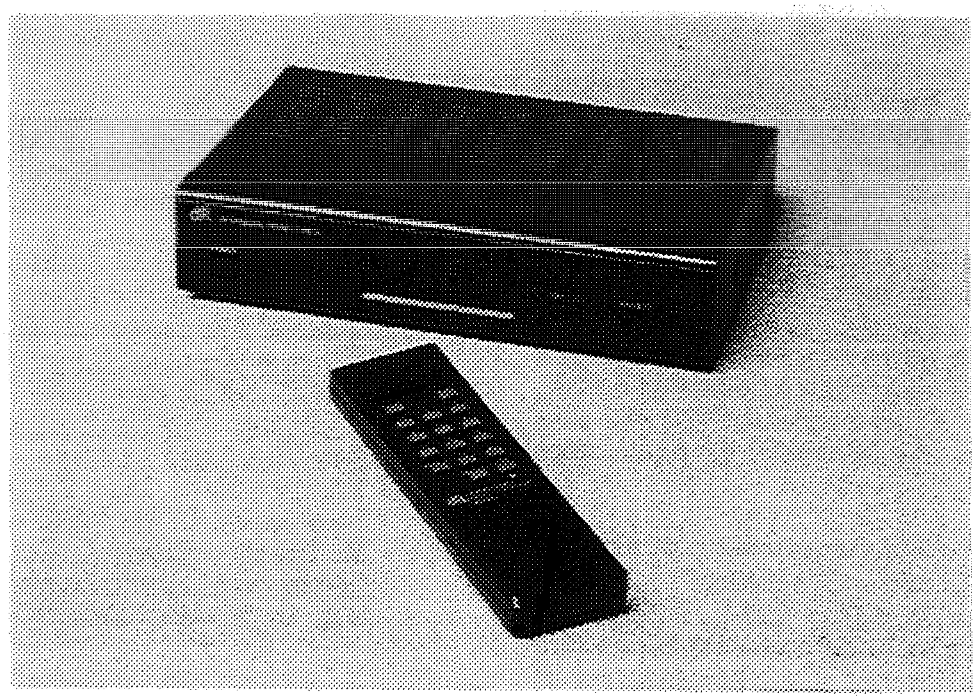

**Fig. 11** Cable TV - digital audio receivers for Digital Planet and other highest quality audio receivers. ASIC implementations by Digital Radio Laboratories of Carson, CA led to a compact size, versatility and low-cost original-proprietary SQPR systems design [6].

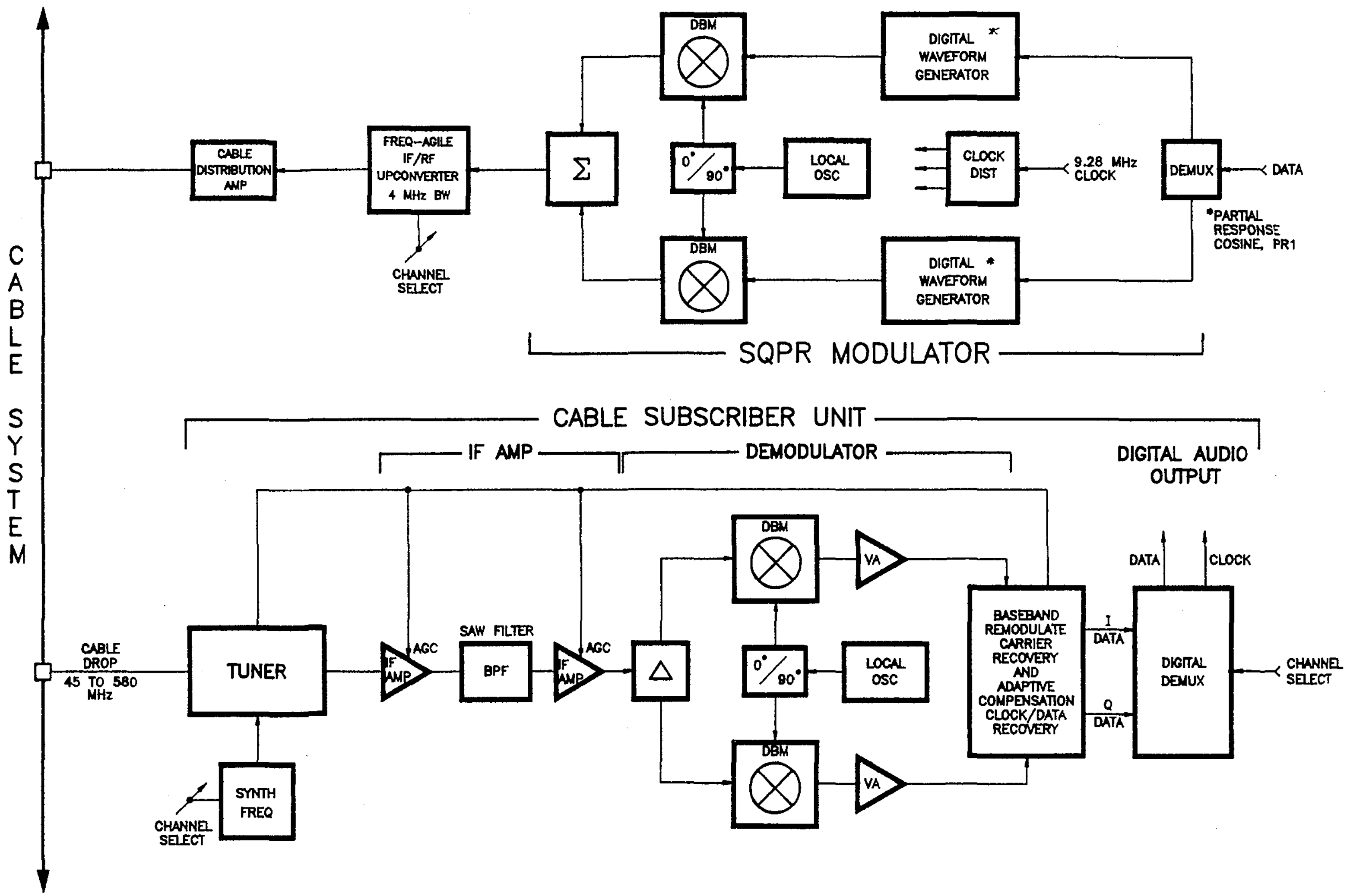


**Fig. 12** Digital Audio Cable System of Digital Radio Laboratories (DRL), Carson, CA. This original ASIC implemented staggered SQPRS already operational cable system outperforms conventional 8-PSK systems, see Table I and [6].

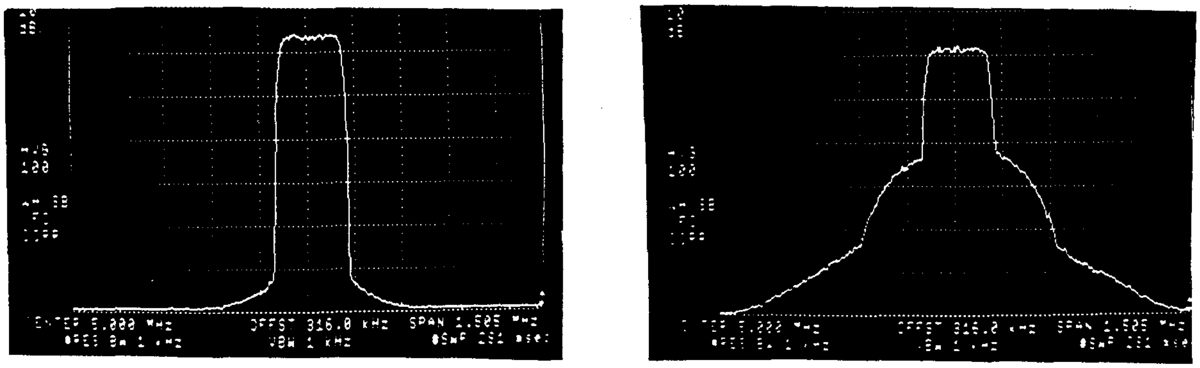

**Fig. 13** Spectrum of a 64-QAM linearly and nonlinearly amplified (NLA) system. Spectral regrowth and interference into adjacent TV channels [7 and 11].

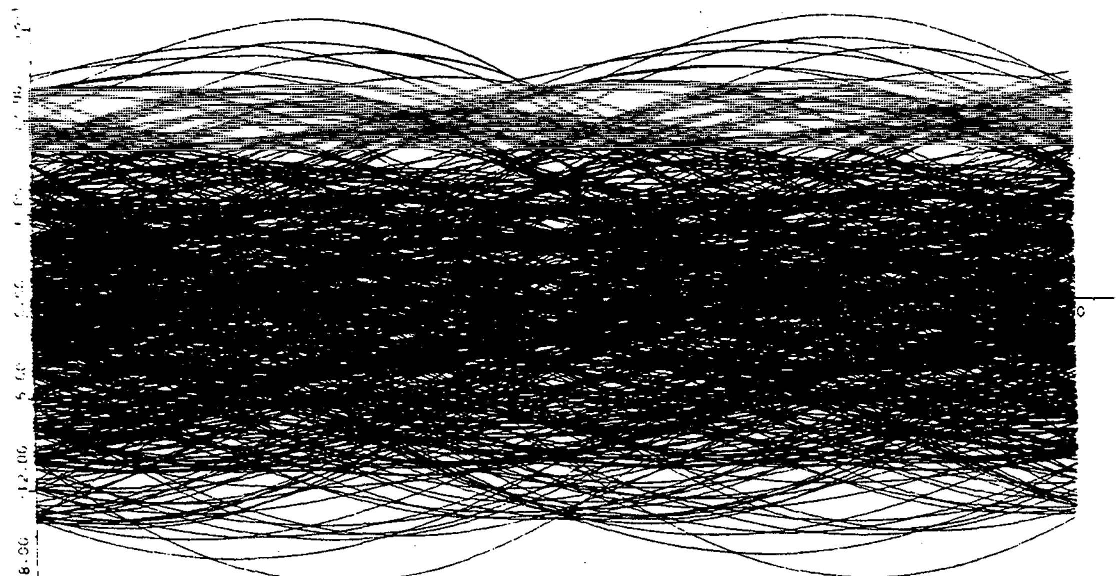

**Fig. 14** Conventional 225-QPRS (15 × 15 baseband) operated at 4% above the Nyquist rate with 6.24 b/s/Hz-computer simulation [1; 12; 13] leads to closed eye diagrams.

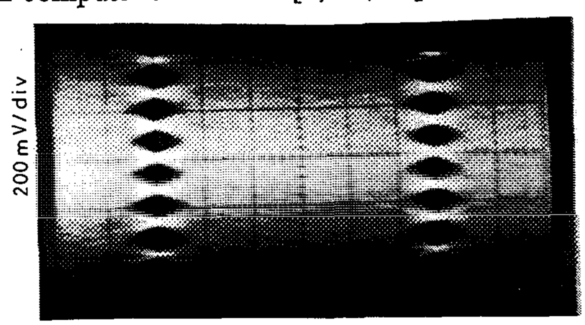


**Fig. 15** New 49-QPR and other QPR filter methods, see Feher et al. Patent [13], enable transmission above the Nyquist rate. In this hardware experiment, instead of 4 b/s/Hz we had 4.32 b/s/Hz (an 8% increase and "open eye diagrams" [1; 4; 11; 12; 13].

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## RESUME OF DR. K FEHER

Kamilo Feher, Professor of Electrical and Computer Engineering, University of California, Davis, and Director, Consulting Group, DIGCOM, Inc., has 26 years of industrial and academic R&D, teaching, management, and consulting experience. He has been Consultant to U.S., Canadian, and to many overseas corporations and to governments and has presented numerous short courses. He is known as a very productive world class systems and design research engineer, consultant and a dynamic, enthusiastic instructor. He has 6 U.S. and international patents. He has coauthored more than 260 original research papers and is the author of five books: Advanced Digital Communications: Systems and Signal Processing Techniques (Englewood Cliffs, NJ: Prentice-Hall, 1987); K. Feher and Engineers of Hewlett

Packard: Telecommunications Measurements, Analysis and Instrumentation (Englewood Cliffs, NJ: Prentice-Hall, 1987); Digital Communications: Satellite Earth Station Engineering (Englewood Cliffs, NJ: Prentice-Hall, 1983); Digital Communications: Microwave Applications (Englewood Cliffs, NJ: Prentice-Hall, 1981); and Digital Modulations Techniques in an Interference Environment, (Gainesville, VA: Don White Consultants, 1977). He supervises large digital communications (university and industry based) research teams. His inventions are used by internationally acclaimed major corporations throughout the world. His major discoveries of emerging digital cellular and mobile radio (modulation-demodulation) communications and digital satellite mobile/braodcasting systems are expected to make significant contributions to the ultimate objective of communications - to enable anyone to communicate instantly with anyone else from anywhere.

For his contribution to digital communications research and development (as a professor at the University of Ottawa, Canada), Dr. Feher was awarded the Steacie Memorial Fellowship in 1981, the most prestigious award of the Natural Sciences and Engineering Research Council of Canada (NSERC), the "Engineering Medal, Ontario, Canada", 1989, and elected Fellow of the IEEE. While at UC Davis he received the 1991 S. Helt Memorial Award for contributing an "outstanding series of papers to the *IEEE Transactions on Broadcasting*". He is very active in advanced technology transfer from university to industry.
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# MULTI-CHANNEL AUDIO DATA COMPRESSION

-ABSTRACT ONLY-

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Dolby Laboratories

The advantages of digital storage and transmission have become so significant that all new entertainment media under development are based on digitally coded representations of picture and sound. This paper concerns the digital processing of multi-channel audio to lower the bit-rate required for storage or transmission. The coding system is referred to as AC-3.

Multi-channel audio is becoming important. First delivered to the consumer in a limited way in the 1950's by the motion picture industry, multi-channel audio is now the norm in the cinema. This has been made practical by the 4-2-4 surround sound matrix process employed on the stereo 35 mm film print. Using this technique, surround sound can be delivered by any two-channel media such as BTSC stereo television or VHS stereo soundtracks. Since nearly all films are produced with matrix surround sound, source material is prevalent. Surround sound equipment for the home is one of the major growth markets in home electronics.

The film industry is preparing to evolve from analog four-channel matrix sound to five-channel discrete digital sound (left, center, right, left surround, right surround, and subwoofer). The evolution will take place for two reasons: the superiority of digital storage over analog storage; and the superiority of discrete multi-channel sound over matrix multi-channel sound. This evolution is important because consumers will be exposed to discrete five-channel sound in the cinema, and large amounts of program material will be available in this format. As in the case of matrix surround sound, consumers will want this at home and the consumer market will undoubtedly follow the film industry with discrete five-channel sound finding its way into the home. New media which are developed should provide for this new sound format, or they will be at a competitive disadvantage.

The state-of-the-art in high quality low-bit-rate sound coding is the coding of audio into 128 k bits/sec/channel. Conventional thinking would indicate that it would take a 2.5 times increase in data capacity to code five discrete channels of sound compared with two, thus requiring 640 k bits/sec for discrete five-channel sound versus 256 k bits/sec for 4-2-4 matrixed multi-channel sound. However, this is not true. Humans only have two ears, and the addition of more loudspeaker channels do not necessarily increase the amount of information that has to be coded for the human listener (although it would be for the 6 eared Martian!). Using advanced techniques to be described in this paper, AC-3 is able to encode discrete 5 channel audio into a total data rate of only 320 k bits/sec, or 1.25 times the rate required for matrix surround sound. This is very significant, as it allows new media to offer a significantly improved sound experience with only a minor increase in data rate (i.e. an additional 64 k bits/sec over two-channel sound).

Numerous desirable features have been designed into AC-3. Although multi-channel sound is encoded, the decoder can reproduce any number of channels between one and five. The lowest cost monophonic or stereophonic receiver need not fully decode all five channels thus minimizing circuit complexity. Along with the five discrete channels, a low bandwidth subwoofer channel is encoded so that program providers can provide low-frequency sound with the same subjective loudness as higher frequency sounds without using up the dynamic range of the DACs (higher levels of LF sound are required for the same subjective loudness). Compression control information is included in the bit stream which allows the consumer the option to enjoy either the original wide dynamic range sound or a compressed low dynamic range version (or anything in between) from the same broadcast. Also included in the bit stream is information about the subjective dialog loudness, which allows different program providers to reserve different amounts of headroom for dramatic impact, while the consumer hears uniform dialog loudness for all channels and programs.

This paper will attempt to make clear the significance of multi-channel audio to the cable industry. The AC-3 coding system will be described to show how high quality discrete multi-channel audio can be efficiently encoded for delivery to the subscriber. The paper will also describe how the use of a coding technology such as AC-3 can solve other problems such as channel loudness uniformity and variable dynamic range requirements.

# NETWORK COMPATIBLE ARCHITECTURES

John Caezza
Magnavox CATV Systems, Inc.

*Abstract*

*In recent years, fiber optic transportation equipment has caused the CATV industry to rethink traditional system architectures and service delivery methods. With digital compression in an explosive development phase and 1 GHz bandwidths rapidly becoming a reality, we are once again re-evaluating architectures and service delivery methods.*

In this paper we will take a look at some of the techniques we can use today to make use of Network Compatible Architectures. These NCAs will allow for customer- and capital-friendly system upgrades in the future at the expense of a little pre-planning today.

## INTRODUCTION

### Having a Plan

The future for communication networks is bright. The emergence of fiber optics and the inevitability of HDTV, digital compression, PCN, Near-Video-On-Demand and more will help insure that bright future. As we look at our networks today, we have begun to ask ourselves if we are prepared for that future. Do we have a plan? In the following pages, this paper will try to provide some thoughts and guidelines for preparing this plan to attain network architectures compatible with the services and technologies of tomorrow.

## DESIGN CONSIDERATIONS

### System Demographics

As we begin to formulate our plan for the future, it's always a good idea to know where we are today. What are the system demographics? Total homes and homes per mile; Architectures will vary significantly depending on whether you have 40 homes per mile or 240 homes per mile. Ethnic diversity; If your market is segmented in such a manner that the primary language or language of preference is not English, then special segmented delivery may be of special concern. Occupational diversity; Not so much whether or not you are providing service to a blue collar or a white collar worker, but whether you service single family dwellings, private campuses (educational and industrial), or local institutions (business and governmental). Viewing diversity; Covers the types of programs and the number of channels necessary to carry them.

### Existing System

Next, an inventory should be compiled on the equipment that is currently in the network and how it is configured. Most systems will consist of a variety of operational bandwidths and amplifier technologies. Understanding what is reusable in the network upgrade and what components can be brokered is important because these components can represent up to 10% of the network rebuild costs. The existing network architecture should also be reviewed (in most cases today it will be a tree and branch network) and as-built maps should be verified.

## Future Network Requirements

Now that we know where we are, we can need to understand where we want to be. And, where we want to be may consist of several phases or stops. Start by reviewing the system demographics and determining whether or not you are providing the necessary services and communication conduits to the proper areas. Understand the application and impact technologies such as fiber optics, HDTV, interdiction and digital compression. Ask questions about how services such as near-video-on-demand, distance learning, inter- and intra-campus communication, alternate access and PCN can be integrated into a network.

## Implementation Requirements

Define project implementation phases. The existing system may be 12 years old, use P3 cable, a combination of 330 and 450 MHz equipment, and provide 32 channels of basic programming and three premium services. Phase I of the project might expand the network to 550 MHz bandwidth and interdiction capabilities and at the same time provide for data communications between local bank branches and intra-campus data, voice, and video communication for a local business. Phase II of the project might call for the bandwidth expansion to 750 MHz and digital compression with HDTV, near-video-on-demand (NVOD) and PCN services. The portion of the band between 450 and 750 MHz will provide for the digital services while the band below 450 MHz will maintain its analog format. Further, existing cable and strand is to be used wherever possible to help contain costs and Phase II implementation should not require major construction efforts. A further requirement would call for existing bridger locations to be maintained.

## Required Network Performance

Define the technical performance requirements of the network and how they will apply to each phase of the project. Phase I may require a 46 dB CNR, with distortions of greater than 53 dBc CTB, 53 dBc CSO, and 50 dBc Xmod with tap levels of 15 dBmV. Phase II of the project will call for a 48 dB CNR for analog programming, 43 dB for digital programming, distortions as in Phase I and tap levels of 15 dBmV. Service areas for Phase II may not exceed 2400 passings which should allow for the implementation NVOD and PCN. Note that tap levels do not have to change between Phase I and II even though the bandwidth expands since the digital signals will be more tolerant of a lower carrier-to-noise ratio.

## Network Implementation

With the existing network defined and the requirements understood, we can now design a network that brings the two together. Experience has shown us that the design process will require us to begin with the 750 MHz design both coaxial and fiber and then "underlay" the 550 MHz design. This will hold true whether we are using fiber-to-the-feeder (FTF), fiber backbone, cable-area-networks (CAN), or traditional coaxial architectures. This is extremely important to do if costs incurred between Phases I and II are to be maintained. Many networks will be a hybrid of several of these design techniques.

Figure 1a shows an example of how the 550 MHz trunk design with future 750 MHz expansion capabilities might be implemented. This particular area has densities of 50 homes per mile. The 550 MHz fiber nodes will always be co-located at a 750 MHz node. This leads to a design which is less than optimal at 550 MHz from a spacing standpoint but it is more than compensated by the ease at which 750 MHz is implemented.

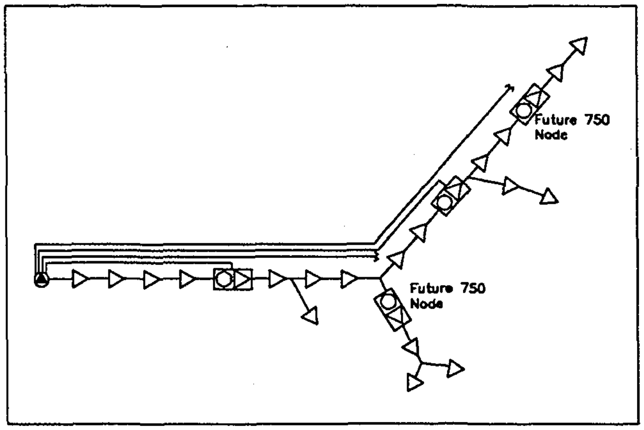


*Figure 1a.*
*550 MHz CAN, 750 MHz FBB Ready*

Figure 2a is an example of the distribution design architecture which makes use of a combination of trunk, network, and line extender amplifiers. Again, amplifier locations are optimized for their 750 MHz performance which detracts from the 550 efficiencies but is eventually recovered at the time of the 750 MHz upgrade.

Figures 1b and 2b show the final 750 MHz design implementations of the 550 MHz areas. In the backbone architectures the 750 and 550 MHz bridger locations are maintained at the original 330 MHz locations. The additional 750 MHz fiber nodes are added, trunk network amplifiers extend reach between original bridger locations, and where necessary trunk station signal flow is reversed through a module upgrade.

Actual network implementation methods now need to be considered. Consideration as to whether or not the fiber for 750 operation is placed in the system during Phase I or Phase II construction, whether or not 750 MHz or 550 MHz amplifiers are initially installed. The development of network amplifiers, also known as distribution amps, mini-trunks or mini-bridgers, has helped provided valuable design tools to extend nodal reach while at the same time reducing active counts and improving network

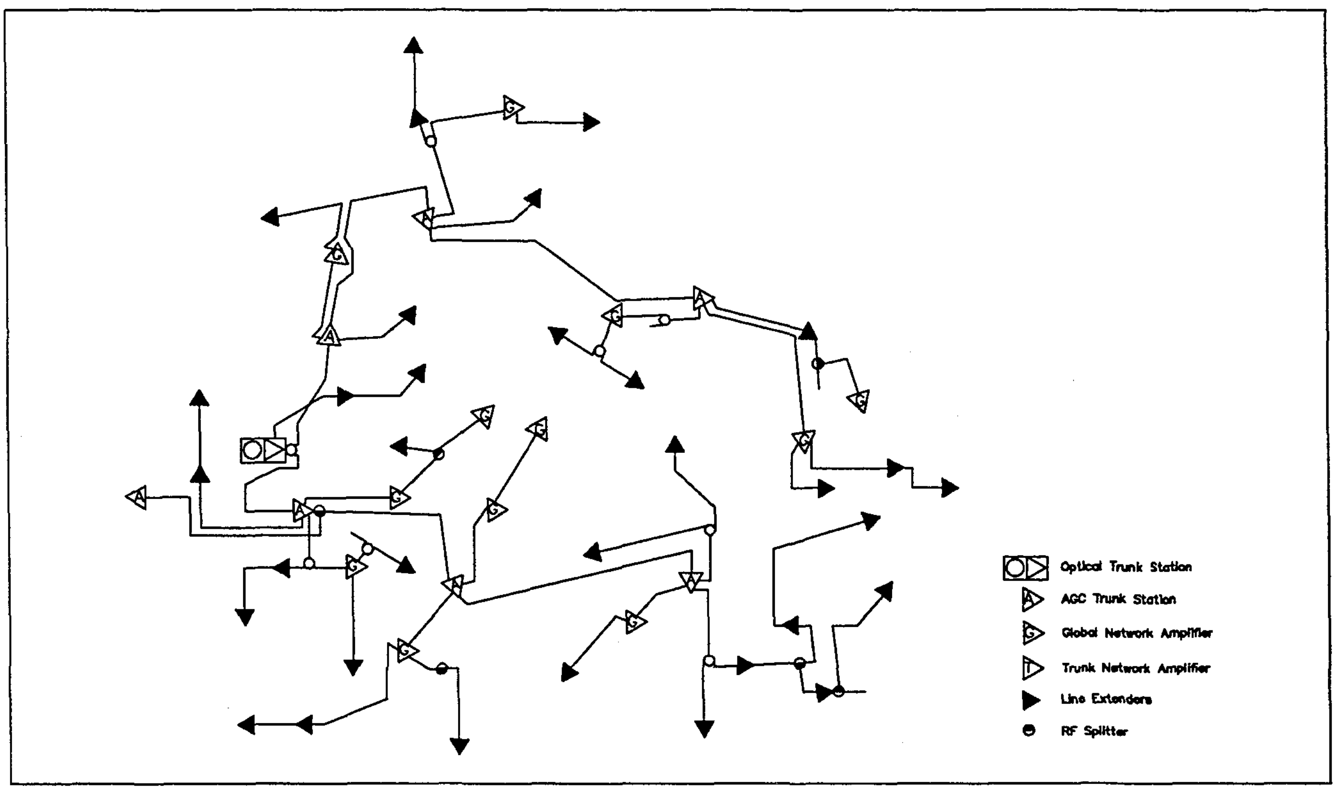


*Figure 2a.*

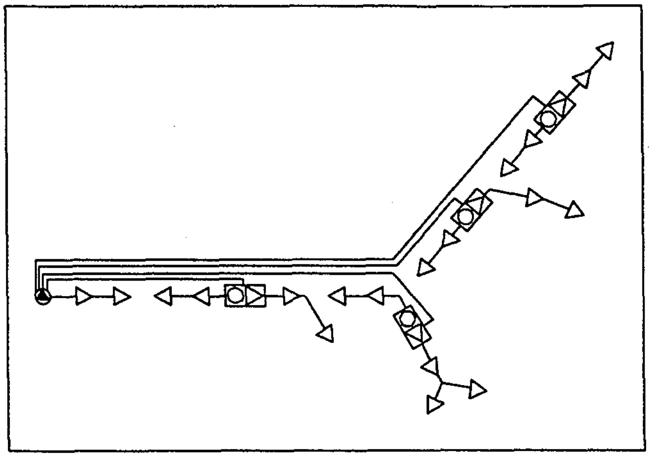

*Figure 1b.*
*750 MHz Fiber Backbone Expanded From*

reliability. Today, most operators would opt to place the extra glass and wait until the 750 MHz amplifier technology matures. A year from now, costs may dictate that the 750 MHz amplifiers also be installed as part of Phase I with operation to 550 MHz.

Tables 1 and 2 show some of the possible cost combinations that might be considered for the implementation of the two sample topologies. The first cost column represents a scenario where the system will be designed to 750 MHz but the fiber and RF electronics will only be installed for 550 MHz operation. All of the fiber will be installed but only the nodes required for 550 MHz operation will be installed. The second column represents the costs of upgrading that 550 MHz plant to full 750 MHz operation. The final column shows the costs of building the system with 750 MHz operational capabilities from the onset. Tables similar to this are useful in the process of deciding when and where to make network investments.

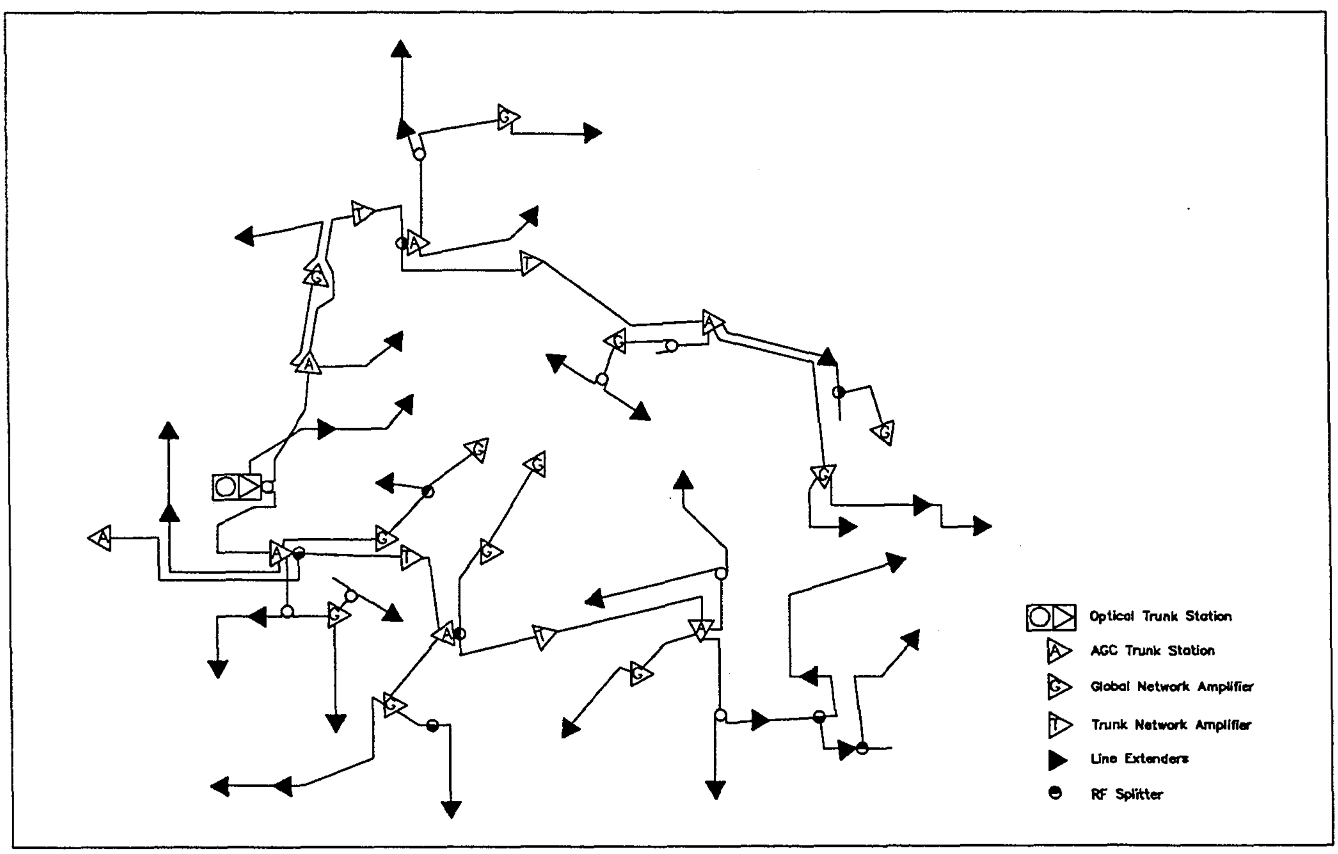


*Figure 2b.*

Table 1.
FTF Upgrade Cost Analysis

| | 550 FTF<br>750 DESIGN<br>550 ELECTRONICS | 750 FTF<br>UPGRADED<br>FROM 550 FTF | 750 FTF<br>750 DESIGN<br>750 ELECTRONICS |
|---|---|---|---|
| Fiber ($/Mile) | | | |
| Electronics | 645 | 530 | 1175 |
| Cable | 1260 | 0 | 1260 |
| Construction | 550 | 0 | 550 |
| | $2,455 | $530 | $2,985 |
| Coaxial ($/Mile) | | | |
| Electronics | 3105 | 2495 | 3510 |
| Cable | 4040 | 0 | 4040 |
| Construction | 3880 | 330 | 3880 |
| | $11,025 | $2,825 | $11,430 |

Table 2.
Fiber Backbone Upgrade Cost Analysis

| | 550 FBB<br>750 DESIGN<br>550 ELECTRONICS | 750 FBB<br>UPGRADED<br>FROM 550 FBB | 750 FBB<br>750 DESIGN<br>750 ELECTRONICS |
|---|---|---|---|
| Fiber ($/Mile) | | | |
| Electronics | 735 | 950 | 1685 |
| Cable | 1920 | 0 | 1920 |
| Construction | 620 | 0 | 620 |
| | $3,275 | $950 | $4,225 |
| Coaxial ($/Mile) | | | |
| Electronics | 2710 | 2375 | 3025 |
| Cable | 475 | 0 | 475 |
| Construction | 1980 | 330 | 1980 |
| | $5,165 | $2,705 | $5,480 |

## SUMMARY

Much looms on the CATV horizon. Advances in both technology and services promise to continue through the decade. As an industry, we can not afford to wait for the advances to slow or stabilize for that will never occur and we will be passed by by others. We must take advantage of what we know, apply it today and prepare for the future. We must develop a plan with contingencies for the future and continue to move forward. This paper, in the space of a few pages, has attempted to shed some light on how we might do this. It makes no pretense of providing a solution for a utopian network but hopes to evoke some thought and effort to that end.

# OPERATING PERSPECTIVES ON THE 1550 nm WAVELENGTH

By

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## ABSTRACT

*Since 1991, cable television operators have been weighing the relative merits of a new transmission window, the 1550 nanometer (nm) wavelength, versus operating at 1310 nm, where much of the industry's recent transmission efforts have been concentrated.*

*In conjunction with this, the feasibility of erbium-doped fiber-optic amplifiers (EDFAs) operating at 1550 nm also have been the topic of much discussion. The high output power and minimal distortions of these devices makes them ideal for amplitude modulation (AM) video distribution. As passive optical components, such as low loss, high port count (ie. 1x4, 1x8, and 1x16) splitters and optical amplifiers become commercially available in high volumes, the economic implications of their deployment in cable TV system rebuilds can be compelling.*

*Fiber-optic amplifiers can reduce the costs of an optical link and allow deeper penetration of fiber into the system. Due to the inherently lower attenuation of optical fiber at 1550 nm, nearly 40 percent lower relative to 1310 nm, amplifiers can permit longer fiber runs and increased sharing of transmitters via optical splitting. In addition, because the 1310 nm operating window is not currently utilized in an optical amplifier based system, operators can design their systems to deliver broadcast signals at 1550 nm today with the ability to offer narrowcast services at 1310 nm or other wavelengths in the future.*

*This paper will discuss the status of passive optical component technologies for splitters and the status of EDFAs. In addition, it will present the current state-of-the-art performance for some of these devices and scenarios for their cost-effective deployment in future applications.*

## BACKGROUND ON ARCHITECTURE EVOLUTION

Prior to 1988, most of the optical fiber used by the cable TV industry was installed for use in super-trunk applications. These applications called for better performance and reliability than could be delivered by coaxial cable or microwave transmission links.

The transmission technology of choice was frequency modulation (FM) over single-mode fiber. With the introduction of very linear, narrow linewidth, distributed feedback lasers (DFB) in early 1989, amplitude modulated (AM) fiber systems were developed and commercialized.

AM fiber-optic equipment is desired by cable TV operators because it is fully compatible with their current standard broadcast format and can be used economically in cable TV plants to upgrade system bandwidth and increase system reliability. However, AM video systems require much higher carrier to noise performance than FM and digital systems, and thus limit the amount of available optical power for link budgets.

Historically, most traditional coax-based cable TV system architectures were designed in a tree and branch configuration to deliver one-way, broadcast video signals. Cable TV designers and operators have constantly sought to get as much power as possible through their networks to the home. Therefore, a cable system's design goal is to maximize the power (RF or optical) presented to the cable plant while inducing the minimum amount of noise and distortion at the lowest cost per subscriber. This design goal typically is achieved by minimizing equipment costs and maximizing the revenue per watt of delivered power (by the laser source or RF equivalent).

Most AM fiber systems today operate at 1310 nm and are limited to distances of less than 25-30 kilometers (km). A major impetus for operating at 1550 nm is to take advantage of optical fiber's lower attenuation at 1550 nm. This could extend the maximum transmission distance of a system out to 40 or 50 km. In addition, optical amplifiers, which operate at 1550 nm, could enable more receivers to be shared by a common laser transmitter, thus potentially reducing the overall system cost.

Corning envisions two applications as appropriate for 1550 nm AM video technology in the near term.

First, initial 1550 nm AM systems are likely to be used for trunking applications, where longer distances can be achieved than with current 1310 nm AM systems. For these applications, it may be appropriate to install optical amplifiers with single-mode dispersion-shifted fiber to reduce significantly the effects of dispersion at 1550 nm.

Dispersion is a key performance parameter affecting current and future capabilities of fiber-based AM video cable TV systems.

In analog transmission, dispersion causes a slightly distorted waveform to become significantly distorted more rapidly. The nature of an AM video signal makes it very susceptible to distortions created by laser chirp. This phenomenon occurs when there is a shift of laser ouput power to slightly different wavelengths.

Because AM cable TV systems rely on DFB lasers with very narrow linewidths, a small amount of laser chirp can significantly distort the ouput video signal. The presence of dispersion enhances this effect and shows up as second order harmonics (or "beats") and is commonly referred to as composite second order (CSO).

Some solutions exist today for this effect. Equipment manufacturers can electrically or optically compensate their 1550 nm AM systems for the effects of dispersion, enabling successful operation at 1550 nm on standard single-mode fiber. Another alternative could be to design channel loading per fiber by octave, thus eliminating the effects of CSO distortions that are enhanced by dispersion.

The second 1550 nm AM system application offers deeper fiber penetration at a lower overall cost per subscriber by employing optical amplifiers. In this case, optical amplifiers would generate high ouput power levels, allowing the laser transmitter to be shared among several optical receivers.

This cost sharing allows for greater penetration of fiber closer to the home and reduces the total installed fiber cost. When optical amplifiers are installed in the headend, then compensation techniques can be used to minimize CSO distortions.

When 1550 nm AM video systems with optical amplifiers are deployed eventually in the field as part of fiber-to-the-subscriber systems, the effects of dispersion on CSO may be minimal, since the maximum fiber lengths would likely be reduced to less than two kilometers (km).

Because this application holds the promise of significant economic benefits for operators, a more detailed examination of two enabling component technologies for 1550 nm -- planar ion exchange couplers and optical amplifiers -- is warranted.

## COMPONENT TECHNOLOGIES

### Planar Ion Exchange Coupler Technology

In addition to offering system cost savings potential, the planar ion exchange technology for fabricating passive optical components also offers performance benefits. This approach is a radical departure from more traditional coupler technologies.

The ability to create higher port count 1xN (ie. 1x4, 1x8, or 1x16) splitters by other methods, commonly referred to as fused biconic taper, usually is achieved by splicing several fused 1x2 devices together to create a unit with the desired number of ouput ports. However, the planar ion exchange coupler fabrication process begins with an optical substrate glass rather than optical fibers (Fig. 1).

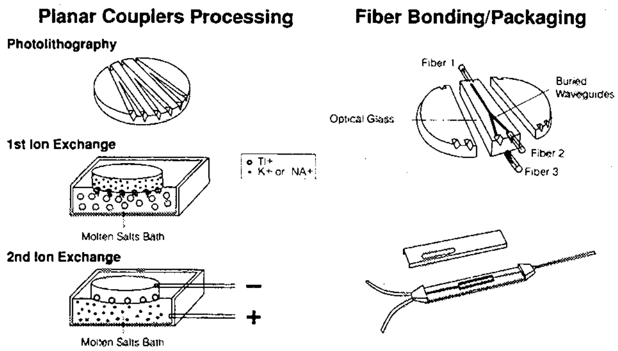


Fig. 1. Planar Ion Exchange Technology

Using photolithography, the pattern of an optical circuit is imposed on the surface of the optical substrate material. Two chemical ion exchange processes then are used to actually create a waveguide structure in the glass substrate itself. Following these steps, the processed substrate wafer, which consists of numerous individual components, is cut into discrete devices or chips. A precision automated technique for mechanically attaching the input and output fibers to each optical chip is used and the resultant coupler then is packaged.

With this technology, it is possible to make low loss, dual window wideband splitters in 1x2, 1x4, 1x8, and, most recently demonstrated, 1x16 configurations (Fig. 2). These devices are extremely small relative to cascaded devices using fused coupler. As an example, a planar 1x8 is approximately 1/250th the size of a cascaded fused 1x8 (Fig. 3).

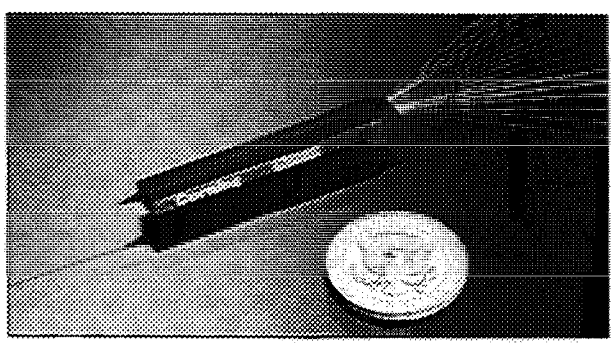

Fig. 2. Miniature 1x16

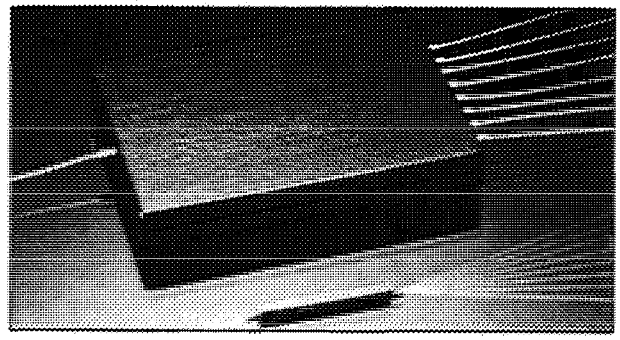

Fig. 3. Size Comparison of Planar and Fused Biconic Taper Couplers

Many new developments in coupler design have been achieved over the past two years through advancements in Corning's planar fabrication technology. Improvements in optical circuit design and waveguide fabrication have resulted in planar couplers with enhanced wide-band achromatic optical performance. Double window 1x4s and 1x8s (1310 nm and 1550 nm) have been achieved with very low insertion losses (Figs. 4, 5).

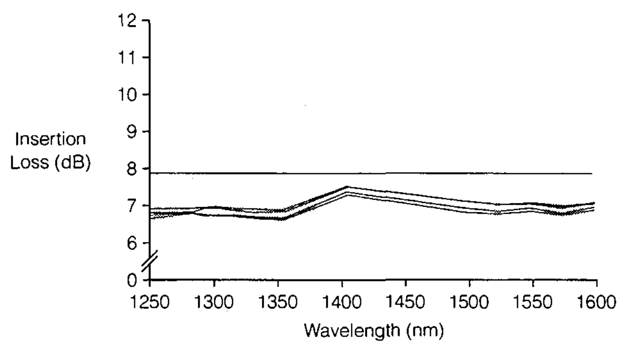


Fig. 4. Typical Insertion Loss - Corning Single-Mode 1x4 Tree Coupler

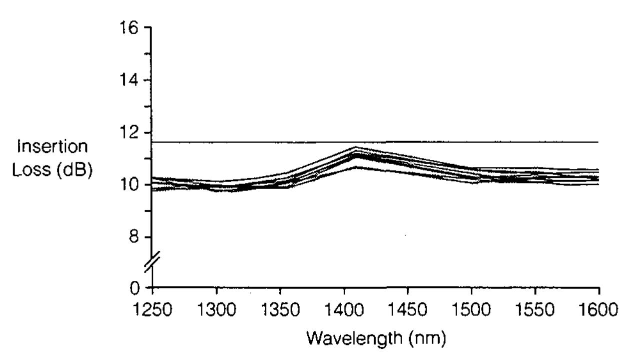


Fig. 5. Typical Insertion Loss - Corning Single-Mode 1x8 Tree Coupler

For 1x4 couplers, typical insertion losses have been measured at 6.8 dB for 1310 nm and 7.0 dB for 1550 nm. In addition, 1x8 couplers have been fabricated with optical losses of 10.3 dB for 1310 nm and 10.9 dB for 1550 nm.

These planar devices have typical port-to-port uniformity of about 0.3 dB for 1x4s and 0.7 dB for 1x8s (Figs. 6, 7).

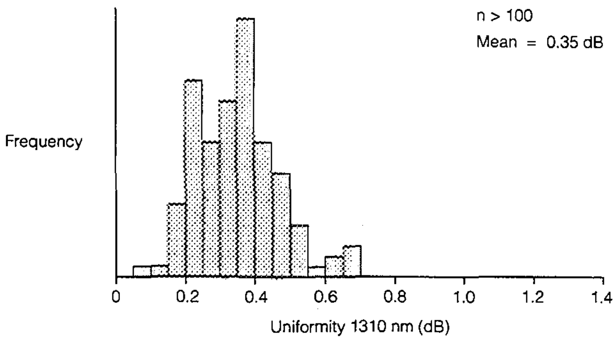


Fig. 6. Uniformity at 1310 nm, 1x4 Double Window Splitter

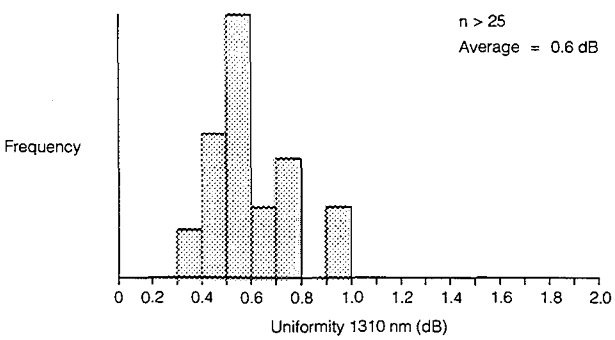


Fig. 7. Uniformity at 1310 nm, 1x8 Double Window Splitter

Planar 1x16 couplers are now in development. Corning's planar technology preserves optical performance and reduces size through passive integration of multiple Y-junctions (Fig. 8). For these developmental products, low insertion losses, on the order of 14 dB and port-to-port uniformity of 1.5 dB, have been achieved (Fig. 9).

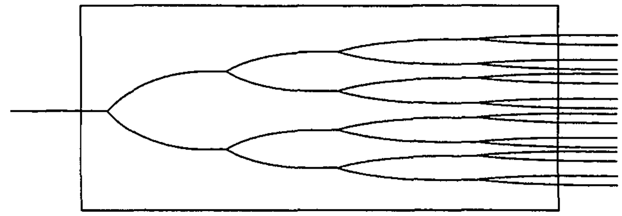

Fig. 8. 1x16 Splitter

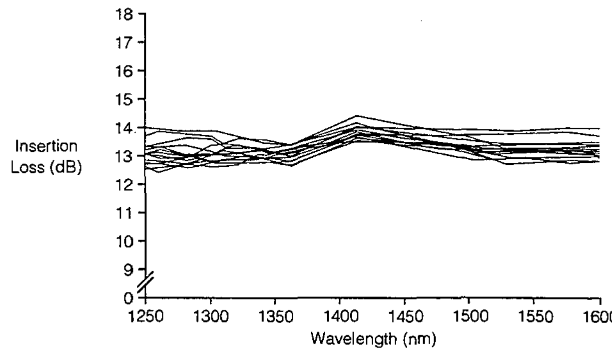


Fig. 9. Typical Insertion Loss - Developmental 1x16 Corning™ Single-Mode Splitter

## Erbium-Doped Fiber Amplifier Technology

This second fiber-optic component technology, when combined with the planar coupler technology previously discussed, presents the greatest opportunity for overall system cost reduction and future upgradeability. The EDFA represents a synthesis of optical components that could have a significant impact on the way cable TV systems are designed.

Although the majority of AM video systems currently operate at 1310 nm, the increased availability of linear DFB lasers operating at 1550 nm, make EDFA technology an attractive alternative for cable operators looking to install optical fiber cost-effectively.

## EDFA Components

An EDFA consists of several distinct elements that can be better understood from a functional standpoint if viewed in block diagram form (Fig. 10). These include: environmental protection, status monitoring, back-up power, and gain control. These functions are absolutely necessary to achieve a fully useful and functional amplifier that will survive in the outside plant environment. While much of this functionality relates to the operation and internal control of the device, particular attention should be given to the element containing the source of the optical gain.

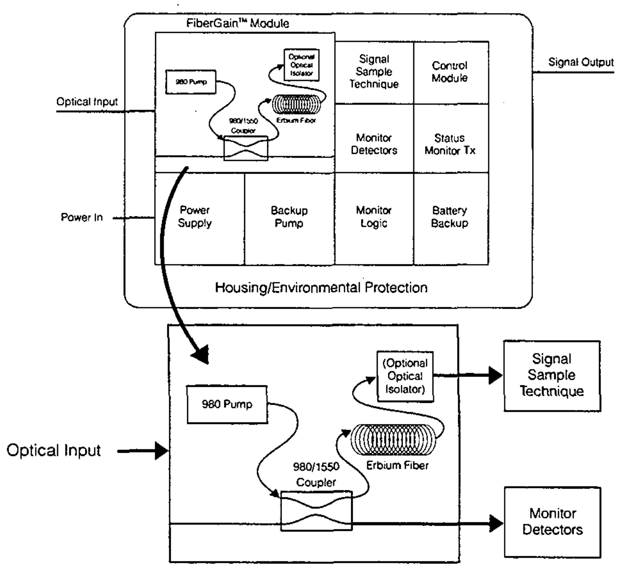


Fig. 10. Erbium-Doped Fiber-Optic Amplifier Schematic

In 1991, Corning introduced its FiberGain™ Module, an optical gain element that consists of one or two 980 nm pump laser(s), one or two wavelength division multiplexer (WDM) coupler(s), and approximately 20 meters of erbium-doped single-mode fiber (Fig. 11). The WDM coupler serves the purpose of combining the input optical signal (1550 nm) and the pump signal (980 nm) into the erbium-doped single-mode fiber where the conversion of pump power light to amplified signal light occurs.

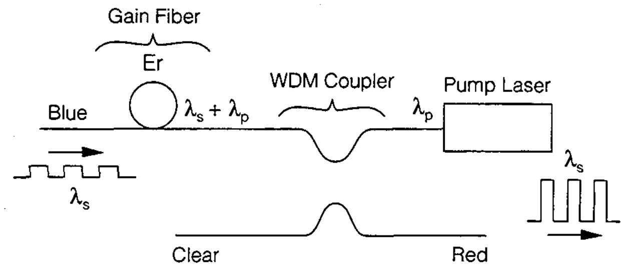


Fig. 11. Fiber-Optic Gain Block

In actual operation, the optical signal would enter the optical gain block (Fig. 12) on an input leg of the WDM coupler. Pump light from the diode laser is delivered on the other WDM input. The input and pump signal is then coupled into the erbium fiber and the input signal undergoes between 15 to 30 dB gain. Saturated ouput power levels as high as 12 to 15 dBm are possible.

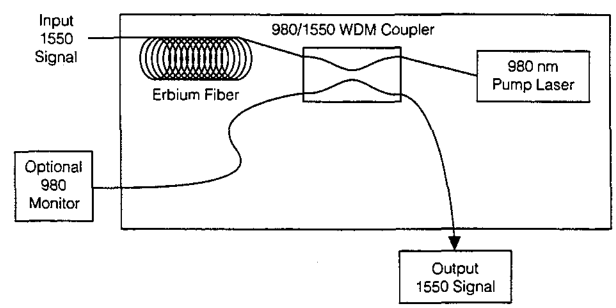


Fig. 12. Corning FiberGain™ Module

## PLANAR COUPLER AND EDFA APPLICATIONS

The most likely applications for the EDFA being considered by cable TV operators are for AM supertrunking and shared distribution. In these approaches, the amplifier is configured as a power amplifier where the laser transmitter's output is immediately boosted and transmitted over long distances or split along multiple paths to serve many nodes.

With 1310 nm AM technology being deployed successfully today, cable operators may ask why they should consider the 1550 nm window. With the development and commercialization of optical amplifiers and passive splitters, the ability to share costly transmitters is greatly enhanced. This will not only permit operators to reduce the initial cost of their optical links, but also allow them to extend fiber's reach further into their systems. In addition, the lower attenuation of optical fiber at 1550 nm, nearly 40 percent lower than the 1310 nm window, will allow operators to lengthen their fiber runs and increase the amount of optical splitting.

In order to take the maximum advantage of the optical fiber and these new optical components technologies, operators will want to consider deploying the EDFA and passive splitters in architectures that vary dramatically from the traditional tree-and-branch. One such approach is the cable TV star architecture (Fig. 13). While the specific details of the actual implementation may vary from system to system, there are two key attributes worth considering.

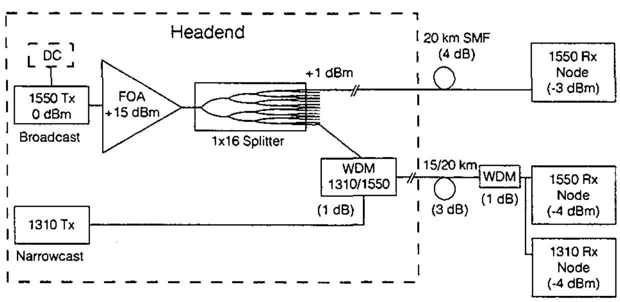


Fig. 13. Cable TV Star Architecture, Using An Optical Amplifier

First, because all splitting is performed in the headend, this architecture will permit operators to maintain home run fibers to and from their receive nodes. Second, while the 1550 nm window is used to deliver AM video, the 1310 nm window remains available to support a host of upgrade possibilities, including narrowcasting, digitally compressed pay-per-view, or video on demand.

To understand the tangible benefits of this cable TV star architecture, it's useful to explore the example shown in figure 13 in greater detail. In this approach, a 1550 nm DFB with an output power of approximately 0 dBm (1 mW) is shown with some form of dispersion compensation (DC). This compensation could be either electronic pre-distortion of the laser transmitter or optical dispersion compensation placed after the transmitter. A key aspect to note is the relatively low output required from the transmitter.

By contrast, 1310 nm DFB systems require almost twice the output power due to higher loss at 1310 nm. Because the EDFA can provide such high output power with very little added distortion and noise, it is possible that lower output 1550 nm devices in volume, may be available at lower prices than 1310 nm devices in the future. The next element of this architecture is an EDFA with a +15 dBm (32 mW) output power.

The output of the amplifier is input to a monolithic 1x16 splitter with a total insertion loss of 14 dB. The resulting output power on the 16 splitter output ports would thus be +1 dBm. Assuming a receiver sensitivity at the 1550 nm receive nodes of -3 dBm to -4 dBm gives an optical budget of 4 dB to 5 dB. At 1550 nm, this equates to link distance of 15 to 20 kilometers (km).

## ECONOMIC BENEFITS ANALYSIS

To understand how the previously discussed cable TV star architecture example offers operators the economic impetus to deploy these 1550 nm technologies, it's useful to review the following rough economic analysis comparing the deployment of a star architecture at the 1310 nm and 1550 nm windows.

Using the link length of 15 km will provide an equalization factor with which to make an "apples to apples" comparison of the 1310 nm and 1550 nm approaches. The optical loss of 15 km of single-mode fiber at 1310 nm is about 5.5 dB. Again assuming a -3 dBm receive sensitivity, one would need a 1310 nm DFB with an output of a 9 dBm (8 mW) with no splitting in order to meet the optical budget for just one link.

A comparison of the prices for this 1310 nm and 1550 nm transmission equipment alone tells the story (Fig. 14). Assuming the small volume pricing for a 1550 nm DFB with 0-5 dBm output is $17,000, an EDFA with +15 dBm output power is $40,000, and a 1x16 splitter with 14 dB insertion loss is $2,500, the total equipment cost is $59,500. Dividing this total by the number of nodes served (16) gives a cost per link of roughly $3,700.

- 1550 nm assumptions
  - 1550 nm DFB: 0-5 dBm output power ($17,000)
  - Double pumped EDFA: +15 dBM output power ($40,000)
  - 1x16 splitter: 14 dB insertion loss ($2,500)

  $\boxed{\text{Cost per link} \sim \$3{,}700}$

- 1310 nm assumptions
  - 1310 nm DFB: +9 dBm output power ($10,000)

  $\boxed{\text{Cost per link} \sim \$10{,}000}$

Fig. 14. Cable TV Star Architecture Economics

For the 1310 nm case, the analysis is straightforward since splitting is not possible. In large volumes, one might assume $10,000 for the 9 dBm transmitter. For discussion purposes, large volumes must be assumed since transmitters capable of this kind of high output power are the exception, not the rule. Thus, the capital outlay for the 1310 nm equipment is nearly three times that of the 1550 nm technology.

While not exhaustive, this analysis does provide an example of how these 1550 nm technologies can permit operators to create low cost, upgradeable, fiber-optic distribution networks today.

Clearly, with higher volumes, the component technologies described above will only continue to further enhance the cost-effectiveness of cable TV star architectures.

In addition, the continued development of such products as planar ion exchange splitters, and fiber-optic amplifiers, which are available today, also will continue to allow for the refinement of advanced system architectures and services that will permit cable operators to bring fiber ever closer to the customer's home.

# Optical Amplifier Basic Properties And System Modeling: A Simple Tutorial

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Abstract
Erbium-doped fiber amplifiers have received much interest for CATV because of their high output power, low distortion, and low noise capability. These devices have only recently become commercially available and probably are not well understood by those who have not been involved with them from a research or applications viewpoint. This paper is intended to present a simple coverage of the principles of operation and characteristics of these amplifiers. Signal gain, saturation output power, and noise properties are discussed. Equations for optical amplifier noise figure are developed and typical data for CATV are given.

## INTRODUCTION

The rapid developments in fiber-amplifier technology now enable employment of these devices for telecommunications and CATV with a substantial advancement in system capabilities. For telecommunications, optical amplifiers can easily compensate for fiber loss in very long repeatered transmission links. Experiments have shown that it is feasible to transmit at multi-Gbit/s rates in ultra-long distance amplifier cascades with losses of a few thousand dB. For CATV, although the expectations are not so astounding, erbium-doped fiber amplifiers (EDFA's) can be used as power amplifiers to boost the output of the laser source and enable a large increase in the optical loss budget. However, EDFA's are expected to have limited application, if any, as repeaters because of the high input signal power required to achieve acceptable carrier-to-noise ratios for AM CATV.

This paper begins with a simple discussion of noise in the electrical and optical domain. Next is a general discussion of the basic principles of operation and characteristics of an EDFA. The noise figure is derived and characteristics for CATV are shown.

## THERMAL NOISE

Thermal noise is an electrical noise phenomena that occurs in a conductor or resistor due to the continual random scattering of free electrons with the molecules. This electron scattering of free electrons is responsible for the electrical resistance of the material. Each flight of an electron between collisions with molecules constitutes a short pulse of current. This random motion of charges results in a thermally-induced noise voltage across the terminals of the resistor. This effect, also called Johnson or Nyquist noise, was first observed by J. B. Johnson in 1927, and a theoretical treatment was given by H. Nyquist in 1928. Accordingly, for a thermal noise source the available power in a 1-Hz bandwidth is given by

$$p_{th} = kT \qquad [\mathrm{W/Hz}] \qquad (1)$$

where $k$ is Boltzmann's constant ($1.38 \cdot 10^{-23}$ joule/°K, and $T$ is the absolute temperature in degrees Kelvin. At room temperature, $kT = 4.0 \cdot 10^{-21}$ W/Hz (-174 dBm/Hz). In a 75 ohm system and for a bandwidth of 4 MHz, the thermal noise voltage is -59.2 dBmV, which is the familiar number used in CATV carrier-to-noise ratio calculations. At room temperature, the power density spectrum is theoretically constant up to ~3000 Ghz and decreases rapidly for higher frequencies. At optical frequencies, quantum noise dominates, as discussed in later sections.

## SHOT NOISE

Shot noise is a noise current caused by the discrete nature of electron flow. It was first observed in the anode current of vacuum-tube amplifiers and was described by W. Schottky in 1918. In a shot-noise source, electrons are generated or released randomly in time. This random flow of electrons results in a mean-square noise current in a 1-Hz bandwidth of

$$i_{sh}^2 = 2qI_{dc} \qquad [\mathrm{A}^2/\mathrm{Hz}] \tag{2}$$

where $q$ is the charge of an electron ($1.6\cdot10^{-19}$ coulomb), and $I_{dc}$ is the direct current through the device in amperes. Shot noise occurs in most active devices, but is not generated in linear resistive and passive networks due to current flow. Unlike thermal noise, shot noise is not a function of temperature.

## OPTICAL NOISE DUE TO THE PARTICLE NATURE OF LIGHT

Light, like radio waves, is a form of electromagnetic radiation. At optical frequencies, the quantum-mechanical effects become important, and the discrete nature of light must be considered. The quantum particle of light is the photon, and the energy of each photon is $h\nu$ (joules), where $h$ is Planck's constant ($6.624\cdot10^{-34}$ joule·sec) and $\nu$ is the optical frequency. The power of an optical signal is thus

$$P = Nh\nu \tag{3}$$

where $N$ is the average number of photons emitted per second by the optical source. However, photons are not emitted at a constant rate, but, in fact, the emission times of the photons vary randomly. From Poisson's probability distribution law, it follows that the variance, or mean-square deviation in the number of photons emitted per second, is equal to the mean number of photons per second:

$$<\Delta N^2> = N. \tag{4}$$

The minimum detected number of quanta is considered to be that for which the rms fluctuation is equal to the average value. Thus,

$$(N_{min})^{1/2} = N_{min} \tag{5}$$

for which $N = 1$. Consequently, for an ideal photon counter, the minimum detectable number of photons per second in a 1-Hz bandwidth is 1, and the minimum detectable power is [1]

$$p_{min} = h\nu \qquad [\mathrm{W/Hz}]. \tag{6}$$

Note that in the following section it is determined that with an ideal photodetector the minimum detectable optical power is $2h\nu$. The factor of 2 difference occurs because the equivalent optical bandwidth (double-sided) is twice the baseband (single-sided) bandwidth. This quantum noise provides the fundamental limitation to performance of optical systems. Because of the high energy of a photon, the minimum detectable optical power is much higher than it is as limited by $kT$ in the electrical domain. At a wavelength of 1550 nm, $h\nu$ = -159 dBm/Hz, whereas $kT$ = -174 dBm/Hz at 25°C.

## PHOTODETECTION PROCESS

The photodetector commonly used in CATV applications is the PIN photodiode. Incident photons are absorbed in the depletion region creating electron hole-pairs. Under the influence of the reverse bias electric field, the electrons and holes drift in opposite directions creating a displacement current in the external circuit. For an ideal photodiode, one electron is emitted for each incident photon. The efficiency of generating electrons is given by the quantum efficiency of the detector, which is defined as

$$\eta = \frac{\text{number of electrons emitted}}{\text{number of incident photons}}. \tag{7}$$

Since optical power = (number of photons/sec)·$h\nu$, and detector current = (number of electrons/sec)$\cdot q$, it follows that

$$I_{dc} = P\left(\frac{\eta q}{h\nu}\right) \tag{8}$$

where $P$ is incident optical power (W), and $I_{dc}$ is detector current (A). Responsivity $R$ of the detector is defined as

$$R = \frac{\eta q}{h\nu}. \tag{9}$$

Noise is inherent in the output of the photodetector. The origin of this noise can be treated in two ways. In the quantum treatment, the optical field is quantized into photons and each photon gives rise to an electron with probability $\eta$. Due to the random occurrence in time of the photons, a noise current is generated which limits the minimum detectable signal as discussed in the previous section. In the semiclassical approach, a constant electromagnetic field interacts with atoms in the photodiode and generates the photodiode current. Shot noise is generated in this process. Each electron-hole pair results in a single pulse of detector current of charge $q$. The total current is the superposition of these pulses occurring randomly in time. Because of this random fluctuation, shot noise results just as Schottky observed it did in vacuum-tube amplifiers. Photodetector shot noise current is

$$\begin{aligned} i_{sn}^2 &= 2qI_{dc} \\ &= 2qPR \qquad [\mathrm{A^2/Hz}] \end{aligned} \tag{10}$$

where $I_{dc}$ is the dc current.

The minimum detectable optical power is defined for quantum efficiency = 1 as the minimum power for which the rms value of shot noise is equal to the dc current. Thus,

$$P_{min} = 2h\nu \tag{11}$$

This is the same as given in the previous section. The question as to whether the limit to sensitivity is imposed by photodetector-generated shot noise or the fluctuations in the incident photon number is academic. The resultant photodiode current is the same in either case [1].

ERBIUM-DOPED FIBER AMPLIFIERS

Amplification occurs in an erbium-doped fiber amplifier (EDFA) due to the photoluminescent properties of the rare-earth element erbium concentrated in the core of the fiber. The atomic energy levels of erbium allow erbium to absorb energy at any of several wavelengths and release energy in the 1550-nm range. In the interaction of photons and atoms, the frequency of absorbed or emitted radiation $\nu$ is related to the difference in energy $E$ between the higher and the lower energy states $E_2$ and $E_1$ by the expression

$$\begin{aligned} E &= E_2 - E_1 \\ &= h\nu. \end{aligned} \tag{12}$$

Discrete energy states correspond to particular energy levels of the electrons within the atom relative to the nucleus. A single electron transition between two energy levels represents a change in energy suitable for the absorption or emission of a photon.

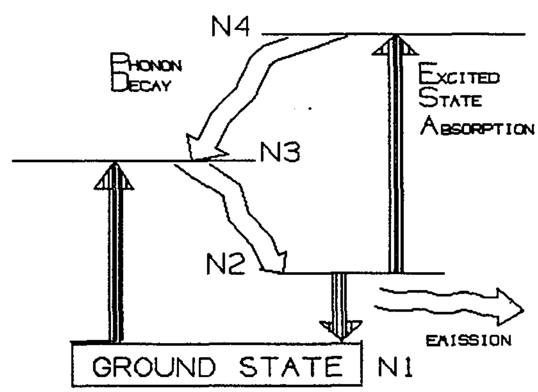


Fig. 1. Erbium three-level system with excited-state absorption.

Erbium has a three level system, the ground level ($N_1$), the metastable level ($N_2$), and a higher level ($N_3$) as shown in Fig. 1. In an EDFA the absorbed energy comes from a pump laser. The optimum pump wavelengths for EDFA's are those for which there is no excited state absorption (ESA). These wavelengths are 980 nm, 1480 nm, or 1047 nm. ESA creates a fourth level ($N_4$) by the process of exciting or pumping the electrons from $N_2$ to $N_4$. This depletes the number of electrons or population at the $N_2$ level creating a need for increased pump power, which will increase the pumping rate allowing the system to maintain the same degree of population inversion. Population inversion occurs when the pumped electrons transition from $N_1$ to $N_3$ and then rapidly decay to the $N_2$ level. The amount of time they spend at the $N_2$ level is controlled by the erbium ion fluorescence lifetime which is approximately 10 ms. Therefore, for complete population inversion to occur, the rate at which the electrons are transitioning from $N_1$ to $N_2$ and then held in the $N_2$ level must exceed the number of electrons transitioning from $N_2$ to $N_1$ at any given time. Ideally, the total number of electrons associated with the erbium atoms available per length of fiber, must remain in the $N_2$ level, and just as quickly as one decays to the lower $N_1$ level another one must already be at the $N_2$ level. Their decay from $N_2$ to $N_1$ can occur in two ways:

(a) by spontaneous emission in which the atoms return to $N_1$ in an entirely random manner, creating optical noise from $\sim 1530$ nm to $\sim 1560$ nm, and:

(b) by stimulated emission in which an incident photon causes the release of a second photon of the same energy. Stimulated emission produces coherent amplification or gain.

By this process the optical input signal and noise generated by spontaneous emission are amplified. At the amplifier output there are two optical components: the desired amplified input signal and amplified spontaneous emission (ASE), which is optical noise.

For the uniform inversion model, which means that the pumping rate throughout the entire length of erbium-doped fiber is sufficient to keep $N_2$ completely populated, the following equations apply:

$$\begin{aligned} p_{ase} &= 2\eta_{sp}(G-1+X)h\nu \\ &\approx 2\eta_{sp}(G-1)h\nu \quad \text{for } (G-1)\gg X \\ \eta_{sp} &= \frac{N_2}{N_2 - N_1}. \end{aligned} \tag{13}$$

$X$ is the excess noise parameter of the EDFA. $p_{ase}$ is the ASE noise power in both polarization modes emitted by the EDFA. Eqn. 13 gives the optical noise spectral density (W/Hz) at the output of the erbium-doped fiber. This noise field has all of the statistical properties of thermal noise; it is additive and Gaussian with respect to the amplified input field [2]. $\eta_{sp}$ is the inversion parameter, and with complete population inversion its value is equal to one. To obtain complete population inversion everywhere along the fiber, the pump power into the fiber must be at least twenty times greater than the pump power threshold at which population inversion begins [3]. Qualifying this further, the product of the pumping rate and the fluorescence lifetime of the $N_2$ level must be much greater than one, which will occur with sufficient remnant pump power. Without complete population inversion, $\eta_{sp}$ increases due to the decrease in number of electrons in level $N_2$. This causes an increase in the amplifier noise figure as given by the equation NF $= 2\eta_{sp}$ as derived later.

## EDFA IMPLEMENTATION

The amplifier gain medium is the erbium-doped fiber, whose length varies from a few meters to about 100 meters. The length can be optimized for either gain or noise characteristics. Also, the optimum length increases with pump power and decreases with the signal power [4]. The fibers used typically have a codopant such as germanium, aluminum, ytterbium, or phosphorus to change various fiber characteristics. The fiber is designed to utilize pump power as efficiently as possible, and has a mode-field diameter much smaller than that of standard single-mode fiber. This results in relatively high splice loss, typically about 1 dB.

A general block diagram of an EDFA is shown in Fig. 2. Two pump lasers are shown, although a single pump laser can be used either at the erbium fiber input (co-propagating pump) or at the erbium fiber output (counter-propagating pump). In either configuration, the pump power should be greater than approximately 100 mW to eliminate the effect of absorption allowing complete population inversion and identical ASE powers in the forward and reverse directions. This is important particularly in a system that utilizes a double pump configuration because if one fails, without sufficient pump power available to the Er-doped fiber, the EDFA noise and gain characteristics will change drastically.

Isolators are used to prevent reflected signals from being re-amplified. The reflections would induce intensity noise and intermodulation distortion by distorting the shape of the ASE spectra. At high pump powers, such as with a double-pump configuration, the reflections can cause laser oscillation or cause the higher ASE powers to self-saturate the amplifier and reduce the pump efficiency [5].

The general input-output signal characteristic of

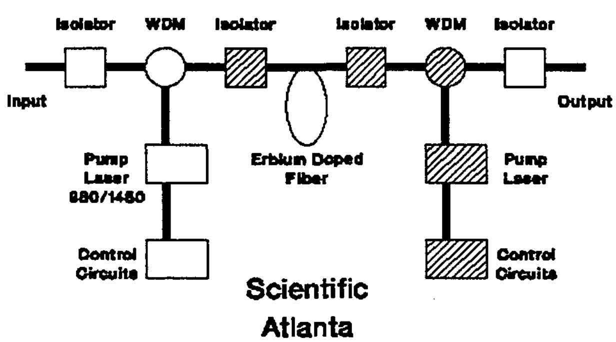


Fig. 2. Block diagram of EDFA with double pump.

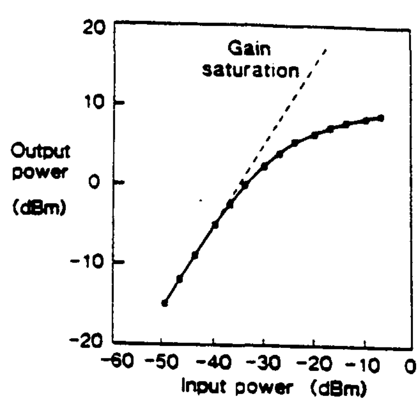


Fig. 3. Amplifier input-output characteristics using a 980-nm pump laser [8].

an EDFA is shown in Fig. 3. As the input signal increases the gain decreases as the amplifier goes into saturation or gain compression. When an EDFA is driven sufficiently hard into saturation the output power tends toward a limiting value ($P_{sat}$) which is the maximum output power [6]. Saturated output increases approximately linearly with pump power. As the EDFA goes deeper into saturation, more and more of the pump light is converted into signal. As booster or power amplifiers, EDFA's are operated deeply in saturation in order to deliver high output power. The saturated output power of commercially-available amplifiers for use as booster amplifiers is in the range of about 10 dBm to a little over 20 dBm. High differential pump-to-signal quantum efficiencies have been obtained. There exists a performance tradeoff for an EDFA to have low noise amplification and high pumping efficiency. To obtain both, the use of short pump wavelengths are preferred in order to lower the ASE power and obtain high efficiencies by strongly saturating the amplifier. At the 980-nm pump wavelength, 47-mW saturation power (16.7 dBm) has been obtained with 100 mW of pump power [7].

Due to the temporal properties of the erbium fiber, the signal may experience intermodulation distortion in a single channel system or saturation induced crosstalk in a multiple channel system. These transient effects are produced when operating in gain saturation. One effect of signal induced gain saturation is to increase the pump attenuation thus reducing the population inversion toward the output of the EDFA which changes the gain spectrum and creates noise. The gain spectra will shift under nonuniform gain saturation conditions depending on such factors as fiber type, fiber length, pump power, pump wavelength, and signal wavelength [6]. For CATV application, these transient gain effects are dampened by the long fluorescent lifetime in erbium which sets a lower frequency limit of approximately 100 kHz by filtering out any high frequency modulation of the EDFA's gain. Fluctuations in the gain spectra are inversely proportional to the pumping rate. Therefore, the stronger the pump power, the smaller the fluctuation or the smaller the gain-recovery time constant which minimizes crosstalk and distortion. The slow gain response combined with the third-order susceptibility of the erbium fiber also prevents the occurrence of intermodulation distortion in multichannel systems [9]. Other techniques useful in eliminating the transient gain effects are by utilizing filters at the erbium fiber to equalize the gain spectrum and by utilizing an all-optical feedback loop to the EDFA [4].

The output spectrum of an EDFA is given in Fig. 4 with no input signal, and in Fig. 5 for a high input signal level (5 dBm). In Fig. 5, the amplifier is operating 20 dB into saturation (small signal gain reduced 20 dB).

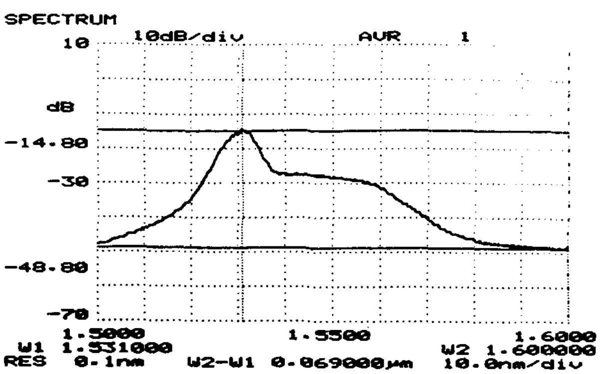


Fig. 4. EDFA output spectrum with no input signal.

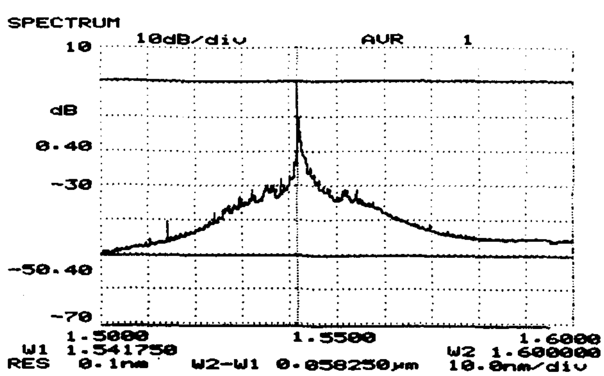


Fig. 5. EDFA output spectrum with 5-dBm input signal power.

SIGNAL AND NOISE CHARACTERISTICS

In direct-detection of an optical signal, the output current of the photodetector is directly proportional to optical power. However, optical power is proportional to the square of the magnitude of the E field and the detector responds to the square of the E field. The detector is in reality a square-law detector, and the fact that the output of a photodetector changes 2 dB per dB change in input power is very familiar. In a system with an optical amplifier, this square-law characteristic causes additional noise currents to be produced at the output of the photodetector: these are called signal-spontaneous (s-sp) beat noise, and spontaneous-spontaneous (sp-sp) beat noise.

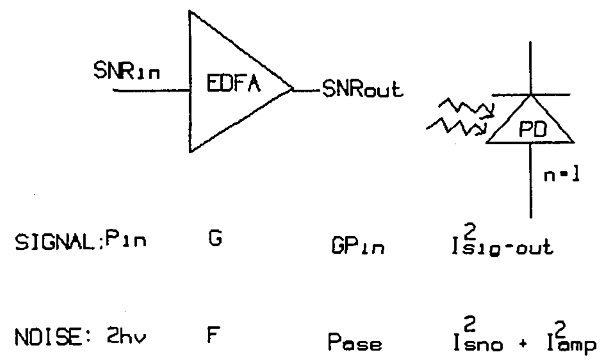


Fig. 6. Diagram for noise figure calculations.

Consider an optical amplifier directly connected to a photodetector as illustrated in Fig. 6. $P_{in}$ is the input signal power and $G$ is amplifier gain. There are two components at the output of the amplifier: amplified signal $GP_{in}$ and amplified spontaneous emission noise $p_{ase}$. Since the photodetector is a square-law detector, heterodyning and mixing occur. The optical noise heterodynes with the signal causing the noise spectrum to be translated to dc. This noise is signal-spontaneous beat noise. In an EDFA, the spontaneous emission is randomly polarized and only the noise that is co-polarized with the signal, $\frac{1}{2}p_{ase}$, is heterodyned. Likewise, components of the spontaneous emission mix with each other in the square-law detection process to produce spontaneous-spontaneous beat-noise currents.

The signal-spontaneous beat-noise current is [10]

$$i^2_{s\_sp} = 2P_{in}Gp_{ase}(\frac{\eta q}{h\nu})^2. \qquad (14)$$

Assuming the amplified spontaneous emission noise $p_{ase}$ is constant over an optical bandwidth $B_o$, the spontaneous-spontaneous beat-noise current is [10]

$$i^2_{sp\_sp} = p^2_{ase}B_o(\frac{\eta q}{h\nu})^2. \qquad (15)$$

Amplifier Noise Figure

The amplifier noise figure as defined and developed herein follows that presented in reference [11] and is the same as in [12]. The noise factor $F$ of an amplifier is defined as

$$F \triangleq \frac{\mathrm{SNR}_{in}}{\mathrm{SNR}_{out}} \qquad (16)$$

where $\mathrm{SNR}_{in}$ and $\mathrm{SNR}_{out}$ are the signal-to-noise ratios at the input and output of the amplifier. The input noise, by definition, is that from an ideal source. At optical frequencies, the input noise is quantum noise ($2h\nu$), whereas in the electrical domain it is $KT$. For the optical amplifier, the SNR's are defined as electrical quantities measured at the output of a ideal photodetector ($\eta$ = 1). Signal currents are expressed in $A^2$, and noise currents are in $A^2$/Hz.

$\mathrm{SNR}_{in}$ is given with the optical source directly connected to the photodetector. The signal measured at the output of the photodetector is

$$I^2_{s\_in} = [P_{in}\frac{q}{h\nu}]^2. \qquad (17)$$

The photodetector output noise is shot noise

$$2q[P_{in}\frac{q}{h\nu}] \qquad (18)$$

and

$$\mathrm{SNR}_{in} = \frac{P_{in}}{2h\nu}. \qquad (19)$$

This result is the same as for $\mathrm{SNR}_{in}$ in the optical domain. With the amplifier connected directly to the photodetector, the output noise currents consist of two components: shot noise due to the amplified input signal, and total noise current due to the optical amplifier. The output signal is

$$I^2_{s\_out} = [GP_{in}\frac{q}{h\nu}]^2 \qquad (20)$$

and the output noise currents are

$$i^2_{sno} + i^2_{amp} = 2q[P_{in}\frac{q}{h\nu}] + i^2_{amp} \qquad (21)$$

where $i^2_{sno}$ is shot noise due to the amplified signal and $i^2_{amp}$ is optical amplifier noise. From the preceding equations, amplifier noise factor is

$$\begin{aligned} F &= \frac{P_{in}}{2h\nu}\frac{i^2_{sno} + i^2_{amp}}{(GP_{in}\frac{q}{h\nu})^2} \\ &= \frac{1}{G}[1 + \frac{i^2_{amp}}{i^2_{sno}}] \qquad (22) \\ &\approx \frac{1}{G}\frac{i^2_{amp}}{i^2_{sno}} \qquad \text{for } G \gg 1. \end{aligned}$$

Noise figure is noise factor expressed in dB:

$$\mathrm{NF} = 10\log_{10}F. \qquad (23)$$

For most practical CATV applications, signal-spontaneous beat noise dominates and the noise factor is then

$$\begin{aligned} F &\approx \frac{1}{G}\frac{4P_{in}G\eta_{sp}(G-1)\frac{q^2}{h\nu}}{2q(GP_{in}\frac{q}{h\nu})} \qquad (24) \\ &\approx 2\eta_{sp}(\frac{G-1}{G}) \\ &\approx 2\eta_{sp} \qquad \text{for } G \gg 1. \end{aligned}$$

The last expression is the definition that is commonly used for EDFA noise factor. The minimum value for $\eta_{sp}$ is one, for which the quantum-limit noise figure is 3 dB. Measured values approaching 3 dB have been achieved with 980-nm pumping, but the minimum achievable value for 1480-nm pumping is about 4 dB. The higher noise figure in the 1480-nm pump band is due to its proximity to the 1550-nm signal emission band, which causes the erbium three-level system to behave differently [13]. This creates a finite thermal population ($N_3$) at the pump level which restricts the fiber from being uniformly inverted. Coupling losses in the amplifier unit increase the amplifier noise figure. External noise figures of 5 to 7 dB appear to be readily achievable.

Measurement of Noise Figure

According to the definition of noise factor given in the preceding section, $F$ is given by currents at the output of an ideal detector with quantum efficiency $\eta = 1$. For practical measurements, particularly when operating at high saturated output power, loss has to be inserted between the amplifier and detector. For simplicity, assume that the amplifier noise $i^2_{amp}$ consists of signal-spontaneous and spontaneous-spontaneous beat-noise currents $i^2_{s\text{-}sp}$ and $i^2_{sp\text{-}sp}$. The noise currents measured at the output of the detector are then

$$\begin{aligned} i^2_{\text{amp\_measured}} &= \eta^2_{\text{eff}}\, i^2_{\text{amp}} \\ i^2_{\text{sno\_measured}} &= \eta_{\text{eff}}\, i^2_{\text{sno}} \end{aligned} \qquad (25)$$

where

$$\eta_{\text{eff}} = \frac{I_{\text{dc}}}{GP_{\text{in}}}\frac{h\nu}{q}. \qquad (26)$$

$\eta_{\text{eff}}$ is the total loss factor from the amplifier output to the detector including the detector quantum efficiency. $I_{dc}$ is the photodiode dc current due to amplified signal power (assumption is that residual pump power is negligible. The noise factor of the amplifier is then given by

$$\begin{aligned} F &= \frac{1}{G} + \frac{1}{G\eta_{\text{eff}}}\frac{i^2_{\text{amp\_measured}}}{i^2_{\text{sno\_measured}}} \\ &= \frac{1}{G} + \text{SNR}_{\text{in}}\frac{i^2_{\text{amp\_measured}}}{I^2_{\text{dc}}} \\ &\approx \text{SNR}_{\text{in}}\frac{i^2_{\text{amp\_measured}}}{I^2_{\text{dc}}} \quad \text{for } G \gg 1. \end{aligned} \qquad (27)$$

CATV CARRIER-TO-NOISE RATIO

For CATV system modeling, the carrier-to-noise ratio (CNR) due to the optical amplifier can be calculated based on the noise factor given in Eqn. 27. In multichannel AM operation, the carrier currents at the output of the photodetector are given by

$$\frac{mI_{\text{dc}}^2}{2} \quad [\text{A}^2] \qquad (28)$$

where $m$ is the modulation index per channel. Thus, from Eqn. 27 and $G \gg 1$, the CNR due to the optical amplifier is given by

$$\begin{aligned} \text{CNR} &= \frac{\text{SNR}_{\text{in}}}{F}\frac{m^2}{2} \quad [/\text{Hz}] \\ &= \frac{\text{SNR}_{\text{in}}}{F}\frac{m^2}{8\cdot 10^6} \quad [/4\,\text{MHz}]. \end{aligned} \qquad (29)$$

These expressions are also valid for amplifier noise from other causes (excluding ASE shot noise) such as interferometric -interference noise. $i_{\text{amp\_measured}}$ includes noise currents from all other sources of noise within the amplifier. Also note that $F$ as given herein is the external noise factor of the amplifier (except for Eqn. 24 which does not take into account input and output coupling losses). CNR expressed in dB for 4-MHz electrical bandwidth is

$$\text{CNR}_{\text{dB}} = 89.9 - \text{NF} + P_{\text{in\_dBm}} + 20\log(m). \qquad (30)$$

An equivalent RIN can also be given:

$$\text{RIN}_{\text{dB}} = -155.9 + \text{NF} - P_{\text{in\_dBm}}. \qquad (31)$$

As an example, if $P_{in}$ = 0 dBm, $m$ = .04, and NF = 5 dB, then the CNR is 56.9 dB and the amplifier RIN is -150.9 dB/Hz. As is evident here, amplifier input power should be in the range of about 0 dbm or greater to prevent excessive degradation of system CNR. The effect of amplifier RIN is to reduce the system CNR for short fiber spans, but the high output power of the amplifier results in a higher CNR for long fiber spans than is possible with today's DFB lasers only. This characteristic CNR performance is shown in Fig. 7.

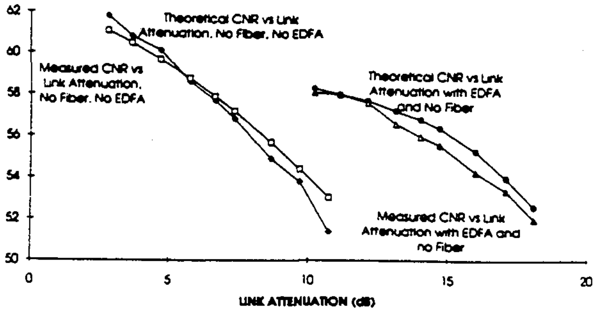


Fig. 7. CNR vs. link attenuation with and without EDFA. 4-dBm input power and 980-nm pumping. 1543-nm DFB source laser with 5.9% modulation index.

## CONCLUSIONS

Erbium-doped fiber amplifiers (EDFA's) have desirable properties for use as power amplifiers for AM CATV. Output power for commercially-available amplifiers is over 10 dBm and as high as about 20 dBm. A realistic noise figure for this type amplifier is about 5 - 7 dB. In this paper amplifier noise figure equations were developed and the relationship of CATV CNR to amplifier noise figure was given. Noise figure is given based on electrical measurements at the output of a photodetector.

In spite of the advantages of 1550-nm operation with erbium-doped fiber amplifiers, because of other factors and limitations of EDFA's not discussed in this paper, the current 1310-nm technology may continue to be the dominant technology for some time. Development of 1550-nm lasers continues to lag that of their 1310-nm counterparts. Dispersion at 1550-nm with standard single-mode fiber causes serious CSO distortion, and although solutions are available, the viability and acceptability of these solutions remains to be seen. However, the EDFA technology has matured and products are now available for AM CATV applications.

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# Optimum Laser Chirp Range for AM Video Transmission

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*Abstract*

*Simple expressions for the impact of DFB laser chirp on the system noise and distortion specifications of AM CATV fiber links are derived. These expressions are found to agree well with measured results. There is a narrow range of chirp for which degradation of both noise and distortion is low. 1310 nm lasers are found to have chirp within the acceptable range. 1550 nm laser are found to have chirp that is unacceptably large for most CATV applications.*

## I. Introduction

High dynamic range fiber optic links have recently been developed for use in CATV systems. CATV systems typically transmit 30 to 80 channels of amplitude modulated video in the VHF and UHF frequency ranges, with trial systems transmitting up to 150 channels. Fiber optic links offer a significant advantage because the link lengths can be 20 km or more compared to less than 1 km for coaxial links. The use of fiber optic links can eliminate the need to cascade large numbers of amplifiers, improving the performance and reliability of CATV distribution systems.

The requirements for CATV systems are very demanding in terms of noise and distortion. CATV fiber links require high optical power, the noise of the fiber links is typically within a few dB of the shot noise limit, and the distortion is very low, even for large modulation depths with peak modulation near 100%. Under the proper circumstances, directly modulated DFB lasers can meet all of these requirements. However, to insure that these demanding requirements will be met, all phenomena which can influence the noise and distortion of AM fiber optic links must be carefully studied. In this paper, the role of frequency chirping of DFB lasers in the noise and distortion of AM fiber optic links is discussed. The basic mechanisms for noise and distortion generation are reviewed, and experimental results for typical links are presented. Simple expressions for estimating the impact on CATV system noise and distortion specifications are presented. In many applications the maximum transmission distance is determined by the impact of fiber dispersion on linearity, rather than by the loss of the fiber. This is particularly true for 1550 nm systems, even when dispersion shifted fiber is used.

## II. Impact of Chirp of Second Order Distortion

Frequency chirping of DFB lasers has been extensively studied [1-4]. Chirp can have a significant impact on bit error rates of digital transmission links, particularly for 1550 nm lasers transmitted through 1310 nm zero dispersion fiber. For digital applications, chirp is most often characterized in terms of the -20 dB width of the lasing spectrum when the laser is digitally modulated from a point near threshold to a specified high level. For analog systems, the small signal chirping characteristic for modulation about a bias point well above threshold is more relevant. This is most often described in terms of the change of optical frequency with current. The chirp for DFB lasers modulated at VHF frequencies typically ranges from 50 to 500 MHz/mA.

The specific mechanisms responsible for chirp in DFB lasers have recently been reviewed [1,2]. At CATV frequencies, the most important mechanisms are spectral hole burning and spatial hole burning. Spectral hole burning results in "blue" shifting, while spatial hole burning can cause either "blue" or "red" shifting. Because the individual mechanisms will sometimes add and other times cancel, the overall spread in the chirp observed for DFB lasers is quite large. Figure 1 shows the measured distribution of chirp for 16 1310 nm

DFB lasers and 6 1550 nm DFB lasers. The chirp was measured using a scanning Fabry-Perot interferometer. The measurement was done at 60 MHz in which case the dynamic spectrum of a DFB laser has two peaks characteristic of wide deviation FM modulation. The differences between the magnitudes of the laser chirp has significant implications for AM fiber optic links as will be discussed later. It is not clear at this time the extent to which the difference between the two distributions is fundamentally related to the laser wavelength, rather than other factors, such as the grating coupling coefficient, K.

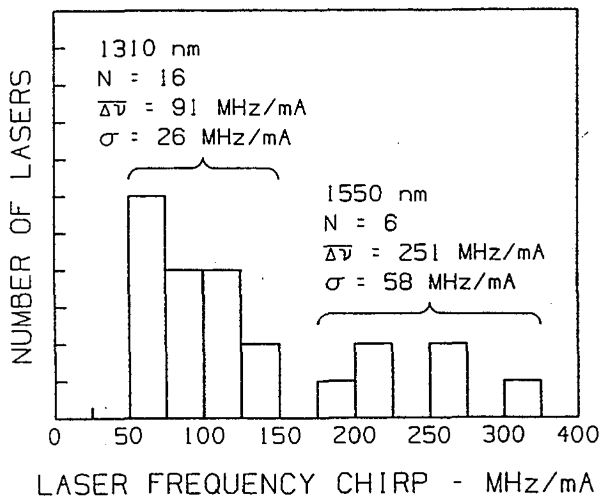


Figure 1
Distribution of measured frequency chirp for 1310 and 1550 nm DFB lasers.

There are several mechanisms by which laser chirp can result in distortion. Any optical component which has loss or delay which varies with optical frequency will convert laser chirp into optical distortion. Multiple optical reflections [5] and optical amplifiers [6] are two examples of situations where frequency dependent loss can occur. In most instances, wavelength dependent loss problems can be overcome by designing components with minimal wavelength dependent loss. A more fundamental problem occurs with fiber dispersion which results in wavelength dependent delay. The distortion due to chirp and dispersion has recently been analyzed [7,8].

If intensity modulated light from a DFB laser is transmitted through a dispersive medium, such as an optical fiber, then the laser frequency modulation is converted to a modulation of the transmission delay, $\tau$, through the link. The output signal waveform, So(t) will be of the form

$$S_o(t) = \frac{S_i(t-\tau(t))}{1+\frac{\partial\tau}{\partial t}} \qquad (1)$$

where $S_i(t)$ is the intensity modulation at the input side of the fiber and $\tau(t)$ is the time dependent fiber delay. The denominator accounts for "bunching" of the signal. The received optical power must be adjusted to account for the fact that light transmitted in a time interval $\Delta t_T$ is received in a time interval $\Delta t_R$, which can be different from $\Delta t_T$. Light transmitted from time $t_o$ to $t_o + \Delta t_T$ is received over the interval $t_o + \tau(t_o)$ to $t_o + \Delta t_T + \tau(t_o + \Delta t_T)$. Thus

$$\Delta t_R = \Delta t_T + \tau(t_o + \Delta t_T) - \tau(t_o)$$

$$= \Delta t_T\left(1 + \frac{\partial\tau(t_o)}{\partial t}\right) \qquad (2)$$

In this paper, we only consider the case where $\frac{\partial\tau}{\partial t} \ll 1$.

If the intensity modulation consists of two sinusoidally modulated signals, then the output signal waveform will contain second harmonics of the input frequencies as well as second order distortion at the sum and difference frequencies. The frequency chirping of DFB lasers at VHF frequencies is approximately linearly proportional to the modulation current and independent of the modulation frequency. For this case, the delay modulation is proportional to the intensity modulation with a proportionality constant, $\delta$, and the second order distortion can be expressed as indicated below:

$$S_i = \cos(\omega_1 t) + \cos(\omega_2 t) \qquad (3)$$

$$\tau = \tau_o + \delta(\cos(\omega_1 t) + \cos(\omega_2 t)) \quad (4)$$

$$S_o = S_i + \omega_1 \delta \sin(2\omega_1 t) + \omega_2 \delta \sin(2\omega_2 t) + (\omega_1 + \omega_2)\delta \sin(\omega_1 + \omega_2)t + (\omega_1 - \omega_2)\delta \sin(\omega_1 - \omega_2)t \tag{5}$$

Equation (5) assumes that $\frac{\partial \tau}{\partial t} \ll 1$, or equivalently, that $\omega\delta \ll 1$.

The RF power in the second order distortion products relative to the power in the fundamental signals is given by the square of the coefficients in (5). Comparing the coefficients in equation (5), distortion due to chirp and dispersion is seen to be most severe for additive second order products at frequencies $\omega_1 + \omega_2$ near the upper transmission frequency.

In an actual CATV system, there will be multiple carriers and many combinations of the various carriers that will produce distortion near the test frequencies. Composite second order distortion (CSO) is due primarily to two tone products at frequencies $f_1 \pm f_2$ because these products are 6 dB higher than the second harmonic and there are many two tone frequency combinations versus a single second harmonic product.

The composite distortion levels for multi-carrier systems can be estimated from two tone second order measurements, or calculations, by using the following method.

1. Adjust the distortion level to account for the number of beats, or frequency combinations that produce distortion near the test frequency. For example, for a 62 channel NTSC frequency plan, there are 22 two tone second order products at 446.5 MHz. This would lead to an adjustment factor of 13.4 dB.

2. Adjust the level to account for inaccuracies in the rf power measurement technique. Spectrum analyzers are commonly used to measure composite distortion products. Spectrum analyzers when used according to common test methods do not accurately measure the total power for many closely spaced distortion products. Our empirical results indicate that the measured power will typically be 2-4 dB below the actual total power.

This estimation method will only be accurate if the distortion contributions from the different frequency contributions are additive. This is generally true when the second order distortion level exhibits the classic 2 dB change for a one dB change in fundamental level.

Following this method, the estimated CSO due to chirp and dispersion are given by:

$$CSO = 20\log(\delta\omega) + 10\log(N_2) - 3dBc \tag{6}$$

where $N_2$ is the numbers of second order products falling at the frequency $\omega$, and a spectrum analyzer correction factor of 3 dB has been assumed. For a 62 channel NTSC system, there are 22 second order products falling 1.25 MHz above the upper carrier (445.25 MHz). The corresponding estimated CSO vs. $\delta$ is shown in Figure 2. The symbols represent measured data of CSO for a 1550 nm laser. The data for small $\delta$ is for transmission through dispersion shifted fiber. The data for large $\delta$ is for transmission through 1310 nm zero dispersion fiber. The deviation between measured and calculated values for large $\delta$ is due to a breakdown in the additivity approximation which requires $\omega\delta \ll 1$.

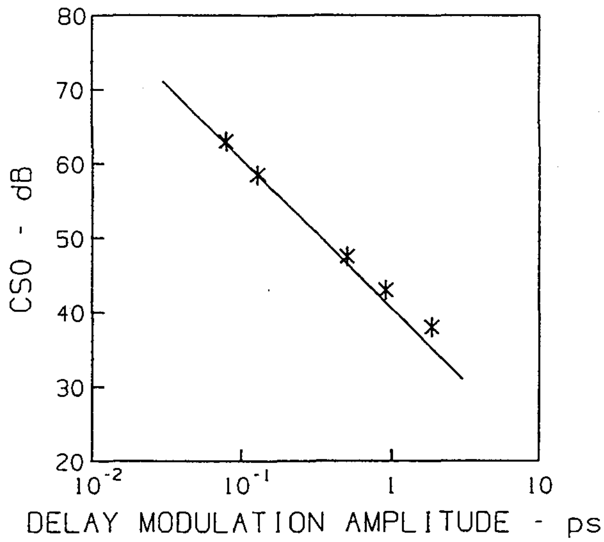


Figure 2
Calculated and measured CSO for a 62 channel NTSC CATV system due to chirp induced delay modulation.

Fiber optic links for AM video distribution, typically must operate at CSO values in the -60 to -65 dBc range. This includes distortion contributions from the amplifier driving the laser, the laser, the optical receiver, and any other mechanisms, such as fiber dispersion. Due to the technical challenges of fabricating DFB lasers with very good CSO, most of the distortion allocation is generally given to the laser. To minimize degradation due to dispersion, the CSO contribution due to dispersion should be no more than -70 dBc, unless the distortion is compensated by pre- or post-distortion. However, the distortion compensation option requires individual adjustment for each laser and fiber used which is often unacceptable for practical systems.

From equation (6), we can see that to meet the goal of -70 dBc CSO contribution due to dispersion, $\delta$ should not exceed 0.034 ps for the 62 channel NTSC example. The corresponding constraint on laser chirp can be determined from the relation:

$$\delta = DL\Delta\lambda \qquad (7)$$

where $D$ is the fiber dispersion, $L$ is the link length in km, and $\Delta\lambda$ is the amplitude of the wavelength chirping, in nm, due to the current modulation from a single channel. Taking $D$ to be 1 ps/nm-km, which allows for a 11 nm mismatch from the zero dispersion point at 1310 nm or a 12 nm mismatch at 1550 nm, and $L$ = 20 km.

$$\delta = 20\Delta\lambda \qquad (8)$$

Alternatively, if the chirping is expressed in MHz, the relation for 1310 nm systems is:

$$\delta = 1.14x10^{-4}\Delta\upsilon \text{ ps} \qquad (9)$$

where $\Delta\upsilon$ is the amplitude of the chirp due to the modulation current of a single channel in MHz.

For 1550 nm systems the relation is:

$$\delta = 1.60x10^{-4}\Delta\upsilon \text{ ps} \qquad (10)$$

For this particular example, we arrive at a maximum acceptable chirp of 296 MHz/ch for 1310 nm and 211 MHz/ch for 1550 nm. These numbers, together with the measurements of actual chirp shown in figure 1 are of tremendous practical significance. In the case of the 1310 nm systems, the laser with the largest chirp could be modulated with up to 2 mA/ch before reaching the chirp constraint, and this is slightly more than the typical modulation current for a 62 channel system of 1.8 mA. However, in the case of 1550 nm, the laser with the lowest chirp is limited to only 1.15 mA/ch which is 4 dB less than typical for 62 channel systems. This results in a corresponding decrease in system C/N. The 1550 nm laser with maximum chirp is limited to 0.6 mA/ch, or nearly 10 dB less than typical. If 1310 nm zero dispersion fiber with a dispersion of 18 ps/nm-km at 1550 nm is used, the 1550 nm laser with the lowest chirp is restricted to 0.074 mA/ch, or a reduction of 29 dB from typical.

In the absence of lower chirp 1550 nm DFB lasers, system designers are forced to make undesirable compromises in order to use 1550 nm DFB lasers. The choices are:

1. Accept system CSO degradation due to chirp.
2. Restrict modulation currents and thereby reduce C/N.
3. Limit transmission distances and thereby sacrifice the advantage of low loss at 1550 nm and the potential for extending transmission distances using Er doped fiber amplifiers.

4. Match laser wavelengths precisely to the fiber zero dispersion point.

5. Distortion compensate individual laser/fiber combinations.

It is important to stress that the serious dispersion problems with 1550 nm DFB lasers discussed above occur when dispersion shifted fiber is used. In the case of 1550 nm lasers with 1310 nm fiber the problem is much more severe.

## III. Impact of Chirp on Noise in Fiber Optic Links

In the preceeding section, the affect of DFB laser chirp on distortion for AM video transmission was discussed. Chirp is undesirable with respect to distortion. However, the opposite is true for noise. The C/N of AM links is unavoidably degraded by double backscattering of light. This effect has recently been analyzed [9,10]. The mixing at the photodiode of light that is transmitted directly from the laser to the photodiode with light that has been twice reflected generates noise which extends over frequencies proportional to the chirped linewidth of the laser. The total amount of noise depends on the fraction of doubly reflected light. Therefore, the more laser chirp, the lower the spectral density of this noise mechanism. It should also be noted that if the laser chirp is much less than the minimum operating frequency of the link, then most of the noise will be out of band. This low chirp case is not attainable with direct modulation of DFB lasers, but can be achieved with external modulation of solid state lasers. The noise that does appear in externally modulated links, however, is at frequencies close to the carrier which is particularly objectionable.

As has been previously reported, the double Rayleigh scattering noise mechanism has the following dependences on fiber length and chirp.

$$\text{Noise} \sim L - \frac{1}{2\alpha}[1 - e^{-2\alpha L}]$$

$$Noise \sim \frac{1}{\Delta\upsilon} \qquad (11)$$

We have measured the noise degradation for many 1310 nm lasers and numerically fit the results to the expressions above. To evaluate the noise increase due to double backscattering, we measure the link C/N for the same laser for transmission through a length of fiber and for an optical attenuator of the same loss. The double backscattering introduces an additive equivalent laser relative intensity noise (RIN). Our numerical estimate for this noise for 1310 nm links is given below.

$$\text{RIN}_{dbs} = \frac{3.6 x 10^{-14}\left[L - \frac{1}{2\alpha}(1 - e^{-2\alpha L})\right]}{\Delta\upsilon_{RMS}} \qquad (12)$$

where $\Delta\upsilon_{RMS}$ is the $RMS$ frequency chirping in MHz, $L$ is the link length in km, and $\alpha$ is the link attenuation (0.80 km$^{-1}$ for the fiber we used). Because this measurement involves estimating a relatively small noise contribution in the presence of other noise sources, the estimate has a potential error of ±1 dB. We do not have a similar estimate for 1550 nm links, but preliminary measurements indicate that the noise contribution is nearly the same as for a 1310 nm laser with the same RMS frequency chirp.

The equivalent RIN due to double backscattering is shown in Figure 3. For comparison, a receiver with a DC photodiode current of 0.5 mA, which is typical for AM links, has noise due to shot noise of the photodetection process that is equivalent to a RIN of -151.9 dB/Hz. The impact of double backscattering noise can also be seen in figure 4, which shows link C/N for a 20 km, 62 channel link for various values of laser chirp. In Figure 4, the following link parameters, which are typical of 62 channel 1310 nm links were assumed. The laser power was 6 mW with a modulation depth of 4.5%/ch. The receiver noise current was taken to be 5 pA/Hz$^{1/2}$ and the responsivity was 0.9 A/W. The link loss was taken to be 0.4 dB/km + 1 dB.

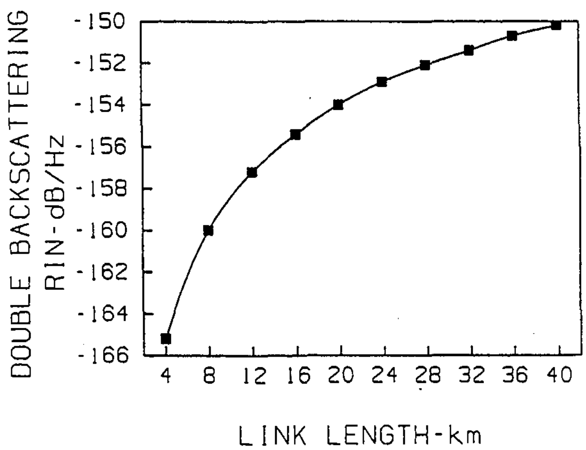


Figure 3
Equivalent relative intensity noise generated by double Rayleigh scattering in 1310 nm fiber link. RMS laser chirp is 1000 MHz.

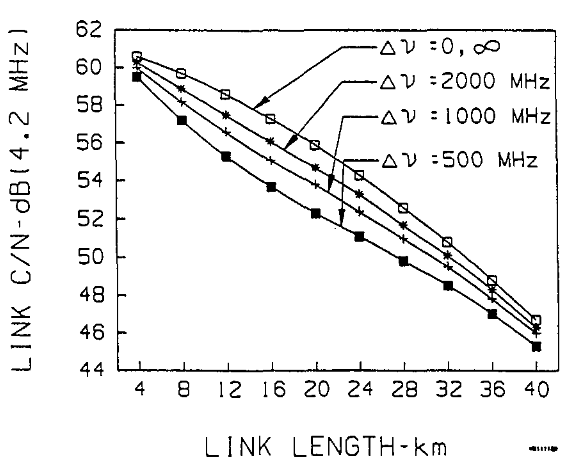


Figure 4
Effect of double backscattering noise on the C/N of a 62 channel CATV fiber link.

In the previous section, an upper limit for 1310 nm laser chirp amplitude of 296 MHz/ch, to avoid excessive distortion in a 62 channel, 20 km link was obtained. This corresponds to an RMS frequency chirp of 1650 MHz from the modulation of all 62 channels. Due to double backscatter noise, a lower limit on the acceptable chirp can also be defined. Figure 5 shows the minimum laser chirp required for 1, 2, and 3 dB system noise penalties as well as the maximum chirp for -70 dBc CSO contribution. As can be seen, for laser RMS chirp around 1000 MHz, the CSO constraint is satisfied and the noise penalty is less than 2 dB for links up to 20 km in length. We believe this represents the best compromise between noise and distortion. Fortunately, the majority of the 1310 nm lasers have chirp near this target level. All of the 1550 nm DFB lasers are significantly above the target for normal modulation currents. It should also be noted that the 1000 MHz target was for a specific set of link parameters. For other link parameters, the acceptable range will change. However, for most AM links of practical interest, laser chirp plays a significant role in determining the C/N and distortion of the link.

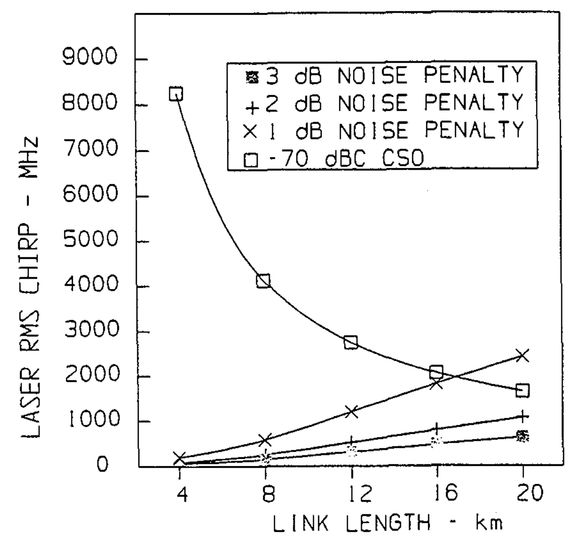


Figure 5
System noise and distortion constraints for laser chirp.

## IV. Summary

In this paper, the basic mechanisms by which DFB laser chirp can affect the noise and distortion of AM fiber optic links has been reviewed. Expressions are presented for estimating the impact of these phenomena on the CATV system specifications of C/N and CSO. These expressions have been found to be in good agreement with measured results. Laser chirp can adversely impact link linearity, but is important in minimizing noise due to double Rayleigh backscattering. There is a relatively narrow range for chirp in which the impact on both noise and distortion is acceptable. 1310 nm DFB lasers consistently fall within this range. Preliminary measurements indicate that 1550 nm DFB lasers have chirp which is greater than the upper limit of the acceptable range. This results in serious linearity problems for 1550 nm links even when dispersion shifted fiber is used.

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# Passive Optical Network (PON) Architectures and Applications

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*Abstract*

*Passive Optical Networks (PONs) have generated considerable interest for telephony applications, and this architecture has been claimed to be very suitable for distribution of video signals as well. The term PON covers an array of design variations (e.g. two fiber versus one fiber), and several trial networks have been built using different designs. This paper describes a number of current approaches to PON design, and typical parameters are given. The characteristics of a typical PON video delivery system are described in terms that allow network operators to evaluate the applicability of this technology to their networks.*

## 1. Introduction

Optical fiber is an almost ideal transmission medium. It has huge bandwidth potential, it is almost inert, it is available in cables which are small, light, and easily handled compared to metallic cables, and it has very low transmission loss which does not vary significantly with temperature. Its major drawbacks are that splicing is more difficult than for copper cables and the cost of optoelectronic transducers is high. These drawbacks are steadily yielding to technological improvements.

Optical fiber transmission has been the technology of choice for long distance digital transmission for a decade, and is steadily becoming cost effective at shorter and shorter distances. In CATV networks, FM fiber optic supertrunk systems have been in use for many years, and AM fiber systems have supplanted coaxial trunking in most new and rebuild construction in the last two years.

In telephony applications, the long term goal is to take fiber all the way to the customer. This approach is usually called "fiber to the home" (FTTH). It will give the transmission benefits of fiber and the "future protection" provided by the bandwidth potential of fiber. However, it is not economically feasible at present. The cost of fiber optic systems makes it necessary for a number of customers to share each fiber network terminal. This approach is usually called "fiber to the curb" (FTTC).

One innovative approach to designing FTTH or FTTC networks is the use of Passive Optical Networks (PONs). This paper provides an introduction to PON technology. Section 2 summarizes the forces motivating the approach and some major network issues, and Section 3 describes some architectural variations that have been proposed for telephony applications. Section 4 discusses broadband PON architectural issues, and Section 5 summarizes the main points.

### 1.1 Terminology

A variety of terms for the two terminal types in a PON network appears in the literature. For readability, the discussion in this paper is written in FTTH terms, but it applies with obvious changes to FTTC networks as well. The term "exchange terminal" is used for the equipment that is located in the telephone exchange (or central office), and the term "customer terminal" is used for the equipment located at the customer site (or at the curb). "Downstream" is from the exchange to the customer, and "upstream" is from the customer to the exchange. In a CATV application, the exchange terminal would be at the head end.

In the literature, the exchange terminal is variously called Exchange Terminal (ET), Central Office Terminal (COT), Subscriber Loop Terminal (SLT), or Optical Line Terminal (OLT). The customer terminal is usually called Optical Network Unit (ONU), but other names such as Network Termination (NT) and Distant Terminal (DT) can be found.

## 2. Motivation and Issues

The PON approach to telephony access networks is an attempt to reduce costs by taking advantage of the following:

1. Most access network links are short (optical loss budget a few dB)
2. Low and medium rate (low power consumption) digital communication systems can accommodate loss budgets much greater than typical access networks.
3. Residential and small business access customers can be served by an average of less than 3 lines per customer. Even medium sized businesses only require a 24 line trunk to connect to a PBX. Low rate digital systems can easily serve tens of customers at this rate.
4. Cost savings can be realized by sharing the exchange transmitter laser over multiple customers.
5. Sharing feeder fibers is also possible if splitters are placed in the field close to the customers.
6. Sharing laser and feeder fiber over more than 20 customers means that network cost is dominated by the customer terminal, so higher splitting ratios yield only small extra savings.

These points suggest an architecture which has a single "low rate" transmitter (say 20-40 Mb/s line rate) broadcasting to 20 or more customer terminals, with each customer using only part of the total bandwidth. The signal travels on a single feeder fiber to a point near the customers, where it is split and routed down separate fibers to each customer terminal (see Figure 1). Clearly, this a point-to-multipoint transmission architecture, which is logically, but not physically, the same as that used in the present CATV network. The classical telephony architecture is a point-to-point structure. The PON equipment design is optimized where possible to reduce the cost of the customer terminal.

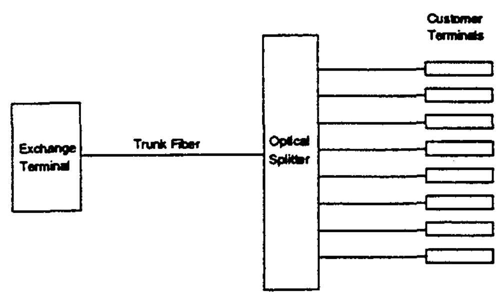


Figure 1. PON Architecture

Many major issues must be addressed to complete the design of a fully functional network:

- The discussion above addresses only the downstream (exchange-to-customer) link. The upstream (customer-to-exchange) link must also be implemented at low cost. The upstream fiber network is assumed to be the same topology (and in some cases the same network) as the downstream network. The loss budgets in the two directions are the same and only a single exchange receiver is needed.

Sharing upstream bandwidth is more complicated than sharing downstream bandwidth, because customer terminal transmissions must be timed to avoid collisions.

- Privacy: signals must be secured, and customer terminals must be monitored and controlled to ensure that data for one customer is very difficult for another person to intercept.

- Operations and maintenance: the whole network must not be taken out of service to add a new customer or to service a faulty customer terminal. This and other operational issues require careful design of control software and operational procedures.

- On-line bandwidth allocation: since customer service requirements change, it must be possible to change the bandwidth allocated to a customer without interrupting service to other customers.

- Power savings: to conserve power, customer terminals must have a "sleep" mode which they enter when there is no active traffic for them. (This is important for telephony services, where battery backup at the customer terminal is usual.)

- Fiber breaks: how do you locate a fiber break between splitter and customer terminal without interrupting service on the network? Ideally, one would prefer to do this from the exchange terminal. This may be feasible using a wave division multiplexer (WDM) to separate optical time domain reflectometer (OTDR) signals from traffic signals on the network. The OTDR dynamic range must be high to cope with the splitter loss. The OTDR display will show superimposed traces because each customer line will generate a separate reflection. If the fault cannot be seen among the superimposed traces, fault location must be done from the customer end. If service personnel do not have access to the customer end, it is necessary to work from the splitter by opening up a splice.

- Measures must be taken to prevent a faulty customer terminal transmitter from jamming the network with continuous transmission. The obvious requirement is global and addressable commands which instruct all (or one) customer terminal(s) to stop transmitting, and redundant transmitter disabling circuitry in the customer terminal.

## 3. PON Architectures

A wide variety of PON architecture implementations have been suggested, and several trial systems have been demonstrated using very different techniques. The main choices (which are interrelated) to be made are:

1. How many fibers should go to each customer?
2. What line protocols should be used to share downstream and upstream bandwidth?
3. What wavelengths should be used for downstream and upstream transmission for narrowband and broadband service?

These questions are probably best understood by enumerating possible answers. While the discussion this far has centered on narrowband (NB) telephony service, any viable architecture must also be capable of carrying broadband (BB) broadcast CATV service. It is desirable that normal AM CATV should be possible, for the same reasons that AM is used in CATV coaxial distribution.

### 3.1 Number of Fibers

Some possible answers to question 1 are:

1. Two fibers: one fiber for each direction, wave division multiplex (WDM) NB and BB downstream.

The broadband signal will suffer loss due to the WDM, but this is only a small part of the loss budget. Isolation of the customer BB receiver from the NB signal must be excellent to prevent degradation of BB performance.

2. Two fibers: one for NB (bidirectional), one for BB (one way).

   There is a choice here with the NB on one bidirectional fiber. The line can be full duplex using a WDM or directional couplers, or half duplex allowing only one end to transmit at a time. If a WDM is used, the isolation must be high enough to prevent the transmitter from interfering with the collocated receiver. If directional couplers are used, the optical reflections must be kept low so that near end cross talk does not interfere with reception. If the optical loss budget is high and the network generates optical reflections, it is possible for the level of the reflected transmitted signal to be comparable to the signal received from the far end [1]. Using a half duplex approach, where the exchange terminal transmits a long burst and then each of the active customer terminals transmits a short reply burst, eliminates most concerns about optical reflections on the NB fiber, but roughly halves the data transfer rate possible for a given line rate.

3. One fiber: bidirectional with WDM of NB and BB to customer.

   Use of one fiber is clearly more economic, but it is also more complicated since it involves both WDM of the downstream signals and bidirectional use of the fiber. In addition to the points mentioned above, at the customer terminal the isolation between the upstream transmitter and the downstream broadband receiver must be excellent. However, all of this is within the capability of existing optics technology.

### 3.2 Line Protocol

Downstream traffic must be multiplexed into a single bitstream. The bandwidth allocated to each customer must be changeable on-line to cope with changing customer needs. To best utilize the system capacity, it is highly desirable that bandwidth unused by one customer be available for allocation to other customers. The protocol also must address many operational details such as bringing a new customer terminal into service, verifying that a new terminal is authorized to be on the network, as well as detecting and isolating faults.

The upstream traffic must be time division multiplexed in such a way that the bursts from the customer terminals do not collide at the exchange receiver. This is accomplished by measuring the range to each customer terminal and giving each terminal a delay time to wait after it receives the end of the exchange transmission before commencing its transmission. The range measurement requires a guard interval in the return path time allocation so that a new terminal of unknown range can be brought into service. Guard intervals may also be needed between transmissions from the customer terminals.

Various degrees of interleave of the customer terminal transmissions have been proposed, from bit interleaving to full burst interleaving. Longer transmissions reduce the number of guard intervals, so there is less dead time (or more data for a given line rate).

Bit interleaving requires very accurate ranging: to within a small fraction of a bit if guard intervals are to be avoided altogether. This is necessary because so many guard intervals would be needed that data throughput would be too low. It also requires control of the customer transmitter laser power by the exchange terminal, together with an advanced exchange receiver, because consecutive bits from different customer transmitters will not be the same amplitude but must be nearly so in order to be received without error.

Burst interleaving requires fast clock acquisition by the exchange receiver, but is otherwise more robust. However, it is less efficient in terms of data rate for a given line rate.

Another tradeoff to be made is in the choice of frame length, where one frame is one complete cycle of transmissions by the exchange and all customer terminals. A short frame reduces the delay through the network. However, a short frame also reduces the efficiency of transmission, because the proportional loss of bandwidth due to network overhead and guard times is higher. The loss of efficiency may require a higher data rate.

### 3.3 Wavelength Choice and WDM

To keep costs low, an uncooled laser must be used in the customer terminal and both the line protocol and wavelength choice must allow for this.

For telephony applications, the normal wavelength for narrowband data is the 1300 nm window. This is a reflection of the maturity (and low cost) of devices at this wavelength. For the exchange laser, the choice is arbitrary, and the 1550 nm window could serve as well, particularly since dispersion on the short links will have no effect on digital signals but may have major impact on AM wide band signals.

Since the customer terminal transmitter cost is a major item in the network cost, 1300 nm uncooled lasers are preferred for this function.

Wave division multiplexing is possible to allow simultaneous use of both windows. At present, dense WDM using several closely placed transmitters in the same window is not economic, but this possibility remains open for future use.

### 3.4 Typical Narrowband PON Parameters

Typical parameter ranges for narrowband PONs are:

- Range: up to 10 km (6 miles)
- Splitting ratio: 8, 16, or 32
- Number of 64 kb/s channels: 150-400
- Optical loss budget: 20-30 dB
- Optical line rate: 20-40 Mb/s

## 4. Broadband PON Design

The optical loss budgets above are suitable for FM or digital CATV transmission systems. The application of these two technologies to PONs is straightforward but use of either of them today would make a customer terminal expensive and provide insufficient channels for a broadcast CATV service. The advent of compressed digital television will make digital CATV delivery on PONs much more attractive.

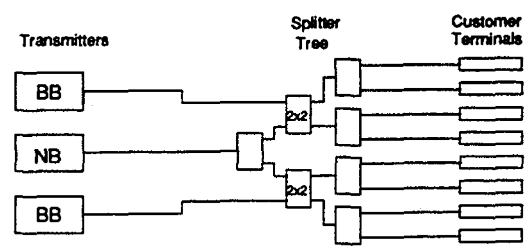


Figure 2. Use of 2x2 couplers to allow different NB and BB splitting ratios

Typical PON loss budgets are far beyond the reach of AM technology using DFB laser transmitters. Lower splitting ratios must be used for broadband distribution with these systems. This is easily achieved if the splitting tree uses 2x2 couplers, with multiple AM systems feeding into the splitting tree at lower splitting ratios than the NB transmitter (see Figure 2).

The need for multiple transmitters and "trunk" fibers increases the cost of the system. The high cost of broadband distribution is fundamental to the use of AM fiber optic systems: such systems require high receiver power levels compared with digital or FM systems.

For a carrier-to-noise ratio of 48 dB at the customer terminal, a typical AM system must have a received optical power of about -7 dBm. (Receiver noise current 6 pA/sqrt(Hz) and 4% optical modulation index assumed.) A typical DFB transmitter operates at 6 dBm, so the optical loss budget of the system is 13 dB. If 6 dB is allocated for connectors, fiber loss, and splice loss, the splitter loss can be 7 dB, which allows a four-way split.

The only option to support higher splitting ratios with AM systems is to increase the transmitter power. This suggests consideration of externally modulated systems or optical amplifiers. These technologies have not been widely deployed to date, and there are some limitations on how they can be used.

External modulation systems currently operate only in the 1300 nm band. They use a narrow line source laser, and stimulated Brillouin scattering (SBS) limits optical power in the trunk fibers to less than 15 dBm for a short fiber and less than 12 dBm for a longer fiber (see Figure 3) [2,3]. Higher power transmitters can be used, but the optical signal must be split at the exchange to keep the level in the multiple trunk fibers below the SBS threshold.

Available optical amplifiers operate in the 1550 nm band. On standard fiber, dispersion will cause excessive second order distortion unless compensation is used [4]. While optical amplifier repeaters could be used in the outside plant, the network would no longer be passive.

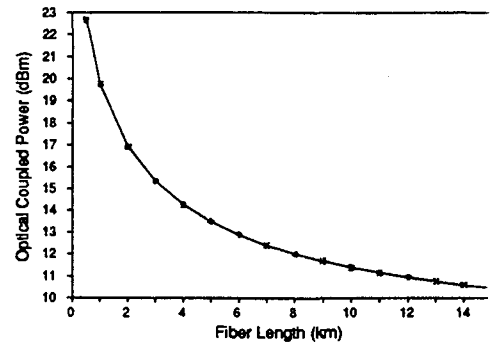

Figure 3. SBS Threshold for CW Laser vs Leng

## 5. Summary

This paper has described the motivat for and typical characteristics of Pass Optical Networks for telephony services. also examined the delivery of CATV serv on networks of this type.

While the optical power demanded AM fiber optic transmission of CATV sigr makes it uneconomic to consider PON serv to the customer, some current applications AM fiber systems in CATV trunking can regarded as PONs, with the terminat points being the optical nodes wh conversion to coaxial distribution takes pla The current splitting ratios are limited by power available from the transmitter to val much lower than those used in dig telephony PONs.

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# PERFORMANCE OF DIGITAL TRANSMISSION TECHNIQUES FOR CABLE TELEVISION SYSTEMS

**Richard S. Prodan, Ph.D.**
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*Abstract*

*An evaluation of complex modulation techniques for the transmission of digital information for advanced television applications is currently underway at CableLabs. The performance of various digital modulation techniques in the cable television transmission medium coexistent with standard NTSC analog video channels has been investigated. Both laboratory and field evaluations on existing cable systems are presented.*

## Introduction

The performance measure customarily utilized in characterizing digital modulation is the bit error rate. The bit error rate (BER) is dependent upon the carrier to noise ratio, or more precisely the bit energy to noise spectral density ratio. Either of these metrics provide the probability of error in terms of the distance between signals in energy space divided by the noise power for the additive white Gaussian noise environment.

The optimum receiver receives a waveform which is comprised of a transmitted signal corrupted by adding white Gaussian noise. The signal can be replaced by an equivalent vector form, and the noise process by a relevant noise process that can also be represented in vector form. This is done by defining a set of orthonormal time waveforms which can be used in linear combinations to represent both the signal and noise vector components (e.g., two quadature modulated carrier phases).

The vector components are derived at the receiver through a correlator or matched filter to the orthonormal time waveforms. The decision as to which signal was sent is made by comparing the distance of the received vector to all possible signal vectors. The receiver decides that a particular signal was sent if the received signal vector is closest to it. An error occurs if the received vector is closer to a signal vector that is different from the originally transmitted one.

For PAM modulated signals with non binary symbols, a rectangular constellation and decision regions result. The constellation diagram represents the signal vectors in a two dimensional space. The rectangular nature of the signals and their resulting decision boundaries are shown for 16 QAM and 4 VSB in Figures 1 and 2 respectively.

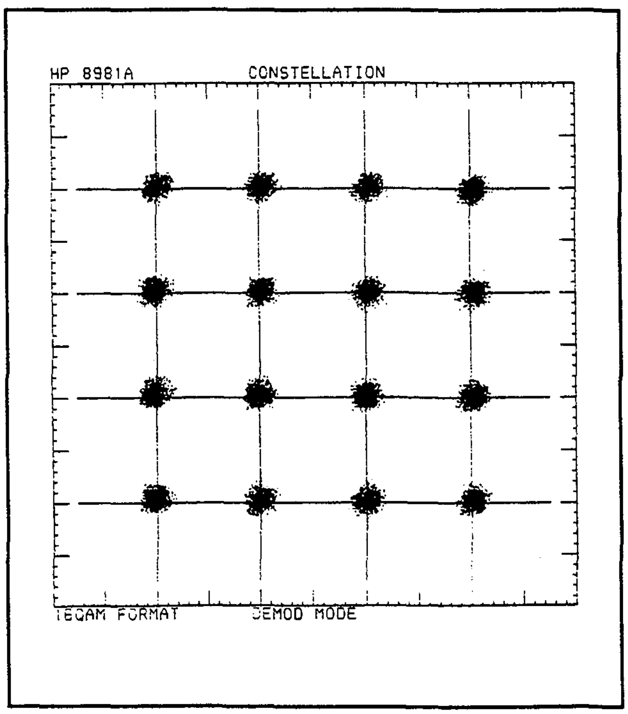


**Figure 1 - 16 QAM Constellation**

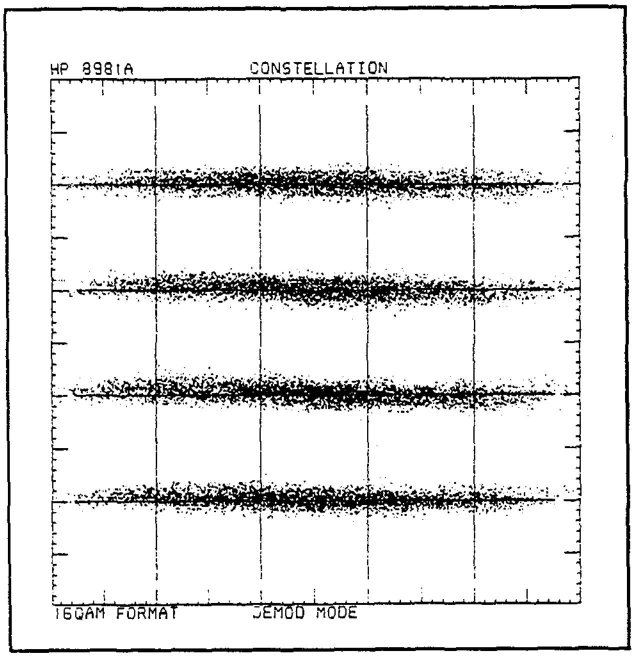


**Figure 2 - 4 VSB Constellation**

This Euclidean distance concept in signal energy in the presence of such noise can be reduced in the presence of additional impairments such as channel nonlinearities, intermodulation, and intersymbol interference due to channel bandwidth restrictions or reflections from impedance mismatches. These impairments reduce the "effective" carrier to noise ratio of an equivalent noisy but unimpaired channel. The bit error rate is calculated as the probability of the equivalent noisy signal erroneously crossing a decision boundary associated with the transmitted symbol (or group of bits) into another region associatedwith a different symbol. The effective carrier to noise with reduced implementation noise margins degraded by channel impairments has been studied to estimate the expected bit error rate of modern terminal equipment with practical performance limitations.

Transmission Testing Method

Two modulation formats were studied. The first format is 16 QAM double sideband pulse amplitude modulation format with quadrature carrier multiplexing and 4 levels (representing 2 bits) per carrier phase with the carrier placed symmetrically in the center of the band. The second format is 4 VSB, which is a (nearly) single sideband pulse amplitude modulation format with a small (5 percent) vestige about the carrier at the band edge. The modulated average signal power was set approximately 10 dB below NTSC carrier level.

The approach suggested for the performance evaluation of digital modulation on the cable distribution plant is vector modulation or constellation analysis. The digital data carrier is discretely modulated in phase and amplitude to convey groups of binary digits (words) as vectors in carrier phase space. The instantaneous switching between carrier phase states requires infinite channel bandwidth. Restrictions on the modulating data signal bandwidth result in intersymbol interference in practical systems.

Evaluation of ISI is possible by examination of the baseband data modulating channel response given by an eye pattern or diagram, where overlapping data symbol periods are superimposed to determine the reduction of noise margin due to ISI. The closure of the eye results in increased bit error rate, since noise in the channel is much more likely to force the signal to cross a decision boundary. The spread in the signal constellation clusters as well as shifts in position due to other channel impairments can be characterized by the constellation diagram. The constellation is a set of sampled points in carrier phase space with the carrier sampling times optimally chosen to coincide with the maximum eye openings in time of the inphase (I) and (for QAM only) the quadrature (Q) channels.

The performance of generalized digital modulation signal sets can be generated and analyzed with vector modulation equipment. Although bit error rate cannot be directly measured, it can be inferred from the constellation and eye pattern parameter measurements. These measurements can be made without constructing modems that require carrier recovery, symbol synchronization, clock recovery, data detection, differential decoding, etc.

Laboratory Tests

Vector modulation equipment available from Hewlett Packard along with prototype digital Nyquist pulse shaping filters and frequency conversion equipment was employed for the digital transmission evaluations. A pseudorandom bit sequence (PRBS) generator provided the data for the digital carrier modulation. The random data modulates I and Q IF carriers in both formats. Channel filters must be designed and inserted to shape the modulation spectrum and limit the modulated carrier bandwidth. A 6 MHz channel is utilized within the 41 to 47 MHz range with an appropriate Nyquist response rolloff characteristic.

A source bit rate of 18 Mbps from the PRBS generator divided between I and Q carrier phases (for 16 QAM) or carried in the I phase only (for 4 VSB) with 10 percent rolloff (excess bandwidth) occupied a 6 MHz channel. The coherent reference from the vector generator was normally used (except for phase noise testing) to demodulate the modulated data IF carrier at the vector modulation analyzer. Constellation and eye pattern measurements were made without the need for carrier recovery.

The modulated IF signal was supplied to a Scientific Atlanta RF modulator IF input with a crystal oscillator selected for the television channel desired for data transmission. A complementary RF demodulator recovers the modulated data carrier at IF, after being degraded by added impairments.

The recovered I and Q data bitstreams can be examined for mean square eye closure, phase offsets, and dispersion in the recovered constellation samples from the vector modulation analyzer. Several hundred thousand points were downloaded via a GPIB interface to a computer for further analysis and estimation of effective carrier to noise ratio and expected error rate.

Some mean square eye closure results from the lab tests done on the CableLabs test bed in the Advanced Television Test Center in Alexandria, VA are given in Table 1 for various cable

| LAB DATA - FROM CableLabs TEST BED | | | |
|---|---|---|---|
| TTS | DESCRIPTION | QAM % | VSB % |
| 1 | Unimpaired | 13.0 | 14.5 |
| 2 | Echo @ 300ns -15.7dBc | 35.0 | 33.4 |
| 2A | Echo @ 300ns -20.7dBc | 21.5 | 20.5 |
| 2B | Echo @ 300ns - 25.7dBc | 15.3 | 15.6 |
| 3 | CW Ingress at -23dBc | 30 | 32 |
| 4 | CTB at -32.5dBc (cw carriers) | 16.6 | 18.8 |
| 5 | Phase Noise -91.6dBc (1Hz) at + 20kHz from carrier | 23.5 | 25.0 |
| 6 | Composite Second Order at -32dBc (cw carriers) | 15.0 | 16.5 |
| 7 | Hum Mod, 120Hz, 4.6% | 16.5 | 17.5 |

**Table 1 - Lab results from the CableLabs test bed in Alexandria, VA.**

impairments. The "unimpaired" eye closure of 14% is due to implementation loss, analog filtering, and the modulation and frequency translation equipment. Data for a short duration reflection characteristic of cable systems is shown for -15, -20, and -25 dB. The composite triple beat level for comparable eye closure is higher than would be present for satisfactory NTSC reception. The same situation applies for composite second order interference. Phase jitter from oscillator phase noise and power supply induced residual FM show a significant degredation at levels that would be unnoticeable on NTSC.

Field Tests

The laboratory tests may be repeated in the field on the TCI cable system in Boulder, CO. An additional complication arises due to the need

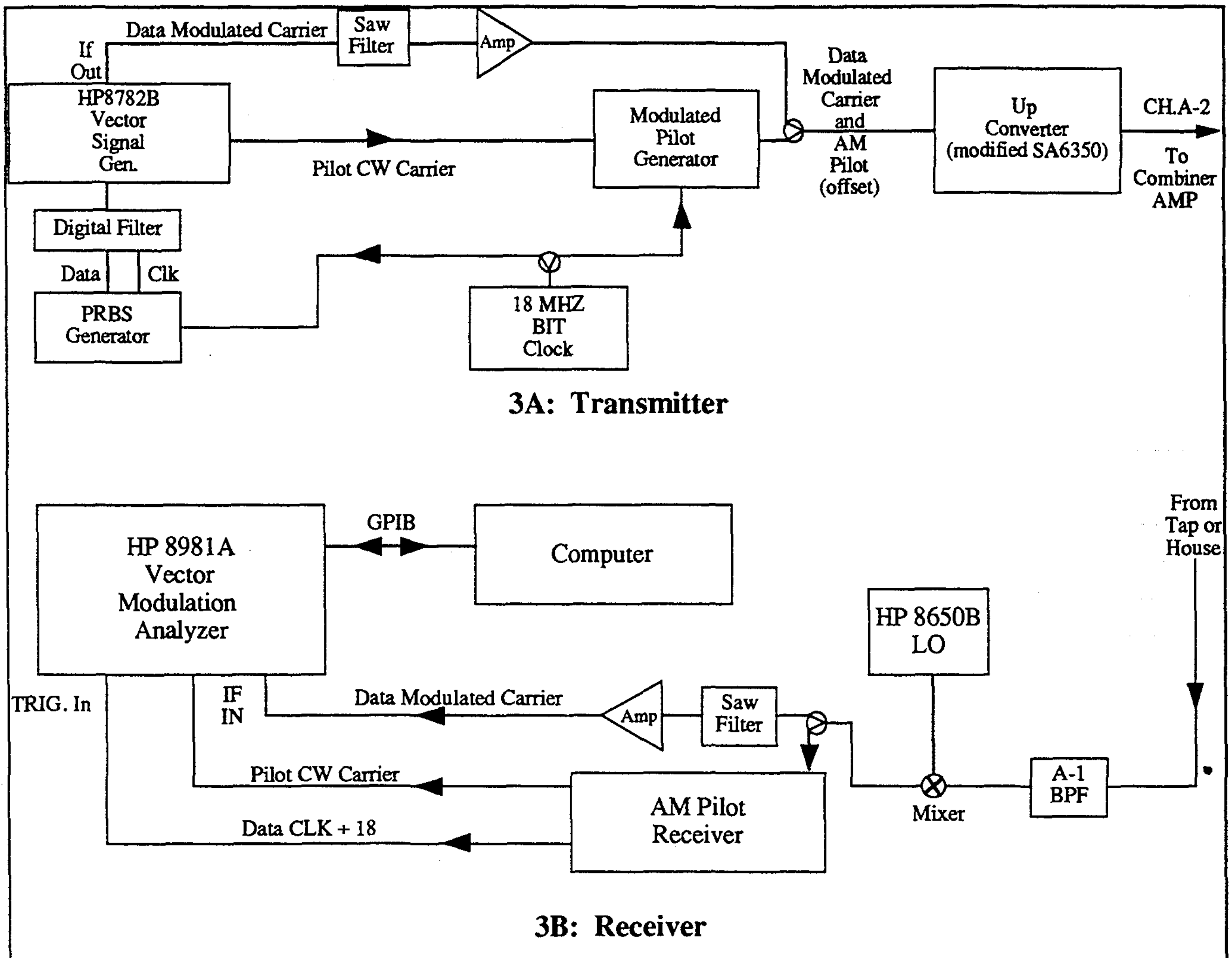


**Figure 3 - Digital Transmission Test Block Diagram**

for a recovered carrier reference and symbol timing for vector demodulation and sampling of the I and Q modulated carrier phases. This unmodulated carrier reference should be phase locked precisely to the data modulated carrier IF frequency.

During the field tests, both 4 VSB and 16 QAM signals were generated to study the effect of cable impairments. The complete test setup for the field (and the lab without the carrier and symbol timing reference recovery portions) is shown in Figure 3. The generation of the I and Q baseband data streams were done in the headend in the same way as in the test bed facility using a pseudo-random binary sequence generator (PRBS) and a digital filter. The digital filter generated the necessary Nyquist shaped, bandlimited data signals which drove the external inputs of an HP8782B Vector Signal Generator. The HP Vector Signal Generator generated both a coherent pilot CW carrier at the output frequency, and a data modulated carrier with the I and Q data.

At the receive site (which was kilometers away in the field test, but less than one meter in the lab test), a coherent pilot CW carrier reference was needed by the HP8981A Vector Modulation Analyzer to demodulate its received data carrier into baseband I and Q data streams. Additionally, the Vector Modulation Analyzer needed an external input data clock to accurately determine symbol timing. This presented a design problem to the recovery of the data over the cable TV plant because the coherent pilot reference occupies the same spectral space as the data modulated carrier.

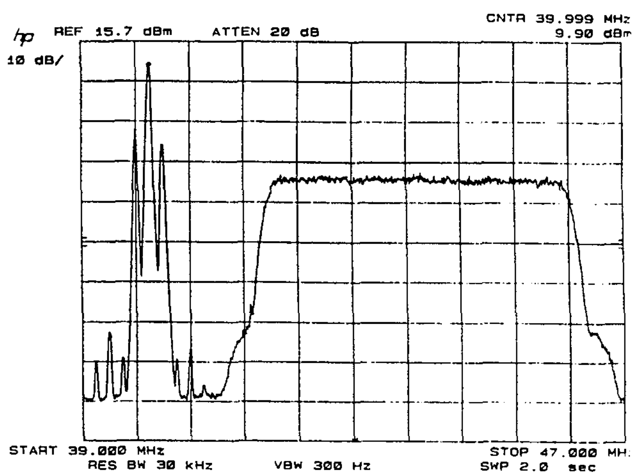


**Figure 4 - Spectrum of Modulated Data Carrier and AM Pilot Tone**

The solution implemented was not to build a modem, but to use an offset AM carrier pilot to achieve both carrier and data timing references. The AM pilot carrier was offset out of band from the modulated data carrier, and a 200 kHz AM modulation tone was put on the offset pilot. The AM pilot carrier was now 1 MHz below the 41-47 MHz IF band used by the modulated data carrier. Figure 4 shows the spectrum of the data modulated carrier and the offset AM modulated pilot tone.

The 200 kHz tone was used in the AM Pilot Receiver to both regenerate the carrier offset, and provide a data clock to trigger the external trigger on the vector analyzer.

At RF the spectrum is inverted from IF. The Modulated Pilot Generator at the headend used the 18 MHz Bit Clock and the pilot CW carrier to generate the offset AM pilot, and it was summed with the data modulated carrier for an input to an upconverter. A Scientific Atlanta 6350 modulator with a frequency agile output converter (FAOC) was used for upconversion of the pilot plus modulated carrier. Normally, the phase noise of the agile upconverter would have been troublesome, but the offset carrier recovery scheme provided immunity to phase noise, as both the out of band AM pilot and modulated carrier undergo identical phase jitter in the conversion process.

It can be shown that any frequency offset in the RF local oscillator will be present in both the modulated IF data carrier and the recovered carrier reference. Hence carrier recovery without phase error is achieved at the remote measurement site, and all the measurements previously described for the laboratory testing may be done in the field.

At the receive site, the signal was amplified and put through a bandpass filter (BPF). Channel A-1 was used in the TCI Boulder, CO system. An HP8656B Signal Generator was used as a downconverter local oscillator (LO), and a double balanced mixer brought the data modulated carrier and the offset AM pilot to the IF frequency band. At IF, the signals were split and the data modulated carrier was bandpassed through a saw filter and presented to the input of the demodulator in the HP8981A Vector Modulation Analyzer. Off of the other split, the AM modulated carrier was put into the AM Pilot Receiver. The AM pilot receiver performed two tasks. The first was to recover an unmodulated carrier reference for demodulation, and the second was to provide a data clock that the Vector Modulation Analyzer could use for triggering, which provided the correct sampling times for the symbols.

The Transmitter and Receiver carrier and data clock reference circuitry was used for both

16 QAM and 4 VSB. With both transmission methods, the AM modulated pilot carrier remained at an IF frequency of 40 MHz. With 4 VSB, the regenerated carrier at 42 MHz was offset by 2 MHz from the 40 MHz AM carrier. In the 16 QAM case, the regenerated carrier at 44 MHz was offset by 4 MHz from the 40 MHz AM carrier. The same 41 to 47 MHz IF band was used for both VSB and QAM modulated data carriers.

Some mean square eye closure results from the field tests done on the TCI system in Boulder, CO are given in Table 2 for various tap and subscriber home locations. The signal received both at the tap and inside the house at the TV receiver input were measured. The mean square eye closure for several locations including

| FIELD DATA | | |
|---|---|---|
| FIELD LOCATION | QAM EYE CLOSURE % | VSB EYE CLOSURE % |
| TCI Headend | 13.2 | 14 |
| CableLabs Lab | 14.2 | 15.7 |
| House 1 | 32 | 34 |
| Tap 1 | 16 | 17.9 |
| House 2 | 16 | (Note 1) |
| Tap 2 | 14.7 | 16.2 |
| House 3 | 14.7 | 15.9 |
| Tap 3 | 14.4 | 17.1 |
| House 4 | 19.9 | 23.3 |
| Tap 4 | 17.2 | 19 |
| House 5 | 22.1 | 22.5 |
| Tap 5 | 20.9 | 24.4 |
| House 6 | 16 | 15.5 |
| Tap 6 | 14.8 | 15.2 |
| Field Fiber HUB | 19.9 | 21.6 |
| Lab Fiber + 12AMPS | 16.1 | 16.2 |

Note 1: Equipment out of service

**Table 2 - Field Results from the TCI System in Boulder, CO.**

a fiber hub are shown. The large variability on the resulting impairment between location in the system and between the tap and the premises wiring in this small sample is significant. The variability of performance for digital modulation within the cable plant at the subscriber drop merit additional investigation.

## Conclusion

It can be noted that rather small reflections causing intersymbol interference results in significant eye closure. This source of interference is most readily caused by cable reflections due to mismatches inside the house. This suggests that adaptive equalization may be required in many receive locations for a uniform level of reliable reception (suitably low error rate) at the lower signal levels that are nominally suggested (and used in this evaluation) for digital cable transmission.

The results obtained in both laboratory and field trials can be used to infer required modem performance in terms of the relative level of importance of the residual impairments at the receiver. A test of actual bit error rate requires the modem, as the carrier recovery, data, clock recovery, and symbol timing and synchronization information (and resulting equipment implementation losses) are needed to recover and evaluate the continuous baseband data stream at the destination. However it is possible to estimate the error rate obtained using the raw data acquired in the present study. This is the subject of a future companion paper.

# Performance of Fiber Optic Cables in AM CATV Systems

by John C. Chamberlain
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*Abstract*

*Because fiber optics has quickly become an accepted technology in AM, CATV applications, the understanding of many component specifications and their impact on system performance has lagged behind. This paper examines current fiber optic cable specifications, and how cable made to these specifications can affect the performance of AM, fiber optic, CATV systems in the field.*

*It is shown that many of the currently applied telephony specifications are suitable for CATV applications. But, some specifications, in particular temperature performance, should be re-examined in order to meet the needs of the CATV AM video market place.*

**Fiber Cable Specifications**

Although there are a number of specifications written by organizations such as REA, Sprint, and GTE that apply to outside plant fiber optic cable, the most comprehensive and critical specification is Bellcore's TR-TSY-000020, or for short, TR-20[1]. This specification was written by Bellcore for the Regional Bell Operating Companies (RBOCs) so that they could reference it when purchasing fiber optic cable from any vendor. The specification was written with the intent of being used to specify fiber optic cable for telephony systems. This paper investigates how those specifications apply to AM, CATV systems.

As mentioned above, TR-20 is the most comprehensive specification that is available to date. TR-20 cable specifications are summarized in Table 1. As shown in the summary, the attenuation measurements for the mechanical tests are all performed at 1550 nm. This is because the fiber is more sensitive to increases in attenuation at 1550 nm than at 1310 nm[2,3]. The bends that most likely occur during mechanical testing are referred to as macrobends. Macrobends range in size from 5-30 millimeters. Macrobends affect the longer wavelengths in fiber transmission before they affect the shorter wavelengths. The smaller the diameter of the macrobend the shorter the wavelength that the macrobend effects. But as shown in Figure 1, the effect on 1550 nm is apparent at far larger bend radii than at 1310 nm. When macrobends are the cause of attenuation increases, the effect is dramatic due to the severe slope of the loss curve. For this reason all systems should be proved into the field at 1550 nm as well as 1310, even if the intent is to only use 1310 nm as an operating wavelength. Potential problems with the long term stability of your system may not show up in a 1310 nm check out, but would if the system were checked at 1550 nm.

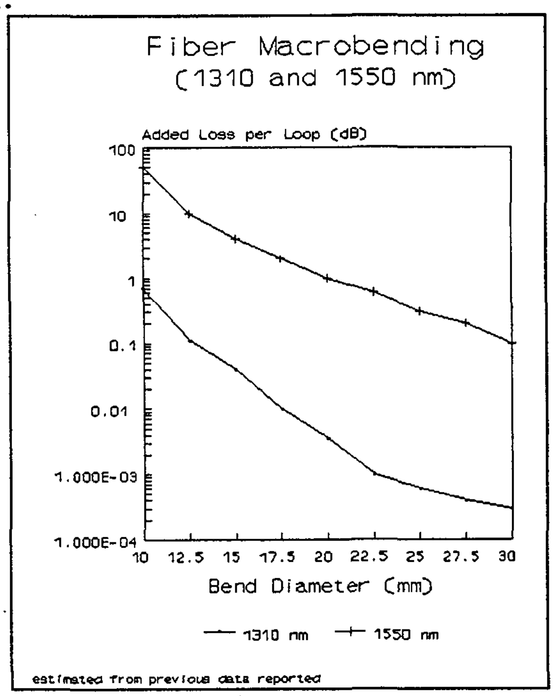


**Figure 1**

| Mechanical & Environmental Tests | | | |
|---|---|---|---|
| Test | EIA-455 Specification | Mechanical Requirement | Optical Requirement |
| Tensile | FOTP-33A | 600 lbs. | <.1 dB @ 1550 nm |
| Compression | FOTP-41 | 1000 lbs. | <.1 dB @ 1550 nm |
| Twist | FOTP-85 | 10 cycles | <.1 dB @ 1550 nm |
| Low Temp. Bend | FOTP-37 | 4 wraps @ -30 C | <.1 dB @ 1550 nm |
| Cyclic Flex | FOTP-104 | 15X cable O.D. | <.1 dB @ 1550 nm |
| Impact | FOTP-125 | 25 cycles | <.1 dB @ 1550 nm |
| Ice Crush | FOTP-98 | 24 hours @ -2 C | <.1 dB @ 1550 nm |
| Temperature Cycle | TR-20 | -40 to +70 C | 100% < 0.2 dB/km<br>80% < 0.1 dB/km |

**Table 1**

**Mechanical Testing**

The mechanical tests consist of impact, crush, twist, flex, and ice crush, among others. Attenuation is monitored before, during, and after the tests. The level of acceptance is 0.1 dB allowable increase in attenuation on each fiber. In general, little or no attenuation increases at 1550 nm are experienced due to these tests. Which means, as per the above discussion, 1310 nm is not affected at all.

If a field problem subjected a cable to conditions similar to these mechanical tests, a resulting 0.1 dB increase in the fiber attenuation would result in a decrease in the CNR of 0.1 dB. Because, the majority of the noise in an AM fiber optic system is created in the transmitter and receiver, CNR varies inversely proportional with increases in attenuation of the passive part of the plant. Therefore, for every 1 dB increase in attenuation in the fiber, the CNR is decreased by 1 dB. In the case of the mechanical tests, a 0.1 dB increase, which, as discussed above is very rarely seen, would result in a 0.1 dB decrease in CNR. Distortions are essentially unaffected by an unpolarized, passive increase in system attenuation.

The conclusion is that the performance levels for the mechanical tests are in TR-20 are adequate for AM systems as well. Although, some consideration should be given to mechanical tests that are more applicable to an aerial plant. Most of the mechanical tests specified in TR-20 are applicable to buried applications, where the majority of telephony fiber optic cable is installed.

**Lightning and Rodent Testing**

There are no lightning or rodent testing requirements that fiber optic cable must pass in order to meet the TR-20 specifications. Although a lightning test must be completed and the results reported,

**Table 2**

Lightning Test Levels

| Test Level | % of Strikes in U.S. below adjacent levels | TR-20 Rating |
|---|---|---|
| 150 Ka | 99% | |
| 105 Ka | 95% | A |
| 80 Ka | 90% | B |
| 55 Ka | 75% | C |
| < 55 Ka | N/A | D |

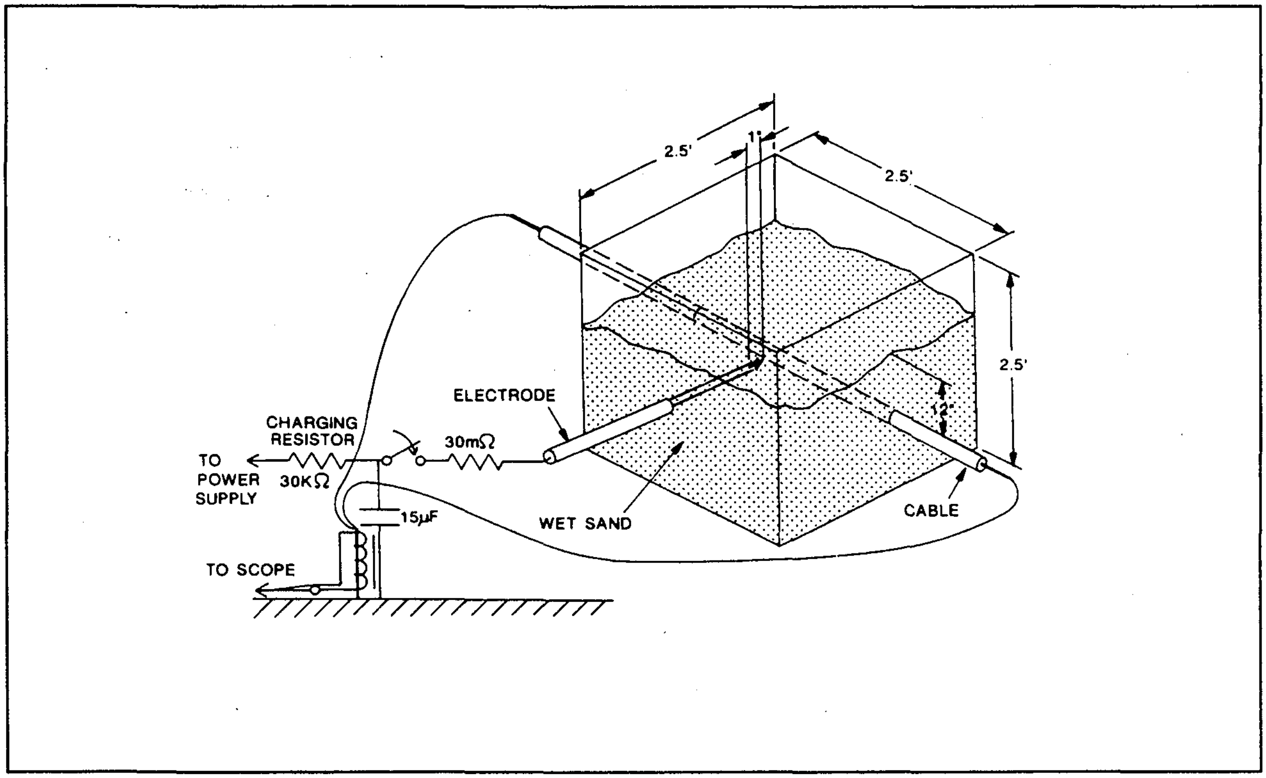


**Figure 2**

plastics in the cable shrink. The distribution of excess fiber in the tubes is such that the induced bend diameter of the fiber varies at low temperatures. As shown in Figure 1 the diameter of the macrobend does not have to decrease much in order to show a large effect on the attenuation of the cable. The effect, because it is a macrobend effect, degrades attenuation at 1550 nm before 1310 nm.

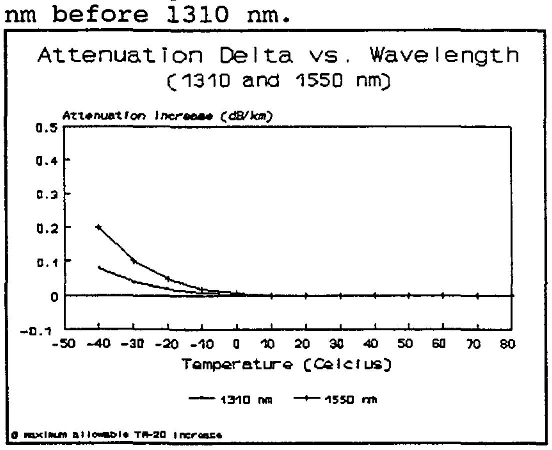


**Figure 3**

The potential effect on AM system performance due to the allowable attenuation increase, per TR-20 is shown in Figure 4[9,10,11,12]. A 0.2 dB/km increase in attenuation in the fiber corresponds to an identical decrease in CNR at the output of the optical receiver. That means for a 20 km run (12.4 miles), if the fiber were to increase in attenuation the allowable 0.2 dB/km, the net attenuation increase would be 4 dB and would decrease an original CNR at the optical receiver of 54 dB to 50 dB. For lower attenuation increases there would be a correspondingly smaller change in CNR.

Although these changes are seen at 1550 nm before 1310 nm, this allowable increase in attenuation may not be acceptable for some AM system performance criteria. If the performance specification for the fiber optic cable where cut in half (100% less than 0.1 and 90% less than .05), the "proposed specification " curve for AM video fiber optic cable performance shown in Figure 4 would result.

no minimum level is required. The results are reported by rating the cable as passing the test at certain amperage levels[4,5,6]. See Table 2. As shown, the highest TR-20 rating accounts statistically for 95% of all lightning strikes in the United States. Based on this information, an all dielectric fiber optic cable might be preferred in prominent lightning areas of the country. But, before rash conclusions are made, the test procedure should be examined in closer detail.

The test is designed to simulate not only the amperage of the lightning hit, but also the hammer effect of a lightning hit in an underground application. As shown in Figure 2 the cable is buried in wet sand before the simulated lightning hit is discharged. This is important because a great deal of damage can be done to the cable just due to the mechanical impact effect of a lightning strike. The question is: How applicable is this information to aerial applications of fiber optic cable? A modification to the sandbox test has been proposed for aerial applications[7]. The test set up is the same, except that there is no sand in the box and the cable is lashed to strand. All the metallic members in the cable, as well as the stand, are grounded to complete the circuit. Testing under these conditions shows that the strand takes the majority of the hit. See Table 3. Therefore, contrary to some claims, the construction of the cable is of diminished importance to lightning susceptibility in aerial fiber optic cable installations.

**Table 3**

Lightning Test Levels
Lashed to Strand

| Fiber Optic Cable Construction | Test Level Passed |
|---|---|
| Core Tube, armored | 200 Ka |
| Core Tube, dielectric | 200 Ka |
| Loose Tube, armored | 200 Ka |
| Loose Tube, dielectric | 200 Ka |
| Coaxial Cable | 200 Ka |

On the other hand, the construction is important to lending rodent protection to the cable. The degree of rodent protection a cable has, is measured at the Denver Wildlife Center with the assistance of gophers. Gophers are used because of the large amount of damage done to buried telecommunications cable by gophers every year. Although a squirrel test might be more applicable to this industry, no such test exists to date. The results of the gopher rodent testing show that a gopher will chew through a non-armored cable during the seven day test, but will not penetrate the steel in an armored product within the same seven day period[8].

In summary, the choice of construction of fiber optic cable should be made with rodent resistance as the determining factor, with less emphasis on lightning resistance.

**Temperature Performance**

The temperature performance specifications for fiber optic cable as per TR-20 (See Table 1) allow an increase in attenuation as high as 0.2 dB/km across the operating temperature range of -40 to +70 degrees Celsius. The reality of the situation is very close to the specification[14,15,16]. Increases in attenuation are sporadic and non-linear. The specification reflects the inconsistency of the situation. One hundred percent of the fibers must have attenuation increases less than 0.2 dB/km and eighty percent must have increase less than 0.1 dB/km. The attenuation increases occur at the low end of the operating temperature range. See Figure 3. Attenuation increases typically begin occurring around -20 Celsius or approximately -4 degrees Fahrenheit. The increase is not the same on all fibers in a cable and also is not linear with respect to temperature. Some fibers may see the full 0.2 dB/km increase at -40 and other fibers in the same cable may see no measurable increase in attenuation.

This temperature dependent increase in attenuation is due to fiber macrobends discussed earlier. The fiber collapses into these macrobends at low temperatures as the

Again, as in the mechanical tests, non-polarized increases in attenuation due to temperature performance do not affect distortion parameters such as CTB and CSO.

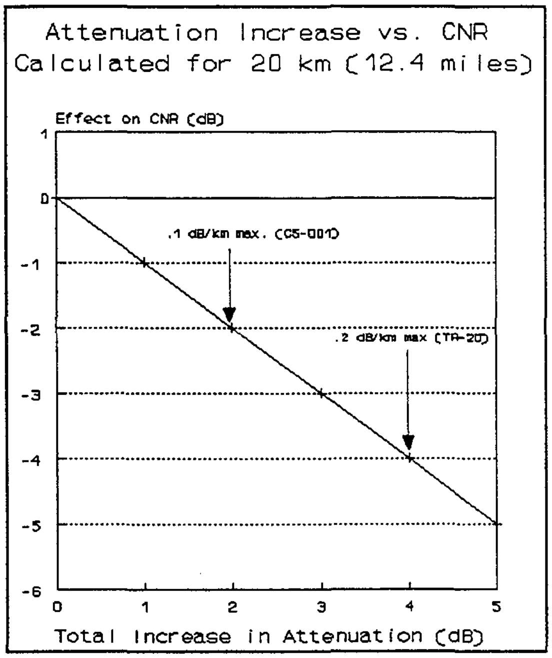


**Figure 4**

**CSO in AM Fiber Systems**

Chromatic dispersion is a general term describing the phenomena associated with pulse broadening of an optical signal in a single mode fiber.

Generally, chromatic dispersion can be broken down into two dispersion mechanisms; material dispersion and waveguide dispersion.

Material dispersion is a function of the chemistry used in the manufacture of optical fibers and waveguide dispersion is a function of the index profile of the fiber. They are both types of chromatic dispersion because both types spread the optical signal due to the finite spectral width of the optical source. Which means, different "colors" of light produced by any real optical source travel at different speeds creating a signal dispersion effect.

This dispersion effect in the fiber, combined with phenomena known as laser chirp, create an frequency modulating effect in AM fiber optic systems that manifests itself as CSO. This is the reason that in AM fiber systems CSO is the limiting distortion factor, not CTB[13].

Laser chirp is defined as the phenomena of the laser optical output wavelength changing linearly with the AM modulating signal input into the laser. This is why essentially no chirp is experienced in externally modulated lasers. The laser is not being directly modulated and therefore does not shift or "chirp" wavelength.

The telephony specifications for single mode fiber dispersion can also be referenced when purchasing fiber for AM fiber transmission. Standard singlemode fiber has it's zero dispersion point centered around 1310 nm. The dispersion at 1550 nm in a standard singlemode fiber is substantially higher. (maximum 18 ps/nm-km) Therefore, in many cases the limiting performance specification in 1550 nm systems is CSO, unless external modulation or precompensation for the chirping effect is used.

**Time Varying CSO Effects**

There have been isolated reports in the field of unacceptable levels of CSO that develop after fiber optic system istallations. These CSO levels can fluctuate in a matter of days or minutes. The effect is caused by a combination of parameters. One being the chirp factor of the laser and the other being polarization maintaining effects of the passive portion of the AM system, which include polarization sensitive loss and polarization dispersion. The passive part of the system consists of connectors, optical couplers and the fiber itself.

The problem occurs when the laser being used has a high chirp factor, and there is some polarization maintaining level of the optical signal in the passive part of the plant.

Two polarizations of the mode in a single mode fiber can be present and have a polarization sensitive loss as well as a polarization sensitive velocity of propagation. The amount of polarization separation is very sensitive to external changes on the fiber cable. Literally, if the wind were to whip a cable in the air, or the outside temperature change, the amount of polarization separation seen in the fiber can change. This change manifests itself as the time varying part of the time varying CSO effect. The actual CSO level is, in part, due to an frequency modulating effect caused by the laser chirp in conjunction with the passive polarization effects. This frequency modulating effect is much like the CSO distortions caused by chromatic dispersion described above.

If the two polarizations continually mix and are not separated, low levels of CSO are seen. As the two polarizations become more distinct for longer periods of time the dispersion levels, and therefore the CSO levels increase. Time varying CSO in AM systems can be controled by minimizing laser chirp, polarization dispersion and polarization sensitive loss.

**Summary**

At this point in time, there is no better specification to reference than Bellcore's TR-20, for the purchase of fiber optic cable for AM CATV systems. It has been shown, however, that the requirements of fiber optic cable for AM CATV systems differ from the requirements of telephony fiber optic cable.

The major installation method in CATV is aerial and the majority of telephony installations are buried. Due to this difference in predominant installation methods, some environmental test procedures, such as lightning resistance should be modified to better simulate application in the CATV industry. In other cases, such as attenuation increases with respect to temperature, and dispersion effects, the requirements for AM video fiber are more stringent than for digital telephony fiber optic cable and should be taken under consideration when specifying fiber optic cable for broadband AM systems.

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# THE USE OF DISPERSION-SHIFTED FIBER FOR CABLE TV APPLICATIONS

By

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## *ABSTRACT*

*As cable television operators deploy optical fiber in increasing volume, single-mode dispersion-shifted fiber has emerged as a viable transmission option in the 1550 nanometer (nm) window. This paper discusses the performance attributes of dispersion-shifted fiber, as well as the relative tradeoffs between it and standard single-mode fiber. The paper also outlines some potential cable television applications for dispersion-shifted fiber.*

## INTRODUCTION

Recent advancements in optical amplifiers and distributed feedback (DFB) lasers suggest that the 1550 nm operating window offers an opportunity for cost-effective cable TV architectures. While 1550 nm transmission over the long term is likely to complement, not compete with, existing 1310 nm technology, it has gained considerable momentum as an alternate operating wavelength.

Simply stated, 1550 nm transmission may help maximize many of fiber's benefits while reducing overall system costs.

Single-mode fiber's low attenuation rate, (the reduction in signal strength over the length of the fiber) at 1550 nm can enable cable operators to extend their link distances. In addition, the commercialization of 1550 nm optical amplifiers could enable more receivers to be shared by a common laser.

However, 1550 nm operation of amplitude modulated (AM) video over standard single-mode fiber presents its own challenges. Chief among these is reducing the effects of dispersion, a key fiber performance parameter.

## UNDERSTANDING DISPERSION

Dispersion is commonly referred to in digital transmission as the key optical parameter that limits the maximum data rate or information-carrying capacity of a single-mode fiber link. It refers to the time-based spreading of each pulse of light as it travels along a fiber.

As the pulse of light travels along the fiber, it spreads due to the different wavelengths that make up the pulse traveling at different speeds. Eventually, the pulses can overlap one another and become unrecognizable, which leads to an increase in the bit error rate. If the bit error rate becomes excessive, the amount of received information or bandwidth becomes severely limited.

In analog transmission, the effect is slightly different. Dispersion can cause an analog waveform to become significantly distorted. The complex nature of an amplitude modulated (AM) video signal makes it very susceptible to any distortion impairment.

Dispersion induced distortion is a result of laser chirp. Laser chirp is created when the wavelength of the laser changes due to injection current modulation of the laser. Typically, as the laser injection current increases, and thus the laser ouput power, the center wavelength of the laser shifts towards longer wavelengths. A small amount of laser chirp can significantly distort the ouput of an AM optical video signal due to the presence of dispersion.

This distortion can be illustrated simply by imagining different wavelengths of light within the waveform, although very close, traveling at different speeds due to the index of refraction characteristics of single-mode fiber. The presence of dispersion significantly alters the traveling speeds of the wavelengths of light within the waveform and thus enhances this distortion effect.

In AM cable TV systems, this distortion shows up as second order harmonics and intermodulation (or "beats") and is commonly referred to as composite second order distortion (CSO). CSO distortion appears on a television monitor as rolling diagonal lines.

Fiber dispersion has been shown to create CSO distortion due to the chirp of a DFB laser interacting with the dispersion. Over standard single-mode fiber, the amount of dispersion is much higher at 1550 nm than at 1310 nm. This results in a reduction in picture quality, unless CSO compensation techniques are used.

For highly dispersive mediums like standard single-mode fibers at 1550 nm, CSO distortion due to dispersion can be compensated for through either electrical or optical means. Electrical compensation techniques typically employ the use of pre-distortion circuits to create CSO equal in magnitude, but opposite in sign to cancel the CSO created by the DFB laser.Optical compensation techniques most often try to offset the amount of dispersion in a system.

Several electrical compensation techniques have been demonstrated, but most are believed to be distance and/or bandwidth limited. One alternative to compensation techniques is the use of single-mode dispersion-shifted fiber at 1550 nm as the transmission fiber, instead of standard single-mode fiber.

## WHAT IS DISPERSION-SHIFTED FIBER?

Corning developed the first commercially available single-mode dispersion-shifted fiber (SMF/DS™) in 1985. Historically, Corning's patented SMF/DS™ fiber has offered significant benefits for long-haul telephony and submarine cable applications.

With recent developments in the commercialization of erbium-doped fiber amplifiers (EDFAs) and DFB lasers operating at 1550 nm, dispersion-shifted fiber is poised to be the fiber of choice for high-data-rate transmission systems over long distances. These same developments in EDFAs and DFBs have led the cable TV industry to consider its use for specific AM video applications.

Dispersion-shifted fiber was designed specifically to capitalize on fiber's inherent lower attenuation at 1550 nm by shifting the zero dispersion wavelength to 1550 nm as well. In order to center the zero dispersion wavelength at 1550 nm and still maintain similar performance characteristics as standard single-mode fiber, Corning manufactures its dispersion-shifted fiber with a unique refractive index profile (Figure 1). The refractive index profile represents the change in refractive index of the core glass relative to the cladding glass and typically defines the fiber's transmission characteristics.

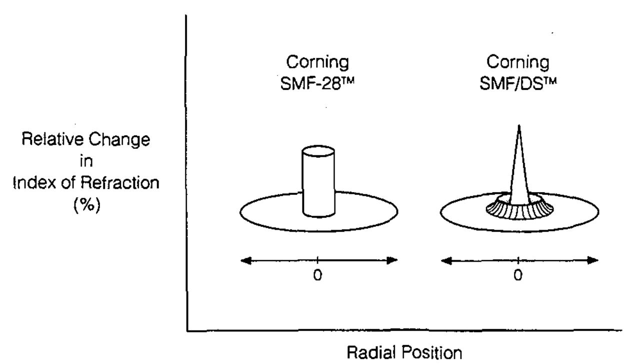


Figure 1. Refractive Index Profiles of Corning SMF-28™ and Corning SMF/DS™ Fibers

This design's advantage is that the attenuation at 1550 nm (~0.2 dB/km) is about half of the 1310 nm value for a standard single-mode fiber. However, a higher dopant level is required to change the refractive index profile for a dispersion-shifted fiber. Therefore, dispersion-shifted fiber theoretically will have a slightly higher attenuation at 1550 nm than standard single-mode fiber.

This is partially offset by SMF/DS™ fiber's improved resistance to macro- and microbending-induced attenuation from cabling, installation and handling. The smaller mode-field diameter (MFD) of the SMF/DS™ fiber - 8.1µm at 1550 nm vs. 10.5µm for SMF-28™ single-mode fiber, along with its unique index profile, results in this performance advantage. Table 1 compares the properties of Corning's standard and dispersion-shifted single-mode fibers.

| | | Corning SMF-28™ | Corning SMF/DS™ |
|---|---|---|---|
| Attenuation | @ 1300 nm | 0.35 dB/km | 0.39 dB/km |
| | @ 1550 nm | 0.19 | 0.21 |
| Mode Field Diameter | @ 1300 nm | 9.3 µm | 6.5 µm |
| | @ 1550 nm | 10.5 µm | 8.1 µm |
| Zero Dispersion Wavelength | | 1310 nm | 1550 nm |
| Dispersion Slope | | 0.09 ps/nm²•km | 0.08 ps/nm²•km |
| Dispersion | @ 1300 nm | ≤ 3.5 ps/nm•km | ~ -19 ps/nm•km |
| | @ 1550 nm | ~ 17 ps/nm•km | ≤ 2.7 ps/nm•km |
| Bend Performance | | Good | Better |

Table 1. Corning Single-Mode Fiber Comparison

## PERFORMANCE TRADE-OFFS

To evaluate the relative performance trade-offs between standard and dispersion-shifted single-mode fiber at 1310 nm and 1550 nm, the following experiment was conducted.

The experimental setup consisted of a matrix generator with 40 channels, 55.25 to 325.25 MHz, offered to the input of the test laser. From the laser, various lengths of fiber were installed into the system to measure the effect on system performance over each length.

A variable optical attenuator was used after the fiber to maintain the same received optical power, thereby eliminating any contribution the receiver might make. The received optical power was kept at 0 dBm. Following the optical receiver were the RF bandpass filters, a post-amplifier, and a RF spectrum analyzer.

Two 1310 nm lasers and one 1550 nm laser were used for the tests. A 1310 nm laser (Laser 1) and a 1550 nm laser (Laser 3) were selected that had similar modulation index (MI), slope efficiency, laser output power, chirp, carrier-to-noise ratio (CNR), and CSO for 6 dB optical loss as shown in Table 2. Since the laser's distortion mechanisms (slope efficiency, linearity, and chirp) were accounted for, this allowed a better performance analysis based on the fiber's dispersion and attenuation characteristics.

| | Laser 1 | Laser 2 | Laser 3 |
|---|---|---|---|
| Modulation Index | 4.1% | 5.0 % | 4.5% |
| Wavelength | 1318 nm | 1313 nm | 1546 nm |
| Slope Efficiency | 0.117 | 0.225 | 0.140 |
| Laser Output Power | 4.5 dBm | 6.1 dBm | 6.0 dBm |
| Received Optical Power | 0 dB | 0 dB | 0 dB |
| Laser Chirp (MHz/ma) | 370 | 92 | 350 |
| CSO (55.25 MHz) | -72.7 dBc | -67.8 dBc | -72.5 dBc |
| CSO (325.25 MHz) | -66.4 dBc | -72.8 dBc | -61.5 dBc |
| CTB (55.25 MHz) | -73.6 dBc | -79.6 dBc | -74.9 dBc |
| CTB (325.25 MHz) | -71.8 dBc | -74.7 dBc | -69.7 dBc |
| CNR (55.25 MHz) | 56.7 dB | 58.2 dB | 55.9 dB |
| CNR (325.25 MHz) | 55.9 dB | 56.0 dB | 56.6 dB |

Table 2. Relative Laser Performance Characteristics

Assuming the absence of any interferometric or receiver effects, the system CSO will degrade in relation to fiber dispersion, fiber length, channel frequency, chirp, and MI [1]. A second 1310 nm laser (Laser 2) was chosen whose chirp was significantly better than the first, and whose system performance is representative of today's cable TV DFB lasers. This was done to determine the impact on the resultant values of CSO and CNR.

The best CNR performance occurred when operating the 1550 nm laser over standard single-mode fiber. This was primarily due to the low attenuation of standard single-mode fiber at 1550 nm relative to any other combination of fiber and operating wavelength. As expected, operating at 1550 nm over dispersion-shifted fiber proved to have the next best CNR. A relative comparison of the CNR performance between standard and dispersion-shifted fiber at 1310 nm and 1550 nm is shown in Figure 2.

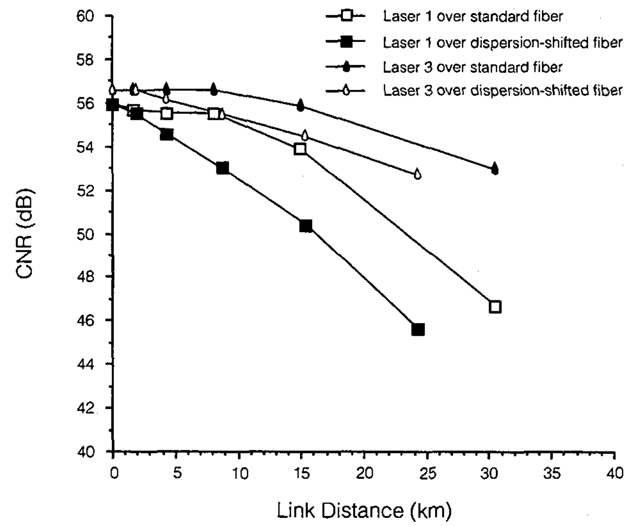


Figure 2. Relative CNR Performance Over Standard and Dispersion-Shifted Fiber

Laser chirp also impacts the overall system CNR. The more chirp a laser has, the lower the spectral density of the noise becomes, which results in a higher CNR [2]. This is a result of the noise due to Rayleigh backscattering within the fiber, coinciding within the same bandwidth as the laser chirp. The results indicated that, given equivalent laser output powers and similar fiber loss per length for the respective fibers, the best CNR performance at 1310 nm will depend upon which laser has the largest chirp. Figure 3 illustrates the effect of laser chirp on CNR performance.

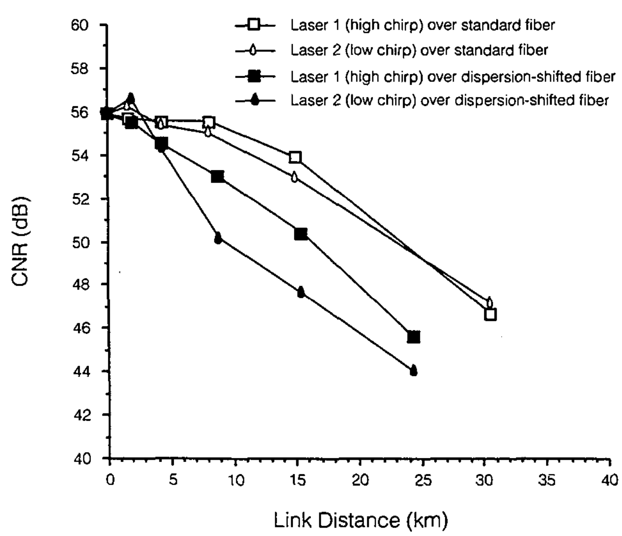


Figure 3. The Effect of Laser Chirp and Fiber Type on CNR

The best CSO performance occurred when operating at 1310 nm over standard single-mode fiber or when operating at 1550 nm over dispersion-shifted fiber. The worst CSO degradation occurred at channel 40 for both standard single-mode fiber operating at 1550 nm, and DS fiber operating at 1310 nm. Figures 4 and 5 show the CSO performance of channels 2 and 40 for standard and dispersion-shifted fiber for the two operating wavelengths, 1310 nm and 1550 nm, respectively.

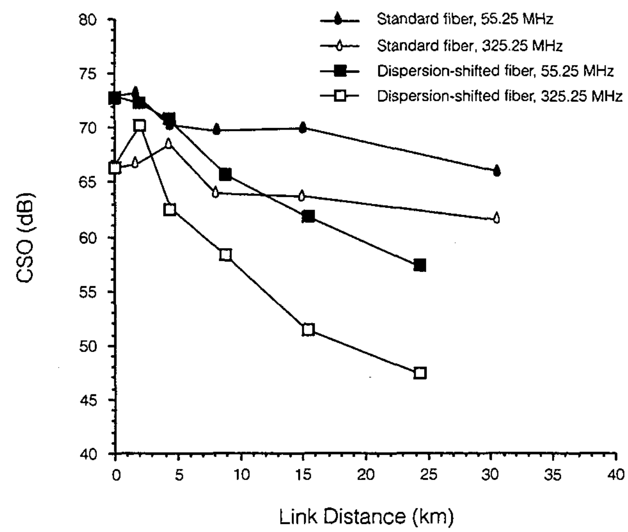


Figure 4. Laser 1 (1318 nm) Frequency Dependence of CSO

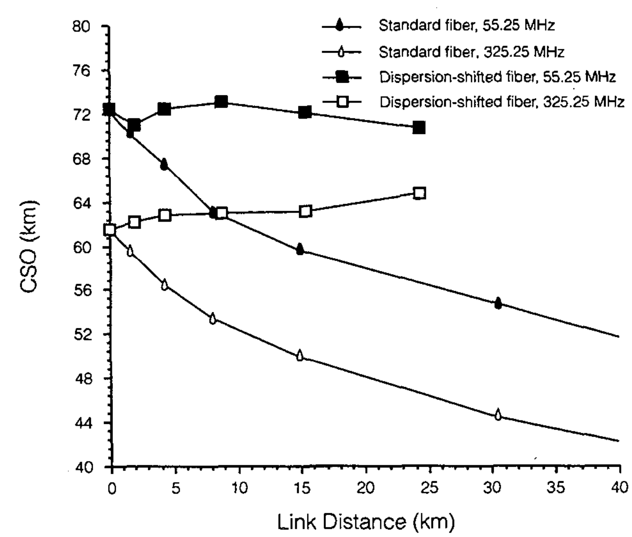


Figure 5. Laser 3 (1546 nm) Frequency Dependence of CSO

It was also observed that some amount of negative dispersion or dispersion compensation is required to improve CSO performance of 1310 nm operation over dispersion-shifted fiber. Figure 6 shows that when operating laser 1 (1318 nm) over dispersion-shifted fiber, the best CSO performance occurs at approximately 2 kilometers (km). It also shows that the CSO performance for laser 3 (1546 nm) over dispersion-shifted fiber appears to increase with length and is estimated to peak at around 50 km. In both cases, it appears that some amount of negative dispersion is helpful to partially cancel the laser or system CSO.

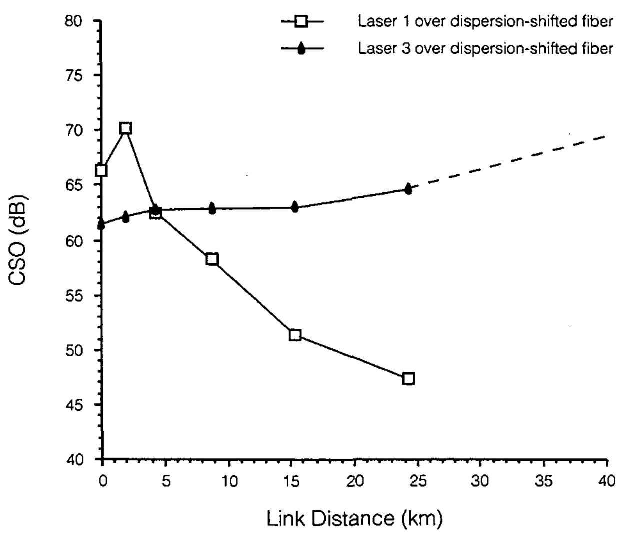


Figure 6. Channel 40 (325.25 MHz) CSO Performance Over Dispersion-Shifted Fiber

In conclusion, the results indicate that a performance tradeoff must be made between CNR degradation and CSO performance when considering 1310 nm and 1550 nm video transmission over standard and dispersion-shifted single-mode fiber.

## POTENTIAL CABLE TV APPLICATIONS

Although, 1310 nm AM video transmission over standard single-mode fiber offers excellent CSO performance, transmission distances typically are limited to less than 30 km due to CNR degradation. Transmission at 1550 nm over standard fiber can extend the reach of cable TV links beyond 30 km, but will require some type of CSO compensation for links beyond approximately 4-5 km in length, depending upon system application.

By using 1550 nm transmission over dispersion-shifted fiber, link distances can be extended beyond 40 km with good CSO and CNR performances. Dipsersion-shifted fiber could also be used at 1310 nm without any form of CSO compensation to achieve good performance links up to 5-10 km and possibly further depending upon laser performance. Employing typical CSO compensation techniques would significantly enhance dispersion-shifted fiber's performance at 1310 nm over even longer distances. The 1310 nm window for dispersion-shifted fiber could conceivably be used to provide "narrow cast" programming, a return signal path, or other services with lower performance requirements.

Dispersion-shifted fiber could be included with standard single-mode fiber within the same cable and used for appropriate sections of the cable TV plant to provide 1310 nm/1550 nm system capability. For this hybrid cable design, the dispersion-shifted fiber could be used with optical amplifiers to provide the low cost solution for broadcast video. While the standard single-mode fiber could be used for narrow casting or other services with conventional 1310 nm equipment.

## SUMMARY

In summary, dispersion-shifted fiber has been shown to be a technically viable option for 1550 nm window cable TV applications, particularly for longer distance AM links. In addition, dispersion-shifted fiber's 1310 nm performance enables good broadband AM video transmission up to 5-10 km or further, depending upon the laser performance, without CSO compensation. Distances could be extended and performance improved by CSO compensation techniques.

Dispersion-shifted fiber's acceptance and use within the cable TV industry will be determined by its performance tradeoffs relative to other technical alternatives.

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# TIME SELECTIVE SPECTRUM ANALYSIS SEPARATES FREQUENCY DOMAIN CHARACTERISTICS OF: VERTICAL INTERVAL TEST SIGNALS, QUIET LINES AND EQUALIZING PULSE TO EVALUATE BANDWIDTH, NOISE AND INTERMODULATION PRODUCTS NON-INTRUSIVELY ON A CATV SYSTEM.

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## ABSTRACT

Complex modulation, crowded frequency spectra, and the demand for higher signal quality require that high dynamic-range measurements be performed on operating systems without interrupting service. The problem of locating the point at which the signal has gone bad is further complicated when testing large systems containing many different functions. The ability to view at a specific instant a signal's spectrum in the presence of modulation is key to non-intrusive diagnosis of system problems.

This paper is organized in a presentation style format with main points summarized in the boxes and corresponding dialog directly to the right of the box.

### Time Selective Signal Analysis Is Not New

- Radar uses timed-return signal analysis for range determination
- Fault locators use a timed return for computing distance to fault
- Some network analyzers use time- and frequency-domain reflectometry

An analogy to gating can be seen in the delayed sweep display mode in an oscilloscope, with its long time record in one trace and delay and width adjustments to intensify and then magnify a portion of the signal in a second trace. The difference is that a spectrum analyzer operates in the frequency domain. Also a stroboscope adjusted to the proper rate will stop a rotating or vibrating object.

### Applications to CATV Shown in 1983

- External circuitry for timing control
- Z-axis modulation on a spectrum analyzer

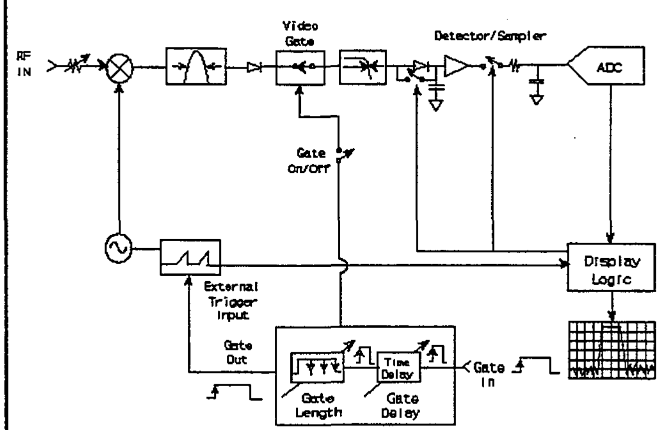
John Huff has used TSSA techniques he developed since 1970 to perform CATV system proofs. However, until recently, manufacturers had not embraced these techniques. Now, TSSA proves to be the only way to view certain types of impairments.

## Use of TSSA in 1992

- Digital communications -- Time division multiple access allows individual transmitter characteristics to be measured
- Disk drives -- Testing sector quality
- VCR -- Testing video-head quality

Signals having more than one contributor to the spectrum or that have a timed sequence of modulation require TSSA for determining system parameters.

## A Video Time Gate




gate16.cgm

Digital signal processing can be performed in several different ways. One method takes a continuous stream of data and later analyzes sections of desired data for its characteristics. This process requires a relatively large data memory, especially for low repetition-rate signals; it also requires specialized firmware processing for the characteristics desired. A second approach controls the start and stop of sweeping, allowing data to be obtained only during a desired time period. This approach is limited by how quickly the sweep can be stopped and started. A third approach takes continuous data and eliminates the undesired data, leaving the desired data for display. While this approach places the burden of control on the user to configure the analyzer for a particular measurement, it makes the fewest requirements on software to develop new measurements. This approach was taken in designing video time-gated spectrum analyzers.

## Limitations on Gated Measurements

- Resolution bandwidth (RBW) settling time
- Video bandwidth (VBW) settling time
- Sweep time
- Span/RBW ratio
- Gate-timing resolution

## Resolution Bandwidth

- Gate delay > 2/RBW for best accuracy
- RBW > 2/pulse modulation on time
- Proper gate delay excludes RBW skirt transient response
- RBW sets noise floor for measurement

The resolution bandwidth filters appear before the gate in the block diagram, so they are subject to all modulation entering the analyzer. The pulse rise and decay times are limited by the RBW setting. The modulation pulse width (time) may dictate the RBW setting to capture the amplitude accurately. These times are meant to cover worst-case situations of pulse modulation and production tolerances of the hardware to obtain the best accuracy. Reducing these times may be in order, depending on the specific situation. The dynamic range of the measurement is controlled by the signal level and the RBW and also is augmented by the correction algorithms used.

## Video Bandwidth

- VBW > 1/gate length minimum
- Not useful for post-detector filtering

The video filters appear after the gate in the block diagram and therefore cannot be used for post-detector filtering. To prevent measured amplitude degradation, the filter must rise from zero to its final value during the gate-length time. A low-pass filter bandwidth is half that of a bandpass filter bandwidth. Therefore to prevent the system bandwidth from being degraded, the video-filter bandwidth should be even wider than the minimum for situations where RBW selection is less than optimum.

## Sweep Time

- Sweep time > number of x-axis data points/signal repetition frequency
- Video averaging increases measurement-acquisition time

The x-axis of the spectrum analyzer display has a discrete number of data points. At least one burst per bucket must be captured. If not, dropouts or zero values will exist in some of the trace elements. Video averaging improves the accuracy of the noise floor measurement at the expense of measurement-acquisition time.

## Span/RBW Ratio

- 40 for 0.1 dB 40 x 300 kHz RBW = 12 MHz span
- 100 for 1 dB
- 200 for 3 dB

Unfortunately, we don't know where the gate pulse will occur relative to the RBW. However, we can ensure that one-half the RBW's width for a desired amplitude error is on the order of the bucket size. Since there are 401 buckets on the x-axis, each bucket must be less than half the RBW's width for the worst-case error allowed.

## Gate Resolution and Range

**Gate delay**

- 1 $\mu$s resolution
- 1 $\mu$s to 65.5 ms range
- ± 1 $\mu$s accuracy

**Gate length**

- 1 $\mu$s resolution
- 1 $\mu$s to 65.5 ms range

Gate resolution and range are determined by the clock frequency used, the length of counters following the clock, and the synchronism of the gate input in relation to the clock.

## Review of Measurement and Computational Techniques

- Bandpass filter/preamplifier
- Composite distortion
- Attenuator test
- Noise-near-noise calculation
- Distortion-near-noise calculation
- Adjacent signals versus RBW

## Bandpass Filter and Preamplifier

- Spectrum analyzer may add distortion and noise
- Preamplifier raises system noise 10 dB above analyzer noise
- Bandpass filter limits signals into analyzer
- Extra hardware and manual control complications

## Composite Distortion

- Number of beats increases drastically with number of carriers
- Beat accumulation based on frequency spacing
- Beat level builds up relatively as 10log(number of beats)
- Measured level based on synchronism and APL of contributors
- Distortion creation mechanisms are not flat versus frequency

## Beat Locations in Standard Channel Plan

**General channel equation for visual carriers**

- (N x 6 + 1.25) MHz for typical channel
- (M x 6 + 1.25 + 4) MHz for channels 5 and 6

**General beat location**

- Third order A ± B ± C & 2A ± b & 3A
- Second order A ± B & 2A

A spectrum analyzer's noise figure, and second- and third-order distortion levels may not be good enough to perform carrier-to-noise, composite-triple-beat (CTB), and composite-second-order measurements without error unless the operator exercises careful control of power input at the first mixer. External hardware is less desirable because of the likely errors introduced.

Creation of composite distortion is complex, as is its measurement. This complexity is due to the simple form (A+B-C), used for ease of analysis and basic measurement setup, coupled to the non-ideal components of diverse system configurations. Only the visual carriers are considered for beats measurement locations of CTB and CSO since they have the largest peak power, but have average power on the order of the audio carrier level. The color subcarriers, FM and digital audio and data services must also be considered for completeness. For measurement of beat products, the NCTA recommends 30 kHz RBW to include all beat products with small frequency differences in the same measurement. The NCTA also recommends 30 Hz VBW to average the peak excursions to a single value. The equation for composite triple beat, CTB = (A+B-C), yields distortion products "on visual carrier" except for channels 5 and 6, which are offset from the others. If (A+B+C) is used, not as many in-band beats are produced, but they accumulate at the high end and are 2.5 MHz above the visual carrier. If (A-B-C) is used, again not as many in-band beats are produced, but they appear 2.5 MHz below the visual carrier, which is very close to the color subcarrier of the channel below. When all carriers are considered, a computer may require days to calculate all the combinations and sum up the occurrences. Yet the process would still not be complete. If a device's distortion generation mechanism is not flat versus frequency and it changes with temperature, then these theoretical curves would be incomplete. Gating allows one to look for intermods and backtrack until the cause is found, without interrupting service.

## Attenuator Test

- Carrier levels don't move with attenuator change

**For distortion generated by spectrum analyzer**

- Second order distortion moves at twice the attenuator step size
- Third order distortion moves at three times the attenuator step size

**If no spectrum analyzer contribution**

- Carrier to distortion ratios remain constant with attenuation or input level change

Carrier frequencies and levels into the first mixer determine the frequency and level of a distortion product. When the attenuator value is increased, IF gain is also increased. The signal level does not change because of this attenuator-to-IF gain coupling. For small signal levels, if the level to the mixer changes, then internally generated distortion changes at the indicated rate. For composite distortion, if the internally generated distortion is more than 10 dB below the system distortion, then the analyzer's contribution can be considered negligible.

## Noise Near Noise Calculation

- Average noise power adds linearly for same RBW
- Total noise = system noise + analyzer noise
- Analyzer measures dBmV for increased dynamic range
- System noise = $10\log[10^{\text{total}/10} - 10^{\text{analyzer}/10}]$

If system noise equals analyzer noise, the total measured noise will be 3 dB higher. This equation allows accurate measurement of system noise closer than 10 dB to the analyzer's noise by backing out the analyzer's contribution.

## Distortion Near Noise Calculation

- Uncorrelated distortion products appear like noise
- Beat level = $10\log[10^{\text{measured beat level}/10} - 10^{\text{measured noise level}/10}]$

This equation allows measurement of composite distortion products closer than 10 dB to the noise level. It also allows the attenuator to be set high enough to keep analyzer distortion well below system distortion. See reference 5.

## Adjacent Signals

- Power of one signal affects measurement of power level of adjacent signal.
- Use narrower RBW if possible.
- Calculate effect if narrower RBW use is not possible.

The operation manuals and application notes for spectrum analyzers include use of the resolution bandwidth to resolve close and unequal signals. Signal modulation will determine what RBW must used and also whether a correction should be applied.

## Gated Quiet Line

- Distortion products
- In-band noise level and slope
- Mismatch, multipath
- Local insertion contribution

Present during the vertical-retrace interval of a typical video signal are horizontal lines with no modulation (quiet lines) as well as specific vertical-interval test signals (VITS) and closed-caption or data signals. As an example, a spectrum analyzer can be set up to build a frequency domain sweep of samples acquired during a quiet line, allowing the signal to be viewed as if no other modulation were present. This setup allows one to see composite second order distortion products at +/-750 kHz and +/-1.25 MHz relative to the video carrier on any channel. In-band noise can also be measured along with noise-floor slope to determine incoming signal to noise, as well as signal to noise at any other RF location in the system.

## Implementation:

**Two spectrum analyzers**

- Trigger receiver: spectrum analyzer with TV line trigger and fast time domain sweep
- Gated receiver: spectrum analyzer with video time gate

**Tuner/demodulator and one spectrum analyzer**

- CATV demodulator with composite video output
- Gated receiver: spectrum analyzer with video time gate and TV line trigger with external baseband composite video input

One approach to perform these gated measurements uses two spectrum analyzers. One spectrum analyzer with AM/FM demod/TV synch trigger and fast time-domain sweep is used as the demod and trigger source. A second spectrum analyzer with video time gate is used to perform the gated frequency-domain measurements. A second approach uses a CATV demodulator with composite baseband video output and an spectrum analyzer with video time gate and TV trigger with a special composite baseband video input. TV trigger out then is connected to gate in. An oscilloscope can aid in viewing the gate setup in relation to the horizontal line: Channel 1, connect aux video out from trigger source analyzer; channel 2, connect gate out from gated analyzer; external trigger or channel 3, connect TV trigger out of trigger-source analyzer.

## Gated Trigger Source Setup

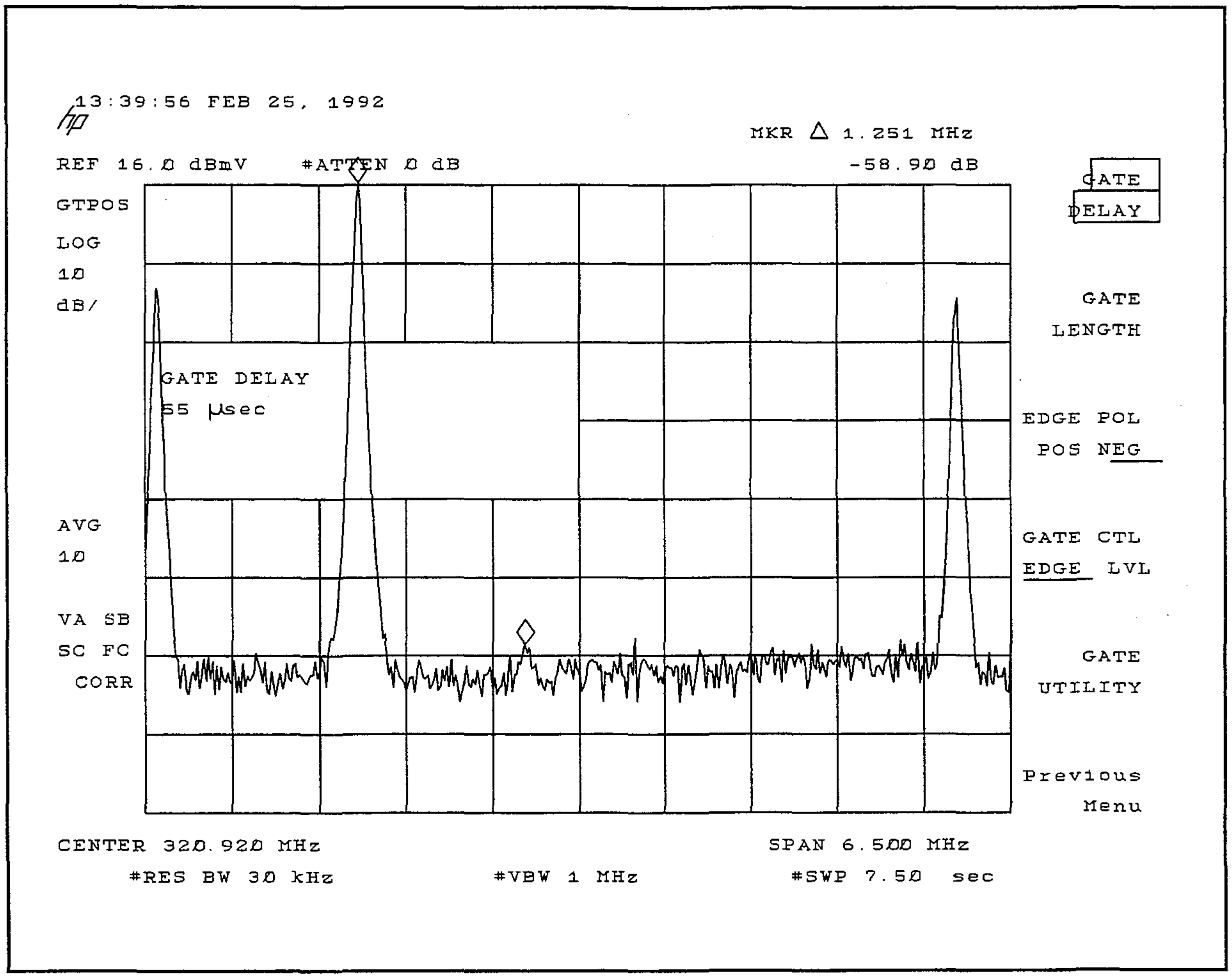
- Visual carrier frequency
- TV line number
- Field identification

For the two analyzer approach, the trigger source set-up is as follows: Set center frequency to desired visual carrier frequency of channel to be tested (55.25 MHz for channel 2). Then push TRIG hard key in control section, and then push TV TRIG softkey, which automatically sets the analyzer to 0 Hz span, 1 MHz RBW, 100 $\mu s$ sweep time, linear detector, sample mode and appropriate reference level. If signal level is low, then manual selection of 0 dB attenuator may be appropriate to get stable triggering. Set the analyzer to trigger on a quiet line, which will be between lines 10 and 20 depending on test signals present and closed caption or data being transmitted during the vertical retrace interval. To reduce sweep time of the second (gated) analyzer, look in both odd and even fields for quiet time (0 IRE level), and then select TV TRIG, vertical interval (VERT INT).

## Gated Analyzer Setup

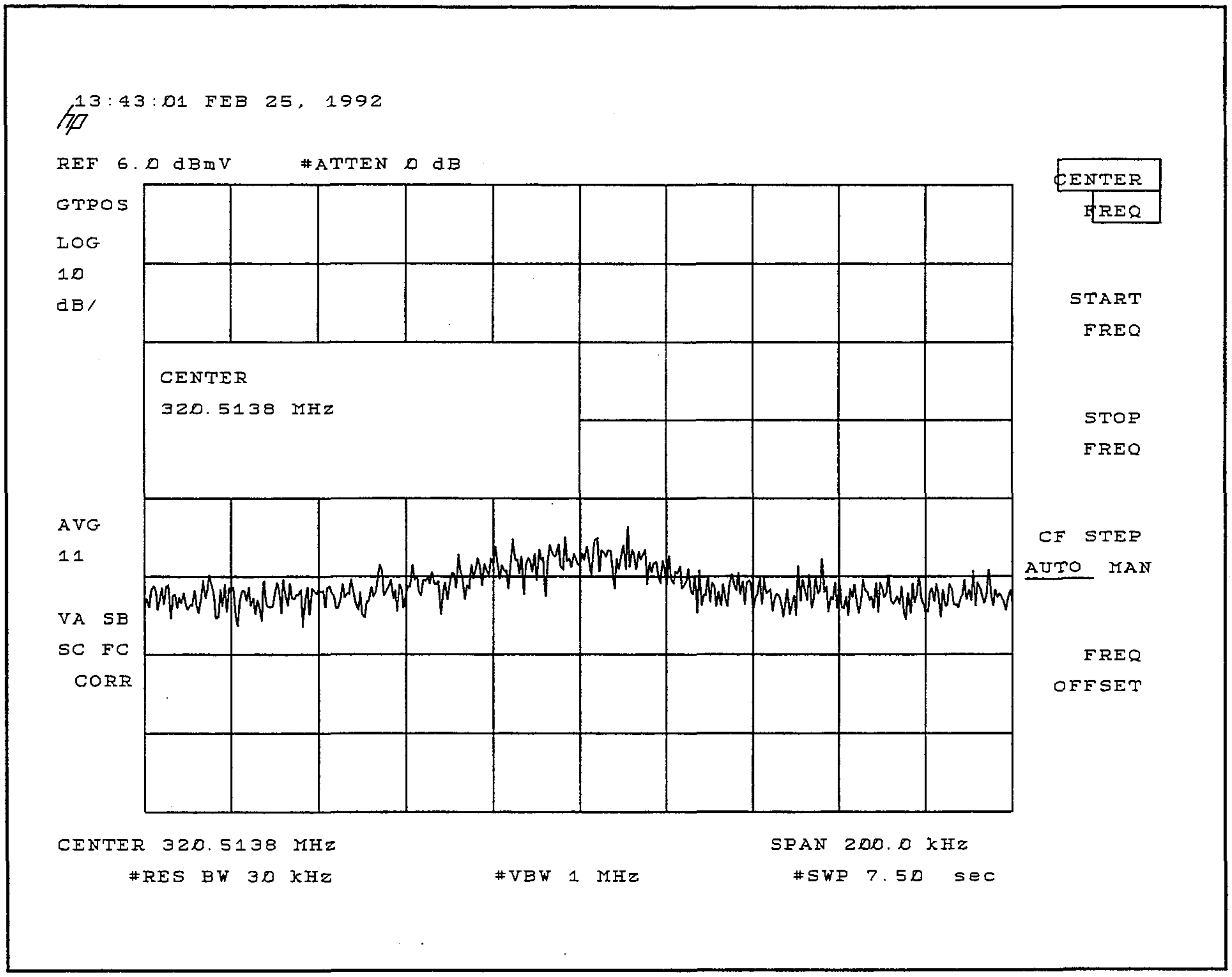
- RBW
- VBW
- Sweep time
- Gate delay
- Gate length
- Negative edge

The gated analyzer is set up as follows: Connect TV trigger output from trigger-source analyzer to gate in on the gated analyzer. Set span to 6 MHz, RBW to 30 kHz, and VBW to 3 MHz. Set center frequency to visual carrier frequency plus span/2. Set sweep time to 7.5 sec. Gate setup: set gate delay to 54 $\mu s$ and gate width to 1 $\mu s$, and set for negative edge operation so reference edge for gate delay is trailing edge of sync. Turn gate ON, video average set to 10.



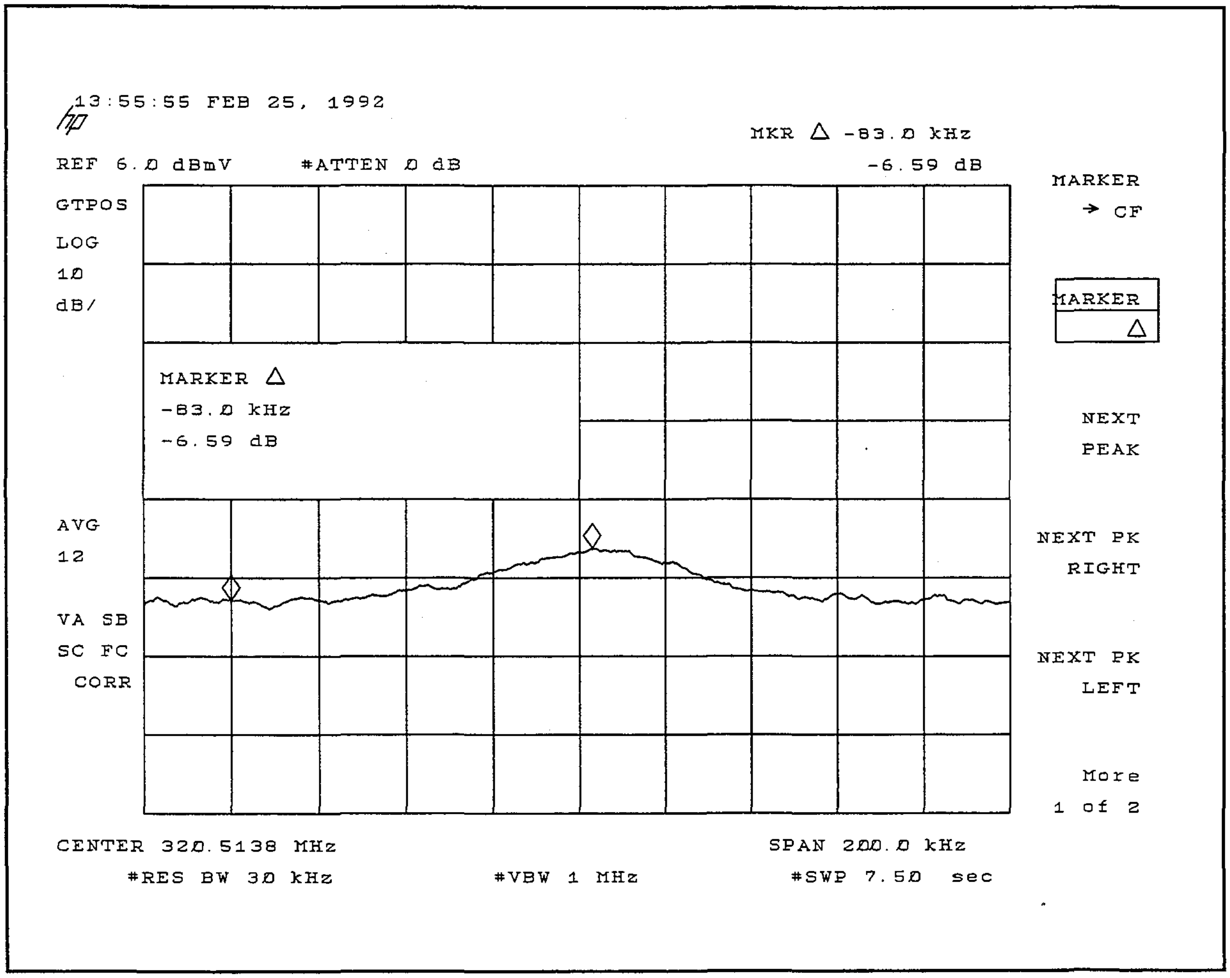

quiet1.hpg

Gated quiet line of channel 39 shows a frequency-domain representation of 0 IRE carrier level to average noise and average intermods. The accuracy of the carrier level is questionable, since the span to RBW ratio is greater than 200. The display does provide a quick look at the whole channel. To properly average noise, the carrier amplitude may have to be set above the top of the screen. If the averaged noise is not above the 60 dB graticule line, then push amplitude and set the reference level to make this so. This allows proper averaging of negative as well as positive peaks. You may have to select attenuator 0 dB if the signal level is low or a large dynamic range is needed. An attenuator test may need to be performed to check that the analyzer is not the cause of distortion. A noise-near-noise calculation will be needed if system noise is less than 10 dB above the analyzer noise. Attenuator should be set to manual, so that reference level changes don't change the attenuator value and the mixer's distortion performance. See the reference 5 for graphs of C/N, CTB or CSO for default attenuator starting point versus carrier level.




quiet2.hpg

Zoom-in on the +1.25 MHz CSO product. The accuracy of this noise measurement is based in part on the number of sweeps being video averaged, so a trade-off exists between time to acquire the data and the accuracy of the data. The gated positive peak detector is being used, but the gating process uses a 1 $\mu$s gate length so it effectively samples the 30 kHz RBW as long as the sweep time is set for one sample per bucket. Logged noise reads high by 2.5 dB, so subtracting 2.5 dB from the absolute noise level is in order. RBW reads low by 0.5 dB when compared to impulse BW, so add 0.5 dB correction. To refer the noise level to another RBW, use 10log(RBW new/RBW old). For noise measured in a 30 kHz RBW and referred to a 4 MHz RBW, the correction factor would be +21.25 dB. Noise-near-noise correction will also be needed.




quiet3.hpg

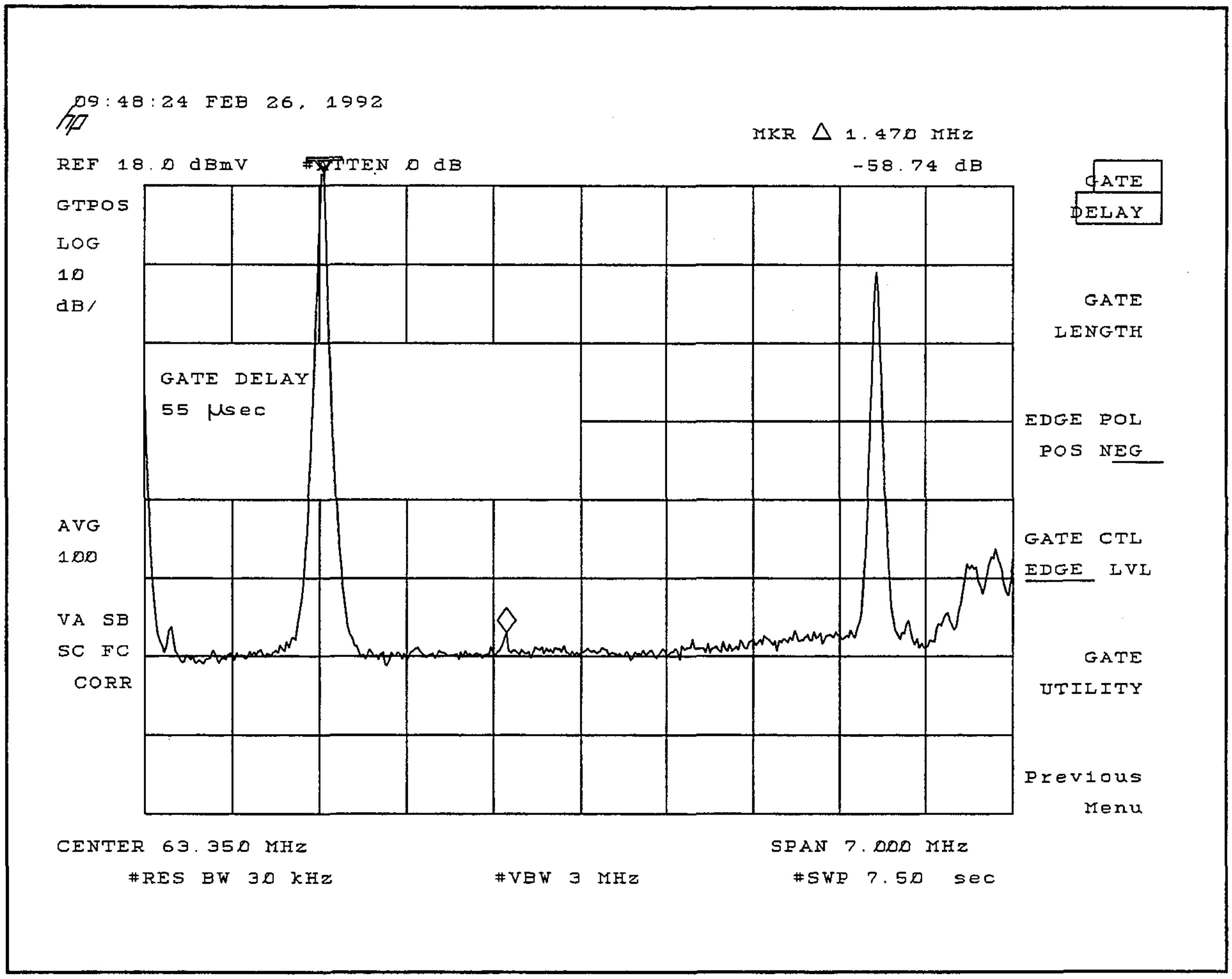
Since the VBW can't be used to average the noise, use (Smooth tra 10;). This command can be executed from the external keyboard. Since the peak is broad, smoothing over 10 buckets (one quarter of a division) is reasonable.

Since the distortion product is closer than 10 dB to the noise floor, its level should be corrected. One way to accomplish this is by looking at the delta level and correcting the peak level accordingly. One also could use the absolute levels to do the correction.

$$\text{Error} = 10\log[10^{0/10} - 10^{-6.6/10}]$$
$$= 10\log[1 - 0.219] = -1.1 \text{ dB}$$

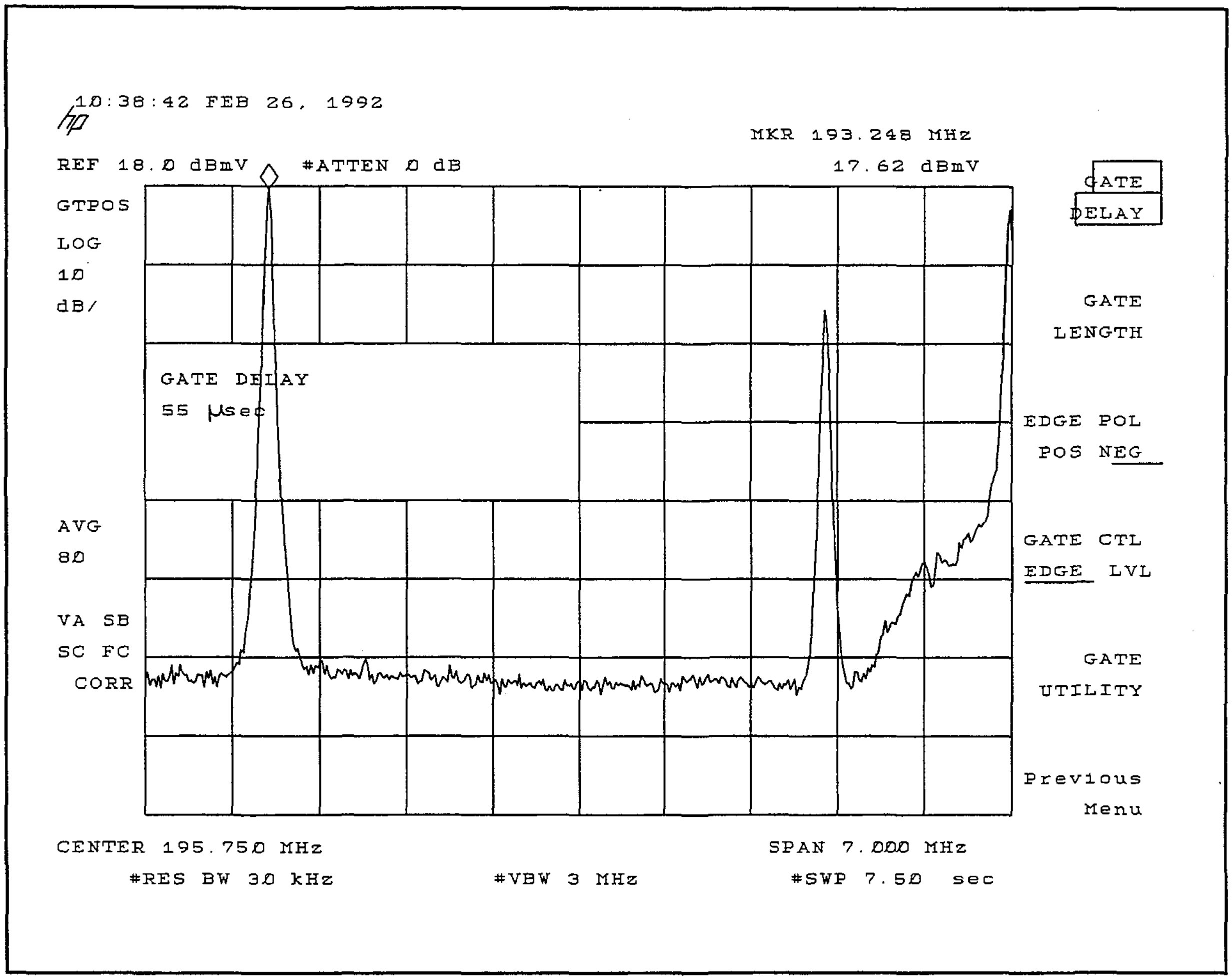
The beat level measures 1.1 dB higher due to the added noise power. Then, subtracting 1.1 dB from the absolute beat level measured would give a more correct beat level.

Most intermods vary in amplitude versus time; therefore, video averaging may conceal this fact. Setting the marker to an intermod and pressing marker to center freq, span 0 Hz and video average off may show its time varying nature, if it is far enough above the analyzer noise floor.



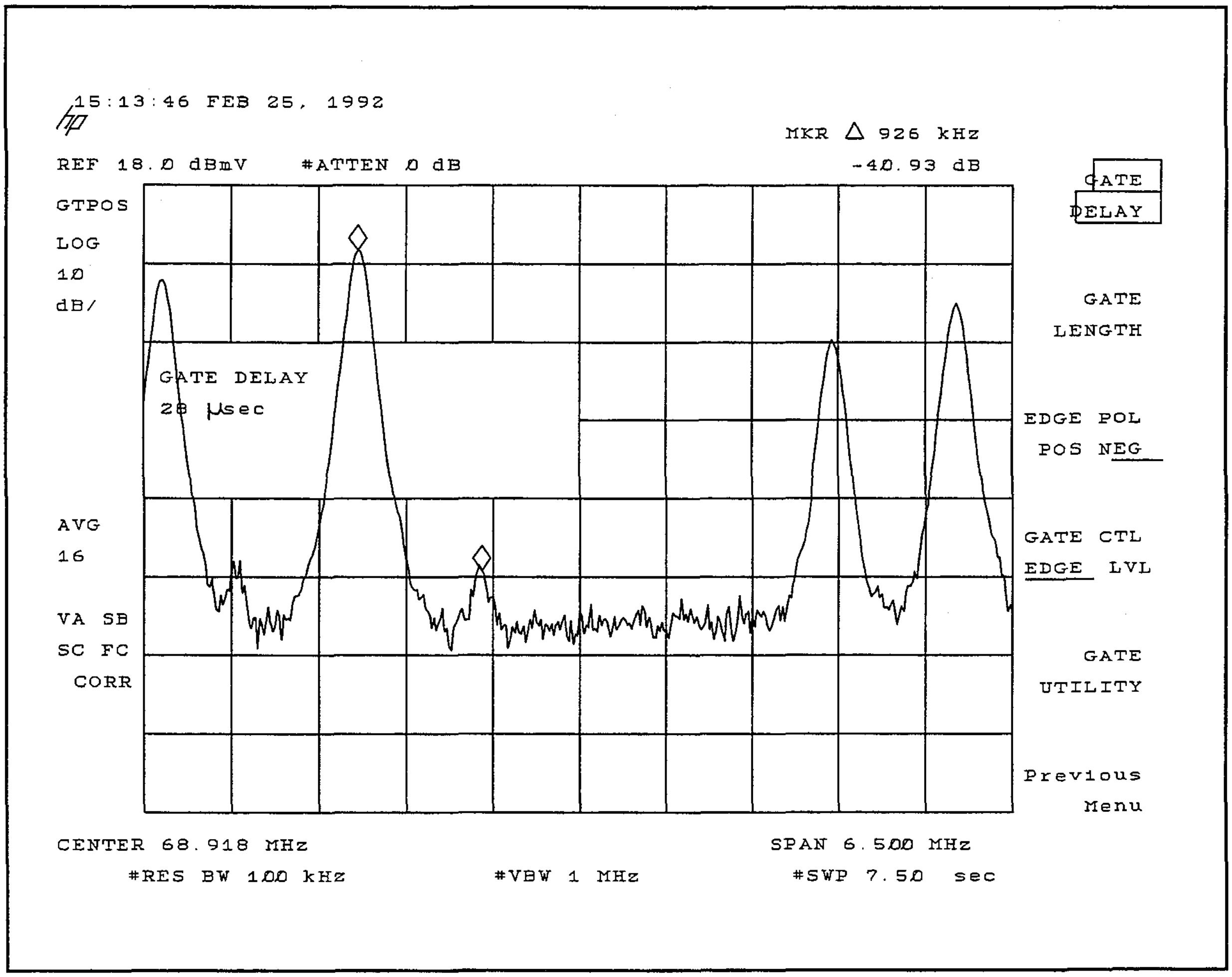
quiet4.hpg

The gated quiet line of channel three shows a +1.5 MHz intermod, possibly from visual - aural + next higher visual in combination with similar beat products. Note also the noise floor slope.




quiet6.hpg

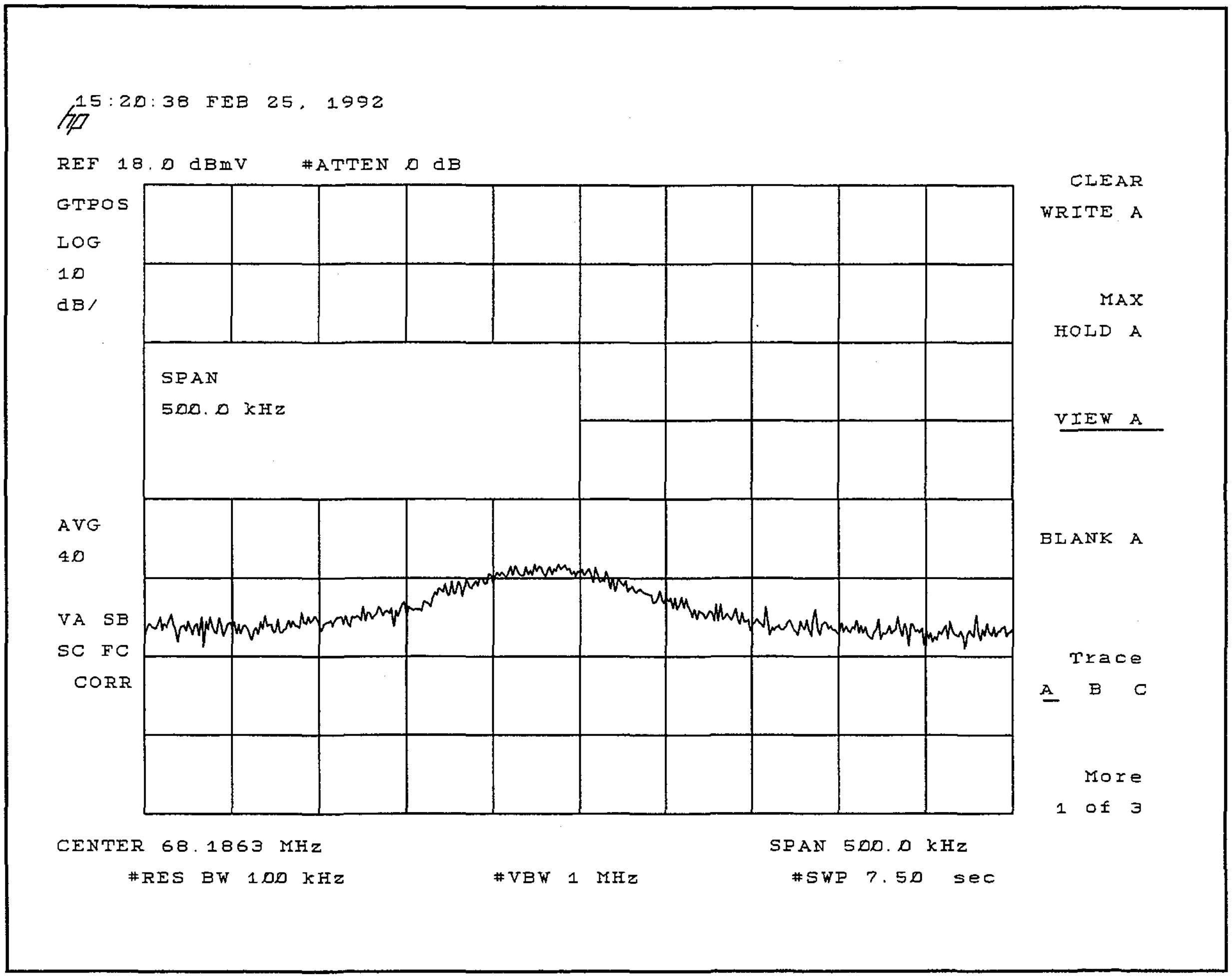
The gated quiet line of channel ten shows a negative noise floor slope. The +750 kHz intermod seen is a CSO product caused by the different spacing of channels 5 and 6 relative to the rest of the channels.




vir4.hpg

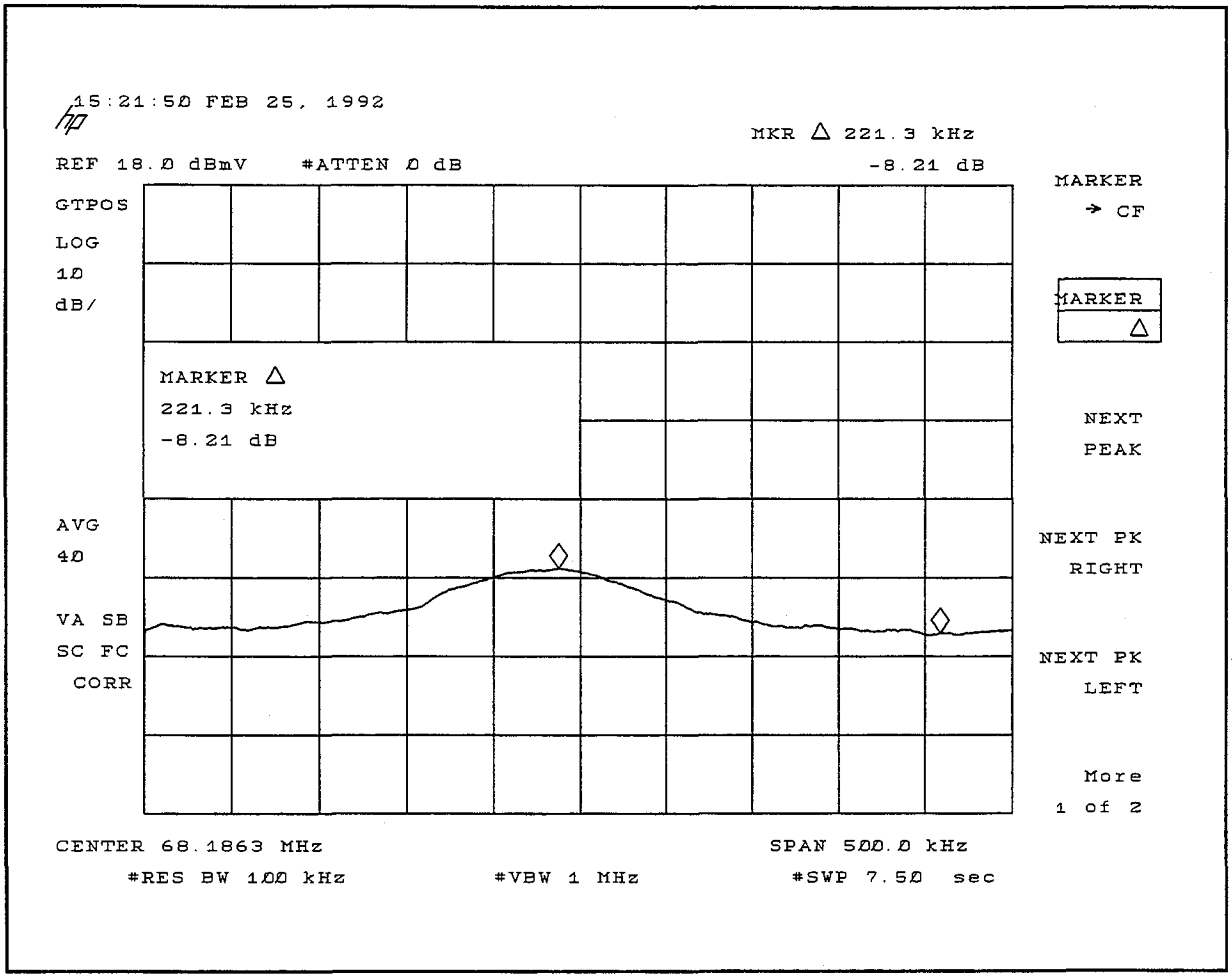
Shown is gated VIR. On the trigger source analyzer, set the TV line number to the vertical interval reference signal. Check if the reference signal is in both fields, and, if so, select vert int (vertical interval). Set the sweeptime to 7.5 sec if the signal is in both fields and 15 sec if not.

Set RBW to 100 kHz so it will respond to less than a full line duration of burst. Set gate delay to 28 μs and gate length to 1 μs. Video average is on. This measurement is a third-order distortion measurement of visual - color + aural carriers at +920 kHz relative to visual carrier. Also seen is a -920 kHz product relative to the visual carrier due to visual + color - aural.



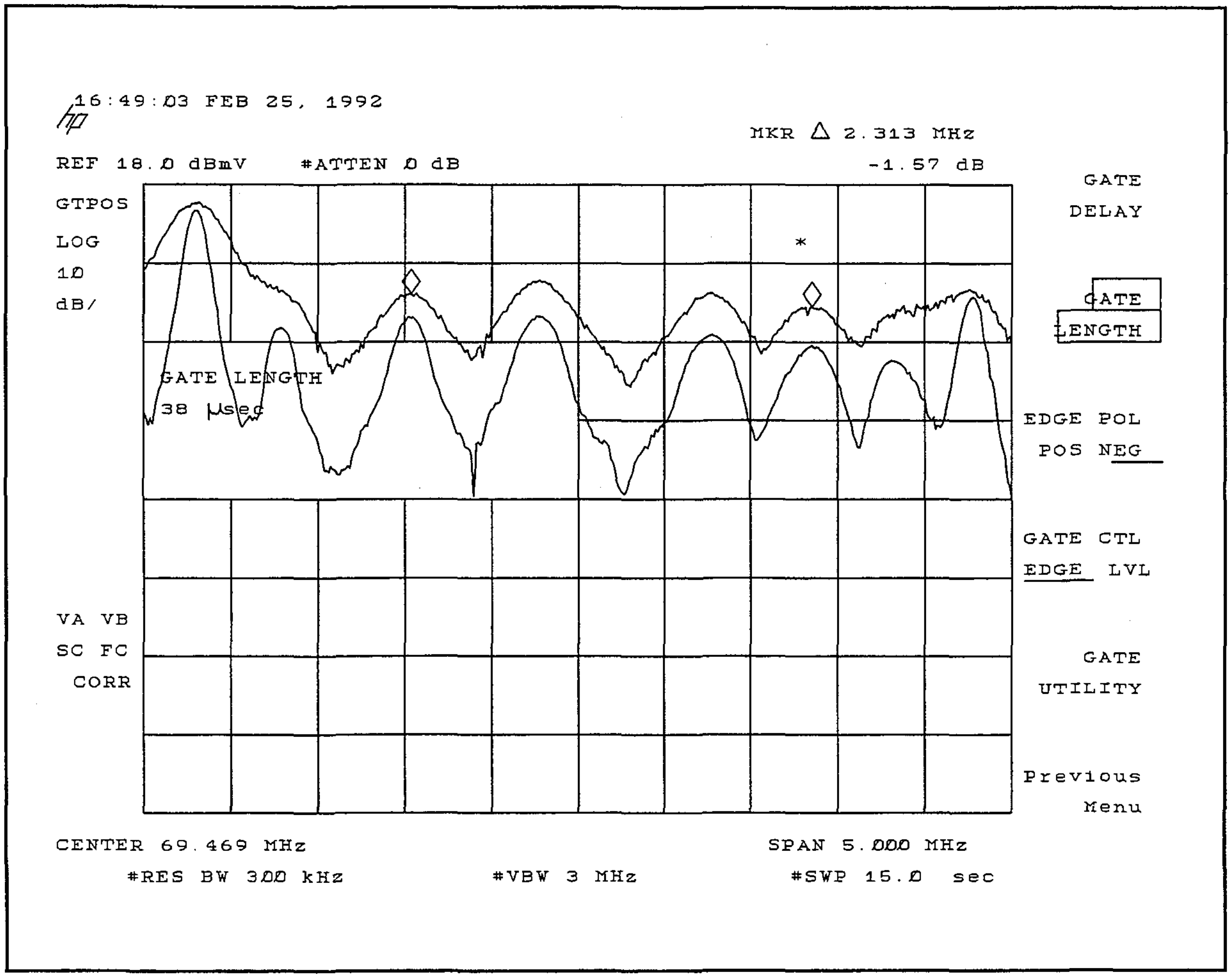

vir6.hpg

Zoom in on the +920 kHz distortion product. This measurement is most useful on broadcast channels or individual channel processors. The intermod level may be very low and difficult to determine, but using a large number of video averages and performing a distortion-near-noise calculation yields a useful result.



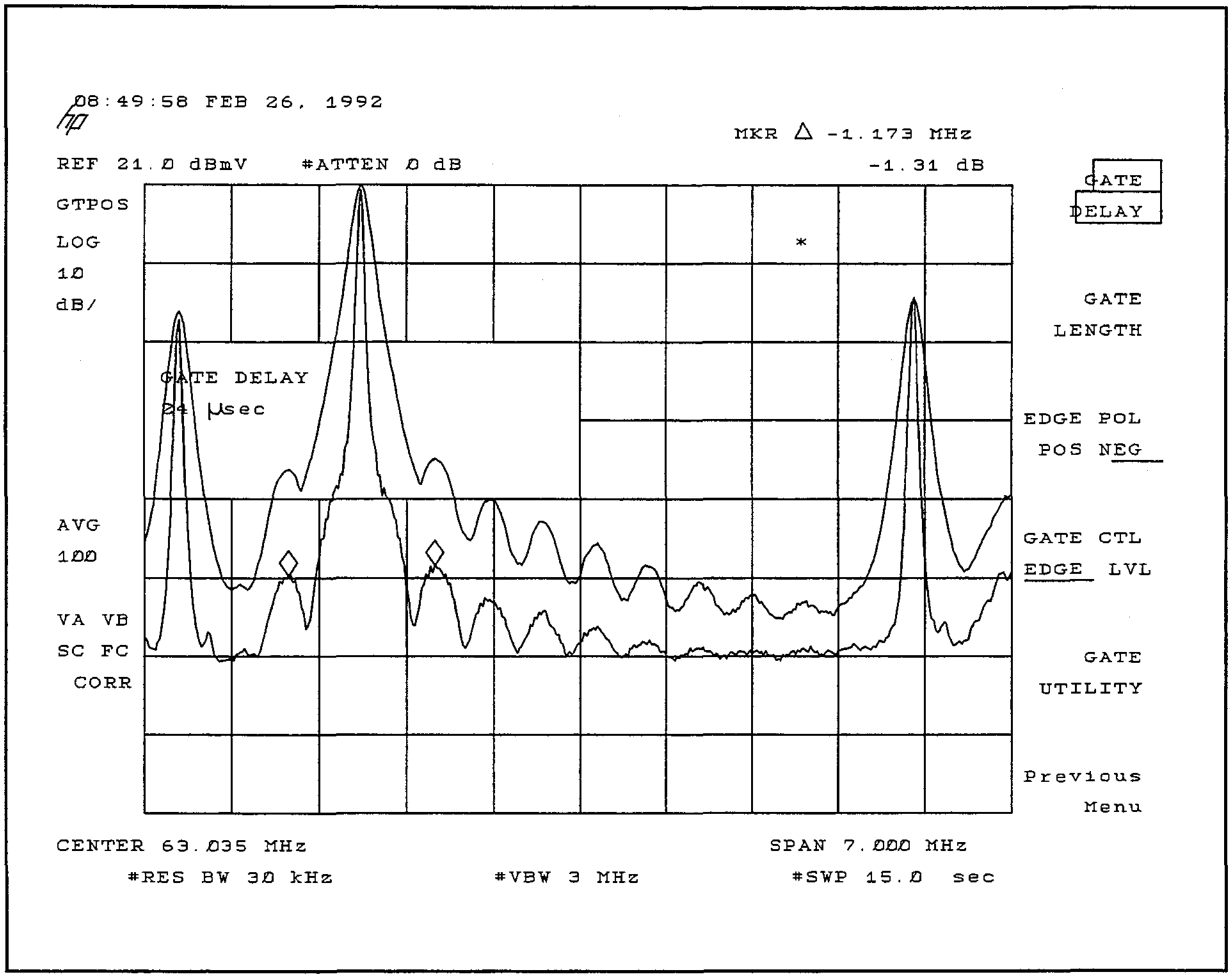

vir6.hpg

Smooth tra 10 also helps resolve the distortion.
Error = $10\log[10^{0/10} - 10^{-8.2/10}] = -0.7$ dB




mb1.hpg

Gated multiburst: Gate delay is 18 $\mu$s and gate length is 38 $\mu$s. Set the RBW to 100 kHz to resolve all bursts, but due to small pulse width of bursts, a RBW of 300 kHz gives better amplitude accuracy. For 100 kHz RBW, the pulse time would have to be 2/100,000, or 20 $\mu$s. But since this is not the case, you would note a reduction of the pulse amplitudes. For 300 kHz RBW, the pulse time would have to be 2/300,000, or 6.66 $\mu$s, and this is not the case. That case would be for a worst-case situation, which is not present here. The difference between the pulse amplitudes between the two RBWs is 3 dB for a 1.25 MHz burst, 4.75 dB for 2 MHz, 5.5 dB for 3 MHz, and 6 dB for a 3.58 MHz burst. Cutting the pulse time requirement vs RBW in half would amount to about a 1 dB loss of pulse amplitude for a typical analyzer. Therefore, 300 kHz for RBW is a reasonable value to use to get an idea of the frequency response. If one were to look at both sides of the visual carrier, vestigial sideband filter performance could be viewed, as long as the -1.25 MHz CSO wasn't added in.

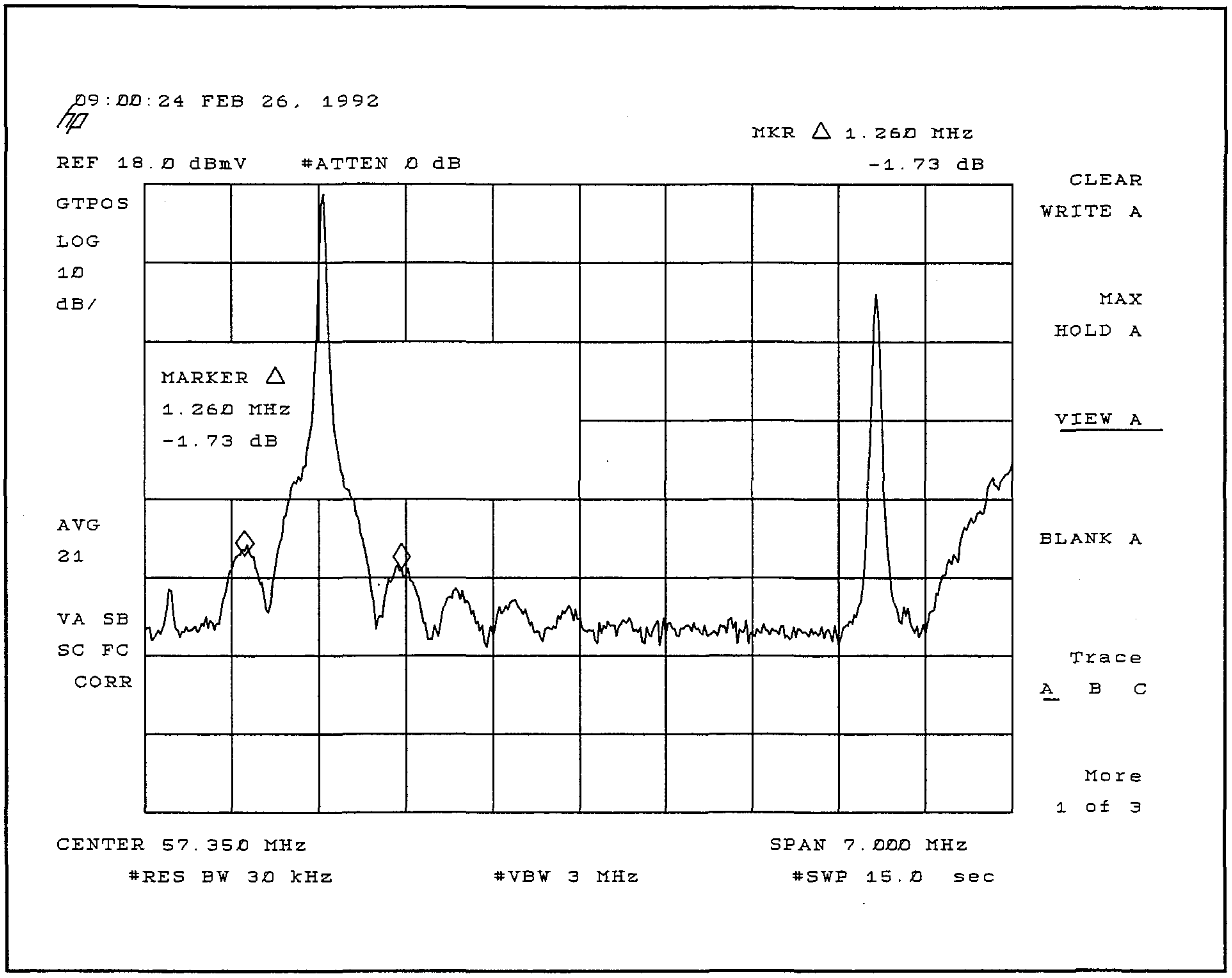



ep2.hpg

**Gated Equalizing Pulse of Line 9 Even Field**

- One sided sinx/x pulse response
- Bandwidth vs. noise
- Vestigial sideband filter performance
- Group delay effects

The RBW filters are not allowed to charge fully due to the shortness of the pulse for the RBW used. For the gated equalizing pulse, gate delay is 24 μs and gate length is 15 μs. Trigger source is set to line 8, even field, so that the gate pulse will window the line 9 equalizing pulse. The line numbers are not correct at this point because two equalizing pulses per line are present. The depth of the nulls between the lobes is a measure of the steepness of the sync-pulse rise and fall times. The spacing between the nulls is a measure of the sync-pulse width, 1/435 kHz is 2.3 μs. The height of the lobes and the rate at which they roll off is a measure of the channel bandwidth. Overall, this pulse response is a measure of picture quality.




ep3.hpg

Mismatch in first lobe heights could indicate filter misalignment, modulator balance, or group delay problems.

## Residual responses on quiet line

- Impedance mismatch
- Cable faults
- Multipath

Responses of delayed horizontal sync or color burst can be viewed by stepping gate delay and a 1 $\mu$s gate length through a quiet line.

## Summary

**Time gating allows one to see**

- In-band noise and intermods non-intrusively
- Any location in the RF chain
- Mismatch, faults and multipath effects
- Problem causes can be isolated before taking apart

## For the Engineer

- Other portions of test signals can be viewed
- Other test signals can be inserted
- Test signals can be designed for specific use

Gating applications can be extended to many other test signals through thoughtful application of the guidelines presented. With gating, there are fewer restrictions to system analysis, and that analysis becomes much easier to perform.

Several useful references follow.

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## About the Author

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# UNDERSTANDING THE INSTALLATION REQUIREMENTS OF THE CD-X DIGITAL AUDIO SYSTEM

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## ABSTRACT

The following paper describes some of the installation considerations necessary in implementing a satellite-delivered digital audio system into an existing cable system. Areas of discussion include satellite reception, headend configuration, signal carriage, distribution architecture, and input requirements at the subscriber locations. It is important to the cable operator that a system provide the flexibility to adapt within the overall architecture of the system. Systems typically want to preserve bandwidth or simply have very little bandwidth in which to work. The digital audio service must operate well in spectrum not necessarily suitable for video channels. Therefore, the challenge for a digital audio service is to find available spectrum that will provide quality signals throughout the system.

## INTRODUCTION

Since the introduction of satellite digital audio to the cable industry, digital audio systems have been installed at an accelerated rate. While the actual implementation of a digital audio system is fairly straightforward, the different types of system architectures and the limited availability of bandwidth is cause for some consideration prior to installation. The first consideration for the cable operator is how the service is to be received from the transmitting satellite. The antenna size and the type of low noise device depends on which satellite the service is carried and the input requirements of the satellite receiver. Once the service is received at the headend, the operator must determine where the system is able to carry the service. Typically, most systems place the service in one or more of the following areas:

- The guard band from 72 MHz to 76 MHz
- The FM band from 88 MHz to 108 MHz
- The roll-off region above the highest video channel

Once the system has determined the placement of the digital audio service, can the service serve the entire distribution network? Since there are a number of distribution architectures in today's CATV environments, the digital audio signal must also be transmitted by these existing means to all subscribers while maintaining a low up-front capital expense. Current architectures range from direct coaxial tree and branch to AML hubs, to more recent AM fiber nodes. In addition, FM fiber and FM supertrunk applications allow operators to transport services to other remote headends for re-processing and distribution. While maintaining compatibility with system architectures, the service must also maintain all minimum input requirements at the terminal device in the subscribers home to insure optimum performance.

## SATELLITE RECEPTION

While the digital modulation techniques used in digital audio systems may vary depending on the actual design criteria, Quadrature Phase Shift Keying (QPSK) is a common modulation technique used for transmitting broadband digital audio signals over the satellite link. QPSK transmission allows the satellite transponder to reach saturation providing maximum power on the output. Since the link is power and noise limited but has sufficient bandwidth (36 MHz per transponder), QPSK is the best compromise.

Multiple audio formats are time division multiplexed (TDM) at the studio and then fed to a QPSK satellite modulator. TDM multiplexes multiple data channels together into a single serial data stream which provides both an economical and bandwidth efficient means of transmitting large amounts of data over satellite. The signal is then uplinked to the satellite for transmission to the cable systems. The satellite receiving equipment at the headend must process the QPSK signal while maintaining the integrity of the digital data.

In order to receive the digital signal from the satellite, a cable system requires a low noise device to block convert the signal from C-band to the band necessary for the digital satellite receiver. Important specifications when considering a low noise device (LND) for digital audio are the frequency stability and phase noise of the local oscillator as well as its effective noise temperature. Minimum phase noise needs to be achieved since QPSK is a Phase Modulation (PM) technique with four phase states. The phase noise requirements for an LND in a digital audio application is typically -85 dBc/Hz when measured 10 kHz from the carrier frequency compared to -75 dBc requirement for analog video. The frequency stability of the local oscillator depends on the frequency uncertainty of the satellite receiver. The cable operator should request prior to installation that the vendor recommend an LND that is suitable for the application. Recommendations may also be necessary for systems that have a receiving antenna that is being occupied by an LND having an output frequency range other than that required by the digital satellite receiver. If block-up or block-down converters are required, they also must meet the same specifications.

With analog video signals, the video receiver must maintain a minimum carrier-to-noise (C/N) to ensure the signal is free of impulse noise or "sparklies." For digital audio reception, the digital broadcast receiver requires a minimum specified Eb/No to ensure a quality link. This quantity is the energy in one bit divided by the noise power density. It is equal to the C/N multiplied by the bandwidth divided by the bit rate. In some cases, a larger antenna than what is available for the service may be required to achieve minimum input requirements. To determine if an existing link is adequate, the cable operator should consult the vendor prior to installing a digital audio system. the satellite link is perhaps the most critical aspect of the installation.

Headend Requirements

In addition to a digital satellite receiver, the digital audio system requires a device to demultiplex the demodulated output of the receiver. The demultiplexer separates the data stream into smaller channel groups for carriage over the distribution plant. Each demultiplexed output is sent to a headend digital audio modulator. Each modulator output is then combined with the existing video services for normal distribution. The bandwidth requirement for each modulator depends on the number of data channels being modulated and the amount of data included in each channel. If TDM is used, a time division demultiplexer will divide the multiple data channels received from the satellite receiver into smaller channel groups. If frequency division multiplexing (FDM) is used, the multiple data channels received from the satellite receiver will be divided into single data channels.

While TDM provides bandwidth efficiency and is more economical in the number of individual modulators that are required to deliver the service, both methods provide flexibility in the placement of the signal.

The most common digital modulation techniques for carriage of digital audio data in cable systems are Quadrature Partial Response (QPR) and QPSK. Both of these techniques consist of suppressed carrier modulation. The RF carrier is digitally modulated at a particular amplitute and phase state which prepresents two bits of information. QPSK consists of four phase states which are 90 degrees apart. This modulation technique is made up of two signal components knows as in-phase (I) and quadrature (Q), each having two amplitude states. With QPR, a third level is generated between the (I) and (Q) axis to produce a total of nine states. While QPSK provides a better noise performance, QPR is more bandwidth efficient.

Simulcast audio as well as FM broadcasts can be digitally encoded to enhance the digital audio service. While some digital audio services provide audio simulcasts and radio formats as part of the satellite package, others give the operator the flexibility to choose the additional services the system wants to carry on a local level. The S/N and stereo separation at the output of a video satellite receiver/decoder is typically better than 60 dB and provides a quality signal for digital encoding. In order to encode the audio from a satellite receiver or other audio sources, the headend requires additional hardware. Analog to digital encoder are required for analog audio sources while digital to digital encoders are required for digital audio sources.

Also required in the headend is some means of communicating addressable commands to the digital audio terminals. Depending on the system, this can be accomplished with a local addressable controller or via a addressable link from the uplink studio. As with any addressable control system, the billing computer must implement the necessary host transactions to update the database of the digital audio control computer. It is important that the cable operator understand the requirements of the interface and whether or not the billing computer can fully support the system. The operator should consult both the product vendor as well as the billing vendor prior to launching a digital audio service.

Additional hardware may also be required for headends with microwave or fiber hubs. With channelized AML, microwave upconverters convert RF frequencies to microwave frequencies. If the cable system requires the digital audio service to operate in a spectrum that is not active over the microwave, additional microwave hardware as well as channel licensing may be required prior to launching the service. With FM fiber and FM supertrunk applications, services are combined with FM video, transported, demodulated and then modulated at the remote headend location. The same requirements are necessary for a digital audio service except that the demodulators and modulators are unique for the digital audio application.

The actual headend configuration will depend on the needs of the operator. The digital audio system must be tailored to serve all service areas and have upgrade potential to support any future demands. Figure (1) shows a simplified block diagram of a full function digital audio system.

Bandwidth Availability

In many cable systems, available bandwidth is very limited and is generally preserved for expanded video services. Therefore, digital audio services must operate well where analog video signals cannot. Digital audio services are typically carried in either the FM band or in the spectrum above the highest active video channel. Many operators will eliminate FM services that are currently occupying the FM band for a digital audio service since the revenue potential for a premium audio service is much greater. The number of digital audio services that can be carried in a particular spectrum depends on the amount of available bandwidth and the bandwidth requirements of the digital audio system. The number of services that can be carried in the "roll-off" region of the distribution plant depends on the frequency response at the end of the amplifier cascade. When selecting a frequency for carriage of

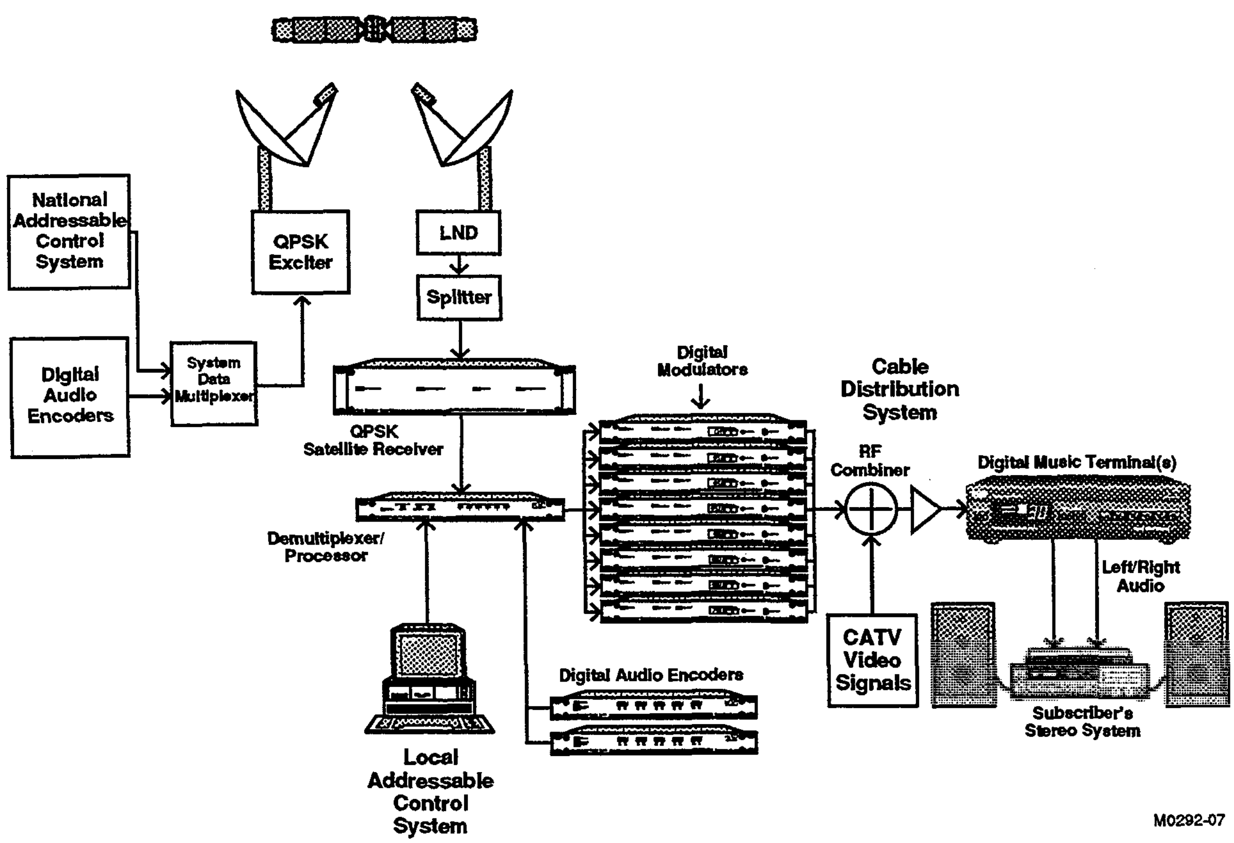


**Figure 1. Digital Audio Configuration**

the digital audio signal over the entire distribution plant, the most important specifications to consider are carrier-to-noise, signal level, and tilt response. These specifications are important to the operation of the digital music terminal.

Digitally modulated signals are more robust than are analog video signals, requiring a lower C/N. Analog signals are also more susceptible to interfering carriers and reflections. Since the interfering source is summed with an AM signal, the receiver has no way to distinguish between the interference and the desired signal. With digital transmission, the receiver only has to determine whether a bit is a "1" or a "0". Therefore, it is easier to reject undesired signals. The result is that available spectrum may be desirable for carriage of digital signals even if it is undesirable for analog video channels.

Smaller portions of a broadband spectrum may also be available. Standard systems have 4 Mhz between channels 4 and 5. This band covers 72 MHz to 76 MHz and is often occupied by a pilot carrier for AML transmission. There is also spectrum below channel 2 (48 MHz-54 MHz) assuming the distribution plant will pass signals in this frequency range. Channel A-1 (114 MHz to 120 MHz) and channel A-2 (108 MHz to 114 MHz) may be available in some systems as well, although both are often used for video channels. Channel A-2 is also used for addressable data carriers and carriers used to detect signal leakage. Since digital audio signals require less bandwidth than analog video channels, they can co-exist with other signals in the same spectrum.

Signal Level

As previously mentioned, both QPR and QPSK are suppressed carrier modulation techniques. Therefore, when modulated, the power is distributed over the entire operating bandwidth. With distributed signals, the level will vary depending on the measurement bandwidth. This is important when notifying the FCC for signals that fall in the FAA navigation and communication bands. The FCC requires offsets if the signal exceeds 38.75 dBmV when measured in a 25 kHz bandwidth and averaged over 160 usec. Since the level of the digital audio signal is typically 5 dB to 10 dB below the video carrier level, the signal level measured in a 25 kHz bandwidth will be much lower than the level of the video carriers and therefore does not require offsets.

The actual signal level may vary depending on where the service is carried. If the roll-off region is used, the actual level may need to be increased to insure that the proper input level and C/N ratio is met at the input of the digital music terminal.

Distribution Architectures

There are a variety of distribution architectures that make up a distribution plant. While conventional tree and branch does well to serve immediate distribution areas, other architectures such as AM fiber and AM microwave allows systems to deliver high quality services to other remote service areas. In addition, FM fiber, FM microwave, and FM supertrunk architectures allow cable systems to transmit signals to remote headends for re-processing and carriage to those service areas as well. Cable systems will use a combination of the above mentioned architectures to deliver cable services throughout the system. It is important that the digital audio service work well under all possible conditions to ensure that the service can reach all service areas.

Conventional tree and branch consists of trunk amplifiers which carry broadband signals to bridger amplifiers, which in turn provide a high level signal to feeder amplifiers. The feeder provides a signal to a series of line extenders, taps, and directional couplers, which provide the proper signal level to the subscriber's home. As mentioned earlier, digital signals are more robust than are analog video signals. Therefore, a digital audio service can survive where video services cannot. With tree and branch, no additional hardware is required for digital audio. The same can also be said for AM fiber. AM fiber improves C/N in a distribution plant by reducing the amplifier cascades. The optoelectronic transmitter takes a broadband input and transmits the broadband spectrum via fiber to an optoelectronic receiver, which provides the same broadband output to the distribution amplifier. Therefore, the digital audio service can be combined in the headend with the video services and fed to the transmitter without any additional hardware.

There are a number of applications consisting of channelized and broadband AML. Channelized AML is made up of individual microwave upconverters that convert the video channel from RF to microwave frequencies. Adding a channel requires an additional upconverter as well as getting licensing for the frequency. The same is true for adding a channel for carriage of digital audio. The hardware and licensing requirements should be discussed prior to installation to insure the service can be transmitted over the microwave link. Alternatives may be to carry the service in existing spectrum that is being carried by the AML. No additional hardware is required for broadband AML since the low power AML transmitter accepts a combined broadband input. This assumes however, that the transmitter has a bandpass that will accept the frequency assignments for the digital audio service.

FM fiber and FM supertrunk require additional hardware considerations as well. Baseband video and audio are typically FM modulated at the origination headend and FM demodulated at the receive headend prior to being AM modulated for distribution. Each FM modulated channel requires 14 MHz of bandwidth. The cable system must find available spectrum to carry the digital audio service over the fiber or supertrunk. At the receive headend, the digital audio signal must be demodulated or reprocessed to filter out other FM modulated services. The signals are then remodulated and combined in the headend for general distribution.

Digital Music Terminal Requirements

It is important for the cable operator to maintain minimum input requirements to the digital music terminal as previously mentioned. The most important input requirements are input level, C/N and tilt response. These specifications are vital to the performance of digital audio service. Once these input requirements are established at the drop location, the installation should insure that the signal quality for the existing video services is maintained. A directional coupler should provide the recommended isolation between ports and have minimal insertion loss. Since the digital music terminal typically requires a lower input level than cable converters, a 6 dB or 9 dB directional coupler is recommended.

The output of the digital music terminal should include a digital output in addition to the left/right analog output. A loop-through option should also be available on the terminal to preserve auxiliary inputs on the stereo amplifier or receiver. The digital output must be compatible with digital audio tape (DAT) recorders, compact disc players and other digital interfaces which use the Sony Philips Digital Interface (SPDIF). A typical subscriber installation is shown in figure (2).

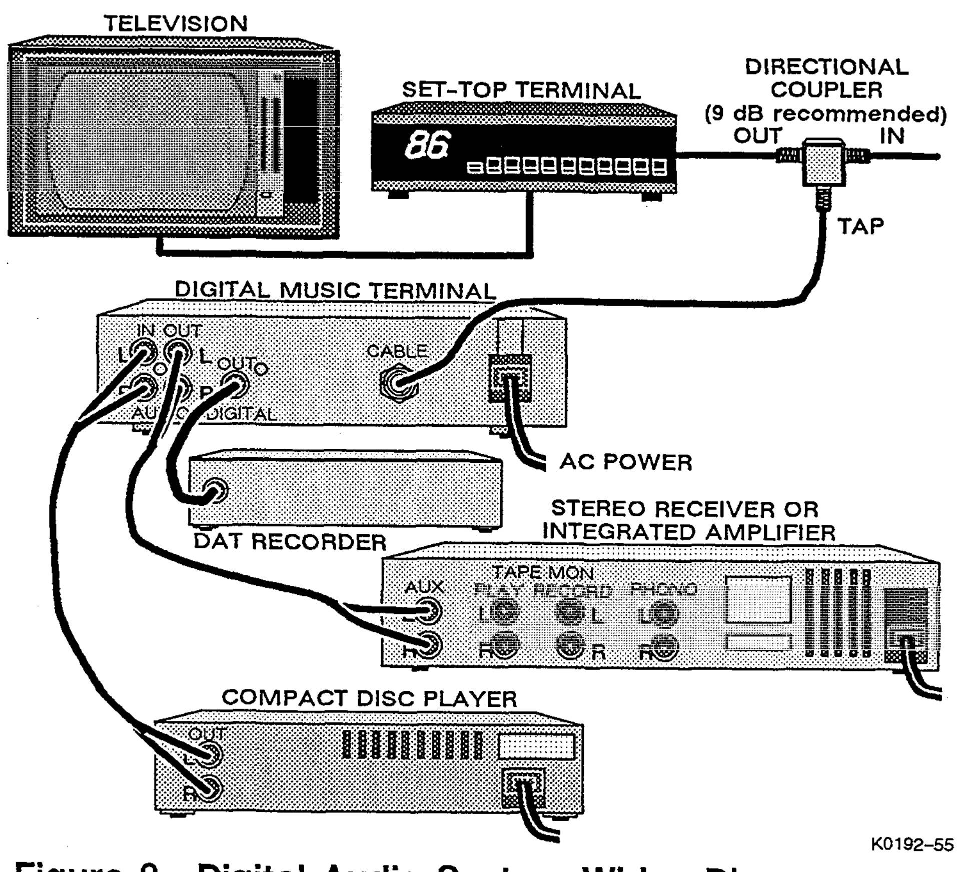


Figure 2. Digital Audio System Wiring Diagram– Subscriber's Home

## CONCLUSION

In conclusion, the digital audio system must be extremely rugged while providing near CD quality audio to subscribers. The system must adapt to existing architectures and be easy to implement. The cable operator demands a system that is bandwidth efficient, reliable and user friendly.

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