THE COMPLETE
TECHNICAL PAPER PROCEEDINGS
FROM:

NCTA
Technical
Papers

A COMPARISON OF PAY-PER-VIEW SOURCES FOR CABLE TV SYSTEMS

Robert H. Shaw
Pioneer Communications of America, Inc.

ABSTRACT

Pay-Per-View programming is becoming an increasingly important source of revenue for the cable operator. To increase the subscriber buy rate, it is necessary for the cable operator to offer a wide variety of programming which will entice the subscriber to buy more frequently. The current use of satellite distribution allows many cable systems to provide quality programming as well as frequent "Blockbuster" events. However, the satellite delivered system does not allow for customization of program offerings. Video tape systems allow for customization, but at the expense of lower equipment reliability and higher maintenance requirements. A Laserdisc based system can offer customized programming with low maintenance, high reliability and a quality signal. It is the intent of this paper to present a comparison between the different technologies available to allow the cable operator to determine which system would work best in his particular application.

OVERVIEW OF PAY-PER-VIEW SOURCE TECHNOLOGIES

Satellite Delivered Programming

Geosynchronus orbiting satellites have been used for many years for the distribution of most CATV programming. This technology is a tried and true means for providing nationwide programming for such services as movie and sports programmers and other "Super-Stations". After initial set-up and alignment, the satellite receive system requires relatively low maintenance and only occasional readjustment of the receive components. Maintenance of other support equipment may require more frequent attention, e.g. compressed air or gas feedline pressure hardware. Signal quality is generally very good although occasional problems may appear, such as rain or snow fade, sun-outages, and satellite drift. Although these problems should not pose a long-term threat for the cable operator, the operator must be prepared to deal with customers that call with related complaints and deal with them in as diplomatic a manner as possible.

Video Tape Programming

Video tape has been used for many years as a source for cable for commercial insertion as an economical means for providing fast, flexible service to the advertisers in their systems. It also was used as a Pay-Per-View video source in the early days of cable before the widespread use of satellite delivered signals. Using video tape players for pay-per-view causes many operational problems for the operator, the main problem being maintenance. Unless the operator can spend many dollars on a CART machine, he is forced to use human labor as the means to keep the VTR's stocked with tapes ready for playing. Tapes must be changed every hour if the machines are to provide an uninterrupted source of video. this requires continuous manning. This can become even more costly over the long-term when salaries for the tape operators is considered.

Preventative maintenance on the equipment is also very high requiring frequent headcleanings, lubrication and mechanical repairing of broken or worn parts. The occasional "devouring" of tapes by the VTR's must also be considered when stocking the movies and ordering tapes. The potential for video noise due to tape drop-outs and/or other problems can cause much customer dissatisfaction with the pay-per-view programs.

Although video tape presents the cable operator with many problems, the one main advantage that is gained is the ability to provide localized programming to the subscribers. This ability to custom tailor the video selection to the market makes video tape a highly desirable means for providing pay-per-view.

Laserdisc Programming

The use of Laserdisc players is fairly new to the cable industry. Laserdisc players can provide the cable operator with the advantages of video tape players without their disadvantages. A clean video signal can be obtained without the degradation caused by repeated playing of the laserdiscs. This is due to the fact that the laserdisc pickup does not contact the disc in any manner, unlike the video tape player head which is constantly spinning in contact with the tape surface. A dirty video tape head can grind the oxide layer off of the video tape causing picture degradation and eventually render the tape useless. Multi-disc players are also available to eliminate the need for the operator to manually remove and insert discs for continuous play.

All of these factors add up to low attending maintenance, consistent program picture and audio quality, and the capability for localized custom programming.

SIGNAL QUALITY COMPARISONS

One of the first considerations almost any operator will make is the overall quality of the signal source available. For a satellite receiver system, the limiting factor of the signal quality is the descrambler unit (assuming a scrambled satellite signal). A Video Cipher II typically has a weighted signa to noise ration of 57dB. This measurement compares to 47dB for a 3/4" (U-matic) VTR,[1] and 57dB for the laserdisc player. When using video tape as the program source, signal degradation caused by the cable plant must be evaluated carefully. Subscriber satisfaction with the Pay-Per-View service will depend on the end point quality of the signal.

Signal to noise comparison:

Satellite Receiver/Decoder	57dB
Video Tape Player	>47dB
Laserdisc Player	57dB

MECHANICAL CONSIDERATIONS

Depending on the particular system being used to provide Pay-Per-View service, the amount of space required for the hardware varies widely. Headend space is usually at a premium and efficient use of the available space must be well planned to accommodate the amount of equipment the operator will be installing for his Pay-Per-View service. Not considered in this comparison is any of the associated headend gear such as modulators and scramblers that would be installed in the headend irregardless of the type of system used.

Satellite System

A satellite system consumes the least amount of space in the headend. Refer to Figure 1. A typical satellite receiver is 3 EIA units high per channel. (An EIA unit is 1.75 inches in height for a normal 19 inch wide rack). Some receiver models, however, can fit two receivers into this same package size. Additional space of at least 1 EIA should be included for proper cooling of the receiver

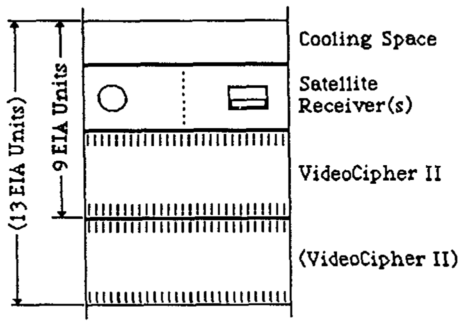


Figure 1. Satellite Hardware
Single Channel / (Dual Channel)

with 2 EIA being the ideal spacing. A VideoCipher decoder is 4 units high and includes its own cooling space with the unit. The design of the decoder allows cool air to circulate over the main circuit board even in systems without forced-air cooling. This totals 9 EIA units per channel (13 EIA units per two channels if a dual receiver configuration is used). If video switching is implemented to switch to an alternate program source in case of receiver failure (fail-safe system), this equipment will require more space. However, the additional space can be considered to be negligible in this discussion since it usually can be used for multiple channels.

Most satellite equipment is bolted directly into the rack where it is mounted. No special hardware is needed to accommodate normal maintenance procedures. Optional rack mount slides can make maintenance easier for the technician, but is not required. Cabling for the satellite receive system is also very straight forward, consisting mainly of video/audio cabling from the receiver and Video Cipher and an RF feed to the receiver. (Figure 2) Occasionally implemented are special cables from the audio channels for cue-tone or special sub-carrier detection by commercial insertion equipment. Powering requirements for the receiver and decoder are modest. The receiver typically draws 100 watts and the Video Cipher 35 watts.

Video Tape System

Modern video tape players do not consume as much space as their predecessors. VTR's now fit into the standard 19 inch equipment rack, some requiring shelves for support, others able to rack mount directly without adaptation. Refer to Figure 3. However, most VTR's are 5 EIA units high and each channel must have a minimum of two VTR's for continuous play. Other associated equipment for the VTR PPV system is a sync generator/timebase corrector, a video/audio switcher, a preview monitor, and a system controller with high resolution video generator. ALL of these components are necessary to enable the VTR system to operate in an encoding environment. (Figure 4) The sync generator provides a "Gen-Lock" signal to the VTR's to keep the video signals from each unit synchronized to one standard. Without this

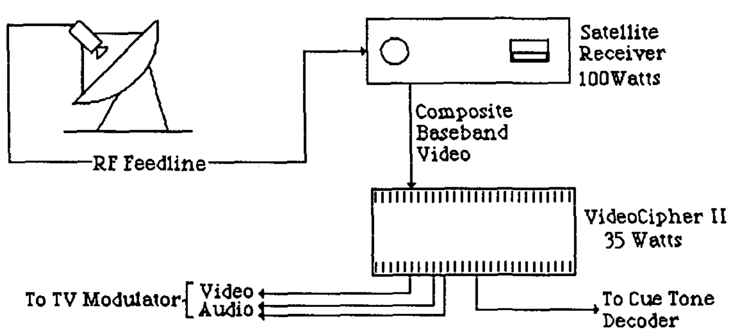


Figure 2. Satellite System Wiring

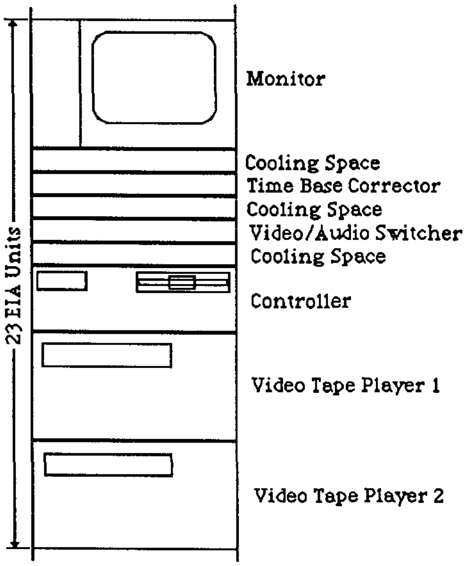


Figure 3. Video Tape Hardware

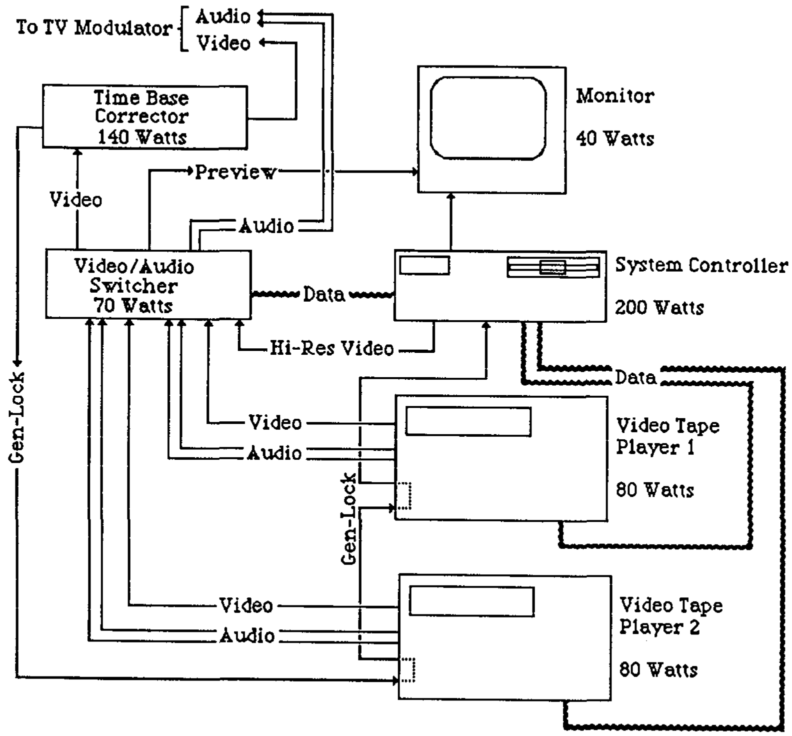


Figure 4. Video Tape System Wiring

the encoder will lose sync with the video signal causing the descramblers to also lose sync, generating an annoying glitch when switching from one VTR to another. A timebase corrector ensures that each video line is of uniform length (in time) and also cleans up any problems encountered with the sync pulses. The video/audio switcher is controlled by the system controller to provide the correct output signal for the channel. The preview monitor, while not necessary, can be used by the operator to verify proper cueing of the next tape before it goes out live on the air. The system controller should incorporate a high resolution video graphics board to enable the system to provide a "Barker" screen to prompt subscribers to purchase, display the program schedule or display a "Please Stand By" screen in case of problems.

The associated hardware can consume up to 14 EIA units of rack space (including cooling space) for a total of 24 EIA units for a system with only two VTR's. Cabling and wiring harnesses for a VTR system can be quite extensive especially if many VTR's are used. Each VTR has a Gen Lock signal (looped through), video and audio output, and control interface cable. All cable outputs must feed the video/audio switcher, which is also connected to the controller by an interface cable. Video from the switcher is fed to the Time-Base Corrector (TBC) where it is buffered and passed on to the modulator.

Powering requirements for a 2 VTR system with the above mentioned hardware is approximately 610 watts and increases by 80 watts for each additional VTR that is added to the system.

Laserdisc System

The Laserdisc system requires some special considerations for location and set-up. Two configurations of the system are possible. The first utilizes two single disc player units. (Figure 5) These units fit in a

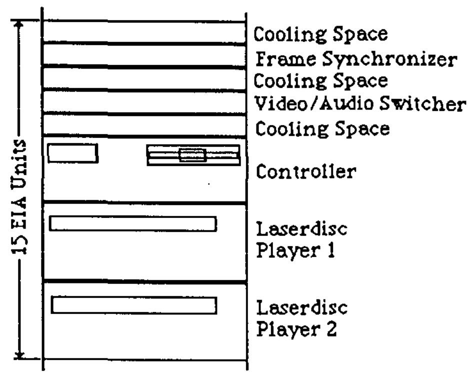


Figure 5. Laserdisc Hardware
Single Disc Player System

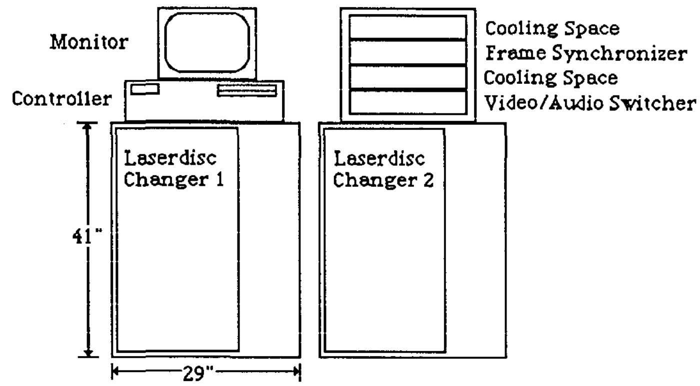


Figure 6. Laserdisc Hardware
Laserdisc Changer Units

standard 19" rack and are 3-1/2 EIA units high. The second system uses two laserdisc changers capable of containing 72 laserdiscs each. (Figure 6) Although the changers cannot be rack mounted in a standard 19" equipment rack, their physical construction allows them to be placed wherever convenient in a headend.

A frame synchronizer is used in a similar fashion as the Time-Base Corrector in the VTR system. It maintains a stable synchronized video signal to feed an encoder and TV modulator when switching between laserdisc players. The frame synchronizer and video switcher take up 1 EIA space each with 1 EIA space needed for cooling for the

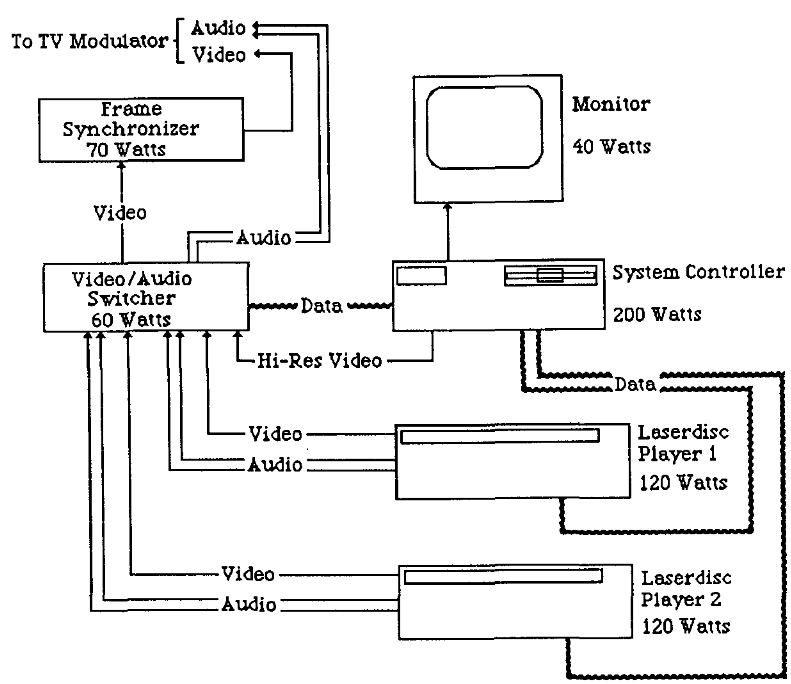


Figure 7. Laserdisc System Wiring

frame synchronizer and video switcher. Cooling space is not needed for the laserdisc changers. A standard amount of cooling for the PC controller is needed although the controller can be placed on one of the changers. It is also possible that the video switcher and frame synchronizer could be placed in a mini-rack and set on top of the other changer eliminating the need for using valuable rack space. (Figure 7) This hardware placement also has the side benefit of eliminating the need for long cables to and from the main rack except for the audio/video cables running from the laserdisc system to the TV modulator.

Power requirements for the Laserdisc system totals approximately 610 watts.

EQUIPMENT MAINTENANCE

Satellite Receiving Maintenance

Even though the satellite receiving system has no real moving parts to wear out, a considerable amount of maintenance may be required if support equipment is used. Most of the support equipment is used to pressurize the RF feedlines from the LNA's to the receivers. The air compressors must be periodically checked for proper lubrication, the air-filters/cleaners must be cleaned to keep contamination from entering the feedlines. This and other required maintenance is common to all channels received by satellite.

Video Tape System Maintenance

Video tape systems for pay-per-view require the most maintenance of all the systems discussed. Even if only one movie is shown per day at regular intervals each day, the tape player will require very frequent cleaning. Regular preventative maintenance should be performed at the intervals recommended by the units manufacturer. This includes extensive cleaning of the tape path, lubrication of moving parts and replacement of worn parts.

A video tape should have a fairly long service life although it won't be useable forever. Each tape should be visually checked for cinching or other problems each time it is inserted or removed from the player. This can help to determine if any potential problems are lurking in the player's mechanism. Eventually a tape player will eat a tape, and a backup contingency will be needed to replace the damaged tape and/or player. A strict schedule of cleaning and preventative maintenance, however, should reduce the probability of this occuring.

A clean environment must be maintained for the tapes also. Any dust or other foreign material (fingerprints) on the tape can damage the tape, be transferred to the VTR and then be transferred to other tapes. Magnetic tape can withstand temperature and humidity extremes for storage, but should be maintained in as stable an environment as practical for playback.

Laserdisc Maintenance

Laserdisc players, by their inherent design, are low maintenance devices. While they incorporate moving parts, the durability of these parts is high and the mechanical designs are sound. This translates into a unit which requires virtually no maintenance.

Laserdiscs have proven to be a sturdy medium. Although "normal" handling of laserdiscs does not affect picture quality or player performance, the utmost of care should be taken to maintain the cleanliness of the discs. The use of the Laserdisc changer helps to keep discs in "like-new" condition by minimizing the amount of handling required to insert and remove the discs. Discs should be inspected before insertion into and after removal from the changer for any signs of damage. A damaged disc inserted into the changer may cause the changer to fail, and likewise, a damaged disc removed from the changer indicates a mechanical failure of the changer. Putting a dirty disc into the changer can contaminate other discs, and the player mechanism itself and cause intermittent problems or even failure of the player. Clean-room practices are not necessary, but cleanliness will extend the life of both discs and player.

HEADEND INTEGRATION

Recent advancements in technology are making cable systems with double the bandwidth of just five years ago possible. The uncertainties of the needs of HDTV will allow operators to fill some of these channels with Pay-Per-View signals. An operator now does not need to commit himself to only one media for his pay-per-view source.

Different channels may be assigned to the different types of pay-per-view available. Nationally distributed programming and special events from satellite can utilize one (or more) channels. Other movies or local events and promotions can be played back from tape and a locally programmed IPPV service can be provided by laserdisc. Even multiple PPV systems can feed a single channel with a mix of different types of programs depending on time of day, events available and market demands.

Recent rulings regarding Syndicated Exclusivity (Syndex) require the operator to 'blackout" certain programming from distant stations when duplicated on local stations. This valuable "dead" air time can be replaced by a locally originated program or other type of program provided by a VTR or laserdisc based system.

SUMMARY

No one system is the answer to all of a cable operators Pay-Per-View needs. Each system discussed has strong advantages and disadvantages. The operator must weigh the market needs with system cost, availability of programming, maintenance requirements and user friendliness. Video quality, headend space consumption and powering concerns should also be examined before committing to a particular system. It is hoped that this paper has helped the cable operator and the cable system engineer to understand the issues related to different Pay-Per-View sources.

ACKNOWLEDGEMENTS

The author would like to thank Mr. John Unverzagt and Mr. Richard Annibaldi, of Pioneer Communications of America, Inc., for their assistance and encouragement which helped to make this paper possible.

REFERENCES

1. National Association of Broadcasters Engineering Handbook. 7th Edition. Section 5.9 3/4" and 1/2" Video Recording Systems. S. Baron, E. Pohl, P. Smith.

A Digital Solution
for
Cable Television Systems

David Jordan * *
Product Manager
C-COR Electronics

Abstract

Digital fiber optic systems have become a strong alternative to analog technology for high performance delivery of CATV signals throughout a cable plant. This paper reviews the performance and characteristics of a currently available digital system that does not use bandwidth compression.

Digital transmission has become the foundation of modern communications networks, and digital signal processing is now the dominant technology in television studio equipment. However, the strong advantages of digital transmission have been offset by a perception that the equipment occupies extensive rack space and expensive in typical multi-channel CATV applications.

Today there are digital systems that are competitive in price, size, and features with current FM analog fiber optic systems, and they are designed around an architecture for a long term "systems approach" to CATV signal distribution. Rather than being limited to an individual, point-to-point optical link that replaces a portion of a coax plant, today's digital system will evolve, in a compatible way, to extend fiber optics deeper and deeper into the plant without fear of signal degradation or equipment obsolescence.

Advances in high speed digital processing will offer the possibility of digital transmission to practically interface with other signal formats. Advances in high speed digital optical components will allow significant increases in signal carrying capacity, without disrupting installed digital plants.

Research indicates the consensus among cable operators favors a long-term evolution toward a digital based architecture. Adopting a digital strategy today will further strengthen the future competitive position of cable operators who look beyond interim fiber optic solutions.

In this presentation I will briefly examine the basic building blocks of a typical digital transmission system and then present a case study. I will describe the Mountain View Cable System and outline the digital fiber optic equipment that could be installed in this system. Finally, I will summarize the presentation with a review of the advantages of digital fiber optics in a CATV application.

I. Introduction of Digital Fiber Optic Systems

Digital transmission of video, with audio subcarrier through optical fiber is achieved by converting the analog video and audio inputs to digital data. The digital data is then "line encoded" (such as scrambled NRZ) for ease of time synchronizing, error monitoring and bandwidth economy. The line-encoded signal then modulates a light source.

Converting analog video and audio to digital is done by precision A/D circuits that are also used in broadcast studio equipment. As in all digital systems, sample rate and accuracy affect the end-to-end signal performance of the system.

High sampling rates with sampling precision of 7 bits or more for video and 12 bits or more for audio ensures high quality signal performance. The resulting data rates of these systems exceed 70 megabits/second for video and 700 kilobits/second for audio, which gives rise to two types of digital systems:

1. Full bandwidth or linear pulse code modulation (PCM) systems that transmit all of the digitized signal information.

2. Bandwidth compressed systems that transmit only a portion of the original digitized signal according to a predetermined processing algorithm.

Uncompressed, linear PCM systems are directed toward high signal quality applications in dedicated links or private networks. PCM coding is ideal for transmission systems where bandwidth is virtually unlimited as in fiber optic systems. Compressed systems are most useful in digital telecommunication networks where standard, fixed bandwidth channels are available (as with twisted pair telephone wiring.)

Digital transmission systems do not require a linear light source, are highly noise immune, and operate with both multi-mode and single mode optical fiber. Thus, light sources for digital systems can have a wide range of non-critical operating parameters.

Because digital PCM is very noise immune, system performance does not degrade with increased transmission distances, and very long ranges are possible without repeaters. If required, digital systems can be regenerated many times without signal degradation.

Digital systems are very tolerant of losses and reflections from connectors, splices and optical devices such as splitters and optical multiplexers (WDM). This provides flexibility in designing point-to-multipoint optical systems and networks.

II. Digital Building Blocks

All commercially available digital fiber optic systems have many similarities in terms of the building blocks (or the boxes) that are used to create a video delivery system. A list of those building blocks is outlined below.

Digital Fiber Optic System Building Blocks

Optical Transmitter / Receiver terminals
Channel multiplexers
Video encoders and decoders
Subchannel multiplexers
Audio encoders and decoders
Data interfaces

A block diagram of a typical digital system configuration is pictured in Figure 1.

I will briefly describe the function of the major blocks in the the system diagram.

The **Encoder** accepts baseband video inputs. Sampling at greater than twice the highest frequency, the encoder digitizes the baseband signal into a linear, full bandwidth data stream. An audio encoder performs the same function by digitizing a baseband audio signal.

The **Time Division Multiplexer** accepts digitized audio, video, and data and converts it to a serial data stream. In addition, a clocking reference signal is generated by the TDM which ensures proper recovery of the digitized signal at the receiver.

The **Optical Transmitter** accepts a serial data stream and converts an electrical signal into an optical signal. Within the transmitter there are typically several alarm outputs which give the operator the ability to monitor the status of the transmitter. Transmitter alarms generally include power supply status, temperature, and laser drive current.

The **Optical Receiver** detects an optical signal and converts it back to an electrical signal. Much like the transmitter, there are several alarms present in the receiver which give the operator the ability to monitor receiver status. These alarms typically include, temperature, power supply status, clock, and optical signal.

The **Time Division Demultiplexer** separates video and audio from data and converts serial information into parallel information. The demultiplexer also sends the audio and video to the decoders and separates the clock from the data signal.

The **Decoder** accepts a digitized signal, converts it back to an analog format and recovers the original baseband signal.

The diagram in Figure 2 shows how a simple digital transmission system might appear. I have chosen to illustrate a system that operates at approximately 780 Mb/s and uses 8-bit coding. This particular system uses a multiplexing structure which gives the user the ability to combine 2 blocks of 4 channels in order to optically transmit 8 channels per fiber. This particular system accepts 8 baseband video inputs and delivers 8 baseband video outputs on one fiber. The typical link loss budget of this system is 29 dB.

As you can see, each of the building blocks is relatively basic in its design and function. Each of the "boxes" in a video transmission link can be built using "off-the-shelf" devices. However, one of the most significant advantages of digital transmission systems when compared against analog alternatives is that digital system performance is not affected by the linearity of the light source. Lasers in digital applications must simply have the ability to turn on and off, and therefore attempting to optimize the linear characteristics of the light is of little concern.

III. System Application

Much has been written about digital fiber optic technology and its usefulness in a CATV system. Most authors accept that digital is indeed the long term solution for cable operators, yet in the same breath they point out that current costs and physical size of the equipment suggest that digital technology is several years away from becoming a viable alternative. The conclusion drawn from these articles is that operators should delay investing in digital transmission systems to allow the technology to "catch up" with user requirements.

This assumption must be challenged. First, the costs for digital fiber optic systems are declining and technology is advancing so quickly that assumptions made even within the last 18 months can no longer be held true. Second, fiber optic transmission equipment available today which occupies no more physical space than comparable analog transmission equipment.

In addition to a digital system's significant advantage that was mentioned above (a much less restrictive operating requirement for the laser diodes) there are several other advantages inherent in digital technology that operators should be aware of when evaluating fiber optic transmission systems.

The high optical budgets available in digital systems, when combined with the fact that it is fundamentally easier to restore and regenerate a digital "word" as opposed to an analog signal, gives many more options to cable operators. For example, in some long distance applications where high quality signal transmission is required, digital has proven to be the most viable and cost effective solution.

Digital fiber optics requires operators to look beyond the "link" or "box" solution offered by analog fiber optic products and examine the use of fiber from a "system" perspective. Figure 3 illustrates one of the system advantages for considering digital technology. The concept introduced here is called digital "fan out" (or drop and repeat). With this concept it is possible for operators to generate two identical blocks of signals with the same transmit and encoding equipment, thus minimizing per channel cost for transmission to multiple hub sites. At the location of the second transmitter the link loss budget would be restored to its original performance specification (in the previous example that is 29 dB).

Many MSO's have concluded that digital technology must be given serious consideration in a cable operation when there is a requirement to interconnect headends or a need to connect multiple receive sites to a single headend.

I would like to outline in detail by describing an actual case where digital fiber optics is being proposed as an effective fiber optic solution. I will examine an actual cable television system and describe a quotation that was submitted to this system. At the request of the customer, I have changed the name of the system.

Mountain View Cable is a typical urban cable television system. They presently serve over 80,000 subscribers in a rather large geographical area. Franchising requirements are forcing Mountain View to rebuild its present system from 330 MHz to 550 MHz (77 channels). The digital system in this study expands in four channel blocks, so the quoted system will be capable of 80 channels. Mountain View Cable has two headends located 52 km apart. They are interested in connecting the East and West headend, dropping signals at the East headend for distribution, and distributing signals from the East headend to four receive sites. The distance from the East headend to the four receive sites are as follows:

Hubsite #1	13.9 km
Hubsite #2	28.0 km
Hubsite #3	18.5 km
Hubsite #4	18.0 km

Mountain View Cable has asked us to determine the cost per received channel, the loss budget margins present at each receive site, and system link performance.

A block diagram of the interconnect of the West and East Headend Site is shown in Figure 4.

The system described is operating at a rate of 780 Mb/s, which is capable of transmitting 8 channels of video per fiber. This particular diagram would be multiplied 10 times to arrive at the system requirement of 80 channels. The laser diodes transmit at -3 dBm and the APD receivers are -32 dBm devices resulting in a loss budget for this link is 29 dB. As an option, to achieve a greater system margin, a transmitter operating at 0 dBm output in conjunction with a receiver with -35 dBm sensitivity yields an optical loss budget of 35 dB.

Using a loss figure for fiber of .5 dB/km, the loss budget (safety margin) is calculated for each receive site. As stated previously, when you use digital technology to fan the optical receiver outputs to secondary decoders and transmitters, a second link can be established with the full 29 dB optical loss budget. The following chart specifies the loss budget remaining at each receive site even after transporting the signal through a series of splitters and connectors. As you can see a minimum of 6.2 dB margin remains in each link, including the East headend site which is 52 km from the West headend.

A block diagram of the regenerator and the distribution section of the proposal is shown in Figure 5.

As you can see we have recommended that Mountain View install both symmetrical and asymmetrical splitters. Obviously the losses are different for each leg of the asymmetrical splitters. Our attempt in proposing this configuration is to optimize the lower splitter losses over the longer fiber links. Once again this diagram portrays the equipment required to receive 8 channels at each of the four receive sites. To arrive at at full complement of 80 channels you must multiply this equipment by a factor of ten.

There are three important issues that are raised in this case study:

Loss budgets-even over 52 km (the distance separating the two head ends, a digital system will transport signals with over 6 dB of system safety margin.

Regenerators-using digital technology an operator is not required to "bring the signal back to baseband" in order to repeat the signal.

Multiple Receive Sites-because of the large loss budgets present in a digital system it is possible to regenerate a digital signal and restore it to an exact duplicate of the original signal and serve many receive sites with a single set of encoders by using optical splitting and digital fanout.

Without going into great detail I would like to discuss the cost analysis for the Mountain View Cable system which is outlined below.

Mountain View Cable Cost Analysis

1.	**West HE - East HE Interconnect**	**$300,000**
2.	**East HE Regenerator**	**80,000**
3.	**East HE - Hubsite # 1**	**131,000**
4.	**East HE - Hubsite # 2**	**126,000**
5.	**East HE - Hubsite # 3**	**126,000**
6.	**East HE - Hubsite # 4**	**126,000**
	Total	**$889,000**

Number 1 - East and West Interconnect The costs included the transmitters and encoders at the West headend site and the receivers and decoders at the East headend site.

Number 2 - East headend Regenerator The costs here include the transmitters, fan outs, and splitters required to send the digital signal to the receive sites.

Numbers 3 - 6 Hubsite Receivers all include the same equipment with the exception of Hubsite 1 which has additional splitters located there so that signals can be fed to Hubsite 2. With that exception each of the equipment is identical at each receive site. The costs for the receive sites include the receivers and the decoders necessary to detect the optical signal and convert it to a baseband signal.

Mountain View Cable System requested a proposal for a baseband transmission link. Therefore, the costs analysis includes only the equipment required to encode, transmit, receive, and decode a baseband video signal with BTSC stereo on a 4.5 MHz subcarrier.

The following list summarizes the proposal for the Mountain View Cable system.

Successfully connected two head ends over 50 km apart, and distributed signals to four receive sites (400 received channels-80 were dropped at the East headend/ regenerator.)

A minimum remaining loss budget of 6.2 dB was present at all the sites.

Mountain View Cable Loss Budget Analysis

Remaining loss budget at each site

West Headend - East Headend Site	**52.3 km**	**6.2 dB**
East Headend - Hubsite # 1	**13.9 km**	**8.3 dB**
East Headend - Hubsite # 2	**28 km**	**9.2 dB**
East Headend - Hubsite # 3	**18.5 km**	**6.9 dB**
East Headend - Hubsite # 4	**18 km**	**7.1 dB**

Performance of RS 250 B Medium Haul Specification was present at all receive sites. A minimum of 55 dB S/N was measured at the output of the decoders. Average S/N when measured over all receive sites was approximately 60 dB.

Physical rack space requirements for the digital transmission equipment was universal at 52.5 inches. The exception in the East headend site regenerator which required a total of 70 inches of rack space. The increased space requirement at the East headend is due to the additional optical transmitters and the digital fan outs.

The total system cost is calculated to be $889,000 which works out to be slightly over $2,000 per received channel.

IV. Summary / Conclusion

When considering fiber optics for cable television systems remember that digital technology's important characteristics make it the dominant choice in certain applications.

Those advantages are:

No adjustments.
Most digital fiber optic systems employ "set up and leave" equipment that is manufactured with off-the-shelf devices.

Drop and Insert.
Signals can be added or removed without degrading signal quality.

Repeaters and Range.
Unrepeated ranges of over 50 km are possible. Even when using regenerators, the repeat process is transparent to the transmission system.

Multiple channels and distribution networks.
Multiple video, audio and data channels are able to be transmitted on one fiber. In addition, the large loss budgets present in all digital systems allow the user to split the optical signal many ways.

These features combined provide the following advantages for cable operators:

- Simple installation and operation.
- Flexible system architectures.
- Better quality long distance transmission.
- Adaptable to future technology advances.

Digital technology encourages new ideas and discussions about its impact in the cable television business. There is a renewed emphasis on delivering a quality transmission to the customer. Digital fiber optic systems provide cable operators the vehicle to transmit very high quality signals very deeply into a cable television plant.

The cable operator who today decides to install digital fiber aggressively moves his system into the future. Cable operators installing digital fiber optics today position themselves to take full advantage of technology advancements and future cost reductions.

* * Special thanks to Ken Regnier from COMLUX for his invaluable assistance is in the creation of this presentation.

Digital Fiber Optic System

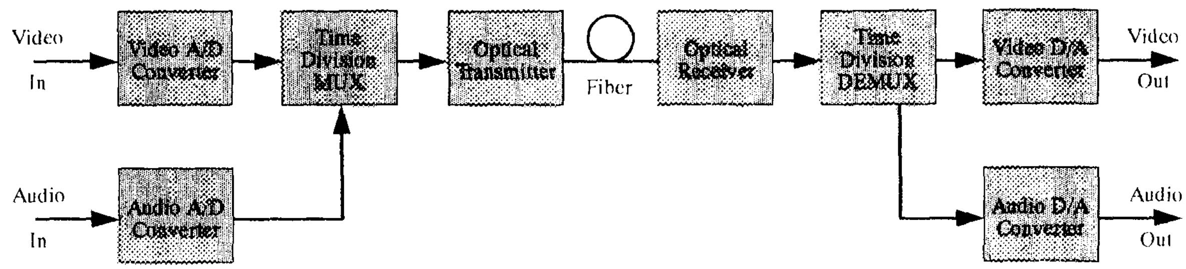
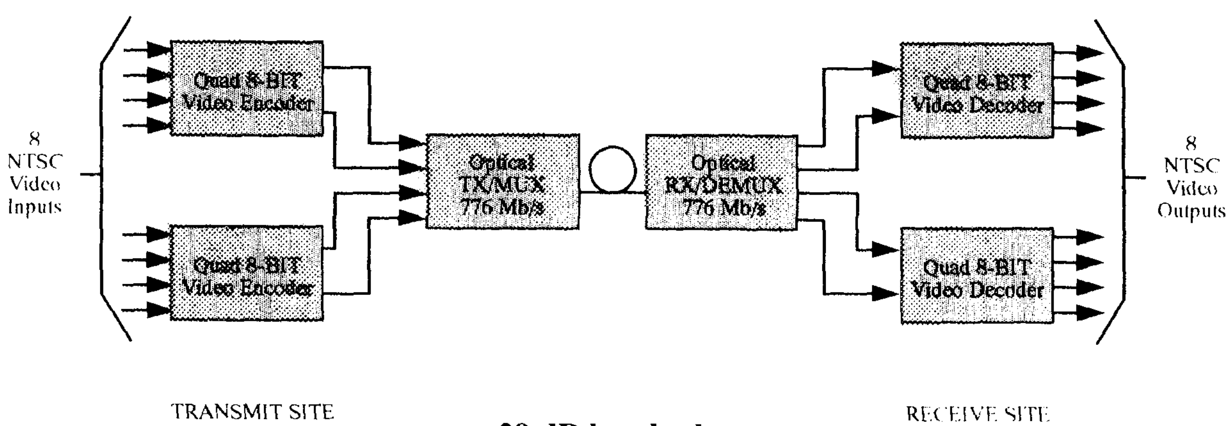


Figure 1

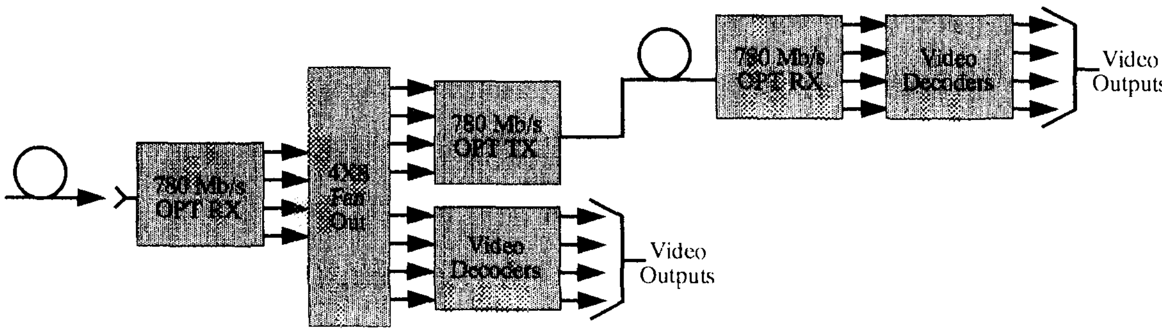
Digital Building Blocks
780 Mb/s TX/RX
8-bit encoding



29 dB loss budget
Typical range up to 58 KM

Figure 2

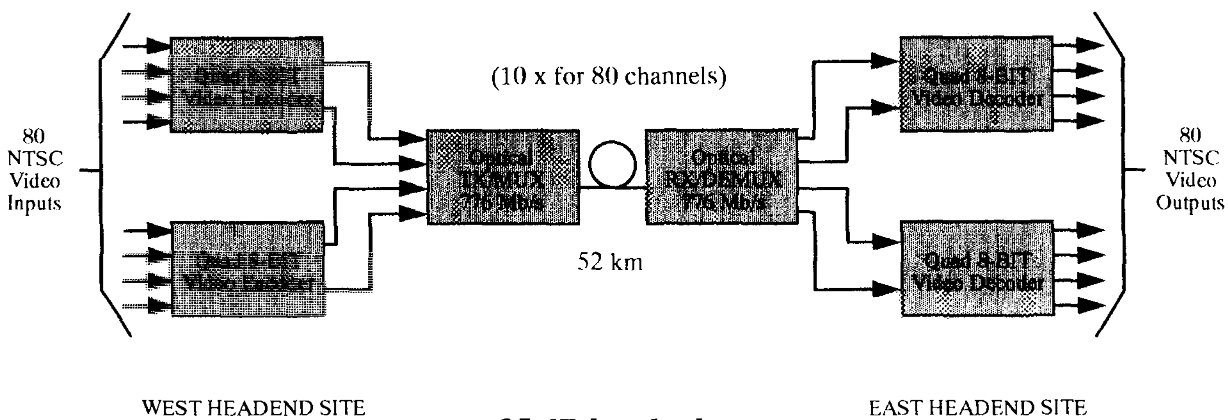
Digital Fiber Optic System



Receive Site with Drop and Repeat
Using 4X8 Fan Out

Figure 3

Digital Fiber Optic System
Mountain View Cable



35 dB loss budget
Typical range up to 70 km

Figure 4

Digital Fiber Optic System
Mountain View Cable

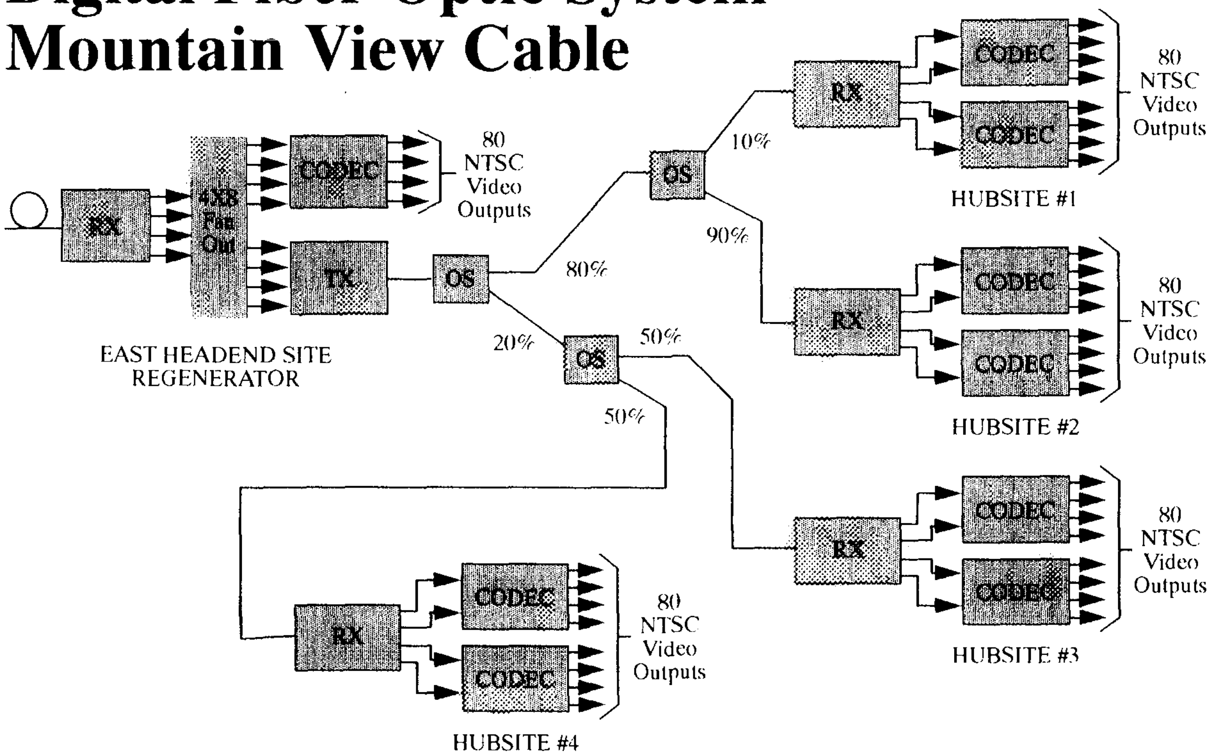


Figure 5

A HYBRID APPROACH TO A UNIVERSAL IR TRANSMITTER

Mack S. Daily, Richard G. Merrell
Zenith Electronics Corporation

ABSTRACT

A multibrand IR transmitter is generally one of two types. One type has a stored library of codes. To program a transmitter, the user selects code sets from the library. The library unit lacks versatility, reproducing only the code sets included in the library. A learning unit can be taught any code set by exposing it, one key at a time, to IR streams from a host transmitter. It can replicate any transmitter, but the learning process is cumbersome and time-consuming.

A hybrid of the two can have both the programming ease of the library unit and the versatility of the learning unit. One such hybrid is disclosed herein. Along with the method for combining the two techniques, a discussion of additional desirable features is also included.

HISTORIC PERSPECTIVE

Library Units

Of the numerous remote control transmitters capable of controlling devices from different manufacturers, several use the library technique. The code formats for a number of different devices are stored in a library and some means is used to specify which code is the appropriate one for the TV receiver, which for the VCR, and so on. In an earlier paper (Ref. 1), some of the different ways to specify the code were examined. The convenience of programming a library unit is further enhanced in that the operation does not require possession of the transmitter that originally was supplied with a device.

Although the library of codes in this type of transmitter may be extensive, it is still finite. A particular code set may be missing from the library because it was unknown during the development, because it was introduced at a later date, or because it represented a small market share. A device requiring that code set, then, cannot be operated by that library unit. Further, the library units typically contain the codes for TVs, VCRs, and cable decoders, but the codes for video disc players, audio equipment, satellite receivers, teletext decoders, and such, are intentionally omitted. Clearly, a library unit is not universal; this is its main shortcoming.

Learning Units

Learning transmitters sample IR emissions from the host transmitter that originally came with a device, pair each IR command with a corresponding key and, generally, a particular mode, and store the information in memory. Subsequently, the device can reproduce any command it has learned. The only limitation is the resolution needed, but this has not been a problem with existing units. Thus, this type of transmitter is universal and not subject to the restrictions noted for the library units. Its principal fault is the learning process itself. These units learn the code sets one key at a time, a distinct annoyance if one is training the device to replicate several dozen commands. The annoyance of the training process is exacerbated if it must be repeated when a battery is changed, particularly if one of the host transmitters has been misplaced.

A HYBRID APPROACH

Overview

When the two methods are combined to eliminate the shortcomings of each, a great improvement results. The hybrid of the two has an extensive library of codes and a convenient selection means so that most programming can be done conveniently and quickly. To accommodate new or altered codes, more devices to be controlled, or even key rearrangement, the transmitter has the capability of learning one or more codes from a host transmitter. These codes can be activated by auxiliary keys or, where appropriate, they can replace functions that were originally programmed from a library.

Selector Switch

The unit has a three-position mode selector switch with TV, VCR, and CATV positions (Fig. 1). In operation, the function transmitted when a key is pressed is determined by the switch setting. As with the transmitter of the previously referenced paper, features are incorporated to minimize the need to move the selector switch. In normal usage:

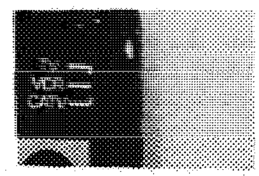


Figure 1. Selector Switch

1. Three separate power keys control the corresponding device regardless of the selector switch setting.

2. For all switch positions, the Volume Up, Volume Down, and Mute keys control the cable box or the TV set as determined during the programming operation.

3. The keys uniquely assigned to the VCR, e.g. Fast Forward, operate the VCR with any setting of the selector switch.

The selector switch labels are for programming convenience only; they do not define restrictions on the unit. Thus, for example, a user might program VCR channel select functions to be active when the selector switch is in the CATV position. This could be advantageous in a lashup where the decoder decodes one channel while the VCR records another. Although there are obvious practical considerations, the user has complete freedom to program any command for any device to be responsive to any key at any selector switch setting.

Keyboard and Displays

The keyboard (Fig. 2) has 42 keys. The Erase and Learn keys are used only for programming and do not produce IR emission when pushed. Because these keys must be activated in any process involving programming the unit, raised bumps below the keys are provided to prevent the user from accidentally depressing either.

The upper 28 keys are labelled for the functions most commonly needed for operating TV, VCRs, and cable boxes. These keys will most commonly be programmed by code selection from the library.

Three rows of unlabelled keys can control up to 36 functions by use of the selector switch. Some of these keys may be programmed from a library, but in many cases, the user will use the learning mode to program the keys to add controls for units other than TVs, VCRs, and cable boxes.

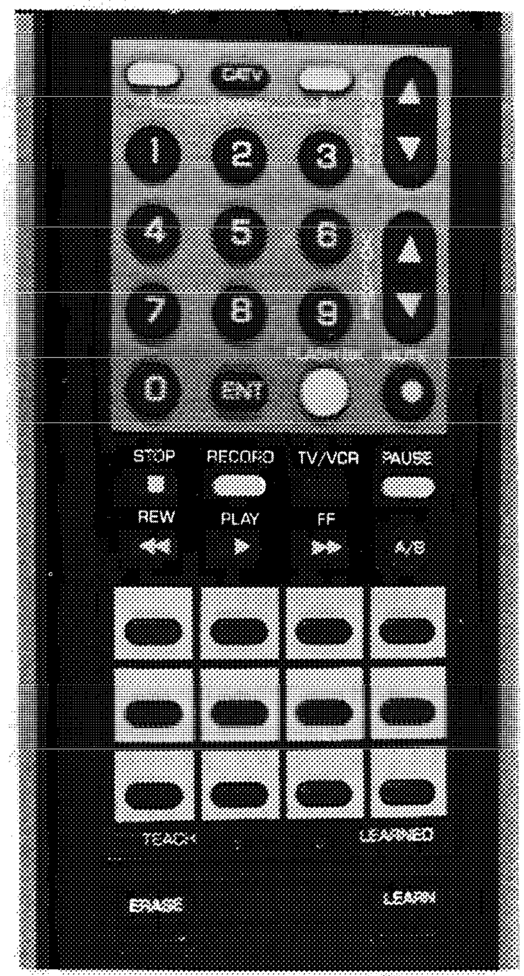


Figure 2. Keyboard

Two LEDs, designated "Teach" and "Learned," serve to prompt the user during a programming operation. Lighting separately or in combination, they signal that the unit is in the programming mode, that an error has been made, that the unit is ready to learn a function, or that a function has been learned.

A writable overlay (Fig. 3) is furnished to allow the user to label the 12 keys.

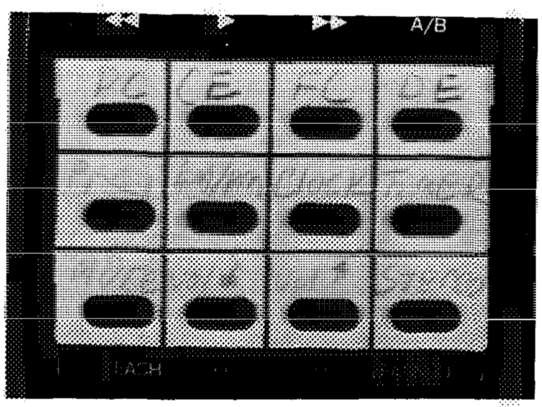


Figure 3. Overlay

PROGRAMMING THE UNIT

Mode Establishment

From the normal operational mode, the unit is switched to the library programming mode by pressing the Learn key for about three seconds. The LEDs flash alternately to indicate the mode.

While in the library programming mode, keying 9-9-9-Enter selects the learning mode. The Teach LED lights to signal entry into the mode and to prompt the user to proceed to the next operation. Entry into the learning mode from the library programming mode is a reminder to the user that programming from the library is always done before head-to-head programming, for a reason to be explained later.

To exit either programming mode, press Learn until both LEDs are off. The device automatically reverts from the library programming mode to the normal usage mode if there is no programming activity during any 50 second time span.

Library Programming

While in the library programming mode, the device can be block-programmed from a library that includes 41 different TV code sets, 39 for VCRs, and 50 for cable boxes. This is easily done by keying in a three-digit number and Enter. The number starts with 1 for TVs, 2 for VCRs, and 3 for cable boxes. Consequently, the position of the selector switch during library programming is immaterial. The unit responds to a library entry by lighting the Learned LED.

If a library entry is made, it overwrites whatever was previously in the memory for that type of unit. If a TV code, for example, is entered, it writes or overwrites the part of memory controlled by the TV position of the selector switch, regardless of the previous contents of the memory. It is for this reason that key-by-key programming in the learning mode must occur after all library programming has been done.

Although the labelled keys define the most common functions for different units, some labelled functions are not needed for certain devices while other devices, particularly cable boxes, require functions not included in the labelled keys. In the former case, library programming blanks the memory for any unneeded function and there is no IR emission if the user pushes that key. In the latter case, all known commands are programmed by a library entry using the unlabelled keys, if necessary. The instruction manual defines any programming of unlabelled keys corresponding to each code set from the library. The user transcribes this to the writable overlay.

Learning Programming

To teach the unit a command from another transmitter, the user selects the learning mode as previously explained. The selector switch is set to the desired position and the learning transmitter and host transmitter are butted together. The user responds to the Teach LED by pressing the key on the host transmitter for the function to be learned and the key on the learning unit that is to control that particular function. Under normal conditions, only one attempt will be needed for each key and the Learned LED will light to indicate acceptance of the function. If the unit requires a second pass, it lights the Teach LED rather than the Learned LED. In either case, after the function has been learned and the user so informed, the Teach LED lights to prompt that the device is ready to learn the next function.

Multiple-Level Programming

For a given key, there are three possible commands that can be transmitted, corresponding to the three positions of the selector switch. When a command is stored during either type of programming, it is always assigned to a selector switch setting. As noted, this is determined by the prefix for library programming or by the switch setting during learning mode programming. Under certain conditions, a command may be assigned to one or both of the other two switch positions.

1. Programming a unit from the library assigns functions of certain keys for all selector switch settings. These include eight keys, including Power, dedicated to VCRs and the Power keys for TV and CATV. This is a convenient method for making these keys independent of the selector switch.

2. Programming a TV or cable box from the library establishes the volume control functions (Up, Down, Mute) in all three positions regardless of previous contents. This allows those keys to operate independ-

ently of selector switch position. It also permits the choice of having volume control functions operate the TV or cable box; whichever of the two was programmed last will respond to the volume control keys.

3. During learning mode programming, a function is assigned to the selector switch position and to either or both of the other two positions for which there had been no previous assignment. This is a convenient technique for making the unlabelled keys responsive independent of the selector switch position.

Master Clear

Under certain circumstances, it may be expedient to de-program the entire memory. This is done by holding down the Erase and Learn keys for several seconds. The Learned LED lights to signal completion.

Individual Clear

The memory for a particular key can be cleared by pressing that key and Clear for three seconds. The Learned LED indicates completion of the activity.

Number of Functions

To minimize the amount of memory needed for storing the functions, the library of codes is in lookup tables in program memory of the microcomputer. The designation of which codes have been selected from the library takes up a small amount of space in external non-volatile memory. The remainder of that memory is used for storing the learned functions. Although data compression is used, there is still a limit to the number of functions that can be stored in the available memory. The number is related to the complexity of the individual commands being stored. It is estimated that the unit will accommodate about 80 average commands in addition to those programmed from the library. The absolute maximum number of commands is 120, derived from 40 keys and a three-position selector switch.

Should the device use up all available memory during programming, both LEDs light to signal an error. At this point, the user must pick and choose the functions which are of most value and can use individual key clearance to reprogram the transmitter.

PROGRAMMED OPTIONS

The transmitter has several programmable options:

1. When the transmitter turns on a cable box, it also sends commands to tune the TV to a programmed channel, generally Ch. 3. The user can program any channel up to Ch. 9 or can disable this feature.

2. When the VCR is commanded to play, the TV is tuned to a programmed channel. This too is user programmable and can be any single digit channel or can be disabled.

3. When the cable box is turned on, it is automatically tuned to a programmed channel. Programming of any turn-on channel up to 99 and installation of the option is done in the factory and cannot be changed by the user.

ELECTRICAL

A combined circuit/block diagram for the transmitter is shown in Fig. 4. The 40 keys that produce IR emissions comprise the keyboard block. The Learn and Erase keys and various other programming options are part of the switches block. The heart of the system is an 8-bit microcomputer. It drives the Teach and Learned LEDs and the IR LED through a buffer. An IR receiver passes along the IR signals from host transmitters.

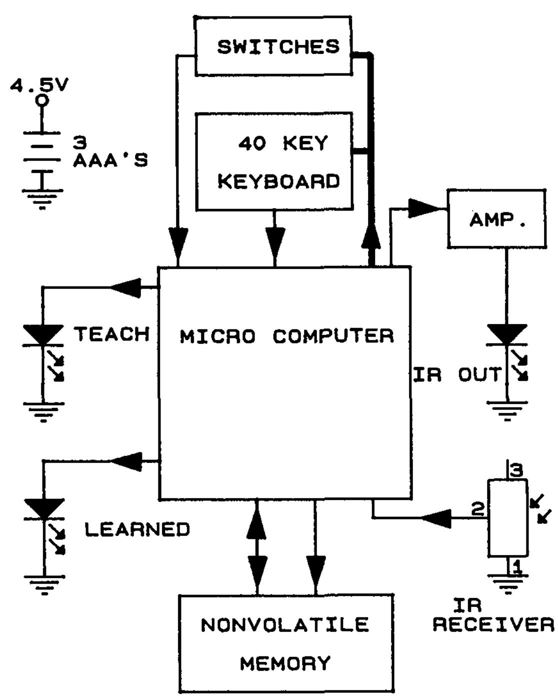


Figure 4. Circuit

The designation of which codes have been selected and the definitions of the functions which have been learned are stored in external nonvolatile memory. The added premium for the non-volatile memory is a worthwhile tradeoff to guarantee the user will not have to re-program the transmitter following a battery change.

MECHANICAL

A survey of various other units on the market makes it clear that manageable physical size is of considerable importance. By eliminating the library programming switches and compressing the keyboard, this transmitter has been made smaller than the device disclosed in the previously-referenced paper. Fig. 5 is a graphic illustration of the size of the transmitter. In view of the features incorporated into the device, particularly the large number of keys, the size of the unit is quite reasonable.

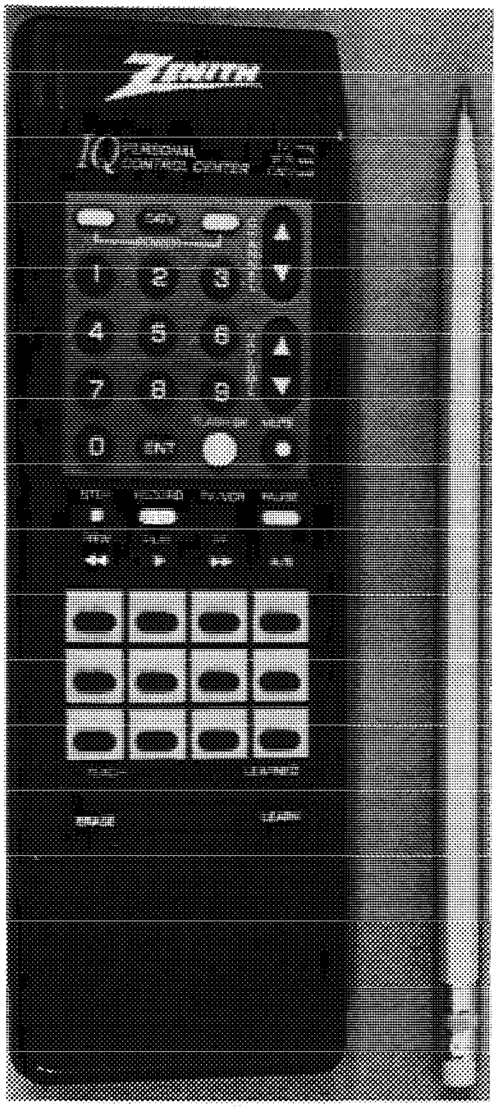


Figure 5. The Transmitter

SUMMARY

As described herein, a very user friendly universal IR transmitter can include the following characteristics:

1. Easy block programming from a library.
2. Convenience of learning from another transmitter.
3. Easy programming to make some of the keys independent of the selector switch.
4. Complete user control of key allocation.
5. Convenient designation of user-programmed keys.
6. Easy to understand prompting.
7. Sufficient memory for a great number of learned functions.
8. No loss of memory with battery change.
9. Programmable macro-instructions for tuning options.
10. Reasonable size.

References:

1987 NCTA TECHNICAL PAPERS, "Multi-Control Remote Transmitters," Richard G. Merrell, p. 213-216

A MULTICHANNEL, BI-DIRECTIONAL SINGLE WAVELENGTH, OPTICAL VIDEO LINK OVER A SINGLE FIBRE

Shane J. Shklov
Manitoba Telephone System

ABSTRACT

Over the past year the Manitoba Telephone System, working with both Catel in California and Cabletel in Toronto has developed a unique optical architecture. In comparison to conventional fibre architectures, this structure has reduced the cost and complexity while increasing the reliability and capacity without sacrificing technical performance.

This multinode, multichannel, bi-directional, single wavelength video conference structure is in use between hospitals in the province of Manitoba. Details of this configuration along with its performance are discussed.

INTRODUCTION

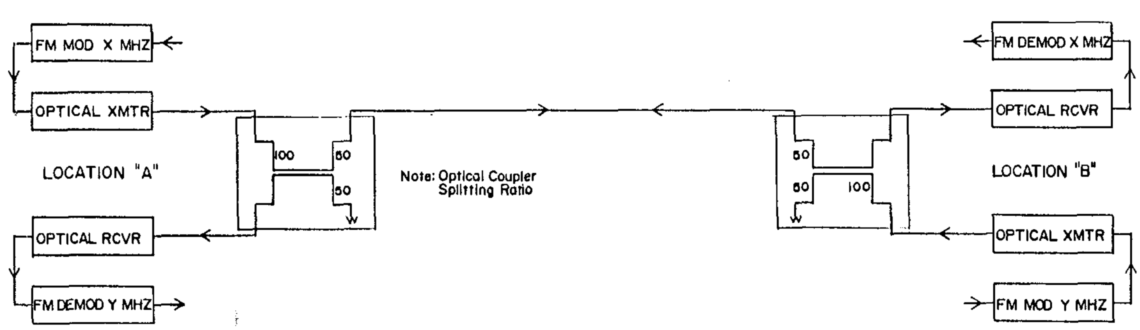
In order to comprehend this multinode, multichannel, bi-directional single wavelength, FM-FDM network architecture , the key building blocks are first examined. A bi-directional concept is combined with an optical structure having more than two nodes to achieve the end result. The theory is exemplified by the practical Hospital Conference Network with realistic system performance specifications.

BI-DIRECTIONAL OPTICAL LINK AT THE SAME WAVELENGTH

One key building block is illustrated in Figure 1. It is a bi-directional optical link at the same wavelength. The transmitter at location "A" is coupled into the fibre facility between "A" and "B" through a passive optical coupler. At "B" the circuit is completed with the receiver connected through another optical coupler. At the same time the transmitter at location "B" is coupled into the

FIGURE 1

BI-DIRECTIONAL OPTICAL LINK AT THE SAME WAVELENGTH



NOTE: ISOLATION BETWEEN INPUT PORTS OF OPTICAL COUPLERS MUST BE HIGH, TYPICALLY BETTER THAN 60dB
CARRIER FREQUENCY "Y" CAN BE THE SAME AS "X"

same fibre through the other port of the optical coupler. And this direction of the circuit is completed with the optical receiver at "A" coupled into the remaining port of the optical coupler at "A".

On the electrical side, the broadband RF signal input at "A" is regenerated at location "B". Simultaneously the broadband input at "B" is regenerated by the receiver at "A".

There are several reasons why the structure will operate in a non-interfering manner.

The photon density in the ten micron single mode fibre core is extremely low. Hence the probability of collisions between photons travelling in opposite directions is also low. Noise generated from such collisions is insignificant. An example of this would be the crossing of two flashlight beams which appear not to interfere with each other.

The other key to the operation of this circuit is the directivity of the optical couplers. Input port isolation is typically better than 60dB. This is important in isolating the receiver from the transmitter at the same location. Ideally the receiver at "A" should see only the optical signal transmitted from "B". Reflections or poor return loss in the optical network can mean that the transmitted signal at "A" will be received at "A" where it would interfere with overlapping RF spectrum. In other words, if the channel input at "A" is the same frequency as the channel input at "B", interference can occur if there is insufficient isolation between the two directions.

TWO TRANSMITTERS AT THE SAME WAVELENGTH LOOKING INTO ONE OPTICAL RECEIVER

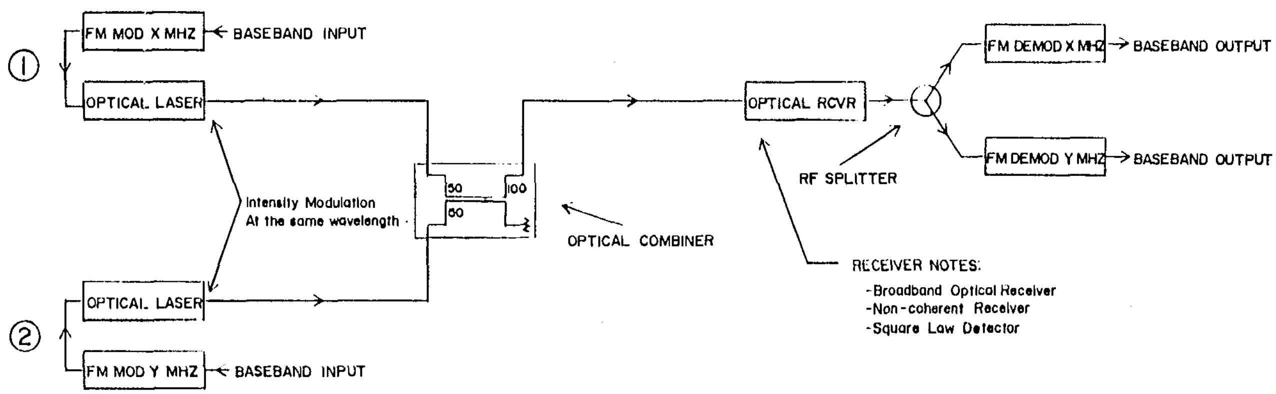
Figure 2 illustrates two optical transmitter signals of the same optical wavelength combined into one single mode fibre by means of a balanced optical combiner. The optical signals then travel through the fibre facility to the same optical receiver. The broadband RF inputs into laser "one" and laser "two" are regenerated by the optical receiver as a combined RF output. If the RF inputs overlap the RF output would also overlap but interference would only occur where the spectrum overlaps. the output of the optical receiver is then tuned by the appropriate frequency demodulator to retrieve the message signal.

On closer analysis, the vestigial sideband video signals are first frequency modulated onto different carriers. The transmitters accept the broadband inputs and directly modulate the current which drives the laser at 1310nm with a spectral line width of approximately 0.4nm. This means that the laser outputs are intensity modulated: there is no phase or frequency reference to the carrier at 1310nm.

The output of laser "one" is an intensity modulated waveform that represents the broadband RF spectrum at its input. Similarly laser "two" outputs an intensity modulated waveform that represents the spectrum at its input. The two optical signals combine through the optical coupler to produce an intensity modulated signal that represents the combined RF input spectrums. The information is contained in the optical power. The

FIGURE 2

TWO TRANSMITTERS AT THE SAME WAVELENGH
LOOKING INTO ONE OPTICAL RECEIVER



NOTE: RF Input into laser number one can not overlap RF Input into laser number two
ie) Carrier frequency "X" can not equal carrier frequency "Y"

combined optical signal is essentially a summation of optical powers. It is important that both of these optical signals are evenly balanced.

The receiver, by means of a square law detector regenerates the RF signals. The receiver is non-coherent; the phase and frequency of the incoming optical signal do not influence the receiver's output. In other words, the receiver does not lock (phaselock or frequency lock) onto any incoming signal. The FDM (frequency division multiplexing) provides the means by which the message signals are separated. To separate the signals at the receiver a simple frequency demodulator tuned to the appropriate frequency will reproduce the correct message signal. The frequency modulation provides the required signal to noise at the expense of more RF bandwidth as compared to direct amplitude modulation.

By combining the two principles of bi-directionality and multiple optical signals at the same wavelength traveling in the same direction on the same fibre the following network is derived.

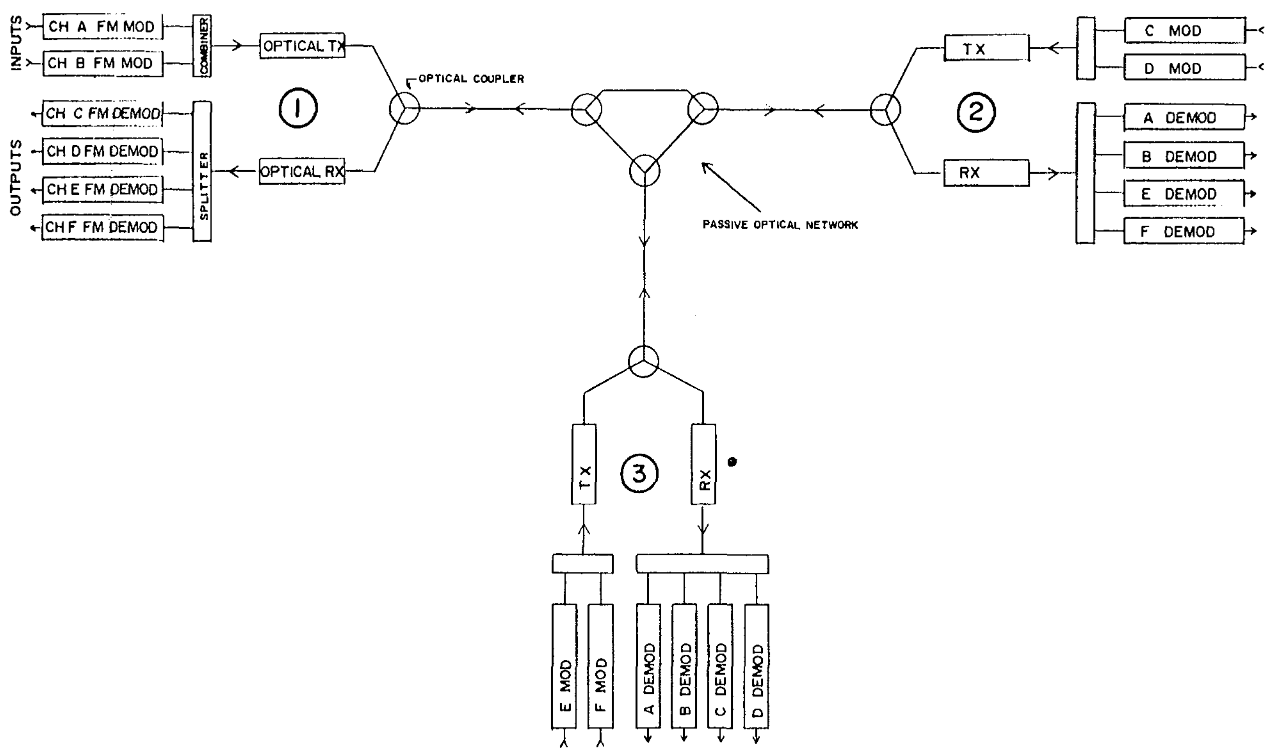
MULTINODE, MULTICHANNEL, BI-DIRECTIONAL, SINGLE WAVELENGTH FM-FDM NETWORK ARCHITECTURE

Figure 3 illustrates a multinode, multichannel, bi-directional, single wavelength, frequency modulation, frequency division multiplexing network architecture. Three locations are shown. Each location has one optical transmitter and one optical receiver. All the transmitters are DFB lasers at the same wavelength. All transmitters are identical

FIGURE 3

MULTINODE, MULTICHANNEL, BI-DIRECTIONAL, SINGLE WAVELENGTH, FM-FDM

NETWORK ARCHITECTURE



off-the-shelf standard wavelength DFB lasers. The receivers, one at each location are also all the same type. Optically the transmitter and receiver at each location are coupled into one single mode fibre. This means that only one optical fibre is required to provide the communication link to and from all the locations. Three fibres, one from each location come to a central passive combiner that is simply a combination of optical couplers designed to optically balance the network. The optical couplers are all fusion spliced to the fibre forming a transmission path entirely free of any active or mechanically connectorized devices other than the pigtails that connect directly to the optical transmitters and receivers.

Electrically, each location is transmitting two message signals; location 1 transmits channels A and B. Each location receives four signals; location 1 receives channels C and D from 2 and E and F from 3. The signals are transmitted as FM-FDM spectrum. This means that each message signal will be frequency modulated onto its own unique carrier. This is the key to differentiating the baseband signals at the receiver. The FDM is achieved through a simple passive RF combiner. On the receive side, the regenerated RF signal is split by a passive RF splitter and is fed into four frequency demodulators that are tuned to the required four channels.

This conferencing network clearly illustrates a Multinode, Multichannel, Bi-Directional, Single Wavelength, Single Fibre, FM-FDM communication facility. With this theory in hand the Manitoba Telephone System was ready for a real application of this form of fibre architecture.

THE HOSPITAL VIDEO NETWORK

The application here is a private conference network for use between hospitals in the Winnipeg and Brandon areas. The customer, The University of Manitoba (teaching hospital) requires conferencing between three locations. This network is illustrated in Figure 4. The three conference locations are: The University of Manitoba, The St. Boniface Hospital and the Brandon Mental Hospital; two in Winnipeg, one in Brandon, 200km west of Winnipeg. The link between Brandon Mental and the Winnipeg Fort Rouge TOC (television operating centre) uses an NEC video codec (digital video to and from Brandon). In Winnipeg, fibre exists between the TOC and the University of Manitoba, and the TOC and the St. Boniface Hospital.

In essence the Fort Rouge TOC becomes the third location, representing the Brandon node. The passive optical network that links all the locations is best located right at the TOC where all the fibres terminate. With the optical passive network at one of the nodes, the TOC, the optical signal levels are unbalanced. The optical combining/splitting network can be configured to balance the three optical links. In our particular case this was achieved by using a 30/70% power split (as illustrated) effectively only dropping off 30% of the light to the TOC and combining only 30% of the light from the TOC. Seventy percent of the power passes to and from the University of Manitoba and the St. Boniface Hospital.

The link contains a total of four video channels; two transmitted from the University of Manitoba, one transmitted from Ft. Rouge TOC (the Brandon signal) and one transmitted from the St. Boniface Hospital. RF channels in the 400-500 MHZ range originate at the University of Manitoba, in the 200-300 MHZ range from St. Boniface and in the 300-400 MHZ range from Ft. Rouge TOC.

The Catel Multichannel Fibre Optic System which incorporates the series 3000 Broadband FM Transmission System with the OT-1010 Optical Transmitter and the OR-1010 Optical Receiver is used throughout the electrical network. BT&D (British Telecom & Dupont) optical couplers are used in the optical network. The optics contain only fusion splices and bypass Telephone System optical patch panels.

The network was turned up and has operated since February of 1990. We are not aware of a comparable network in existance to date.

SYSTEM PERFORMANCE

Following is the network performance summary. Table one is an overview of measured video parameters at each location. The measurements were made with the Tektronics VM 700 automated test

FIGURE 4

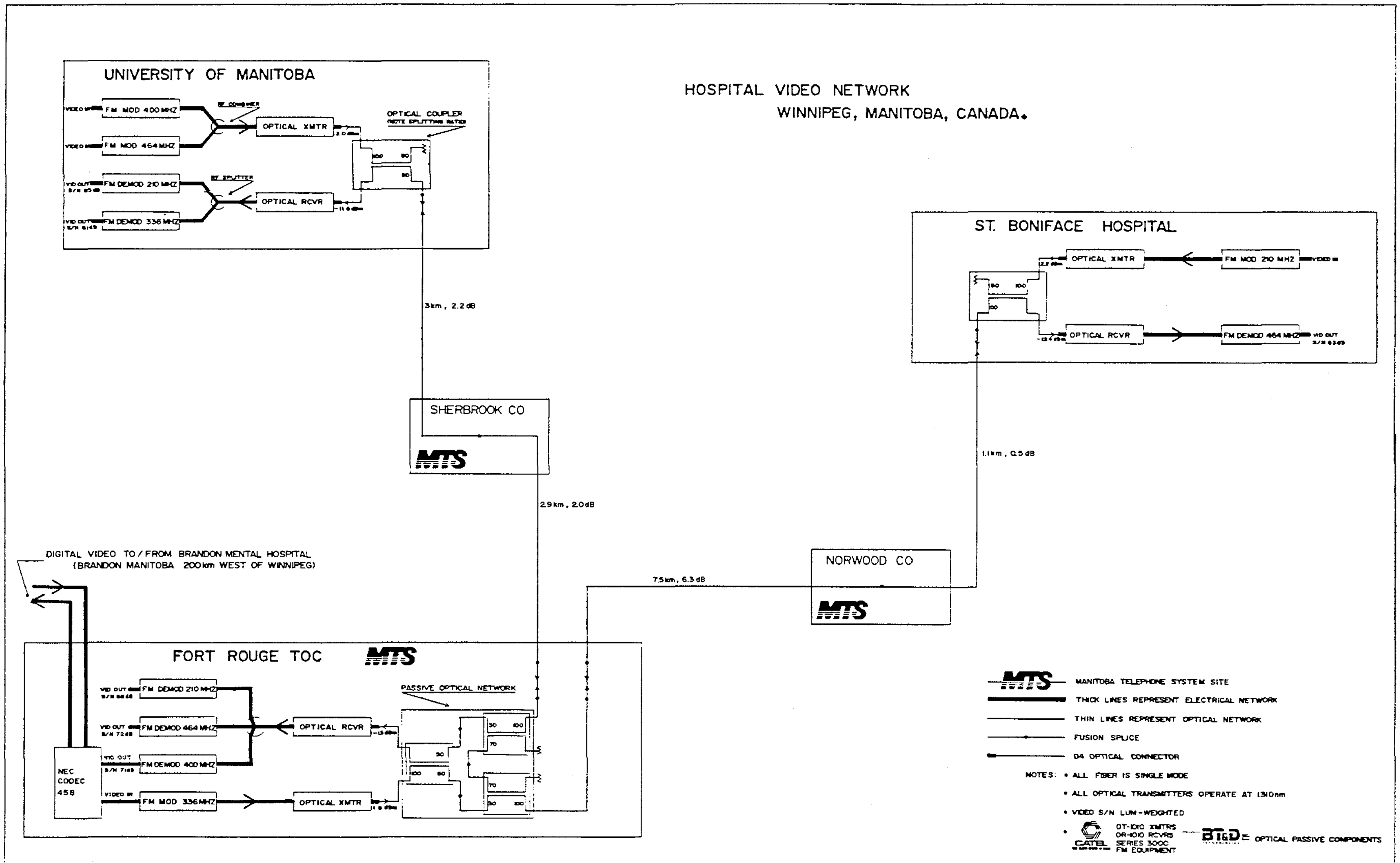

HOSPITAL VIDEO NETWORK
WINNIPEG, MANITOBA, CANADA.
UNIVERSITY OF MANITOBA
FM MOD 400 MHZ
FM MOD 464 MHZ
FM DEMOD 210 MHZ
FM DEMOD 336 MHZ
OPTICAL XMTR
OPTICAL RCVR
OPTICAL COUPLER
3km, 2.2dB
ST. BONIFACE HOSPITAL
OPTICAL XMTR
FM MOD 210 MHZ
OPTICAL RCVR
FM DEMOD 464 MHZ
SHERBROOK CO
MTS
2.9km, 2.0dB
1.1km, 0.5dB
NORWOOD CO
MTS
7.5km, 6.3dB
DIGITAL VIDEO TO / FROM BRANDON MENTAL HOSPITAL
(BRANDON MANITOBA 200km WEST OF WINNIPEG)
FORT ROUGE TOC
MTS
FM DEMOD 210 MHZ
FM DEMOD 464 MHZ
FM DEMOD 400 MHZ
FM MOD 336 MHZ
OPTICAL RCVR
OPTICAL XMTR
PASSIVE OPTICAL NETWORK
NEC CODEC 458
MANITOBA TELEPHONE SYSTEM SITE
THICK LINES REPRESENT ELECTRICAL NETWORK
THIN LINES REPRESENT OPTICAL NETWORK
FUSION SPLICE
D4 OPTICAL CONNECTOR
NOTES: • ALL FIBER IS SINGLE MODE
• ALL OPTICAL TRANSMITTERS OPERATE AT 1310nm
• VIDEO S/N LUM-WEIGHTED
CATEL
BT&D
OPTICAL PASSIVE COMPONENTS

TABLE 1

Video Parameter	University of Manitoba 210 MHz	University of Manitoba 336 MHz	TOC, Fort Rouge 400 MHz	TOC, Fort Rouge 464 MHz	TOC, Fort Rouge 210 MHz	St. Boniface Hospital 400 MHz	St. Boniface Hospital 464 MHz	St. Boniface Hospital 336 MHz
Bar Amplitude (IRE)	100.6	101.4	99.4	101.5	98.9	100.2	99.9	100.0
Sync Amplitude (% Bar)	40.2	40.0	40.1	40.0	40.1	40.4	40.5	40.1
Line Time Distortion (%)	0.5	0.3	0.5	0.1	0.9	0.5	0.9	0.4
Pulse / Bar Ratio (%)	95.8	96.1	92.6	95.2	97.1	92.1	89.1	90.0
2T Pulse K-Factor (% K-f)	0.6	0.6	0.5	0.6	0.4	0.7	0.9	0.4
S/N Unweighted (dB)	55.3	50.9	63.5	64.4	57.0	56.2	56.3	55.1
S/N Lum-Weighted (dB)	64.5	61.2	71.2	71.5	67.5	66.3	66.3	67.1
S/N Periodic (dB)	50.0	44.7	58.3	58.9	50.9	51.1	51.3	55.4
Chroma-Luma Delay (ns)	30.5	10.5	15.1	6.3	19.6	21.5	27.1	18.8
Chroma-Luma Gain (%)	97.9	98.8	98.4	95.1	99.8	101.9	97.5	98.6
Differential Gain (%)	0.99	3.36	4.60	2.46	1.67	0.50	0.77	0.80
Differential Phase (Deg)	0.77	1.11	1.53	3.04	1.03	1.41	0.89	1.32
Lum Non-Linearity (%)	1.58	0.20	1.22	0.67	1.48	2.46	3.80	2.91
80 IRE Chroma (IRE)	77.8	80.3	77.4	78.0	77.0	78.2	77.7	78.0
Chr NL Phase (Deg)	3.0	1.6	2.8	1.4	3.0	0.9	0.5	1.2
Chr-Lum Intmd (IRE)	-1.8	-0.4	-1.1	-0.4	-1.6	-2.4	-3.7	-2.5
Field Time Dist (% Bar)	0.64	1.01	0.85	0.79	0.64	1.02	1.11	0.95

set. Measurements were made with simultaneous modulation at each location on each video channel.

Table two shows optical transmit levels. Table three shows all combinations of optical receive levels.

Table four shows received RF levels at each location.

CONCLUSIONS

The network described in this paper uses commercially available off-the-shelf components.

The structure is ideal for video conferencing and educational networks over moderate distances. The Hospital Video Network with three locations proves to be economical, reliable, and simple without sacrificing performance. The maintenance of this network is also very straightforward. Spare parts are compatible at all locations. Only 1310nm transmitters are used, FM modulators and

TABLE 2

Optical Transmit Power (dBm)		
University of Manitoba Laser "A"	St. Boniface Hospital Laser "B"	Ft. Rouge TOC Laser "C"
2.0	2.2	1.6

TABLE 3

Optical Receive Power (dBm)			Laser Status	
University of Manitoba	St. Boniface Hospital	Ft. Rouge TOC	ON	OFF
0.0	0.0	0.0		A,B,C
-36.2	-12.2	-12.5	A	B,C
-11.2	-44.5	-10.2	B	A,C
-13.5	-10.9	-40.2	C	A,B
-12.0	-12.2	-12.3	A,B	C
-13.5	-8.2	-12.5	A,C	B
-13.4	-10.9	-10.2	B,C	A
-11.6	-13.4	-13.0	A,B,C	

Note: Optical Receiver Sensitivity, -25dBm

TABLE 4

LOCATION	RF RECEIVE LEVELS (dBmv) 210 MHZ	336 MHZ	400MHZ	464MHZ
University of Manitoba	28	30	-56	-54
St. Boniface Hospital	-50	30	32	32
Ft. Rouge TOC	30	-52	30	30

demodulators are all frequency agile and there are no active devices or optical connections outside of the three locations. The spare parts for this network are as follows:

One 1310nm transmitter
One 1310nm receiver
One FM modulator (agile)
One FM demodulator (agile)
One power supply

The optical passive devices used in the network are BT&D couplers which are minor cost items.

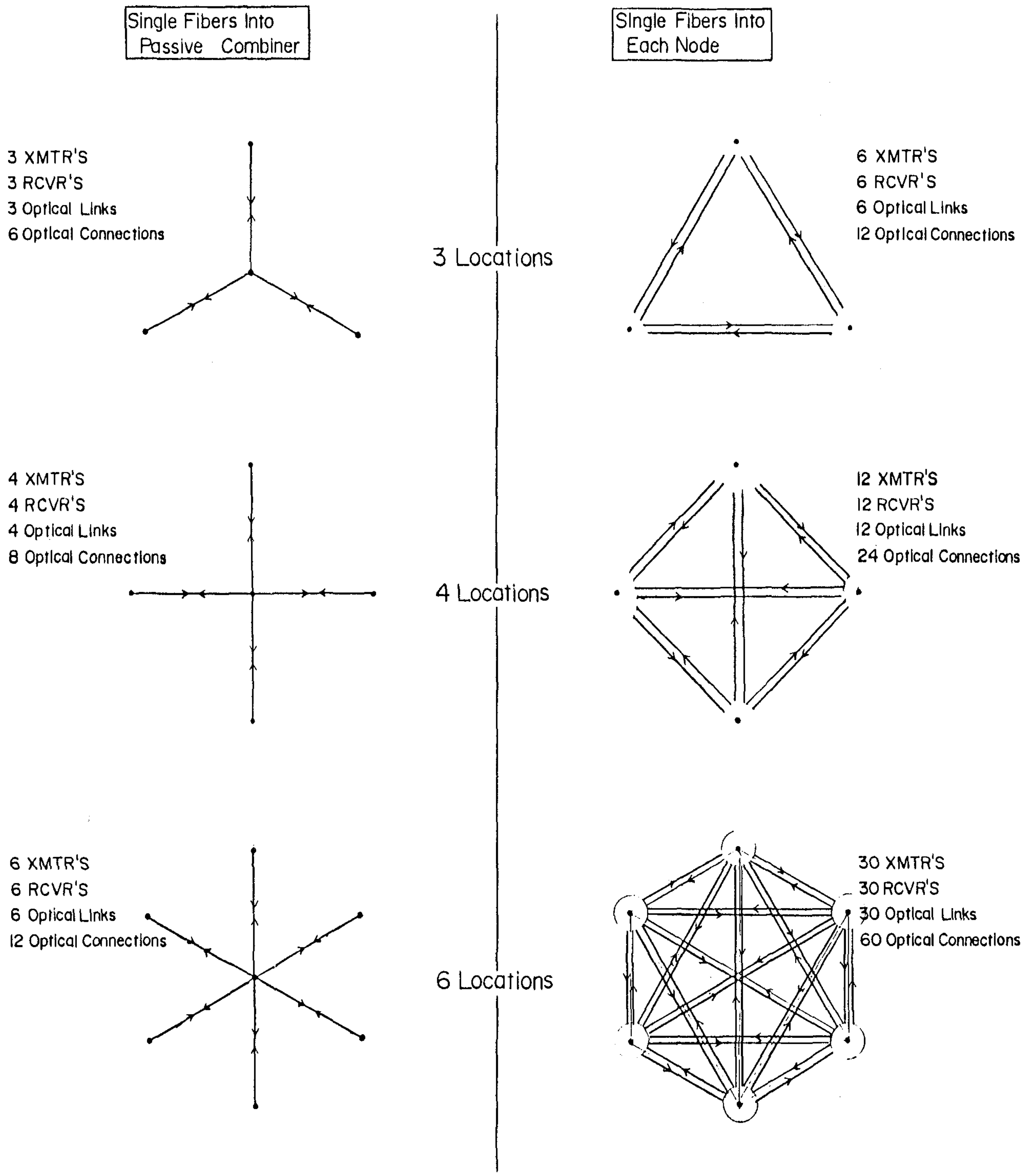
Other structures could be used to link three locations for video conferencing. One such structure would be individual fibres to and from each location with separate transmitters and receivers. With the Hospital Video Network this would have meant twice the number of transmitters and receivers or in other words over a 1/3 increase in the total network cost.

Another structure would involve using wavelength division multiplexing (WDM). WDM devices which are optical prisms that separate the wavelengths are fairly expensive. Also transmitters and receivers would be at different wavelengths requiring twice the number of spare parts. This type of network would be more expensive, more complex and harder to maintain.

Figure 5 shows direct comparisons of equipment and complexity based on different fibre architectures. Clearly as locations are added savings can be substantial. The performance of this multinode, multichannel, bi-directional, single wavelength, FM-FDM network with more than three nodes and more than four channels is a question that requires further investigation.

FIGURE 5

VIDEO CONFERENCE FIBER ARCHITECTURES



A TUTORIAL ON DIGITAL VIDEO COMPRESSION TECHNIQUES

Joseph B. Waltrich
Jerrold Communications
Applied Media Lab

ABSTRACT

Digital video transmission offers significant advantages to the cable operator because of its reduced susceptibility to channel impairments. However, a penalty is exacted in terms of the bandwidth required to transmit an uncompressed digital video signal. In order to achieve a reasonable bandwidth for digital video transmission, some degree of video bandwidth compression is necessary.

This paper discusses the relationship between compression requirements and transmission bandwidth and presents a review of some of the more commonly advocated video compression schemes. The applicability of video compression for delivery of both NTSC and HDTV is discussed. Implications for both cable and fiber are considered.

INTRODUCTION

Because of its reduced susceptibility to transmission channel impairments, digital video transmission offers significant advantages over analog transmission. However, a penalty is exacted in terms of bit rate and the associated bandwidth required for transmission. For example, if a composite NTSC signal is sampled at four times the color subcarrier with 8 bit quantization, then the active video portion of each line would contain 768 samples. Since there are 480 active video lines in an NTSC frame and the NTSC frame rate is 29.97 frames/second, the total video bit rate can be calculated as follows:

$$8 \times 768 \times 480 \times 29.97 = 88.4 \text{ Megabits/second} \quad (1)$$

It should be noted that the bit rate calculated from Equation (1) is for transmission of active video only. In practice, additional data such as digital audio (500 Kbits/sec or more), synchronization data and error corection data must also be transmitted. For cable applications, additional data must be transmitted for encryption and addressabiility. If all these factors are considered, the total bit rate for digital video transmissionn could easily exceed 100 Mbits/sec.

The required bandwidth for digital transmission is a function of the total data rate and the modulation technique. The bandwidth may be calculated as follows:

$$W = R_d/E_s \quad (2)$$

where:

W = bandwidth (Hz)
R_d = total data rate (b/s)
R_s = spectral efficiency (b/s/Hz)

Table 1 lists the spectral efficiencies for various modulation schemes. Additional information regarding the data in Table 1 may be found in Feher [1].

From Equation (2), it is seen that transmission of uncompressed video at bit rates on the order of 100 Mb/s would require a transmission bandwidth of 30 - 60 MHz. Transmission of HDTV in RGB form (approximately 1.5 Gb/s) would require a bandwidth of about .5 -1 GHz. This is obviously impractical for cable and, although it might be argued that the use of fiber optics would alleviate bandwidth constraints, the cost of electronics for fiber based systems increases with increasing bit rate. Therefore, some form of bandwidth compression must be employed in order to achieve both spectral and cost efficiencies for digital video transmission.

COMPRESSION REQUIREMENTS

Prior to discussing compression techniques, the degree of compression as a function of bandwidth should be examined. The degree of compression is best expressed in terms of the average information or entropy of a compressed video source, expressed in terms of bits/pixel. (A pixel or pel, as it is sometimes abbreviated is defined as a picture element or sample). As previously stated, the total data rate includes not only digitized video but also digital audio and a certain amount of overhead data. The total data rate, as a function of video entropy and other data, is given by:

$$R_d = (R_v H_v + R_a)(1 + OH/100) \qquad (3)$$

where:

R_d = total data rate (b/s)
R_v = video data rate (pel/s)
H_v = video entropy (b/pel)
R_a = audio data rate (b/s)
OH = overhead (%)

In order to fit a signal into a given bandwidth, the amount of compression must be determined for a particular modulation technique. That is, the entropy must be calculated, given the other parameters. Using (2) and (3), the following expression may be obtained for the entropy:

$$H_v = (1/R_v)(WE_s/(1 + OH/100) - R_a) \qquad (4)$$

For successive calculations, it will be assumed that digital audio is transmitted at 500 Kb/s and that a 25% overhead is required for ancillary data.

In practice, video compression is usually performed on the signal in component form. Although the signal may be in RGB format, it is more efficient to perform signal processing on the luminance and chrominance components (e.g. - Y, I, Q) since the chrominance components may be sampled at half the rate used for the luminance, thereby reducing the uncompressed data rate. (The missing samples are reconstructed by interpolation at the receiver). Three sampling schemes will be considered:

- NTSC sampled at 4 times subcarrier
- NTSC sampled at 3 times subcarrier
- HDTV in Common Image Format

For the first case, there are 768 luminance pels per line and 480 active lines per frame. The luminance data rate is therefore:

$$R_Y = 768 \times 480 \times 29.97 = 11.05 \text{ Mpel/s} \qquad (5)$$

And, since each chrominance component is sampled at half the rate for luminance:

$$R_C = 384 \times 240 \times 29.97 = 2.76 \text{ Mpel/s} \qquad (6)$$

Therefore:

$$R_V = R_Y + 2R_C = 16.57 \text{ Mpel/s} \qquad (7)$$

For NTSC sampled at 3 times subcarrier, there are 576 luminance pels/line. Using the same sampling scheme, one obtains a value of 8.29 Mpel/s for R_Y and 2.07 Mpel/s for R_C. This yields a value of R_V = 12.43 Mpel/s.

The Common Image Format for HDTV is currently specified as 1920 pels/active line by 1080 active lines. Applying the same sampling scheme to this format yields an R_V = 93.31 Mpel/s.

Fig. 1 presents a plot of H_V as a function of bandwidth for the three above-mentioned cases using a modulation technique with a spectral efficiency of 1.66 b/s/Hz. From this plot, it is seen that, in order to transmit digital NTSC in a 6 MHz channel using 2PAM or 4QAM, an entropy of .45 - .6 b/pel is required, depending on the sampling frequency. Assuming 8 bit source quantization, this amounts to a compression ratio of 13 -18. If 4PAM or 16QAM is used, the entropy is doubled and the compression ratio is halved. Transmission of digital HDTV in 6 MHz would require an entropy of .08 - .16 b/pel, corresponding to a compression ratio of 50 - 100.

COMPRESSION TECHNIQUES

During the past decade, considerable effort has been expended on the development of a variety of digital video compression techniques. Although much of this research was spurred by non-entertainment applications of television (e.g. - teleconferencing, military applications, etc.), some of these techniques have recently found their way into commercial television. Increased interest in HDTV has also generated a corresponding interest in compression techniques.

Currently, a number of video compression techniques are being used in a variety of applications. These include the following:

- o Predictive Coding (e.g. - DPCM)
- o Transform Coding
- o Vector Quantization
- o Subband Coding

This paper will concentrate on discussion of those techniques which are currently thought to be best suited to broadcast and cable applications.

Differential Pulse Code Modulation

Differential Pulse Code Modulation (DPCM) is a technique in which the value of a given pixel is estimated, based on the values of preceding pixels. This estimate or predictor is a linear function of preceding pixel values. For a pixel having a value X_N, the general form of the predictor is:

$$\hat{X}_N = \sum_{i=0}^{i=N-1} a_i X_i \qquad (8)$$

The predicted value is then subtracted from the pixel value to generate an error signal:

$$e_N = X_N - \hat{X}_N \qquad (9)$$

The error signal is encoded and transmitted. At the receiver, the pixel value is recovered by adding the error signal to the predictor which the receiver has determined from previously recovered pixels. The equation for the recovered signal X'_N is:

$$X'_N = e_N + \hat{X}'_N \qquad (10)$$

where

$$\hat{X}'_N = \sum_{i=0}^{i=N-1} a_i X'_i \qquad (11)$$

The bit rate reduction for DPCM is due to the fact that the variance of the error signal e_N is significantly less than that of the original image and therefore the error signal lends itself well to bit rate reduction via variable run length coding.

Fig. 2 presents an example of DPCM using a simple 2-dimensional predictor (i.e. - pixels from both the present and previous lines are used to estimate the value of X_N).

A block diagram of a DPCM system is shown in Fig. 3. System complexity depends on the nature of the prediction algorithm. Predictors may be either one, two or three dimensional, requiring from one line to one frame of memory. In practice, it is advantageous to use combinations of multi-dimensional predictors which are adaptively switched so as to minimize the value of the error signal. A more detailed discussion of adaptive prediction schemes may be found in publications by Ng and Hingorani [2] and Knee [3].

A comparison of the distribution of quantization levels for an uncompressed image and its associated error signal is shown in the histograms of Fig. 4. The variance of the error signal is approximately 1/50th that of the original image. Because of this, the error signal can be encoded using a variable run length code such as Huffman coding, thereby achieving a substantial reduction in entropy. This technique yields entropies on the order of 2 -5 b/pel.

Since the value of a reconstructed pixel depends on the value of previous pixels, transmission errors in DPCM can cause streaking effects over one or more lines of the picture, depending on the nature of the prediction algorithm.

Huffman Coding

Huffman coding is a variable length coding technique which reduces the average bit rate required to represent a set of quantization levels. This is accomplished by assigning short codewords to those quantization levels which have the greatest probability of occurrence and longer codewords to less frequently occurring quantization levels. Obviously, this scheme works best on signals having relatively little spread in the distribution of quantization levels (e.g. - the DPCM error signal). An illustration of the coding technique is shown in Fig. 5. The coding procedure is accomplished by starting with the two least frequently occurring quantization levels. These are combined into one node (a7 of Fig. 5). logic 0 and 1 levels are arbitrarily assigned to the two branches of this node. This process is repeated until we are left with a single node having a probability of 1. The codeword associated with each probability is determined by tracing back through the branches, starting with the last node.

The coding scheme of Fig. 5 is, of course, an oversimplification, since, in practice, the number of quantization levels is larger than six. However, a group of codes with very low probabilities of occurrence can be assigned a single codeword. The actual value of the quantization level is then transmitted following the codeword.

Obviously, a variable length code such as Huffman coding lends itself to other compression schemes as well as DPCM.

Transform Coding

In transform coding, the image is transformed from the time domain to a different domain (e.g. - the frequency domain). The transform coefficients are then encoded and transmitted. An inverse transformation is performed at the receiver to recover the original image.

A number of transforms have been used for various video compression applications. Among these are the Karhunen-Loeve, Walsh-Hadamard, Haar, Discrete Fourier and Discrete Cosine transforms. Detailed information on all of these transforms is given by Stafford [4]. Currently, the Discrete Cosine Transform (DCT) appears to be the most widely used of the above-mentioned transforms. Therefore, this paper will limit discussion of transform coding to the DCT.

The DCT is typically performed on blocks of pixels ranging in size from 4 x 4 to 16 x 16. The transform is similar to the real part of a Fourier transform. The contents of each pixel block are converted to a series of coefficients which define the spectral content of the block. The general form of a two-dimensional DCT performed on an N x N pixel block is given by the equation:

$$Y_{mn} = (1/2N) E_m E_n \sum_{k=0}^{N-1} \cos((2k + 1)n\pi/2N)$$

$$\times \sum_{j=0}^{N-1} X_{jk} \cos((2j + 1)m\pi/2N) \quad (12)$$

where:

Y_{mn} = DCT coefficients at coordinates m, n
X_{jk} = pixel amplitude at coordinates j, k
$E_m, E_n = 1/\sqrt{2} \quad (m,n = 0)$
$E_m, E_n = 1 \quad (m = n \neq 0)$

The inverse transform (IDCT) is given by the equation:

$$X_{jk} = (2N/E_m E_n) \sum_{m=0}^{N-1} \cos((2j + 1)m\pi/2N)$$

$$\times \sum_{n=0}^{N-1} Y_{mn} \cos((2k + 1)n\pi/2N) \quad (13)$$

Fig. 6 presents some examples of DCT's of various pixel patterns. For pixel block patterns which are typical of live video, the DCT reduces the pixel block data to only a few non-zero coefficients. Therefore, it would be expected that a bit rate reduction would result solely from the transform process. In practice, however, the presence of noise in the signal can generate spurious transform coefficients. Filtering the input signal prior to generating the transform is, therefore, necessary in order to reduce the occurrence of these coefficients.

In order to maintain a desired entropy, it is often necessary to discard some of the DCT coefficients during the coding process. This is usually not a problem since, in most cases, only a few coefficients make a major contribution to the signal's spectral content. The coefficient selection process is typically determined by one of two types of coding: zonal coding and threshold coding. Zonal coding discards all of the coefficients except those within a selected zone (which always includes the low frequency components). Threshold coding discards those coefficients whose values are below a given set of threshold levels. (In some forms of threshold coding, these coefficients are coarsely quantized rather than simply discarded).

Since, in zonal coding, the high frequency coefficients of the DCT are discarded, edge blurring in the reconstructed image sometimes occurs. This effect is less noticeable in threshold coding since coefficient selection depends on the magnitude of a particular spectral component as well as its position. However, threshold coding carries a bit rate penalty since coefficient locations are not predetermined and, therefore, coefficient address information must also be transmitted.

Unlike a DPCM encoded picture, transmission errors when using DCT coding are confined to individual pixel blocks. This would cause a "spotting effect" in the decompressed picture, the nature of which depends on which DCT coefficients were corrupted.

Commercially available DCT codecs are currently being used for transmission of broadcast quality NTSC video at 45 Mb/s [5]. Several IC manufacturers will soon be offering DCT processors in chip form. Among these are the INMOS IMSA121, SGS-Thomson STV3200 and the LSI Logic L64730. These chips are capable of operating at clock rates in the 13.5 - 40 MHz range and computing both forward and inverse transforms on 8 x 8 pixel blocks.

The DCT is capable of generating reasonably good picture quality at entropies on the order of .5 - 2 b/pel. It is possible to combine the DCT with other compression schemes such as DPCM to achieve greater bit rate reductions.

Vector Quantization

In vector quantization, an image is divided into a number of non-overlapping blocks. Each block is regarded as an N-dimensional image vector $\mathbf{X}$ where N is equal to the number of pixels in the block. Each vector is compared with a set of N_C stored reference patterns or codevectors $\mathbf{Y_1}\ldots\mathbf{Y_{Nc}}$ in order to find the codevector $\mathbf{Y_k}$ which most closely matches the image vector $\mathbf{X}$. Once found, the index k of the codevector is transmitted to the receiver which then uses a lookup table to reproduce the codevector $\mathbf{Y_k}$.

The entropy for vector quantization is dependent on the size of the vector and the number of vectors stored in the codebook. If the codebook size N_C is a binary number which can be represented by b bits, then a vector of N pixels is represented by one of 2^b codevectors and the entropy is simply:

$$H_V = b/N = \log_2(N_C)/N \qquad (14)$$

Conversely, the codebook length N_C may be determined from the desired entropy:

$$N_C = 2^{(NH_V)} \qquad (15)$$

The codevector $\mathbf{Y_k}$ is chosen to minimize the error (commonly referred to as the distortion) between $\mathbf{Y_k}$ and the image vector $\mathbf{X}$.

The key elements for effective vector quantization are codebook generation and codebook search. Codebook generation may be based on statistical distribution of image vectors or on the use of training images. The LBG algorithm [6] is a popular method for optimizing codebook design.

The method of searching the codebook affects the number of computations required to determine the best choice of codevector. An exhaustive search involves computation of the distortion between the image vector and every vector stored in the codebook. By using a tree-structured codebook and binary search, the number of computations is decreased significantly. However, more memory is required for storage of a tree structured codebook. Details of codebook design and search techniques may be found in Lim [7].

In vector quantization, transmission errors can have a significant effect on the reproduction of individual pixel blocks since an incorrect codevector index results in selection of the wrong codevector by the decoder. As in the DCT, however, errors are confined to individual pixel blocks. The effect of occasional errors would be a random pattern of light and dark spots in the picture.

Subband Coding

In subband coding the image is divided into frequency subbands using various combinations of horizontal and vertical filtering and subsampling techniques. One form of subband coder, in which the image is divided into octave bands, is shown in Fig. 7. Prior to subsampling, the image is divided into frequency bands by combinations of lowpass and highpass filters. Following frequency division, the image is subsampled or decimated by an integral factor M as shown in the blocks represented by the down arrows. (For division into octave bands as shown in Fig. 7, M = 2).

At the receiver, the subbands are interpolated by an integral factor L by inserting L -1 zeroes between received samples of the subband data. (For octave bands, L =2). The interpolation process is indicated by an up arrow. The interpolated subbands are then filtered and recombined to reconstruct the received image.

Image compression is obtained by coding each subband using compression schemes such as DPCM and/or by discarding some of the high frequency information or transmitting it at a lower rate. This technique has been proposed by both Schreiber [8] and Zenith [9] to compress HDTV into a 6 MHz channel.

The effects of subband coding in the frequency domain are shown in Fig. 8. If ideal filtering is assumed, the low frequency information is not affected by the decimation process. However, decimation of the high frequency information results in aliasing of the data with associated spectrum folding. The interpolation process creates spectral images which allow the original information to be recovered by filtering the imaged spectra within the desired frequency range. Detailed discussion of the effects of decimation and interpolation may be found in Crochiere and Rabiner [10].

Since it is not possible to construct ideal filters, the decimation process will produce some degree of aliasing in all bands. The filters used in subband coding must be chosen such that the synthesis filters (i.e. - the filters following the interpolators) cancel aliasing generated during the decimation process. This may be accomplished by the use of Quadrature Mirror Filters.

Quadrature Mirror Filter design considerations are described by Vaidyanathan [11].

CONCLUSIONS

If a suitable compression technique can be developed, digital video could be transmitted over existing cable systems. However, new headend and subscriber equipment would be required and different modulation schemes may require re-structuring of carriers. Fiber offers a more likely medium for digital transmission since a cost-effective compression technique would probably prove more economical than adding fibers and associated electronics to achieve increased system capacity. In addition, fiber would be free of most of the problems which could cause errors in digital transmission over cable.

Since video compression is generally performed on component signals, digital transmission offers the opportunity to eliminate NTSC artifacts via comb filter encoding and decoding.

Most compression schemes are more effective on clean signals. It is likely that, for signals such as satellite feeds, noise reduction techniques will have to be developed. If digital transmission becomes popular, compression and noise reduction techniques may spur the development of low cost frame stores for both headend and subscriber equipment.

REFERENCES

[1] K. Feher, "Telecommunications Measurements, Analysis and Instrumentation", Prentice-Hall, 1987.

[2] S. Ng and R. Hingorani, "Adaptive Spatio-Temporal DPCM for Videoconferencing", GLOBECOM 1986, Houston, pp 8.7.1-5.

[3] M. Knee, "Adaptive Prediction for High Quality Television Transmission Coding Based on the LMS Algorithm", SMPTE J., 98, 580-589, Aug. 1989

[4] R. Stafford, "Digital Television", Wiley, 1980

[5] S. Cucchi and F. Molo, "DCT Based Television Codec for DS3 Digital Transmission", SMPTE J., 99, 640-646, Sept. 1989

[6] Y. Linde, A. Buzo and R.M. Gray, "An Algorithm for Vector Quantizer Design", IEEE Trans. on Comm., Vol. COM-28, No. 1, 84-95, Jan. 1980.

[7] J. Lim, "Two Dimensional Signal and Image Processing", Prentice Hall, 1990.

[8] W. Schreiber et al, "Channel Compatible 6 MHz HDTV Distribution Systems", SMPTE J, 98, 5-13, Jan. 1989.

[9] W. Bretl, R. Citta, R. Lee, P. Fockens, "Spectrum Compatible High Definition Television Systems", SMPTE J., 98, 748-753, Oct. 1989.

[10] R. Crochiere and L. Rabiner, "Multirate Digital Signal Processing", Prentice Hall, 1983.

[11] P. Vaidyanathan, "Quadrature Mirror Filter Banks, M-Band Extensions and Perfect Reconstruction Techniques", IEEE ASSP Magazine, 4-20,

TABLE I

MODULATION TECNIQUES AND SPECTRAL EFFICIENCIES

Modulation Type	Theoretical RF Spectral Efficiency (b/s/Hz) [1]	Practical Efficiency (b/s/Hz) [2]
2PAM	2	1.66
3PRS	2	1.66
4QAM	2	1.66
4PSK	2	1.66
4PAM	4	3.33
16QAM	4	3.33
8PAM	6	5
64QAM	6	5
16PAM	8	6.66
256QAM	8	6.66
32PAM	10	6.66
1024QAM	10	6.66

[1] Assumes an ideal brick-wall filter

[2] Assumes a Nyquist filter with 20% rolloff

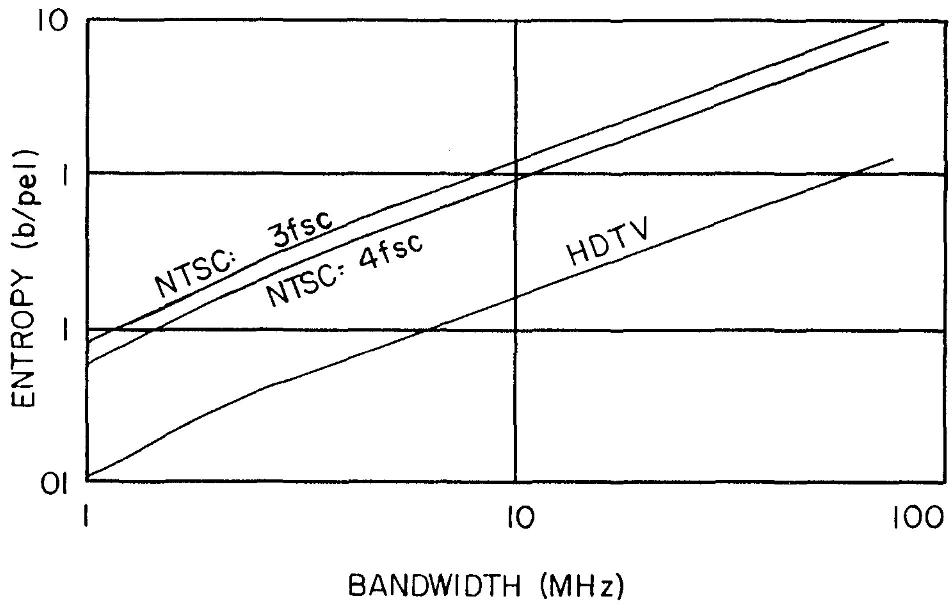


FIG. I REQUIRED ENTROPY AS A FUNCTION OF BANDWIDTH. SPECTRAL EFFICIENCY = 1.66 b/s Hz.

Predictor: X = .8X(x-1,y) - .6X(x-1,y-1) +.8X(x,y-1)

Transmitter:

line	Pixel Values		
N - 1	104	130	156
N	125	156	187
X	125	156	187
$\hat{X}$		142	172
e		14	15

Receiver:

line	Pixel Values		
N - 1	104	130	156
N	125	156	187
$\hat{X}'$		142	172
e		14	15
X'	125	156	187

Fig. 2 DPCM Example

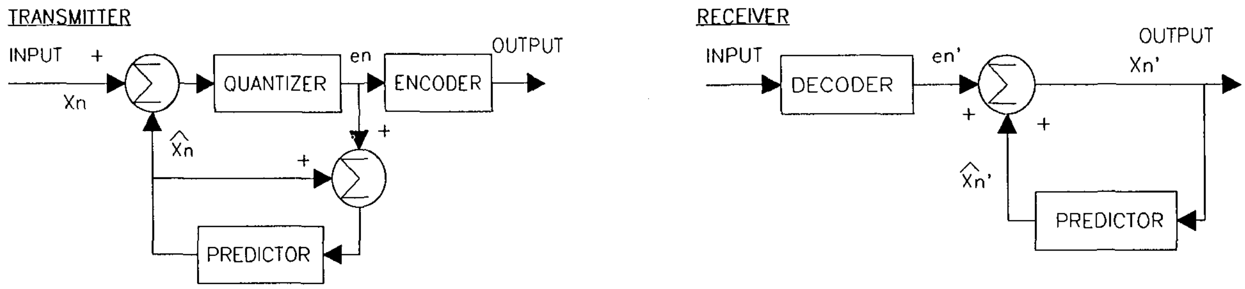


FIG. 3 DPCM SYSTEM BLOCK DIAGRAM

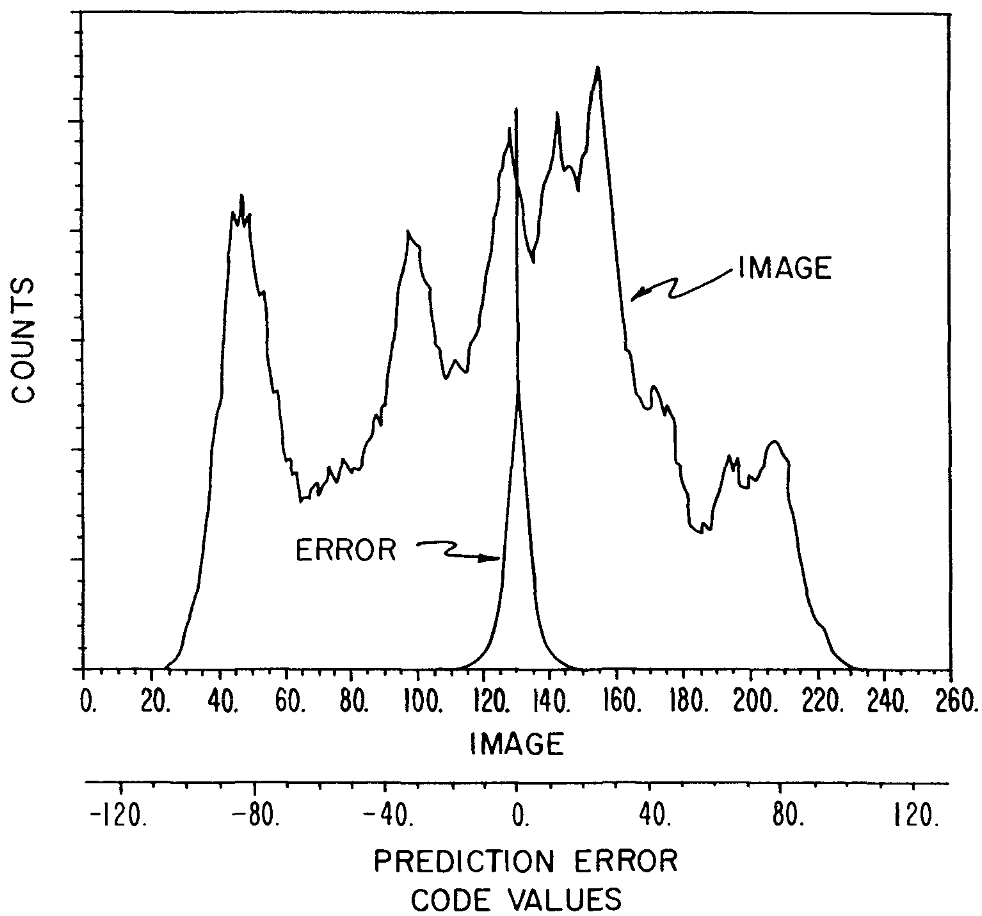


FIG. 4. COMPARISON OF IMAGE AND PREDICTION ERROR DISTRIBUTIONS

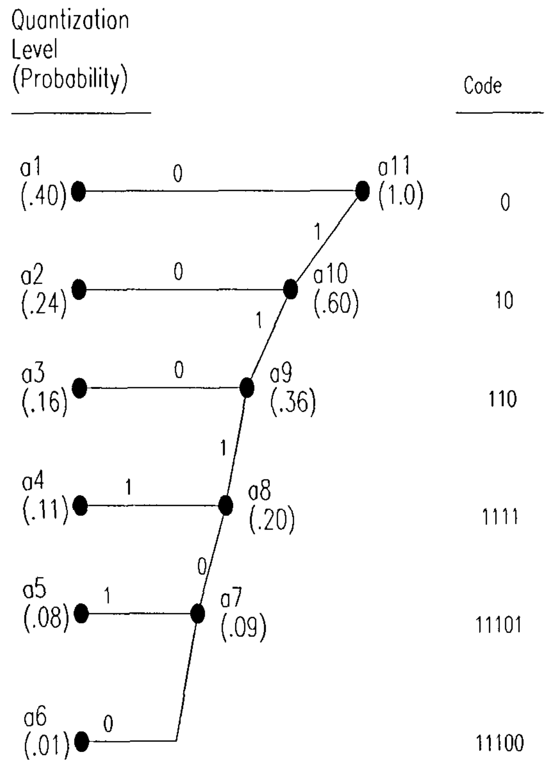


FIG. 5 Example of Huffman Coding Probabilities of Each Level are Shown in Parentheses

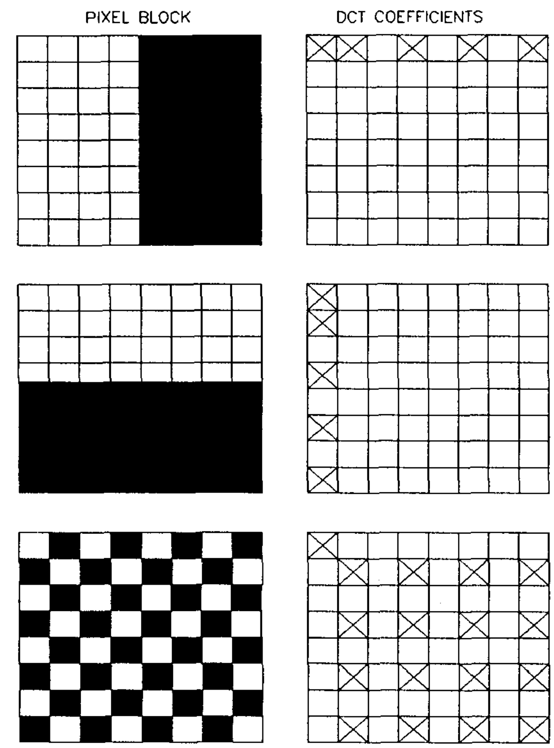


FIG 6. DCT EXAMPLES

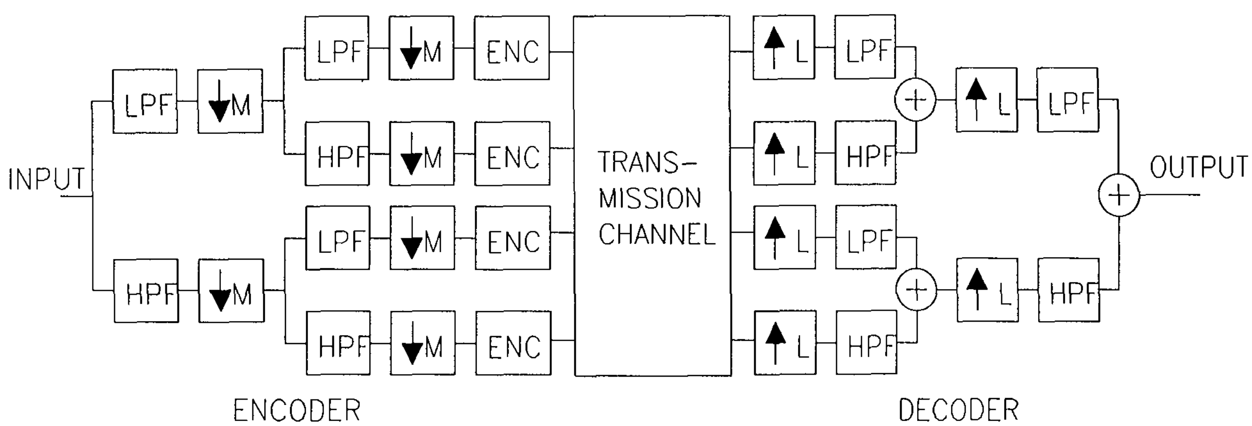


FIG. 7 SUBBAND CODING BLOCK DIAGRAM (M,L =2)

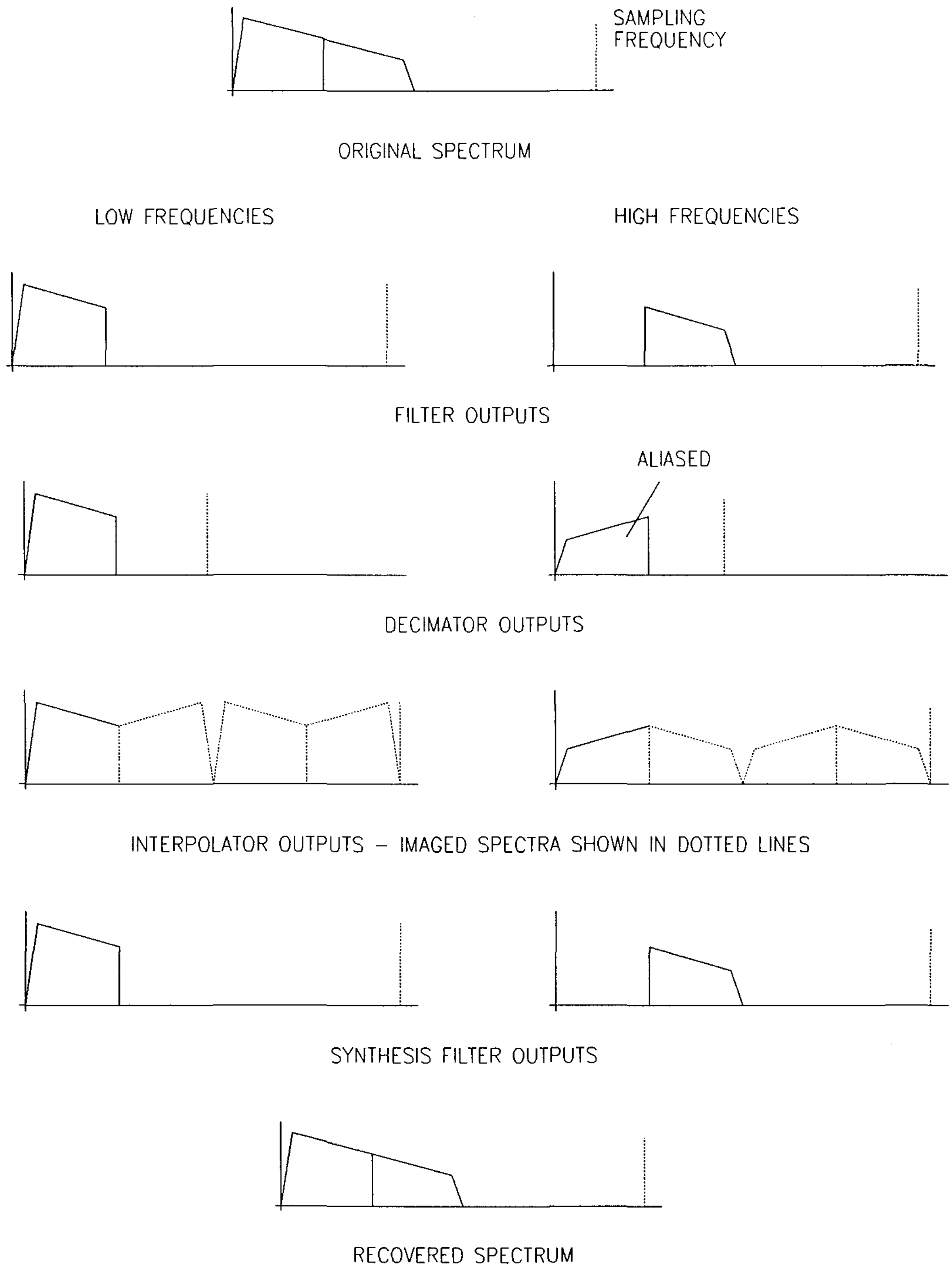


FIG. 8 SUBBAND CODING SPECTRA

ACHIEVING FIRST GENERATION NTSC QUALITY IN A RECORDABLE LASER DISK COMMERCIAL INSERTION SYSTEM

By Dominic D'Allessandro and Gilbert Moreira, Arvis Video Information Systems

C. Robert Paulson, AVP Communication

Abstract

Recent introductions of recordable and ***track-erasable*** *(TE) laser disks have improved the versatility, storage capacity, operational flexibility, applications and image and sound reproduction quality of commercial insertion systems by orders of magnitude over the original 3/4-inch videotape cassette systems. Image quality which is subjectively the equivalent of 1-inch Type C recordings results from the FM recording of the chrominance signal as line-sequential C_R and C_B, and FM recording of the luminance signal at a higher center frequency.*

Incorporation of TE disk drives in these systems will also dramatically reduce their Total Life Cost (TLC). Their applications can be extended to include storage and playback of ***program*** *material in cablecasting and broadcasting, and a variety of image, sound and data material archived in libraries, schools, museums, MIS systems, government data bases, etc.*

Use of a Component Analog Video (CAV) videotape recording format, such as S-VHS (S-Video), for shooting and postproduction editing of original footage before transfer to either a WORM (Write Once, Read Much (or Many - times) laser disk, enables each playback of a segment from the automation system to be ***first-generation NTSC!***

The 1980s evolution of automated commercial insertion systems

When local cable system operators first began to generate income by inserting local commercials into their programming, only 3/4-inch videocassette recording and playback equipment would both fit their budgets and deliver a minimum acceptable level of picture and sound quality. This format had become the de facto standard for broadcast industry ENG (Electronic News Gathering) in the mid 1970s. Because SMPTE (the Society of Motion Picture and Television Engineers) had subsequently documented its specifications[1], it was a "safe" format for the cable industry to adopt as a de facto standard for commercial insertion systems. Both recording and playback equipment and videotape cassettes were available from competing vendors with guaranteed interchangeability .

During the 1980s, signal output quality was regularly improved through upgrading of recording head, tape and processing electronics designs. However, the limit of these improvements is imposed both by the 1969-introduced electro-mechanical design, and the recorded video and audio footprint on the ferrous oxide tape frozen in the above-referenced "ANSI/SMPTE Type E" US NTSC 3/4 U standards.

On-line spot commercial insertions into cable channels began as a process of manually cueing and rolling spots from "day reels" on a single playback deck. By 1989, however, the commercial insertion process was typically accomplished on computer-controlled automation systems containing upwards of four decks per network. Its software can manage the tightly-

[1] SMPTE 21M-1986, Video Recording--3/4-in Type E Helical Scan--Records
SMPTE 22M-1986, Video Recording--3/4-in Type E Helical Scan--Cassette
SMPTE 31M-1989, Television Analog Recording--3/4-in Type E--Small Video Cassette

scheduled playback of hundreds of on-line spots to multiple active channels, initiating each commercial break either against a master control clock command, or by responding to cues embedded in the current on-line program. It can also select playbacks from external sources.

On-line system functioning could be controlled by a separate traffic department computer, issuing cue and play commands against insertion orders stored in its data bank. Optional software in the automation system computer today can also monitor the playback system's state of health, shuffle playback sequences to prevent dead air, switch defective tape decks off line and sound alarms, generate logs, command the printing of invoices in a billing department computer, and schedule and automatically run make goods.

Constant software upgrading driven by competitive pressures insures that current automation systems will keep pace with 1990s-emerging demands for more and more operations sophistication. However, the same cannot be said of cable operators' ***and viewers'*** continually more critical demands for improvements in received signal quality. Further, the 1990s will see a constantly growing need to maintain larger and larger quantities of spots on line, for computer-directed access against clients' rapidly changing scheduling needs.

Five concerns about the current 3/4-inch VCR systems are being voiced by cable operators:

1 - Insufficient on-line storage capacity to accommodate future needs, made worse by software limitations on capacity expansion by adding tape decks;

2 - Limits on picture and sound signal quality improvements to keep pace with the constantly improving signal quality being distributed by the cable networks and syndicated program vendors;

3 - Quality deterioration which results both from multiple generation dubbing of NTSC format recordings and repeated tape shuttling and playbacks;

4 - Tape shuttle time limitations, which often preclude back to back scheduling of selected spots except by manual "podding" (re-recording them at a new tape location);

5 - Tape deck maintenance costs required to meet manufacturer-prescribed schedules for regular replacement of mechanical components, at the risk of on-line failures and resulting revenue loss, advertiser unhappiness and viewer irritation.

1990s commercial insertion systems features

Arvis 7000 series system features

Automated head-end operations software developed by Arvis Video Information Systems during a decade of service to the cable industry eliminates the first two cablecasters' concerns listed above. Arvis also offers two recording system hardware options in new systems or system add-ons, which eliminate the first concern without adding tape decks, and eliminate the quality ceiling, quality deterioration, scheduling flexibility, maintenance cost and system reliability problems which in fact characterize the four remaining concerns.

Features in the current generation of Arvis playback automation software provide access, cueing and and glitch-free playback of spot commercials stored on up to sixteen on-line cassette tape decks and laser disk drives. These can include both the 3/4-inch decks in an existing Arvis system ***plus*** a mixture of added high-density sources, including both S-VHS format playback-only tape decks and laser disk players. The on-line storage capacity and worst-case access times of both of these formats are substantially better than the 3/4-inch format. Therefore, adding either to an existing system results in a reduction of the number of 3/4-inch decks needed on line.

Typically, a single laser disk drive can replace at least three and sometimes four 3/4-inch decks. Two or three S-VHS decks can replace four 3/4-inch decks. These expansions dramatically improve system operating capabilities and reliability, while substantially reducing labor and spare parts costs now required for 3/4-inch system maintenance.

S-VHS format features and benefits[2]

Several 1/2-inch professional video cassette formats introduced in recent years have potential as de facto standard successors to 3/4-inch systems, to eliminate the last four cablecaster concerns. They all offer offer substantial improvements over 3/4-inch in bandwidth (resolution), color fidelity, noise margin, tape deck shuttle speeds, tape handling and tape life. They all have image and sound quality specifications approaching 1-inch Type C recorders. The component analog professional S-VHS format is of particular interest to cablecasters because it offers these quality advantages in economically priced hardware. It further offers substantial increases in on-line capacity without tape deck additions, and reduced access time, which increases scheduling flexibility.

S-VHS cassettes hold 120 minutes of tape in the smaller, simpler and lower-cost consumer VHS cassette. This cassette provides twice the storage capacity of the 60-minute 3/4U cassette, doubling the inventory of spots which can be stored on line in any size of automation system without any increase in worst case access time. Further, the S-VHS tape transport shuttles at twice the 3/4-inch tape deck speed, providing access to over 40 spots in 30 seconds, compared to 12 in a 3/4-inch system. On any tape system, however, the available tape length cannot be filled with recorded spots. Even the fastest-shuttling tape deck requires five to ten seconds of separation between recordings (16 to 33 percent of maximum available tape recording time) for pre-roll, and as a margin of safety against tape cueing errors and running out of tail-end black.

S-VHS signal processing electronics provide improvements over 3/4-inch in both image and sound quality. Video is recorded and reproduced as a two-channel "Y/C 3.58" signal (also referred to as "S-Video"), eliminating cross color and cross luminance artifacts common to 3/4-inch NTSC playbacks. Luminance resolution is maintained at 400 lines, compared to 240 lines in 3/4-inch. Video signal input and output interfaces are both NTSC and S-Video. This makes it possible to shoot, transmit, record, edit and rerecord images without ever losing the bandwidth or adding the noise created by signal processing in NTSC.

The S-VHS format provides space on the video track for recording two FM-quality signals synchronous with video, leaving the two standard linear audio tracks available for time code, audio cues and housekeeping data. A digital bit-serial tape transport control signal port provides straight-forward interfacing to standard automation system computers and programmable controllers in either EIA-232C or EIA-422 standards. S-VHS automation systems can therefore be programmed to maintain larger on-line spot inventories for playback on larger numbers of channels.

Laser disk format features and benefits

If laser video disks are used for on-line spot storage and playback, the tape's unusable storage space required for pre-roll is eliminated, because the readout device is a non-contact laser laser subsystem which "flies" above the disk surface. Each spot plays back instantly on demand without a pre-roll cue. Each disk holds 50 30-second spots on each side. Hundreds of spots or the entire active library can be stored on line in a multi-drive system.

Worst-case access time to a spot on the other side of a disk currently being played is less than twenty seconds. Access to another spot on the same side is under one second. The Arvis laser disk system can therefore handle commercial break schedules calling for any reasonable combination of five and ten second promos and IDs intermixed with fifteen and thirty second spots.

If the spots are recorded direct from 1-inch tape or film, image and sound quality are qualitatively identical to the quality of cable network programs which are typically shot and edited on film or one-inch tape.

Further, there is no quality deterioration caused by shuttling and cueing and unlimited playbacks, because there is no physical contact between the laser reading head and the recording surface. Image and sound quality do not de-

[2] SMPTE standards-drafting committees are now working on the upgrading of the existing VHS standards documents, including the addition of the S-VHS (S-Video) format, at the request of one of the S-VHS equipment manufacturers.

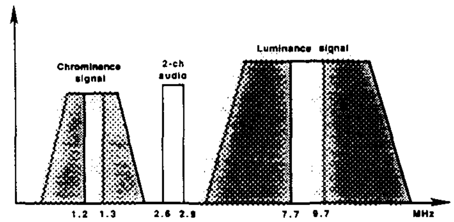
disk and its cassette package, reduce both the physical size of the automation system and the lineal shelf space required for off-line storage.

Recordable laser disk features and benefits

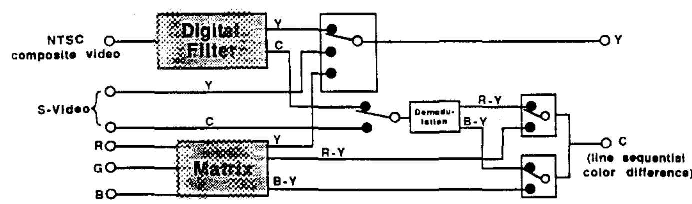
Using laser disk drives in a cable head end commercial insertion system would be operationally and economically unjustifiable, had not the capability to record intermittently on a new-design 12-inch (30 cm) laser disk become a reality. This first level of improvement over the capabilities of the original playback-only Video Disk systems introduced in the mid 1970s is popularly described as "WORM," -- Write Once, Read Much (or Many -- times). Arvis's first generation of **OptiCaster**™ 7000 Series Laser Disk Commercial Insertion Systems uses a WORM laser disk recorder available from Panasonic and Teac to record program materials and spot commercials off line. Playback-only drives in the automation system reproduce the signal in NTSC with more than 450 TV lines of luminance resolution and more than 45 dB chroma Signal to Noise Ratio (SNR). Two discrete channels of audio have better than 70 dB dynamic range at a bandwidth of 20 Hz to 20 kHz.

This substantial performance quality benefit over the 3/4-inch format's color-under NTSC is achieved by recording the input video signal in "Components." Separate frequency bands in the nominal 15 MHz recording bandwidth spectrum are allocated to the chrominance signal, each of two program audio signals and the luminance signal. The two color difference signal chrominance components are time-processed for line-sequential C_R and C_B recording as an FM signal. The video input signal to the recorder may be either NTSC, S-Video (Y/C 3.58 analog components), analog RGB, or a direct dub of an S-Video signal from the playback pre-amp of another drive.

Judicious allocation of spot commercials, IDs, PSAs and promos to WORM disks minimizes the number of disks that must be changed out as the inventory grows and changes over time. Separate disks can be allocated to storage of seasonal items. Individual clients whose spots would never be played back to back, such as competing automobile dealers, can be grouped on two or more disks which are kept on-line for long periods. Experienced Arvis operations analysts are available to provide each automation system operator with a customized recording allocations policy and procedure.



A. Recorded signal frequency spectrum



B. Video input signal block diagram

Video Resolution	More than 450 TV lines
Chroma S/N ratio	More than 45 dB
Audio response	20 Hz to 20 KHz
Audio dynamic range	More than 70 dB
Video inputs/outputs	NTSC, S-Video, RGB, Y/C dub
Operational control	EIA 232 C, EIA 422 optional

C. Performance and control specifications

Panasonic recordable laser disk functioning and specifications

Track-erasable/re-recordable laser disk features and benefits

At the 1990 NAB Conference and Exhibit held also in Atlanta, several laser disk drive manufacturers debuted production models of ***track-erasable*** laser disk recorders and companion players. Their appearance is too recent to have resulted in a single popular acronym to describe them, but for brevity they can be referred to as "TE" laser drives and disks.

These units provide a commercial insertion system with the storage flexibility and spot replacement benefit of an all-tape configuration, plus all the above-listed capacity and quality advantages of laser disk drives.

TM A service mark of ARVIS Information Systems, Waltham MA.

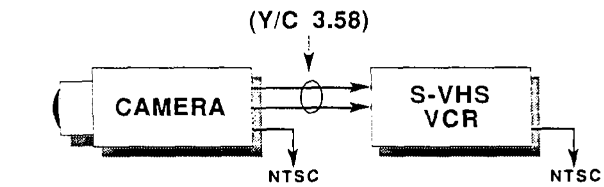
Arvis is debuting this newest commercial insertion system option at this NCTA conference. As quickly as production models of these TE drives are priced and available, the OptiCaster 7000 Series family will include them. Agreements are being worked out with several drive suppliers to provide trade-in options to near-term purchasers of WORM systems whose long-term needs can better be satisfied by a mixture of tape, WORM and TE systems.

Achieving first generation NTSC quality in spot production and playback (See block diagram)

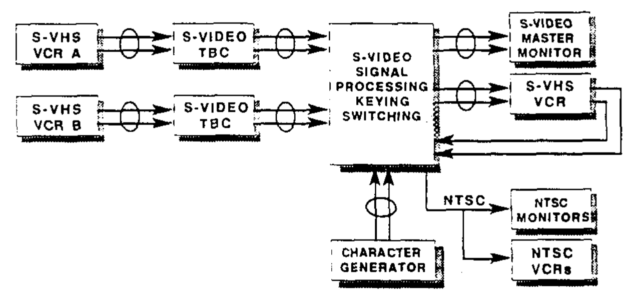
Editing and dubbing in the Y/C 3.58 format common to the laser disk drives or S-VHS tape decks eliminates the video signal degradation so objectionable in editing and dubbing in the 3/4-inch NTSC color-under format. Quality-minded cable operators can completely eliminate this degradation by standardizing on S-VHS equipment in both production shooting and postproduction editing systems.

Many portable and studio camera manufacturers now provide both NTSC and S-Video outputs. All S-VHS tape recorders have input connectors to accept an S-Video feed from these cameras. Their S-Video outputs can be fed to a wide selection of video switchers and special effects units now also equipped with S-Video Input/Outputs. Many monitor manufacturers also market high-resolution master monitors equipped with S-Video inputs. Therefore, an edited master tape can now be produced without ever recording or transporting or viewing the video signal in its composite analog NTSC form.

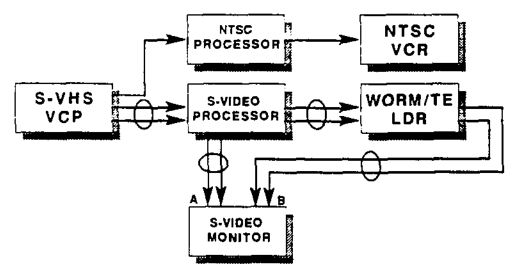
The S-VHS tape decks and laser disk recorders marketed by Arvis in their 7000 Series Commercial Insertion systems are similarly equipped with both NTSC and S-Video inputs. If the edited master tape has been created in an all-Component Analog Video (CAV) environment, then any spot resident on an S-VHS tape or laser disk on-line in the Arvis system is truly "zero-th" NTSC generation. Thus its playback through an analog component NTSC Time Base Corrector (TBC) or NTSC encoder creates a ***first generation NTSC*** signal each time it is played. To the home viewer, the quality of this signal will be indistinguishable from any cable network feed into which it is inserted!



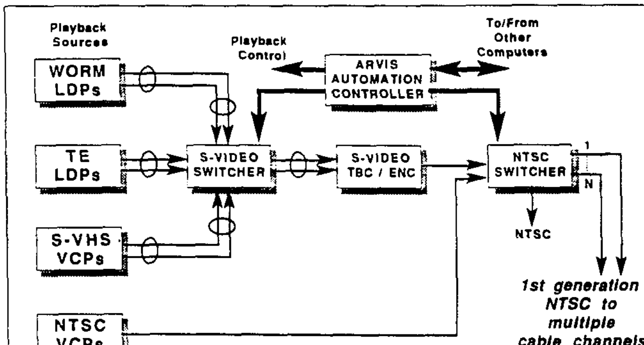
A. CAV (Y/C 3.58) shooting system



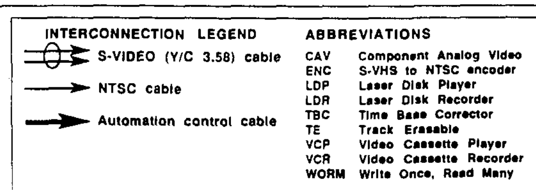
B. CAV postprodution editing system



C. S-VHS master dubbing system



D. ARVIS 7000 series commercial insertion system



Economics of S-VHS tape and WORM and TE laser disk formats

Pricing of both 3/4-inch and S-VHS recording and playback decks and blank cassettes is currently in a state of turmoil. Pricing of WORM laser disk drives and disks has been relatively stable for some time. The recent announcements of TE drive and disk availability, features and specifications have been accompanied only by "Preliminary" pricing. These numbers are in general twice or more times higher than those currently quoted for WORM drives and disks.

It is rumored that TE pricing will fall quickly if the product generates early buying enthusiasm. This will trigger significant decreases in WORM pricing. Further, S-VHS pricing may be forced downward as marketing of the HI-8 8 mm component analog format steps up.

Arvis is developing "total life cost" comparison information on the alternative integration of S-VHS, WORM and TE playback sources into existing and new 7000 Series Commercial Insertion systems. These will be available from Arvis as soon as possible in one or a series of economic considerations bulletins.

ADVANCED STATUS MONITORING AND CONTROL OF FIBER OPTIC SYSTEMS

John Iles and Dennis Donnelly
Orchard Communications, Inc.

Abstract

Advances in fiber optic systems now allow the development of new system architectures, frequently utilizing combinations of modulation formats and multiplexed wavelengths to allow optical signals to be carried deep into the system. As these systems become more complex and to insure the reliability that the system promises, the CATV operator requires a method to monitor and control the various system components and operating parameters. This paper describes such an integrated system, MM-Net.

INTRODUCTION

Fiber optic systems are rapidly becoming one of the most effective tools used by CATV operators to provide improved quality and reliability within the subscriber delivery system. From super trunking between headends to the latest uses of AM systems to extend optical signals deeper into the system, fiber is rapidly integrating with the other elements of the CATV system.

One of the key reasons for incorporating this new technology is increased reliability. A single fiber optic system can frequently replace a dozen or more amplifiers, thereby reducing the number of system components that might cause a subscriber outage. While reliability may increase because of a numerical reduction in the system, that alone does not guarantee the operator a good night's sleep.

Unfortunately, fiber optic systems are still built out of the same parts as any other electronic device, namely power supplies, ICs and transistors, capacitors and resistors. While the life of the laser and detector may be long, the rest of these parts have lives of their own. Any one can be the source of failure, just as it could have been in the amplifier cascades the optical systems have replaced.

As the architecture of a typical cable system expands to encompass more fiber plant, the operator faces a real challenge in finding a suitable method of monitoring the fiber system to assure high standards of reliability.

At the same time, consolidation of headends and changes in programming methods may also require the operator to have control over programs and channels sent to various areas of the system or even to different towns that are now linked by fiber.

The purpose of this paper is to present the details of a system that provides both status monitoring capability and control of many of the individual pieces of the system.

Of course, neither status monitoring nor system control is new to fiber optic systems. Many operators today utilize one of the status monitoring systems currently available to keep tabs on their system.

However, fiber systems may present a different set of problems. In the near future, an operator could have a system that encompasses several different types of modulation schemes over fiber and maybe even several wavelengths of light on the same fiber.

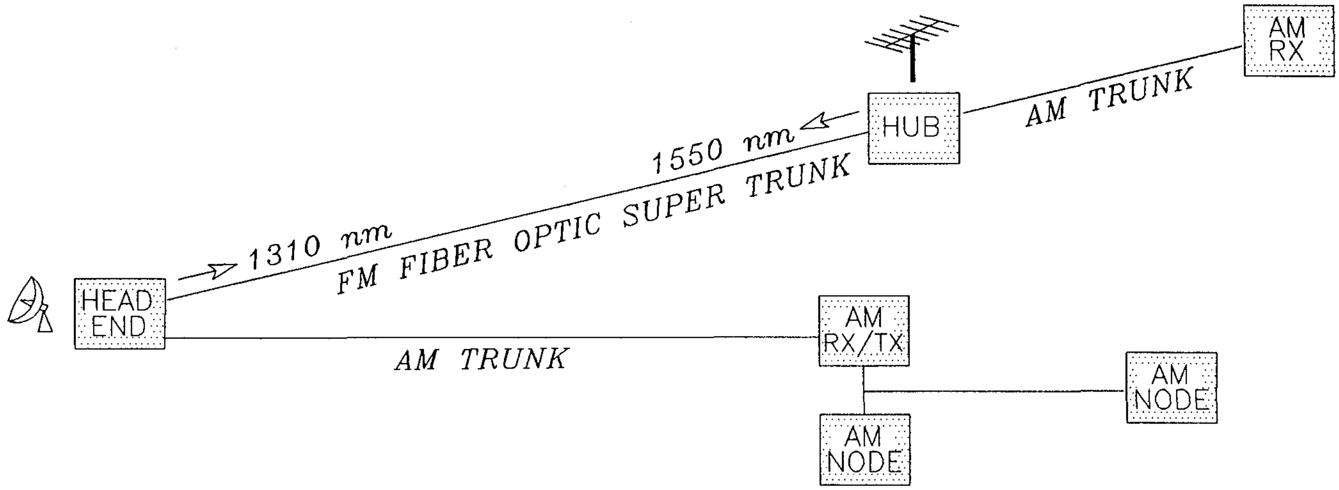


Figure 1 CATV Fiber Optic System

Figure 1 illustrates what a future cable system might look like, containing FM modulated trunking from the main headend to a hub in another town, AM modulated trunking using the new externally modulated AM system, and standard AM modulated fiber nodes. The FM portion uses 1310 nm and 1550 nm signals, wavelength division multiplexing both over the same fibers. Because of the distance, there is also a repeater in the super trunking system.

In developing MM-Net, it was necessary to plan ahead, so that not only would it take care of today's needs, but would also look forward to systems like the one shown in Figure 1.

WHAT MM-Net DOES

To borrow some words from the computer industry, MM-Net is an interactive status monitoring and control system that operates as a token ring network. Within the constraints of current personal computer technology, the system has been designed to be simple to operate. The system is menu driven, with all screens in plain English (Figure 2). It can be learned quickly and easily.

What does MM-Net do? The system diagram in Figure 1 illustrates some of the tasks that it will accomplish.

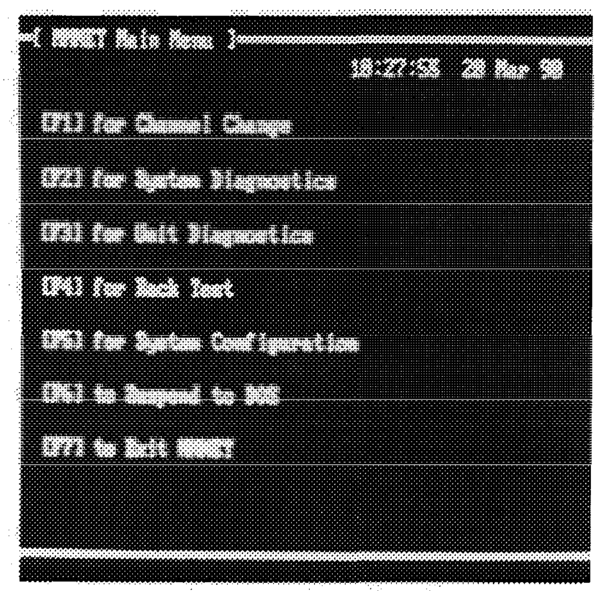


Figure 2 MM-Net is Menu Driven and Easily Mastered

System Diagnostics

To start, when the system is activated, the software will poll each location to find out what kind of hardware resides there. From this information, it develops an inventory of all the units in the system by location (Figure 3). By performing this task, the software automatically learns the configuration of the system.

While the software identifies each piece of equipment by a unique address, it also knows what kind of equipment it is, such as a transmitter, receiver, modulator, etc., and provides that information along with the address. An editor built into the software allows the operator to assign a plain English name to the unit. For instance, the system might identify address 0 as a transmitter. For easier identification by personnel, the unit could be named Headend TX 1. That way, each unit can be identified in the simplest and most direct terms possible.

Once the software has "learned" the system configuration, it is ready for operation. Using the system for diagnostics or status monitoring is as easy as selecting the Automatic Diagnostics feature on the main menu. In this mode, the software continuously polls the entire system, so that any change of status in any piece of equipment will provide an immediate alert to the operator that there is a problem.

The system identifies the equipment experiencing the problem by both location number and name. Using the Manual Diagnostics menu selection, the operator can ask the software to display, in detail, what fault has occurred (Figure 4). Therefore, the operator knows what has happened and where before dispatching a repair truck to the site.

Because the system has been designed with an eye to the future, the software functions not only with existing FM technology, but is also compatible with emerging technologies, such as the new externally modulated AM schemes or digital systems. In each

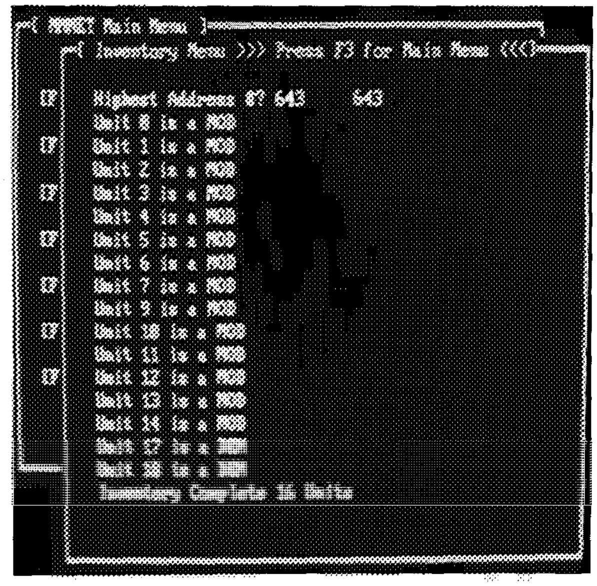


Figure 3 Screen Displaying System Inventory

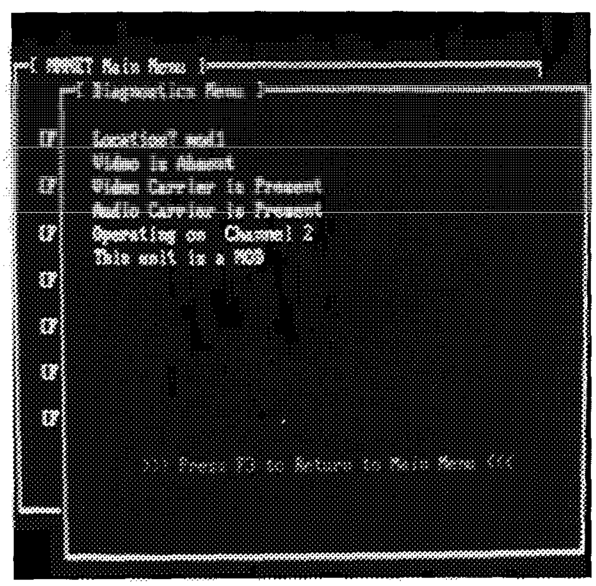


Figure 4 Screen Displaying Diagnositics for a Unit

case the necessary parameters of the equipment are measured and accurately reported.

Diagnostics, though an important function of the system, is only one of the tasks the software can accomplish.

FM Channel Selection

In addition, it is possible to assign channel selections for the FM modulation equipment at either headend or hub locations. This flexibility in selecting channels allows the program lineup to be varied by area or town. For instance, to allow a town council meeting of local origination to be shown only in the areas interested and not system wide. This feature also allows spare equipment to be assigned in the event of a failure, or as a replacement unit during routine maintenance.

Scheduled Automatic Switching

Another feature allows for scheduled automatic switching. This feature allows advance scheduling of channel changes that may be required as a result of shared services, SYNDEX, or other reasons. The scheduled switch times may be entered in the system for multiple events over any time frame from 1 minute to well over a year.

In addition to the automatic system, the operator has control over channel selection and switching on an instantaneous basis at any time.

Additional Features

Access to MM-Net is password protected. In addition, the system can be set up to provide varying levels of access, for example, allowing system maintenance but not scheduling or switching activity.

All commands issued by the MM-Net software are verified by a return signal from the equipment, notifying the master controller that the event actually occurred. This can be critical in areas such as channel selection or scheduled switching. All information can be routed to a printer to provide a permanent record of daily events.

In addition to direct control from a master computer, the system can be interfaced to a modem and controlled from a remote location over conventional telephone lines. A feature such as this allows personnel not on location, but authorized to access the system, to obtain immediate information on system performance and status. It also allows corrections and switching information to be entered from a remote location.

HOW MM-Net DOES IT

The MM-Net system is a small software package, designed to operate from an AT-class personal computer that has a color monitor, 640K of memory, a serial data port and a parallel port if a printer is used. In short, a simple computer.

While the computer need not be dedicated to the MM-Net system, the use of the automatic diagnostic feature does demand a dedicated computer. The complexity of such a system certainly makes a dedicated machine desirable.

What is this computer and its software communicating with? Beginning with the individual pieces of equipment, each unit in the system contains a microprocessor capable of monitoring up to 8 analog or digital inputs. Analog measurements are converted to an 8 bit approximation and are compared to an average parameter to reduce the volume of data traffic on the system. As long as the approximation stays within a window, no communication is required.

Each unit, designated a transponder, is capable of 8 ON/OFF control outputs. The transponder contains non-volatile memory, so that in the event of a loss of power, the system parameters are immediately restored when power is reapplied.

The unique address for each unit is field programmable using a DIP switch located on the unit's processor card. This switch is located inside the unit to prevent unauthorized or accidental changes.

Each transponder, in a group of up to 32 units, reports to a group controller device. The group controller communicates with its respective group of transponders over an EIA-485 interface operating at 9600 baud. Each group controller also has a unique address. Group controllers at any one location are chained to provide a single input/output port. This port communicates with the master PC through a full duplex modem. Signals for the modem can be carried by the fiber optic system or by conventional telephone lines.

The basics of the system network provide for communication using an EIA-485 communications interface, operating at 9600 baud. The system is capable of addressing up to 65,504 unique addresses.

The network controller is capable of issuing commands to the entire system, to an individual group controller, or to an individual transponder. In practical terms, the system operates as a token ring. Status requests originate at the controller and are sent to all group controllers in the system. If the status of the group controller's transponders is normal, the system continues to the next group controller for interrogation. If any group controller responds indicating an abnormal indication, then all of the transponders in that group are polled to find out where the abnormality is. When there is no abnormality in the system, the information flows from the controller in a ring. The ring is opened when a group controller indicates an abnormality and that information is added to the flow of information as it is returned to the network controller.

The total time for a command/reply sequence is approximately 25 milliseconds. In an average fiber system, the poll time would be less than 10 seconds. In a system using all 65,000 addresses, the full poll cycle time would be less than 1 minute.

Interfacing Other Equipment

In addition to the equipment and system described, MM-Net has a general purpose interface (MGPI), developed to provide input/output control and monitoring for various types of devices in the system, such as power supplies with status monitoring capability.

An external unit as described will appear to the MM-Net system as a normally addressable device. The interface outputs provide a range of standard input/output protocols, including EIA-232C, EIA-485, 16 bit parallel control lines and on/off contacts. Most devices in a cable system are capable of operating within the parameters of one of these protocols.

SUMMARY

As can be seen, the use of fiber optics to enhance reliability in cable systems can be improved through the use of a high quality, computer controlled diagnostic and management system.

The system described can provide full automatic diagnostics of all elements within the system, regardless of the modulation scheme being used, providing the basis for both current and future equipment additions.

In addition to the automatic diagnostic features of the system, MM-Net provides for scheduled program switching and allows immediate operator intervention to override commands or create late changes in scheduling.

A general purpose interface provides for integration of other devices that use status monitoring or command controls. This allows management of multiple pieces of equipment within one software system.

AN ECONOMICAL HIGH SPEED SATELLITE DATA TRANSMISSION SYSTEM

Charles M. White and Clyde Robbins
General Instrument, Jerrold Communications Div.

ABSTRACT

This paper will present and review the techniques and considerations used for delivery of 28 channels of CD quality digital audio via one satellite transponder. Subjects include bandwidth, tolerable bit error rates, forward error correction, and carrier to noise ratios.

INTRODUCTION

The design and implementation of a high speed satellite data transmission system requires optimizing all the parts of the system to achieve good, reliable performance as well as economy. The satellite distribution system used for Jerrold's "Digital Cable Radio" (TM) service was developed with these goals in mind. Developing a satellite distribution system requires selecting data rates, data format, error correction scheme, allowable bit error rate, modulation scheme, and channel bandwidth. Each of these factors is influenced by the others.

"Digital Cable Radio" service distribution requires the distribution of 28 digitally encrypted audio channels in a single satellite transponder channel. Each audio channel consists of a stereo audio pair plus control data. In order to fit all this information into a single satellite transponder bandwidth of 30 MHz, data compression and an efficient scheme of modulation is required. The reason for using a minimum number of transponder channels is cost. Like any scarce resource, the leasing of satellite channels is very expensive. The use of multiple satellite channels would also significantly increase the costs at the CATV head end. Finally, it has to be recognized that it costs money to get high Carrier to Noise ratios on the receiver end. This cost factor is due primarily to the need for larger Satellite Receiver Dishes to obtain higher C/N ratios. The "C" band was chosen because it affords adequate C/N (>12 dB) with the prevailing 3.5 meter dishes and good resistance to fading from rain and clouds. In order to minimize terrestrial interference, the data is transmitted in the center 20 MHz.

SYSTEM

Although this paper is primarily about the Satellite Link itself, the characteristics of the rest of the components of the "Digital Cable Radio" system set much of the link requirements. Therefore I will briefly describe the "Digital Cable Radio" system ("DCR" (TM) for short).

The DCR system consists of the following main blocks:

1 Source Material
2 Encoding and Multiplexing
3 QPSK Modulator
4 Satellite Transmitter
5 Satellite Receiver
6 QPSK Demodulator
7 Head End Transcoder
8 CATV Distribution System
9 DCR Receiver
10 Home Stereo System

The source material consists of CD disks and simulcast channels from premium CATV services. These audio sources go to Dolby (TM) Encoders which use adaptive delta modulation to maintain CD quality audio (95 dB Dynamic range, 20 Hz to 20 KHz response) with a data rate of 294 K bits per second per stereo pair.

Forward error correction is used to increase the corrected error rate by 3 orders of magnitude (from 1E-3 to 1E-6). This involves adding additional bits to the data stream. This plus the encryption and the additional data brings the channel data rate to 694 K bits per second. 28 Channels plus framing and sync bits gives a net data rate of 29 M bits per second. The use of error correction allows for a higher tolerable Bit Error Rate at the Downlink and thus a lower required C/N at the Downlink. Listening tests have indicated that in this format, a corrected Bit Error Rate (BER) of 1E-6 is imperceptible from the original CD. Thus, an uncorrected BER of 1E-3 is the requirement for the Link.

The QPSK Modulator is fed by two 14.5 M bits per second inputs (each the equivalent of 14 Audio Channels) into its "I" and "Q" data inputs. The QPSK Modulator produces a Quadrature Phase Shift Keyed (QPSK) signal on a 70 MHz carrier. This IF signal now contains the 29 M bit/s data stream in a bandwidth of 20 MHz. The QPSK modulator feeds the satellite uplink chain.

The Satellite relays the RF signal back down to the Downlink site. The receiver tunes to the RF signal, down converts to 70 MHz, AGC's the signal amplitude, and AFT's the output frequency.

In the Transcoder frame, the QPSK Demodulator recovers the original data stream and the data clock from the 70 MHz QPSK IF input. These signals are fed to the Processor and Channel Transmitters. The processor controls the Channel Transmitters. The Channel Transmitters pick off their individual channel data and modulate an individual CATV RF output channel (typically in the FM band). These individual channels (at different frequencies) are summed for distribution through the CATV System to the home DCR Receiver.

The DCR Receiver tunes to the desired channel and demodulates the RF to recover the channel data stream. Then, the channel data stream is decrypted and converted to audio. This audio output drives the subscriber's stereo and speakers.

QPSK MODULATOR

Differential QPSK Modulation encodes the data stream as phase changes to the carrier. For example, if both the I and Q data inputs were all zeros, then there would be no phase changes and a bare carrier would result. If both I and Q data inputs were all ones, then both would change every data period. The I and Q data is first differentially encoded so that the differential QPSK demodulator will correctly demodulate the data. Then, pre-emphasis is applied to the I and Q signals. The purpose of the pre-emphasis is to correct for the spectrum shape of rectangular random data (rolls off) and thus to "flatten" the resulting spectrum in order to optimize the transmitted signal energy. The demodulator has a corresponding roll-off filter to equalize the response. Then a 70 MHz carrier is QAM modulated by the differentially encoded and pre-emphasized I and Q signals. Next, the QPSK signal is band limited in a SAW filter and amplified to +3 dBm output level.

The reasons that Differential QPSK was chosen are that it is bandwidth efficient, can deliver less than 1E-3 BER for 12 dB C/N, is robust (tolerates compression and limiting), and can be demodulated in a cost effective manner.

SATELLITE RECEIVER

The main purpose of the Satellite Receiver is to provide amplification and frequency down conversion of the signal from the Satellite. The output of the Satellite Receiver is 70 MHz at 0 dBm. The tuner down converts to 130 MHz, and then this is down converted to 70 MHz. The second down conversion allows for AFT so that the final (most narrow) SAW filter on the QPSK Demodulator is centered on the IF signal. This allows for some drift and error in the LNB at the antenna. The AGC holds the signal amplitude constant so the QPSK Demodulator operates optimally. An AUX input is provided for a 70 MHz input from an existing LNA installation.

QPSK DEMODULATOR

The QPSK demodulator utilizes delay type differential QPSK demodulation. The result is relative design simplicity and virtual immunity to incidental phase noise. First the 70 MHz IF input is bandpass filtered and then mixed down to 29 MHz. The demodulation takes place at an IF frequency of 29 MHz. The demodulator also provides for AFT. This AFT mainly corrects for any slight drift or error in the delay lines. The detected I and Q signals are low pass filtered in a filter which also complements the pre-emphasis in the QPSK Modulator. Then, the data clock is recovered from the detected I and Q. This recovered data clock is then used to sample the detected I and Q to produce the recovered I and Q data.

RESULTS

Testing of uncorrected Bit Error Rate verses Carrier to noise has confirmed results approaching theoretical for C/N in the region of 12 dB. It has been shown that for C/N of 12 dB the uncorrected BER is less than 1E-3 and the corrected BER is less than 1E-6.

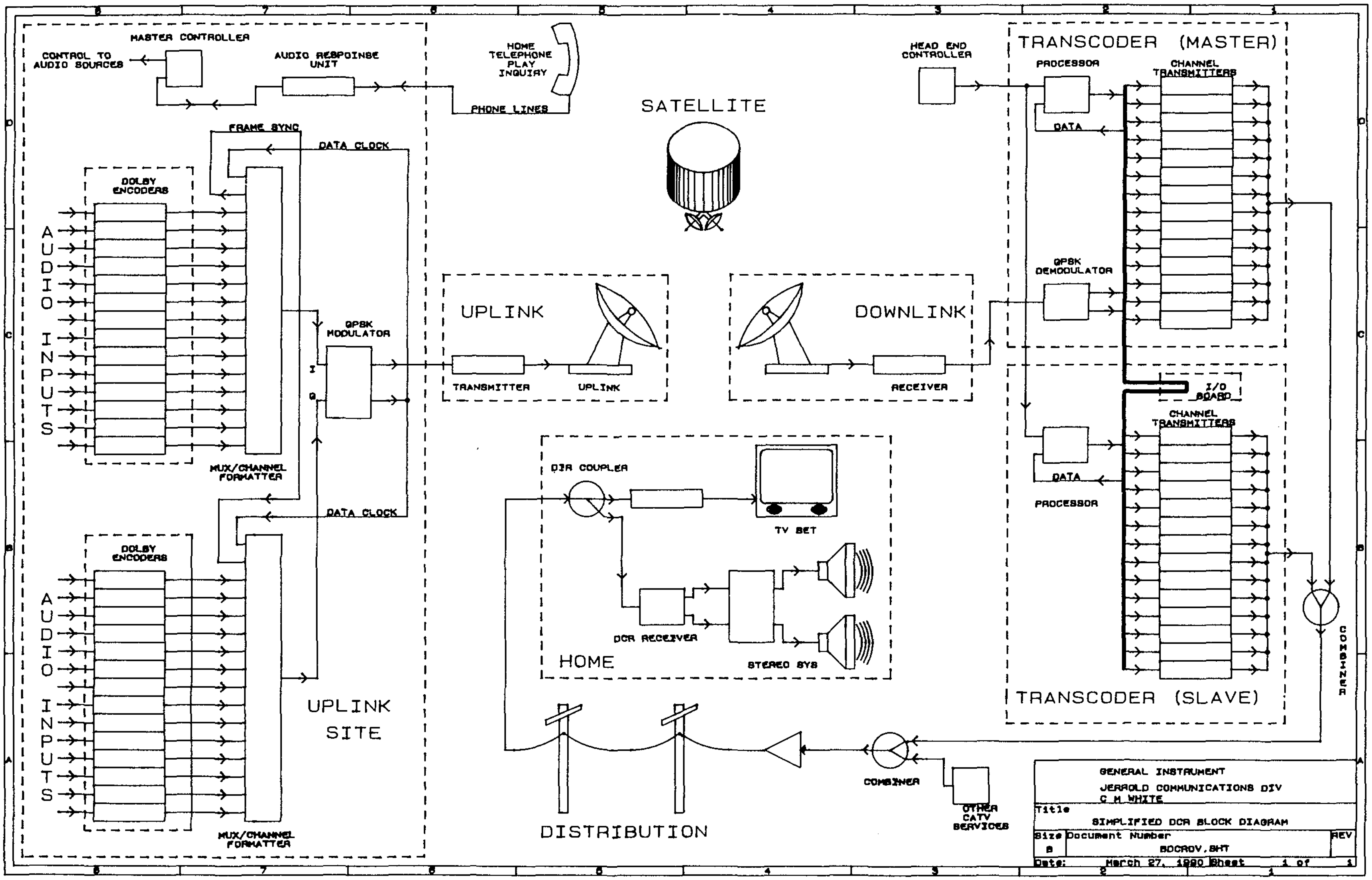
SUMMARY

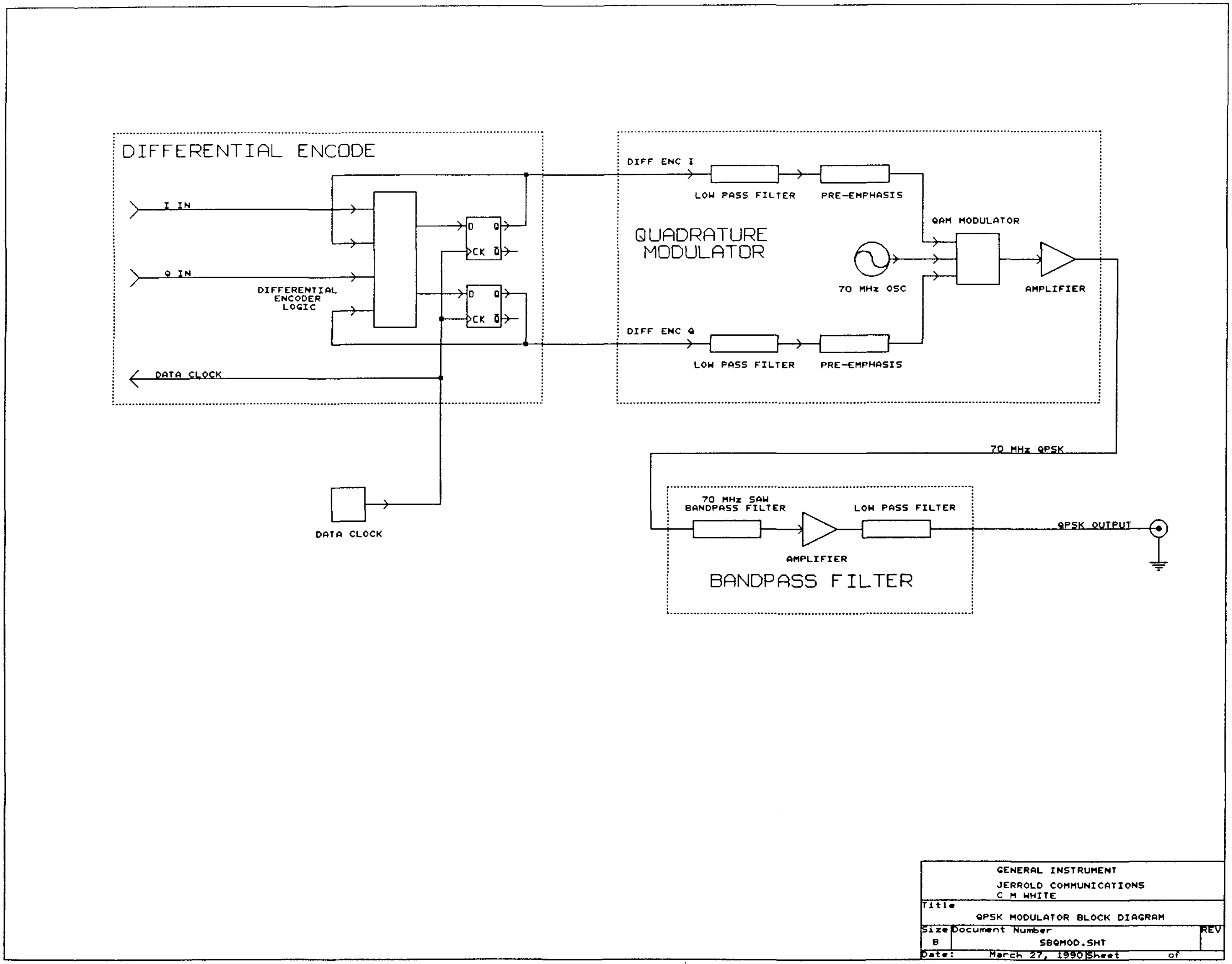
By the careful choice of data coding, forward error correction, modulation/demodulatiom scheme, and implementation it is possible to distribute 28 stereo audio channels over a satellite link while maintaining quality levels equivalent to the CD source. And, it is possible to do so in an affordable fashion.

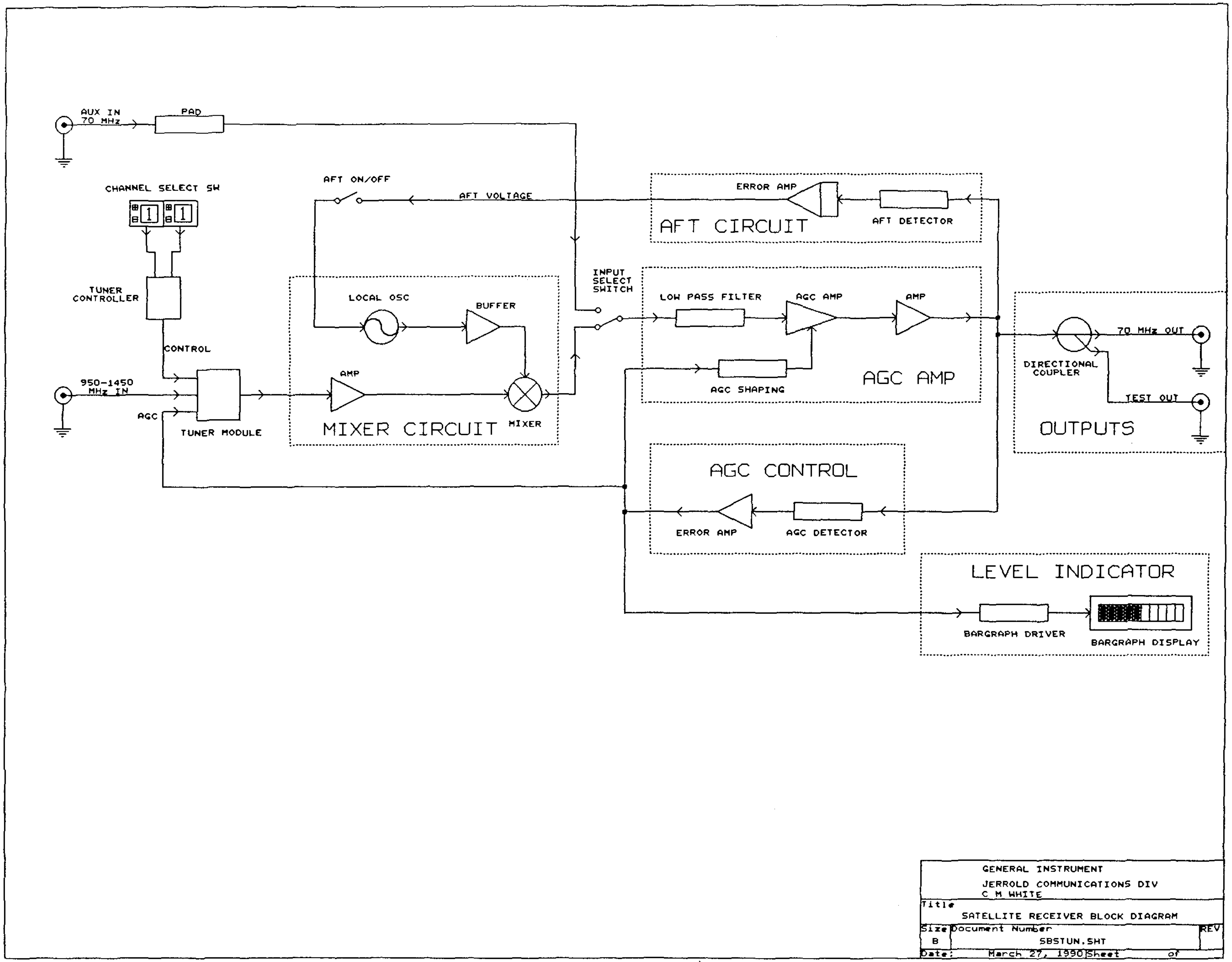
TRADEMARKS

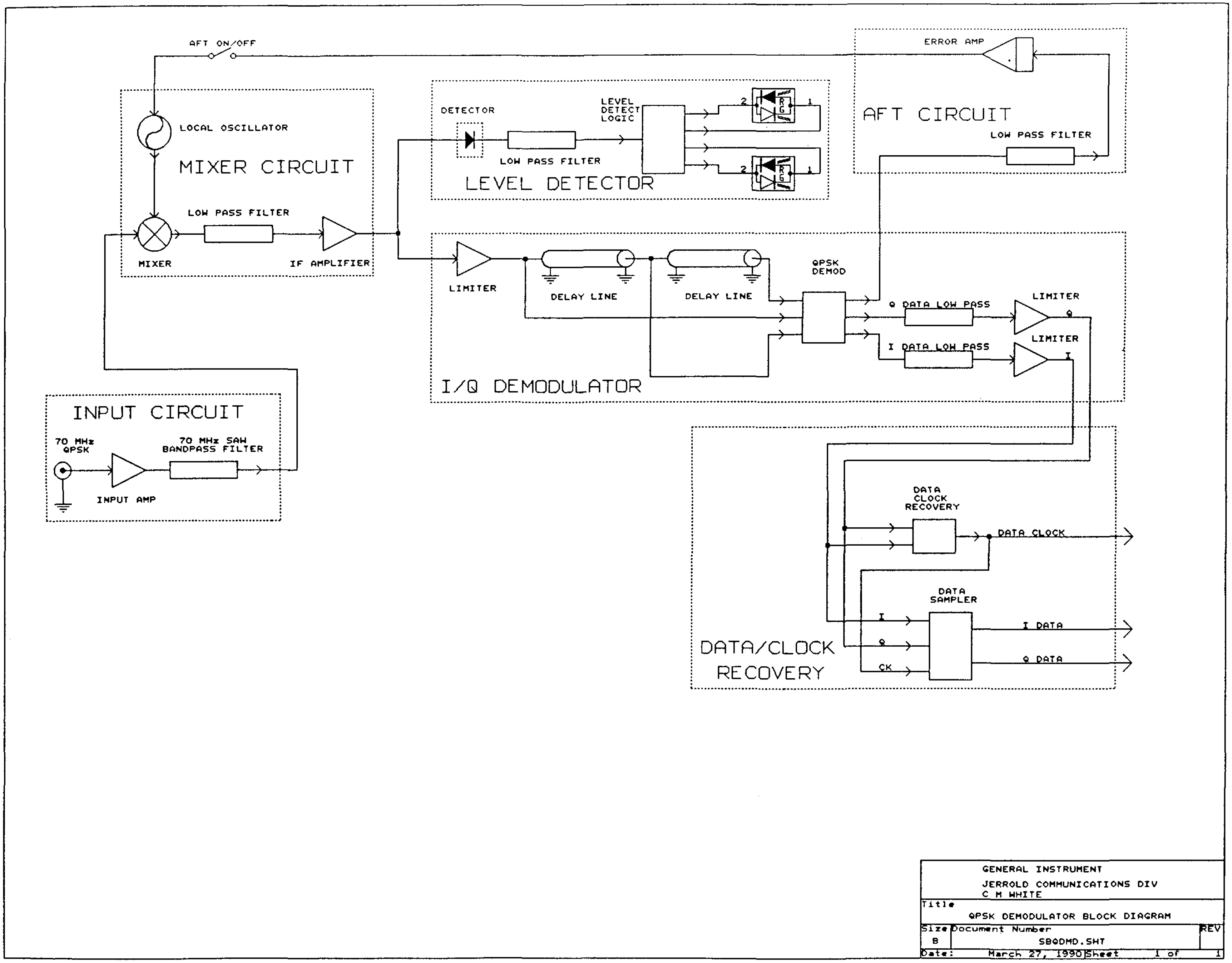
"Dolby" is a registered trademark of Dolby Laboratories.

"Digital Cable Radio" and "DCR" are registered trademarks of General Instrument Corporation, Jerrold Communications Division.


MASTER CONTROLLER
CONTROL TO AUDIO SOURCES
AUDIO RESPONSE UNIT
HOME TELEPHONE PLAY INQUIRY
PHONE LINES
SATELLITE
HEAD END CONTROLLER
TRANSCODER (MASTER)
PROCESSOR
CHANNEL TRANSMITTERS
DATA
FRAME SYNC
DATA CLOCK
DOLBY ENCODERS
AUDIO INPUTS
QPSK MODULATOR
I
Q
MUX/CHANNEL FORMATTER
UPLINK
TRANSMITTER
DOWNLINK
RECEIVER
QPSK DEMODULATOR
I/O BOARD
CHANNEL TRANSMITTERS
DATA
PROCESSOR
DIR COUPLER
TV SET
DCR RECEIVER
STEREO SYS
HOME
DATA CLOCK
DOLBY ENCODERS
AUDIO INPUTS
UPLINK SITE
MUX/CHANNEL FORMATTER
COMBINER
TRANSCODER (SLAVE)
DISTRIBUTION
COMBINER
OTHER CATV SERVICES
GENERAL INSTRUMENT
JERROLD COMMUNICATIONS DIV
C M WHITE
Title
SIMPLIFIED DCR BLOCK DIAGRAM
Size B
Document Number
BOCROV.SHT
REV
Date: March 27, 1990 Sheet 1 of 1


DIFFERENTIAL ENCODE
I IN
Q IN
DIFFERENTIAL ENCODER LOGIC
D Q
CK Q̄
DATA CLOCK
DATA CLOCK
QUADRATURE MODULATOR
DIFF ENC I
LOW PASS FILTER
PRE-EMPHASIS
DIFF ENC Q
LOW PASS FILTER
PRE-EMPHASIS
QAM MODULATOR
70 MHz OSC
AMPLIFIER
70 MHz QPSK
70 MHz SAW BANDPASS FILTER
AMPLIFIER
LOW PASS FILTER
BANDPASS FILTER
QPSK OUTPUT
GENERAL INSTRUMENT
JERROLD COMMUNICATIONS
C M WHITE
Title
QPSK MODULATOR BLOCK DIAGRAM
Size
B
Document Number
SBQMOD.SHT
REV
Date:
March 27, 1990
Sheet
of


AUX IN
70 MHz
PAD
CHANNEL SELECT SW
AFT ON/OFF
AFT VOLTAGE
ERROR AMP
AFT DETECTOR
AFT CIRCUIT
TUNER
CONTROLLER
LOCAL OSC
BUFFER
INPUT
SELECT
SWITCH
LOW PASS FILTER
AGC AMP
AMP
CONTROL
950-1450
MHz IN
AGC
AMP
TUNER MODULE
MIXER CIRCUIT
MIXER
AGC SHAPING
AGC AMP
DIRECTIONAL
COUPLER
70 MHz OUT
TEST OUT
OUTPUTS
AGC CONTROL
ERROR AMP
AGC DETECTOR
LEVEL INDICATOR
BARGRAPH DRIVER
BARGRAPH DISPLAY
GENERAL INSTRUMENT
JERROLD COMMUNICATIONS DIV
C M WHITE
Title
SATELLITE RECEIVER BLOCK DIAGRAM
Size
B
Document Number
SBSTUN.SHT
REV
Date:
March 27, 1990
Sheet
of


AFT ON/OFF
ERROR AMP
AFT CIRCUIT
LOW PASS FILTER
LOCAL OSCILLATOR
MIXER CIRCUIT
LOW PASS FILTER
MIXER
IF AMPLIFIER
DETECTOR
LEVEL DETECT LOGIC
LOW PASS FILTER
LEVEL DETECTOR
LIMITER
DELAY LINE
DELAY LINE
QPSK DEMOD
Q DATA LOW PASS
I DATA LOW PASS
LIMITER
LIMITER
I/Q DEMODULATOR
INPUT CIRCUIT
70 MHz QPSK
70 MHz SAW BANDPASS FILTER
INPUT AMP
DATA CLOCK RECOVERY
DATA CLOCK
DATA SAMPLER
I DATA
Q DATA
CK
DATA/CLOCK RECOVERY
GENERAL INSTRUMENT
JERROLD COMMUNICATIONS DIV
C M WHITE
Title
QPSK DEMODULATOR BLOCK DIAGRAM
Size B
Document Number SBQDMD.SHT
REV
Date: March 27, 1990
Sheet 1 of 1

CABLE HEAD END OPERATION WITH COPY-PROTECTED SIGNALS

Graham S. Stubbs
Executive Vice President
Eidak Corporation
Cambridge, MA

ABSTRACT

Pay Per View is emerging as a business capable of delivering programming which has substantial revenue potential from later markets. In order to preserve the value of later showings of program material, technology has been developed to inhibit taping of PPV signals. Copy-protected programming may be distributed by satellite or originated locally. In either case, some special provisions have to be made in the cable head end. This paper examines head end operation with Eidakized copy-protected video signals. Consideration is given to reception of copy-protected satellite signals, and to the origination of signals locally from tape or from copy-protected laser video discs.

Addressable scramblers used in cable head ends generally require some special attention when used to encode copy-protected video signals. Similarly, hub operation of PPV systems may require the use of special interfacing equipment. Following a discussion of the method of signal modification to achieve copy-protection, the paper describes the operation of head end equipment for various satellite and local origination distribution scenarios.

INTRODUCTION

The revenue potential for later markets for PPV programming has created a demand by program providers for some form of control over unauthorized copying of program material. Electronic copyright protection, already in use in a different form in the videocassette pre-recorded medium, is fast becoming a necessity to assure access to product, with timely availability, desired by cable operators to fuel the anticipated growth of PPV. Movies and some forms of live events have significant market revenue potential following a PPV exhibition, and thus are candidates for copy-protection of transmitted signals.

Electronic copy-protection is achieved by modification of the video signal in such a way that a program can still be readily displayed on a standard receiver or monitor, but an attempted recording using a video cassette recorder has no commercial or entertainment value. Copy-protection methods rely on differences in sensitivity on the part of VCR's and television receivers to modifications of the video waveform [1,2,3]. For the transmitted PPV signal, the Eidak technique uses modification of the television frame rate. It has been developed and tested specifically for operation in cable systems, with emphasis on security and compatibility with other equipment.

Projected application anticipates both satellite delivery to cable systems and standalone (locally originated) operation. The copy-protection process can be applied in any of the following ways:

- o At a satellite uplink--to the transmitted/encrypted signal.
- o At a cable head--to a signal received by satellite or originated locally on tape.
- o On a pre-recorded laser disc for use at a head end.

Because the copy-protected signal does not conform strictly to the NTSC 525 lines per frame standard, there are specific technical guidelines for proper head end operation. Reception of encrypted satellite signals, local program origination, and addressable scrambling all require attention when dealing with copy-protected programming.

THE COPY-PROTECTION METHOD[4,5,6,7,8]

The copy-protection method to be described exploits differences in the sensitivity of television receivers and VCR's to small changes in vertical frame rate. In particular, the electromechanical nature of the VCR causes it to be more sensitive to such disturbances. In order to most efficiently make use of the surface of magnetic recording tape, the VCR records video waveforms in stripes diagonally across the moving tape. Physically, this is accomplished by locating two or more recording heads on a rapidly spinning drum, around which the tape is wound in a helical fashion. At any one time only one recording head is actively recording and in contact with the tape. As the head moves diagonally across the tape it records one field of video. When one head reaches the edge of the tape a second head starts to record a next diagonal stripe corresponding to the next field. This operation is critically dependent upon careful synchronization of the rotational speed of the drum and the field rate of the video signal being recorded. A servomechanism system is used to achieve this precise synchronization.

The Eidak copy-protection technique varies the field rate in such a way that proper synchronization is upset, the servo loses lock, and the video signal is improperly applied to the recording tape. The effect on playback is to create gaps in the program material (i.e. goes to snow), onscreen artifacts due to non-synchronized head switching, and other video distortions. The variation in field rate is achieved by adding or deleting horizontal lines. For maximum effect, the technique is applied periodically as shown in Figure 1.

The all-electronic picture scanning system of the television receiver is able to respond properly to these variations in vertical scanning rate and thus provide a normal display. In order to maintain interlace and proper positioning of the displayed picture, horizontal lines are added or deleted in pairs, i.e. one line to or from each of the two fields in a frame. Additionally, adjustment is made to the timing of the first vertical sync pulse to assure proper display interlace when the line count is changing. The location of the active picture lines is adjusted dynamically within the field in order to keep the displayed picture centered on the vertical axis of the television screen. This centering compensation is accurate to within about +/- 1 line. In order to mask even this minimal effect, the time varying profile — as shown in figure 1 — is applied when significant changes in program content occur ... typically at scene changes. Other minor modifications are made to the vertical blanking interval to assure compatible operation of television receivers with digital and count-down synchronization signal processing.

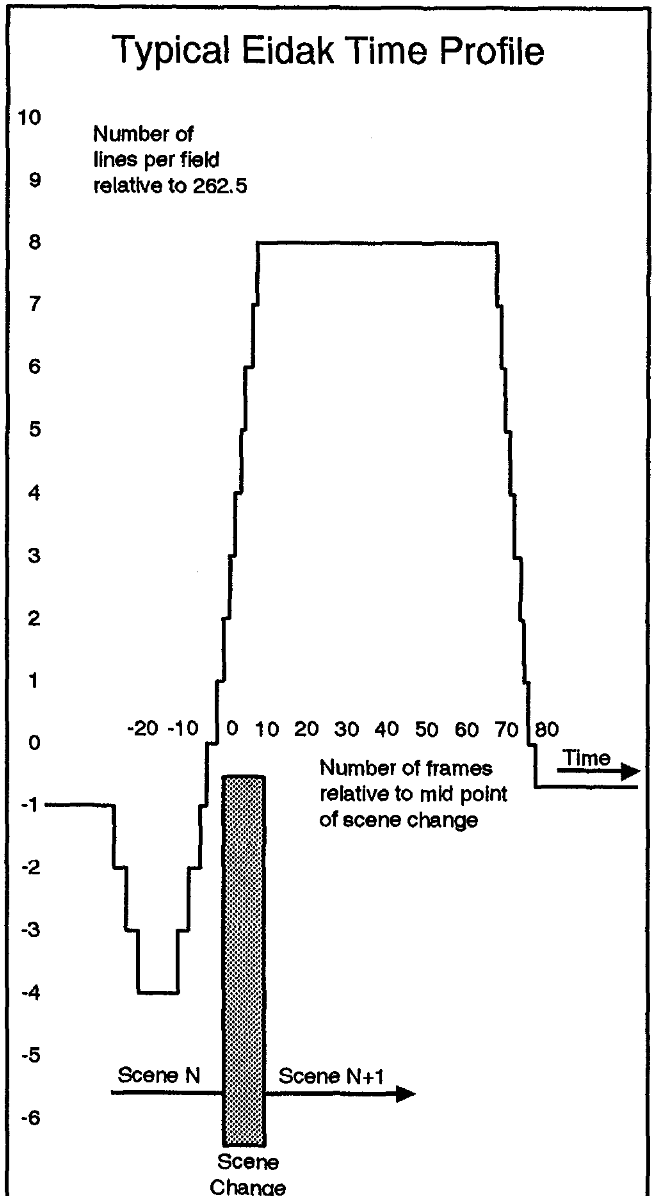

Typical Eidak Time Profile
Number of
lines per field
relative to 262.5
10
9
8
7
6
5
4
3
2
1
0
-1
-2
-3
-4
-5
-6
-20 -10 0 10 20 30 40 50 60 70 80
Time
Number of frames
relative to mid point
of scene change
Scene N
Scene N+1
Scene
Change

Figure 1.

Figure 2 compares the vertical blanking interval of the copy-protected video signal with NTSC per RS 170. The video field waveform differs only slightly from standard NTSC; normal horizontal line structure is maintained and color burst is locked to sync. The following are the differences between the copy-protected video signal and RS 170 NTSC:

1. <u>Horizontal line count</u>. The line count on any field may vary from 254.5 to 272.5 lines. Interlace and color phase coherence is maintained throughout. During the transit from one line count to another, the rate of change is one line more or less per field. Any resulting line count (except 525) may be held for an arbitrary period of time.

2. <u>Vertical interval</u>
   A. The last vertical sync pulse on a half line boundary is deleted from every field. The equalizing pulse on the half line boundary is deleted from lines 7, 8, and 9 in field 1 and from lines 6, 7, and 8 in field 2.

   B. The start of first vertical pulse in field 1 is advanced approximately 20 us while the field line count is increasing. The start of the first vertical in field 1 is delayed approximately 20 us while the field line count is decreasing.

3. <u>Compensation</u>
   The start of the active vertical display is delayed one TV line for each increase of two in the line count. The start of display is advanced one TV line for each decrease of two in line count. The advance or delay occurs in field 1 and never exceeds one line per change.

Analysis of program material to determine the optimum profile timing (e.g. scene changes) occurs prior to transmission -- for pre-recorded material. A data file is created which associates this information with the program's time code track. This information is then either added to a data track (or inserted in the VBI), on the pre-recorded medium, i.e. tape, or kept on file for use with the copy-protection processing equipment. For live events the scene change analysis is performed automatically in real time.

Variation of frame length is accomplished digitally by changing the rates at which frames of digitized video are written into and out of a multiple frame store buffer memory. Within the Eidak processor (figure 3), the control code reads the profile timing data and uses it to control the variation in number of lines per frame. (Alternately, the control code reader extracts this information from a data file and matches it to time code). Another important function of the processor is the scrambler interface. Because the copy-protected signal inherently contains non-standard vertical sync, some re-synchronization must be provided for cable scramblers which typically are dependent on precise VBI timing. For remote hub operations of scramblers, frame length data is encoded into the VBI, and at the hub location a scrambler interface (ESI) converts the frame length information into re-synchronization signals for the scrambler.

Although the copy-protection process for PPV is performed in real time for a transmitted signal, it can also be applied to a pre-recorded medium e.g. laser disc. In this case the copy-protection process is applied to the video signal during the mastering of the disc. When such a copy-protected disc is used later for program origination, the resulting video signal is is already copy-protected.

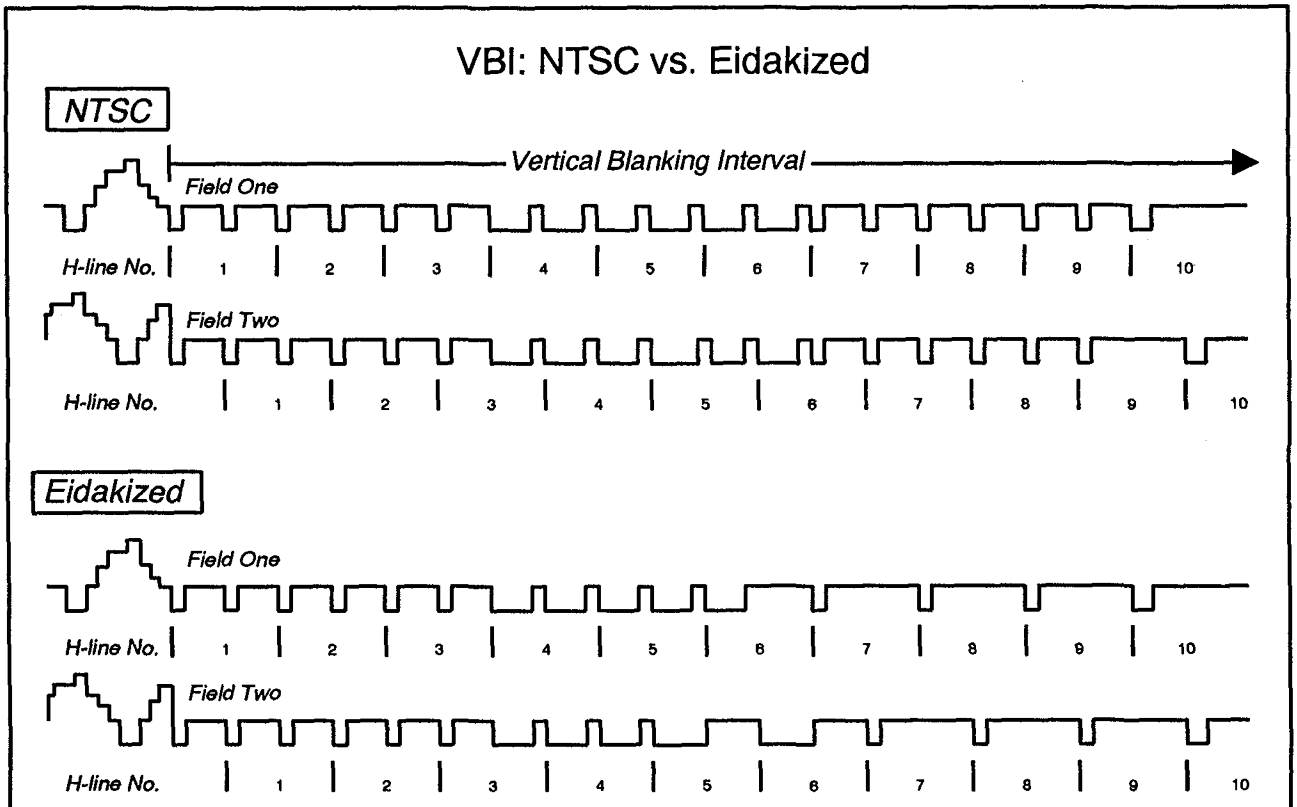

VBI: NTSC vs. Eidakized
NTSC
Vertical Blanking Interval
Field One
H-line No.
1 2 3 4 5 6 7 8 9 10
Field Two
H-line No.
1 2 3 4 5 6 7 8 9 10
Eidakized
Field One
H-line No.
1 2 3 4 5 6 7 8 9 10
Field Two
H-line No.
1 2 3 4 5 6 7 8 9 10

Figure 2.

Eidak Processor Diagram

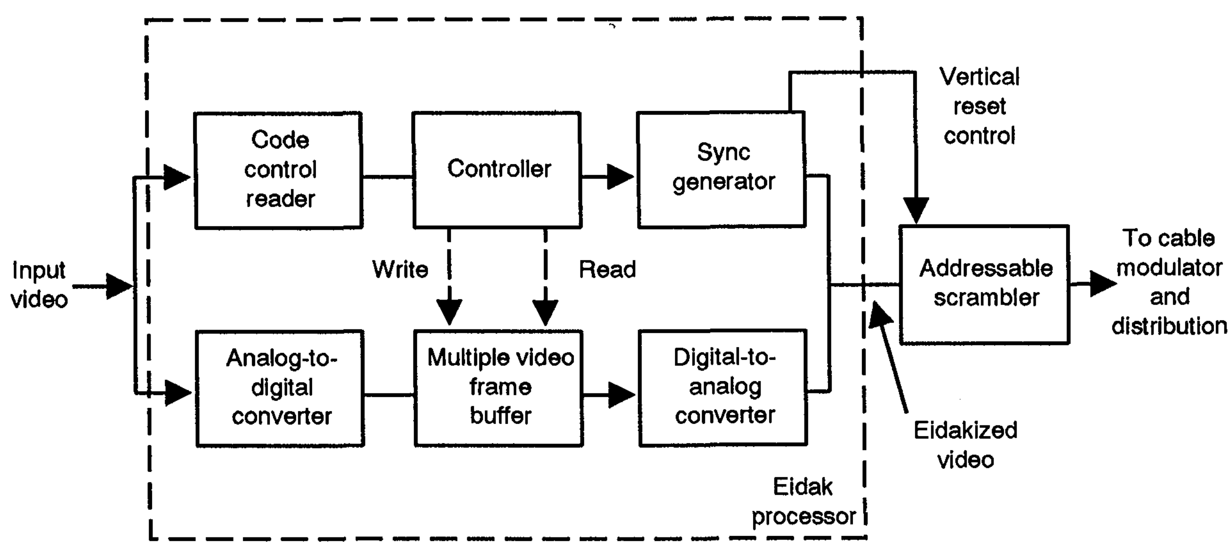


Figure 3.

SIGNAL DISTRIBUTION/COMPATIBILITY

System operation with video signals which depart in any way from the NTSC standard of 525 lines per frame, requires attention to compatible operation of equipment. System segments in which particular care must be taken to assure compatibility are:

Satellite

- o Satellite Link -- encoders and decoders.

Cable System

- o Head end -- video processing equipment and addressable scramblers.
- o Hubs -- link transmission equipment and addressable scramblers.
- o Subscriber equipment -- addressable converters and television receivers.

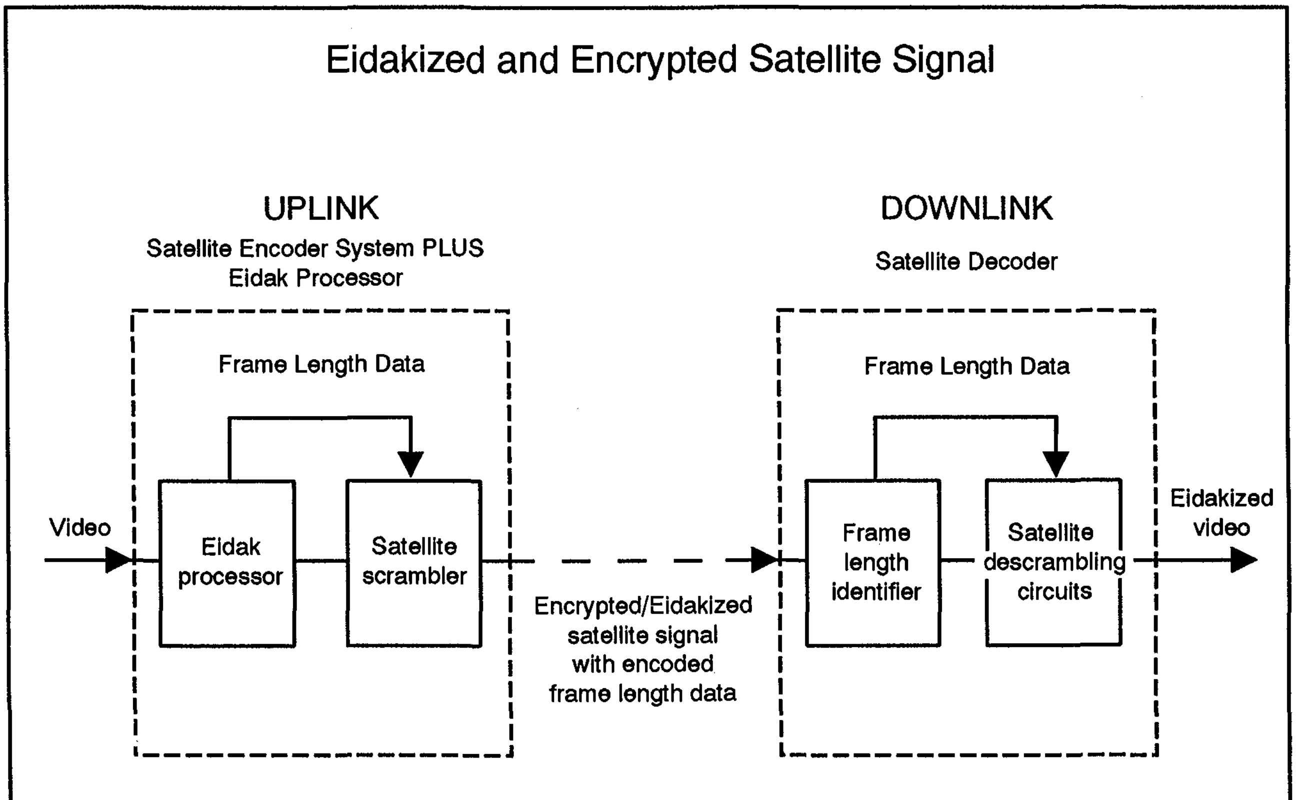
Copy-protection of satellite-distributed PPV programming may be accomplished in either of two ways, depending upon the encryption system employed. If the encryption system is designed for compatibility with copy-protected video signals, then it is most efficient to apply the copy-protection treatment to the program material prior to satellite transmission. In the case of an encryption system which is not capable of passing copy-protected programming, the copy-protection process is applied at the cable head end.

Satellite encryption systems generally are sensitive to field or frame rate variations for two reasons. First, descrambling of video and associated sync is performed field by field and normally assumes the use of standard 525 lines/frame video. Secondly, control signalling is often synchronized to the standard video frame rate. These sensitivities are readily disposed of if frame length information accompanies the encrypted signal, provided the descrambler is appropriately equipped. Figure 4 shows the uplink configuration with a copy-protection processor providing frame length data for encoding into the encrypted signal. The satellite decoder includes provision for recognizing video frame length, and adaptively adjusting the descrambling function.

Operation with a satellite encryption system which is not designed for use with variable field length video requires the use of a copy-protection processor at the head end downlink. In this case, the satellite encryption and decryption equipment is standard. In order to ensure that PPV programming is copy-protected as required at the head end, the transmitted signal is subject to a scrambling overlay which can only be removed by passing the received video signal through a copy-protection processor (figure 5). The processor thus serves a dual role -- the scrambling overlay is removed at the same time the copy-protection processing occurs. In this scenario, the signal at the head end prior to the copy-protection processor is scrambled with the overlay (and is therefore not usable in the cable system); after processing it has the scrambling overlay removed and copy-protection applied.

Compatible operation with head end equipment is the key to satisfactory distribution of copy-protected signals within the cable system. Cable distribution equipment has been found to be transparent to protected signals. Cable converters generally operate satisfactorily provided certain precautions are taken with the head end scrambler. The copy-protection process is optimized to work satisfactorily with subscriber television receivers.

Addressable scramblers almost always use vertical field rate timing for one or both of two purposes. Frequently address-control and/or program-



Eidakized and Encrypted Satellite Signal
UPLINK
Satellite Encoder System PLUS
Eidak Processor
Frame Length Data
Video
Eidak processor
Satellite scrambler
Encrypted/Eidakized satellite signal with encoded frame length data
DOWNLINK
Satellite Decoder
Frame Length Data
Frame length identifier
Satellite descrambling circuits
Eidakized video

Figure 4.

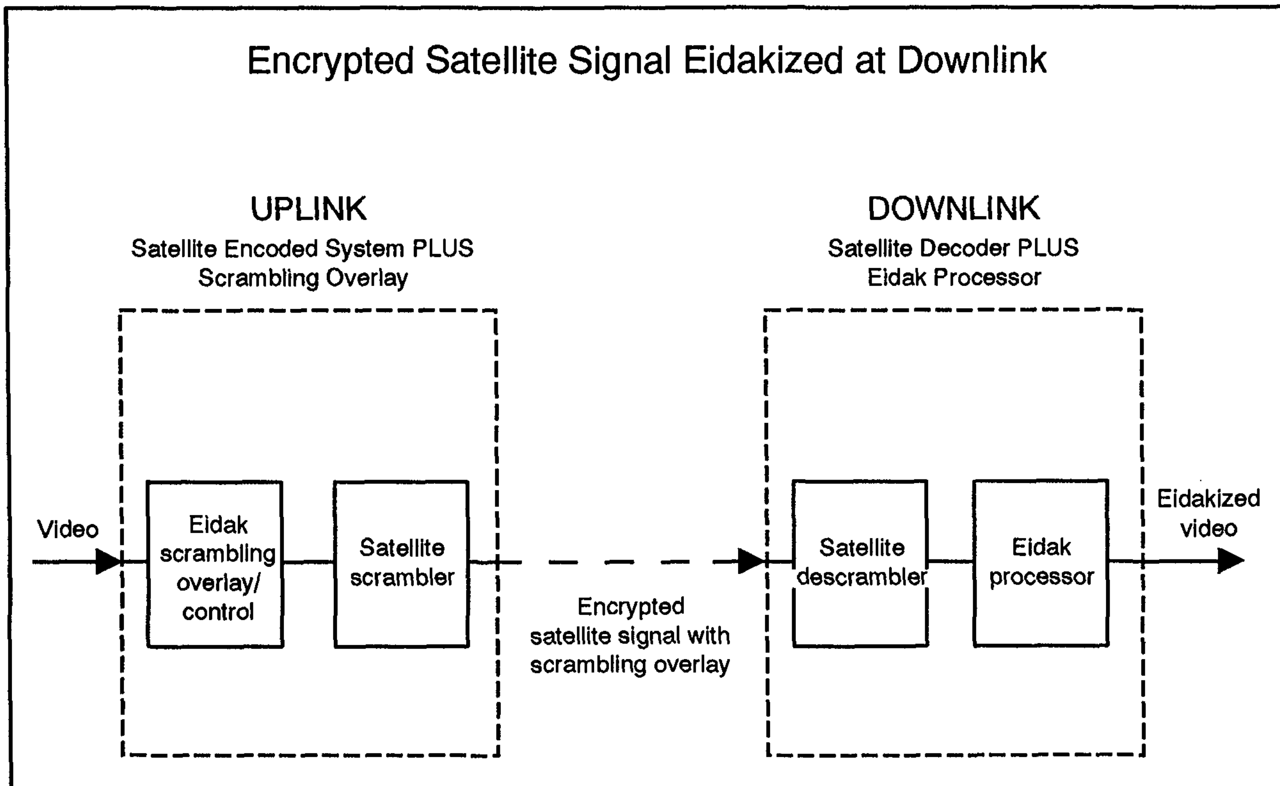

Encrypted Satellite Signal Eidakized at Downlink
UPLINK
Satellite Encoded System PLUS
Scrambling Overlay
DOWNLINK
Satellite Decoder PLUS
Eidak Processor
Video
Eidak scrambling overlay/ control
Satellite scrambler
Encrypted satellite signal with scrambling overlay
Satellite descrambler
Eidak processor
Eidakized video

Figure 5.

identification data is transmitted either within the VBI, or applied to the sound carrier and timed to coincide with the VBI. It is common practice for the scrambler to determine VBI timing by counting lines or clock pulses from a previous VBI, with the assumption that frame rate timing is standard 525 lines/frame. In order to accommodate video with variable field length, it has been found necessary to provide the scrambler with field reset information, either in the form of a reset timing signal, or by interruption of the scrambler's internal timing clock.

When the copy-protection processor is co-located with the scrambler, this vertical reset control output is provided by the processor -- configured for each specific scrambler type.

In cases when the copy-protection is applied remotely (e.g. satellite delivery of copy-protected signals, hubs, or origination from copy-protected laser disc), a separate scrambling interface device -- the ESI (fig. 6) -- is used to derive the reset control signals from the copyprotected video signal. The ESI consists of a data receiver which extracts the frame length data encoded in the VBI. The frame length data is used to generate properly timed reset or clock interruption pulses for the scrambler; the ESI interface is also figured for each specific scrambler type.

When a copy-protection processor is employed at a cable head end, information is required to cause the profile timing to coincide with scene changes. This data is provided either automatically by VBI signalling, or by means of a data storage device (either EPROM or floppy disk) sent to the head end site for use with program tapes or discs.

Standalone operation with a video laser disc player can be readily achieved by use of copy-protected discs (fig. 7). When the copy-protected disc is played at the head end, the resulting signal is already copy-protected, and is provided with the VBI data necessary to activate the ESI scrambling interface.

HEAD END EQUIPMENT

Configuring head end equipment to work with copy-protected video depends upon the program source and the location at which the copy-protection treatment is applied. If the program material is copy-protected prior to arrival at the head end (e.g. at a satellite uplink, at another head end, or on a pre-recorded copy-protected laser disc), only the scrambler interface is required. Should the copy-protection treatment be applied at the cable head end, then the copy-protection processor is used -- which also includes the scrambler interface function. In each case the addressable scrambler must be equipped for use with variable field length video signals. Wherever the Eidakized signal passes through a secondary hub equipped with another scrambler, the ESI scrambler interface is used.

Copy-protection of an encrypted satellite signal requires a compatible satellite descrambler at the head end receiver locations.

Table 1 shows the head end equipment requirements for operation with copy-protected PPV signals for each of these configurations.

In each configuration, operation is simple and automatic. Monitoring indicator lights on the copy-protection processor indicate its status (e.g. copy-protected video mode). Operation and adjustment of the cable scrambler and modulator are the same as with standard NTSC video. Operation of the laser disc player for local origination requires no additional adjustments or controls. **Important Note:** Once the video signal is copy-protected, it should not be passed through any other sync-sensitive or sync-restoring video processing equipment, for example video proc. amplifiers.

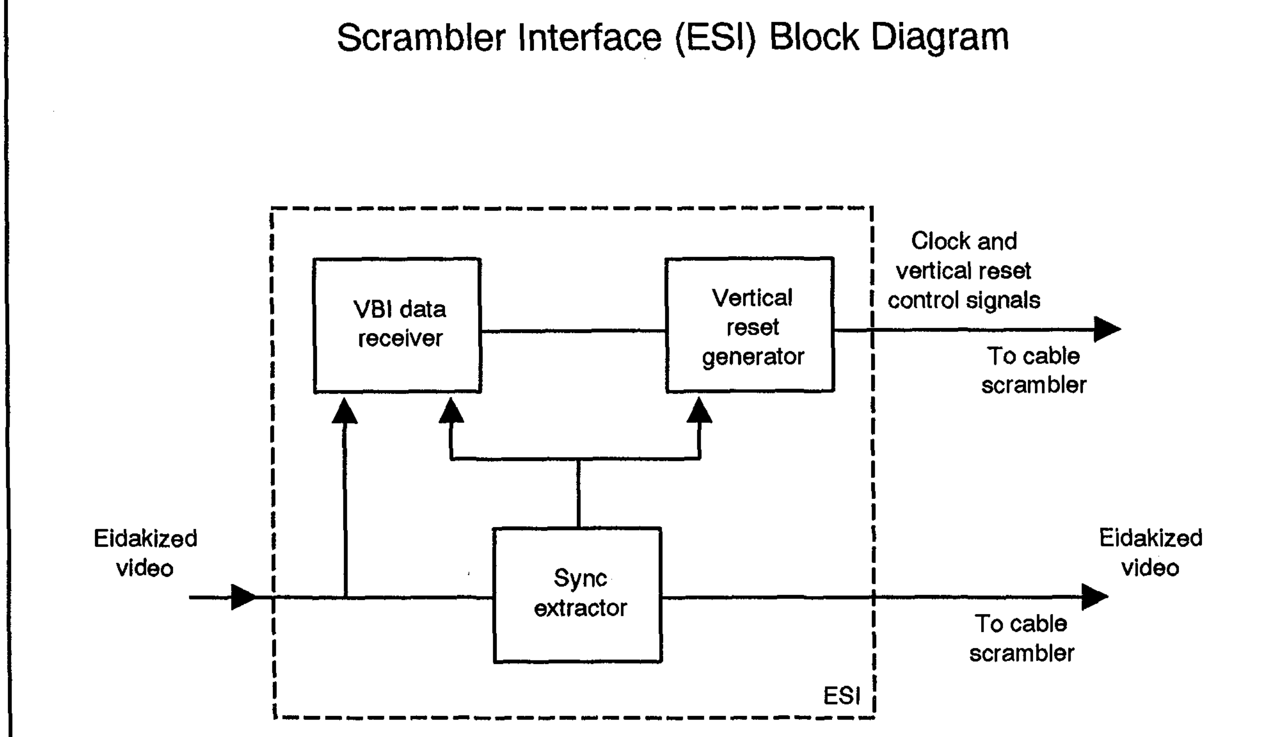


Figure 6.

Programming Source	Satellite Receiver / Descrambler	Scrambler Interface: Head End	Scrambler Interface: Hub
1. Satellite distribution — copy-protected at uplink	Eidak compatible	ESI	ESI
2. Satellite distribution — copy-protected at head end	Standard	Eidak Processor	ESI
3. Standalone — copy-protected laser disc	—	ESI	ESI
4. Standalone — Non copy-protected tape or disc	—	Eidak Processor	ESI

Table 1.

CONCLUSION

The ability to prevent copying of PPV video signals is becoming a necessity to assure access to the kind of programming required to fuel the growth of PPV. Technology developed specifically for this purpose has been shown to work for cable distribution of PPV scrambled signals. In order to assure proper systems operation, special attention has to be given to equipment at cable head ends and hubs. Cable scrambling equipment generally requires some adaptation in order to permit operation with a video signal which has non-standard synchronization. As cable increasingly looks to PPV for revenue growth the cable PPV universe can now be equipped to provide access to those new program sources which require control of home copying to assure timely availability.

REFERENCES

1. "Unrecordable Video" A. Lippman & T. Hogue (MIT). IEEE Trans Broadcasting. Sept. 1987.

2. "Pay-per-view Copy-Protection", G.S. Stubbs, Communications Technology, March, 1989.

3. "A working Anti-Taping System for Cable Pay-Per-View" B. Perlman, IEEE Trans. on Consumer Electronics, August, 1989.

4. "Copy-Protection: What's in Store for PPV", G. S. Stubbs, CED Magazine, February 1989.

5. "PPV Copy-Protection in the Cable System Environment" G.S. Stubbs. NCTA Convention, May, 1989.

6. "Technical Issues for Copy-Protection in the Cable Television Environment.", B. Perlman. Paper to be presented at the 1990 ITE Annual Convention, Tokyo, Japan, July 25-27, 1990.

7. U.S. Patent Number 4,673,981 A. Lippman, et al. Issued June, 1987.

8. U.S. Patent Number 4,914,694. E. Leonard, et al. Issued April 1990.

Standalone PPV with Eidakized Laser Disc

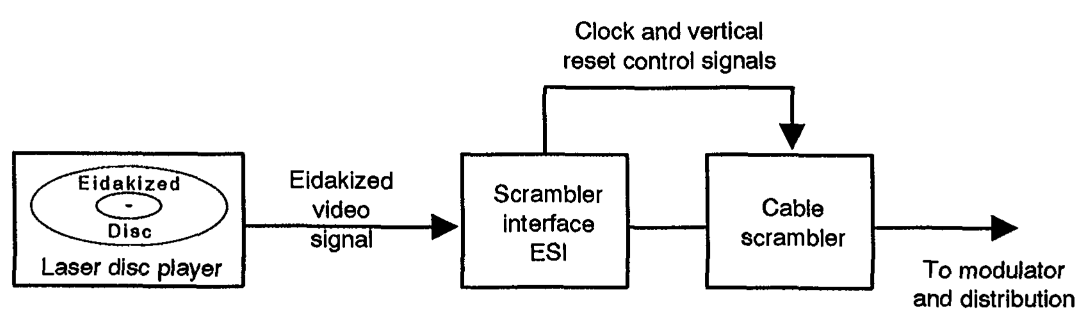


Figure 7.

Cable's Excellent Position in HDTV

Walter S. Ciciora, Ph.D.

American Television & Communications
Stamford Connecticut 06902-6732

ABSTRACT

The message of this paper is that HDTV will not happen suddenly. Experience shows that developments such as this will take a considerable amount of time. In the mean while we need to take care of our present customers. To do so, cable should deliver the best NTSC available. The need for this is based on the ever-improving performance of consumer electronics equipment.

When we review the mechanisms which cause picture impairments, we realize that there is a limit to what can be accomplished in the receiver. The only alternative is to improve the cable plant. There are a number of cost-effect methods which can be phased into practice in an evolutionary manner. When we examine the technical demands for HDTV carriage, we realize that they are essentially the same as for quality NTSC delivery. Since quality HDTV delivery is mandatory for competitive reasons in the short run, it becomes clear that cable will be ready for HDTV long before consumers are ready for HDTV.

INTRODUCTION

There are three reasons why HDTV is important to the cable industry. First, our friends in the telephone industry continue to say that they are the only ones who will be able to deliver true HDTV to the consumer. They say this must be done digitally and over fiber to the home. This is simply not true. Cable must make it clear that HDTV works well on the kind of coaxial cable systems currently in operation. Fiber backbone strategies and similar approaches make the job easier and the picture even better. But they are technical options, they are not required for cable delivery of HDTV. The second reason for cable involvement in HDTV is that HDTV will be with us for at least fifty years. That alone makes it important. We must be involved in shaping something that will have such a great impact on our future. The third reason is that we need to know how to plan rebuilds and upgrades. Waste can occur in two ways. First we can spend too much too soon. Secondly, if what we build becomes obsolete too soon, it too will have been wasted.

RATIONAL EXPECTATIONS

There was a lot of hype over HDTV and over how quickly it was to sweep across the country. Fortunately, most of that has died down.

Testing of HDTV proponents was originally scheduled to begin in December of 1989. Then it was postponed until June 4, 1990. It is currently hoped that this testing will begin sometime in the Fall of 1990. Nine testing slots were created from the original batch of proponents who qualified. Other proponents have come forth wishing to have slots. It is only human nature to expect that some of those who were tested early in the schedule will have made further inventions and progress and wish to be re-tested.

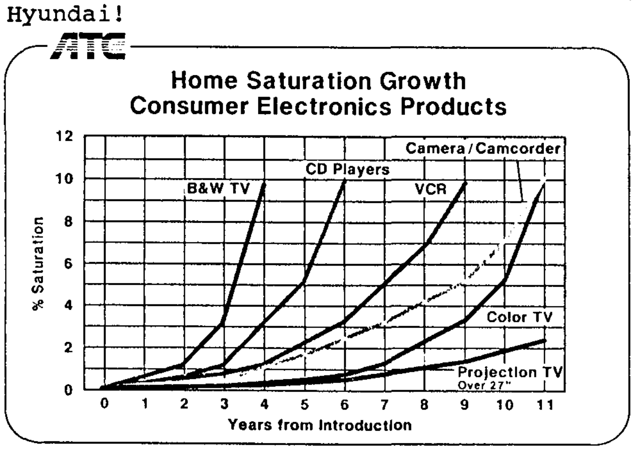
After all of the testing is complete, the advisory committee may need up to a year to digest the data and prepare its conclusions. Then, the FCC will need an additional year to eighteen months to decide.

This all boils down to the conclusion that a terrestrial standard may be available in early 1994. The first HDTV receivers may go on sale in late 1994 or in 1995.

Possible short circuits to this lengthy process would occur if the proponents themselves decide to merge into one system. The David Sarnoff Research Center and Philips Laboratories have agreed to do just that. While this action was necessary, it is by no means sufficient. Nearly all of the proponents need to merge for this short circuit to be successful. That is unlikely for several reasons. Many of the proponents are fundamentally different from each other and cannot merge for technical reasons. There are some very large egos involved which provide political impediments to merger. The "little guys" would cry foul if the big guys joined forces. Finally, there are too many involved in the FCC Advisory Committee to let too much happen without extensive testing.

HBO and MIT both did consumer research which shows that a viewer cannot tell the difference between NTSC and studio quality HDTV when seated more than five time the picture height. The HDTV signal used was not compressed by any of the proponents' systems. It was artifact free. There would be even less difference between NTSC and most proponent systems. Home viewing distance is primarily determined by room size and furniture placement. To satisfy these two constraints, the HDTV receiver must be a large screen device, three to four feet high. HDTV is a large screen phenomena! Large screens of any type, NTSC or HDTV, are expensive. In addition, there are only a limited number of homes which can accommodate a large screen.

History has some useful lessons that can help us project how the HDTV market will be penetrated. It took color TV seven years after introduction to reach one percent penetration. It took color TV eleven years to achieve ten percent penetration. The first black and white television sets cost as much as a compact car of the time. The first color television receivers likewise cost about the same as a compact car. It is reasonable to expect that the first HDTV receivers will cost about as much as a Hyundai!



I believe it is reasonable to predict that HDTV will be at one percent penetration in seven to ten years after introduction. Ten percent penetration will take thirteen to fifteen years after introduction. Thus if introduction takes place in 1994, expect one percent penetration in 2004 and ten percent after 2007. Any more aggressive projections than this demand an explanation. The challenge is this: why would HDTV be more of an improvement over color than color was over black and white? Will this degree of improvement justify the price of a large screen TV? Just as with color, programming will initially be scarce. The broadcast industry is mature and not likely to grow because of HDTV. Its motivation to quickly add programming in an era of less than a few percent receiver penetration will be very low. Certainly, advertisers will not pay more for programing in HDTV if the population of receivers is small.

Another reason for slow penetration of broadcast programming is the very practical problem of new tower construction. It is nearly universally agreed that a new 6 MHz channel will be required for HDTV. This means another transmitter and antenna. But current antenna towers are fully loaded. Few can accommodate another antenna. Since many of these towers are shared, they would have to accommodate several more antennae. The problem of tower construction involves not only cost but getting construction permits. These are difficult to obtain since land is scarce, environmental impact statements need to be filed, and environmental activists' objections overcome. Add to this the growing fear of the potential for electromagnetic radiation to cause cancer, and it is easy to appreciate that spectrum availability is not the only constraint on broadcasters. But this issue must not be slighted. Not only do broadcasters have a spectrum scarcity, more importantly they have a scarcity of <u>quality</u> spectrum. HDTV spectrum must be free of ghosts, co-channel interference, and noise. Otherwise, the HDTV picture will be unacceptable. For these reasons, even if broadcasters could justify the cost, the rush to HDTV transmission simply won't materialize.

GOALS AND OBJECTIVES

Cable should have four principal objectives for HDTV: 1) preserve cable's ability to compete, 2) deliver broadcasters' HDTV signals, 3) serve the NTSC population, and 4) accommodate cable's unique needs. In all of this, cable must find cost-effective solutions.

Clearly, the environment for cable is becoming more competitive. Pre-recorded media, Direct Broadcast Satellite, MMDS, and even digital delivery of video via fiber to the home by the telco's are strong potentials. The most immediate and possibly highest quality delivery may come via baseband prerecorded media with bandwidths of 15 MHz to 20 MHz. We must not be embarrassed when our subscriber turns off his VCR and turns to HBO.

Cable must be able to deliver the HDTV signal broadcasters choose. Broadcast signals are likely to remain very important to our subscribers. It is only good business to deliver what subscribers want. Fortunately, broadcast technologists recognize the importance of cable. While acrimonious rhetoric is employed by some in the upper levels of broadcast and cable management, the engineering community is working together to ensure that things will work from a technical perspective.

Cable has a number of unique needs. These include scrambling, encryption, and addressability. Also, the cable signal must be delivered via satellite to cable headends. The energy in cable signals is important because of the number of signals carried. Unnecessarily high signal energy means amplifiers, laser diodes, and other devices will reach into their non-linear regions of operation and generate undesirable distortions.

Cable should not lose sight of the population of NTSC receivers. There are currently nearly two hundred million NTSC sets and nearly seventy five million NTSC based VCR's. In excess of twenty million new NTSC receivers are added to this population each year. These NTSC receivers have a lengthy life expectancy, typically twelve to fifteen years. Many survive well beyond that. They will be with us for a long time to come.

Given the cost and size of an HDTV receiver, it is likely that for a couple of decades, only one HDTV receiver will exist in most homes. The rest of the home will be sprinkled with NTSC receivers. Thus the technical standard for HDTV must do nothing to impair NTSC receivers.

From a cost perspective, cable must be sure it does not have to raise rates for the NTSC majority to cover the costs of HDTV delivery to a small, less than one percent, opulent minority. This would make for bad business and very bad politics. We believe that this will not be necessary. We believe that cable's best HDTV strategy is to improve our plant to deliver the best NTSC possible. We believe that this will allow HDTV carriage without economic penalty.

IMPROVING NTSC

For decades, NTSC was better than consumer electronics. In that situation, it makes no economic sense to deliver a signal that could not be displayed. Resources are better spent making sure more subscribers have access to signals and have this access at reasonable rates.

But things have changed in the last five years or so. Now consumer electronic equipment can display more quality than the NTSC standard can support. It is now time for the transmission path to catch up to the capability of consumer electronics.

There are two fundamental constraints to the improvement of NTSC: 1) compatibility with existing receivers, and 2) the existing huge population of receivers.

There are two kinds of picture impairments: 1) those due to the NTSC baseband standard, and 2) those due to the manner in which the NTSC modulation structure interacts with the transmission path. The design of the NTSC modulation scheme makes it particularly vulnerable to transmission path problems. A modulation scheme based on modern communications theory would be substantially less subject to transmission path deficiencies. Unfortunately, such a scheme would be incompatible with the existing population of TV receivers.

Baseband impairments include dot crawl and cross color. Dot crawl is the movement of tiny dots along the edges of colored objects. Cross color is the name of the spurious rainbow that appears on monochrome detail such as Johnny Carson's checkered jacket. These are the consequence of the color information being shoehorned into a black and white signal. The two kinds of information get in each others' way. The process for combating this is called "comb filtering". There are a variety of techniques for improving the performance of comb filters. While all make some compromises, smart comb filtering does a very good job of minimizing these artifacts. Other NTSC artifacts include scan line visibility and flickering of bright image areas. These can be minimized by line doublers. They double the number of scan lines by estimating the information not transmitted between the lines. While this does not double the vertical resolution, it does minimize these NTSC artifacts.

Transmission path impairments are much more difficult to deal with. They come in two varieties, coherent and non-coherent. Coherent impairments are more severe. They include ghosts, micro-reflections, cross-modulation, co-channel interference, and beats between carriers. Non-coherent impairments include random noise and impulse noise.

The principal weaknesses of the NTSC standard are its vestigial sideband modulation scheme, the use of simple amplitude modulation, the inter-leaving of the color signal, the separate audio carrier, and the addition of stereo. Vestigial SideBand, VSB, modulation was used to save bandwidth. As a side effect, when the signal is distorted, VSB introduces complexities which are very difficult and expensive to remove. This is the biggest impediment to ghost cancelling. The biggest impediment to cross-modulation and beats is the number of high energy carriers in the signal. The simple method of amplitude modulation makes the NTSC signal vulnerable to noise in the transmission path. These defects are extremely difficult and costly to attack while still remaining compatible with the NTSC standard. The NTSC standard has our hands tied. We have very few options.

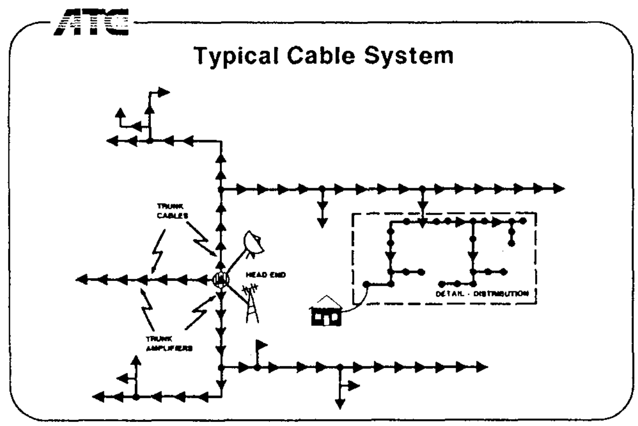
There is one last improvement to NTSC which subscribers demand: more channels. More programing becomes available almost yearly. Subscribers are anxious to have it.

If we insist on remaining compatible with the large population of existing receivers, there is little we can do to improve system performance. After we exhaust the few tools available to us, we must upgrade the plant itself. Fortunately, this can be done in a cost effective and evolutionary manner.

Two of the many techniques for improving NTSC which directly support a strategy for HDTV are the fiber backbone coupled with amplifier upgrades and super distribution. The first of these techniques yields more channels and less noise while the second reduces miro-reflections and non-linearities.

In the typical cable system, long cascades of amplifiers build up noise and limit bandwidth. While the cable itself is often capable of transmitting one gigaHertz and more, the amplifier cascades are limited to 650 MHz or so. In addition, the long cascades are a serious reliability hazard.

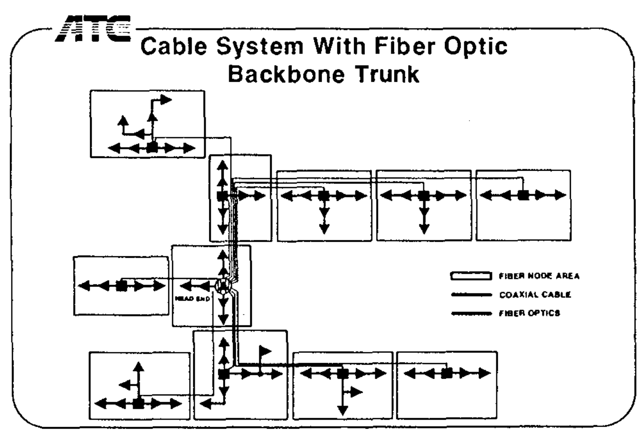
The fiber backbone approach breaks the cable system into a multitude of much smaller cable systems with amplifier cascades limited to four to six amplifiers. Each of these small cable systems is fed with a fiber link to the headend. The advantages include significantly lowered vulnerability to amplifier outages, reduced bandwith restrictions and noise build up due to amplifiers in series, and greatly reduced ingress. The latter effect makes two-way cable practical. A major attraction of the fiber backbone is that its implementation cost is low. Fiber is "over lashed" onto the existing trunk plant. The in-place cable is broken into segments and used for the small scale cable systems. Some of the amplifiers are reversed in direction. Nothing is wasted.



Typical Cable System

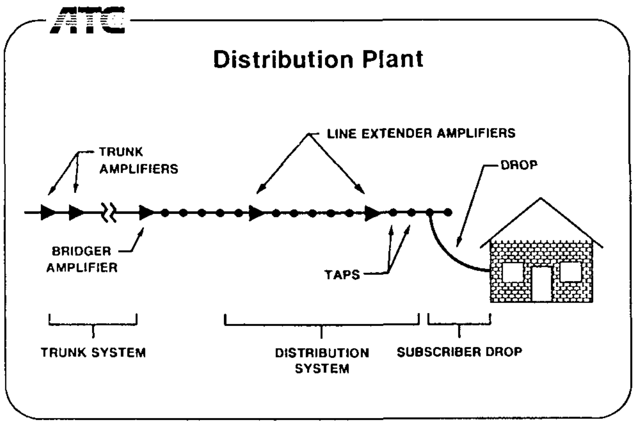
In the typical cable system, 10% of the cable footage is in the trunk, 40% of the footage is in the tapped distribution, and 50% is in the drops. The cost effectiveness of the fiber backbone technique stems form the fact that it involves only 10% of the plant footage. In some design studies, the cost of implementing this upgrade came to less than $25 per subscriber.

The fiber backbone effectively cures most of the ills of the trunk part of the plant. This improves NTSC delivery as well as prepares the trunk plant for HDTV.



Cable System With Fiber Optic Backbone Trunk

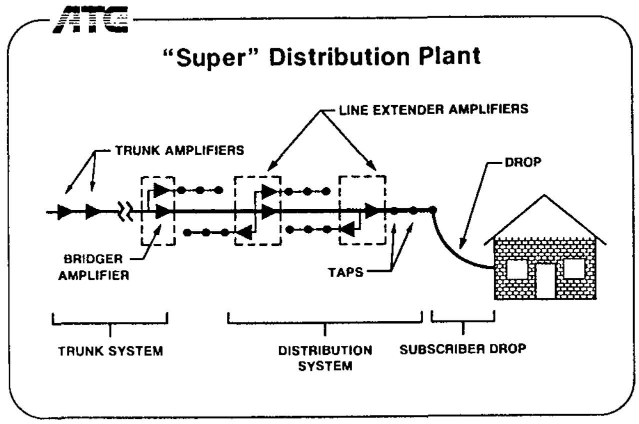
If we now turn our attention to the distribution plant, we find that we must run higher signal levels to support the tapping of energy to serve drops to customers. The higher signal levels mean that we begin to reach into non-linear areas of the amplifiers' operating characteristic. Non-linear distortion builds up. In addition, the taps are not perfectly impedance matched to the cable. Consequently, the signal is reflected back and forth between the taps resulting in a smearing of the picture. This phenomena is called "micro-reflection" for two reasons. The strength of the reflections is low and the time delay of the reflections is short.



Distribution Plant

Rogers Cable of Canada has suggested the answer to these difficulties. They have called their technique "Super Distribution". In one form of its implementation, line extender amplifiers are structured to have up to three hybrid amplifier chips. One feeds the next line extender amplifier, one feeds half of the taps back to the previous line extender, and the third feeds half of the taps to the next line extender. The existing tapped feeder cable is cut in half between line extenders. New, untapped cable is over lashed to connect line extenders. The consequences are that the signal level between amplifiers is lower since that cable is not tapped. The signal level on the tapped runs is lower because they are shorter. Non-linearities are reduced and they do not build up as in the previous structure. Also, signal leakage may be less of a problem because of the lower signal levels. In addition, the number of taps in series in any cable is drastically reduced thereby reducing the amount of micro-reflections experienced by any one subscriber. This technique effectively cures the ills of distribution portion of the plant. Rogers estimates the cost of this upgrade to be less than $25 per subscriber.

With fewer amplifiers in series, the constraints on their design and operation are reduced. Higher bandwidths become practical.



"Super" Distribution Plant

Recently, Dave Pangrac of ATC has developed an extension of the fiber backbone approach which takes fiber farther into the plant. In one version, passive splitters are added to the fiber run from the headend. Then shorter runs into the neighborhood bring the optical plant closer to the home. Even fewer amplifiers are interposed between the subscriber and the headend. In another implementation, low cost lasers are used as repeaters to feed the branches at the end of the trunk run. The potential of optical amplifiers promises to yield further evolution of this technique.

CONCLUSION

If the assumptions on HDTV timing are correct and if the assumptions for the need of better NTSC in the shorter term are correct, cable will be ready for HDTV long before consumers are ready to spend the money these new receivers will demand. This will happen in an evolutionary manner over many years. It will be a cost effective approach which will generate compensating revenues and economies of operation.

THE AUTHOR

Dr. Ciciora is Vice President of Technology at American Television &Communications, ATC, in Stamford Connecticut. Walt joined ATC in December of 1982 as Vice President of Research and Development. Prior to that he was with Zenith Electronics Corporation since 1965, He was Director of Sales and Marketing, Cable Products, from 1981 to 1982.

Earlier at Zenith he was Manager, Electronic System Research and Development specializing in Teletext, Videotext and Video Signal Processing with emphasis on digital television technology and ghost canceling for television systems.

He has nine patents issued. He has presented over seventy papers and published about thirty, two of which have received awards from the IEEE. Walt writes a monthly column titled "Ciciora's Page" for Communications Engineering and Design magazine.

He is currently chairman of the National Cable Television Association, NCTA Engineering Committee, Chairman of the Technical Advisory Committee of Cable Labs, and President of the IEEE Consumer Electronics Society. He is a past chairman of the IEEE International Conference on Consumer Electronics. Walt is a Fellow of the IEEE, a Fellow of the Society of Motion Picture and Television Engineers, and a senior member of the Society of Cable Television Engineers. Other memberships include Tau Beta Pi, Eta Kappa Nu, and Beta Gamma Sigma. He served on several industry standard-setting committees. Current interests center on competitive technology, the consumer electronic interface with cable, and HDTV.

Walt received the 1987 NCTA Vanguard Award for Science and Technology .

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 fro 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, and a hope to someday become more active in amateur radio (WB9FPW).

CONTROLLED SUBJECTIVE TESTING OF CABLE SYSTEM IMPAIRMENTS TO PICTURE QUALITY USING PSYCHOPHYSICAL METHODS

Michael F. Jeffers
Jerrold Communications
Applied Media Lab

ABSTRACT

For the first time, the Cable Industry will take on an organized critical evaluation of the subjective effects of typical impairments to a television picture generated in a CATV system. In the past, this information was gathered from independent sources without verification or firm documentation. This upcoming study is sponsored by Cable Labs under the direction of Tom Elliot, Vice President of Science and Technology. It will be conducted at the Jerrold Communications Applied Media Laboratory in Hatboro, PA. Bronwen Lindsay Jones, an Audio-Video expert in the field of psychophysical testing, has been contracted by Cable Labs to establish this criterion for the subjective measurements and to record all data. The detailed results of this study will be published by Cable Labs.

System Description

The impairments to be measured are:

- Video Signal to Noise
- Composite Second Order
- Third Order Distortions
- Chroma/Luma Delay Inequality (Envelope Delay)
- Phase Noise
- Reflections - (Echoes)

A test system will be set up at the Jerrold facility. There will be a headend with the capability of generating sixty (60) video modulated radio frequency (RF) carriers. These television channels will carry twenty (20) or more clean NTSC signals received off-air , from satellites, and video test pattern generators. The reference channel will be modulated with special pictures chosen to show most readily the various impairments. The reference channel can be assigned to any frequency in the RF. spectrum. Both still and moving scenes will be evaluated. Still scenes will be evaluated in the first set of experiments. The sixty (60 channel headend can be operated in the Standard, the Incremental Related Carrier (IRC), and the Harmonically Related Carrier (HRC) mode. The headend output is connected to the input of a balanced CATV system located in one of Jerrold's temperature control chambers. It includes complete stations with equalizers and the full trunk spans of coaxial cable. Ganged attenuators at the test system input and output allow the system to be driven to various levels of distortion while maintaining a constant RF signal input at the TV set displays. The following displays will be used: a Cable Ready TV set; a TV set or video monitor with a video output port; and an IDTV set to view the impairments in a 525 line progressive scan mode. These displays utilize comb filter technology to produce high resolution pictures while minimizing NTSC artifacts. These subjective tests for distortion will be conducted in the Standard mode and will be repeated when the headend is phase-locked in the HRC mode and the IRC mode. The IRC mode is best suited to examine Composite Second Order (CSO) distortion. The "System Under Test" will be completely characterized before and throughout the testing period for Carrier to Noise (C/N), Composite Triple Beat (CTB), Cross Modulation (Xmod) and Composite Second Order (CSO) at various RF levels across the system operating range. Test measurements procedures from the NCTA Recommended Practices, second edition, will be used for the above objective measurements.

In addition to the noise and distortion impairments, subjective tests will be made on phase noise and micro reflections. For phase noise we intend to overdrive a headend modulator with noise or low frequency components to the point where phase noise is observable, to allow expert and non-expert viewers to evaluate the annoyance of this distortion. Similarly, the test for micro-reflections will allow us to pick various magnitudes and delays of the resultant echoes to ascertain subjectively the annoynance of the impairment. The phase of the echo will also be adjusted for worst case interference.

It is also our intention to generate Envelope Delay. This is the distortion that occurs when the chroma information is delayed relative to the luminance (or vice-versa). The degree of the delay can cause smear, and when the delay is severe it can cause an "out of register" situation some times seen in the Sunday comics wherein the color of hair, eyes and lips of a character do not line up with the image of the character's face. This delay can be measured by instruments and associated with the subjective evaluation of the impairment. As a first step, expert viewers will establish a threshold level of subjective impairment. This is the level at which the impairment can just barely be seen. Once experts agree on threshold, we will establish levels of impairment in each distortion category to which the non-expert viewers will be tested.

It is seen in the diagram (Figure 1) that the various levels of impairment will be controlled by programmable attenuators and coaxial relays. All these devices are functional to well above one Gigahertz. An extensive software program will be developed by Bronwen Jones' staff to control various levels of impairment, ranging from threshold to significant distortion These will be used to test the response or judgement of non-experts to different degrees of picture impairment. Each person viewing the subjective test will have a control so that they may conveniently register their opinion of the quality of the picture they see. These subjective judgements, as well as the specific level of the distortion for each test, will be automatically recorded. The accumulation of this data from the many non-expert viewers will be analyzed and published in the final report as a group of curves similar to the old TASO study conducted before color TV was available.

The subject matter used will be mostly still scenes chosen to highlight the various impairments. Some moving programs may be presented at a later stage. Each participant will be pretested for color blindness and visual acuity to eliminate those not qualified.

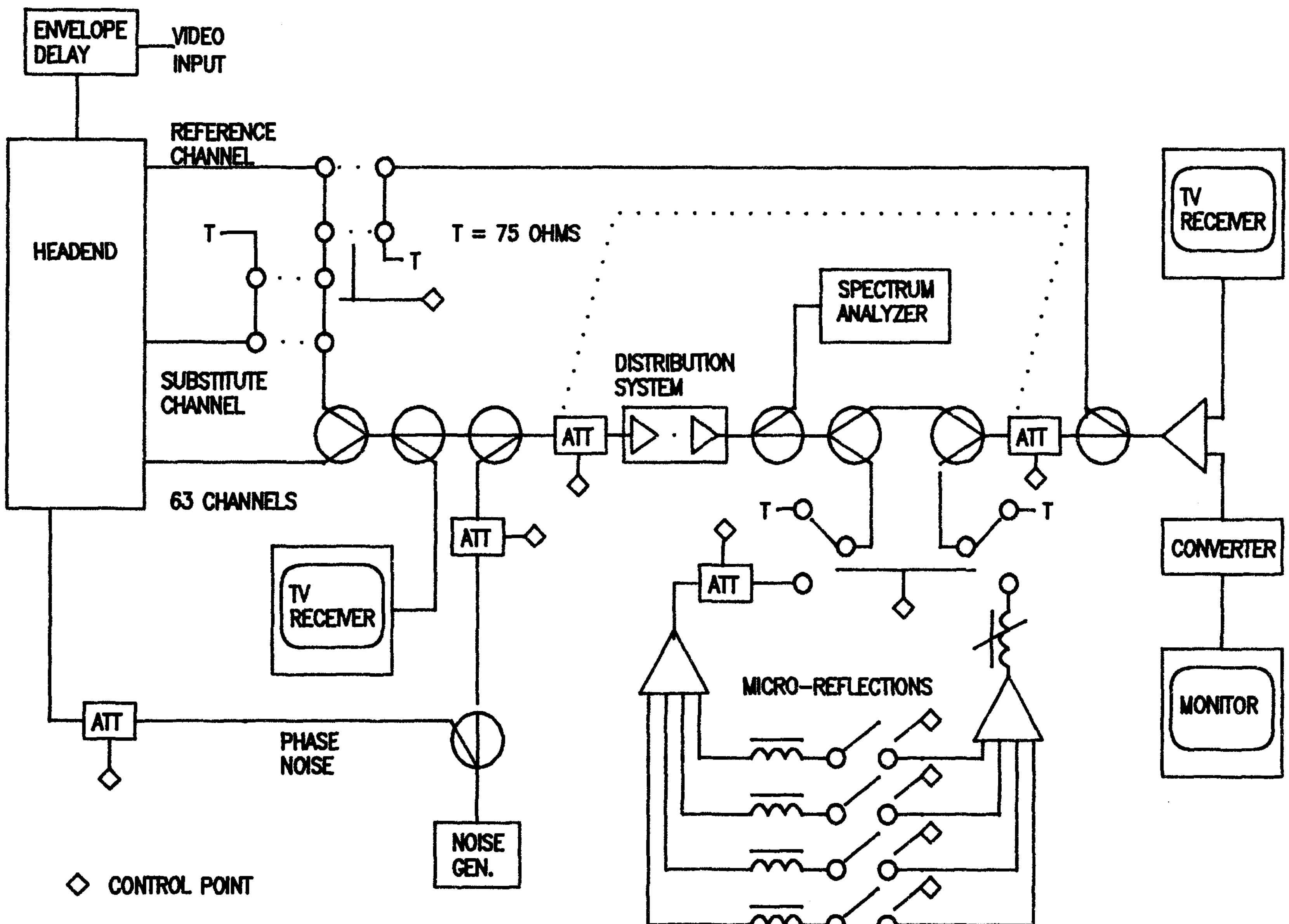


FIG. 1 – TEST SYSTEM BLOCK DIAGRAM

Conclusion:

The true conclusion of this effort will be the final report to be published by Cable Labs. This is estimated for the last quarter of 1990. We expect the results of the impairments that CATV has established by trial and error over the years to give no surprises. These are Carrier-Noise, Cross Modulation and Second and Third Order Intermodulation. The remaining impairments have not been thoroughly tested before and we should be able to establish meaningful data on thresholds. In all cases the subjective results of non-expert viewers should establish for the first time the picture quality levels acceptable by Cable TV subscribers. The goal is to give to the cable operator a cross reference between acceptable picture quality and the objective measurements made on cable television systems using standard instruments.

It might appear to some that this testing can be completed more expeditiously. It is to Tom Elliot's credit that he insisted on the automatic testing and recording of the data to insure accurate and sustaining results. We feel the time and money spent to achieve this goal is well worth it.

CONVERTER NOISE MEASUREMENT DEFINITIONS AND CONVERSION DERIVATIONS

Blair Schodowski

Scientific - Atlanta

ABSTRACT

When evaluating noise introduced by cable television converters, terms such as carrier-to-noise ratio, signal-to-noise ratio, and noise figure are commonly used. These terms are measures of noise performance. The ability to use these terms interchangeably has become increasingly important with the advent of baseband converters. Baseband converters have precluded the use of traditionally used noise figure equipment which is widely used for evaluating RF converters. Therefore, a complete understanding of noise terms is essential. This paper will identify the differences between the commonly used noise terms along with the derivation of the conversion factors.

INTRODUCTION

For years, noise figure has been the unwritten standard for specifying the noise performance of RF cable television converters. However, the substantially different architecture of a baseband converter precludes the use of a noise figure meter for measuring noise performance of baseband converters. One method used for measuring noise performance in a baseband converter is illustrated in figure 1. In this configuration noise performance is measured in the form of signal-to-noise ratio of the baseband video signal.

Even though video S/N ratio represents the true picture quality delivered to a customer's television set, it does not clearly indicate the level of noise the converter adds to the transmitted signal. Noise figure is a more familiar way to represent noise added by a converter. Fortunately, through mathematical manipulation, there is a conversion factor that can be added to the unweighted S/N measurement for translating video S/N to noise figure. The following equation shows this conversion,

$$NF = C_p(in)(dBmV) + 59.21(dBmV) - S/N(dB) - 6.86(dB) \qquad (1)$$

where:

$C_p(in)$ = input carrier power level into the converter
S/N = unweighted signal-to-noise ratio of baseband video signal
Noise BW = 4.0 MHz
Reference Impedance = 75Ω

To help understand this equation, the mathematical derivation for translating video signal-to-noise to noise figure will be presented. Also, several terms that are commonly used when describing noise performance will be defined.

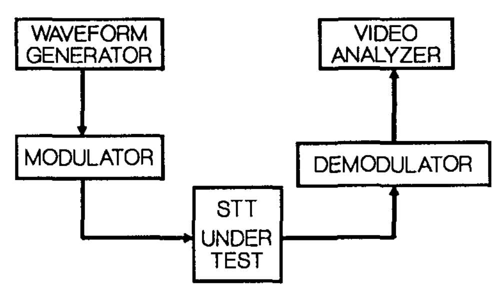


Fig. 1 Typical Signal-To-Noise Test Configuration

DEFINITION OF NOISE TERMS

When measuring noise performance in CATV systems, several terms are used to define the relative amounts of noise produced. The most common terms are: signal-to-noise ratio (S/N), carrier-to-noise ratio (C/N), thermal noise, noise figure, noise equivalent bandwidth, and weighting filters. The following is a brief definition of these commonly used noise terms.

Signal-to-Noise Ratio

Signal-to-noise ratio as defined by the International Radio Consultative Committee (CCIR) is a measurement of noise performance made on a baseband video signal. CCIR defines S/N as the difference in voltage between blanking level and peak white to the RMS noise voltage weighted by a prescribed noise filter.

$$\left[\frac{Sv}{Nv}\right]_{CCIR} = \frac{\text{Blanking to peak white}}{\text{(RMS) Noise voltage in a reference bandwidth}} \quad (2)$$

In decibels,

$$\frac{Sv}{Nv}(dB) = 20 \cdot Log\left[\frac{Sv}{Nv}\right] \quad (3)$$

It should be noted here that other institutions such as EIA and Bell Telephone Laboratories (BTL) define signal-to-noise ratio differently than CCIR. However, for this discussion only the CCIR's definition of S/N ratio will be presented.

Carrier-to-Noise Ratio

Carrier-to-noise ratio as defined by the National Cable Television Association (NCTA) is the power of the carrier signal during the sync pulse to the noise power in a bandwidth of 4 MHz. This bandwidth is used because it is approximately the bandwidth of the baseband video signal.

$$\left[\frac{Cp}{Np}\right]_{NCTA} = \frac{\text{Carrier power during sync pulse}}{\text{Total noise power in a reference bandwidth}} \quad (4)$$

In decibels,

$$\frac{Cp}{Np}(dB) = 10 \cdot Log\left[\frac{Cp}{Np}\right] \quad (5)$$

Similar to the S/N definition, the Television Allocation Study Organization (TASO) defines C/N differently than the NCTA. However, for this discussion only the NCTA's definition for C/N ratio will be discussed.

Thermal Noise

Thermal noise is due to the random fluctuation of electrons in any conducting medium whose temperature is above absolute zero. Thermal noise is also referred to as white noise because it has been shown experimentally and theoretically to have a uniform spectrum up to frequencies on the order of 10^{13} Hz. In a CATV system, the level of thermal noise with respect to the signal or carrier determines the amount of visible snow viewed on a television receiver. As the definition implies, a resistor is a thermal noise source, with the level of noise being dependent on the physical temperature of the resistor. By definition of noise figure a reference temperature of 290°K will be used for determining the level of thermal noise generated by a resistor. The mean squared noise voltage at the output of a resistor R can be shown to be,

$$n_e^2 = 4 \cdot R \cdot B \cdot T \cdot K \quad (6)$$

where:

- n_e = the RMS noise voltage
- R = resistance of circuit
- T = temperature in Kelvin, (°K = °C + 273.15)
- B = bandwidth of measurement system, 4 MHz is the bandwidth of most video systems
- k = Boltzman constant (1.38 x 10^{-23} watts/°K)

By the maximum power theorem, the maximum available noise power from the resistor is given by,

$$n_p = \frac{n_e^2}{4 \cdot R} \quad \text{or} \quad n_p = K \cdot T \cdot B \quad (7)$$

Consider the following example shown in figure 2, where the resistance of R is equal to 75Ω and the temperature is equal to 290°K. After substituting the appropriate numbers into equation 6 the mean squared voltage is equal to 4.802×10^{-12} Volts^2, or

$$n_e^2 = 4 \cdot 75\Omega \cdot 4\mathrm{MHz} \cdot 1.38^{-23}(\mathrm{watts}/{}^\circ\mathrm{K}) \cdot 290{}^\circ\mathrm{K} \tag{8}$$

After applying the maximum power transfer theorem to the 4.802×10^{-12} Volts^2 noise source and converting to dB referenced to a millivolt (dBmV) the available noise power from the resistor equals -59.21 dBmV.

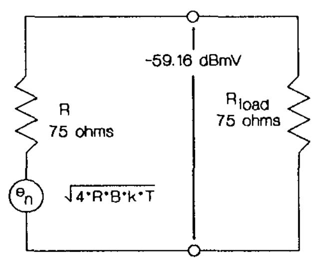


Fig 2. 75Ω Thermal Noise Source

Noise Figure

The noise figure (NF) of a two-port network gives a measure of the degradation of the S/N or C/N between the input and output ports. Noise figure compares the ratio of the carrier power to noise power at the input, to the carrier-to-noise power at the output, or expressed in dB

$$NF = 10\mathrm{Log}\left[\frac{\frac{Cp(in)}{Np(in)}}{\frac{Cp(out)}{Np(out)}}\right] \tag{9}$$

By definition, the input noise power, N_p(in), in equation 9 is equivalent to the thermal noise power generated by a resistor matched to the input and at a temperature of 290°K. In other words the input noise power is equal to, $K \cdot 290 \cdot B$. The noise figure of an ideal device be it a television converter, amplifier, or other circuit would equal 0 dB. As the converter or device introduces unwanted noise the noise figure will increase to some power ratio greater than 0 dB.

Consider a cable television converter that has a noise figure of 12dB. A 12dB noise figure will increase the output noise floor by 12dB over the input thermal noise floor. In other words there will be a 12dB degradation of the noise performance at the output of the measured converter.

Equivalent Noise Bandwidth

When flat broadband noise is transmitted through a communication system, the total noise power at the output becomes a function of the bandwidth and shape of the system's transfer function. Consider a filter whose transfer function can be related by the mean square voltage A_v^2. The frequency response of this filter is shown by the solid line in figure 3. If the input to this filter is white noise with mean-square voltage V_{in}^2/Hz, the corresponding output voltage in a 1-hertz interval at frequency f is,

$$\frac{V_{out}^2}{Hz} = \left|A_v(f)\right|^2 \cdot V_{in}^2 \tag{10}$$

Integration over the entire frequency band yields,

$$\int_0^\infty V_{out}^2(f)\, df = V_{in}^2 \cdot \int_0^\infty \left[\left|A_v(f)\right|\right]^2 df \tag{11}$$

The dashed line in figure 3 represents the equivalent noise bandwidth resulting from the integration. Basically, the equivalent noise bandwidth is the value that gives equal areas under the solid and dashed curves such that,

$$B = \frac{\int_0^\infty \left[\left|A_v(f)\right|\right]^2 df}{\left[\left|A_m\right|\right]^2} \tag{12}$$

Understanding equivalent noise bandwidth is essential when taking into account vestigial filters and baseband weighting filters in the noise analysis of a television system. Consider the unified baseband weighting filter shown in figure 5. The equivalent noise bandwidth of this filter integrated from 0 Hz to 4.2 MHz is equal to 0.881 MHz. This filter will reduce the level of broadband thermal noise power, which is flat vs. frequency in a 4.2 MHz video bandwidth by 6.8 dB.

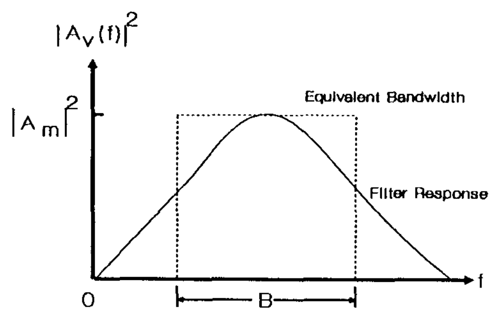


Fig. 3 Filter and Noise Equivalent Bandwidth

Weighting Filters

Studies have shown that the perceptibility of noise in television pictures varies with frequency. Noise at high frequencies is less noticeable to the viewer than noise at low frequencies. Weighting filters are intended to reduce the noise at higher frequencies in the same proportion as the noise perception to the viewer is reduced. The shape was determined by introducing controlled amounts (at different frequency bands) of noise into television pictures and having a group of viewers rate degradation.

Several institutions such EIA and CCIR have adopted weighting filters. The most commonly used weighting are those adopted by CCIR. The weighting filter of CCIR recommendation 423-1 is shown in figure 4. This was developed for System M (NTSC). Other weighting have been developed for other systems such as PAL. More recently, a weighting filter referred to as the CCIR Unified was adopted which is somewhat of a compromise between the original NTSC weighting and that of other systems. This weighting is shown in figure 5. It is intended for use in both NTSC and other systems. All of these comments apply to the luminance signal and do not consider chrominance noise.

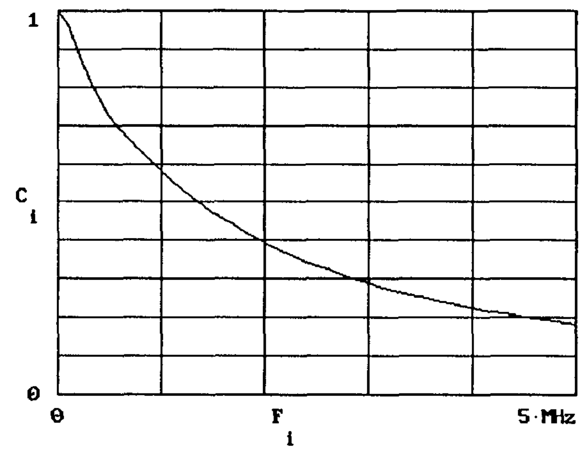


Fig 4. Weighting Filter, CCIR 423-1

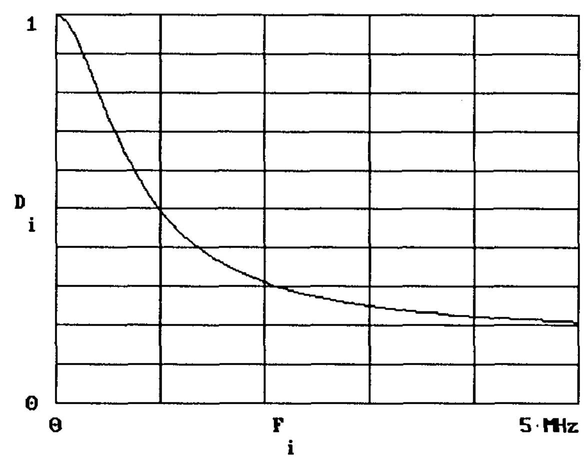


Fig 5. Weighting Filter, Unified

SIGNAL-TO-NOISE TO NOISE FIGURE DERIVATION

Noise Figure Expanded

It was previously shown that the noise figure of an amplifier is a measure of the ratio of carrier power to noise power at the input, to the carrier-to-noise power at the output. In terms of voltage in dB, noise figure can also be expressed as,

$$NF = 20 \cdot Log \left[\frac{\dfrac{Cv(in)}{Nv(in)}}{\dfrac{Cv(out)}{Nv(out)}} \right] \tag{13}$$

After applying some well know logarithmic identities, NF can be expanded as follows,

$$NF = 20 \cdot \text{Log}\, C_V(in) - 20 \cdot \text{Log}\, N_V(in) - 20 \cdot \text{Log}\left(\frac{C_V(out)}{N_V(out)}\right) \quad (14)$$

The term $20 \cdot \text{Log} C_V(in)$ represents the input signal level in dBmV, and $20 \cdot \text{Log} N_V(in)$ represents the input noise voltage in dBmV. Equation 14 is valid only if the input noise is exclusively due to thermal noise at 290°K. Since we are considering a 75Ω system, the thermal noise power deliverable by a 75Ω resistor at 290°K is equal to -59.21 dBmV. After replacing 20 Log $N_V(in)$ in equation 14 with -59.21 dBmV, NF becomes,

$$NF = 20 \cdot \text{Log}\, C_V(in) + 59.21\ (dBmV) - 20 \cdot \text{Log}\left(\frac{C_V(out)}{N_V(out)}\right) \quad (15)$$

The third term in equation 15, $20 \cdot \text{Log}\{Cv(out)/Nv(out)\}$, represents the output C/N ratio of the converter under test. However, when analyzing baseband television converters the output C/N ratio is not readily available. Again with a little mathematical manipulation C/N ratio can be expressed in terms of S/N ratio.

S/N to C/N Relationship

The relationship between RF carrier-to-noise ratio and baseband video signal-to-noise ratio can be shown by expanding and rearranging terms in the detected AM wave equation. Before expanding the detected wave equation lets first consider a carrier that has been amplitude modulated by a baseband video signal. A simplified expression showing the time domain representation of a video modulated RF carrier is shown as,

$$V_{AM} = V_c(1 - D + D \cdot \text{Cos}\, w_m t)\cos w_c t \quad (16)$$

where:

V_c = amplitude of carrier
w_c = frequency of carrier
w_m = frequency of video signal
D = modulation ratio

The modulation ratio for a video modulated signal expressed in percent, is referred to as depth of modulation (DOM). DOM is the difference between the maximum and the minimum level of the RF envelope amplitude expressed as a percentage of the maximum RF envelope level, or

$$DOM = \frac{A - B}{A} \cdot 100 \quad \% \quad (17)$$

Where A in figure 6 is equal to the peak-to-peak amplitude of the RF carrier and B is equal to the peak-to-peak amplitude of the RF carrier during white level modulation. Therefore, the modulation ratio, D, in equation 16 is simply equal to:

$$D = \frac{A - B}{2 \cdot A} \quad (18)$$

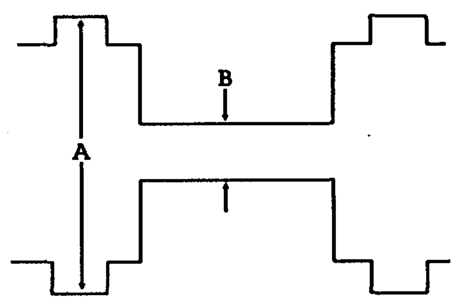


Fig 6. Video Depth Of Modulation

The predominate noise in most RF systems is usually generated at RF. The sidebands of this broadband noise with respect to the carrier are uncorrelated. The uncorrelated noise voltage sidebands will add in a RMS fashion upon detection. The following equation shows the AM wave equation with constant amplitude broadband noise added:

$$V_{AM} = V_c(1 - D + D \cdot \text{Cos}\, w_m t)\text{Cos}\, w_c t + V_n \cdot \text{Cos}[w_n t + \alpha(t)] \quad (19)$$

Where, V_n is the noise voltage and $\alpha(t)$ is the phase angle of the noise which varies randomly from 0 to 2π. This noise model expresses noise voltage in a narrow band of fixed sinusoidal amplitudes but variable phase.

Now consider synchronously detecting the video information and noise in equation 19. This is performed by multiplying equation 19 by a frequency and phase coherent signal equal to Cos w_c, or

$$V_{det} = \{V_c(1 - D + D \cdot \text{Cos } w_m t)\text{Cos } w_c t + V_n \cdot \text{Cos}[w_n t + \alpha(t)]\} V_o \cdot \text{Cos } w_c \tag{20}$$

After low pass filtering, V_{det} becomes

$$V_{det} = \frac{V_c V_o}{2}(1 - D) + \frac{V_c V_o D}{2} \cdot \text{Cos } w_m t + \frac{V_n V_o}{2} \cdot \text{Cos}((w_c - w_n)t + \alpha(t)) \tag{21}$$

The first term in equation 21 represents a DC term which can be ignored in the S/N to C/N translation. The second term represents the detected peak signal voltage. At peak modulation when Cos $w_m t$ is equal to negative one, the detected peak-to-peak signal can be shown as,

$$S_v = V_c \cdot V_o \cdot D \tag{22}$$

Since we are dealing with video modulation the detected peak-to-peak signal during peak white modulation is defined from sync tip to peak white. Recall that when expressing S/N ratio, CCIR defines signal from blanking to peak white. The CCIR signal level correction is made by multiplying the detected peak-to-peak signal, sync tip to peak white, by the ratio of blanking level to peak white with respect to sync tip to peak white. The following equation shows the correction factor ratio,

$$A = \frac{\text{Peak White - Blanking}}{\text{Peak White - Sync Tip}} \tag{23}$$

After correcting for CCIR's definition of signal, the detected signal can be expressed as:

$$S_v = V_c \cdot V_o \cdot D \cdot A \tag{24}$$

The final term in equation 21 is the detected peak noise voltage. The detected peak noise can be expressed in RMS noise density as,

$$N_{vd} = \frac{V_n \cdot V_o}{2 \cdot \sqrt{2}} \qquad \frac{\text{Volts}}{\text{Hz}} \tag{25}$$

Now that we have an expression for noise density, the total noise voltage can be determined by integrating the noise voltage over the appropriate RF bandwidth times the baseband response. In the case of a baseband converter the equivalent RF bandwidth will be determined by the Nyquist filter response shown in figure 7. Where as the baseband response will be determined by bandlimiting filters or weighting filters. For now we will presume an ideally flat baseband frequency response without any bandlimiting or weighting filters. Since the uncorrelated noise sidebands add in a RMS fashion, the equivalent noise bandwidth is determined by the following integration:

$$B_{NYQ} = \sqrt{\int_0^{f_h} NYQ(f)^2 \, df - \int_0^{-f_n} NYQ(f)^2 \, df} \tag{26}$$

Where:
- f = the frequency offset from the carrier
- NYQ = the Nyquist filter voltage frequency response
- f_n = lower Nyquist filter cutoff frequency
- f_h = upper Nyquist filter response

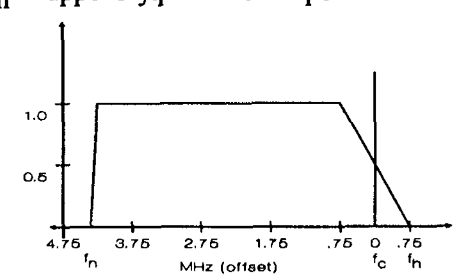


Fig 7. NTSC Nyquist Filter Response

By knowing the equivalent noise bandwidth, the total RMS noise voltage can be expressed as:

$$N_v = \frac{V_n \cdot V_o}{2\sqrt{2}} \cdot B_{NYQ} \tag{27}$$

After reducing the carrier voltage, V_c, by a factor of two, as a result of the Nyquist slope, the detected signal to noise voltage can be expressed as:

$$\frac{Sv}{Nv} = \frac{\dfrac{Vc \cdot Vo \cdot D \cdot A}{2}}{\dfrac{Vn \cdot Vo \cdot B_{NYQ}}{2 \cdot \sqrt{2}}} \tag{28}$$

Simplifying,

$$\frac{Sv}{Nv} = \frac{Vc}{Vn} \cdot \frac{2 \cdot \sqrt{2}}{2} \cdot (D \cdot A) \cdot \frac{1}{B_{NYQ}} \tag{29}$$

Notice that the first term in equation 29 represents the ratio of the carrier voltage to noise density. Since we need to derive an expression for carrier to noise power in a reference bandwidth, the noise density needs to be multiplied by a reference bandwidth, B_{ref}. Without disturbing the integrity of equation 29, S/N voltage can be rewritten as:

$$\frac{Sv}{Nv} = \frac{Vc}{Vn \cdot \sqrt{B_{ref}}} \cdot \frac{\sqrt{2}}{2} \cdot (2 \cdot D \cdot A) \cdot \frac{\sqrt{B_{ref}}}{B_{NYQ}} \tag{30}$$

Expressing S/N voltage as S/N power in dB, equation 30 can be rewritten as:

$$\frac{Sp}{Np} = 20 \cdot \mathrm{Log}\left[\frac{Vc}{Vn \cdot \sqrt{B_{ref}}} \cdot \frac{\sqrt{2}}{2} \cdot (2 \cdot D \cdot A) \cdot \frac{\sqrt{B_{ref}}}{B_{NYQ}}\right] \tag{31}$$

Notice that the first term in equation 31, $V_c/(V_n\sqrt{B_{ref}})$, represents the carrier-to-noise ratio as defined by the NCTA.

S/N to Noise Figure Relationship

As previously shown noise figure can be expressed as,

$$NF = 20 \cdot \mathrm{Log}\, C_v(in) + 59.21\ (dBmV) - 20 \cdot \mathrm{Log}\left(\frac{C_v(out)}{N_v(out)}\right) \tag{32}$$

Also recall that the expression $20 \cdot \mathrm{Log}\{Cv(out)/Nv(out)\}$ needs to be expressed in terms of S/N ratio. Therefore, after solving for C/N in equation 31, C/N can be expressed as

$$\frac{Cp}{Np} = \frac{Sp}{Np} - 20 \cdot \mathrm{Log}\left[\frac{\sqrt{2}}{2}\right] - 20 \cdot \mathrm{Log}(2 \cdot D \cdot A) - 20 \cdot \mathrm{Log}\left[\frac{\sqrt{B_{ref}}}{B_{NYQ}}\right] \tag{33}$$

Substituting C/N in equation 32,

$$NF = 20 \cdot \mathrm{Log} Cv(in) + 59.21\ (dBmV) - \frac{Sp}{Np} + 20 \cdot \log\left[\frac{\sqrt{2}}{2}\right] + 20 \cdot \mathrm{Log}(2 \cdot D \cdot A) + 20 \cdot \mathrm{Log}\left[\frac{\sqrt{B_{ref}}}{B_{NYQ}}\right] \tag{34}$$

Now let's consider a detected NTSC video signal where the modulation ratio, D, is equal to .4375 and the CCIR correction factor, A, is equal to .7142. The noise power will be measured in a reference bandwidth of 4 MHz and a Nyquist equivalent bandwidth of 3.8 MHz. The noise figure of this signal can be shown as:

$$NF = 20 \cdot \text{Log}\, C_V(in) + 59.21\,(dBmV) - \frac{S_p}{N_p} - 3\,(dB) - 4.08\,(dB) + .222 \quad (35)$$

Simplifying the noise figure expression:

$$NF = C_p(in)(dBmV) + 59.21(dBmV) - S/N(dB) - 6.86(dB) \quad (36)$$

where:

$C_p(in)$ = input carrier power level into the converter
S/N = unweighted signal-to-noise ratio of baseband video signal
Noise BW = 4.0 MHz
Reference Impedance = 75Ω

Conclusion

It has been shown that noise figure can be determined from the baseband signal-to-noise ratio measurement. However, the reader must be cautioned that the numbers used in the S/N ratio to noise figure translation are dependent on several factors. Factors such as depth of modulation, Nyquist slope, reference bandwidth, and the level of thermal noise will influence the accuracy of the conversion factors. Care must be taken when measuring baseband signal-to-noise in order to have reliable and repeatable results.

REFERENCES

1. T.M. Straus. "The relationship between the NCTA, EIA, and CCIR definitions to signal-to-noise ratio" NCTA 1974 Technical Proc.

2. Krauss, Herbert L., Charles W. Bostian, and Frederick H. Raab. Solid State Radio Engineering. New York: John Wiley, 1980

3. Alley, Charles L., Kenneth W. Atwood. Electronic Engineering. New York: John Wiley, 1973

4. CCIR. "Recommendations And Reports Of The CCIR", 1978

5. Robinson, Gerald H. Private communication and internal document.

6. Farmer, Jim. Private communication

Digital Audio for NTSC Television

Craig C. Todd
Dolby Laboratories
San Francisco

0. Abstract

A previously proposed method of adding a digital carrier to the NTSC broadcast channel was found to be marginally compatible with adjacent channel operation. The technique also has some problems unique to broadcasters. Digital transmission techniques are reviewed, and a new set of digital transmission parameters are developed which are thought to be optimum for digital sound with NTSC television.

I. Introduction

It is very feasible to compatibly add digital audio to the NTSC television signal as carried on cable television systems. Besides the marketing advantage which can accompany the use of anything "digital", there are some real advantages to digital broadcasting, especially where the transmission path is imperfect. Digital transmission is inherently robust. While the coding of high quality audio into digital form theoretically entails a loss of quality, there is no further loss of quality when the digits are transmitted through an error-free channel. This is in stark contrast to the transmission of analog audio, in which a "perfect" channel is required to avoid degradation. Unfortunately, the intercarrier sound channel used with broadcast television is somewhat less than perfect, and inherently limits the sound quality of the analog BTSC stereo system.

Digital transmission techniques have matured to the point where they can be economically applied to broadcast audio. Previous work[1,2] led to a proposal for the addition of digital audio to the NTSC television broadcast signal. That proposal borrowed heavily from work performed in Sweden[3] and Finland where a 512 kb/s QPSK carrier was extensively tested with PAL system B. The similarities between B-PAL and M-NTSC indicated that the Scandinavian test results would apply in the U.S. Our original 1987 proposal was:

A. QPSK carrier with alpha=0.7 filtering.

B. Carrier frequency 4.85 MHz above video carrier.

C. Carrier level -20 dB with respect to peak vision carrier level.

Compatibility testing of that system has been performed, and some television sets have been found on which the data carrier causes detectable interference to the upper adjacent video channel in a clean laboratory setting. It should be noted that with these problem sets, the FM aural carrier also caused noticeable interference to the upper adjacent picture. The interference from data occurs into luminance, and manifests itself as additive noise between approximately 1 MHz and 1.4 MHz. The noise level in a problem TV is subjectively similar to the noise level resulting from a video carrier-to-noise ratio (CNR) of approximately 47 dB. While this level of interference may not be detectable given the current quality level of cable signals in consumers' homes, the potential of optical fibre to raise the quality level of cable TV warrants a reduction in the level of interference by approximately 6 dB to be safe.

The system should be compatible with broadcast television as well as cable television. We must consider broadcast problems which can result from existing transmitter plant, and broadcast spectrum allocations. These considerations also call for some changes to the previously proposed system.

II. Digital Modulation Basics

Figure 1 shows the spectrum of a random data stream with a data rate of 250k bits/sec. In a digital transmission the data is always intentionally scrambled so that the transmitted data appears random. Note the spectral nulls at multiples of the data rate (250 kHz, 500 kHz, etc.). From Nyquist sampling theory we know that all information in this signal may be gleaned from the first 125 kHz, i.e.

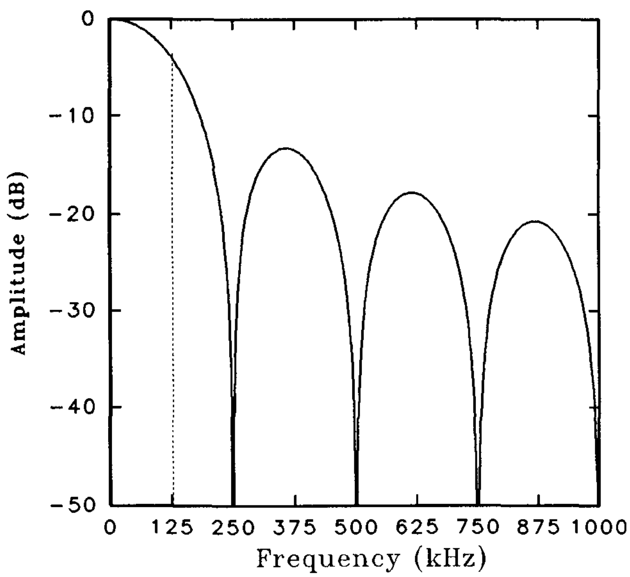


Figure 1 Spectrum of 250 kb/s data

the Nyquist frequency which is ½ of the 250 kHz sample rate. If we brick wall filter this signal at the Nyquist frequency of 125 kHz and use it to modulate a carrier, we will have a double sideband modulated bandwidth of 250 kHz, which will carry 250k bits/sec of data (1 bit/sec/Hz). If we add a second similar channel at the same frequency but in quadrature, we will double the information in the 250 kHz bandwidth RF signal to 500k bits/sec, or 2

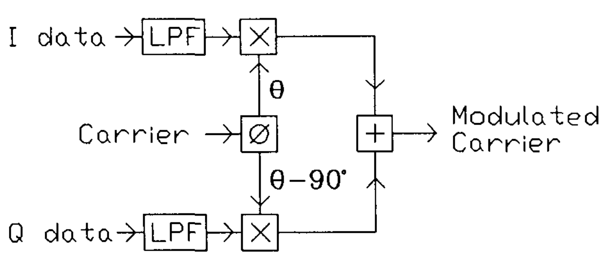


Figure 2 QPSK Modulator

bits/sec/Hz. Fig. 2 shows a block diagram of this modulator which is simply a pair of double sideband AM modulators in quadrature. The double sideband modulator can be considered to be a phase modulator, generating phases of 0 and 180 degrees. The second modulator can be considered to generate phases of 90 and -90 degrees. The combination of the two phases will generate one of four phases, ± 45 and ± 135 degrees. This form of modulation is typically called **Q**uadrature **P**hase **S**hift **K**eying or **QPSK**.

Fig. 3 shows the constellation diagram for QPSK. There are four possible states of the RF carrier phase. We refer to a chosen state as a symbol. Since there are four possible states per symbol we are transmitting two bits per symbol with a symbol rate of 250k symbols/sec, for a total data rate of

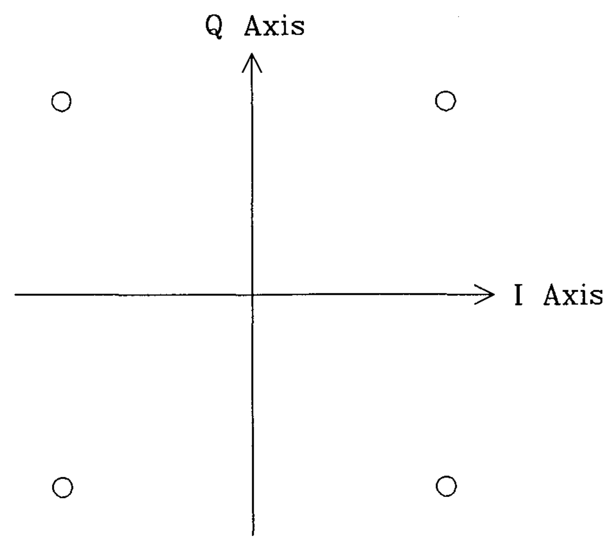


Figure 3 QPSK Constellation

500k bits/sec. If the low pass filters were not present, the phase would be instantaneously jumping between the four states shown, at a rate of 250 kHz. In practice the actual signal will be continuously traversing between these points and the receiver will sample the signal at the time when the signal passes through the constellation points.

The QPSK receiver is shown in Fig. 4. A carrier recovery loop coherently regenerates the transmitted carrier, which was suppressed in the modulator. Quadrature mixers demodulate the I and Q signals, low pass filters limit the effective receive bandwidth,

$$H(j\omega) = \begin{cases} 1, & \left[0 \le \omega \le \frac{\pi}{T_s}(1-\alpha)\right] \\ \cos^2\left\{\frac{T_s}{4\alpha}\left[\omega - \frac{\pi(1-\alpha)}{T_s}\right]\right\}, & \left[\frac{\pi}{T_s}(1-\alpha) \le \omega \le \frac{\pi}{T_s}(1+\alpha)\right] \\ 0, & \left[\omega \ge \frac{\pi}{T_s}(1+\alpha)\right] \end{cases} \quad (1)$$

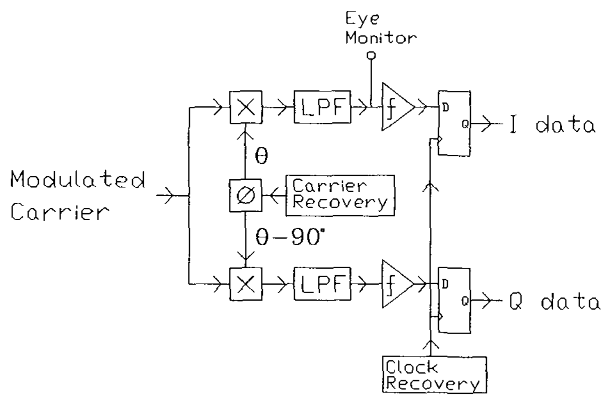


Figure 4 QPSK Demodulator

and comparators and clocked latches recover the data. The latches are driven by a symbol timing recovery circuit which reconstructs the transmit data clock at the proper phase so that the sampling time is correct.

While QPSK can theoretically achieve a spectral efficiency of 2 bits/sec/Hz, this is not achievable in practice since a brick wall filter is not realizable. The steepness of the filter determines the spectral efficiency. Steep filters ring, and as the spectral efficiency of QPSK is raised by sharpening the filters, the ringing of the individual data pulses increases. This ringing can cause the pulses to interfere with each other (intersymbol interference) unless the ringing is controlled. If the filtering is done properly, each individual pulse waveform, except for the pulse being detected, will pass through zero at the instant the receiver samples the waveform. Nyquist filters provide the desired characteristics. A commonly used class of Nyquist filters are the "raised cosine" filters which have a frequency response as described in Eq. 1, and shown in Fig. 5. The alpha term determines the fractional excess bandwidth over a perfect brickwall (alpha=0) filter. A filter meeting the amplitude response of Eq. 1, and having constant group delay (linear phase) will have no intersymbol interference.

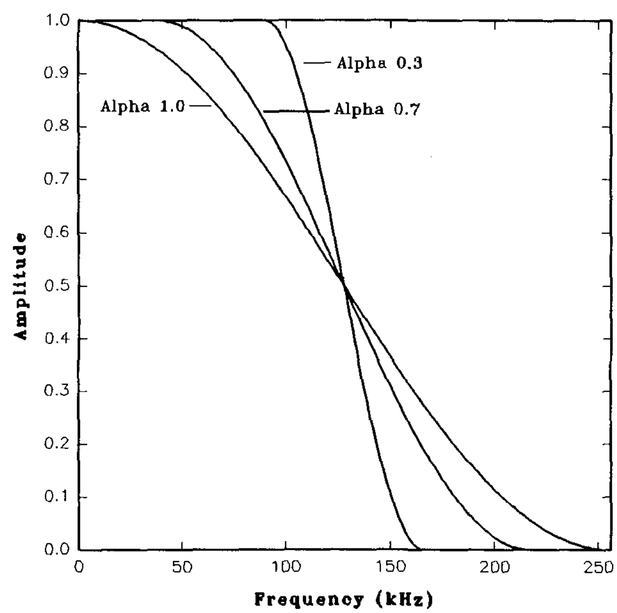


Figure 5 Raised Cosine Nyquist Filters

When we speak of the channel filter, we mean the filter function of the complete channel. This includes the transmit baseband lowpass filter, any transmit RF filtering, receiver preselection and IF filtering, and receive baseband lowpass filtering. It also includes the effect of the transmission path which might have a non-flat response (or selective fade). Typically, all of the intended filtering is designed into the system at IF or baseband, and all other circuitry is made wideband. The most precise control over filtering is achieved with baseband lowpass filtering. It is much easier to make a precision 125 kHz lowpass filter than a precise 250 kHz IF filter.

$$QPSK\ Spectral\ Efficiency = \frac{2}{1+\alpha}\ bits/Hz \qquad (2)$$

As alpha is reduced, spectral efficiency of QPSK is improved. Equation 2 shows the relationship between spectral efficiency and alpha. As alpha is reduced, the data waveform will have more ringing and the "eye" pattern will become more complex. The eye pattern is observed at the point where the analog waveform is sampled to recover the data. In the QPSK decoder block diagram (Fig. 4), the eye monitor point is shown for the I channel, and is just after the baseband lowpass filter. In Fig. 6 we see some examples of eye patterns for alpha=0.3, alpha=0.7, and alpha=1.0.

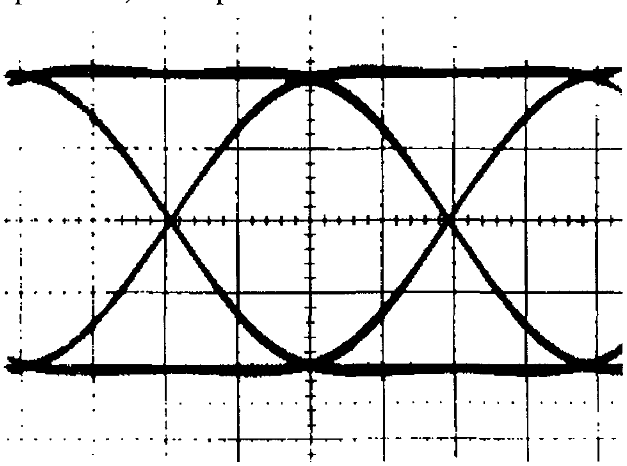


Figure 6a Eye Pattern, Alpha = 1.0

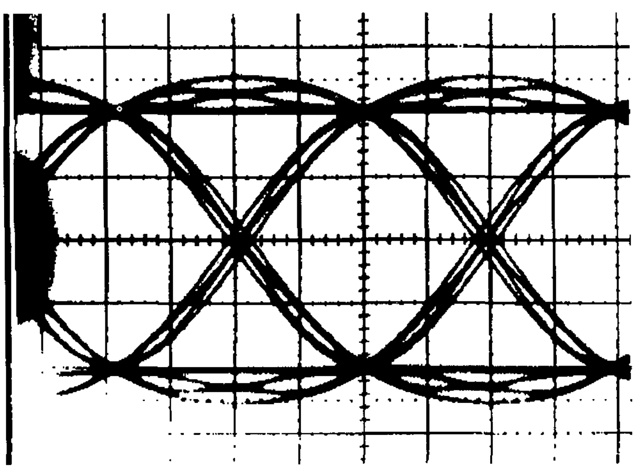


Figure 6b Eye Pattern, Alpha = 0.7

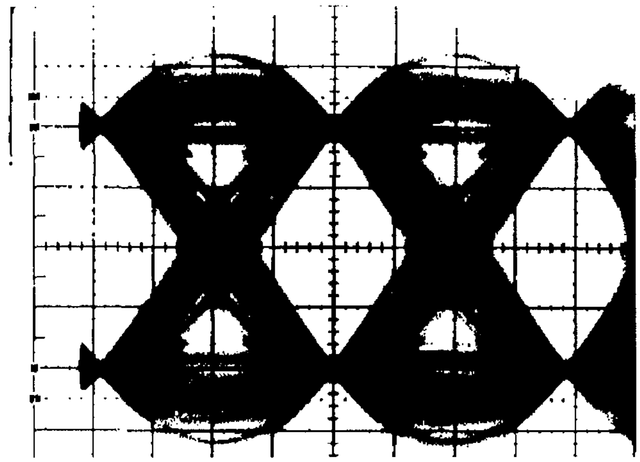


Figure 6c Eye Pattern, Alpha = 0.3

The eye patterns become more complex as alpha is lowered and spectral efficiency is improved. More accuracy is required in implementation of low alpha systems. Any error in amplitude or phase linearity will rapidly degrade a complex eye. The eye will appear to close. Fig. 5d shows an alpha = 0.3 eye with a lot of closure due to poor filtering. Partial eye closure due to imperfect filtering leaves less margin against noise and interference.

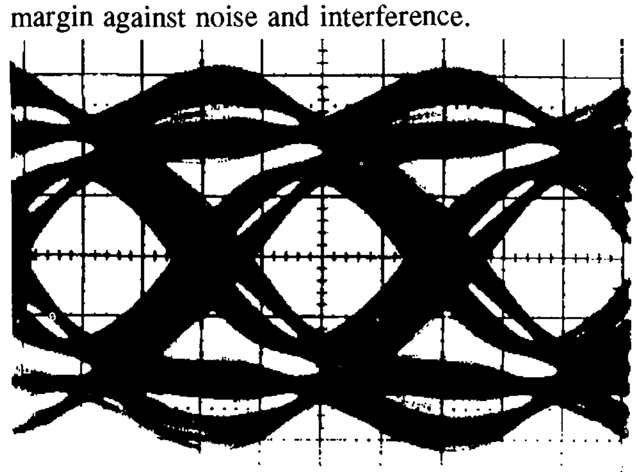


Figure 6d Alpha = 0.3, Poor Filter Accuracy

For optimal performance over a noisy channel, the amplitude portion of the filtering should be partitioned in equal portions between the transmitter and receiver i.e. the transmit filter and receive filters should both have an amplitude response equal to the square root of the nyquist filter response. The cascade of the two filter characteristics will then be a Nyquist filter. The combination of the filters should have linear phase (constant group delay). If both the transmit and receive filters individually have linear phase, the combination will as well. However, it is also acceptable for the receive filter to have group delay variations and to compensate for these in the transmit filter.

Attempting to achieve good spectral efficiency with low alpha filtering requires the use of more accurate filters with good phase compensation. Filters may be implemented with either analog or digital circuitry. Since we are starting with digital data, the use of a digital filter in the transmit side is attractive. The receiver is more easily implemented with an analog filter which does not require the use of A-D converters with high sample rates and anti-aliasing pre-filters. In a broadcast application we want to keep the receiver cost low, so we want to use the minimal receive filter (without phase compensation). If the receive filter magnitude response accurately matches the square root of a Nyquist response, then the transmit filter can be implemented as an FIR (finite impulse response) filter with an impulse response which is simply the time reversal of the impulse response of the receive filter. The cascade of the filters will then have a symmetrical impulse response with constant group delay, and will have an amplitude response equal to that of a Nyquist filter. The matched filter criterion is satisfied, so performance in the presence of noise will be optimal. All phase compensation is done in the transmit filter so receiver cost is minimized.

In order to achieve spectral efficiencies of 2 bits/sec/Hz or greater, QPSK must be abandoned and a higher level modulation method used. This can be done by hitting the modulator filters (Fig. 2) with multilevel data symbols. With QPSK, the I and Q data are simple binary symbols with two levels (1 and 0, or +1 volt and -1 volt). If pairs of bits are fed to 2 bit D-A converters, we can form four level data (0, 1, 2, 3; or +3 volts, +1 volts, -1 volts, -3 volts). QPSK extended to 4 level symbols is known as 16 QAM (Quadrature Amplitude Modulation). 16 QAM has a theoretical (with ideal brick wall filtering) spectral efficiency of 4 bits/sec/Hz. Unfortunately, the added spectral efficiency is achieved at the expense of ruggedness. The eye pattern of 16 QAM is similar to that of three QPSK eyes stacked on top of one another. For the same absolute eye opening (which would give the same performance in a noise
or interference environment), the peak level is 3 times higher, or about 9.5 dB. For the same carrier level, the 16 QAM signal is 9.5 dB less rugged than QPSK.

III. Partial Response Signaling

Another attractive technique which can achieve improved spectral efficiency is known as "partial response signaling". This method involves the use of a controlled amount of intersymbol interference that is removed by an extra processing step in the receiver. The simplest form of partial response involves performing a running average of adjacent pairs of bits before driving the modulator. If we average a pair of bits which can each be +1 or -1 we get symbols with three levels: +2, 0 or -2. This 3-level signal still only contains 1 bit of information per baseband symbol, but the spectral characteristics are changed. The spectrum of the data is changed

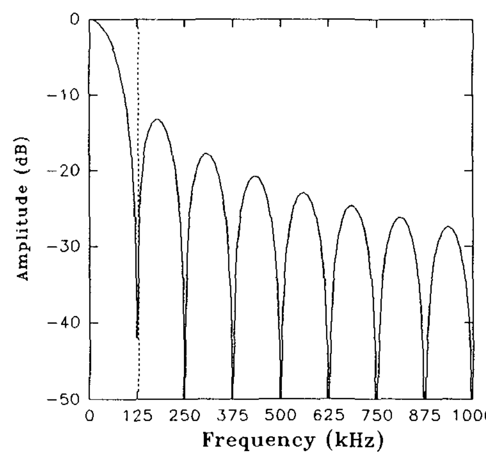


Figure 7 Partial Response Spectrum

from that shown in Fig. 1 to that shown in Fig. 7. The first spectral null has moved from 250 kHz down to 125 kHz. If we filter off all of the energy beyond 125 kHz (which is relatively easy), we will achieve a spectral efficiency of 1 bit/sec/Hz at baseband. The use of a pair of these partial response system modulators in quadrature is known as QPRS. The constellation diagram for QPRS is shown in Figure 8.

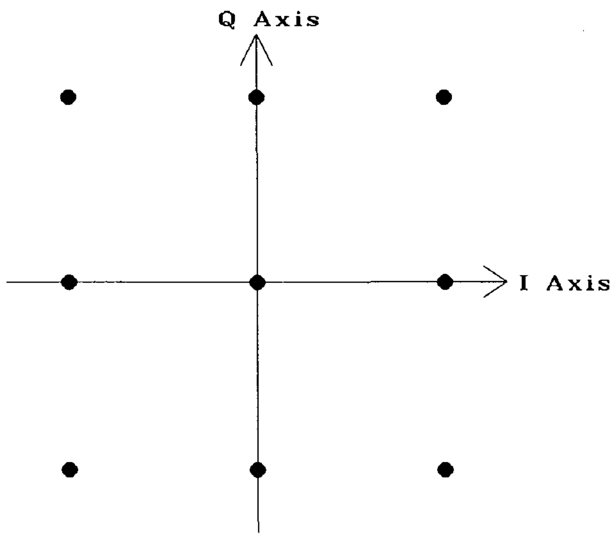


Figure 8 QPRS Constellation

The pairwise bit averaging is equivalent to a filter with an amplitude response equal to a cosine function, as shown in Fig. 9. In order to have matched filtering, the bit averaging should not be done at the transmitter, but the cosine filter function should be divided equally between the transmit and receive filters. That is, both the transmit filter and the receive filter should have an amplitude response equal to the square root of the first quadrant of a cosine. The transmit filter may be a digital FIR filter with an impulse response which is the time reversal of that of the receive filter.

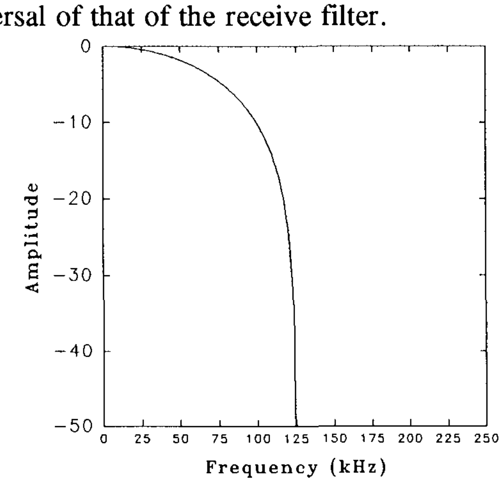


Figure 9 PRS Cosine Filter

Fig. 10 shows the spectrum of the transmitted data carrier for several modulation schemes. The narrowest spectrum is for QPRS, the second narrowest is alpha=0.3 QPSK, and then we have

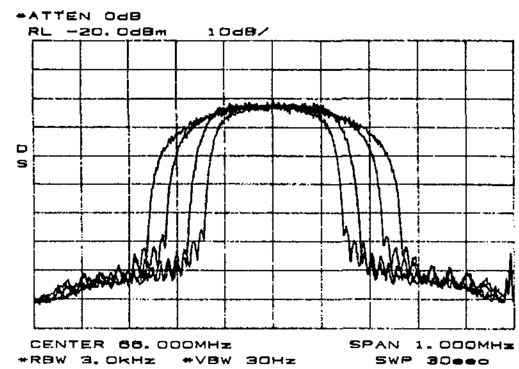


Figure 10 Comparative RF Spectra

alpha=0.7 and alpha=1.0. The transmit filters used are FIR filters with a magnitude response equal to the square root of the raised cosine curve (for the QPSK cases), or the square root of a cosine curve (the QPRS case). The spectrum of QPRS is very attractive, but its eye pattern (Fig. 11) shows it to be less rugged than QPSK. Since there are two eyes, in order to achieve the same absolute eye opening (and thus the same ruggedness), the peak power must be increased by a factor of two, or 6 dB. For the same peak power, QPRS should be 6 dB less rugged. There are some mitigating factors though.

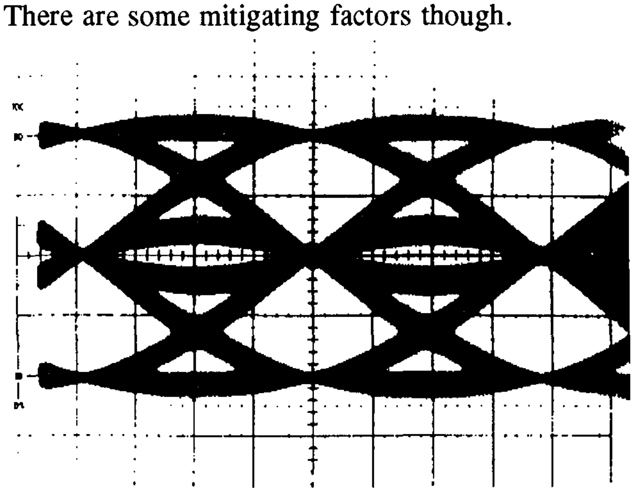


Figure 11 PRS Eye Pattern

First, the PRS data stream spends half its time at the 0 level (data = 1,0 or 0,1), and half at the +2 or -2 level (data = 0,0 or 1,1). This means that half of the time the carrier is modulated by zero, and half of the time it is modulated by the peak value. QPRS will have an RMS power level half that of QPSK for the same peak level. For the same RMS power level, QPRS will thus lose only 3 dB of ruggedness compared to QPSK, and not the 6 dB that is lost with equal peak power levels. Second, since there is some redundancy in the PRS 3-level data stream, it is possible to detect and correct some errors without intentionally including redundant FEC data in the data multiplex. Coding gains of 1.5 to 2.0 dB can be achieved economically. The combination of these factors indicate that for comparable carrier-to-noise ratios (CNR), QPRS is only about 1 to 1.5 dB less rugged than QPSK against gaussian noise. QPRS is an attractive candidate if its narrow spectrum would solve our interference problem.

While QPRS requires some added complexity in the receiver, the amount is within reason for a consumer product. If the narrowness of the QPRS signal significantly reduces the potential for interference, QPRS could be adopted for digital TV sound broadcasting without seriously impacting the receiver cost.

IV. Practical Broadcast Considerations

In the case of adjacent channel operation, the proposed data carrier is placed above the FM sound carrier of one channel, and below the lower vestigial sideband of another channel. Figure 12 shows the RF spectrum of TV Ch 3 and 4, with an Alpha = 0.3 QPSK carrier inserted between the channels. In the case of adjacent channel operation, it makes little difference whether we say that the data signal is assigned to the upper or the lower channel. The BBC[4] choose to allocate the signal to the lower channel. There are advantages to this choice.

Depending on the actual vestigial sideband filtering in the video modulator, there may be some overlap of video energy into the spectrum occupied by the data signal. This will allow video to interfere with the data, possibly causing data errors. In the case of broadcasting without adjacent channel operation, placing the data carrier at the top of the channel means that ones own video will not interfere with ones own data.

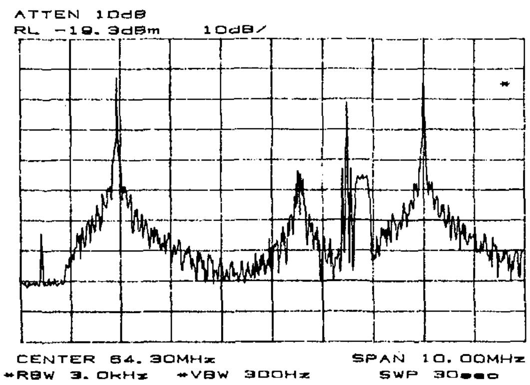


Figure 12 Ch3, Data, Ch4 Spectrum

There is also an advantage in receiver complexity. If the first sound IF is widened slightly, the inter-carrier mixer used to recover the 4.5 MHz FM sound will also recover the data sound carrier at 4.85 MHz. This point can be tapped and fed to the QPSK demodulator. If the data carrier is associated with the upper channel, the inter-carrier offset will be -1.15 MHz, and additional IF filtering and another inter-carrier mixer will be required. Filtering will have to be sufficient to reject the image at 1.15 MHz above vision. The choice of placing the carrier above the channel is a good one, and we originally agreed with it.

Unfortunately, there are some serious problems with this choice for broadcasters in the U.S. The most serious is that in the case of TV Ch. 6, the data carrier would lie at 88.1 MHZ, making both 88.1 and 88.3 unusable for FM broadcasting. The second is that many transmitter installations are unable to easily broadcast the signal due to the notch diplexers in use. Stations which use separate amplification for sound and vision would introduce the signal into the sound transmitter. The sound and vision transmitters are combined in a device known as a "notch diplexer" in which tuned cavities (creating notches) reflect the sound signal into the antenna. There are no cavities tuned to the frequency of the data signal, and the diplexer would need expensive modifications in order to add them. While this is not a technical problem, it is an economic one and could seriously hamper the acceptance of the digital system by broadcasters.

Another problem, unique to cable, has to do with sync suppression scrambling. When sync information is carried on the FM sound carrier as pulse modulation, the spectrum of the FM sound carrier is widened and overlaps the data spectrum. The presence of the data may cause misbehavior in the descrambler, and the presence of the sync information on the FM sound carrier may cause errors in the demodulated data.

Choosing to place the data carrier below the vision carrier appears to be the correct choice for the U.S. Broadcasters may add the signal to the vision transmitter just after the vestigial sideband filter, and it will be amplified linearly along with the picture. At this frequency there will be no problem with the notch diplexer. With the carrier below the channel, there is no problem with a scrambled channel, although the same problem will exist if the FM sound carrier on the lower adjacent channel is pulse modulated.

V. Interference of Data into Picture

Placing the data signal on the lower sideband of our own channel makes it even more critical to avoid interference into video because now we will interfere with ourselves. The mechanism of interference is imperfect VSB Nyquist filtering in the TV receiver. While it is easy to make a TV receiver reject the data signal there are a wide variety of receivers in use, some of which have poor rejection. Fig. 13 shows the ideal receiver IF filter response. Also shown is the kind of response that could be susceptible to interference. The problem is the 'tail' in the rolloff. If the filter goes right down to near zero response at 1 MHz away from the vision carrier the data signal will be fully rejected. If the filter has a tail, the data will not be rejected and may be visible in the picture.

The only way to reduce the potential interference is to: 1) reduce the data carrier level, and 2) move the data carrier farther away from the video carrier (so as to get farther down on the slope of the receive filter). The data carrier can be moved by either moving the center frequency or narrowing the data spectrum, or both. Moving the upper bandedge of

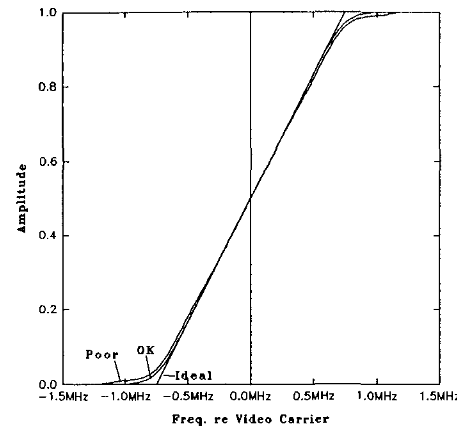


Figure 13 Receiver VSB Nyquist Filters

the data spectrum down by 100 kHz reduces the interference into the upper video picture by approximately 6 dB on problem receivers.

VI. Interference of Data into FM Sound

Our initial studies of this system concentrated on the interference of data into the BTSC sound signal. We found that with QPSK and alpha=0.7 filtering, the data carrier could be placed 350 kHz above the sound carrier. With alpha=0.3 filtering, the carrier could be moved down another 50 kHz so that it is only 300 kHz above the FM sound carrier. The mechanism for interference into FM sound is demodulation of the data signal as noise by the FM detector. The frequency of the demodulated noise is equal to the offset between the FM carrier and the data signal. Since the data signal is wideband, the demodulated noise is also wideband.

Fig. 14 shows the demodulated composite BTSC FM signal from DC to 200 kHz with no modulation on any of the BTSC channels. The 15.7 kHz pilot is clearly visible, along with L-R noise (produced by the BTSC compressor working at full gain) and the SAP carrier at 78.7 kHz. Overlaid is a plot with the data carrier turned on (using the parameters to be specified in section VIII). The data carrier causes a slightly higher noise level starting at about 90 kHz, and at 140 kHz the noise level rises abruptly. A higher noise level at these frequencies should be of no consequence in a properly designed mono, stereo,

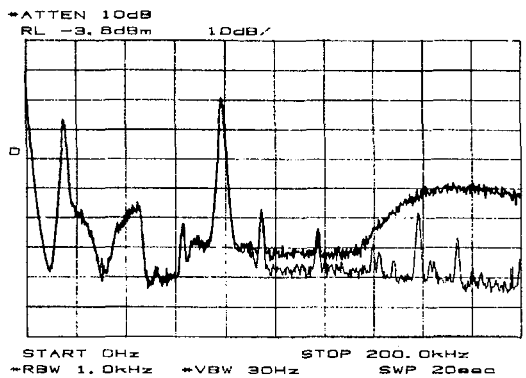


Figure 14 Spectrum from FM Demod, No Channels Modulated

or SAP decoder, and of only minor significance to a Pro channel decoder. A poorly designed decoder which inadvertently mixes components above 100 kHz down to baseband might suffer a slight degradation due to the addition of the data carrier, but this type of decoder would also suffer a premature noise degradation as CNR is reduced. Figure 15 shows the same plots, but with the Left and SAP channels fully modulated with a 400 Hz sinewave. This figure confirms that the noise induced into the composite baseband signal is similar whether the FM carrier is fully modulated or not. The presence of non-linear FM modulation does not appear cause the noise to spread down to lower audio frequencies.

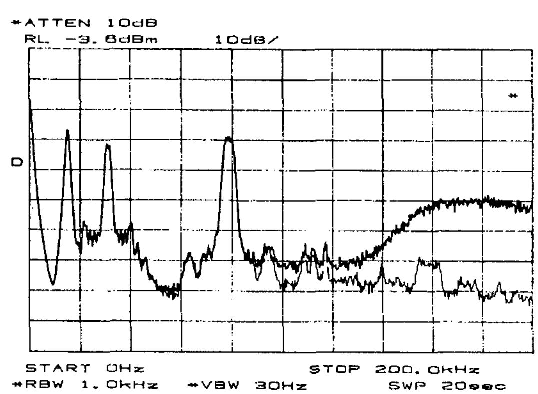


Figure 15 FM Demodulated Spectrum, All Channels Modulated

VII. Interference of Picture into Data

The system adopted in Scandinavia, which our first proposal was based on, has problems with interference of upper adjacent vision into data. The lower edge of the vestigial sideband energy extends low enough in frequency to be received by the data demodulator tuned to the lower channel. Their solution was to modify the spectral mask of allowed energy in the lower sideband. Basically, this involves changing out all VSB filters in cable modulators for new ones with the cutoff moved up approximately 200 kHz in frequency. Our original proposal would have required this change in modulator filtering as well.

When choosing new parameters to make the system more compatible with adjacent channel operation, it is very desirable to make the system operate with existing modulator filters. The steps taken to lower interference of data into picture will help to lower interference of picture into data.

VIII. A More Compatible System

Our original proposal of alpha=0.7 QPSK placed 350 kHz above the audio carrier at a -20 dB level is adequately compatible with BTSC audio, but could benefit from about 6 dB of improved compatibility with video. This could be achieved by dropping the carrier frequency by 100 kHz, but that would compromise compatibility with audio. Another way to achieve the 6 dB improvement is to narrow the carrier bandwidth as the frequency is dropped, keeping the lower data bandedge the same distance from the FM sound carrier. In theory, a 50 kHz drop in carrier frequency accompanied by a 100 kHz narrowing in bandwidth should meet our target. The narrowing could be done by staying with QPSK and changing the filtering from alpha=0.7 to alpha=0.3. The bandwidth will narrow by 0.4 x 250 = 100 kHz, and the carrier frequency can be dropped 50 kHz.

Unfortunately, when this is tried, the improvement in video interference is only about 3 dB. This is because the more tightly filtered data has a higher peak level, and the peaks are visible in the picture. The higher peak level has cost us 3 dB of the expected improvement. Changing from QPSK to QPRS, using the same RMS signal level, does not help, because QPRS inherently has a peak level which is 3 dB larger than the average level.

In making these comparisons, we have found it useful to normalize our signal levels to the nominal peak level, ignoring the overshoots. The eyes shown in Fig. 6 are normalized in this way, with a nominal peak level of $\pm$ 2 divisions. We measure the level of the modulated carrier with a static data pattern which produces a pure carrier modulated with this level. This is how we set the level in our original proposal. Using this method of measuring level, we find that we can meet our target by lowering the carrier frequency by 50 kHz and using either alpha=0.3 QPSK with a level of -23 dB, or QPRS with a level of -20 dB. These two signals will have comparable RMS levels, and using non-redundant error correction in the QPRS demodulator, similar performance in the presence of gaussian noise (the QPSK would actually be about 1.3 dB more rugged).

In order to choose which method to pursue, we have to look at the relative costs, and the performance in the presence of interfering signals. QPRS will require a more expensive decoder. Since the eye is multilevel, two comparators instead of one are required to convert from the analog waveform back to digital data. The level of the QPRS waveform at the comparator inputs is also critical, so a better AGC is required.

The data is subject to interference from both the FM audio carrier, and lower sideband visual information. The receive data filtering can adequately reject any interference from the FM carrier. Interference from video cannot be fully filtered out because it can be inband, depending on the VSB filter in the modulator. In order to improve rejection of video information, we have increased the order of the data receiver baseband low pass filtering from a 3 pole to a 5 pole filter. This change, along with the lowering of carrier frequency, allows our eyes to remain open even with the worst case interfering vision modulation present. With the worst case signal present our margin against error is reduced, but we can still operate error free.

If the QPRS and QPSK carriers were used with the identical carrier levels as defined above, the QPSK eye would start out twice as open as the QPRS eye. Since we can operate QPRS with 3 dB more level (as defined above, or at the same level on an RMS basis), the QPRS eye in the receiver will be 70% (-3 dB) the size of the QPSK eye. This smaller eye will be more susceptible to interference from lower sideband video information. Careful measurements show the -20 dB QPRS system suffers a 2 dB interference penalty compared to the -23 dB QPSK system. QPRS looses out on both circuit costs and ruggedness. Despite the attractively narrow spectrum of QPRS, QPSK is the superior solution. Therefore, should stay with QPSK.

The new QPSK carrier frequency will be 1.2 MHz below the vision carrier frequency. This value is very close to 1/3 of the NTSC chroma subcarrier frequency, which is 1.193182 MHz. This is a convenient value to choose, because it can allow future receiver circuits to use the chroma oscillator as a reference to demodulate the data, instead of a requiring a separate crystal oscillator to be locked to the incoming carrier.

The data rate may also be locked to video. We need approximately 512 k bits/sec of data to use a low cost digital audio coding method based on adaptive delta-modulation. It is desirable to use the same clock frequencies that are used in satellite implementations of this audio system such as B-MAC and HDB-MAC. Those systems place integral numbers of audio bits on horizontal lines and so have a direct relationship to video. It turns out that using 1/7 of the chroma frequency as a data clock gives us a total bit rate of 511.363 kHz. With the simplest conceivable multiplex structure, the audio clock rate will be 13 times the horizontal scan rate, which is identical to clock rate used by the BMAC systems. Using the chroma oscillator in the receiver as a data clock reference will save another crystal in the data timing recovery circuitry.

IX. The 1990 Proposal

The following seems to be the optimum set of parameters for an NTSC compatible digital carrier:

A. QPSK carrier with alpha=0.3 filtering.

B. Transmit filtering phase compensates for a specified receive filter.

C. Carrier frequency 1.193182 MHz (locked to 1/3chroma) below video carrier.

D. Carrier level -23 dB relative to peak vision carrier level.

E. Data rate 511.363 kHz (locked to 1/7 chroma).

IX. Conclusion

Minor modifications have been made to the NTSC compatible system described three years ago. We explored a number of methods of improving compatibility and settled on a new set of specifications. The new parameters improve compatibility, reduce receiver cost, and simplify transmission of the signal for both broadcasters an cable operators.

References

1. Craig C. Todd, *A Compatible Digital Audio Format for Broadcast and Cable Television,* IEEE Trans. on Consumer Electronics, Vol. CE-33, No. 3, Aug. 1987, pp. 297-305.

2. Craig C. Todd, *Digital Sound and Data for Broadcast Television - A Compatible System*, NAB Proceedings, 41st NAB Engineering Conference, 1987.

3. Anders Nyberg, *Digital Multi-Channel Sound for Television*, ICCE Digest, June 1987.

4. A. J. Bower, *Digital Two-Channel Sound for Terrestrial Television*, ICCE Digest, June 1987.

DISTORTION ACCUMULATION IN TRUNKS

Doug McEwen, CUC Broadcasting Limited
Norm Slater, Comlink Systems Inc.
Rezin Pidgeon, Scientific-Atlanta

ABSTRACT

The conventional analysis of distortion in a cable television system assumes that composite triple beat (CTB) accumulates on a voltage basis, which is a worst case situation. This assumption, although it may be valid for long cascades of identical devices, such as push-pull trunk amplifiers, may not be valid for feedforward trunk cascades. This is because feedforward amplifiers, which use distortion cancellation, may not be identical to other feedforward amplifiers in the cascade.

Test results for pairs of push-pull and feedforward CATV equipment are presented, as well as a theoretical analysis which explains these results.

INTRODUCTION

Many papers have been published discussing the cascading of CTB in cable television trunks (1-4). These papers indicate that CTB cascades with voltage (20logN) addition, the worst case possibility. CTB is generally considered to be the limiting intermodulation distortion in a cable television trunk. Consequently, the 20logN assumption has a large impact on cable system architecture and cost.

In order for 20logN addition to occur, two basic conditions must be met in an amplifier: 1) a linear phase vs. frequency response and 2) identical distortion phases for all amplifiers. Both of these criteria are met in a push-pull cascade, and, therefore, 20logN cascading results. Although feedforward amplifiers have a linear phase vs. frequency response, it is not clear that they all have equal distortion phases. Unlike the distortion phase of a push-pull amplifier, the distortion phase of a feedforward amplifier is heavily dependent on the amplitude and phase matching of signals when the distortion from the main amplifier is cancelled by signals from the error amplifier, so it is expected that distortion phase is not constant from device to device. If distortion phases vary significantly, the 20logN assumption is very pessimistic.

CASCADING OF CTB

Consider the repeater section in Exhibit 1

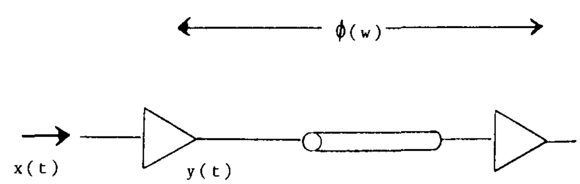


EXHIBIT 1 - Repeater Section

The input signal is x(t) and the output is y(t) which is represented by:

$$y(t) = K_1 x(t) + K_2 x^2(t) + K_3 x^3(t) + \dots$$

Note that frequency dependent non-linear characteristics are ignored. A more rigorous treatment is possible using a Volterra series, but this makes the analysis much more complex. Higher than third-order distortions are also ignored.

Let the input function x(t) consist of three amplitude-modulated sinusoidal voltages:

$$x(t) = A(t)\cos w_a t + B(t)\cos w_b t + C(t)\cos w_c t$$

The output y(t) consists of many terms, including a triple beat term:

$$3/2\, K_3\, A(t)B(t)C(t) \cos(w_a t + w_b t - w_c t + \Theta_1)$$

where Θ_1 is the phase of the triple beat in the first repeater at frequency a + b - c.

In a cable television system, many of these triple beats are formed, and their frequencies are nominally those of a video carrier frequency. The accumulation of beats is CTB.

Assume that the repeater section has a linear phase vs. frequency characteristic as is shown in Exhibit 2

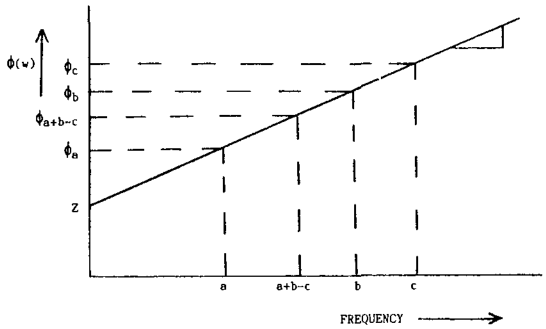


EXHIBIT 2 - Linear phase-frequency characteristic of a repeater section

The phase shift ϕ from the repeater section is:

$$\phi(w) = mw + Z$$

The signal at the output of the first repeater will include the terms:

$$\begin{aligned} &K_1 A(t)\cos(w_a t + \phi_a) \\ &+ K_1 B(t)\cos(w_b t + \phi_b) \\ &+ K_1 C(t)\cos(w_c t + \phi_c) \\ &+ 3/2 K_3 A(t)B(t)C(t)\cos \\ &\qquad (w_a t + w_b t - w_c t + \Theta_1) \end{aligned}$$

The signals at the output of the second repeater will include the terms from the first repeater:

$$\begin{aligned} &K_1 A(t)\cos(w_a t + \phi_a + \phi_a) \\ &+ K_1 B(t)\cos(w_b t + \phi_b + \phi_b) \\ &+ K_1 C(t)\cos(w_c t + \phi_c + \phi_c) \\ &+ 3/2 K_3 A(t)B(t)C(t)\cos \\ &\qquad (w_a t + w_b t - w_c t + \Theta_1 + \phi_{a+b-c}) \end{aligned}$$

It will also include a term for the triple beat generated in the second repeater:

$$\begin{aligned} &+ 3/2 K_3 A(t)B(t)C(t)\cos \\ &(w_a t + w_b t - w_c t + \Theta_2 + \phi_a + \phi_b - \phi_c) \end{aligned}$$

Comparing the phases of the two triple beats shows a difference Θ_D of:

$$\begin{aligned} \Theta_D &= \Theta_1 + \phi_{a+b-c} - (\Theta_2 + \phi_a + \phi_b - \phi_c) \\ &= \Theta_1 - \Theta_2 + \phi_{a+b-c} - \phi_a - \phi_b + \phi_c \end{aligned}$$

But, from Exhibit 2,

$$\begin{aligned} \phi_{a+b-c} &= m(a+b-c) + Z \\ &= ma + mb - mc + Z, \\ \phi_a &= ma + Z \\ \phi_b &= mb + Z \\ \phi_c &= mc + Z \end{aligned}$$

Therefore,

$$\begin{aligned} \Theta_D &= \Theta_1 - \Theta_2 + ma + mb - mc + Z \\ &\quad - ma - Z - mb - Z + mc + Z \\ &= \Theta_1 - \Theta_2 \end{aligned}$$

If Θ_1 and Θ_2 are equal, then the beats will add on a voltage basis. If Θ_1 and Θ_2 are not equal, then the beats will add with a phase difference. In a cascade of many amplifiers, this will lead to MlogN addition where M depends on the statistical distribution of phases and may vary over a wide range.

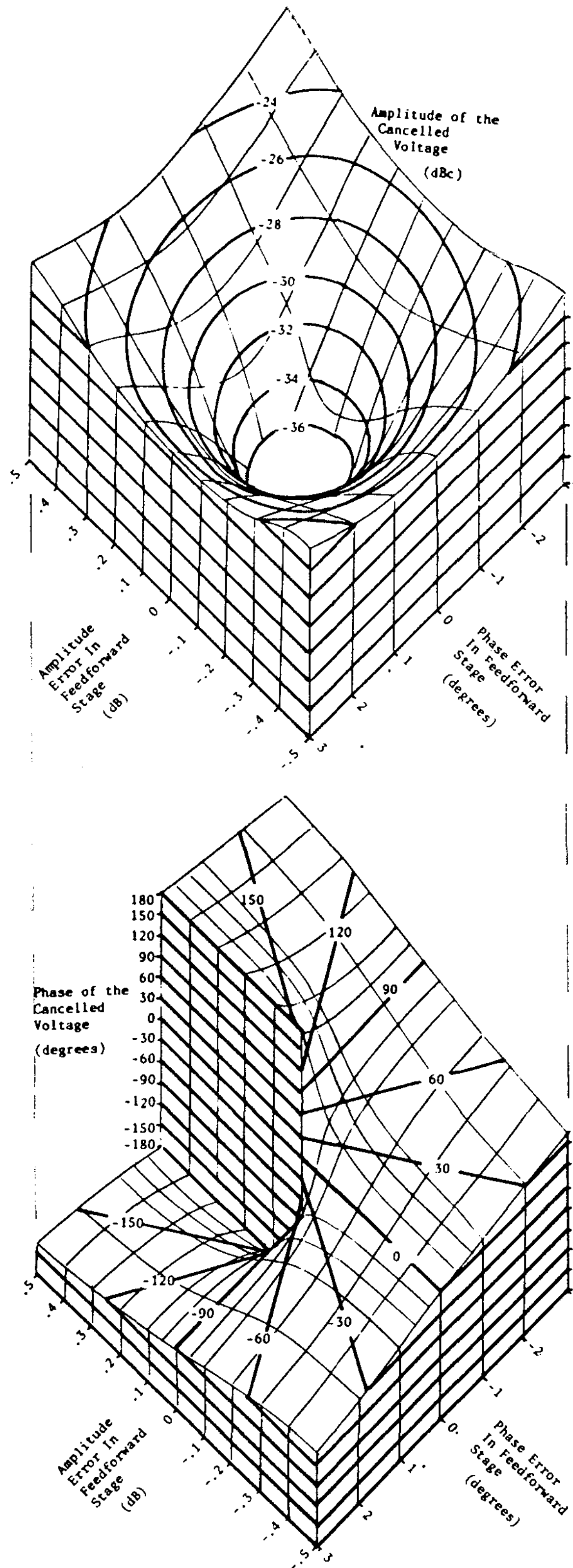
The relationship between Θ_D and MlogN for two devices in cascade is shown in Exhibit 3.

Θ_D (Degrees)	Mlog2 Addition	Type of Addition
0	20.0log2	voltage addition
30	19.0log2	
60	15.8log2	
90	10.0log2	power addition
120	0.0log2	no addition
150	-19.0log2	partial cancellation
180	-∞ log2	complete cancellation

EXHIBIT 3 - Phase Difference vs. Mlog2

FEEDFORWARD DISTORTION PHASE

Unlike push-pull amplifiers, which are generally similar to each other and which consequently generate beats with nearly equal phases, feedforward amplifiers cancel the distortion, typically with a minimum of 20dB cancellation. As is shown in Exhibit 4, small amplitude or phase changes in the feedforward amplifier cancellation circuitry can lead to large changes in the amplitude and/or phase of the resultant distortion.



Distortion Amplitude and Phase vs. Cancellation Amplitude and Phase Errors

Exhibits 5A to 5C show average CTB levels plus or minus one standard deviation as a function of temperature at 55.25, 295.25 and 547.25 MHz respectively (5). The CTB level of feedforward hybrids varies significantly from unit to unit, and varies on individual units with temperature. Temperature sensitivity is lower at 55.25 and 295.25 MHz than at 547.25 MHz. The unit to unit variation is lower at 55.25 MHz than at 295.25 and 547.25 MHz.

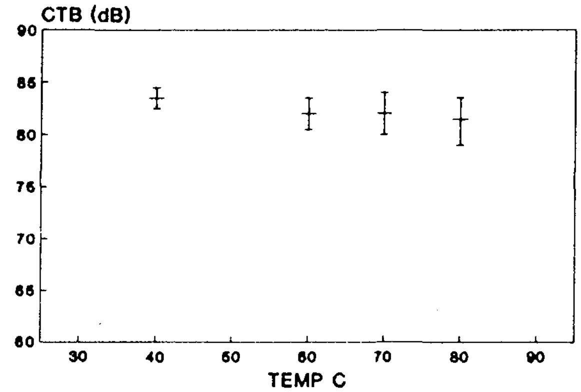


EXHIBIT 5A
CTB vs. Temperature 55.25 MHz Feedforward

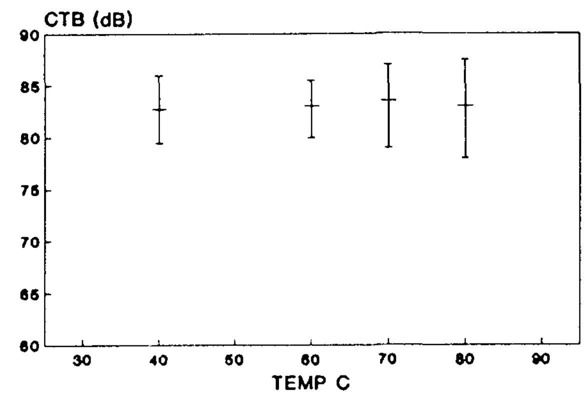


EXHIBIT 5B
CTB vs. Temperature 295.25 MHz Feedforward

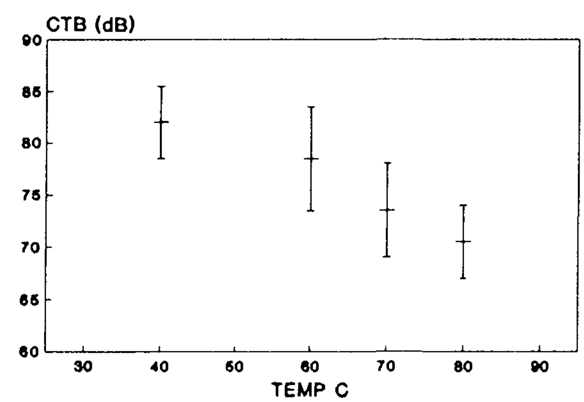


EXHIBIT 5C
CTB vs. Temperature 547.25 MHz Feedforward

Exhibits 6A and 6B show second loop cancellation as a function of termperature and frequency for two sample feedforward hybrids (6). The cancellation of feedforward hybrids varies more at 450 and 550 MHz than at 50 MHz, accounting for the greater change in CTB with temperature at higher frequencies. Note that a change in cancellation will cause an equal change in distortion level.

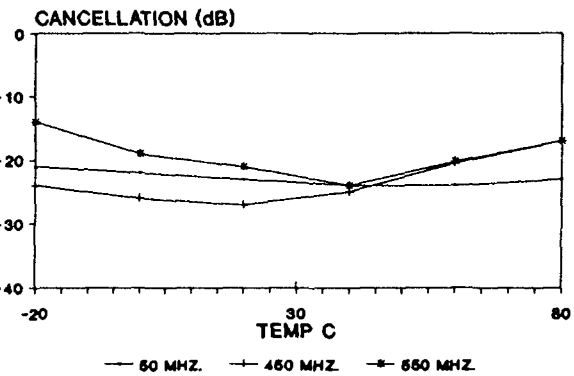


EXHIBIT 6A
2nd Loop Cancellation vs. Temp
Feedforward Sample A

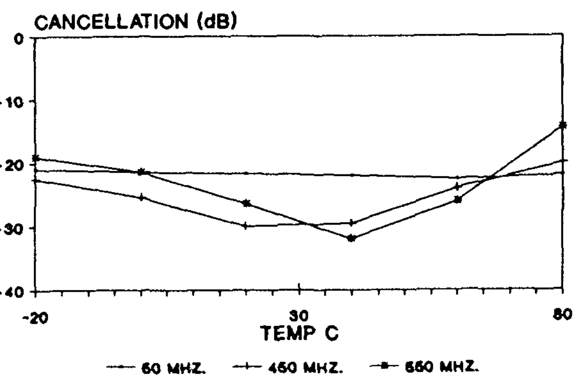


EXHIBIT 6B
2nd Loop Cancellation vs. Temp
Feedforward Sample B

The absolute distortion level is also lower at 55 MHz than at 550 MHz because the push-pull main amplifier in the feedforward circuit generates lower distortion at 55 MHz than at 550 MHz.

Most relevant to this paper is the fact that, at high frequencies, cancellation varies from unit to unit and with temperature. This leads to distortion phase variations from unit to unit and with temperature and means that feedforward amplifiers will not cascade on a 20logN basis. This concept is demonstrated with experimental data in the next section.

TESTING OF DISTORTION PHASE

Distortion phase was measured indirectly by measuring the CTB level of one amplifier, of a second amplifier, and of the two amplifiers in cascade. Since small errors in CTB measurements can cause large errors in calculated phase values, great care was taken in making accurate measurements. Amplifiers were operated at output levels where higher than third order distortions were negligible. The output levels of amplifiers were monitored and maintained within approximately 0.1dB, whether measured singly or in cascade. The noise floor was measured and CTB levels were corrected for these noise levels.

Eight feedforward amplifiers were tested. These were Scientific-Atlanta feedforward trunk amplifiers with Motorola feedforward hybrids. The push-pull input hybrid was replaced in each module with a passive through block. As a control test, six standard Scientific-Atlanta push-pull trunk amplifiers were also tested. These amplifiers had push-pull input and output hybrids.

For each set of test results, the difference in distortion phase Θ_D was calculated. From Exhibit 7 it can be seen that

$$CTB_{TOTAL} = \sqrt{(CTB_1 + CTB_2 \cos \Theta_D)^2 + (CTB_2 \sin \Theta_D)^2}$$

where CTB_{TOTAL}, CTB_1 and CTB_2 are expressed as voltages, not in decibels. Manipulating this equation provides

$$\Theta_D = \cos^{-1}\left(\frac{CTB_{TOTAL}^2 - CTB_1^2 - CTB_2^2}{2\, CTB_1\, CTB_2}\right)$$

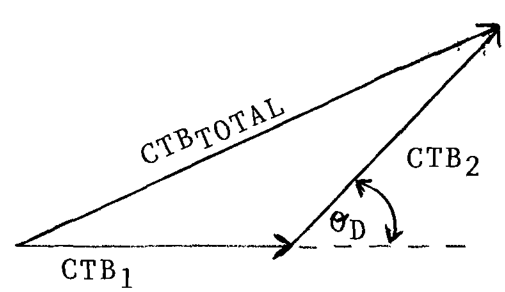


EXHIBIT 7
Vector addition of CTB

Exhibits 8A and 8B show the relative distortion phase O_D of five pairs of push-pull amplifiers, at 295.25 and 445.25 MHz respectively.

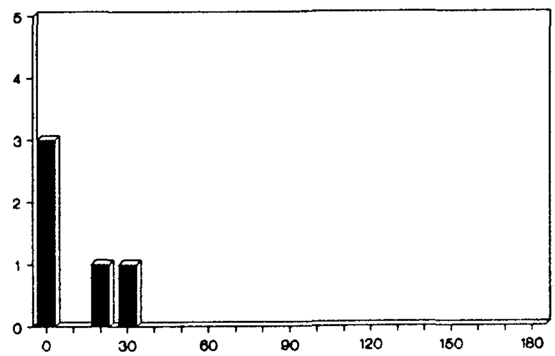


EXHIBIT 8A
295.25 MHz Push-Pull Pairs

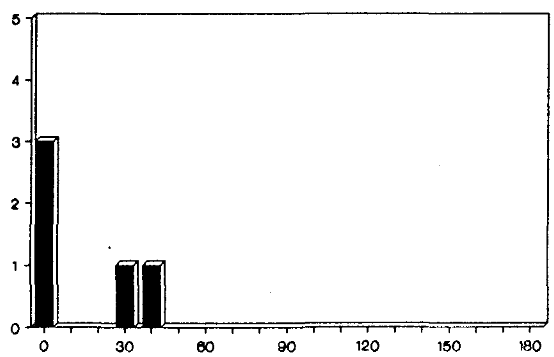


EXHIBIT 8B
445.25 MHz Push-Pull Pairs

At both frequencies, all pairs of amplifiers measured very low values of Θ_D, as expected. Note that a 0.2dB error in the measured CTB will cause a calculated phase error as large as 24° near $\Theta_D = 0^\circ$, but less than 1° at $\Theta_D = 150^\circ$. Consequently, experimental error in the reported values of Θ_D will be greater when Θ_D is low.

Exhibits 9A and 9B show Θ_D at 67.25 and 445.25 MHz respectively, as measured on push-pull pre-amplifier and feedforward post-amplifier pairs. The distortion phases of the push-pull pre-amplifiers are all nearly identical as demonstrated above. From Exhibit 9A, the Θ_D of the feedforward amplifiers is relatively constant at this low frequency. This is because the feedforward amplifiers exhibited no deep cancellation nulls or CTB variation with temperature at low frequencies. It must be noted that although there was little variance in Θ_D at 67.25 MHz, the average value was about 120°, which caused no CTB degradation when push-pull and feedforward hybrids were cascaded.

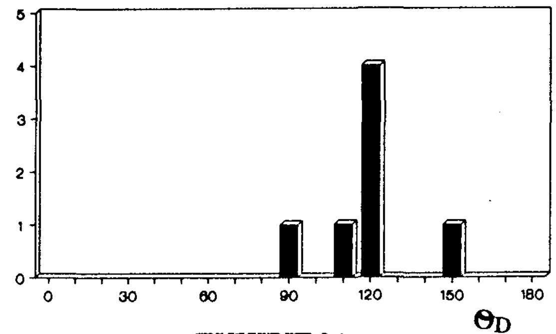


EXHIBIT 9A
67.25 MHz Push-Pull Feedforward Pairs

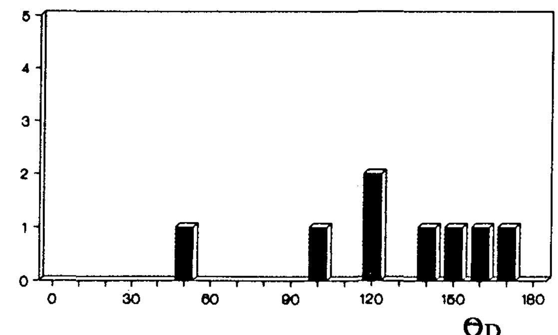


EXHIBIT 9B
445.25 MHz Push-Pull Feedforward Pairs

Results for push-pull/feedforward hybrid pairs at 445.25 MHz are shown in Exhibit 9B. Unlike at 67.25 MHz, the feedforward distortion phases vary over a wide range (from 51° to 169°), and the units are neither generally similar to each other nor to push-pull hybrids.

From these results, one would expect that the Θ_D between feedforward hybrid pairs would have the following characteristics: at low frequencies Θ_D would be consistently low, but at higher frequencies Θ_D would tend to vary over a wide range. This was tested in the next set of results.

Distortion phase differences for pairs of feedforward amplifiers are shown in Exhibits 10A, B and C, at 67.25, 295.25 and 445.25 MHz respectively. At 67.25 MHz, Θ_D is always less than 90°; while at both 295.25 and 445.25 MHz, Θ_D varies over wide ranges. The results shown in Exhibits 10A to 10C demonstrate that, at intermediate and high frequencies, feedforward amplifiers will cascade on significantly less than a 20logN basis. At low frequencies, feedforward amplifiers do cascade on a near 20logN basis; however, distortion levels at this frequency are significantly lower than at higher frequencies.

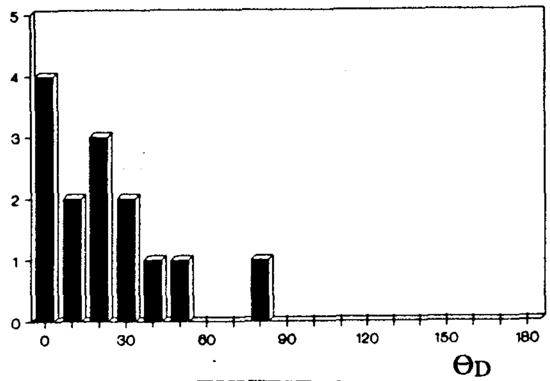


EXHIBIT 10A
67.25 MHz Feedforward Pairs

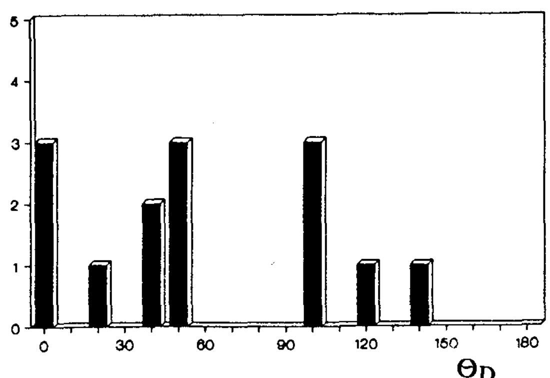


EXHIBIT 10B
295.25 MHz Feedforward Pairs

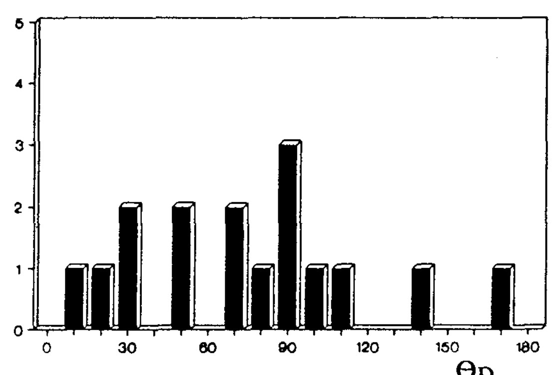


EXHIBIT 10C
445.25 MHz Feedforward Pairs

It must be noted that the number of data points for the Θ_D of feedforward amplifiers is artificially high since all the data points are not independent. The data may include the Θ_D of amplifier pairs 1 and 2, 2 and 3, and 1 and 3. If the Θ_D for two of these pairs are known, then the Θ_D for the third pair can be predicted. For example, at 445.25 MHz amplifier pairs, pairs 6 and 7 had a Θ_D of 98° and pairs 7 and 8 had a Θ_D of 71°. Pairs 6 and 8 should, therefore, be approximately either $98-71=27^\circ$ or $98+71=169^\circ$. In fact, pairs 6 and 8 measured 174°, within 5° of one of the predicted values.

A comparison of the predicted distortion compared to measured results shows that the angles were measured accurately (typically with less than 15° cumulative error). Since the results were taken over many hours of elapsed time, they also indicate that at a constant temperature, Θ_D remains stable in a feedforward amplifier.

It is interesting to note that these phase additions apply to CTB, which is comprised of many hundreds of beats. The concept of phase addition can be applied to CTB in feedforward amplifiers because Θ_D is essentially identical for all the beats comprising the CTB. The beats generated in the main amplifiers are all similar in nature. Feedforward cancellation is performed at the final beat frequency, and all individual beats at the same approximate frequency will be cancelled identically, and will have identical phase shifts from the beats generated by the main amplifier.

Other work on cascading of CTB in dissimilar devices (e.g. AML microwave, AM fibre, and trunk amplifiers), has shown that the phase additions of CTB do not apply (7). This is the case because Θ_D varies over a wide range (70°) for the individual beats comprising a CTB. Truly dissimilar devices have fundamentally different non-linear transfer characteristics and generate beats with a wide range of phases.

In feedforward amplifiers, since all beats are treated identically by the feedforward circuitry, it is possible to measure any value of Θ_D, from 0 to 180°. Between dissimilar devices, extremely low or extremely high phase differences never occur because different beats at the same frequency add with different phases (7).

IMPLICATIONS OF TEST RESULTS

Since CTB in feedforward amplifiers has been demonstrated to cascade with a large variation in distortion phases at high frequencies, a 20logN design rule for cable systems is excessively conservative. Worst case and expected CTB performance in a cascade of stations with feedforward post-amplifiers and push-pull pre-amplifiers is discussed below.

The push-pull pre-amplifier and feedforward post-amplifier generate approximately the same distortion level which will be taken as a reference in a worst case analysis. The single station will generate a CTB ratio 6dB worse than the reference, and a cascade of twenty stations will generate a CTB ratio 32dB worse.

Based on the previous test results, it is expected that the worst case scenario will not occur. At high frequencies, push-pull pre-amplifiers will tend to cascade randomly with feedforward post-amplifiers, generating a typical single station CTB ratio only 3dB worse than the reference. The feedforward post-amplifiers will tend to cascade randomly with each other so the cascading of the pre-amplifiers and post-amplifiers must be considered separately. A cascade of twenty pre-amplifiers will generate a CTB ratio 26dB worse while a cascade of twenty post-amplifiers will generate a CTB ratio 13dB worse. The total cascade will generate a CTB ratio 26.2dB worse, a 5.8dB improvement from the worst case calculation. Note that almost all of the system distortion is due to the push-pull pre-amplifiers.

The expected CTB ratio can also be calculated for low frequencies, where push-pull pre-amplifiers and feedforward post-amplifiers cascade with $\Theta_D = 120^o$, but where both pre-amplifiers and post-amplifiers cascade on a 20logN basis. The expected performance of individual amplifiers at low frequencies is 10dB better than at high frequencies.

These worst case and expected CTB ratios are summarized in Exhibit 11. The expected results are not recommended as a system design rule since cascade performance will demonstrate statistical scatter. More data is required to derive a system design rule.

A similar analysis of a cascade of trunk stations, with feedforward, not push-pull pre-amplifiers, is shown in Exhibit 12. With this station configuration, CTB cascades randomly at high frequencies. At low frequencies, however, 20logN cascading occurs on both the pre-amplifiers and post-amplifiers, as well as between pre-amplifier and post-amplifier pairs. The 10dB low frequency improvement still applies. Expected improvements from worst case predictions are even larger for this station configuration than for the push-pull pre-amplifier configuration: 10 and 13.8dB.

A station with a feedforward pre-amplifier would have a 3dB higher noise figure than a station with a push-pull pre-amplifier because of input losses on the error amplifier of the pre-amplifier stage. After station CTB performance has been altered by 6dB to provide a station with equivalent carrier to noise ratio (CNR), the two amplifier configurations can be compared as in Exhibit 13.

CTB RATIO RELATIVE TO REFERENCE

	Worst Case	Expected High Freq.	Expected Low Freq.
Pre-amp	Ref.	0	+10
Post-amp	Ref.	0	+10
1 Station	-6	-3 **	+10 ***
20 Pre-amps	-26	-26 *	-16 *
20 Post-amps	-26	-13 **	-16 *
20 Stations	-32	-26.2 **	-16 ***
Improvement from Worst Cast (dB)	--	5.8	16

* Assumes $\Theta_D = 0^o$
** Assumes $\Theta_D = 90^o$
*** Assumes $\Theta_D = 120^o$

EXHIBIT 11
CTB Ratios of a Cascade of Push-Pull Pre-Amplifiers, Feedforward Post-Amplifiers

CTB RATIO RELATIVE TO REFERENCE

	Worst Case	Expected High Freq.	Expected Low Freq.
Pre-amp	+20	+20	+30
Post-amp	Ref.	0	+10
1 Station	-0.8	0 **	+9.2 *
20 Pre-amps	-6	+7 **	+4
20 Post-amps	-26	-13 **	-16 *
20 Stations	-26.8	-13 **	-16.8*
Improvement from Worst Cast (dB)	--	13.8	10

* Assumes $\Theta_B = 0^o$
** Assumes $\Theta_D = 90^o$

EXHIBIT 12
CTB Ratios of a Cascade of Feedforward Pre-Amplifiers, Feedforward Post-Amplifiers

	Push-Pull Pre-amp Feedforward Post-amp	Feedforward Pre-amp Feedforward Post-amp
Low Freq.	Ref. -16	Ref. -22.8 *
High Freq.	Ref. -26.2	Ref. -19 *

* Corrected for equal CNR

EXHIBIT 13
Expected CTB Ratios from Cascades of 20 Amplifiers

After CNR corrections, a dual feedforward cascade is expected to perform over 3dB better in CTB than a conventional cascade. Performance of such a cascade would be limited by the CTB ratio at low frequencies, because cascading tends to be 20logN at low frequencies.

CONCLUSIONS

The phase differences between CTB distortion generated by feedforward amplifiers have a wide variation at high frequencies, causing cascading which appears to be random. At low frequencies, the phase differences are small, causing near voltage addition of CTB between feedforward amplifiers. Feedforward amplifiers and push-pull amplifier pairs typically generate CTB with 120^{o} phase difference at low frequencies.

The performance of a cascade of feedforward amplifiers will be significantly better than that predicted using 20logN addition, even in stations with push-pull pre-amplifiers. At high frequencies better performance is due to random cascading, while at low frequencies it is due to individual station CTB performance significantly better than specification.

Because of the random cascading of feedforward amplifiers at high frequencies, feedforward technology is more attractive than was previously thought. For example, a cascade of trunk amplifiers with feedforward pre-amplifiers, as well as feedforward post-amplifiers, will generate significantly less CTB than a cascade of amplifiers with push-pull pre-amplifiers and feedforward post-amplifiers. A dual feedforward cascade would be limited by low frequency CTB performance.

More data is required to define CTB distortion phase at various frequencies so that a system design rule can be generated.

1 Bell Telephone Laboratories. "Transmission Systems for Communications", 4th Edition, Western Electric Co. Ltd. Technical Publications, 1971.

2 Bell, Richard; and Rebeles, P. "Third Order Distortion Build Up In A Multi-Channel Cascade", NCTA Convention Papers, 1973.

3 Prochazka, A. "Improving CATV System Performance Through Amplifier Design", CCTA Convention Papers, 1975.

4 Bauer, Dieter; and Reichert, Harry. "400 MHz CATV System Performance Studies", NCTA Convention Papers, 1981.

5 Cowen, Marty - Scientific-Atlanta correspondence.

6 Do, Steven - Motorola correspondence.

7 McEwen, Doug; Slater, Norm; Straus, Tom; Pidgeon, Rezin. "Distortion Characteristics of Integrated Fibre, AML and Cable Amplifier Systems", CCTA Convention Papers, 1990.

Doug McEwen, P.Eng.

Doug McEwen is a Staff Engineer with CUC Broadcasting Limited. He is responsible for many areas of network design, and provides engineering support for over 50 cable and multiple FM and AML systems.

Prior to joining CUC in 1986, Doug worked in the Network Development group of Cablesystems Engineering with Rogers. His responsibilities there included network design and calculating the technical performance and cost of various design options.

Doug received his Bachelor of Engineering Science from the University of Western Ontario in London in 1983. He is a member of the Association of Professional Engineers of Ontario.

Rezin Pidgeon

Rezin Pidgeon is a Principal Engineer for Scientific-Atlanta and currently supports the fiber optic development group. Rezin has been with Scientific-Atlanta since 1962, and was with the Antenna and Telecommunications Instrument groups until 1980. Rezin has been with the Broadband Communications Business Division since then.

Prior to joining Scientific-Atlanta, Rezin was a senior engineer at the Georgia Tech Engineering Experiment Station. Rezin holds BSEE and MSEE degrees from Georgia Institute of Technology, and is a member of I.E.E.E., the Society of Cable Television Engineers, and ETA Kappu Nu.

Norm Slater, P.Eng.

Norm joined Comlink Systems in January 1986 as Manager of CATV Systems Engineering.

Educated at the University of Western Ontario, Norm received his B.E.Sc. degree, Electrical Engineering in 1975 and also the Harry Cross Medal for Electrical Engineering.

Commencing in 1975 up to the end of 1985, Norm was employed by Cablesystems Engineering latterly as Manager, Network Engineering where his responsibilities included the development and implementing of network designs for all Rogers' systems.

Norm has submitted several technical papers to the Canadian Cable Television Association and the U.S. National Cable Television Association. He is a member of the Association of Professional Engineers of the Province of Ontario.

DR. STRANGELEAK RETURNS

Ted E. Hartson

Post-Newsweek Cable, Inc.

ABSTRACT

Dr. Strangeleak first appeared at the 1983 convention, telling the story that the regulations for Cable Television leakage were inconsistent with other potential interference sources. The Good Doctor takes a look at what has happened in the last five years, including Docket 85-301 regarding set top converters, and offers a prescription for unregulated SMATV systems using aeronautical channels.

Just now, before the July, 1990, deadline for CLI compliance, it's a good idea to make an appointment with the Doctor.

Dynamic Feedback Arrangement Scrambling Technique

David Scott Brown
General Instrument Corporation, VideoCipher Division

Abstract

The Dynamic Feedback Arrangement Scrambling Technique (DFAST) is a method of generating keystream for use in scrambling binary data to prevent unauthorized listeners from recovering that data. The algorithm was initially developed and modeled using the "C" programming language, and then implemented in discrete digital hardware, assembly language code, and several custom integrated circuits. A United States patent was granted in August 1989 for the DFAST Keystream Generator [1]. Two VideoCipher® systems using this technique have received export licenses from the State Department.

INTRODUCTION

Environment

The keystream generated by this technique is added modulo-2 to the binary data on a bit-by-bit basis, creating an encrypted data stream. This encrypted data stream is recoverable if an identical keystream, generated synchronously by an identical keystream generator that begins with an identical initialization number (the key), is added modulo-2 to it. DFAST falls in the general category of "stream ciphers" [4].

Design Constraints

DFAST was developed in order to provide a scrambling technique usable in exportable products. All previous VideoCipher® systems utilized the Data Encryption Standard (DES) [2,3], which is not approved for export by the United States Department of State. This resulted in three major design constraints:

1. The algorithm must be secure against real-time attacks.

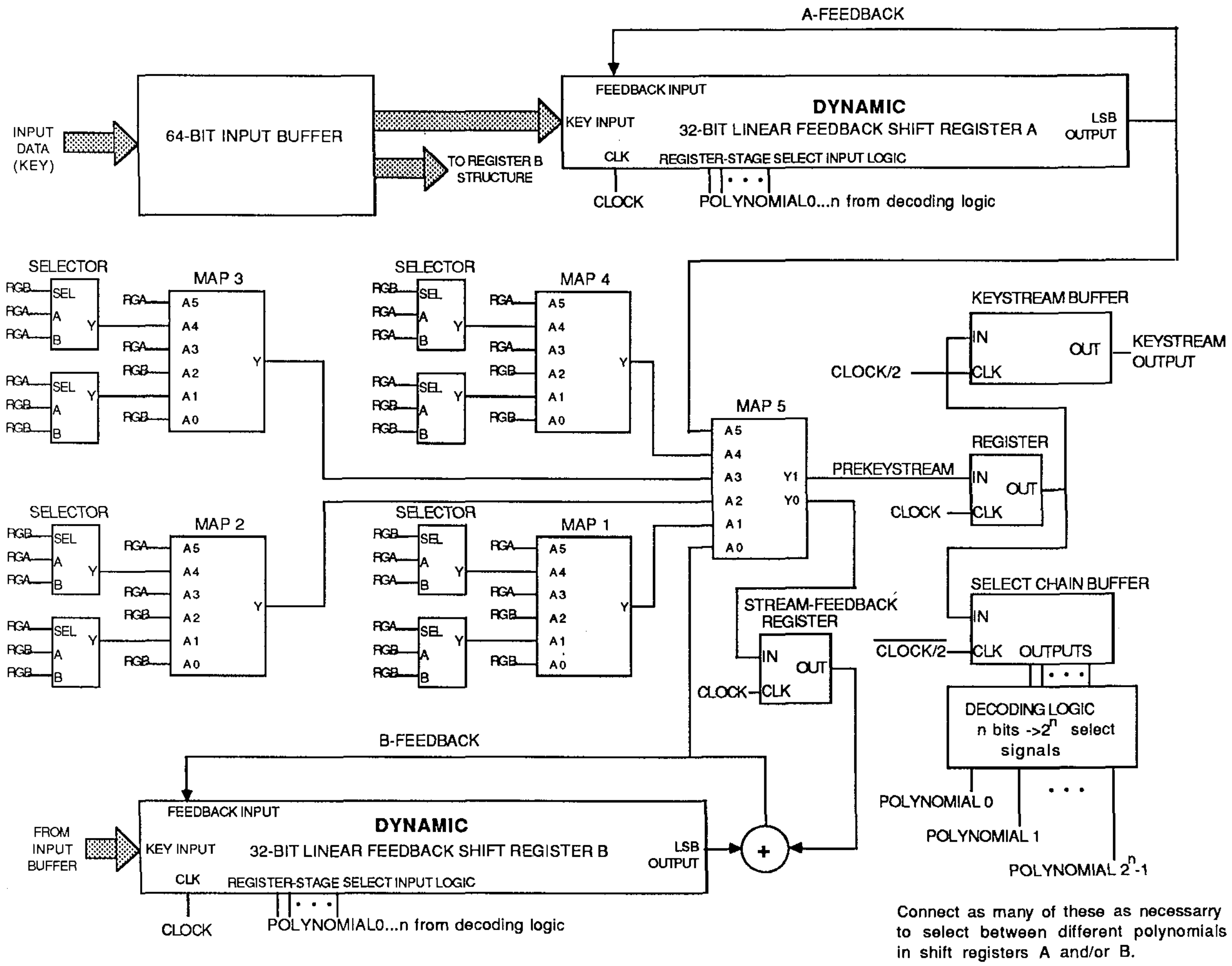
2. It must be readily implementable at high speeds in both hardware and firmware (assembly language).

3. It must be readily implementable in a high-level language so that an export license can be obtained.

GENERAL STRUCTURE

The general form of DFAST is shown in figure 1. There are two 32-bit feedback shift register structures, labeled A and B. Either or both may be a dynamic structure capable of implementing any desired number of polynomials. It is recommended that one of the structures be static and that it implement a primitive, irreducible polynomial of degree 32 to generate a maximal-length binary sequence of length 2^{32}-1. This insures that the length of the keystream output sequence will also be at least 2^{32}-1 before repeating [5].

Twenty out of the 32 register outputs from each of these structures (40 outputs total) are tapped as inputs to the selectors and mapping functions. Each tap is unique; that is, it only connects to one point in the Selector-Map structure. Further rules concerning the connection of the shift register taps are discussed in subsequent sections.



NOTES:

1. The selector places the A or B input on the output Y based on the value of the select (SEL) input (1=A, 0=B).
2. The select input of each selector (SEL) always comes from the register opposite to the inputs A and B.
3. RGA and RGB are all outputs from unique stages of the A and B registers, respectively.
4. There may be any number of different degree-32 polynomials implemented in the A and/or B registers; the select chain buffer will expand to provide enough select bits.
5. The rules for connecting the taps from the A and B register stages to the Selectors and Map functions appear on subsequent pages.
6. The rules for determining the mapping functions (MAP1-5) appear on subsequent pages.
7. Only one of the feedback shift register structures MUST be dynamic, although both may be.

Figure 1

There are five mapping functions, designated MAP1 through MAP5. MAP1 through MAP4 produce one output bit each, while MAP5 produces two output bits at a time. Each MAP function is preceded by two selectors, one of which selects the A4 input to the MAP and the other the A1 input.

There is a holding register at the Y1 output of the MAP5 function, for the purpose of splitting up the PREKEYSTREAM bits. The structure shown here places alternate bits of PREKEYSTREAM into the Keystream Buffer, with the remaining bits stored in the Select Chain Buffer. Many other subdividing schemes are possible; the idea here is to avoid placing contiguous PREKEYSTREAM output bits into the KEYSTREAM. The Select Chain Buffer stores as many successive alternate bits from the PREKEYSTREAM output as necessary to select between the different polynomials available in the

Dynamic Register(s).

There is also a holding register at the Y0 output of the MAP5 function for the purpose of storing the STREAM-FEEDBACK bits. This pseudo-random bitstream is exclusive-OR'ed with the LSB output of the B-register structure and the result is used as both the feedback bit of the B-register structure and as the A0 address input of the MAP5 function. This causes further randomness in the sequence of states that appears in the B-register structure, making that sequence even more difficult to predict. The holding register is necessary to avoid an unstable feedback path around the MAP5 function.

Finally, there is a block of Decoding Logic that uses the stored Select Chain Buffer bits to decode a particular polynomial, setting a single signal corresponding to a particular polynomial true while keeping all other polynomial signals false.

COMPONENT DESCRIPTIONS

Static Feedback Shift Register Structure

If a Static Feedback Shift Register Structure is used and it implements a primitive, irreducible polynomial, it will generate a maximal length sequence of length $2^{32}-1$. This will insure that no matter what polynomial is used in the dynamic register, the overall keystream generator for DFAST will have a sequence length of at least $2^{32}-1$ before repeating. Figure 2 shows a more detailed view of a Static Structure.

Selector Structure

There are two Selectors in front of each Mapping function, selecting the middle input to that Map from each register structure. This provides a means to use additional taps from the shift register structures into the mapping functions and provides greater nonlinearity in the overall scrambling process. Each selector chooses between its two data inputs, labeled A and B, based on the value of its SEL (select) input. If SEL is a logic TRUE (1), the A input is selected; if SEL is FALSE (0), the B input is selected.

Select Chain Buffer/Decoding Logic Structure

The Select Chain Buffer in simply a serial shift register (with no feedback) which must be long enough to provide the capability to select between the different polynomials used in the Dynamic Register structure(s). The general formula is:

```
Size of chain =  highest n such that
```

2^{n-1} `< number of polynomials`
`and` 2^{n} `> number of polynomials`

For example, if 7 different polynomials were desired, one would use a Select Chain Buffer of length 3 since $2^3=8$ which is greater that 7 while $2^2=4$ is less than 7. Figure 2 shows the structure.

Structure of Register Select Logic

The Register Select Logic is simply a logical-OR function. For any given Register-Stage within a Dynamic Register structure, selection of the feedback to the selector for that stage is accomplished according to the following formula:

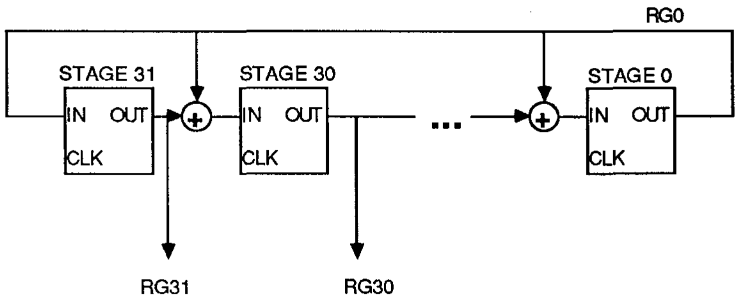
```
Select = TRUE if that register stage
         is part of the polynomial

         FALSE if not.
```

So each SEL signal is a logical OR of however many POLYNOMIAL signals that register stage is involved in.

Dynamic Feedback Shift Register Structure

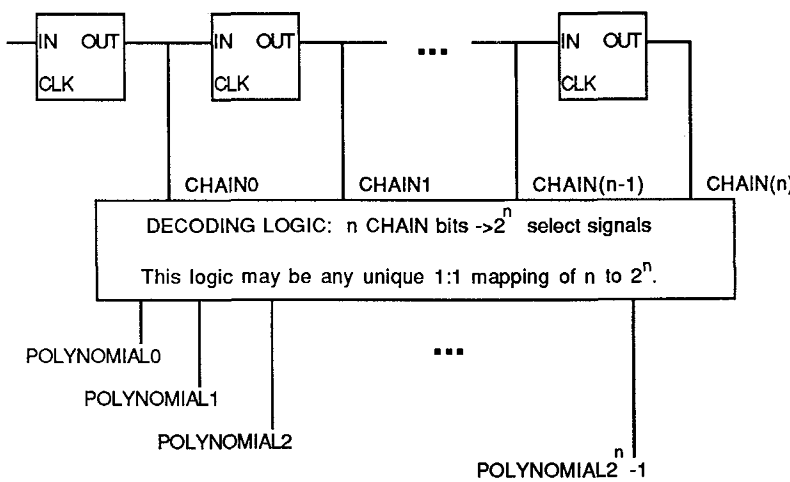
The structure of a Dynamic Feedback Shift

STRUCTURE OF STATIC FEEDBACK SHIFT REGISTER:
(ALL CLOCKS ARE COMMON AT SYSTEM CLOCK RATE)

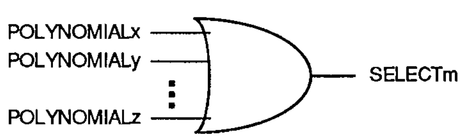


THIS EXAMPLE SHOWS MODULO-2 SUMS (EXCLUSIVE-ORs) AT INPUTS TO 31,30 AND 0. THE SUM AT THE INPUT TO 31 IS IMPLICIT IN THIS STRUCTURE.

STRUCTURE OF SELECT CHAIN BUFFER AND DECODING LOGIC:
(ALL CLOCKS ARE COMMON AT 1/2 THE SYSTEM CLOCK RATE)



STRUCTURE OF REGISTER-STAGE-SELECT INPUT LOGIC:



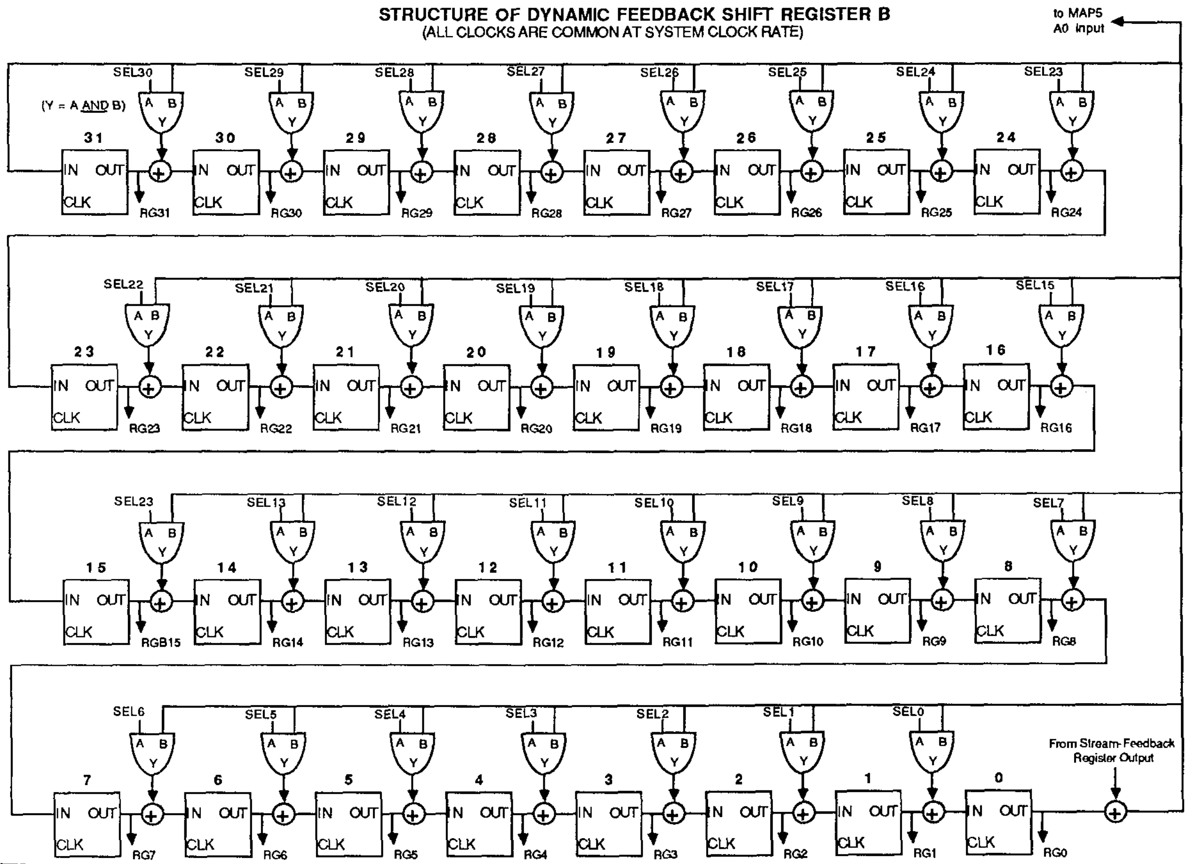
NOTES:

8. The SELECT signal for a given Register-Stage inside the dynamic B-register is the logical-OR combination of all POLYNOMIAL signals that include that given register.

Figure 2

Register is shown in figure 3.

STRUCTURE OF DYNAMIC FEEDBACK SHIFT REGISTER B
(ALL CLOCKS ARE COMMON AT SYSTEM CLOCK RATE)



NOTES:

9. The select signals, SEL31...0, are logical 1 (true) if that particular register stage is in the selected polynomial, and that polynomial is selected by the decoding logic. This allows the feedback bit through the logical AND gate to the modulo-2 adder. If a select signal is false, the logical-AND output is also false, and the exclusive-OR is effectively removed from the input to that Register-Stage. Each individual signal is of the form: SELn = POLYNOMIAL0 + POLYNOMIAL1 + ... POLYNOMIALm; where n=the register-stage number,m is the number of different polynomials used, + is logical OR, and only polynomials that the particular register-stage is a part of are included in the sum. The POLYNOMIAL signals are generated by decoding bits from the Select Chain Buffer.

Figure 3

This structure can implement any number of different polynomials of degree 32. The most-significant Register-Stage is numbered 31 and the least is numbered 0. The feedback signal is either:

- The output of the 0-register, or
- The output of the 0-register exclusive-OR'ed with the "STREAM-FEEDBACK" output of MAP5. In figure 1, this is the option chosen for the Dynamic Structure, which is register-structure B.

This feedback signal is connected to the input of the Register-Stage 31 regardless of which polynomial is used (in other words, all polynomials have a 31 term in them, in order to be of degree 32).

Between each pair of Register-Stages is an exclusive-OR (modulo-2 addition) function, with inputs from the preceding Register-Stage and a selector gate and an output to the next Register-Stage. The selector gate (a logical AND) provides one of the following to the exclusive-OR:

1. A logical FALSE (0) if the SEL signal for that stage is FALSE (0).

2. The feedback term if the SEL signal for that stage is TRUE (1).

When the FALSE is selected (SEL=0), the exclusive-OR function is effectively disabled, and the term corresponding to the following Register-Stage is removed from the polynomial. When the feedback term is selected, the following Register-Stage is part of the polynomial; thus by connecting the SEL inputs as the output of a logical-OR of all desired polynomials, that Register-Stage is included in those polynomials.

SUMMARY OF DFAST STRUCTURE RULES

At least one of the Feedback Shift Register Structures must be **Dynamic**, implementing any number of degree 32 polynomials, with a modulo-2 sum capability between each of the 32 Register-Stages. The sums are invoked depending on the polynomial selected, and this selection takes place on a clock-by-clock basis.

No tap (individual register output) from either the A-Register Structure or the B-Register Structure is used more than once as a Selector or Map function input.

The Selectors are arbitrarily arranged as inputs to the Maps. The Selector Select inputs, however, must be taps from the Register Structure opposite to that supplying the Selector Data Inputs. The number of Selectors used is only limited by the above rule restricting taps to be used once.

Any two inputs to a particular Map function must have at least one exclusive-OR function between them in their source Register-Structure. This limits the polynomial combinations that may be chosen for implementation by that Register-Structure. This includes taps that enter the Map function through a selector.

The second-level Map function must use all of the first-level Map outputs as inputs as well as the feedback signals from both the A and B register structures.

FUNCTIONAL DESCRIPTION

The operation of DFAST is as follows. A 64-bit initial value, called a key, is loaded into the two feedback shift registers. The registers then shift with every subsequent cycle of CLOCK, and in fact the entire machine is synchronous to this CLOCK. During each clock cycle, or state of the machine, an individual PREKEYSTREAM bit is computed. Various points in each feedback shift register structure are tapped into the Selectors and MAPs as shown in figure 1. Each Selector selects its output between its two data inputs based on the value of its SEL input, and is wired so that the SEL signal comes from the feedback register *opposite* to that of the A and B data signals. Note further that no tap is used in more than one place.

Each MAP function is a pseudo-random distribution of logic 0 and logic 1 values, selected according to the previously defined rules. The outputs of the first-level MAPs (MAP1 through MAP4) are used as the middle four inputs to the second-level MAP. The top and bottom inputs to the second-level map come from the feedback bits of the A and B register structures, respectively. The outputs from this second-level MAP are the PREKEYSTREAM and the STREAM-FEEDBACK, each of which are generated every CLOCK cycle.

The STREAM-FEEDBACK is modulo-2 summed (exclusive-or'ed) with the 0-register output (LSB) of the Dynamic B-Register Structure (RGB0) to produce the B-FEEDBACK signal. This puts further uncertainty in the pattern that the B-Register Structure will produce. The PREKEYSTREAM, which is the result of all these computations, is then subdivided to produce actual KEYSTREAM for use in encryption and to produce the contents of the Select Chain Buffer to select between the various A and B Polynomials.

Conclusions

DFAST is a hardware-efficient, cryptographically strong keystream generation algorithm that has received export license approval. By fixing one of the two feedback shift register structure polynomials, it is possible to guarantee a minimum cycle length for the keystream without repetition. After this, the key may be changed to maintain nonrepeatability. The general structure allows a large number of possible specific implementations (varying the polynomials, the taps, etc.), each of which will meet the previously outlined design constraints while remaining distinct from each other. This allows production of deliberately incompatible systems with only minor changes to the hardware or software implementations.

References

1. "Dynamic Feedback Arrangement Scrambling Technique Keystream Generator", David Scott Brown, U.S. Patent Number 4,860,353, issued August 22, 1989.

2. "Data Encryption Standard", Federal Information Processing Standards Publication 46, U.S. Department of Commerce, January 1977.

3. "DES Modes of Operation", Federal Information Processing Standards Publication 81, U.S. Department of Commerce, December 1980.

4. "Cryptography - A Primer", Alan G. Konheim, John Wiley and Sons, New York, 1981.

5. "Error Correcting Codes", W.W. Peterson and E.J. Weldon, Jr., MIT Press, Cambridge, MA, 1972.

ELECTRONIC IN-HOME INTEGRATION

DAVE WACHOB

JERROLD COMMUNICATIONS

ABSTRACT

The 1990's promise to be an exciting time for the communications' industry, with rapid technical advances supporting a wide variety of new subscriber services. Electronic in-home integration of entertainment, telephony, home automation, control and security, interactivity and data services will all be feasible in the coming decade. CATV is well positioned to provide this integration through expansion of its technology, and the existing sales, installation, billing and support infrastructure. This paper will explore some of the technical, regulatory and historical issues surrounding integration opportunities.

DEFINITION

Electronic in-home integration refers to the integration of video, voice, and data to, and in, the residential environment. It includes delivery, control, and support of both conventional CATV entertainment services and non-conventional services such as telephony, home automation, monitoring and control, and data networks. Analog, digital, or hybrid integration of a variety of distributed information systems are all germane to the concept.

BACKGROUND

Historically, attempts at integration have not readily flourished, for a variety of reasons. Systems such as Homenet, CEBus, Smarthouse, Qube, Omnitel, Prodigy, X*Press, Teleaction, Viewtron, Picturephone and others have had varying degrees of success.

While some of the earlier failures were caused by imperfect technologies, present systems have overcome most of these technical limitations. Human interface issues contributed in part to some of the earlier failures, where system interaction required numerous screens or commands to access the appropriate information or service. Regulators also imposed constraints on other systems and the types of information delivered, hardware manufacturability or return on investment. Commercial viability and market acceptance, however, were most often cited as the major contributor to the success or failure of earlier integration approaches.

ASSUMPTIONS

Video, voice and data will continue to integrate in the 90's, through a diversity of analog, digital or hybrid approaches. Specific implementations, applications and timing will be driven by technical, market, regulatory and commercial forces.

Integration can occur in either the distribution system and / or the residential environment independently or concurrently. Independent integration can occur through a common interface at the subscriber's residence.

Industry integration standards will continue to be refined and accepted as a means to expand the integration worldwide. Two of the more visible standards to be promoted are CEBus and SONET.

MOTIVATION

Cable is in a unique position to be an integrated services provider. With over 49 million homes, representing 55% of the TV households already connected, the numbers are projected to grow even further. Growth will occur not only as additional video subscribers are added to the existing base, but also as new non-video services such as digital audio, interactive services, etc. are added. Expansion into PPV will further expand cable's competitive position.

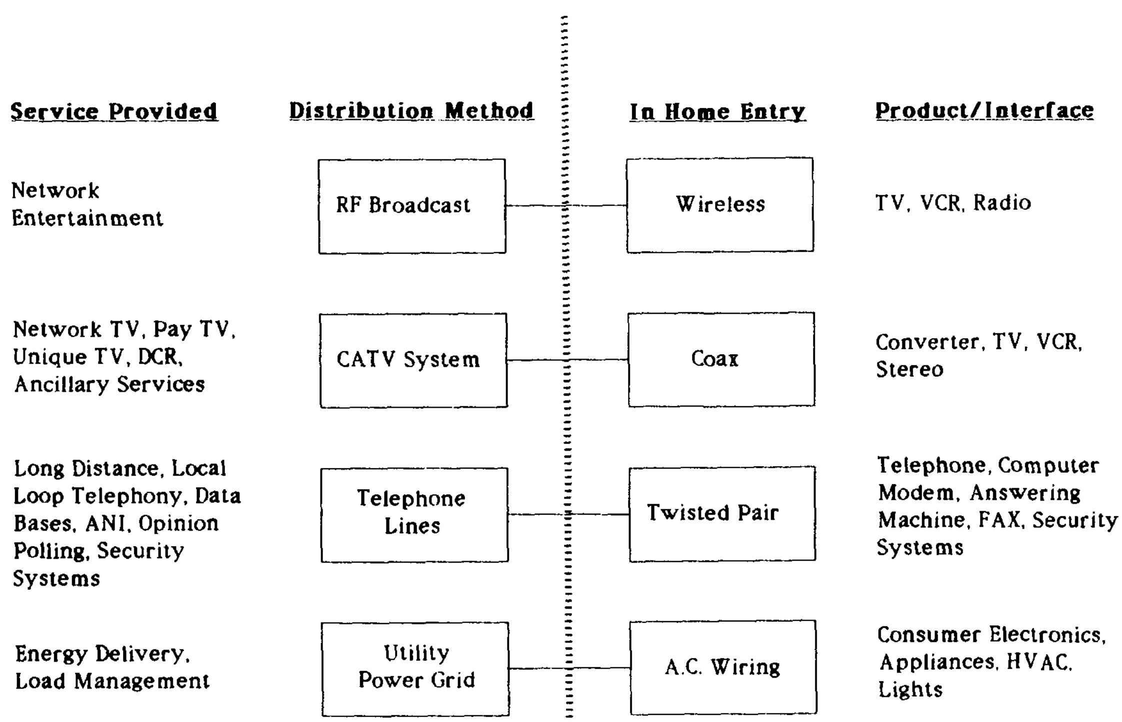
The current cable infrastructure is also well positioned to provide the necessary sales, installation, service and support for electronic in-home integration. Furthermore, the current CATV billing, conditional access and addressable control systems could be easily expanded to include the additional services.

Technically, cable is already delivering first class video signals in state-of-the-art, one-way and two-way cable systems worldwide. This capacity for information transmission and processing, including expansion into 1 GHZ and fiber optics technology, provides the foundation for a true wideband information pipeline in and out of the home.

The existing network interface (Figure 1) details the present level of service integration into the home. Currently, the four main information interfaces into the home include the CATV coax, broadcast RF, telephone lines and the utility AC power lines. AC power lines have not generally been considered an information media, but recent applications have indicated their viability for overlayed signal transport.

FIGURE 1

Existing Network Integration

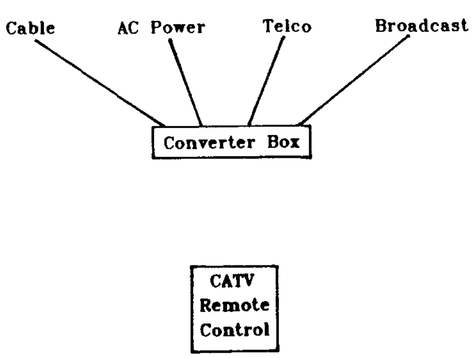


Each distribution method carries information relative to the particular media and interfaces directly with the consumer hardware noted. While there is some degree of interface and information overlap in the home, for the most part they are separate non-integrated media.

The ability to support electronic in-home integration in a cable environment exists now, via addressable converters. As shown in Figure 2, all of the current information distribution systems pass through existing addressable converters. Clearly, coax and AC power are interfaced through conventional connections. With telephone return path modules, the telco phone network is also available. Finally, with the FCC mandated A/B switch, broadcast RF is also accessible. Similar CATV integration would be possible with on / off premise, or multiport technology.

FIGURE 2

Current Converter Interface

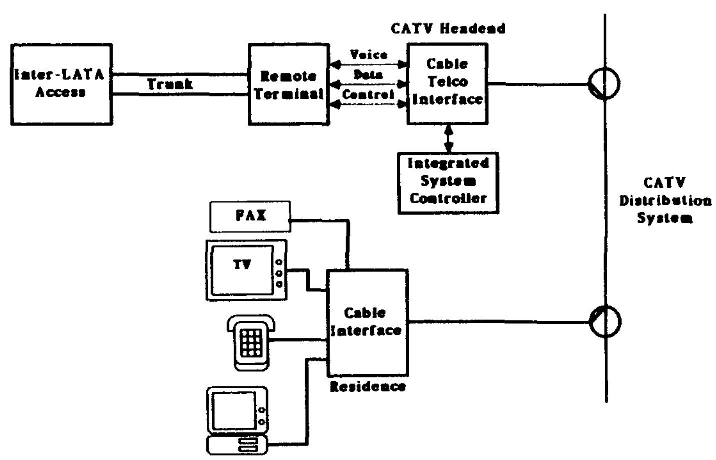


One of the advantages of the common CATV integration points described above includes the use of existing hardware within the device. The addressable receiver, descrambler, micro-processor, memory, remote control, data transmitter and on-screen display capabilities can all be employed to provide integrated services. Interconnection with the existing television also provides a natural vehicle for service selection, display, interaction and feedback.

POSSIBILITIES

Given the above capabilities, some interesting possibilities for integration open up. One concept combines a remote control and cordless telephone, communicating with either the converter or on-premise device. Common components in both the remote control and cordless telephone (keypad, communications path, processor, battery, housing) allow for cost-effective and simple integration. Adding some simple appliance control capabilities (X-10) to the concept would further integrate the concept, and allow for the ultimate "couch potato"!

FIGURE 3



Interconnection to the phone network with the remote control / cordless telephone approach described above could either be through a local residential interconnect, or local loop bypass detailed in Figure 3. In addition to voice, this system would bypass other conventional telephony services, including fax, data base, and security services.

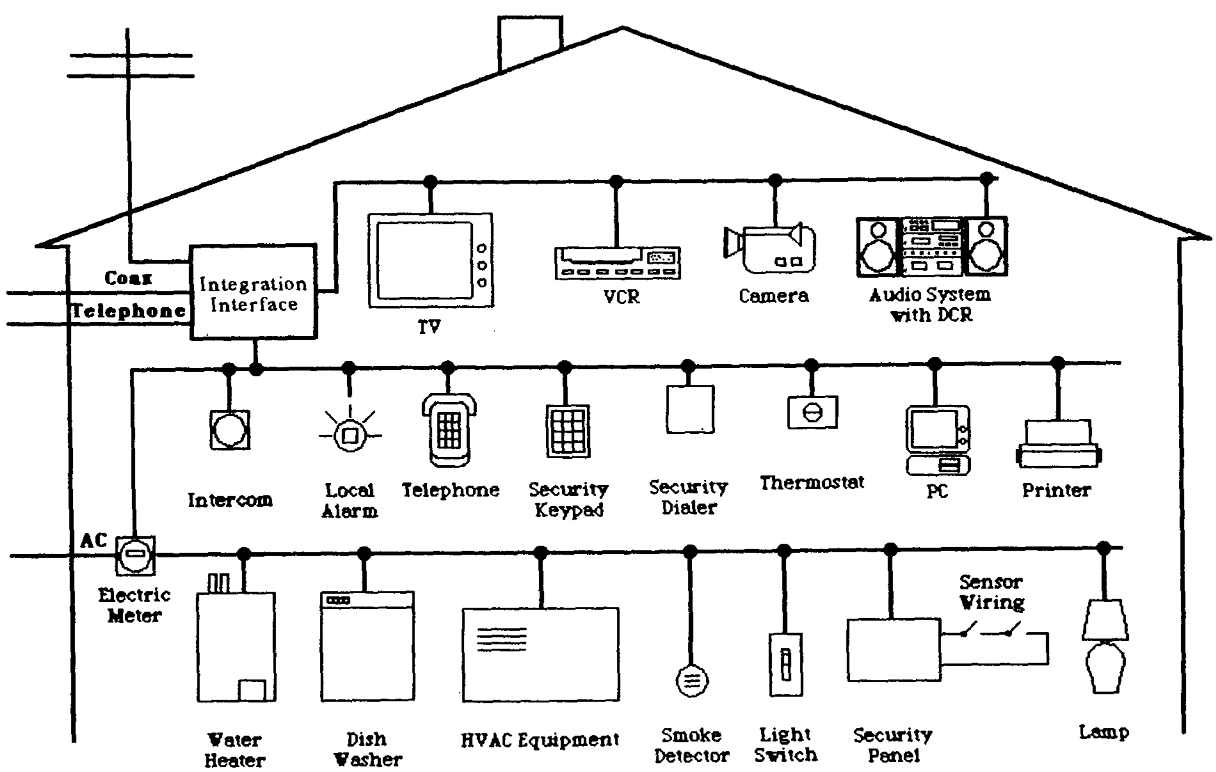
At the other extreme is complete in-home integration as detailed in Figure 4, where all available media are available for communications. This system allows for total integration flexibility, at the expense of complexity and cost. Human interface is also inherently more complex, due to the level of integration provided.

CONCLUSION

Integration of video, voice, and data in the 90's will continue to expand, through a variety of technical approaches and market applications. Complete digital integration is a natural progression of this technology as we move toward the 21st century. Cable, with its technical base, existing infrastructure and strategic beachhead in the home is uniquely positioned to provide this integration, for the upcoming expanded information society.

FIGURE 4

Integrated Home Block Diagram



Fiber Optic Supertrunking: A Comparison of Parameters and Topologies Using Analog and or Digital Techniques

Vincent R. Borelli, President and Hermann Gysel, Vice President
Synchronous Communications, Inc.

Abstract: Fiber optic supertrunking has a well established place in larger CATV networks. Supertrunks have to be transparent to the signal quality and they have to have no impact on the system reliability of the total network. 60 dB video SNR has to be achieved and down times of less than one hour per year are desirable. AM on fiber cannot achieve these goals (yet). FM and digital with a 20 dB higher loss budget than AM and better inherent reliability are well suited for high transmission quality as well as for redundant supertrunk topologies, achieving excellent system availability numbers.

Introduction:

Supertrunks need to be transparent to the signals that are normally fed from a satellite dish through the CATV network into a set top converter and/or the TV set. The achievable quality of a satellite link has to be determined first. Then the degradation caused by the coaxial network and by the set top converter and/or TV set needs to be estimated. Then parameters of a transparent supertrunk can be found. Down time of a supertrunk can be estimated as a function of the topology of the supertrunk.

1. Noise accumulation in various sections of a CATV network:

1.1. The set top converter

Assuming that the set top converter has a noise figure of approximately 15 dB and that the signal level at the outlet is 7 dBmV the set top limitation for signal to noise is approximately:

$$\text{Video SNR} = 174 - 66 - 15 + (7\text{-}49) = 51 \text{ dB}$$

This is approximately the limit of perceptibility of noise in an average TV picture.

1.2. Trunk and distribution

In a CATV system that uses fiber optic supertrunks the target picture quality after trunking and distribution is normally high, the video SNR planned to be 48 to 50 dB at the outlet. Together with the set top converter the video SNR limitation is then 46.2 to 47.5 dB.

1.3. Supertrunk

Supertrunks are designed to produce a video SNR of 60..65 dB. Together with trunk, distribution, and set top converter the video SNR limitation is now 46.0 to 47.3 dB with a 60 dB supertrunk and 46.1 to 47.4 dB with a 65 dB supertrunk. Obviously supertrunks that are better than 60 dB in video SNR do not improve total system performance significantly.

1.4. Satellite link

Using the numbers from [1], let us compare three dishes, each of them with three different LNB noise temperatures:

Dish size:	4.5m	7m	10m
Noise temp.			
90K	50.3	54.7	58.7 (dB)
80K	50.7	55.1	59.1 (dB)
35K	51.7	57.6	61.5 (dB)

Combining these satellite video SNR's with the SNR of the cascade supertrunk, trunk, distribution, and set top converter (46 to 47.4 dB) the following SNR ranges are found:

Video SNR of total system:

4.5m dish with 90K LNB:	44.6 to 45.6dB
4.5m dish with 35K LNB:	45.0 to 46.0dB
7.0m dish with 80K LNB:	45.5 to 46.7dB
7.0m dish with 35K LNB:	45.7 to 47.0dB
10 m dish with 35K LNB:	45.9 to 47.2dB

1.5. Conclusions on system noise performance:

A 4.5m dish with a noisy LNB is not adequate for the use in a high quality CATV system. A 7m dish with a very low noise LNB is probably the best compromise between cost and signal degradation. Higher signal levels into the set top converter or a lower noise figure for the set top converter would improve the picture quality.

Assuming that the set top converter can handle full channel loading at 10 dBmV (without a noise figure change because of AGC and no set top CTB) then the following numbers could be achieved:

Set top converter	54.0 dB
Trunk/distribution	50.0 dB
Supertrunk	62.0 dB
Satellite receiver	57.6 dB
System Total	48.0 dB

The picture quality for a 48 dB video SNR is approximately 3dB worse then the limit of perceptibility of noise. The quality number in accordance with CCIR Report 959 and CCIR Rec. 500-2 is 4.75, 5 being a perfect picture without any impairment.

2. Performance parameters of different modulation schemes used in fiber optic supertrunks:

2.1. AM techniques

Amplitude modulation does not offer the possibility of a trade off between CNR and occupied bandwidth like FM or digital. AM is about 30 dB more susceptible to noise than FM. It requires therefore the use of very advanced techniques in lasers and fiber optics in order to achieve CNR's that are typically 10 dB worse than what can be achieved with FM or digital. Nevertheless the absence of modulation conversion equipment makes AM on fiber attractive. The basic AM system calculations as they are known today are:

2.1.1. Noise in AM systems

Two years ago, DFB lasers that were designed for digital applications typically achieved RIN numbers of -150 dB/Hz. Today's DFB lasers, that are designed for AM applications, are reported to have RIN's of -155 to -160 dB/Hz. Nevertheless the production yield of lasers with RIN's better than -155dB/Hz is only 5% [2]!

The TV channel CNR produced by the laser RIN is:

$$CNR_{(laser\ RIN)} = -RIN + 20\log(m) - 3 - 10\log(4.2MHz)$$

$$= -RIN + 20\log(m) - 69$$

The receiver noise consists of shot noise (or quantum noise) of the detection process and of the noise in the following RF amplifiers. A convenient way to describe this is:

$$CNR_{(receiver)} = 152 + 20\log(m) + P_{opt}(dBm) - 10\log(4.2\ MHz) - N_a$$

$$= 86 + 20\log(m) + P_{opt} - N_a$$

N_a is the differential between total receiver noise and shot noise. This differential is a quality number for various receiver designs. One of the best receivers, Ortel's 2605A, is 1dB at 0dBm, 1.5dB at -5dBm, 4dB at -10dBm, and 10 dB at -15 dBm.

Another noise source is the fiber itself. McGrath [3] reports that laser phase noise can be converted to intensity noise by reflections in the fiber link. Assuming a 4 GHz bandwidth of the laser (chirping because of modulation) and a fiber reflectivity of 29 dB one can calculate an equivalent fiber RIN of -152 dB/Hz. The fiber limitation to CNR is therefore:

$$\text{CNR(fiber)} = 20\log(m)+152-3-66 = 20\log(m)+83$$

Assuming a laser RIN of -155 dB/Hz (which is optimistic) the following link performances for -3 dBm, -6 dBm, and -10 dBm optical received power can be estimated:

Estimated AM link performance:

	Number of channels: 10	20	40
Modulation index m:	0.12	0.075	0.06
Laser noise:	67.6	63.5	61.6 (dB)
Fiber noise:	64.6	60.5	58.6 (dB)
Total noise laser & fiber:	62.8	58.7	56.8 (dB)
Receiver noise at -3dBm:	63.6	59.5	57.6 (dB)
Total noise at -3dBm:	60.2	56.1	54.2 (dB)
Receiver noise at -6dBm:	59.6	55.5	53.6 (dB)
Total noise at -6dBm:	57.9	53.8	51.1 (dB)
Receiver noise at -10dBm:	53.6	49.5	47.6 (dB)
Total noise at -10dBm:	53.1	49.0	47.1 (dB)

Obviously only a 10 channel loading is useful for supertrunking. Even then the received optical power should not be less than -3 dBm. Using a 4mW laser a loss budget of up to 9 dB can be achieved.

2.1.2. Usefulness of AM supertrunks

Above numbers indicate that AM fiber optic systems produce substantially lower SNR's than FM or digital. AM is therefore more useful in applications in the trunk and/or feeder section of a CATV network, where it can outperform coaxial techniques. In AM supertrunks channel numbers of ten or less have to be used and optical receive powers of -3 dB are minimum, limiting its usefulness considerably.

2.2. Frequency modulation

FM has been used successfully for many years. Using a deviation of 8 MHz sync tip to peak white the FM improvement over AM is 30 dB. Received CNR's of 30 dB (in 4.2MHz bandwidth) still produce a 60 dB video SNR. It is useful to use APD's in the optical receiver to achieve very good loss budgets with receivers of reasonable complexity. A Germanium APD receiver needs about 5µW (-23dBm) of optical received power to produce 60 dB video SNR with 16 channels. Supertrunks using FM therefore outperform ones using AM by about 20 dB of optical loss.

2.3. Digital techniques

So far video compression techniques have rarely been used for digital supertrunks and are therefore not considered here. Early systems used 7 bit resolution, achieving video qualities far below the ones achievable by using FM. Today's designs use 8 or 9 bits. The quantization noise is:

$$\text{SNR} = 6n + 1.8 \quad \text{(dB)}$$

n is the number of bits. This is an rms number and needs correction, when applied to video. Assuming no overhead and a range of conversion of the video signal with -40 IRE corresponding with 00..0 and +100 IRE with 11..1 then 9 dB can be added because video SNR is referenced to a peak to peak number (black to white). 3 dB has to be subtracted because video SNR uses 100 IRE as the signal reference and not 140. Video SNR is:

$$\text{SNR}_{(\text{video})} = 6n+7.8 \quad \text{(dB)}$$

and weighted video SNR is:

$$\text{SNR}_{(\text{video,weighted})} = 6n+7.8+7.4 \quad \text{(dB)}$$

(assuming flat quantization noise). Several mechanisms can produce noise that increases with video frequency. Therefore the above equation is too optimistic by 1 to 3 dB.

The theoretical numbers are:

Resolution:	7	8	9	10 (Bits)
Video SNR:	57.2	63.2	69.2	75.2 (dB)
Practical Video SNR:	55	61	67	73 (dB)

8 bit is a sufficient resolution for supertrunks. Going to 9 bits increases cost more than proportionally and improves the system video SNR after the set top converter by only 0.19 dB.

2.3.1. TDM systems

Time division multiplex (TDM) is attractive because of relatively inexpensive high speed digital multiplexing and demultiplexing IC's [4]. A system with 8 bits resolution and approximately 2 bits overhead for BTSC stereo transmission and synchronization needs at least a data rate of 107 Mb/s. A 16 channel system runs at a 1.8 Gb/s data rate.

2.3.2. Digital modulation of RF carriers

A well established way of transmitting digital data over phone lines, satellite links etc. is using digitally modulated carriers. Typical modulation formats are FSK, PSK, ASK etc. [5]. A lot of work has been done to find modulation formats that make an efficient use of the available spectrum as well as of the power capability of the transmitters. In satellite down links it is of great importance not to waste power by using up to 10 dB of backoff [6]. Modulation schemes that produce non constant envelopes require backoff because operating a transmitter near compression restores sidebands that have been filtered. QPSK modulation is well known for this phenomena. A multi channel fiber link is very different in that respect. Bandwidth of up to approximately 2 GHz is readily available. The only device operating close to compression is the laser, which is periodically driven into clipping. The nonlinear distortions are well known CSO, CTB etc. They can be considered additional noise and limit system performance well before the unwanted sidebands of the individual channels are restored. QPSK is therefore a very good candidate for digital carrier systems on fiber because of its simplicity. It has a bandwidth efficiency of 2bits/Hz, so that a 107 MB/s channel occupies approximately 54 MHz. A 16 channel system can therefore be realized with 900 MHz of RF bandwidth, slightly higher than FM.

2.3.3. Comparison TDM/QPSK

TDM-NRZ and QPSK have theoretically the same bandwidth efficiency. In practice TDM needs slightly more bandwidth than QPSK. A TDM system requires a very flat amplitude and group delay response of the transmission path, down to very low frequencies. A 3 dB roll off over the 1 MHz to 1.2 GHz transmission path can cause nearly a 30% reduction of the eye opening. That means in practice that the bit error rates are higher than calculated from the received CNR. A 3 dB amplitude error is what FM system FO links typically achieve when they are well maintained. The 3 dB roll off does not affect FM or QPSK link performance at all.

Another very important difference is reliability (or better availability) of the supertrunk. A TDM link that has a problem means that all channels are down whereas in an FM or QPSK link only the channel which has a problem is affected.

QPSK links are basically analog links in the RF section. It is therefore very easy to add other analog channels like FM stereo, satellite IF signals etc.

Although, TDM signals can be repeated nearly endlessly, we have not found the need for more than 2 to 3 repeats even in advanced redundancy schemes. Multichannel QPSK signals can easily be repeated by that number.

A 12 channel QPSK link can be realized using approximately 700 MHz of RF bandwidth. It can easily be expanded to a 24 channel link by optically combining 12 channels that modulate one laser up to 750 MHz and 12 more channels modulating a second laser from 900 tp 1700 MHz. TDM would have to multiplex in the time domain, a task that is difficult above 2Gb/s.

Figure 1. shows how a 24 channel QPSK system can be configured.
An other important difference is the ease of maintenance of a QPSK system. A standard spectrum analyzer is enough to locate problem channels. No Gbit test equipment is needed as would be the case for TDM.

2.3.4. Comparison AM/FM/QPSK

The following table shows the most important differences when 4mW transmit power is used (FM and QPSK after conversion to AM channels):

	AM	FM	QPSK	QPSK
Number of channels:	10	16	12	24
Video SNR:	60	60	60	60 (dB)
Loss budget:	9	29	30	26 (dB)
RF bandwidth:	0.06	0.7	0.7	1.4 (GHz)
CTB:	65	70	70	70 (dB)
CSO:	70	70	70	70 (dB)
Maintenance:	high	med.	low	low

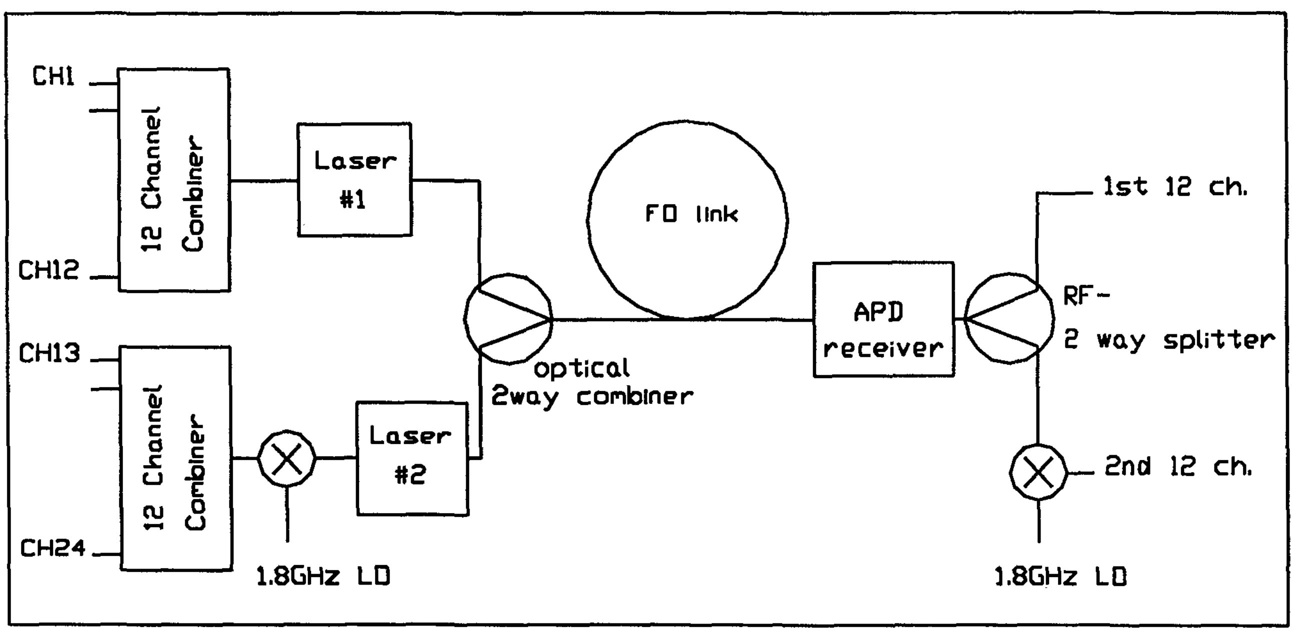


Figure 1.
A 24 channel QPSK system

Fiber optic systems have like other systems a certain hidden cost in maintenance. Fiber optic systems in general have bigger variations in CNR, CTB, and CSO than coaxial links. Especially short systems like AM systems depend heavily on the dynamic performance of the laser. CNR, CTB, and CSO can vary substantially with reflections, temperature changes in the optical isolators, etc. Ironically AM systems need to be planned therefore with more margin than FM or digital systems. Maintenance cost is a function of that margin. With no margin at all a system has to be maintained on a daily basis. 3 dB margin in optical power brings maintenance cost into a reasonable range. 6 dB makes maintenance cost negligible. The real numbers are difficult to obtain. Figure 2. shows approximately how maintenance cost of AM, FM, and digital compare.

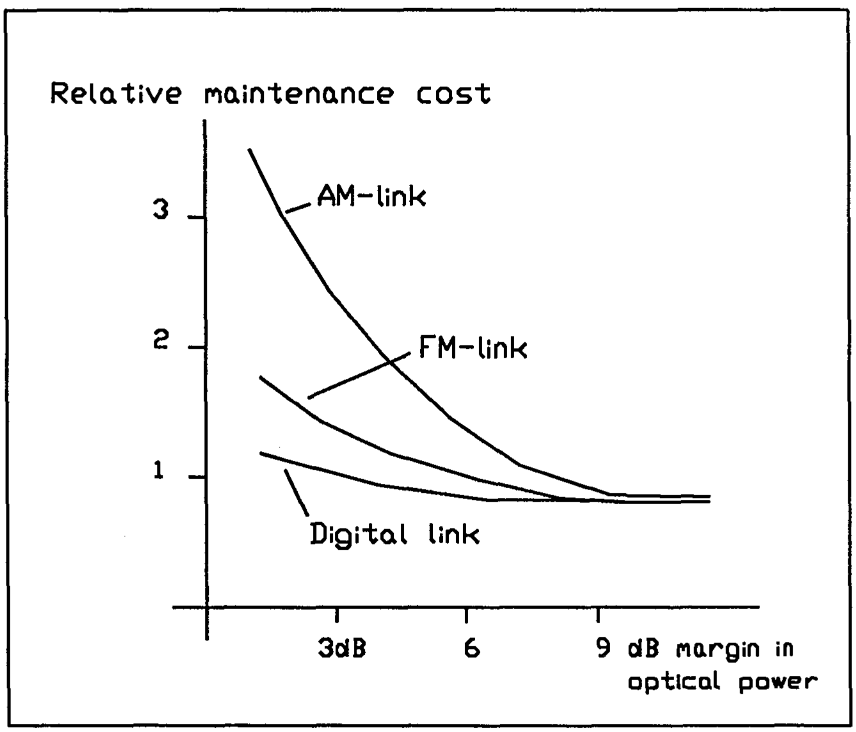


Figure 2.
Maintenance cost of AM, FM, and digital as a function of optical power margin

3. Supertrunk concepts using digital

The biggest advantage of digital supertrunks in comparison to AM supertrunks is their nearly 20 dB higher loss budget as well as their lower maintenance. The down time of a critical supertrunk system can be made to be nearly zero, when each hub is reached by two fibers that do not have common paths (fiber breaks normally cut all fibers in one cable). Figure 3. shows such a system.

Eight Supertrunks deliver the signals from the Headend to eight Hub Sites (A,B,C,D,E,F,G and H). The signals reach Hub Site A through one main fiber. If there is a fiber break in the main path, Hub Site A switches to a redundant fiber coming from Hub Site B.

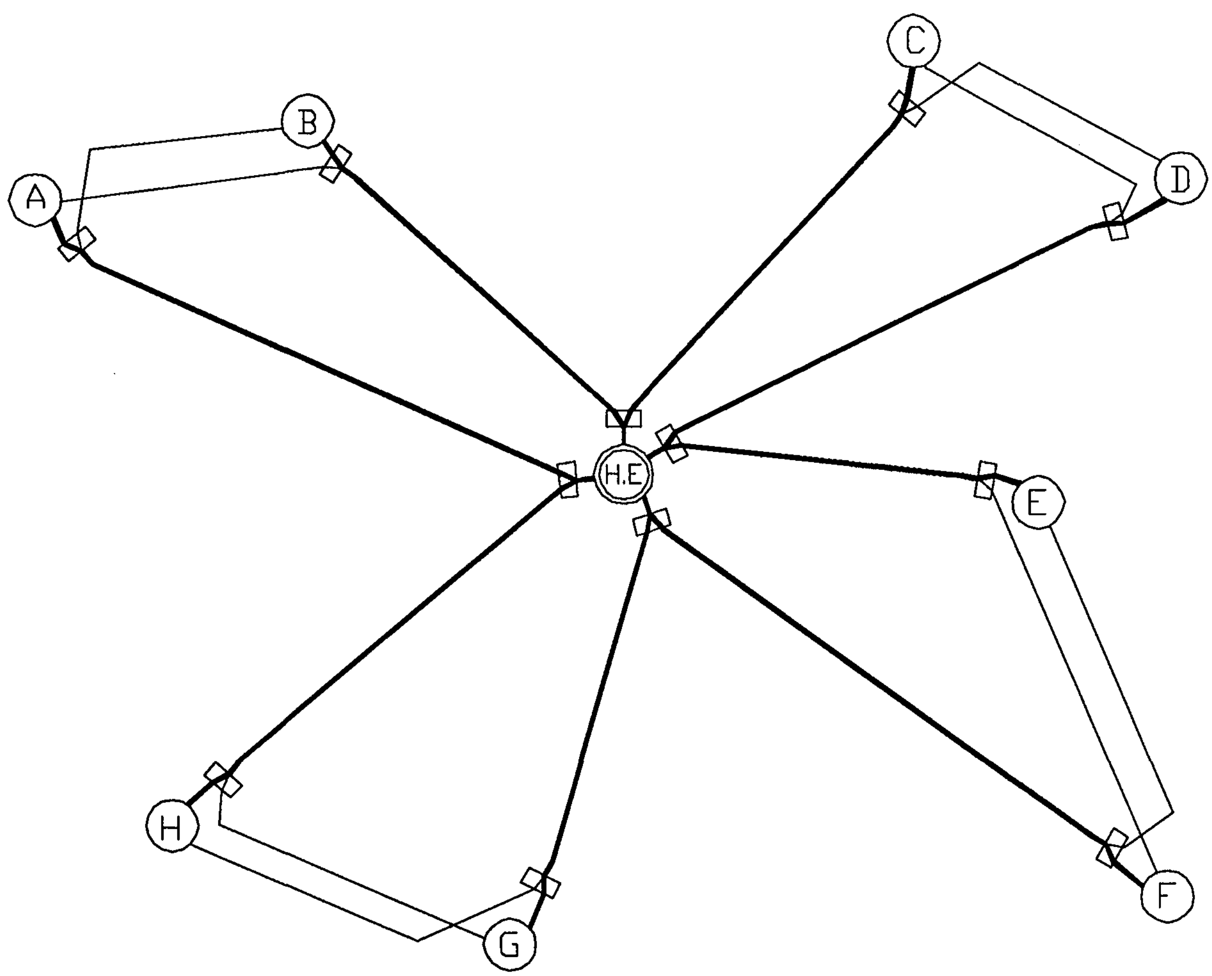


Figure 3.
A redundant supertrunk using independent redundancy paths

Supertrunk availability can be defined as:

$$A = 1 - \text{down time}/\text{total time}$$

With three days down per year that availability would be 0.992. If a redundant supertrunk is used the new availability number is:

$$Ar = 1-(1-A)^2$$

or in the above example 0.99993 or an average down time of only 36 min/year. By going to a redundant (and independent) path the reliability of the system has therefore been improved dramatically.

The price to pay is additional fiber installation as well as higher optical loss budgets. Digital (as well as FM) can perform with the higher loss budget, AM cannot. But AM has a reliability advantage because no modulation conversion equipment is needed.

4. BTSC stereo

In digital links video and audio is normally encoded in a base band format. Special schemes have been developed for the transmission of BTSC stereo so that the hubs do not need BTSC stereo encoders. The cheapest way is to add a 4.5 MHz subcarrier to the video before digitization. This method has some drawbacks:
A. The resolution that is available for video is reduced.
B. Video overshoots cause 920 kHz beats as well as audio buzz when the A/D converter is overdriven.

In FM we used discrete audio carriers very successfully. There was no interaction between audio and video whatsoever. We did the same with digital. The BTSC stereo signal produces an independent bitstream that is digitally multiplexed to the one produced by video.

5. What about scrambling

One of the biggest advantages an AM supertrunk has is that there is no need to treat scrambled signals separately. We have developed (and applied for a patent) a scheme that takes Baseband or RF scrambled signals down to a video baseband signal that can be transmitted by FM or digital. In the case of RF scrambling timing information on the sound IF carrier has to be transmitted as well. Our BTSC transmission scheme does this. All scrambling methods can be transmitted, keeping in mind that what we really transmit is the in phase component of the VSB envelope, or the information that the TV receiver really needs. Even a phase modulated signal (Zenith PM) can be handled. It will produce positive and negative envelopes, therefore reducing the SNR by 6 dB.

Digital transmission of these signals is very advantageous when dynamic video inversion is used, where DC stability is critical.

Conclusions:

The achievable picture quality in a carefully designed CATV system can be very high. The limiting elements are set top converter (and/or TV set) and to some degree the satellite dish and receiver. Supertrunks should be designed to achieve 60 dB video SNR or better. AM on fiber can do that with 10 channels when the 5% best lasers are selected and when no margin is needed. FM and digital can achieve that number, digital achieves 60 dB consistently, FM can achieve even better numbers in shorter supertrunks. AM has a limited loss budget that is approximately 20 dB less than what can be achieved using FM or digital. Supertrunks are critical for the reliability of a CATV network. Redundant schemes can easily be implemented using FM or digital but not with AM. The scrambling advantage of AM is irrelevant, FM and especially digital can transmit scrambled signals as well.

References:

[1] Communication with Ken Cannon, Scientific Atlanta

[2] E.J. Flynn et al: Performance and reliability of laser devices for CATV, Proceedings SCTE fiber optics 1990, Monterey

[3] Carl J. McGrath: Broadband AM lightwave transmission systems, a technology and applications review, Proceedings SCTE fiber optics 1990, Monterey

[4] John T. Griffin: Practical realization of a 16 channel fiber optic digital supertrunk for CATV, Proceedings SCTE fiber optics 1990, Monterey.

[5] John D. Oetting: A comparison of modulation techniques for digital radio, IEEE transactions on communications, Vol. COM27, N0.12, December 1979

[6] K. Sreenath et al: QAM and QPRS digital broadband cable systems, International journal of digital and analog cabled systems. Vol.2, pp. 139-148, 1989

FIBER OPTICS AND UHF IN A CATV NETWORK

Bert Henscheid
Texscan Corporation

INTRODUCTION

This paper describes a system that utilizes the extremely wide bandwidth of a fiber optic network by applying only UHF carriers in the range of 470 to 850 MHz. Such a system has many advantages and few disadvantages and would suffice as an international amplifier system.

DESCRIPTION

A fiber/UHF network will consist of the fiber network, an optical receiver with UHF amplification, and UHF trunk and distribution amplifiers.

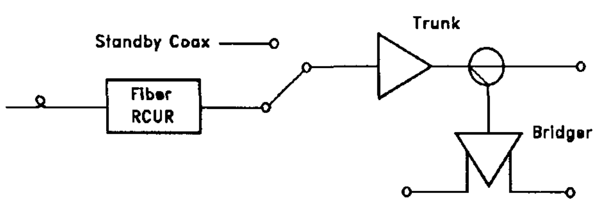


Figure 1 Fiber Optic Receiver Station

The performance of lasers and detectors is essentially constant over the band up to 1 GHz. That is, the lasers and detectors have the same distortion performance for modulation frequencies from 47 MHz to 850 MHz. The distortion performance is only sensitive at this time to the number of carriers being used. Therefore, no improvement in optics is needed for operation up to 1 GHz other than that which is desired for use up to 550 MHz.

The UHF spectrum is single-octave, and therefore, no second-order distortion is generated in the band of interest. The amplifiers could, in fact, be single-ended (i.e., no push-pull). Given a bandwidth of 470 to 850 MHz, all second order beats fall outside this passband. The sum beats fall above 940 MHz and the difference beats fall below 380 MHz. Since the push-pull hybrids are so commonly used and since they provide some improvement in third-order distortion, it is likely that the amplifiers will contain push-pull units. For either single-ended or push-pull amplifiers, good hybrid modules to 850 MHz do not exist and will have to be designed.

The ALSC attenuator circuits may be much simpler than VHF amplifiers, since the cable footages are shorter, the bandwidth is narrower, and the slope differential (i.e., tilt between 470 and 850 MHz) is smaller.

The tilt in cable attenuation from 470 to 850 MHz is about 6.4 dB in a 22 dB span at 850 MHz. Over the temperature range of $\pm 70^{\circ}$ F, this tilt would vary less than ± 0.7 dB. With an ALSC module at every station (a current trend), the slope control attenuator needs only ± 0.7 dB range, thus simplifying the slope circuitry.

Cable Loss		Loss	Slope	Gain
850 MHz	470 MHz	Diff.	Range	Range
24 dB	16.8 dB	7.2 dB	1.2 dB	0 dB
22 dB	15.4 dB	6.6 dB	0.6 dB	-2 dB
20 dB	14.0 dB	6.0 dB	0.0 dB	-4 dB
18 dB	12.6 dB	5.4 dB	-0.6 dB	-6 dB
16 dB	11.2 dB	4.8 dB	-1.2 dB	-8 dB

Figure 2 Gain and Slope Range with ALSC at Every Second Amplifier

The gain control attenuator requires a ± 2.5 dB range (or a total of 5 dB) which is fortunate, as it allows the attenuator to track better over the frequency range. Again, since the bandwidth is less than one octave, good tracking will be easier to obtain.

Cable Loss		Loss	Slope	Gain
850 MHz	470 MHz	Diff.	Range	Range
24 dB	16.8 dB	7.2 dB	0.6 dB	0 dB
22 dB	15.4 dB	6.6 dB	0.0 dB	-2 dB
20 dB	14.0 dB	6.0 dB	-0.6 dB	-4 dB

Figure 3 Gain and Slope Range with ALSC at Every Station

ADVANTAGES

There are many advantages to using only UHF in a fiber system. If single-ended circuits are used, the cost and power consumption is lower. There are no second-order beats or CSO.

There are NO CLI PROBLEMS, since there are no other radio services in this band and particularly, no aircraft frequencies. The only worry about signal leakage is theft of service. There are no FCC regulations on signal leakage in this band in the U.S.

Return filters will be extremely simple. The current usage of return signals exists in a relatively narrow bandwidth. If these return carriers are left in the 5 to 30 MHz band, very little filtering is required to separate the bands. Zero ripple Butterworth filters would work very nicely instead of the complex elliptic filters in use today.

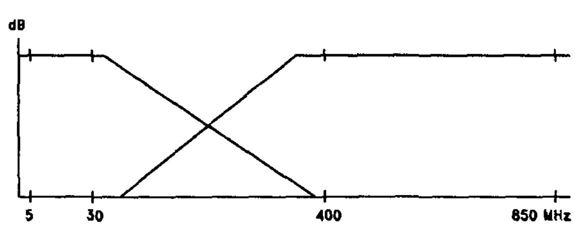


Figure 4 Forward and Reverse Filter Passbands

Equalizers are easier to design in single-octave applications because of the narrower bandwidth. It is likely that few, if any, control adjustments would be necessary. Again, due to the single octave bandwidth, response ripple will be more easily controlled.

DISADVANTAGES

UHF channels would mean UHF converter/descramblers for pay channels. In the U.S., TV tuners have inadequate stability, signal handling capabilities, and selectivity. The UHF tuners in Europe have satisfactory stability and signal handling capability, but also suffer from poor adjacent channel selectivity. In a 20 channel system with carriers spread every second or third space, the selectivity may be adequate.

A major problem at this time is the fact that hybrid modules do not exist that can handle over 10 or 20 channels. However, these few channels may be adequate for many applications.

Hi-Q pilot carrier filters will be more difficult to design; however, helical filters are smaller and readily available. SAW filters may also be usable, but their insertion loss is high.

Passives and taps in the UHF region are in common use in Europe, but would have to be developed for use in the U.S.

Drop cable would, no doubt, be RG-6 or equivalent because of its lower attenuation at UHF compared to RG-59 type drop cable. Type 6 cable has an attenuation of about 5.9 dB at 850 MHz and type 59 cable has an attenuation of 7.4 dB. The difference is 1.5 dB.

Trunk amplifier spacing will be relatively short. Assume 22 dB trunk spacing with 0.750 cable. At 850 MHz, the spacing will be about 1375 feet or roughly 4 amplifiers per mile. (Running miles not strand miles). Using 0.875 cable will increase the spacing about 14 % or to 1570 feet in this example.

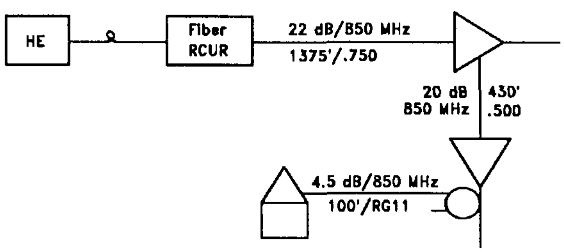


Figure 5 System Diagram

With a typical line extender gain of 20 dB (probably higher in the future) and a typical span of 10 dB of cable and 10 dB of loss due to taps and passives, feeder spans will average about 430 feet with 0.500 cable.

Typical cable losses per 100 feet at 850 MHz are:

0.875	1.4 dB
0.750	1.6 dB
0.500	2.3 dB
Type 6	5.9 dB
Type 59	7.4 dB

HYBRIDS

There are currently only two hybrid amplifier modules available from two manufacturers, respectively, that are designed for 850 MHz, but they were designed for a different application. The distortion performance of these devices is marginal for this application. It is anticipated that with the demand for this kind of product, a suitable device will be produced by the hybrid vendors.

Be that as it may, one can calculate (actually estimate, since hard data is not available) the system performance with currently available technology. Figure 6 lists estimated system performance for the fiber link, fiber receiver, one trunk, one bridger, and one line extender. The operating levels were chosen for best dynamic range with, hopefully, reasonable levels.

Note that the carrier-to-noise ratio (C/N) is set by the fiber link at about 51 dB and the composite triple beat (CTB) for 20 channels is set by the bridger and line extender at about 55 dB. C/N is adequate, but CTB is marginal, depending on the application.

Two trunk amplifiers would degrade the CTB by about 1 dB (i.e., -54 dB instead of -55 dB) and would not appreciably alter the C/N.

	FO Link	Trunk Ampl	Bridger	1 LE	EOL
Level (dBmV)Flat		+30	+40	+40	--
C/N (dB)	51	60	66	66	50
CTB (dB)	71	-80	-63	-63	-55

Figure 6 Estimated performance analysis with 20 channels

Sixty channel applications will require a 9.5 dB improvement in hybrid CTB performance. Perhaps some of this improvement can be gained by changing the transistor die to a current generation. The remainder could be achieved by designing a power doubling version. Alternatively, suitable devices could be obtained by pushing current 550 MHz units to 850 MHz and optimizing performance between 470 and 850 MHz.

COSTS

No cost figures are available, since some of the key items are not yet available. It is anticipated that individual equipment costs would not be significantly greater than an equivalent VHF system. Some costs would be greater, but some savings would also result as described above. However, as discussed, the shorter spacing will require more equipment than an equivalent VHF system.

CONCLUSION

A single-octave fiber/UHF system has many advantages; however, all of the technology is not yet available. The missing technology is within reach, but some work remains to achieve it. All that is needed is the desire to build it. The first practical application would probably be in Europe where UHF TV channels are more prevalent and in widespread use.

FIBER TRUNK AND FEEDER -- THE CONTINUING EVOLUTION

BY: David M. Pangrac & Louis D. Williamson

American Television and Communications Corporation

ABSTRACT

Cable Television technology has been in a constant state of evolution since the first crude mountain top installation many years ago.

Since the beginning of this industry, we have seen the introduction of hard line coaxial cable, solid state amplifiers, directional taps, satellite delivered programs, addressable converters and, in 1988, the introduction of the "Fiber Backbone" architecture which made use of VSB-AM modulated optics and a significant amount of fiber optic cable.

This paper will focus on ATC's evolution of the "Fiber Backbone" into the new "Fiber Trunk and Feeder" architecture designed for new builds and rebuilds.

The basic concept will be examined as well as the technical specifications.

INTRODUCTION

In 1988 when the "Fiber Backbone" architecture was introduced, its primary purpose was to enable cable operators to cost effectively upgrade their plants, improve reliability and picture quality. The key to making the project a reality was the ability to use VSB-AM modulation to drive the lasers used for fiber optic transmission.

The first lasers used were marginally acceptable with CNR of 48 dB to as high as 51 dB and CTB approaching -63 dB. CSO numbers of -54 dB to -57 dB were also common. Channel loading ranged from 12 to 40 per laser.

In the last two years, a significant improvement in the performance of lasers has occurred. With many laser manufacturers developing products specifically to be used for AM applications, it is now possible to obtain equipment capable of 55 dB CNR, CTB and CSO at -65 dB or better and power budgets as high as 10 dB. Channel loading at these levels have been as high as 42 and some "new" lasers have been close to these specifications with as many as 80 channels.

This performance has allowed the next evolution of the "Fiber Backbone" known as "Fiber Trunk and Feeder".

"FIBER BACKBONE" LIMITATIONS WHEN USED WITH "NEW BUILD" OR "REBUILD" SCENARIOS

The original concept of the "Fiber Backbone" architecture was to allow a system operator to reuse the most expensive part of his plant during a bandwidth expansion project, his cable.

ATC has determined that about 58% of the cost of a cable plant is made up of the cable, strand, hardware and labor to install it. The balance, 42%, includes the plant passives and electronics.

By using fiber to transport the signals from the head end to points deep in existing amplifier cascades, we are able to develop small "neighborhood" cable systems. (See Figure 1) The heart of the small cable system is the optical node where the "light" is converted back to "RF" which then feeds the small coaxial tree and branch system. The short amplifier cascades create an increase in our distortion head room budget that can be used in various ways.

New broadband electronics are installed in the same locations as the original equipment and in some cases, physically turned around. (See Figure 1) The increased distortion head room budget created by the short amplifier cascades can then be used to overcome the cable loss at the higher frequencies and improve picture quality at the same time. (See "Off Premises Broadband Addressability: A CATV Industry Challenge", by James A. Chiddix and David M. Pangrac, 1989 NCTA Technical Papers.) This then results in a cable TV system that has a greater bandwidth, better picture quality and improved reliability over the original system, but costs about 50% less than a new, conventionally built tree and branch system of the same bandwidth. (See "Lake City Cablevision: A Case Study in the Fiber Optics Application", by Ronald W. Wolfe, Proceedings Manual: Collected Technical Papers, SCTE Fiber Optics 1990.)

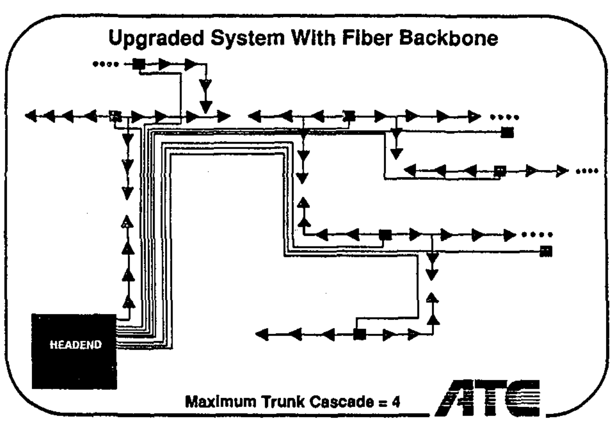


FIGURE 1

While the "Fiber Backbone" architecture works well for a system "upgrade", it is not financially well suited for a new or rebuild application. When a new or rebuild scenario is looked at, using "Fiber Backbone", it can be seen that almost the same amount of trunk cable and amplifiers are needed as would be used for a conventional build.

In addition, fiber and fiber optic electronics must also be installed. Since there is no cost savings available from reusing existing cable as in an upgrade, the cost of the fiber, fiber construction and fiber optic electronics become a significant incremental cost to a conventional plant. This can add as much as $2,600 per mile to the build.

"FIBER TRUNK AND FEEDER"

"Fiber Trunk and Feeder" (FTF) makes use of the technology that was developed for the "Fiber Backbone" but changes the basic cable TV architecture to simplify it and to make more efficient use of fiber. While the "Fiber Backbone" achieves its economics by leaving the coaxial cable in place during an upgrade, the FTF system achieves its economics by eliminating the labor and material needed to build the trunk portions of a rebuild or new build. But the system does more than eliminate the coaxial trunks, it also reduces the number of actives to a maximum of five for any subscriber in the system. These features make the system both economical and reliable.

As can be seen in Figure 2, the concept takes fiber deeper into the system than was previously possible. The optical equipment needed in the system is very similar to the current equipment being built for the "Fiber Backbone". It requires an optical transmitter/receiver pair capable of 54 dB CNR, -65 dB on the other distortions and a 10 dB power budget.

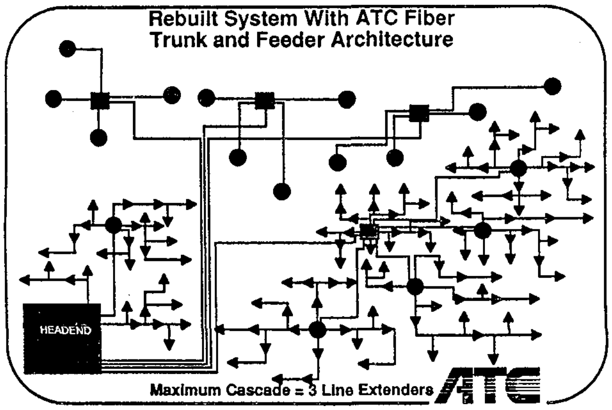


FIGURE 2

A CLOSER LOOK AT THE TECHNICAL SPECIFICATIONS

The signal performance that ATC desires at the tap is shown below.

Desired Tap Performance

	Today	Future
CNR	46 dB	49 dB
CTB	-53 dB	-54 dB
CSO	-53 dB	-54 dB

The RF portion of a FTF plant, which consists of three push pull line extenders, will contribute the following:

CNR	56.4 dB
CTB	-59.5 dB
CSO	-66.8 dB

To meet ATC's future desired tap specification, the fiber portion of the plant must have the following minimum specification at the output of the receivers:

CNR	50 dB
CTB	-59 dB
CSO	-60 dB

As long as the performance at the receiver meets the specification shown above, the tap performance will be met. This allows for a wide variety of configurations from 8 way optical splitting to optical repeating. By using various combinations of laser and splitting networks, all of a CATV system can be served with the FTF architecture.

DEPLOYING THE "FIBER TRUNK AND FEEDER" ARCHITECTURE

The area immediately around the headend is served by three line extender cascades that originate from the headend. The reach of the three line extender cascades is approximately one mile.

After this initial area, the remainder of the plant is served by optical systems. The optical system that is required is assumed to have the following specifications:

CNR	54 dB
CTB	-65 dB
CSO	-65 dB
Loss Budget	10 dB
TV Channels	60

Since this level of performance is not required at the tap, some of it is traded for a larger loss budget. If the loss budget is increased to 12 dB, the CNR of the system will be decreased to 50 dB. The distortion should be unaffected. This allows you to serve the distance and still achieve the desired optical specification at the receiver. (See Figure 3)

FIBER TRUNK AND FEEDER CONFIGURATIONS

DISTANCE (Miles)	NODES Served	SPLITTING Loss (dB)	PATH Loss(dB)
1	NA	NA	NA *
1.6	8	10.7	1.3
3.5	6	9.2	2.8
6.1	4	7.1	4.9
10.8	2	3.3	8.7
14.3	4	7.1	11.75 **

* 3 line extender cascade
** optical repeater needed

FIGURE 3

As is shown in Figure 3, most of a system can be served with passive splitting of lasers that are kept at the headend. It is not until you reach a distance of more than 10.8 miles that an active optical repeater is required.

The optical repeater is shown in Figure 4. Since the optical repeater has to convert the light back to RF to re-modulate the laser, the area immediately around the repeater is fed by three line extender cascades that are driven with this RF signal. The remainder of the RF signal is used to feed the laser for the second optical path. This laser feeds a four port optical coupler, which then feeds four secondary receivers. The secondary receivers can be a maximum of 1.75 dB from the repeater. (See Figure 5)

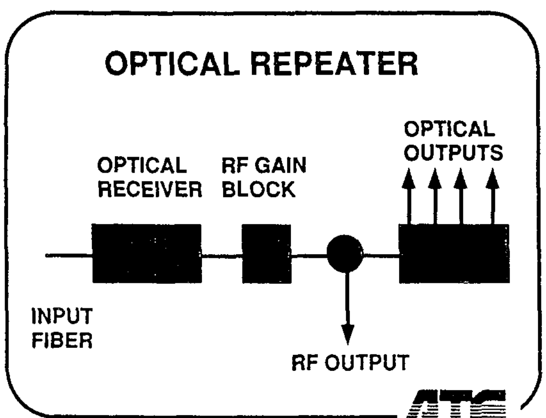


FIGURE 4

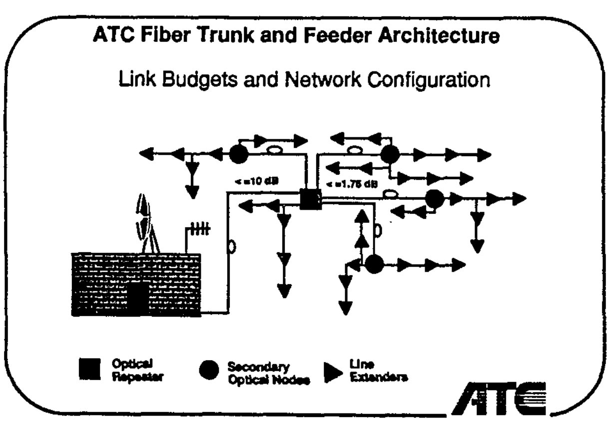


FIGURE 5

In test designs completed by ATC, it appears that each secondary node/three line extender cascade combination can serve an area covering about the same geographical size as four conventional bridgers designed to the same bandwidth.

The diagram in Figure 6 shows an eleven hundred mile plant that was designed using only passive links. The actual design is not shown (for the sake of clarity) but rather the concept that was used to cover all parts of the system with the passive links. The system had two headends that were connected via AM super trunks and the passive links were then served from these locations.

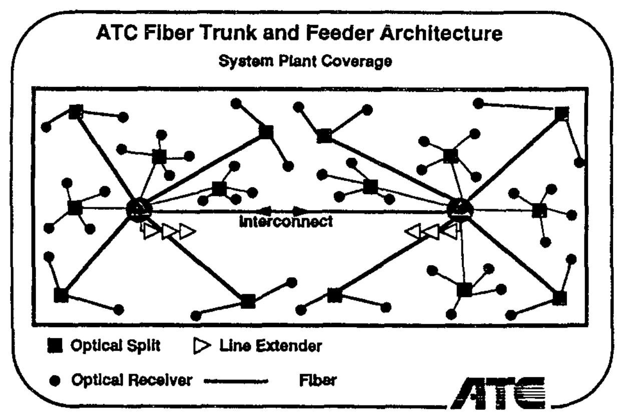


FIGURE 6

CONCLUSIONS

Calculations made by ATC have indicated that when using the "Fiber Trunk and Feeder" architecture to rebuild a 450 MHz plant, the cost for this fiber intensive system will be the same or a little less than a conventional tree and branch coaxial plant.

Nothing has to be invented to make this architecture a reality. The performance of many of the required lasers need be only mediocre by today's standards. The most challenging piece of equipment needed is the inexpensive optical receiver which needs very little development work to become a production item.

With all the benefits of fiber and no increase in cost over a conventional system, the development of this new architecture, "Fiber Trunk and Feeder", will now be the design of choice for ATC, starting this year.

Dave Pangrac is the director of engineering and technology for American Television and Communications Corporation (ATC), the country's second-largest cable television operator.
Pangrac has been in the cable television business for 23 years. He joined ATC in 1982 as Vice President and chief engineer for American Television of Kansas City and in 1987 joined the ATC corporate staff as director of engineering and technology.
Pangrac is a member of the Society of Cable Television Engineers and past president of the Hart of America Chapter. In 1989 he was awarded the NCTA Vanguard Award for Engineering and Technology.
Pangrac is currently involved in ATC's effort to develop the use of fiber optic technology in cable television plants.

Louis D. Williamson received his B.S.E.E. degree from Virginia Polytechnic Institute and State University in 1980. From 1980 to 1983 he was employed at Martin Marietta Aerospace in Denver, CO as a Design Engineer.
Since 1983, he has worked for American Television and Communications Corporation as a Member of the Technical Staff. His responsibilities involve applying new technology to CATV. His is also involved with HDTV and serves as Vice Chairman of the Standard Working Party of the System Subcommittee of the FCC Advanced Television Committee.

FIBER-OPTIC PASSIVE COMPONENTS FOR FUTURE SYSTEMS

by

Scott A. Esty

James E. Matthews III

Corning Incorporated

Abstract

As fiber proves its worth in a variety of cable TV system architectures, system engineers can make greater use of fiber-optic wavelength division multiplexers, splitters or couplers, and directional coupler taps to reduce costs and provide for increased system utilization and expansion in the future.

This paper will focus on the basics of coupler technology with specific explanations of the functional parameters that need to be specified when purchasing these components. The presentation also will discuss the reasons for this emerging trend, specifically geared to the needs and interests of cable television operators.

I. UNDERSTANDING COUPLER TERMS AND PARAMETERS

Before looking at how cable TV systems can use various types of couplers, it is important to understand the basics of couplers and a little bit about how they are made.

Couplers come in many shapes, types and flavors. Luckily, they can be grouped into a few basic categories:

* Trees or Even Ratio Splitters: Power is evenly divided into, or combined from, two or more ports.

* Variable Ratio Directional Couplers or Taps: Power is unevenly split or combined between two or more ports.

* Multiplexers/Demultiplexers: Power is combined or separated according to wavelength (frequency) of light.

* Stars: Light from two or more inputs is combined and then split into two or more outputs.

As we will discuss later, these types of couplers can be used as discrete components, or several can be combined in one location to provide multiple capabilities over a single fiber.

Coupler Parameters

The most important functional parameter for any coupler, regardless of its type, is insertion loss. Insertion loss, usually expressed in dB, is how much optical power is lost from the transmission system when the component is added to the system. It refers to the total optical loss from an input port to an output port. In order to make an informed choice when comparing devices, it should be noted that insertion loss is not necessarily specified the same way from one manufacturer to another.

True insertion loss should be specified over the operating passband of wavelengths of the system, rather than at one nominal wavelength value. Many devices have what appear to be very low losses at a nominal wavelength, but exhibit somewhat higher losses over the wavelength range where systems typically operate.

This can be hidden by specifying a second term, such as wavelength dependence or coupling ratio tolerance. Both of these approaches are misleading to the user because optical power budgets are calculated in dB; the user, who must make the conversion anyway, often is surprised to find the true insertion loss is higher over the passband than originally thought. The loss over a passband is essential because optical sources are never sold or specified for applications at a single wavelength. Laser sources of the type used for cable television AM transmission often are specified with a nominal distribution of ± 10-15 nanometers (nm) from a nominal wavelength, with an additional allowance of about 5 nm shift in the device operating wavelength over time, due to aging and temperature-induced change.

In addition to insertion loss over a passband, a prospective coupler user also should insist on a true maximum loss specification, rather than an average or maximum average insertion loss. Devices that are specified by averages seem to have a number that appears as a lower loss. The prospective buyer should remember that half of the couplers he or she buys will have losses higher than the specified number, perhaps with no real limit on the upper bound. Couplers specified to a maximum insertion loss over a passband is the specification method most friendly to users, and most compatible with the way couplers are used in real life.

After insertion loss, the parameter of most concern to cable TV systems is backscattering. Backscattered or back reflected light is optical power that is reflected by the device back toward the optical source on the input fiber. In many systems, especially AM cable TV transmission with narrow optical linewidth Distributed Feedback (DFB) lasers, the backreflected energy can cause the source characteristics to change. This results in more system noise or signal distortion. There are two ways backscattering should be considered: backscattering inherent to the device, and backscattering at the fiber pigtail to system joints. The inherent backscattering characteristic of the device itself is measured with all ports but one terminated optically in index-matching gel. This is done so that the backscattering measured is due only to the performance of the coupler itself, rather than a reflection from a fiber end.

When the coupler is installed in a working system, the pigtails either are fusion spliced or connectorized. Either way, the amount of light reflected from the fiber-to-fiber joint usually is much more than the light reflected from the coupler. For most couplers available today, the backscattering is specified to be better than -40 dB. This level of backscatter will not functionally degrade a system's performance.

Some applications will use bi-directional transmission over a single fiber. When two signals are transmitted at the same wavelength, a 1x2 coupler is used to split light between the transmitters and receivers as shown (Fig. 1). In this application, low backreflection is important throughout the optical system, but directivity also is an important factor. The directivity refers to the amount of light going out an adjacent port in the non-coupled direction. In figure 1, this would be the amount of light going from the transmitter, shown attached to Port A, through the coupler to Port B. Typical coupler directivity numbers are well below 40 dB from the input signal.

Another approach to bi-directional transmission over a single fiber could involve the use of two wavelengths, transmitting in opposite directions, or a combination of the two techniques above. In these cases a multiplexer/demultiplexer type of coupler is used. These couplers, known as WDMs, combine or separate signals by their wavelengths of light. The analagous device in the electrical world combines the function of a bandpass filter and a mixer. The functional limit to system performance is known as crosstalk. Crosstalk is the amount of light seen at a port where it is supposed to be blocked. Near-end crosstalk is where light from the transmitter can interfere with the adjacent receiver. Far-end crosstalk occurs where misdirected signals from the wrong wavelength could interfere with a distant receiver's operation. Like insertion loss above, it is essential to specify crosstalk over the passband of operation, rather than at a single wavelength.

As important as the optical characteristics of a coupler are, environmental and mechanical robustness should not be overlooked. Coupler insertion loss, usually in the form of coupling ratio, can change over a range of operating temperatures. Figure 2 shows the change in insertion loss of an achromatic single-window Corning™ coupler during repeated cycling between -40°C and +85°C. Note that the graph is blown up in size so that each division represents 0.1 dB change. Changes in insertion loss over the operating window of less than 0.2 dB are considered acceptable for environmental performance. Although each coupler application may not demand such rugged performance, the general trend is toward less sheltered, outside plant passive-network applications. This trend will mean that tomorrow's devices, many of which are being installed today, will have to operate over a wider range of temperatures, environments and optical passbands than ever before. Table 1 shows an example of typical parameters to be specified for a 1 x 2 splitter and a 90%/10% variable ratio directional coupler.

The potential quantity of couplers required to satisfy widespread fiber network implementation may present the most significant challenge to the industry, however. With hundreds of millions of potential subscribers worldwide, the implementation of fiber-based distribution networks will require industrialization of passive components and opto-electronic components and systems on an unprecedented scale. In fact, fabrication technologies must be optimized for high-volume production of higher-performance components, while ensuring quality and reliability.

II. TECHNOLOGIES FOR COUPLER MANUFACTURING

Over the years, a number of coupler manufacturing technologies have been developed by industry. While techniques and capabilities are diverse, fabrication technologies have not been amenable to large volume manufacturing. The challenge to manufacturers is to deliver in large volumes high-performance products such as low-loss achromatic (wavelength-indepedent) multi-window devices required by future coupler-intensive fiber-to-the-subscriber architectures.

Fused biconic taper

The most common manufacturing technique now in use is the fused biconic taper (FBT) technology. Two fibers are axially aligned, heated and then stretched until they become joined in a thin cross-sectional region that permits light to travel from one fiber to the other (Fig. 3).

The thin coupling structure typically is affixed to an intermediate package with epoxy to prevent mechanical strains on the coupling region. An external package over the coupling structure protects the coupling region from moisture and other environmental and mechanical effects. Additional packaging may be required to provide strength/retention for the fiber pigtails.

The FBT technique is best suited for the manufacture of couplers that display low losses at a discrete wavelength or over a narrow wavelength range, typically +/-10 nm (narrowband). FBT couplers generally have not met the achromaticity requirements of future networks. However, recent work on FBT couplers has improved their performance somewhat. Double window couplers operating at 1310 nm and 1550 nm can be fabricated by etching or tapering the fibers prior to fabrication, but this additional process step adds expense and complexity to the production process.

A variation of the FBT process recently developed by Corning addresses many of limitations of the single-mode FBT process. In this method, the coupling region of the fused fibers is embedded in a specially fabricated low-index glass, creating a hermetic seal. This larger, more robust coupling region results in a device more resistant to environmental and mechanical conditions.

This simpler process yields low insertion loss without the need for pre-etching or tapering the fibers in order to control double-window optical performance. Control over the fiber and sealing glass compositions allows fabrication of various product designs with precise optical characteristics.

Fusion technologies can, within practical limits, join only two fibers at a time with the required control of optical performance. Higher-order 1xN devices are made through a process of sequential splicing, or cascading and repackaging 1x2 couplers. Aside from the increased component size, the optical performance of the resulting devices is somewhat compromised by the splices.

Planar fabrication

Recognizing the inherent limitations of standard fused biconic technologies, another very different fabrication process is gaining interest.

Fabrication technologies are being implemented to produce couplers that meet the optical performance and mechanical requirements of future systems architectures, as well as volume demands by making an optical circuit in a planar piece of substrate material.

Couplers are produced much like electronic integrated circuits using standard photolithographic techniques to transfer mask patterns of numerous coupler structures onto a wafer. The substrate wafer can be made of glass, silicon or a Group III/IV compound (Fig. 4). The wafer then is processed to create actual waveguides in the pattern of the mask structures. The processed wafer is cut into discrete coupler chips, to which fiber pigtails are added before packaging in rugged housings suitable for outside plant use.

The optical guide profiles in the substrate generally are created by diffusion or material deposition processes. Corning and other companies have developed planar technologies using ion exchange techniques to create the waveguides. Others, such as Photonic Integration Research Incorporated, employ chemical vapor deposition techniques.

Calling on its engineering and production technology base from working with glass and its raw materials, Corning has developed a planar fabrication technology that creates optical waveguides in a glass substrate. The process yields optical circuits of varying complexity in a small, monolithic sealed glass structure.

Based on this ion-exchange planar technology, planar couplers are being developed in a number of designs, including single-mode 1xN tree couplers. The design of these units produces inherently low and level optical losses over the complete range of wavelengths planned for cable television fiber-based subscriber distribution systems, because true Y-branching of the light is used. Figure 5 shows the optical performance of a 1x2 coupler from 700 nm to 1660 nm. The solid line segments represent the three passband ranges that are likely for each operating window in future services applications. As shown, the coupler can operate simultaneously in the three bands. Within each band, insertion loss is constant across the operating window.

The coupling ratio of a planar Y-branching coupler, i.e. the ratio of power split between output ports, is uniform and remains consistent across the passbands, even in the 850 nm wavelength region where the guides actually are multimoded. However, in the case of short wavelength operation, some cautions apply. The use of highly coherent sources should not be used or modal interference effects can occur in the system.

Non-coherent sources such as LEDs or self-pulsating coherent laser sources do not exhibit these interference effects. Therefore, the coupler is compatible with combinations of current and planned source wavelengths.

Devices with multiple output ports on a single small piece of glass can be made through passive integration of sequential Y-junctions (Fig. 6). Using the approach, tree couplers with any number of output ports can be fabricated.

In addition to manufacturability in large volumes, planar-based couplers offer the potential to integrate several coupler functions into a single chip, resulting in a smaller overall package. For a 1x8 coupler, the size difference between a planar device and a packaged set of cascaded fused devices can be as much as 10 times or more (Fig. 7).

Having a wide variety of low-cost devices from which to choose is important to facilitate the implementation of a variety of architectures. Each architecture places a unique set of demands on coupler performance.

III. APPLICATIONS FOR CABLE TV

The specific requirements of cable TV transmission fit very well with the concepts of passive splitting of optical signal power. The cable TV world has pioneered coaxial/electrical technology from infancy to just about its maximum potential. Perfection of the coax tree and branch architecture has been a technological challenge well met. The next generation cable TV systems will be looking to do the same with lightwave technology, and passive optical components promise to play a major role.

Together, the industry can originate some creative ideas to optimize system performance while minimizing costs. The eventual objective is bringing the specific advantages of optical transmission deeper into the system and closer to the subscriber. Much of this early thinking has been done by telephone companies in their fiber-to-the-home drive. However, the telco imagination has been constrained by their existing service commitment to provide as much upstream capacity as downstream, and a copper-wire architecture that required a "hard-wire connection" to every customer, resulting in their existing real-estate commitments .

The earliest fiber architectures that emerged were star-structured designs with dedicated individual paths between each subscriber and his or her switching center. Economic analyses quickly pointed to the cost advantages of sharing the most expensive network elements, primarily the electronics, between several customers. An underlying trend in present and emerging designs for optical distribution systems at the trunk level and beyond (what we will call fiber-beyond-the-trunk or FBTT for simplicity's sake) are architectures using fiber-optic splitters or couplers to share electronics among many users. The common thread is the extensive use of passive optical components to reduce total installed system costs and allow for future system evolution.

Coupler-Intensive Architectures

The most coupler-intensive FBTT architectures currently under consideration fall into two generic categories: active and passive stars.

Active star FBTT architectures most often use intermediate electrical-to-optical (E/O) and optical-to-electrical (O/E) conversions to capture, remodulate and transmit signals to create channels actively or logically between the headend and subscriber. One example is an active double star architecture, with an intermediate hub site terminating an FM trunk and feeding AM video into fiber feeders out through the neighborhood. Depending on the need, the architecture can be configured for double-window (two wavelength) operation or single-window (one wavelength) operation. The second window would double the capacity for additional channels, provide on-line system diagnostics or some new interactive services that can be overlaid on the second wavelength.

For all distribution architectures, a pair of 1x2 tree couplers can be used to provide two-way transmission capacity over a bi-directional single fiber. One coupler is located in an optical network interface (ONI) module located at the front end of the link, and the other in an ONI module at the far end. The ONI executes the optical, opto-electronic and electronic functions necessary to interface with the coaxial cable plant for the drop into the home.

Additional subscriber services such as interactive home shopping or customized programming options can be supplied in the initial installation or phased in later as part of system upgrades. One method to add services and provide a partition between basic and enhanced services is to add optical channels by means of wavelength division multiplexing couplers. These couplers/splitters selectively channel most of the light power associated with a specific wavelength to the respective O/E (optical-to-electrical) interface. Such WDM couplers most likely would be located in the ONI at the headend and in an ONI near the subscriber premises.

Another active star design is a triple star architecture. In this design, fiber could be used between the headend and a pole or pedestal-mounted terminal, as described above.

Coax drops to the subscriber would deliver the basic entertainment services. Later, the architecture could evolve to fiber drops to subscriber premises through the use of 1xN tree couplers located in the pedestal (Fig. 8). This implementation, in essence, is a hybrid active/passive star architecture.

Passive Alternatives

A significant cost element in AM systems often is the linear AM laser transmitter. This must launch a high level of optical power, maintaining a linear relationship between the drive current and optical power. The tight performance specifications on these AM lasers results in a cost premium. Increasingly, 1x2, 1x4, or 1x8 couplers are being used to share or spread the cost of this transmitter between routes. AM cable television systems usually have a very limited loss budget, which places low-loss requirements on couplers and other system components. Another approach to optimizing the power balance among links in shared systems is to use variable ratio directional couplers, which are 1x2 couplers with an unequal power-splitting ratio. Wavelength division multiplexers also may be used in these systems to increase the number of channels per fiber or to enable on-line diagnostics. These couplers may be located either out in the cable plant or locally, nearer to the AM transmitter.

Table 2 lists the potential use of couplers by active, passive and cable TV backbone architectures.

IV. SYSTEM DESIGN TRADEOFFS

Implementation of coupler-intensive FBTT architectures must meet the system design goal of achieving the lowest installed cost per subscriber, while keeping open the technical potential for new revenue opportunities. These cost and design considerations being faced by system designers have a direct impact on the couplers required.

Multiple wavelengths can be used in FBTT systems to enable enhanced broadband services and an optical partitioning of the individual service levels. Multiple-window systems using WDMs also provide a means of cable plant sharing, conserving the fiber resource and helping to control costs. Furthermore, today's short-wavelength (~780 nm), CD-type lasers are lower in cost than their long-wavelength counterpart lasers. Use of short-wavelength CD-type lasers for low-bandwidth applications such as signaling and upstream communications provides a direct cost saving, but necessitates multiple-window system operation. Couplers must, in turn, meet system requirements for multiple-window operation.

V. COUPLER ENVIRONMENTAL REQUIREMENTS

The FBTT environment is characterized by harsh conditions, such as operating temperature extremes of -40°C to +85°C and humidity that reaches saturation levels. FBTT systems also face a variety of real-world intrusions such as flooding of leaky pedestals, exposure to unskilled personnel, or unintentionally left open ONI or pedestal doors, leading to exposure hardware to environmental extremes and animal and insect intrusion.

A number of tests have been established to evaluate the performance of couplers in the types of environments described above. The prospective user should insure that his or her coupler product has been fully tested for behaviour during exposure to the adverse conditions as well as after. This is essential because products in working systems must continue to function when the outside environment is hot, cold, or damp.

VI. COMPONENT CHALLENGES

The major significant challenge to fiber-optic component manufacturers, however, is the potential scale of FBTT deployment. There are hundreds of millions of potential subscribers around the world, so implementation of FBTT and telco subscriber loop activities will require production of passive components, optoelectronic components and systems in massive volumes. Manufacturing technologies must be optimized for high-volume production of high-performance components that exhibit the quality and reliability needed to take advantage of the enormous bandwidth of optical fiber and meet the projected long-life requirements of the coming systems. Moreover, the subscriber density leads to hardware space limitations, so require that systems and components must have small footprints to be accommodated in this high-density environment.

Thus, implementation of coupler-intensive FBTT architectures and the conditions under which these systems will operate impose severe system design constraints. These constraints, in turn, directly impact the requirements placed upon couplers.

As Table 3 illustrates, some single-mode architectures will place further demands on couplers, including wavelength independence (achromaticity); low insertion losses over a wide passband range; high isolation over wide passbands; short-wavelength operation; the ability to operate in multiple windows; and the ability to be fabricated in compact, monolithic structures. In addition, high-quality couplers will have to be manufactured in high volume, at low-cost. A number of vendors are developilng components that meet these challenges, and novel new technologies and devices are being introduced.

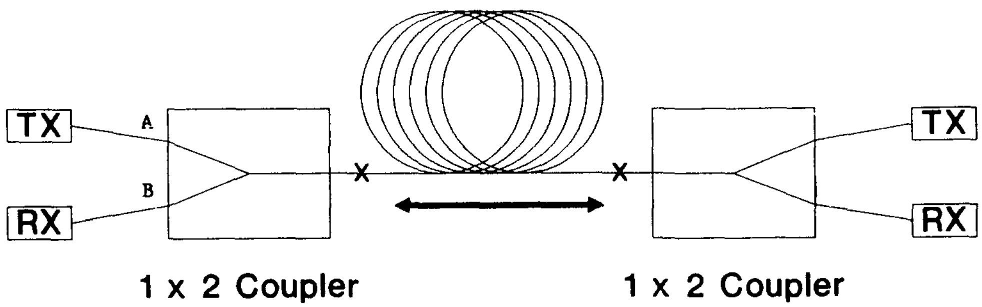


Figure 1. Use of couplers for bidirectional transmission over a single fiber.

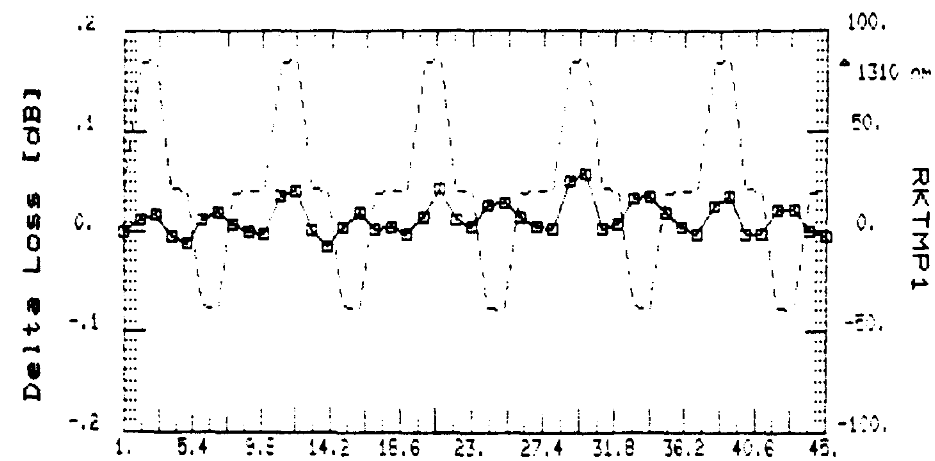


Figure 2. Change in insertion loss for a coupler port during temperature cycling.

Table 1. Coupler Specification Example for 1x2 50/50 Even Ratio Splitter and 90/10 Directional Coupler.

Element	Specification
Insertion Loss Over Optical Passband	3.8 dB for 50/50 Even Splitter 11.5 dB / 1.0 dB for 90/10 Dir. Coupler
Operating Temperature	-40°C to + 85°C
Maximum Insertion Loss Change over Temperature	± 0.2 dB
Change in Insertion Loss due to Polarization Effects	± 0.2 dB
Directivity	Better than 60 dB
Backscattering	Better than 55 dB
Tensile Strength on Fiber Pigtail	5 N
Tensile Strength on Tube/Cable	>10 N

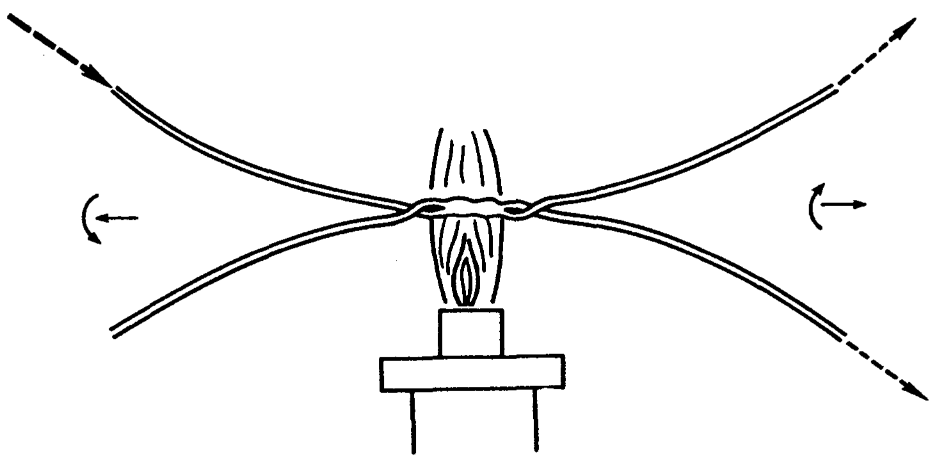


Figure 3. Fused biconic taper method of coupler fabrication.

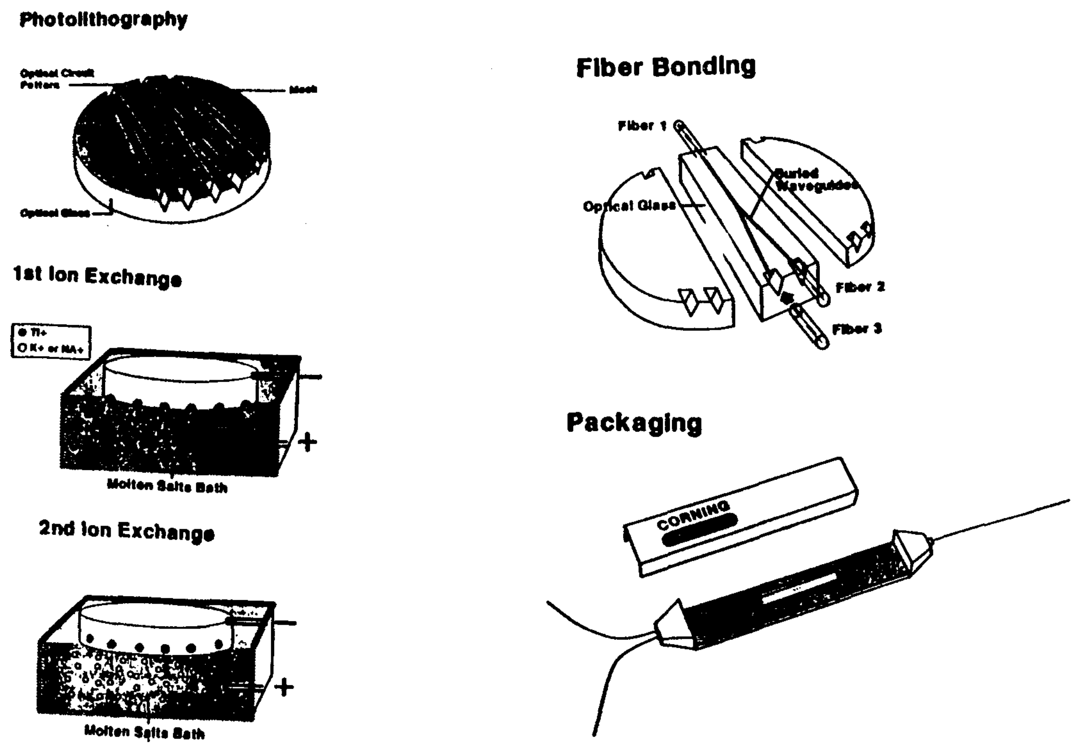


Figure 4. Coupler fabrication by Photolithography

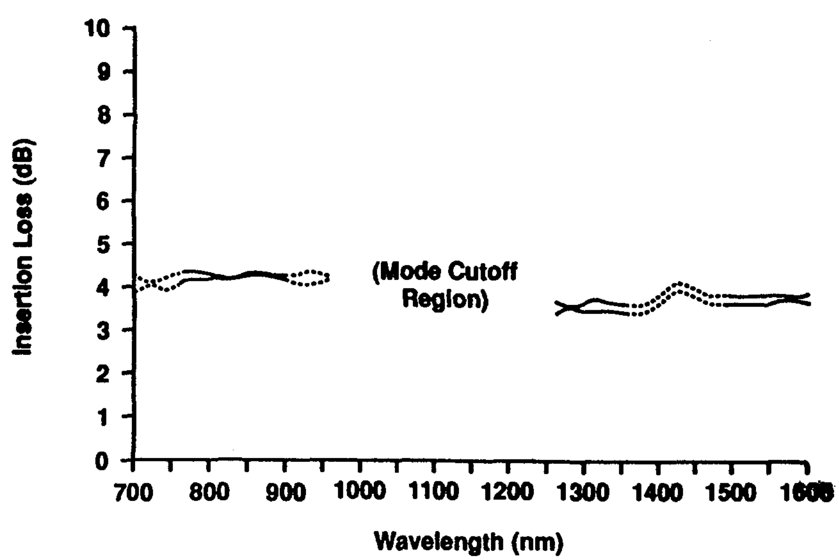


Figure 5. Insertion loss over wavelength for an achromatic coupler

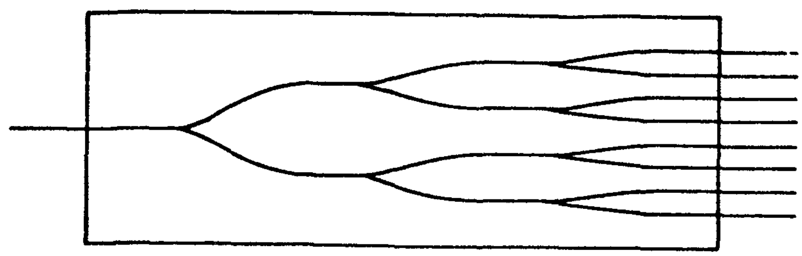


Figure 6. 1 x 8 Coupler made with passive integration of Y-Junction devices

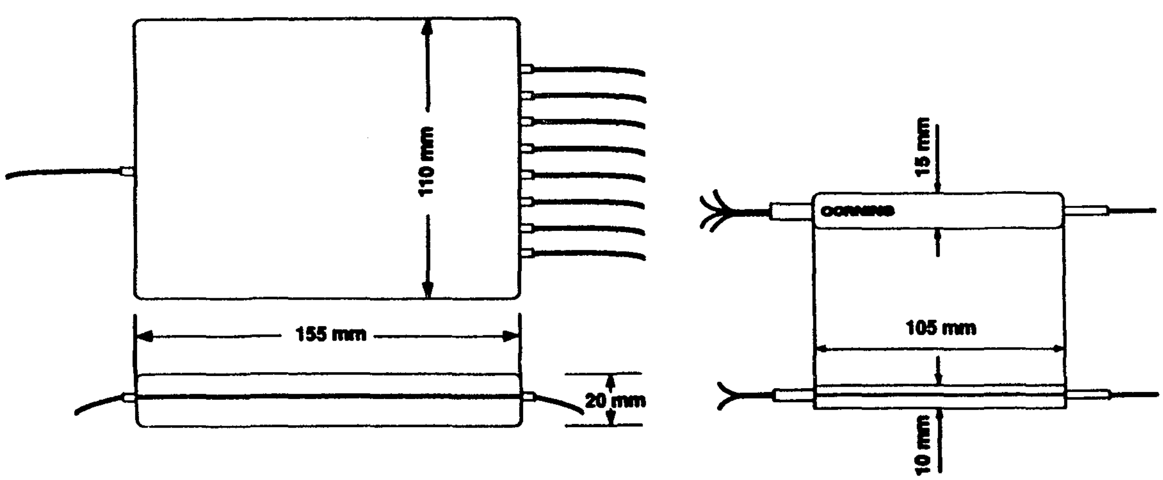


Figure 7. Comparison of size of FBT Cascaded 1 x 8 coupler and planar passive integrated Y-Junction coupler.

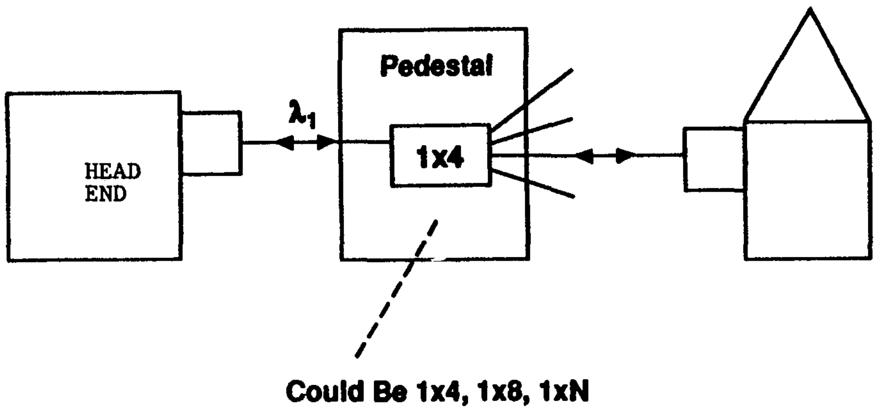


Figure 8. Pedestal network architecture example

Table 2. FTTS Architectures and Potential Coupler Uses

Architecture	Tap	1x2	WDM	1xN	2xN	NxM
Active Star		X	X			
Passive Star		X	X	X	X	X
Cable TV Backbone	X	X	X	X		

Table 3. FTTS System Design Constraints and Coupler Requirements

Implementations Requirements	System Constraints	Coupler
Uncooled lasers or LEDs, with wide source wavelengths.	Power budget over broad wavelengths.	Low uniformity and insertion loss over passband(s).
Low-launch power sources.		
Low-responsivity detectors.		
Low-dynamic-range detectors.		
Short wavelength sources.	Multiple window operation.	Achromaticity in all windows, low far-end crosstalk over passband.
Broadband avenue/upgrade.		
Optical service partition.		
Bi-directional transmission.	Reflections.	High directivity.
AM laser transmitter		Low backscatter.
Environment.	-40 to +85°C.	Mechanicals.
	Saturation humidity.	Packaging.
	Salt spray, other chemicals.	
	Real-world environment.	
Millions of subscribers.	High density, limited footprint.	Monolithic, compact structures, passive integration.
Millions of subscribers.	High volume.	Industrialization.
	Repetitive quality.	High reliability.
	Lifetime cost.	
	Installed cost.	

Michael Labiche
Physicist
Scientific-Atlanta

Mark Newhouse, Ph.D
Sr. Scientist
Corning, Incorporated

FIBER-OPTIC SWITCH FOR CATV SYSTEMS

ABSTRACT

We describe a single-mode fiber-optic switch for use in an AM CATV transmission system. This paper will discuss several applications, including the incorporation of a standby laser to fulfill the need for redundancy. This switch allows for system testing, such as OTDR traces and optical power testing, without breaking splices or losing fiber transmission. In all of its applications it introduces no measurable system degradation. A detailed description with diagrams as well as performance characteristics are given.

The Corning fiber-optic switch is well suited for several CATV applications. Some examples are transmitter/receiver back-up, system testing, and networking. Three performance parameters become critical when dealing with optical switching: far-end crosstalk, insertion loss, and backreflection. An understanding of these parameters and the switch function is necessary before the switch's performance can be fully appreciated.

A switch is a four port two state device with the capability to connect either input port to either output port.(FIG.1) In the bar state, port 1 feeds port 2 and port 4 feeds port 3. In the cross state, port 1 feeds port 3 and port 4 feeds port 2. It is the transition between bar and cross states that constitutes switching in this device.

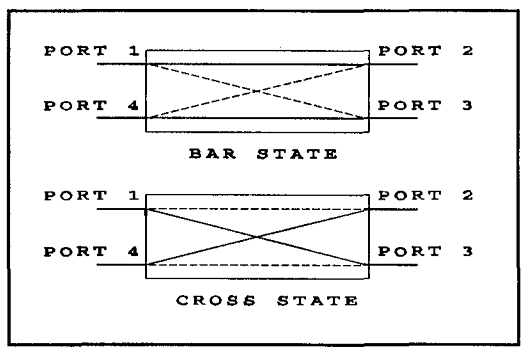


Figure 1 Optical Switch States

Far-end crosstalk is a measure of the amount of light reaching the undesired output port. Specifically, if the switch is in the bar state, and light is input into port 1, then the far-end crosstalk is given by 10 log(P3/P2), where Pi is the optical power at port i. For the cross state the ratio is simply inverted. This "bleed over" of power can cause signal interference. In many AM applications crosstalk values of less than -20dB are required to sufficiently suppress this interference.

Insertion loss is a measure of the amount of power lost between the input and the desired output port of the device. With the switch in the bar state and light launched into port 1, the insertion loss is given by 10 log(P2/P1). High insertion losses can reduce the signal to noise ratio in an AM optical transmission system. Insertion losses of 0.5 dB or less are acceptable for these systems.

Back reflection or return loss is the ratio of optical power reflected back out of the input port to the power input. Directivity or near-end crosstalk similarly characterizes the power leaving the other input port. For example, directivity is given by 10 log(P4/P1), if power is launched into port 1. These reflected signals in an AM fiber-optic system can contribute to noise in the transmitter. It is best to keep both back reflection and directivity below -40 dB to preserve signal quality.

The Corning fiber-optic switch performs well in all three of these critical areas. It typically exhibits far-end crosstalk of -20dB, insertion loss of 0.3dB, and backreflection of less than -55db. This allows the introduction of the switch into AM fiber optic CATV systems with minimal degradation of signal quality.

Switching is achieved in the Corning optical switch through the perturbation of a 2x2 fiber-optic coupler(i.e. a coupler with two input fibers and two output fibers). Fiber optic couplers are an increasingly common element in fiber-optic networks. Typically they are used to split the optical signal among multiple paths. Many of these couplers rely on the coupling achieved between two fibers brought into close proximity, usually by tapering and sometimes by etching. The coupling between the two proximate fibers can be stopped by bending the coupling region so that the signal continues to propagate solely in the input fiber.

One of the technologies that can be used for coupler fabrication utilizes a three-index tapered glass structure, made up of two fibers inserted into a tube (of a third refractive index) which is then necked down to the particular radius required to achieve a particular degree of coupling. For the switch application, a coupler is made such that 100 percent of the light is coupled from the input fiber to the second fiber. With bending, the percent of light coupled to the second fiber can be varied from 100 percent to 0 percent. The bend radius required to reach 0 percent coupling is on the order of 20 centimeters. This bend can be obtained with roughly a 1mm displacement of one end of the approximately 50mm long coupler. Glass, being an elastic material, does not fatigue due to repeated bending, which has been confirmed by switching many devices each over a million cycles without breakage. The three-index coupler technology provides a stable, ruggedized structure in which the coupling can be easily and reproducibly controlled with bending. Therefore, switching can be achieved in a device in which the light never leaves a waveguiding structure (in contrast to moving fiber and moving prism optical switches), thereby yielding a robust switch with low optical losses, low back reflection and high reliability.

A useful application for the switch is in system back-up or redundancy. A "hot" stand-by back-up transmitter can be switched on line if a failure occurs in the primary system.(FIG.2) The switch can be activated manually or by remote logic from a status monitoring system. A system such as this was displayed by Scientific Atlanta at the 1989 Western Cable Show with great success. In this system, the isolation of the switch prevented the back-up and primary transmitters from interfering with each other, eventhough both devices were fully active. This aplication can be expanded to accomodate multi-laser systems.(FIG.3) Receivers also can be switched on line as back-ups in the event of a failure.(FIG. 4) If the ideas in figures 2 and 4 are combined by stacking switches in series, either transmitter can access either receiver. This also allows the use of a back-up fiber which could be accessed if a break were to occur in the primary path.(FIG. 5)

Another application allows system testing without ever losing fiber transmission.

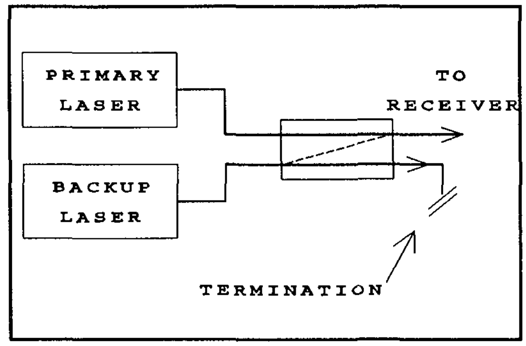


Figure 2 Transmitter Backup

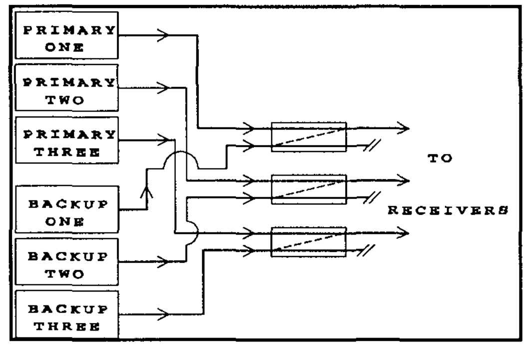


Figure 3 Multi-Transmitter Backup

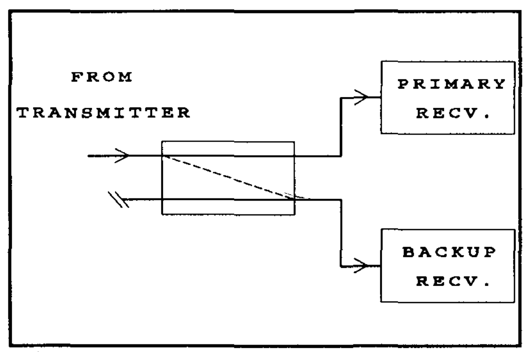


Figure 4 Receiver Backup

(FIG. 6 & 7) Suppose it is desired that an OTDR trace be generated as well as having the laser power level determined. In order to do this, switches must be stacked so that a back-up system can be accessed. The first switch(S1) sends an output to the primary receiver and a power meter. The second switch(S2) sends an output to

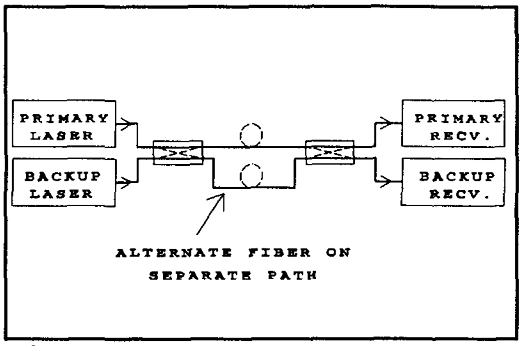


Figure 5 Backup Path

S1 and the back-up receiver. (FIG. 6) When both switches are activated (FIG. 7), the back-up laser will feed the back-up receiver to give fiber transmission. An OTDR trace can be run through S2 to S1 and then to the primary receiver. The primary laser is switched to a power meter to get a power measurement. This application will prevent any downtime during system testing.

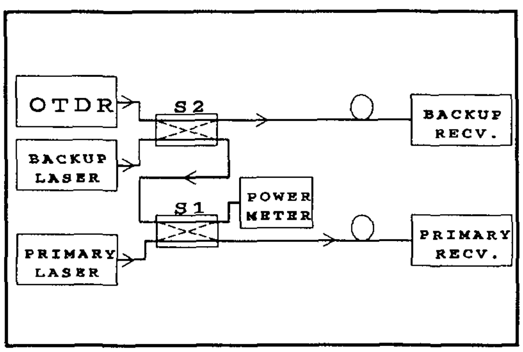


Figure 6 Before System Test

Another application for the switch is networking. By cascading several switches, the number of access points increases exponentially. This could be useful for status monitoring datalinks or a reverse video link for measuring signal quality directly. With various combinations of switch states, every hub could be accessed.

Switch isolation is sufficient to prevent the effects of noise funneling and interference caused by having many sources feeding one receiver. (FIG. 8)

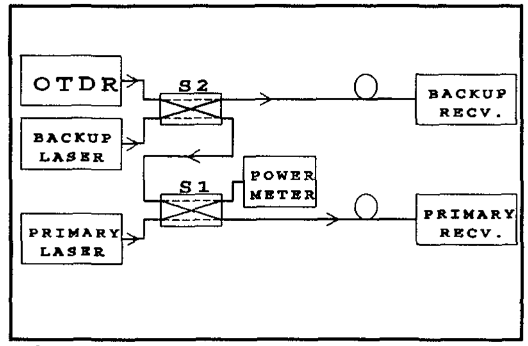


Figure 7 During System Test

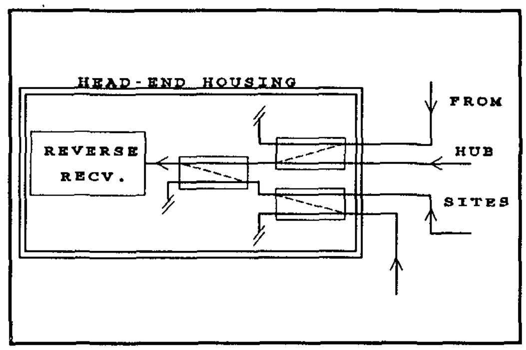


Figure 8 Reverse Receiver

The switch's versatlility lends itself to many applications in fiber-optic AM CATV systems. The performance parameters allow it to be introduced into these systems with minimal degradation of signal quality. With the growing interest in system redundancy and status monitoring, the Corning fiber-optic switch promises to be a significant development in AM CATV.

FLYOVER CALIBRATION PROCEDURES

Robert V.C. Dickinson
Dovetail Systems Corporation

In qualification of a cable system according to the FCC Leakage rules the flyover is often preferred since it is a direct measurement of the leakage signal strength in the airspace above the cable system. As in any other measurement it is necessary to establish standards against which to compare the actual data taken. In the case of a flyover one must establish the threshold level of 10 microvolts per meter (uV/m) at 450 meters above the average terrain in order that the pass/fail requirements of the regulations may be tested. A rather specific method for producing this reference field is given in part 76.611(a)(2) of the FCC Rules and Regulations. Even this well defined procedure is subjected to certain inherent inaccuracies and some aspects that may be improved upon. This paper deals with these subjects and suggests some possible modifications. Before proceeding with the discussion it must be stated that anything said below which is not consistent with the current Rules and Regulations has NOT BEEN APPROVED by the FCC and therefore cannot be assumed to be acceptable for flyover calibration.

CALIBRATION BASICS

The fundamental concept of flyover calibration must first be understood. In the survey aircraft we have a total measurement system which involves a receiver and antenna in addition to the data gathering, storage, and analysis equipment. The receiver and the other components in the path of the received signal may be conveniently and accurately calibrated in the laboratory. Not so the antenna. Since the antenna pattern is affected by its mounting and its environment (the aircraft) it becomes a very complex system which is extremely difficult to quantify. Measurement of the pattern of an antenna on an aircraft is done only in extremely large, elaborate, and expensive antenna ranges. These procedures are not appropriate or required by cable TV signal leakage measurement demands. Even if detailed calibration was achieved this information would not be of great value since the signal leakage data taken is the composite of the signals received from numerous leaks in the system.. The intent of the leakage measurement rules is to evaluate the threat of interference to aircraft flying through the airspace. These aircraft will also receive simultaneous signals from multiple sources and will doubtlessly have a somewhat different antenna configuration than the measurement aircraft. The program then is to find the measurement system response to a standardized field from a single source and use this as a basis for surveying in the airspace. The results from such a survey will be adequate to assure protection of the aircraft using this airspace and to check the limits established for cable signal leakage.

The Commission has required that sensing of cable signal leakage in the airspace be done with a horizontally polarized antenna. There are points of discussion as to whether horizontal polarization is a necessity or even the best choice, but it is certainly a reasonable choice and it is the law. It is well to note that standard aircraft antenna are seldom purely horizontally polarized even though they exhibit a good deal of structure which is horizontal. They are often, in the frequencies of use, largely vertically polarized with horizontal sections employed as end loading for the vertical element.

Establishment of the 10 uV/m field at 450m for calibration of the horizontally polarized aircraft antenna, is achieved by use of a pair of horizontally polarized dipoles on the ground. The configuration specified in 76.611(a)(2) (the paragraph which covers all aspects peculiar to flyover measurements) consists of two resonant half wave dipoles mounted at right angles to each other, parallel to and one quarter wavelength above a ground screen of at least two meters in diameter. These two dipoles are fed with radio frequency energy equal in amplitude but differing in phase by 90 degrees. For those unfamiliar with antenna theory, in this configuration the resulting electric field vector rotates about the vertical axis producing circular polarization. This will be either righthand or lefthand polarization depending upon which current is leading. For this case the polarization sense is unimportant. Ideally the rotating vector maintains the same amplitude at all angles in the horizontal plane. Practically speaking, this seldom happens since any current inequality or phase error between the dipoles will cause the polarization no longer to be circular but elliptical. Hence the antenna pattern has a property called "ellipticity" or "axial ratio". We will examine the importance of ellipticity further.

Considering a practical flyover calibration maneuver an attempt is made to fly directly over the calibration antenna in order to pass through the maximum field which is preset to a 10 uV/m intensity at 450m above the antenna. In the general case some amount of cross wind can be expected which means that the aircraft, although flying directly over the antenna, will have its longitudinal axis and hence the axis of the sense antenna at an angle to the line of flight. If the antenna on the ground were a single dipole it then would be necessary to ensure a flight path with no cross wind component and hence no crab angle in order that the two antenna elements be parallel so that the maximum field is received. On a practical level this is extremely difficult to achieve. It may be seen, however, that with the circularly polarized field not only can a crab angle be tolerated but the approach can be made from any angle and still achieve the desired results as long as the pass is made directly over the antenna. This is due to the fact that the circularly polarized field appears to be linear and parallel to the receiving dipole at any angle of approach. Hence, the crossed dipole circular polarization scheme required by the Commission is an excellent choice. Practically speaking, such an arrangement can be constructed to produce circular polarization with less than 1 dB ellipticity.

Another aspect of the calibration run is the precision with which the pass is made over the calibration antenna. Even though it sounds simple a high degree of pilotage proficiency is required to fly directly over the antenna at a 450m altitude. This is due to many factors including the ability to visually judge lateral offset from the aircraft. Even with electronic navigation aids errors of a few hundred feet are not uncommon.

Considering the calibration antenna configuration and its sensitivity to misalignment, it should be noted that the 1 dB beamwidth of a simple dipole (which is the equivalent of the circularly polarized antenna as approached from any angle) is in the vicinity of 90 degrees, that is 45 degrees either side of center. At an altitude of 450m altitude the 45 degree angle would allow misalignment up to 450m with only 1 dB of reduction in the calibrating field strength. Assuming a similar drop off in the aircraft receiving antenna pattern this misalignment could result in a - 2 dB total error. The one-half dB beamwidth of a dipole antenna is greater than 80 degrees therefore misalignment of 1200' laterally would result in no more than 1 dB of total error. While other factors do affect the situation this example is given to illustrate that reasonable misalignments will not materially distort the calibration results.

There are other effects which bear on the calibration procedure. The presence of other signals within the bandpass of the measurement receiver may cause erroneous results. A typical situation occurs when attempting to calibrate on a frequency which is also used by a nearby cable system. It is generally true that the calibration generator and the signal from the other cable system will not be exactly the same in frequency, in which case the airborne receiver will see the power addition of the calibration signal and the spurious signal. The presence of a spurious signal 10 dB below the calibration signal will result in an indication in the aircraft receiver which is approximately 1 dB too high. This receiver will then be calibrated to the wrong level. This receiver sensitivity miscalibration will interpret the signal leakage measured in the flyover to be 1 dB lower than actual. The interfering signal might also be noise from the power local system or spurious signals from a host of electronic emitters including large signals at great distances. As a rule any interfering signal within the passband of the receiver should be no greater than -20 dB relative to the calibration signal level.

The same power addition effect exists when there is a cochannel signal during a measurement flight, but with the opposite result. In the case of masking noise such as power line interfence of level equal to the leakage signal, the sum of the two noncoherent signals is 3 dB higher than either therefore the leakage indicated would be 3 dB higher than actual. This is equivalent to 3 points in a ground based CLI calculation and could well fail a passing system if the effect existed over large portions of the area surveyed. The moral to this story is that when overflying systems with substantial spurious cochannel signals and no ability to select a better frequency, constant monitoring to identify the leakage signal must be done to verify that actual signal leakage is present. It might also be well to consider post-flight calcuations to eliminate the weighting effects of the noise. This is not always possible since quantification of the noise signal level in the presence of the cable signal leakage may not be within the capability of the measuring equipment or procedures.

CALIBRATING THE CALIBRATION SETUP

Generally speaking compliance with the specific details of the calibration rules in paragraph 76.611(a)(2) will satisfy the FCC. However, in an effort to have a high degree of confidence in the validity of the test procedure and its results, one must devise some method of verifying the actual field radiated by the calibration system. Indeed this is virtually mandatory since the price for failure to qualify due to flawed data is so high. Basically the radiation can be quantized by using a probe antenna in the field of the calibration antenna and measuring the signal level with a well calibrated receiver. This same setup may be used to check ellipticity as well. There are, however, some sticky problems in doing such measurements. For instance, feed lines to the probe antenna must be routed in such a way as to not affect the pattern of the probe antenna or the calibration system. The probe antenna may be rotated on axis to measure the ellipticity but the same cautions apply. In addition neither the test equipment nor the technician should be close enough to either antenna to distort the radiation pattern(s).

Measurement of the ellipticity of the calibration antenna pattern has been mentioned and the question may arise as to why one would expect significant errors in such a simple system. Briefly stated there are numerous reasons including physical and electrical parameters such as stray capacitance, ground plane irregularities, and probably most importantly, imperfections in the power division and 90 degree phase shifting networks in the crossed dipole antenna. These networks can be rather simple but must be quite extact to maintain the tolerance that is necessary. For instance a 1 dB difference in drive levels can cause a 1 dB difference in ellipticity resulting in an uncertainty of 1 dB in the actual calibration level depending upon angle of approach, crab angle, etc.

In situations where measurements must done at varying frequencies the calibration antenna must be capable of standardizion at each frequency used. It is desirable to have a single unchanging physical and electrical configuration which can be excited with any required frequency. This, however, is very difficult to achieve and presents a challenge to the design engineer.

Another important consideration is the actual calibration site. The qualifications for this site include a flat open area without structures which can affect the pattern of the calibration antenna. For instance, if the calibration antenna where set up near a large reflecting structure, reflections from that structure could affect the energy arriving at the aircraft at altitude while not altering the calibration antenna pattern sensed locally by the probe antenna. This is a difficult situation since even though the calibration antenna itself checked out well the operator would be unaware of the change occuring in the airspace. Any resonant structures or large conducting objects in the field of the antenna can induce such perturbations. Reraditation by a tower, guy cables, or a metal or steel reinforced building near the site could seriously distort the calibration pattern. It is therefore important that a clean, flat, open area be selected for the calibration. All calibrations should be done at the same properly selected site(s) resulting in better and more stable results. Although this suggestion is contrary to the FCC requirement that calibration be done in the area to be flown, it can well be a step toward significant improvement in calibration accuracy.

The calibration antenna ground system is also a matter of some concern. The FCC requires a ground plane of at least two meters in diameter beneath the calibration dipoles. This does a lot towards stabilizing the antenna impedance and the radiation pattern. As a matter of fact, this insures that the field directly above the antenna is fairly well defined since, by tracing rays from the antenna, it can be seen that all of the power that goes vertically toward the ground is reflected upward by the ground screen which is highly conductive. On the other hand the two meter ground screen does not intercept all the near field currents of the antenna. Those which are not intercepted by the ground screen must return through the local earth ground whose conductivity can vary with location. This is a second order effect in terms of ray reflections and does not significantly alter the field directly overhead but does impact upon the exact impedance of the antenna system. Use of the same calibration location will at least stabilize this effect.

Impedance match to the calibration antenna is also an important concern. This involves the return loss of the antenna system. Return loss, as we are aware, is a measure of the amount of power which is reflected back from a device and in this case is not used in the process of radiation. If one is expecting all of the power introduced into the calibrating antenna system to be radiated the resulting field will be reduced when some energy is reflected and cause incorrect calibrations. A return loss of 10 dB results in 1 dB less power delivered. This is a good reason to measure the actual radiation from the antenna system rather than simply calculate the theoretical value. In the same vein, it is well to check the match of the final system for each calibration run to make sure that nothing has changed.

Questions are often asked about the power required to produced a 10 uV/m field at 450m. The method of calculation of that power runs along these lines. The field strength in free space is related to the power density by the following formula:

$E = \text{sq.rt.}(30 \times P_t \times G_t)/A$

where

E = field intensity - uV/m

P_t = power transmitted - watts

G_t = gain of transmitting antenna

A = altitude - meters

Solving this equation for P_t produces the power required to produce the desired field intensity (E) at the desired altitude (A) which is 10uV/m at 450 meters. The calibration antenna using orthogonal dipoles can be thought of as two independent systems. Therefore the computation may be made on the basis of a single dipole with equal power required by the second dipole. It is then necessary to compute the losses in the power dividing and phasing networks, the cables and any other elements introduced into the system. Remember that the power level must be correlated to a secondary standard which can usually to supplied by an organization that does test equipment calibration. The standardized signal generator can also be used to calibrate receiving equipment, signal level meters, spectrum analyzers, and other equipment used in the process.

After we have done a careful job of setting up the calibration system we must consider the accuracy of the calibration achieved by use of the generated field. A careful analysis would have to include the matters such as precision of power generation, loss measurement, antenna gain, etc. not to mention the uncertainties of the aircraft antenna and receiving system plus the aircraft attitude during the calibration and measurement process. It seems likely that the uncertainty of the calibration field could well be plus or minus 2 dB from the desired level even with good engineering practice and careful control. It is doubtful whether a certainty better than plus or minus 1 dB can be claimed with anything but the most elaborate instrumentation and setup. Although not a point for detailed discussion here, it is abundantly clear that the field established by these methods in the airspace is far better controlled than the field used to calibrate ground based CLI measurement equipment. The typical case is the antenna near which you drive your truck to set your 20 uV/m threshold. Here it is clear that there are so many nearby uncontrolled reflecting objects that to expect precision calibration is not realistic.

CONCLUSIONS

It can reasonably be said that aerial calibration and subsequent measurement is plainly the most direct and the most accurate method of surveying cable signal leakage since it is done by direct measurement made in the environment where protection is desired rather than estimation from ground data. Airborne calibration is subject to fewer errors but, all in all, is not a laboratory situation where 0.1 dB precision can be expected nor for that matter is even important. To properly setup for and conduct the necessary calibration requires great care and a system of checks and balances to assure accuracy and repeatability. Good calibrations are necessary to uphold the dedicated efforts of the ground repair teams and assure timely qualification of the cable system.

Handheld Direction Finder for Cable Leakage Location

By Clifford B. Schrock, President
CableBus Systems Corporation

ABSTRACT

A handheld direction finding system has been developed to aid in the rapid location and pinpointing of leakage sources in a cable system. This paper covers the development of a practical, handheld unit, including the antennas, circuitry and display. Also covered are some of the problems and solutions to the measurement of radio signals in the VHF spectrum.

THE PROBLEM

Pinpointing the exact location of a cable leak has been a continuous problem for the cable technician. Vehicular and portable equipment is available that can detect areas where a leak exists, and near field-probes can be held next to the cable to check for leakage from a connection or housing. In between, a bewildering array of dipoles and directional antennas have been used with signal level meters and portable receivers to try to find the general direction of a leak from the ground. Technicians have, in some cases, spent hours trying to locate a single leak.

My involvement in a solution to the problem of locating leaks in cable systems actually started about four years ago with a tragedy in the Northwest. A group of high school students were attempting to climb Mt. Hood and got caught in a storm. Eight of the group died before they were found three days later in a snow cave. At that time, I started working on an idea I had for a handheld direction finder and coded radio tags that could be worn by climbers.

In the summer of 1989, I decided that the direction finding technique could be applied to the cable leakage problem, and the outcome of this research resulted in the development of a handheld direction finder and field strength measuring instrument, called the Leakage Locator System.

ANTENNAS FOR LOCATING

A number of units with various kinds of antennas have been traditionally used by the cable technician to measure and find leaks. These include the simple dipole and various multiple element gain antennas such as the Yagi antenna. They are analyzed below.

The simplest antenna, the dipole, has a very broad symmetrical pattern as shown in Figure 1. The maximum signal strength orientation lobe is very broad, typically over 100 degrees wide between the 3 dB points. It is symmetrical, which makes it difficult to tell whether the leak is behind or in front of the user without taking multiple readings at different positions, and triangulating. The dipole IS good for measuring field strength since it is easy to characterize the antenna factor, and the directionality is so broad that little error is encountered if the antenna is not exactly oriented to peak.

Various gain antennas which use multiple elements have been proposed or sold for improved beam-width and to make the main lobe directional, eliminating the arbitrariness of the simple dipole. The pattern for a four element Yagi is shown in Figure 2. One lobe of the simple dipole is superimposed for reference. The particular Yagi tested showed a significant improvement in directionality and almost 11 dB of gain compared to the dipole, however, the 3 dB points are still greater than 50 degrees. The Yagi can also be quite cumbersome due to its size, especially if cut for the 108 to 136 MHz lower FCC compliance band.

Other antennas were tested, and the general conclusion reached was that they could barely provide enough directionality to find "the broad side of a barn".

DIRECTION FINDERS

Direction finding techniques have been used since the early 20's, and involve the use of the null that can be obtained from the end of a dipole element or a loop antenna rather than the peak of the main lobe. Early systems were operated by rotating an antenna by hand, and taking multiple readings from different locations to eliminate the front/back arbitrariness.

Modern electronic direction finders with indicators or displays used an antenna array combined with a mechanical or electronic switcher, and a sense antenna to resolve and pinpoint the source of a radio emission.

The significant improvement in resolving power of a direction finder is derived from the null characteristic, which has a rapid rate of change close to the null position,

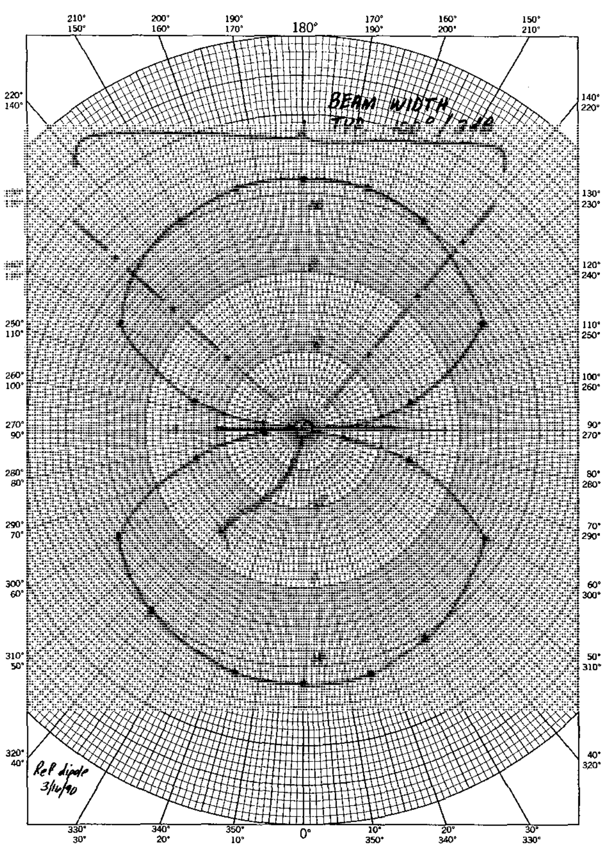


Figure 1 - Antenna pattern for a simple dipole antenna. Typically the 3 dB points of the beam will be over 100 degrees apart.

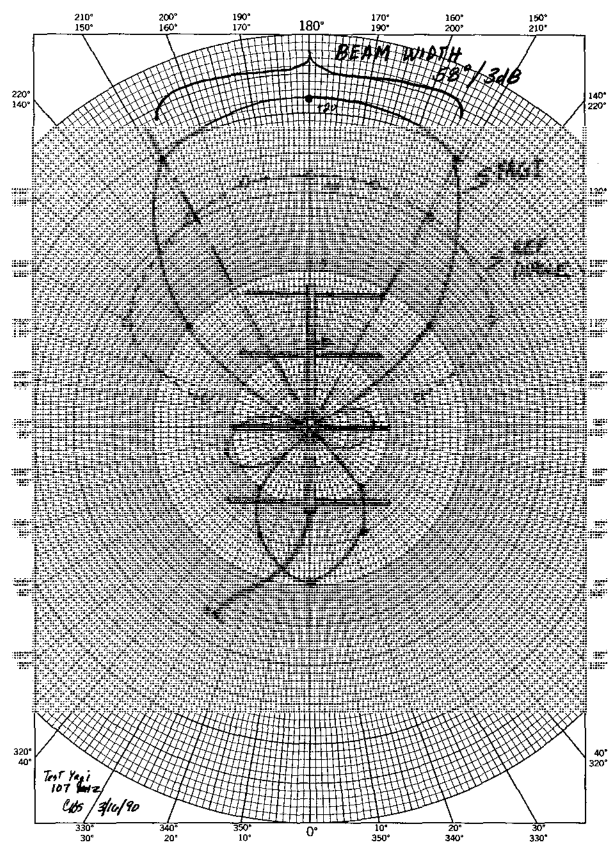


Figure 2 - Antenna pattern for a multiple element Yagi type antenna which shows a 3 dB beam-width of 58 degrees.

as shown in figure 3. The figure shows the pattern of the null and sense elements of the antennas of the actual handheld unit that was developed. Null 3 dB beam-widths of less than 5 degrees are typical, and the unit easily resolves 1 dB changes, giving better than 2 degree of locating resolution.

The handheld direction finder developed by the author, which includes leakage measurement capability, is called the Leakage Locator System. It consists of a receiver, antenna array, direction finding circuitry with a visual peaking display, and a precision field strength meter, allowing a cable technician to quickly pinpoint and measure a leak.

A block diagram of the direction finding circuit is shown in figure 4. The antenna array consists of two dipole antennas, tuned and loaded to frequencies in the 108 to 136 MHz band. One of the dipole elements is the boom, which includes the effects of the human body holding the unit. The photo in figure 5 shows this clearly. The system is horizontally polarized, since this appears to be the predominant orientation of leakage in overhead cable plants.

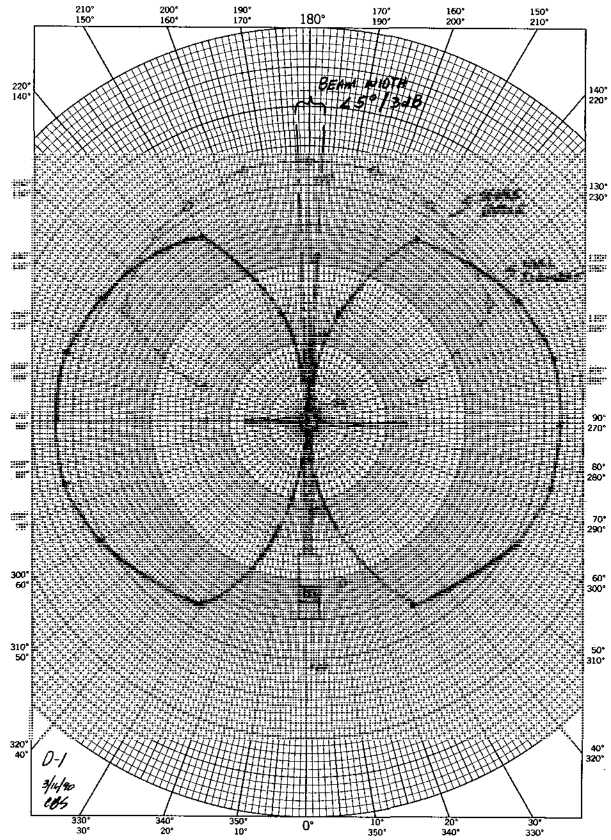


Figure 3 - Null and sense patterns for the Leakage Locator System show a 3 dB beam-width of less than 5 degrees.

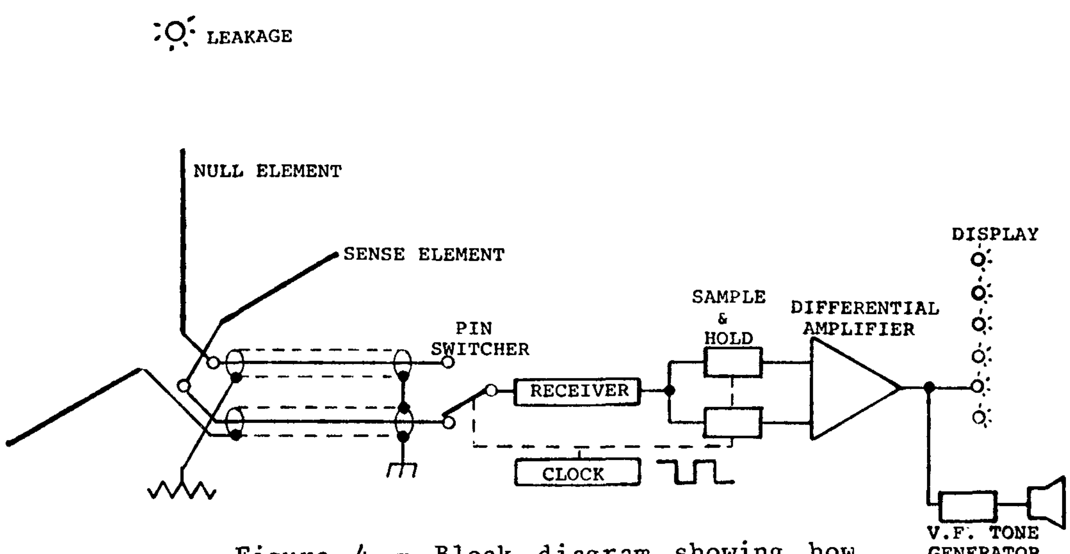


Figure 4 - Block diagram showing how the direction finder portion of the circuit works.

The Leakage Locator operates by measuring the difference between the peak output of the sense dipole, and the null of the second antenna. The unit peak-to-null ratio does not vary with signal strength within the dynamic range of the system. When the array is oriented 90 degrees from a leak, the sense element is at null. This causes the output of the differential amplifier to be negative, causing the display to read zero. When a leak is 180 degrees from the pointed direction, behind the operator, the human body disturbs the null pattern significantly. Therefore, the operator cannot falsely peak the display and will know that he is not pointed at the leak.

The Leakage Locator, as described, is a horizontally polarized device. However, the null antenna actually produces a cone shaped, three dimensional null pattern. This allows the unit to pinpoint in both the vertical and horizontal axis. The actual calibration, as measured,is 2 degrees horizontal and 3 degrees vertical, for each of the last four red dots of the light display.

Figure 5 - Photo of the Leakage Locator shows the position of the human body relative to the unit, which disturbs the rear null pattern.

The block diagram in Figure 4 shows the antennas being toggled by a PIN diode switch into a common receiver. The output is sampled in two hold circuits, then amplified in a differential amplifier. The output of the amplifier feeds a calibrated display and an audio Voltage-to-Frequency converter to aid the user in peaking the display.

The direction finder display and the measurement meter on the unit is shown in figure 6. The actual Leakage Locator System has two modes: the Locate mode which enables the direction finder circuitry, and a Measure mode, which uses the Sense antenna and the receiver system, and is calibrated to read directly in uv/m at 10 feet.

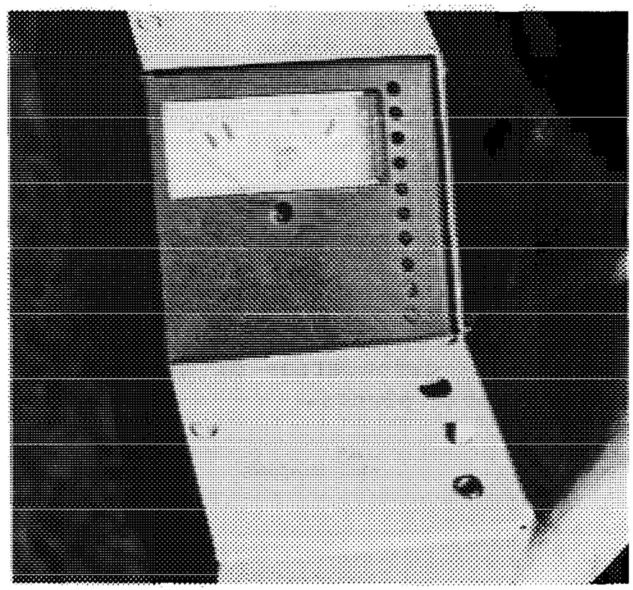


Figure 6 - Photo of the meter face and light display which allow the operator to pinpoint and measure leak sources.

Originally, the meter was going to provide only a simple GO/NO GO type reading with a 20uv/m calibration line in the center of the scale. Subsequently, in field tests, it was found to be most useful if the scale was calibrated to measure the actual field strength, and the 20 to 200 scale and a X10 switch were added to allow the measurements to be made from below 10uv to 2000uv/meter.

Additionally, it was found useful to be able to measure from a distance greater than 10 feet. For example, if the leak were on the side of a residence, or on a rear easement, the technician would not want to enter the property unless there was a problem. The receiver sensitivity of the unit was increased to allow measurement of a 20uv at 10 foot leak, up to 160 feet from the actual leak. The photo in figure 7 shows the calibrated potentiometer that the operator can use to dial the estimated distance to an apparent leak located with the Leakage Locator, and get an idea of the field strength.

The actual design of the Leakage Locator as a product involved the building of a highly sensitive receiver, capable of being calibrated and remaining stable for the Measure mode, and having a high dynamic range for the Locate mode. The sensitivity required to measure and direction find on a leak with the equivalent field strength of a 20 uv leak, 160 feet away, including the antenna correction factor, and further loss for electrically shortening the elements, required a receiver sensitivity of -76 dBmV or - 124 dBm. This dictated the use of a narrowband design to remain comfortably above thermal noise (KTB). A 3 kHz dual conversion receiver was designed, with the theoretical KTB being -144 dBm, allowing adequate margin for front end switching, filter losses, and a high noise figure for the RF preamplifier.

The operating frequency of the Leakage Locator was chosen to utilize the TV carriers already present on the cable. The typical unit can be switch selected for either midband channels B or C, at 127.25 MHz and 133.25 MHz respectively. The offset of +/- 12.5 kHz is also switch selectable. Optional HRC frequencies and offsets can be accommodated by changing the crystals. The 3 kHz narrowband receiver has a specially compensated detector and a 4.2 dB correction circuit to provide signal strength measurements on the TV carriers "equivalent to the RMS value of synchronizing peak" as required in FCC Rules Part 76.609(h).

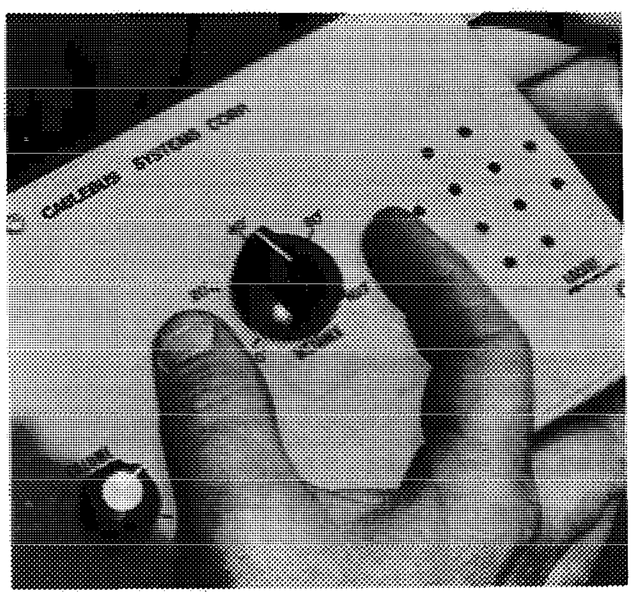


Figure 7 - Photo showing the distance potentiometer that converts the uv/m reading to a 10 foot equivalent.

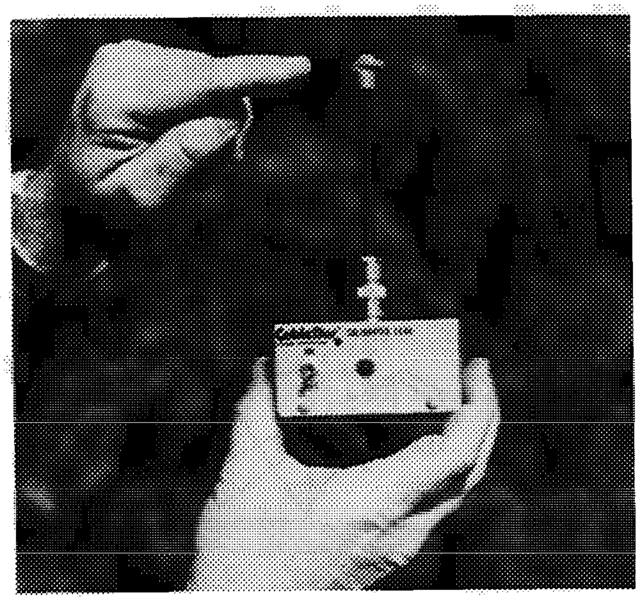


Figure 8 - Photo showing the 20 uv Calibrated Leak supplied with the Leakage Locator System.

During the development of the Handheld Leakage Detector, continuous use was made of a small transmitter that simulated a single point leak of 20uv/meter at 10 feet. The author developed a small point source horizontally polarized antenna to use with the transmitter. This was Calibrated Leak, shown in Figure 8, was found to be a very useful accessory and is supplied with the Leakage Locator System.

FIELD EXPERIENCE

The final part of this paper covers the leakage location field experience gained in both overhead and underground installations, along with a discussion covering some of the myths of leakage location such as multiple leaks causing the detection of "phantom" leak locations. One of the authors main concerns throughout the development of the Leakage Locator System was the question of usefulness. Assuming the unit was capable of pinpointing a single point radio source, would the unit be useful in cable leakage situations?

Field testing bore out the answer. Yes, indeed, the unit has proven to be a very effective tool in locating and measuring leakage. Some interesting situations were encountered, but every situation has an explanation that does not "mysteriously" violate any laws of physics.

One of the first topics to be understood is the ground reflection phenomena, and it's relationship to any field measurements (not limited to those with the handheld direction finder). Figures 9 and 10 illustrate this effect. Within close proximity to the ground, be it wet or dry, sand or grass, the radio wave is reflected off the surface of the ground very efficiently at VHF frequencies.

In the case of measurements being made perpendicular to the ground, such as when the unit is used from the street, aimed at the side of a house, the effect is most pronounced. Figure 9 shows that the direct wave, and the ground wave, follow a path of similar length. Therefore, both waves are of

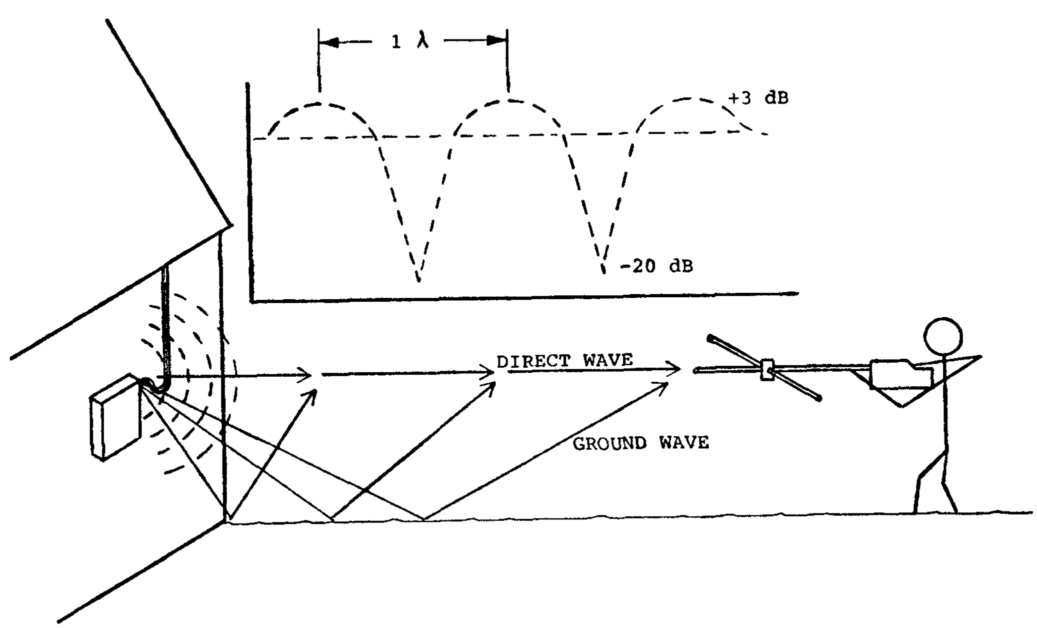


Figure 9 - Illustration showing the ground reflection effect when the measurement is parallel to the ground.

similar strength, particularly as the distance is increased. If they arrive in phase, they can add together, the sum being as much as 3dB greater, and if they are 180 degrees out of phase, they can cancel, sometimes almost perfectly. The error therefore, can be +3dB to - 20db or greater. In practice, moving back and forth a few steps will often cause the meter to vary through the peak and null readings.

The second condition shown in figure 10 is a measurement on an overhead cable. In this case, the ground wave follows a longer path, losing 3dB of strength for each doubling of the distance over the path of the direct wave. If, for example, the strength of the ground wave was attenuated by ldb, the additive and subtractive effect would be much less, in the order of +/- 2dB. Still, it is important to move the antenna enough to verify the peak to valley variation.

Murphy's Law fully applies here. You will always stand in the null to measure and think your system passes, and the FCC will always stand in the peak and fail that location when they measure. Be aware of the problem, particularly when extrapolating distance to a leak from more than 10 foot, and especially when looking parallel to level ground.

Another phenomenon that must be mentioned is the ground proximity effect. As a radio source or test antenna is moved closer to the ground, the signal strength will reduce. Within one wavelength, a drop of 1 dB for each 6 inches is not uncommon. This effect is easily demonstrated if the Calibrated Leak is set on the ground. The measured field will be as low as 1 to 2 uv. Therefore, the calibrator should be used only on top of a 5 or 6 foot wooden post. Alternately,

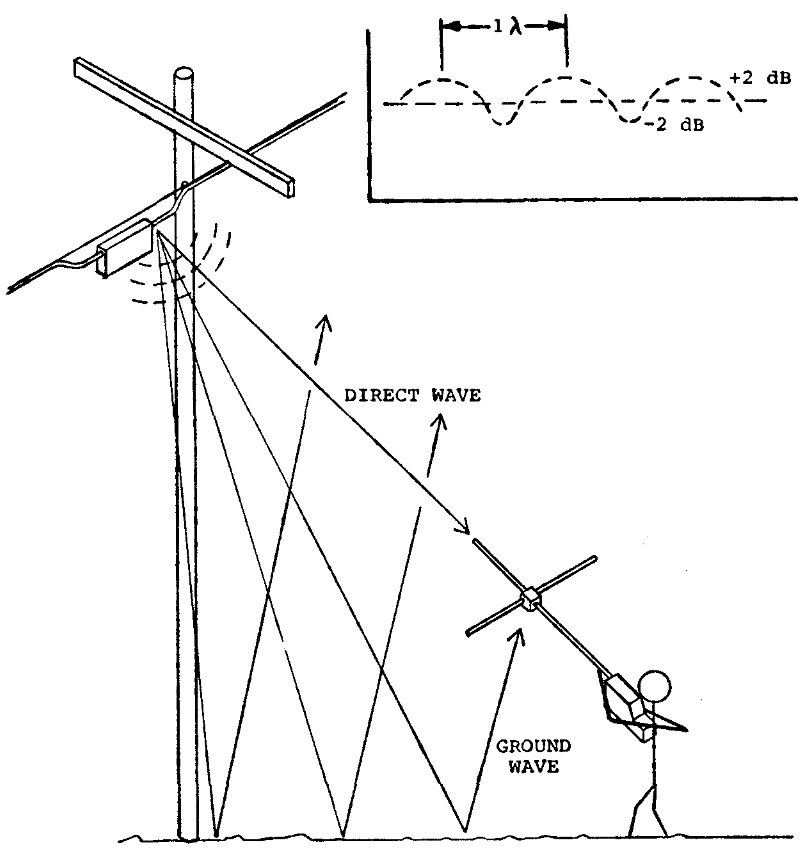


Figure 10 - Illustration showing the ground reflection effect when measuring overhead from the ground.

lowering the test antenna close to the ground, such as squatting to take measurements, will yield low readings.

The final group of field experiences that I will summarize involves common sense issues. Radio waves bounce, bend, and travel in strange but always explainable ways. The Handheld Leakage Locator, because of it's narrow beam-width, will seem to exaggerate problems that would normally be masked by a wide beam-width antenna.

Listed below are some of the more common situations that might be encountered:

- Searching for leaks beside or in front of a vehicle can bend or reflect the leak field. Move at least 15 to 20 feet away from vehicles.

- Guy wires are also a problem. The user should move away from and around the down-guy wires, and take multiple "shots" from a few positions to confirm the probable leak source.

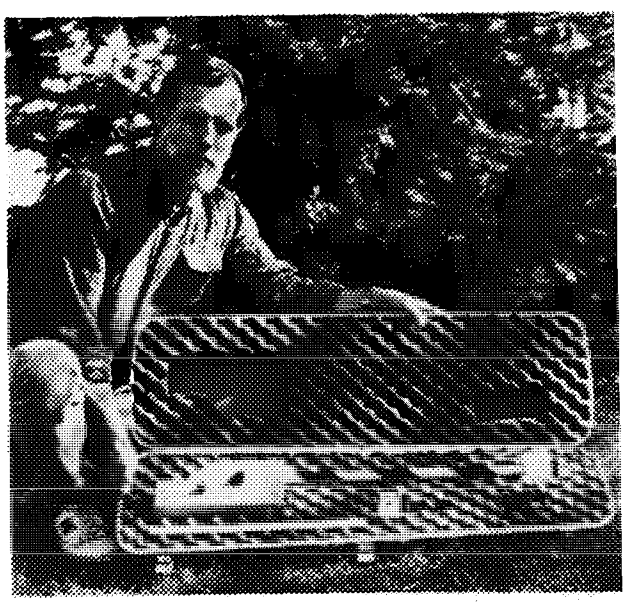


Figure 11 - Photo showing the complete field system which includes the Leakage Locator, Calibrated Leak, AC and Auto chargers, and the carrying case.

- Follow the orientation of the wires, keeping the sense antenna parallel to the wires. If, for example, a power supply is mounted on a pole, rotate the antenna 90 degrees and sweep the vertical section of the coax from the power supply to the power adder.

- The same recommendation applies to "shooting" a house from the street. Rotate the unit 90 degrees and check both the horizontal and vertical fields.

- Multiple leaks, within one wavelength of each other, may appear to give a "phantom" target at some point between the two leaks. A full ring break in a sheath may also have a false peak within a wavelength of the actual break. This is where you apply a near field probe. A wavelength is only seven feet at 130 MHz, so don't panic!

- As you move closer to multiple point source leaks, they will usually begin to resolve into separate leaks that will each peak on the direction finder.

CONCLUSION

The Handheld Leakage Locator has been shown to be a powerful and effective tool to be used in locating leaks in both aerial and underground plants. It is felt that the unit can be part of the cable operator's arsenal of equipment, and will assure a complete and successful Cumulative Leakage compliance program.

A patent application has been prepared for filing on the Handheld Leakage Locator System.

HDTV MUSE Signals on Cables and Optical Fibers

Yozo Utsumi, Hiroo Arata, Mikio Maeda

Science and Technical Research Labs. of
Japan Broadcasting Corporation (NHK)

ABSTRACT

We performed a MUSE (Multiple Sub-Nyquist Sampling Encoding) transmission experiment through two transmission systems in cascade, comprising a communications satellite and coaxial cable facilities, with the cooperation of NCTA and HBO in April 1989. At that time, the MUSE-FM signal was uplinked at the HBO Communications Center in Long Island, New York, to SATCOM K1 and downlinked to two cable headends in the Washington suburbs. At the cable headends, this signal was transformed into the MUSE-VSB•AM signal, and transmitted to remote subscriber locations. Typical results of this experiment were as follows: the unweighted SN ratio is greater than 39 dB and the picture quality is better than 4 on a 5-grade subjective evaluation.

We also developed a demand access optical fiber CATV system for HDTV MUSE signals. On a trunk line, 34 MUSE-FM signals can be transmitted over about 20 km, and at the hub, any 4 MUSE-FM signals can be selected by channel request signals from an individual subscriber, and are transmitted on each subscriber line of about 2 km. The CN ratio through this total system is greater than 17.5 dB, which is the perceptible noise limit of MUSE-FM transmission, whose bandwidth is 27 MHz.

INTRODUCTION

The 1125 scan line/60 field rate HDTV system has demonstrated its capability of producing high quality pictures and sound. In fact, television programs have already been produced in many countries utilizing 1125/60 cameras and equipments. NHK has been broadcasting HDTV experimental programs for an hour per day via broadcasting satellite BS-2 since last June, and plans to start full-fledged HDTV satellite broadcasting via BS-3 in 1990 by using the MUSE system.

In addition, the MUSE system has been successfully tested over communications satellites under the auspices of INTELSAT and other organizations, for cable television transmission, and for optical fiber transmission. In Japan, a space-cable-net. project demonstrated MUSE-FM and MUSE-VSB • AM transmission on cable facilities via communications satellite in October 1989. These experiments have proved the system's feasibility and versatility for a variety of distribution modes. Thus, HDTV has extensive applications to video media.

Presently, there are growing expectations in the United States as to the potential of HDTV services via cable networks and cable television systems. In May of 1988, NHK performed an initial demonstration of HDTV over a cable system in Los Angeles during the NCTA Convention. Additional testing was performed in January of 1989 over two Washington D.C. area cable systems: a 120-channel, state-of-the-art system owned by Media General, and a smaller, more conventional system owned by Jones Intercable. These tests demonstrated that an HDTV picture can be provided today through a typical coaxial cable system without significant impairment due to cable propagation characteristics.

In order to further evaluate the feasibility of HDTV services on cable, the MUSE-VSB•AM transmission including the satellite link in the cable distribution network, was tested in April 1989.[1] The first half of this paper describes the outline and experimental results of this transmission test.

In addition, remarkable progress has been achieved in FM-FDM transmission of video signals for optical fiber CATV systems.[2],[3]

We recently developed an HDTV optical fiber CATV system employing a demand access technique. It can be used by low cost systems, because it does not need expensive wideband optical receivers at the subscribers. Any 4 MUSE-FM signals can be selected among 34 MUSE-FM signals transmitted on trunk lines, at the hubs without demodulating. A compact and inexpensive hub system can be realized by using a heterodyne technique. We adopted commercial low cost LDs developed for compact disk players. Moreover, we utilized conventional BS receivers for Japanese satellite broadcasting at the receiving ends. The second half of this paper describes the outline and experimental results of this demand access optical fiber CATV system for HDTV.

HDTV CABLE TRANSMISSION EXPERIMENT VIA SATELLITE

MUSE-VSB•AM Transmission on Coaxial Cable

A MUSE transmission system compresses the baseband bandwidth of HDTV signals, which have five times as much information as current TV signals, to just double, without deteriorating picture quality. This means an HDTV picture can be transmitted through a narrow bandwidth, and therefore MUSE can be effectively applied to CATV systems.

The transmission spectrum for MUSE-VSB•AM is shown in Figure 1, and a VSB filter is used on the transmitting side and a Nyquist filter on the receiving side. It has 12MHz bandwidth, and can be transmitted in a channel adjacent to current TV channels as shown in Figure 2.

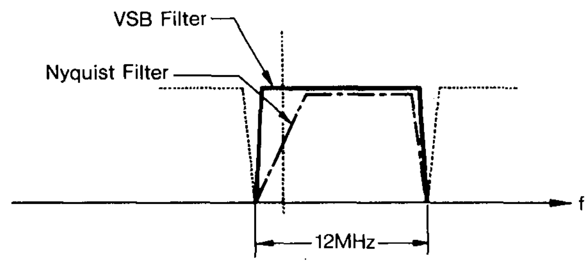


Figure 1 MUSE-VSB•AM Transmission Spectrum

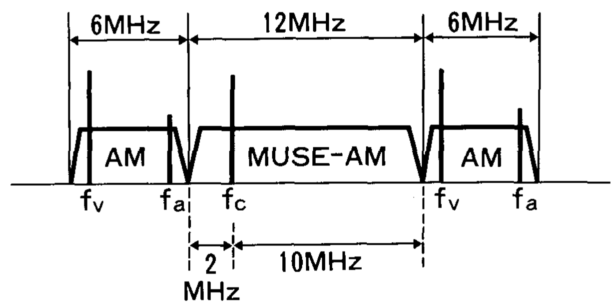


Figure 2 Example of MUSE-VSB•AM Adjacent Transmission

The required CN and CTB (Composite Triple Beat) ratios are the main factors in limiting the number of cascaded amplifiers for coaxial cable systems. The relationship between the number of amplifiers and the required CN (4MHz bandwidth) and CTB ratios for VSB•AM transmission of MUSE and NTSC signals are shown in Figures 3 and 4.

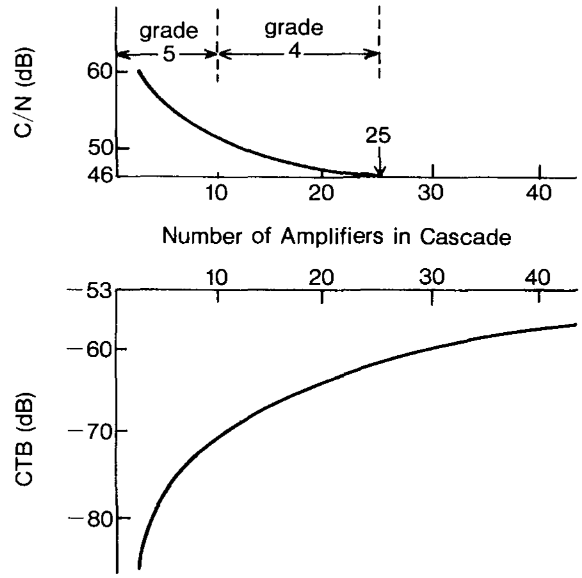


Figure 3 CN and CTB Ratios in a Cascade (MUSE-VSB•AM)

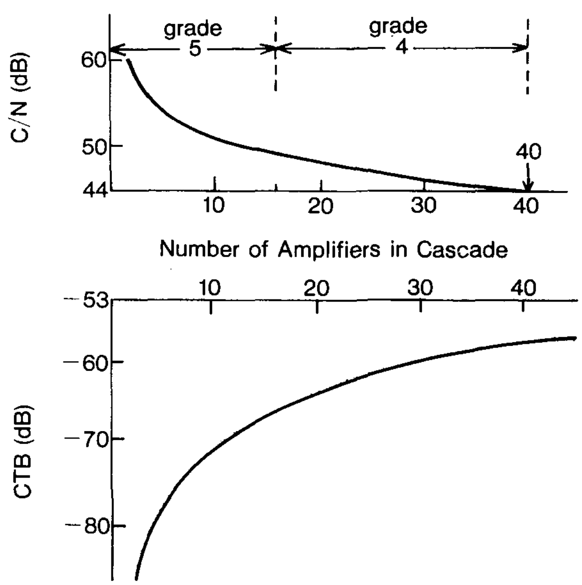


Figure 4 CN and CTB Ratios in a Cascade (NTSC-VSB•AM)

Also shown in the figures are the regions where the CN ratio gives a picture quality of grades 4 and 5 using a five-point comment scale. For the calculation, the performance of trunk amplifiers is assumed as follows: CN ratio = 60 dB, CTB ratio = -90 dB. It can be seen from the figures that a picture quality of grade 4 requires a CN ratio of about 46dB for MUSE, and about 44 dB for NTSC; thus, HDTV transmission requires a higher CN ratio.

Experimental Setup

The setup of this transmission experiment is illustrated in Figure 5. In this experiment, a MUSE signal is transmitted through two transmission systems in cascade, comprising a communications satellite and cable facilities.

At the satellite transmitting end of the HBO Communications Center in Long Island, New York, an HDTV studio signal is encoded into the MUSE signal, which is transmitted by frequency modulation to the communications satellite, SATCOM K1.

At the cable headends, the MUSE-FM signal received from the satellite is transformed into the MUSE-VSB • AM signal. This signal is multiplexed with many channels of conventional broadcast waves transmitted via the other satellites and ground microwaves, and these multiplexed signals are transmitted to the cable receiving ends through multistage amplifiers and coaxial cables. The cable systems involved in this transmission experiment were Media General which owns state-of-the-art system in Fairfax, Virginia, and Jones Intercable, which owns more conventional system in Anne Arundel, Maryland.

At the cable receiving ends, the received MUSE signal is decoded and displayed on an HDTV monitor. The MUSE signal is also received by a conventional TV set, with a simple and low cost MUSE-NTSC converter.

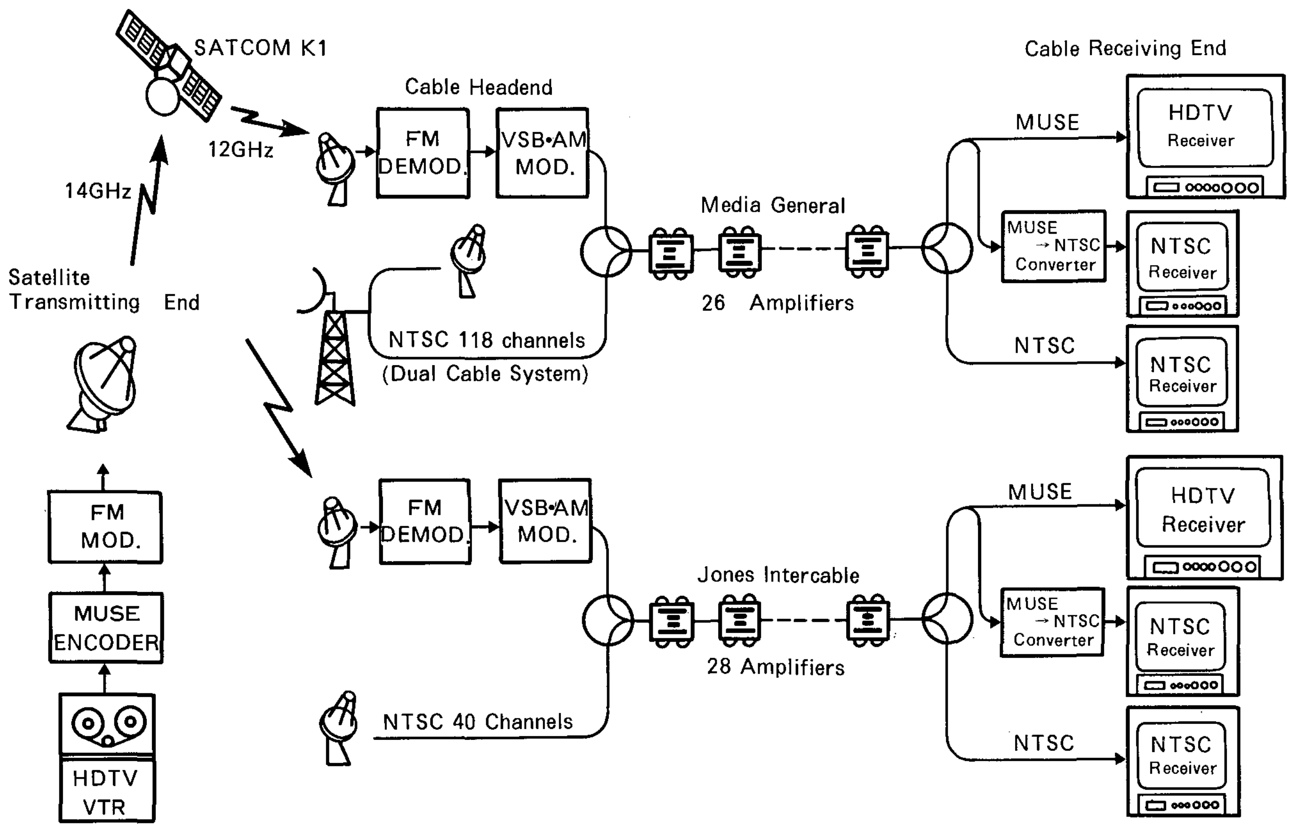


Figure 5 Experimental Setup of HDTV Cable Transmission via Satellite

Items	Data	Comments
Downlink EIRP	50.0 dBW	
Receiving Antenna Gain	49.5 dB	Antenna Diameter 3.1m
Uplink CN Ratio	30.0 dB	
Total CN Ratio	23.7 dB	
Transmission Bandwidth	36 MHz	
Frequency Deviation	17 MHz	Without Energy Dispersion
FM Improvement	17.7 dB	
Emphasis Improvement	9.5 dB	
Received Unweighted SN Ratio	50.9 dB	

Table 1 Transmission Parameters of Satellite Link

Items	State-of-the-art System	Conventional System
Company	Media General	Jones Intercable
Number of NTSC Channels	118(Dual Cables)	40
NTSC Transmission Method	HRC*	Standard
Number of Subscribers	167,000	32,000
Modulation Method	VSB•AM	VSB•AM
Transmission Bandwidth	12 MHz	12 MHz
Video Carrier Frequency	330.30 MHz	331.25 MHz
Channels for HDTV	42, 43	42, 43
Adjacent Channels	41, 44	41
Number of Amplifiers	26	28
Cable Length	13 km	20 km

* Harmonically Related Carrier

Tabel 2 Outline and Transmission Parameters of Cable Facilities

The typical transmission parameters of SATCOM K1 are shown in Table 1. The outline and the typical transmission parameters of cable facilities are shown in Table 2.

Experimental Results

The experimental results of this MUSE cable transmission via satellite are shown in Table 3. The received CN ratios of the satellite link at both cable headends are greater than 23 dB, and the received CN ratios at both cable receiving ends are greater than 44 dB (8.1 MHz). Thus, excellent HDTV pictures whose picture qualities are grade 4 at least, and whose unweighted SN ratios after demodulating are greater than 39 dB, can be obtained.

Thus, this transmission experiment has proved that excellent HDTV pictures can be achieved by MUSE-VSB•AM cascaded transmission comprising a satellite and cable facilities.

Facilities	Satellite System		Cable system				
	Frequency Deviation (MHz)	Received CN Ratio (dB)	Received CN Ratio (dB)	Modulation Depth (%)	Unweighted SN Ratio (dB)	CTB (dB)	Grade of Picture Quality
State-of-the-art System	16.8	23.5	44.5	83.3	41.0	-44	4
Conventional System	16.8	24.0	44.3	70.0	39.1	-53	4

Table 3 Experimental Results

HDTV OPTICAL FIBER CATV SYSTEM EMPLOYING DEMAND ACCESS TECHNIQUE

FM Frequency Allocation and Parameters

The frequency allocation of MUSE-FM signals on the trunk line is shown in Figure 6. At the trunk line, 34 MUSE-FM signals are allocated with the frequency interval, 38.36 MHz, used in Japanese satellite broadcasting. At the subscriber line, any selected 4 channels are transmitted on the BS-IF band (1-1.3 GHz). The FM parameters of HDTV MUSE-FM signals are shown in Table 4.

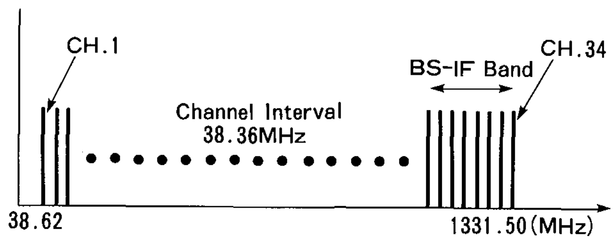


Figure 6 Frequency Allocation of MUSE-FM on Trunk Line

Items	Data
Video Bandwidth	8.1 MHz
Frequency Deviation	10.2 MHz
FM Bandwidth	27 MHz
FM Improvement	11.9 dB
Emphasis Improvement	9.5 dB

Table 4 Transmission Paramenters of MUSE

System Configuration

The system configuration of an optical fiber CATV system for HDTV is shown in Figure 7.

At the CATV headend, MUSE signals are frequency modulated and combined with the rebroadcasting MUSE-FM signals. Thirty four channels of FM signals are transmitted over 20 km to the hub through a single mode fiber by intensity modulation of an LD (1.3 um).

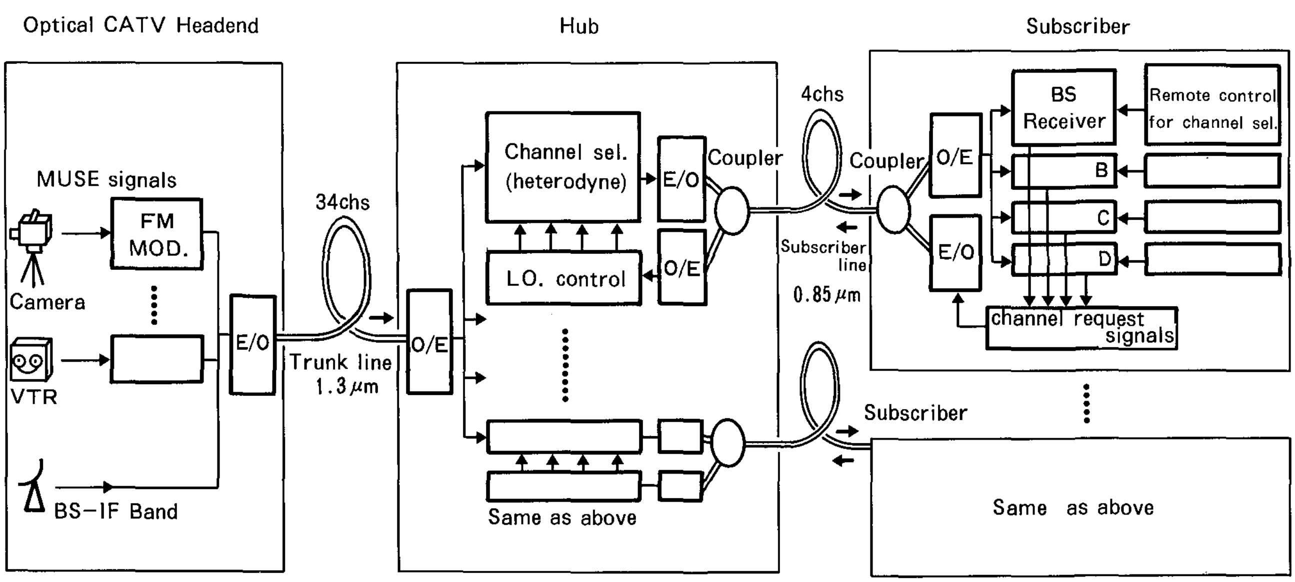


Figure 7 System Configuration of Demand Access Optical Fiber CATV System

At the hub, these signals are transformed into electrical signals. Any 4 signals can be selected by channel request signals from an individual subscriber employing a heterodyne technique. The selected 4 signals allocated in the BS-IF band, modulate a low cost LD (0.85 μm), and are transmitted to each subscriber through an ordinary single mode fiber.

At the subscriber, these signals are transformed into electrical signals, and distributed to 4 conventional home receivers used for satellite broadcasting, and then demodulated. The channel request signals are sent to the hub through the same fiber used for a down stream. The digital signal format of the remote controlled channel selector at the television sets is adopted for the transmission of channel request signals.

The optical transmission parameters are shown in Table 5. Optical fiber couplers with a core diameter of 6 μm, are used at both ends of the subscriber lines to obtain bi-directional transmission. These couplers also eliminate the higher order mode generated in the fibers when an ordinary single mode fiber is used at a short wavelength.

Experimental Results

The relationship between the received CN ratio and the received optical power, P_R, when multiplexing 34 MUSE-FM signals on a trunk line, is shown in Figure 8. In this figure, the dashed line indicates the CN ratio when all channels are not modulated, and the solid line indicates the effective CN ratio including distortion power as noise when all channels except the measured channel are modulated. In the frequency allocation chart shown in Figure 6, we can avoid the second-order distortions, but third-order distortions drop into the FM transmission band. The CN ratio degrades effectively as the optical modulation depth m increases, because many third-order distortions are also frequency modulated and added in random frequency and phase, and behave like random noise. Therefore, the optical modulation depth giving the maximum CN ratio can be determined when P_R is given. In the design for 20 km transmission on the trunk line, P_R is -12 dBm since the output of the optical transmitter is -2 dBm, and the optical loss of the fiber is assumed as 0.5 dB/km. In the case of Figure 8, the maximum CN ratio can be obtained as 28 dB.

Items	Trunk Line	Subscriber Line (Down stream)	Subscriber Line (Up stream)
Optical Source	InGaAsP-LD	GaAlAs-LD	GaAlAs-LD
Wavelength	1.3 μm	0.85 μm	0.85 μm
Threshold Current	5 mA	15 mA	12 mA
Bias Current of LD	40 mA	42 mA	16 mA(Max.)
Optical Power	-2 dBm(10/125)	+2 dBm(6/125)	-10 dBm(6/125)
Optical Isolator	Isolasion: 60 dB	not used	not used
Optical Fiber	10/125 SMF	10/125 SMF	
Optical Coupler	-	2 x 2 type (0.85 μm) 6/125 SMF	
Optical Receiver	InGaAs-APD	Si-APD	Si-PD

Table 5 Optical Transmission Parameters

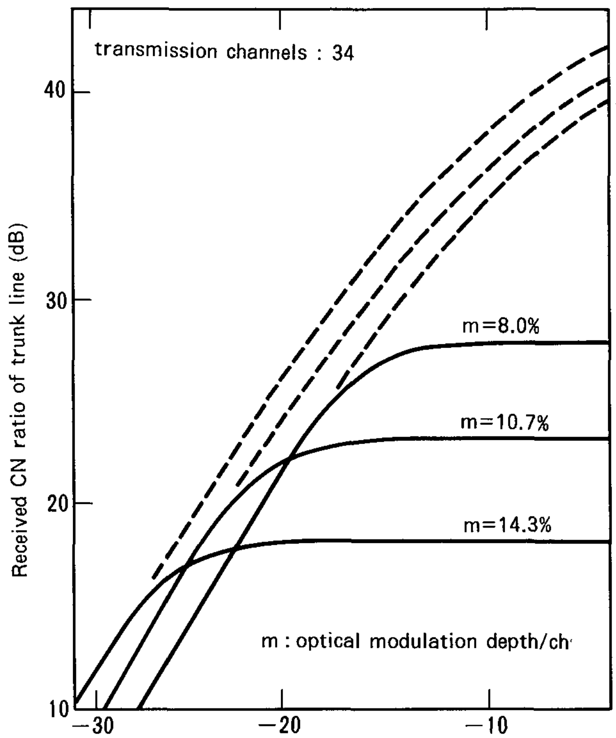


Figure 8 Received CN Ratio on Trunk Line
(dashed line : unmodulated
solid line : modulated)

In the case of rebroadcasting HDTV from a satellite on CATV, the CN ratio at a subscriber, $(C/N)_{total}$ is given by the following equation:

$$\frac{1}{(C/N)_{total}}=\frac{1}{(C/N)_{sat}}+\frac{1}{(C/N)_{trunk}}+\frac{1}{(C/N)_{sub}}$$

, where $(C/N)_{sat}$ is the received CN ratio of the satellite link at the headend, $(C/N)_{trunk}$, $(C/N)_{sub}$ is the CN ratios of the trunk and the subscriber lines, respectively. We obtained $(C/N)_{sat}$ of 25 dB with a parabolic antenna of 1.2 m diameter and a RF converter of 1.3 dB noise figure. If we design the system with $(C/N)_{total}$ of 17.5 dB, which gives the perceptible noise limit, the required CN ratio for the cascaded transmission on the trunk and subscriber systems is greater than 18.4 dB.

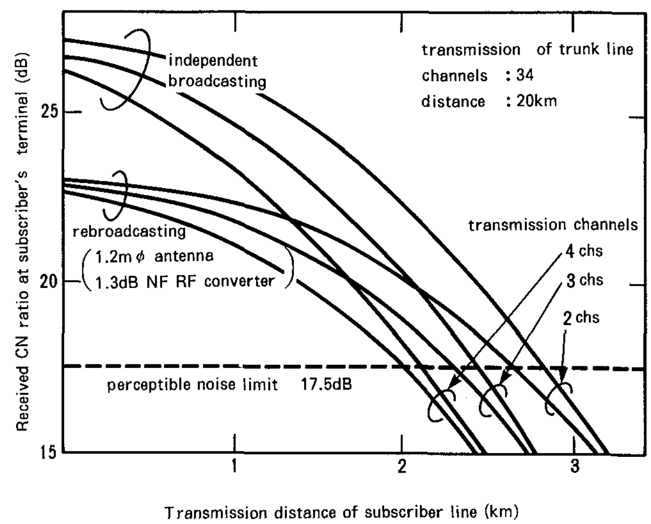


Figure 9 Total CN Ratio of Demand Access Optical Fiber CATV System

The relationship between $(C/N)_{total}$ and the transmission distance of the subscriber line after 20 km transmission of the trunk line, is shown in Figure 9 with the parameters of the number of channels on the subscriber line. Four MUSE-FM signals can be transmitted over more than 2 km, as shown in Figure 9.

CONCLUSION

The feasibility of HDTV services on coaxial cables has been demonstrated by this MUSE-VSB•AM transmission experiment. A demand access optical fiber CATV system for HDTV MUSE signals has been developed, and the possibility of the FTTH (Fiber To The Home) of HDTV has also been recognized.

References

(1) Y.Utsumi, S.Yamazaki, H.Arata, T.Susaki, M.Honda, Y.Iwadate, H.Miyazawa, T.Tokumoto: "HDTV Transmission Tests on Coaxial Cable via Satellite", NHK R&D, No.6, pp.34-47, Aug. 1989.

(2) R.Olshansky, V.Lanzisera, P.Hill: "Subcarrier multiplexed lightwave systems for broadband distribution", IEEE J. Lightwave Technol., LT-7, 9, pp.1329-1342, 1989.

(3) M.Maeda, M.Yamamoto: "FM-FDM optical CATV transmission and system design for MUSE HDTV signals", to be published in IEEE. J. Selected Area of Comm., 1990.

IMPROVED CUSTOMER SERVICE THROUGH AUTOMATION AND ENHANCED RESPONSIVENESS

NICK HAMILTON-PIERCY & JOHN ANDERSON

ROGERS ENGINEERING

ABSTRACT

This paper describes how the foundation of good customer service (quality, reliable distribution of pictures) is enhanced by a fully automated network management monitoring system with alternate signal source and transmission path switching. Also described are the integration of automated repair crew dispatch (digital dispatching and terminals within the trucks) and the voice responses of the business office and repair answer with the network management systems to further enhance quality service.

The foregoing approach is described through a "systems approach" using block schematics and examples of applications to implement such an automated approach.

BACKGROUND

How customers are taken care of will become the ultimate management opportunity in a world that is rapidly becoming Customer driven. Companies that will be successful in the 90's treat customer service as a key component in their strategic planning process.

INTRODUCTION

Like many other companies within the Cable Television Industry, Rogers has established lofty customer service goals in response to rising customer expectations. Part of our Corporate Mission Statement says that "we shall be known consistently as providers of outstanding customer service." However, there is an enormous amount of effort required between simply making these statements and turning them into reality, particularly if we rely on existing operating methods and technologies. Rogers recognized quite some time ago that if we were to realize dramatic changes in our levels of customer service, then simply adding more staff or processes to the existing operating system would not bring about these changes. We would have to make fundamental changes to our operations and bring in new technologies to create new functions and opportunities for improvement that previously did not exist.

To accomplish this we inaugurated a project called RACE which stands for Rogers Advanced Customer Environment. RACE's goal is to supply the tools to our staff to provide an improved level of customer service that is quick, comprehensive and economical. We will provide our staff with the proper information and support so that they can deal immediately with customer problems and resolve them with minimum call-backs or repeat visits. At the same time, we want to provide this enhanced level of service without increasing existing staff levels and operating costs.

Two major initiatives have been taken to bring this about:

1) Repetitive, mundane functions were identified and targeted for automation (i.e. computerization) as much as possible.

2) Interfaces were established to integrate existing customer service technologies with new ones in order to leverage more effectivity from each of the stand-alone systems.

The technologies in existence prior to RACE included a Computerized Billing and Customer Service Data Base, a Computer-Aided Drafting and Design System (CADD) and Automatic Telephone Call Distributors (ACD). However, each of these were isolated systems with no interconnection for shared facilities or functionality. New technologies under consideration or in the process of implementation are Voice Response Units (VRU's) to automate incoming and outgoing phone traffic; Digital Truck Dispatching which creates a "paperless" work order cycle and makes information available on CRT screens in each service vehicle; and Network Management (Status Monitoring and Control) to provide feedback on performance and control of the components within the microwave, fibre optic and co-axial networks. Each technology had been fully justified based on its own individual merits and contributions to the Rogers customer service plan. However, through interconnection of these technologies there is the potential for enormous additional benefit at a small incremental cost.

Figure 1 is a block diagram of the interconnection of the various customer service technologies using a local area network. Each technology has an integral translator that converts its proprietary protocol to open (ISO/OSI) data packet standards. Once on the network, data can be transferred freely to any device. The network is compatible with distributed components and remote terminals. New technologies can be added or deleted without affecting the functionality of others. The Integrator/Server is the "intelligence" in the network. It is programmed to take specific actions and set priorities under certain combinations of conditions being reported. It adds a level of sophistication to make the configuration more than just a data exchange network.

CASE STUDIES

The following fictitious "Case Studies" are offered to provide a vision of both Pre and Post RACE environments. Although hypothetical, the pre-RACE case study is based upon current practices.

Case Study 1/Pre-Race

It is a Saturday evening at 7:30 p.m. during prime time viewing hours. A blown amplifier module has caused an outage somewhere in the system. Suddenly there are 200 phone calls to the office. Part-time dispatchers are on duty this evening, and can see no correlation between the streets, and call out the entire standby fleet available (four trucks) to the affected homes in time stamped order.

The four trucks proceed through the list of calls and gradually begin to piece together a hypothesis of a major outage.

Meanwhile customer homes have been visited, and restoration of service was not complete, although the intrusion is.

ROGERS ADVANCED CUSTOMER ENVIRONMENT

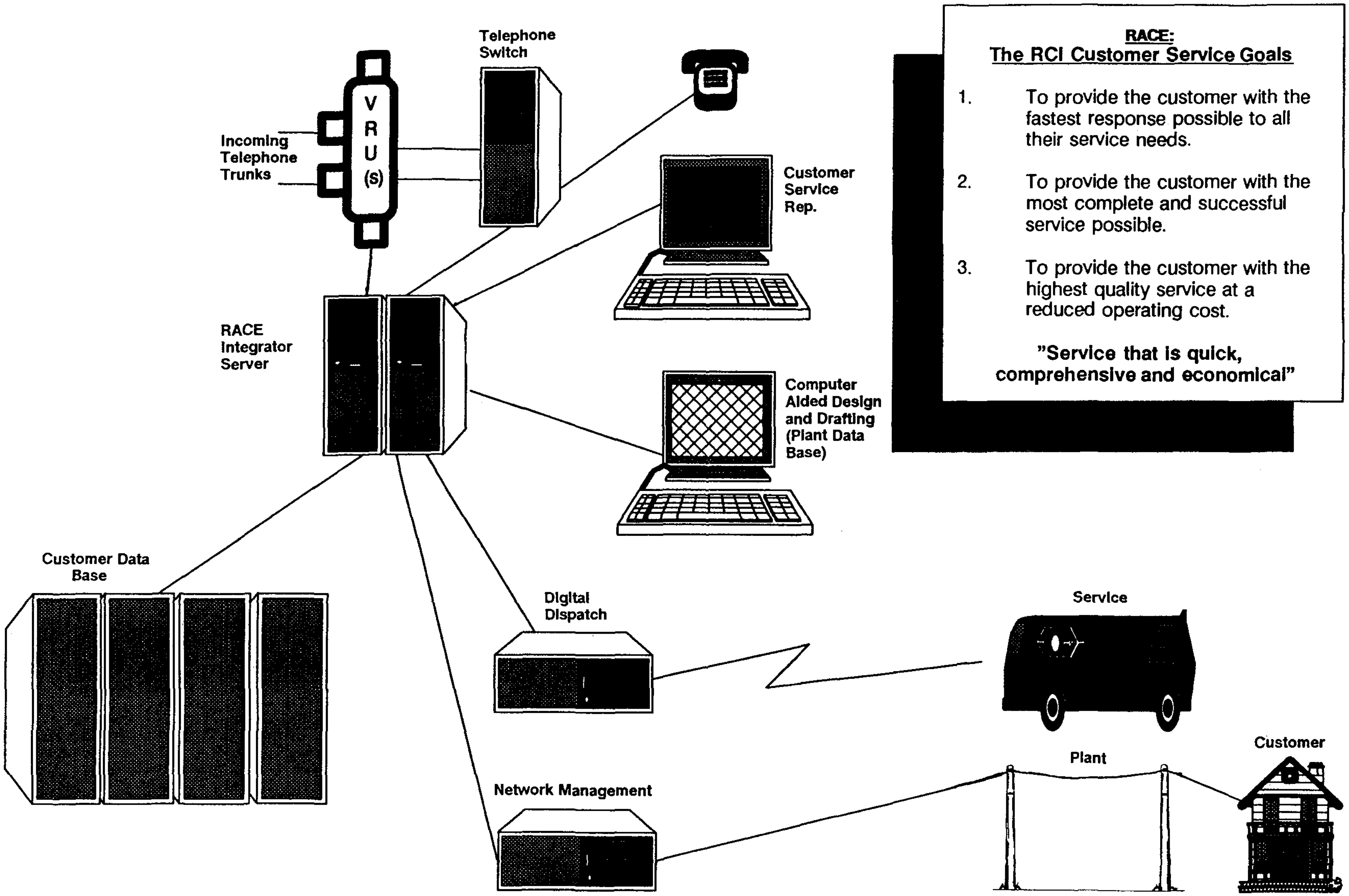


FIGURE 1.

After a couple of hours, and the review of several plant maps by the dispatchers, a common point of failure is assumed and a truck is dispatched to this location. (This assumes that the plant maps are current and reliable).

A service vehicle arrives at the common point of failure two hours from the time of outage. Quickly he is able to determine that he has had no prior experience with the unit in question. However, working with the radio he is able to obtain assistance in correcting the problem. Although his service vehicle has thousands of dollars worth of inventory, he is without the necessary module for repair. Fortunately, the service technician on the other end of the radio (to whom was providing him with assistance) has the module necessary and races for the outage destination.

Finally after three hours of outage, the problem is corrected and service is restored.

Pre-RACE Result

There were so many problems in Case Study #1 that it would be just as lengthy to point them out as to repeat the story. One point that may be hidden, is that even with millions of dollars of "state-of-the-art" technologies, simple location and correction of this outage could only be minimally shortened.

There is a weakness because there is no integration of the systems. The technologies are functioning, but only on the strengths of their own abilities. One could spend millions of dollars on new technologies, and not realize the maximum "hidden" potential benefits.

RACE is designed from the inception of new technologies to provide a comprehensive customer service environment regardless of the human resources available.

Case Study 2/Post-RACE

The same outage situation has occurred, but RACE technology is in place.

Thirty percent of the incoming calls are answered by the VRU's. ANI provides the subscribers home phone number, and VRU's run through the traditional CSR diagnostic check list, prior to posting the call as a viable outage.

CSR's perform the same function as the VRU's, where the customer either has no touch tone phone, or simply wants to deal with another human. Since the CSR's load has been reduced by the VRU partner, he/she is able to offer personal assistance to even the nastiest of callers.

As outages (homes or business customers) are confirmed by the CSR's and VRU's, the information is routed to the Network Management System which provides a detailed graphic display of the plant and recorded outages. By using stored "knowledge" of the plant, the Network Management System is able to "look for" common points of failure. It does this by traversing "up the tree", looking for the components common to major outages.

Network Management locates a common point of failure on the screen and displays this to the dispatchers. At the same time the Network Management System issues a message into the network asking for status information on the suspected amplifier. The response from the amplifier is compared to standard parameters stored within the Network data base for further analysis.

Network Management automatically issues a message to the Digital Dispatch System for a support vehicle to proceed to the identified common point.

Network Management has determined the common point of failure, the probable equipment failure and the current and ideal operational parameters.

Before leaving for the call, the technicians knows:

- where the problem is
- what the problem is
- what he should have to correct it

Also available is a work order history for the malfunctioning amplifier.

Utilizing an integrated stores and vehicle inventory data base, the system has selected a truck known to contain the parts required to repair or replace the damaged component. With the integrated resources data base, the system has selected the service technician who is most able to correct the given outage in as short a time as possible. This would also allow us to reduce the duplication of expensive inventories between the stores and service vehicle.

More advanced mobile terminal technology can be integrated and enhanced to provide detailed plant maps to help the service technician locate the damaged equipment. With fully portable and powerful terminals, it is possible to provide the service technician with an "automated assistant" rehearsed in diagnostic procedure for the suspected device.

The dispatchers now know where the outage is, and its scope (area of outage). This information can be relayed BACK to the CSR's and VRU's for presentation to the incoming subscriber call.

For example (via VRU); "...Please enter your home phone number...We currently have an outage bordered by York Mills and Lawrence, and Leslie and Don Mills. A service vehicle is on location, and should have the service restored within approximately twenty minutes...."

Twenty minutes later service is restored.

The status screens are updated with call cleared, the VRU and CSR console messages are automatically dropped.

Except for the actual outage, the customer was not further inconvenienced by an in-home visit. Service was restored to multiple reported outages by a single service vehicle.

Post RACE Result

What has changed? The technologies are essentially the same, yet the results were much better than Pre-RACE, with less staff.

The difference is the integration provided by RACE. RACE combines the strengths of the autonomous technologies, driven by the Customer Service Goals.

No subscribers need be bothered in their home. Only one truck rolled. Customers were informed of the situation. Part-time staff are able to perform like seasoned professionals. The customer's service was restored far faster than in traditional modes.

PREVIOUSLY UNATTAINABLE LEVELS IN CUSTOMER SERVICE BECOME ROUTINE UNDER RACE.

One "hidden asset" is that the Network Management Systems normally would have identified a failing part long before actual failure. Under normal RACE operation, the Network Management System cyclically inspects all the plant components looking for potential points of failure. Under normal RACE operation, a work order would have been issued to inspect the amplifier prior to total failure.

Network Management

Effective management of our coaxial, fibre optic and microwave networks is crucial for the delivery of excellent customer service. Traditionally, cable TV systems have employed, if at all, a single-vendor "status monitoring" system which simply reported on the current status of individual trunk amplifiers.

With the implementation of the Rogers Fiber Architecture (ref. 1989 NCTA Technical Papers) with its hierarchy of Primary Fibre Hubs, Secondary Hubs and co-axial plant, we recognized the need for a more comprehensive network management system that not only monitored the status of the networks, but also provided a means to control switching functions for backup facilities should failures ever occur. The network management system would also need to function in a multi-vendor environment with an hierarchy of network technologies (i.e. FM fibre, Digital Fibre, AM fibre, AM co-ax, microwave links, TVRO's). Rogers currently has such a system under development with multiple networked graphical work stations. The system is being developed with standard OSI network management translators to provide the access link to the various vendors' proprietary hardware platforms. The current Fibre Network Management System is scheduled for completion in 1990 and will offer true cross-system management facilities to all of the fibre components, including laser transmitters, receivers, modulators/demodulators, and redundancy switching (See Figure 2).

The Network Management applications available in the system include:

a) Monitoring - Automated monitoring of the network's transmission, received signal processing, and equipment building facilities provides a continuous assessment of the infrastructure's operation. It provides immediate feed-back on a malfunction prior to an effect being noticed by the customers. Each part of the distribution infrastructure needs its own unique monitoring function tailored to the types of malfunctions expected in that part of the network:

1) Coaxial Trunk Network
2) Fibre Supertrunk and Trunk Network
3) Equipment Powering and Auxiliary Powering

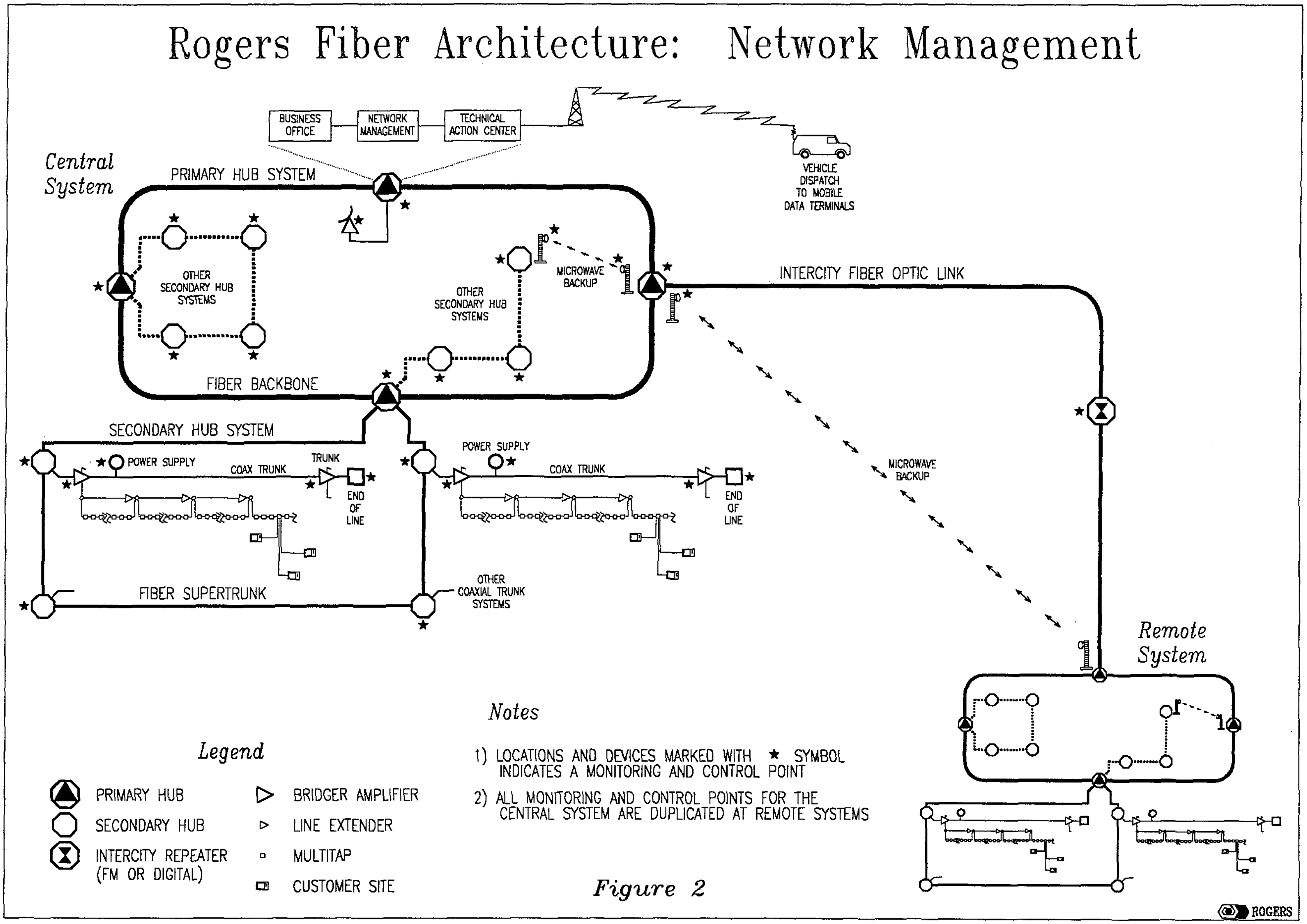

Rogers Fiber Architecture: Network Management
BUSINESS OFFICE
NETWORK MANAGEMENT
TECHNICAL ACTION CENTER
VEHICLE DISPATCH TO MOBILE DATA TERMINALS
Central System
PRIMARY HUB SYSTEM
OTHER SECONDARY HUB SYSTEMS
OTHER SECONDARY HUB SYSTEMS
MICROWAVE BACKUP
INTERCITY FIBER OPTIC LINK
FIBER BACKBONE
SECONDARY HUB SYSTEM
POWER SUPPLY
COAX TRUNK
TRUNK
END OF LINE
POWER SUPPLY
COAX TRUNK
END OF LINE
MICROWAVE BACKUP
FIBER SUPERTRUNK
OTHER COAXIAL TRUNK SYSTEMS
Remote System
Notes
1) LOCATIONS AND DEVICES MARKED WITH ★ SYMBOL INDICATES A MONITORING AND CONTROL POINT
2) ALL MONITORING AND CONTROL POINTS FOR THE CENTRAL SYSTEM ARE DUPLICATED AT REMOTE SYSTEMS
Legend
PRIMARY HUB
SECONDARY HUB
INTERCITY REPEATER (FM OR DIGITAL)
BRIDGER AMPLIFIER
LINE EXTENDER
MULTITAP
CUSTOMER SITE
ROGERS

Figure 2

4) Headend Hubs and Microwave
5) Equipment Shelters and Buildings
6) Satellite Uplinks, Downlinks
7) Data Modems and Multiplexors
8) Extending Monitoring Visibility into Interconnected Networks Operated by Others
9) Providing Visibility of the Network Monitoring Functions for use by the other operator of an Interconnected Network or by specialized Customers connected to the Network.
10) Ability to make available remote objective measurements from any principal node or extremity of the network.
11) Ability to make available remote subjective assessment of video and audio quality from any principal node or extremity of the network.

In addition to those benefits discussed, comprehensive performance and functional monitoring of all parts of the network allows maintenance functions to be dispatched only to where needed. This eliminates need for random patrolling of the network and its elements by maintenance crews and the many non-productive adjustments and measurements they routinely conduct. A substantial reduction in maintenance expenses are potentially available.

b) Alarm Rationalization - A significant shortcoming of most existing "Status Monitoring" Systems is their propensity to overwhelm the operator with immense amounts of data; for example if a trunk amplifier fails, then it affects the transmission through all subsequent trunk amplifiers and the status monitors within each report a faulty condition. There may be as many as 60 amplifiers downstream of the faulty unit, and this represents an overwhelming amount of data being offered to the operator. The real fault gets buried. The situation is aggravated with an intermittent fault. It is because of this shortcoming that Status Monitoring has, until now, been considered a very marginal tool for assisting operations and has not gained popularity. The new Network Management System, through software intelligence and interaction with the computer aided design network map database, filters out the excess data and presents the operator only the pinpointed fault.

This is a practical, useable, alarm display which will allow efficient responsiveness, appropriately directed maintenance and repair, and improved transmission reliability. Reduced operating expenses and a greater level of customer service will result.

c) Automatic Diagnostics - This application combines the rationalized monitoring function with intelligent software, and through automatic interaction with the digitized network maps and equipment inventories, provides problem diagnosis in plain working English User-friendly network monitoring minimizes the time to isolate a problem, eliminates incorrect judgments on the source of the problem, and minimizes network down time.

d) Automatic Circuit Restoral - The various monitoring applications pinpoint the faulty portion of the signal processing equipment or distribution network. Command and switching signals are automatically generated through the application of additional diagnostics and software intelligence. These commands invoke a number of automatic functions:

1) Switches in an alternative transmission route if part of the fibre, microwave or coaxial trunking network become faulty.

2) Switches in an alternative signal source if severe reception difficulties are encountered on TV and broadcast radio signals.

3) Switches in back-up RF, video or optoelectronic equipment should any channel processing modulator, demodulator or data equipment fail.

The switching process is either totally automatic or is a two step process where a prompt message is sent to the Technical Action Centre (TAC). The message describes the problem and instructs which switch should be activated. This option enables the TAC personnel to ignore switching commands that may have resulted from deliberate maintenance activities on the network or rearrangement of equipment.

Automatic Circuit Restoral minimizes down time and interruption of service. It operates 24 hours per day without the need for dedicated switching personnel and minimizes the chance for incorrect switching.

e) **Assisted Repair** - This enhancement builds on the strengths of the monitoring and diagnostic applications previously discussed. When a technician is dispatched to correct a fault, the mobile data terminal assists the technician by displaying the nature of the fault and its precise location. It also provides information on the most likely corrective action to take while executing the repair. Once the repair is completed the technician is prompted to either confirm the remedy was correct or provide additional information through the terminal key pad on what action was necessary. The database accepts this new information and updates its rule base on appropriate corrective action.

In its more advanced form this service aid not only provides expert advice on the likely cause and corrective action for the reported problem but on request it walks technicians through more difficult diagnosis and automatically provides further information on appropriate corrective actions. Through successive usage and updates, the technical service aid delivers recommendations with a greater likelihood of being correct.

The most significant benefit is that the level of training of field personnel can be quite modest yet the personnel will be able to handle a diversity of equipment types and technical problems. Furthermore, the automatic guidance in problem correction will significantly speed up the repair process and minimize down time. An overall reduction in repair staff expenses should result.

f) **Automatic Issue of Trouble Tickets** - The monitoring, VRU and diagnostics applications previously discussed provide all of the contents necessary to issue a trouble ticket. This application collects together this information and determines to which group or specific person the compiled ticket should be passed.

Prompt and more accurate trouble

tickets, will result in more efficient utilization of personnel provides a higher level of customer service.

g) Automatic Dispatching - The previous application automatically compiled the trouble ticket. This application automatically issues the trouble ticket to field personnel through various dispatching channels

1) Digital Dispatching directly to the truck or portable terminal, or
2) Dispatching to the field personnel via a pager, or
3) Automatic telephone call to a technician on standby with the VRU providing specifics of the trouble ticket or work order.

An enhancement to this application utilizes the physical plant and geographic database information to pinpoint the most recent known location of field personnel. The skill sets of each technician are automatically assessed through interaction with the human resource data base and the technician with the appropriate skills located closest to the trouble is automatically dispatched.

Prompt and more accurate dispatching of field personnel provides for more efficient utilization of staff. It also minimizes non-productive driving time and is particularly valuable in the dispatching of specialist technicians for business customers.

h) Alert and Advisory Screens for CSR/TSR - The diagnostic knowledge and intelligence of many of the above applications results in valuable information that needs to be known by CSR's/TSR's as they interface with the customers. In this application information associated with the geographic location of the calling customer automatically appears on the help screen as the customer call is accepted or as the customers file is retrieved from the database. Additional information from this enhancement includes graphic displays of the customer areas affected and identification of the boundary streets. It also provides the local weather conditions by accessing the local Environment Canada Data Base. This enables the CSR/TSR to provide a more personal local focus with the customer although the CSR/TSR might be located hundreds of miles away.

These customized screens provide a higher level of customer service, more accurate interpretation of a customer's problem, and minimize inaccurate information being given to customers in a Regional Service Office.

i) Automatic Alert to Business Customers - In the telecommunications competitive environment it is becoming more and more important that the customer be immediately advised if a transmission problem is developing and if failure occurs. It is also important to provide the customer progress results on the restoration status. Furthermore, the sophisticated customer who perceives they have a problem with the network needs to know the status of their particular circuits to facilitate their own diagnostics. This enhancement to the previous application extends certain monitoring, configuration and help screens to the customer and provides the necessary level of customer service to remain competitive in the telecommunications environment.

j) Graphical Information in Vehicles -This application makes use of the Digital Dispatching telecommunications channel and terminal equipment to provide the field technician access to the physical plant data at a macro level. It provides access to the CAD plant maps and on the micro level it provides the technician with both graphical and text information on the electronic and electro-optic equipment at hubs along with detail on fibre allocations, maintenance responsibilities and restoration procedures. Additional vehicle equipment such as a higher resolution terminal screen and hard copy printer are necessary for the technician to use this application.

This application will eliminate the massive amounts of paper maps in the truck and totally up-to-date information available for the technicians at any time, and it will also result in more rapid and accurate equipment replacement, repair or fibre restoration during transmission malfunctions. On-line feed-back ensures the physical database is continually updated as field personnel uncover inconsistencies between the physical plant in place and the database records and enter corrections through their terminals.

k) Automatic Collection of Statistics - Statistics related to the reasons for technical service visits, equipment failures and the number and frequency of transmission interruptions are very questionable in their accuracy at best when gathered manually. Many outages for instance go unrecorded. This application collects together all of the automatic monitoring, dispatching, diagnostics, data and arranges it into statistical records for use by the Management Information System.

It provides usable accurate data for measuring levels of service and efficiency in the utilization of staff. It provides objective information for future planning and technical upgrading of the network or rearrangement of equipment. Also, measured transmission performance satisfies business customers that the contractual obligations have been met.

l) Automatic Customer Advisory on Receipt of Trouble Call - On receipt of a trouble call the VRU offers touch-tone equipped customers four inquiry options:

1) Total Loss of Cable TV Service
2) Loss of Only Pay TV Services
3) Other Reception Problems
4) Stay on the Line for a Technical Service Representative

For options 1) - 3) Automatic Number Identification (ANI) has already indicated the customers home phone number and the VRU advises whether the problem is known to exist in that customer's area, the actual nature of the problem, and the status of correction progress. If the VRU itself has not been advised that a problem exists, then it immediately routes the customer to a Technical Service Representative, and simultaneously posts an alert to the Network Management System. This in turn awaits further postings to determine whether the problem may have been unique to that customer. Then once it has been cleared, the VRU would phone back the registered home telephone number and ask the customer whether they are satisfied that the problem has been resolved and to invite the customer to signify this by hanging up the phone or to stay on the line for help by a Technical Service Representative.

This system provides an improved level of customer service while presenting less calls to technical service representatives, hence enabling a lower level of TSR staffing for a given telephone service quality level.

m) Post-ServiceSatisfaction Calls - At a predetermined time, after any installation or service visit, the customer is called by the VRU and their satisfaction with the recent visit verified. A satisfied customer is prompted to hang up the phone; an unsatisfied customer is invited to stay on the line and is routed to a Technical Service Representative. These calls are orchestrated with predictive dialler equipment keyed to the availability of the existing TSR pool and the incoming customer call load on the telephone lines.

This provides a vehicle for customers to advise of dissatisfaction or satisfaction, provides feed-back on the quality of service being provided, and provides the above benefits without additional Technical Service Representatives.

n) Digital Dispatch Call-Ahead to Customers - This application is available to those operations employing digital dispatching of their service vehicles. Upon leaving one customer and travelling to the next, the technician presses the "Call-Ahead" key on the vehicle terminal. The digital dispatching system instructs the VRU to phone the customer in conjunction with the predictive dialler equipment as in the above application. If the call is not answered by the customer, the visit is automatically cancelled and placed for rescheduling. The technician meanwhile proceeds to the next scheduled visit. If the call is answered, the VRU asks the customer to accept the appointment by hanging up the phone, or staying on the line to advise a Technical Service Representative for a change in the appointment. The VRU also invites customers with a touch-tone phone to press the #1 key on telephone if the reason for the service is no longer required.

This call-ahead alerts the customer of the imminent arrival of the technician so that they may prepare themselves and not be inconvenienced.

It also saves non-productive truck visits if the customer is unavailable or the reason for the visit no longer needs attention, and provides a much more efficient utilization of field technician time.

o) Elimination of the Need for Touch-Tone Responses - True Voice Recognition technology is becoming well advanced. The accuracy of identifying simple spoken commands such as 0-9, yes or no, etc. in thousands of different dialects is now very high, and the equipment is becoming fairly inexpensive. Substituting the spoken word in place of touching the telephone keypad on a prompt from the VRU substantially simplifies many of the above applications, and makes them accessible to all of the customers.

This is very customer friendly, and enhances the effectiveness of many VRU prompted applications.

SUMMARY

The Rogers Advanced Customer Environment (RACE) is being developed to provide a quantum improvement in the level of service offered to our cable TV subscribers and business customers. RACE is a valuable tool to offset the problems of growing complexity of coaxial, fiber optic and microwave networks, the demands for better customer service, and the scarcity of qualified staff.

The Network Management System has been highlighted as the key element in providing exceptional customer service. However, in itself, Network Management will not provide all of the facilities necessary to meet the customer service goals. Heavy reliance is also placed upon the integration of the Network Management System with the Customer data base, the plant data base (CADD), telephone technology such as advanced switches and voice response units (VRU), as well as digital dispatching to mobile data terminals in the service vehicles. This integration of the technologies leverages substantial additional benefits and value out of previously isolated systems. The examples and applications presented in this paper are all feasible using this approach and a number of them are currently under development.

ACKNOWLEDGEMENTS

The authors wish to thank their colleagues at Rogers Engineering for their insights and contributions to the development of the solutions to the applications presented in this paper.

REFERENCES

1989 NCTA Technical Papers, pp 104-110, G. Hart, N. Hamilton-Piercy

CURRICULUM VITAE

NICK F. HAMILTON-PIERCY, P. ENG.

Nick Hamilton-Piercy is the Vice-President, Engineering and Technical Services for Rogers Cablesystems, his duties include managing the corporate engineering group, Rogers Engineering, as well as being responsible for the engineering and capital programs within the Company operating systems. Nick has been with Rogers for 15 years.

Prior to this, Nick was with the Canadian Marconi Company where he managed an analogue and R. F. engineering department responsible for the design development and manufacture of a wide range of telecommunications equipment including microwave radio systems, multi-channel UHF tactical relay systems and microwave R.F. hardware.

Nick graduated in Engineering from the Medway College of Technology in Chatham, England, during 1961 where he specialized in electrical engineering and qualified for full Chartered Engineering status. He is a member of the Institute of Electrical Engineers (IEE) U.K.; a member of the Association of Professional Engineers of Ontario (APEO); a Senior Member of the Institute of Electrical and Electronic Engineers (IEEE); a Senior Member of the Society of Cable Television Engineers (SCTE); Chairman - Futures Committee, Canadian Cable Television Association (CCTA); Chairman ATV Subcommittee of Cable

Television Laboratories Technical Advisory Committee, and Chairman of the National Cable Television Association Engineering Committee - HDTV Subcommittee.

JOHN C. ANDERSON, P. ENG.

John C. Anderson is currently Vice-President, Project Engineering for Rogers Engineering, the Corporate Engineering division of Rogers Cablesystems. In this role he is responsible for a number of corporate projects which are targeted towards improving customer service. Since joining Rogers in 1981, he has held a number of Engineering positions, most recently as Vice-President Engineering for Rogers Cable TV - Toronto. Prior to joining Rogers, John spent three years with Maclean-Hunter Cable TV in various engineering roles and two years with the Jerrold Division of General Instrument.

John is a graduate of the University of Waterloo with a B.A.S.c in Electrical Engineering. He is a registered Professional Engineer in the Province of Ontario (APEO).

IMPROVING CONSUMER FRIENDLINESS WITH ON AND OFF PREMISE ADDRESSABLE EQUIPMENT

JOHN E. BURKE, PRODUCT MANAGER
TOM MARTIN, DIRECTOR R & D
JERROLD COMMUNICATIONS

ABSTRACT

This paper will outline in detail the various On- and Off-premise approaches to addressability including: Interdiciton, Switches, and Hybrids. The current state of techniology for each of these approaches will be discussed, as well as what the future may hold in store.

The paper will also compare and contrast on- and off-premise from a conceptual perspective. The differences may be subtle, but are important in understanding why operators select one approach to another. Topics that will be covered include: Cost Sub vs Cost Per Port, Powering, and Deployment. Possible solutions to the limitations of both approaches will be presented.

Let's first define on- and off-premises. Both provide addressable control of services with equipment located outside the home; the difference is where you locate the equipment. On-premises approaches attach the equipment to a sub-scriber's residence in a secured, weatherized housing. Off-premises locates the equipment in a weatherized housing on the pole, strand or in a pedestal.

The reasons why an operator might want to consider on- or off-premises are virtually the same and can be segregated into three key areas: addressable control capability, consumer friendliness and improved operating efficiencies.

On- and off-premises equipment both contain data receiving and microprocessing memory circuitry necessary for addressable control. This circuitry enables the equipment to be remotely controlled by a computer located at the headend, office or some other remote site. While the functions performed by the circuitry will vary from one approach to another, the goal is the same: remote control of services and features. These generally include service connect or disconnect, pay-per-view and premium service upgrades or down-grades. Additionally, both approaches offer remote control of security devices, including interdiction (jamming) and traps (control through switches).

It may be helpful at this point to discuss an aspect of CATV technology that has long been misunderstood: The words addressability and security are not synonymous. Addressability, as outlined in the previous paragraph, can be defined as "the ability to remotely control services and/or features." Security, in the CATV world, is provided by the actual device used to supply or deny services or features. Most likely, the misunderstanding has come about as a result of the widespread use of addressable descrambling set-top converters, an in-home approach to security that utilizes a separate RF tuner and descrambling circuitry to supply or deny services and features.

On- and off-premises approaches are similar to the addressable descrambling set-top converter approach in that they both offer addressable control. The difference is in the type of security device used (jammers or switches vs. descrambling circuitry), the tuning function and the location of the device (inside vs. outside the home).

Having clarified "addressability" and "security" we can better understand the next reason why an operator might consider on- or off-premises technology: consumer friendliness. These technologies achieve the ultimate in friendliness by providing a transparent cable service delivery system (from the consumers' standpoint) to the entire house. The security devices utilized in on-/off-premises approaches (jammers or switches) do not require the inclusion of a separate RF tuner as is used in descrambling set-top converters. This enables the entire spectrum to be passed on to a cable-compatible television or VCR in the clear, thereby allowing all of their functions to be fully utilized (i.e., the TV's or VCR's handheld can be used to tune channels, viewing/taping combinations are unrestricted).

The potential savings arising from improved operating efficiencies is a third reason to consider on- or off-premises. These savings result from the fact that once the equipment is installed, one truly has an "addressable home." Since in-home equipment is not utilized, truck rolls are not necessary for connects or disconnects. Additionally, by minimizing the need for in-home equipment, equipment theft and abuse are minimized.

Turning now from the reasons why a cable operator may want to implement on-or off-premise equipment, we will attempt to outline the various approaches to on-/off-premise security.

Trap Switching

Positive and negative traps have been used for years to supply or deny a particular channel to a subscriber. Adding addressability to traps yields a low cost on-premise control mechanism.

A typical addressable control module for trap switching is based on a series of double-pole, double-throw (DPDT) RF switches. These switches must pass the entire 50-550 MHz spectrum, with flat frequency response. One leg of each switch connects to two "F" connectors, called a port. The trap (or series connected traps) connects to this port - signals are routed out one connector to the trap, then through the trap and back to the second connector. Therefore, whenever the DPDT switch associated with that port is in the "trap" position, all signals are routed through the trap.

The other leg of the DPDT switch is a through connection called the "bypass" path. With the switch in this position, the signals pass without modification to the next port. Another RF switch in a series connection allows disconnect of the drop.

Products currently on the market offer from four to eight ports, with corresponding differences in size and cost. The smallest device available measures only 1.5 by 2.5 by 7.8 inches, and is designed for use either in a single dwelling unit plastic box, or in a larger metal multiple dwelling unit enclosure.

The RF switches can be either relays or PIN diode circuits. Advantages of PIN diode switches include enhanced reliability and reduced current consumption. PIN switches also allow incorporation of the disconnect function in the basic switch, by making each of the RF switches double-pole, double-throw, center-off. This gives excellent disconnect isolation without the added size or expense of a separate switch.

The state of each switch is controlled by a microcomputer which receives addressing and tagging data from an out-of-band FSK data receiver, in a manner very similar to addressable converters. Authorized channels or events are stored in non-volatile memory to avoid any problem with power interruptions.

Because the RF switches have some loss, an input amplifier may be used to overcome this loss and provide net gain for the module. This gain is especially beneficial in light of the increasing number of TV's and VCR's in the subscriber's home. This amplifier must be designed with low distortion and low noise figure for optimum results.

Powering for the addressable control module in an on-premise, single dwelling unit application typically is from the home. A small, plug-in wall transformer near the TV provides low-voltage AC which is routed to the addressable control module over two-conductor wire. This power wire can be messengered with the coax for a cleaner installation. The addressable control module would then have internal rectification, filtering, and voltage regulation, assuring clean and stable DC voltage to operate the unit, regardless of voltage drop between the transformer and addressable control module.

Limitations of Trap Switching

While the switched-trap approach to addressability has many benefits, a review of its limitations is necessary in an objective evaluation. There are three significant limitations with this technology - first, the limitations of the traps themselves; second, the physical size problem as more traps are added; and third, the inflexibility of addressable channel lineup.

Traps, being high-frequency passive filters, have limited "Q" or quality factor. This is the measure of filter sharpness, expressed as the filter center frequency divided by its bandwidth. For a given "Q" (which is a function of physical constraints in a passive filter), the higher the operating frequency, the wider the bandwidth of the filter. In the case of a negative trap (notch filter), this means the notch width will become so wide that it will affect the adjacent channel. Because the negative traps are typically centered on the picture carrier, the lower adjacent channel, only 1.25 MHz away, is affected first. Similarly, a positive trap, which is simply a notch filter to remove an interfering signal injected at the headend, distorts the frequency response of the channel because of its finite "Q". This inherent limitation of traps has limited their use to frequencies well below the upper limit of today's 550 MHz cable plants.

Physical configuration of an addressable control system using traps poses a difficult packaging problem. The volume available in either the single dwelling unit enclosure or multiple dwelling unit enclosure is quite limited, and four to eight traps and associated cabling occupies most of that volume. Therefore, the addressable control module must be as small as possible to be effectively used.

Channel lineup flexibility is compromised somewhat by a switched-trap approach, since the traps on each port are fixed-frequency filters for a specific channel. Once the traps are installed, there is no way to change the frequencies (channel number) of the controlled channels. The addressable control is over whether or not a channel is authorized; there is no way to redefine those channels short of replacing traps.

Interdiction with Jamming Oscillators

An approach to on-premise or off-premise addressability that overcomes most of the limitations of switched traps is interdiction with jamming oscillators.

The video, and to some degree the audio, on a channel can be severely disrupted by summing a jamming carrier into the channel at the subscriber location. This approach, like a negative trap, is a form of deny security. Jamming gives a high degree of masking or concealment, as well as high signal security. The limitations of trap switching are also overcome.

The problem of limited-Q traps and their effect on adjacent channels is not an issue with an interdiction system. The jamming carriers are well controlled as to frequency and spectral content, so jamming energy can easily be contained within the channel being jammed. This is true regardless of frequency of the channel, so jamming of channels anywhere in the spectrum is possible.

Physical configuration can be less of a challenge with interdiction. Typically eight to sixteen channels can be controlled using a device considerably smaller than an addressable control module plus eight to sixteen traps. There are also far fewer RF interconnects with interdiction - two (input and output) versus eighteen (for an eight-port addressable control module). There is also the obvious advantage of not having the expense of eight or more traps.

Perhaps the biggest advantage that interdiction offers is the flexibilty it gives an operator. As with the switched-trap system, addressable control of each of many channels is available. Unlike the switched-trap approach, the frequency of each jammed channel can also be addressably controlled (within certain limits). The typical oscillator used in an interdiction device can cover approximately a 1.4 time range. For example, an oscillator might be designed to cover the entire midband, from channel A (121.25 MHz) to I (169.25 MHz), a 169.25 / 121.25 (1.4) range. This one oscillator could then be addressably moved to any channel within the midband, by a simple download of data from the headend.

Various approaches to interdiction are possible, with the biggest differences being in the oscillator deployment. A limited form of interdiction would simply use a crystal controlled, non-agile oscillator for each controlled channel. These oscillators would be switched on or off in a manner similar to the switched-trap. The advantage of no interference to adjacent channels would remain, but there would be no flexibility to change controlled channels.

A second approach to oscillator deployment is to have a limited number of oscillators available for each subscriber. Each oscillator can then cover a fairly wide frequency range (hence, number of channels). Each of these oscillators is considerably more expensive and complex than a fixed-frequency oscillator, so typically no more than four are devoted to a drop. To control more than four channels, each oscillator is quickly hopped from channel to channel, in a move, turn on, turn off, move sequence.

The advantage to the hopping oscillator is that one relatively expensive oscillator can be used to cover many channels, as long as they are within the tuning range of the oscillator. The disadvantage is that, as more channels are jammed by one oscillator, the dwell time (duty cycle) on each channel gets shorter. The masking or concealment of the video becomes less than excellent after (typically) four channels are jammed. Another potential problem with hopping oscillators is the sidebands that are generated by the rapid on/off switching of the oscillator. These modulation sidebands, if not carefully controlled in the design, will cause interference in adjacent channels.

A third approach to interdiction is to use more oscillators per drop. To do so cost effectively, each oscillator must be low in cost yet retain frequency agility. With eight to sixteen oscillators per drop, each agile over a wide range, great flexibility for deployment is possible. Those channels requiring absolute concealment can be served by dedicated oscillators. Other channels which require less masking can share an oscillator. If channels sharing an oscillator are adjacent, an entire tier of service might be controlled with a single oscillator.

An alternative to control of a tier of service is a hybrid approach, combining an interdiction jammer with switched control of a limited number of traps. For example, a hybrid with eight oscillators and two ports for tier traps could be an excellent combination of cost-effectiveness and flexibility.

Now that we have defined on-/off-premise technology and outlined the reasons why a cable operator may want to implement the technology, it is appropriate to point out the differences between the two.

The first area of difference, cost per subscriber, results from the single-home design of on-premises vs. the multihome design of off-premises. With off-premises, devices typically have four or more ports, each port serving a single sub. Typically they consist of shell or base electronics and plug-in modules for each sub that it is capable of serving. The cost generally relates to the base electronics, in that the per subscriber cost is minimized only if all the ports are utilized (100 percent penetration). Conversely, the cost per subscriber rises if all the ports are not fully utilized. For example, one off-premises device currently being offered has four ports and $180 in base electronics, with each plug-in module costing $65. Assuming a 100 percent penetration, the cost per sub is $110. Now assume a 50 percent penetration level; the cost per sub rises to $155. Given that the national average is around 55 percent penetration, the economics become very critical.

Assume the equipment is to be deployed in a system that passes 40,000 homes and has a 65 percent penetration level (26,000 paying subs). The initial capital outlay for the off-premises equipment would be $3.5 million.

With on-premises, the cost per sub is minimized since equipment is initially installed for paying subs only. So with 65 percent penetration, the initial capital outlay would be $2.8 million (assuming comparable equipment costs of $110/sub). On-premises offers a savings of $700,000 in initial capital costs over off-premises. However, in either approach, the average cost per sub may be higher, depending on whether or not the equipment is left in place or redeployed when a paying sub disconnects.

The second area of difference is the issue of powering. Off-premises equipment is typically powered from the cable plant. This raises two concerns: First, since the operator pays for the power that the cable plant uses, a substantial increase in annual power costs will occur. Second, at a minimum, the feeder lines of the system will have to be rebuilt (repowered and adding appropriate power passing capability) to accomondate the additional power requirements. In the previous example, 10,000 active devices would be added to the system.

On-premises equipment, on the other hand, is typically powered from the sub's home by a low voltage wall transformer. The power is either added to the coax or run on separate power wires. The advantage here is that the cable system's power bill is unaffected. The potential disadvantage (at least with coax powering) is the possiblility of damaging other consumer electronics equipment connected to the coax.

The third area of difference is deployment. This is directly related to the level of commitment by the operator toward deploying one approach over the other. A much higher level of commitment is required when deploying off-premises, in that significant construction or rebuild activities are required to perform a field trial, let alone a full-scale deployment. These activities are not only costly in terms as dollars and cents but also in distruption of services. Assume, for example, that you are going to field trial 25 off-premises devices. In the best-cast scenario, all 25 devices would be installed on the same feeder line. The installation process will automatically cause service disruptions to all subs on that line-test and non-test subs alike.

On-premises devices offer clear advantages in terms of deployment. Target subs can be selected regardless of where they live on the system. Installation does not affect any other subs, since the equipment is connected to the drop line, as opposed to the feeder.

Today, the on-premises approach has the advantage in cost per subscriber, powering and ease of deployment. However, ongoing engineering design efforts are expected to produce reductions in the cost of base electronics, reduced power consumption and easier system integration. This will then make it possible for these two approaches to be much more competitive in the future.

In-Home Wiring, Problems and Potentials

Walter S. Ciciora, Ph.D.

American Television & Communications
Stamford Connecticut 06902-6732

ABSTRACT

The two other wire based service companies, the power company and the telephone company, have a very different policy regarding the wires that run thru the home. As with every thing, there are advantages and disadvantages to their approach. It is time to review the situation and come to some conclusions.

The Power Company

Today, no one would think of calling the power company to install wiring in their home or to fix the problem when a fuse blows or a circuit breaker trips. The phone book has pages of electrical contractors who will help the unskilled or the intimidated with repairs and installations. Do-it-yourself stores have well stocked aisles of hardware for those who are (or think they are) handy with tools. Books, video tapes, and Saturday home owner seminars provide a degree of education. Procedures and methods are standardized and well accepted. Supplies bear approval stickers. Local building codes cover the legal and accepted ways of doing things. A system of permits and inspectors to enforce the laws is in place.

Another significant difference in policy is that the power company does not attempt to charge for each separate power outlet. You can have as many plugs and lights in your home as you like, but you only pay for the power you use and for having it available at your house.

It may be that the power company is in the most enviable position regarding in-home wiring. If this becomes our conclusion, then the power company model should be our ultimate goal.

The Phone Company

The phone company is a relative new comer to the concept of subscriber ownership of wiring. In fact they were forced into accepting the idea. For decades they resisted. Part of the resistance was a genuine fear that unskilled or even malicious subscribers would damage the telephone network. This would be expensive, cause other customers to complain, and impair public safety. The other part of the telephone company's resistance was the loss of an attractive revenue stream for the rental of in-home hardware. The charges collected for extension phones was no small issue. If subscribers became comfortable with doing their own wiring, they'd probably add their own extension phones. The phone company even had a mechanism for checking from its office to determine how many extension phones you had. Since the bell in rung by capacitively coupled alternating current, the phone company measured the capacitance of your circuit to determine how many ringers you had. This was done automatically at night when you probably wouldn't be using the phone. It took just a few seconds. More knowledgeable subscribers realized that if they disconnected the bell in their self installed extension phones, they could avoid detection. If the telephone repair man had to make a visit, you disconnected and hid your extension phones.

Subscribers had a bit more reverence for the telephone. They recognized it as their link to a doctor, the fire department, or the police. The felt that even if they didn't need to make an emergency call at the moment, their neighbor might. Years of being told to relinquish the party line in an emergency taught subscribers to view the phone line as almost sacred. For some folks, there was the suspicion that since the telephone was such a wonderful communications device, just maybe the phone company could tell what you were doing by listening in on the other end. This has all changed now.

As labor rates continued to increase and as subscribers added extension phones anyway, the phone companies realized that continued control of in-home wiring was no longer attractive or enforceable. The grand old revenue streams from in the home no longer even covered the costs. The phone companies have now turned over the responsibility to the subscriber and made a virtue of having done so. Significant charges accrue to the subscriber who needs his in-home wiring serviced by the phone company. Service has become a profit center in many telephone companies rather than just a cost center. After just one of these expensive service calls, most subscribers will view the monthly option of an in-home wiring service contract as an attractive alternative. In most cases, the contract is more attractive to the phone company since the next service call is likely to not come for years.

The telephone companies worked with the FCC to institute a series of standards. Manufacturers of telephone customer premise equipment must comply with these standards. The products must be registered with the FCC and the designs "type approved". The manufacturer himself certifies that the product is in compliance. The FCC does not verify every design. In addition, there is a "ringer equivalence number" which the subscriber is expected to supply the phone company. Almost certainly the vast majority of subscribers are not even aware of this obligation. Those who are aware, don't treat it seriously. Have you reported all your ringer equivalence numbers? Do you know anyone who has? There does not appear to be an effective means of monitoring subscriber compliance with these standards.

Cable's Differences

The primary difference between the power company, the phone company, and the cable company's wiring is cable's potentially dangerous signal leakage. The rules on Cumulative Leakage Index, CLI, which we all must observe, have serious consequences for an in-home wiring policy. Since cable is an enclosed, self contained spectrum, it is allowed to re-use spectrum normally occupied by other services in the external environment. Two problems occur when the cable system is not completely sealed. Cable signals leak out into the environment and cause interference and environment signals leak into the cable and also cause interference. The interference with other signals in the environment is more serious because some of these frequencies are used for emergency communications and others are used for aircraft navigation and communications.

The Signal Leakage Issue

There appears to be substantial disagreement over the likely contribution of leakage from subscriber installed in-home wiring to the Cumulative Leakage Index, CLI. On the one hand, a vocal group of engineers feels that CLI is almost entirely dominated by leaks in the large signal portion of the plant. This group feel that contributions from drops and in-home wiring won't add up to much and can be ignored. On the other hand, an equally vocal group feels we haven't yet begun to see the full scope of trouble that can come from subscriber in-home wiring. While it is true that many subscribers now do their own wiring, they usually do it knowing their cable company doesn't approve. Therefore they limit their extra outlets to just a few. If subscribers were told that the in-home wiring was their responsibility, more would do it and likely more rooms would be wired. If the hardware used was of low quality, and / or the workmanship was inadequate, leakage would certainly result. Anecdotal evidence from fly-overs indicates that apartment buildings and especially college dorms yield leakages that are not insignificant. While it is not known how many of these incidents are necessary to cause a cable system to fail the CLI, it is certainly true that a "background level" of leakage is created which reduces the tolerance for leaks in the rest of the system.

The Signal Quality Issue

Picture quality is becoming more and more important as consumer electronics hardware evolves. The interest in High Definition Television will only emphasize quality. When we allow the subscriber to do his own wiring, video quality is likely to suffer. Wiring installations that may leak will also suffer ingress. Excessive splitting of the signal will cause noisy pictures. Subscriber installed amplifiers will degrade noise figure, contribute to composite triple beat, cross modulation, and second order effects. Cheap amplifiers may oscillate. The headaches are almost limitless.

CeBus and Smart House

The EIA has been working diligently on a standard for home appliances to communicate with each other. This standard is called the Consumer Electronics interface Bus, CeBus. Communication occurs over four possible media: via radio, over twisted pairs, over coaxial cable, or by infra red links. Fiber may be in the future. Two important issues are raised by this EIA project. Consumers are led to expect that their TV's should be interconnected. Subscriber responsibility for in-home wiring will be expected. Service companies will be calling to sell installation of wires and connectors for CeBus. This must not lead to a CLI hazard or to degraded video quality. Secondly, the CeBus standard intends to distribute in-home video to the television sets connected to the bus. In-home video comes from VCR's, satellite receivers, front door and baby sitting cameras, and graphics generators associated with security systems and home automation systems. Some of the early work involved trapping out a piece of the spectrum for use by the in-home video equipment. Cable involvement in these efforts is mandatory if we are to protect our interests. Specifically, we need to prevent the trapping of our channels to make room for in-home generated video.

A similar but incompatible effort is underway by the National Association of Home Builders. Their project is called "Smart House". The goal is to have new homes pre-wired with an interconnection scheme that would include in-home coaxial cable. Again cable involvement is necessary to insure that cable's interest's are protected. The NCTA Science and Technology Department has representation on the Smart House board. This may not be enough. More cable industry participation would be helpful.

Standards

One possible approach to minimizing problems from subscriber in-home wiring is to establish a set of hardware standards and an approval labeling system. These would be supported by training booklets, video tapes, and seminars. Subscribers who use these resources would likely do a job that minimizes the CLI hazards and protects their image quality. The cable company could reserve the right to disconnect installations which do not measure up.

The Multiple Outlet Issue

Nearly all cable operators charge for multiple outlets. Depending on the amount and the degree of enforcement, this can be a significant bottom line contributor. The financial positives of subscriber responsibility for the in-home wiring will have to out-weigh the loss of multiple outlet income.

There is also a strategic issue. The single most important trend in consumer electronics is the proliferation of TV's and VCR's in the home. These products have become impulse purchases. Multiple receivers in the home are very common and will become more so. If the subscriber only has cable on the set in the principal viewing room, he remains tied to over the air broadcast in the rest of the home. Strategically, it would be best if cable is available on nearly all of the TV receivers and VCR's in his home. This become more important as we face competition from telco and possibly Direct Broadcast Satellite, DBS.

Initiatives

The NCTA Engineering Committee has a new subcommittee on In-Home wiring chaired by Larry Nelson, Executive Vice President of Comm/Scope, Inc.. It's purpose is to explore these and other issues and to reach an industry position on them. In addition, Cable Labs is considering what might be appropriate for it to undertake. Broadly speaking, the differences between these two groups are that the NCTA is generally responsible for regulations and congressional matters while Cable Labs has the funding and full time staff to undertake technological projects which require the expenditure of resources for their accomplishment.

A third group which has the potential to contribute is the Electronic Industries Association, EIA. Cable has had joint efforts with the EIA for at least seven years. The relationships developed with the EIA can be used to reach common goals.

You are encouraged to contact Larry Nelson (704/324-2200) and become actively involved in this important issue.

THE AUTHOR

Dr. Ciciora is Vice President of Technology at American Television &Communications, ATC, in Stamford Connecticut. Walt joined ATC in December of 1982 as Vice President of Research and Development. Prior to that he was with Zenith Electronics Corporation since 1965, He was Director of Sales and Marketing, Cable Products, from 1981 to 1982.

Earlier at Zenith he was Manager, Electronic System Research and Development specializing in Teletext, Videotext and Video Signal Processing with emphasis on digital television technology and ghost canceling for television systems.

He has nine patents issued. He has presented over seventy papers and published about thirty, two of which have received awards from the IEEE. Walt writes a monthly column titled "Ciciora's Page" for Communications Engineering and Design magazine.

He is currently chairman of the National Cable Television Association, NCTA Engineering Committee, Chairman of the Technical Advisory Committee of Cable Labs, and President of the IEEE Consumer Electronics Society. He is a past chairman of the IEEE International Conference on Consumer Electronics. Walt is a Fellow of the IEEE, a Fellow of the Society of Motion Picture and Television Engineers, and a senior member of the Society of Cable Television Engineers. Other memberships include Tau Beta Pi, Eta Kappa Nu, and Beta Gamma Sigma. He served on several industry standard-setting committees. Current interests center on competitive technology, the consumer electronic interface with cable, and HDTV.

Walt received the 1987 NCTA Vanguard Award for Science and Technology .

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 fro 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, and a hope to someday become more active in amateur radio (WB9FPW).

KAUAI: ADVENTURE IN PARADISE

Peter N. Smith
Vice President - Engineering
Rifkin & Associates, Inc.

ABSTRACT

The paper will describe the evolution of Garden Isle Cablevision from acquisition to present day. Special attention will be paid to the use of fiber optics and the innovative use of UHF transmission to solve a difficult off-air problem on the western-most of the Hawaiian islands.

HISTORY

In the late 1960's, a Los Angeles auto parts manufacturer named Ray Derby Sr. owned a vacation home in Kauai on the eastern side of the island. Mr. Derby was unable to receive television at his location due to terrain obstruction and the 90+ mile distance from Honolulu. Many tests were made by a wide variety of people in attempts to receive Honolulu television stations. All tests results indicated only one small accessible area in Kalaheo was suitable. Mr. Derby decided to build a headend on the site in Kalaheo and run a cable to his home. Thus, a cable system was formed.

Apparently Mr. Derby had some dispute with the local utilities and decided to install all cables and equipment underground in ducts and vaults. Originally the vaults were on the side of the road, but due to road widening, some of the vaults now sit in the middle of one of the main roads, which circles approximately 75% of the island. Kaiser Phoenician single ended amplifiers spaced at 220 MHz (approximately 2,000' with 1st generation GIP cable) were installed from Kalaheo and eventually reached Kapaa. This required 60 amplifiers in cascade. By the mid 1980's, the system was carrying 12 channels consisting of 5 Honolulu broadcast stations, 3 pay services, 3 satellite basic services and a bulletin board channel.

PROBLEMS

In 1983, Rifkin and Associates, Inc. became aware that Derby Cable was for sale. We made an offer which was accepted in 1984. Approximately two years later we were able to get the franchise transferred and in April, 1986, the name was changed to Garden Isle Cablevision.

During the time interval required to purchase the system and transfer the franchise, we discovered a few problems with the operation that needed attention:

- Satellite reception in Hawaii is difficult due to low power levels. The Galaxy satellites feed Hawaii with a side beam which is about 6 to 8 db lower than normal in level. Some channels are not available as Satcom FIIIR only sends 12 of 24 channels on the Hawaii beam. Some eastern satellites are not visible due to blockage by the earth.

- Off-air reception is only available at one point on the island. The Honolulu broadcast stations operate in the downtown area and ghosting from buildings is a problem. The path to Kauai requires the signal to pass over a mountain ridge and about 90 miles of water. This results in low signal levels which are more subject to electrical interference which is more prevalent in coastal environments due to salt spray corrosion. This corrosion covers insulators allowing paths for arcing and it degrades connectors, causing them to loosen and arc.

- Underground plant in vaults is not necessarily a problem, but Kauai is the wettest spot on earth, with some areas of the island receiving over 400 inches of rain per year. Some vaults literally have rivers running through them after a rain shower. We opened a splitter once in the trunk and found the inside completely encrusted with salt corrosion. Amazingly, even though the cables were pointed up and submerged in water much of time, the water had not penetrated the cable more than an inch or so. No waterproofing had been used.

- A 60 amp cascade of single ended amplifiers was installed to reach the furthest subscribers. No standby power was used and much of the system was still at 30 volts, requiring many power supplies. This was compounded by a small local power company that experiences frequent outages. Invariably, upon restoration of power after an outage, a large surge would flow through the cable plant, causing fuses and equipment to blow. Finding an outage in a 60 amplifier cascade is very time consuming. System channel capacity was limited to 12 channels.

Along with this were a variety of other technical and operational problems, such as:

- A one-of-a-kind billing system.

- Newest vehicle was 10 years old with well over 100,000 miles and held together with wire and tape.
- A very small office located at one end of the system.

- A program of daily tape playback requiring at least a 3 hour long round trip to the headend.

- Existing manager left two months after purchase.

- Only an unreliable mobile phone system for communication. There were many dead areas where there was no communication possible.

- All Hawaii systems are State regulated by a professional staff that oversees most areas of operation. This system could not meet State minimum technical standards.

- The staff had little formal training due to the distance from most of the industry. Fortunately, this was positively offset by good attitudes and a willingness to work hard.

- Everything costs 20% to 100% more due to shipping and tax, and it sometimes takes 4 to 8 weeks to receive equipment after it is shipped.

All of these items add up to dissatisfied customers. The rate for eight basic channels was $11.95 or $1.50 per channel. In a classic market with no other choices, this is less of a problem, but because Mr. Derby was unwilling to serve the entire island, a second franchise was granted and the other operator offered more channels for the same rate. There is also an overbuild area of about 300 homes.

SOLUTIONS

During the franchise transfer, Garden Isle Cablevision made commitments to the State to fix many of the problems and bring the system into technical compliance. We wanted to add channels to reduce the cost per channel

We were also aware of the customer complaints and, like most operators, we intended to run a quality operation. Therefore, we were committed to solving the problems and only two other items were necessary; money and creativity. This is a stand alone operation that is separately financed, so creativity became the focus.

Satellite Reception

The satellite pictures at time of purchase typically had some impulse noise due to the use of a dual feed 7 meter earth station with 120 ° LNA's. The earth station was realigned along with a realignment of the dual feeds. New LNA's of 50 - 60 ° were installed and impulse noise is now rare. When Galaxy 3 channels were added, a new 7 meter was purchased. The dish cost about $20,000, but it cost $20,000 to ship it and another $10,000 to install it. The look angle to Galaxy 3 is 13.9° elevation, so site planning was very critical.

Channel Capacity

While we wrestled with ways to cut the cascade, we also knew that the single ended amplifiers would have to be replaced. The vaults were custom built and are difficult to work in, so we wanted to find a drop-in upgrade. With some experimenting we discovered that Kaiser Phoenician II modules would fit the old housings and the Phoenician II's could be retrofit with modern chips. These use a single slope compensated pilot AGC system and were not considered a long term solution but did allow the addition of some mid-band channels.

After substantial research and testing we found one amplifier that would fit the vaults, although it was tight. The main trunks have been replaced with dual pilot AGC high gain power double amplifiers. All old line extenders have also been replaced with power doubled units. The main trunk is now capable of 300 MHz operation and the feeder lines are now capable of about 250 MHz. The system is now carrying 5 broadcast, 5 pay, 11 satellite basic channels and an access channel. The basic rate is $16.95 or $1.05 per channel, for a 33% reduction from purchase.

During these upgrades the system was converted to 60 volt to reduce power supply connections. Strategic locations also employ standby power. Much attention has been paid to surge suppression. Outages presently are at about 10% of their previous level. Subscribers can now turn on their television at night with reasonable confidence that they will get pictures.

60 Amp Cascade

The off-air reception had to stay on the south end of the island and we had to continue to provide service to the Kapaa area, so our only choice was to relay the signals with high quality to the middle of the system. We considered several alternatives.

Microwave was difficult because the mountainous terrain required two hops as there is no line of sight from the headend to any portion of the east side of the island. Use of microwave also would involve use of at least one site with difficult access and difficulty in obtaining power. While AML would be less expensive, it is subject to fades because of rain and this is the wettest spot on earth. A repeatered low power AML system did not seem to be the answer. FM microwave was considered but was deemed too expensive.

Consideration was given to adding the off airs to the existing coaxial system using FM to improve quality. The earth stations would be moved to a new site. This option was rejected because of the cost and the continued reliance on a long amplifier cascade with potential outages. We would also have long term capacity problems as we continue to serve subscribers directly from the headend. FM requires at least 12 MHz per channel and the original cable only has about 300 MHz of capacity. This would limit normal capacity to about 240 MHz. A separate coaxial system was rejected because of reliability and cost concerns.

We finally settled on fiber optics using FM because it offered the reliability we desired due to no repeaters and cost was equivalent or less than other options. Our major concern was introducing a relatively new technology into an area far removed from normal support systems. We selected Synchronous FM equipment because of cost, experience, and the reputation the equipment has for being very reliable.

A primary concern with fiber is the mean time to repair. Few people question that it will fail less often, but if it takes many hours to repair, then overall reliability may be equivalent to coaxial systems. The key to quick repair is technical training and reducing the inherent fear of the new technology. For that reason, we selected AT&T LXE cable with six fibers. Using 16 FM channels per fiber, the ultimate capability is 96 channels using the 1300 nanometer band. Use of the 1550 nanometer band could potentially double this. AT&T was selected primarily due to the availability of rotary mechanical splices. We wanted our technicians to install the splices to reduce their fear of the technology. The link distance was 24 kilometers and had 12db of loss upon completion. If we figure .4db per kilometer of loss for the fiber, then 12 splices averaged about 0.2 db each. They were not yet optimized with a TDR or other devices but <u>were</u> installed by our technicians. When the transmitter was turned on, we had excellent pictures at the receive end.

Admittedly, we did have experts from AT&T and Anixter on hand for training and supervision during the initial splicing. However, about 9 months later the fiber had to be cut for rerouting around a new bridge. All the splicing was again done by our technicians and the link loss is still in good shape. The point is that fiber is not to be feared and the same technicians that maintain your coaxial system can maintain your fiber system. With proper prior preparation, a cut fiber can be restored in less than two hours.

The fiber hub was installed at the 38th amplifier location and by reversing the middle amplifiers, the longest cascade was cut from 60 to 22. End of line signal to noise improved by an average of 6db. Obviously, this is part of the overall reliability improvement. We have had no outages on the fiber to date.

Now that the fiber is in and amplifiers have been replaced, we have a clean transportation system. Unfortunately, we have poor off air signals and we fall into the "garbage in-garbage out" syndrome.

<u>Off-Air Problem</u>

Over the years, many sites have been used for off-air reception on Kauai. While certain sites yielded slightly higher signal levels, none of the sites could provide a picture free of ghosts and electrical interference. The ghosts are a problem even in Honolulu and Oceanic Cable, which serves Honolulu, takes a direct feed from the stations to avoid the problem. They also relay these feeds via FM microwave to Mauna Kapu which is a mountain on the west side of Oahu. From there, they use AML to feed to various hub sites. The off-airs exist in excellent shape on the mountain due to the very high quality transmission paths.

Near Oceanic's facility is a Hawaiian Telephone microwave site that relays signal to Kauai. We asked them for a bid to transport two channels and the price was near $10 per subscriber per month. However, we believed that if they could do microwave, we could also do it. We discovered, however, that they have two separate paths, both with frequency and space diversity. They also back it up with an underwater cable. The problem with microwave over a long water path is reflections from the water arriving at the receive antenna out of phase with the main signal and causing fades. A microwave signal has a short wavelength and ocean water can at times look like a mirror to these wave lengths.

Other solutions such as a satellite or underwater were much too expensive. It occurred to us, however, that if high frequencies and low frequencies have problems, that maybe middle frequencies might be suitable. UHF is far enough away from power line frequencies to avoid electrical interference and yet has a long enough wavelength so as not to be as subject to reflections. UHF also has the ability to be focused well to allow an antenna to project a reasonably narrow beam with fairly high gain. Unfortunately, these frequencies are licensed only to broadcasters and there were no previous cases of cable usage of those frequencies (at least legally).

We ran the calculations and it appeared that a 100 watt UHF transmitter with a 12' transmit and an 8' receive dish would offer a good signal to noise. Calculations are shown in Exhibit A. We were unsure however, about reflections and fades. Therefore, we decided to ask the FCC for an STA (Station Temporary Authority). We paid a visit to the FCC and presented our case and were very favorably received, based on our desire to improve customer service.

About a month later we were able to clear channel 38 for use and we began our test two months later. Our input was a video feed from Oceanic and we installed our transmitter in an old building of theirs and the antenna was placed on their tower. From the inception and alignment there have been no multi path problems. The signal has been rock steady since turn on. There are no ghosts, no electrical interference, and no fades. After the tests were complete, we presented evidence of our success to the FCC and were subsequently granted a full license. The license is secondary to any other full power user and, of course, is subject to restrictions concerning harmful interference.

It should be emphasized that this license is unusual because the circumstances were unusual. There was no other economically feasible method and there is not a high probability of interference as this is a remote place without high UHF usage. There are certainly other parts of the US where this method would apply, but UHF congestion would prevent finding a clear channel. Generally, rules preventing adjacent channel and seventh adjacent channel operation due to local oscillator interference will not allow additional channels in many cases. It is very possible that, even with the aforementioned restrictions, there are other places where this technology would apply.

Summary

All of the other aforementioned problems with billing systems, vehicles, etc. have been solved though replacement. All of these actions have reduced service call rates to about 50% of their previous levels. All of the improvements in Garden Isle Cablevision have resulted in much more satisfied subscribers. While money and manpower have certainly played their part, we believe the major improvements have come through creativity and training. Improved customer service is not gained by simply spending more money and hiring more people. It is gained by listening to your subscribers and through training and creativity finding solutions to their concerns.

EXHIBIT A

OAHU TO KAUAI UHF PATH ANALYSIS

1. DETERMINE IF THERE IS LINE OF SIGHT

IF DISTANCE IN MILES IS LESS THAN OR EQUAL TO THE SUM OF THE SQUARE ROOTS OF TWICE THE ANTENNA HEIGHTS IN FEET, THERE IS LINE OF SIGHT.

TRANSMIT ANTENNA HEIGHT	2700 FEET	MAUNA KAPU, OAHU
RECEIVE ANTENNA HEIGHT	459 FEET	KALAHEO, KAUAI
SUM OF SQUARE ROOTS	103.8 MILES	
TRANSMIT COORDINATES	21D 24M 13S NORTH 158D 06M 06S WEST	
RECEIVE COORDINATES	21D 55M 01S NORTH 159D 22M 21S WEST	

USING GREAT CIRCLE CALCULATIONS DISTANCE IS 88.98 MILES (D)

SINCE D IS LESS THAN THE SUM OF TWICE THE SQUARE ROOTS, THERE IS LINE OF SIGHT.

2. CALCULATE PATH LOSS USING LINE OF SIGHT

FORMULA IS $A = 37 + 20\log(D) + 20\log(F)$

	A IS ATTENUATION IN DECIBELS	
	D IS DISTANCE IN MILES	88.98 MILES
	F IS FREQUENCY IN MHZ (CH 38)	615.25 MHZ

CHANNEL 38 AT 88.98 MILES

FREE SPACE ATTENUATION	131.8 DB
MULTIPATH ATTENUATION	20.0 DB
TOTAL ATTENUATION	151.8 DB

3. CALCULATE RECEIVE SIGNAL STRENGTH

	WITH MULTIPATH	WITHOUT MULTIPATH
TRANSMIT POWER 100 WATTS	+ 50.0	50.0 DBM
TRANSMIT ANTENNA GAIN 12' MARK	+ 24.7	24.7 DB
FEEDLINE LOSS	- 1.2	- 1.2 DB
PATH LOSS	- 151.8	- 131.8 DB
RECEIVE ANTENNA GAIN 8' MARK	+ 21.2	21.2 DB
RECEIVE ANTENNA OUTPUT	- 57.1	- 37.1 DBM
CONVERSION TO DBMV	+ 48.8	48.8
	- 8.3	11.7 DBMV

4. CALCULATE SIGNAL TO NOISE

USING PREAMPLIFIER WITH 4 DB NOISE FIGURE

RECEIVE LEVEL	- 8.3	11.7 DBMV
THERMAL NOISE, 75 OHM, 4 MHZ	- 59.3	- 59.3 DBMV
PREAMP NOISE FIGURE	- 4.0	4.0 DB
SIGNAL TO NOISE	47.0	67.0 DB

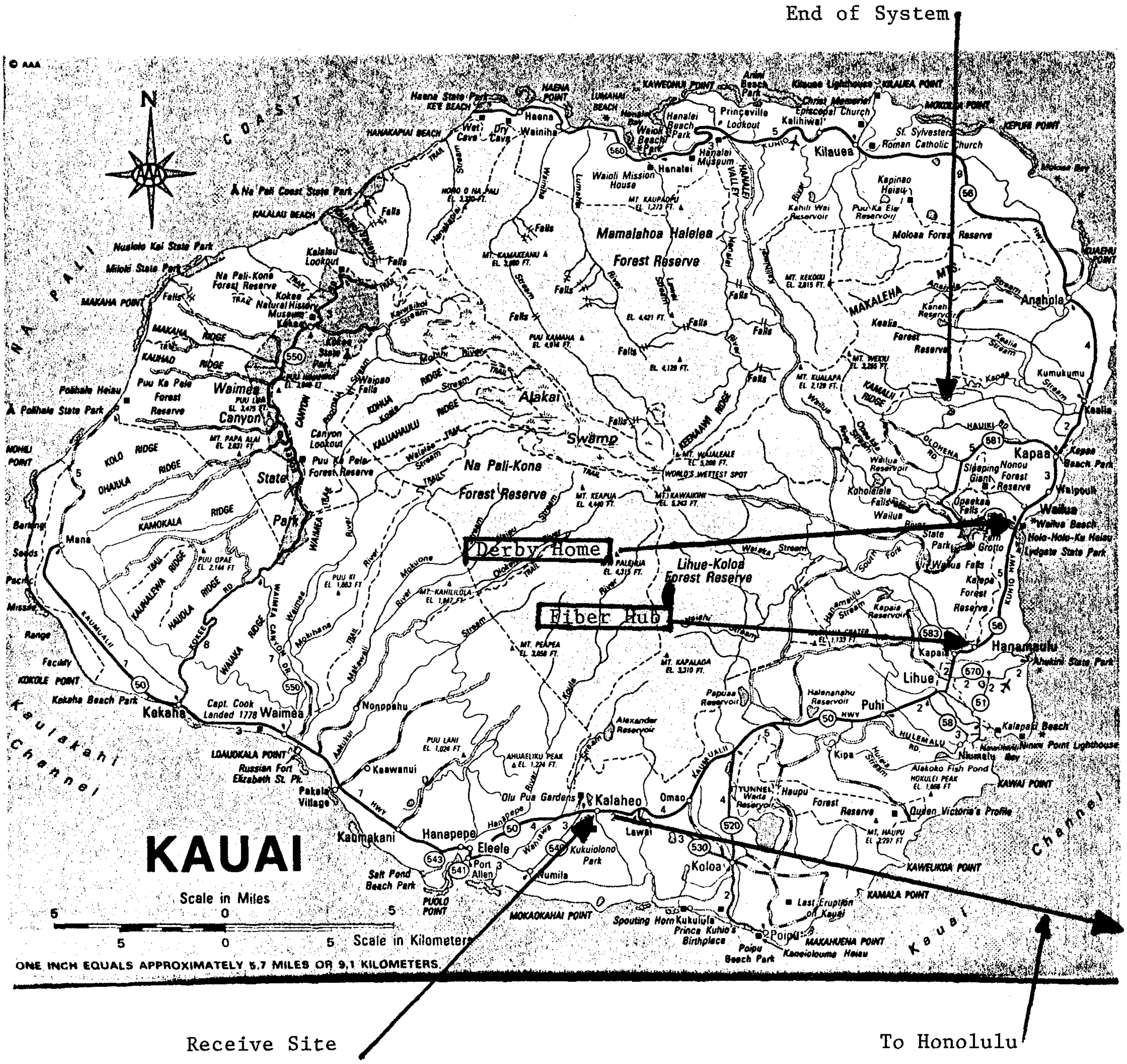
5. SIGNAL TO NOISE AT VARIOUS TRANSMIT POWERS HOLDING ALL OTHER FACTORS CONSTANT.

TRANSMIT WATTS	DBM	WORST S/N(DB)	BEST S/N(DB)
1.0	30.0	27.0	47.0
5.0	37.0	34.0	54.0
10.0	40.0	37.0	57.0
20.0	43.0	40.0	60.0
50.0	47.0	44.0	64.0
100.0	50.0	47.0	67.0
1000.0	60.0	57.0	77.0
10000.0	70.0	67.0	87.0

6. ACTUAL RESULTS

		S/N	
LOWEST LEVEL	-10 DBMV	45.3	VERY RARE
HIGHEST LEVEL	10 DBMV	65.3	
NORMAL LEVEL	6 DBMV	61.3	GREATER THAN 99% OF THE TIME

EXHIBIT B


End of System
Derby Home
Fiber Hub
Receive Site
To Honolulu
KAUAI
Scale in Miles
Scale in Kilometers
ONE INCH EQUALS APPROXIMATELY 5.7 MILES OR 9.1 KILOMETERS
Kaulakahi Channel
Kauai Channel
Mamalahoa Halelea Forest Reserve
Alakai Swamp
Na Pali-Kona Forest Reserve
Lihue-Koloa Forest Reserve
Waimea Canyon State Park
Kekaha
Waimea
Hanapepe
Eleele
Kalaheo
Koloa
Lihue
Kapaa
Hanamaulu
Anahola
Kilauea
Hanalei
Poipu
© AAA

MAINTENANCE AND EMERGENCY RESTORATION OF FIBER OPTIC CABLE SYSTEMS

by Mike Genovese and Becky S. Frye
SIECOR Corporation

Abstract

As the deployment of fiber optics grows in the CATV industry, the need for a comprehensive and effective fiber optic cable system maintenance and restoration plan will become critical. A basic maintenance and restoration plan should include recordkeeping, personnel, training, equipment/materials and emergency restoration procedures. A well-defined plan will improve the operator's ability to swiftly identify faults and act on them, thus restoring service to the customer quickly.

Introduction

The CATV industry has quickly embraced the technology of fiber optics. With this new technology there are new requirements to maintain and ensure the reliability of the CATV operating system. Although the concept of having a maintenance and restoration plan for one's system is not new to the industry, there are unique issues/needs that must be addressed when dealing with a fiber optic cable system.

Fiber optic systems are becoming more economical and practical for CATV applications, but how does the use of fiber optics affect system maintenance and reliability? Fortunately, the CATV industry does not have to wait for their cable systems to age 10 years to get this information. Fiber optic cables have been used in the telecommunications industry for over 12 years and several studies have been made on installed fiber optic cable showing them to have excellent reliability.{3}-{5} Almost all of the cable system failures were due to extrinsic influences like digups, breaking poles, collapsing ducts, etc. One study concluded "that the existing data on the frequency of copper cable cuts in a given service area can be applied to fiber optic cable cuts".{3} This type of logic should be transferable to the CATV industry and can give the operator an indication of the minimum level of reliability for the cable system. In fact, fiber optic cables generally are more rugged than their coaxial cable counterparts and will in all likelihood provide better reliability.

But what about the electronics? We do know that in recent years DFB lasers have been used extensively in telephony digital applications and have shown outstanding reliability. But there is little experience with using lasers in the current CATV AM applications. Recently one lab study focused on the effects of accelerated aging on DFB lasers. The study indicated that the DFB lasers currently used in AM fiber systems can provide the performance and reliability needed by the CATV industry for a median life of 25 years.{1} If this is indeed true, then the biggest reliability issues for the company's fiber optic system are uncontrollable extrinsic failures like pole breaks, digups, etc. Given this assumption, having a well defined system maintenance/restoration plan could significantly improve a system's reliability.

First, the operating company must decide what level of maintenance and restoration capability to have. The company may want to maintain and restore their fiber optic system themselves, or may wish to contract out certain portions to contractors who specialize in fiber optics. This decision is a philosophical one as well as economic. For the rest of this paper we will assume that the CATV company will have a complete in-house capability.

There are two phases to consider with regards to maintenance and reliability of a fiber optic cable system: Pre-installation and Post-installation.

PRE-INSTALLATION

Although the system's equipment and materials are not always considered a part of the maintenance plan, the selection of system products can have a long-term effect. The operator should choose products that meet industry standards to ensure that the installed system will perform consistently over its expected life and provide the needed reliability. Standards for optical fiber, cables, and associated products have been developed and used in the telecommunications industry for over 10 years. Most of these standards are available and should be considered by the CATV industry.

While performance and cost are usually considered the key parameters of planning a fiber optic system, the system design can have a major impact on the system reliability. With today's watchwords being "customer service"," reliability and quality", it is imperative to design CATV systems that enhance reliability. Such features could include advanced monitoring and control systems, backup coaxial systems/AML systems (usually already installed), redundant optical cable routes and 1 x N protection with optical switches and electronics. Once the system is designed and installed there are other factors that need be addressed by the operator and come under the guise of "The System's Maintenance/ Restoration Plan".

POST-INSTALLATION

A CATV system's "Maintenance/Restoration Plan" should address five key elements: records, personnel, equipment/materials, training and restoration procedures. {2} We will examine each individually.

RECORDS

Complete and accurate records are invaluable for the troubleshooting and restoration of an operating system. It is important that the records be functionally organized and located in a designated place for easy access. Duplicate records should be kept in an alternate place to prevent accidental loss of this valuable data. Typical fiber optic system records should include:

1) Route/cable plan with cable feet or meter marks. Be sure to record any places where excess cable slack is stored. This will be a factor when using an OTDR to determine the geographical location of a cable fault.

2) Splice plan detailing fiber assignments, splice locations and restoration priorities.

3) Splice loss data

4) Transmitter output levels

5) End-to-End system attenuation

6) OTDR signature traces

For AM fiber systems that are to be hard-wired (without optical connectors) into the electronics the OTDR signature trace can also provide system attenuation.

7) Checks and calibration of test and troubleshooting equipment.

PERSONNEL

The operator must designate which personnel are responsible for maintaining and more importantly, troubleshooting and restoring the fiber optic system. As with any new technology there will be gaps of knowledge and experience within the company. Properly trained personnel should be given specific task responsibility for testing, troubleshooting and restoration of the fiber optic system. In addition to these responsibilities, the designated personnel should have responsibility for maintaining the emergency restoration materials and equipment, as the primary users of it. Also, the maintenance plan should have provisions for how designated personnel will be notified when an emergency occurs.

It is important to note that it is very easy to end up with an elite "fiber team" in the company. Although this may be a functional necessity at first, in the long run, it will be to the operator's advantage to cross-train as many personnel as possible in fiber optics to give flexibility to scheduling and emergency restoration task assignment. Thus, training of the personnel is the next critical link in the maintenance/restoration plan.

TRAINING

The initial training required by a restoration team begins with ensuring each team member practices proper safety procedures. During a time when speed and accuracy are critical, it is also very important that team members remain injury-free. System repair begins with the identification and location of the system failure. To do this, team members must learn the skills necessary to operate equipment such as OTDRs and power meters.

Once located, the fault must be repaired. This may involve replacing a damaged connector, replacing a link of cable or anything in between. So the team must be well versed in all possible solutions. Proper cable handling procedures, hardware preparation, mechanical splicing and fusion splicing are vital and necessary skills in the event a system repair is needed. A maintenance/restoration plan isn't worth the paper its written on, without well trained personnel to implement it. It is recommended that the initial classroom training include some basic theory as well as extensive hands-on training.

Cross training of personnel is also important. The ability to perform more than one task will prove beneficial in the event that one team member is not available when an emergency occurs.

Refresher training is equally important and should not be overlooked. As personnel changes occur, the replacement personnel will require the same sequence of training to ensure a smooth transition of responsibilities as well as stability of the system. Refresher training ensures that old skills remain sharp and new skills also are implemented. Refresher training for emergency restoration is best done with a combination of classroom training and mock restoration drills once every six months.

EQUIPMENT/MATERIALS

Everyone knows how difficult it is to perform a job without the proper tools —— with an emergency restoration this difficulty increases ten-fold. But having the proper tools is just part of it, having the tools readily accessible is just as important. One method for accomplishing both of these objectives is to have an Emergency Restoration Kit (ERK)™ contain all the needed tools and materials for repairing an optical cable system at the restoration site. A typical kit should include:

1) Optical Cable: The restoration cable should have the same fiber count as the highest fiber count cable in the optical cable system. It should also be of sufficient length to span the longest pole span distance in the system, plus slack for splicing, (typically add 100 feet). If the system contains ducts or manholes, the restoration cable should be of sufficient length to span the largest distance between manholes, plus slack for splicing, (typically 75 feet). With buried cable you have more flexibility in locating the new splice points. The cable needs to be of sufficient length to span the damaged area, plus slack for splicing.

Note: Keep in mind that, certain types of depressed clad single-mode optical fiber are susceptible to modal noise if the two adjacent splice points are less than 20 meters apart.{6} Consult your optical cable manufacturer to determine if this is a concern for your system.

Typically, there is a restoration cable reel that is used to supply the cable for the kit. The restoration cable reel, should contain about 1000 - 2000 feet of optical cable. The restoration cable reel should be centrally located and clearly marked so that it is not installed into another portion of the operator's system by mistake.

2) Splice Closures: Two closures should be included. The operator should choose splice closures that have sufficient capacity for the restoration cable, are easy to splice in, seal and reenter.

3) Splice Trays: There should be a sufficient number of single-mode splice trays placed in each splice closure to route the buffer tubes of the restoration cable. The operator may wish to consider using splice trays that are designed for both mechanical and fusion splices. This way the transition from the temporary mechani-

cal splice to the permanent fusion splice will be much easier.

4) Mechanical Splices: The splice should be fast and easy to use. Typically mechanical splices that accept cleaved fibers and require no epoxy or polishing are the fastest and easiest to install. The goal is to get optical continuity as quickly as possible while achieving a reasonably low splice loss. The advantage of the mechanical splice is that the restoration cable can be stripped, loaded into the splice trays and have the splice installed on each fiber prior to an emergency ever occurring.

5) Optical Power Meters: At least one optical power meter should be included in the kit so that the output power of the damaged cable can be measured to ensure that all the damaged cable has been cutout. Inexpensive, handheld power meters are readily available that can measure both 1300 and 1550 nm. To measure the output power of a bare fiber the power meter should be equipped with a bare fiber adaptor.

6) Optical Fiber Cleavers: The kit should contain two fiber cleavers, one for each restoration crew member. The cleavers should be handheld, rugged and require no external power supplies.

7) Miscellaneous Tools/Materials: This would include items such as: sheath knifes, isopropyl alcohol, tissues, electrical tape, tywraps, silicone RTV, etc.

The restoration cable should be fully prepped and installed into the two splice closures. The buffer tubes should be routed to the splice trays and the optical fibers should be prepared and installed into one side of the mechanical splice parts. This preparatory work will save about 1.5 man-hours of on-site restoration work and will enable the operator to bring up the fiber optic system that much faster. Restoration kits are available from several optical cable manufacturers including Siecor.

There are other tools/equipment needed for troubleshooting the fiber optic system from the headend or hub:

* Optical Time Domain Reflectometer (OTDR): This is a very versatile piece of equipment that operates on the same principle as an electrical TDR, but is used only on optical fiber. The OTDR can measure splice loss, detect system faults, measure system length and estimate the system's link loss. This is an essential piece of equipment for any operator having a fiber optic system.

* Optical Power Meter: As described before, this piece of equipment measures optical power and is used to measure the output power of the Laser Transmitters.

RESTORATION PROCEDURES

This is where the fun starts. Its 1:30 PM on Super Bowl Sunday. The headend technician has just discovered that the status monitor system indicates a loss of optical and backup coaxial transmission to the hub that just happens to be the largest one in the system and provides service to the city mayor. What should happen?

Step 1: Commence Troubleshooting: The headend technician should inspect the appropriate system electronics to see if the failure has occurred there. If the RF electronics/laser indicates it is operating properly, the next step is to inspect the optical cable system.

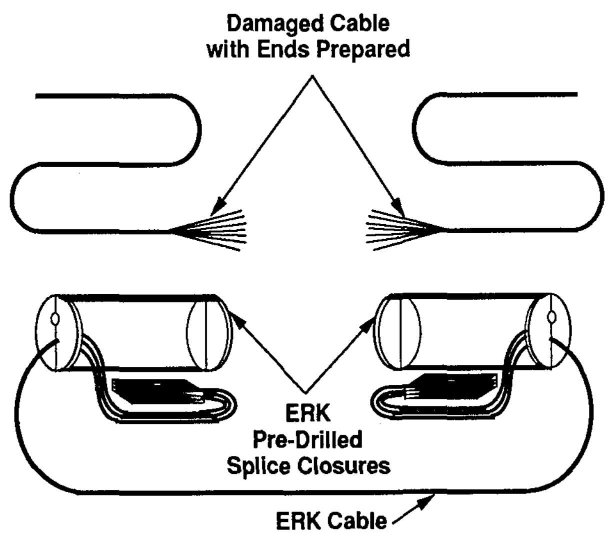
Step 2: Test Cable System: The technician should take the OTDR located in the headend and obtain an OTDR trace of the optical fiber. The trace should be compared to the original signature trace to see if any abnormalities exist. Let us assume that the OTDR indicates a fiber fault. The OTDR will give an approximate distance from the headend to the fault. By comparing the distance given by the OTDR to the cable route diagrams, the technician can estimate the geographic location of the fault. At this point, the headend technician should notify the designated emergency restoration team of the cable system fault and its approximate location.

Step 3: Site Evaluation: The emergency restoration team arrives on site with its restoration kit, safety equipment, access ladders, etc. They find that a little old lady, from Pasenda, with a 1969 Cadillac has knocked down a large tree and although the lady and the Cadillac are fine the aerial coaxial and fiber optic cable have been severed. Its now 3:30 and the Super Bowl starts in less than 3 hours.

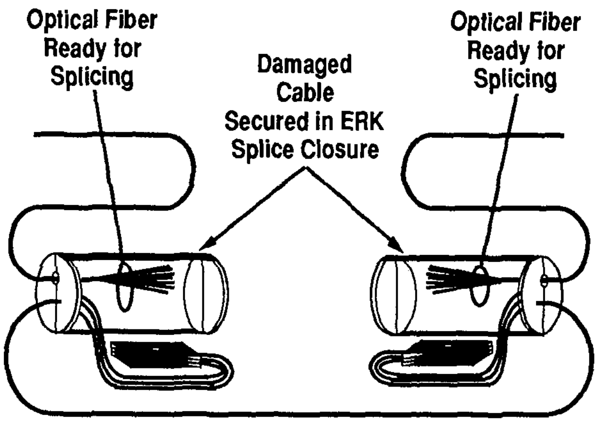
Step 4: Cable Restoration: The restoration team should inspect the damaged cable and cutout the damaged section plus an additional 10 feet on each side in case there are any offset breaks. In this case, the team decides to run their restoration cable along the ground next to the street until the broken steel messenger is replaced.

The following is a sequence of cable restoration using a restoration kit.

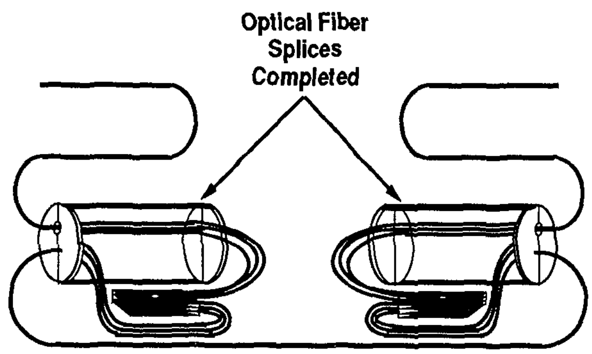
Prepare the damaged cable ends for installation into the splice closures.



Route the buffer tubes of the damaged cable into the splice trays. Measure the received optical power of the damaged cable end coming from the headend to ensure fiber continuity.



Consult the emergency restoration plan and mechanically splice the high priority fibers



first. Check with the headend technician to see if the status monitor alarm has cleared. Once cleared, continue splicing the remaining fibers. The splice closures can be temporarily sealed until the permanent fusion splicing can take place.

Its now 5:50 P.M. and the Super Bowl starts in ten minutes, and your customers are celebrating the quick return of their CATV service.

SUMMARY

As the use of fiber optics gathers wider acceptance and application in the CATV system, the importance of a maintenance/restoration plan will also grow. The 12 plus years of fiber optic experience in the telecommunications industry indicates that fiber optic cable systems are very reliable and that most system failures can be traced to uncontrollable extrinsic failures. By having a well defined maintenance/restoration plan, the operator will have the ability to swiftly identify system faults and act on them, thus quickly restoring service to the customer.

References

(1) Flynn, E.J., et al.,"Performance and Reliability of Laser Devices for CATV." SCTE Proceedings March 1990: 167.

(2) "Maintenance/Restoration Training Course." Siecor, June 1989.

(3) Szentesi, Otto I."Reliability of Optical Fibers, Cables, and Splices." IEEE J. Select. Areas Commun., vol. SAC-4, No. 9 December 1986: 1502 - 1508

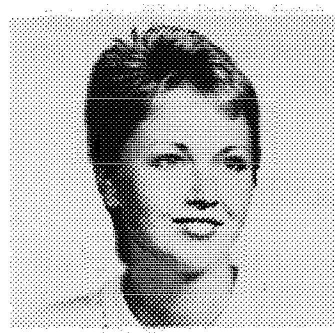
(4) Kiss, L. "Performance of Optical Fiber Cables and Operations and Maintenance Procedures in the Australian Telecommunications Network." Intl. Wire & Cable Symp. Proceedings March 1983: 212

(5) Ishio, H. "Japanese field trials and applications in telephony." IEEE J. Select. Areas Commun., vol SAC-1 April 1983: 404-412

6) Sear, F. M., et al., "Probability of Modal Noise in Single-mode Lightguide Systems." I00C-EC0C 1985: 823-826

Mike Genovese

Mike received a B.Sc. degree in Mechanical Engineering from The Pennsylvania State Univ., State College, PA in 1978 and was commissioned as a Naval Officer in the Nuclear Submarine Service. In 1984, he joined Siecor where has held positions as Project Engineer, Field Engineering Manager and is currently Product Manager for the CATV and Utility market.



Rebecca S. Frye

Rebecca is Training Supervisor at Siecor Corporation, where she is currently responsible for the development, coordination and implementation of all Technical Training. She holds both B.S. and M.A. degrees in Mathematics with concentrations in Education, Physics and Psychology. A member of SCTE, Ms. Frye is actively involved in the development of courses to further education the CATV industry in use and applications of fiber optic technology.

Nd:YAG Light Source and External Modulator for A VSB-AM Video Transmitter: A Performance Evaluation

Y. Trisno and D. Grubb III
General Instrument Corporation
Jerrold Communication Division

ABSTRACT

A miniature Nd:YAG laser and an intensity modulator based VSB-AM video signal transmitter is described. With a light source having RIN level below -170 dB/Hz, the system allows a relatively low modulation index to be used. For a 40 channel system using a standard MZ, a CTB= 60 dB corresponds to m_i= 1%. With 75 mW Nd:YAG laser source, a system with power budget of 5.5 dB and 50 dB CNR is realizable. Several promising techniques of modulator linearization are also given.

INTRODUCTION

The development of optical transmitter for VSB-AM video transmission has been mostly based on direct modulation of DFB lasers. An alternative design approach is to use an external modulator and the relatively high power Nd:YAG laser as the CW optical source. In this way, the stringent linearity requirement on the laser can be removed.

This paper presents the analysis of an optical VSB-AM video transmitter based on the Nd:YAG laser and the intensity modulator. It is divided into four separate sections which are outlined next. We begin by describing the characteristics of a Nd:YAG laser as an optical source for this application. The analysis of the intensity modulator will then follow. After describing the components of the transmitter, the system level performance is analyzed. Finally, different techniques of modulator linearization to improve the CTB performance are described.

LIGHT SOURCE

While the source linearity is no longer an issue in this transmitter design, many of the laser parameters such as noise, optical spectrum, and linewidth together with the available power are still critical. For the Nd:YAG laser measurement of the aforementioned parameters was performed. First, the noise component originating from the laser unit is addressed. Then the effect of the optical spectrum on the amount of power coupled into the fiber will be discussed.

The RIN is measured under three different measurement conditions. First a back to back (with attenuator and minimum amount of back reflection) measurement is done from 50 MHz to 1GHz. Under this condition, the laser RIN is measured better than -173 dB/Hz which is our measurement system limit.

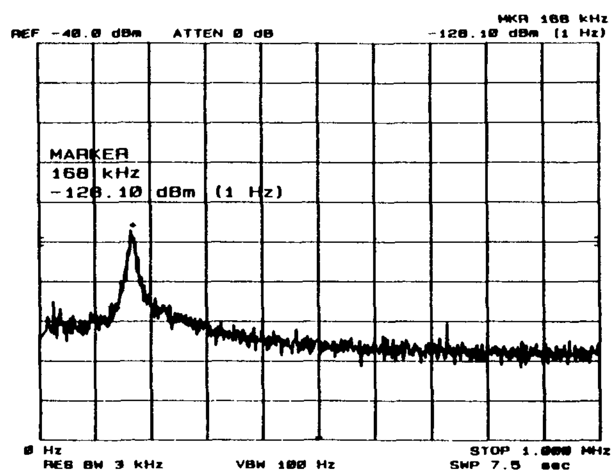


Figure 1. The relaxation oscillation of the Nd:YAG laser unit used in this experiment

Next a 35 kM link of standard SM fiber is used. In this case no degradation in RIN is measurable with negligible external back reflection (BR). However, with the presence of BR laser RIN degrades non uniformly. The worst measured value is -162 dB/Hz. Based on this result an optical isolator is necessary for this application. There is also no mode partition noise observed in the last measurement.

The second noise source arises from the relaxation oscillation frequency (ROF) of Nd:YAG laser system which occurs at around 170 kHz for the unit on hand, Figure 1. After RF modulation, this ROF frequency is shifted up to the RF domain and is located at ± 170 kHz around the modulating RF carrier. Figure 2 shows the shifted ROF around a modulating carrier at frequency of 250 MHz. The relative level of the carrier to the noise is ≈65 dBc. It is clear from the figure that the translated ROF will be an additional noise rather than an addition to CTB (the spectrum beneath the carrier) level.

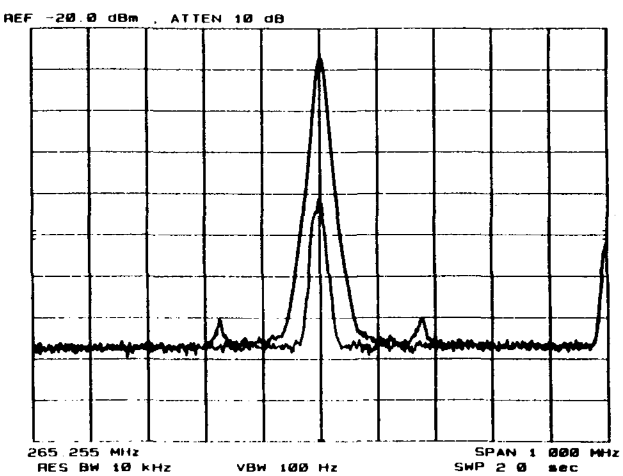


Figure 2. The up converted laser ROF
after being modulated by a tone at 250MHz

Further measurements were performed for different modulation index and different frequencies (450 MHz, m= 0.3). The result is shown in Figure 3. The relative level of the carrier and the noise stays relatively constant [1]. At this level, this noise is not expected to affect the picture quality in an VSB-AM video transmission application. If it is absolutely required to remove this type of noise, an optoelectronic feedback loop could be used to cancel it [2].

In the diode pumped laser, relatively high level of current is involved to drive the pump laser diode, therefore it is necessary to reduce the excess electronic noise from the power supply unit. Otherwise, the noise could propagate into the Nd:YAG laser output if it falls into the proper frequency range. Figure 4 shows the measurement result of this type of excess noise. As shown in the figure, it will contribute significantly to the CTB. However, this type of noise is not a fundamental phenomena and is easily corrected.

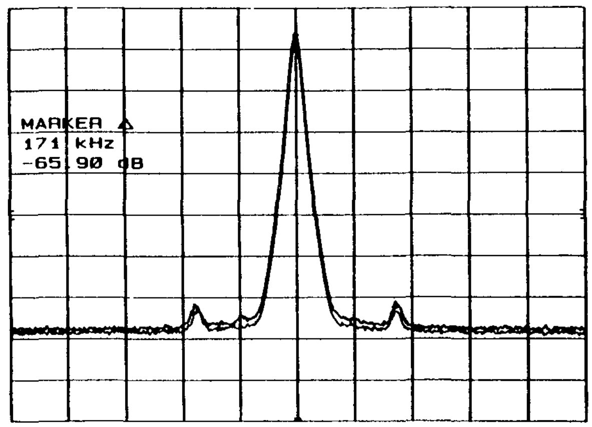


Figure 3. A similar measurement as in
in the Fig. 2 for m= 0.3 and F= 450MHz

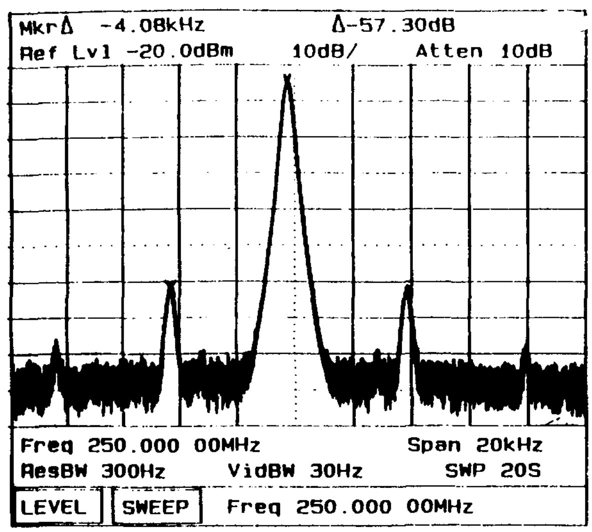


Figure 4. A low frequency excess noise
originated from the pump laser diode circuitry

One of the potential advantages of using Nd:YAG laser is to extend the available power budget for the AM link. Currently the available power ranges from 50 to 200 mW. For long distance (greater than 20KM) application the optical spectrum of the laser becomes important to avoid nonlinear SBS threshold of the fiber [3]. Among the Nd:YAG products available, that with a single line at 1319 nM with several longitudinal modes (total spectrum of ≈4Å) is the optimum candidate.

In summary, a Nd:YAG laser is an attractive choice of light source for high power VSB-AM video transmission system. It has many of the desirable characteristics sought for this application. Unfortunately, an external modulator is required to impress the video information. Therefore a linear optical intensity modulator must be developed. This is the focus of the next section.

INTENSITY MODULATOR

There are many different categories of optical intensity modulators [4] to [6]. At present the Lithium Niobate based modulators are readily obtainable as advanced R&D product from few different suppliers. Consequently this section mainly focusses on the Lithium Niobate based modulator. In particular, the performance of a Mach Zehnder (MZ) type modulator will be analyzed.

A MZ intensity modulator is basically an implementation of a Mach Zehnder interferometer in a planar wave guide form. The optical output power as a function of the modulating carrier voltages can be expressed as in equation (1).

$$P_o = P_b \left[1 + Cos \left\{ \phi_i + \phi_b + \frac{\pi}{V_\pi} \cdot \sum_{i=1}^{N} m_i V_i \, Sin(\omega_i t + \theta_i) \right\} \right] \qquad (1)$$

Instead of directly attempting to analyze the distortion components from equation (1), it is simpler and more elegant to work using single carrier modulation. Based on the known relation between the composite distortion and the harmonic, the expression for CTB will then be derived. For linear analog applications, a MZ modulator is biased at the quadrature point of its response curve. If this point is stable, the first two arguments in the cosine function will be equal to integer multiple of $\pm\pi/2$ depending on the bias point selected. Performing trigonometric expansion on equation (1) and then writing the result in term of the first three components of its series approximation yields the next two equations.

The fundamental component is:

$$\left[KV_m + \frac{(KV_m)^3}{8} \right] \cdot Sin \ \omega t \qquad (2)$$

The third harmonic component is:

$$\frac{(KV_m)^3}{24} \cdot Sin 3 \omega t \qquad (3)$$

where

$$K = \frac{\pi}{V_\pi}$$

$$V_m = m \cdot V_p$$

The parameters K and V_m are the device and modulation index dependent variables respectively. To convert equation (3) into a CTB expression two additional variables must be known. The first is the relation of the third harmonic to the CTB. It can be shown that each CTB is 15.56 dB higher than the third harmonic. The other is the number of triple beats that fall onto the particular frequency being monitored. The result is equation (4) [7].

$$CTB|_{Freq} = 20 \cdot \left(\frac{Fundamental}{3^{rd} \ Harmonic} \right) + 10 \cdot Log(N) + 15.56 \quad (dBc) \qquad (4)$$

where N is the number of triple beats that fall onto the desired frequency.

In practice, the quadrature point may not be stationary due to several drifting mechanisms in the device. This drift causes an increase in the second order distortion. The permissible level of drift depends on the CSO being specified. Further, a large drift will also contribute to some change in the fundamental amplitude. These relations can be seen clearly from the following set of equations, which are derived from the full series expansion of equation (1).

As before the expansion is also based on a tone modulation and carried out up to third order components.

The fundamental component is:

$$\frac{K_m V \cdot Sin\,\overline{\Phi}\left[-1 + \frac{3}{8}\cdot(K_m V)^2 - \frac{5}{96}\cdot(K_m V)^4\right]}{1 + Cos\,\overline{\Phi}\cdot\left[1 - \frac{1}{2}\cdot(K_m V)^2 + \frac{3}{32}\cdot(K_m V)^4\right]} \quad (5)$$

The second harmonic component is:

$$\frac{\frac{1}{4}\,(K_m V)^2\cdot Cos\,\overline{\Phi}\left[1 - \frac{1}{3}\cdot(K_m V)^2\right]}{1 + Cos\,\overline{\Phi}\cdot\left[1 - \frac{1}{2}\cdot(K_m V)^2 + \frac{3}{32}\cdot(K_m V)^4\right]} \quad (6)$$

The third harmonic component is:

$$\frac{\frac{1}{24}\,(K_m V)^3\cdot Sin\,\overline{\Phi}\left[-1 + \frac{5}{16}\cdot(K_m V)^2\right]}{1 + Cos\,\overline{\Phi}\cdot\left[1 - \frac{1}{2}\cdot(K_m V)^2 + \frac{3}{32}\cdot(K_m V)^4\right]} \quad (7)$$

where

$$\overline{\Phi} = \Phi_i + \Phi_b$$

When the angle ϕ is equal to $\pi/2$, the above set of equations reduce to the previous equation set (2) to (3). To compute CTB equation (4) is still valid provided it is replaced by equation (5) and (7) respectively.

Before proceeding to present the theoretical plot of MZ distortion as a function of modulation index, we would like to discuss the experimental characterization of a MZ. The parameters of interest at this state of evaluation are the third order distortion performance and the second order distortion vs bias point. For a CTB measurement, the MZ is biased in the neighborhood of its quadrature point. The modulation index of the modulating RF signal is set by a computer controlled RF attenuator. Figure 5 shows a typical plot of the CTB versus modulation index. The abscissa represents the individual modulation index assuming the carrier has a Gaussian distribution [3]. A comparison between the experimental result and the theoretical calculation is displayed in Figure 6 for 40 channel carriers. From

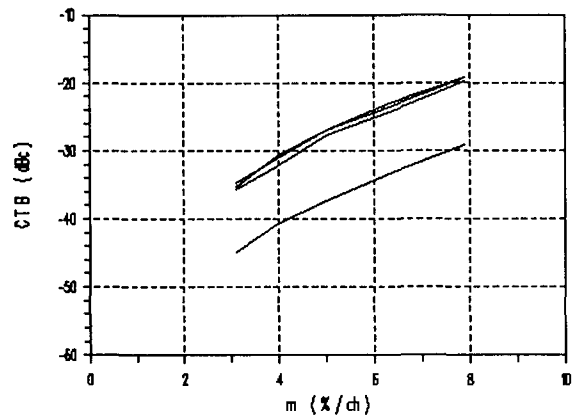


Figure 5. A typical CTB level of a MZ under 40 channel modulation

the figure it is apparent that good agreement is achieved between them. This eliminates the need of multi carrier generator requirement to perform this measurement. Therefore much simpler experimental set up can be used.

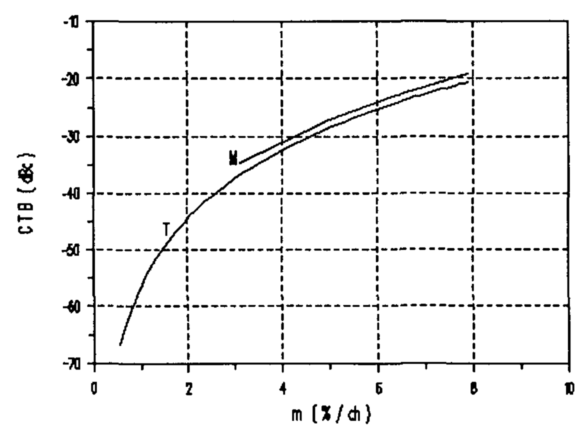


Figure 6. Comparing the theoretical calculation (T) and measurement (M) values of CTB vs m.

Measurement of CSO is done rather differently. For this measurement the RF attenuator is set at a fixed value. The MZ bias level is swept between a given voltage range. The result is shown in Figure 7. The upper graph is the carrier and the lower one is the composite second order components at a given frequency. Each pair of graphs (carrier and distortion) represents a specific frequency band being tested. The bias voltage is stepped in 10 mV increments. The best CSO is approximately 65 dBc which can be seen directly from the plot. For this case, the bias drift tolerance is in the order of 20 mV.

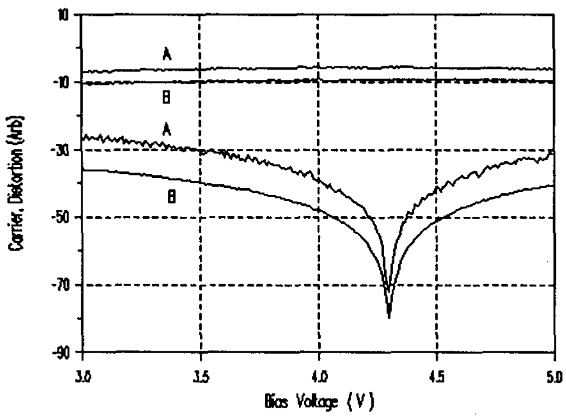


Figure 7. The second order distortion versus bias voltage of a MZ

SYSTEM LEVEL PERFORMANCE

In this section the various parameters of both the light source and the modulator which are related to the system performance are detailed. Specifically, the influence of the RIN, the modulation index, the CTB, and the system power budget on the system performance will be considered. Finally, an example of system design calculation based on two graphs is presented.

Assuming the MZ is perfectly linear, then for modulation index exceeding a certain threshold, the only distortion components are due to the non linear clipping induced distortion (CID) at both lower and upper portion of a MZ transfer function. For single clipping region and further assuming large number of channels (greater than 10) Saleh [8] and Grubb and Trisno [3] show the threshold of the RMS modulation index is 0.246. This number corresponds to a 5.5 % modulation index for each channel in a 40 carriers transmission system. For the MZ case, the RMS modulation index will be 3 dB lower.

The central issue of the analysis is to find the relation between the CNR and the CTB for a power budget, laser output power, and laser RIN. Figure 8 is a plot of the CNR versus power budget. It is calculated based on the parameters given below. The laser output power is 75 mW. The laser RIN is 170 dB/Hz. The modulator parameters are given as follows. The fiber to fiber insertion loss is 6 dB. The experiment uses 40 channel modulating carriers. A standard MZ modulator device is used. The receiver thermal noise is 7 pA/√Hz. Each pair of graphs corresponds to a specific modulation index. The upper curve of each pair represents the shot noise limit for that particular modulation index. Graph A, B, and C correspond to modulation index of 4, 3, and 1 % per channel respectively.

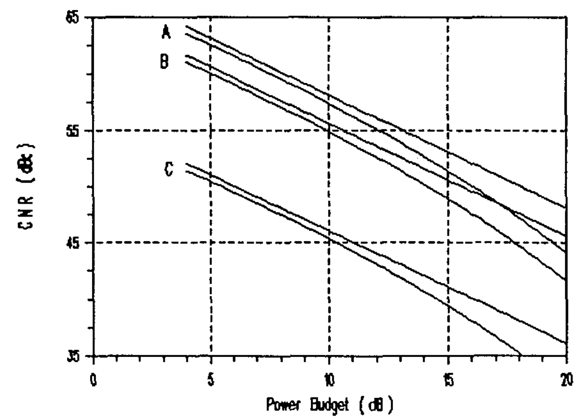


Figure 8. The CNR plot as a function of system power budget for different m.

Using the above graphs and the distortion (CTB) versus modulation index plot (Figure 6), the desired system performance can be easily estimated. For example, a system using a standard MZ and Nd:YAG source having parameters mentioned above, m= 1% (Figure 6) is required to reach a CTB level of 60 dB. Using Figure 8, a CNR= 50 dB corresponds to a 5.5 dB system power budget. For a linearized modulator, the pair of graphs can still be used provided the CTB versus m plot is updated for the corresponding condition.

MODULATOR LINEARIZATION

Comparing the measured (also the theoretical) third order distortion to the generic VSB-AM system specification, some forms of modulator linearization will be required if more efficient device utilization is desired. Fortunately, many compensation techniques can be used because of the repeatability of the MZ response. Basically, there two different approaches. First is the use of electronic techniques such as feed forward and predistortion methods. The other is optical compensation approach. In this section each of these methods based on published experimental data are reviewed.

Johnson and Roussell of Lincoln lab [9] showed that by adjusting the amount of TE and TM mode content in the MZ and setting the proper biasing for each of the modes for optimum cancellation, a respectable 20 dB of compensation can be obtained. While their result was derived from rather low frequency measurements, it is believed that similar result can be readily obtained with measurement in the CATV frequency band. One potential cause of system performance degradation is during the transmission over long fiber. The fact that the TM mode is a slightly higher order mode than the TE mode implies that the TM mode will be more susceptible to bending loss. The actual effect due to this potential problem must be measured experimentally.

Another simple linearization techniques is to use an electronic predistortion circuit. Childs and O'Byrne [1] have used a pair of diode strings to generate the required level of predistorted signal to compensate the MZ response curve. Their measurement is done under modulated video conditions. For 3.3% of modulation index per channel (of 50 channels), a 60 dBc CTB level is achieved. This corresponds to a 16 dB improvement from the standard MZ curve.

Still another approach is to employ a hybrid ooptoelectronic feed forward network. Ridder and Korotky [10] have experimented with one version of this approach. Basically, it uses a pair of identical MZ to first generate the distortion component which will then be used to compensate the second MZ. The CTB improvement in this case is in the order of 15 dB across the 50 to 300 MHz range, the same order of magnitude of improvement as in the previous case.

In summary, many different approaches of third order distortion compensation can be done for the MZ modulator. By combining some of this techniques, a better compensation level could be obtained. To reach the optimum modulation index operation, another 10 to 15 dB improvement is required.

SUMMARY

In this paper components making up a VSB-AM transmitter based on the Nd:YAG laser and the intensity modulator have been described. Additionally, various trade offs among parameters which influence the system level performance were discussed. Finally, a comparison among different techniques of modulator linearization have been presented. The key to the success of this type of transmitter is the availability of linear or linearized intensity modulators.

ACKNOWLEDGEMENT

The authors wish to thank Hank Achor for valuable discussion and his time for proof reading this paper.

REFERENCES

1. R.B. Childs and V.A. O'Byrne, "50 Channel VSB-AM Transmission Employing a Linearized External Modulator," Proc. OFC 1990, Jan 22 - 26 1990, San Francisco, CA, paper PDP-23.

2. T.J. Kane, "Diode Pumped Laser as Sources for Fiber and Space Communications," Proc. 1989 LEOS Ann. Meeting, Oct 17 - 20, 1989, Orlando, Fl, paper ELT3.4.

3. D. Grubb III and Y. Trisno, "AM Fiber Optic Trunks - A Noise and Distortion Analysis," Proc. NCTA 1989, May 21 - 24 1989, Dallas, TX.

4. R.V. Johnson, "Recent Developments in Integrated Optics at Crystal Technology," Proc. Integrated Optics and Optoelectronis, Sept 5 - 8 1989, Boston, MA, SPIE Vol. 1177, pp 327-33.

5. L. D. Hutcheson, ed., Integrated Optical Circuit and Components, 1987, ch. 7, pp 169-228.

6. T. Tamir, ed., Integrated Optics, 2nd ed., Springer Verlag, NY 1985

7. Y. Trisno, "A Brief Summary on Ti:$LiNbO_3$ Mach Zehnder Modulator Devices Characterization," Unpublished internal report.

8. A.A. Saleh, "Fundamental Limit on Number of Channels in Subscarrier Multiplexed Lightwave CATV System," Elect. Lett., vol 25, no. 12, pp 776-777, June 1989.

9. L.M. Johnson and H.V. Roussell, "Reduction of Intermodulation Distortion in Interferometric Optical Modulators," Optics Lett, vol. 13, no. 10, pp 928-30, Oct 1988.

10. R.M. Ridder and S. Korotky, "Feedforward Compensation of Integrated Modulator Distortion," in Proc. OFC 1990, San Franscisco, CA, Jan 22-26 1990, paper WH5.

NEW MICROWAVE AND FIBER OPTIC SUPERTRUNKING SYSTEM CONFIGURATIONS

By T. M. Straus, R. T. Hsu, and L. A. Kaufman
Hughes Aircraft Company

ABSTRACT

Rapidly changing fiber optic and AML microwave technologies have opened up a new range of possibilities in implementing CATV supertrunks. These considerations apply not only to new builds, but also to many existing AML microwave systems whose reach could be extended. The same technologies can be applied to improve system reliability either through redundancy configurations or by dividing overly long microwave paths into smaller segments. Combinations of the two technologies are explored. System performance improvements possible with higher power transmitters and on-frequency repeaters are delineated. Advantages of configurations employing fiber to AML, AML to fiber, "low power" channelized AML with high power microwave repeaters, and fiber in parallel with AML are set forth.

INTRODUCTION

Within the last few years, extraordinary changes have occurred in both fiber optic and microwave technologies which are applicable to CATV transportation systems. At a time when there is increased concern with the quality and reliability of TV pictures available to the customer, these changes have tended, at least in the near term, to strengthen the hand of the advocates of delivery of multiple VSBAM signals all the way from the head-end to the home. Thus, although FM has been used in supertrunking applications, whether by cable, microwave, or fiber, this paper will restrict itself to AM systems which are not encumbered by the need to reprocess each channel with the attendant complexity, cost, and maintainability considerations.

The advances in fiber technology have been well publicized. Starting with the demonstration of a 40 channel, 6.5 km, AM fiber link at the 1987 Western Cable TV Convention, the spotlight of interest and attention both in the trade press and at the shows has been firmly on fiber. Deservedly so! The achievement of close to 50 dB C/N through direct modulation of Fabry-Perot diode lasers with a standard 40-channel VHF TV format[1] was an important milestone. It further stimulated development of the fiber backbone concept[2] and was followed within a year by the introduction of Distributed Feedback (DFB) lasers with 40-channel C/N capability of 52 dB at distances of 12 or more kilometers[3]. Improvements in power output, Relative Intensity Noise (RIN), and linearity of DFB lasers have most recently led to reports of 56 dB C/N with 6 dB optical loss[4]. These are indeed spectacular accomplishments.

The recent advances in microwave technology are in some respects equally significant to the cable industry even though the publicity spotlight has not been focussed on them. One could hardly expect that it would be otherwise since AML Microwave® is a technology which has been widely applied within the industry for almost 20 years. For many, the familiar and ubiquitous "high power" and "low power" channelized AML systems were the only yardsticks against which fiber system cost and performance could be measured. Although recent technology development has led to an all solid state high power channelized AML transmitter [5], with which 60 dB C/N microwave links can be supported, the greater part of recent microwave technology advances have been evidenced by low-cost broadband block conversion types of transmitters and active on-frequency repeaters. These latter equipment categories are similar to AM fiber transmitters in that power output, at given channel loading, is limited by CTB and CSO performance. The first generation of such transmitters was capable of only -9 dBm/channel output for 40-channel loading and 65 dB C/CTB.[6] However, within four years, a more than sixty-fold increase in power was made possible through the application of microwave feedforward technology[7] and the introduction of a higher power FET amplifier and power doubling within the feedforward loop[8]. The combined noise and distortion performance is equivalent to that of a 60 watt low noise FET power amplifier if such a device were to exist at 13 GHz. Rigorous CATV system requirements are indeed conducive to the development of state-of-the-art microwave and fiber performance.

AM FIBER LINK PARAMETERS

Table 1 updates a similar summary published in 1989[9]. The changes reflect the increased laser output power, the lower RIN, and a higher per-channel modulation index, m, due to improved laser linearity. The net effect is that quantum (shot) noise dominates as before for optical loss in excess of 6 dB, but the overall link C/N is now much closer to the objectives established for fiber backbone systems. Significant additional improvement may be possible with still greater optical power output (7 dBm has been mentioned) and better RIN and linearity, but even the parameters assumed in the table have yet to be verified in production quantities.

The higher the modulation index, the greater the need for stable input level since the laser performance will degrade rapidly if the voltage swings below the laser threshold voltage. Another factor to consider is that DFB laser linearity typically degrades as the modulation frequency increases.[10] For 2-laser 80-channel systems, this is counterbalanced by avoidance of in-band second order distortion in the upper 40-channel grouping. Indeed, the second order distortion has typically dominated to such a degree that asymmetric frequency plans which minimize in-band second order are common in multi-laser systems. As with distortion, DFB laser RIN degrades with frequency so that 40-channel test C/N results are not necessarily indicative of what can be achieved between 300 and 550 MHz. Another factor which can play a limiting role at such excellent laser RIN is multiple reflection on the fiber link resulting in conversion of laser phase noise to AM noise[11]. The lower frequency channels could be affected most strongly when this phenomenon is present.

Recent advances in external modulation type fiber optic links[12,13] have renewed interest in this form of optical communication. A high power solid state Nd:YAG laser provides the optical carrier, which is then intensity modulated by a $LiNbO_3$ Mach-Zehnder interferometer. The modulator transfer characteristic is of the form

$$I = I_o/2\ [1 - \cos\ (\theta_o + \frac{\pi}{V_\pi}\ v(t))]$$

where V_π is the half-wave switching voltage and θ_o is set to $\pi/2$ by the dc bias voltage so as to avoid second order distortion. Since the transfer function is clearly not linear, some form of linearity correction circuit is required to avoid excessive third order distortion at even a modest modulation index. Moreover, in addition to the 3 dB loss intrinsic to the bias condition, the $Ti{:}LiNbO_3$ waveguide material is lossy and the match between the optical field pattern in the fiber and the modulator is imperfect. Finally, there are losses associated with focussing the available laser power into the fiber so that the overall optical power available to the fiber link is typically

TABLE 1
ASSUMED 40-CHANNEL DFB LASER LINK PARAMETERS

Optical Loss (dB)	C/N Source (dB)	C/N Quantum (dB)	C/N Receiver (dB)	C/N Link (dB)
2	59.9	63.2	77.6	58.2
4	59.9	61.2	73.6	57.4
6	59.9	59.2	69.6	56.3
8	59.9	57.2	65.6	54.9
10	59.9	55.2	61.6	53.2

Transmitter:

m = 5%

RIN = -155 dBc/Hz

P_{Laser} = 4 mW (into fiber after isolater)

Receiver:

Responsivity, R = 0.85 A/W

Noise equivalent current, i_N = 5 pA/$\sqrt{Hz}$

attenuated by 12-13 dB relative to the laser output. This may still be improved by 2-3 dB.

Table 2 models a 40-channel fiber link based on 15 dB cancellation of third order distortion and a requirement for 65 dB C/CTB. The transmitter output power is consistent with reference (12). We have also confirmed that laser RIN is on the order of -170 dBc/Hz. The measurement was made by plotting total link noise output versus receiver photo current. At high receiver input level, shot noise dominates and one can, therefore, clearly establish the RIN from the deviation of the total noise from pure shot noise.

One advantage of the Nd:YAG laser is that its line width is much narrower than that of a DFB diode laser. Consequently, the effect of optical reflections on link C/N should be much less. On the other hand, there remain serious questions related to the laser life. The Nd:YAG laser is "pumped" by an array of semiconductor lasers operating at high current. It is the long-term reliability of this optical pump source which still needs to be established. A secondary issue relates to power limitation due to the narrow spectral linewidth which could result in Brillouin scattering in the glass fiber at higher optical powers. A further property of the YAG laser is a low-level relaxation oscillation which modulates its output. In our measurements, we found this to result in 175 kHz sidebands at -62 dBc on either side of the VHF TV carrier. Although this does not effect the C/N, it can limit the achievable baseband S/N at high optical receiver input levels.

Another key issue relating to external modulation is long-term stability. Drift in the properties of the $LiNbO_3$ material could result in serious degradation of both second and third order distortion. At the very least, some form of active control must be established to maintain the modulator at its optimum bias point.

Comparison of Tables 1 and 2, for which both CTB and CSO are assumed to be 65 dB, shows that if the external modulation approach can overcome the above problems, it will offer slightly better C/N for optical loss under 10 dB. It must show a bigger advantage to overcome the cost burden of its more complicated transmitter. The most promising avenue would be increase of the modulation index made possible by further linearization of the transmitter.

AML MICROWAVE LINK PARAMETERS

For maximum link distance and optimum C/N, a high power solid-state channelized transmitter is

TABLE 2

ASSUMED 40-CHANNEL EXTERNAL MODULATION FIBER LINK

Optical Loss (dB)	C/N Source (dB)	C/N Quantum (dB)	C/N Receiver (dB)	C/N Link (dB)
4	69.9	59.8	76.5	59.3
6	69.9	57.8	72.5	57.4
8	69.9	55.8	68.5	55.4
10	69.9	53.8	64.5	53.3

Transmitter:

m = 2.8%

RIN = -170 dBc/Hz

P_{Trans} = 10 mW (into fiber after modulator)

Receiver:

R = 0.85 A/W

i_N = 5 pA/$\sqrt{Hz}$

currently utilized. The output power capabilities of these transmitter arrays are only 1 to 4 dB less than high power klystron-based units. The overwhelming advantages of greatly reduced power consumption and size of solid-state units when combined with the by now proven high reliability of the 5-watt FET amplifiers has, for the most part, turned the design choice for the highest power systems in their favor. A 40-channel system can span more than 32 km in each of 8 directions while providing 60 dB C/N and better than 65 dB CSO and CTB. The performance advantage of such microwave systems over AM fiber is further compounded by the fact that the 32 km is a line-of-sight distance which for a typical suburban fiber run might stretch to over 40 km.

A closer challenge to microwave system capability occurs if the system is restricted to low-cost block conversion transmitters. With this constraint, a 56 dB C/N can be obtained at a line-of-sight distance of 19.2 km. This still represents a sizable advantage over the best fiber system.

Recent developments in broadband microwave power output capability can best be summarized by Figure 1 which shows the relative capabilities of three microwave active on-frequency repeaters[14]. The initial 10 dB jump in power capability was due to the introduction of feedforward technology. While widely used within the cable industry, this technology had not previously been applied at such high frequency (13 GHz). The further 8 dB increase was due to utilization of higher power (5W) FET amplifiers as well as power doubling within the feedforward loop. Overall performance of the IBBR-124 is summarized in Table 3. Note that the combined noise figure and gain imply a noise power output of -51.5 dBm/4 MHz, which in turn results in a 61 dB C/N contribution from the repeater for 40 channel loading at 65 dB C/CTB. Comparison to the two earlier repeaters shows that the IBBR-124 has respectively 3 and 7 dB better C/N, and 5 and 10 dB higher gain, than the FFR-123 and the MWB-122 for the same channel loading and C/CTB. Thus, in addition to the very large increase in power output capability, the latest microwave repeater technology offers substantial improvement in C/N and gain. These parameters are critical to achieving significantly better overall microwave link performance on extended supertrunk paths.

SUPERTRUNK CONSIDERATIONS

As with any other portion of the CATV system, cost, performance, and reliability are the considerations which determine the design of the supertrunking system. The choice of optimum design is a complex problem, not only because these three parameters are closely interrelated, but also because trade-offs can be made in allocating distortion budgets between the supertrunk and the other parts of the CATV system. In particular, it has been

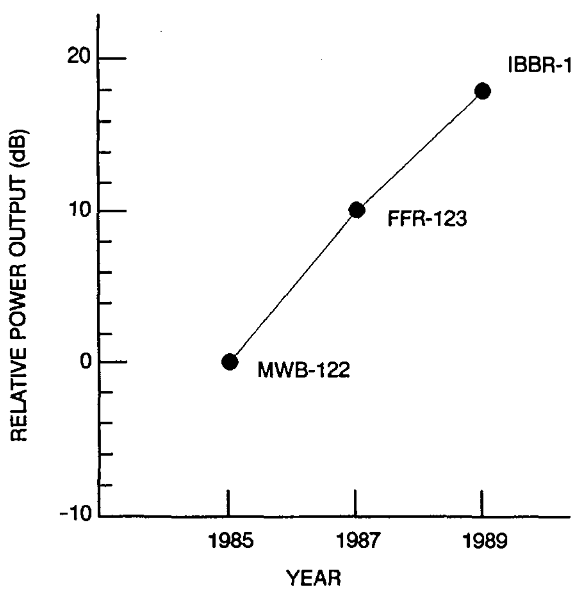


Figure 1 Microwave Repeater Development.

TABLE 3
IBBR-124 PERFORMANCE SUMMARY

Noise Figure	6.5 dB
Gain	50 dB
Capability at 65 dB C/CTB	
Channel Loading	Output Power/Channel (dBm)
12	15.5
21	13.5
35	10.5
60	7.5
80	5.5

pointed out[15] that in some smaller system designs it is more cost effective to allocate a larger portion of the distortion budget to the supertrunk than the traditional approach of demanding that it be essentially "transparent". In addition, there is the question of how the distortion on the supertrunk will add to the distortion in the cable trunk and distribution. The fact that composite triple beat need not add on a voltage basis is now documented[16],[17] and discussed further below.

In any case, the supertrunk services all or a very large portion of the cable system customers. For customer satisfaction, extended or repeated outages cannot be tolerated. With the increased emphasis on quality of service, it is mandatory to closely consider all aspects which affect service reliability. If the system is one that already exists, substantial cost saving may be possible with an upgrade which takes advantage of the latest technology without sacrificing improved reliability and performance goals. Extensions to the reach of the supertrunk should also, in many cases, be possible using either fiber, microwave repeaters, or a combination of the two.

Cost

Any discussion of cost comparisons is tremendously complicated by the large variability of this parameter. Microwave transmitter costs can vary from less than thirty thousand dollars for the lowest power broadband equipment, to over a million dollars for an 80-channel high power channelized transmitter. Other cost factors affecting microwave systems include questions relating to the necessity for towers, transmit and receive site property rentals, if required, and variations in antenna size and receiver noise figure. In terms of potential coverage area, and therefore, the number of customers who are serviced through the supertrunking system, the most cost effective systems are the large ones. On the other hand, the lower cost broadband systems lend themselves to greater flexibility in matching capability to limited numbers of receive sites at various distances from the transmitter.

With fiber, the greatest cost variability arises not from the electronics cost but from the cable length and installation costs. Pole attachment or duct utilization fees must also be included in the calculation, if applicable. Overlash onto existing plant and new underground construction in urban areas represent the extremes of installation costs. In any case, the longer the total cable length the less favorable the cost will be relative to microwave. The variability in the cost factors is so great that each case must be separately analyzed. However, even for the lowest construction cost situation, a microwave alternative will generally be more cost effective if the sum of cable lengths (i.e. supertrunk paths) exceeds 10 miles.

Performance

As previously stated, FM is excluded from consideration in this discussion because of the increased cost and complexity of reprocessing each channel at each of the hub sites. However, a special case exists when the satellite receive antennas are located at a considerable distance from the principal headend. In such instances it is possible to transport the FM signals via a fiber optic link after they have been downconverted to the 950-1450 MHz band, but before the second conversion and demodulation. If the system utilizes AML microwave after reprocessing the signals at the headend, one has, in effect, an FM fiber supertrunk feeding a VSBAM microwave supertrunk. As an alternate to the fiber, it is also possible that these FM signals could be transmitted via AML microwave. The FM portion of the supertrunk should, in such cases, be essentially transparent to the S/N established by the satellite down-link.

The performance of several microwave/fiber systems is summarized in Table 4. Although the combination of microwave and fiber is most probable in the context of an integrated AML/fiber backbone system, this need not necessarily be the case. It is possible that the fiber serves only to extend the reach of the supertrunk. Generally the link performance would be better if a direct line-of-sight were possible between the microwave transmitter and ultimate hub site, but in severe climate zones such as Central Florida this may not be the case. In other cases, the desired hub site is simply not compatible with a clear path, but an alternate microwave receiver site is available at some modest distance from the preferred site. In such cases, addition of the fiber supertrunk extension makes possible the addition of a receive site which might otherwise be

TABLE 4
40-CHANNEL INTEGRATED MICROWAVE/FIBER SYSTEMS

	AML Transmitter Type			
System Parameters	SSTX-145	MTX-132	IBBT-116	IBBT-116/ IBBR-124
Power Out, P_0/Channel, dBm	16	9	8.4	9
Number of Outputs at P_0	8	8	1	1
Path Length, km$^{(1)}$	30.4	21.6	19.2	34.4
C/N, dB	58	58	56	55
C/CTB, dB	71	71	65	61
Fiber Tail Length, km	18	18	16	10
Supertrunk C/N	52	52	52	52
C/CTB	64	64	62	60
Effective Reach, km$^{(2)}$	58	46	41	55

(1) Path calculations assume 10-foot antennas, 4-dB total transmit and receive waveguide loss, average multipath and rain (CCIR, Zone D2, and 1 hour/year fade below 35 dB C/N, except for the last column which is based on 1.5 hours/year.

(2) Microwave line-of-sight length x 1.3 plus maximum fiber backbone length.

burdened by the appendage of an excessively long coax cable trunk run.

With increased channel loading, the performance will, of course, degrade. However, in a recent laboratory experiment designed to investigate addition of CTB between dissimilar devices it was found that with 80-channel loading the combination of an IBBT-116 transmitter and an IBBR-124 repeater, each operating at +7 dBm/channel output, with a COR-299 receiver adjusted for -44 dBm AGC threshold, resulted in the expected 53 dB C/N but with C/CTB ranging between 59.2 and 64.2 dB across the frequency band. To partially explain these rather good results, it should be noted that a 3 dB CTB margin relative to published specifications is required in factory test of the IBBT and IBBR to allow for some drift in performance with temperature. Even so, to explain the measured CTB one cannot stick with voltage addition.

More particularly, when a standard CATV hybrid amplifier was added to the chain at the microwave receiver output and its output level adjusted so as to generate 61 dB C/CTB at the highest channel (547.25 MHz), the combined microwave plus CATV hybrid C/CTB was 60.9 dB. The microwave system by itself also measured 61 dB C/CTB on this channel. Therefore, at this frequency, for this particular pair of subsystems (i.e. the complete microwave system and the hybrid), the CTB added with an effective phase angle of 119°. "Normal" voltage addition is based on 0°. As frequency decreased, the phase angle gradually dropped below 90°. At the lowest frequency channels where the worst system CTB of 58.5 dB was measured, the hybrid CTB was so good that experimental error made it difficult to determine an exact phase angle, although it was clearly smallest (closer to voltage addition) at this end of the spectrum. On average, a 90° phase angle (power addition) was most descriptive of the CTB addition.

That this should be the case with dissimilar devices, particularly distortion cancellation devices such as the feed forward circuits within the IBBT and IBBR, should not be surprising. In this particular case, no attempt was made to pretune these circuits beyond the standard factory procedure. However, in a separate experiment[16] it was shown that tuning of the feedforward circuit does effect the phase angle. One point which remains to be investigated is the possibility of deliberately tuning the circuit for best overall C/CTB while simulating the rest of the cable system with an overdriven hybrid. The key question to be answered is that of long term stability. If it turns out that this is a successful technique, significant benefit would accrue to the microwave feedforward system since signal levels could be increased for better C/N and path reliability.

The above-described CTB addition measurement system was next modified by the substitution of an experimental 12-km fiber optic link for the CATV hybrid. The experimental arrangement is shown in Figure 2. With 40-channel loading, the IBBT and IBBR output levels were raised to +10 dBm and the receiver AGC threshold set for -43 dBm. The microwave and fiber systems were separately characterized and then combined for the system measurement. Results are tabulated in Table 5. The system C/N was better than 52 dB. The point to be made is not that this represents the best that can be achieved - the optical link parameters were not as good as in Table 1 and its composite second order distortion was only 58 dB - but rather that the microwave and fiber CTB addition was even more favorable than power addition. The overall favorable addition is generally similar to that reported[16] with a totally different pair of microwave and fiber systems.

Supertrunk performance can be improved by reducing channel loading. In particular, many fiber optic systems utilize this technique to avoid the generation of otherwise limiting second order products. Similar techniques can be applied at microwave. For instance, in a 60-channel system, the channels could be equally divided among 3 block conversion transmitters so that the output power can be raised by 5 dB. Each transmitter is then connected to a separate antenna but all three antennas are aimed at the receive site where all 60

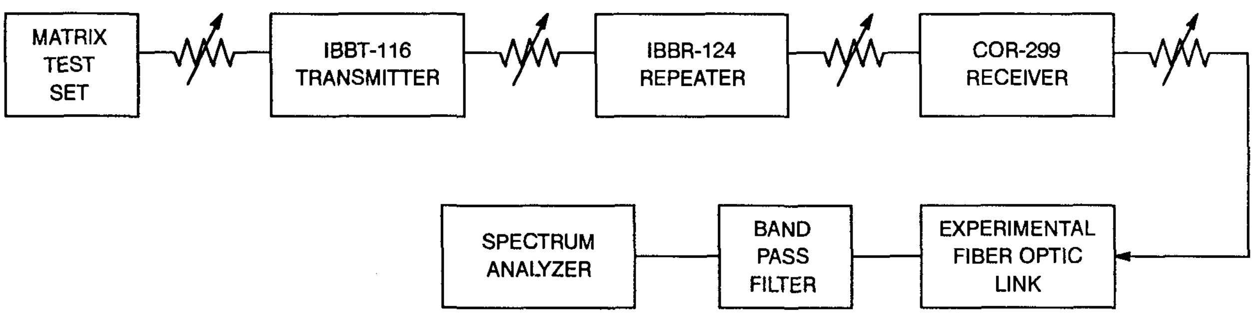


Figure 2 CTB Addition Experimental Setup.

TABLE 5

CTB ADDITION OF MICROWAVE AND EXPERIMENTAL FIBER LINK

Frequency (MHz)	Microwave CTB (dB)	Fiber CTB (dB)	System CTB (dB)	Effective Phase Angle (degrees)
61.25	64.7	67.3	63.5	99
193.25	65.4	66.0	64.9	114
307.25	63.5	66.5	63.9	115

channels are extracted from a single receiver. In the fiber system, 3 separate optical receivers would be required, and unless wavelength division multiplexing is used, 3 glass fibers are utilized. The only other difference is that the microwave transmitters must be locked to the same reference oscillator.

When multiple receive sites are involved, the technique of paralleling transmitters lends itself naturally to the addition of a combining network which simultaneously acts as a splitting network. Figure 3 shows such an IBBT-116 transmitter array. Here each of the 9 outputs carry all 60 channels, and thus each of the 9 transmit antennas would be trained at a separate receive site. Directional couplers are utilized to tap off power to the shorter paths. The transmitter contribution to both C/N and C/CTB is not any different than with the above described space combination mode, but with the multiple receive sites, transmit antennas need not be duplicated to obtain the benefit of improved output power. A further advantage is that this configuration lends itself to a graceful degradation redundancy

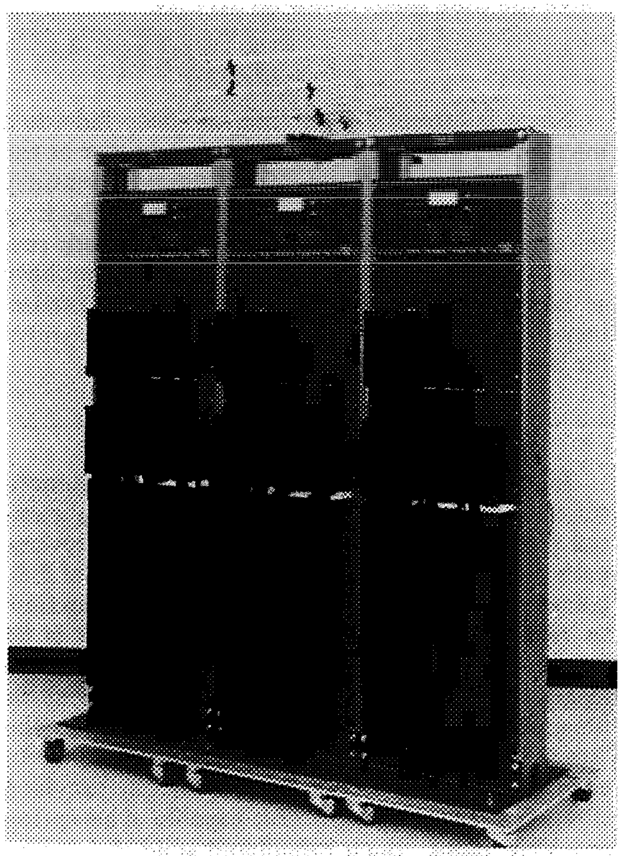


Figure 3 Fail-soft redundant IBBT-116 Transmitter Array.

mode. In the event one of the transmitters should fail, the VHF input to that transmitter is redistributed to the remaining two units such that each would operate at 30-channel loading. This feature would not be available in fiber systems which rely on filters to remove out-of-band distortion and noise at the output of each optical receiver.

Performance of existing microwave supertrunks can often be upgraded at relatively low cost. In klystron-based high power AML systems, replacement of the solid state source with a higher power source allows 3 dB increase in output power and consequent improvement in both C/N and path reliability. Regardless of the transmitter type, if the received signal is weak, addition of a tower-mounted LNA can substantially improve both C/N and path reliability. However, there is nothing new in either of these options.

What is new is that installation of a broadband repeater can now substantially improve performance of a MTX-132 transmitter link. Consider the example given in Table 6. Many older systems have expanded by adding receive sites which in some cases were more distant than originally intended. An LNA was then added to the receive site to obtain the best possible noise figure. Even so, predicted path reliability is only 99.88% for the 25-mile example. As before, an average multipath and rain condition is assumed. The table shows the result of interposing an IBBR-124 repeater at an intermediate distance. The repeater must be located such that the direct ray from the transmitter to the distant receiver is attenuated at least 45 dB through antenna angle discrimination. In this example, path reliability is improved to 99.97% and the normal C/N can be raised to 55 dB.

Microwave repeaters are particularly attractive as extenders of supertrunk reach when the supertrunk can be allocated a large share of the total CATV noise and distortion budget. In one such lightly populated region located in a benevolent B1 rain zone, a 4-hop 36-channel microwave system spans a total distance of 78 miles. Predicted end of line performance is only 49.7 dB C/N and 54 dB C/CTB (cw) with 99.83% path reliability. However, the broadband system which includes lesser length branches and a total of 15 receive sites at various intermediate points was the most economic solution for providing CATV service to the extended community.

TABLE 6
25-MILE MICROWAVE PATH UPGRADE

	Without Repeater	With IBBR-124
MTX-132 Output 40 Channels (dBm/ch)	9	9
First Hop Distance, Miles	25	10.8
Repeater Output in AGC (dBm/ch)	N.A.	8.1
Second Hop Distance, Miles	N.A.	14.2
LNA Input, dBm	-45.7	-40.7
System C/N, dB	53	55
System C/CTB, dB	77	65
Hours/Year Below 35 dB C/N	10.3	2.3

Reliability

Overall communication link availability depends on both the electronics equipment and on the intervening path. In the case of fiber, rain and multipath fades are not a problem, but the cable connection can nevertheless fail. The failure can be due to either natural causes or man-made. The latter category includes both accident and intentional sabotage. Whatever the cause, and however seldom a break occurs, the time to restore service can be quite lengthy. The news has provided numerous examples of horrendous outage situations in the communications industry. For this reason ring fiber architectures have been proposed and in some cases are being implemented. The primary drawback is one of cost.

In contrast to path failures in fiber links, CARS band microwave fading is a relatively common occurence, especially in areas of high rainfall rates. Path availability predictions are generally based on fading to a 35 dB C/N, at which point the pictures are noticeably noisy but still watchable. Deep fades beyond this point are generally of quite short duration and the link usually restores itself within a few minutes.

The only way to protect against such fading is to provide additional link margin through higher transmitter power, lower receiver noise figure, and reduced waveguide loss. The use of active repeaters to increase link availability has also been illustrated in the preceding section. The efficacy of such measures depends on a number of factors including whether the fade is due to rain or multipath. In many cases a mere 3 dB greater fade margin will halve the time spent below 35 dB C/N.

Since failure of a fiber link and deep fade on a microwave path are highly unlikely to occur at the same time, one method of providing essentially 100% reliability would be to use the one to back up the other. Such a solution could be particularly cost effective since the temporary back-up need not have as high a quality as the primary link. A further benefit of this solution is that protection is provided for both the path and the electronics.

In situations where parallel paths are not practical, redundant electronics can still make a substantial difference in overall reliability. A microwave path designed to have less than 1 hour/year of fade below 35 dB C/N makes little sense if the mean time to repair an unprotected electronic failure is 24 hours. As a minimum, adequate spares should be locally available. Fail-soft configurations such as illustrated by Figure 3 are attractive since the "spare" is fully utilized during normal operation. In channelized transmitters, back up can be provided through a frequency agile unit capable of accepting any VHF input. Such broadband solid state units have recently been developed for both "low" and "high" power AML.

Duplication of electronics to protect against broadband equipment failure is possible with both fiber and microwave systems. Automatic switching in the event of a failure, such as provided in the microwave receiver redundancy unit (RRU), is also possible.

CONCLUSION

Recent advances in both fiber and microwave technology have enlarged the design choices available to CATV supertrunk designers. While the highest performance systems still require channelized microwave, utilization of broadband fiber and microwave links in various combinations can provide attractive, cost effective solutions with good performance and reliability. The development of high power, low noise broadband microwave repeaters provides the means for extending supertrunk reach and improving path reliability in some existing systems.

REFERENCES

(1) F. Dawson, "Skeptics Become Fiber Optic Believers as Breakthroughs Demystify Technology," Cable Vision, Dec 21, 1987, p. 47.

(2) J. A. Chiddix and D. M. Pangrac, "Fiber Backbone: A Proposal for an Evolutionary CATV Network Architecture," NCTA Technical Papers, 1988.

(3) CED - In the News, "Anixter Intros Fiber Product: ATC Test Called a Success," Oct 1988.

(4) Multichannel News, "Jerrold Achieves Major Breakthrough with AM Laser," Dec 1989.

(5) T. M. Straus et al, "Solid State High Power AML" CCTA Technical Record, 1988 (Halifax)

(6) T. M. Straus, "Tradeoffs in Multichannel Microwave System Design", NCTA Technical Papers, 1985

(7) T. M. Straus and R. T. Hsu, "Application of Feedforward to CARS-Band Microwave Transmitters", CCTA Technical Record, 1987 (Toronto)

(8) T. M. Straus and R. T. Hsu, "New Higher Power Microwave Broadband Transmitter Design and System Application" CCTA Technical Record, 1989 (Toronto)

(9) T. M. Straus, "Integrated AML/Fiber Backbone Systems", NCTA Technical Papers, 1989.

(10) S. Y. Huang, et al., "Frequency Dependent Distortions of Composite Triple Beat? in Lightwave CATV Transmission Systems," Optical Fiber Communications Conference, San Francisco, Jan. 1990, WH4.

(11) M. M. Choy, et al., "Interferometric Conversion of Laser Phase Noise to Intensity Noise by Single-Mode Fiber Optic Components", Electronic Letters, 23, 8 Oct. 1987

(12) R. B. Childs and V. A. O'Byrne, "50-Channel VSB-AM Video Transmission Employing a Linearized External Modulator," Optical Fiber Communication Conference, San Francisco, Jan. 1990, p. 23.

(13) Multichannel News, "Orchard Debuts External Modulator System," February 12, 1990, p. 28.

(14) E. G. Pastell and T. M. Straus, "Boosters for CARS Band Microwave Systems," CCTA Technical Record, 1986 (Vancouver).

(15) I. R. MacFarquher, "Distortion Balancing with Broadband AML Microwave Systems," CCTA Technical Record, 1988 (Halifax).

(16) D. McEwen, et al., "Distortion Characteristics of Integrated Fiber, AML, and Cable Amplifier Systems," CCTA Technical Record, 1990 (Edmonton).

(17) D. McEwen, et al., "Distortion Accumulation in Trunk," NCTA Technical Papers, 1990.

OPTICAL AMPLIFIERS FOR VIDEO DISTRIBUTION

J.B. Glaab and D.R. Huber
Jerrold Communications Applied Media Laboratory
General Instrument Corporation

ABSTRACT

The promise of a broadband fiber communications network has attracted the interest of CATV and Telco concerns. Various networks and Topologies have been discussed. Cost is a drawback in some of the proposed networks, particularly those networks which require switching. This paper discusses a low cost broadcast tree and branch network which utilizes optical amplifiers to extend the network penetration. Recent work in the development of optical amplifiers suggests that the Erbium fiber amplifier may be compatible with VSB-AM. The characteristics of such an amplifier will be discussed. This scenario allows an orderly transition from present day AM backbone system to the tree and branch fiber architecture. Once the broadband fiber plant is in place new services can be implemented that exploit the broadband nature of fiber.

INTRODUCTION

Optical Amplifiers have been studied for fiber optics since 1973. Significant work has been expended in the last ten years to develop practical devices. The last three years have shown an exponential increase in the development of these components. Most of the major makers of laser diodes have programs to develop semiconductor amplifiers.

The major telecommunications companies have demonstrated that the optical amplifier can be used to increase the link budget in digital applications.[1] During the past two years a number of demonstrations have shown that the link budget in subcarrier FM[2] fiber video systems can be significantly extended via optical amplifiers. These experiments have utilized FM or digital to solve the problems of adequate signal to noise ratio for a large number of channels, as well as for adequate system linearity. The technology has successfully demonstrated that 90 channels of video can be subcarrier multiplexed on a single laser using FM modulation. A sketch of a Bellcore[3] experiment where two optical amplifiers were used to send 90 channels of microwave sub-carrier multiplexed FM modulated video over a tree-and-branch network serving 2,048 subscribers is shown in figure 1. Commercial microwave satellite electronics equipment was used in the testing. The signal-to-noise (SNR) obtained in the referenced 90 channel experiment was 55 dBc. Composite second order (CSO) and composite triple beat distortion (CTB) are not a problem with FM transmission since the amplifier need only be capable of about 20 dBc CSO and CTB for top quality video. These demonstrations are significant since they show that a high quality, high channel capacity, electrically passive, video distribution network can be constructed using fiber. Optical amplifiers have the potential to lower the cost and improve performance in fiber video distribution systems. As volume increases it is likely that the cost of the optical amplifiers will drop. An optical amplifier capable of delivering AM video services would seem to provide the basis for the most cost effective distribution.

Analog Video Loop Distribution Using Two-Stage Cascade Optical Amplifiers

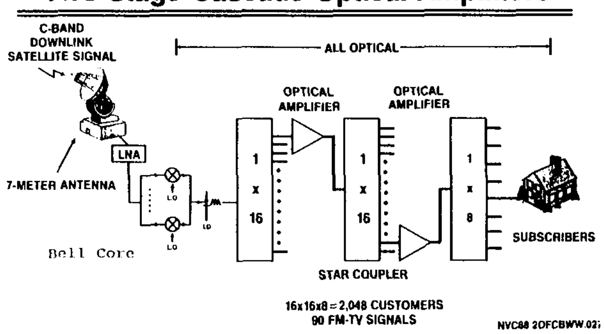


FIGURE 1

The optical amplifier is a designer's dream. The optical semi-conductor amplifier has a gain bandwidth product of about 300 THz, about 100 times that of the best microwave amplifiers. In addition, analysis has shown that optical fiber amplifiers having noise figures of about 3.5 can be realized[4]. Semi-conductor optical amplifiers with noise figures of 5.2 have been realized[5]. The optical amplifier is required to make a tree and branch structure practical for video distribution. It is believed that the tree-and-branch structure can be utilized to implement the fiber backbone or other similar architecture. See figure 2.

BRANCHING FOR AM FIBER BACKBONE

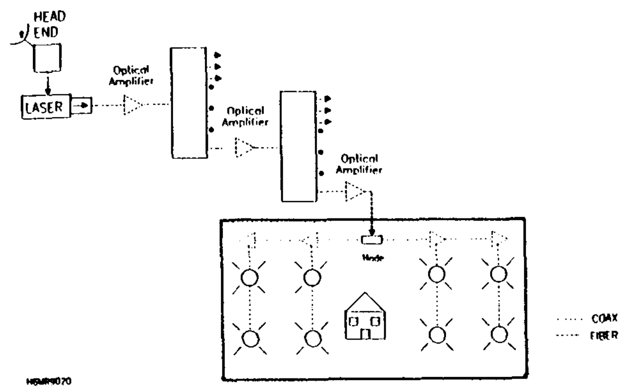


FIGURE 2

The tree and branch structure is needed to reduce the required amount of fiber. This is extremely important since the fiber plant is the most expensive component in the distribution scheme. Fiber is an ideal medium for video transport; it is a low loss medium, and it has a large bandwidth. A fiber tree and branch network will significantly reduce the number of active components between the headend and the subscriber, when compared to present video distribution networks. At present there may be more than forty amplifiers between the headend and the subscriber. The fiber backbone approach can be used to achieve a reduction in the number of cascaded electronic amplifiers. This would be advantageous since the SNR degradation due to amplifier cascade would be greatly reduced. The reduction in the number of active components in the distribution network would also lead to a more reliable network. An additional advantage is that the fiber plant supports virtually unlimited signal bandwidth. This excess bandwidth capacity can be utilized in the future as bandwidth demand grows. The optical amplifier is an inherently simple device with low power consumption. The semi-conductor amplifier is a monolithic component (see figure 3). The fiber amplifier requires an optical pump in the form of a monolithic semi-conductor laser, a fiber coupler for combining the pump and signal beams, the rare-earth doped fiber and possibly optical isolators (see figure 4).

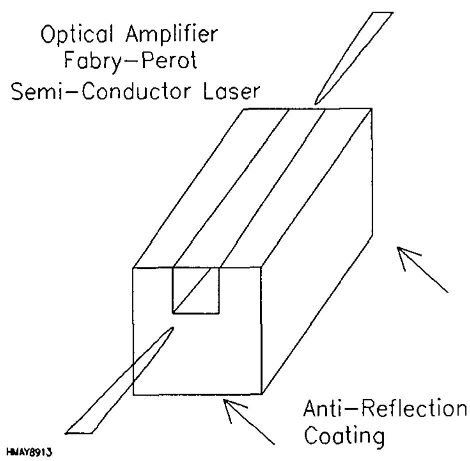


FIGURE 3

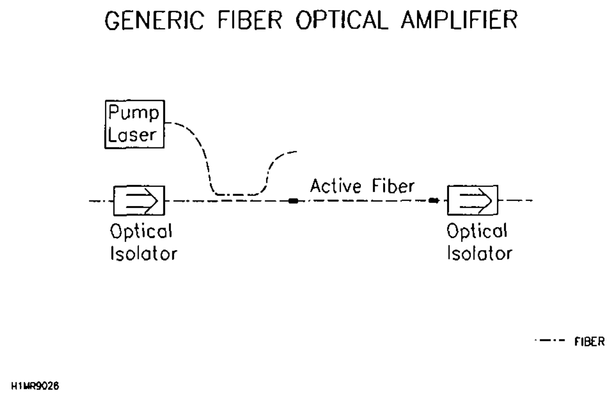


FIGURE 4

The demand for video bandwidth will grow due to increased numbers of broadcast channels, higher channel bandwidth requirements of HDTV and new services such as switched or bidirectional video. Fiber networks with optical amplifiers can transparently support changes in signal bandwidth and modulation format. The most popular modulation and multiplexing formats for fiber transmission are AM, FM, digital pulse code modulation (PCM), optical wavelength division multiplexing (OWDM) and optical frequency division multiplexing (OFDM). OFDM differs from OWDM in that it requires a coherent local oscillator for recovery of the transmitted information. With the exception of high density multi-channel AM, laboratory experiments have demonstrated the compatibility of the optical amplifier with all of the above modulation formats. Tests are under way in laboratories to determine the compatibility of the Erbium fiber amplifier with VSB-AM.

The low noise characteristics of the optical amplifier make it desirable for use as the first gain stage in a conventional high bandwidth fiber optic receiver. The optical amplifier can also be used to construct a fiber optic switch[6]. See figure 5. Recently, four-wave optical mixing has been demonstrated in semi-conductor optical amplifiers[7]. In this case, four-wave mixing was used to transfer the modulation from one optical frequency to that of another optical frequency. Optical frequency exchange could be important for future switched networks that are able to discriminate optical frequency.

OPTICAL AMPLIFIER USED
FOR BIDIRECTIONAL SWITCH

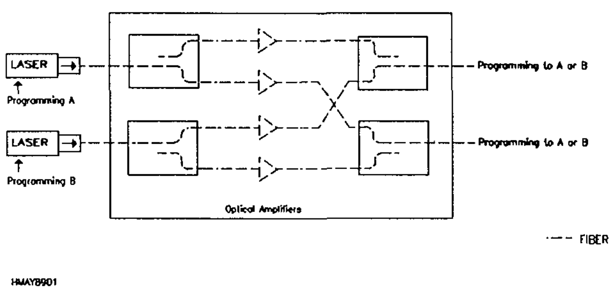


FIGURE 5

Both fiber and semi-conductor optical amplifiers are bi- directional devices, so a bi-directional network is possible in theory. A system architectural analysis of a specific fiber plant would be required for practical bi-directional systems. Reflections and optical noise terms limit practical systems.

In volume, the optical amplifier has the potential for relatively low cost. The compact disk player is the first consumer product to incorporate a laser. The path to lower cost in optical components is found by increasing the installed base of that component.

STRATEGY

The introduction of fiber plant overlay will improve performance, reliability and provide a means of offering new revenue producing services. The introduction of fiber plant may also be required to maintain a competitive posture.

COMMERCIAL MARKET STATUS

Within the last two years a number of vendors have begun offering commercial semi-conductor optical amplifiers. Most of these devices are developmental devices. A few companies are presently qualifying semi-conductor optical amplifiers for undersea telecommunications applications. Laboratory work is underway to develop the Erbium fiber amplifiers for telecommunications and CATV applications. Doped Erbium fiber and semi-conductor amplifiers are the technologies that are viable for near term development.

PENETRATION

Penetration of fiber into the subscriber network will lead to improved system reliability and higher quality video. The first thrust into the network is to deliver video to the node as in the fiber backbone architecture. The optimum placement of the optical amplifiers depends on a number of system parameters, which would be determined by component performance. Figure 2 is an AM network topology that could be supported by optical amplifiers having a modest fiber-to-fiber gain. Optical amplifiers have demonstrated 23 dB gain at a saturation output power level of 20 mW. Amplifiers having this level of performance allow a cost effective network to be built. The next evolutionary step would be to cascade an additional stage of optical amplifiers.

TYPES OF OPTICAL AMPLIFIERS

There are a number of types of optical amplifiers that are compatible with fiber systems. These are shown in figure 6. Earlier in the discussion it was indicated that one great advantage of fiber is that the fiber attenuation is very low. That is the good news. The bad news is that it is difficult to achieve the high signal power necessary for VSB-AM CATV distribution. This is an important point since VSB-AM tree and branch is very desirable so that the cost of distribution may be kept low. Optical amplifiers may aid the distribution of video via the fiber backbone approach.

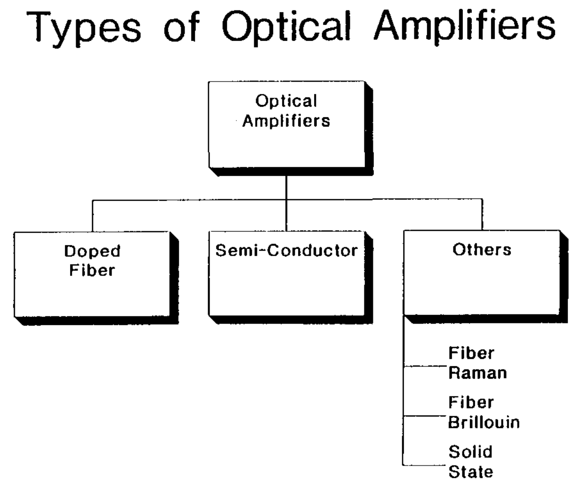


FIGURE 6

Semi-conductor Optical Amplifiers

Most of the semi-conductor optical amplifiers that have been built to date are standard lasers with a precision anti-reflection coating applied to the facets. The anti-reflection coating must be sufficiently good that even residual lasing action is inhibited. The first problem lies in the removal of the mirrors. It is desirable to achieve reflectivity[8] on the order of 10^{-5}. As expected, it is very difficult to achieve these results with multi-layer dielectric coatings. Unfortunately, these devices are produced one at a time by individually monitoring each laser during the coating process. Work is also being done to reduce the stringent requirements placed on the anti-reflectivity coating. The most promising technique is to place the waveguide at an angle relative to the cleaved facet. This technique coupled with a modest anti-reflection coating promises to yield effective facet reflectivity near 10^{-5}. One or the other of these two techniques is required to produce reflectivities on the order of $R=10^{-5}$. Reflectivities this low are required to produce an optical amplifier with high saturation output power. The slanted facet concept coupled with a modest anti-reflectivity coating offers promise for mass production of optical amplifiers.

The material gain peak in some semi-conductor amplifiers is not aligned with the standard communications wavelength. This problem occurs in amplifiers that are based on standard lasers. An increase in threshold current brought on by lowering the facet reflectivity causes a corresponding increase in injection current density, this in turn causes the center of the material gain peak to shift to shorter wavelengths. One way around this problem without changing the material doping or optical cavity is to cleave the amplifier chip at a longer length. A standard laser chip is about 300 um in length. Chips that are destined to become optical amplifiers are cleaved at about 500 um, the device can then be operated at a lower current density, thereby shifting the gain curve peak back toward the standard operating wavelength.

Current devices have polarization sensitive gain. The gain sensitivity to polarization decreases to about 3 dB maximum at a reflectivity near $R=10^{-4}$. Devices with $R=10^{-5}$ exhibit some residual polarization sensitivity which stems from the design of the optical cavity. Paper studies examining the optimization of the gain medium for the optical amplifiers application have been published. Major laboratories have programs to begin producing amplifiers incorporating these ideas as part of their optical amplifier development programs.

A major problem affecting the semi-conductor optical amplifier is the efficiency of the fiber coupling. Current commercial devices have as much as 6 dB loss per facet. This 12 dB loss reduces available gain from 27 dB to 15 dB. The poor input coupling efficiency also increases the noise figure. If the input coupling efficiency were 100%, semi-conductor amplifier noise figures near 3.8 dB could be obtained.

Semi-conductor optical amplifiers are based on mature technology. Relatively slight modifications to the existing Fabry-Perot device structure, an anti-reflection coating and an additional fiber pigtail produces an optical amplifier. The semi-conductor optical amplifier is also directly pumped, leading to a very simple and reliable device. These devices have been commercially developed for undersea applications[9]. It is expected they will be deployed in three years. The limiting factor in undersea deployment is the qualification program which requires three years.

Another problem with semiconductor optical amplifiers is that there is a large optical loss associated with coupling the light from fiber to amplifier and back to the fiber. At present the combined input and output loss for commercial semiconductor optical amplifiers may be as high as 12 dB. Even with this high loss, amplifiers with gains as high as 15 dB can be realized. Recent devices with 3 dB optical saturation powers as high as 45 mW have been achieved. These devices offer a broad 3 dB gain bandwidth. Standard laser structures offer 40 nm bandwidth, while recently developed quantum-well devices may offer 100 nm of optical bandwidth. One drawback of the semiconductor optical amplifier is that the excited state life-time of the optical carriers is short compared to video modulation rates. This leads to large intermodulation distortion products at power levels and subcarrier frequencies that are compatible with VSB-AM. The semiconductor amplifier works well with digital and FM subcarrier modulation.

Brillouin Fiber Amplifier

The Brillouin fiber amplifier utilizes the non-linear properties of standard fiber. In this amplifier, a narrow linewidth pump laser is co-propagated with the signal laser in a fiber. The gain bandwidth of the Brillouin amplifier is about 100 MHz. Gain is present at an optical frequency that is offset by about 11 GHz from the pump laser. Because of the narrow gain profile, the usefulness of this fiber amplifier is limited to special applications. One application is as an all optical phase lock loop for the recovery of a vestigial local oscillator in coherent phase shift keyed systems.

Raman Fiber Amplifier

The Raman fiber amplifier is also made with standard fiber. An incident pump laser photon is scattered to a lower optical frequency by a vibrational state of a silica molecule. Unlike Brillouin gain, the gain bandwidth of the Raman amplifier is 40 THz, with a sharper satellite peak which has a 13 THz bandwidth. The Raman pump power threshold at 1.55 um is 600 mW. At present it is difficult to achieve this level of pump power with semiconductor pump lasers. This amplifier could become important when higher power semiconductor lasers are developed.

Doped Fiber Amplifiers

A functional schematic of a fiber doped amplifier is shown in figure 4. Early work utilized Neodymium (Nd) doping for operation near 1.0 um or 1.3 um. These amplifiers have not found use as practical amplifiers because they suffer from excited state absorption (ESA) of the signal when operated at 1.3 um. In the Nd amplifier ESA of the signal laser is the dominant transition from the excited state. Current interest is centered in Erbium doped silica fibers. This laser system is free from excited state absorption under certain conditions.

Erbium Fiber Amplifier

Erbium fiber amplifier is made by doping a fiber with the rare-earth Erbium. A co-propagating pump laser excites the Erbium laser system. This laser system is shown in figure 7. The excited state lifetime of the $^4I_{13/2}$ line is very long, about 12 mS. This long excited state lifetime yields an amplifier with characteristics that may be compatible with VSB-AM. It is believed that CSO and CTB levels compatible with VSB-AM can be achieved. Erbium fiber amplifiers with gains higher than 40 dB have been demonstrated. Separate experiments have demonstrated amplifiers with output saturation powers higher than 20 mW. The Er doped fiber is a three level laser system that may be pumped at a number of wavelengths. Each of these pumps has attributes which affect its desirability as a pump laser. A number of pump lasers have been considered. A summary of some of the practical considerations follows.

532 nm: This source would be frequency-doubled Nd:YAG. KTP could be the doubling material. The Nd:YAG requires a semi-conductor pump in the area of 800 nm. This could show promise as guided- wave non-linear optics are developed. This is a desirable laboratory wavelength because these components are all available commercially.

807 nm: This is available from high power injection locked GaAlAs diode arrays. Both 532 nm and 807 nm are multimoded in the Er doped fiber so pumping efficiency suffers because of poor mode field overlap between the pump and signal wavelengths. Both the 532 nm and 807 nm pump wave suffer from excited state absorption of the pump.

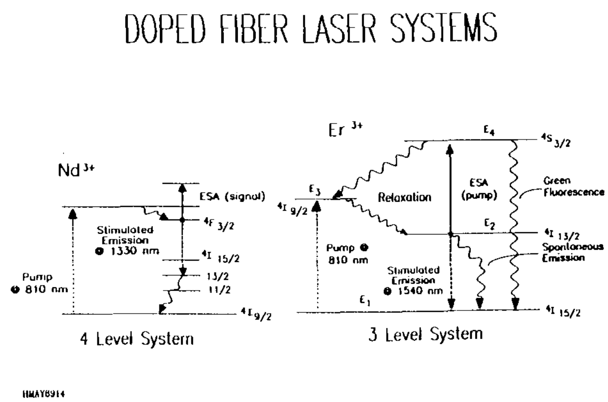


FIGURE 7

980 nm: Ion pumped dye lasers have demonstrated this pump wavelength. This wavelength is free from excited state absorption. Indium doped GaAs semi-conductor laser operation at 980 nm has been demonstrated in research environments. Development work indicates that high power strain[10,11] layer lasers may produce a reliable high power laser source for use as a commercial pump. The 980 nm pump produces an amplifier with a lower noise figure than the 1480 nm pump and is a more efficient pump.

1480 nm: This is an advanced structure, high power GaInAsP laser diode. This is an attractive pump since there is no excited state amplification and the mode fields of the pump and lasing fields have good overlap. Table 1 compares the efficiencies of these various pump lasers.

One of the advantages of the doped fiber amplifier is that low reflectivity splicing to CATV fibers is easily accomplished. Unfortunately the ideal mode field diameter for a doped fiber amplifier is not the same as for a CATV fiber; however, the mode field is Gaussian so a simple Gaussian transformation in a tapered fiber should lead to a very low loss interface between the amplifier fiber and transmission fiber. There are a number of researchers[12] that are active in developing these transformation devices. High coupling efficiency is required to achieve an optical amplifier with a low noise figure. Fiber amplifiers have been demonstrated to have a saturation power level of 20 mw[13]. Fiber-to-fiber gains of 50 dB have also been reported.

The spontaneous emission spectra of an Erbium fiber amplifier with and without an amplified signal is shown in figures 8 and 9.

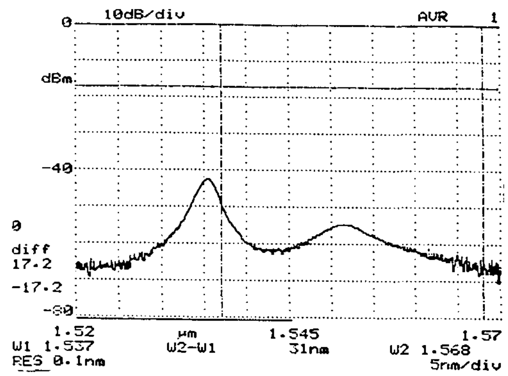
TABLE 1

GAIN PER UNIT OPTICAL PUMP POWER

Pump Wavelength	Gain dB/mW
665 nm	.26
514 nm	.22
528 nm	.31
647 nm	.23
807 nm	.30
980 nm	2.20
1480 nm	.30

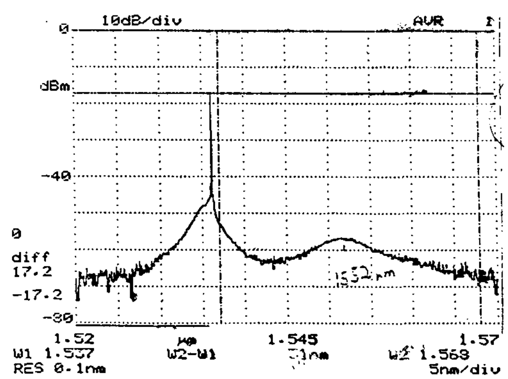
Er Fiber Amplifier

Output of Optical Amplifier Showing Amplified Source



Spontaneous Emission of Erbium Fiber Amplifier

FIGURE 8



Gain dc 19 dB
ac 14 dB

Distortion CTB >57 dB

FIGURE 9

The characterization of amplifier CTB and CSO is extremely important in the transmission of multi-channel analog intensity modulation. Characterization of CTB and CSO is an area that researchers at Jerrold Applied Media Lab are examining. All distortion tests to date have been limited by laser probe performance, not amplifier performance. The following is a summary of the results that have been obtained. Single tone distortion tests have been made that extrapolate to better than 65 dBc for a 40 channel system. 20 channel CTB levels of less than 55 dBc have been verified. Work is under way to improve the measurement system so lower levels of distortion can be measured.

CONCLUSION

The tree-and-branch network shows promise as a cost effective method of bringing fiber closer to the home. The optical amplifier provides the required gain to offset branching and transmission loss. The bandwidth provided by standard CATV fiber is essentially unlimited. The optical amplifier provides a low-noise, extremely large gain-bandwidth product to support wideband transmission of information. The literature does not indicate any fundamental reason that fiber optical amplifiers cannot support amplitude modulation. frequency modulation with microwave sub-carrier modulation has already been demonstrated successfully with semi-conductor and Erbium fiber amplifiers. Recent work indicates that an output power of 20 mW can be delivered by a fiber optical amplifier.

Fiber optics technology is developing at a rapid rate. As these optical technologies develop, costs will reduce and CATV will evolve to utilize the new technology. The evolutionary approach using AM-VSB signals has the best chance of driving fiber further into the system. With the advances in linear AM lasers, power splitting is becoming practical, thereby lowering the cost of a link. On the near horizon is the promise of optical amplifiers capable of supporting AM-VSB signals. The positive aspect of this is that true optical tree-and-branch architectures are possible terminating at bridger stations, then line extenders and perhaps, with more implementation, at the tap. With the present cost of terminal equipment, it will be a long time until it is economical to take fiber to the home. The present goal should be to get fiber further into the system. VSB-AM is the most difficult but most cost-effective modulation format. As cost effective digital systems become available in future years, the installed network will evolve to support them.

REFERENCES

1. Shigeru Saito, et al., "An Over 2,200 km Coherent Transmission Experiment at 2.5 Gbits/s Using Erbium doped fiber Amplifier," Optical Fiber Communications Conference 1990, San Francisco, CA paper PD2.

2. R. Olshansky, E. Eichen, "Microwave Multiplexed Wideband Lightwave System Using Optical Amplifier for Subscriber Distribution," Electronics Letters, vol. 24, no. 15, pp. 922-923, 1988.

3. W.I. Way et al., "Carrier-to-noise ratio performance of a ninety-channel FM video optical system employing subcarrier multiplexing and two cascaded traveling-wave laser amplifiers," Optical Fiber Communications Conference 1989, San Francisco, CA paper WC3.

4. R. Olshansky, "Noise Figure for Erbium Doped Optical Fibre Amplifiers," Electronics Letters vol. 24, no. 22, pp. 1363-1365, 1988.

5. T. Mukai, T. Saithoh, "5.2 dB Noise Figure in a 1.5 um InGasAsP Travelling-Wave Laser Amplifier," Electronics Letters, vol. 23, no. 5, pp. 216-218, 1987.

6. E. Eichen et al, "Lossless, 2x2, All-Fiber Optical Routing Switch," Optical Fiber Communications Conference 1990, San Francisco, CA, paper PD20.

7. K. Inoue, N. Takato, "Optical Frequency Exchange Utilizing Semiconductor Laser Amplifiers and Mach-Zehnder Filters," presented at Conference on Laser and Electro-optics (CLEO), 1989.

8. N.A. Olsson, M.G. Oberg, L.D. Tzeng, T. Cella, "Ultra-Low Reflectivity 1.5 um Semiconductor Laser Preamplifier," Electronics Letters, vol. 24, no. 9, pp. 569-570, 1988.

9. K.H. Cameron, et al, "Packaged Laser Amplifiers at 1.5 um for Submarine Systems," presented at Conference on Lasers and Electro-optics 1989.

10. M.E. Givens, C.A. Zmudzinski, R.P. Bryan and J.J. Coleman, "High-power Nonplanar Quantum Well Heterostructure Periodic Laser Arrays," Applied Physics Letters, vol. 53, no.13, pp. 1159-1161, 1988.

11. P.K. York, K.J. Beernink, G.E. Fernandez, and J.J. Coleman, "InGaAs-GaAs Strained-Layer Quantum Well Buried Heterostructure

Lasers by Metalorganic Chemical Vapor Deposition," Applied Physics Letters, vol. 54, no. 6, pp. 499-501, 1989.

12. D.B. Mortimore, J.V. Wright, "Low-Loss Joint Between Dissimilar Fibers by Tapering Fusion Splices," Electronics Letters vol. 22, no. 6, pp. 318-319,1986.

13. F. Fontana, A. Fighetta, "Performance Characterization of Erbium Doped Optical Fiber Amplifiers," Conference on Lasers and Electro-optics, 1989.

OPTICAL RETURN LOSS

R. TODD HUDSON

SIECOR CORP.

ABSTRACT

Because CATV applications are pushing the limits of fiber transmission equipment, a high quality optical fiber system is essential for the required performance. In most systems to date, the sole measure of this quality has been the attenuation of the link. However, a primary concern in CATV systems is the sensitivity of most AM fiber transmitters to reflections.

An understanding of what causes reflections in fiber systems, measures taken to minimize these, and methods to specify and test return loss for individual components and installed systems will provide the designer/installer with the tools to ensure a fiber optic system with optimum signal quality and flexibility.

OPTICAL RETURN LOSS

The benefits of transmission using fiber optics have led to great interest and aggressive plans for the design and implementation of lightwave systems for cable TV applications. In the past, conventional fiber optic systems were designed almost completely based on optical attenuation, specified as a link loss budget. However, with the recent development and installation of equipment for transmitting high-quality analog signals, potential limitations associated with the reflections of a system require attention also.

Reflections are a concern for analog transmission in particular since they can lead to significant degradation of signal quality. When light is reflected back into the laser cavity it causes interference which can create instabilities in signal output power and spectral behavior. The resultant noise will introduce power penalties and reduce the system signal-to-noise ratio.

The system designer's goal, therefore, is to minimize the source of this optical feedback and prevent its impact on laser performance. A well-planned system will maintain optimum signal quality and allow for future growth and flexibility. This goal is achieved by: a) specifying components based, in part, on their reflective qualities, and b) testing the installed system for its combined reflection or system return loss.

The Root Of All Reflections

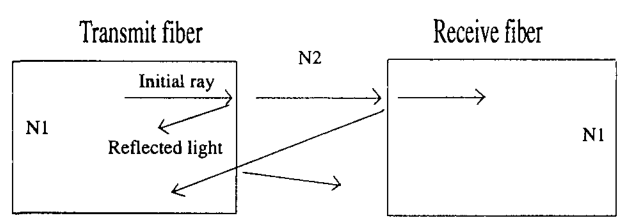
In order to optimize a system's design and selection of its components it will be helpful to understand the two basic types of reflections:

(1) **Fresnel Reflections**

These reflections occur at points in the cable where the continuity of the glass is interrupted, e.g. at connectors, mechanical splices, splitters, couplers, and other fiber optic components. These points of localized change in the light's medium are commonly seen on an optical time domain reflectometer (OTDR) as spikes.

FIGURE 1

PRINCIPAL OF REFLECTION



N1 = Index of refraction of glass

N2 = Index of refraction of air

(2) **Rayleigh Backscattering**

Backscattering is low-level reflection from the fiber itself. It is inherent to the glass structure and minimized by the design and manufacturing process of the fiber. This backscattered light is distributed over the entire length of fiber and seen as a linear trace on an OTDR.

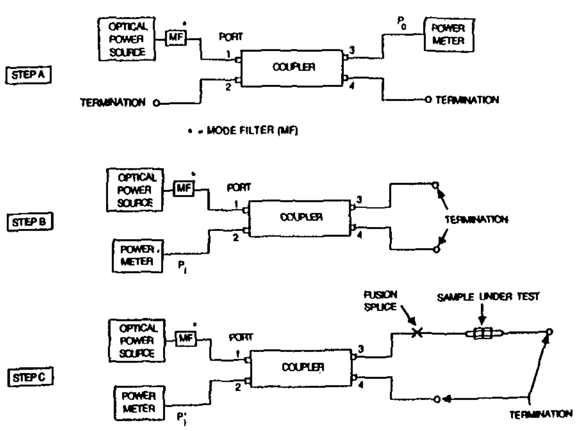
Quantifying Return Loss

In order to examine or evaluate these reflective components either individually or collectively in an installed cable system, it is necessary to quantify the amount of reflected light. This is referred to as **return loss** - simply a ratio of how much light is reflected back towards the transmitter compared to how much is transmitted out of it.

The Electronic Industries Association has recently published a Fiber Optic Test Procedure, FOTP-107, "Return Loss" that defines its test method (Figure 2). Bell Communications Research also outlines this same procedure in its Technical Reference TR-TSY-000326 on fiber optic connector specifications.

FIGURE 2

RETURN LOSS TEST SETUP



Return loss is calculated as follows:

$$\text{Return Loss} = -10 \log_{10} \frac{P_t - P_r}{P_o} + c$$

Where P_o = Optical Power Incident On Fiber

P_r = Reflected Optical Power - Reference

P_t = Reflected Optical Power - Test

And c = Constant Determined By The Characteristics Of The Test Setup

Notice that return loss is measured in dB and defined so that the smaller the amount of reflection, the larger the return loss. As a result, efforts to minimize reflections and construct a better system will actually maximize the return loss reading in dB.

Minimizing Reflections

Putting together a system with the least amount of reflection is accomplished by first evaluating and selecting individual components with small reflections. Efforts to minimize reflections are focused on the primary contributors: connectors and mechanical splices. Table 1 lists typical return loss values of components used in fiber optic cable systems.

TABLE 1

TYPICAL OPTICAL RETURN LOSS

• Connectors	
Conventional	15-25 dB
Physical Contact (PC Or Super PC)	30-50 dB
• Mechanical Splices	20 To >40 dB
• Fusion Splices	>50 dB
• Single-Mode Fiber	>50 dB

Actual System Return Loss:

1. Is based on combined effects of the above system components.

2. Cannot be mathematically added or calculated but is dependent on the number, magnitude and locations of the reflections.

3. Must be measured to accurately determine the combined effect of the individual components.

Development of modified polishing techniques, anti-reflection coatings, and methods of reducing the air gap between two connectors has resulted in "PC" (Physical Contact) and "super PC" connectors with higher return loss as is seen by the high upper range for connectors in Table 1.

Several mechanical splices now utilize index-matching materials, tighter tolerances, and polishing steps similar to connectors in attempts to maximize return loss.

There has also been developmental work done recently on optical isolators. An isolator attempts to mask the amount of light actually reflected back into the laser cavity. As it stands, even when isolators are used, the combination of reflections can result in noise that can significantly degrade the signal quality of the lightwave transmission system.

Determining System Return Loss

Each of the system components can be evaluated individually for its return loss performance. However, the overall system return loss is the combined effect of each Fresnel reflection, e.g. connectors, mechanical splices, and the backscatter of the fiber itself.

These reflections cannot be simply added together or calculated. Their combined effect is interdependent upon the number, magnitude, and location of all the reflections, as well as the attenuation of the fiber.

The amount of Rayleigh backscatter is based on the fiber's intrinsic scattering factor, attenuation, and response to the optical signal energy. These variables along with the effects of:

(a) Attenuation of reflections on the return trip.

(b) Multiple reflections between fiber joints, make measurement of the actual system return loss a key to predicting system performance.

Note that Rayleigh backscattering caused by the fiber plays a minor role in overall system return loss when connectors or mechanical splices are present. Therefore the system designer can maximize return loss by attempting to minimize the number and magnitude of Fresnel reflections. This is most easily and best accomplished by utilizing fusion splicing and PC connectors at fiber joints. (Reference Table 1 for a comparison of splicing and connection methods.)

System Performance

Specification of return loss for a given system is a function of the transmission equipment, indicating how much reflection it can handle while maintaining its standard of signal quality. Vendors of the equipment being used should be able to provide guidance in establishing a system specification. Measurement of system return loss would allow reliable prediction of its performance based on comparing test values with the system specification.

In addition to meeting the current return loss specification, consideration should be given to adding margin for potential future enhancements such as link extension, signal splitting, wave division multiplexing, and other enhancements requiring a minimum level of signal quality.

Summary And Recommendations

Reflections in fiber optic cable TV transmission systems can cause noise that leads to significant degradation of signal quality. Evaluation of system components and measurement of installed systems can help ensure the proper operation of the system with the required standard of signal quality. After examination of return loss - its causes, effects, and measurement - the following practical guidelines should be considered for system design and testing:

(1) Evaluation and specification of return loss of individual components prior to installation. One of the criterion for selection should be return loss.

(2) Testing the overall system return loss after installation as an acceptance test for designed system return loss specification.

(3) Considering the impact of potential future enhancements on overall system reflection in the original design and specification.

Taking these steps will lead to designing and implementing a system with optimum signal quality while maintaining flexibility for smooth growth and enhancement.

REFERENCES

1. Choy, M.M., Gimlett, J.L., Welter, R., Kazovsky, L.G., and Cheung, N. K.: "Interferometric Conversion Of Laser Phase Noise By Single-Mode Fiber-Optic Components," Electron. Lett. 23, pp. 1151-1152 (1987).

2. Gimlett, J.L., Young, J., Spicer, R.E., Chung, N.K.: "Degradations In Gbits/s DFB Laser Transmission Systems Due To Phase-To-Intensity Noise Conversion By Multiple Reflection Points," Electron. Lett., 24, p. 406 (1988).

3. Mazucyzk, V.J.: "Sensitivity Of Single-Mode Buried Heterostructure Lasers To Reflected Power At 274 Mbit/s," Electron. Lett. 17, p. 143 (1981).

4. Hirota, O. Suematsu, Y. and Kwok, K.S.: "Properties Of Intensity Noises Of Laser Diodes Due To Reflected Waves From Single-Mode Optical Fibers And Its Reduction," IEEE J. Quantum Electron., QE-17, p. 1024 (1980).

5. Peters, J.W., "Field Measurements Of Fiber Optics Systems," submitted to 1988 National Communications Forum.

6. Mulkey, O.R., "Return Loss," EIA FOTP-107, FO 6.3/6.4, pp. 1-7 (1988).

7. Bell Communications Research, "Fiber Optic Connectors," Technical Reference TR-TSY-000326.

8. Young, W.C., Shah, V.S., and Curtis, L.: "Optimization Of Return Loss and Insertion Loss Performance Of Single-Mode Fiber Mechanical Splices," submitted to International Wire And Cable Symposium '88, Proceedings pp. 395-400 (1988).

9. Wagner, R.E., Sandahl, C.R., "Interference Effects In Optical Fiber Connections," Appl Opt. 21, p. 1384 (1982).

10. Suzuki, N., and Nagano, O.: "Low Insertion Loss And High Return-Loss Optical Connector For Use In Analog Video Transmission," IOOC '83, 30A p. 3-5 (1983).

USING FIBER TO INTERCONNECT DIFFERENT SYSTEMS, REDUCE CASCADES AND CHANGEOUT ADDRESSABLE CONVERTERS

Larry N. Lehman
Cencom Cable Associates, Inc.

ABSTRACT

Over the past five years, Cencom Cable Associates, Inc., has acquired three different cable systems in St. Louis County, Missouri. These systems have about 3000 miles of plant and serve over 100,000 subscribers today with five headends and three AML microwave receive sites.

This paper will discuss the plan that has been proposed and is being implemented to interconnect the systems using fiber optics with control from one headend, eliminating the microwave systems and reducing all trunk runs to less than nine amplifiers in cascade. Also involved in this plan is the changeout of addressable converters in one of the systems without adding channel capacity, but using AM fiber nodes to break up the trunk runs. After the plan is completed, within three years, the system will be served from one headend with three hubs fed by FM fiber links and 90 AM fiber nodes.

BACKGROUND

Cencom Cable Associates, Inc., is a MSO headquartered in St. Louis County, Missouri with cable systems in eleven states serving over 420,000 customers. CCA was founded in 1982 and by December 1984 had acquired systems in five states with around 40,000 subscribers. The next year saw Cencom grow from 40,000 subscribers to 150,000 subscribers largely due to the acquisition of three systems in St. Louis County. Each system uses different outside plant technology and different addressable converters.

The management and customer operations of the systems have been consolidated with all customer service and repair service functions being handled from one location. This plan will integrate the technical networks and allow the systems to be managed and operated as one system.

EXISTING

Existing Network

The existing systems and network configuration is shown in figure 1. Cencom I is the system acquired from Warner Amex. It is a 400 MHz system utilizing Qube two way interactive addressable converters and C-COR electronics. The headend is located in Olivette with three AML receive sites using Hughes high power AML microwave equipment. Cencom II is the system acquired from Group W. It is a 400 MHz system utilizing Zenith Z-Tac one way addressable converters and Jerrold electronics. Its territory is served with three separate headends. Cencom III is the system acquired from Storer. It is a dual 330 MHz system utilizing Tocom addressable converters and Texscan electronics served out of one headend.

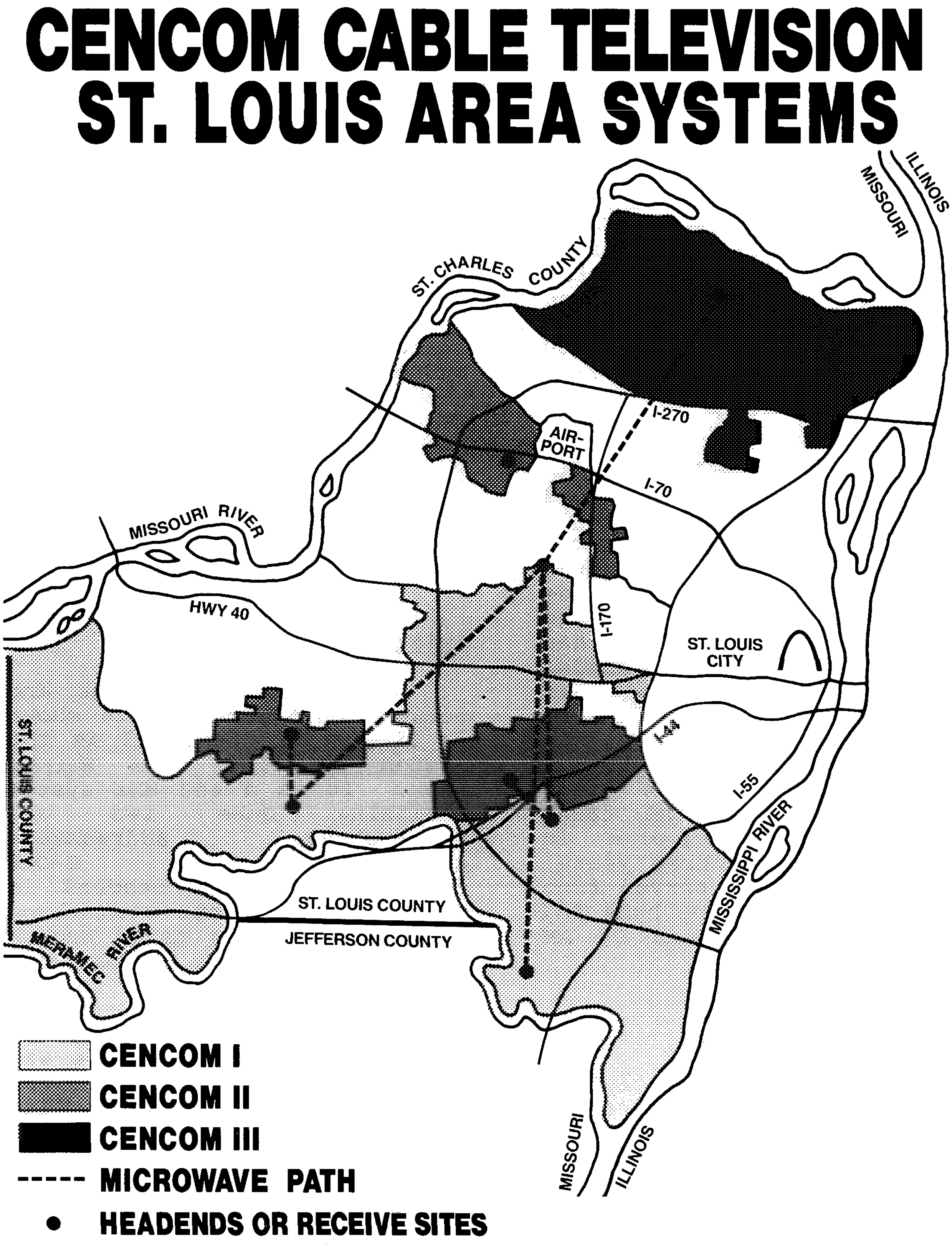

CENCOM CABLE TELEVISION
ST. LOUIS AREA SYSTEMS
ILLINOIS
MISSOURI
ST. CHARLES COUNTY
I-270
AIR-PORT
I-70
MISSOURI RIVER
HWY 40
I-170
ST. LOUIS CITY
FRANKLIN COUNTY
ST. LOUIS COUNTY
I-44
I-55
MISSISSIPPI RIVER
ST. LOUIS COUNTY
JEFFERSON COUNTY
MERAMEC RIVER
MISSOURI
ILLINOIS
CENCOM I
CENCOM II
CENCOM III
MICROWAVE PATH
HEADENDS OR RECEIVE SITES

Figure 1

The systems' five headends have been interconnected on a limited channel basis utilizing AML, AML line extenders and FM over coax. These links are used to deliver Pay Per View channels, barker channels, and local access channels.

Qube converters

As stated earlier, the system acquired from Warner Amex, Cencom I, uses the Qube two-way interactive converters. These converters have not been manufactured in six years and additional Qube converters that would work in this system are not available. The system was running out of converters! Attempts to find a satisfactory Qube compatible converter or decoder were unsuccessful. The decision was made to change out the converters using Jerrold IPPV converters retaining the RF return path over the cable system.

There are six encoded channels on the system. In order to change out the converters, the Qube and Jerrold encoding schemes needed to exist simultaneously. There was not excess channel capacity and no plan to increase the channel capacity of the system. Replacing six existing channels with the Jerrold encoded channels for the length of time it would take to replace 38,000 converters is not acceptable from a customer or political standpoint. If however this period of time is reduced to a few weeks, the time to replace 1500 or so converters, it would be acceptable. The use of AM fiber nodes is an excellent way to break up the system with each node capable of having a separate channel lineup feeding a unique group of customers.

Local Origination, PPV, Ad Insertion

Each of the three systems in St. Louis County is addressable and offers three channels of Pay Per View programming. Two of these channels are locally originated. In addition there is a county wide Cooperative School District program that delivers programming over the cable systems to local schools.

In order to deliver this programming, two of the three Missouri II headends receive limited channels utilizing Hughes AML line extender equipment from two of the AML receive sites in Missouri I. The third headend is connected using a FM over coax link. The Missouri III headend is connected using an AML path.

Local ad insertion is currently being done on four channels being inserted at each headend. Inserting local advertizing on any additional channels will require additional equipment at all five headends.

Plant Extensions

The western section of St. Louis County is the current growth area. A significant portion of this area is not currently being served by cable. Most of this area could not receive satisfactory service by extending the existing plant. The longest length of the existing plant has 38 feed forward amplifiers in cascade after an AML receive site. A different means to service this area is required before plant can be extended.

PROPOSED PLAN

In addition to the above specific situations that needed solutions, there existed the desire to improve the overall performance and the reliability of service provided to our customers. The result is the Cencom Cable Television Missouri Fiber Optic Plan as shown in figure 2. The following will discuss the plan and how we made our decisions.

Backbone Architecture

There has been much written over the past two years regarding the fiber optic backbone architecture proposed by ATC and the different variations that have been proposed and utilized since then by different cable operators. Most of these alternatives were considered.

One of the first decisions to be made was if we would turn around amplifiers. Route diversity for the fiber optic cable was not feasable for most of the area. Since the fiber cable would be overlashed to the existing coaxial cable, a redundant backup switch would provide backup only in the case of optical failure. A loss of continuity, fiber break, pole knockdown, etc., would affect the coax as well as the fiber. Not turning around amplifiers would require about 40% more AM fiber modes. We decided that this additional cost could not be justified to just provide backup in the case of optical failure.

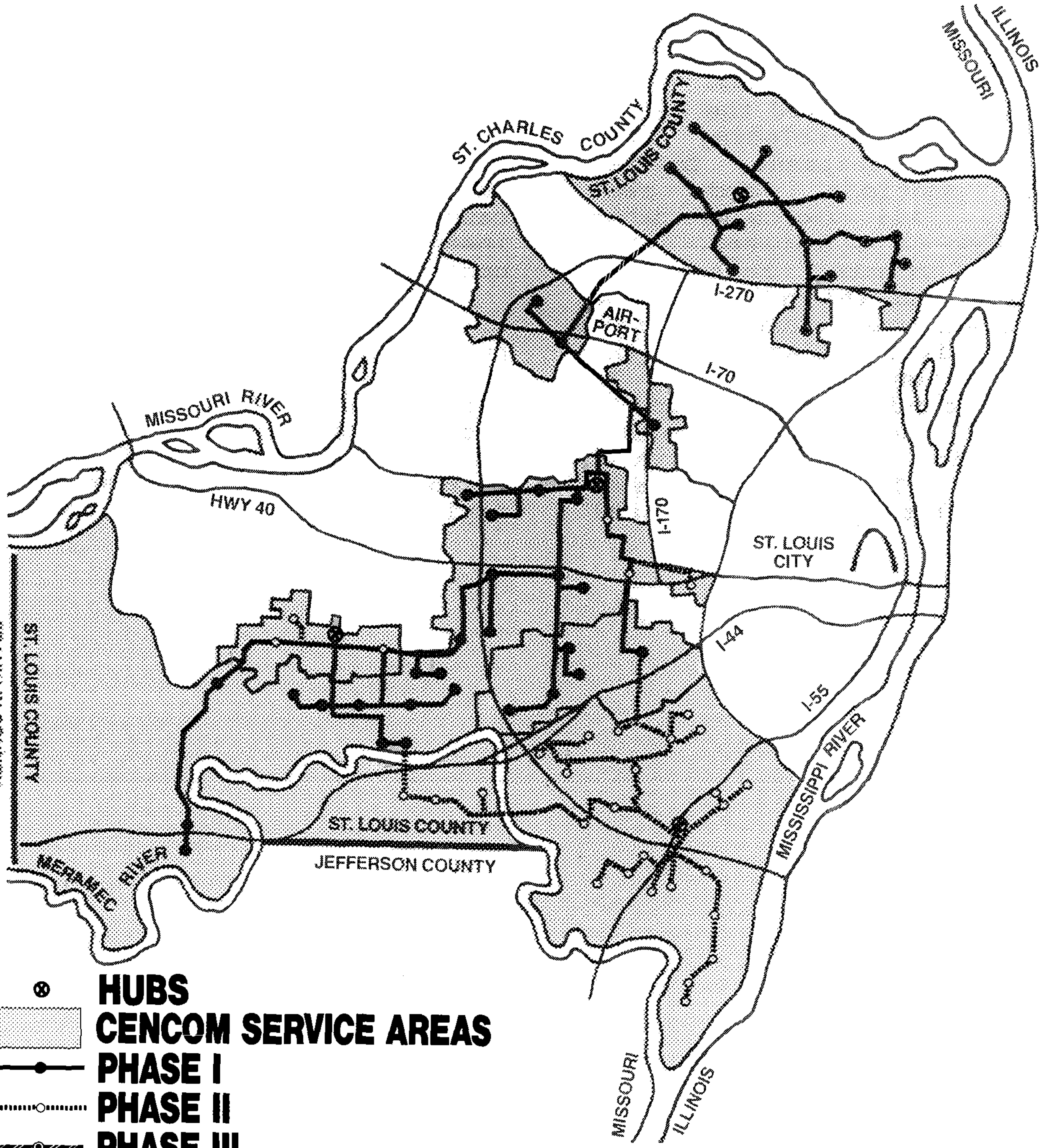
Another major decision is the number of amplifiers that will be left in cascade behind the fiber nodes. A fiber backbone plan was first designed for four amps in cascade. Then we looked at a design with nine amps in cascade. The nine amp design required only about one third as many nodes as the four amp design. Since the systems are 400 MHz with HRC headends, the performance with nine amplifiers in cascade will meet all performance requirements. Each node will serve less than 2000 customers which will work well with the addressable converter changeout process.

Hubs

As mentioned earlier, a main objective of the plan is to be able to control all the systems in St. Louis County from one location. This location is in Olivette where all customer and repair service representatives are located, where the studio and pay per view insertion equipment is, and where the billing computer is located.

The AM fiber technology is limited in the distance it can serve. Our first design anticipated AM nodes leaving six headends, Olivette and five fiber hubs. Each hub had to receive "headend quality" and is connected back to Olivette using FM over fiber. As the quality of AM fiber products has improved, we feel confident that excellent quality can be achieved on links up to 23KM in length using single laser transmitters. Because of the lengths possible with the AM technology, all of Cencoms St. Louis county systems, including currently unserved areas, can now be served by four hubs, the

CENCOM CABLE TELEVISION FIBER OPTICS PLAN



HUBS
CENCOM SERVICE AREAS
PHASE I
PHASE II
PHASE III
AM FIBER NODES

Figure 2

three fiber hubs served by FM over fiber from Olivette and the Olivette headend.

All programming sources, addressable security encoding and commercial ad insertion can now be done at Olivette. This design will enable us to eliminate two of the three headends that served the Cencom II system including the buildings, equipment, towers, earth stations, and real estate. All microwave paths will also be eliminated. This will eliminate rain fade problems and the need for the towers and real estate at each receive site. In addition, the ad insertion equipment now in each headend can be moved to Olivette and used to insert advertizing on additional channels. The required number of satellite receivers and VideoCipher decoders is also reduced dramatically.

CONCLUSION

When completed by the end of 1992, the Cencom Cable Television Missouri Fiber Optic Plan will have placed around 200 miles of fiber optic cable containing around 2400 miles of fiber. There will be 90 AM fiber nodes each serving less than nine amplifiers in cascade. All programming and insertions can now be done and controlled from one location.

All customers, including potential customers currently unreachable, will receive improved picture quality and much better reliability with the elimination of the microwave and no more than nine amplifiers in cascade.

WHAT SHANNON REALLY SAID ABOUT COMMUNICATIONS AND ITS IMPLICATIONS TO CATV.

Robert S. Burroughs

Panasonic Technologies
Communication Systems Technology Laboratory

ABSTRACT

HDTV and the application of fiber optics is causing engineers to rethink the way television pictures are being delivered to the home. In discussions about new delivery methods engineers occasionally will quote Shannon, on what the theoretical limitations are for communication systems.

This paper reviews Shannon's fundamental theorems and premise, and postulates a method for evaluating communication systems. Based on the analysis, future digital CATV distribution systems are contemplated.

INTRODUCTION

The U.S. has had the same basic television transmission system for the past 50 years. It has served the U.S. well, but in the meantime modern communications theory has matured. We are now at a transition point where the television system, for possibly the next 50 years, is about to be decided. It would be a shame if the system of the future was constrained by the technologies of the past.

Several proponents are now proposing ATV systems for testing by the FCC's Advanced Television Test Center, (ATTC). Most of the proponent's systems are extensions of the existing NTSC system. In November of 1988 the System Analysis Working Party of the FCC, Systems Subcommittee for Advanced Television Systems, met for a week with the proponents of ATV systems, at a Days Inn, just outside of Washington D.C.. Each proponent presented their system to the committee. During one of the presentations the committee was not able to evaluate or analyze one of the proponent's system. The system did not look like a typical NTSC system, it was more digital in appearance. It was even remarked by committee members that they thought the system violated "Shannon's Limit". This comment struck me as odd, until I realized that I was in the company of primarily analog engineers.

After the meeting was over I contemplated the comment about "Shannon's Limit". If you want move from a classical communication approach to a modern approach you need to go back and review the fundamentals. And the fundamentals started with Shannon. It was then that I decided to go back to Communication Systems 101, and review "What Shannon Really Said About Communications ....".

SHANNON'S CONTRIBUTION

Shannon's fundamental theorem for a discrete channel with noise, (Theorem 11), is the basis by which all systems should be judged - it is the *ideal.*

Prior to Shannon's classic 1948 paper [1], "A Mathematical Theory of Communications", it was universally accepted that the accuracy of a transmitted signal was irrevocably altered by noise. This thinking was only natural. If random noise, n(t) is added to a signal, s(t), the result is a new signal r(t) = s(t)+n(t), which is also a random signal, for which an accurate replica of s(t) can not obtained.

Shannon, however, proved the contrary; a signal s(t) can be recovered to any desired accuracy, in the presence of noise N, if the bandwidth of the signal W is constrained and the signal magnitude S is restricted. Then the effects of noise can be combined with S and W in a parameter called "Channel Capacity" C, in the following form :

$$C = W \text{ LOG } (S+N/N) \qquad (1)$$

This shows that the rate,

$$W \text{ LOG } (S+N)/N \qquad (2)$$

measures the capacity of a channel for transmitting information. Shannon defined capacity C of a noisy channel as the maximum possible rate of transmission when the source is properly matched to the channel. He used a new measure of information which he called entropy, H to define Channel Capacity :

$$C = \text{Max } \{H(x) - H(y|x)\}, \qquad (3)$$

where the maximum is averaged over all possible information sources.

The implications of Shannon's Channel Capacity theorem were quite revolutionary to communication theory. Consider the situation where a number of message possibilities M increases as a function of the signal duration T, slowly enough so that;

$$M < 2^{(CT)} \qquad (4)$$

then, although perfect accuracy can not be attained, one can get arbitrarily as accurate as one wishes by choosing T large enough, by using sufficiently long signals. Shannon also showed the converse was true - reliable communications is not possible, regardless of signal-processing schemes, when

$$M > 2^{(CT)} \qquad (5)$$

For a source rate R < C it is possible to make the probability of an error in transmission as small as desired by properly choosing the set of :

$$M = 2^{(RT)} \text{ signals.} \qquad (6)$$

Or conversely, for a source rate R > C it is not possible to make the probability of an error arbitrarily small with any choice of T or any choice of signals.

The theorem is extremely general and is not restricted to Gaussian or discrete channels. Note that the theory does not say what form the transmitted signal should have or how one should go about finding signals which will achieve "Capacity" C.

Note the remarkable aspects of this theorem. If for any value of S, (signal power), greater than 0, a value of W can be picked such that one can transmit virtually error free messages at a rate, R < C. Or, theoretically , we can recover any reasonable signal buried in noise, given the proper code sequence. In actuality, close to these conditions exist in communication with deep space probes. Compare this situation with a typical S/N that is used to send TV pictures over a Cable system to the home. Current Cable targets are to get approximately a 50 dB S/N to the home vs. a 0 dB S/N used in space communications.

The price we pay for getting arbitrarily close to zero errors is long durations T. The consequences of this will be discussed in the section titled, "Geometrical interpretation of signals".

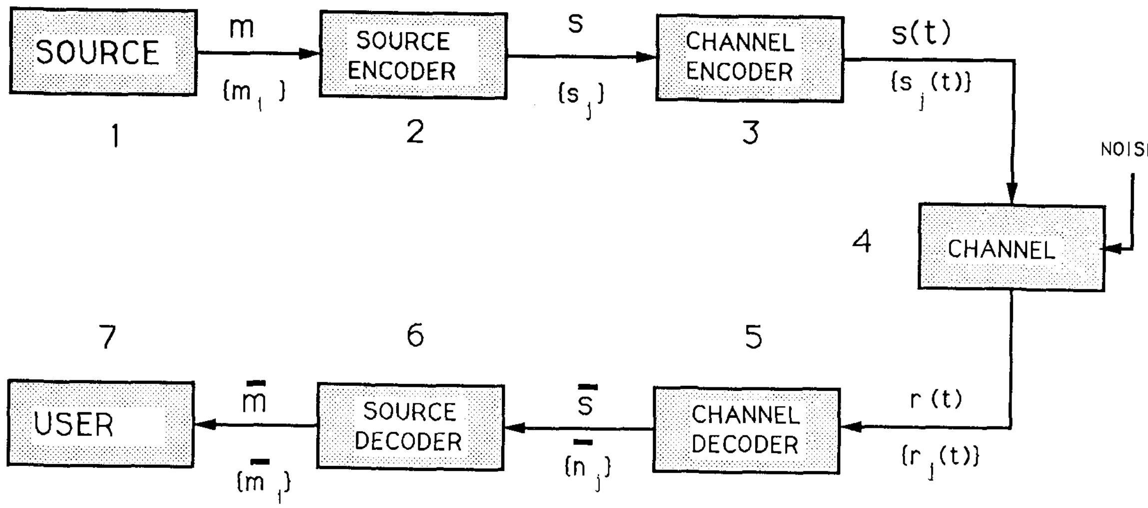
Even though most of the theory presumes discrete signals, Shannon showed that any continuous signal can be represented by a discrete source and the Channel Capacity theorem holds.

System block diagram.

A block diagram of a general communication system is shown in Figure 1. The first element is the 1. information source. The output of this source may be a sequence of discrete symbols, letters, or numbers in which case the source is referred to as a *digital source*. If the source output is a waveform or sequence of continuous valued variables, the source is referred to as a *continuous source*. In this paper, *digital sources* are the focus, since the concepts generally apply to *continuous sources* as well. An example of a *digital source* might be a sequence of "ones" and "zeros", or the text of this paper which is stored on the disk of my computer. These symbols could be expressed by the 128 symbols of the ASCII code. The essential feature of any source is that their output is generated by a *random* or *probabilistic* mechanism. This randomness is required, for if the output was known before the source generated it,there would be no need to communicate the source output to anyone.

The next block in the system is the encoder which is broken into the, 2. source encoder and the, 3. channel encoder. The reason for this separation is because of the different coding requirements required for the source and the channel.

The 5. channel decoder, 6. source decoder, and 7. destination perform the inverse operations of 1,2, and 3. The difference is that the received signals are only approximations to what was sent. Block 4. Channel is the particular medium used such as fiber, wire, free-space, etc. It is also the point where external noise is introduced to the system.



Block diagram of communications system with encoder and decoder each split into two parts.

Figure 1.

INFORMATION AND ENTROPY.

Although the Channel Capacity theorem in the presence of noise, is Shannon's main contribution, he is also responsible for his insight and pioneering work into the definition of Information and its subsequent application to the communications problem. Many of Shannon's concepts were not totally new, but he brought a fresh approach to explaining the fundamental concept of communication - "what is information and how best can one communicate it" ?

Information can have at least three levels of meaning :

1. Technical: how accurate can symbols be communicated ?

2. Semantic: how precise is meaning of symbols communicated ?

3. Effectiveness: how effectively does received meaning affect conduct in desired way?

Shannon concentrated on the technical level, even though the generality of his results also apply to levels 2 and 3.

The use of the term R, rate, and message possibilities, M were used to define "Channel Capacity". Shannon also uses the word *information* in a very special sense that should not be confused with meaning. And R is the rate at which *information* can be communicated.

Shannon once stated that the "semantic aspects of communication are irrelevant to the engineering problem". Note that the opposite is not necessarily true.

Information

When one message is selected from a set of possible messages, the information produced when this message is chosen can be quantified, (under certain conditions). As suggested by Hartly and Nyquist the logarithmic function is a convenient measure to use.

The meaning of "message" is quite general. Message can be a simple yes or no , (1 or 0), or a message can be a two hour television program. The information contained in a message of two possible choices is one (1), because of our choice of logarithms to measure information:

$$\log_2(2) = 1 \quad (7)$$

If we had 4,8,16,... choices, the information would be 2, 3, 4, ... bits respectively.

The content of information is typically measured in bits, a contraction of "binary digits".

For a typical communications source, we do not make a single choice, but a series of choices, one following the other as letters in a word or words in a sentence.

Shannon used a probability measure to define information. Using probability theory he defined three types of information :

1. Mutual information -

$$I(x,y) = \log P_{x|y}(a_k|b_j)/P_x(a_k) \quad (8)$$

the information provided about the event $x = a_k$ by the occurrence of the event $y = b_j$.

In terms of Figure 2, event $x = a_{1or2}$, $y = b_{1or2}$

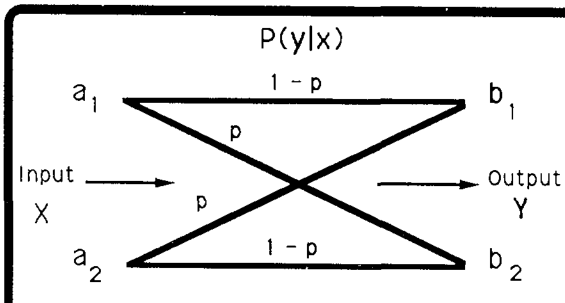


Figure 2 Binary symmetric channel.

1-p is probability that output letter is replica of input, and p it is the opposite of the input letter.

2. Self information -

$$I_X(a_k) = \log [1/P_X(a_k)] \quad (9)$$

the mutual information required to specify $x = a_k$

3. Conditional self-information

$$I(x|y) = \log 1/P_{x|y}(a_k|b_j) \quad (10)$$

the self-information of an event $x = a_k$, given the occurrence of $y = b_j$.

Self information, mutual information, and conditional self-information are all random variables.

Entropy

The entropy of an ensemble (x,y) is defined to be the average value of the information, or in the case of Self-information :

$$H(x) = \sum_{k=1}^{K} P_x(a_k) \log 1/P_x(a_k) \qquad (11)$$

The average mutual information between x and y is the difference between the entropy of X and the conditional entropy of X given Y or :

$$H(x) - H(X|Y), \qquad (3)$$

the form used in Shannon's coding theorem. Where $H(x)$ is the average information of the source x and $H(x|y)$ is the average information required to specify x,(input), after y, (output), is known. Or $H(X|Y)$ is the uncertainty in y as to which x was transmitted. Shannon refers to this uncertainty as equivocation.

Shannon defined "entropy", similar to the thermodynamic definition which connotes the random character of nature. (Shannon once said that the mathematician John von Neumann urged him to use the term entropy, since no one really knows what it means, Shannon would have an advantage in debates about his theory.)

Although an understanding of the mathematics of entropy is not essential to the purpose of this paper, a simple explanation is warranted. If from the, 1. source , we have a set of n independent symbols or messages, whose probabilities of occurrence are $p_1, p_2, \ldots, p_n$, then the corresponding entropy is:

$$H = -[p_1 \log p_1 + p_2 \log p_2 + \ldots + p_n \log p_n]$$

or

$$H = -\sum_{\text{all } i} p_i \log p_i .$$

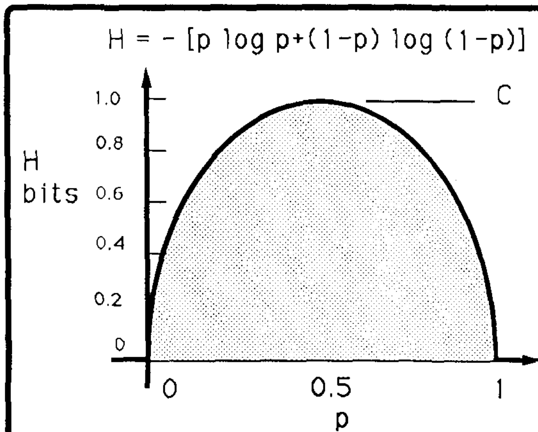


Figure 3.

Entropy in the case of two possibilities with probabilities p and (1-p)

Figure 3 is an example of the entropy of the binary symmetric channel of Figure 2. The maximum entropy can be seen to occur for the case where the input probabilities p_i are all equal, or the most uncertain condition. The condition of zero entropy is when one of the two p_i is certain. The implications are that when the inputs are know for certain, then no information is communicated, or that when the inputs are equally likely then the maximum transfer of information occurs and the channel achieves capacity, C. The first condition is obvious, if we know the inputs with probability 1, then there is no need to transmit them. The second situation is paradoxical. The maximum information is transferred when the inputs are equally likely, or completely random, but this is similar to the definition for noise. So, when is the transmitted signal, information and when is it noise ? I leave this paradox to a later paper, after I have had a chance to by some books newer than 1968.

DISCRETE SOURCES.

Although Shannon showed that the "Channel Capacity" theorem applies equally well to continuous time functions, it is the concept of a discrete source which is fundamental to the development of the theorem. The source outputs individual messages, each message a point in the message space of all possible messages. The transmitter, from a geometrical standpoint, maps the message space into the signal space. Many of the possible messages in the message space are redundant and do not convey any more information, than if they were not transmitted.

If we want to represent a continuous signal s(t) as a discrete message, we can represent this signal, if it is limited in frequency, to no frequencies greater than W_o hz.; then s(t) can be exactly represented by taking 2 W_o independent samples. This is the well know sampling theorem. These independent samples are discrete in time, such that a finite number of sample values are needed to define s(t) over a period T seconds. To each sample, if we can assign a number, we have an ensemble of numbers which are sequentially tied together. We now have a discrete digital source. But, unless we are very lucky, we can not always assign the exact value to the sample. We can, however, assign a value which is as close to the actual value of s(t) as we wish. This is the concept of quantization. Typically the accuracy we assign to this value is some multiple K of the RMS noise voltage N_o at the input to the decoder, Figure 1, 5. Channel decoder. This defines the resolution to which we have approximated the signal s(t) to a value:

$$q = KN_o. \qquad (12)$$

The important point is that we have defined the source s(t) to be a sequence of numbers

$$s_i(1), s_i(2), \ldots s_i(n). \qquad (13)$$

We also have defined the rate at which these discrete symbols occur, 2W. We now have a set of numbers and the communication problem comes down to transmitting this series of numbers to the user as close to the original sequence as possible. Since these are just numbers, and not a fixed single signal, s(t), we can perform almost any mathematical operation on these numbers we desire, as long as from the resulting sequence of numbers we can decode the original sequence. This is the basis of modern communications theory. Note that in Figure 1, we are not restricted from using conventional "analog" modulation methods such as FM, AM, PM, etc. One possible use of coding would be for more effectively use of spectrum. Consider normal TV signals where most of the energy is located near zero frequency. We could develop a coding scheme which would select codes such that the frequency characteristics of the sequence would spread the energy more evenly across the band. This transformed set of signals would fit within the same bandwidth as the original s(t), but would distribute power more efficiently such that less peak power would be required by the transmitter and less distortion products in the channel. This technique would reduce any benefits gained by companding or pre and de- emphasis schemes such as are typically used for FM systems. This scheme would be a natural scrambling scheme for CATV signals with the added advantage that it could improve system performance rather than reduce performance.

It should be emphasized that digital does not mean binary as is generally assumed. Digital, in communications theory, means that the source is a discrete source.

GEOMETRICAL INTERPRETATION OF SIGNALS.

A set of three numbers can always be used as the co-ordinates of a point in three dimensional space. In mathematics the concept of n-dimensional space is common. Similarly we can use the 2WT sample values, from the sampling theorem, to be the co-ordinates of a 2WT dimensional space. All of the points in this 2WT space represent all of the possible messages of length 2WT samples.

The size of this space is quite large. For a typical television program lasting an hour, with a bandwidth of 6 mhz. this space will have about

$4.3\text{x}10^{10}$ dimensions. And the total possible messages which can be transmitted in this space will fill all of the possible points in the 2WT dimensional space.

When considering the length of a sample T, simple PCM systems do not use more than one word of 7 to 10 bits. The previous example of T equal to one hour would make for an extremely complicated and slow system. But, Ts which encompass several symbols are quite common in communication systems which work with very low signal to noise ratios.

The importance of this representation is that the mathematics of geometry can be used in discussing and solving communications problems.

If the co-ordinates of this space are at right angles, (orthogonal), then the distance from the origin to one of the points can be interpreted as 2W times the energy of the signal,

$$d^2 = 2WE \qquad (14)$$

$$= 2WTP$$

where P is the average power over the time T.

When noise is added to a signal, this corresponds to a new point in the space which is proportional to the RMS value of the noise.

Different co-ordinate systems can be used. A specific co-ordinate system which is used in many communications problems uses sines and cosines, such as used in the Fourier series expansion.

In modern communication theory, the vector representation of signals is typically used, [4]. In the theory, a set of orthonormal functions is selected. Each waveform $\{s_i(t)\}$ will be completely

determined by a vector and its coefficients :

$$s_i = (s_{i1}, s_{i2}, \ldots s_{iN}); \quad i=0,1,\ldots,M-1 \tag{15}$$

We now have M vectors $\{s_i\}$ defining M points in an N dimensional vector space, called the signal space, with N mutually perpendicular axes. If the set of unit vectors defining the space are

$x_1, x_2, \ldots, x_N$, then the signal

can represented as:

$$s_i = s_{i1}x_1 + s_{i2}x_2 + \ldots + s_{iN}x_N. \tag{16}$$

The key benefit to being able to visualize transmitter signals geometrically is illustrated in Figure 3, which shows four signals in a two-dimensional signal space.

The points s_0, s_1, s_2, s_3,

are all a distance:

$$d = E_s^{1/2} \tag{17}$$

from the origin, where

$$d = \int s_i^2 \, dt \quad i=0,1,2,3 \tag{18}$$

is the energy dissipated in a 1-ohm resister if the voltage is $s_i(t)$.

Note that X_1 and X_2 are:

$$X_1 = (2/T)^{1/2} \sin 2pi\, f_o t \tag{19}$$

$$X_2 = (2/T)^{1/2} \cos 2pi\, f_o t \tag{20}$$

respectively for

$0 < t < T$, and 0 elsewhere.

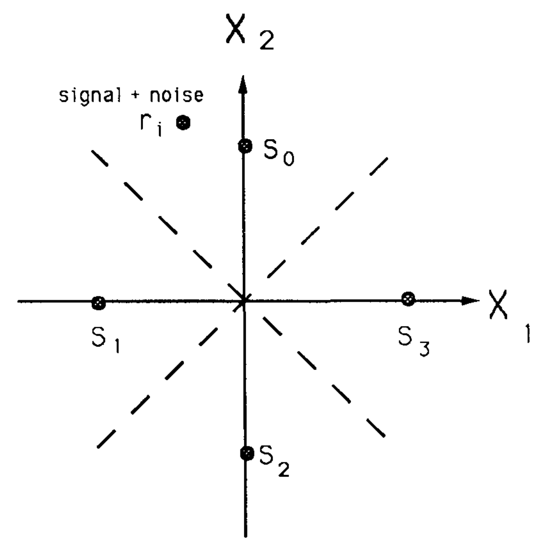


Figure 4 : Bi-orthogonal signal set

$d = \sqrt{E_S}$

(distance of signal point from origin)

Vector diagrams, of the type shown, are convenient ways of modeling signals for design-O*p+90Xmitted. ing various signal sets for transmission. From these diagrams the optimum decoding algorithm can be developed. Point r_i, in Figure 4, represents a signal s_i to which noise has been added. Equally spaced regions have been defined by the dashed lines. The signal points enclosed within these dashed lines define the respective optimum decision regions for each signal, when the a priori probabilities of the input signals are not know. This is known as a *maximum-likelyhood* receiver. If the received signal vector r_i is in the region associated with a particular s_i, then the s_i in that region is selected as the signal transmitted. It can be shown that choosing in this manner is optimum in the sense that the probability of an error is minimized.

In the general case, if the a priori probabilities are known, then the shape of the decision regions are shaped such that the optimum receiver, on observing the received vector r, sets the estimate of the received signal m = m_k whenever the *decision function* :

$$P[m_i]\ p_r(r|s=s_i)$$

is maximum for all i = k.

In the case where the input probability functions are known, the symmetrical placement of the signal vectors, s_i, are
not optimum. This could imply that a specific receiver might readjust the *decision function* depending on the specific television program to be transmitted.

COMPARISON OF TELEVISION TRANSMISSION SYSTEMS FOR CATV.

We are now armed with tools to evaluate various systems to be used for transmitting A/V signals over a CATV network.

Motivation

Why should we not just keep with the current methods ? Primarily , as was stated previously, we have had essentially the same television transmission system for the last 50 years. The new ATV and HDTV television systems being proposed are changing the requirements for transmission systems. Unlike the early days of television, where there was only the broadcast channels, there are now several television delivery methods capable of delivering the new ATV pictures - DBS, fiber telephone systems, video tapes, CATV, etc.. The media which is capable of providing the best possible pictures, with the most consumer convenience, at competitive prices, will have an edge in the competitive market. But, if traditional analog approaches are used, for new ATV systems, compression of a 30 mhz. HDTV baseband signal to a 6 mhz. band could be stretching the capabilities of the channel. This in itself is not bad, because there are several other channels available with considerably more bandwidth available, and the consumer will choose the one they like best. The near term dilemma is the fixation on making the new television systems compatible with existing receivers, as opposed to the opposite, making the old NTSC system compatible with the new HDTV receivers. Cable has a long tradition of making interface boxes to consumer electronics equipment, for the purpose of bringing more services to the consumer.

SYSTEM EVALUATION.

How can different television transmission systems be compared on an equal basis ? How efficiently is the transmission channel utilized ? What is the net effect at the television receiver, in terms of picture quality as measured by S/N and cost ?

What are the physical limitations of the channel and what is an ideal model by which we can use to measure all systems? Of course, Shannon's "Channel Capacity" comes to mind.

To test this method, a QPSK system and a typical FM system, used for "Super Trunking", are compared to illustrate how the concepts of information theory and "Channel Capacity" can be used to rate the two systems.

Analog systems do not lend themselves well to discrete analysis, primarily because, once a particular discrete model is selected for the analog system, changing parameters of the discrete system can dramatically alter the fundamental characteristics of the modeled analog system, such that it may no longer be the same system. To get around this difficulty, the analog systems will be compared to an equivalent digital system of the same bandwidth and information transmission characteristics at the receiver. The comparison will then be made between CNR required in the channel, with the *channel encoder* of the digital system being selected to give approximately the same bandwidth in the channel as the analog system. The difference in required CNR will be an indication of how well each system utilizes the channel. Or more specifically how much power we need in order to get a specific level of performance.

An example is given to illustrate the method. Most of the calculations are approximations. The important aspects are the trends and relative magnitudes, not the exact numbers.

FM System:

This system is designed to use 40 mhz. channel bandwidth, 65 dB video SNR, 35 dB channel CNR, and it also uses pre- and de-emphasis which accounts for 12 dB increase in video SNR.

In the comparison, 12 dB is subtracted from the video SNR because the digital system used, did not use this technique, but it could achieve the same effect by using variable sample sizes. This would complicate the calculations and would not give any more insight into the comparison.

Digital System:

A digital system is selected to match the parameters of the FM system.

Sampling rate of 12 Mhz. Although this is lower than what is typically used for digitizing video signals, much lower rates can be achieved by pre and post processing and using statistical sampling methods.

A 7 bit word was selected to give a video SNR of 53 dB. This is about the same as the FM system with 53 dB, when the 12 dB of pre-emphasis is subtracted.

This then gives a channel bit rate of:

$$7 \times 12 = 84 \text{ mhz.}$$

We can use a bandwidth equal to the bit rate.

We then select a Bi-Orthogonal signal set, (QPSK) to reduce the channel bandwidth to 42 mhz. If the bandwidth of the FM system were greater, giving more video S/N, we would use more bits in our word. If the FM system used less bandwidth, with the same S/N, then we would use a higher dimensional signal set, such as an 8-phase system.

<u>Information comparison.</u>

The information communicated by both systems is the same since the video S/N and the video bandwidth were chosen to be the same. The Channel Capacity is determined by the dimensionality W, and the maximum CNR of the channel.

If the FM parameters are used, it is seen that the Capacity for this channel is :

$$C = W \log_2(1 + CNR)$$

$$= 40 \text{ mhz. } \log_2(1 + 56)$$

$$= 233 \text{ million bits.}$$

$$(35 \text{ dB} = 56)$$

At the receiver :

R (rate of information) =

$$4.2 \text{ mhz. } \log_2(1+ 128)$$

$$= 29 \text{ million bits}$$

An efficiency can be calculated:

$$R/C = 29/233$$
$$= 12.4\%$$

For the QPSK case selected, R is the same as in the FM case, since it was chosen that way. The C, however, is different since less CNR is required in the channel.

$$C = 42 \text{ mhz. } \log_2 (1+25)$$

$$= 197 \text{ million bits.}$$

$$R/C = 29/197$$
$$= 14.7\%$$

The percentage difference is slight, but the digital system was constructed to match the equivalent FM system with a corresponding 7 dB_v less power required in the transmitter. This means that lower cost lasers can be used, or signals can go twice as far, or one more level of splitters deeper into the fiber system can be accommodated.

	FM	QPSK
CHANNEL CHARACTERISTICS		
Carrier to Noise (CNR)	35 dB	28 dB (assumes P{e} = 1x 10^{-7})
Band Width	40 mhz.	42 mhz.
Capacity	233 m bits	197 m bits
RECEIVER CHARACTERISTICS		
Video signal to noise (SNR)	53 dB	53 dB
Video bandwidth	4.2 mhz.	4.2 mhz.
Information rate of video - (R)	29 m bit	29 m bit

The key point is that if the 7-bit system is changed to an 8-bit system, the bandwidth will increase by 1/7 th., but the video S/N will double. This is the exponential tradeoff between bandwidth and SNR that is inherent in PCM and digital systems that is not present in analog FM systems, whose tradeoff is only linearly related. To get an equivalent performance increase in the FM system would require doubling the bandwidth.

S/N_{FM} = function of (log n)

S/N_{PCM} = function of n

(where n is key parameter which gives S/N improvement, deviation in FM, more quantization levels in PCM).

Note, that no compression or coding schemes were used to reduce bandwidth requirements or to improve P(e) of receiving the digital messages. both systems presumed white Gaussian noise. Typical CATV systems have higher levels of coherent noise such as cross-mod and inter-mod. The effects of coherent noise can be substantially eliminated in digital systems by coherent detection and the proper design of the encoders and decoders.

The sampling method can also be improved such that both the input sampling and output recovery are processed digitally.

CONCLUSION.

It is presumed that the entertainment delivery system of the future will have more bandwidth capability than coaxial systems, such as fiber or satellite.

Although it can be shown that certain FM systems can outperform certain digital systems, the choice for the future is clear. Even if cost is the primary reason for not going the digital path today it does not make sense not to make the investment in digital for the future. If the costs and performance of analog and digital systems are almost equal today, the greatest return on investment can be achieved with a digital approach for the future.

The digital CATV transmission system of the future will look much like the block diagram of Figure 1, where

The, 2. source encoder will:

reduce the redundancies in the source.

The 3. channel encoder will:

select transmission schemes, (like Bi-orthogonal signals),

add error correcting codes,

and add program encryption.

Many of the communications problems associated with digital systems have already been solved for the telecommunications, military, and aerospace industries.

The key to providing digital to the home isn't fiber optics to the home, it is digital inputs into the new ATV receivers. And just as in the past, Cable will provide digital to NTSC encoders for NTSC televisions and VCRs. And at the same time getting rid of many of CATVs signal quality problems such as ghosts, and various other modulation effects.

It now remains for the industries which deliver home entertainment to come into the 20 th. Century, before the 21 st. Century is upon us.

REFERENCES

[1] C.E. Shannon "A Mathematical Theory of Communications" Bell Systems Tech. Journal vol. 25 (Jul. 1948), pp 379 to pp 423, vol.(Oct. 1948), pp 623 to pp 656

[2] B.M. Oliver, J.R. Pierce, C.E. Shannon, "The Philosophy of PCM", IRE, 36, 1324-1331, November 1948.

[3] C.E.Shannon, "Communication in the Presence of Noise", Proceedings of the IRE, 37 No.1, 10-21, January 1949.

[4] J.M.Wozencraft&I.M. Jacobs, "Principles of Communication Engineering", John Wiley & Sons, 1967.

[5] R.G. Gallager, "Information Theory and reliable Communication", John Wiley & Sons, 1968.

[6] A.J. Viterbi, "Principles of Coherent Communication", McGraw-Hill, 1966.

[7] M. Schwartz, W.R. Bennett, S. Stein, "Communication Systems and Techniques, McGraw-Hill, 1966.

ISBN 0-940272-01-6; 0-940272-08-3; 0-940272-10-5; 0-940272-11-3; 0-940272-12-1; 0-940272-14-8; 0-940272-15-6; 0-940272-16-4; 0-940272-18-0; 0-940272-19-9; 0-940272-20-2; 0-940272-21-0; 0-940272-22-9; 0-940272-23-7; 0-940272-24-5; 0-940272-25-3; 0-940272-26-1; 0-940272-27-X; 0-940272-28-8; 0-940272-29-6; 0-940272-32-6; 0-940272-33-4; 0-940272-34-2; 0-940272-35-0; 0-940272-36-9; 0-940272-37-7; 0-940272-38-5; 0-940272-39-3; 0-940272-40-7; 0-940272-41-5; 0-940272-42-3; 0-940272-43-1; 0-940272-44-X; 0-940272-45-8; 0-940272-46-6; 0-940272-47-4; 0-940272-48-2; 0-940272-49-0; 0-940272-50-4; 0-940272-51-2; 0-940272-52-0; 0-940272-53-9; 0-940272-54-7