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A COMPATIBLE IN-BAND DIGITAL AUDIO/DATA DELIVERY SYSTEM

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0. Abstract

A system for adding digital data to an M-NTSC 6 MHz broadcast channel is described. The data capacity is sufficient to carry high quality digital stereo audio. The new system is shown to be compatible with the existing BTSC stereo system.

I. Introduction

The past few years have seen a rise and fall in the level of interest in high quality digital audio systems for cable television. At the 1983 convention a single paper was presented describing the basics of an efficient digital audio coding method.[1] In 1984, the final form of the ADM coding method was presented[2], and several papers either described the desireablility of digital audio, or prototype systems. 1985 saw more descriptions of prototype systems, and discussion of the advantages of digital audio. The '86 convention saw little discussion of digital audio; the emphasis was decidedly on making BTSC work, although one paper did suggest the replacement of the FM audio carrier with a digital audio carrier[3]. All of the prototype systems seemed to have faded away. There are two reasons for the fading of interest; the practical reality of having to deal with BTSC, and the impracticality of the digital audio system prototypes.

Digital audio offers the advantages of transparency and encryptability. Transparency means that the audio quality is unaffected by its travel from the cable headend to the subscriber (assuming all is working correctly). Minor effects like noise, IM distortion products, and reflections which would have a degrading effect on analog audio will have no effect on the digital audio (assuming they are within reason!). It is a simple matter to encrypt digital audio for program security.

The disadvantage of digital audio is that a lot of data must be put somewhere on the cable. Most systems have put the data out-of-band, at some frequency away from the associated video channel. This requires additional tuning, and channel tags for one tuner to follow another, etc. Some systems have put the data in-band, by utilizing the horizontal interval of the video signal. These systems are not compatible with normal receivers (they will no longer sync) so all subscribers have to be outfitted with new equipment. In order to achieve high quality, digital audio systems have had to use a very high data rate so that much bandwidth has been required. High precision D-A converters and sharp cutoff low pass filters required with PCM coding have increased the cost of systems.

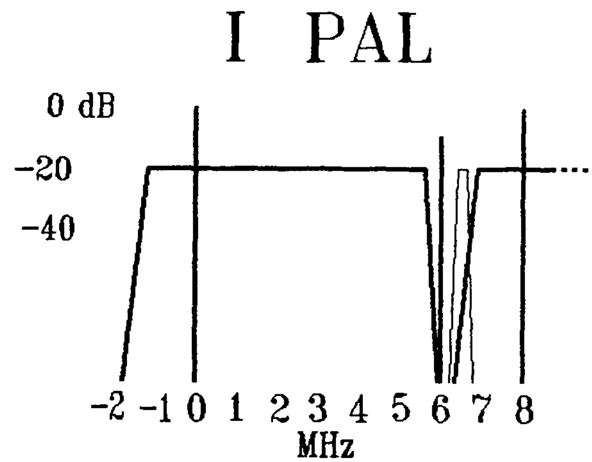
The system described here is in-band and compatible. The system may be used for broadcast or on cable. Stereo digital audio is incorporated into every 6 MHz video channel without removing (or even impairing) the BTSC audio signal or altering the sync interval. ADM audio coding is used, and the cost of incorporating reception hardware into a tv set is on the order of $10.

The BTSC stereo system which is now being incorporated into many CATV systems is theoretically capable of reasonably good performance. Unfortunately, the realities of intercarrier sound prevent this potential from being realized. The output of a BTSC stereo decoder invariably contains numerous 'birdie' and 'buzz' products. BTSC stereo quality is markedly inferior to all other common consumer audio formats: compact disc, compact cassette, FM radio, analog disc, VTR and video laser disc. The quality of BTSC is not a major issue at present. After all, it is "STEREO", and some tests have shown that consumers can't hear anyway[4]. The adoption of a stereo tv system has stimulated improvements in all areas of tv sound production. More care is taken during the production of stereo audio, and the use of audio noise reduction is becoming more common on the studio VTR. Eventually, the better quality source material and the quality conscious consumer will begin to meet in mass, and the BTSC stereo system will become a very noticeable limitation.

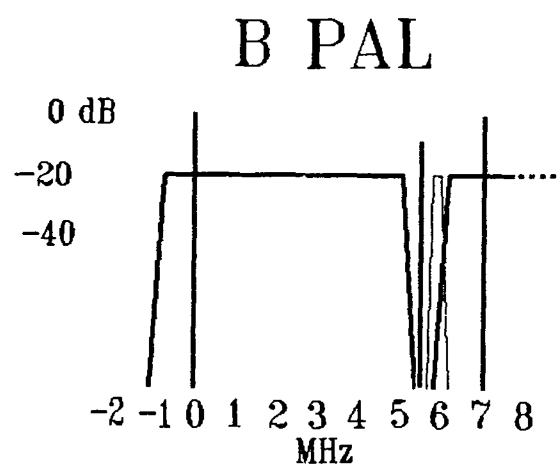
In 1983, the BBC conducted a series of tests at Wenvoe in South Wales of an experimental digital stereo sound system for television system I-PAL[5] (fig. 1a). The tests showed that the QPSK digital carrier which had been added was rugged, and could be received even under adverse reception conditions. In 1984, tests conducted in London showed good compatibility with existing tv receivers. Further tests have since been conducted in Hong Kong (I-PAL), and in Stockholm and Helsinki with system B-PAL (fig. 1b). The narrower bandwidth of the B-PAL broadcast signal makes it more difficult to insert the digital sound carrier without creating interference. In order to avoid interference, the Scandinavian broadcasters were attracted to the Dolby adaptive delta-modulation coding method which allows a 30% bandwidth reduction compared to a companded PCM system. Dolby laboratories became involved in supplying sound coding and digital multiplexing equipment for broadcast tests in Hong Kong, Sweden, and Finland. While observing the work being done in Scandinavia, and comparing the B-PAL broadcast spectrum to that of M-NTSC (fig. 1c), it became apparent that the results of those tests should be applicable here in N. America. Both the B-PAL and M-NTSC signals have the same 750 kHz spacing between the FM sound carrier and the corner of the adjacent channel lower sideband. Work thus began to determine whether a digital carrier could be included into the broadcast M-NTSC signal without creating interference to either the existing vision or BTSC sound signals.

Fig. 1 RF Spectra of TV Systems showing additional QPSK data signal.

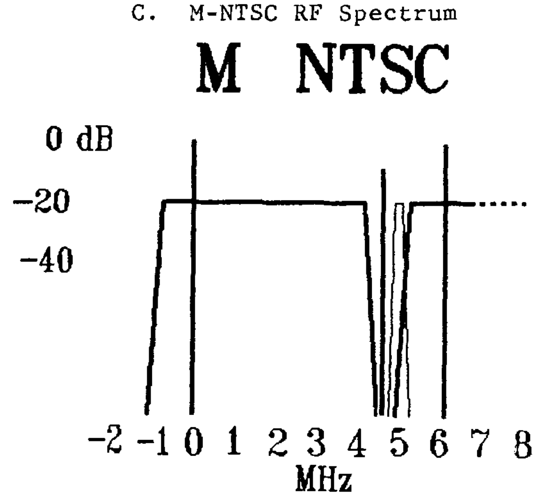
A. I-PAL RF Spectrum



B. B-PAL RF Spectrum



C. M-NTSC RF Spectrum



II. Basics of a New System

Based on work done in Scandinavia, and that reported on here, these are the tentative parameters of the new system:

Carrier Frequency	4.85 MHz above vision carrier
Carrier Level	-20 dB relative to vision carrier
Modulation	Differential QPSK
Bit Rate	512 kbits/sec
Audio Coding	Adaptive Delta-Modulation

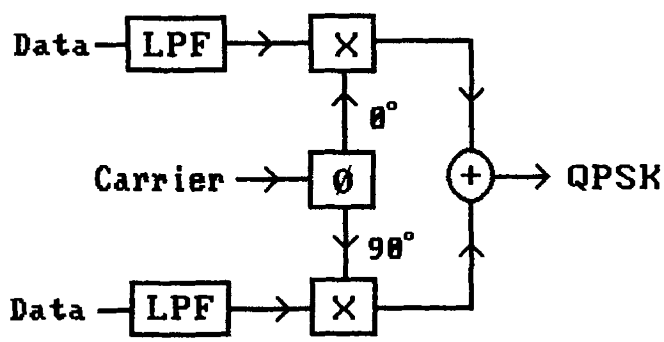
The most important requirement for the new system is that of compatibility; the new digital carrier must not interfere with the existing vision or sound signals. The spectrum of the new signal will resemble a band of noise since the transmitted data will appear random (if it isn't naturally random, scrambling with a pseudo random sequence will make it random). Any interference will appear as noise. There is the potential for interference into the BTSC stereo and SAP signals, and into the adjacent channel vision signal. The carrier frequency and amplitude are the equivalent to those chosen for use with B-PAL. The BTSC signal has a wider RF bandwidth than the mono FM signals in either I or B PAL, and thus there is more potential for interference from a new digital carrier here in the US than in Europe. In order to minimize the interference potential, the new carrier should have the minimum bandwidth necessary to convey stereo audio. A data rate of 512 kb/s has been chosen. Using an advanced form of adaptive delta-modulation coding, this rate is sufficient for two very high quality audio channels, along with synchronization and mode signalling bits, and 64 kbits/s of auxiliary data.

There are a large number of digital modulation methods available which range from simple and inefficient to complex and highly bandwidth efficient. There are inherent tradeoffs between efficiency, ruggedness, and complexity. A good compromise between these factors is offered by quadrature phase-shift keying (QPSK). Bandwidth efficiency can approach 2 bits/Hz and IC chips to demodulate QPSK are now available from at least two sources.

III. QPSK Modulation

QPSK modulation involves sending a carrier which takes on one of four phase states (0, 90, 180, or 270) degrees during each data symbol period. Since there are four states to choose from, each data symbol carries two data bits. It is perhaps easiest to consider a QPSK modulator as a pair of bi-phase modulators working in quadrature as shown in figure 2. Each of these modulators handles half of the data, or 256 kbits/s.

Fig. 2
QPSK Modulation

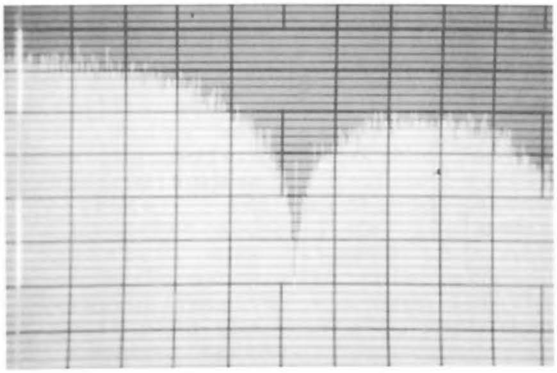


A digital data stream is low pass filtered to constrain the data spectrum. The spectrum of a random 256 kbits/sec data stream is shown in figure 3a. The spectrum extends to infinity with nulls every 256 kHz, and repeating lobes of diminishing amplitude. Only the first two lobes are shown. In theory, it is only necessary to transmit half of the main lobe, or up to 128 kHz (achieving 2 bits/Hz bandwidth efficiency). In practice it is necessary to send somewhat more, and a parameter known as Alpha specifies the fractional excess bandwidth. In figure 3b, the effect of two different data filters are shown; Alpha = 0.3 (the narrower spectrum) and Alpha = 0.7. The sharper the filtering used, the narrower the RF spectrum which will be transmitted and the less interference which will be caused.

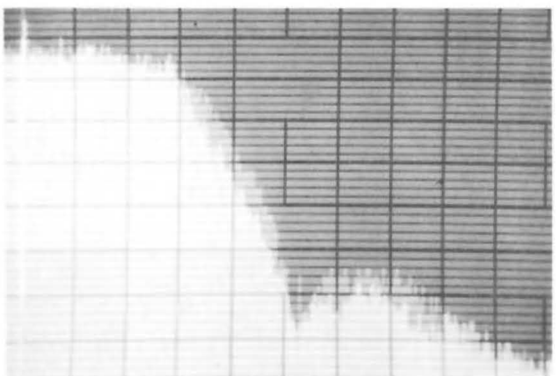
Sharply filtering the data has the undesirable effect of making the system less tolerant of some transmission impairments, and requires tighter tolerances on circuit elements. This is due to the time domain effect of filtering on the transmitted digital pulses. Figure

Fig. 3 Baseband data spectra of 256 kb/s psuedo-random data. Vertical scale 10 dB/div. Horizontal scale 50 kHz/div.

A. No filtering.



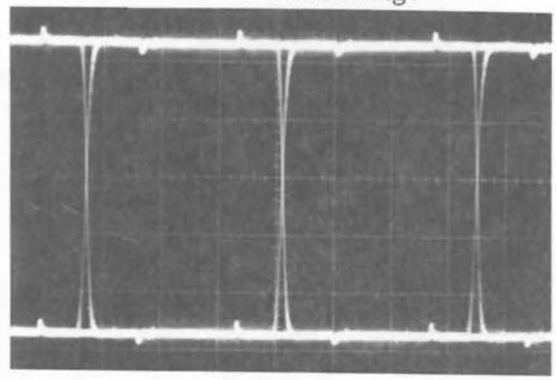
B. Alpha=0.3 (bright) and Alpha=0.7 filtering.



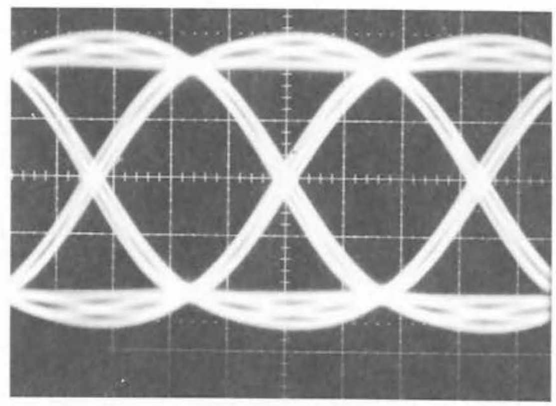
4a shows the 'eye pattern' (the eye pattern is the overlap of many pulses) of an unfiltered baseband data stream. This is what a decoder looks at in order to demodulate the data. This pattern has a 'wide open' eye and no data transition jitter. Figure 4b shows the effect of an Alpha = 0.7 data filter. The eye is now only maximally open at one point in time, and there is some data transition jitter. The eye pattern of an Alpha 0.3 filter is shown in figure 4c. The pattern is quite a bit more complex, the maximum opening is narrower in time, and there is a lot of data transition jitter. The reason that the eye pattern becomes more complex with tighter filtering is that each data pulse 'rings' for a longer time, and thus has the potential to effect the demodulation of more of the other pulses. With ideal filtering this ringing can be controlled so that it does not create a problem, but any deviation from the ideal degrades a narrower band system sooner than a wider band system. Another problem with narrow filtering is that the longer ringing pulses create more havoc if multipath is present, because more pulses can interfere with the pulse being detected. The complexity of the eye pattern correlates

Fig. 4 Eye patterns at data demodulator.

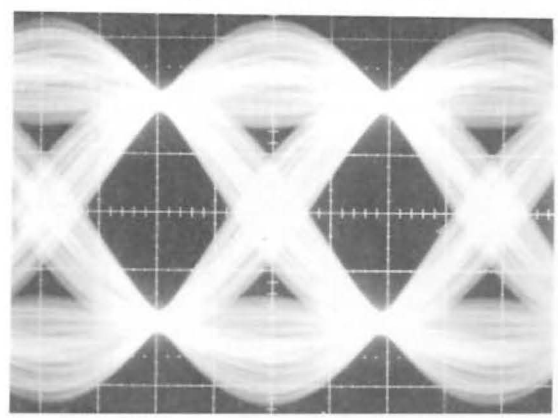
A. No filtering.



B. Alpha=0.7 filtering.



C. Alpha=0.3 filtering.



directly to the ruggedness of the data when passed through channels with imperfect amplitude flatness and non-linear phase response. The more complex eye pattern will degrade rapidly when exposed to any additional filtering, or effects such as multipath. For a consumer system, it is desireable to keep things simple and non-critical and thus the Alpha = 0.7 filter is preferred.

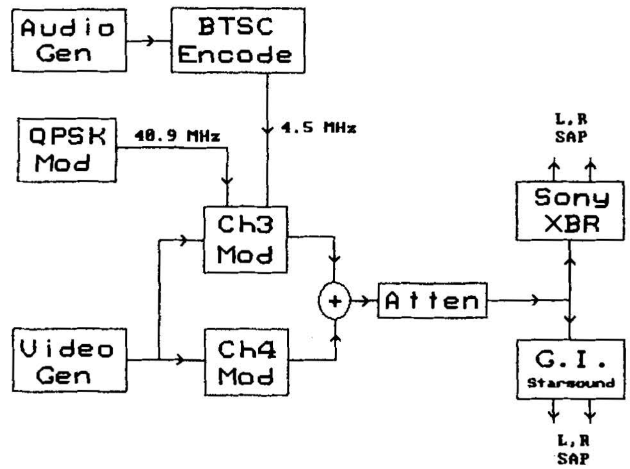
In the QPSK modulator of figure 2, one filtered data stream modulates a 0 degree carrier which results in a 0 or 180 degree phase shifted output. The original baseband data spectrum is shifted up to the carrier frequency and becomes doubled sided. Since each sideband contains the same information, half of the bandwidth is wasted. The other data stream modulates a 90 degree quadrature carrier. Its' output is either 90 or 270 degrees. When the two carriers are summed, the result will have one of 4 phase states. The two quadrature carriers occupy the same bandwidth, and the composite signal now has different upper and lower sidebands, thus fully utilizing the spectrum. The absolute phase is not available to the decoder. The decoder can only detect phase changes, so the data must be encoded differentially. The phase change from one data symbol to the next (a change of 0, 90, 180, or 270 degrees) carries the two bits of information. This requires some additional digital circuitry at each end.

IV. BTSC Compatibility Tests

The effect of the new signal on the BTSC stereo and SAP channels will be one of additive noise. This is due to the fact that the transmitted data spectrum is noiselike. The data multiplexer contains a pseudo-random noise sequence generator. All data is scrambled by the PN sequence and thus no matter what the original data spectrum, the transmitted spectrum is essentially noiselike. At the receiving de-multiplexer the same PN sequence is used to unscramble the data to recover the data. The interference into BTSC will be worst into SAP since the SAP signal spectrum extends farthest away from the 4.5 MHz FM sound carrier (the 'pro' channel is being neglected here). Interference will be less severe into the L-R signal, and still less into the L+R signal. Testing for BTSC compatibility involves measuring the additional noise caused by the new signal.

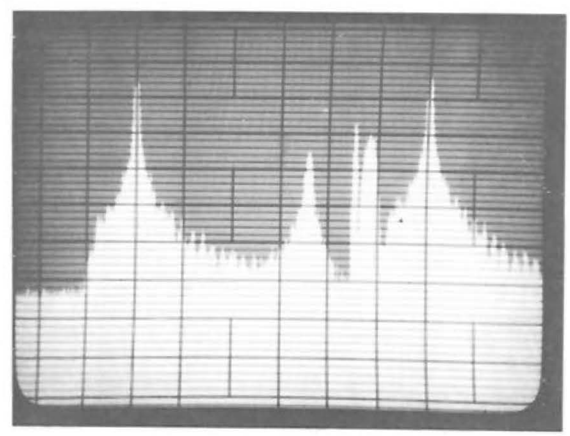
The test setup is shown in Figure 5. Adjacent channels 3 and 4 are generated by a pair of high quality CATV modulators (GI/Jerrold Commander IV with SAW filters). The RF level is controlled by

Fig. 5 Test Setup



an attenuator before entering the receivers. The tested receivers include a very high quality consumer tv set (Sony XBR) and the best quality BTSC decoder available (GI/Jerrold Starsound). Figure 6 shows the spectrum delivered to the receivers.

Fig. 6 RF spectrum of signal at receiver showing Ch3 vision carrier, color subcarrier, FM sound carrier, QPSK data carrier, Ch4 vision carrier.



Because the SAP and L-R channels are heavily companded, we can't simply measure the no-signal noise level to determine the extent of the interference caused by the new digital carrier. We must 'open up' the companding with a test signal and then measure the noise in the presence of the test signal. Since more companding is done at high frequencies, we use a test signal of 5 kHz at full amplitude. So that we are most sensitive to the QPSK caused interference we want to use the best reception equipment with a relatively high RF level. The RF level chosen for these tests was 4 mV RMS or +12 dBmV. Figure 7 Shows the SAP output spectrum of the Sony XBR with video modulation (color bars) on (upper trace) and off (lower trace). Since the audio is significantly degraded by the presence of video modulation, we will be able to see the effects of interference most easily if the video is left unmodulated. All BTSC compatibility measurements were performed with no video modulation. Figure 8 compares the performance of the Sony XBR (upper trace) to the GI StarSound receiver (lower trace). The Starsound receiver uses a separate detector for the L-R and SAP signals[6], and is able to achieve superior performance with this technique. The Starsound receiver was used for these tests as it gave the best BTSC performance of any receiver tested.

Figure 9 shows what happens when the QPSK signal is turned on (upper trace) and off (lower trace). Since the distortion components still dominate, it is not really feasible to measure noise with a SINAD (notch out the fundamental and read what's left) measurement. With an HP3561A FFT based analyzer it is feasible to set up a band noise measurement. The vertical dotted lines in Figure 9 delineate a frequency band from 6200 Hz to 9600 Hz. There are no distortion or 'birdie' products in this band, so the total energy in the band can be used as a relative noise indicator. The FFT analyzer will conveniently display the total energy in the band.

To test for interference from the digital signal we measure the noise in the band 6200 Hz to 9600 Hz as a function of:

1) QPSK carrier frequency [4.80, 4.85, 4.90 MHz above vision carrier]
2) Data filtering applied [Alpha = 0.7, 0.3]
3) QPSK carrier level. [-10 dB and lower, relative to vision carrier]

The absolute numbers obtained are not significant. What we are looking for is the amount of noise increase caused by the presence of the QPSK signal, which will be a measure of compatibility. If the noise is only increased a little bit, then the new signal is compatible. If the noise is increased a lot, then the new signal is incompatible. This judgement is a subjective one, and different judges will likely come to differing conclusions given the same data.

Fig. 7 Sony XBR SAP output. SAP modulated 100% with 5 kHz. Video modulated with color bars (top trace) and un-modulated (lower trace).

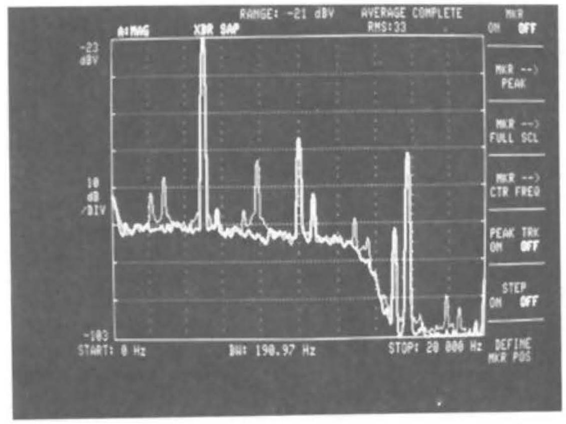


Fig. 8 SAP output of Sony XBR (upper trace) and GI Starsound (lower trace). SAP 100% modulated with 5 kHz. Video unmodulated.

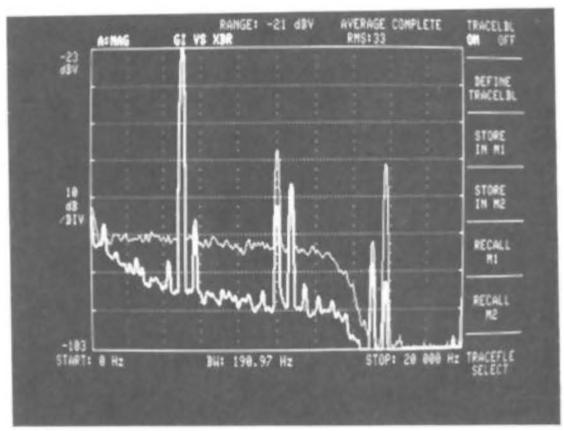
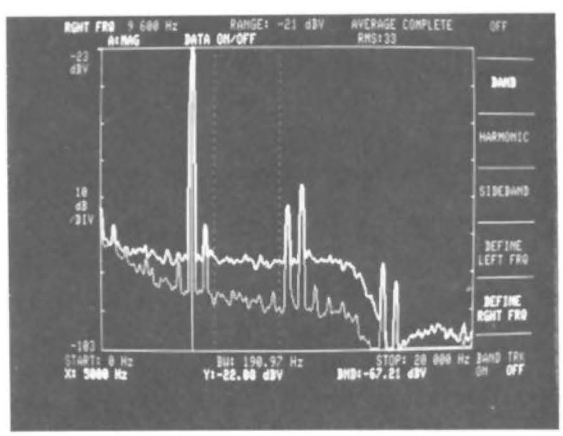


Fig. 9 SAP output of GI Starsound. SAP 100% modulated with 5 kHz. QPSK data signal on (upper trace) and off (lower trace).



The SAP channel test results are shown in Figure 10. As expected, the noise level is worst with the QPSK carrier closer to the FM sound carrier (4.8 MHz) and with the wider data filter (Alpha = 0.7). As the QPSK carrier level is reduced, the noise level drops. Using a narrower data filter (Alpha = 0.3) with the 4.80 MHz offset moves about half way to the curve of the wider filter (Alpha = 0.7) with a 4.85 MHz offset. The difference between Alpha 0.7 and 0.3 filtering would imply a narrowing of the spectrum by about 50 kHz, but since the filtering is split between the transmit and receive filters, the actual narrowing in the channel is more like 25 kHz. At the target operating point of 4.85 MHz, Alpha = 0.7, -20 dB, the degradation to SAP is approximately 5 dB. If the carrier frequency is lowered, the interference rises rapidly. 4.85 MHz appears to be the closest the QPSK carrier can be placed to the 4.5 MHz sound carrier. Operation at 4.90 MHz would only create about 2 dB of degradation, but may create problems with adjacent channel operation.

The 5 dB noise penalty must be put in perspective. The noise penalty only occurs when:

1) A relatively high RF level is used (+12 dBmV).
2) A special receiver is used (Starsound).
3) No video modulation is present.

Figure 11a shows the SAP with the QPSK on (4.85 MHz, -20 dB, Alpha=0.7) and with video modulation on (upper trace) and off (lower trace). Even with the noise made 5 dB worse, the output spectrum is still totally dominated by distortion products. This might lead one to question whether the extra noise is even audible. Figure 11b shows the same situation but with the video left on and the QPSK turned on (upper trace) and off (lower trace). The increased noise in the spectral holes between the distortion products is apparent and can be heard. It is not as objectionable at the distortion.

The penalty for the addition of the QPSK signal is a noise penalty in the SAP channel which is audible under ideal conditions. Under practical conditions the noise change would probably not be noticeable.

Of more concern is the change in noise level in the stereo signal. This is shown in figure 12. While the individual curves at the bottom are hard to make out, the significant finding is that if the carrier frequency is 4.85 MHz or higher, there is no effect on the BTSC signal. At the target operating point, the interference causes less than 1 dB increase in noise level.

Fig. 10 SAP noise level in band 6200 Hz to 9600 Hz. GI Starsound receiver. FM sound carrier level -15 dB rel vision carrier. SAP 100% modulated with 5 kHz. Video unmodulated. QPSK modulated data at 512 kb/s. QPSK carrier frequency at 4.80, 4.85, 4.90 MHz above video carrier. Data filtering Alpha=0.7, 0.3.

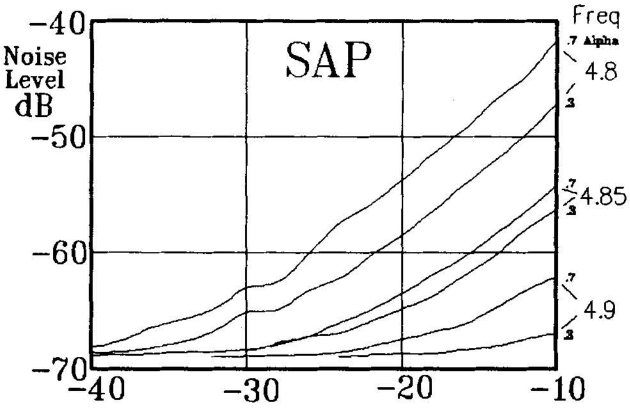
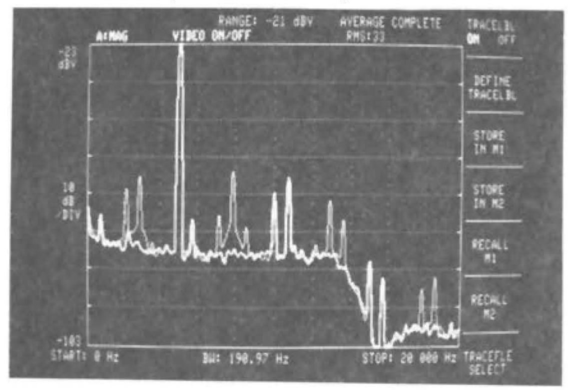


Fig. 11 SAP output of GI Starsound receiver. SAP 100% modulated with 5 kHz.

A. QPSK data on. Video modulated with color bars (top trace) and video un-modulated (lower trace).



B. Video modulated with color bars. QPSK data on (upper trace) and off (lower trace).

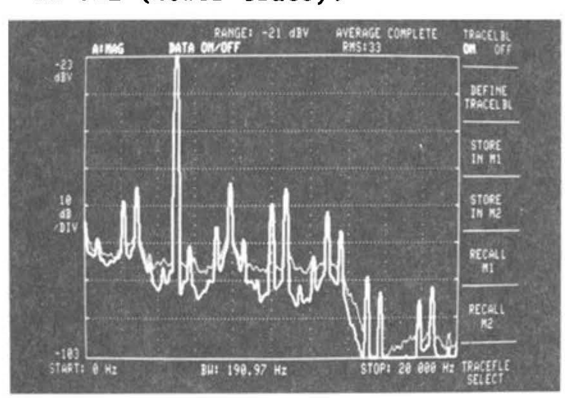


Fig. 12 Left channel noise level in band 6200 Hz to 9600 Hz. GI Starsound receiver. FM sound carrier -15 dB rel vision carrier. L channel 100% modulated with 5 kHz. Video un-modulated. QPSK carrier frequency at 4.80, 4.85, 4.90 MHz above video carrier. Data filtering Alpha=0.7, 0.3.

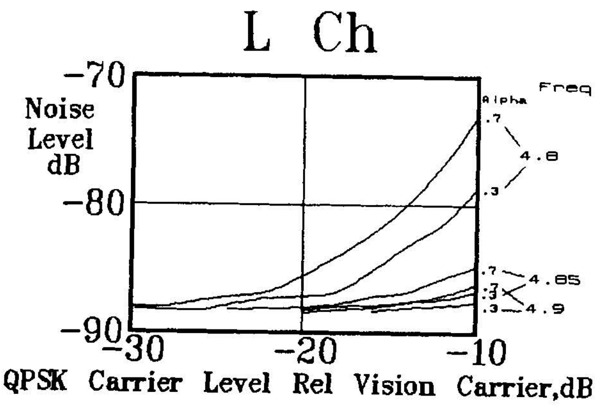
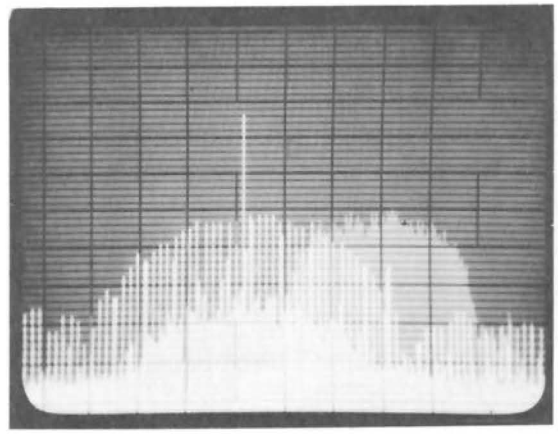


Fig. 13 Spectrum of pulsed FM sound carrier and QPSK data signal. 10 dB/div vertical. 100 kHz/div horizontal.



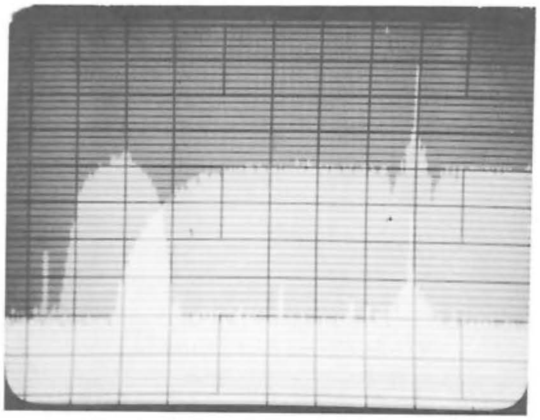
V. Compatibility with Video Scrambling

The new system has not been fully tested with video scrambling systems. Because of the low level and noiselike spectrum of the data signal, it is very unlikely that any additional interference into video or analog audio will occur. It is possible, however, that the techniques used in some video scrambling systems may create some interference into the digital signal. The worst case would be sync suppression scrambling systems which pulse the FM aural carrier to convey sync information to the descrambler. The spectrum of such a system is shown in figure 13, along with the QPSK data spectrum. The pulsing of the FM carrier level at the horizontal rate creates a comb type spectrum which is bandlimited by a band pass filter in the scrambler. The high side of the comb overlaps the QPSK spectrum. Even with this substantial overlap, the data was correctly received without errors, though the margin against error was significantly reduced. In order to handle this situation with sufficient margin against errors, it will probably be necessary to either raise the QPSK carrier level a few dB, or to filter the FM carrier with a notch type filter to attenuate some of the energy above 4.65 MHz.

VI. Adjacent Channel Video

There is the potential for interference from the upper adjacent channel video into the digital signal. There is an overlap of the upper adjacent channel lower vestigial sideband and the QPSK signal spectrum as shown in figure 14. This display was produced by modulating the upper adjacent channel video with a 100 IRE frequency sweep, and setting the spectrum analyzer to peak hold. The shape of the modulators' SAW vestigial sideband filter is revealed. The QPSK data spectrum is displayed on a background trace, and the bright area is the overlap. High level video signals with a frequency of 1.0 to 1.1 MHz do intrude on the QPSK spectrum. With this particular modulator this did not cause data errors, but the margin against error was reduced. It may be necessary to add some additional filtering of the lower sideband to assure adequate margin against error.

Fig. 14 Overlap of adjacent channel lower sideband and QPSK signal. 10 dB/div vertical. 200 kHz/div horizontal.



There is also the potential for the QPSK data signal to interfere into the adjacent channel video signal. TV receivers can be separated into two classes: those that can handle adjacent channel operation and those which cannot. Newer sets with SAW filters handle adjacent channels, and the presence of the digital carrier with its noiselike spectrum, without ill effects. Older sets do not yield excellent video quality with adjacent channel operation, and the appearance of a very slight bit of noise from the new signal is difficult to detect; the patterns and noise caused by the adjacent channel video, chroma, and FM sound carriers totally dominate. Older sets can give acceptable performance when used with a cable converter which contains filtering to remove adjacent channel energy. Since these filters have not been designed to remove the new digital carrier, they will only attenuate it a small amount. This is an area which will require a lot of additional study. It is probable that only tests on a number of real cable systems will reveal whether this is a significant problem. If only a few sets are affected, the addition of a trap at 60.10 MHz (for a Ch 3 output converter) in those installations would solve the problem.

VII. Conclusion

An in-band digital carrier may be added to every cable tv channel in a way which is compatible with the BTSC stereo audio system. Further tests are required to prove compatibility with adjacent channel operation with all tv sets and converters. Further tests are also required to determine if there is adequate margin against error when the FM carrier level is pulsed by video scrambling systems. The digital carrier may contain purely digital data, or a combination of one or two channels of high quality audio and axillary data. The audio may be program audio related to the video signal, or may be a completely unrelated audio program.

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A NEW OPTION IN SUBSCRIBER CONTROL

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INTRODUCTION

We introduce in this paper a new concept in subscriber control based on program denial outside the home. The technique is rooted in positive trapped systems, but is believed to offer improvements in both security and in quality of recovered signal. It is optimized for protection of services having relatively low but significant penetration.

Two novel ideas are presented. The technology used for program denial is new, as is the packaging technique. The packaging is not an integral part of the program denial technology, nor is the denial technology integral to the packaging. However, the two together do form a nice package for the intended application.

BACKGROUND

Early cable television systems transmitted only on the 12 standard VHF channels. Later non-standard channels were added and a set-top converter was used to convert them to a standard channel.

With the introduction of premium services, traps were added to the system as a way to control which subscribers recieved premimum services. Negative traps were used to remove premiums for basic subscribers, but carried a significant cost disadvantage when the pay to basic ratio was low. Positive trap systems overcame this cost disadvantage by requiring traps only at the homes of the premium customers (who were already paying extra). Positive traps, however, remove part of the desired signal along with the interfering signal. This resulted in poor picture quality and violated one of the first premises of cable television, which was the delivery of a good quality signal. Also, as their technology became known, easy defeats (twin lead and aluminum foil) were discovered by pirates.

In an effort to make premiums more secure, cable operators moved to more sophisticated scrambling techniques, placing the descrambler in the subscriber's set-top converter. This limited interface continues to serve well. Additional functionality is now required by the the video revolution, with its proliferation of remote controls, VCRs and multiple TVs. Several ideas have been proposed to ease the interface between the cable system and the cable subscriber. One new system is described here-in.

This system is based on a new and improved positive trap technology. It employs an interfering signal that is much harder to remove and a trap technology that removes only a small, redundant part of the desired TV signal. this makes compensation easier. Advantages to the cable operator include removal of equipment from the subscriber's premises, a lower cost when serving multiple sets, a capital cost proportional to pay subscribers, and subscribers who perceive a higher value to their cable service. Advantages to the subscriber include easier interface to multiple sets and VCR's and the full use features included in remotes and cable-ready sets.

The system removes the cable operator's equipment from inside the home and provides a broadband outlet to the cable subscriber. This broadband signal contains all purchased premiums descrambled simultaneously and is easily routed to all TV sets and VCRs.

INTRODUCTION TO THE NEW SYSTEM

Positive trap technology has been used in the CATV industry for many years. This technology is based on the simple concept of inserting an interfering signal within a television channel at the headend and removing that interference with some form of filter at the subscriber's home. The goal of the design of this filter is to remove the interference without producing perceptible artifacts in the recovered television signal.

This technique has several attributes that may be desirable to the CATV system operator. One of these attributes is that the secured channels are scrambled on the distribution plant. Secondly, capital investment is required for filters only at the subscriber locations where premium services are being purchased.

The conventional embodiment of this technology involves the use of an L-C notch filter that is located at the subscriber's home. A notch filter (or band reject filter) is a frequency selective device that attenuates a band of frequencies while passing all others. Unfortunately L-C technology has some inherent limitations when used for this application. In the positive trap scenario it is desirable to make the notch filter as narrow as possible so that the interfering signal may be removed without removing or significantly affecting the desired television information.

One figure of merit for quantifying the performance of a notch filter is Q. Q is defined as the ratio of the notch center frequency to its 3 dB bandwidth:

$$Q = \frac{F_o}{F_{3dB}}$$

See Figure 1. Typical maximum values of Q for a conventional L-C notch filter are around 30. This means that a trap for channel 7, near 177 MHz, will have a 3 dB bandwidth of about 5.9 MHz, essentially the bandwidth of a TV channel.

In order to minimize the effect of a notch filter on the recovered signal it is important to maximize the Q (i.e. decrease the 3 dB bandwidth). It is difficult to accomplish this goal with conventional L-C technology, especially as center frequency is increased. This constraint has limited the use of positive trap technology to the low end of the CATV spectrum.

Q characteristics limit the number of choices for placement of interfering signals in positive trap scenarios. In the conventional configuration the interfering signal is placed midway between the picture carrier and the sound carrier (see Figure 2). This area of the television channel was chosen due to the relatively low energy concentration occurring during normal television programming. Any information removed from this area of the spectrum has a minimal effect on the quality of the recovered signal.

Despite this location, L-C notch filters remove significant amounts of useful information so as to noticably degrade the quality of the recovered signal. Additionally this location makes it possible for the pirate to recover an acceptable signal using components that are readily available.

The design goal for the new system was to develop a positive trap that is significantly more secure and provides an improved recovered television signal when compared to the existing L-C notch system. The heart of this new system is an improved version of the notch filter. This filter has been developed utilizing SAW (Surface Acoustic Wave) devices as resonant elements in the notch filter design. The SAW devices provide filter Q's far in excess of that of L-C technology (around 450 with SAW based notch filters as compared with 30 from L-C based filters). The resulting filter has a minimal effect on recovered signal quality. The improved notch filter supports the placement of the interfering signal in a part of the television channel that makes it much more difficult for a pirate to remove the interference using techniques generally at his disposal, without corrupting the recovered signal to the point of being virtually unusable.

NEW NOTCH

The development of an improved notch filter has necessitated the development of a new and unique SAW device as well as the utilization of this SAW device in a new filter topology.

Figure 3 shows measured data comparing the response characteristics of a SAW based notch filter to a conventional L-C based notch filter. Both filters are centered close to 200 MHz. Typical performance characteristics of a SAW based notch are given below:

1.	SAW Notch Center Frequency	199.022 KHz
2.	3 dB Bandwidth	450 KHz
3.	6 dB Bandwidth	350 KHz
4.	40 dB Bandwidth	60 KHz
5.	Passband Insertion Loss	<1 dB Max. 0.5 dB Typ.
6.	Passband Bandwidth	DC to >800 MHz
7.	Frequency Drift vs. Temp from -40°F to +140°F	-15 KHz + 5 KHz

Passband characteristics are shown in Figure 4. The low passband insertion loss makes it possible to cascade several notch filters at various frequencies. Classically, one associates SAW bandpass filters with high insertion loss. The present notch filters are not based on the same principles and the high loss conditions do not apply.

LOCATION OF JAMMING CARRIER

The drastically reduced width of the notch filter allows placement of the jamming carrier much closer to the picture carrier. This makes the pirating job much more difficult, since simple traps which might remove the jamming carrier also remove the picture carrier and essential sidebands. While traps may be constructed which yield some sort of recovered signal, considerable degradation results. Figure 5 shows the location of the jamming carrier on the vestigial sideband of the protected channel. This location was chosen because it allows reasonable jamming coupled with reasonable ability to recover a quality signal.

Several conflicting requirements are placed on the jamming signal location. A location as close as possible to the picture carrier is desired in order to improve the robustness of the system against the onslaughts of the pirates. On the other hand, a location as far as possible from the jamming carrier improves the ability of the SAW trap to recover a quality signal. The placement chosen represents the best engineering compromise between these requirements. Three considerations led to the final choice of a jamming frequency, once the approximate limitations of the technology were determined.

In order to minimize the visibility of artifacts generated by incomplete attenuation of the jamming carrier (for a legitimate subscriber), the same technique is applied as is used to minimize effects of the color subcarrier beating with the picture carrier. The offset between the jamming and picture carriers is locked to an odd harmonic of one half the horizontal line rate. It can be shown that doing so causes any beat which does result, to be stationary on the screen, and a stationary beat is considered less visible than is a moving beat. A second consideration is that the jamming carrier, being equal in amplitude to the picture carrier, must fall in one of the required offsets should the protected carrier be on one in the aeronautical band. Assuming that the picture carrier has already been offset to fall at one of the permitted offsets, it then becomes necessary to offset the jammer by a multiple of 25 KHz (for most channels) from the picture carrier, with a tolerance of 5KHz. A third criteria is that the artifacts relating to the notch must be minimized.[1] One way of doing this is to place the notch at such a point that the energy contained in the horizontal sync is minimally disturbed. Figure 6 shows the horizontal sync signal and the spectrum which results. Not shown are the 15.734 KHz components resulting from the frequency of the sync signal. These components may be shown to follow an envelope, which is shown, having a sin(x)/x shape. This envelope exhibits nulls in the spectrum at frequencies equal to the reciprocal of the width of the sync pulse, W. A standard NTSC sync pulse has a width of 4.7 microseconds, so the nulls in the spectrum occur every 212.8 KHz.

Thus, in finding frequencies suitable for the jammer, one looks for offset frequencies equal to an odd multiple of one half the line rate, which also fall very close to an aeronautical offset (in this case, we look for frequencies removed from the picture carrier by a multiple of 25 KHz.). If the chosen frequency falls close to a null in the sync spectrum, so much the better. One suitable frequency is an offset equal to 29/2 of the horizontal line rate, about 228.1 KHz from the picture carrier. This is close to an aeronautical offset, though tighter than normal torerancs would be required to utilize an aeronautical frequency.[2] Further, it is acceptably close to a null in the sync spectrum.

MODULATION

The jamming carrier is modulated with a sine wave locked to the horizontal line rate and phased such that the envelope of the jamming signal is maximum during the middle of the TV line, when the amplitude of the picture carrier is lower. The jammer is then at minimum amplitude during the sync tip, when the picture carrier envelope is maximum. Thus, the modulation reduces the system loading. We have found that the scrambling effect is considerably enhanced by the modulation. This appears to be because the sync separator in the TV frequently takes the peak of the jammer to be sync. We are also experimenting with other forms of modulation on the jamming carrier, which appear to create additional irritation for the pirate, without being visible to the legitimate subscriber.

GENERATION OF THE JAMMING CARRIER

Figure 7 shows one possible method of generating the jamming signal. The jamming is done at the modulator IF in order to ease selection of the protected channel. Depending on the modulator, interface may be the same as interface with RF sync supression scrambling. A bandpass filter, F1, selects the picture carrier from the modulator. It is mixed with the jamming carrier from OSC 1, which in turn is locked to an offset derived as shown below. The difference frequency is recovered in lowpass filter F2, and applied to a phase detector, PD 1, along with a frequency at the correct offset. The error from the phase detector is integrated and applied to the jamming oscillator such as to keep the offset between the jamming and picture carriers equal to the frequency of offset oscillator OSC 2.

The offset oscillator is controlled within a second phaselocked loop, the reference for which is horizontal sync derived from a sync separator. Video is looped through the sync separator before being applied to the modulator. Predistortion of the video amplitude and delay is also performed, in order to compensate for the errors introduced by the trap. The horizontal sync frequency is divided by 2 and applied to phase detector PD 2. Output from OSC 2 is divided by 29 and applied to PD 2 as the other input. The error from PD 2 is integrated and used to correct the frequency of OSC 2.

Finally, the output from the jamming oscillator is modulated with a horizontal rate signal and added to the picture carrier before up conversion to the desired output channel. We should note that, in practice, the interface with the modulator may be different, but the above illustrates the technique involved. In order to set the level of the picture and jamming carriers with simple equipment, the jamming carrier generator includes provisions for individually turning off the picture and jamming carriers.

MECHANICAL ARRANGEMENT

An important feature of the system is its plastic enclosure - a UV stabilized, impact and weather resistant, tamper resistant housing in which the modular components are mounted. The primary mounting intended is to the side of a subscriber's house, similar to the interface provided by the telephone and electric companies. Other mounting methods may be used, such as mounting on a ground rod or on a post. Hasps are provided to permit locking with either a lock or a lead seal of the type used to protect power meters. Approximate outside dimensions of the box are 10 by 9.5 by 3.5 inches.

Internally, the unit includes a mounting post grid on 1.5 inch centers and is designed to accomodate suitable cable components such as ground blocks, splitters, and cylindrical traps in addition to the components of the new system. Cable entry and exit is provided for by gasketed holes in the enclosure bottom. RG59 or RG6 cable with connector installed may be inserted through the gasket while maintaining a weather and insect resistant seal. Figure 8 is an illustration of the internal component layout. Modules may be placed at any convenient location in the housing.

The modular components are designed for ease of installation and mechanical and electrical integrity. They are packaged in die-cast, tin plated zinc enclosures. Three modules have been developed to date. A filter module mounts directly to the mounting grid array and features two male push-on connectors for input and output. It contains a positive trap filter for a specific channel. Input/output and jumper modules mount to the filter modules and connect by female push-on connectors. Input/output modules provide a conversion from standard fittings to push-on and contain transient protection circuitry. Jumper modules serve simply to connect one filter to another. The push-on connectors have an integral O-ring seal to provide a pressure tight (to 10 PSIG) connection between modules. Center conductor contacts are precious metal plated and signal ground connections are tin to tin. Ground straps to connect the modules to a ground block are also available to insure a good earth ground. Captive screws are provided with each module for mounting ease.

EXPANDABILITY

Consideration has been given in the design to future applications and growth. It is anticipated that an amplifier module will be developed. Other modules are possible also. An addressability module has been discussed that would replace the input/output module and jumpers and would allow the use of existing filter modules. The addressability module would be mounted to the mounting grid array and existing filter modules would be inserted into it. The cost of an addressable upgrade would be minimized in this manner. Power would be supplied to the unit from the subscriber's residence through the center conductor of an output cable or through a separate set of wires.

Mechanically, the unit is designed to expand in two ways. First, a deeper lid would allow the mounting of electronics in the lid. The cable operator could remove the existing lid in the field and replace it with the new one. Secondly, provisions have been made for a subscriber accessable enclosure to be mounted directly below this enclsure. This enclosure would allow a subscriber or contractor access to the output cable to perform the signal splitting function or for connection to a pre-wired residence.

EVALUATION OF THE ISSUES

A number of issues must be evaluated when considering a new system like this. Cost is a major issue. The system obviously bears an initial cost for the housing. Since individual traps are employee for each channel, the cost is going to be related to the number of channels protected, in contrast to a set-top terminal which is generally insensitive to the number of channels protected. Should one configure this system with a number of channels under addressable control, the cost could become higher than that of an addressable set-top terminal, though the ability of one box to serve the entire subscriber premise is an obvious advantage.

Mounting to the side of a house as currently planned is an unusual concept in the CATV industry. This has been practiced for years by the electric and gas companies, and recently telephone companies have been using a box which performs an identical interface function. The box is a convenient point of demarcation between the subscriber's equipment and the cable company's equipment.

As currently configured, the equipment is passive, so powering is not an issue. However, future enhancements will require power. This power logically comes from the subscriber's home, but getting it to the box raises issues of safety and practicality.

Security is a big issue with any program denial technique. The system is certainly more secure than is a conventional positive trap, as a result of the proximity of the jamming carrier to the picture carrier. On the other hand, one would be reckless to assert that pirating was impossible. Rather, this system is intended to complement the contemporary notion that security must be evaluated in the context of probability of turning a pirate into a legitamate subscriber. If positive trapping has been considered a viable protection alternative, then this new system should surely allow improved confidence.

Quality of the recovered signal is thought to be better than that of a conventional positive trap system due to the narrow spectrum removed by the jamming signal. Further, since the removed signal components are duplicated in the full sideband, good compensation is possible. This compensation is added at baseband in the jamming carrier generator, and is based on measured amplitude and group delay of the baseband to baseband signal path.

As shown above, a conventional positive trapping system favors the lower part of the spectrum, generally the low and mid bands. The SAW technology employed in this new trap favors a higher portion of the spectrum. We have developed traps for channels 10 and 11 so far, and the next developments are expected to be in the lower portion of the superband spectrum, just above 216 MHz. The limit of the technology is not well known at this time, but clearly, we prefer to go towards higher frequencies.

These filters are compatible with scrambling systems, which may function on any channel including the trap-protected channel. This would be important to the operator wishing to make a transistion from scrambling to this system.[3] He may continue to operate a scrambler and descrambler while installing positive traps. After all subscribers receive the traps, he turns off the scrambler and turns on the jammer. The set-top converters should continue to function during the change-over period and after the scrambler is turned off. They may then be reclaimed by the operator at his convenience.

SYSTEM CONSIDERATIONS

While modulator interface is similar to that of an RF sync supression scrambling system, the requirements on the modulator are somewhat different. Since a second carrier equal in amplitude to the picture carrier is added, the modulator must have adequate linearity to handle it. At this early stage of development, we are unable to comment on the suitability of all modulators in existance. But, for example, a recently produced Model 6350 modulator is suitable, though we recommend some simple modifications for optimum results.

Addition of the jamming at a processor is technically feasible given the availability of a demodulated signal for phaselocking. However, baseband predistortion of the video must be accomplished somewhere in the system. Video delivered to subscribers not using a trap should not be predistorted, except possibly during a change-over period.

Transmission of a jammed signal over FM microwave links is not feasible, since the jamming signal must be added at RF. An AM microwave system theoretically can handle the scrambled signal, though generation of the transmitted signal may be complicated by the lack of sufficient upconverter linearity to handle the extra signal. This can be overcome by individually upconverting the picture and jamming signals and combining at the transmitted frequency.

Modulation on the jamming signal reduces the loading effect of the jamming carrier, though it is not eliminated. With the parameters currently planned for the system, addition of the jammer creates a peak voltage in the channel which is 4.6 dB higher than the voltage normally present on the picture carrier alone. Another way of looking at the increase is that it is equivalent in terms of system loading to adding another picture carrier, but at a level 3.1 dB below other picture carriers on the system. In most systems the increase in loading is probably not a big factor, but this needs to be examined in marginal situations.

CONCLUSION

A new positive trapping system has been shown which is believed to offer significant advantages when compared with conventional positive trap systems. These advantages accrue from the use of very narrow traps realized through the use of SAW resonators. Security is enhanced by placement of the jamming carrier very close to the picture carrier. Quality of the recovered signal is enhanced by the narrowness of the trap and the ability to place pre-corrected signals in the opposite sideband. Independent of this new technology but offered with it is a new packaging technique utilizing mounting to the side of the subscriber's house, providing a line of demarcation between the CATV system and the subscriber.

ACKNOWLEDGEMENTS

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[1] The effects of the notch may be compensated in the full sideband, but since compensation cannot be perfect, one is moved to minimize the effect of any residual errors.

[2] Presently only channels 10 and 11 are being used, so the question of aeronautical offsets is academic. In order to provide for future expandability, such offsets have been considered.

[3] The scrambling system manufacturer should be contacted before the change-out is started.

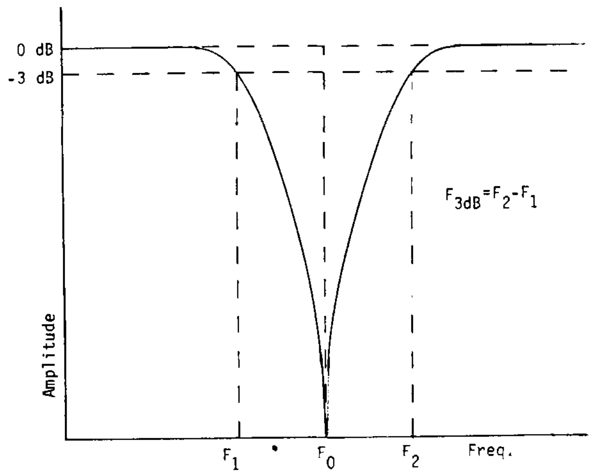

FIGURE 1 - FIGURE OF MERIT FOR A NOTCH

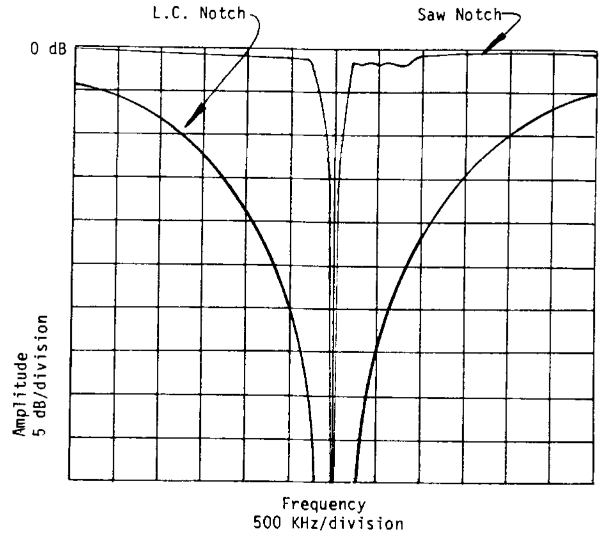

FIGURE 3 - COMPARISON OF NEW AND TRADITIONAL NOTCH PERFORMANCE

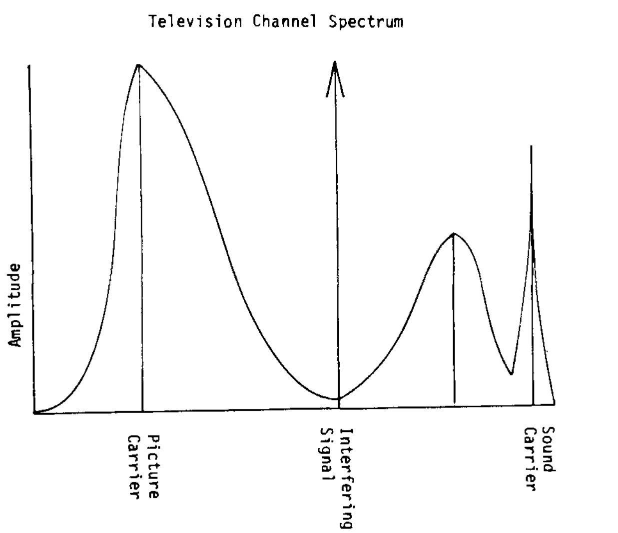

FIGURE 2 - TRADITIONAL JAMMING CARRIER PLACEMENT

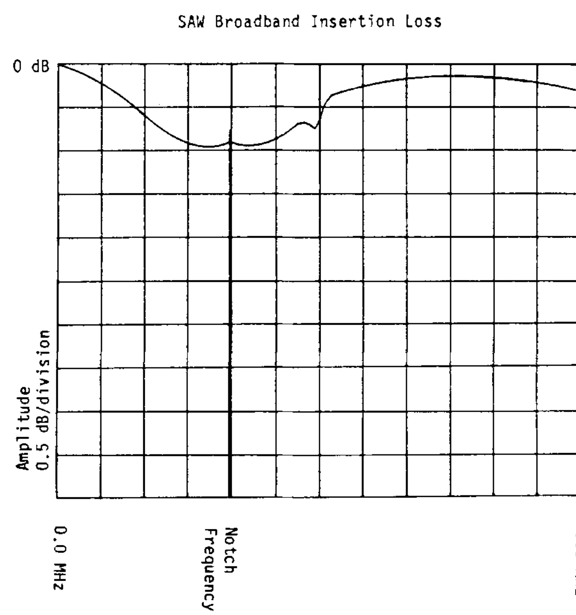

FIGURE 4 - NEW NOTCH PASSBAND CHARACTERISTICS

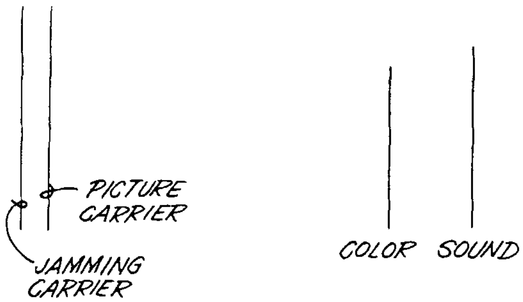

FIGURE 5 - LOCATION OF JAMMING CARRIER

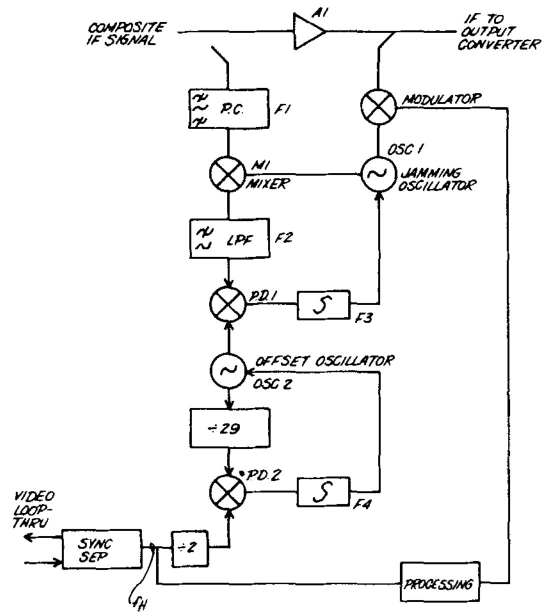

FIGURE 7 - GENERATION OF JAMMING CARRIER

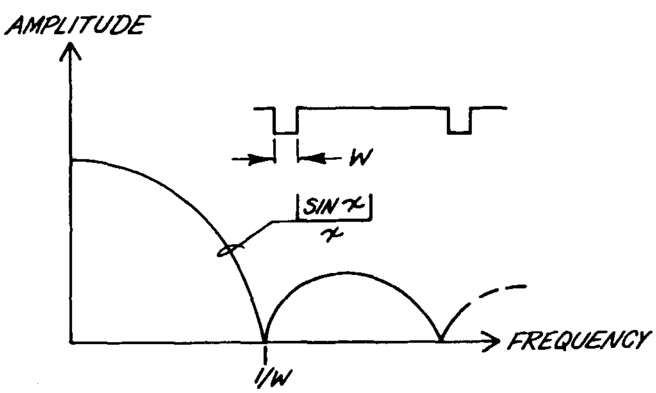

FIGURE 6 - HORIZONTAL SYNC SPECTRUM

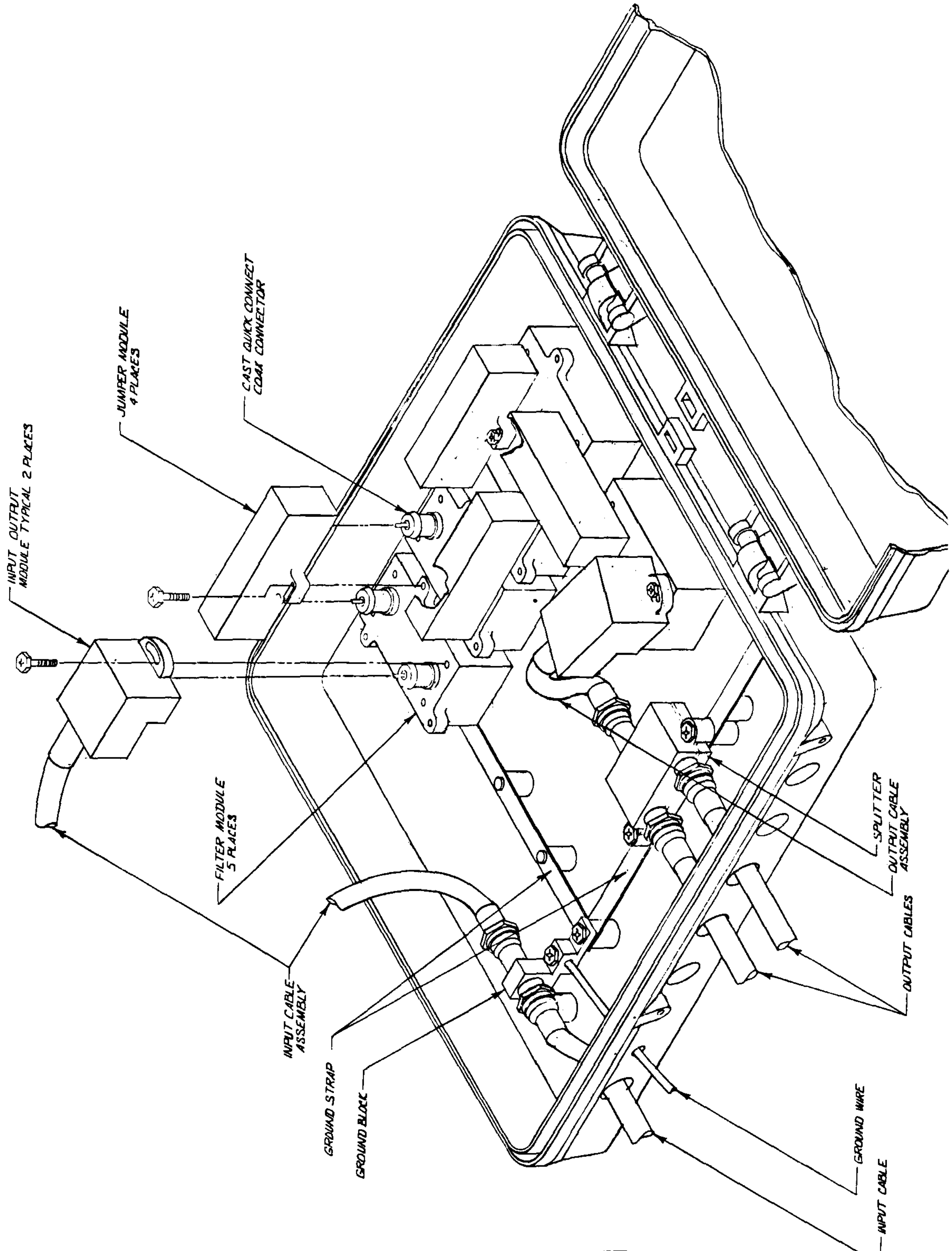

FIGURE 8 – INTERNAL COMPONENT LAYOUT

A REMOTE STATUS MONITORING SYSTEM FOR ONE WAY PLANT

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ABSTRACT

A system for transmitting data in an upstream direction on broadband cable plant equipped only with downstream transmission is described.

A multiple-frequency repeater technique allowing transmission of low frequency signals over long cascades in an upstream direction is used. This technique overcomes the need to undertake any significant modifications to existing plant and eliminates the problems of crosstalk and attenuation occurring in a single-frequency approach.

The resulting modem designs facilitate simple and economic addition of remote monitoring capability to any cable system using existing software and power supply interfaces.

INTRODUCTION

With increasing system bandwidths, complexity and operating costs there is a growing need to optimize maintenance procedures to reduce system downtime. Remote status monitoring has been a valuable adjunct to organized maintenance programs for over ten years. Status monitoring capability has been developed for the broadband industry by the major manufacturers of broadband amplifiers and standby power supplies.

The power supply monitoring systems designed and developed by Alpha Technologies during the early eighties included a complete system dedicated to power supply monitoring and a second system designed to interface Alpha's power supplies to the major amplifier monitors.

Alpha's system, like the others, required a two-way plant, hardware interfaces, transponders (or data modems), head-end controllers and dedicated monitoring software. Given that the installed base of two-way plant is a very small fraction of total CATV plant, few systems could hope to implement these techniques. Clearly a system which could function as a monitor in a one-way plant would have applicability to all CATV systems currently in use.

In early 1986 a decision was made by Alpha Technologies to proceed with a project to investigate the feasibility of two-way data transmission in one-way plants. A technique was proposed for use of a low-frequency return signal using the broadband system's power path and product development work was initiated.

BACKGROUND

A proprietary power supply monitoring system had been developed years earlier by Alpha Technologies, allowing CATV operators to implement a completely separate ('stand-alone') monitoring system for standby power-independent of any other monitoring system already operating in the system. The monitoring of power system status and the ability to remotely test or exercise critical power supply functions can expose power supply and battery malfunctions which affect large amounts of plant simultaneously.

The Alpha system, known as "RSM"*, incorporates a head end computer, head-end modem, local power supply modems, addressable power supply interfaces, and requires two-way plant. This well established system provides the following capabilities:

1. Automatic, constant scanning of up to 4095 locations, and

2. 'Single-unit mode' allowing detailed interaction with a single power supply,

3. Choice of up to eight operator-configurable test sequences for each address, with up to 14 selectable test parameters per test.

4. Automatic alarm functions configured by the user around locally defined fault conditions.

5. Stored or printed records of fault data by location and time.

6. Remote access to battery voltage values, charger operating modes, failure modes, inverter status, standby times, a.c. line status, and tamper switch status.

7. Remote control of battery charge modes, test function, test reset, alarm modes, and single or multiple unit scanning.

The development of the one-way plant communications technology is intended to take advantage of Alpha's existing monitoring technology - utilizing the same power supply interfaces, the

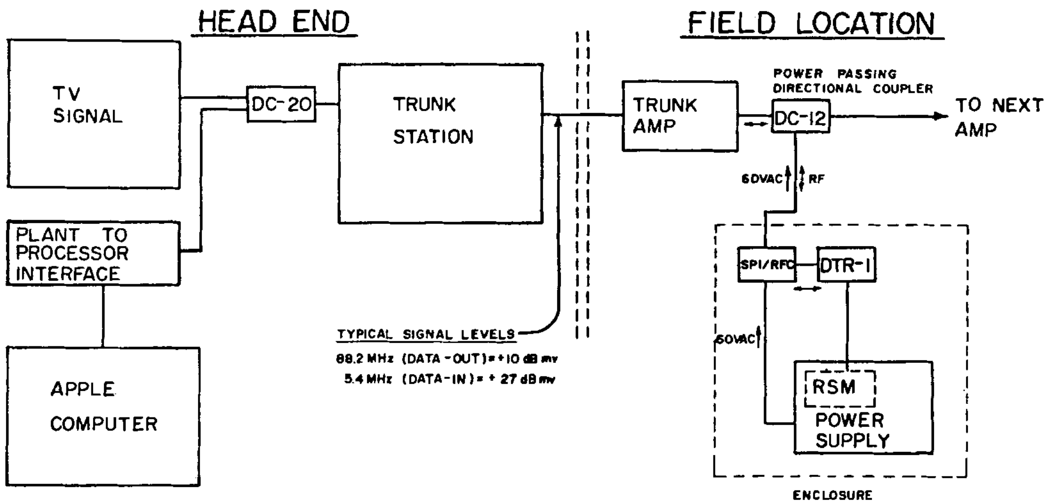


Figure 1. Alpha RSM System Layout

same software and the same physical characteristics. (Fig. 1.)

PROJECT "LIFELINE" IMPLEMENTED FOR LOW FREQUENCY RETURN

Selection of suitable return frequencies for data transmission is restricted to the range between 40KHz and 200KHz. This is based on the bandpass characteristics of typical plant components as illustrated in Figure 2. The power pass band below 40KHz is particularly noisy, so current channels begin in the 50KHz range and go up.

Initial designs for the low-frequency return were based on a single return frequency in the 115KHz range. This approach required adding a second semi-rigid cable to the power inserters used at standby locations. The two cables provided low frequency signal paths to and from the power-supply mounted repeater module. See Fig. 3.

The single-frequency design required change to or extensive modification of the power inserters, in addition to the modification of the power supply enclosure and 60VAC output circuitry. Further, adequate isolation between the two signals was difficult to achieve.

It was concluded that a multiple frequency approach was needed, providing the advantage that no power inserter change was required and no modifications were necessary to the power supply. In addition, repeaters may be implemented at any open distribution port between any widely-spaced powering locations. Open power directors require shunting with fairly large capacitors (between .22 and .47 microfarad) to complete the return path. This is the only system modification required.

SIX FREQUENCIES PER SYSTEM

Typically, six frequency pairs can be used to cover a complete system. Adjacent repeater stations use adjacent low-frequency channels with the last frequency before the head-end using the same frequency as the head-end receiver. See Figure 4.

Eight channel frequencies have been designated between the horizontal sync pulse harmonics. Adjacent channel rejection is approximately 66dB. Frequency pairs would be assigned as 1-3, 3-5, 5-7, 7-1, and so on, or 2-4, 4-6, 6-8, 8-2, etc.

Systems long enough to require more than three frequency pairs would repeat the same group of three as the duplicated frequencies would be sufficiently far apart so as not to interfere. The alternate channels are used to connect branches to the trunk.

Each power supply location supports a modem/repeater combination connected to the cable directly, via the 60 Volt power connection. At the power insertion point, four signals are multiplexed into and out of the cable:

1. *60Hz power at 60VAC*
2. *The downstream VHF data carrier*
3. *The low-frequency signal received from adjacent downstream repeater*
4. *The remodulated low frequency transmission to the next upstream repeater.*

The multiplexing network is a proprietary transformer hybrid network, mounted in the modem housing and is connected directly in line with

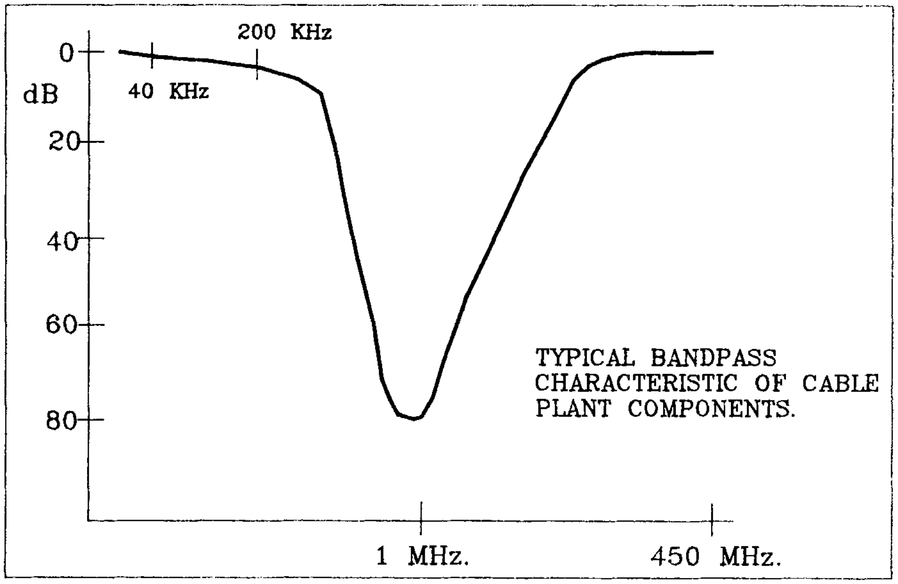


Figure 2. Typical Bandpass of Cable Plant Components

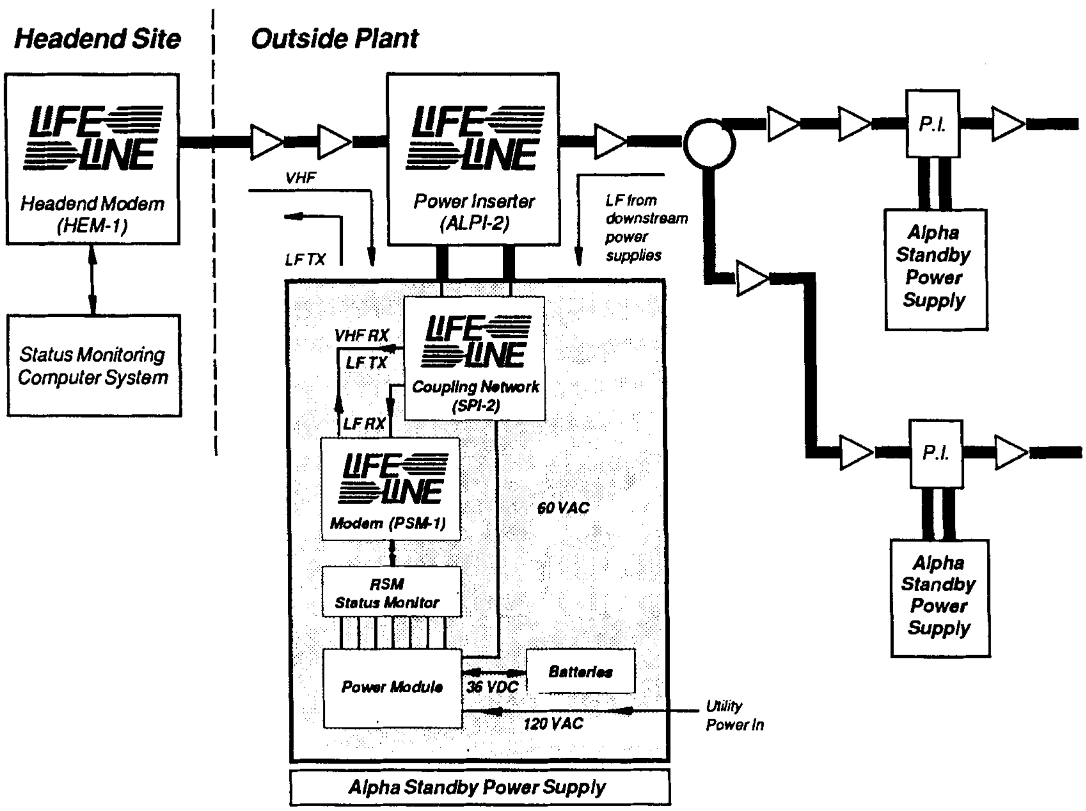
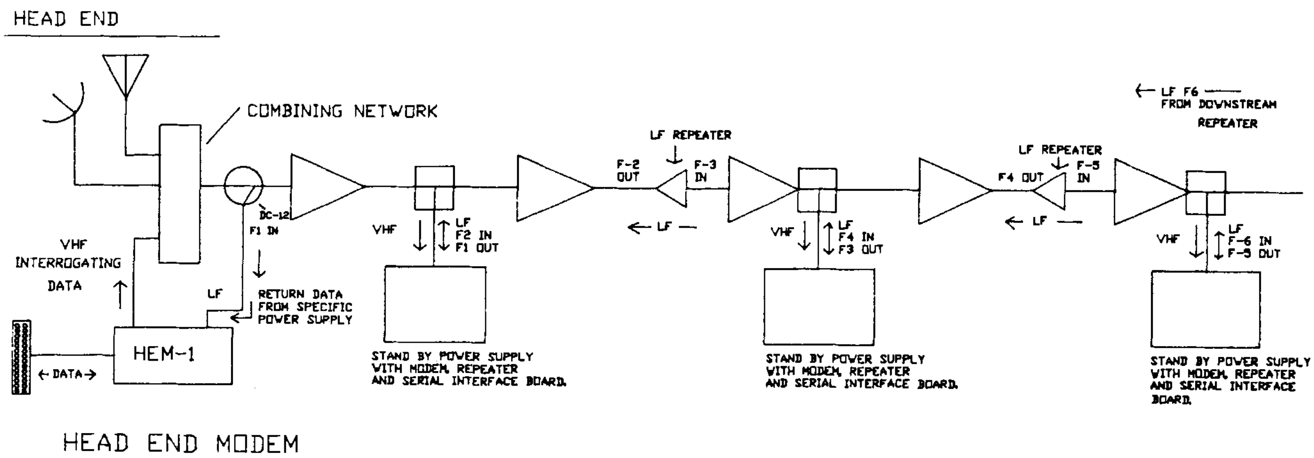


Figure 3. Single Frequency Lifeline Architecture



NOTE: LF REPEATERS BETWEEN TRUNK STATIONS ARE ONLY USED WHEN LF SIGNAL IS ATTENUATED SEVERELY BY TRUNK STATION MOTHER BOARDS.
OTHERWISE, REPEATERS ARE LOCATED ONLY IN THE POWER SUPPLY ENCLOSURE.

Figure 4. Typical Monitored One-Way Plant

the 60VAC output of the standby power supply. The addition of the modem to the powering location can be effected without interruption in power to the cable. See Figure 6.

Low frequency carriers are transmitted at approximately 70dBmV. With maximum attenuation between powering locations of 60dB, 10dBmV minimum signal is available to each receiver. For wider separations, an in-line repeater is added at any convenient open distribution port. These units are housed in standard line-extender hardware.

The power supply modems are full duplex, handling the downstream VHF carrier from the head-end and the upstream low-frequency carrier. Modems contain both VHF and LF circuits in the same housing, with the multiplexing circuitry needed to combine the three signals with the 60Hz power feed as described above.

When polled by the head-end computer, the addressed modem will begin a transmission by initializing the upstream repeaters. Carrier-detect circuitry in each demodulator turns on the repeater transmitters and data is passed upstream. Data is FSK encoded on the LF carriers, which are frequency-synthesized from a crystal reference. Data speeds up to 4800 BPS can be supported.

Once the repeater structure is installed in a CATV trunk, this "party-line" system enables extension of the same frequency pairs into any system branch.

CONCLUSION

The low-frequency return system implemented in the Alpha PSM-1 and HEM-1 modems will allow complete power supply monitoring in one-way plant. Use of existing controllers, software, and power-supply interfaces is facilitated, and no significant system modifications are required other than capacitive shunting of open power directors.

The technique may be used for any data communications in an upstream direction on broadband cable, including amplifier monitoring and control, alarm or security systems or subscriber services, without the need for expensive, power-hungry return electronics.

ACKNOWLEDGEMENTS:

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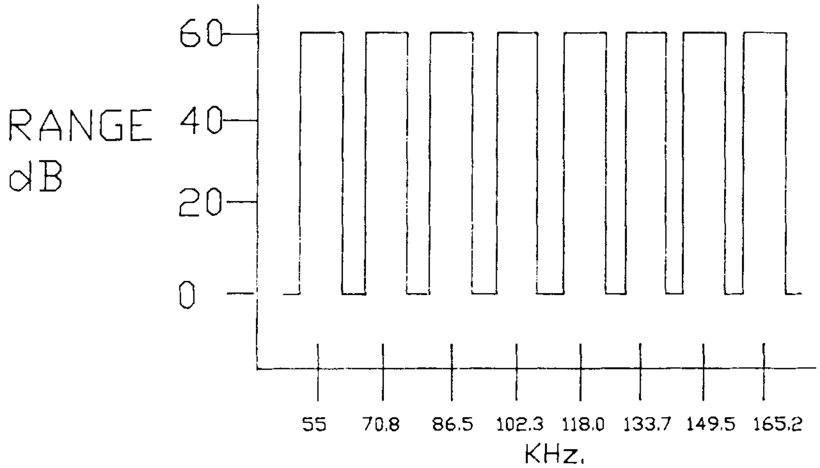

Figure 5. Lifeline Channel Frequencies in a Multiple Frequency Repeater System.

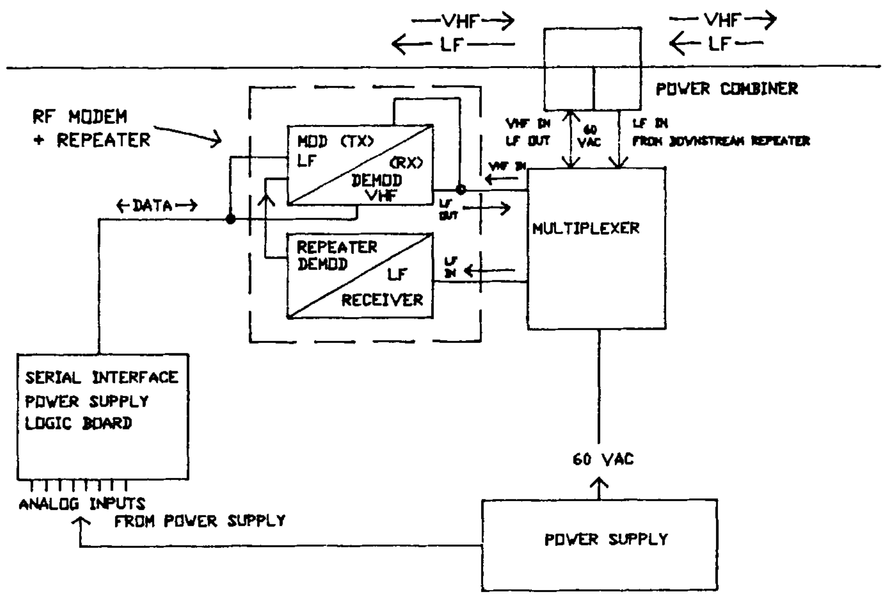

Figure 6. Signal Multiplexing at Power Inserter

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ADVANCES IN AML TRANSMITTER AND RECEIVER TECHNOLOGY

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ABSTRACT

New technologies have recently led to a substantially wider choice of both transmitters and receivers that can be utilized in local distribution microwave CATV systems. One area of particular interest to smaller CATV systems involves the increased power available from low cost multichannel block upconverter type transmitters. The highest output power with this type of transmitter is now made possible by the development of a microwave feedforward amplifier, which provides up to 10 dB improvement relative to the standard AML® Microwave Line Extender transmitter. At the other end of the application scale, the continued expansion of the number of channels carried by CATV systems in major markets has led to the expansion of AML capability to 550 MHz. In addition to incorporating the latest low noise VHF hybrids to fulfill this need in AML receivers, circuit modifications, which include a built-in microwave LNA and VHF AGC, have led to significant improvements in distortion performance without sacrificing system C/N. Design criteria are provided to guide the proper application of these new microwave components in CATV systems.

INTRODUCTION

The first application of AML microwave to provide local distribution of multiple television signals in the 12 GHz band was in 1971.(1) The transmitter consisted of a 2-bay MTX-132, with 16 channelized high level upconverters and passive multiplexing of 14 TV channels, an FM broadcast band channel, and a pilot tone. The receiver was phaselocked to the pilot tone(2) and utilized a ferrite attenuator to provide a microwave AGC, which maintained a constant carrier-to-noise ratio (C/N) and synchronous cross-modulation distortion performance for input signals above the AGC threshold. The outdoor mountable receiver housing was temperature controlled through the use of a gravity-gradient cooling system,(3) which maintained the internal temperature constant over a wide range of external ambient conditions.

Over the years, significant changes were made in both the transmitter and receiver to enlarge the system capacity to 60 channels. Although improvements in receiver noise figure, including the introduction of microwave low noise amplifiers (LNAs),(4) compensated to some extent for the increased multiplexing losses associated with the increase in channels, the application of AML to longer path distances led to the wide use of the channelized STX-141 high power AML transmitter first introduced in 1974. More recently, a much lower power block upconversion type transmitter, referred to as the Microwave Line Extender,(5) was developed to meet the needs of smaller CATV systems, in which isolated pockets of potential subscribers could not otherwise be economically serviced.

The inherent differences between the channelized and block conversion transmitters led to a large 18 dB performance gap between the MTX-132 and the Line Extender transmitter. Recent developments that substantially narrow this gap are described in the following section. These improvements in the block conversion type transmitter, while leaving the inherent multiple output advantage of the channelized transmitter unchallenged, can lead to respectable path length capability and wider applicability of the block conversion approach.

The second section of the paper deals with the new approaches employed in the 550 MHz system. Traditionally, AML systems have generally incorporated an Interface Unit (IFU) between the receiver output and the trunk cable. This unit, originally developed in 1973, serves a number of purposes. Key among these functions is the provision of an IF AGC to optimize total cable system C/N during microwave fades and a trap for the 74 MHz AML pilot tone. These functions are now incorporated in the 550 MHz receiver. But most importantly, the new design leads to substantial improvement in the distortion performance of the overall receiver. This is particularly of interest in CATV systems carrying 80 channels, but could also be an important factor when fewer channels are carried.

The final section of the paper provides a typical example that may serve as a guide to proper application of these more recent AML transmitters and receivers.

ALTERNATIVES FOR HIGHER POWER BLOCK CONVERSION TRANSMITTERS

The key to the performance of the block upconversion transmitter is its high power output FET microwave amplifier. The problem is essentially the same as with CATV hybrid amplifiers. Thus, one obvious solution is power doubling. The output amplifier in the AML Microwave Line Extender utilizes this technique. Two 1-watt FET stages are paralleled, and the outputs are combined in the internal microstrip circuit of the 2-watt amplifier. Could such power doubling be extended to 4 or even 8 branches? In principle, the answer is yes; but there are some important drawbacks. Aside from the added cost and power dissipation, the technique is limited by the losses associated with the microstrip combiner. Moreover, the FET stage gain is quite low (5 to 6 dB) for high power K_u-band FETs. Thus, if the driver stage is to avoid becoming the limiting distortion generator, it also must be doubled. Thus, circuit complexity multiplies even as the microstrip combining losses eat up a substantial portion of benefits obtainable from power doubling.

If power doubling is not as attractive as it might otherwise seem, what about utilizing higher power FETs? To some extent, this option suffers the same drawbacks as power doubling. This is due to the approach taken by the transistor manufacturer in obtaining the higher power capability. Internal to the packaged FET, in essence the same power doubling techniques are employed, although on a much more intimate scale. Combining losses are therefore somewhat less. Figure 1 shows a photograph of an internally matched 4-watt IMFET 1-1/4 inch wide package, in which two high power FETs are paralleled. Note the FET that has been placed on top of the package (near the letter H) to illustrate the relative size. Such FETs employ many parallel gate fingers, which serve to increase the power handling capability of the device. A key design consideration of a high power FET is thermal impedance, which largely determines the operating temperature of the FET gate channel. For long life (>10^6 hours), the channel temperature must be less than 120°C. The reliability deteriorates exponentially with temperature, so care must be taken throughout the transmitter design to provide for adequate cooling. An appreciation of the problem may be gained by recognizing that for each transistor die, approximately 6 watts are being dissipated in an active area smaller than 0.4 sq. mm. Nevertheless, the present state-of-the-art high power FET technology has shown itself capable of providing 5 to 6 dB more output power than the devices utilized in the standard AML Line Extender.

In contrast to the somewhat brute force techniques described above for increasing the available transmit power, a more subtle approach is to improve the transmitter linearity. This option results in greater available power without increasing the saturated output power capability, since less backoff is required for acceptable distortion performance. Both predistortion and feedforward techniques have been investigated. The results obtained with the former approach were no better than those obtainable with high power FETs. On the other hand, feedforward is not only well known by the CATV industry, albeit at VHF frequencies, but also proved to be by far the most attractive in terms of its performance capability.

Figure 1 Ku-Band FET chip with 4-watt IMFET package.

Figure 2 shows the block diagram of the microwave feedforward amplifier. The input signal is split by directional coupler C1 into two paths. The top path passes through main amplifier modules A1 and A2. Amplifier A2 is identical to the output amplifier used in the standard AML OLE-111 transmitter. Part of the amplified signal, along with the distortion generated by A2, is sampled by directional coupler C2. By adjusting attenuator R1 and phase shifter P1 for minimum power at test point TP, leakage of signal into distortion amplifiers A3 and A4 is suppressed by at least 20 dB. Therefore, amplifier A4 need only have half the output capability of A2 while still performing its amplification function without introducing any of its own distortion. Attenuator R2 and phase shifter P2 are adjusted to cancel out the distortion in the main signal path and therefore optimize C/CTB at the output.

The feedforward circuit is constructed in order to match the absolute delays encountered in the main and distortion amplifiers, respectively, by the delays T_1 and T_2. These are implemented in waveguide in order to minimize loss that would otherwise be excessive at these high frequencies. Although the resulting circuit, shown in Figure 3, is somewhat cumbersome, it was possible to package it within the temperature controlled AML outdoor housing. This is vitally important, since at these frequencies, the feedforward circuit is particularly sensitive to change in temperature. Other lower frequency microwave feedforward investigations have been reported,(6,7,8) but to our knowledge, the present application is the first at K_u-band. The availability of the temperature-controlled AML outdoor housing provides the key element, which makes this feedforward amplifier a practical part of an LDS microwave system.

Table I summarizes the performance parameters of the amplifier. When compared to the standard AML Microwave Line Extender, the power output capability of the feedforward amplifier is approximately 10 dB greater, even though the identical FET amplifier is used in both units. Indeed, in the feedforward unit, an additional 2 dB insertion loss follows amplifier A2 so that distortion

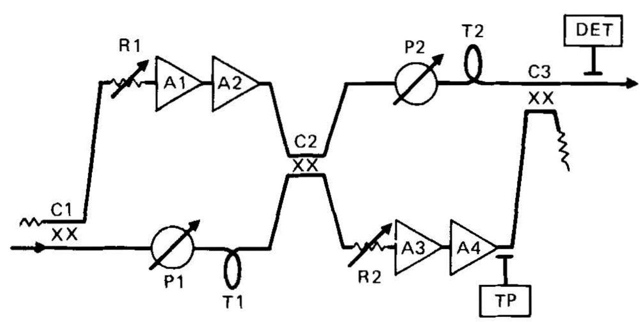


Figure 2 Microwave feedforward block diagram.

a new transmit monitor design was utilized to extend the range up to the full 550 MHz. This last was an outgrowth of the new 550 MHz receiver.

Figure 5 shows a block diagram of the 550 MHz indoor phaselock receiver. It differs from previous AML receivers in a number of key parameters summarized by Table II. Aside from the obvious extension to 550 MHz, the most significant change is the inclusion of a VHF AGC in addition to the standard microwave AGC. The VHF AGC, which also operates off the pilot tone level, provides a minimum of 12 dB additional AGC control if the receiver input signal drops below the microwave AGC threshold. The 74 MHz AML pilot signal is trapped out at the receiver output. As with previous receivers, the microwave AGC threshold is adjustable, but is normally factory set for 53 dB C/N. Since at minimum VHF gain the noise figure is 8 dB, this corresponds to -47 dBm receiver input. The noise figure improves to 7.3 dB as the PIN attenuation in the VHF AGC module decreases as would happen in a deep microwave fade. This noise figure is established by the combination of 1 dB input attenuation, a nominal 3.5 dB noise figure of the LNA, and 5 dB noise figure of the VHF preamplifier following the combined 6 dB filter plus mixer conversion loss. Although lower overall noise figure is possible by putting the LNA outside the AGC loop or by use of a 2-stage LNA, the receiver sensitivity improvement is generally not worth the impact on second and third order distortion performance. Exceptions may occur when the maximum received signal is low or if distortion is in any case limited by a block converter type microwave transmitter. Note that when calculating the maximum expected signal level, the so-called "field factor" should not be utilized. Note also that the maximum input level can be influenced by so-called multipath up-fades that can occur on some paths.

TABLE II

AML RECEIVER CHANGES

54 to 550 MHz output frequency range
Built-in VHF AGC and pilot-carrier trap
Standard 1-stage LNA in microwave AGC loop
Nominal +21 dBmV output level
8 dB NF at microwave AGC threshold (7.3 dB at small signal)
80 channel C/CTB 80 dB
2nd order beat -82 dBc
Phase-lock in indoor rack-mount configuration
AGC and phase lock disable switches for troubleshooting

Figure 6 shows C/N, 80-channel C/CTB, and second order beat performance as a function of received signal level for four different 550 MHz receiver configurations. The mixer is the largest contributor to both second and third order distortion. Therefore, the microwave AGC is in all cases essential to satisfactory performance at higher input level. When the LNA is outside the AGC

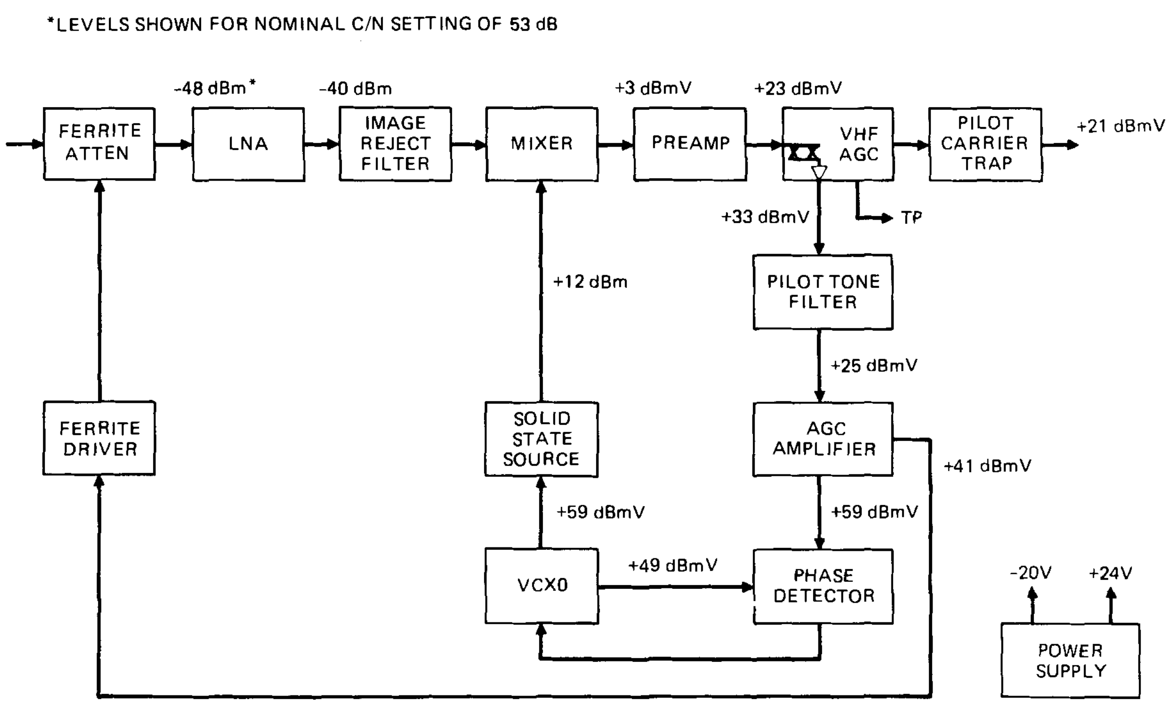


Figure 5 550 MHz receiver block diagram.

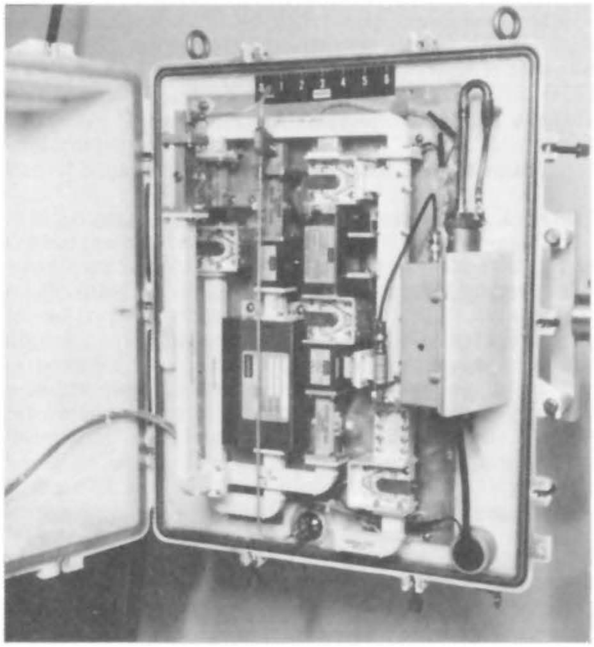


Figure 3 Microwave feedforward implementation.

cancellation in the output loop is on the order of 24 dB. Because of its relatively narrow percentage bandwidth, the constant temperature provided by the AML housing, and also because it would typically not be employed in long cascades, the microwave feedforward amplifier does not suffer some of the limitations pointed out by Preschutti(9) for CATV feedforward units. However, just as with VHF amplifiers, the C/CTB degrades somewhat faster than 2 for 1 as the total power in the amplifier is increased. The typical degradation is 3 for 1 with increased power so that, for instance, a C/CTB of 53 has been measured for 30 channels each at 8 dBm. The power out versus input transfer function approaches that of an ideal limiter, so even at +29 dBm output, a single channel VSBAM TV signal appears essentially free of distortion. This last point has implications with regard to the potential utilization of the feedforward amplifier as a frequency agile backup to substitute for any failed channel in a high power AML STX-141 transmitter array.

TABLE I

FEEDFORWARD AMPLIFIER PERFORMANCE SUMMARY

Frequency Range	12.7 -13.2 GHz
Nominal Gain	20 dB
Noise Figure	15 dB
Power Output/Channel at 65 dB C/CTB	
No. of Channels	Po (dBm)
12	+7
21	+5
35	+2
60	-1

Figure 4 summarizes the relative output capabilities of multichannel AML transmitters now that the gap between the channelized MTX-132 and the block conversion OLE-111 is somewhat filled in. Note that the comparison is for "transparent" 65 dB C/CTB operation of the lower powered units compared to a single output of the MTX-132 and STX-141. These latter automatically provide increasing numbers of multiple outputs as the increased number of channels leads to a greater multiplexing complexity. However, as the number of outputs increases, the per channel level at each output decreases. In the block conversion type amplifiers, the per channel power output drops to maintain required C/CTB for an increased number of channels. The reasons are different, but the results of increased channel loading on per channel power output capability are essentially the same, so the comparison made in Figure 4 holds over a wide range of channel loading.

550 MHz AML

The multiple output capability of the MTX-132 transmitter is well illustrated by the 16 outputs provided with the recent expansion to 80 channel loading. Each output is rated at +6 dBm per channel. Achieving this power, following transmitter multiplexing losses, required that the high level upconverter be redesigned for operation up to 550 MHz input. The design modification involved extensive change to the upconverter VHF input circuitry, but the operating principle remains unchanged from lower frequency modules. The transmitter VHF driver amplifiers also required complete redesign to operate with required linearity at the higher output power necessitated by reduced gain in the upconverter. Finally,

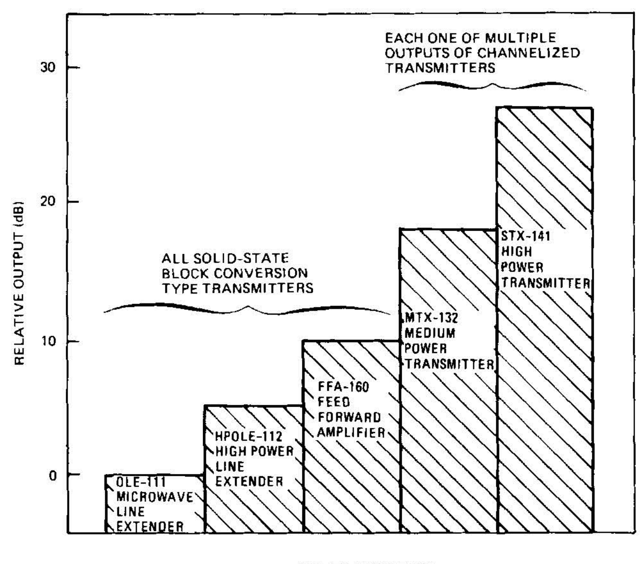


Figure 4 Relative output capability of AML transmitters.

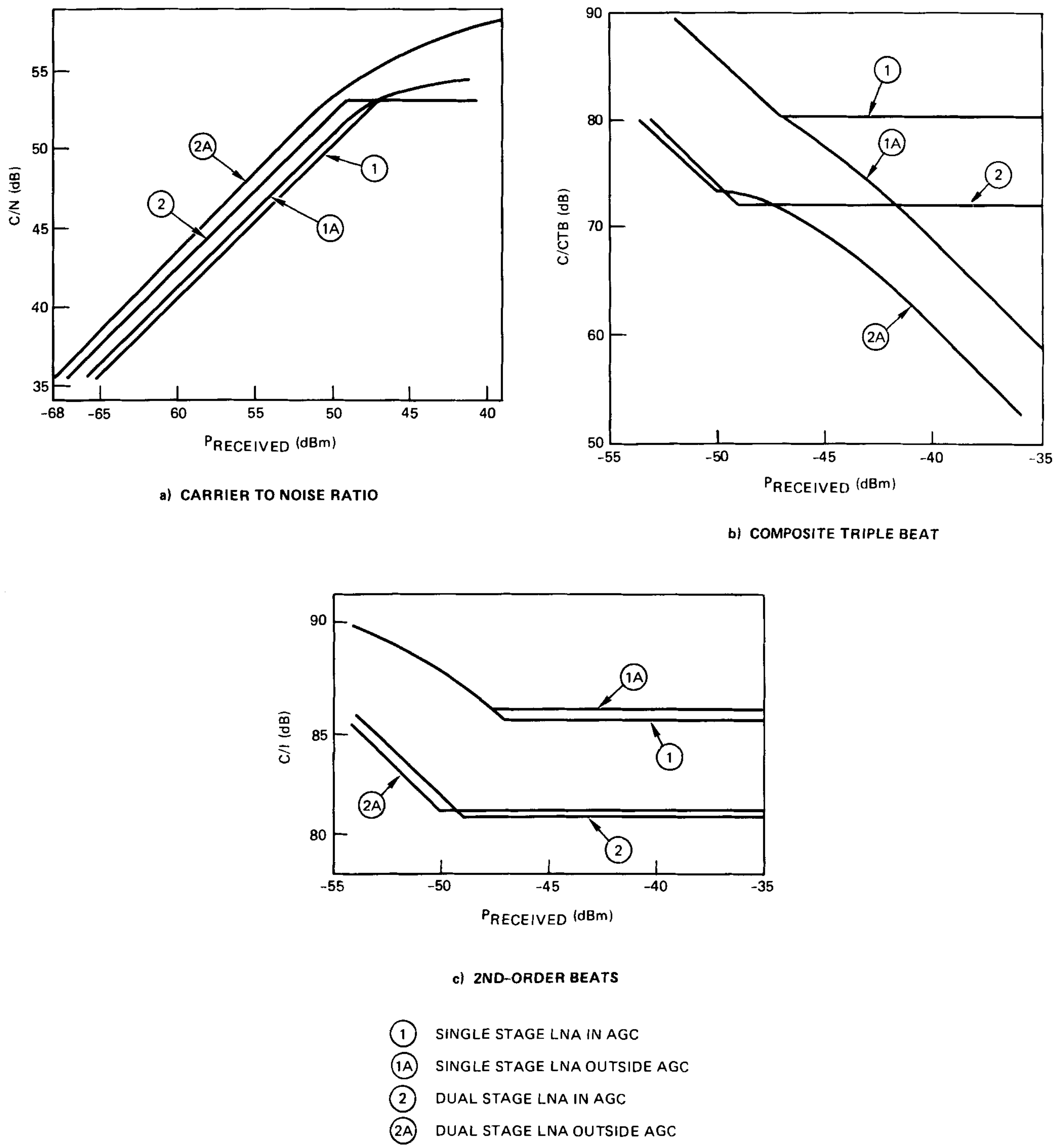


Figure 6 80-channel performance of various 550 MHz receiver configurations.

loop, the shape of the C/N and C/CTB curves is influenced by the choice of microwave AGC threshold. For the 2-stage LNA, the threshold was selected to yield 56 dB C/N at -45 dBm input; while for the single-stage LNA outside the loop, the illustration shows 53 dB C/N at -47 dBm input. If the LNA is inside the AGC loop, the tradeoff between C/CTB and C/N is a straightforward 2 for 1 function of the AGC set point.

Figure 7 shows the indoor 550 MHz phaselock receiver as it normally appears and with the front panel dropped to provide access to the phase detector balance adjustment. The receiver is designed to facilitate maintenance by making all controls accessible through either the front or back panels. To further aid in troubleshooting, AGC and phaselock disable switches are provided. The latter automatically sets the VCXO to a free-run mode.

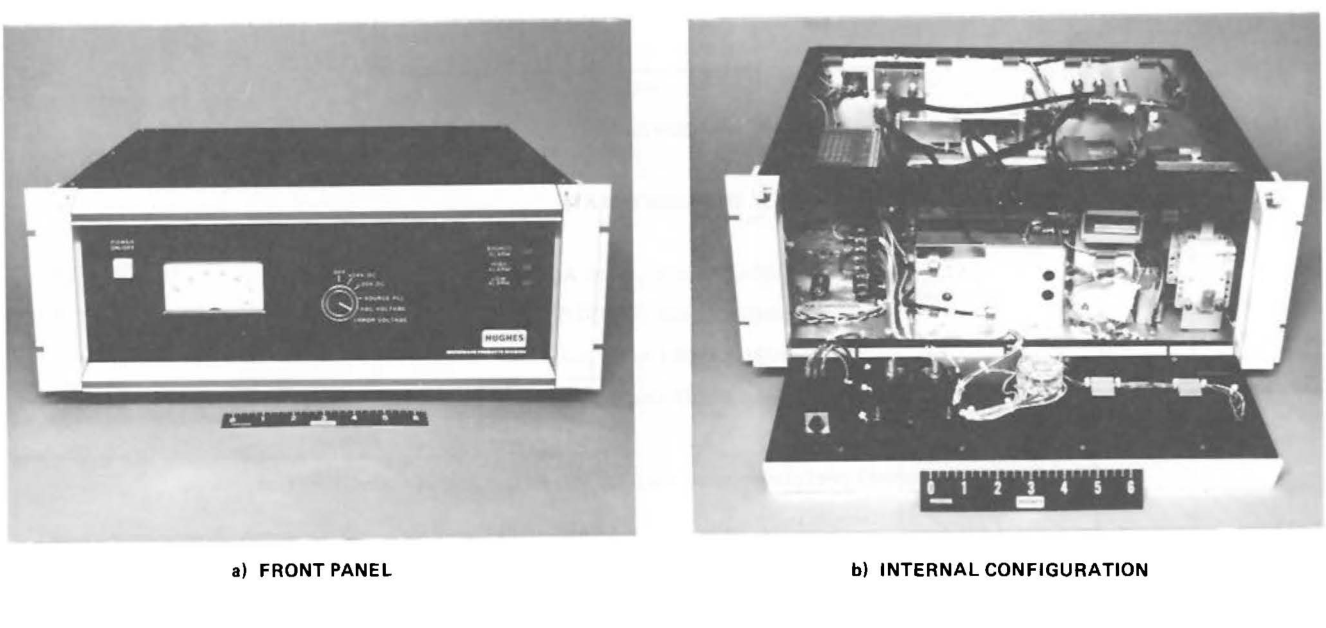
The main features of the indoor 550 MHz receiver are designed into the outdoor receiver retrofit kit, which permits upgrading of existing phaselock receivers. In particular, the VHF preamp and AGC modules and LNA are incorporated. One change not required by the indoor unit is the AC/VHF diplexer which had to be completely redesigned to accommodate 550 MHz. Insertion loss of this module is less than 0.2 dB and return loss better than 20 dB.

APPLICATION CONSIDERATIONS

Figure 8 shows how the feedforward amplifier is typically used in tandem with the Microwave Line Extender. In establishing the desired operating levels, the first question to ask might be what level of C/CTB is acceptable, since block conversion transmitter output levels are generally limited by this parameter. If the microwave feedforward amplifier is to be the primary contributor to C/CTB, the OLE-111 must be backed off from its normal output level. The difference in level between the feedforward input and OLE-111 output represents the permissible loss that might take the form of a long waveguide run if the FFA-160 is to be tower mounted to achieve the greatest possible path length. With this allowance, the OLE-111 might still be at the base of the tower, where maintenance should be much easier. In any case, the next concern is to reduce the noise power output of the OLE-111, since its output level is quite low. By setting the interstage attenuator between the LNA and 2-watt amplifier to 14 dB, the primary source of noise is then the 2-watt amplifier. Finally, one must check to see whether third order distortion generated by the LNA and by the upconversion mixer, as well as second order distortion of the mixer, are still acceptable. Since a high power local oscillator is utilized in the OLE-111, mixer distortion is minimized. The LNA is also designed with distortion in mind - having a 3 IM intercept point of +27 dBm. The noise and distortion contribution of each stage is shown in Figure 8 assuming 40-channel loading. The overall transmitter performance is also summarized. If the LNA is replaced with a piece of waveguide and the interstage attenuator reset to 0 dB, a further 1 dB improvement in transmitted C/N is possible.

Note that power addition, rather than voltage addition, is applied to C/CTB. This is because each contributing stage is different from the others, resulting in a randomness of the relative phases of the intermodulation products. The necessity to back off the driver stage and the concomitant concern with C/N is a feature common to all post-amplifier installations, not just when the post amplifier is of the microwave feedforward variety. However, if the post amplifier is a standard power amplifier, the intermodulation distortion from the intermediate power amplifier may more likely add in-phase with that of the output amplifier, since their characteristics are very similar.

Consider next an application involving a 13-mile path with 10 foot antennas at both receive and transmit ends. Even with only 0.3 dB waveguide loss



a) FRONT PANEL

b) INTERNAL CONFIGURATION

Figure 7 Indoor 550 MHz phase-lock receiver.

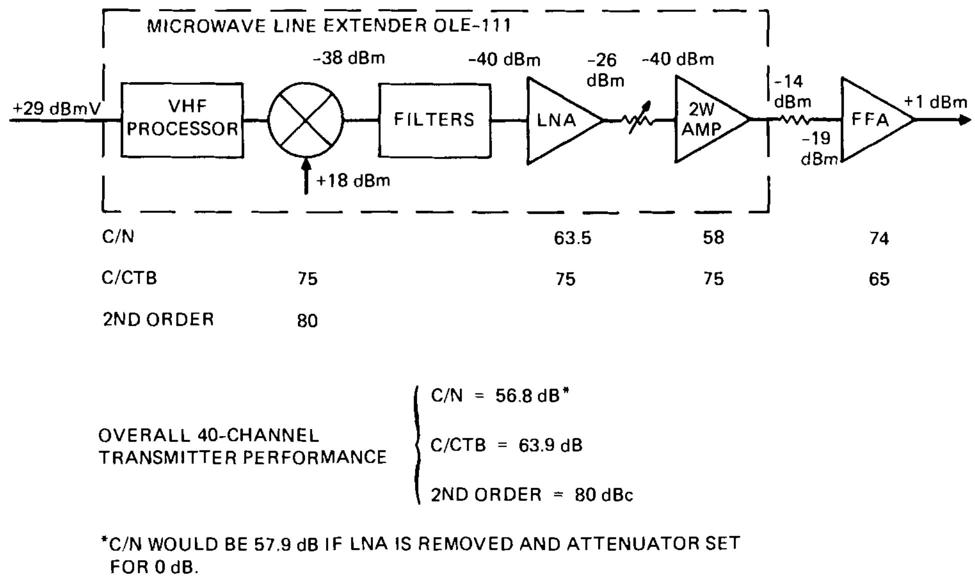


Figure 8 Typcial feedforward application.

interconnecting to each antenna, the maximum clear weather signal input to the receiver would be -43 dBm. Despite allowing for some multipath up-fade, this weak signal example suggests that consideration be given to a 2-stage LNA outside the AGC loop. For this 40-channel case, the receiver C/CTB will be 6 dB better than that shown in Figure 6b, so even in the maximum signal condition the total microwave link contribution to C/CTB is 63 dB. The second order beat for the total link is 77.5 dBc. The C/N provided to the last subscriber is determined by the sum of the contributions of transmitter, receiver, and cable system. This is illustrated by Figure 9, in which it has been assumed that the normal cable system contribution is 47 dB. Even allowing for a 2 dB "field factor," an 8 dB fade margin exists to a very respectable total system 45 dB C/N. The margin to 35 dB C/N is 22 dB, which for average rainfall and multipath conditions leads to a predicted 99.7 percent path reliability. Note, in Figure 9, the influence of the receiver VHF AGC down to -62 dBm input signal. Here, it is assumed that once this AGC runs out, no further AGC is available in the cable system to counteract the drop in signal level.

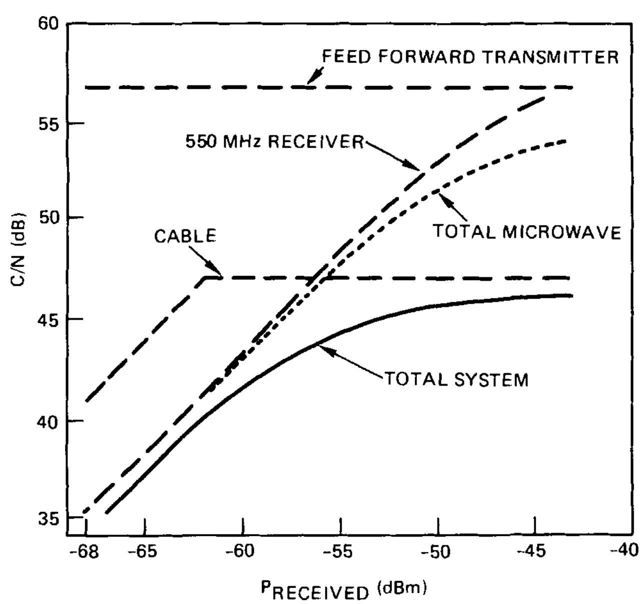


Figure 9 40-channel performance for 13-mile microwave feedforward system.

SUMMARY

The increased output capability of block conversion type transmitters utilizing feedforward post amplifiers has further expanded the potential range of application of cable-fed broadband microwave transmitters. In many cases, the maximum received signal level will dictate the use of a receiver with a 2-stage LNA outside the microwave AGC loop. Nevertheless, this microwave AGC is still important to controlling second and third order distortion of the receiver mixer, while the VHF AGC extends the range of the constant level output signal fed to the cable system. Improvements in the receiver design with alternative types of LNA implementation facilitate a wide range of AML applications, including those involving up to 80 channels.

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AN ENHANCED RF TELEVISION SCRAMBLING SYSTEM USING PHASE MODULATION

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ABSTRACT

A new scrambling and descrambling system for television RF signals is presented which represents a new generation of technology for CATV service security. The new system has been designed to be low cost, easily manufactured, reliable and more secure than conventional RF technology. Through the use of complex response Surface Acoustic Wave (SAW) filters, the system employs RF Carrier Phase Modulation to enhance the economy of RF sync suppression techniques with some of the video and audio signal security attributes of Baseband systems. Additionally, the new encoding system provides an integral means of reliable, noise insensitive addressable data communication to subscribers' descramblers without the use of a separate data channel.

INTRODUCTION

Since the advent of Pay TV in both broadcast and CATV environments, a secure, effective and economical method of scrambling and descrambling Television signals has been sought. Many RF techniques, while inexpensive, have proven over time to be weak in signal security and vulnerable to program piracy. Baseband scrambling systems, by supplementing basic sync suppression with video and audio manipulation, can increase security substantially but with a corresponding increase in cost.

Over the last few years the market demands for Addressable Home Terminals have been changing. Signal security is still an important issue but not as important as is cost and user features. What the market really desires is an economical, fully featured, secure CATV Converter-Descrambler.

REQUIREMENTS

In order to study and evaluate past and present system design shortcomings and to plan development of a new enhanced CATV Converter to meet present and future market demands, a "wish list" of requirements was composed. The following list includes many of the the desirable attributes and features an ideal Home Terminal should have:

- Secure
- Reliable
- Addressable
- Reliably Manufacturable
- BTSC Stereo Compatible
- Downloadable Configuration
- High Quality Signal Restoration
- Fully Featured
- Future Friendly
- Low Cost

Most of the above features seem straightforward enough, but in order to achieve both low cost and a high degree of security a totally new TV Scrambling system must be developed.

A NEW SCRAMBLING SYSTEM

Security and low cost at first seem to be mutually exclusive parameters. Conventional RF scrambling techniques are economical but their designs are well known and have made them easy targets of "black-box" decoders used to steal CATV services. Also, many RF systems require critical timing adjustments and have performance problems with BTSC Stereo due to data and timing AM modulation of the Pay TV channel's Sound Carrier.

An ideal system would have the security attributes of a Baseband System and the economy of an RF system without performance degradation. Toward that goal, a new TV scrambling system was developed using Video Carrier Phase Modulation (PM) over a defined frequency range to scramble the TV signal and transmit data and descrambling information to subscriber terminals.

Phase Modulation - Encoding

As with other RF scrambling methods, the PM Encoder processes the Visual IF signal of the modulator and returns it for conversion to the CATV channel frequency as shown in figure 1. Modulator video is also looped through the encoder to be used as a timing reference in the scrambling process. An audio loop-through is used for the purpose of enhancing the system's Audio Masking effect. Since the modulator's sound carrier is not modified in this scheme, an Aural IF carrier is not connected through the encoder.

Figure 1

P. M. SYSTEM CONFIGURATION

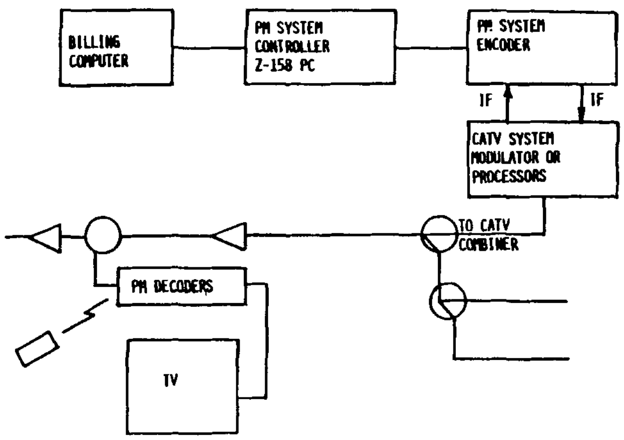
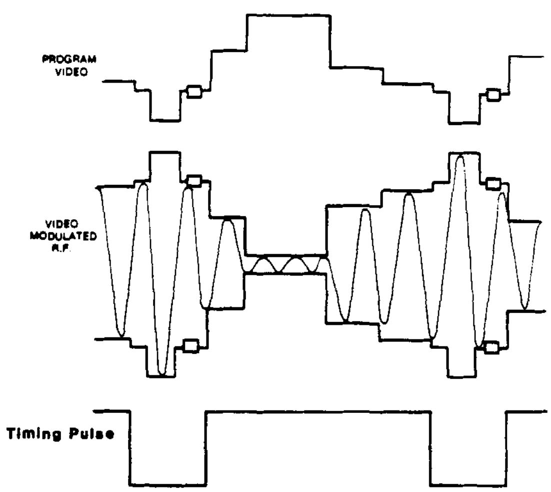


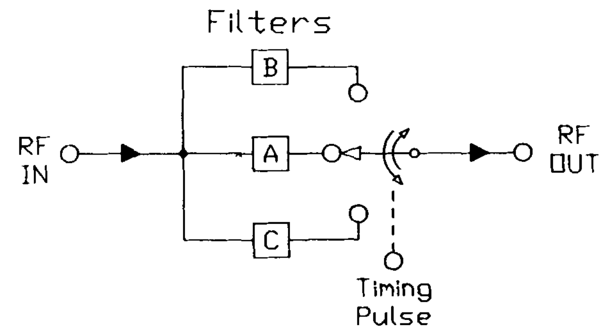
Figure 2 shows the typical IF modulation process performed in the CATV headend channel modulator. From the modulating video a timing pulse is generated, synchronized with each Horizontal blanking period.

Figure 2
IF Modulation and Timing Pulse Generation



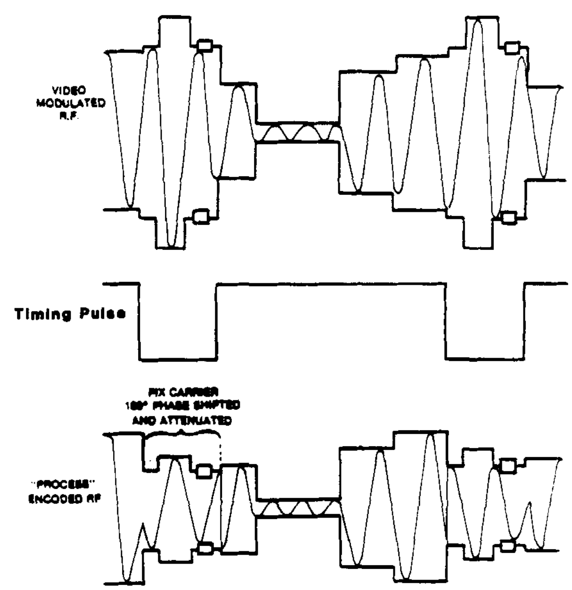
The generated timing pulse is used to dynamically switch the Video IF signal from passing through a flat response bandpass filter (A) to one of two alternate filters (B or C) as shown in figure 3.

Figure 3 RF Switch



Each of these alternate filters, corresponding to two different scrambling modes, have the characteristic of not only attenuating the IF signal at the picture carrier frequency, but to completely phase reverse it 180° as shown in figure 4. The resulting IF output looks as if it is encoded with standard sync suppression but also contains a carrier phase reversal. This Phase and Amplitude modified Video IF is then returned to the CATV modulator for conversion and distribution to subscribers over the cable plant. This combination of sync attenuation and non-linear phase modulation provides non-authorized viewers with both video scrambling and audio distortion.

Figure 4
Phase Modulation - Sync Suppression

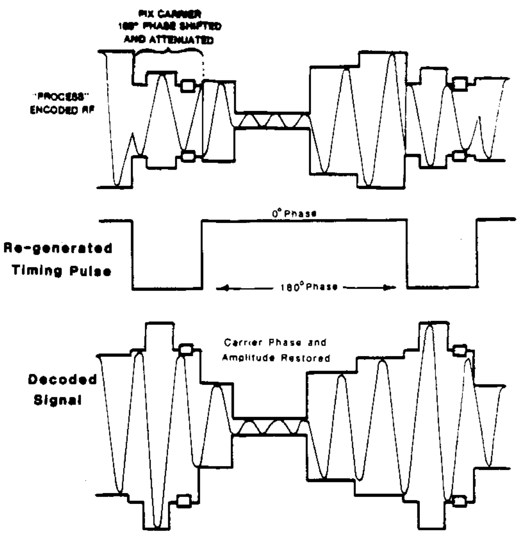


Phase Modulation - Decoding

At the subscriber's home terminal, the PM encoded channel is tuned and converted to a low VHF channel (Ch 2, 3 or 4) for processing and transfer to the home TV receiver. The phase modulated VHF signal is first filtered and passed to a PLL synchronous detector designed to be stable for both a 0° and 180° phase lock. The output of this detector, shown in figure 5, is a replication of the timing pulse originally used to encode the signal at the head-end. Since the detected timing pulse is derived directly from the encoded signal, any timing errors are insignificant.

Figure 5

Decoding - Detection of Timing



The regenerated timing pulse is then used to control an RF switch which selects the routing of the encoded VHF signal through either of two filters complementary to those used in the head-end encoder. The result at the output of the decoder RF switch is a VHF TV signal restored to its original Amplitude and Phase and passed on to the TV receiver.

Data Transmission

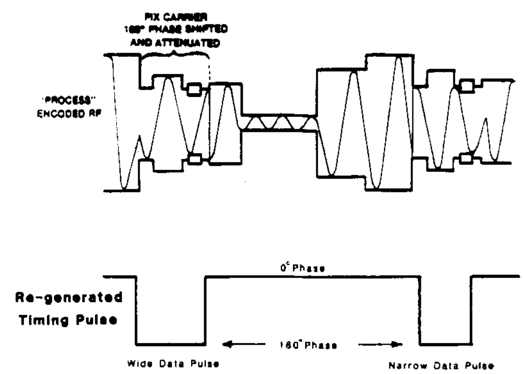
One deficiency of conventional RF TV scrambling systems is the method of data transmission. In-band timing and program tags are normally AM modulated on the channel's Aural Carrier making them prone to be noise sensitive, critical in timing adjustment and to cause performance problems with BTSC Stereo encoded audio. Out-of-Band FSK data provides a continuous communication link from the head-end but takes up valuable CATV spectrum and requires decoder safeguards to protect against loss of the data carrier. Out-of-Band Data also requires each decoder to include the extra circuitry, cost and potential unreliability of a separate data receiver.

The new Phase Modulated RF scrambling system provides a secure, reliable and relatively noise-immune in-band data transmission system inherent to the scrambling system without the need for a separate data receiver.

PSK modulation of data is widely recognized as an extremely noise insensitive method of communication. Since the basis of the new PM scrambling system is the phase modulation of the carrier 0° and 180°, data transmitted additionally in this fashion could constitute a very reliable communication link. As shown in figure 6, the PM system data is encoded as pulse-width modulation of the timing pulse itself creating 262 bits of data for every vertical field of video, one bit for each horizontal period.

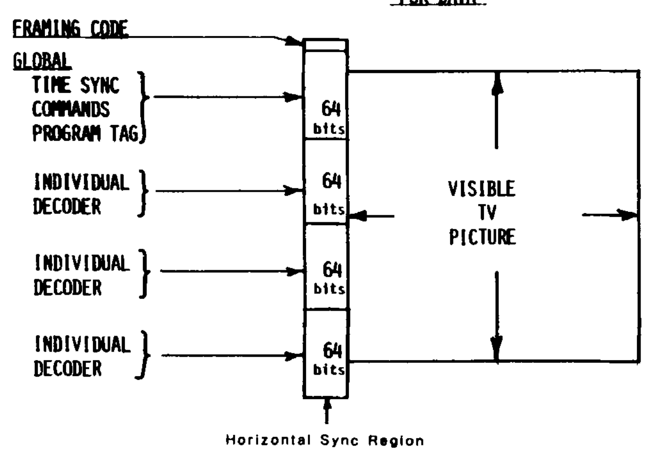
Figure 6

PM - Data Transmission



The data, encoded into the timing pulses and BPSK (Bi-phase PSK) modulated on the RF picture carrier is transmitted on all PM scrambled channels. The data is formatted into four packets of 64 bits each per video field. This is shown graphically in figure 7 as horizontal blanking interval data alongside the visible TV picture.

Figure 7 PM IN-BAND PSK DATA



The first of the four data packets is used for global command data, time sync and channel related program tag information. This global packet is encrypted for data security with a variable "session" key. The session key is one of several decryption keys periodically downloaded individually to subscriber's decoders.

The other three data packets contain information addressed to individual decoders and are encrypted with a unique "address key". The address decryption key is different for each decoder and has been factory programmed and "locked", along with the decoder's address, into each decoder's non-volatile memory. Enough address data is allocated in the individual data packets to allow for a potential of over 67 Million addresses.

With an addressing rate of three subscribers per field, up to 10,800 addressable decoders per minute can be processed. Preceding each group of four data packets is a framing code used to identify each new data group.

In addition to the rugged PSK transmission system, the data is further protected by three additional "shells" of security. First, each data bit received by the decoder's synchronous detector must meet specific timing requirements in order to be accepted by the decoder as good data. If only one bit in a packet falls out of the acceptance "window", the entire data packet is discarded by the decoder.

Secondly, every packet received by the decoder must match one in the decoder's "library" of recognized packet types in order to be accepted. For the final "shell" of data security, to further enhance noise immunity, each packet includes 16 bits of CRC (Cyclic Redundancy Code) for error detection. Noise tests of the system have shown accurate decoding at Carrier-to-Noise ratios below 14 dB.

Multi-Mode Scrambling

In order to ensure scrambling security, encrypted data and signal scrambling must be linked in some fashion. The system must be designed such that if the data link is broken, the TV signal cannot be descrambled. A good design requires that certain information must be present in the data necessary for the decoder to properly function.

In the PM system, there are two distinct levels of scrambling corresponding to two different attenuation levels of the horizontal sync interval. Each scrambling mode requires different signal switching between three complex filters (complementary to those in the encoder) at the correct time and in the correct sequence. Either mode can be used independently in a fixed format or dynamically random-switched at the Cable Operator's discretion. Encrypted downstream data is used to indicate to the individual decoders which mode is being used.

Included in the first data packet of each field is the encoder's scrambling mode definition for the next field. This data is transmitted to all decoders and encrypted with the CATV system encoder's variable "session" key. If the decoder is authorized and has the previously downloaded decryption key in its memory, it can decipher the program level and scrambling mode definition. This information is used with other available data to correctly descramble the program. If the decoder cannot decrypt the data packet or if the packet's program level does not match the corresponding downloaded decoder authorization level, the signal is not allowed to be descrambled.

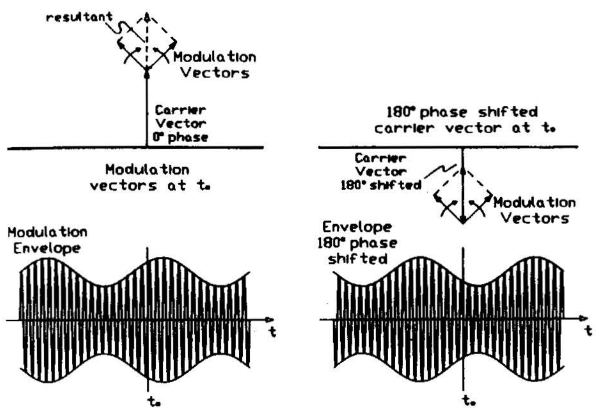
Thus, the scrambling mode and authorization system is protected with three levels of security: (1) the scrambling mode is defined by difficult-to-detect downstream data, (2) the mode definition data is encrypted with a variable "session" key, (3) in order to have access to the "session" key, the decoder address must be in the head-end data base, and (4) all information downloaded to addresses, including "session" keys, is encrypted with a unique "address key" for each decoder.

Chroma Inversion

As described, the PM scrambling is accomplished by switching the CATV modulator visual IF signal through complex filters which both attenuate and phase reverse it during each horizontal blanking period. In fact, the scrambling filters (and descrambling filters) used are complex multi-element SAW (Surface Acoustic Wave) devices of a proprietary design. Because of the particular design of these SAW filters only the channel's visual carrier, and not it's Chroma and Aural carriers, are phase and amplitude modified during the horizontal period. This technique produces some interesting effects on the TV detected video.

The major effect of the PM scrambling on a TV receiver is the loss of horizontal synchronization due to sync suppression into the luminance video range. But in addition to this standard RF scrambling effect, the detected video chroma becomes inverted. This is due to the fact that with common intercarrier TV detection processes, phase modulation is transferred to the modulating subcarriers if only the visual carrier is phase modulated. This can be shown graphically using the simplified AM modulation vector diagram of figure 8.

Figure 8 Chroma Phase Inversion



In figure 8, a carrier is being AM modulated by a pair of sideband modulation carriers. At the point of time described by the vector diagram, the resultant main carrier is nearing its peak in amplitude due to the addition of the modulation sideband vectors. This point in time is shown in the accompanying carrier diagram by a vertical line. If at the same point in time the main carrier vector (and not the modulating sideband vectors) is 180° phase reversed, the result is that the sideband vectors <u>subtract</u> from the main carrier envelope causing it to be near its <u>minimum</u> amplitude. The resulting envelope of the 180° phase shifted modulated carrier is a sinewave of the opposite phase from the case of a 0° phase carrier.

Thus, during the PM encoded horizontal blanking interval, when the Visual carrier is 180° phase shifted from normal, the color burst within that interval detected by a TV receiver would be phase inverted from normal. Since the TV color circuitry phase-locks to the color burst, this phenomenon would then invert the TV video chroma. In this fashion the PM scrambling system incorporates video manipulation, an attribute of Baseband Scrambling, into a low cost RF system.

<u>Audio Masking</u>

Just as the Chroma carrier becomes phase inverted during the PM encoded carrier reversal period, the Aural carrier, detected in an intercarrier TV receiver, becomes phase inverted. During the fast transitions from normal phase to inverted phase (and then back again), the TV FM sound detector experiences large frequency variations as the detector tries to track the huge carrier phase shifts. This effect produces large audio transients in the TV FM detector output, driving the detector into non-linearity. The resulting effect of the scrambling on the TV receiver audio is severe audio distortion and buzz.

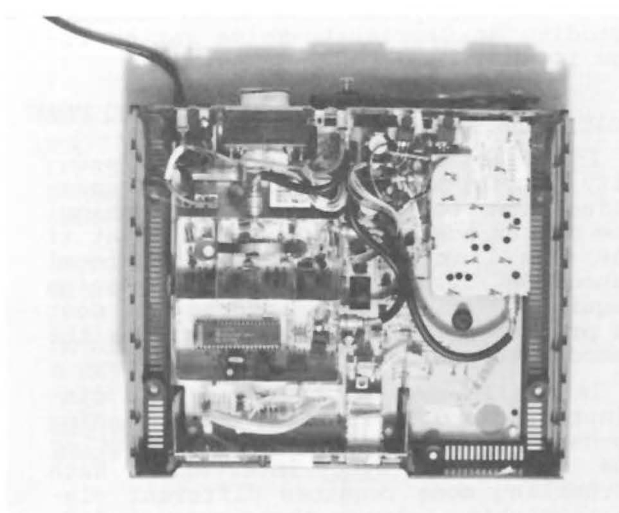
In an effort to enhance this audio distortion for unauthorized viewers a modulated audio subcarrier, similar to the BTSC stereo "SAP" subcarrier, is added by the PM Encoder to the audio signal fed to the CATV Channel modulator. Due to the severe nonlinearities that the TV Audio detector is driven to by the scrambling phase transients, the modulation of the added subcarrier becomes mixed with the buzz and detected baseband audio produced by the TV audio circuitry. This additional "garbling" effect on the TV audio is referred to as enhanced Audio Masking.

If BTSC stereo is used on the channel, effective audio masking occurs with only the standard BTSC SAP (Second Audio Program) signal. When a SAP signal is detected on the encoder audio, the Audio Masking Subcarrier is automatically turned off.

DECODER DESIGN

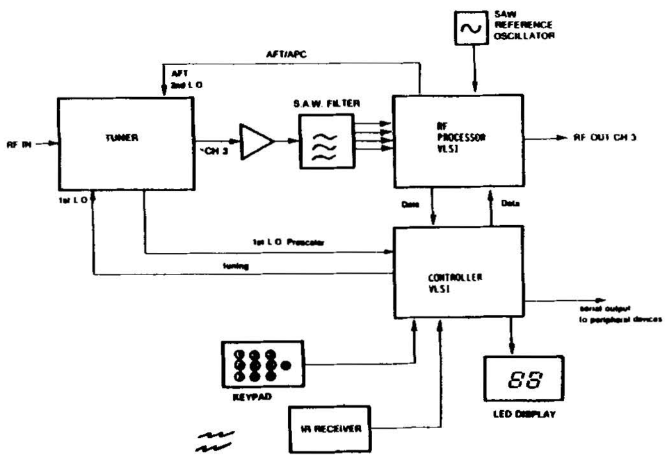
For low cost, reliability and manufacturability, the decoder design for the PM system is kept as simple as possible with few parts and minimal alignment requirements. For security, the design includes custom VLSI circuits, custom complex response SAW filters and data protection. Figure 9 is a photograph of the internal design of the decoder.

FIGURE 9
PM Decoder Internal Photo



The design uses a one board approach of minimal parts count requiring no factory electrical adjustments. In addition to the two power supply regulators, there are only three ICs on the module. The centralized microprocessor control and serial data output port in the design allows for extensive automated factory testing before shipment to ensure quality. Figure 10 shows a basic block diagram of the decoder design.

Figure 10
PM DECODER BLOCK DIAGRAM

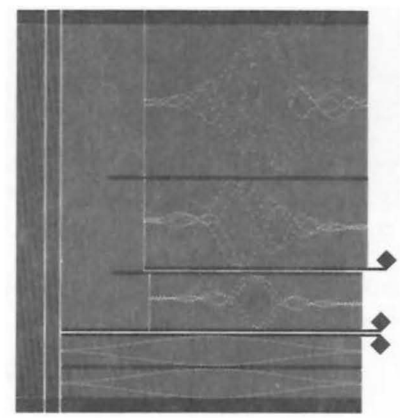


Custom Multi-Element SAW Filter

The prime decoder component necessary to restore the PM encoded RF carrier to its proper amplitude and phase is the SAW filter. The use of SAW filters for both encoding and decoding was decided upon due to the complex nature of the required amplitude and phase response for the scrambling system, precluding any unauthorized replication. Additionally, SAW filters do not require alignment. Once the SAW design is finalized, production devices are virtual clones of each other.

In the development of the SAW filters used in the PM system, an iterative process was used to perfect the final designs. Of primary importance was to be certain that the decoder SAW amplitude and phase ratios between modes matched those of the encoder SAW as closely as possible. To ensure tight mode matching tolerances, the SAWs actually incorporate 4 independent filter elements on the same substrate. The four elements are used for data channel filtering, unencoded filtering, and dual scrambling mode filtering. A die photograph of the decoder SAW, showing the multi-filter elements, is shown in figure 11.

FIGURE 11
PM Decoder SAW Filter



Good decoded signal quality requires tight tolerances between modes in both encoding and decoding. Not only do the video sync pulses need to be restored to the proper levels but the decoding side-effects on the color and audio signals must be minimized. Processing time delays must also be compensated for in order to accurately match-up decoding to encoding in the time domain. If scrambling dynamic mode switching is used, matching between encoder modes and decoder modes must also be accurate in order to avoid annoying flicker in the descrambled TV picture.

All of these requirements have been accomplished through the use of the system's custom proprietary SAW filters. The specific SAW characteristics and the lack of "off-the-shelf" parts make attempts to manufacture acceptable "pirate" decoders impractical.

Figure 12 diagrams in simple terms the amplitude responses required of the complementary encoder and decoder SAW filters. In the encoder, the relative amplitude of the response at the picture carrier frequency is attenuated at a level (B or C) dependant upon the defined scrambling mode, while the response at color and sound carriers is flat with no attenuation. The decoder SAW provides a relative amplification at the picture carrier compensating for the encoder attenuation (B or C), while maintaining color and sound at 0dB.

Figure 12
Encoder - Decoder
SAW Amplitude Response

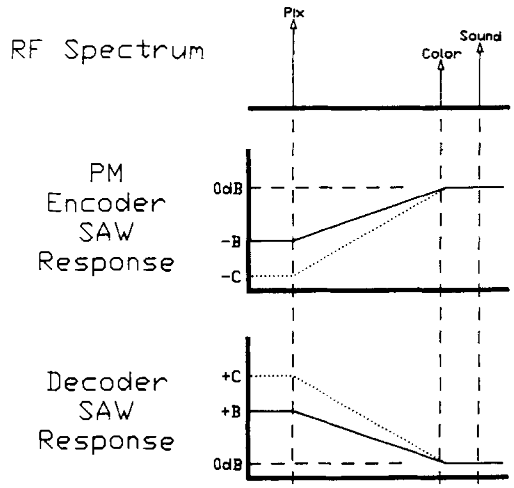
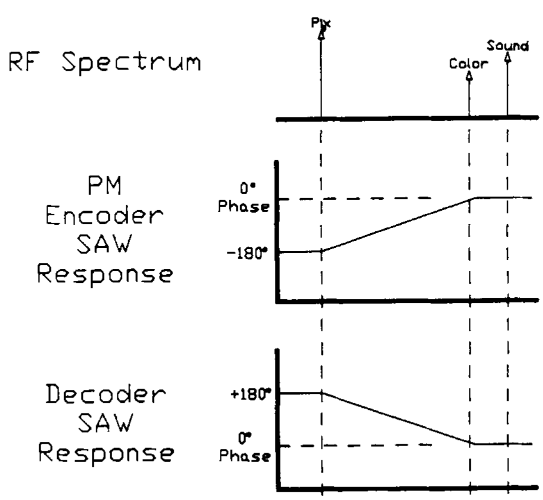


Figure 13 shows the encoder and decoder filter phase responses. Note that the encoder phase response adds a 180° phase shift at the picture carrier frequency while maintaining a 0° phase shift at color and sound carriers. The decoder response is such that it corrects for the encoder response.

Figure 13
Encoder - Decoder
SAW Phase Response



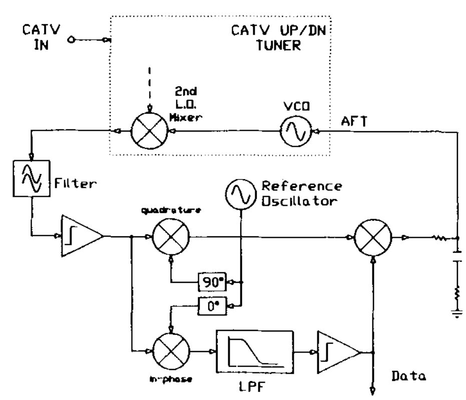
Custom RF Processing IC

The PM scrambling system phase shifts the encoded channel's Visual carrier by 180° during the video horizontal blanking interval. In order to recreate the timing signal necessary for accurate descrambling and to receive the BPSK headend data, the decoder needs to monitor and accurately track the phase of the incoming RF signal. Common PLL (Phase Lock Loop) technology has no difficulty with locking to normal signal modulation, but a signal having rapid 0° to 180° phase changes represents a problem. A Phase Lock Loop capable of being stable at both 0° and 180° phase states is necessary.

Such is the case in the custom, proprietary RF processing IC incorporated in the PM system decoder. Included in this IC is a patented FPLL (Frequency and Phase Lock Loop) which is Bi-Phase stable. The IC's FPLL serves two purposes: it not only detects the PM signal's phase modulation for timing and data but it also provides AFT for the converter CATV tuner.

Figure 14 is a simplified block diagram of the FPLL system.

Figure 14 FPLL Block Diagram



In this diagram, the RF signal tuned by the CATV tuner is amplified, limited and supplied to two multipliers. The upper multiplier, used as a normal PLL phase detector, is compared to a 90° (quadrature) phase shifted reference oscillator at the picture carrier frequency. The lower second multiplier acts as an in-phase detector comparing the input signal to the 0° phase shifted reference oscillator. If the input signal is nearly in-phase with the 0° reference oscillator, the output of the lower (in-phase) multiplier produces a "+1" signal to the third multiplier in the chain. The output of this third multiplier then linearly reproduces the phase error from the upper (quadrature) multiplier and supplies it to the tuner 2'nd local oscillator as an AFT/APC.

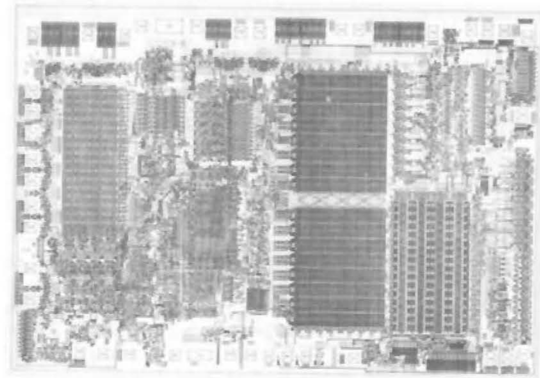
This system effectively locks the tuner output frequency to the reference oscillator, a stable SAW resonator design. If the input signal shifts 180° in phase, as in the PM encoding scheme, the second (lower) multiplier produces a "-1" signal to the third multiplier in the chain. That multiplier then inverts the phase error control signal supplied to the tuner maintaining phase lock stability. The +1 and -1 output signal from the in-phase multiplier is in fact a re-creation of the timing pulse originally used in the head-end encoding for PM scrambling. This output from the FPLL is used then as the received PM data for further processing.

Custom Micro-Controller IC

The processing of the PM data is one of the tasks of the third key component in the decoder design, the microprocessor. Largely due to security reasons, a custom microprocessor-controller was designed for the product and manufactured exclusively for Zenith.

A primary concern in the decoder design was tamper protection for the authorization and decoder address memory. "Off the shelf" parts, standard external memory ICs and battery back-up systems can all be security weaknesses in an addressable Pay-TV Descrambler. For this reason, the PM decoder microprocessor incorporates its own internal non-volatile read-write memory, not requiring a battery back-up. Part of the memory is user accessible for features such as "favorite channel scan memory". However, most of it is used for Pay TV and IPPV authorization level storage, downloadable converter characteristics and decoder address memory. As an additional security measure, the section of memory used to store the decoder address and address decryption key are permanently sealed off from future modification after initially being loaded in the decoder factory.

In addition to secure information storage, the microprocessor has the functions of PM data processing, decryption and error detection along with subscriber interface and head-end interface through Store-Forward IPPV data communication. Keyboard, IR remote control and LED display processing is controlled for customer features and converter tuning functions. An internal tuning PLL and a self diagnostic routine help assure reliable service. In order to make the product "future-friendly" several input/output ports are provided for feature flexibility through head-end control of future peripheral equipment, for Master-Slave operation and for a data communication link.



CONCLUSION

A novel approach to low cost RF addressable Pay-TV scrambling has been shown which enhances premium service security over conventional RF scrambling techniques. Through the use of carrier Phase Modulation (PM) and proprietary SAW filter responses a scrambling effect is produced having video and audio alteration in addition to typical sync suppression properties. The full-featured PM decoder design, incorporating considerations for reliability, addressing data security and noise immunity, signal quality and BTSC stereo compatibility is meeting expectations from initial installations in Cable Systems.

BIOGRAPHY

Michael Long graduated from the University of Illinois in 1969 with a BSEE and has a Master of Electrical Engineering degree from Midwest College of Engineering. He is presently Manager of RF Product Engineering for the CATV Engineering department of Zenith Electronics Corporation. In his 17 years with the company, he has held positions in Research and Development, Color TV Engineering, and VCR-Video Disc Engineering. Mr. Long holds four U.S. Patents and has three additional patents pending.

Richard Citta obtained his BSEE degree in 1969 from Illinois Institute of Technology and his MSEE degree in 1971 from the University of Washington. He is presently a Manager in the Electronic Systems Research and Development department of Zenith Electronics Corporation. Mr. Citta has published several technical papers and holds 12 U.S. patents with 6 patents pending.

AUTOMATED SIGNAL LEAKAGE MEASURING SYSTEMS

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Abstract

This paper deals with the FCC requirements for monitoring and measurement of signal leakage in conjunction with the cable operator's need for a leak free system. Moving from the basically cumbersome methods now employed, possible integration of various techniques and hardware is considered to accomplish ground and air monitoring, measurement, data reduction, display and hard copy generation. The paper suggests a "better way" to approach this very important aspect of cable system maintenance.

Introduction

During the mid 70's, the attention of the cable industry was forced to focus upon signal leakage as a result of the concerns of various users of over-the-air communication services as well as a few instances of interference by signal leakage from CATV systems. The now famous Docket 21006 was generated and FCC/FAA/Industry "Advisory Committee on Cable Signal Leakage" was formed, worked and reported. Part 76 of the FCC Regulations was modified several times leaving us today with specific regulatory responsibility for controlling CATV leakage, This requires constant vigilance in the form of continuous monitoring to detect the same, plus the task of ongoing repair. This is all in order to protect not only safety-of-life services, but all over-the-air users of the frequencies utilized "within the cable". It took the cable industry some time to come up to speed on this new perspective, however, today most cable operators have instituted monitoring and repair programs which have resulted in reduction of cable system leakage. The job of clean-up is far from complete. In 1990, the regulations require qualification of performance by way of the "Cumulative Leakage Index" or fly-over measurement of the leakage fields within the air space above the cable system. This move toward protection of over-the-air communications services will result in better cable systems since where less energy leaks out, less interference to CATV product can be expected to leak in.

It is most sobering to consider that even the best CATV components operating in the outside environment, cannot be counted upon to maintain leakage integrity indefinitely. The bottom line is that monitoring and repairing (and probably verification for the FCC) will go on indefinitely in the forseeable future. A great deal of money will be spent upon continuing labor plus equipment to perform this quality control task. Due to the labor intensive nature of this job, it would behoove us to seek more efficient methods of leak detection, location and repair as well as qualification of our results.

Perhaps, a better way...

This is certainly a situation where we need the proverbial "better way" to ease our task and improve our results. It is well to attempt to quantify that way and direct our attention toward better methods lest we become "slaves of the system". It is our objective here to suggest directions to proceed in weaning ourselves from signal level meters, handheld dipoles and tedious routines.

We seek a systematic approach, not only for monitoring and leak location but for data-taking to satisfy the FCC requirements. Such an approach must be one which takes meaningful and accurate data but does not become highly labor intensive due to the care required to take the data or the volume of data which must be assembled. In other words, we need a system which can automatically sample all necessary parameters and store large volumes of data for later reduction. The equipment must be easy to handle, install and operate and leave little room for omission and error. We are talking about a task which needs to be done both in the air and on the ground and combines leakage detection and location with the FCC qualification measurements.

After the measurements are done, it is absolutely necessary to produce good records and documentation along with flexible displays so that the leakage situation may be viewed from the perspective of the entire system as well as observation of single leaks.

While we are about it, we should leave some flexibility for "esoteric" items like multifrequency measurements which will surely enhance our understanding of leakage phenomona and hopefully lead to better practices in both construction and measurement.

Approach

With the above considerations in mind, let's dream a little about the configuration of such an integrated signal leakage measurement package. There certainly is a need for an automated and continuous data gathering device which can be used in vehicles utilizing DC power, etc. There are really two distinct vehicular requirements. The first is in an aircraft to accomplish the measurement of leakage fields in the airspace above the cable system, while the second is a ground vehicle configuration.

Airborne measurements must be done with a carefully thought-out and executed procedure. In order to satisfy the FCC requirements (Section 76.611), the overflight must be made at approximately 1500 feet above the average terrain. A grid pattern should be flown with a spacing of approximately one-half mile between passes. When you are flying in the airspace above the cable system, you must employ some system of geographical reference in order to assure the desired flight path. There are various ways of accomplishing this, the traditional one being to preselect landmarks which can be used to establish the grid pattern and visually fly by these landmarks. It is reasonable to assume that careful scrutiny of your data may be undertaken by the FCC and would require definitive notes on the landmarks employed, how the passes were flown and any discrepancies or anomalies in the flight path which might be caused by terrain obstacles, drifting off course, changing course to avoid other aircraft traffic, etc.

Ultimately, your flight path should be overlaid on a map of the area. Presentation of early flight tests included a transparent overlay of the flight path attached to a map of the system.

In terms of the flight path, a constant check on the altitude should also be maintained. Since the flight path is specified relative to the average height above the terrain, flying a constant barometric altitude (the altitude indicated by a standard calibrated aneroid altimeter) is indicated and a running check or repeated spot check of the actual altitude should be recorded. Careful consideration of the above requirement shows that maintaining the prescribed flight path and recording exactly where you've been is more than a trivial task under realistic flight conditions. It certainly would be nice to have some magic position indicating device which would feed in the actual position, including the altitude, for each datapoint taken thereby eliminating all of the problems associated with the above.

Considering data points, it behooves us to have a lot of them so that there is little chance of missing a peak in the leakage intensity during the run. The FCC regulations provide for either analog or digital data collection. Analog collection might be obtained by using a continuous chart recorder and logging the amplitude of the AGC voltage from a calibrated receiver while a digital method might utilize the same detection scheme with at least several samples per second for each frequency being monitored. In either situation Section 76.611 instructs us how to handle the data. Analog data should be smoothed "by good engineering practice" while the digital scheme uses the 90th percentile for the criterion, i.e., the level equal to or greater than 90% of the data points.

When you look at the above tasks, including the full time job of flying the airplane, it does not seem necessary to exhaustively display or attempt to analyze the data while in the air. It is also far more convenient to bring it all back to your office and then work it over from several different angles and to arrive at specific conclusions as to the magnitude and distribution of the leakage fields.

It is, however, necessary to have enough indication in the aircraft to assure you that measurements which you are taking are really measurements of the proper carrier and that you are not being overridden by spurious noise or interference. To avoid overlooking these conditions, an audio output is very useful since the ear can often detect changes in the sound which are related to spurious effects. A chart recorder output can also be useful so that unusual and unexpected occurrences can be noted and those areas be reflown if there is sufficient doubt that the data is valid. For instance if there are long periods of no measurable leakage, or extremely high signal there are reasons to question the results and probably to take another look.

The last several paragraphs describing in-flight measurements illustrate that there are many important tasks to be accomplished in airborne gathering of data. Many of these, if attempted manually, can be expected to give at best, crude results. Now sit back and imagine a "magic box" which can be hand carried to the airplane, quickly connected to power and antenna and flown in a "hands-off" manner. This box should have the audio and visual indications to assure that the data being taken is valid, but other than that, there need be little more than a "start" and "stop" button.

Position indication is magically derived and inserted along with the data, as well as indication of current altitude. If you really want to do it right, imagine a connection to the aircraft autopilot which accomplishes flying of the predetermined course at the predetermined altitude. The entire setup can be operated by a pilot who has very little knowledge of the technicalities of the test and spends virtually all of his time flying the aircraft.

This hypothetical box which we have been describing would also be very handy to install in a service vehicle and drive throughout the system. We must understand that the conditions on the ground are somewhat different from those in the overflight situation. Here we deal with single or a few leaks at any one time rather than the combined effects of many leaks. In the air, with the exception of a very clean system with a few remaining leaks, there is little indication of the location of any single leak. On the ground, however, the only leakage signals received are generally those within the immediate proximity of the vehicle. It is normal to have frequent periods of no leakage signal and distinct peaks as one drives near to specific leaks. However, problems arise in connection with those leaks which are not directly on the right-of-way which you are driving, i.e. those leaks which occur in the alleys, easements, high-rise buildings, etc. There is a dual problem here. First, a leak of minimum intensity may be 30 meters or even 100 meters away rather than the 3 meters specified as the measuring distance. The received signal then varies inversely as the distance so that a 20 microvolt per meter at 3 meters leak will produce only 2 microvolts per meter at 30 meters making the requirements for receiver sensitivity much more stringent.

When monitoring leaks on the ground, the job is incomplete unless the leak can be found. When the leak occurs on the strand overhead of the vehicle, finding it is fairly routine. However, when the leak is in the alley, easement or high-rise, some means of direction and range finding is required for location or else the whole exercise becomes futile. This is the place for a little more magic to aid in actual leak location. The availability of position data to tie to the leakage data is extremely valuable along with distance and directional information for location of remote leaks. In the ground case, immediate readouts of distance and direction will facilitate efficient leak location.

Results

Now that we've taken this magic box and run it through its paces both in the air and on the ground, what can be derived from the mass of data accumulated? We must first verify compliance with the FCC regulations and second aid in the overall assessment of the cable signal leakage. Obviously we can easily compile tables of leakage versus location in terms of the route (simply the chart recorder plot) and attempt to (laboriously) relate this back to the actual system map. Given location information for each data point, it is possible to program a computer to reduce this data and produce a contour presentation of the system leakage patterns. Why not pop this whole thing into the PC on your desk and be presented with plots or signal leakage contours relative to an actual system map? You can even produce a color display to highlight levels, etc.

It is easy to see that such a presentation is easier said than done particularly in matters of "where does the map come from?", or "how do you get it into the computer?" In general, "how do you put the whole thing together so that the presentation is useful and informative?" It is evident that hardware and software can be assembled to achieve all of these tasks. Described above is sort of "what the world needs" which may not be too far from practical realization.

BROADCASTING HIGH DEFINITION TELEVISION

Ben Crutchfield
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Department of Science and Technology
National Association of Broadcasters

ABSTRACT

High definition television has been broadcast over the air in both the UHF TV and 13 GHz bands. This first demonstration broadcast, sponsored by the National Association of Broadcasters and the Association of Maximum Service Telecasters, used the 1125 line, 60 field/sec. system developed by NHK. For transmission, the wideband video signal was compressed to 8.1 MHz using the MUSE system, also developed by NHK. This signal was transmitted by vestigial sideband AM in the UHF band and FM at 13 GHz.

Receiving points include the NAB, the FCC, and Congress. Programming includes a variety of materials produced on high definition video tape and other materials produced on 35mm film and transferred to tape.

The purposes of this project are to: promote development of a system for terrestrial broadcast of HDTV; assess the performance of proposed systems, and; demonstrate to the FCC and Congress the need for adequate spectrum to allow terrestrial broadcast of HDTV.

This paper describes the demonstration setup and results and discusses the plans for future activities.

INTRODUCTION

On January 7, 1987, experimental television station WWHD-TV signed on in Washington, DC. The operation of this station demonstrated that high definition television programs could be delivered to viewers by means that many had dismissed:

- Terrestrial (not satellite) broadcast,
- In the UHF television band,
- Using conventional, spectrum-efficient vestigial sideband amplitude modulation (VSB-AM).

This historic broadcast was commemorated in a ceremony at the Federal Communications Commission where broadcast HDTV programs were received and displayed on direct view and rear projection equipment. There were also operating versions of a 1/2 inch video cassette recorder-player and a video disc player.

After more than three weeks of daily operations, WWHD-TV signed off at the end of January, completing the first in a planned series of demonstration broadcasts by the National Association of Broadcasters and the Association of Maximum Service Telecasters.

OBJECTIVE

The objective of this project is to demonstrate to the broadcast industry, its regulators, and the public that significant advances in the technical improvement of television have reached the point where they may be delivered, at least on an experimental basis, via terrestrial broadcast methods.

At present, HDTV is being used to produce program materials ranging from commercials to feature length "films." Results have been good, both aesthetically and technically, and producers report significant savings of money and time compared with the use of 35mm film. HDTV combines the image quality of 35mm film with the speed and flexibility of electronic production and post production.

While these productions must, at present, be converted to standard television or film for distribution, we expect to see consumer HDTV equipment appear on the market in the U.S. in about five years. Prototypes of video cassette and disc players have been demonstrated; work is now beginning on the design for VLSI chips needed for mass production of affordable units. Chip development will take about three years and product design, another two years.

For broadcasters, it is particularly important to begin working on a system for HDTV delivery immediately. Any system will take time to develop, but more urgently, it is essential to determine the spectrum requirements and then secure the necessary spectrum. Other services are pressing for more spectrum, including spectrum already allocated to television broadcast. Without spectrum, broadcasters will probably be at a competitive disadvantage with respect to other media. We are acutely aware of what happened to AM radio when FM "caught on." In a decade, a major shift occurred as audiences turned to "high definition radio." We have also seen the sudden shift from vinyl audio disc to CD, abated only by anticipation of the arrival of digital audio tape (DAT) recorders. The common thread is consumer demand for higher quality.

ADVANCED TELEVISION SYSTEMS

Production

There is a wide range of improved television systems. Scientists and engineers in the U.S., Japan and Europe are studying or actually testing various approaches. At one end of the technological spectrum are improvements to the existing NTSC (or PAL or SECAM) standard. These apply to one or more phases of the imaging, encoding, processing and display systems and are generally compatible with existing equipment.

Because of compatibility, these improvements may be tested with existing broadcast facilities. No special test setup or authorization is required. Furthermore, much of the work being done is in the area of receiver improvement and appears on the market in the form of the latest model receiver.

The next big step in improvement of picture quality, however, is expected to come in the form of a system incorporating some of the following features:

- Increased number of scanning lines (possibly double or more) for higher vertical resolution,
- Wider bandwidth for higher horizontal resolution,
- Higher frame rate to reduce flicker,
- Progressive scanning,
- Wider aspect ratio,
- Separation of luminance and color signals

Probably the most important factor, from the point of view of the broadcaster, is the increase in bandwidth of the final output signal. It now appears that making a major improvement in picture quality means, inescapably, transmitting more information per unit time, which means more bandwidth. The HDTV production system used in the demonstration has a bandwidth over 20 MHz.

This 1125 line system was developed by NHK. Design criteria were based on the performance of 35mm film, including the wider aspect ratio, 5.3:3. An important overall design objective, however, was relative picture size for the viewer. It has been found that people tend to view NTSC pictures from a distance of six to seven times the picture height. For a 19-inch (diagonal) picture, the height is about 12 inches and the average viewing distance is about six to seven feet. For 35mm film, people tend to select a seat at a distance of three to four times picture height. Since the relative distance is half, the picture covers twice the visual angle for the viewer or four times the area for a picture of same aspect ratio. Most 35mm film uses an aspect ratio on the order of 5:3 so a 35mm picture actually covers closer to five times the visual area of a conventional television picture.

The design objective was, then, not to produce simply a clearer picture but to produce a picture clear enough that viewers would sit closer and experience a greater sense of reality. It is not coincidence that making a picture with five times the relative area requires around five times the bandwidth: over 20 MHz instead of 4.2 MHz.

Transmission

Delivering to the consumer a signal as wide as four or five conventional television channels is not practical for today's media. Those that could, such as CATV or multichannel multipoint distribution services (MMDS), would probably find that one channel of HDTV programming would not generate enough revenue to warrant the trade-off of the four or five conventional channels. Tape is not a likely medium in the foreseeable future, either, since 20 MHz is well beyond the capacity of professional one inch video tape recorders currently used in studios.

For terrestrial broadcasters, the possibility of finding enough spectrum to accommodate such wide signals is practically nil.

Fortunately, the technology that gave us laptop computers gives us a way to process picture information as data. It is no longer necessary that the television system be the same from studio camera to home receiver, as it is now. High speed processors and cheap memory provide the means to convert picture information to digital data and perform sophisticated processing on that data. After the program production and post production are completed, using full bandwidth recordings, the finished product can be transmitted by a system which transmits only enough information to satisfy the eye.

Researchers working with video have known for a long time that much of the information in a video signal is redundant. Unless there is much rapid action or a sharp transition between scenes, it is often hard to distinguish one frame from the next. Nonetheless, current television systems, including NTSC, update every part of the picture every time a frame is scanned, and this is true from camera to receiver.

Another relevant fact which has been known for some time is that the human visual system is less sensitive to detail in a scene or part of a scene which is moving. If, for example, the eye (or camera) is fixed on a scene and a car drives through, the eye will distinguish more detail in the unmoving scenery and less detail in the car.

This means that it is possible to build a television transmission system which updates visual information on moving parts of a scene more frequently but with less detail and updates static parts less frequently but with more detail. While there are ways to do this, the net result is a reduction in the amount of information which must be transmitted in a given period of time, and that, of course, means a reduction in bandwidth.

The most highly developed of these systems is MUSE, developed in Japan by NHK. MUSE stands for MUltiple Subnyquist sampling and Encoding. The encoder at the transmitter takes the wideband video signal and reduces it to 8.1 MHz, baseband. At the receiving end, the decoder reconstitutes the signal for display on an HDTV monitor. Sampling and processing are done digitally, but the transmitted signal is analog for further spectrum efficiency. See Table 1 for characteristics of MUSE.

The MUSE Demonstration

In the first demonstration, we used the MUSE system, developed by NHK for satellite, cassette and tape delivery of HDTV. An important, non-technical advantage of MUSE is that it is a working system. It has been tested by NHK in both satellite and terrestrial broadcast applications. The NAB-MST demonstration used vestigial sideband amplitude

Table 1 Characteristics of the MUSE system

System		Motion-compensated multiple subsampling system (Multiplexing of Y and C signal is by TCI format.)
Scanning		1125/60 2:1
Bandwidth of transmission baseband signal		8.1 MHz (-6 dB)
Resampling clock rate		16.2 MHz
Horizontal bandwidth	(Y)	20-22 MHz (for stationary portion of the picture) 12.5 MHz (for moving portion of the picture)*
	(C)	7.0 MHz (for stationary portion of the picture) 3.1 (for moving portion of the picture)*
Synchronization		Positive digital sync
* Values for a prototype receiver: these values should be 16 MHz and 4 MHz, if a perfect digital two-dimensional filter could be used.		

modulation (VSB-AM) for the UHF channel and frequency modulation (FM) for a 13 GHz channel. The latter was operated as a backup and as to demonstrate the possibility of using the SHF band where UHF may not be available.

The results were very satisfactory: there were no major difficulties in setting up or operating the system and clear pictures were delivered to all receiving sites. Where ghosts were found, they were removed by antenna positioning and a ghost-cancelling circuit built on the site by NHK engineers. Please see Table 2 for a summary of system data.

There was some concern about the audio signal. In tests where the signal was attenuated, the PCM audio was lost while the picture was still viewable. The sudden cutoff is characteristic of a PCM when the decoder cannot get sufficient data to operate. This would not normally be a problem for satellite use, for which the MUSE system was designed, but further study and work will be necessary to use the MUSE audio for terrestrial broadcast.

In general, MUSE proved to be a very "robust" signal. It looked good at all receiving sites and showed no unusual susceptibility to noise or interference. We were concerned that the motion sensing and compensating circuits might be easily disrupted but there was no evidence that this was the case. We were also concerned that it might take considerable time to set up and adjust the encoder and decoders after a long air and truck trip, particularly since these were developmental rather than production units. That everything worked was a very pleasant surprise, given our tight schedule. In fact, the large and complex encoder, consisting of four tightly-packed racks, was operational within hours of delivery.

CONCLUSIONS AND PLANS

The project described here is to demonstrate the feasibility of terrestrial broadcasting of HDTV. This first phase has successfully demonstrated one possible system and has called the attention of the television industry to the need to be prepared for the arrival of the next generation of television technology.

Several other systems have been proposed for broadcasting HDTV. One which may be available soon is being developed by the New York Institute of Technology. It uses two television signals which may be transmitted on separate channels. This approach involves transmitting a standard NTSC signal on one channel with additional picture information being transmitted on the second channel. Conventional receivers will work with the first channel while high definition receivers will combine the two channels to produce a high definition picture.

To analyze these and other systems, the Advanced Television Systems Committee (ATSC) has established a Technical Specialist Group on HDTV Transmission. ATSC is composed of representatives of the broadcasting and cable industries as well as consumer equipment manufacturers. They are seeking a single standard or family of standards which will serve the needs of all areas of the television industry and, ultimately, the consumer.

This work is considered very important as it will likely set the technical parameters for the development of television well into the next century.

Table 2 WWHD-TV MUSE Demonstration, Washington, DC

Path Description

Transmission Site:	WUSA-TV (Gannet, channel 9, Washington DC), auxilliary tower.
Path Length:	4 to 6 miles.
Receiving Sites:	Federal Communications Commission, 1919 M St. NW.; NAB, 1771 N St. NW; Russell Senate Office Building, 1st & D Sts, NE

UHF System

Encoder:	NHK MUSE II HDTV encoder, analogue video output 8.1 MHz including 2 PCM audio channels.
Transmitter:	ITS Model 20H/257H, vestigial sideband AM, UHF TV channels 59-59, nominal 50 watts (PEP) output, broadbanded for 8.1 MHz baseband input.
Transmission Line:	Cablewave 1 5/8 inch foam, corrugated copper, 350 ft., attenuation approx. 2.5 dB.
Transmitting Antenna:	Micro Communications special design UHF horn, gain 17.8, VSWR across band 1.03:1, ERP approx. 1300 watts.
Preamplifier:	Maspro 30U, broadband, 30 dB gain.*
Receiving Antenna:	Scala PR-450U parabolic reflector, gain 16, VSWR 1.2:1, received signal (typical) -48 dBm free space, -32 dBm into downlead.
Downlead:	Belden 8281 precision coax, 75 ohm.
Receiver:	Nihon Tsushinki precision TV demodulator, special design provided by NHK, output is 8.1 MHz to MUSE II decoder. input adjusted for nominal -37 dBm for design 55 dB S/N.

* Note: Attenuators and filters were used as needed, depending on the location, to prevent overload from high powered UHF-TV signals in the area.

SHF System

Transmission System:	Harris FV13MP 13 GHz portable microwave transmitter and receiver, filters bypassed for broadband operation, nominal 63 mW, 2 ft. dish., approx. -55 dBm to receiver for 56 dB S/N.

BTSC PERFORMANCE MEASUREMENT IN THE LAB AND FIELD

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ABSTRACT

The U.S. television viewing public has recently been turned onto the phenomenon of stereo TV thanks to the development of the BTSC multi-channel sound system. The BTSC system offers the ease of inband tuning, backward compatibility with monaural receiving equipment and true high fidelity stereo performance. Cable television viewers are in a favorable position to enjoy stereo TV even more because of the wealth of high fidelity stereo programming available on many of the basic and premium services (19 cable channels are in stereo already).

The great potential that BTSC stereo over cable offers will not be realized if cable operators provide anything less than true, high quality BTSC stereo. There is a crying need for reasonably priced test equipment that a cable operator can utilize in obtaining a quantitative measure of the signal quality he is delivering. There is also an immediate need for the development of training programs to bring system engineers and technicians up to working speed with this new audio signal processing area.

All to often, BTSC encoding equipment is installed in the easiest manner rather than by a method which will yield the highest quality the system is capable of and which subscribers deserve. This paper will describe the recommended test equipment and techniques which encoder manufacturers and cable system operators may use to ensure that the best possible signal is being provided to the subscriber, within the confines of a typical cable system environment. A discussion of signal quality concerns from the source through the satellite link, FML, AML, headend distribution, and subscriber equipment will be included.

BTSC TODAY

BTSC stereo is here to stay. At the present, stereo TV sales are the fastest growing segment of the TV market. The growth curve for stereo TV is similar to that of the compact disc player, a product which has rejuvinated the music industry. The National Broadcasting Network (NBC) has contributed greatly to the growth of stereo TV by converting the majority of its prime time programming over to stereo, much as it led the way with color broadcasting.

The Cable TV world has seen premium programming services such as HBO and Showtime/TMC as the major promoters of stereo broadcasting via satellite. In this area it has grown quickly and its application on satellite delivered channels will be commonplace this year. The performance that the BTSC system can deliver in the CATV environment is nothing short of excellent. Achieving the best performance, however requires an understanding of the system and its potential pitfalls.

Although there have been numerous technical papers published outlining the difficulties with BTSC stereo, this paper will present test data showing new levels of performance which have been achieved, describe how the measurements are made, what test equipment is required and what the significance of the measurements are. Recommended test equipment will be described ranging from the sophisticated to the simple to fit the budget of the most cost conscious cable operators.

In addition, a step by step approach to checking performance in the field will be outlined, along with techniques for correcting commonly encountered problems. Specific topics discussed include: video modulation; video bandwidth; group delay; ICPM; VITS; relative carrier amplitudes; hum; buzz, THD; S/N; dynamic range; deviation adjustment and stereo separation.

GARBAGE IN, GARBAGE OUT

Cable operators have a responsibility to their subscribers to provide the highest quality signals feasible. However, if the service provider does not start with a quality source, there is little that a cable operator can do. It is up to the major MSO's to pressure the service providers to live up to the capabilities of the transmission system. These service providers must remember that the growth in popularity of stereo and HI-FI VCR's have established a new level of home video entertainment expectations.

If a film, for example, was originally recorded in stereo (and is available on videocassette in stereo), it should appear in stereo on the premium service. If the programmer was not able to obtain a stereo version of a current movie, selective, tasteful stereo synthesis at the uplink should be used as a service to customers equipped for stereo reception. High quality synthesis is economically justifiable at an uplink, and indeed, should not be the responsibility of the cable operator who is paying a monthly fee for the service.

Another example of poor performance can be found on a number of music video services. Many of the tapes they play regularly have totally unacceptable levels of hiss on them. This is 1987 not 1967! These services should be reminded that their growth in popularity was due to a desire for enhanced video/audio entertainment. Many viewers have invested considerable sums in their entertainment systems only to be bombarded with poor quality audio tracks. Satellite Services which use VideoCipher scrambling have at their disposal a transmission link capable of providing a quality audio signal ideally suited to BTSC stereo encoding. VideoCipher incorporates 14 Bit Digital Audio companded to 10 Bit, yielding 84 dB dynamic range and 60 dB instantaneous signal to noise.

GETTING IT THERE SAFELY

Once a satellite delivered service is received, it is necessary in some situations to transfer or retransmit it to distant locations. If FM modulation is utilized (i.e. FML or FM supertrunk) it is strongly recommended that the discrete left and right audio signals modulate separate FM subcarriers. Using a 4.5 MHz modulated BTSC signal as a subcarrier of the video is generally a poor choice for FML. Degradation of audio signal to noise and dynamic range will definetely occur due to video interference (buzz). Placing the BTSC stereo encoder at the receiving end of the FM link will give the proper performance.

Other remote applications might require AML. This transmission technique is broadband and nearly linear. Reducing the sound carrier level to -17 dB with respect to the video carrier will have a small, negative effect on BTSC noise performance compared with -15 dB. But the dbx companding, which is a required part of the BTSC stereo system, does an excellent job of eliminating transmission hiss.

A common pitfall in interfacing BTSC encoders to modulators is to try to combine 4.5 MHz and video into the video modulator as a single input. This will result in reduced video or audio quality, or both. This is because the 4.5 MHz trap in the video path, before the video chopper modulator, is not wide enough or deep enough and as a result some of the stereo sidebands will pass through the video path causing interference. In order to avoid this interference the 4.5 MHz sound carrier must be kept low relative to the video; however, with the sound carrier low, the same problem as mentioned with FML will occur. The high frequency video components cross over the audio spectrum and create buzz in the audio. This type of interface is also likely to reduce stereo separation. Not only is the 4.5 MHz trap too narrow, but the 4.5 MHz bandpass filter is often not wide enough. The final IF audio spectrum which is a result of the video and audio paths would by no means be flat in frequency or group delay response.

When faced with interfacing with a modulator that does not have either a 4.5 MHz or a 41.25 MHz Input, it is best to either: 1) modify the unit according to the manufacturer's procedure by adding a 4.5 MHz or 41.25 MHz input, or; 2) replace the modulator with another unit which has separate inputs.

The distribution system, being broadband and linear has very little effect on BTSC audible performance. System noise is well masked in BTSC by the dbx noise reduction system. Customers will complain about noisy pictures long before the signal degrades because of the system introduced noise (hiss) in the stereo audio. Stereo separation is essentially unaffected by the cable distribution system.

On the individual subscribers side, RF converters are generally transparent to BTSC stereo. First generation baseband decoders generally do not pass stereo through because of the de-emphasis and low pass filtering used. Second generation baseband converters, however, have been designed for wide audio bandwidth, thus allowing stereo pass-through. Volume control, which is a very popular feature on baseband converters, is accomplished by changing audio deviation which changes stereo separation. Whenever the volume control is operated in the vicinity ($\pm$6 dB) of the unity gain point (where deviation in equals deviation out) relatively good stereo separation can be achieved. Thus, reasonable although less than ideal, stereo performance can still be achieved when using the volume control feature of a baseband converter. A point worth noting here is that recognizable directionality is maintained to as low as 6 dB separation, which means that many subscribers will continue to use the volume control feature of their baseband converter with their new stereo TV receivers and be pleased with the performance.

With regard to stereo separation, a general rule of thumb for acceptable performance would follow the following scheme. The stereo encoder manufacturers should aim for 40 dB of separation across the full bandwidth (50Hz to 14 KHz), operators should delivery 30 dB in practice and TV manufacturers (as well as stand-alone decoder manufacturers) should deliver product with 20dB or better separation. Following these guidelines will yield separation that is perceived to be perfect as far as audio detectability in the typical home viewing/listening environment is concerned.

Separation is only a small fraction of what determines the quality of the delivered sound. Stereo TV's and VCR's generally have not set any milestones in audio quality. However, in comparison with the audio system of a typical monaural TV, they can be quite impressive. Since there have been indications that quality audio helps sell high end sets it can be expected that more emphasis will be placed on audio reproduction in the future. We have already seen the introduction of a unique Bose audio system for TV's and there will be several sets produced with Dolby™ surround sound decoders built-in later this year. The dual detector receiving system developed by General Instrument will become available for licensing later this year, thus allowing TV receiver manufacturers to substantially reduce buzz levels.

WHAT EXACTLY AM I DELIVERING?

After purchasing and installing BTSC encoding equipment, the first question that comes to mind is how good is the equipment I have just installed and how well is it functioning in my particular application. Figures 1 and 2 illustrate the test equipment required and its interconnection for making pertinent BTSC measurements. BTSC set-up is an exercise in analog precision and as such, the test equipment to authoritatively make numerical measurements is fairly expensive. As shown, the price is in excess of $100000. The equipment listed is not meant to be an exclusive list, rather it is just one example of equipment which has proven to be effecitve.

The first test most often desired is stereo separation. In order to perform testing to 40 dB accuracy, the baseband input level to a reference decoder must be set within a few tenths of a percent. This requires calibration at the start of each testing session. The only method of deviation adjustment offering this level of precision is with a sine wave, monitored by a frequency counter and a bessel null greater than 60 dB observed on an RF spectrum analyzer. The measuring set-up must be calibrated before testing can begin. The best technique is to apply a very slowly swept (20 to 50 sec.) audio tone to one channel of the stereo encoder and measuring the output on that channel from the stereo decoder. Use the desired channel output as a reference and then measure the output of the undesired (not driven) channel. It is very important to sweep the audio at a low enough level so that as pre-emphasis occurs, overmodulation will not. Specifically, this requires a sweep at about 20 dB below the full scale point for a low frequency signal such as 300 Hz.

Frequency response is another test which is a useful measure of the level of performance of a stereo encoder. Generally, lower priced encoders will not provide the full 15 KHz bandwith specified for BTSC. In order to test the performance of only the encoder and not the combination of the encoder/decoder, it is best to switch the encoder into mono operation, disable the pre-emphasis, and connect the encoder baseband output directly to the measuring device (i.e. the network analyzer). The displayed sweep will indicate the response of all filtering in the L plus R path which must be identical to that in the L minus R path in order to achieve stereo separation. Full 15 KHz bandwidth is desireable in order to keep as much detail of the original sound source as possible. Although most stereo TV receivers today do not reproduce 15 KHz, it is reasonable to assume that high end products will do so in the very near future.

Total harmonic distortion (THD) is a measure of the maximum received signal level versus the level of all signals present in the audio band except the fundamental. With professional equipment and extreme care in eliminating ground loops, it is possible to get meaningful readings of harmonic distortion with a distortion analyzer, expecially with separate sound detection or video removed. Usually, THD is measured to check for non-linearity. If the measurement is affected by spurious signals, an audio spectrum analyzer will be easier to use for reading THD. This is generally the case with BTSC. Note that THD will vary with frequency and may become non-harmonic, especially at high frequencies, due to intermodulation effects of the L - R pilot and L and R signal components.

Dynamic range is the measure of the ratio of the largest signal received (i.e. full modulation) to what is left over when no signal is transmitted. In the BTSC system, dynamic range is not a single number, but rather a curve which varies wth frequency. Signal-to-noise ratio is another measurement which is often confused with dynamic range. In a companded system such as BTSC, signal-to-noise is measured by applying a signal and viewing the noise present with the signal on an audio spectrum analyzer. Then, you must calculate the difference between signal and noise based on resolution bandwidth and correction factors.

Subjectively, dynamic range indicates how quiet the noise and interference will be in between audio passages. Signal-to-noise ratio indicates how clean, or pure, the audio will sound. Due to video interference, this is the weak link in BTSC stereo. Dynamic range and S/N are the areas where the cable operator, not the stereo encoder manufacturer, has control and his or her actions can have the greatest positive or negative effect.

TABLE I: RECENT TEST DATA

with TEST CONFIGURATIONS OF FIG. 1 & 2

CHARACTERISTIC	MEASUREMENTS	
FREQUENCY RESPONSE	15 KHz	-1.5 dB
	15.734 KHz	-73 dB
DYNAMIC RANGE ● 1 KHz	BASEBAND	>85 dB
	RF	78 dB
THD ● 1 KHz	BASEBAND	0.03 %
	RF	0.15 %
STEREO SEPARATION	50 to 14 KHz	RF MEASURED
	-WORST POINT	35 dB
	-BEST POINT	70 dB

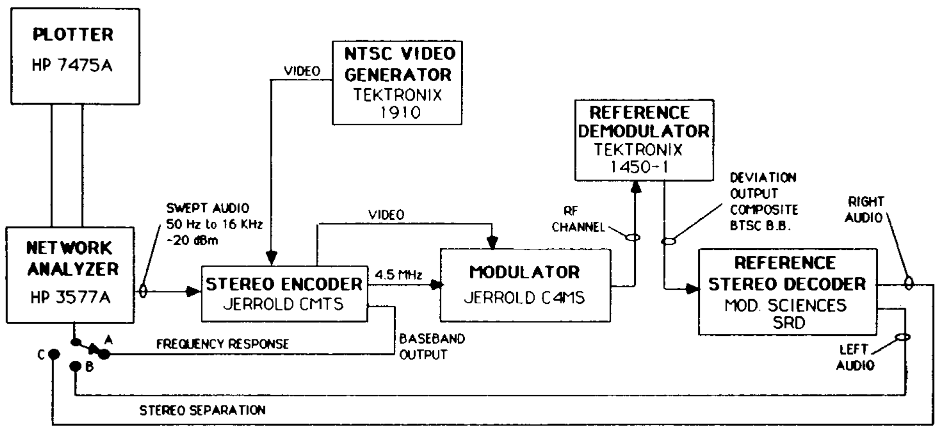

FIGURE 1: AUDIO PERFORMANCE TEST CONFIGURATION
-SEPARATION & FREQUENCY RESPONSE

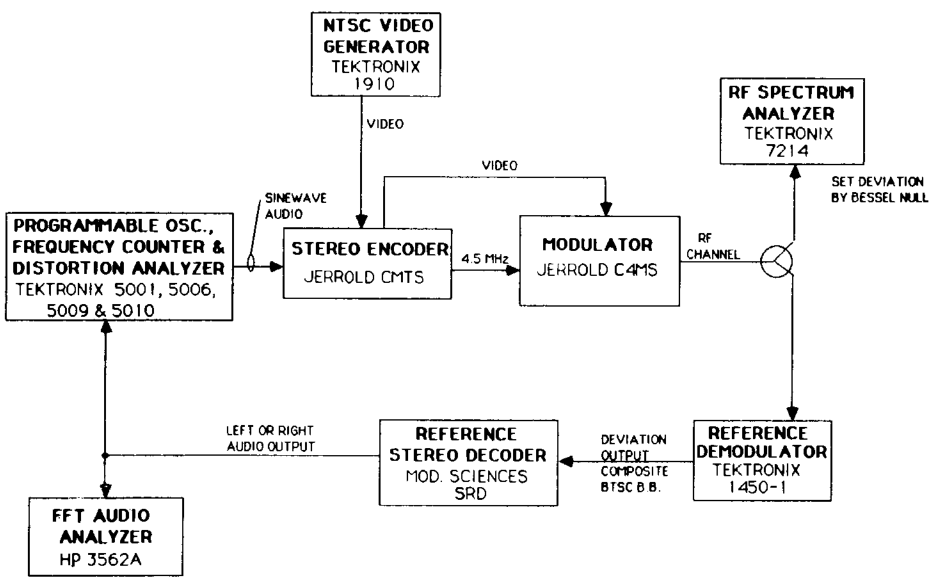

FIGURE 2: AUDIO PERFORMANCE TEST CONFIGURATION
-DEVIATION, DYNAMIC RANGE, THD & S/N

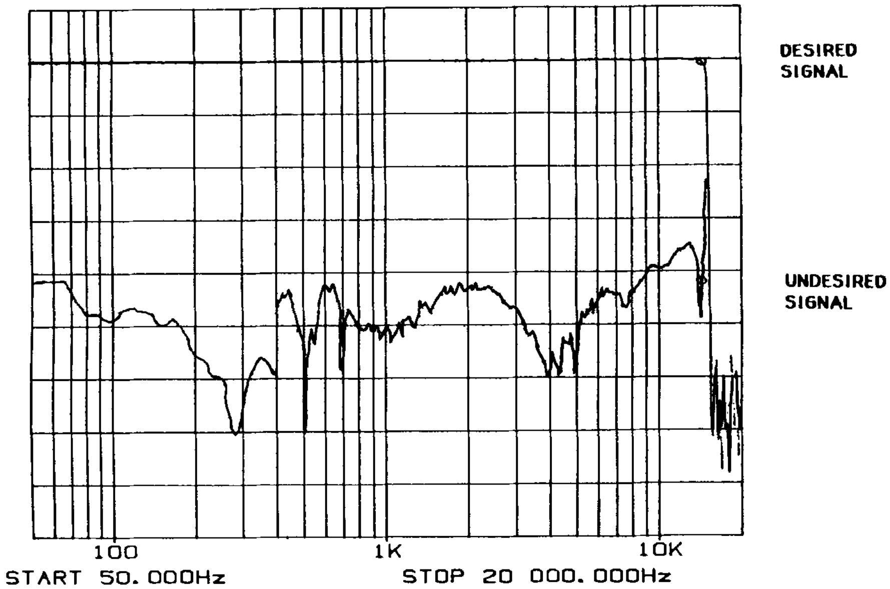

FIGURE 3: RESULTANT STEREO SEPARATION VERSUS FREQUENCY
with TEST CONFIGURATION SHOWN IN FIGURE 1.

Obviously, the test set-ups described in Figures 1 and 2 are not practical for normal on-site testing since this kind of equipment does not usually exist at the typical system headend. What follows is a step-by-step procedure for installing and testing BTSC encoders to insure optimum performance on-site, concentrating on the area where the cable engineer is in control - the elimination of unnecessary video interference. The primary measuring instrument here is a pair of human ears. Getting in the practice of using them correctly is the first step in providing quality stereo TV.

HOW TO PREP A CHANNEL AND INSTALL A STEREO ENCODER

The following is a minimum list of test equipment required for this procedure:

1. RF Spectrum Analyzer
2. Dual Trace Oscilloscope
3. Stereo Amplifier
4. Stereo Headphones
5. Television Stereo Receiver (BTSC)

Step 1 - RF Interference Check

Connect the RF spectrum analyzer to a system test point. Tune to the channel under test. Switch the modulator power off. Check for interference within the channel limits. Any signals present should be no greater than 65 dB below the video carrier level. If interference is found to be present, locate the source and eliminate the problem before proceeding.

Step 2 - Video Interference Check

Disconnect the audio input to the modulator. Switch the modulator power on. Measure the video depth of modulation. Do not trust panel meters or indicator lights. Be certain that maximum white level, such as during VITS signals, does not exceed 87.5% modulation (18 dB on a log scale or 7 out of 8 divisions on a linear scale). If overmodulation exists at any time during the video fields (including VITS) reduce the modulation. It is better to be slightly under rather than slightly overmodulated. Overmodulation will cause buzz in stereo TV receivers.

Tune the spectrum analyzer to the desired channel sound carrier. Observe the area on both sides of the sound carrier. There should be no video modulation sidebands at a level higher than 60 dB below peak video carrier level within +/-150 KHz of the sound carrier. If video sidebands are present in the sound carrier area greater than -60 db, a video low pass filter should be installed in the video path before the modulator. To prevent picture quality degradation the low pass filter should be delay equalized within 50 nanoseconds to 4.08 MHz.

Step 3 - ICPM Check

Connect the television stereo receiver audio outputs to the stereo amplifier. Connect the TV stereo receiver to the system test point. Note: Do not overload the TV receiver, +10 dBmV is a recommended level. It may be desirable to use an RF converter in front of the TV receiver in order to tune all channels or eliminate cross-mod problems. Tune to the channel under test. With video present, but no audio, listen on the headphones for buzz, increasing the amplifier volume until buzz becomes clearly audible. Remove the headphones. Reconnect the audio source to the modulator and verify that proper deviation occurs. Carefully start to replace the headphones. If the audio is too loud to listen to, then the video modulator ICPM is acceptable. If the audio is not too loud, then this indicates excessive ICPM and adjustment is necessary to remove audible buzz during quiet passages. To adjust for minimum ICPM; remove the audio input. Listen to the buzz at high volume level. Adjust the trim capacitor(s) on the diode chopper modulator starting from minimum capacitance and increasing slowly until minimum buzz level is heard. Fortunately, best video performance occurs at the same point as minimum ICPM buzz. If you are not comfortable making this adjustment, arrange for servicing by a qualified service center. Be sure to indicate that minimum ICPM (<2 degrees) is required.

Step 4 - Left and Right Audio Source Check

Connect the L and R audio outputs of the satellite subcarrier receiver or decryptor to the auxilliary inputs of the stereo amplifier. Listen to the sound quality on the headphones. The audio should be free of hiss, hum, buzz, pops, clicks, or obvious distortion. Move the balance control all the way to each side and listen to each channel carefully. If audio problems exist, troubleshoot and correct the problem before proceeding.

Connect the left channel subcarrier receiver output to both channels of the dual trace oscilloscope, Adjust the gain and position controls so that the identical waveform is observed on both channels. Connect one scope channel to the right audio output. If level adjustments are available, adjust the outputs for about 2 Vpp on audio peaks. Check that both channels are of the same polarity, that is rising and falling together most of the time on the low frequency portions of the audio. The polarity of both channels must be the same for proper stereo operation.

Step 5 - Connecting Audio to the Stereo Encoder

Connect the L and R audio from the subcarrier receiver or decryptor to the left and right inputs of the stereo encoder. Balanced connections are desirable and should be used where available. Correct polarity must be maintained. After making connection according to the encoder installation manual, verify polarity with the oscilloscope as in Step 4. Adjust the input level controls for proper bar graph indications. Switch the stereo encoder to mono operation. Connect the stereo encoder composite audio baseband output to one of the auxiliary inputs of the stereo amplifier. Switch the subcarrier receiver power off. Increase the volume of the amplifier and listen on the headphones for hum indicating ground loop problems. If excessive hum is present, experiment with connecting and disconnecting the audio cable ground at the stereo encoder, while listening for minimum hum. If possible, power the amplifier, subcarrier receiver, stereo encoder, and modulator from the same power strip. After minimizing the hum, reduce the stereo amplifier volume and switch the subcarrier receiver power back on. Listen to the L and R (mono portion) of the stereo encoder output. It should not be distorted or have hum, buzz, or hiss.

Step 6 - Modulator - Encoder Interface

Connect the 4.5 MHz or 41.25 MHz from the stereo encoder to the modulator as indicated in the encoder installation manual for your desired configuration. Set the sound carrier level by observing at the system test point with the RF spectrum analyzer. For stereo performance, the higher the sound carrier the better within FCC limits. For example, -13 dB from video carrier is preferable to -17 dB.

Step 7 - Performance Verification

Connect the TV stereo receiver to the stereo amp and to the system test point. Tune to the desired channel. Listen to the program audio. It should be free of hum, buzz, hiss or distortion. Now remove the left channel audio input to the stereo encoder. Listen to the headphones. The sound should be entirely on the right channel. Connect the oscilloscope probes to the left and right TV stereo receiver outputs. Measure the relative amplitude of the peak to peak signal present on the two channels. If you are using a consumer grade receiver, the level should be about ten times larger on the desired channel than the undesired. A commercial grade receiver should indicate thrity times larger or more. Reconnect the left input. Your are now ready for stereo operation.

B T S C   S T E R E O

Implementation on Normal and Scrambled Channels

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Houston, Texas

ABSTRACT

The decision to begin transmission of stereo programming to subscribers involved a process that required several years. Construction of the Houston system began in 1980. In 1982 we began to review the options available for carriage of stereo programming, including the choice of not transmitting stereo in any form. One important factor was the impact that stereo would have in the areas of increased subscriber satisfaction and retention.

BTSC, introduced in 1984, was determined to be the most practical choice due to:

1. the capability to deliver an acceptable stereo product, and
2. maximum convenience for the subscriber.

Remaining to be answered was whether BTSC was viable in an environment that includes scrambled channels. If we were to begin transmitting stereo, we wanted to go beyond the usual practice of limiting stereo to music services, and include all premium channels; the premiums being scrambled using a gated sync suppression system. Evaluation began at the Houston system in April, 1986.

Our efforts culminated on December 10, 1986 with the commencement of BTSC transmissions on nine (9) satellite services, and retransmission of BTSC from the three (3) local TV stations carrying part-time stereo. Marketing surveys have since indicated that not only are present subscribers more satisfied with our service, but many people have become subscribers because of the availablility of stereo.

BACKGROUND

There are presently more than 400 television stations utilizing BTSC stereo, and an increasing number of cable systems are distributing stereo in this format. Add to this the emphasis of the consumer electronics market on stereo television sets, stereo VCR's, and separate TV stereo decoders and it is easy to see the direction cable system operators will be forced at least to consider if not fully implement.

Our de_ re is to qualify the use of BTSC in an environment that includes a majority of scrambled channels on which stereo service must be provided. At the time our research began, no definitive information as to the advisability of utilizing BTSC on scrambled channels was available. As is usually the case, there were many 'theories' about this application, but few 'realities' seemed to exist in this area.

METHODS OF DELIVERING STEREO

Before deciding to use the BTSC system, the alternatives were considered. Those determined to be viable were:

1) FM broadcast band

Using the 88-108Mhz standard FM band, stereo signals are generated at the head-end and transmitted through the cable system to a subscriber's FM tuner. At that point, the subscriber tunes to a frequency carrying the stereo audio which corresponds to a channel that has been tuned on the television set. This is one of the simplest ways to transmit stereo to subscribers since most cable systems do not use the FM band for anything but delivery of FM radio signals; however, in Houston these frequencies are utilized for distribution of video channels. While this method will deliver very good stereo, it is inconvenient for the subscriber and virtually impossible to secure.

2) Out-of-band FM

The stereo signal is transmitted on frequencies that cannot be tuned by an FM tuner; i.e., a channel that is otherwise restricted for the area, or a channel assignment for which there are no other plans. A converter is installed at the subscriber's home to change the frequency back to the 88-108Mhz FM band to enable tuning the stereo as described above. This provides some additional security, but the converter is easy to build or purchase as a black market item.

3) Out-of-band FM with 'MTS' adaptor

Similar to (2), but additional head-end equipment is required to encode the normal TV channel with an identifier which will cause the MTS adaptor to tune automatically to the out-of-band signal and provide the subscriber the left and right audio signals to be fed to a stereo system. The adaptor also has volume control. Black market converters as described in (2) will also work with this system.

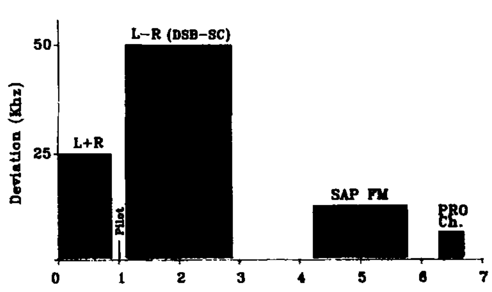
4) BTSC Stereo

This most recent development transmits the stereo information as subcarriers on the normal TV sound carrier, and has been implemented by many broadcast television stations throughout the nation. In addition, it is being utilized by a number of cable television systems on basic services such as MTV. The convenience of transmitting stereo information on the associated channel, rather than occupying additional bandwidth, along with the increasing number of stereo-capable television sets makes BTSC an attractive method for use in cable TV. The system requires a BTSC generator at the head-end location and a stereo TV or adaptor at the subscriber terminal.

BTSC TESTING

Before implementation, a variety of tests were performed to insure compatability with our addressable converter system (Pioneer BA5000). At the time we began our evaluation, research into the effects of utilizing BTSC in an encoded-channel environment was not available.

Below is a chart depicting the spectral components of the BTSC system. Let's assume we are familiar with how the system operates and continue into the actual test results.



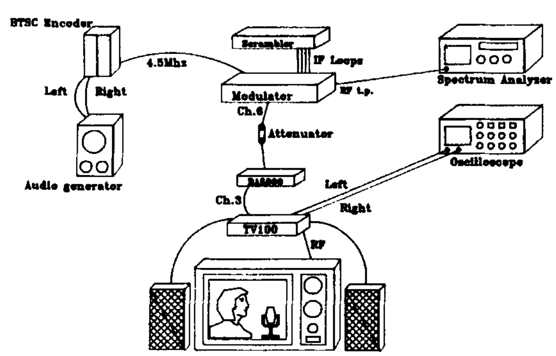
The BTSC system utilizes a pilot signal at a frequency equal to the horizontal sync rate (15,734Hz) as does the Pioneer scrambling system. Theoretically, there should be no interaction between the BTSC pilot signal which is frequency-modulated onto the main audio carrier of the television signal, and the scrambling system 'key' signal which is amplitude-modulated onto the same carrier. Imperfections in the FM detector (TV) and the AM detector (descrambler), however, can create an AM component in the FM detector and an FM component in the AM detector, thus creating concern that interaction between the two could interfere with proper recovery of the scrambled video and with separation of the left and right component of the stereo signal.

Other areas of concern are that the FM component produced in the descrambler could be of sufficient amplitude to activate the stereo circuitry of a BTSC adaptor or television set and open the sound system to noise rather than the expected stereo subcarrier, and that the AM component of the stereo detection process could create a problem with sync buzz in the stereo signal.

Evaluation and measurement of the aforementioned effects of BTSC vs encoded (scrambled) channels was vital to the process of determining whether or not to recommend its use. The Pioneer BA5000 home terminal is being used in the Houston system to replace basic converters in order to gain efficiencies in the areas of upgrades, downgrades, non-pays, etc. Scrambling of premium and certain other services is utilized to reduce the problem with illegal connections. Coupled with the operational factors of using the BA5000 system is the increasing desire of subscribers for more programming in stereo.

For testing, following equipment was utilized:
Wegener 1791 BTSC Generator
Scientific/Atlanta 6350 Modulator
Hewlett-Packard Tone Generator
Hewlett-Packard 8569B Spectrum Analyzer
Tektronix 465 Oscilloscope
Radio Shack TV-100 Stereo Adaptor

INTERCONNECTION DIAGRAM
(Evaluation)



Equipment set-up and calibration was performed with a 1 Khz tone feeding the left and right channels of the BTSC generator. This enabled us to make an accurate determination of the input and output levels at each point in the system without regard for variations in program content. All equipment was calibrated to assure proper operation at all points.

Initial measurements indicated separation between left and right channels was in the 10 to 12 db range. Previous subscriber evaluation by other system operators demonstrated that individuals would consider the stereo to be good with as little as 8db separation. We felt the 2-4 db differential to be too small a margin for error, so further investigation was made. Conversations with others who had researched the carriage of BTSC on clear channels revealed that a modification to the audio module of the S/A 6350 modulator might yield an increase in the left-right separation. When this modification was made to our test module, a repeat of our measurements indicated a 50% increase in separation to an average of 18 db.

Test #1 was made with the Pioneer encoder in the standby (unscrambled) mode. For tests #2 through #5 the encoding was activated in various combinations of operating modes; i.e., zero, six, and ten db suppression of sync, in combination with modes 1-4 of the Pioneer dynamic phase shifting.

The results were as follows:

Test	Mode	Separation
1	Standby	20 db
2	Pioneer 1	
	Clear	17.8 db
	6 db	17.8 db
	10 db	17.8 db
3	Pioneer 2	
	Clear	18.3 db
	6 db	17.6 db
	10 db	17.3 db
4	Pioneer 3	
	Clear	18.3 db
	6 db	17.6 db
	10 db	18.9 db
5	Pioneer 4	
	Clear	18.3 db
	6 db	18.1 db
	10 db	18.9 db

Termination of both inputs to the BTSC generator allowed observation of the residual components of the scrambling signal. Total distortion from these components ranged from 0.8% to 1% depending on the mode of operation.

Also of concern was the possibility of activation of the BTSC stereo detector by the key signal component on the audio carrier of scrambled channels. The key signal being a horizontal rate pulse might be interpreted as the BTSC pilot signal. This was not a problem on the receivers used in our evaluation. Since introducing stereo to subscribers, we have encountered a very small quantity of devices exhibiting this characteristic, and these also appear to trigger their stereo circuitry on clear, non-stereo channels.

LISTENING TESTS

In order to determine the end result of the entire process in a typical subscriber setting, listening tests were made in the homes of several employees and at a TV/Video equipment retail outlet. For evaluation in employee homes, the Radio Shack BTSC adaptor was used as both a stand-alone item and a supplier of left and right audio signals to larger stereo systems; different addressable converter/descramblers were used in each home to add to the variety of conditions for evaluation.

Perfection was not expected in the final product delivered to the subscriber considering the various forms of degradation of the original signal that can occur. But in all of the situations utilized for listening tests, the end product was determined to be acceptable.

LONG-TERM TESTING

Both laboratory and cable system environments were utilized during the evaluation process. The final test of the BTSC system began June 1, 1986 with the full-time transmission of stereo on The Movie Channel. We felt it was important to have an opportunity for feedback from a larger audience before the decision to proceed could be made.

Contacts were made with operators of stereo retail outlets to encourage them to evaluate our product and provide their comments. Cable company employees also began to provide information from their experience as well as from other contacts. Most of the information received was of a very positive nature. Only rarely did we hear any negative comment about stereo.

DECISION TIME

In October, a final decision was made to proceed with implementation of BTSC on the seven premium channels, two music channels, and the local off-air channels offering a stereo product. Stereo is now available on the following services on Warner-Houston:

Off-air:

KPRC-TV	NBC
KUHT-TV	PBS
KHOU-TV	CBS

(the ABC affiliate does not offer stereo)

Satellite-delivered Services:

MTV	VH1
HBO	TMC
Showtime	
Cinemax	
Disney	
Viewer's Choice 1	
Viewer's Choice 2	

The locals, MTV, and VH1 are transmitted in the clear. All seven of the premium services are scrambled.

CONNECTING TO THE SYSTEM

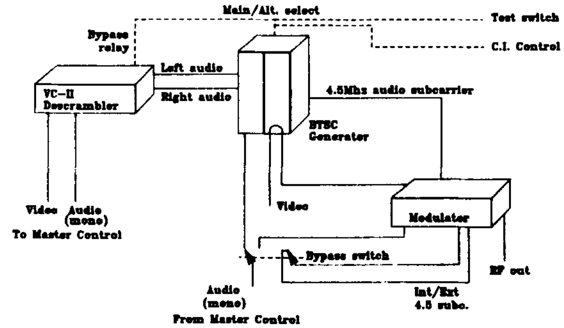
Interfacing the BTSC generation equipment to the cable system can be fairly simple in the basic configuration. In our case, we had to modify Scientific/Atlanta 6350 modulators to accept the external 4.5Mhz subcarrier from the BTSC generator. This process required approximately 20 minutes per unit with about $2 in materials.

Stereo is available directly from the VideoCipher II descrambler on the services whose satellite feeds are scrambled. Subcarrier demodulators are used to recover the stereo signals on services that are not scrambled.

Another factor in Houston was the requirement for support of commercial insertion on MTV and VH1. We also wanted the capability to switch away from the stereo source to an alternate audio source (such as a tone) for testing.

Diagrammed are the two arrangements we utilize for stereo in Houston. The first shows the connections for operation with the VideoCipher II; the next shows connections for operation with subcarrier demodulators.

INTERCONNECTION DIAGRAM
(Operation with VC-II Descrambler)

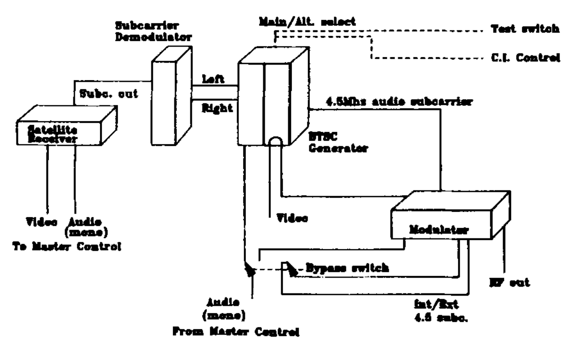


Stereo left and right audio is fed from the VideoCipher II directly to the BTSC generator. Program video and mono audio are routed through Master Control. Video is looped through the BTSC generator and terminated at the modulator.

A manual bypass switch normally supplies the mono audio (program or commercial) to the alternate audio input of the BTSC generator. It also supplies the closure to the 6350 modulator to cause it to choose the external 4.5 Mhz stereo subcarrier. For testing, or in the event of failure of the BTSC generator, the bypass switch can be toggled to the bypass position which places mono audio on the modulator and selects the internal 4.5 subcarrier.

Control lines are run from the main/alternate select connection on the BTSC generator to the VideoCipher II bypass relay, the commercial insertion controller, and a test switch. A closure from any of these sources causes the BTSC generator to change from the stereo inputs to the mono audio input.

INTERCONNECTION DIAGRAM
(Operation with subcarrier demod.)



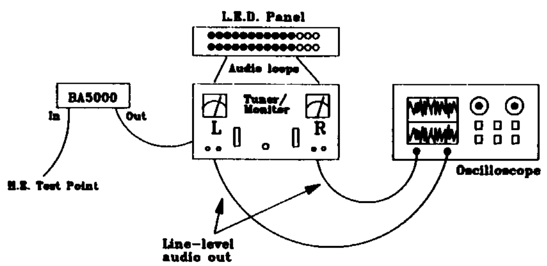
Connections when using a subcarrier demodulator are similar to the VideoCipher II connections except for the source of the left and right audio signals.

MONITORING

Maintenance of proper operation of each of the components of the stereo system is aided by the following equipment:

Marcom 730 Metered Stereo Receiver
Audio Technology 510 LED Display
Hewlett-Packard 1222A Oscilloscope

Head-end Monitoring



Output signal from a test converter is fed to the stereo receiver. High-impedance line-level audio output is routed to the LED display and to the oscilloscope. Modulation levels of both the left and right channels can be adjusted very accurately by observing the LED display. The oscilloscope is used for checks of separation and any scrambling components that may be present.

Adequate monitoring must be available. However, the stability of each component in the chain has, thus far, required little time to be spent making adjustments.

SUMMARY

Delivery of stereo sound to subscribers via the BTSC system will not be void of problems. However, with proper monitoring and reasonable maintenance, it is capable of providing an acceptable final product. When compared to the various other methods, it is certainly the most convenient for the subscriber. Initial investment in head-end equipment ($2,500 per channel) is higher than other systems, but no company-owned equipment is required at the subscriber terminal. The subscriber is required to provide his own means of recovering the stereo signal, whether it be a stereo television set or a BTSC adaptor. Long term, because of the influence of consumer electronics retailers, we believe this method of stereo delivery to be the most viable to our subscribers.

RESULTS

We must do all we can to maximize the perceived value of our product since we are in a market with competition from four VHF and four UHF television stations, along with MDS operations and videotape outlets. Entry into stereo was viewed as a step in that direction. Retention of existing subscribers was a primary goal.

Marketing surveys indicate a greater impact than we expected. Not only are present subscribers more satisfied with our product, the availability of stereo has had a positive effect on sales to new subscribers. In the sampling of subscribers interviewed, 28% are TV stereo capable. Of those equipped for TV stereo, 23% said stereo availability was a major factor in their decision to become cable television subscribers.

CABLE T.V. SATELLITE DISTRIBUTION C OR KU BAND

Andrew G. Setos
Paul F. Beeman

VIACOM NETWORKS GROUP

ABSTRACT

Satellite distribution of cable television programming has been a key element in the overall success of the industry. Conversely our continued success is reliant upon uninterupted availability of adequate satellite capacity. Commercial domestic communications satellites generally have a 10 year lifetime. We are therefore given the opportunity from time to time to reflect on our experiences and make decisions about replacement satellite capacity. The cable industry is now at such a point in time. Decisions made now will be with us into the 21st century.

TODAY

Within 12 years of its beginnings satellite distribution of cable television programming is using third generation C-band satellites and a ground segment estimated at over 15,000 TVRO sites serving multiple customers and over 1,500,000 TVRO sites used by individual customers. It would be an understatement to characterize this system as anything but highly successful. Commercial quality TVRO's have shrunk from 10 meters in size to 3 meters due to more powerful C-band transponders and better understandings about adjacent satellite interference. Locations of TVRO's in urban centers, once thought impossible, are now routine because of a better understanding of terrestrial microwave interference. The installed based of commercial TVRO's receiving cable programming has a replacement value well over $250 million, more than the cost of a modest satellite system in orbit.

TOMORROW

Currently two operational satellite types are in orbit for use by domestic communiations users. One, first launched in 1975 operates in the C-band and the other first launched in 1980 operates in the Ku band. Each of these satellites uses similar technologies and are made up of receivers and transponders which together with antennae, power supplies, station keeping systems, and other equipment retransmit signals received from uplinks to the area of coverage, or footprint.

The difference between them is not so much the equipment but the band of frequencies used. This difference compells us to make a choice between C-band or Ku-band for distribution of programming to the cable industry. To make that choice a careful comparison of the characteristics of each frequency band and the constraints placed on both space and ground segment is necessary. In addition, as in any high stakes decision such as the building and launching of future satellite systems, issues not purely based in technology must be evaluated, such as business climate and launch availability.

THE LAUNCH CRISIS

Today there are no operational commercial launching systems available in the Western world. Regularly scheduled commercial launch availabilities are not expected to be back with us until the early 1990's. The insurance industry has also been severely affected by the recent spate of launch failures. Until regular launches begin to create a steady stream of premiums the insurance underwriters cannot generate the revenue necessary to spread risk. Further compounding the insurance crisis is an unwillingness on the part of any U.S. governmental agency to hold harmless, as NASA had, commercial launch operations, from liability of all sorts, including damage to governmental facilities and civilian property. All these factors make it impossible to forecast the cost and date of replacement satellite capacity.

It is therefore crystal clear that capacity must be secured which is currently in orbit and can outlast the launch crisis with enough margin to safely ensure uninterupted distribution.

It is also necessary to arrange for backup satellite capacity already in orbit that can stand ready to replace the prime satellite if a single point failure occurs, such as loss of stationkeeping or power supply. Doing otherwise is not prudent.

There is today not enough medium power Ku-band capacity aloft to supply the total cable industry with reliable distribution; as measured by transponders or spacecraft. Because of the launch/insurance crisis more capacity is unavailable until the 1990 time frame.

By arranging for the use of Hughes Communications Galaxy III satellite, with backup provided by its sister Galaxy II Viacom will have uninterupted service on a constellation whose end of life is currently predicted as late 1994/early 1995.

Viacom's decision bridges the launch crisis, but we are still faced with making a decision for the next constellation that will serve us through the year 2005.

THE BUSINESS CLIMATE

Building and launching satellite constellations is a costly and risky business. The first round of commercial satellite launches in this Country were totally speculative. No orders were written or deposits taken for over a billion dollars worth of spacecraft launched in the first epoch of that new industry. Each endeavor backed by such companies as RCA Americom, Western Union, and SBS hoped that space traffic would develop around the traditional terrestrial traffic models of message, data, and video. But as in many speculative ventures changing conditions, overbuilding in a highly competitive atmosphere, and competing technologies such as fiber optic cable, created a huge oversupply. Today approximately half of all U.S. commercial transponders aloft generate revenue insufficient to return satisfactorily on investment. One would expect that satellite operators would be wary on the next round of launches. However, the list of FCC grants for new satellites reveals that speculation continues although a subtle shift has occurred. 8 equivalent C-band satellites have been granted construction and launch permits and 14 equivalent Ku-band spacecraft have been granted.

The ratio of C-band and Ku-band is a sign at once of new realism and continued speculation in the market place. C-band has become a well understood business, suited primarily to Broadcast and Cable Network video program distribution. As such the market size for C-band can be more accurately guaged. Ku-band on the other hand is the only potential growth satellite industry and, just as C-band was in 1975, shows evidence of speculative activity by its high number of launch commitments. The C-band licenses for future launches in the early 1990's are held by such companies as Hughes Communications, AT&T, and RCA Americom. In addition, recently John Koehler, President of Hughes Communications has stated his desire to apply for additional construction permits to replace the current Galaxy constellation.

The established carriers are sending a clear signal that follow on C-band capacity will be available in the 21st century if firm orders materialize. Indeed, it is becoming clear even among Ku-band satellite operators that follow on capacity will on the whole only be available if there are firm orders Therefore, no matter what type of capacity is needed by the industry, that type will be launched.

THE EARTH SEGMENT

The total investment in commercial TVRO's receiving Viacom's cable program services is at least $250 million. This equipment will be serviceable into the 21st century. Virtually none of that equipment is transferrable to Ku-band use. The main reflectors employed in most dish antennae are not smooth enough to achieve adequate gain. The Low Noise Amplifiers, or Block converters and feedhorns are totally unusable. Of the total video receiver universe installed over the last ten years only a small minority are able to switch to Ku-band. Making obsolete such a large installed and operating investment can be justified only with the most compelling arguments.

THE DIFFERENCES

There are differences between the C and Ku-bands. Some arise from physical laws, others from laws passed by governments.

Rain Antenuation is the most well known difference between Ku and C-band. Indeed this is the primary reason Ku-band satellites followed C-band by so many years. In order to provide adequate power levels to be received through rain by reasonably sized antennae the development of large solar power arrays and high power transponders had to be awaited. Early Ku-band transponders included 20 watt traveling wave tubes while the first C-band transponders were 3.5 watts. Current generation Ku-band transponders have 45 watt traveling wave tubes. This represents a 3.5 db improvement in EIRP for footprints of similar size. An additional 1.3 db improvement has been requested by RCA Americom of the FCC for its future spacecraft K-3. For adjacent satellite interference reasons the Commission has yet to approve that request and so we cannot count that improvement. To put these improvements in power level in perspective it is interesting to note that good practice for most of the United States is that rain fade margin should be at least 10db. Unfortunately the rain fade during heavy thunderstorms can be 15 to 20 db. As the cable industry's experience with CARS band reminds us there is no economical way to protect a link under such severe conditions.

Antenna Size has absolutely nothing to do with which band is in use, given the same transponder power and footprint. For instance an excellent signal can be received from Galaxy III on C-band with a 3 meter antenna. Galaxy III employs 9 watt transponders. If a Ku band transponder's signal of 9 watts were received by a 3 meter antenna an equally good signal would be received. Unfortunately in that case the least amount of rain or snow would reduce the quality of the signal, eventually obliterating it entirely. It is for this reason that Ku-band transponders must be higher powered, to overcome rain attentuation. Using this comparison RCA Ku-band satellites such as K-1 achieve a rain fade margin of approximately 9db with like antenna size to Galaxy III type C-band satellites. The suggestion here is that for truly reliable service which the cable industry today enjoys on C-band larger antennae will be required for Ku-band.

Spacecraft Reliability is always a concern when unexpected outages deprive customers of programming. Because of the aforementioned need to overcome rain attentuation Ku-band spacecraft must rely on high power vacuum tubes and their companion high voltage power supplies. The latest generation of C-band spacecraft utilize solid state power amplifiers in their transponders. Transitors operate at much lower voltages, putting less strain on components in their power supplies. It should be noted that high voltage power supplies have accounted for more transponder failures than any other single cause of outage. Higher signal powers also puts a strain on other satellite components, including waveguide networks and cooling systems. As a result of these factors calculated reliability for current Ku-band spacecraft is materially inferior to solid state C-band spacecraft.

Cost of spacecraft, including launch are much higher for Ku than C-band. Regardless of frequency band the more powerful a transponder the more weight which will be required. Several systems' weights are affected, including the transponder itself, cooling equipment, eclipse batteries, solar cells, power supplies and waveguide components. Weight in the launch business means expense. Indeed some launch vehicles restrict the total number of transponders per launch because of their total weight. As a result the cost per transponder of constructing and launching a number of Ku-band spacecraft to form a constellation of similar capacity and backup to Galaxy's is much higher - up to three times as high.

Cost of an uplink for Ku-band is more than double that of C-band. One thunderstorm over the uplink can disrupt service to the entire Country. It is not practical to build enough margin into the uplink to compensate for such rain fades. Therefore two widely separated uplinks need to be built. In addition a very high reliability microwave or cable link must be established between the two locations.

Terrestrial Interference on Ku-band does not exist in any material amount. The band has been held exclusively for the use of satellites. This is the only enhancement Ku-band delivers to satellite communications beyond C-band. C-band satellites share the band with terrestrial microwave links carrying primarily message traffic. As a result and as long as there is line of sight to the satellite a Ku-band link can be established to the customer's premises. Two applications specifically made possible because of this feature are Direct Broadcast service to homeowners and Private Business networks to coorperate headquarters in downtown urban areas.

In the early 1970's terrestrial microwave interference was expected to limit C-band TVRO's to extremely rural environments. Over the last decade the art of coordinating TVRO sites in the C-band has become much more sophisticated. Engineering models now take into account terrain and man made shielding while interference models have become much more realistic. Filter techniques and hardware have also added signficntly to the body of knowledge and tools which has resulted in thousands of TVRO's throughout the United States in urban and suburban areas. Indeed in most cases the limiting factor for coordination is line of site, not terrestrial interference.

An interesting possiblity for the future of C-band is that greater reliance on fiber optic cable for terrestrial traffic will cause coordination to be easier as fewer and fewer microwave links are in use.

Adjacent Satellite Interference is an effect which will have increasing importance on Ku-band and less on C-band. Because of the increased speculative launches on Ku-band it is more likely to experience such interference than on C-band, where fewer satellites will be operational, and most likely farther apart.

CONCLUSION

Viacom has taken several factors into consideration in choosing what type of satellite shall distribute its program services to the cable television industry. Above all our decision was driven by the most reliable, cost-effective technique, now and for the foreseeable future. We have found that Ku-band has nothing in its favor except the ability to reach directly into the customer's premises. Its inferior reliability, greater cost, and uncertain availability in the near term stand in stark contrast to today's functioning C-band system.

COMPATIBLE HIGH DEFINITION CABLE TRANSMISSION TECHNIQUE

William E. Glenn, Ph.D.

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High definition television using 1125-line scans is now under advanced development in Japan. The image quality is comparable to 35mm motion picture film. Within the next few years HDTV receivers with video disc or cassette players will undoubtedly be available on the market.

The developers of HDTV recognized the impracticality of the full bandwidth (30 megahertz) distribution. Therefore, they have developed a bandwidth reduced transmission format called "MUSE" that requires 8.3 megahertz bandwidth in a continuous channel. This signal cannot be received by a standard NTSC receiver.

This laboratory has under development a transmission technique that is compatible with NTSC transmission. The design of this system has depended heavily on studies of the visual system in order to determine what information can be removed from transmission without degrading the displayed image.

In the system two channels are used. One is a standard 525-line NTSC channel. The other channel contains the detail information in both luminance and chrominance that is necessary to upgrade this "base" channel to 1125-line resolution. This detail channel has about 3 megahertz bandwidth which does not have to be adjacent in frequency to the "base" channel. A standard 525-line receiver tuned to the "base" channel will reconstruct a normal 525-line image without any modification to the receiver.

An HDTV receiver can combine the base and detail signals to reconstruct an 1125-line image. Even though the base channel has 3x4 aspect ratio, the HDTV reconstructed image has 3x5 aspect ratio.

A bandwidth reduced two-channel system similar to this has been demonstrated at the last two NAB conventions in "closed circuit" form. It has achieved over 800 line limiting resolution from a test chart, both vertically and horizontally in the reconstructed image. Motion rendition with the system is excellent.

For the cable operator this system has several significant advantages:

1. Only one additional half channel is required to upgrade an existing NTSC transmission to HDTV.
2. Since the detail channel need not be adjacent in frequency, the present cable spectrum does not have to be reallocated to upgrade a channel to HDTV.
3. Cable has a characteristic 6 MHz "picket fence" interference spectrum. This would be a serious problem for an 8.3 MHz continuous signal unless the HDTV channels are all located at the bottom of the spectrum at 12 MHz intervals. It is not a problem in a bi-channel system of the above design. Both channels can locate their carriers so that the 6 MHz signature causes no visible interference.

COMPUTER ASSISTED DESIGN OF CATV ANTENNA-TOWER/ANTENNA-ARRAYS

STEVEN I. BIRO

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ABSTRACT

Most CATV engineers are well aware of the difficulties in producing functional, interference repellent, and cost-effective antenna-tower/antenna-array designs, and in completing the project on time. The Computer Assisted Design (CAD) program, developed by Biro Engineering, optimizes array configurations, their dimensions, and proper location on the tower.

The paper will discuss the major benefits of the program, such as improved co-channel, adjacent channel and second harmonic FM interference rejection, as well as the advantages of a computer drafted and printed tower/array design, as applied to a 400' guyed CATV antenna tower project.

Most CATV engineers are well aware of the difficulties in producing functional, interference repellent, and cost-effective antenna-tower/antenna-array designs, and in completing the project on time.

For those who consider the array design a simple task, just take into account the different VHF-UHF antenna models on the market, their gain, beamwidth, Front/Back ratio specifications, the different shape and physical size of the radiators, mounting features, the length of the antenna boom or the diameter of the parabolic dish, and last but not least, the antenna manufacturer. Then, focusing on the tower design, one must choose between selfsupporting and guyed antenna towers, deal with tower height and/or guy-wire layout restrictions, reserve space for microwave dishes or two-way radio radiators, the mounting of the arrays on tower-legs, crossarms or V-Gate structures, vertical and horizontal separation between arrays,--just to mention a few variations.

In effect, the design may involve a vast number of possibilities, too many to be chosen by intuition or experience, and too many to be analyzed in an ordinary way.

Off-air TV receiving antennas have to fulfill two distinctive functions. First, to provide maximum gain in the direction of the desired signal. Second, to repell signals arriving from the undesired (interference) directions.

CATV receiving antennas cannot operate in open space. They are mounted on gates, crossarms, or on the tower legs of massive steel towers. Or, by choice, the antenas are co-located on the tower. Nearby reflecting surfaces may destroy the radition pattern of the arrays, and any such interference should be avoided or reduced to a tolerable level. That formulates a third objective: to generate a "clean" tower and antenna-array design. No matter whether the engineer follows the old fashioned "manual design", or applies the recently developed Computer Assisted Design program, he needs design specifications, such as:

1. Listing of the desired VHF-UHF stations to be carried on cable.
2. Projected or actual signal levels of the desired stations.
3. Predicted or measured interference levels, such as:
   a. Co-channel interference
   b. Adjacent channel interference
   c. Second harmonics of local FM stations
   d. AC interference sources
4. The direction of reflections (ghosting)
5. Type of tower.
6. Tower height limitations.
7. Microwave dish allocations
8. Preference of antenna-type and/or antenna manufacturer.

To comply with the signal level and interference protection requirements of the design specifications, the design must be preceeded by an on-site Signal Survey or by a Computerized TV Reception Study. Both require the exact coordinates of the proposed antenna site. They can be obtained either from 7 1/2' U.S. Geological Survey Maps (Topo-Maps), or with the aid of a LORAN receiver during the on-site Signal Survey.

The printout of a Computerized TV Reception Study (Fig.1) lists all major technical and program parameters of all desired and most of the undesired TV stations, including computer calculated great circle distances and directions. The computer projected signal levels should provide a reasonable basis for antenna-size and antenna height calculations, as well as for co-channel and adjacent channel considerations.

*** COMPUTER AIDED TV RECEPTION STUDY ***

BIRO ENGINEERING
PRINCETON,NEW JERSEY

LOCATION = TUCSON, AZ

COORDINATES = 32/12/52 110/56/57

THIS IS A LISTING OF ALL VHF AND UHF STATIONS WITHIN 250 MILES

DIST	CHAN	CALL	LOCATION	STATE	NETWK	POWER	OFST	HAAT	AZIMUTH	LEVEL
10.06 MI	40	KPOL	TUCSON	ARIZ	IND	1534.	0	2029.	283.9	18.8
19.41 MI	9	KGUN	TUCSON	ARIZ	ABC	110.	-	3720.	44.3	9.5
19.48 MI	6	KUAT	TUCSON	ARIZ	ED	36.	+	3630.	44.5	9.7
19.56 MI	13	KOLD	TUCSON	ARIZ	CBS	110.	-	3610.	44.6	8.8
19.56 MI	4	KVOA	TUCSON	ARIZ	NBC	35.	-	3680.	44.6	10.8
35.23 MI	11	KMSB	NOGALES	ARIZ	IND	150.	0	1570.	177.6	5.4
48.75 MI	18	KDTU	TUCSON	ARIZ	REL	2490.	-	1966.	86.8	8.7
61.46 MI	2	XHFA	NOGALES	ARIZ	IND	8.		231.	179.9	-20.3
100.67 MI	45	KUTP	PHOENIX	ARIZ	IND	3020.	0	1792.	320.6	-35.6
100.77 MI	5	KPHO	PHOENIX	ARIZ	INO	100.	-	1770.	320.6	-23.7
100.79 MI	21	KPAZ	PHOENIX	ARIZ	INO	251.	0	1600.	320.6	-45.6
100.79 MI	10	KTSP	PHOENIX	ARIZ	CBS	316.	-	1700.	320.6	-28.8
100.80 MI	12	KPNX	PHOENIX-MESA	ARIZ	NBC	316.	-	1780.	320.5	-28.6
100.81 MI	3	KTVK	PHOENIX	ARIZ	ABC	100.	+	1670.	320.6	-22.4
101.39 MI	33	KTVW	PHOENIX	ARIZ	SP	2290.	0	1710.	320.8	-37.1
101.39 MI	15	KNXV	PHOENIX	ARIZ	INO	524.	-	1710.	320.8	-40.0
101.42 MI	8	KAET	PHOENIX	ARIZ	ED	316.	+	1756.	320.8	-28.0
184.72 MI	7	KUSK	PRESCOTT	ARIZ	IND	9.	0	2810.	339.1	-103.8
192.97 MI	2	KNAZ	FLAGSTAFF	ARIZ	NBC	100.	0	1540.	350.6	-76.8
232.14 MI	22	KRWG	LAS CRUCES	N.M.	ED	1450.	-	450.	87.6	-269.0
232.88 MI	13	KYEL	YUMA	ARIZ	NBC	316.	+	1560.	286.1	-162.7
232.91 MI	9	KECY	EL CENTRO	CAL	CBS	316.	+	1601.	286.1	-154.6
243.67 MI	48	KASK	LASCRUCES	N.M.	INO	79.	+	25.	86.8	-395.0

OUR SIGNAL LEVEL CALCULATIONS ARE BASED ON SINGLE YAGI ANTENNAS AT 30 FEET ABOVE GROUND EXHIBITING 6 DB GAIN ON VHF AND 9.5 DB GAIN ON UHF.

FIG. 1

But the computer cannot take into consideration:

a. Terrain features, affecting desired and undesired signal levels

b. Reflections, causing ghosting

c. The intensity and direction of any AC or RF interference.

Therefore, the efficiency and reliability of the tower/antenna-array design can be significantly improved if it is followed by an on-site ground or helicopter survey, confirming signal level projections, as well as the direction and level of interference transmissions.

THE SYNTHESIS PROCESS

There are as many approaches to meet a set of specifications as there are design engineers. Given plenty of time, paper, patience, and a scientific calculator, an expert could produce all the unique solutions that meet the design specifications. However, there must be a better way. A computer can perform repetetive operations with ease. While not the ultimate solution, computerized synthesis is still the best tool to satisfy the complexities of the tower/array design.

Synthesis has been widely used in connection with filter designs. The "G" or "K" parameters, often applied in "cookbook type" filter designs, represent no more than tabulation of solutions for a number of synthesis possibilities.

The CAD program for towers and antenna-arrays, as a matter of fact, uses the same procedures in the computer aided mode that the experts use in the manual mode. The difference is that the computer constantly monitors many aspects of the design process and adapts ifself to changing specifications. Then, when a single or a set of specifications are judged inconsistant, the program may refuse to accept the input. Or, in the case of potential inaccuracy, the program flashes a warning note on the screen.

```
                    SUBROUTINE
                       FOR
           ADJACENT CHANNEL PROTECTION

               TOWER:   GUYED TOWER
           DESIRED CHANNEL: CHANNEL 12
          UNDESIRED CHANNEL: CHANNEL 13

 ANTENNA: YAGI                  NUMBER OF ELEMENTS:12

        DIRECTION OF DESIRED: 74° AZIMUTH
      DIRECTION OF UNDESIRED:138° AZIMUTH

 TWO-BAY:  YES                          FOUR-BAY: NO
              DIAMOND ARRAY: NO

                   LIMITATIONS

      HORIZONTAL CLEARANCE  :  2 METERS
      VERTICAL CLEARANCE    :  5 METERS
      GUYED TOWER           :  YES
      SELFSUPPORTING        :  NO
      GUY WIRES             :  AT 180'
      STAR-MOUNT            :  NO
```

FIG. 2

Fig. 2 shows the synthesis display for the ADJACENT CHANNEL PROTECTION subroutine. The upper display is the "SPEC.PANEL", while the bottom lines are a list of "COMMAND PARAMETERS", obtained from the preliminary specifications.

Once the minimum parameters have been selected, the designer has to enter only the "PFMPRES" command to synthesize all possible array configurations and antenna types that match the specifications.

The usefulness of the CAD program extends beyond the basic tower/array design function. There are powerful diagnostic tools built into the program, which allow the designer to investigate the effects of certain components. For example, the designer can probe Front/Back ratio conditions while switching from Yagis to Log-Periodics, or from a horizontally stacked, phased-array configuration to the vertical stagger stacking approach.

ARRAY SYNTHESIS SUBROUTINE

This subroutine of the CAD program focuses on one of the array synthesis techniques which has proven useful in determining the excitations of arrays that are to generate the desired radiation patterns.

To force a deep radiation pattern null into the direction of an interference source is a common practice in antenna engineering. Less frequently analyzed, however, is the protection capability of the phased array into a number of other directions, namely into the directions of the secondary interference sources.

The radiation pattern of any array is the product of the individual antenna pattern, multiplied by the ARRAY FACTOR.

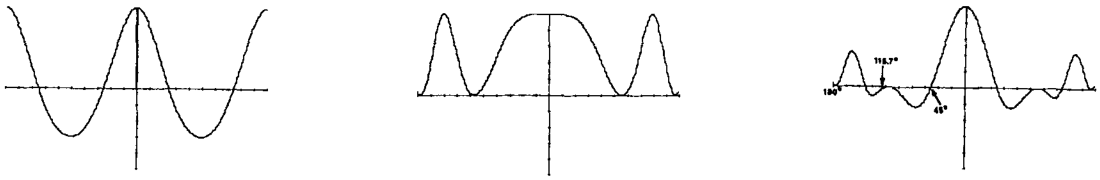
In the simple case of two, horizontally stacked antennas, fed by equal amplitude and phase currents, the array factor is:

$$\frac{\sin(S \sin\Theta)}{2\sin(S/2 \sin\Theta)}$$

where S is the electrical spacing between the antennas in radians.

The radiation characteristics of a short Yagi antenna can be described by the following equation:

$$Y = \cos\left(\pi/2 \sin 0.5\,\Theta^2\right)$$



ARRAY FACTOR times YAGI PATTERN = ARRAY PATTERN

FIG. 3

FIG. 3. shows the forming process of the radiation pattern. The phased-array was designed to create a deep null into 45° AZIMUTH, the direction of the identified co-channel offender, as well as at 115.7° AZIMUTH, the direction of the nearby FM radio station.

As expected, the array exhibits nulls at +/-45° of the main beam, as well as at +/-135°.

Note the excellent Front/Back ratio of the pattern (28.6 dB), not only exactly at 180°, but in the wide range of 171° to 189°.

At 115.7°, which is the direction of the secondary (local FM station) interference source, the two-bay array is ascertaining 23 dB protection.

The Computer Aided Design provided a fast solution and enhanced protection against a multitude of interference sources.

TOWER DESIGN SUBROUTINE

For the tower design portion of the program an iteration algorithm was developed which takes into account the following tower geometry parameters.

1. The width and shape of the tower.
2. The position (height above ground) of the guy-wires and star-mounts.
3. The height, width and vertical separation between the antenna crossarms (gates).
4. The separation between the inner tip of the antenna elements and the tower.
5. The minimum vertical separation between adjacent antenna-arrays.
6. Partial or total blockage caused by guy-wires and startmounts.
7. Minimum clearance between microwave dishes and antenna-arrays.

The program's interactive data-base manager provides a flexible means of manipulating and examining very large amounts of input data. The data-base manager also lets the designer create, delete and revise critical parameters, provided that the revisions do not alter excessively the number of radiating elements, the size of the array or the antenna-gate configuration. For example, one may change the size of the side-mounted gate from 14'x4' to 20'x4', but not to a V-GATE configuration with 10'x4' side arms.

When the computer program requires more information to continue a routine, it will prompt the operator a new table on the screen that must be edited and completed. These tables contain default warnings on tower/array configurations, in order to prevent the designer from wasting time with unacceptable architectural formations.

Once the designer is satisfied with the results, the program automatically recomputes antenna gains, array factors, radiation pattern nulls and beamwidths, as well as the critical horizontal/vertical clearances, and reopens the program options for a printout record or canceling the data, making additional changes or exiting the program.

DRAWING ERRORS, DRAWING MODIFICATIONS

The computer generates error-free drawings. It shows the tower and antenna-array layouts exactly as they were designed.

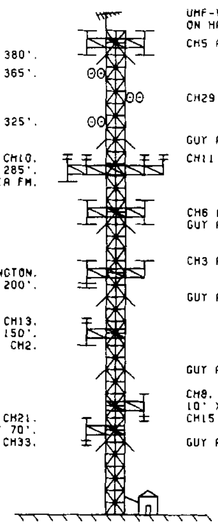
Not only is the time required for a complete printout (drawing) many-many times shorter than the customary manual drafting, but any later modifications or changes, provided these were properly fed into the program, will produce precise and instant new printouts.

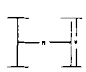
Since there is no human element introduced into the drafting process, there is little room for human mistakes. The customary checking and double-checking of the drafted product has been also eliminated, representing an additional savings in time and costs.

A DESIGN EXAMPLE

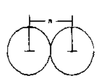
400' GUYED CATV ANTENNA TOWER

FOR HERSHEY, PA.

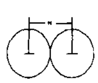




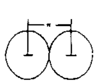
CH5. WAC. WASHINGTON. IND. - OFFSET
92.85 MI. 194.0 AZIMUTH ANGLE
H=261° V=145°
ARRAY PHASED FOR CO-CHANNEL PROTECTION
AGAINST CH5. NEW YORK. 144.68 MI.
USE 4 PCS OF LINDSAY MODEL 9AY-5 YAGIS.



CH17. WPHL. PHILADELPHIA. IND
76.73 MI. 100.8 AZIMUTH ANGLE
H=54.4°
ARRAY PHASED FOR PROTECTION
AGAINST LOCAL CH15. LANCASTER. 10.7 MI.
USE 2 PCS OF JERROLD 4' PARABOLIC DISHES.

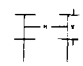


CH29. WTAF. PHILADELPHIA. IND
76.80 MI. 100.9 AZIMUTH ANGLE
H=57.25°
ARRAY PHASED FOR PROTECTION AGAINST
ADJACENT CHANNEL 28. WILKES-BARRE. 76 MI.
USE 2 PCS OF JERROLD 4' PARABOLIC DISHES.

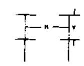


CH48. WKBS. PHILADELPHIA. IND
76.58 MI. 100.8 AZIMUTH ANGLE
H=83.3°
ARRAY PHASED FOR PROTECTION AGAINST
ADJACENT CHANNEL 49. RED LION. 22.9 MI.
USE 2 PCS OF JERROLD 6' PARABOLIC DISHES.

BALTIMORE/WASHINGTON FM ARRAY
63.92 MI. 186 AZIMUTH ANGLE
V=120°
VERTICAL SPACING CALCULATED FOR MAXIMUM GAIN.
USE 2 PCS OF LINDSAY MODEL 10AY-FM YAGIS.

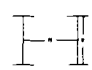


CH10. WCAU. PHILADELPHIA. CBS. 0 OFFSET
76.88 MI. 100.7 AZIMUTH ANGLE
H=83° V=60°
ARRAY PHASED FOR CO-CHANNEL PROTECTION
AGAINST ALTOONA. PA. 95.87 MI.
USE 4 PCS OF LINDSAY MODEL 12AY-10 YAGIS.

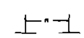


CH11. WBAL. BALTIMORE. NBC. - OFFSET
63.99 MI. 179.3 AZIMUTH ANGLE
H=91.6° V=58°
ARRAY PHASED FOR CO-CHANNEL PROTECTION
AGAINST CH11. NEW YORK 144.68 MI.
USE 4 PCS OF LINDSAY MODEL 12AY-11 YAGIS.

PHILADELPHIA FM ARRAY
77 MI. 100.8 AZIMUTH ANGLE
V=120°
USE 2 PCS OF LINDSAY MODEL 10AY-FM YAGIS.



CH6. WPVI. PHILADELPHIA. IND
76.67 MI. 100.7 AZIMUTH ANGLE
H=213° V=135°
APPLY VERTICAL STAGGER STACKING FOR PROTECTION
AGAINST JOHNSTOWN. PA. 122.27 MI.
USE 4 PCS OF LINDSAY MODEL 9AY-6 YAGIS.



CH3. KYW. PHILADELPHIA. NBC
76.69 MI. 100.8 AZIMUTH ANGLE
H=251°
ARRAY PHASED FOR PROTECTION AGAINST
CH3. CLEARFIELD. PA. 110.21 MI.
USE 2 PCS OF LINDSAY MODEL 9A6-3 YAGIS.

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2. J.A. GREENFIELD, S.E. HANSEN & R.W. DUTTON, "TWO-DIMENSIONAL ANALYSIS FOR DEVICE MODELING", STANFORD UNIVERSITY PRESS, STANFORD, CA., 1980

3. STEVEN I. BIRO, ADVANCED DISTANT SIGNAL RECEPTION TECHNIQUE, TV COMMUNICATIONS, OCT. 1979.

4. THE BIRO TECHNIAL BULLETINS, JULY, 1986.

EIA IS-15 INTERFACE COMPATIBILITY WITH RF SYNC SUPPRESSED SCRAMBLING

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ABSTRACT

The video descrambling portion of IS-15, EIA MULTIPORT, can be compatible with present day RF Sync Suppression schemes. Industry consensus seems to indicate that for this standard to survive, a "phasing-in" period will be required. During this time, compatibility with RF Sync Suppressed scrambling is imperative. This scheme, in one form or another, is presently used by a majority of CATV plants. As a result, millions of RF descramblers are currently in use. Furthermore, as bandwidth in existing systems is at a premium, duplication of service, for IS-15 compatibility, is economically prohibitive.

The ideal scenario for a Cable Operator is to modify or replace the scramblers in his headend, retain all his descramblers in the field and achieve IS-15 compatibility,without compromising security.

INTRODUCTION

The Electronic Industries Association, in July 1986 released Interim Standard (IS-15) titled:- "Standard Baseband (Audio/Video) Interface Between NTSC Television Receiving Devices and Peripheral Devices." Baseband descrambling techniques, compatible with RF Sync Suppression systems will be developed. A merging of the two technologies, and how they can both survive in an already existing cable system is discussed.

This standard emerged as the result of a need to make CATV services friendlier to future television sets, VCRs, etc. A compatible Cable System will be able to deliver Broadband RF service (with scrambled premium channels) to a compatible receiver. The receiver will demodulate the signal and loop it through a baseband video descrambler. Clear signals will pass through unaltered. Scrambled and authorized channels will be descrambled. Channels not authorized will not be descrambled.

Advantages of this standard to a TV set (or VCR) capable of tuning the entire CATV spectrum would be the full use of it's advanced features (remote control, picture-on-picture, VCR switching, programmability, etc). The Cable Operator will benefit from lower cost in hardware and less investment in the subscriber's home.

A disadvantage is that, given the long lifespan of an average television, it could take several years for a significant percentage of compatible sets to reach the subscriber. Also, there is a considerable amount of investment in RF descramblers currently in the field.

One feasible approach to IS-15 would be to make it "backwards" compatible. That is, to make it work in an RF Sync Suppressed system, without major and costly modifications.

This report will start by a brief overview of one RF Scrambling/Descrambling system currently in use. An IS-15 Baseband descrambler is presented, with pertinent addressing required. Modifications needed at the headend will be considered.

RF SYNC SUPPRESSED SCRAMBLING/DESCRAMBLING AN OVERVIEW

RF Sync Suppressed Scrambling is a standard method to secure video transmission in the CATV environment, and has been successfully used for many years.

The system includes a video scrambler in the headend operating in conjunction with a modulator. A converter (or Set-Top Terminal) tunes the scrambled channel to Channels 2, 3 or 4. An appropriate descrambler is placed inline to decode the signal. The clear signal is presented to the television set.

A method traditionally used to scramble a video signal is to suppress approximately 12 uSec of the horizontal blanking interval including the horizontal sync and the color burst. (See Figures 1 & 2). Suppression is achieved by switching in a 6dB attenuator during this time. As the signal is Amplitude Modulated, the carrier is "envelope suppressed." As shown in Figure 2, the horizontal sync has been suppressed, and the TV will attempt to find sync in the black level. Also note the "pivot" point of this modulated signal--the zero carrier level.

To descramble this signal at the Set-Top Terminal, synchronization information is needed. This information is carried on the Sound Carrier (fpix + 4.5MHz). As the Sound Carrier is Frequency Modulated to carry the audio, synchronization information may be Amplitude Modulated on it. In addition to this, many addressable systems carry data in or near the FM band. This data is in the Frequency Shift Keyed (FSK) format. Addressable data is decoded in a stand alone receiver.

BASEBAND DESCRAMBLING AN RF SYNC SUPPRESSED SIGNAL

In the IS-15 Multiport environment, a multichannel broadband RF signal is received by the TV. This signal is tuned to a particular channel and demodulated. Demodulated video and audio are separated, and the video is looped through the baseband descrambler. Figures 3 & 4 show

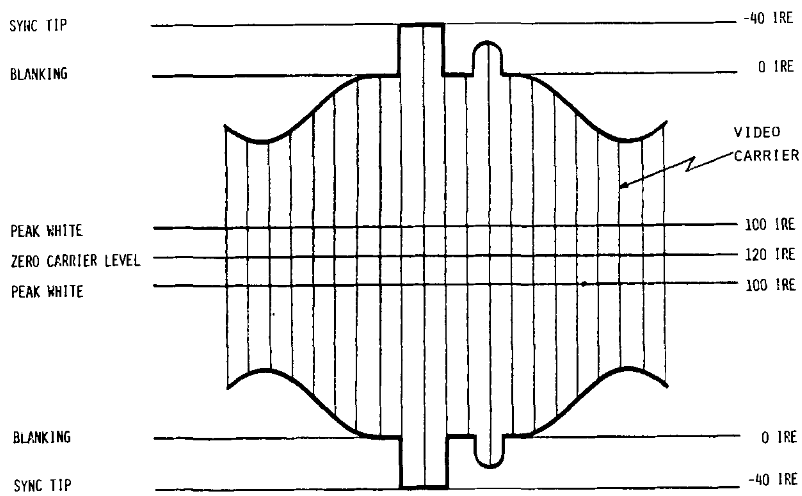

Fig. 1 MODULATED NON-SCRAMBLED HORIZONTAL BLANKING INTERVAL

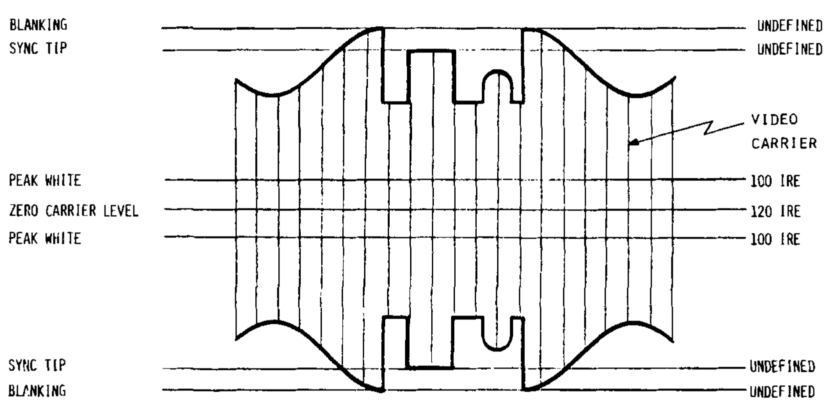

Fig. 2 MODULATED SCRAMBLED HORIZONTAL BLANKING INTERVAL

a non-scrambled and scrambled video signal as it appears to the descrambler. The non-scrambled signal is passed unchanged. In the case of a scrambled and authorized signal, the descrambler will restore the sync.

The 6dB Attenuator

Sync restoration at baseband presents a different set of problems than does restoration at RF. In an RF descrambler, sync would be restored by switching in a 6dB attenuator during active video. This, in effect, causes the horizontal blanking interval to have 6dB additional gain. Note that this is done at RF, and the additional gain is with respect to zero carrier.

A switched gain amplifier will not necessarily amplify with respect to the zero carrier level, which corresponds to +120IRE. Depending on the video content, an attenuator on this signal may or may not restore sync. Even if the sync is restored, video levels would be incorrect. (Figure 5).

To clarify, Figures 1 & 2 show a modulated video signal. In baseband, only the portion below the zero carrier level (120IRE) is available. Therefore, for proper video restoration using the 6dB attenuator method, the zero carrier level would have to be at 0.0VDC. This can be used as a fixed reference level.

The "Zero Carrier Level"

As has been noted above, a fixed reference level is available, which remains stationary during scrambled and non-scrambled video. This is the "Zero Carrier Level." In IRE scale this level occurs at 120IRE in baseband video. (See Figures 1 through 4).

IS-15 requires the Receiver baseband output in non-scrambled mode to be 2.00V +/- 0.1V at 100IRE (Paragraph 4.19.11 EIA IS-15, July 1986). Also, sync tip to peak white will be 1.00V +/- 0.16V (140IRE), (Paragraph 4.19.8). (See Figure 3).

As, Sync tip to Blanking = 0.29V -> 40IRE

And, Blanking to Peak White = 0.71V -> 100IRE

0.29V + 0.71V = 1.00V -> 140IRE

Therefore, 1IRE = 1/140 = 0.007V

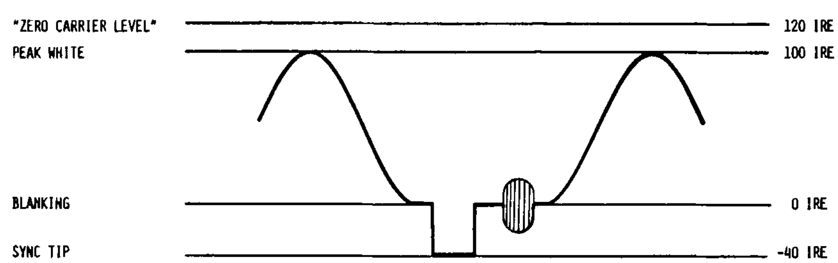

Fig. 3 BASEBAND NON-SCRAMBLED HORIZONTAL BLANKING INTERVAL

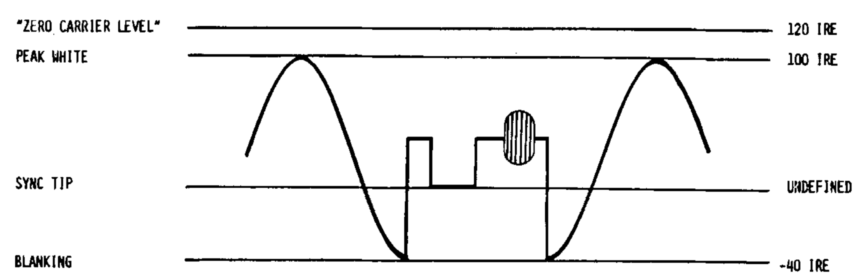

Fig. 4 BASEBAND SCRAMBLED HORIZONTAL BLANKING INTERVAL

And, 20IRE = .007V x 20 = 0.143V

In IS-15 Peak White at 100IRE = 2.00V

And, Zero Carrier at 120IRE = 2.00V + 0.143V = 2.143V

The incoming video signal "zero carrier level" must be clamped at 0.0VDC, using a precision reference. This may be a fixed reference in the vertical interval or an internal AGC loop.

Descrambling at Baseband

Once the video signal is clamped such that the 2.143V, 120IRE zero carrier level is set to 0.0VDC, the attenuation method can be applied. A 6dB attenuator is switched in during active video. This results in the horizontal blanking interval to be restored by 6dB. (Figure 6). Descrambled video is amplified, and the DC level is then restored using the same reference.

This signal is also supplied to the TV AGC circuit in the form of Decoder Restored Sync.

Timing Information

In traditional RF scrambling, timing pulses are transmitted on the audio sub-carrier. This is not available at baseband. Other "reliable" pulses , the sync tip and the color burst are suppressed. The color burst amplitude will also depend on the television manufacturer, and cannot be relied upon. The vertical interval is also suppressed, and is not a good reference point.

A reliable source had to be generated. This is in the form of a 16 bit Manchester encoded dynamic word, placed on lines 11 and 12 of each field. The information sent by the scrambler is decoded in the descrambler. Successful decoding starts a clock which generates the timing for horizontal and vertical restoration. For additional security this word may be spread through any of the seven or eight lines in the vertical interval not currently assigned. Channel or program identification is included to allow the decoder to know what channel has been tuned.

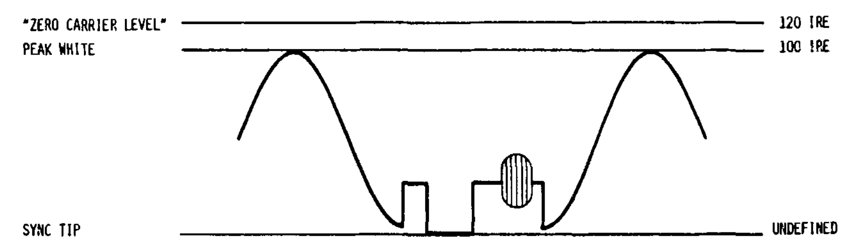

Fig. 5 SYNC RESTORATION AT BASEBAND WITHOUT ZERO CARRIER CLAMP

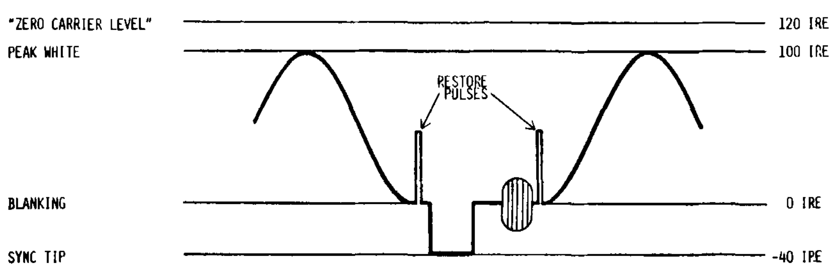

Fig. 6 BASEBAND DESCRAMBLED HORIZONTAL BLANKING INTERVAL

ADDRESSABILITY

Addressability can be achieved by sending information on a separate carrier as in the RF System. A directional coupler placed in the broadband RF line would provide signal to an address receiver. The address receiver will decode the information and provide the necessary inputs to the baseband decoder. (See Figure 7)

Information needed for addressing a baseband descrambler should be identical to the data used in RF Sync Suppressed schemes.

AUDIO

IS-15 Multiport provides a wideband audio output. It also gives the descrambler a choice between looping the audio through or processing it. This gives the Cable Operator an opportunity to scramble/descramble the sound.

As wideband audio is provided, a future option to provide high quality stereo is available.

MODIFICATIONS NEEDED AT THE HEADEND

The Scrambler

The Scrambler for an RF Sync Suppressed system generates the timing pulses for descrambling, and amplitude modulates them on the audio carrier. This is usually done in the IF section.

Baseband video is also looped through the scrambler, to provide sync information. This signal may be used to insert information needed to descramble in IS-15 format. Data required would be the dynamic word inserted in the vertical interval. Information would include channel tags, descramble authorization and

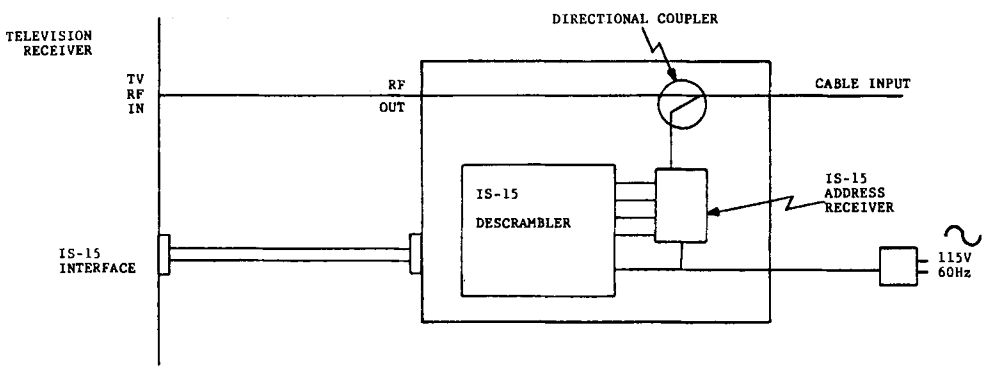

Fig. 7 IS-15 EIA MULTIPORT DESCRAMBLER WITH ADDRESS RECEIVER IN RF LOOP

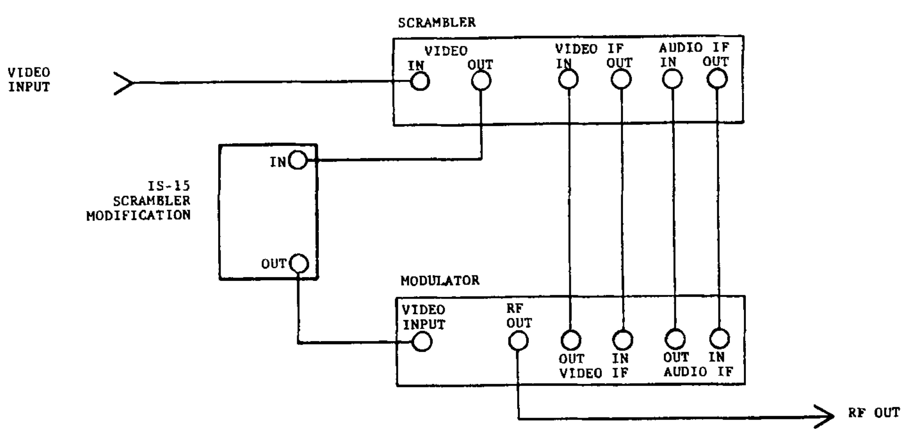

Fig. 8 SCRAMBLER/MODULATOR WITH IS-15 MODIFICATION

timing information. Figure 8 shows a possible interface. Economics would dictate whether the IS-15 board is "stand-alone" or built into the scrambler.

Addressable Systems

Addressable systems in RF sync suppressed scrambling poll each descrambler to determine presence and status of the service provided. This is accomplished by an Address Transmitter (ATX) in the headend. The ATX modulates data on a carrier in or near the FM band, which is coupled into the downstream Broadband RF.

In IS-15 EIA Multiport, a similar scheme would be employed. The amount and speed of data may require the use of a separate carrier.

CONCLUSION - MERGING OF THE TWO TECHNOLOGIES

RF Sync Suppression and IS-15 Multiport systems can co-exist in the same channel. This can be accomplished without the expense of replacing all the descramblers and Set-Top terminals in the field. Use of additional bandwidth to carry the same service would not be necessary.

Scientific-Atlanta has developed a system where compatibility is achieved by sending baseband descrambling information in the vertical interval. The data is continuously moved and/or updated, providing adequate security. This has no effect on the performance of RF descrambling in the Set-Top Terminal.

Similarly, timing information in the Audio Carrier used for RF descrambling is not available to the IS-15 descrambler.

Address receiver information may be contained on separate carriers in the FM band or elsewhere, and may be the exact data transmitted to the Set-top terminals.

ACKNOWLEDGEMENTS

The author would like to express sincere gratitude to Mr. James O. Farmer, Division Technical Manager, Scientific-Atlanta. Jim's help, patience and encouragement made not only this paper possible, but also the design effort that lead to it.

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ENCRYPTION-BASED SECURITY SYSTEMS
What Makes Them Different and How Well are They Working?

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OAK Communications Inc.

Introduction

Over the past half dozen years a new genre of secure audio/video transmission equipment based on the application of encryption technology has developed. The use of encryption has been made possible by advances in technology in several areas. Developments in semiconductors and microprocessors and the evolution of low cost digital communications and processing techniques has been married to newer industry trends such as addressability, satellite television broadcasting, and pay-per-view, which demand enhanced security.

This paper will discuss some of the fundamental aspects of how encryption technology can be used to secure program delivery, why cryptographic methods are different from traditional scrambling techniques, and how their proper application offers long term solutions where other approaches fail. We will then test these arguments by looking at how products featuring encryption-based approaches to security are doing in the marketplace.

Historical

A microcosm of what later became an industry trend took place within Oak Industries in the early 1980's. This was the battle against signal theft on both legal and technological fronts. We were rapidly expanding our subscriber base in STV in 1980 and, while vigorously pursuing pirate activity on the legal front, were experiencing severe piracy problems in our large Los Angeles and Chicago markets. It may be surprising, but the scrambling approach being used in those days was more complex than many types still going into new cable installations today. Yet we estimate at least 50,000 illegal pirate boxes existed in our subscriber base of some 380,000 in Los Angeles alone. Illegal STV boxes were selling for $200 to $400 and were generally built from scratch. This for one channel of premium programming, when over-the-airs were readily available for free! So at Oak we learned early to respect the degree to which pirate activity could organize and be technology-wise.

We sought a technology solution to the piracy problem through the application of encryption, and in 1982 launched our new Dallas, Texas and Portland, Oregon STV operations with the world's first commercial use of a security system based on true encryption principles. Also that year we introduced ORION, Oak's broadcast-quality satellite security system. The STV markets didn't survive long enough for a true test of their encryption systems (they were on the air for only about a year), but ORION is still in use today, with approximately 20,000 units in both commercial programming and private network use. We now market our Cable Sigma product line which applies similar encryption technology to the cable environment.

With the industry looking to protect satellite broadcast of premium programming services by the adoption of a de facto standard, and the private sector (networks) also widely utilizing scrambling, we see a technology solution to the theft problem on a broader scale today. Pirate activity is also being combatted legally and procedurally by companies (e.g. General Instrument vs. "Cooper et al") as well as by collective industry efforts with groups such as the Coalition Opposing Signal Theft (C.O.S.T.).

Today, a cryptographic approach to securing television transmissions is accepted as the contemporary method. Yet all products described as using encryption principles are not created equal. As we're seeing today, encryption systems can and have been compromised. The test of longevity will be those which are able to recover from a compromise once it has happened. So how does one know if a system is technically secure, or secure enough, or by what margin? How does one get past buzzwords or generalisms in developing a figure of merit on a subject as esoteric as encryption? It's not as difficult as it might seem, once a few basic themes are examined.

Three Tests for True Security

The discussion will center around three areas, each a fundamental prerequisite before the system can qualify as "cryptographically" secure. These are:

1. What is being secured? That is, what aspects of the total information transfer process are being (or, more importantly, are not being) protected by the use of encryption?
2. What are the actual encryption algorithms, and how are they used?
3. How was key management problem solved?

Unfortunately one of the areas that gets most attention in security studies is many times the least important, at least in entertainment applications. That is number (2) above. It happens to also be the most esoteric, that is, most difficult for the layman to evaluate. For the moment let's just say an "algorithm" is the lockbox that mathematically envelops or encloses the information such that it cannot be recovered without a "key". The encryption algorithm all by itself must be evaluated in terms of its ability to withstand "breaking". Once encrypted by, or through the use of the algorithm, the information must not be able to be recovered by analyzing the algorithm, or the resultant encrypted information. With today's computers and high speed logic, it is straightforward to design and implement algorithms which are low cost, and extremely "hard" or difficult to break; although you may need some expert help in this part of the evaluation. The most popular algorithm in commercial use today is the "DES" or Data Encryption Standard. DES is only an algorithm. It's use does not a secure product make by any means.

Items (1) and (3) above ensure that the algorithm is put to work properly. It is the objective of an encryption-based system to 1) "bottle" up or secure the information (in our case, programming and subscriber management information) by encrypting it, and 2) ensure that no back doors exist allowing the information to be recovered by any means other than <u>de</u>crypting the <u>en</u>crypted information at legitimate receiver sights by 3) using a secure key management scheme alongside our secured signal.

When studying encryption systems one always comes back to key management, as we'll see when we look at real world systems. Since we said that by definition the only way to recover the secured information is by using the algorithm key, the system must provide for convenient, dependable and secure methods of distributing decryption keys to legitimate receivers to recover the broadcast information.

Test Examples

Let's now look at some specific examples of the above concepts in cable or satellite programming distribution networks. First, we'll look at why traditional "scrambling" methods fail the first of our three tests for total systems security.

Consider a contemporary addressable scrambling system having scrambled programming and one or more control or addressing channels. When considering the piracy issue, which includes any kind of unauthorized access to programming, note that the control channel or channels have no relationship (as far as the pirate is concerned) to the service being purchased. One of the first questions to ask then about a scrambling system is what is the function of the control/authorization channel? That is, how is it related to the scrambling approach if at all?

In most systems the control channels direct the decoder to decode or not to decode as a function of either the channel tuned, or the tier of a given program. Critical to the issue is whether any information contained in the control channel is <u>required</u> in the decoding process. If not, the control channel can be ignored when attempting to pirate the signal. Likewise, if the scrambling technique or decoder circuitry easily succumbs to one-time defeats (e.g. a hardwired defeat), the control channel content is of no interest. Such is the case when descrambling can be accomplished by observation of the scrambled signal alone, while ignoring the control channel information.

What about "time-varying scrambling"? Time-varying scrambling adds a dimension of change to the scrambling process such that the decoder will not properly decode at all times unless it appropriately follows the change. Is this better security? To a degree, yes. But if the attribute that changes has few or trivial differences, then no real barrier to defeat of the system is actually created. Consider the pirate entrepreneur who wishes to build a "universal decoder". Most positive scrambling systems use one of several techniques of suppressing the horizontal sync pulse. ("Positive" systems are those that actively scramble the premium signal, and thus require a decoder. "Negative" systems remove the signal from the unauthorized viewer through traps or signal path switching). Whether the system's scrambling is at RF or baseband, our pirate's universal decoder can quite easily be designed to reconstruct the sync pulse and completely ignore all control channel information, time-varying or not.

This discussion is gearing us toward a theme: In programming distribution, security is a systems issue. The simplest method of defeat will be the path followed by the would-be pirate. The system must therefore be viewed from several angles and an adequate threshold against compromise developed for each. In so doing, one must ask what information (timing, control data, circuitry, etc.) is available at the receive location that can be used to get around (not through) the secured or encrypted material. There is usually an amazing amount of data available to tap. How much added security is afforded by random video inversion of the picture for example, if a simple-to-detect "flag" exists in the vertical interval indicating current polarity? Is any security afforded in an addressable system simply because it's addressable? Not if it's easier to address (authorize) the box yourself than it is to open the box up and tamper with circuitry. This is our first test then, the notion of the "back door" entry to information. Why worry about breaking the encrypted audio out of your satellite receiver, for example, if it's coming out of your local cable drop in the clear?

The Encryption Algorithm

We'll now digress momentarily to discuss some details of what constitutes the encryption algorithm, and how keys or key variables are used. The encryption algorithm executes a "digital processing" function. The actual entity that undergoes encryption must be in a digital format. The output of the algorithm can then be used to perform other random processes, if desired.

CLASSICAL CRYPTOGRAPHIC SYSTEM

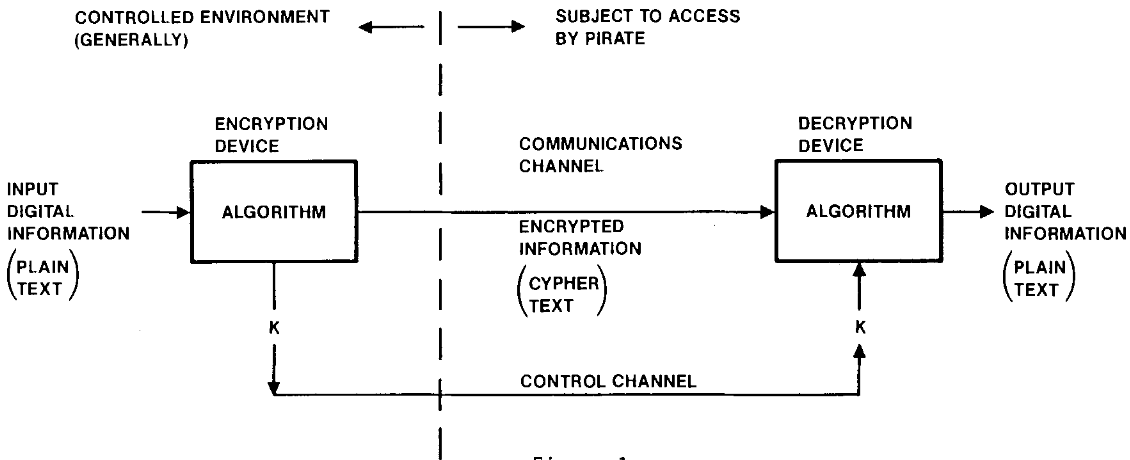


Figure 1.

In a conventional encryption system (Figure 1), a digital bit stream (the information) is passed through the algorithm that transforms the input into a seemingly unrelated output bit stream. The transformation that is performed is a function of the "key variable", and in a conventional system the same key is used at both the transmit side where encryption is performed and the receive side where decryption is performed. A different transformation is implemented whenever the key changes. The key is a digital word of many bits (generally in the range of 24 to 64 bits), so 2^n (where n is the number of bits in the key) different transformations are possible by varying the key. In a properly designed algorithm, all keys are equally strong (i.e., resistant to "cracking") and no detectable relationship exists between the input data, output data or key variable. Each combination of key bits represents a completely different scrambling "mode" and there is no such thing as "almost having the correct key". The key must be exactly correct or no decryption is possible.

The process of encryption must, of course, be reversible. That is, applying the same key at the receiver must restore the original information. The original, non-encrypted data is called clear or plain text, the encrypted data is called cipher text. So during transmission (i.e., between headend and decoder), only non-intelligible cipher text is available to the would-be tamperer. If the decoder doesn't have the proper key, no message or clear text will be obtainable, even if the pirate has the hardware. Further, in a properly designed system based on cryptographic security principles, we can give the pirate just about anything he wants: hardware, access to, and knowledge about the control channel, schematics, any firmware, and even the crypto-algorithm itself. The only doorway to information access, in our case programming, should be through the key variable (no back doors, right?). Controlling access to the key variables is thus essential. This is called "key management" and is the basis for what ultimately makes or breaks the security of a cryptographically-based system. The cryptographic or encryption algorithm, therefore, can be thought of as a lockbox. The message is encrypted or locked by the algorithm, and can only be unlocked by the same algorithm, which means the identical digital key must be used for decryption.

Now that we have discussed some essentials, the value of encryption as a mechanism for security will be more readily evident. For encryption simply enables a complex security problem, in which many variables (audio, video, control) must be secured, to be reduced to simple protection of a few digital keys. Figure 2 summarizes these and other advantages of digital encryption.

Key Management Problem

The third of our three tests for security asked about the key management problem. Encryption alone will not assure the security of information in any network in which it's used unless the key management problem is carefully addressed. In the broadcast scenario, the problems of key variable distribution are particularly challenging (in comparison

ADVANTAGES OF DIGITAL ENCRYPTION AS A BASIS FOR SECURITY

- IMPLEMENTED WITH INEXPENSIVE DIGITAL HARDWARE
- ENCRYPTION REQUIREMENTS INTEGRATE NICELY INTO THE ADDRESSABLE CATV ENVIRONMENT
- EASILY AND NATURALLY BECOMES TIME VARYING
- SECURITY IS NO LONGER MANIFESTED IN PROPRIETARY CIRCUITS
- LEVEL OF SECURITY ACHIEVED CAN BE ORDERS OF MAGNITUDE ABOVE ANALOG SCRAMBLING APPROACHES OF EQUIVALENT COST

Figure 2.

to applications where only point-to-point situations exist). It probably has occured to the reader by now that, if access to working hardware is given the pirate, it is little trouble to determine what digital key is being used for decryption. Recall that previously it was stated that one-time defeats won't be allowed. Therefore, the encryption/decryption keys must be changed from time to time. The time interval depends on the key length, the ability of the encryption algorithm to resist analysis by computer, the expected accessibility of keys and the motivation of the system's enemy.

In an addressable system, the control channel is the obvious choice for a key distribution path. (Alternate methods might be by courier, mail, etc.). But one can't just go broadcasting the new keys throughout the network. They must remain private to all but authorized decoders. The solution for controlling key access is to encrypt the keys for transmission. By transmitting decryption keys in an encrypted form throughout the system, we have not really solved the key distribution problem, however, because to decrypt these keys requires yet another key. Such is the notion of "multilevel key distribution" (Figure 3). Various information exchange networks utilize different solutions to a multilevel approach. In the television broadcast environment, either satellite or CATV, the requirements dictate that: 1) when the keys are changed (updated), all decoders (and encoders too) must do so at the same time; 2) the system operation must ensure that all decoders have had the new keys properly delivered, decrypted and prepared prior to engaging them; and 3) only authorized decoders are able to perform (1) and (2).

In fact, many types of information passing through the control channel are candidates for encryption. Authorization or tiering data, for example, also should be considered "sensitive" information since, as pointed out earlier, it can easily be locally synthesized and fed to the decoder by simple digital hardware or any home computer. Such control channel manipulation by other than the legitmate network controller is just as dangerous a form of tampering as hardware tampering. Attempts to subvert the system by such address channel tampering is called "spoofing". Integrated within the operational framework of the system must be a fully developed methodology for key distribution and protection against spoofing.

Spoofing can take several forms, depending on how the pirate is attempting to fool the box. Control channel mechanisms must be in place to:

1) Prevent the insertion of illicit control data.
2) Prevent the deletion of valid control data.
3) Prevent the modification of valid control data.
4) Prevent the replay of control data.
5) Prevent box swapping between systems and geographic areas, and ensure stolen boxes are, or become useless.

If the first two tests or prerequisites to a secure system have been properly attended to, one has a totally secure system -- for the moment. It's the ability of a system to refresh key variables securely that will then enable it to be secure over time.

The "Key Question"

In a general sense, all control data having to do with enabling or directing box functions can be thought of as a form of keys. Virtually all systems in use today rely on a hierarchical or multilevel key distribution process as previously described. At the time boxes are installed they are brought on line by being given certain information (current keys) which allows them to join the network. Thereafter, having joined or signed up, they

MULTILEVEL KEY DISTRIBUTION

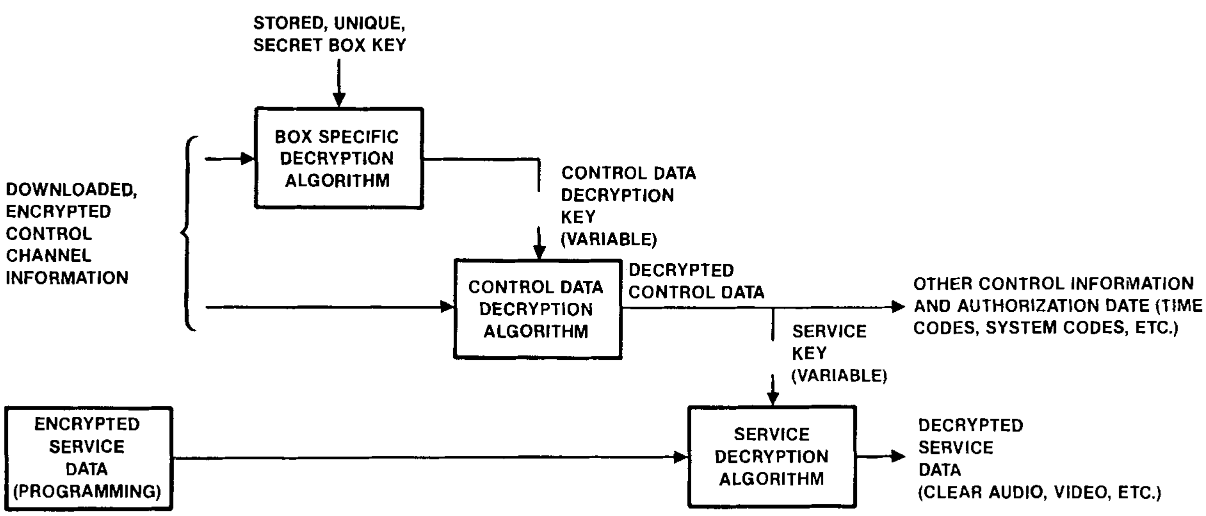


Figure 3.

are kept on line by being included whenever new network enabling codes (authorization codes, security codes, time codes, channel codes, program codes, keys, etc.) are distributed. The updating process must be performed securely, otherwise illegitimate boxes can listen in and update themselves too. Remember, having satisfied our first two tests, all boxes, legitimate or illegitimate, must have these decryption codes or key information before the signal can be recovered. If the network is distributing new keys encrypted, they must then be decrypted under existing keys already in the box. If existing boxes are suspected of being copied (cloned), then one has to consider the probability that the cloned boxes also received the updated codes or keys.

Now one of the most important points of this paper: Since all systems operate in multilevel key distribution formats, where new authorization data is distributed encrypted under "old", then how is it possible to really maintain security at all? It isn't, unless each and every decoding box has something fundamentally different about it from all other boxes. (Now we're back to the "one-time-defeat" idea). The multilevel key distribution process must begin at its most fundamental level by distributing information uniquely to each box, and then build further layers on that initial unique information. This requires a unique code or "box key" for every box, which is unknown by any other box. Thereafter, the network can always fall back to rebuilding its key levels by starting over with each box, leaving out of the redistribution or rebuilding those boxes which are known to have been cloned, stolen, etc. With the cloned box not reauthorized, all of its offspring are also up the proverbial creek.

One final comment about this notion of box uniqueness. When looking at any system purporting to have "encryption-based" security, don't let its manufacturer side-step the question of box differentiation. Without the equivalent of a box key, which is never broadcast over the control channel (because it must be encrypted under "something" to do so), the system is fundamentally, and by cryptographic standards, absolutely insecure. That the process is "too complex" or "proprietary" is the usual argument when such questions are asked. These notions are basic to any sound crypto system and they are not complex at all. Now let's look at our industries' experience with encryption products, and see why good system designs which take into account all three of our tests are so important.

The Real World Test

This paper began by noting that encryption-based systems have been commercially utilized only since 1982. Within that period, however, over a half dozen major types of products have seen extensive utility in several marketplaces. However, only three products have had exposure to the extent that significant piracy efforts have been mounted; these are the Oak ORION and General Instrument VideoCipher 2 (VC2) satellite products and the Oak Cable SIGMA system. The Oak STV SIGMA, SA B-Mac, Oak/Leitch Video Polaris, and GI Starlock systems are not believed to have ever been compromised, but have not had the exposure time and/or appropriate audience to have been really tested.

We'll make a distinction now between a system compromised and a system broken. A "compromise" is a temporary condition which can be expected to develop, and has developed, for both the ORION and VideoCipher products. "Breaking" the system would be a condition where the headend no longer has the ability to overcome the compromise, or deauthorize a decoder. This has not happened. After four years of operation and over 160,000 units installed, our Cable SIGMA system has yet to have shown evidence of any compromise.

In the cat-and-mouse game between manufacturers and pirates, the compromise of ORION and VideoCipher have been much ballyhooed in the press. It makes great gossip! But you will not hear companies like Oak and General Instruments actively responding to claims and challenges by individuals or organizations involved with the illegal activity of stealing programming. We will be quietly going about the business of ensuring that appropriate measures are taken to update keys and deauthorize modified or cloned boxes as they are discovered.

Let me now describe what has been in process in Oak's ORION system with our major customer CANCOM, for several months. CANCOM, the Canadian Satellite Communications company, chartered by the Canadian Radio and Telecommunications Television Commission, has been using ORION since 1982 to secure eight channels of television programming to CATV systems and individual homes. Approximately 15,000 decoders are currently on line.

Our knowledge of modified ORION decoders at the time of this writing indicates that most approaches have caused the boxes to simply ignore tiering alteration commands. This is a trivial compromise to overcome from the headend, and not nearly as sophisticated as either the cloning or "Three Musketeer" attacks that VC2 has seen. We are currently in the process of performing a "cycle change" on the CANCOM system. In Oak vernacular, this means the complete rebuild or redistribution of the network decryption keys. ORION has the attribute that each box stores a secret, and unique box key under which this is done. This has never before been executed on the CANCOM system, as the headend control system has only recently been upgraded to perform this function. The original design, however, planned for this exercise and there are no decoder modifications required. Any illegal decoders still operational after the cycle change can be assumed to be clones. A subsequent cycle change will then be performed, with clones thus identified eliminated from the redistribution process. Why this simple technique can work effectively is because it's computer-controlled, passive (that is, a background function) and very easy to invoke. Pirate boxes can always crop up, but once discovered can always be shut down.

This total redistribution is not possible with any system that does not have the equivalent of a box key. If, at the deepest or most fundamental level, boxes are manufactured with hardwired keys or key seeds, once uncovered, these seeds will cause the redistribution process to be insecure and thus piratable. It may take a while -- we didn't see any significant piracy in ORION for three years -- but pirates will eventually break any system with hardwired or hardcoded keys.

Summary

The level of sophistication and organization behind the attacks currently being mounted against VC2 should lend credence to arguments that "OK" security is really not OK any longer. The cable industry should in fact take a lesson from what is happening in the satellite arena and understand why. The why is really economics. As cash flow from services increases, either through new revenue generators (IPPV and home shopping!) or increased audience, the motivations for system subversion (not just signal theft) will also increase. Along with the tests for security reviewed above, Figure 4 outlines additional considerations that relate to areas such as internal threats from employees, increased sophistication of the enemy, and advances in the state of the art.

Oak and our equipment manufacture competitors have spent a great deal of time and energy over the past two or three years educating our industry with respect to the merits of encryption. There is a tendency on the part of some manufacturers to confuse the issue by jumping on the bandwagon, claiming encryption processes are employed when what is really being done is trivial to undo under some of the examinations we looked at earlier. When such products are defeated, together with the publicity about products featuring true encryption getting compromised, the public and our industry gets naturally confused, and misled, and it is the consumer who eventually gets duped in the process.

The success that Oak, and other companies may enjoy (literally, sometimes) in combatting piracy is due to proper attention to theory and practical considerations in the application of true encryption principles. This paper has discussed some basic attributes and prerequisites of those principles, and reviewed how they can be employed to take advantage of the resultant system strengths.

GRADING A SYSTEM'S SECURITY

- KNOW THE FULL RANGE OF POSSIBLE ATTACKS
- DO NOT UNDERESTIMATE THE ENEMY
- BE AWARE OF TECHNOLOGY ADVANCES
- UNDERSTAND THE BASICS OF ENCRYPTION
- UNDERSTAND THE APPLICATION OF ENCRYPTION

Figure 4.

FEASIBILITY OF MULTI-CHANNEL VSB/AM TRANSMISSION ON FIBER OPTIC LINKS

Jack Koscinski

GENERAL OPTRONICS CORPORATION

ABSTRACT

Fiber optic supertrunks have already demonstrated the capability of transporting 8-16 video channels/wavelength over path distances greater than 20 km (repeaterless). These systems have traditionally utilized wideband FM as the RF modulation technique. Wideband FM provides the dual benefits of 1) signal to noise improvement and 2) high immunity to intermodulation distortion.

While the performance of these fiber optic (FO) links is impressive, they are also relatively costly (>$5K/channel not including FO cable). Additionally, the use of FM modulation requires a potential modulation conversion (AM/FM/AM) at both ends of the FO link with the attendant costs for AM modulators and demodulators. For many broadband transportation applications these costs are prohibitive and have prevented the widespread usage of FO links.

The ability to use multi-channel VSB/AM transmission is highly desirable. The ideal scenario would be to accept a broadband signal (bridger output etc.) and transport these signals directly over the fiber optics links without any modulation conversion.

The scope of this paper reviews the following areas:

- A Multi-channel VSB/AM Fiber Optic System Block Diagram.
- Basic VSB/AM System Considerations.
- Performance of a Multi-channel VSB/AM FO Link.
- Feasibility of Optimizing the Performance and Number of Channels on VSB/AM FO Links.

INTRODUCTION

Fiber optic systems have been finding more and more applications where they provide the most cost effective solution. The majority of these applications pertain to digital communications. However, there are a large number of analog system applications for which fiber optic links are also being selected. The most common requirement for analog transmission on a fiber link is transportation of video signals. Although the video information could be converted to a digital signal prior to transmission, this is usually cost prohibitive. In many point to point applications, where repeaters are not required, fiber optic analog (FO) transmission systems can prove to be the most cost effective approach.

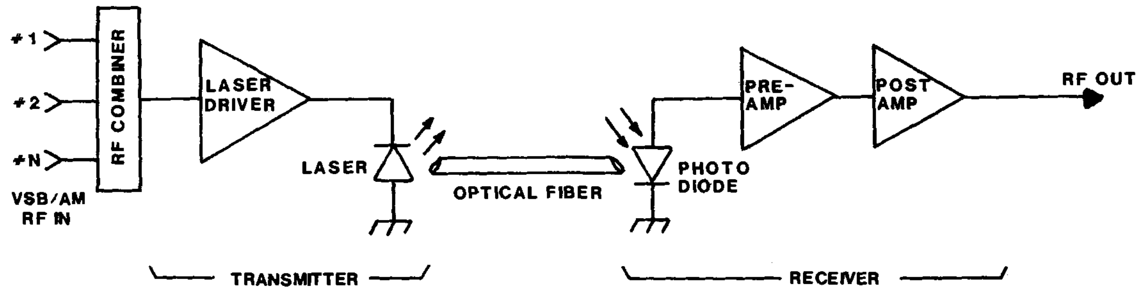
This paper focuses primarily on FO analog VSB/AM transmission applications for carrying multiple video signals over a fiber link. Cable television supertrunks and campus networks would be typical applications where these multichannel analog transmission systems would be utilized.

Although there are several types of lasers and fiber available, this discussion will deal exclusively with 1300 nm single mode lasers and single mode fibers. These selections are becoming de facto standards for optimum multi-channel FO analog systems since they provide low loss links which are immune from multimode noise phenomena (modal noise).

MULTICHANNEL VSB/AM FIBER OPTIC SYSTEM

A block diagram for a multi-channel VSB/AM fiber optic link is illustrated in Fig. 1. The major system components are:

1) the optical transmitter, 2) fiber optic cable, and 3) the optical receiver.



VSB/AM MULTI-CHANNEL FIBER OPTIC LINK

FIG. 1

Optical Transmitter

An optical transmitter accepts the individual RF VSB/AM analog inputs and provides the signal conditioning necessary to drive a semiconductor laser diode. Fig. 2 shows the major optical transmitter components.

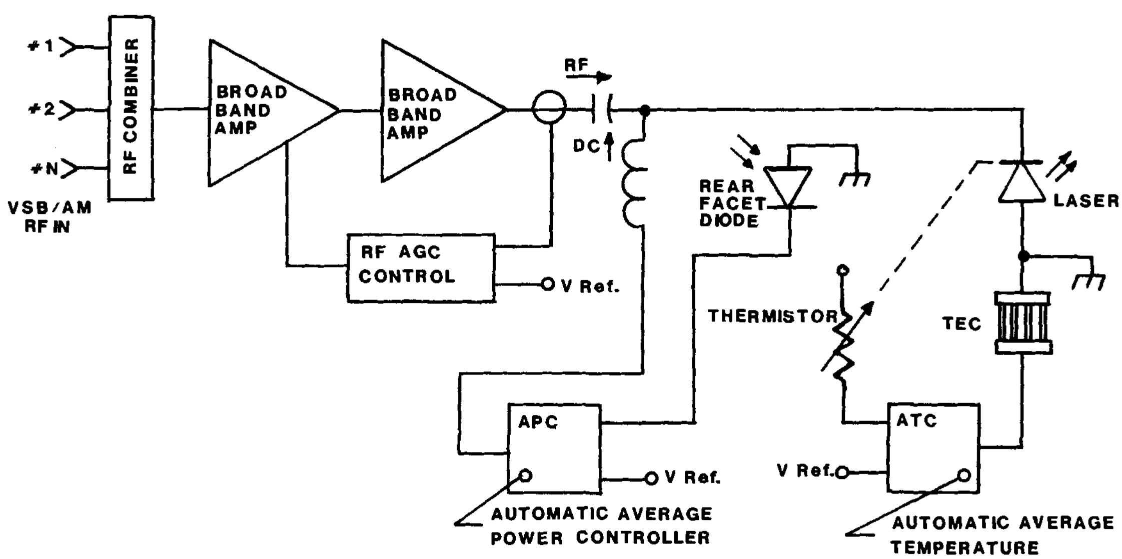
First, an RF combiner sums the multiple analog inputs which are to be transmitted. The RF levels for each carrier should be equalized prior to the optical transmitter. Otherwise, optimum noise and distortion performance for each carrier will not be achieved. However, a slope compensation stage may be provided after the combiner stage to adjust for normal cable slope effects.

Broadband amplifiers, AGC controlled, provide the necessary signal level to drive the laser diode. The RF drive level to the laser must be precisely controlled to realize the optimum system noise and distortion performance.

A d.c. bias is applied to the laser to provide a linear operating point. This d.c. bias current will determine the average optical output power out of the laser diode which is typically 0.5 mw for single mode 1300nm lasers.

Laser power is sensitive to changes in temperature and laser aging. To preserve a constant average optical output power, two control circuits are commonly provided in the laser transmitter; a laser temperature controller and an automatic optical power controller.

A photodiode monitors the rear facet of the laser as a sample of the transmitted optical power and uses this information to control the laser d.c. bias current. Thus, if the laser average optical power changes due to time or temperature, the laser bias is automatically adjusted to maintain constant average optical power.



RF ANALOG OPTICAL TRANSMITTER

FIG. 2

Laser life is adversely affected when operating at higher temperatures. Temperature control is accomplished by using a thermistor to monitor the laser temperature. A control circuit then drives a TEC (thermal electric cooler) to which the laser heatsink is mounted to maintain the laser at a constant temperature, typically 20 C).

Fiber Cable

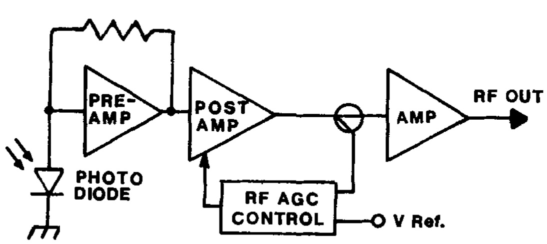
Single mode, 1300 nm, fiber cable is preferred for multi-channel analog systems. This fiber has a core diameter of only 9 um with an overall cladding/buffer diameter of 950 um (typical). These fibers may be assembled into various cable assemblies which provide multiple fibers, strain relief, and jacketing options.

Fiber cable is available in lengths up to several kilometers per reel. For distances greater than several kilometers, the fibers are typically fusion spliced to minimize path insertion losses. Single mode fusion splices have typical optical losses of only a few tenths of a dB. Single mode connectors have losses of 0.5 dB. The optical transmitter and receiver are normally connectorized for convenience in servicing equipment.

Optical Receiver

The primary function of an analog optical receiver is to reconvert the light power into an RF signal with a minimum contribution of noise and distortion.

An analog optical receiver block diagram is shown in Fig 3. The optical detector commonly employed for 1300 nm analog applications is either an InGaAs pin-diode or a Ge avalanche diode The major distinction between them is that the Ge avalanche diode has gain available (approx. 10) whereas the pin-diode has unity gain.



RF ANALOG OPTICAL RECEIVER

FIG. 3

The photodiode current drives a transimpedance pre-amp which provides high input sensitivity and converts the diode current into a voltage at its output. These preamps are available as DIP package devices with fiber pigtails attached.

A post amplifier, with AGC, follows the pre-amp to provide sufficient gain to obtain unity system gain for the entire FO link. AGC is utilized to maintain a constant output level independent of optical input power which may change due to fiber resplicing, fiber loss versus temperature etc.

BASIC VSB/AM
SYSTEM CONSIDERATIONS

The primary advantages of multi-channel VSB/AM transmission over multi-channel FM or digital FO systems are bandwidth efficiency and cost. A standard VSB/AM channel occupies a 6 MHz bandwidth whereas wideband FM uses 30 to 40 MHz of bandwidth per channel. Digital video requires even larger channel bandwidths for high quality video transmission.

System costs/channel for a VSB/AM FO system are also generally lower than a comparable FM or digital FO link. The major reason for this is the elimination of FM modems and digital codecs which are required when FM and digital FO links are employed. Fig. 4 illustrates the basic hardware differences for these three system approaches.

One might assume from this discussion that VSB/AM is being used extensively in multi-channel analog FO systems. However, this is not yet the case.

In spite of the availability of very wide bandwidth FO systems, the maximum number of VSB/AM channels which may be transmitted is limited by the optical source (laser transmitter) due to power loading and distortion constraints.

Semiconductor lasers currently provide maximum average optical output powers of -3 dBm. When a single carrier modulates the laser, all of the laser light power is dedicated to this one channel. As more carriers are loaded onto the laser, the corresponding optical power available per channel will be reduced since they all must share the finite available laser optical power. Assuming a constant average depth of modulation on the laser, the available optical power (per VSB/AM channel) degrades by 3 dB for each doubling of the number of channels transmitted. The result of lower transmitted carrier power is a corresponding reduction in carrier/noise ratio.

Depth of modulation of the laser is defined as the amount of current shift in the laser due to the modulation signal relative to the laser bias current. Fig. 5 illustrates this by showing a typical laser LI (light/current) curve. Increasing the depth of modulation of the laser will increase the effective (RF) transmitted laser power. This can be seen from the equation for total power launched by a semiconductor laser diode.

$$P(t) = Pb \left[1 + ms(t)\right] \qquad (1)$$

where:

m = modulation depth
Pb = average laser optical power
s(t) = modulating signal (composite)

Increasing the modulation depth of the laser will also tend to increase the signal distortion. To control distortions in the optical output signal, the modulation must be confined to the linear region of the laser LI curve (Fig. 5). Typical m values for analog applications range from .25 to .50 depending on the linearity of the laser and system performance requirements.

Typical values of distortion for single mode lasers operating at 50% modulation depths are:

2nd orders : 30-45 dB
3rd orders : 45-60 dB

The spread in these values indicates that lasers do not necessarily have consistantly good linearity. Careful specification criteria and selection of lasers for analog parameters are very important to obtain high performance in multi-channel analog FO links.

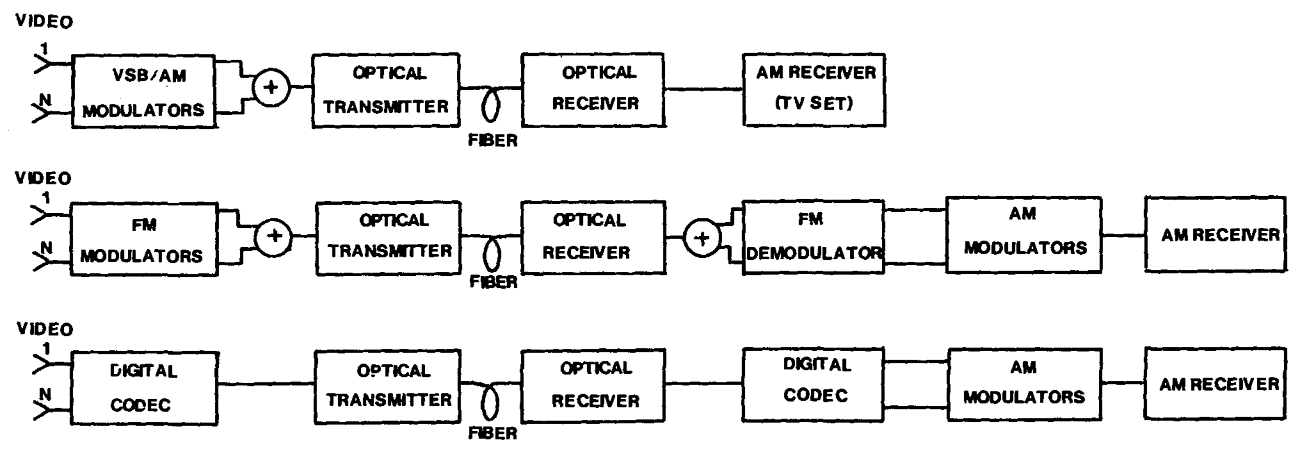
Due to the relatively high levels of distortions present (especially 2nd order) in semiconductor lasers, multi-channel VSB/AM systems may require careful spectrum planning. Schemes which avoid channel transmission where second orders are present may be required. One approach is to utilize only the upper octave of transmission bandwidth. Limiting the transmission bandwidth to the highest octave will allow all the second orders to land above or below the desired transmission spectrum (Fig. 6).

As can be understood from this discussion, a direct tradeoff exists between system distortion and carrier/noise performance as contributed by the optical source. Typical system performance will be reviewed by discussing a VSB/AM FO system which has been implemented recently by General Optronics.

VSB SYSTEM APPLICATION

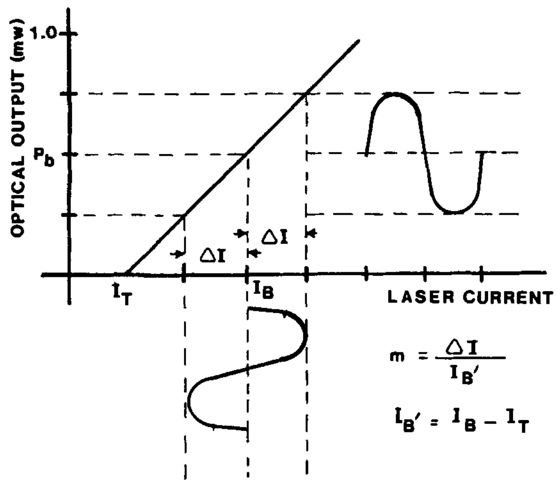
VSB/AM fiber optic systems have successfully transmitted 4 to 8 channels of video up to 12 km. Figure 7 illustrates the block diagram of a system installed for Seattle City Power & Light.

The application required transmission of channels 4, 5, 7, 9, 11 and the complete FM band over a distance of 12 km. Signal inputs to the fiber system were fed from a bridger amplifier output of a CATV system. The accessibility of the terrain required a repeaterless link which immediately ruled out extending the coax system. Microwave was also ruled out due to not having a line of sight available. Thus, a fiber optic link was selected as the implementation approach.



VSB/AM - FM - DIGITAL FO LINKS

FIG. 4



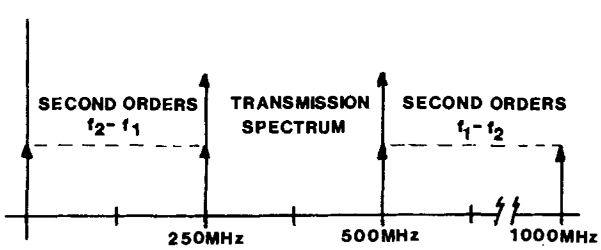
LASER L/I CURVE

FIG. 5

Two fiber optic links were utilized to transmit the 5 VHF channels and the FM radio signals. As shown in the block diagram (Fig. 7) the input signals were separated into the low VHF and high VHF bands with a VHF band splitter. Effectively three channels were carried on each fiber optic link. One link carried channels 7, 9, and 11. The second link carried channels 4, 5, and the FM band.

The optical transmitters are broadband (5-250 MHz) which accept RF input levels from 30 to 50 dBmV. AGC is included within the transmitter to provide a stable RF drive level to the laser. These transmitters employ single-mode semi-conductor laser diodes operating at 1300 nm wavelengths and launching average optical output powers of greater than -3 dBm.

The optical receivers employ wideband high sensitivity pin-fet preamplifiers. InGaAs pin-diodes are used as the opto-electronic conversion component. Post amplifiers with AGC provide stable output levels within the range of 30 to 50 dBmV.



SPECTRUM PLAN
TO AVOID SECOND ORDERS

FIG. 6

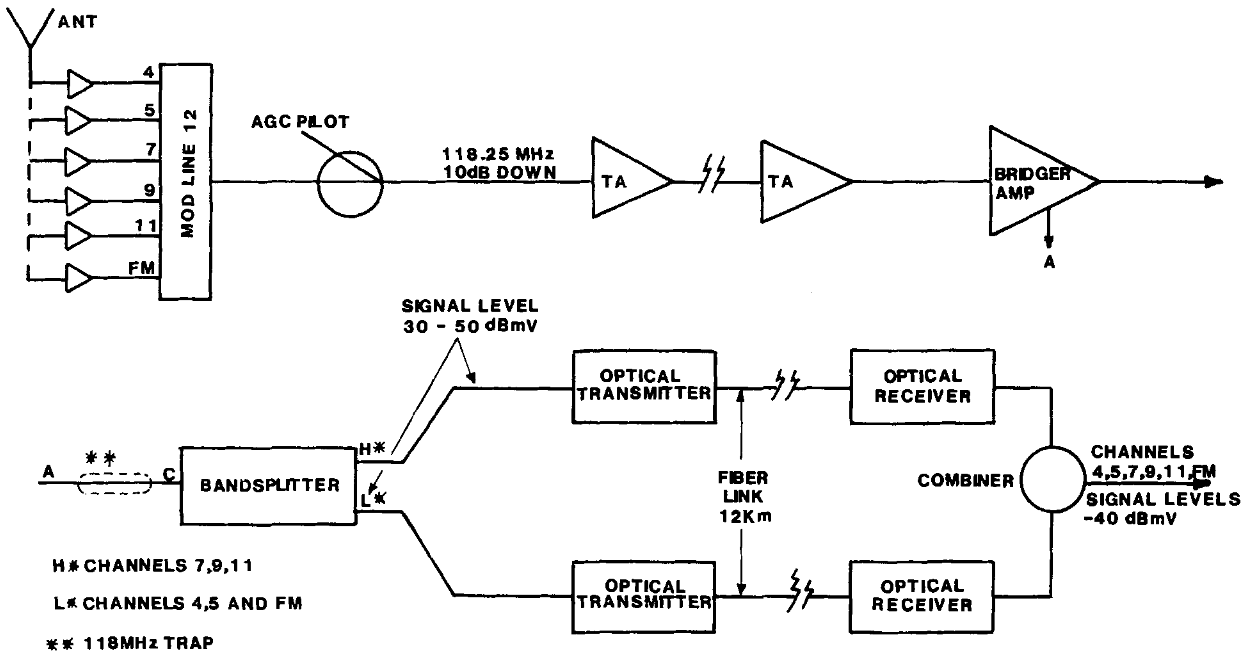
The key performance parameters for these VSB/AM FO systems are carrier/noise and intermodulation distortion. Figure 8 illustrates the typical performance obtained from these VSB/AM FO links. Four carriers at channels 8, 9, 11, and 12 were used for a proof of performance of the links. Carrier/noise (BW=4MHz) per channel measured greater than 45 dB. Third order (worst case) intermodulation distortion measured lower than -60dB. Second order distortion contraints were avoided by choosing each link to transmit less than an octave of bandwidth. Second order distortion measured approximately (40 - 45) dB.

These results represent the current capability of multi-channel VSB/AM FO systems. The lasers used in these FO links are General Optronics high performance Dip 6300 semiconductor laser diodes which have been carefully specified and screened for optimum performance in analog FO systems.

Since the number of VSB/AM channels must be limited (4 - 8) to maintain good performance, these links presently will not accommodate CATV link requirements where channel loadings of 30-50 channels are standard. However, an attractive application for these FO links would be the transport of video in the university or industrial campus environments. FO links are already providing transportation of voice and data in these environments. If multi-channel video distribution is also required, using the same fiber cable (different fibers) for video could result is measurable cost savings.

POTENTIAL ADVANCES IN VSB/AM FO SYSTEMS

As mentioned earlier, the number of channels which may be transmitted as VSB/AM on a fiber link is currently limited to (4-8) channels depending on the performance objectives. Currently General Optronics has undertaken a study to further improve the performance and increase the number of channels. For this to happen, the key system performance parameters (carrier/ noise and distortion) must be further improved. The following discussions review these FO parameters and suggest how they may be further optimized.



SEATTLE CITY POWER & LIGHT SYSTEM BLOCK DIAGRAM

FIG. 7

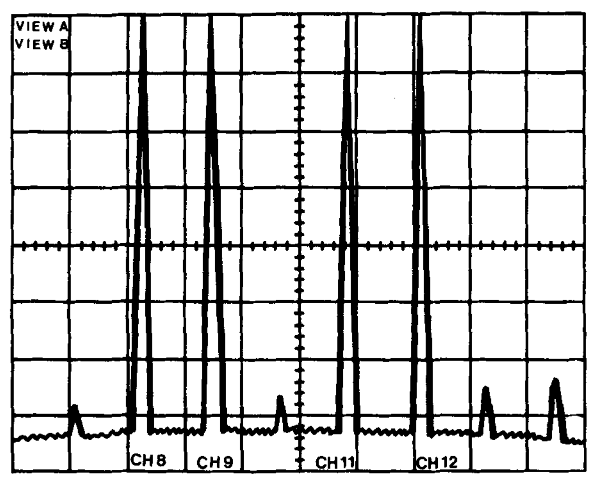
Carrier/Noise

Carrier/noise available at the output of a multi-channel analog FO link (pin-diode optical receiver) is described by:

$$\frac{C}{N} = \frac{(1/2N)(m\ Ro\ Pb)^2}{\underbrace{(RIN\ Ro^2 Pb^2 B)}_{(SOURCE)} + \underbrace{2q(Ro\ Pb + Id)B}_{(QUANTUM)} + \underbrace{(4KTB/Req)Ft}_{(RECEIVER)}} \quad (2)$$

Ro = diode responsivity (A/W)
m = modulation depth
Pb = average optical power rec'd (W)
q = electron charge
K = Boltzmann's constant
Id = diode dark current (A)
B = bandwidth of receiver (Hz)
T = temperature (K)
Req = equivalent resistance of photodiode load and amplifier (OHMs)
Ft = noise factor of preamplifier
RIN = source relative intensity noise (dB/Hz)
N = number of FDM channels

As seen from the equation, there are three independent sources of noise present in a FO system: 1) source, 2) quantum, and 3) receiver. Source noise is the inherent residual noise of the laser diode which is referred to as RIN (residual intensity noise). RIN of semi-conductor lasers used for these applications is approximately -120 to -140 dB/Hz. Quantum noise is the "shot noise" of the pin-diode which is directly proportional to average optical power received and bandwidth. Receiver noise consists of the "KTB" noise power of the receiver and the noise figure of the pre-amplifier electronics.



CF:	195 MHz	BIAS:	33.1 ma
HORIZ:	5 MHz/div	OPTICAL POWER:	500 uw
VERT:	10 dB/div	MOD. DEPTH:	60%
RES B.W.:	10 KHz	CARRIER F:	CHANNELS 8,9,11,12

VSB/AM FO LINK PERFORMANCE

FIG. 8

As can be seen from the equation, the carrier/noise versus optical power received (loss budget) will have several breakpoints.

When the optical power incident on the photodiode is low the "receiver" noise term dominants the system noise, so that

$$\frac{C}{N} = \frac{(1/2N)(m\ Ro\ Pb)^2}{(4K\ TB/Req)FT} \qquad (3)$$

Here the carrier/noise ratio is directly proportional to the square of the average received optical power. Thus, for each 1 dB change in optical power received, the carrier/noise ratio will change by 2 dB.

For larger optical signals incident on the photodiode, the quantum noise associated with the signal detection process dominates (assuming Id negligible), so that

$$\frac{C}{N} = \frac{(1/2N)(m^2 Ro\ Pb)}{(2qB)} \qquad (4)$$

Since the carrier/noise ratio in this case is independent of circuit noise, it represents the fundamental or quantum limit for analog receiver sensitivity. In this optical power range the carrier/noise ratio will change 1 dB for each 1 dB change in received optical power.

For very high optical power levels the carrier/noise ratio will be limited by the laser source.

$$\frac{C}{N} = \frac{(1/2N)(m^2)}{(RIN\ B)} \qquad (5)$$

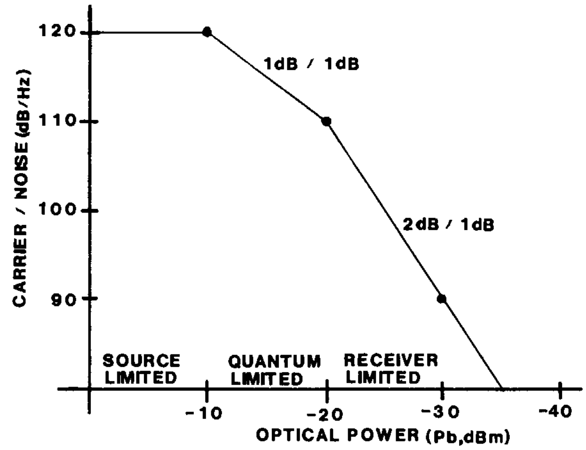
Thus, for very high optical powers, the carrier/noise ratio is constant at the maximum obtainable from the laser transmitter.

Fig. 9 illustrates the carrier/noise obtainable at the receiver output as a function of optical input power when these noise sources are present.

Several conclusions can be from Fig. 9.

1. The ultimate carrier/noise is limited at approximately -10 dBm of received optical power. Thus, for FO loss budgets of 7 dB (10-14 Km), the carrier/noise will be limited by the optical source.

2. The quantum limit will not be achieved due to the RIN noise of the laser.



CARRIER/NOISE VERSUS RECEIVED OPTICAL POWER

FIG. 9

3. Between -10 to -20 dBm (quantum limited), the carrier/noise ratio will vary 1 dB/1 dB of optical received power.

4. Beyond -20 dBm the carrier to noise ratio degrades 2 dB/1 dB of received optical power.

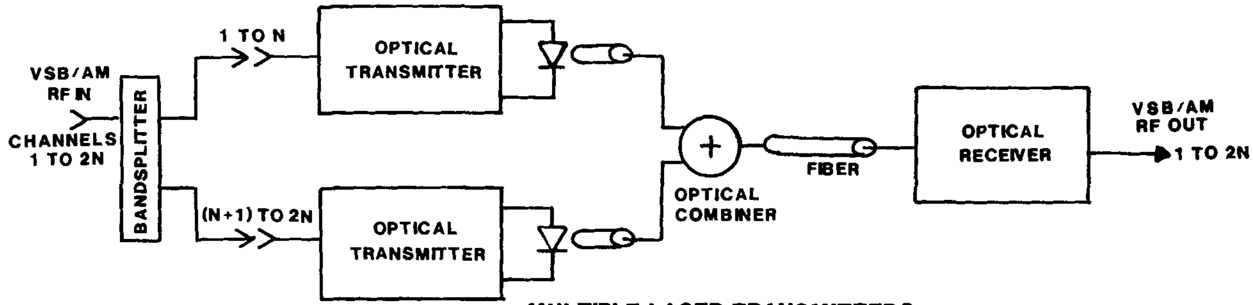
Having reviewed the carrier/noise expression, what options are available to further optimize VSB/AM FO link performance for carrier/noise? The primary candidates are to reduce the RIN of the laser diodes and to increase the modulation index.

For typical lasers emitting several mw of optical power, the RIN lies between -120 to -140 dB/Hz. However, the RIN of the laser is not constant. It depends on the optical power level to which the laser is biased. The RIN varies as the inverse cube of the bias optical power. General Optronics has provided single mode laser diodes which launch optical output powers up to 2.5 mw peak. This can provide up to 10 dB of carrier/noise improvement as compared to 1 mw lasers.

Increasing the modulation depth will also provide a carrier/noise improvement. As "m" is increased by a factor of 2, the carrier/noise ratio will improve by a factor of 4 (6 dB). However, as discussed earlier, laser distortions usually dictate the maximum modulation depth to which the laser is modulated.

Distortion

Augmenting the finite linearity of the semiconductor laser is the key to significant performance improvements.



MULTIPLE LASER TRANSMITTERS

FIG. 10

The benefits of improved linearity are:

1. Higher values of modulation index may be used to drive the laser (thus improving system carrier/noise ratios).

2. More VSB/AM channels may be transmitted on the FO link without being limited by second and third order distortions.

There are several approaches that provide linearity enhancements in analog FO systems. Since the lasers are the limiting distortion component, all these techniques center about the optical source.

Multiple Laser Transmitters

This approach is illustrated in Fig. 10. Two laser transmitters are used to double the channels transmitted onto a single fiber and into a single optical receiver. The RF VSB/AM input signals are FDM (frequency division multiplexed) which allows a simple optical combiner to couple the two laser light signals together. If the RF inputs were at the same frequencies, then two distinct laser wavelengths and a wavelength division multiplexer would be required to combine the two laser signals onto a single fiber.

This approach allows a doubling of the number of VSB/AM channels transmitted with the tradeoff of a 3-4 dB loss in the optical loss budget due to the optical combiner.

There are two optical sources required in this approach. However, the costs of lasers are decreasing as a result of competition and product manufacturing maturity.

The RF input signals need to be filtered off and isolated from each other prior to driving each laser transmitter. This may be accomplished with bandsplitting filters.

Linearized Optical Transmitters

There are two basic approaches which provide linearization of optical transmitters; 1) feedforward compensation and 2) negative feedback.

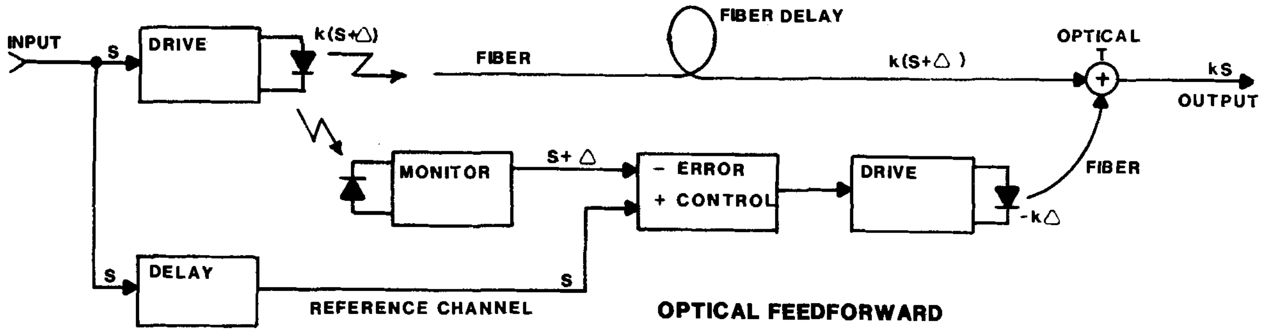
Feedforward Compensation

Feedforward compensation is achieved by isolating the distortion produced in a nonlinear circuit and subsequently injecting the processed error signal back into the circuit. This is the same principle utilized in CATV RF feedforward amplifiers.

The optical feedforward system shown in Fig. 11 requires a monitoring photodiode, an error processing circuit, and a second optical source to generate a compensating optical signal. This signal is then coupled to the original optical signal through an optical combiner. The practical feasibility of this technique has been enhanced by the development of low loss single fiber optical couplers and combiners.

Optically matched sources must be available to provide significant distortion improvements (20 -30 dB). However, these may be provided by matching LI curves closely. Long term tracking may be a problem if the laser characteristics do not age uniformly. Further investigation in this area is warranted.

Quasi-feedforward compensation (Fig. 12) combines elements of feedforward and predistortion techniques. The incoming signal S modulates two matched optical sources, both of which generate an equal amount of distortion . With the aid of the reference signal path, the distortion from optical source "1" is isolated, inverted and brought to the level required to create a compensating signal equal in amplitude and opposite in sign to the distortion generated by optical source "2". Compensation is achieved by predistortion of the modulation signal S to S-Δ. By accurate error leveling and proper control



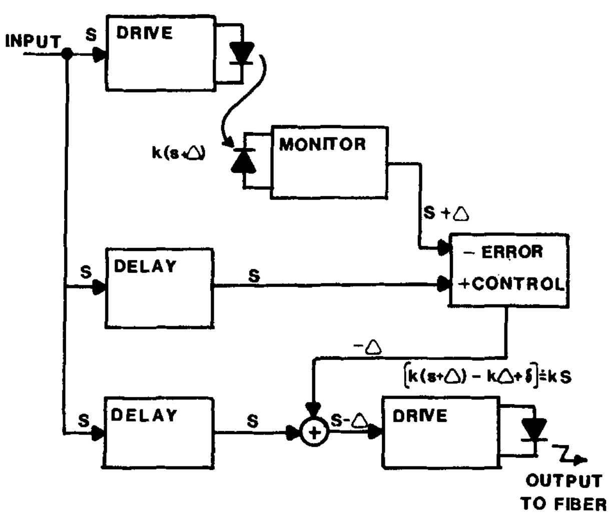
OPTICAL FEEDFORWARD

FIG. 11

of delays, distortion may be cancelled across a wide range of modulation levels. Second order distortions have been reduced by 35 dB and third orders distortions by 20 dB utilizing this technique.

Negative Feedback

Negative feedback (Fig. 13) relies on a photodiode to monitor the optical signal and provide the necessary feedback signal. The amount of distortion compensation depends on the feedback gain. Although the application of negative feedback is straightforward, large bandwidth requirements may create problems at high frequencies. The utmost available feedback bandwidth depends on the component bandwidths and the time delay caused by the finite length of the feedback loop. This suggests the feedback circuits need to be integrated into the laser package for optimum performance.

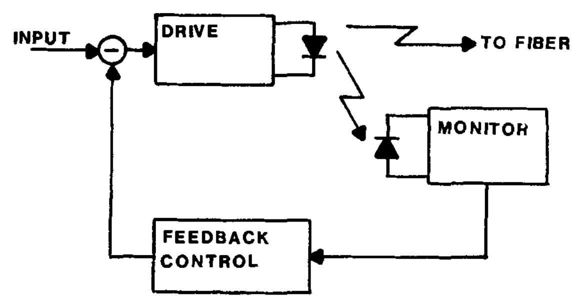


OPTICAL QUASI FEEDFORWARD

FIG. 12

Negative feedback is preferred over feedforward as a linearization approach because it eliminates the requirement to have matched sources and the cost of two optical sources.

Computer simulations have indicated the capability of achieving 20-30dB of feedback distortion reduction over several hundred MHz bandwidths.



NEGATIVE-FEEDBACK COMPENSATION

FIG. 13

SUMMARY

Multi-channel VSB/AM fiber optic links have provided good performance for transmission of (4-8) channels per fiber. To provide CATV performance for 30 or more VSB/AM channels per fiber requires selection of lasers for analog applications and development of linearization circuits for the optical transmitter.

General Optronics has already implemented several analog VSB/AM FO systems which have provided CATV performance for up to 12 TV channels/fiber.

Further investigation into linearization techniques are presently underway to further maximize the performance and the number of VSB/AM channels which can be transmitted on a fiber system.

Fiber Optics Broadband Systems Present and Future

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1. Introduction

In the search for an optimal system for CATV broadband networks ,certain key objectives to be met include:

a)Achievement of highest quality of the delivered video, audio or data
b) Reduced or no sensitivity to EMI, crosstalk etc.
c) No adverse environmental effects
d) Low equipment and installation cost
e) Low system maintenance cost
f) Information security

Degradations in a typical CATV system are caused by the cable itself and the associated repeater amplifiers. They limit the number and the level of the transmitted carriers and the plant bandwidth. As a result, the performance of the signals delivered to the farthest subscriber suffers. Other limitations include :

I) FCC-dictated frequency allocation aimed at minimizing the interference between signals.
II)The system susceptibility to external noise and interference.

For broadband supertrunking signal transportation it is imperative to be able to deliver video **transparently** (with virtually no system-added degradation) from geographically separated locations to distant (tens of miles away) head-ends. Typical CATV delivery systems having repeater amplifiers every 1/2 mile cannot achieve this goal.

Some of approaches that can be used to achieve higher quality of delivered signals include use of microwave links, different modulation over a coaxial cable ,or both. One of the earliest approaches used by Catel was to use FM instead of VSB-AM over coaxial cable or microwave links, with excellent performance results achieved.

With the rapid development of cost - effective optical sources and fiber technology, an attractive and economical approach to the delivery of high quality signals would be to use fiber optics links as transmission media. It would allow the system designer to take full advantage of the wide bandwidth available and of the insensitivity to EMI. Also, no repeaters are required for fiber links less than 25 miles.

2. Why bother with fiber optic - based supertrunks?

The necessity for transparent transmission of video in supertrunks demands that the delivery system performance be extremely high - the performance specifications typically require that the overall video transmission system must meet or exceed the RS250B broadcast industry standards for short or medium haul video transmission .To satisfy these stringent performance requirments **economically,** a careful system design is needed.

The topology of a supertrunk include path lengths extending from several miles to several tents of miles with one or more branches along the way.The number of video channels and their associated audio channels that each supertrunk is required to carry can vary - (depending on the SNR required and the path lengh) - with maximum achievable (using wide deviation FM) being 16.

Coaxial cable - based supertrunks must use booster amplifiers in order to maintain the proper signal levels and CNR in all of the path sections. Although the coaxial cable itself has a wide bandwidth, the booster amplifiers bandwidth is limited to 500 - 600 Mhz. To guarrantee that after a cascade of booster amplifiers the required SNR can be maintained at the farthest receiver location , FM modulation is frequently selected . With the proper modulation index chosen for optimum SNR improvement vs bandwidth expansion , large improvements factors can be achieved at the expense of the number of channels the system can support. If more channels would need to be added either now or in the future, more cables (and booster amplifiers) will be required and the system cost would tend to increase due to the added cost of the cable, additional amplifiers and the maintenance involved.

Microwave supertrunks use Frequency-Division Multiplex methods to combine several FM - modulated video and audio carriers. The combined carriers (typically occupying up to 550 Mhz of bandwidth) are then upconverted and transmitted at microwave frequencies. Although the microwave equipment costs more ,it requires a licenced frequencies to operate and tends to suffer from fading, but it is indispensible whenever it is very difficult or imposible to use a wired system.

Fiber optic supertrunk delivery system offer wide bandwidth, low losses, very small size and weight (and therefore many fibers can be packed into a small space), security of communications and rapidly dropping cost. Because of these properties the system designer can select the frequency plan ,the modulation method and the number of fiber links offering the best and the most economical way of packing the number of required channels into a fiber optic - based supertrunking system.

For long haul paths especially, the benefits derived from a broadband fiber optic supertrunk delivery system can be substantial - both in up front cost as well as in maintenance. Because FM modulation can offer large SNR improvement factors and the bandwidth expansion associated with it can easily and inexpensively be accomodated in a fiber optic link, a broadband fiber optic supertrunk can carry a large number of high quality video channels inexpensively.

3. Elements of an optical fiber broadband system

A block diagram shown in Fig. 1 illustrates the basic components of a fiber optic - based supertrunk. It consists of an RF and a fiber optic subsystem portions and a broadband single mode fiber serving as the transmission medium.

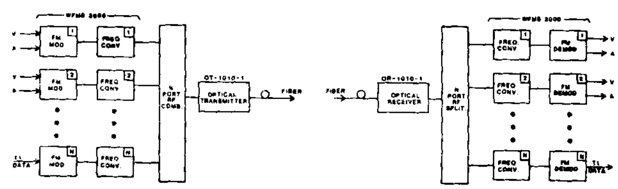

Figure 1: Block diagram of a fiber optic supertrunking system

4. Fiber medium

An optical fiber is a dielectric waveguide that operates at optical frequencies. It is cylindrical in form and it confines the electromagnetic energy in the form of light to within its surfaces and guides the light in a direction parallel to its axis.

One of the principle optical fiber characteristics is its attenuation as a function of wavelength.Three bands which have minimum attenuation to light signals are shown in Figure 2.

Early applications have made exclusive utilization of the 800-900 nm wavelength band , since the fibers made at this time exibited a local minimum in the attenuation curve and optical sources and photodetectors operating at these wavelenghts were available.

Improvement in the fabrications processes by

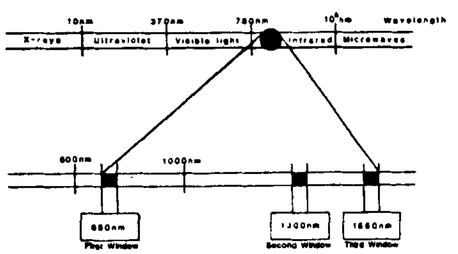

Figure 2: Fiber optics communications windows

the fiber manufacturers have permitted them to produce optical waveguides with very low losses in the 1100 - 1600 nm region (this spectral wavelength band is referred to as the long-wavelength region). Since the 1300 nm wavelength region presents minimum signal dispersion to signals in pure silica fibers, it has been widely used as the choice wavelength in present day uses of single mode fibers

The transmission properties of an optical waveguide are dictated by its structural characteristics. The structure basically establishes the information-carring capacity of the fiber and its environmental response.

The most widely accepted structure of an optical waveguide is the single solid dielectric cylinder of radius a and index of refraction n_1 - this is known as the core of the fiber. The core is surrounded by a solid dielectric known as the cladding having a refractive index n_2 which is less than n_1 .The cladding facilitates the total light reflection into the core, it adds mechanical strength to the fiber and protects the core from surface contaminants.

Variation in the material composition of the core give rise to two commonly used fiber types shown in Figure 3.

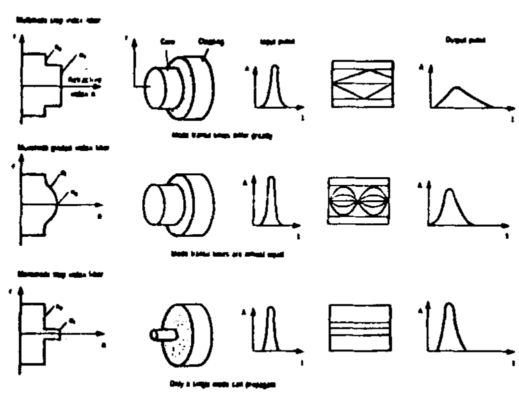

Figure 3: Fiber types and mode of operations

In the first case the refractive index of the core is uniform and undergoes an abrubt change (step) at the cladding boundary. This is called a step-index fiber. In the second case the core index is made to vary as a function of the radial distance from the center - this is known as a graded-index fiber.

Both the step and the graded-index fibers can further be divided into single mode and multi-mode classes. As the name implies, a single mode sustains only one mode of wave propagation, whereas multimode fibers support many hundreds of modes.

Since multimode fibers have larger core radius, it is easier to launch optical power into the fiber and to interconnect similar fibers. Another advantage is that light can be launched into the multimode fiber using a LED source.

A disadvantage of multimode fibers is that they suffer from internal dispertion. When an optical pulse is launched into the fiber, the optical power in the pulse is distributed over most or all modes of propagations that the fiber supports. Each of the modes propagating through the fiber travels at a slightly different velocity, causing the signal to spread out in time as it travels along the fiber. This is called modal dispersion.

A measure of the information capacity of an optical waveguide is usuallly specified by the bandwidth-distance product in Ghz . Km.

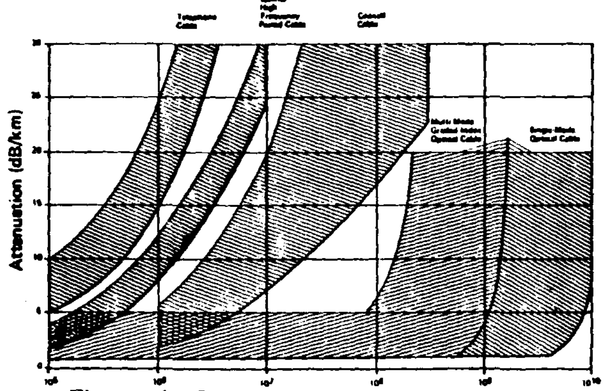

Figure 4: Frequency characteristics of different media

Since intermodal dispersion effects are not present in single mode fibers, they can have very wide signal bandwidths transmission capabilities.

Figure 4 compares the attenuation vs frequency of various media (telephine cable, coaxial cables and different fiber types). It can be determined from the curves that single mode fiber bandwidths exceed 10 Ghz . Km.

5. Optical sources

The principle light sources used for fiber optic communication applications are the laser diodes and light-emmiting diodes (LED). These devices are suitable for fiber transmission systems because they haves adequate output power for wide range of applications, they can be directly modulated by varying the input current to the device, they can have high efficiency and their dimentional characteristics are compatible with those of the optical fiber.

A major difference between LED's and laser diodes is that the optical output from an LED is incoherent , whereas the laser diode output is coherent.

Laser sources generate the optical energy in an optical cavity. The resulting output is highly monochromatic (single wavelength) and the light beam is very directional. (the output has high spatial and temporal coherency). The emmision spectrum of the lasers is narrow (typically 4 nm), they have modulation capabilities of up to 1 Ghz and their radiance is high (1 - 2 mw coupled into the fiber).

In an incoherent LED source no optical cavity exists for wavelength selectivity and the output radiation has a broad spectral width (typically 50 nm) and has modulation capabilities up to 200 Mhz.. In addition, the incoherent optical energy is emitted into the hemisphere according to a cosine power distribution and has a large beam divergence.

The spatially directed coherent optical output from a laser diode can be coupled into either a single mode or multimode fibers. However, sufficiently large incoherent optical power for it to be useful (10 - 50 μw) from an LED can only be coupled into multimode fibers.

Figure 5 shows the effect of resultant fiber optic plant bandwidth as a function of the light source spectral width. It can be seen that very wide bandwidths are possible if a precise control over the optical source spectral width and wavelength can be effectively accomplished.

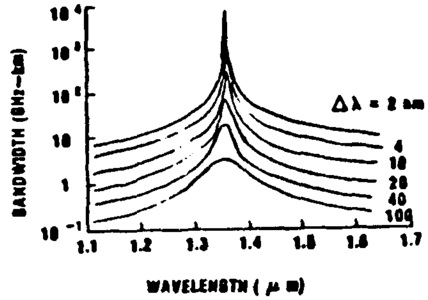

Figure 5: Effect of source spectral width and wavelength deviations

An important factor to consider in the application of laser diodes is the temperature dependence of the threshold current as shown in Figure 6. Consequently, if a constant optical power and undistorted signal outputs is to be

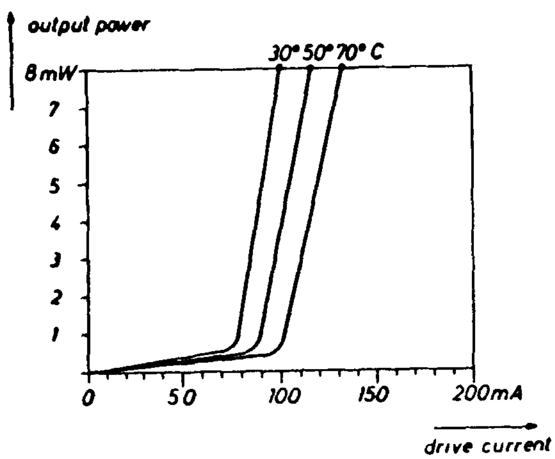

Figure 6: Effects of temperature on lasers

maintained with time, it is necessary to use precise dc bias and temperature control techniques.

For broadband supertrunking applications, intensity modulation of the laser diodes is carried out by making the its drive current above threshold vary about the bias point in proportion to the modulation signal. as shown in Figure 7.

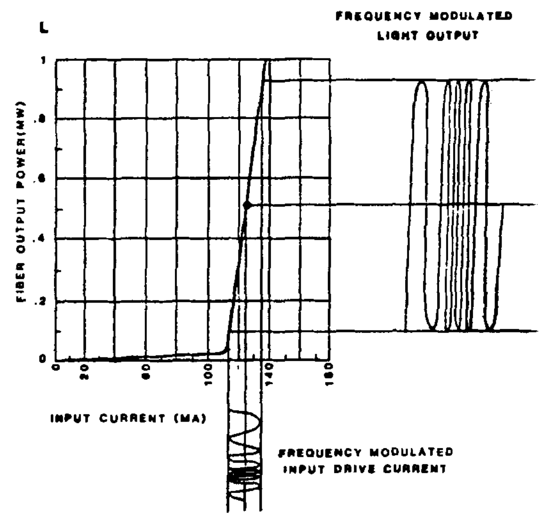

Figure 7: Linear modulation of a laser diode

A requirment for this modulation scheme is that a linear region exist between the light output and the current input. Signal degradations resulting from the nonlenearities in the transfer characteristic of the laser diodes make the implementation of the analog intensity modulation susseptible to both intermodulation and cross-modulation effects if not accounted for.

Methods of compensation for the nonlenearity of the optical sources include different linearization techniques (complementary distortion, negative feedback quasi-feedforword compensations) or use of modulation schemes less sensitive to those distortions such as PPM or wide deviation FM.

6. Optical detectors.

At the receiving end of an optical transmission line there must be a receiving device which interprets the information contained in the optical signal. The first element of this receiver is a photodetector.

Of the semiconductor-based photodetectors, the photodiode is used almost exclusively for fiber optic systems. The two types of photodiodes used are the PIN photodetector and the avalanche photodiode (APD).

The PIN photodiode generates electrical surrent in response to incident light. Two important characteristics of a photodiode are its quantum efficiency and its response speed.

The quantum efficiency η can be defined as

$$\eta = \frac{Ip/q}{Po/h\nu} \qquad (1)$$

Here Ip is the average photocurrent generated by a steady-state average optical power Po incident on the photodetector, q is the electron charge and $h\nu$ is the photon energy.

In practice, 100 photons will create between 30 and 95 electron-hole pairs ,thus yeilding quantum efficiency ranging from 30 to 95%.

The performance of a photodiode is often characterized by its responsivity R. This is related to the quantum efficiency by

$$R = \frac{Ip}{Po} \qquad (2)$$

This parameter is quite useful since it specifies the photocurrent generated per unit optical power.

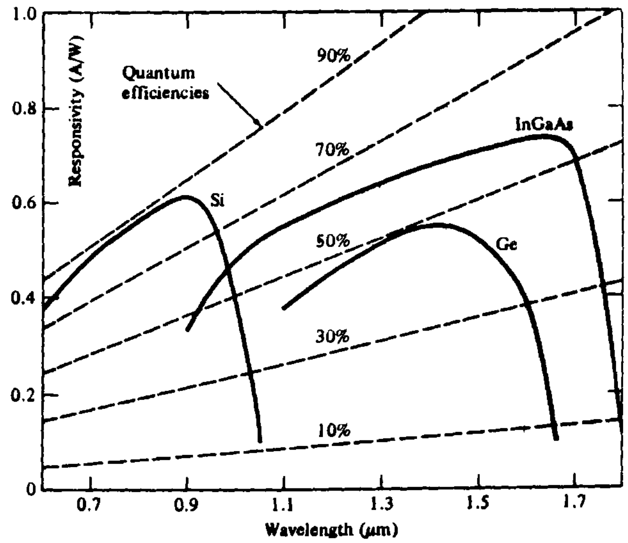

Figure 8: Responsivities of typical detectors

Typical PIN photodiode responsivities as a function of the wavelength can be seen in Figure 8.

PIN diodes are simple to use and they have low dark currents. However, the PIN diode receivers have low sensitivity and low dynamic range too.

The Avalanche photodiode (APD) internally multiplies the primary photocurrent before it enters the input circuitry of the following amplifier.This increases the receiver sensitivitiy since the photocurrent is multiplied before encountering the thermal noise associated with the receiver circuitry. In order for current multiplication to take place, the photogenerated carriers must traverse a region where very high elelctric field is present. The photogenerated electrons can now gain enough energy to ionize forward bound electrons before colliding with them. The newly created carriers are also accelerated by the high electric field ,thus gaining enough energy to cause further impact ionization. This phenomena is the avalanche effect.

The APD's require high bias voltages (Si - 300V, Ge - 30 V), the multiplication factors are statistical and can be temperature-dependent. APD devices also tend to have high dark currents and excess noise for long wavelength devices. However, they permit the receiver to have high sensitivity and dynamic range.

7. Carrier-to-noise ratio of optical photodetector output signals.

As shown in Figure 7, analog modulation of the laser diodes is used for broadband supertrunking applications.

In this scheme the time-varying electrical signal s(t) is used to modulate directly the laser diode about some bias point I_b .

The transmitted optical power P(t) is therefore of the form

$$P(t) = P_t [1+m.s(t)] \qquad (3)$$

Where P_t is the DC optical power, s(t) represents the combined analog FM signals and m is the modulation index defined as

$$m = \frac{\Delta I}{I_b - I_{th}} \qquad (4)$$

Where ΔI is the peak modulating signal and I_{th} is the threshold current

At the receiving end the photocurrent generated by the intensity modulated optical signal is given by

$$i_s(t) = Ip.G[1+m.s(t)] \qquad (5)$$

Where G is the photodetector gain and Ip again is the (unmultiplied) photocurrent generated.

If s(t) is a sum of N sinosoidally FM modulated signals, then the mean square signal current is

$$\langle i_s(t)\rangle = (1/2).\{G.m.Ip/N\}^2 \qquad (6)$$

It can be shown (Reference 1) that the mean square noise current for a photodiode receiver is the sum of the mean square quantum noise current, the equivalent resistance thermal noise current, the dark noise current and the surface leakage noise current. Therefore, the total mean square noise current $\langle i_N(t)\rangle$ is given by

$$\langle i_N(t)\rangle = 2.q(Ip+I_D).G^2.F(G).B+2.q.I_L.B+ (4.K_b.T.B).F_a/R_{eq} \qquad (7)$$

Where F(G) is the excess photodiode noise factor $= G^x$ $(0<x<1)$, B is the equivalent noise bandwidth of the detector, R_{eq} is the equivalent resistance of the photodetector load and amplifier, F_t is the noise figure of the low noise preamplifier, I_D is the (unmultiplied) dark current, I_L is the surface leakage current, T is the equivalent noise temperature of the preamplifier and K_b is the Boltzman constant.

The carrier-to noise ratio of the FM-modulated analog signals at the output of an optical detector (and before FM demodulation) is given by

$$C/N = \langle i_s(t)\rangle \,/\, \langle i_N(t)\rangle \qquad (8)$$

The term $(4.K_bT.B).F_a/R_{eq}$ represents the circuit noise and the term $2q(Ip+I_D).G^2.F(G).B$ the quantum noise (and dark current) associated with a photodetector.

When an avalanche photodiode is employed at low signal levels, and with low values of G, the circuit noise term dominates. At a fixed low level, as the gain is increased from a low value, the C/N increases with the gain until the quantum noise becomes comparable to the circuit noise. As the gain is increased further, the C/N decreases as $F(G)^{-1}$.Thus, for a given set of operating conditions, there exists an optimum value of the avalanche gain for which the C/N is maximum.Since an avalanche photodiode improves the C/N for small optical signals, it is the preferred photodetector for this situations.

For very large optical signals the quantum noise dominates the receiver noise. In this case the avalanche photodiode can decrease the C/N of the received signals if the gain is not decreased or optical attenuation inserted.

8. FM system performance in fiber optic supertrunks.

If analog FM modulation is used to transmit video, we can define the weighted output SNR_{ow} (Ref.2) of a video-modulated carrier as a function of the input $(C/N)_{IF}$, the modulation index β, the IF bandwidth B_{if}, and the highest baseband video frequency F_m as follows

$$SNR_{ow} = [3.\beta^2 B_{if}/(2.Fm)] + (C/N)_{IF} + W \qquad (9)$$

Where $(C/N)_{IF}$ is measured in the IF bandwidth and

$$B_{if} = 2 \times (\Delta F + Fm) \qquad (10)$$

B_{if} is the Carson's rule bandwidth and the modulation index is

$$\beta = \Delta F/Fm \qquad (11)$$

where ΔF is the peak deviation of the video carrier and W is the video weighting improvement resulting from using preemphasis and deemphasis (3.6db with CCIR-405-1 characteristic) , CCIR noise weighting (11.5db) and P-P/RMS conversion (9db).

Note that ΔF is the peak deviation of the carrier by a **sinusoidal** signal with no preemphasis included.Other deviation definitions, used by equipment manufacturers (including Catel) include **sync tip to peak white deviation** $\Delta F_{st\text{-}pw}$. It can be shown (Ref. 4) that the two deviation definitions are related as follows:

$$\Delta F = \Delta F_{st\text{-}pw} / (2x0.3) \qquad (12)$$

If we refer the noise generated by the receiving equipment to the input, the carrier to noise ratio C/N becomes

$$(C/N)_{IF} = Pr/(kT^0_{eq} x B_{IF}) \qquad (13)$$

Where k is the Boltzman constant, T^0_{eq} is the equivalent noise temperature given by:

$$T^0_{eq} = T^0_0 x(F-1) \qquad (14)$$

in which T^0_0 is the ambient noise temperature (300^0K) and F is the noise figure of the receiver.

To estimate the theoretical achievable performance of a multichannel FM video modulation system, the SNR_{ow} will be calculated with the following assumptions:

Carrier deviations :

$\Delta F_{st\text{-}pw}$ = 4Mhz , 6Mhz

(corresponding to ΔF = 6.67Mhz, 10Mhz)

IF Bandwidth B_{IF} = 30 Mhz, 40Mhz

Worst case NF : 20 db (to account for multichannel operation),

Worst received power : -26 dbm (From the optical link power budget),

Substituting in equations (9) to (14), the SNR_{ow} becomes:

For $F_{st\text{-}pw}$=4Mhz and IF bandwidth=30Mhz, SNR_{ow} = 72 db.

For $F_{st\text{-}pw}$=6Mhz and IF bandwidth=40Mhz, SNR_{ow} = 75 db.

Although not shown in the analysis, it has been demonstrated that FM can reject interference from other sources including adjacent FM channels, intermods, crossmods and any other interference not coherent with the in-channel video. This feature permits the system designer to select the modulation to cost-effectively design high performance multichannel video FM systems over broadband supertrunks

Early multichannel FM video systems using peak carrier deviation of 1.6 Mhz and RF

Table 1: Link budget for 16 channel CATV supertrunk

<u>**Transmitter**</u>	
A. Minimum optical power	-1.5 dbm
B. Connector loss	0.8 db
C. Launch power	-2.3 dbm
<u>**Receiver**</u>	
D. Receiver sensitivity	-26.0 dbm
E. Degradation allowance	1.0 db
F. Minimum Input power	-25.0 dbm
<u>**Fiber link loss**</u>	
G. Fiber loss (0.4 db/km)	16.0 db
H. Splice loss (0.3 db/km)	3.3 db
I. Total link loss	19.3 db
<u>**System**</u>	
J. System gain	22.7 db
K. System margin	3.4 db

bandwidth of 16 Mhz have been used in coaxial cable supertrunks and achieved S/N_0 improvement factors of 10 db over the C/N_i.

9. Link budget for analog FM video fiber optic system

A typical link budget for a 16 channel CATV supertrunk is given in Table 1.

The laser transmitter average optical power is -1.5 dbm. Up to 0.8 db can be lost in the connector, resulting in total launched power of -2.3 dbm.

The receiver sensitivity is defined as the average input power at the receiver which yields a weighted video signal to noise ratio of 60 db with full 12 channel loading. From the previous section and allowing for multichannel derating, this input power is -26 dbm. An allowance of 1 db is allocated to degradations due to temperature and aging, resulting in required input power of -25 dbm.

A fiber link of 25 miles (40 km) , using 0.4 db/km and cable sections each 4 kilometers (slicing loss of 0.3 db/splice, 11 splices) will result in 19.3 db of optical attenuation and will yield a system margin of 3.4 db.

An example of present state-of -the art high performance wide deviation video FM systems delivered by Catel is shown in Figure 9. The system is carrying 12 channels, each 40 Mhz wide with multiple subcarriers at a distance to 40 km and achieving more than 65 db video S/N_0 per channel.

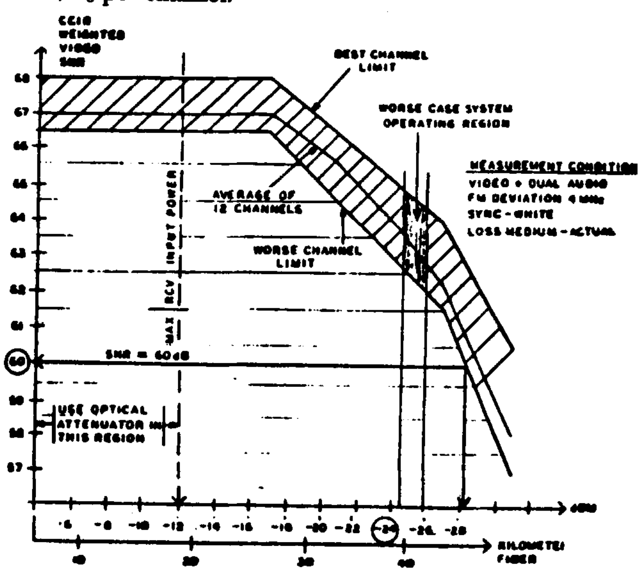

Figure 9: S/N_0 of a 12 channel FM system over a single mode fiber media

The chart shows that worst case of 60 db S/N_0 per channel at 40 km distance can be achieved with worst case received power of -28 dbm and dual audio channels.

10. Flexibility of fiber optic video FM systems

In addition to its high performance, video FM system can also be quite flexible.

For example, the system can be easily adapted to transmit data. Since multi-level PCM data is a video-like signal, PCM multiplexers can be readily interfaced (as shown in Figure 9) with the FM modulator and demodulator permitting high data rate signals to be transmitted over broadband networks.

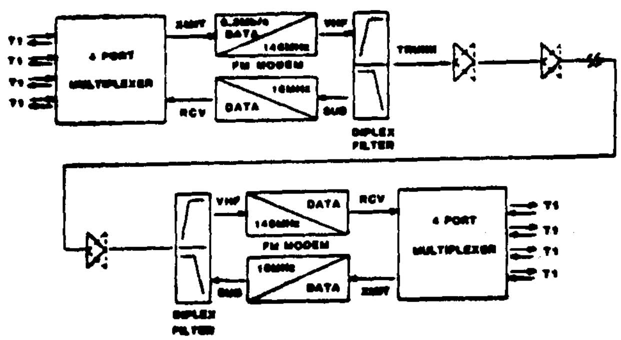

Figure 9: PCM multiplexer interfaced to an FM system

Other type of complex signals (such as BTSC - stereo audio signals) can be conveniently and efficiently carried as subcarriers above the audio subcarrier signals.

11. Future directions of fiber optic systems

What direction of the future for fiber optic systems will take depends mainly on the applications being developed today.

For long haul analog or digital data transmission, 1550 nm seems a natural extension of the present 1300 nm systems.

For the local loop application, it seems that low cost repeaters and fiber optic to coax cable converters would be very desirable. Future developement of the laser transmitters/receivers allowing direct FM modulation of the light beam would help lower the cost and increase the performance. Finally, development of coherent demodulation methods for lightwave signals will extend the range of the transmissions by several orders of magnitude. Lets hope and work on it to bring it closer to reallity!

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Dr. Mircho A.Davidov has received his PH.D. in 1981 from the University of Southern California and Msc and Bsc in 1976 and 1974 respectively from Tel Aviv University.

Dr. Davidov presently directs all of the engineering activities at Catel Telecommunications Inc. developing products in the areas of FM modulators and demodulators for video transmission, VSB-AM modulators and demodulators, FDM-FM fiberoptics transmitters and receivers and frequency translators for LAN data signals.

He was Director of Corporate Research at Oak Industries from 1981-1985. His responsibilities at Oak included developing systems for securing the transmission of high quality video signals. He worked for Honeywell and Brunswick Corporations between 1979-1981 and was responsible for development of RF communication systems for wireless monitoring of energy and control systems and wireless transmission of language translations. Between 1976-1979 he was a consultant to LinCom Corporation in Los Angeles, performing studies in the area of PLL, synchronization and signal processing for satellite communications.Between 1970-1976 he was an engineer for the Israeli Governmental Communication Office,designing circuits for the International Telephone,Telegraph and Telex lines. Between 1967 and 1970 he managed a military communication lab designing communication equipment for the Israeli Defence Forces.

He was an assistant professor at Cal State Northridge in 1981 and teaching assistant between 1974-1979 at Tel Aviv University and the University of Southern California. He speaks 5 languages fluently and is a member of IEEE.

Dr. Davidov has numerous publications and 7 patents pending.

FIELD EXPERIENCE WITH FEEDFORWARD AMPLIFIERS

Mark Adams

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ABSTRACT

The demand for an amplifier that gives a high level of distortion immunity while providing large amounts of amplification has driven the CATV industry over the last three to four years. The introduction of feedforward technology presented a viable solution to this problem. During its infancy, feedforward presented a manufacturing challenge to the CATV suppliers who sought to participate. The development and introduction of the integral feedforward package approximately three years ago however, offered the industry an excellent opportunity to maximize cascade lengths for optimum performance while maintaining superior distortion results.

This paper will look into the areas of reliability on the integral feedforward package from the standpoint of heat transfer, and mean time between failures (MTBF). This paper will also investigate the conditions under which feedforward amplifiers are being used. Areas in this section include the economics of feedforward and how field personnel know that feedforward is offering the distortion improvements they need for their systems to function properly.

INTRODUCTION

Since the idea of feedforward was first conceived nearly twenty years ago, many indepth articles have been published on the mechanisms that make this technology so important in the CATV industry. This paper will not delve deeply into these mechanisms but provide more of an overview into where feedforward is today.

The first application of feedforward presented itself approximately four years ago when two discrete hybrid amplifiers were matched with two delay line circuits and associated tuning circuitry to form a feedforward stage. This unit presented difficulties not only for the equipment manufacturer in both gain and phase matching, but for the cable operator as well.

No longer was the cable operator allowed to luxury of field replacement of hybrid modules. If one section of a feedforward stage failed, the unit had to be returned to the factory for re-alignment.

The introduction of the integral feedforward package around three and a half years ago, however offered many advantages over the discrete approach several of which are:

- Lower die temperature than standard CATV die
- Better temperature tracking of the 4 individual sections of the feedforward stage (i.e. 2 gain blocks, 2 delay lines)
- Better and more controlled loop cancelation
- No fine tuning by the cable operator
- Better and more predictable flatness
- Better and more predictable distortion improvements
- Smaller size

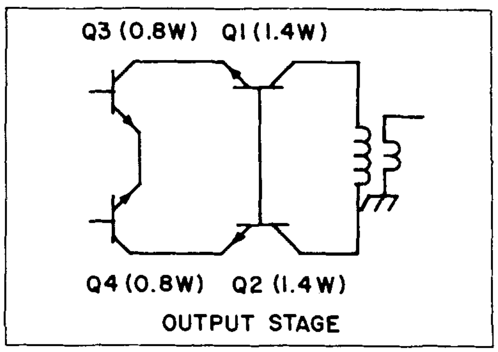
These features and others are what attracted equipment manufacturers to this concept, which revitalized feedforward. In turn, this allowed the cable operators the flexibility to realize the extra distortion headroom many of today's systems demand. However, many questions arose during the introduction of the integral feedforward package and in the feedforward concept in general. These questions consist of such concerns as the thermal properties of both the feedforward package and trunk station; how can feedforward be maximized to obtain the optimum performance versus price and finally; how are feedforward amplifiers checked for proper operation? It is these areas where we will now focus our attention.

FEEDFORWARD GAIN BLOCK AND TRUNK STATION THERMAL CHARACTERISTICS

The integral feedforward package offers a large thermal advantage over the discrete feedforward concept. The entire concept of feedforward operation is based on two RF loop cancelations. These loops consist of both amplitude and phase characteristics and any misalignment may result in reduced distortion cancellation. In the case of discrete feedforward the four individual components (2 gain blocks and 2 delay lines) could all exhibit different thermal expansion over temperature which could cause this misalignment.

The integral feedforward package however, offers thermal compensation to protect the amplitude and phase alignment. Common heat sinking of both amplifiers and delay lines are added insurance that provides the stability needed for proper cancellation.

With the mounting of all the components of a integral feedforward package to a common heatsink, the question of power dissipation of the transistor dice is brought to bear. Figure 1 shows a simplified schematic of the output stage of one of the amplifier gain blocks of a feedforward amplifier.



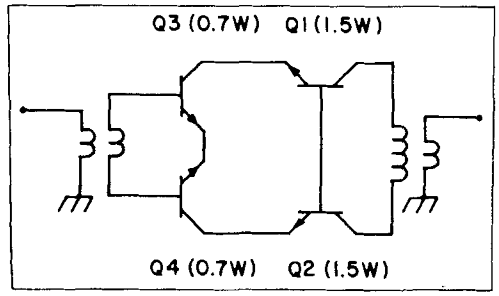
	+65° C Case Temperature	θjC
Q1	+100° C	25° C/W
Q2	+105° C	29° C/W
Q3	+ 89° C	30° C/W
Q4	+ 89° C	30° C/W

	+100° C Case Temperature	θjC
Q1	+140° C	29° C/W
Q2	+142° C	30° C/W
Q3	+124° C	30° C/W
Q4	+128° C	35° C/W

FIGURE 1

POWER DISSIPATION OF THE OUTPUT STAGE TRANSISTORS

As can be seen from the following chart, two different case temperatures were recorded for transistors Q1-Q4. When the maximum case temperature reaches +100° when the die temperature reaches +142°C. In comparison, data was taken on an 18dB push/pull hybrid utilizing the same transistors as the feedforward unit. Figure 2 shows this simplified schematic.



	+65° C Case Temperature	θjC
Q1	+115° C	33° C/W
Q2	+120° C	36° C/W
Q3	+ 89° C	34° C/W
Q4	+ 85° C	28° C/W

	+100° C Case Temperature	θjC
Q1	+151° C	34° C/W
Q2	+159° C	39° C/W
Q3	+125° C	36° C/W
Q4	+120° C	28° C/W

FIGURE 2

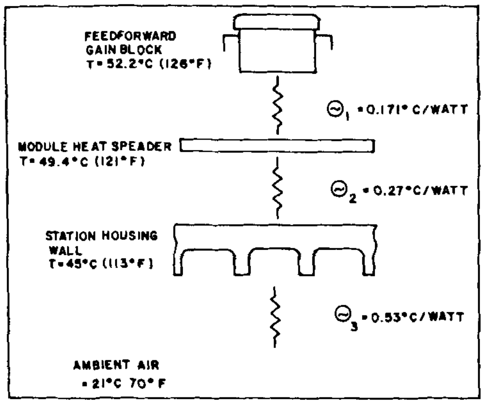
POWER DISSIPATION IN STANDARD CATV AMPLIFIERS

This data shows that on an average the thermal resistance of feedforward is 4° C/W lower in feedforward than in a push/pull package and that with similar case temperatures, feedforward shows a lower die temperature of 14°C over push/pull.

The next consideration that must be given is provide a path to convey the heat produced by the feedforward package to the external air. As with any active components the reliability is based on the average component operating temperature. Since feedforward results in a larger power dissipation than push/pull circuits, equipment manufacturers had to pay special attention to trunk station thermal design.

In the case of a Scientific Atlanta trunk station, the feedforward block is mounted to a heatsink located on the amplifier module. This in turn is mounted to the finned outside station housing wall. Figure 3 shows this mounting configuration

FIGURE 3



With an outside ambient temperature of 21° C (70° F) the feedforward heatsink temperature will be 52.2° C (126° F). If the assumption is made that a constant temperature difference of 31.2° C (56° F) holds between the outside ambient air and the inside of the station then the maximum feedforward heat sink temperature will be 91.2° C (196° F) when the outside ambient temperature reaches 60° C (140° F).

Reliability data accumulated over a three year period shows that with a junction temperature of 150° C the mean time between failures (MTBF) results in a lifetime in excess of 142 years. Since the worst case junction temperature seen in a Scientific-Atlanta trunk housing is far less than 150° C excellent reliability can be expected.

FEEDFORWARD PERFORMANCE VS. PRICE

The introduction of the feedforward technology has opened up a new arena for hardware comparisons where distortion parameters are concerned. The most common distortion limitations are Composite Triple Beat (CTB) and System Noise. The feedforward concept offers improvements in the area of Composite Triple Beat, but shows a slight degradation in noise. A trade-off in distortion parameters can be utilized by the cable operator in two ways: first in a supertrunk application where levels can be run higher to make the noise not a limiting factor and second in a combination of feedforward and push/pull amplifiers which provides a good alternative to Parallel Hybrid Amplifiers at lower costs.

In the case of supertrunk applications the operator can choose three different gain combinations of feedforward trunks allowing for higher operating levels which in turn results in a lower number of actives needed. The following example shows a system price versus end of line performance comparison with three different gain feedforward trunks (22,26 and 30dB) in conjunction with three different cable sizes (0.750,0.875 and 1.000"). The desired end of line performance is 45dB C/N and 57dB CTB.

TABLE 1

Typical Trunk Amplifier Specifications
450 MHz, 62 Channel Loading

Trunk Amplifier	Gain(dB)	CTB(dB)	NF(dB)
22dB PP Trk.	22	81	9.1
28dB PP Trk.	28	82	9.3
22dB FF Trk.	22	99	12.0
26dB FF Trk.	26	99	10.0
30dB FF Trk.	30	99	9.0

Note: Specifications Include All Loses.
All Numbers Are Referenced To 33dBmV.
All Distortion Numbers Within This Paper Are Derived From Table 1.

TABLE 2

22 dB Gain Feedforward (450MHz)

0.750"
Cable Total = 110,880 ft.
Cable Cost = $40,000
FF Trunk Total = 55 (22dB)
FF Trunk Cost = $54,000

0.875"
Cable Total = 110,880 ft.
Cable Cost = $53,000
FF Trunk Total = 50 (22dB)
FF Trunk Cost = $48,000

1.000"
Cable Total = 110,880 ft.
Cable Cost = $77,000

FF Trunk Total = 46 (22dB)
FF Trunk Cost = $43,000

System Cost With 22dB Spacing
0.750" = $ 94,000
0.875" = $101,000
1.000" = $120,000

TABLE 3

26dB Gain Feedforward (450MHz)
0.750"
Cable Total = 110,880 ft.
Cable Cost = $40,000
FF Trunk Total 48 (26dB)
FF Trunk Cost $48,000

0.875"
Cable Total = 110,880 ft.
Cable Cost = $53,000
FF Trunk Total = 42 (26dB)
FF Trunk Cost = $42,000

1.000"
Cable Total = 110,880 ft.
Cable Cost = $77,000
FF Trunk Total = 38 (26dB)
FF Trunk Cost = $38,000

System Cost With 26dB Spacing
0.750" = $ 88,000
0.875" = $ 95,000
1.000" = $115,000

TABLE 4

30dB Gain Feedforward (450MHz)
0.750"
Cable Total = 110,880 ft.
Cable Cost = $40,000
FF Trunk Total 42 (30dB)
FF Trunk Cost = $45,000
0.875"
Cable Total = 110,880 ft.
Cable Cost = $53,000
FF Trunk Total = 36 (30dB)
FF Trunk Cost = $38,000
1.000"
Cable Total = 110,880 ft.
Cable Cost = $77,000
FF Trunk Total = 33 (30dB)
FF Trunk Cost = $35,000

System Cost With 30dB Spacing
0.750" = $ 85,000
0,875" = $ 91,000
1.000" = $112,000

Now that the financial models are in place, Table 5 provides a comparison of the price of feedforward versus end of line performance.

TABLE 5

	C/N(dB)	CTB(dB)	Cost
22dB Spacing			
0.750"	44.7	55.9	$ 94,000
0.875"	45.2	57.0	$101,000
1.000"	45.7	57.6	$120,000
26dB Spacing			
0.750"	44.4	55.4	$ 88,000
0.875"	45.0	56.4	$ 95,000
1,000"	45.4	57.4	$115,000
30dB Spacing			
0.750"	43.0	54.5	$ 85,000
0.875"	43.7	55.9	$ 91,000
1.000"	44.0	56.9	$112,000

As can be seen from the data in order to meet the desired 45dB C/N and 57dB CTB while maintaining the lowest cost possible, the selection of the 22dB gain trunk in combination with the 0.875" cable would be the most appropriate.

Feedforward also provides the cable operator the ability to mix and match this technology with push/pull technology to achieve a attractive economic model while still providing quality end of line performance. This next example shows how a forty percent feedforward and sixty percent push/pull cascade provides a end of line performance of 43dB C/N and 61dB CTB for the total system.

TABLE 6

Feedforward Specifications
C/N = 60.2dB
CTB = 91.0dB
Output = 37dBmV

TABLE 7

Push/Pull Specifications
C/N = 54.9dB
CTB = 85.9dB
Output = 31dBmV

TABLE 8

Cascade Analysis

1. Feedforward Segment (8 Amplifiers)
   CTB(Csc) = (-91.0) + 20 log(8)
   = -72.9
   C/N(Csc) = (-60.2) + 10 log(8)
   = -51.2
2. Push/Pull Segment (13 Amplifiers)
   CTB(Csc) = (-88.0) + 20 log(13)
   = -65.7
   C/N(Csc) = (-54.9) + 10 log(13)
   = -43.8
3. FF(8) And PP(13) Combined
   CTB(Csc) = 20 log($10^{-72.9/20}$ +

$$10^{-65.7/20})$$
$$= 62.5\text{dB}$$
$$C/N(Csc) = 10 \log(10^{-51.2/10} + 10^{-43.8/10})$$
$$= 43.1\text{dB}$$

Table 9 next shows the price of this forty percent feedforward and sixty percent push/pull combination.

TABLE 9

40 Percent FF And 60 Percent PP

1.	Price of FF Amplifier	= $ 500/ea.
	Total Price Of FF	= $4,000
2.	Price Of PP Amplifier	= $ 250/ea.
	Total Price Of PP	= $3,250
3.	Total Price Of 21 Amplifier Cascade	= $7,250

As can be seen from the proceeding data, the mixture of feedforward and push/pull technologies offers the operator quite an arsenal to optimize his cable plant for the best performance versus cost.

FEEDFORWARD TRADE-OFFS

As with any new technology that in introduced trade-offs must sometime occur in order to realize the maximum benefits of that technology. In the case of feedforward the trade-offs are represented in the forms of flatness and in the ability to check the distortion improvement that is offered.

Where flatness is concerned the combination of two gain blocks within the same circuit, each having its own flatness, creates a unit that cannot match the flatness of the push/pull units that preceded it. When this is introduced into a trunk amplifier module a degraded module flatness specification is realized. When a cascade of these units are combined with the other irregularities of a cable plant (i.e. cable,connectors,passives etc.) the operator is hard pressed to meet the N/10 +1 (N = number of amplifiers in cascade) flatness specification that is generally used in the industry for acceptable flatness. To combat this problem, system trimming is often needed in increased numbers over a push/pull amplifier cascade.

The ability to check distortion improvements provided by feedforward can sometimes be cumbersome. Many operators feel that this level of testing is not necessary and in most cases they are right. Others however, like to keep tabs on the operation of both gain blocks within the package to truly know if the distortion improvement they paid extra money for is really there.

In the case of Scientific-Atlanta feedforward amplifiers an external test set can be utilized to check both the error and main amplifiers of the package. This is accomplished by sampling the RF signal from the output test point while providing a 4KHz square wave modulation to the feedforward power supply. If the unit is functioning properly, a pass indicator is illuminated on the test set and a fail indication if not. The test set also allows the operator to turn off and on the +24VDC supply to the individual error and main amplifiers within the package in order to see this modulation effect. With this device on similar tests, there can be no question or not of the feedforward amplifiers operation.

CONCLUSION

While not new, feedforward still confuses many people. Since its early implementation of discrete circuitry, feedforward has made great strides. The integral feedforward package offers excellent performance in terms of:

- Thermal stability and heat transfer
- Reliability
- Ease of operation
- Distortion immunity
- Economics

There are drawbacks to feedforward however, these include:

- Reduced flatness
- Larger power consumption

Overall, feedforward offers the cable operator a nice solution to today's problems faced in terms of economy versus distortion improvements. Weather in supertrunk applications or in a mix and match scheme with push/pull, feedforward has proven that it is a technology here to stay in the CATV industry.

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HIGH FREQUENCY/HIGH DENSITY DESIGN CRITERIA

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UNITED CABLE TELEVISION CORPORATION

ABSTRACT

United Cable Television won the franchise to build the system in Baltimore, Maryland. This plant will have extremely high density in some sections of the city. In addition, the plant was bid and will be built at 550 MHz. These factors combine to present difficult and unique design challenges.

This paper will address the approach used by United's Engineering Department, in conjunction with Jerrold Electronics, to solve these problems. A detailed set of requirements will be presented and the techniques used to meet these requirements will be given. In addition, some of the more unusual aspects of the design criteria will be discussed.

INTRODUCTION

The Baltimore franchise was one of the last of the major urban areas to be bid by CATV operators. City officials there were familiar with the recent history of extravagant franchising proposals followed by operator failure to follow through completely on the promises made. The request for proposals issued by Baltimore specified that the bidders present a realistic and financially viable plan which would represent the plant as it would actually be built.

This requirement necessitated a good deal of engineering research on which to base the bid. This research was undertaken by United's Engineering Department along with the Franchise Department and utilized the services of several outside consultants. Combined with the information provided by the City in it's RFP, the research revealed some very interesting facts which would have to be considered in both the proposal and the actual build. These included items which we had not encountered before or seen only rarely. For instance, a large portion of plant in the center of the system was termed the 'street crossing area'. This referred to the fact that no aerial cable of any kind was permitted to cross the streets from one block to the next. Naturally this necessitated the use of risers and underground plant to maintain continuity. The additional footage, not to mention the extra costs, had to be considered in the design.

Another item of particular concern was the concentration of row homes in this same general geographic area. Row homes in Baltimore represent a significant portion of the total homes passed so they are an important part of the system. However, at least in this particular city, they tend to be clustered only in certain areas. This leads to small geographic areas which contain a very large number of homes and which are right next to areas with a more normal urban density. Methods for locating and identifying the row home areas, as well as how to properly include them in the design, became a topic of prime interest.

Along this same line we found that Baltimore does not have as high a percentage of apartment houses or condominiums (MDU's) as most of the other urban systems we have built. Those that do exist tend to be very large and many are multiple story buildings as opposed to only two or three stories. Again, this leads into a very high number of residences contained within a relatively small area. Methods for designing and building vertical plant have been refined by the Industry over the years, but this was the first time we had tried this type of plant with 550 MHz technology.

Additional items which were uncovered by the research included an open wire fire alarm circuit, built aerially in some areas of the City, which could affect the routing of large portions of plant. Since the City was desirous of minimum street damage they hoped that underground plant could be built using existing conduit. This approach, while it had some cost benefit, would limit routing flexibility, but had to at least be considered. All in all, we were faced with designing and constructing a major new system with some very unusual features.

SYSTEM DEMOGRAPHICS

The City of Baltimore, as franchised, includes some 302,680 homes and has been described as the City of Neighborhoods. While suffering from of the same troubles experienced by many large cities, Baltimore has a fairly strong economy and many innovative social programs. The weather is mild, as a rule, and should have no impact on the cable design.

The City is shaped roughly like the state of Nevada and fronts on the Patapsco River and Chesapeake Bay. It is divided approximately in half by the Jones Falls

Expressway which provides a natural design boundary. Existing utility poles are placed in streets or alleys, resulting in few backyard easements. Street composition varies but some roads consist of several layers of cobblestone, asphalt and thick (up to 12") concrete. Some streets contain buried trolley tracks.

Baltimore can receive eight local off-air signals plus four from Washington D.C. Hilly terrain causes pockets of poor off-air reception. The residents are active television viewers and have a very high percentage of VCR penetration. There is no history of a previous pay TV sevice in the city, such as MDS, so receptivity to cable services is expected to be good. Due to the density of dwelling units and lack of open areas there are few home TVRO antennas. There is some SMATV activity in the area and most of the surrounding County is served by Cable. The city has virtually no room for outward expansion so future growth will probably consist of high density MDU's.

SYSTEM LAYOUT-GENERAL

Due to the general size and shape of the city, and taking the economics of the build into consideration, a decision was made during the franchise process to serve the system from two headend/hub locations. Each hub would feed about one-half of the system. Hub boundaries were defined by the city limits and the Jones Fall Expessway.

The main downtown area, referred to as the Core Area, consists mainly of businesses with a few residences scattered throughout. City governmnent offices are generally located in this area. For this reason subscriber plant will be extended only to those areas with dwelling units. The balance of the Core Area is to be fed with a high-split Institutional Network.

After testing off-air reception and analyzing potential terrestrial interference levels, a site in the East hub was selected for the main headend location. Also considered during the site selection process was the need for a fairly well centralized location within the hub boundary so that trunk cascades could be held to a minimum. Prior to final equipment selection it was felt that maximum trunk cascade would have to be kept within the 20 to 24 amplifier range. A similarly located site for the hub was tentatively identified in the West half of the city. The main headend was designed to be the primary reception facility. It would feed all signals to the hub facility for distibution in that hub area. High levels of microwave interference and poor off-air reception limited the practicality of placing more than one reception facility within the franchise area. The main headend and hub are to be interconnected via a 2-way link. Single mode fiber was eventually selected for this purpose over FM coax or microwave due to its relatively low cost and excellent technical parameters.

Like any large city, Baltimore has areas which are not as wholesome as others. Discussions with the other utilities convinced us that it would be beneficial to exercise caution in the placement of line electronics. To this end, two separate design methods were proposed. The first called for line electronics to be located wherever they actually fell in the design. The second method called for electronics to be placed only on street poles rather than down inside an alley or backyard easement. The second method would have to utilize high gain power-doubling or quadra-powered amplifiers so as not to shorten the overall effective length of the feeder lines. Higher plant and associated powering costs had to be included in considering this approach. The actual design will be a combination of both approaches, rather than exclusively one ot the other, to provide the most efficient system possible for the physical circumstances in each area.

SYSTEM LAYOUT-AERIAL

Initial on-site inspection of the Baltimore area indicated that the vast majority of the plant would be aerial. In fact, about 85% of the plant will be built above ground. As previously noted, most of the existing utility plant is built on poles located on streets or in alleys.

What starts out as a fairly straightforward aerial cable plant is quickly complicated by a number of factors. The first of these, and a major item in the overall design, is a lack of continuity. Many of the alleys are 'L' or 'T' shaped. Feeds into these alleys must come from a lateral street. The lateral runs must be placed at least every other street, and sometimes at every street, in order to pick up all the alleys. The end result is a design which is not as efficient as one with straight alleys.

A second complicating factor is the lack of aerial crossings where the alleys <u>are</u> straight. As previously noted, in one particular part of the city there is a complete restriction against aerial cables connecting the alleys over the street. At the last pole in each alley the cable, if it needs to continue across the street, must be placed in a riser and transverse the road underground. On the other side it is brought up in another riser and continues aerially. This process adds a lot of cable footage to each run for the risers. Each alley will contain only three or four poles. The end poles are anchored, so they must be set in from the street fifteen to twenty feet. Since few of these poles are set on the street there are limited numbers of lateral routes available. Again, design efficiency suffers from a lack of flexibility due to the physical configuration of the existing utility system.

The third complicating factor is the age, condition and size of the existing utility poles and plant. These items directly affect the design and must be included in the overall planning considerations for the system layout. The existing aerial utility plant in Baltimore has been in place for many years. The poles are relatively short and in the kind of shape one would expect for the age of these installations. The designer must keep these factors in mind since there is not a lot of room available on the poles for Cable TV equipment. Rearrangements on this plant are costly and time consuming so great care must be exercised, where possible, in the placement of multiple devices at any one location.

All of these factors occur in the area of the city with the highest percentage of row homes. This means that plant in this area will be passing 600 to 800 homes or more per mile. With a system wide average of

48 poles per mile, each pole location will be feeding 12.5 to 16.6 passings. Some isolated cases exist where 50 or more telephone drops are required at an individual pole. Similar numbers of Cable drops can be anticipated unless techniques besides dedicated taps are used.

SYSTEM LAYOUT-UNDERGROUND

The underground portion of the Baltimore plant is a much smaller part of the overall build but no less complicated than the aerial plant. The City, from the outset of the project, has been concerned about damage to the streets resulting from underground construction. For this reason they originally requested that the winning bidder consider using existing conduit within the franchise area. Attempts were made to determine the exact location and condition of any such conduit systems. Indications were that the City and the Power and Telephone companies all had unused conduit in place. The City's conduit was determined to be unusable for various reasons. The Power company did not have sufficient conduit available and the use of what was there was eliminated by grounding and isolation strictures required by the City and the Phone company. The Telephone company had conduit available but placed so many conditions on installation procedures that this was not a viable alternative. Any of the existing conduit systems, if usable, would have placed severe restrictions on the ability to route the plant efficiently.

The City's concern over street damage arose from the composition of the streets themselves and the methods used to do underground construction up to that time. Many of the streets consist of several layers of various materials including cobblestone, asphalt and concrete. Some streets also contain buried trolley tracks. As information is developed regarding the exact location of the streets with heavy substrata this data will be supplied to the designer so that these roads can be avoided wherever possible. The actual construction will utilize a rocksaw to cut 4" wide slits rather than the 2' foot wide trenches which had always been done before. The City's acceptance of this construction method has freed the designer to route the system as needed for the most efficient design.

EQUIPMENT SELECTION

Cable

The initial decision to build the Baltimore plant at 550 MHz narrowed the equipment considerations to the latest available gear. Cable selection was based on the need for very low attenuation due, to the high design frequency, and the requirement that any cable used be extremely easy to handle because of the physical difficulties expected to be encountered during construction. Other factors taken into account for cable selection included availability and cost. Considerable attention was paid to the cascade lengths which would result given the attenuation of various brands and sizes of cable. Working from scaled maps and using various routing schemes and spacings, the cables were compared to see which would yield the lowest total amplifier usage. The cable which was finally selected combined a low attenuation with good handling characteristics and which was available in sufficient quantities in the time frames needed.

Drop Cable

Drop cable was analyzed in much the same manner as the plant cable. A drop budget was devised which detailed the signal requirements for a typical installation. Each drop had an average length of 150' and had to provide a specified signal level to the converter. Each installation also included the loss for a 2-way splitter. This method of drop budget planning was also used to determine the optimum tap output levels and tilt for feeder design. Drop budget considerations included overall feeder line configuration in the analysis since different tap outputs, as required by the various drop cable attenuations, determined overall line extender requirements and usage levels.

In addition to these considerations, the shielding effectiveness of the different cables was carefully analyzed. This was considered to be an especially important factor in a major urban market such as Baltimore. Naturally, the cost and availability of the cables (and the connectors as well) also entered into the selection decision. The cable type finally chosen was a tri-shield construction RG-6 which used a commonly available connector.

Active Devices

The selection of the system amplifiers was perhaps the most difficult aspect of the entire design criteria. The first step in this area was to decide on an overall design approach. For various reasons, a more or less standard tree and branch approach was selected over switched-star or other configurations. Once this decision was made, several amplifier manufacturers were contacted to obtain detailed specifications of their gear.

As in the case of the cable, a thorough analysis was done on paper of the various brands of equipment. The analysis was done in several stages , the first being to find amplifiers with sufficient gain to make the spacings assumed during the cable selection process.

The second stage consisted of drawing up a list of minimum operating specifications for the system. These included carrier to noise, composite triple beat, cross-modulation, single second order, composite second order and peak to valley response for several different cascades. All of the amplifiers were subjected to a mathematical analysis of various combinations of types (ie. standard, feed-forward, power-doubling, etc), with varying amounts of gain and operating levels to determine performance.

The third step was the preliminary selection of the amplifier supplier. This was followed by physical testing of the amplifiers in environmental chambers to determine actual as opposed to theoretical performance. The results of these tests led to the final refinement of operating levels and system specifications. Included in the final spec callout, in addition to the items noted above, were hum modulation, thermal control limits and powering details.

Coincidental to the development of forward system operating specifications, the return system was also specified. The same steps were used in the analysis and the same specification categories detailed. An operating spec for the Institutional network was also devised based on the subscriber system equipment and cable vendors. In all instances, the derivation of system operating specifications was based on worst case conditions.

Passives

The selection of system passives, including taps, was based primarily on the determination of the amplifier vendor. Taps had already been analyzed in some detail in the drop budgets. The main criteria for passives was insertion loss at the high frequencies and the relative flatness of attenuation over the passband. Again, lab tests were performed on the passives to confirm published specs. The vendor supplying the system actives was also picked to supply the passives.

Power Supplies and Powering

Power plays a much larger role in expanded bandwidth systems than it ever has before. In order to carry a high number of channels with acceptable distortions an operator must use one of the newer amplifier technologies. Feed-forward, power-doubling and dual power-doubling (quadra-power) amplifiers provide the performance edge needed but all of these devices use high amounts of power compared to standard IC technology. In addition, most current franchises call for the inclusion of standby power capacity. Even without this requirement, most operators will at least consider using standby units to protect potential Pay Per View revenue.

In order to determine powering requirements, sample designs were done for several areas, of various densities, using the cable and amplifiers selected for the system. The test designs showed that, at best, a power supply would be required for each three miles of plant. In the very dense areas, and depending on feeder routing as determined by local physical requirements, a power supply per mile may not be uncommon.

A high output power supply is needed for this type of system, so consideration in this category was limited to those supplies capable of 14 or 15 amperes of power. Due to the pole clutter of existing aerial utiliy plant the physical size of the supply also had some bearing on the final decision. The unit which was selected had a fifteen amp rating and held three batteries in the lower portion of the cabinet. This cabinet design was picked over an upper battery design to minimize potential damage to the power supply electronics by leaky batteries.

In powering a system, United will usually load a power supply to 80% of its rated output. This approach is used because it is felt that the supply is most efficient when loaded in this range and also because it allows a little cushion should the future addition of a line extender or two be needed. Powering and power supply placement are both critical items in this particular design for a number of reasons. First, overall costs, both initial capital and ongoing operating expenses, can be significantly reduced by efficient powering of the system. Second, system passives, especially amplifier chassis, are not capable of passing high amperage so extreme care must be used in selecting power supply locations within the plant. The total load of a fifteen amp supply must be nearly evenly split between the two outputs of the power inserter so that high amperage on the line does not damage passive devices. Power supply placement in Baltimore can be complicated by pole limitations. Consequently, the designer must be provided with exremely accurate strand map data to minimize the cases of power supply relocation due to utility limitations.

STRAND MAPPING CONSIDERATIONS

In a design of the size and complexity of Baltimore, the importance of strand mapping cannot be overemphasized. If the success of the overall design hinged on any one factor it would be this one. It has been previously noted that there are many unusual situations and difficulties facing the design of this plant, from extremely high densities to severe routing limitations to prohibitively expensive underground construction areas. The street crossing area requires very accurate pole placement and footage information to be designed.

The strand mapping specification was designed to take into account all of the special problems of the Baltimore design. All footages had to be wheeled where possible or obtained by optical tape measure. Pole heights, where risers were known to be required, were to be indicated on the maps. Accuracy for all measurements had to be within 2%.

All streets, street names and alleys were required to be shown in their entirety. Street widths, as drafted, were to represent the actual right of way. All bodies of water, parks, cemeteries or any other physical features which could hinder routing were to be indicated. Railroads, highways and existing or potential crossings for both were to be shown.

All homes were designated by address with lot lines indicated. Row homes were to be identified and actual house counts placed near the feeder pole. Businesses, churches, schools and government buildings were all to be shown on the maps as well. The drafting specifications for all items were called out in great detail so that the final maps would be consistent and as readable as circumstances allowed. Accuracy in the drafting process, overlaps and border matching for example, was emphasized in the strand mapping spec. Because of the density and clutter, and because the greatest degree of accuracy possible was desired, a scale of 1"=50' was designated, even though this would mean a very large number of maps. Key maps were specified for a 1"=500' scale.

Routing information was to be provided on blue lines of the strand maps. All possible routes, aerial and underground, were to be indicated to allow the greatest degree of flexibility to the designer.

Additionally, certain pole information, such as existing transformers or heavy drop clutter, was to be shown so that redesign could be minimized. The strand mapping effort was begun well in advance of the design so that a solid base of information could be established.

THE DESIGN CRITERIA

Because of the amount of information required to do the Baltimore design correctly and due to the numerous special considerations involved it was decided that a single document was necessary to provide a comprehensive reference for the designer. It was hoped that this criteria would be thorough enough to give the answers to any question which might arise. Using this resource the designer should not encounter time consuming delays when an unusual situation comes up. Further, the depth of the strand mapping information should minimize field verifications which would also delay the design.

Design delay could not be tolerated in this project. Makeready engineering was to be based on design so that only contacted poles needed to be engineered and rearranged. The smooth flow of makeready depends on the accuracy and consistency of the design. The business plan and projected revenue flows are dependent on the scheduled completion of construction and design becomes the pivot on which all of this turns.

The criteria had to incorporate a section which dealt with MDU's and row homes. Initially, row homes were to be treated as individual dwelling units and tapped accordingly but this soon proved to be impractical in some areas. In the high density sections of Baltimore, dedicated tapping would have required upwards of 48 tap ports on each of a high percentage of the poles. The row home design method was changed to treat them similarly to apartment building, which is in essence what they were. Based on the strand map information, each row home was to fed with a single 'hot drop' which would supply enough signal to a splitter box to meet the minimum level requirements for the number of units involved. Maximum level on any one drop was limited to 24 dBmV to help minimize future leakage problems on the high level drops. MDU's were handled in a similar manner. A chart was devised which called out the tap and splitter configurations required for any given number of units. Depending on size, MDU's were fed with either high level drops, feeder line extensions or trunk extensions. With all of this information available, the designer should be able to move through the dense population areas and MDU's with an absolute minimum of delay.

The design criteria also called for specific treatment of the underground design. The designer was directed to add specified amounts of footage to the design to account for the numerous risers in the street crossing area. Additionally, vault, manhole or pedestal locations were detailed as to cable looping requirements so that this extra footage could also be added to the design. Wherever such information was available the strand maps were to indicate which streets contained heavy subsurface materials so that these streets could be avoided if possible. Again, the idea behind this requirement was to minimize design and subsequent construction delays due to uncuttable streets.

The aerial portion of the criteria dealt specifically with the two design approaches. In areas where the second approach was to be used, as designated by the construction manager, all actives were to be located so as to be accessable from a bucket truck. Specific items included in this section, to allow for the second approach, included backfeed limits, the use of and levels for high gain bridgers and line extenders, the use of trunk sized cables in feeder lines to make reaches and power supply placement.

The entire design criteria document, when completed, consisted of some 35 pages divided into 3 sections plus a 16 page Bill of Materials which detailed the allowable equipment for each design approach.

The first section of the criteria dealt with general information and was prepared in the format of a summary of operating levels, performance specifications, cable types and attenuations and some powering data. This section also contained a question and answer portion which supplied the designer with specific answers to certain situations which could arise.

The second section was a detailed listing of all operating levels. These were given for the input and output of each type of amplifier in both the forward and reverse directions. Feedermaker levels, tilts, tap levels and feeder line equalizer specifications were given in this section. Also detailed are the amp equalizer specs. Equalizer values are called out for each type of amplifier and a footage window is provided for each trunk equalizer value. Similarly, a table giving total passive losses for each value of line extender equalizer is given. Through losses for taps and other passives is detailed for both forward and reverse design.

Section two also provides detailed powering assumptions for each station configuration. Cable attenuations are called out for forward and return frequencies. All attenuations used in the criteria are maximum values rather than nominal values. Trunk cable attenuations are increased by a factor of 5% to allow for future aging of the plant. The use of trunk sized cables in approach two feeder design is covered in detail in this section and appropriate losses, without the 5% aging factor added, are given.

The second section then provides the MDU chart for reference. Thermal equalizer usage and relevant technical data is detailed. The second section is completed by inclusion of the published specs for all passive devices.

The third section of the criteria begins with a narrative system description. This details exactly what the system is and the overall configuration to be used. Amplifier types are called out and specified for location within the cascade or design approach. System performance specifications are given for the forward and reverse of both subscriber and institiutional networks. Finally, performance testing and recording requirements are called out for the system. Procedures for the review and approval of strand and design maps are given. A problem resolution procedure, specifying contacts and lines of authority, is the final step in the criteria.

The appendix contains two 8 page BOM's which provide the designer with an exact list of eqipment which is to be used in both design approaches. Pricing is given so that accurate BOM's can be drawn from finished design.

SUMMARY

Faced with a complex and technically advanced system in the city of Baltimore, a design specification which provides the designer with as much information as possible is an absolute necessity. The design criteria which developed as a result and described in this paper meets that requirement. The Baltimore system can be designed in a timely and consistent manner using the methods specified in the criteria. Accurate design of this system will result in a smoother and more cost efficient construction effort. The ultimate result will be a cable system which will function correctly and provide the highest quality signals possible for the life of the plant.

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IMPROVING NTSC IN A CABLE TELEVISION FACILITY

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INTRODUCTION

In the early 1950's, when the parameters of the NTSC color television system were established, the theoretical performance of the system was completely unachievable by the television technology existent at that time. Unstable vacuum tubes, drifting components, bandwidth limitations and hue shifts due to phase errors in the equipment, or the transmission path, were quite objectionable. Constant readjustment of the home receiver to maintain a tolerable, if not perfect, image was a fact of life then. However, the visionary pioneers, who laid down those brilliantly conceived NTSC parameters, gave to color television a remarkable potential that, in today's world of solid-state precision, can finally come into its own.

NTSC's basic advantages are almost self-evident. It is the most efficient use of the available spectrum through the interleaving of the luminance and chrominance signals, and it was and continues to be fully compatible with monochrome television. This inherent simplicity obviously makes it the most cost effective system in use as well. Millions of viewers, who live in areas where NTSC is the Cable TV color standard, can now receive acceptable quality color images in their homes, coming from NTSC transmissions.

However, if one sees an original RGB image on a studio monitor, sitting next to the best home NTSC receiver with a cable feed of the same picture, the difference in image quality is still significant, even if the basic colorimetry is accurate. These differences are due to an approach to the NTSC encoding and decoding process which, in spite of the last decade's improvement, still suffers from intermodulation between the luminance and chrominance components in the video signal, NTSC's major disadvantage. The spectrum overlap, which plagues the present encoding process, produces this unwanted cross color and cross luminance, and in no small measure relies on the psycho-physical characteristics of the human eye/brain interaction to act to diminish these annoying NTSC artifacts by adaptive mental filtering.

This paper describes new approaches to NTSC encoding and decoding which virtually eliminate the effects of cross color and other undesirable artifacts, thus rendering a final NTSC image to the home viewer of near RGB quality. The use of this completely compatible technique will not only render better NTSC images on existent color receivers, but when combined with other currently proposed techniques, will enhance future TV services, and will ultimately rival presently proposed HDTV systems without imposing the higher line rate and wider transmission paths they require.

The goal of this system is to make full use of those technological advances that are available today, namely, more intelligent processing circuitry, and inexpensive memories in line and frame store form. The new hardware can replace the eye/brain filter in removing artifacts, and can even eliminate visual fatigue by providing an NTSC image that looks like RGB. In the opinion of the authors, this can be accomplished without increased demands on our least available resource, RF spectrum bandwidth.

Cable companies employing these new encoding techniques will soon be delivering better NTSC images to their subscribers, regardless of the type of color television receiver they now have.

OBSERVANCE OF SPECIFIED NTSC RULES AT THE HEAD END

There are two general areas where observance of NTSC rules usually apply in varying degrees.

At the transmission level adherence to rules is almost automatic because of the built-in limitations of the equipment. Little complex signal processing is done at the transmitter, and as a result the quality of the distributed NTSC signal may be a function of the few elements that modify it at that site, the final encoder and the hopefully precision sideband filters that shape the selected channels response curve.

However, the NTSC signal path through the studio is a much more convoluted one. Here the signal is routed, switched, recorded, and processed through successive devices in both its analog and often digital form.

Within the studio or production facility, there may not be any regulations for common practices to maintain the NTSC baseband video at its optimum, and each of the devices used to manipulate the signal may indeed contribute some unwanted deterioration to the overall system.

As an example of some studio devices which contribute to this growing problem, consider the proliferation of character generators, computer-graphic systems, color keyers and digital effects systems. Most of these systems, which have entered into the mainstream of program production, use such fast rise time video signal edges that they generate illegal sidebands when they are applied to the chroma channels of an encoder. It is not unusual when watching a cable channel carrying news and data, in the form of scrolling text, to see breakup or tearing due to overmodulation from these hard-edged signals.

The I & Q (or R-Y and B-Y) chroma bandwidth bounds are exceeded, and the resultant signals, full of intermodulation overlap, can never be properly decoded by even the best comb filter decoder in a monitor or receiver. Obviously the careful control of these factors at the source would contribute greatly to the betterment of the NSTC images received by the home viewer.

EXPRESSION OF IMPLIED RULES

Throughout the complex studio chain, the NTSC signal finds itself submitted to a variety of sampling mechanisms which are subject to the Nyquist criteria. The techniques for minimizing these effects have been described by researchers in America, Japan and West Germany. In fact recent articles in the SMPTE Journal, by Dr. Wendlund and his colleagues, and in the IEEE Communications Journal, by Takahiko Fukinuki, et al., give good theoretical analyses of these phenomena.

The answer to the problem lies in the use of very careful multi-dimensional Nyquist pre- and post-filtering. This would result in the elimination of aliasing caused by line scanning, and by the 2:1 interlace in normal television. It would also reduce the motion artifacts and the chroma/luminance spectral overlap that creates visual disturbances in the image when certain kinds of fine details are present.

When NTSC was first developed the cameras of that era were incapable of rendering any useful MTF at luminance frequencies above 4.0 MHz, therefore, the notch filter used in the home receiver to filter out the color subcarrier did little harm to the overall resolution of the image. That is no longer the case, and modern cameras do have useful high frequency output above 6.0 Mhz, therefore, requiring that the upper luminance frequencies be recovered after comb filtering of the chroma signals. Again, this implies adherence to operational practices that retain the full quality of a proper NTSC signal.

COMPATIBLE RULE CHANGES

It is also possible to make some beneficial changes to the current NTSC rules which still maintain full compatibility with the present system. To make this clear, "compatible" in this case means full forward and reverse compatibility with no degradation if any new element is introduced in the chain, either at the transmitter or the receiver. The level of any improvement will increase if both sides are implemented.

It has already been adequately demonstrated that the use of 2H comb filtering in the encoder and decoder, in combination with some additional adaptive logic circuitry, can indeed produce a near RGB result in the home receiver. That technique has by no means been taken to its limit, and laboratory tests have shown that further improvements can be made with more than 2H combs. So far, the 2H comb is a cost effective means of getting a considerable improvement in NTSC images without burdening the system with greater complexity and cost.

Another proposal is to replace the currently differing I and Q bandwidths with equal 1.0 MHz baseband channels, which would also have very sharp roll off characteristics incorporated into them. This would definitely help to eliminate the effect of cross luminance in the decoder. If at the same time the transitional characteristics of the chroma channel were very strictly defined, there would be a significant reduction in those well known deficiencies of NTSC, chroma ringing, chroma/ luminance delay and the fast rise time pulse handling. These high quality NTSC signals, radiated by a transmitter that has good filters, will certainly make a visible improvement even on old notch filter receivers.

Speaking of transmission, that may very well be the weak link in the chain, and we would propose very strict performance regulations for filters used in the video channel of the transmitter. This would assure that what is actually transmitted on the air would not negate all of the good practices adopted in the studio to improve NTSC image quality.

SUMMARY

In summary, it is clearly evident that the implementation of these steps for the originated and trasmitted NTSC signal would greatly improve the quality of the signal, and would greatly simplify the design of more efficient home receivers.

Line doubling at the receiver is made easier and better if the transmitted signal has been properly pre-filtered in the scanning process; sophisticated comb filter decoding, making use of adaptivity in the horizontal and vertical domains, and of chroma bandwidth expansion, is made easier if chroma luminance spectral overlap is not allowed in the transmitter.

In conclusion, it is believed that an NTSC signal, fully compatible with the present standard, can be displayed as a 525 line, 60 Hz progressive scan image. This image will be free of ringing, cross color, cross luminance, and have all appearance of a 7 MHz RGB signal.

These changes in quality, reasonably easy to implement, will give to NTSC a longer lease on life as it was previously expected, and will give to this industry enough time to design, without unnecessary haste, the proper high definition television system of the future.

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IMPROVING POWER SUPPLY EFFICIENCY

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ABSTRACT

Power supply efficiency has become very important to all CATV system operators as electrical power rates continue to rise. Efficiency can be improved in several different areas of a plant powering system. Even a 2 to 3 percent increase in efficiency can add up to a surprising cost savings over time.

INTRODUCTION

New developments in CATV technology have brought increased capabilities and transmission quality at the expense of increased power consumption. As cable systems become more "power hungry", system designers have focused more attention on powering efficiency. Efficiency has to be considered carefully in new plant design as well as system rebuilds. This paper will suggest some ways to improve the efficiency of the typical CATV powering system.

AC POWER SUPPLIES

By far the majority of North American CATV systems use AC power supplies that provide conditioned and regulated 60 volt ac power to the active devices in the system. Most of these power supplies are based on the "Ferroresonant" power transformer topology. The Ferroresonant transformer design has been around for over 40 years and has proven to be extremely reliable as well as providing other important advantages.

A ferroresonant transformer differs from a regular linear power transformer in many ways. First of all, a ferroresonant transformer is a two component magnetic regulator.

It consists of a specially designed lamination core with separate windows for the primary and secondary coils. Unlike a linear transformer, a ferroresonant core is designed to go into magnetic saturation. The second component is an AC capacitor that with a resonant winding on the transformer secondary forms a resonant tank circuit. (See fig. 1.)

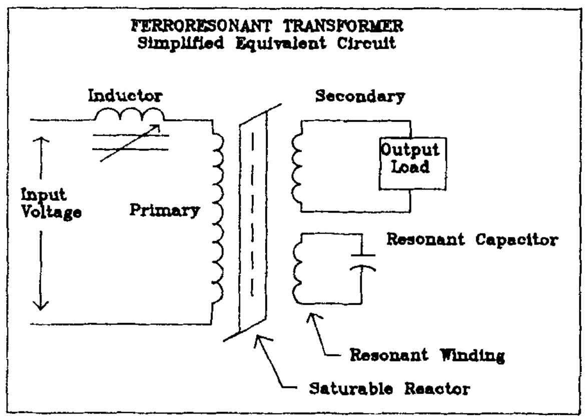

FIG. 1

A ferroresonant transformer can be easily understood as an LC low pass filter with a corner just above the line frequency (60 hz) followed by a roll off of 40 dB per decade.

In typical operation, voltage present at the primary will excite the main magnetic flux path which in turn excites the secondary winding which is tuned by the resonant capacitor (usually several microfarads). As the secondary goes into resonance high circulating currents flow in the resonant tank circuit which drive the main flux path into saturation. Once the core is in saturation, normal voltage fluctuations at the primary will not pass through and increase the secondary voltage. Any decrease in primary voltage will not affect the secondary voltage as long as the core remains in saturation. This transformer once it is saturated will provide line regulation over a wide range of input voltage, (usually 80 to 140 VAC).

Load regulation is provided by the use of a shunt magnetic path which has air gaps between it and the main magnetic path which is operating in saturation. The air gaps limit the flux in the shunt path preventing saturation in the shunt portion of the magnetic circuit providing a good linear response. If the output load current in the secondary is increased, the resonant circuit "Q" drops and the circulating currents will then decrease. The shunt flux also will decrease allowing an increase in the main magnetic flux path, that in turn transfers more energy from primary to secondary thus compensating for the increase in the secondary load current. This provides load regulation.

The transformer regulation can be improved by using a "compensation coil", otherwise known as a "buck" winding. This is a part of the primary winding that is physically wound on the secondary to aid in regulation. This design is not optimal for CATV systems because it compromizes the isolation and noise attenuation capabilities of the transformer.

Most ferroresonant transformers can withstand a dead short on the secondary for an extended time without damage due to the foldback characteristics of the shunt circuit. (see Fig. 2.)

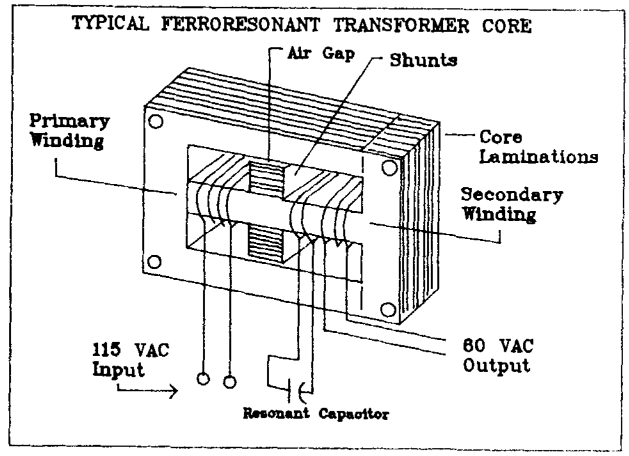

FIG. 2

It is important to note that the output waveform of a ferroresonant transformer is clipped, it is often described as a "Quasi-square wave". This wave shape is caused by the transformer core saturation and is desirable for CATV systems because of its lower peak voltage which is easier to rectify and filter in the Dc power supplies in trunk stations and other active devices. This output wave shape can be corrected by an additional winding on the transformer which is appropriately called a "correction coil". The output can be corrected with this winding to become a sine wave with low distortion if desired, but is rarely used in CATV plant powering.

The ferroresonant design offers both line and load regulation which provide protection from voltage surges and sags at the primary. The majority of utility power problems are characterized by dropouts or sags caused by momentary line faults, utility switching operations and heavy equipment such as motors and compressors coming on line in close proximity to the CATV AC power supply. A ferroresonant transformer can provide output power for as long as a one cycle dropout (depending on load). This is known as hold-up time which is provided by the energy stored in the resonant tank circuit described previously.

The most important advantage of ferroresonant based power supplies is their outstanding isolation characteristics. The primary and secondary windings are in separate window areas of the core and thus are physically isolated from each other. This minimizes capacitive coupling from primary to secondary and greatly reduces the possibility of voltage spikes and noise being coupled to the secondary and then to the load. It is common in certain parts of the country to see utility powerline transient voltage spikes up to 1,500 volts regularly and spikes up to 5,000 volts occasionally. The ferroresonant transformer does an excellent job at attenuating these spikes and protecting the output load from damage. Typical noise attenuation is 120 dB for common mode noise referenced to ground, and over 60 dB for transverse mode (line to neutral).

NO FREE LUNCHES

The ferroresonant based power supply is a natural for the CATV application, but it does have its disadvantages. As the saying goes "there are no free lunches", you can't get something for nothing. Ferroresonant transformers are not as efficient as linear power transformers. The textbook maximum efficiency for a Ferroresonant is about 94 % but typical designs run as low as 80%. There are two main causes for inefficiency in a ferroresonant transformer; core loss and I^2R drop (otherwise known as copper losses). The losses can be directly related to temperature, as the operating temperature increases, so will the losses. Copper has a positive temperature coefficient, its resistance will increase about .4 % per degree C. Core losses caused by eddy currents will increase over temperature and make up the majority of the loss in the transformer. There is another variable that relates to efficiency and that is the regulation tolerance. A ferroresonant transformer can be designed to meet less than a +/- 1 percent combined line and load regulation specification, but it will be very inefficient due to the large circulating volt-amps in the resonant tank circuit. The tank circuit will have to be running at a high energy level to maintain the tight regulation over line and load changes.

Efficiency will be the best at close to full load and low input line because most of the circulating VA is being delivered to the output load. Efficiency will be worse at nominal or high input line with less than full output load, again, because of the large amount of circulating VA in the tank circuit that is not being used by the load and is subject to core loss and winding resistance losses. This scenario is common in new CATV installations where system designers often load the power supply to 75% capacity worst case. This offers a safety factor and room for future expansion, but it does so at the expense of efficiency (see figure 3.) A ferroresonant transformer will always run more efficient if operated close to its rated full load output current.

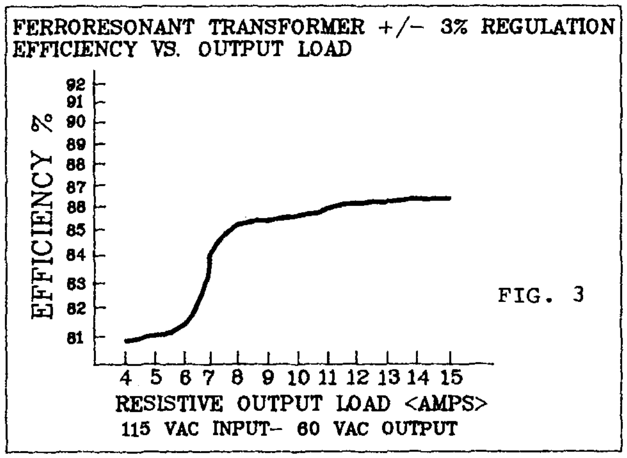

FIG. 3

DC SWITCHING POWER SUPPLIES

There is a relatively new family of DC power supplies known as "switching power supplies". There are several topologies including boost, buck, forward, and flyback converters whose main job is DC to DC conversion, either step up or step down. Switchers offer much higher efficiency than the older linear "series pass" regulators. Some switchers can run up to 98% efficient compared to 40 to 50% for an equivalent linear supply. Switching power supplies are being used in CATV trunk stations and other actives because of the massive gain in efficiency as well as reduction in physical size (watts per square inch) and weight. The switching power supplies in the full featured trunk stations offered by the major manufacturers have efficiencies close to 90% and more importantly, have an input line regulation tolerance of up to +/-30% ! Their variable duty cycle design offers this wide line regulation range while maintaining very high efficiency.

This takes us back to the ferroresonant AC power supplies, why maintain +/- 3% output regulation (which is the typical specification) when the active devices to be powered by the AC supply can accept up to a +/- 30% variation in input voltage ?

It can be argued that I^2R drops in the cable spans between power supplies can eat up some of the power so that a lower voltage is presented to each Active further down the span. But with systems using 450 MHz, the trunk amplifiers are closer together and are linked by larger cable which has lower 60 hz loop resistance, so that the 60 VAC I^2R drop is lower from the power supplies to the loads.

If the regulation specifications for the AC power supplies were relaxed to a plus or minus five percent or even a plus or minus 7 percent, transformers could be designed with less energy in the tank circuit and thus higher overall operating efficiency (see figure 4.) This could mean an increase in efficiency of up to 7 percent for some power supplies.

This wider regulation tolerance would not cause any ill effect to the DC power supplies in the active devices in the system and would provide increased efficiency.

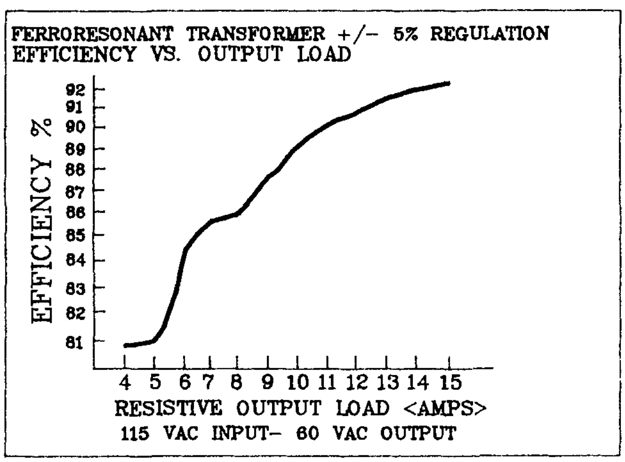

FIG. 4

If a reduction of 7 percent power consumption was multiplied by the number of power supplies in the system, multiplied by the Kilowatt Hour rate that is being charged by the local utility. It will become obvious the magnitude of cost savings that is possible. This can be achieved by using the Ferroresonant transformer power supply primarily for its extremely effective protection against transients, utility line noise, dropouts, lightning strikes and as a pre-regulator for the DC power supplies in the Active devices in the CATV plant.

The other obvious advantage to the Ferroresonant AC power supply is the battery back-up standby mode for protection from complete utility power outages. The ferroresonant transformer uses hold-up time as mentioned earlier, to cover the brief interruption in power as the power supply transfers in phase from utility power to battery backed inverter mode. This is so effective that the load sees no interruption in the AC waveform during transfer in either direction. The transformer will also regulate the change in battery voltage as presented to the AC inverter stage from high battery to low battery while maintaining a steady AC output voltage. With the wider regulation tolerance implemented, the transformer efficiency would be higher and thus the standby time would be longer for the same battery pack.

RELATED TOPICS

It is important to note that the load power factor can influence the regulation characteristics of a ferroresonant transformer. Switching power supplies look like a partially capacitive load to any AC power supply. This is due to the large DC filter capacitor after the input rectifier in the switcher. There is also a complex relationship with the effect of the DC output filter capacitor as the switching duty cycle changes with fluctuations in AC input voltage and output load variations.

This makes for a complex load to model accurately. As the load power factor in some systems could approach .7 P.F. (leading) due to the load capacitance. The effect this has on a Ferroresonant power supply is to effectively add the load capacitance in parallel with the resonant tank circuit capacitance, which will de-tune the tank slightly. This will increase the output voltage by a certain amount depending on the actual load power power factor. If the transformer was loaded to the other extreme with a .7 P.F. inductive load (lagging), the tank circuit would be de-tuned in the other direction producing a drop in output voltage.

Thus it is important to remember that in testing a Ferroresonant AC power supply, a purely resistive load or a load using incandescent light bulbs is not necessarily an accurate simulation of the characteristics of the real world loading in the cable plant. Some power supply manufacturers are aware of this phenomenon and design their transformers accordingly.

Another possible way to improve powering efficiency in a CATV system, is to use solid copper center conductor trunk cable instead of the more popular copper clad center conductor cable. It is possible to reduce the cable loop resistance by at least 20%, and this would reduce the I^2R loss in the cable between the power supplies and the loads. A plant designer would have to evaluate the estimated power savings over a certain amortization period versus the initial extra cost of the solid copper center conductor cable.

A related topic to the discussion of efficiency, is the issue of utility billing for CATV plant power usage. The method used varies with each utility and CATV company. Some systems are billed by the nameplate rating of the power supply. It is important to measure the true power in watts that is being used by the input of the power supply as it operates in the field. This will take into account the actual loading of each supply including losses. If XYZ power Co. reads the nameplate output rating of the power supply as 14 amps and then multiplies that by 115 volts supplied to the power supply input, they would charge the cable system operator for 1,610 VA or 1.61 KWH multiplied by 24 hours for a daily consumption of 38.64 Kilowatt hours multiplied by the going rate of, lets say $.15 per KWH. That power supply would cost the CATV operator $5.79 a day to operate.

This example assumes the mistaken conclusion that the power supply is actually fully loaded to the 14 amp rating (are most power supplies in the system fully loaded ?) XYZ Power co. also made an error in figuring the true power usage of the supply. It should be calculated as 14 amps output load multiplied by the output voltage of 60 VAC which is 840 VA output load. Next calculate the power lost due to the inefficiency of the power supply. Assuming, for the sake of example, the power supply was only 84% efficient. 840 VA divided by .84 equals 1,000 VA, This is 1 kilowatt hour. Next multiply by 24 hours to arrive at 24 Kilowatt hours per day. This is the true power usage of the supply per day. To calculate the operating cost, the system operator must multiply the total daily usage by the cost per kilowatt hour charged by the local utility. For example, $.15 per kilowatt hour. This KWH cost multiplied by the 24 KWH per day consumption rate equals the true power supply operating cost of $3.60 per day!

I suggest the use of a clamp-on True Rms wattmeter to measure the input power of each power supply in the system to get an accurate indication of the real power consumption of the system. Instead of the power supply nameplate ratings. Then compare the measured consumption with what you are being billed for by the utility. Some cable systems get billed at a very low commercial rate, in which case measuring each power supply might not be worth the effort. If the system is measured and there is a substantial over billing by the Utility, A system operator might want to consider installing metering at each power supply to ensure a fair billing representing the true power usage of the system. The potential cost savings could pay for the cost of installing the meter hardware in a few months, after which there would be a noticeable reduction in operating costs.

SUMMARY

Efficiency is important to CATV system operators because of the direct relationship to operating costs. Even small improvements made in the system to gain power efficiency can amount to a substantial savings over time, depending on the total system power consumption and the cost of electricity.

A relaxation of Ferroresonant power supply regulation specifications, solid copper center conductor cable, switching power supplies in active devices, and accurate power consumption measurement by the utility are all possible ways to reduce operating costs for a CATV system.

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INNOVATIVE ASPECTS OF A SWITCHED STAR CABLED TELEVISION DISTRIBUTION SYSTEM

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ABSTRACT

A fully integrated 'switched star' cabled television system is described, with particular emphasis on the more innovative aspects of the design.

Among the most important factors affecting the design was the need for complete transparency in the distribution of present and likely future wideband services. This dictated that a combination of Space and Frequency Division Multiplex be adopted on the Primary Distribution network. Consequently a hybrid system of matrix crosspoint switches and frequency agile converters in combination is employed at the switchpoint.

INTRODUCTION

Historically, cable television in the U.K. and Europe, although in existence for over 30 years, has been confined until recently to the relay of national broadcast services.

In recent years a liberalisation of the regulations was anticipated and this encouraged British Cable Services to formulate a broadband cable system which would serve it well into the future. (1)

Drawing on its 50 years of cable experience, and with some foreknowledge of likely licensing requirements, design criteria were laid down.

MAIN SYSTEM DESIGN CRITERIA

Capacity of the system should support initially 30 TV programmes and up to 20 FM radio programmes downstream. There should be capacity for two way data transmission to support interactive services immediately. Furthermore the level of interactivity should be upgradeable without costly changes in hardware.

Licensing requirements also stipulated that at least one upstream vision signal should be possible from each network sector.

Transparency in the distribution system was essential, in order to comply with 'must carry rules' for all present and likely future wideband services such as stereo TV sound, MAC standard signals from Direct Broadcast Satellites (DBS) and possibly High Definition TV (HDTV). These stipulate that broadcast signals must be delivered to the subscribers in such a way that completely normal domestic equipment can be used for their reproduction.

Again, it was considered important that catering for future services should not require major re-engineering of system or hardware.

Modularity should be adopted as the design principle for all hardware, allowing low cost start-up options and progressive investment by the cable operator linked to subscriber base growth. Additional benefits would accrue in maintainability and upgradability with minimum downtime.

Flexibility in programme tiering and pay per view facilities was very important to accommodate easily marketing requests for changes.

Simplicity of operation, especially at subscriber level, was paramount. The cost of the subscriber equipment was also to be minimal.

Reliability was essential to ensure maximum subscriber satisfaction especially when interactive services are operated.

Topology In addition to the above criteria it was widely anticipated that licensing requirements would favour switched star secondary distribution network topology as a means of promoting interactive services. For the reader who is not aware of the claimed advantages of switched star networks these can be summarised as follows:-

- The number of TV programmes available to subscribers can be increased indefinitely without disturbance to the secondary network.

- Conditional access is achieved without expensive, usually addressable, equipment in the home, and it is impossible for the subscriber to gain unauthorized access. The alternative of scrambling, with its attendant drawbacks, loss of signal fidelity and incapacity to accept a change of TV standards, is avoided.

- The delay on pay-per-view access is minimal compared with the delays in tree and branch networks as illustrated in (2).

- A large effective bandwidth for communications is available, due to data processing in switching points. The message queueing which occurs with tree and branch systems due to node switching is avoided.

- Text signals can be inserted in switching points.

- Neighbours cannot eavesdrop on private communications.

- The topology lends itself to adoption of optical fibre transmission.

Satisfying the above design criteria in the system design has resulted in a switched star cabled distribution system with the parameters listed below.

SYSTEM 8 SWITCHED STAR SYSTEM PARAMETERS

System Reach The trunk network serves an area up to 30 km in diameter. A single head end can provide service to up to 192,000 outlets. Larger conurbations can be covered using super trunk connections to additional hub sites. Trunk amplifiers are spaced at intervals of up to 440m.

Channel Capacity The system delivers up to 30 TV channels; all of which can accommodate MAC signals in any form, 6 of which can be extended to carry HDTV transmissions. 1300 MHz of trunk bandwidth is available.

Switching Points Up to 512 switching points can be served by one front end processor. Each switching point serves up to 95 outlets - up to three independent outlets from a single subscriber drop cable. Max drop length is 300m. Switching point processors are bus-based for ease of expansion with software easily downloaded from the head end.

Frequency Bands and Data Rates Distribution frequency bands and data rates have been planned for primary (trunk and sub-trunk) and secondary (subscriber drop) networks as follows:

TV channels

Primary: 50-200MHz on each of six cables taking due account of prohibited frequencies. Only 5 clear channels required. Channel spacing >19 MHz.

Secondary: UHF, three frequencies in the band 470-860MHz.

Upstream Video Channel One upstream channel per sector.

Primary: On 39.5 MHz carrier on cable 7.
Secondary: On 39.5 MHz.

FM channels

Primary: Bank II 88-108MHz on cable 7

Secondary: Band II 88-108MHz

Up to 20 stero channels can be delivered.

Upstream Interactive Data Channels
Primary: 250 Kb/s in the band 2-4 MHz on up to 4 cables per sector.
Secondary: In the band 0-10kHz

Downstream Interactive Data Channels
Primary and Secondary: Embodied as teletext inserts on entertainment or dedicated TV channels. Also "in vision" text responses provided from switching points.

Transaction Peaks Peaks of 1,000 transactions per second can be dealt with on a 100,000 home network giving individual responses to subscriber commands and messages.

System Response Time Response to subscriber channel change commands - less than 250ms.

Additional Data Capability Up to 24 Mb/s each way on primary network in the band 4-10 MHz, using TDM and FSK modulation. By data concentration at S.P's, any two way data requirement to outlets can be provided, using modems and the 1-10 MHz available spectrum.

Upstream Video Channel One Upstream channel per sector.

Primary: On 39.5 MHz carrier on cable 7.
Secondary: On 39.5 MHz.

Overall system complies with the performance requirements of BS6513 and the radiation limits of MPT1510/1520.

The following sections give an overview of the overall system but only the more innovative aspects of the design and their interplay with the above design criteria are examined in greater detail.

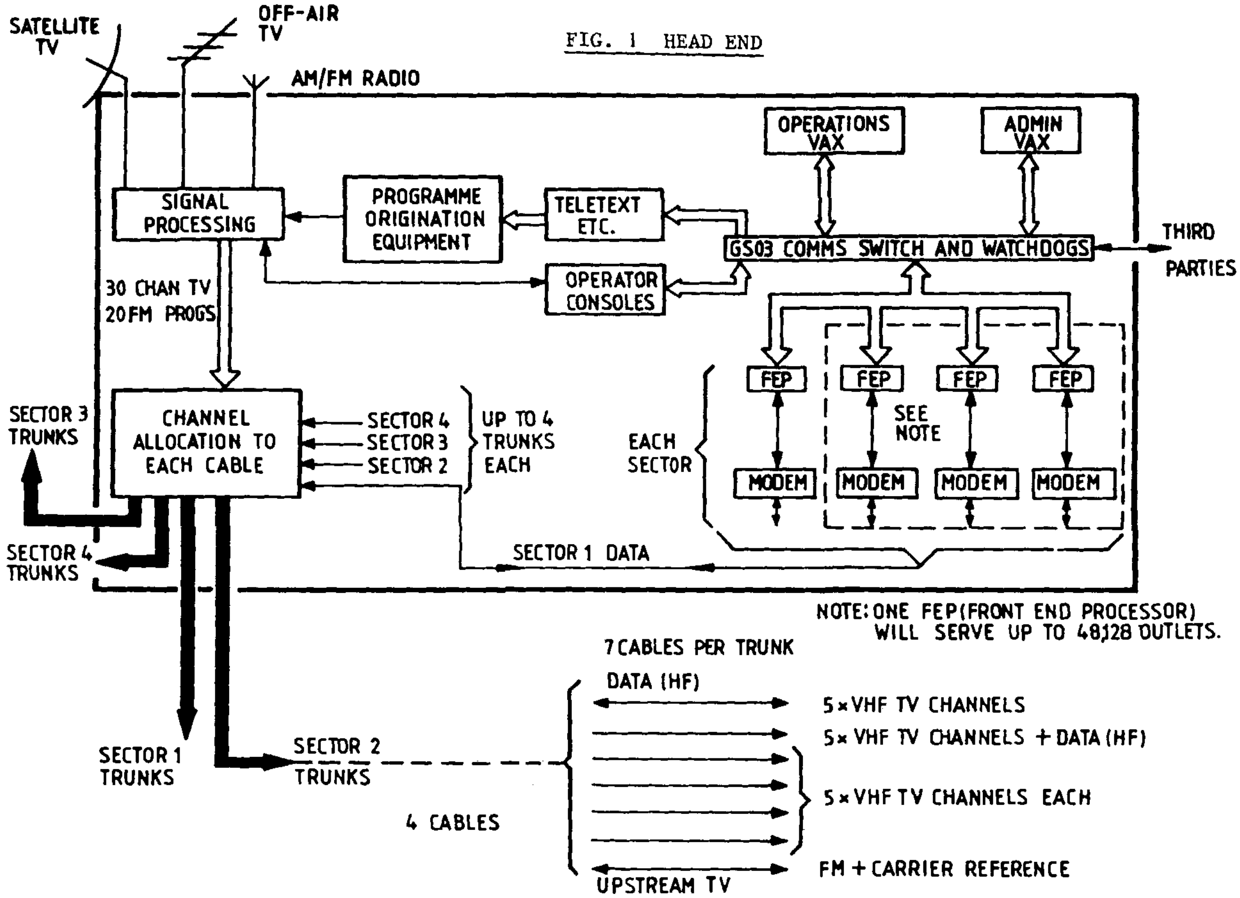
HEAD END

The head end accommodates all the usual signal processing equipment required for the distribution of television and radio signals from terrestrial and satellite transmitters. The aerials and dishes may be located at the head end or sited elsewhere for optimum reception.

The head end can also incorporate studio facilities for local video programme origination, video tape recorders, editing consoles and cabletext signal generators.

The linked computers which control routing of data, access to services, automatic subscriber billing and network status monitoring are also housed at the head end. These computers have interfaces for connections via the PTT to information providers and national and international data networks. This enables the system to provide such interactive services as teleshopping, telebanking and telebetting.

FIG. 1 HEAD END



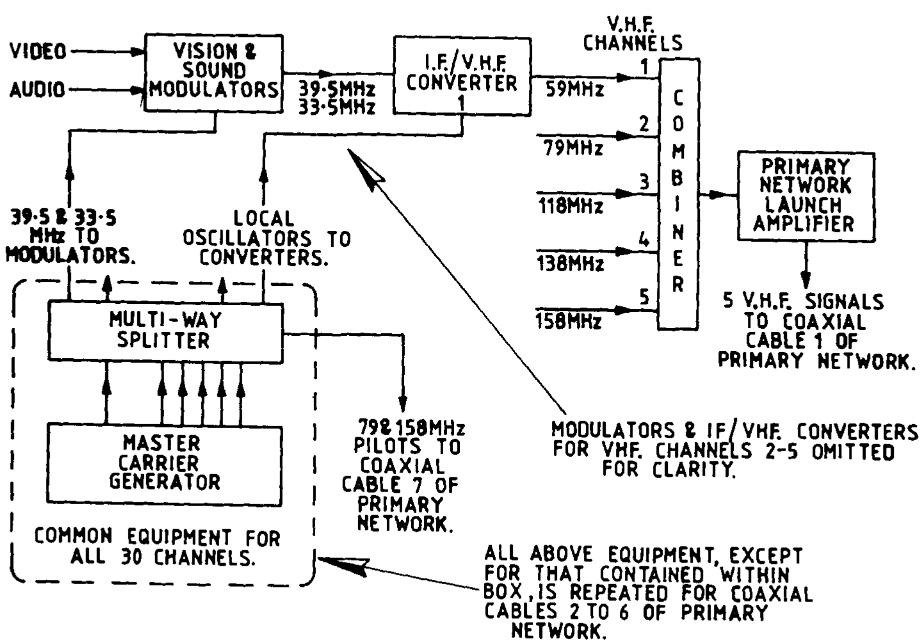


FIG. 2 ANALOGUE SIGNAL PROCESSING.

The equipment at the head end is broadly divided into,

Analogue Signal Processing
Operations Control and Administration.

Analogue Signal Processing

The equipment consists of UHF/IF converters for processing the UHF terrestrial TV signals, These embody phase locked loop circuitry which ensures that the output IF at 39.5 MHz from each processor is phase locked to the station Master Carrier generator.

The Master Carrier generator creates all the carriers, local oscillator frequencies and reference signals required by the system and by means of a multiway passive splitter makes these signals available to all the appropriate equipments, both at the head end and on the network. This approach has the advantage of concentrating all the frequency accuracy and stability requirements into one unit. It also leads to considerable simplicity in the design of modulators and converters which do not need to contain individual oscillators and phase locked loops.

Satellite derived signals and all other baseband signals are processed by IF sound and vision modulators.

All IF signals are converted to VHF, organised into groups of five channels, amplified and combined with the data signals prior to launching on the separate coaxial cables of the trunk network.

The VHF carriers are harmonically related to I.F/2, i.e 19.75 MHz.

This wide frequency separation between channels enables adequate bandwidth to be made available on all channels to ensure transparency to all present and foreseeable transmission standards.

Fig.1 illustrates the overall head end arrangement and Fig.2 the more detailed Analogue signals processing arrangement.

In particular it can be seen how the VHF carriers in each cable of the primary network are synchronous to each of other thereby markedly reducing the effects of any crosstalk between cables and equipments.

Upstream Vision Systems

It is a licensing authority requirement that at least one upstream vision channel must be capable of transmission from a subscriber in each of the four sectors of the network.

To meet this requirement a simple portable origination equipment is available. This uses the same IF modulators as in the head end. The output at 39.5 MHz is fed via the subscriber FM outlet at a sufficient level to reach the first upstream vision version of the trunk amplifier.

The use of 39.5 MHz as the carrier has the following advantages,

1. It fits into the spectrum space below the Band II programmes on coaxial 7 with ample margins for the crossover filter design not to be too demanding.

2. On arrival at the head end, synchronisation of the upstream carrier with other IF signals is ensured since it is derived by division by 2 of the 79 MHz reference signal available at every subscriber.

3. Again at the head end, patching into any downstream channel is easily achieved at the IF interface without further processing. The upstream vision signal thereby becomes available for distribution generally or to a closed user group.

Operations Control and Administration can be performed initially on a single VAX 1130 computer. Larger or linked machines for Operations and for Administration would be introduced as the sub- scriber base grows and more serves are provided.

The Operations computer performs the following main functions:

- System monitoring
- Downloading subscriber parameters to switchpoints (SPs)
- SP configuration
- Statistics provision for admin. system
- Fault reports to admin. system
- Conditional access control
- Interactive functions (vote counting, etc)
- The routing of data signals to service providers.

The Administration computer peforms the following main tasks,

- System configuration
- Setting credit flags
- Automatic billing
- Pay-per-view order processing
- Preparation of fault reports

The Administration computer has no direct interface with the trunk cables and functions via the operations computer. The operations computer interfaces with the network via front end processors (FEP), each of which can provide a

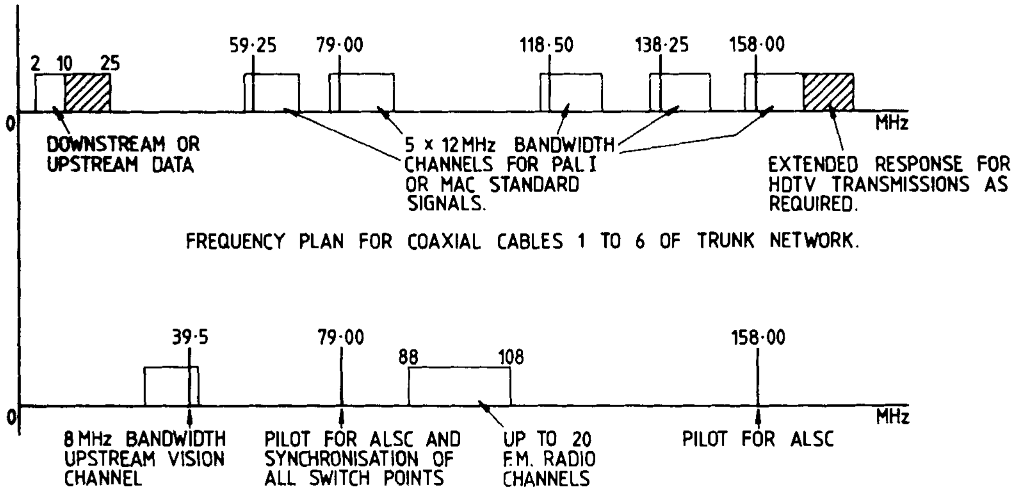


FIG. 3. SPECTRUM OCCUPANCY

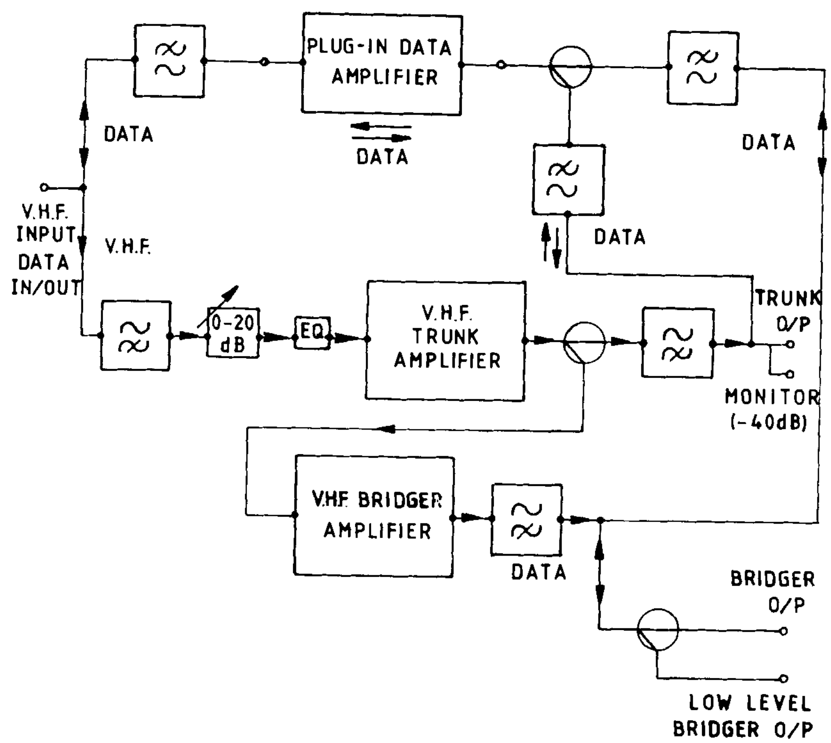


FIG. 4 VHF/DATA TRUNK/BRIDGER AMPLIFIER

microprocessor based cable handler for up to four trunk cables in one sector. Downstream data, from the FEP, is FSK modulated (HF) by a modem and applied to one of the cables carrying television signals. Upstream data from a second trunk cable is routed to the FEP via the modem.

Again a modular construction allows cable handlers to be added as the demand dictates.

PRIMARY DISTRIBUTION NETWORK

The primary network from the head end to the switching points is normally divided into four sectors. Each sector is fed by a trunk system comprising seven coaxial cables, each 8 mm in diameter.

The transparency and capacity requirements dictated that a multi-cable trunk system be adopted. It can be seen that contending with 30 downstream channels of at least 12 MHz bandwidth and some extendable to 20 MHz or more, one 8 MHz upstream channel, 20 FM radio channels and additional bandwidth for two way data communications requires a single coaxial system to provide a total bandwidth well in excess of 550 MHz, taking into account the need to avoid forbidden bands. This bandwidth exceeds the capability of known amplifier designs. Furthermore there would be no room for future enhancement as more wideband services needed to be provided.

The decision to instal more than one cable was inevitable and immediately posed another question. How many cables should be used?

A factor which has an appreciable bearing on this decision process is that in the U.K., new build broadband system networks are required by the licensing authority to be laid underground. Since a very significant proportion of the build cost is in the civil engineering work the additional cost of laying extra cables is not a very significant factor. The decision process is also influenced by the fact that as more cables are used, to achieve a given total bandwidth, individual cable bandwidth can be reduced resulting in smaller cheaper cables and simpler amplifier design.

These technical and economic considerations led to the decision to use seven cables in the primary distribution network of System 8.

This extension of modularity to the primary network allows for low cost start up options, e.g. only four cables and amplifiers would be installed if the operator only required a 15 channel start up.

The spectrum occupancy of the system is illustrated in Fig.3.

Consequential benefits ensuing from the above decisions are:

1. Very simple low cost and reliable trunk/bridger amplifier design is possible without sacrificing system reach and performance.

2. Redundancy in the primary network is achieved so that if a trunk amplifier fails only 16% of the service is affected. When a high value programme is being carried its position in the trunk can be re-allocated on command with minimal subscriber disturbance.

3. Ample bandwidth is available to accommodate any future requirements for two way data transmission.

These benefits incur little additional cost compared to a single coaxial system, assuming the latter was achievable.

For reference purposes the overall performance of the system with a primary network cascade of 35 amplifiers in each of four directions and including all secondary network and subscriber equipment distortion is within the requirement of British Standard BS6513 Part 3, i.e. the carrier/noise ratio is in excess of 43dB and the signal to cross- modulation ratio is in excess of 46dB.

Primary Network Hardware

The block diagram of Fig.4 illustrates the design of the integrated VHF/HF data amplifiers. An interesting feature is the plug in daughter board approach adopted for the data amplifier design. This small subassembly uses Surface Mounted Component (SMC) technology with edge connectors. This allows the integrated amplifier to be configured for use with either downstream or upstream data transmission by simply unplugging the subassembly turning it through 180° and plugging it back. Front panel LED's indicate the configuration chosen.

A version of the amplifier is available with Automatic Level and Slope Control (ALSC). The vision carriers at 79 and 158 MHz are used as the reference levels for the control circuits. 59.25 MHz is avoided since it is absent on coaxial 7 due to the requirement of the Upstream Vision system.

Versions of the amplifiers are available for Upstream Vision on coaxial 7. In these, the data amplifier and low pass sections of the input and output crossover filters are modified to cope with the 39.5 MHz upstream vision carrier.

As explained earlier the use of 79 MHz as a reference signal simplifies the Upstream Vision origination equipment design.

The upstream vision amplifiers embody signal routing relays which can be operated manually or under switchpoint processor control to configure, from the head end, the reverse signal path from the source subscriber to the head end. Contributory noise from unused reverse vision amplifiers is thereby eliminated.

SWITCHED STAR CONFIGURED SECONDARY NETWORK

In a switched star system the primary network is usually in a branching configuration and it interfaces with the secondary network at switching points beyond which the distribution is by a discrete cable to each dwelling, carrying only the selected programmes to the subscriber. This idea originated from:

a) A rural requirement for a long drop cable which, for cheapness, should only carry the few TV programmes being viewed by the family.(3)

b) The concept of an ever increasing number of TV programmes, which could be dealt with by augmenting the primary network and the switch whilst the secondary distribution (which comprises a high proportion of the total cable required) could remain undisturbed.

There are 3 principles on which the switching point can be based. Firstly, by providing a total of crosspoints which is the product of the number of incoming channels times the number of independent outlets. This was used in the USA and Holland and is the basis of the BT switched system installed in Westminster.(4) Secondly, the "switch" can be a frequency-agile converter, i.e. effectively the receiver tuner positioned in the switching point but with an output frequency suitable for the star network. In this case there is one converter per independent outlet. This principle is in use in the UK by GEC and Cabletime. Thirdly, a hybrid arrangement of crosspoint switching between a plurality of incoming cables with each output from the crosspoint matrix connected to a frequency-agile converter. This principle has been in use by BCS for several years.(5)

Each of the above arrangements has its merits but, in the first case, there can be a lack of transparency if the switching is done at baseband because a demodulation/remodulation process is involved.

Having decided that the total bandwidth requirement on the primary network exceeded the practical limits of a single cable, the second "switching" option, i.e. the exclusive use of a frequency agile converter was precluded.

A novel aspect of the secondary distribution network is that programme delivery to the subscriber is carried out at UHF. There is a licensing authority requirement that delivery to the subscriber's equipment must be at UHF and in some systems this is accomplished by installing a simple converter in the subscriber's home.

Factors which influenced the decision to distribute UHF channels on the secondary network were,

1. The desire to reach a sufficiently low cost for the subscriber's active equipment that recovery on disconnection would be uneconomic.

2. The difficulties in meeting the frequency accuracy and stability requirements of the British Standards for the converter in the home.

3. The desire to synchronise similar final delivery channel frequencies to minimize possible interference problems under fault conditions.

4. The availability of inexpensive devices that would produce adequate output level and linearity to meet a secondary network reach of 200 -300 metres, using cable diameters between 5.5 and 10mm. This reach dovetails conveniently with the 440 metres span between primary network amplifiers.

5. The easier realisation of an SP converter with no image channel rejection problems.

UHF Channel Selection Procedure

The choice of 3 UHF channels for programme delivery to the subscriber is governed by the following constraints:

1. They must not coincide with channels in use for broadcast in the area.

2. They must be separated by at least 6 channels to allow inexpensive filtering at the subscriber installation as explained under 'Parental Control' below.

3. They should not lie in the image channels of local broadcast channels. ($\neq N \pm 9$).

A computer program has been devised which will list all the available clear channels given information on local broadcast channels.

SWITCHING POINT

The Switching Point (SP) is the interface between the subscriber and the trunk. Its main functions are to:

- Poll subscriber channel selectors
- Control (permit/deny) subscriber channel selection, including P.P.V.
- Route selected television channel to appropriate subscriber outlet
- Pass interactive data from subscriber to head end
- Route FM radio to all subscribers
- Power subscriber channel selectors

The basic switching point houses common equipment which includes DC power supplies, bus amplifiers, a microprocessor system, and a single frame of equipment which includes a data handler, polling generator, reference generators and power supply filter.

Following the modular approach, switches and subscriber converters are added only when subscribers are connected. A second frame of equipment is fitted only when the 48 outlets on the first frame are fully committed.

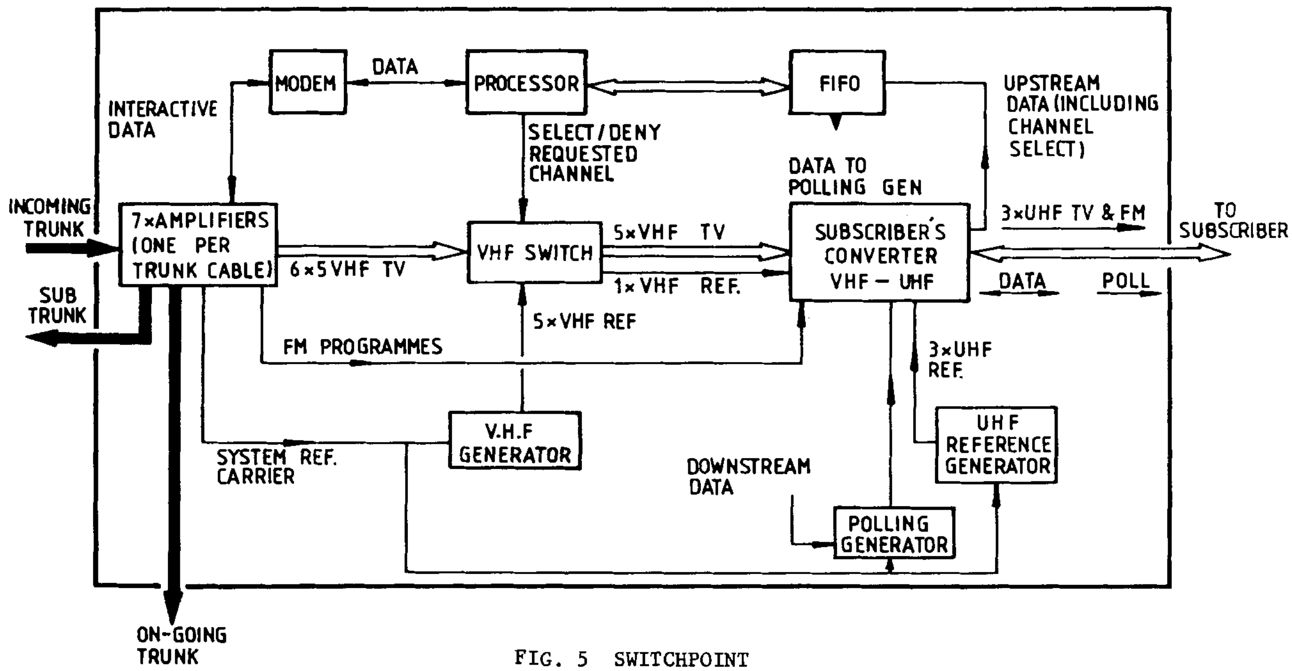


FIG. 5 SWITCHPOINT

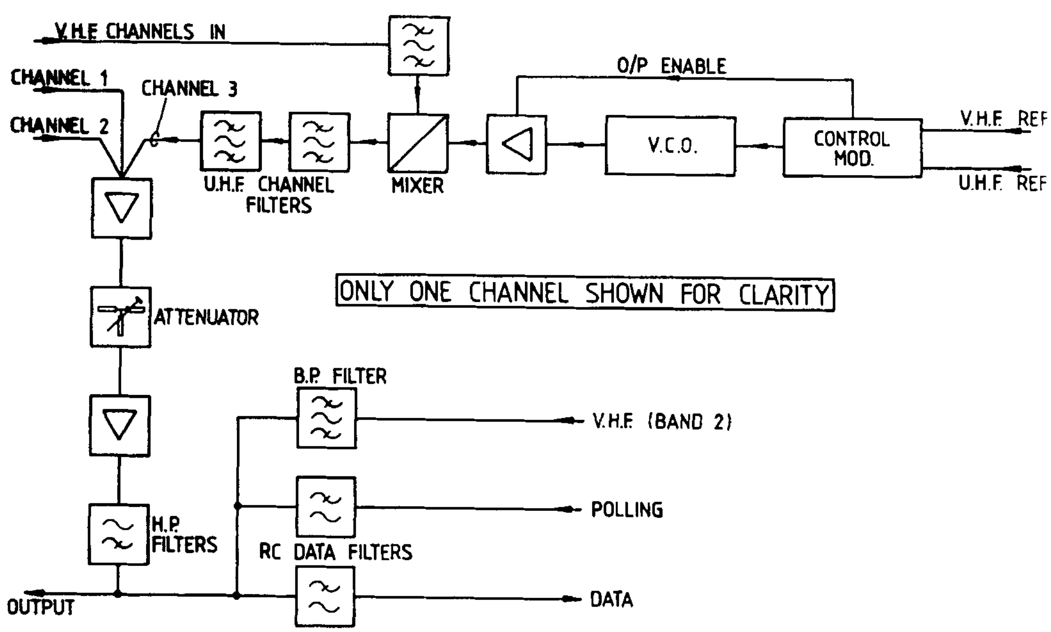


FIG. 6 S.P. SUBSCRIBER CONVERTER

Switching points can also house trunk amplifiers.

The following description and block diagram in Fig.5 illustrate a configuration where a subscriber receives 3 UHF television channels on a single input cable and has three independently controlled outlets.

All incoming signals on each cable are amplified. Downstream data is separated, the FM being applied to the subscriber converter and the reference signal to the reference generators.

The six cables each carrying five VHF TV channels are routed via bus amplifiers and splitters to a number of VHF switches, each being a 6-in 12-out matrix with logic control. The bus amplifiers and splitters have been omitted in the diagram for clarity.

A channel selection is entered at the subscriber's keypad and held in the subscriber channel selector's memory. The channel selector is polled sequentially at approximately ten times per second from the switching point and the data is read into the switching point processor.

The processor checks the selection against the reference file of access rights for each subscriber that has been downloaded from the head end computer. If the request is valid the VHF switch is operated to select the cable and VHF reference associated with the required channel. It then applies them to the subscriber's converter.

If the request is not valid e.g. including a P.P.V. offering without authorisation, the VHF switch and subscriber converter default to a predetermined channel.(6)

The subscriber may have up to three independently controlled outlets, and in this case the VHF switch can select three separate channels to feed identical processing circuits in the subscriber converter.

The subscriber converter performs the following tasks:

- Selects channel
- Upconverts VHF to UHF
- Separates upstream data
- Combines on to one output cable -
  - UHF TV channels (up to 3)
  - FM radio programmes
  - Downstream data
  - Polling/power signal

Converters are available to serve 1, 2 or 3 independent outlets from a single cable.

Block Diagram in Fig.6 gives a more detailed picture of the operation of the subscriber's converter.

The VHF and UHF reference frequencies are generated in two boards and shared among all the subscriber converters.

These generators in turn take their input reference from the 79 MHz reference on coaxial 7 of the trunk.

In this way complete synchronism of similar output UHF channels on the secondary network is achieved. The system can however run asynchronously from the network, whilst retaining synchronism between similar UHF channels, should equipment on the trunk cable 7 fail.

Although at first sight synchronous operation of the UHF channels may appear an unnecessary complication it must be realised that it is relatively simple and inexpensive to achieve since the reference generators are shared by up to 95 outlets.

The benefits of synchronous operation are,

1. The problems of interference patterns due to the proximity of equipment in the S.P., cables in the ducts and subscriber's equipment in blocks of flats are reduced by 10-12dB.

This allows a subscriber converter design which does not need to be very heavily shielded and therefore helps with packing density and in turn street furniture size reduction.

2. It becomes very easy to meet the very tight frequency accuracy and stability requirements laid down by the British Standard BS6513, to which all new build networks must conform.

Text Generators are provided in the switchpoint which, under processor control, can insert messages into the picture the subscriber is viewing. Examples of messages displayed could be "your VCR is receiving Channel X" in response to a Recall request from the subscriber or "Thank you for your order" to end a Teleshopping or P.P.V. interactive transaction.

Supervisory Equipment is contained in the SP for status monitoring. Data is transmitted via the SP processor and modem to the head end computer.

SUBSCRIBER INSTALLATION

A typical installation is shown in the block diagram of Fig.7. A double outlet is located in one area and a single outlet in the other. The VCR/TV receiver combination is independently controlled from one channel selector.

The equipment plugged into each outlet is tuned (preset) to one of the available UHF channels.

The channel selector is supplied with a signal from the switching point and uses this to generate d.c. power supplies. The same signal is momentarily interrupted to poll individual channel selectors. When polled the channel selector returns the keypad selections to the switching point in the form of 7 bits superimposed on the signal. Downstream data to the channel selector is superimposed on the signal as notches representing '1' or '0'.

By connection of a simple modem, 2-way high speed data can additionally be carried on the subscriber's cable.

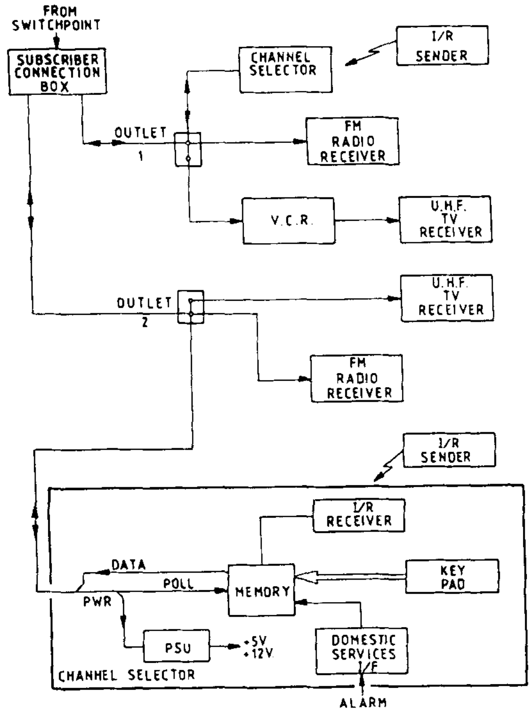


FIG. 7 SUBSCRIBER INSTALLATION

Parental Control

In the example depicted the normal conditional access to programmes, inherent in all switched star systems, is available at all outlets in the house.

Although the switchpoint configuration may call for some programmes to be denied at a particular outlet, if these programmes are available at another outlet on the same subscriber drop then retuning of the receiver to the appropriate UHF channel would allow access to these programmes.

To overcome this potentially undesirable situation a second mode of denial, termed 'parental control', is available. For this purpose versions of the subscriber's connection box are available with built in UHF channel stop filters, ensuring that only the UHF channel/s intended for a particular outlet are present there.

To simplify the design of these filters, the UHF channels on the secondary network are selected to have at least a 6 channel separation between them (48 MHz).

Extended Reach

Inevitably situations arise when a switchpoint may be out of reach of a few potential subscribers and yet due to natural boundaries, etc., it is not considered viable to instal another switchpoint.

To cope with these exceptional cases a version of the Subscriber Connection Box containing a small amplifier and data bypass exists which extends the switchpoint reach to 450 metres.

CONCLUSION

A switched star cabled distribution system has been described whose design was largely determined by the requirements of licensing authorities in the U.K. and Europe, in terms of transparency and capacity.

The outcome of meeting these requirements however has been a system with application wherever a large number of channels and interactive services are required to be provided, and which by virtue of its modular design allows for future expansion without major re-engineering.

Maximum flexibility and upgradeability has been achieved by adopting the principle of distributed intelligence. Incorporation of RAM-based microprocessors in switching points provides for near-instantaneous visual responses to requests and reduction of real-time data handling at the H.E.

ACKNOWLEDGEMENTS

The author wishes to place on record the invaluable pioneering work of R.P.Gabriel and E.J.Gargini, and to express his gratitude to the Management of B.C.S. for permission to publish this paper.

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INSIDE THE LEAD ACID CELL

Martin H. de Alminana

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With the ever increasing use of stand-by power and consequently lead acid storage cells, the need for a clearer understanding of this device and it's components is necessary. No one battery is universal. Therefore, a basic knowledge of cell constructions and their corresponding performances are required. With this, one can then appreciate the manner in which batteries respond when exposed to various charge and discharge parameters. This paper will deal with the design and application characteristics one should be aware of when called upon to make a choice of today's available products.

A brief look back shows that in response to research conducted during the years 1859 to 1879 by Gaston Plante, considered to be the father of the lead acid cell, a wealth of ingenuity and imagination was illustrated by a rapid growth of design and application variations of early products thru the turn of the century. Since that time, the improvements have been essentially refinements of these basic designs using more modern materials and techniques.

However, even yet there is no ideal battery. The ideal battery, were it available, would exhibit infinite energy, handle all power levels, operate over a full range of temperatures, have an infinite shelf life and be consumer proof. Since this is not the case, we must approach this device with its many factors in mind.

In the design of a lead acid cell, the function of the various components must be considered in relation to the actual requirements of each application. It is worth noting that any design is a compromise. Each function has been considered in terms of its importance in the particular application. Thus, there will be different optimized designs depending on the use into which the cell will ultimately be placed. Cells that are designed for example for optimum capacity at relatively low or moderate discharge loads contain maximum quantities of active material. On the other extreme, cells capable of high-rate, large current delivery performance, are designed with reaction surfaces and other features to minimize internal resistance. The later is done often at the expense of total capacity. It will be found that cells designed for high-rate service will exhibit a more constant performance, ie: current output, to those of the greater capacity category.

The typical lead acid battery is comprised of three basic elements. The positive plate, the negative plate and the electrolyte. There are, of course other nonreactive components such as the separators, connecting straps and case. We will only consider the active components and there interrelated effects.

PLATES AND GRIDS

The plates should be considered as a compound element in that they are a combination of two items. First, a grid is formed in one of several methods. The most popular, for a vertical plate design, being a casting process where the lead is poured into a mold to form the basic structural component. The second most popular process employed today is a wrought technique where a solid lead sheet is drawn through an expander which places slots in the sheet and then stretches these slots into openings much like the common expanded mesh sheets available from most hardware suppliers. This system reduces the cost of manufacture as there is less energy required to process a continuous sheet of lead as opposed to that of maintaining supplies of molten lead for the casting operations. The advantage of individual casting is that in process changes can be made in the lead mixture at any time where as in the wrought process the entire lead supply must be reworked. The claimed advantage of lower gassing rates in the wrought process, due to a more consistent and finer grain structure in the lead, will most likely not be a factor of concern in CATV applications as this reduced gassing is such a low value of change that other factors in the electrochemistry and particularly the application will serve to overshadow these gains.

After the grid is prepared, a mixture of highly reactive sponge lead for the negative electrode, (anode) or lead dioxide for the positive electrode, (cathode) is mixed with sulfuric acid to form a paste. Included in this mixture are sufficient expander materials to achieve and maintain a high level of porousity. This is necessary so as to achieve an optimum ion conductivity in and out of the plate active material. The paste is then spread onto the grid which will provide the necessary support structure and electrical conductivity.

There are, however, several points regarding the lead which should be addressed here. If pure lead were attempted to be used for the grid structure, which by far is the most desirable, it would not exhibit enough rigidity to withstand handling during manufacture, not to mention the actual in service requirements. This then dictates that an alloy be introduced into the pure lead to add strenght. Historically an antimonial alloy was used as this strength adding agent. The presence of this compound though, exhibited two major down side effects. The first to be discussed is that of excessive gas production.

The gassing is a direct result of electrolysis depletion of the water at the end of charge. This electrolysis produces hydrogen at the negative plate and oxygen at the positive plate. The main contributor to this activity is the presence of this antimony alloy. In the past, low cost vertical plate design batteries were routinely maintained therefore frequent watering intervals were not of great concern. It is interesting to note that the production of oxygen starts as soon as the recharge cycle is commenced and continues through full charge. The evolution of hydrogen at the negative plate does not start until it has reached 90 to 95% recharge. An interesting point is that the negative plate accomplishes it's recharge must faster than the positive. This is why the "End of Charge Rate" must be closely controlled to achieve total recharge of the positive plate without excessive over charge of the negative which would result in excessive hydrogen evolution.

In the mid 1970s, the maintenance-free lead acid battery became popular. This design featured a grid alloy that basically was a nonreactive component. The antimonial alloy, being a reactive component, and as stated before, elevated the gassing rate and contributed to plate corrosion which also results in increased water depletion as the cell ages. By using a nonreactive strengthening agent, a more nearly pure lead grid could be achieved. This resulted in a dramatically reduced end of charge gassing rate. This marked the introduction of the Lead-Calcium battery. The inclusion of calcium resulted in a reduction of the antimony to generally no more than 2%. A maximum strength yield is achieved with about .11% calcium content.

The inclusion of calcium brought on it's own set of problems mainly related to the manufacturing process. It seems that while maintaining the lead in the molten stage rapid loss of calcium through oxidation would result. Many attempts to control this were tried.

Floating lids of various types were tried by placing them on the surface of the molten lead. One attempt of a floating layer of dross was even attempted. None of these were successful. It was then found that by adding Aluminum to the mixture that this condition could be controlled effectively. Along with the inclusion of AU it has been found that by utilizing nucleating agents such as copper and selenium a very uniform and fine grain structure results. This allows for a superior active material to grid adhesion there by reducing loss of the active material through sheadding. This also reduces the ill effects of corrosion as the corrosion tends to be more uniform on the grid wires rather than forming pits and fractures as a result of inconsistency in the lead crystallization.

The second ill effect of using antimony for grid alloy is a high self discharge rate. The self discharge rate (loss of battery capacity on storage) is dependent on a number of factors, including the type of lead alloy used, the concentration of electrolyte, the age of the battery and storage temperature. The greatest contributor to this activity however is the antimony lead alloy. Self discharge is caused by local reaction of the plate materials and occurs almost entirely in the positive electrode. The rate of self discharge is about 15% per month for antimonial-lead batteries at 25'C. Batteries using lead-calcium grids have substantially lower rates of self discharge. A reduction in this activity by 50% or more is not unrealistic. For best practice, a battery on stand should be recharged every 3 to 6 months, since prolonged storage can cause irreversible damage and make recharging difficult, owing to sulfation of the plates. (Fig-1)

Stationary batteries utilize a thick plate design that reflects the lack of need for high energy and power as in the case of starting, lights, and ignition (SLI) types. The typical overcharge operation of stationary batteries requires a large electrolyte volume and non-antimonial grids, all to maximize intervals between waterings. This over or constant recharging, by any of the various on-off, dual-rate or closed loop methods used today, causes some positive grid corrosion. This is manifested a "growth" or expansion of the grid structure and must be allowed for during the design of the cell so as to provide room for this normal expansion to take place during the useful life of the battery. Excessive overcharging can accelerate this activity to a point where, if allowed to continue over time, it may cause case rupture to occur. The normal tolerated growth is calculated to be about 10% over the life of the cell. (Figure 2)

ELECTROLYTE AND SPECIFIC GRAVITY

The selection of a specific gravity used for the electrolyte depends on the application and service requirements. The concentration must be high enough for good ionic conductivity and to fulfill electrochemical requirements, but not so high as to cause separator deterioration or corrosion of other parts of the cell which would shorten life and increase self discharge. In some cases where a battery is required to operate in high ambient temperatures, the electrolyte concentration may be deliberately reduced to off set the effects of temperature elevated chemical activity resulting in accelerated plate corrosion and lowered overvoltage gassing potential. The concentration for most lead acid batteries to be used in temperate climates is usually between 1.26 - 1.28 sp.gr. Higher concentrations tend to attack the separators and other components. Lower concentrations tend to be insufficiently conductive in a partially charged cell and freeze at low temperatures. In standby and or stationary cells with larger proportional electrolyte volumes and no high rate demands, concentrations as low as 1.12 sp.gr. are used.

It is important to maintain not only the correct sp.gr. at full charge per the manufacturers specifications, but also, especially in the case of calcium alloyed grids, not to contaminate the acid solution. A preventative measure, from the manufacturers standpoint for automobile batteries, has been to permanently attach the vent caps. This is done understanding that under normal conditions the electrolysis of the water will be at a rate far lower than the overall aging of the cell. Therefore, cell failure due to other factors will occur prior to damage caused by water loss. If the electrolyte is allowed to become contaminated , as in the case of replenishment with other than distilled water, an increase of the gassing rate will result. This obviously will increase watering intervals and if at these times the same practice of using non-distilled water is employed, the gassing rate will again be increased. This practice , if allowed to continue, will most likely result in premature failure of the cell as it is highly unlikely that the watering schedules will be adjusted to compensate for this increased loss.

VOLTAGE AND SPECIFIC GRAVITY

The nominal voltage of the lead acid cell is 2 V. The open circuit potential is a direct function of the sp.gr. ranging from 2.12 V for a cell with a sp.gr. of 1.28 to a potential of 2.05 V at 1.21 sp.gr. Figs. (C,D & E) present typical discharge curves for the lead acid cell. The end voltage is typically 1.75 V but can be as low as 1.0 V at extremely high rates such as in automotive starting service. The 1.75 V point is the standard cut-off voltage which manufacturers design too. During discharge the sp.gr. decreases about 0.125 to 0.150 points from a fully charged to a fully discharged condition. The change is proportional to the ampere-hours discharged. The sp.gr. is thus an excellent means for checking the state of charge of the battery.

A short period should be allowed prior to measurement after completion of the discharge for equalization of the concentration throughout the cell. On charge, the change in sp.gr. should similarly be proportional to the ampere-hours accepted by the cell. There will be a lag in the sp.gr. change if the cell incorporates a high concentration electrolyte as complete mixing of the concentration will not occur until overvoltage gassing occurs and is sustained at the end of charge.

The variation of the performance of the lead acid cell at different temperatures and loads is given in Figures 6 and 7. Although the battery will operate over a wide range of temperatures, continuous operation at high temperatures may reduce cycle life as a result of the aforementioned increased rate of chemical activity and subsequent corrosion.

It should be understood that in applications where the battery will be housed outside, subject to very low temperatures for long periods of time (dissipating the internal heat needed for full capacity delivery), some form of warming provision should be considered. If long-term storage is going to take place, an area should be chosen that will provide a low mean temperature but will not fall below freezing as this could damage a flooded free electrolyte battery which is only partially charged.

PHOSPHORIC ACID EFFECTS

The utilization of phosphoric acid as an electrolyte additive has been patented as far back as 1929 for the claimed purposes of strenghtening of the positive active material and preventing harmful sulfation during long discharged stand conditions. In the case of cell designs using tubular positive plates, a reduction in active material shedding and in the rate of positive grid corrosion have been claimed.

It has been demonstrated that cycle life of lead acid batteries using plates with lead-calcium grids is increased when the electrolyte contains small amounts of phosphoric acid, at the cost of some reduction in the positive plate capacity, of the order of 5 to 10%. The mechanism by which phosphoric acid increases cycle life is one of modification of the pattern of the lead sulfate as it forms on the plate active material surface. The result is a more conductive interface, with less of a sulfate barrier and less interference with the discharge process.

There are also chemical and or electrochemical reactions which occur between the phosphoric acid and the positive plate active material, with phosphoric acid incorporation during charge and release during discharge. However, the formation of lead phosphate compounds in the positive plate during charge is the likely candidate for responsibility of the observed reduction in capacity in such cases. Other effects due to the presence of phosphoric acid in the electrolyte solution are lead "Mossing" and dendrite formation, both of which are factors which tend to reduce capacity.

DISCHARGE LOADS

A battery can be discharged under different modes depending on the equipment load. Three typical modes are constant resistance (the equipment resistance remains constant during discharge), constant current, and constant power (the current load on the battery increases as the voltage drops to maintain a constant power,I x V). Assuming that the discharge current is the same at the start of the discharge, the current will be different during the discharge under different discharge modes as shown in Fig.(G-a). The constant resistance curve reflects the drop in the battery voltage. Fig.(G-b) shows the voltage vs. time discharge curves for the three modes. Under the conditions shown, the service time is longest in the constant resistance mode.

Fig.(G-c) and (G-d) show the same relationships assuming the same average current during the discharge. Under these conditions, the service time is about the same, but the voltage regulation for the constant resistance mode is best. The constant power mode has the advantage, however, of providing the most uniform equipment performance throughout the life of the battery and, hence, makes most effective use of the battery's energy. It is worth noting at this point that standby equipment powering line equipment will regard this load as a more nearly constant power load. In the case of a pulsed discharge load, these conditions will exhibit the recovery effects that occur when the cell is left open circuit for a period of time after discharge. This sawtooth type of response will, in some cases, lead one to thinking that a battery has sufficient charge to operate when in fact it has reached cutoff under load but has had time to recover above the specified exhaustion voltage. See Fig.(8)

BATTERY CHARGING

Proper recharging is important to obtain optimum life from any lead acid battery under any condition of use. Some rules for proper charging are given here and apply to all types of lead acid batteries. The charge current at the START of the recharge can be of ANY value that does NOT produce an average cell voltage in the battery string greater than the gassing voltage which is typically 2.39 V/cell. During the recharge and until 100% of the previous discharge has been returned, the current should be controlled to maintain a voltage lower than the gassing voltage. If one wishes to minimize charge time, this voltage should be just below the gassing point which is directly related to the sp.gr. and cell electrolyte temperature. When 100% recharge is accomplished, the charge rate will have to decay to the "Finish" rate. This rate is described as a constant current no higher than 5A per 100 ampere hour of the rated capacity at the 5 hour rate. As a result of testing done at temperatures of 80' on new cells, a finish rate for batteries available to the cable television industry of no more than 100ma float current will overcome the effects of self discharge and keeps the end-of-charge gassing to a tolerable limit.

Some of the more popular methods to achieve recharge are listed here.

1. Constant current
2. Constant potential, modified constant potential
3. Taper
4. Pulse
5. Trickle
6. Float

1) CONSTANT CURRENT

Constant current recharging at one or more rates is not widely used for lead acid batteries. This is because of the need for current adjustment unless the charging current is kept at a low level throughout the charge cycle. This would, however, result in an extremely long recharge time.

2) CONSTANT POTENTIAL, MODIFIED CONSTANT POTENTIAL

In normal industrial applications, the modified constant potential is employed. In this case the charging circuit has a current limit, and this value is maintained constant until a predetermined voltage is reached. Then the voltage is maintained constant until the battery is called upon to discharge. It has been found that more than 50mv positive negative over potential is necessary to prevent self discharge. So that .005a float current per 100 ampere hour of capacity is required for lead calcium batteries. Lead antimony cells require at least .06a per 100 ampere hour, but this increases to about .6a per 100 ampere hour as the battery ages. This higher current also increases the water electrolysis depletion rate. This can be attributed to plate grid corrosion and accelerated sulfation. Decisions must be made regarding the current limit and the constant voltage value in accordance with the manufactures specifications.

3) TAPER

The taper charging is a variation of the modified constant potential method using less sophisticated controls to reduce equipment costs. The initial rate is limited, but the taper is such that, if special precautions are not taken, the 2.39 V/cell at 25§C may be exceeded prior to the 100% return of charge. This method could result in gassing at the critical point of recharge. The degree of gassing is a variable depending upon the charger design. Battery life can be degraded from excessive gassing. the gassing voltage decreases with increasing "Electrolyte", not ambient, temperature. A correction factor for the actual temperature of the electrolyte should be employed. The end of charge is often controlled by a fixed voltage rather than a fixed current. therefore, when a new battery, which has a high counter EMF is used, this battery often does not receive sufficient charge. Conversely, an older battery who's CEMF is low, will now receive a higher than normal finishing rate resulting in excessive gassing.

4) PULSE

A pulse system periodically disconnects the batteries from the charge circuit and performs a high impedance, no load, voltage check. If the open circuit voltage is above a preset value, depending on the reference "Electrolyte" temperature, the charger is not called upon. When the open circuit voltage decays below that limit, the charger delivers a dc pulse for a fixed period of time. The duration of the open circuit and charge pulses are chosen so that when the battery is fully charged the time for the open circuit voltage to decay is exactly the same as the charge pulse duration. When the charger senses this condition, it is automatically switched over to the finish rate current and short charging pulses are delivered.

5) TRICKLE

This is a constant current simplistic system which delivers a very low current to the battery and is used mainly to overcome the effects of self discharge. There are no provisions for any compensation techniques.

6) FLOAT

This is a system used to deliver a low rate, constant potential charge. In it's purest form it will be found in applications where stationary batteries are used in constant temperature environments.

CONCLUSIONS

As can be seen from the information given here, wise application of the proper battery for a specific need is of most importance. The cable television industry has basically three cell configurations to choose from. The traditional flooded lead acid battery, the gelled flooded battery and now the SLA configuration. Of the three mentioned, list price comparisons will show that the traditional flooded type has the lowest cost, with the gelled and SLA being competitive with each other. The obvious disadvantage of the flooded type is the need for water replenishment and the overhead cost there of. The gelled types can be thought of as a can't maintenance cells as replenishment of dissipated moisture is impossible. They do have however very low gassing rates so as to provide an improved life performance before replacement is required. The SLA technology, as applied to catv uses, would seem to indicate the best of all available performance characteristics. It can be spoken of as the only true no maintenance product. It's inability to vent the internal gases obviously will prevent it from dissipating the electrolyte moisture. Improved internal resistance characteristics provide for greater capacities at lower temperatures as well as more desirable recharge results. Another interesting feature is in the case of freezing of the cell. This type of treatment will not effect the cell adversely as in the case of flooded types. The only result will be that the cell will not operate until the internal temperature has been raised above the electrolyte freezing point allowing ion activity to resume.

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He comes to Lectro with background in computer based automated test systems for the aviation industry, system design and application of rf communication equipment and having served as an instructor on military microwave communication installations.

Figure 1: Capacity/power loss during storage at 25°C

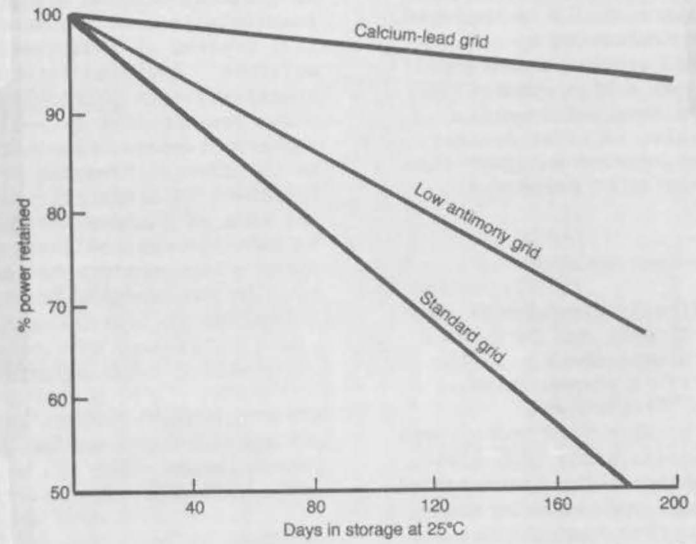
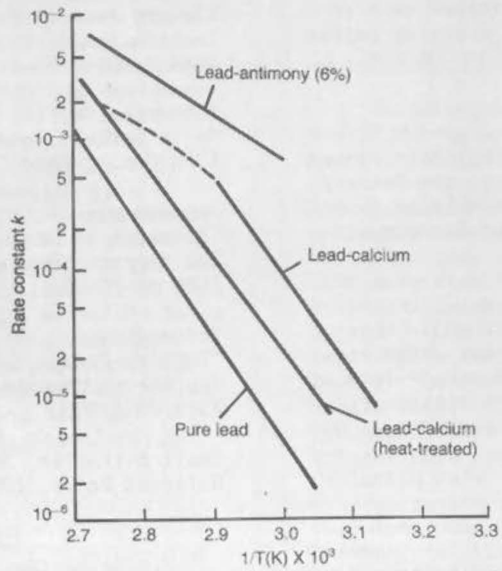


Figure 2: Plate corrosion rate chart



Corrosion rate constant *k* on log scale vs, *i/t* (*K*) for different lead-alloy grids.

Figure 3: Self-discharge

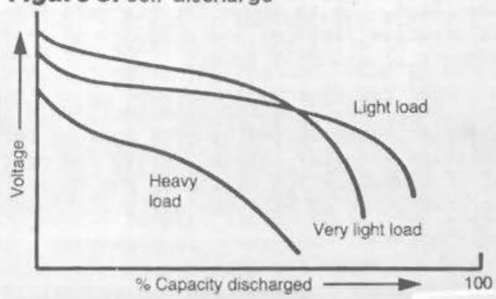


Figure 4: Discharge curves at hourly intervals

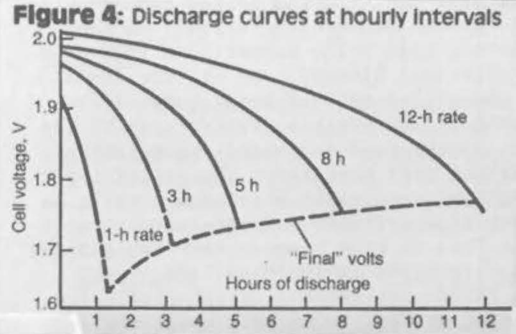


Figure 5: Discharge/charge characteristics

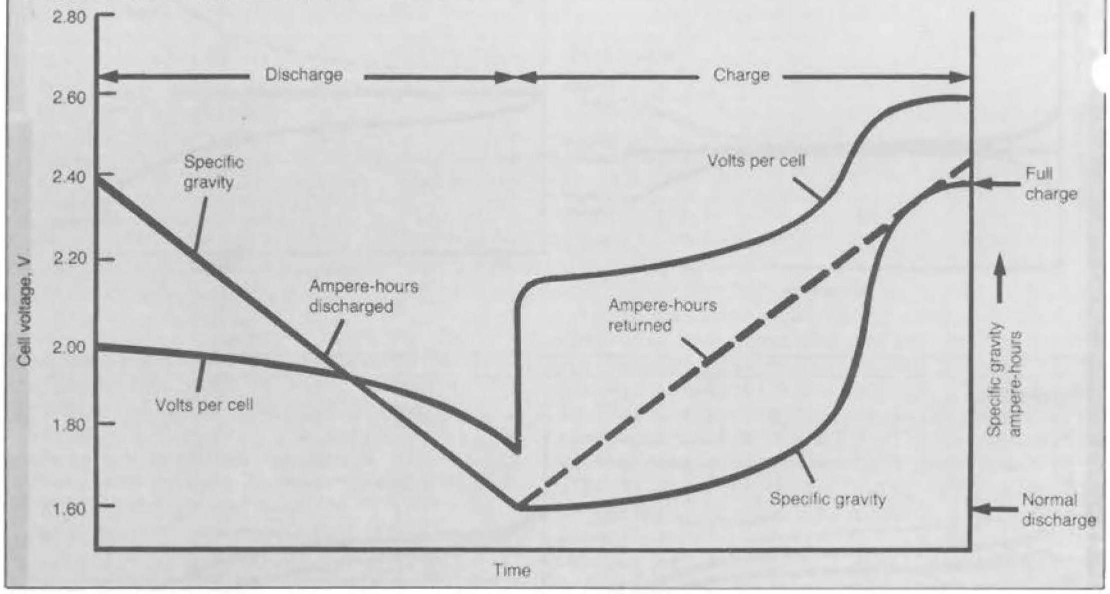


Figure 6: Temperature effects on electrolyte

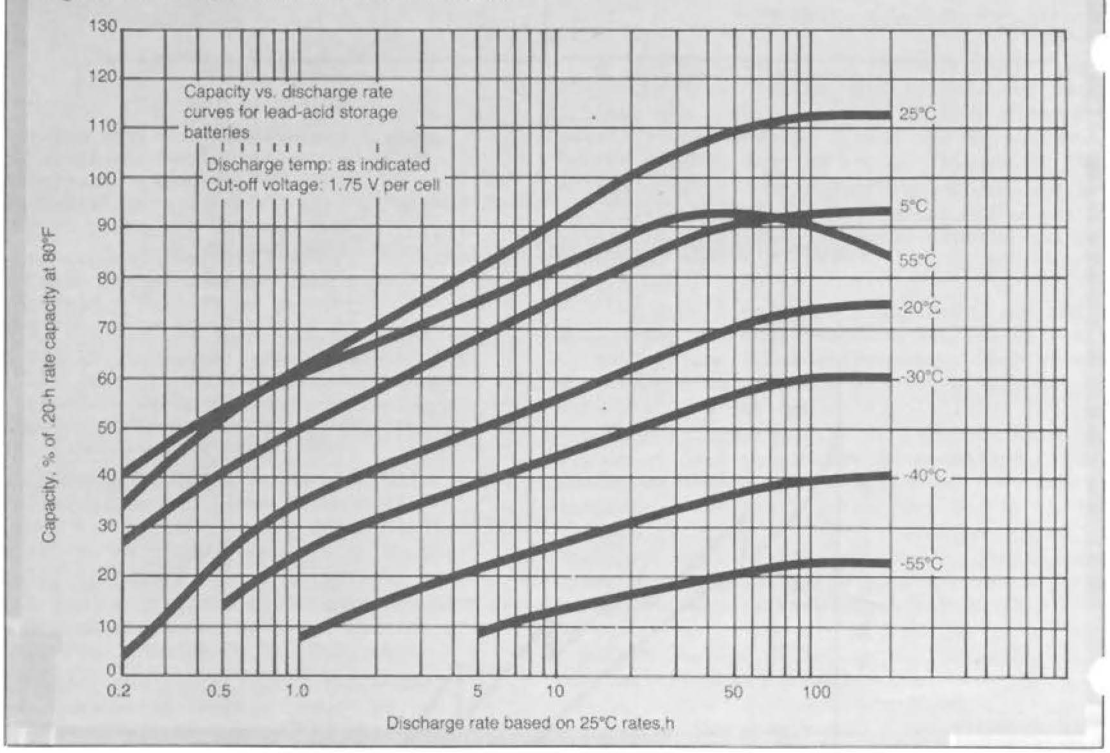


Figure 7

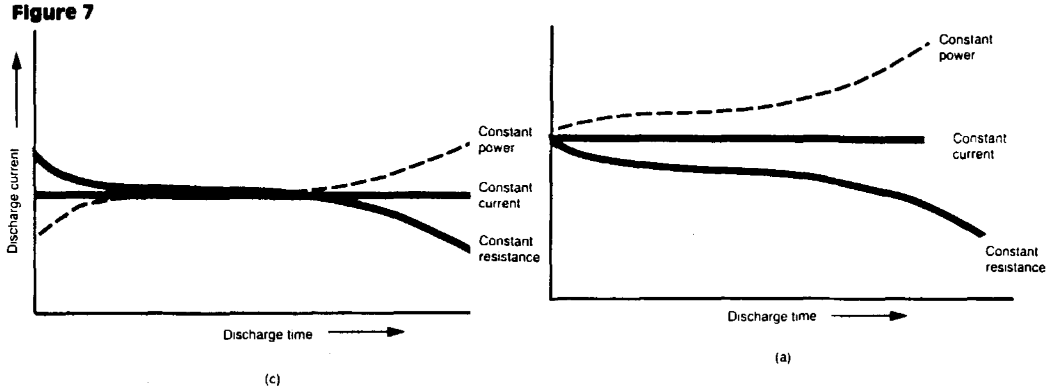


Figure 7

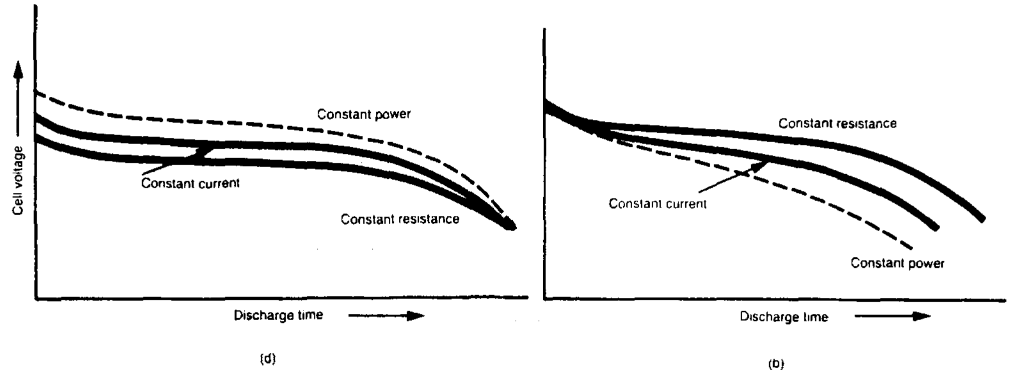
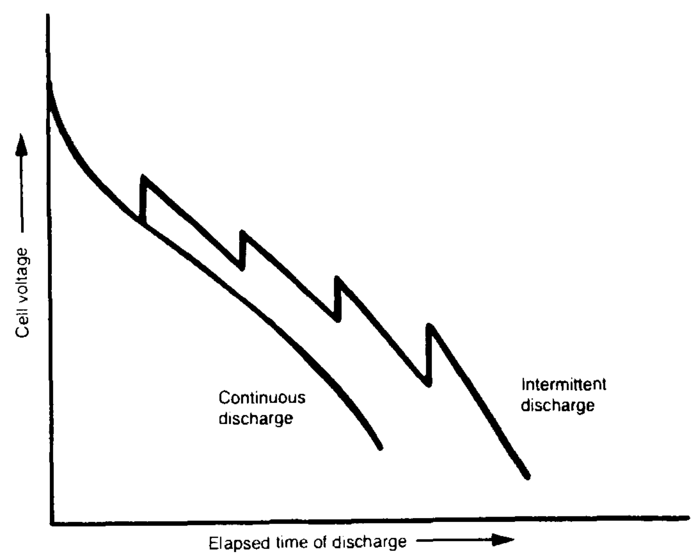


Figure 8: Effects of intermittent discharge on battery capacity



KU-BAND DISTRIBUTION OF TELEVISION PROGRAMMING FOR THE CABLE INDUSTRY

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HOME BOX OFFICE, INC.
and
RCA AMERICAN COMMUNICATIONS, INC.

INTRODUCTION

Ku-band for satellite delivery of television programming has become accepted worldwide. Abroad, European and Japanese satellite systems are based on the Ku-band frequency to take advantage of (1), higher power signals, which result in a less costly ground segment, and (2), relative freedom from RFI, which facilitates antenna siting.

In the United States, Ku-band has been selected for network television program distribution and for the establishment of nationwide satellite newsgathering services by several entities. It is also the preferred transmission technology for VSAT networking to establish private voice/data, transaction and video conferencing systems.

By 1992 67 percent of the satellites world wide will be operating at Ku-band. And of the 6 domestic U.S. launches planned between now and 1990, five are either Ku-band or hybrid.

The use of Ku-band for program distribution to CATV and SMATV systems is an evolutionary process, not a revolutionary step. Ku-band distribution allows affiliates to serve subscribers considered unreachable at C-band, eliminates the problems of C-band terrestrial interference encountered at some primary receive locations, provides a 2nd totally redundant signal distribution path thus protecting against program disruption and, over the next 3 years, allows an unprecedented increase in the efficient use of a satellite orbital slot and spectrum.

The success of Ku-band distribution requires the optimization of the satellite system including satellite performance, in orbit and ground protection to insure continued operation, and the frequency plan to ensure maximum utilization of the orbital resouce. On the ground careful planning and operation of the Ku-band receiving earth stations is needed to insure superior quality and reliable reception of the signal.

In July, 1986 subsidiaries of HBO and RCA formed a joint venture called Crimson Satellite Associates to build, acquire, and launch a Ku-band satellite called K3. Crimson also has the option to acquire and launch a second Ku-band satellite, called K4, and, subject to FCC approval, to co-locate it with K3. Launch of K3 is scheduled for Mid-1989.

An existing RCA Ku-band satellite, K1, serves to provide interim capacity until 1989 and allows the cable industry to build confidence in Ku-band distribution by providing a time to experiment and learn. Thereafter K1 will serve as an integral component in the satellite system protection plan.

WHY KU-BAND

The use of Ku-band for program distribution to CATV and SMATV systems is a natural step in our industry, as illustrated by examining future satellite systems.

A review of launch schedules reveals that the next generation of satellites are predominately Ku satellites.

Between now and 1992, 59 proposed communications satellites are authorized for launch worldwide. Of these, 52 will be Ku-band or Ku-hybrids, six will be C-band and one (Italy's) will operate at another frequency.

Between 1987 and 1990, Ariane has reservations or contracts to launch 32 communications satellites on 24 rockets. Of these 32 satellites, only one does not have a Ku-band payload. This inescapable fact merits emphasis: Every communications satellite launched between now and 1990 will be K-band, save one.

Ku-band has become the clear choice for future capacity for four important reasons:

1. Business Benefits to the Cable Industry:
   The benefits of Ku-band stem largely from its regulatory heritage. C-band satellites are secondary users of a shared-frequency assignment and consequently must not interfere with the terrestrial microwave links also using the spectrum. The interference is held to a minimum by operating C-band satellites at low power levels. Also, reception of C-band signals can be difficult or impossible because of interference from these same terrestrial microwave links.

   The Ku-band satellite service, on the other hand, is a primary user and does not need to protect any other radio service. Consequently, Ku satellites can be more powerful than those using C-band, permitting the use of smaller, less expensive equipment to receive the signal. This naturally provides business benefits to both programmers and cable affiliates, particularly as they look to serve areas not yet passed by cable plant. These business benefits include:

   - The ability to situate a Ku earth station at an urban headend regardless of the presence of terrestrial microwave signals, which will facilitate construction as the major urban markets are cabled.
   - In most cases, economical 1.8 meter Ku dishes for rural cluster cable systems not yet receiving satellite programming, thereby providing new services to homes heretofore unreached by programmers.
   - Ku earth stations costing approximately $1300 can be used by operators to reach unpassed multi-family units (SMATV) prior to an urban cable build or in lieu of long cable runs to reach a rural pocket such as a trailer park or apartment complex.

2. Superior Protection Against Failure:
   Ku-band offers the opportunity to create a superior level of protection from catastrophe that is cost-effective, flexible, and adaptable to the evolving business environment. The objective is to maintain the business with minimal disruption to program continuity or the installed receiving system.

   Protection at Ku-band makes possible a protection configuration ensuring minimum disruption of service by eliminating the need to repoint earth stations at a new satellite or disperse programming across several satellites in the event of catastropic failure.

   Unlike any protection plan at C-band, Ku-band protection is achieved by on-board redundancy, an in-orbit spare satellite and a ground spare satellite. The growing base of Ku satellites means the cost of protection is reduced to each user while the number of C-band systems decrease, the cost of protection will be shared by fewer users, and thus cost to each user will increase.

3. Technical Evolution:
   The early Ku-band satellites were primarily designed for the transmission of high volume voice and data traffic from point A to point B using relatively large antennas. Ku satellites being constructed today and in the future will be

more suited to the predominant uses of satellite capacity -- video and point-to-multipoint small dish data transmission.

Ku-band satellite systems have been operational since 1980. The technology and performance characteristics have been studied and well-defined. The past 7 years of experience allows this new generation of Ku-band satellite to overcome many of the problems experienced in the past.

The early problems of degraded performance due to rain attentuation exhibited in the first generation Ku satellites have been overcome through the development of more powerful satellites capable of providing increased signal margin. Using the antenna sizes our industry employs, rain no longer degrades service for any significant time.

4. Channel Capacity:
Only 24 channels of video programming can be transmitted from each C-band orbital location. Therefore, headends are now "antenna farms" with an ever-increasing number of dishes looking at a variety of C-band satellites. The proposed channelization plan of the Crimson Satellite Associates Ku-band satellite system would allow each TVRO to receive a total of 32 program channels from a single orbital location.

CHARACTERISTICS OF THE K3 and K4 SATELLITES

The present FCC rules governing Ku-band satellites operating in the FSS band allow more flexibility in satellite design: i.e. higher power densities are permitted (compared with C-band) since interference with terrestrial networks is not a problem.

The design of the K3 and K4 satellites incorporate many advanced features that will enable the high performance requirements of the CATV and SMATV industry to be met across the country.

The Ku-band satellite system operates in the 14.0 to 14.5 GHz band on the uplink and in the 11.7 to 12.2 GHz band on the downlink. Each satellite includes sixteen transponders with 47 watt TWTAs (60 watts with FCC approval). Eight of the transponders will be horizontally polarized and eight will be vertically polarized. For high reliability, these sixteen transponders are protected by a total of six spare TWTAs on board the satellite.

The transponders will be fully protected against eclipse outage. This insures full operation and power for 24 hours a day, 365 days a year for the 10 year design life of the satellite.

The high power available from the TWTAs on board the satellite will be directed to the earth in one of three modes that are reconfigurable on orbit by ground command. These modes will allow any individual transponder to be configured as a CONUS beam or an East beam or a West beam. This allows the entire transponder power to be concentrated in the areas of the country where the signal is desired. Selectable coverage ensures that each transponder has flexibility for the user and can readily accomodate varying traffic usage patterns.

An EIRP pattern showing the minimum EIRP valves for 60 watt transponders expected for each of the transponder modes is shown in Figures 1, 2, and 3. The East beam provides coverage to the Eastern and Central time zones while the West beam covers the Pacific and Mountain time zones. The CONUS mode covers the 48 contigious states. Hawaii is covered in the West beam and CONUS modes. Two video signals per transponder can be accommodated, allowing up to 32 television channels to be distributed by one satellite.

Another innovative feature of this satellite is the on board dynamic limiter amplifiers (DLA). When the DLA is operating in the limiter mode, the transponder output is maintained at full downlink power even when the uplink signal fades. This insures that the overall video signal to noise ratio will be affected minimally by uplink fades and will obviate the necessity for expensive uplink power control at the transmit earth station. The inclusion of the DLA on board the satellite greatly enhances overall link robustness and provides a superior video signal in the event of propagation effects on the uplink.

The planned launch of K3 is in Mid-1989. An additional satellite (K4) can be launched at a later time and co-located with K3 (subject to FCC approval). The two satellites (each with sixteen active transponders) would then provide a total of 32 channels at a single orbital location. The satellite design is such that each of these 32 channels utilizes the full power of its own transponder. This is accomplished primarily by narrowing the bandwidth of each transponder from its nominal 54 MHz to a nominal 27 MHz upon ground command. This frequency plan, which is depicted in Figure 4 accomplishes an extraordinarily efficient utilization of the orbital location.

The satellite system design has been optimized for the transmission of television programs. The high power TWT's (operating at Ku-band) provide increased margin against propagation effects thereby permitting a substantial reduction in ground segment costs as well as a substantial increase in the number of locations where an earth station can be placed. The innovative frequency plan, employing transponders with switchable bandwidths, permits a single orbital location to provide 32 full power television signals into a single relatively inexpensive receive antenna. This ability to reconfigure each transponder individually and transmit the full transponder power in either full CONUS, Eastern or Western modes, the protection against the effects of uplink rain fade afforded by the DLA, the six spare TWT's and other on board satellite equipment redundancy, the use of switchable attentuators on the uplink to accommodate a variety of earth station transmit power capabilities and a satellite protection plan (described below) that calls for spare satellites (in-orbit and on the ground) to protect against the unlikely event of a catastropic satellite failure are all additional features of the satellite and satellite system that were designed to meet the needs of the CATV and SMATV industry.

K3/K4 PROTECTION PLAN

An earlier section (Why Ku-band) referred to the superior protection from catastrophe offered by a well designed Ku-band satellite system. The proposed K1, K3, K4 system discussed here will provide a level of protection unheard of in other satellite systems.

Satellite-protection requirements have grown along with our industry's revenue stream. In the early days of satellite distribution, protection involved relocating programming to other satellites in the event of failure. Later satellites enhanced reliability by incorporating on-board spare parts.

Today, our industry uses dishes at more than 25,000 commercial locations to receive satellite programming. The protection scheme for the 1990s must provide programmers with the ability to stay in business in case of catastrophic failure without repointing thousands of dishes overnight.

Components of a sound, cost effective protection system include (1) on-board protection (spare components); (2) in-orbit restoration capability where the satellites move, not the dishes; (3) ground spares. Such protection will be achieved with the K1, K3, K4 satellite system.

Once K3 is operational, the K1 satellite will serve as an in-orbit spare, which can be rapidly relocated to the K3 position in the event of a total K3 failure. Consequently, service restoration occurs in space, obviating the need to repoint thousands of dishes.

Furthermore, sometime after K3 is operational, K4 can be co-located with K3 (subject to FCC approval), operating 32 transponders from the same orbital location, achieving an even higher level of protection. In order to co-locate K3 and K4, the satellites need to be clones. Alone, each satellite has the highest ratio of redundant spare-parts to active-parts of any satellite placed in-orbit to date. When K3 and K4 are co-located, the sum of their spare TWT amplifiers, receivers, etc., is virtually the equivalent of a spare satellite. If a problem occurs, on-board spare equipment can be called upon to replace practically 50% of the critical parts on either satellite.

If it becomes necessary to replace any of the orbiting satellites, a compatible ground-spare satellite, constructed along with the active satellite, will be readied for launch.

The costs of the in-orbit and ground-spare capability are shared among all the users of the entire RCA satellite system, thereby minimizing costs to each.

KU-BAND CHARACTERISTICS AND SYSTEMS CONSIDERATIONS

For a given power radiated from the satellite, the earth station size and receive noise temperature (G/T needed) is determined by the performance required (the video signal-to-noise ratio) and the availablity or the fraction of time this performance will be provided. The availability is a function of equipment availability and propagation effects. The major propagation effects on satellite links at Ku-band are due to rain. (The largely ignored effects of sun outage at C and Ku will be discussed later.) These effects lead to the attentuation of the signal propagated through the atmosphere and an increase in the receive system noise temperature due to the warm rain.

The total attentuation of the signal (expressed in dB) over the earth-space path is of the form

$$A = \alpha L \qquad (1)$$

Where α is the rain attenuation in dB/km and L is the effective path length through the rain.

The effective path length depends on the elevation angle of the receiving antenna and on the rain rate. For low rain rates, such as 5 to 10 mm/hr (light rainfall), L is the true path length through the rain, while for high rain rates such as over 40 mm/hr (heavy rainfall), the non-uniform nature of the rain storm requires an empirically determined correction factor to compute the effective path length.

The degradation of the received C/N is a result of the signal attentuation caused by the rain along the signal path and the increase in system noise temperature caused by the warm rain. This total signal degradation due to rain is called Rain Loss and is given by:

$$\text{Rain Loss} = A + 10\log\left[1 + \frac{290}{T_s}\left(1-10^{-.1A}\right)\right] \text{dB} \qquad (2)$$

where T_s is the clear sky system noise temperature (°K) and A is the signal attentuation (in dB) caused by the rain as calculated in Equation (1).

The rain regions of the world and the United States have been identified and characterized as shown in Figure 5.

For a given earth station site, the rain rate statistics appropriate to the rain region in which the site is located are used. The elevation angle to the satellite is determined and the rain attentuation and the Rain Loss are computed as in equations 1 and 2 above.

The carrier-to-noise ratio under rain conditions is related to the clear air carrier-to-noise ratio by the following:

$$C/N_{rain} = C/N_{clear} - \text{Rain Loss}$$

Figure 6 shows the rain loss for a number of cities for a system noise temperature of 290°K for the satellite located at 85° West Longitude.

The information contained in the figure is the margins needed to accomodate the atmospheric effects of signal fading and increase in the receive system noise temperature at various locations throughout the country. The margins needed have been combined with the EIRP values for K1 and K3 to obtain the video signal-to-noise ratio versus availability across the United States for a given antenna size.

These results have been plotted for K1 on the contours shown in Figures 7. The results for K3 (with 60 watt transponders) are plotted in Figures 8 and 9.

SIGNAL OUTAGE AT KU-BAND

Discussions of Ku-band delivery systems inevitably turn to the topic of rain outage or rain fades. The general perception is that Ku-band is plagued by significant rain outages (program distruptions) and that C-band is not. In fact, this is the most often used argument against using Ku-band frequencies for distribution to CATV and SMATV systems.

Most of the force of this argument has been eliminated by the design of the K3/K4 satellites and the inclusion of sun outage into overall availability or outage calculations for C and Ku-band systems.

Most of the experience with Ku-band rain outage was gained on first generation Ku-band satellite systems. These systems used 10 watt or 20 watt satellites and spread this relatively low power across all 48 states. Receiving systems, using first generation electronics, incorporated low noise amplifiers with noise figures of 3dB or larger (the smaller the noise figure, the better the performance). The result was very low rain loss margins resulting in the rapid onset of signal degradation under rain conditions.

The K3/K4 satellites are designed to improve the rain loss margin. Each satellite carries sixteen 47 watt transponders (60 watt transponders with FCC approval). The result is about 4-7 dB of improvement in margin over the 10 or 20 watt satellites when K3 or K4 are in full conus mode and about 6.5 - 10dB of improvement when K3 and K4 are in half conus mode. A typical receiving system tôday includes antennas with an efficiency of 65% or better, resulting in a higher antenna receive gain, and a LNB with a typical noise figure of 2.5 dB or better.

These improvements in satellite and TVRO performance have dramatically increased the amount of rain fade that the system can tolerate before signal degradation becomes a problem.

A factor almost always overlooked when discussing signal availability is sun outage. Sun outage occurs twice a year in the Fall and Spring for a total of 18 - 20 days. Sun outage occurs when a given receiving TVRO, the satellite and the sun line up. During these occasions, random noise from the sun, to varying amounts degrades the received video signal. Sun outages are a fact of life, affecting C and Ku-band systems to differing degrees.

Table 1 shows the number of minutes over a year the received video signal to noise will fall to 45dB or less for a 4.5 meter C-band TVRO receiving a signal from RCA Satcom F3R and 1.8 meter and a 3.1 meter Ku-band TVRO receiving a signal for RCA K1 in full conus mode. As defined for rain outage, a video S/N of 45 dB or less caused by sun degradation is considered an outage. Using K1 in CONUS mode is a worst case situation since K3 provides over 1.25 dB more margin in CONUS mode than K1 and an additional 2.5 dB of improvement over CONUS when operating in the East or West beam modes.

TABLE 1
MINUTES OUT OF THE YEAR THAT S/N FALLS TO 45 DB OR LESS DUE TO SUN OUTAGE ONLY

C-band (4.5 meter TVRO, RCA Satcom 3R)	138 minutes
Ku-band (1.8 meter TVRO, RCA K1 CONUS)	22 minutes
(3.1 meter TVRO, RCA K1 CONUS)	20 minutes

Table 1 clearly shows that the C-band TVRO experiences significantly more sun outage than the K-Band TVRO. The availability for this C-band system can be shown to be 99.97%. The availabilities shown in Figures 7, 8 and 9 for various Ku-band receiving systems demonstrate there is very little difference between C and Ku-band availability.

KU-BAND RECEIVING SYSTEM DESIGN

System Design

Given the inescapable benefits of Ku-band delivery, the optimized satellite operating characteristics and the unrivaled protection plan offered by the satellite system, close attention must be paid to the TVRO system to ensure that the installed ground segment does not become the weak link in the overall system.

Program distribution to CATV and

SMATV systems requires very careful system design, installation and operation to ensure continued high performance.

Crimson Satellite Associates have identified certain performance requirements to be achieved by TVROs utilized to receive the K1 and K3 distributed programs. See Table 2. The TVRO systems distributed to cable systems by HBO and RCA American Communications, Inc. ("RCA Americom") were selected to conform to these performance requirements.

TABLE 2
SIGNAL AVAILABILITY AND OUTAGE DEFINITION FOR RAIN LOSS ONLY

Signal availability CATV :
99.99% (goal)
Signal availability SMATV:
99.9%

Signal outage occurs when:
S/N_{Video} equals or falls below 45dB

Clear sky S/N_{Video} :
50 dB or better

Signal availability, as previously described, is the percentage of a year that the signal received by a CATV or SMATV TVRO exceeds the specified outage threshold level. From the above table, a CATV system should provide to the extent possible a video signal with a S/N greater than 45 dB 99.99% of the year and an SMATV system must perform to that extent 99.9% of the year. In other words, a CATV system should have an outage as defined in Table 2 of no more than 53 minutes out of the year and an SMATV system for no more than 530 minutes out of the year. Two very important points must be recognized and considered when evaluating the signal availability requirements in Table 2. First, the availablity here refers only to signal degradation caused by rain or atmospheric conditions. It does not include other system induced degradations such as sun outage. Secondly, the term outage as used in this analysis is misleading. For the purpose of conservative system design and to insure system margin, a signal outage is defined by Crimson Satellite Associates as a video signal exhibiting a S/N of less than 45 dB. A 45 dB video signal is still a very acceptable picture and provides plenty of margin from that point to where the picture is actually unusable. If a lower S/N is used as the outage criteria, system availability improves.

The selection of Ku-band TVROs distributed by HBO and RCA Americom was made as follows:

From the requirements of Table 2, the Rain Loss for multiple sites in each rain zone was calculated. This Rain Loss described the amount of margin between clear sky performance and the defined outage to be provided by the TVRO. Equation 3 and 4 were then used to calculate the required TVRO gain needed to provide that margin.

$$C/N = S/N_V - RTF \quad (3)$$

$$G_A = C/N - EIRP + Ls + Lp - 228.6 + 10\log Ts + \text{Rain Loss} + 10\log B \quad (4)$$

Where:

RTF = Receiver Transfer Function
G_A = Antenna Receive Gain
Ls = space Loss
Lp = Pointing Loss
Ts = System Noise Temperature
B = System Noise Bandwidth

Note that the required gain was calculated for the worst case operational conditions; K1 EIRP in full CONUS mode on the lowest powered transponder. Half CONUS mode operation will only improve availability. Also, K3 provides an increase in CONUS and half CONUS EIRP compared to K1, thus improving performance further.

If the calculated gain did not achieve at least a 50 dB Video S/N under clear sky conditions, the gain of the system was increased accordingly. Available TVRO components were then selected to build the required TVRO system.

Figures 7, 8, and 9 show the required antenna sizes for a given availability for K1 and K3.

The 22 minute sun outage discussed earlier reduces the availability shown by .0042%.

System Installation and Operations

Ku-band TVROs require slightly more

care with installation and operation than a C-band TVRO. While the special requirements are not onerous, they must be recognized if the systems are to perform at maximum capacity.

The wavelength of a Ku-band signal is about 1/3 that of a C-band signal. This means that the main beamwidth of a Ku-band antenna is about 1/3 the beamwidth of a C-band antenna of the same size. The reduced beamwidth imposes a requirement to aim the Ku-band antenna more accurately than a comparable C-band antenna and to hold the antenna as still as possible. Movement of a CATV or SMATV Ku-band antenna can cause a signal degradation of 2 or 3 dB. This amount of degradation is not critical under most operating conditions but it does reduce the system rain loss margin. Additionally the reduced wave lengths of Ku-band frequencies requires the antenna reflector surface smoothness to be held to a tight tolerance and the shape of the antenna be as close as possible to the required parabolic shape.

The installer of a Ku-band TVRO must take steps similar to the one he would take when installing a C-band TVRO to ensure the pad and all reflector mounting hardware is rigid and capable of holding the reflector as motionless as possible in expected wind conditions. Also, during reflector assembly, care must be taken not to distort the reflector surface by over torquing bolts or denting or bending the reflector.

Antenna pointing and polarization alignment of a Ku-band TVRO is similar to that of a C-band TVRO. However, because of the very high S/N of the video signal received under most conditions, the installer can be misled about the accuracy of his alignment efforts. A Ku-band TVRO, under most operating conditions, produces a video signal far superior to that produced by a comparable C-band system. In fact, the quality of the received picture is so good that even a 2 or 3 dB degradation in the received signal C/N due to antenna misalignment is not noticable by viewing the picture on a TV set.

However, the result of antenna misalignment is to cause the system margin to be reduced. The reduced margin in turn causes the defined outage level to be reached quicker than predicted, thus reducing the signal availability.

It is recommended that the operator align his Ku-band TVRO using a field strength meter or if possible a spectrum analyzer instead of viewing the received picture quality on a TV set. Additionally, as with C-band TVROs, the antenna alignment should be peaked periodically when the satellite is in the center of the box.

CONCLUSION

This paper has attempted to familarize the reader with Ku-band satellite requirements, the advantages of Ku-band distribution and system operating considerations.

Ku-band provides: (1) the opportunity to operate a TVRO in an interference free environment thus opening up additional business opportunities; (2) unheard of protection against business disruption and ground segment disruption in the event of a satellite failure; (3) video signal performance superior to C-band for comparible antenna sizes; and (4) signal availability equivalent to C-band when both rain fade and sun outage effects are considered.

The C-band satellites used by the cable industry reach the end of their design lives by 1992. Therefore, 1987 is the year to focus on the transition to the next generation of satellites. A short term re-location to a satellite with 2 or 3 years additional life is not a solution. In fact, it puts the industry at a competitive disadvantage with the evolving technology and actually presents the risk that a launch opening will not be available when a successor satellite needs to be launched.

The satellites of the future will be state-of-the-art Ku-band. They will be capable of affecting both cable operators' and programmers' businesses and may be operated by companies with little or no investment in our business. The cable industry cannot afford to restrict itself to yesterday's technology. Today, with Ku-band technology, we have the opportunity to maintain a competitive technological edge - an opportunity that has come at a time that is a natural decision juncture for satellite capacity planning.

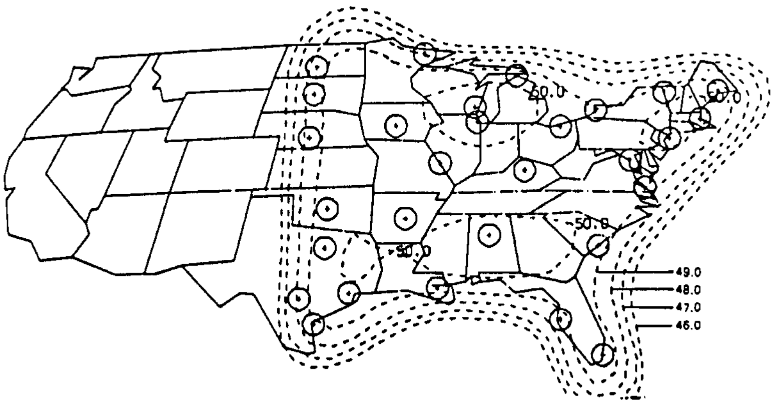


FIGURE 1*
EIRP CONTOURS (dBW) FOR EAST BEAM

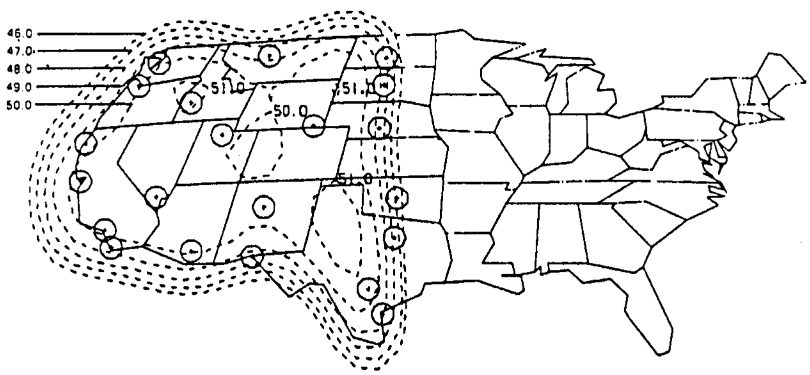


FIGURE 2*
EIRP CONTOURS (dBW) FOR WEST BEAM

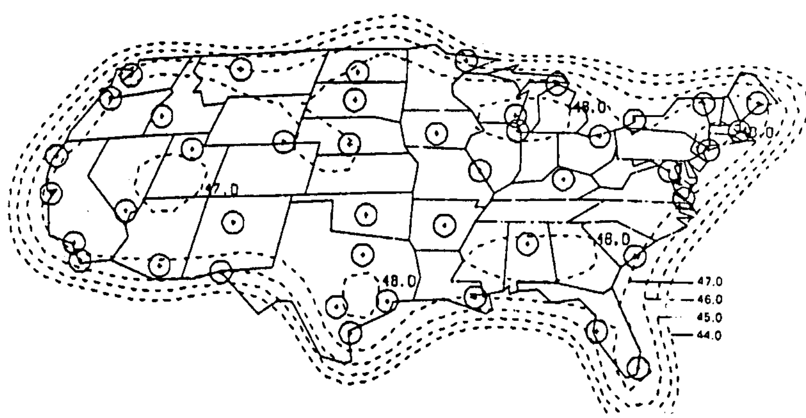
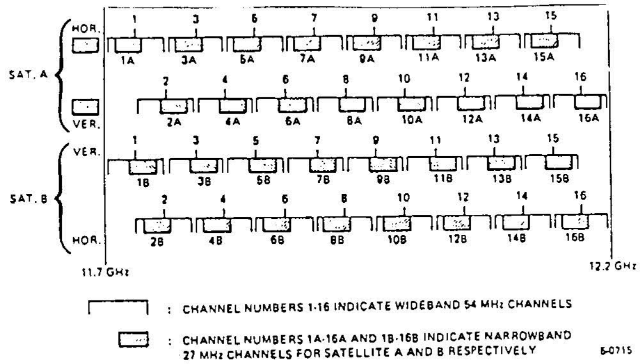
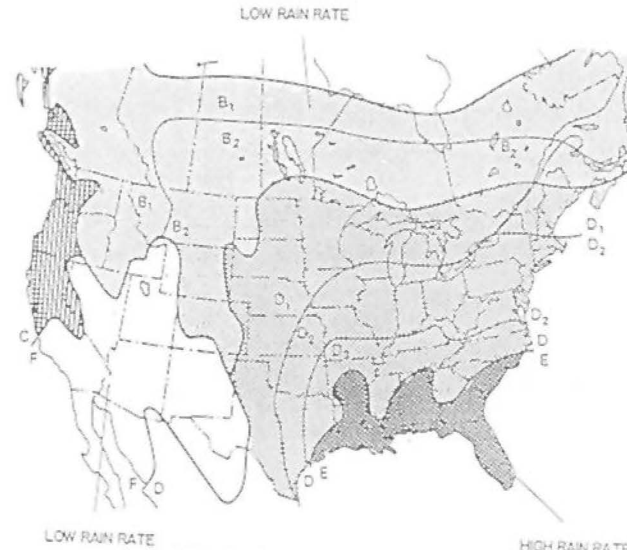


FIGURE 3*
EIRP CONTOURS FOR CONUS

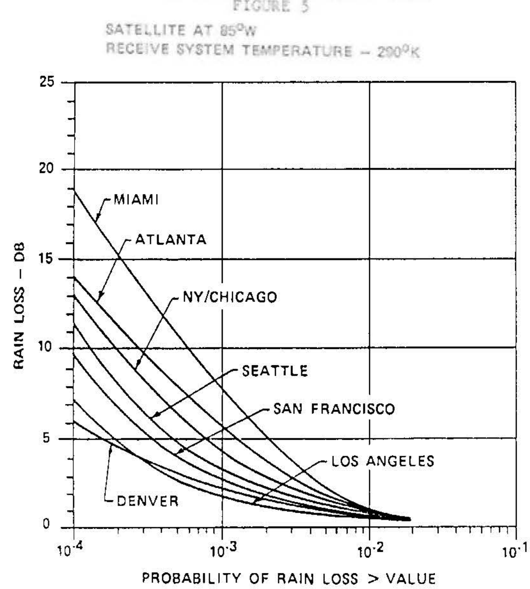
*CONTOURS FOR 60 WATT TRANSPONDERS (PENDING FCC APPROVAL). 47 WATT TRANSPONDERS' CONTOURS ARE REDUCED BY ABOUT 1 dB FROM THOSE SHOWN.



K3/K4 CHANNELIZATION PLAN
FIGURE 4



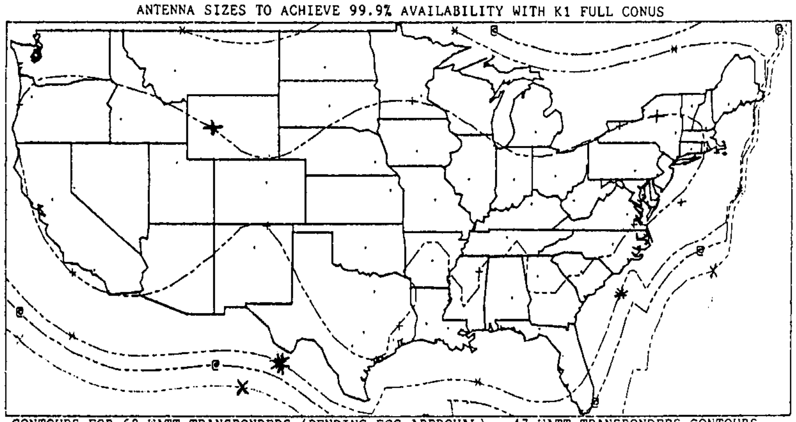
RAIN REGIONS SOURCE: NASA
FIGURE 5



RAIN LOSS AT VARIOUS CITIES
FIGURE 6

FIGURE 7

ANTENNA SIZES TO ACHIEVE 99.9% AVAILABILITY WITH K1 FULL CONUS



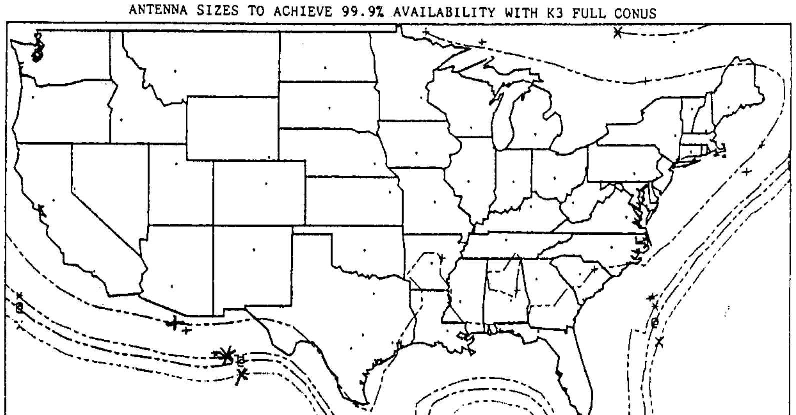
CONTOURS FOR 60 WATT TRANSPONDERS (PENDING FCC APPROVAL). 47 WATT TRANSPONDERS CONTOURS ARE REDUCED BY ABOUT 1 dB FROM THOSE SHOWN.

+ 1.8 METERS @ 3.7 METERS

* 3.0 METERS X 5.5 METERS

FIGURE 8

ANTENNA SIZES TO ACHIEVE 99.9% AVAILABILITY WITH K3 FULL CONUS



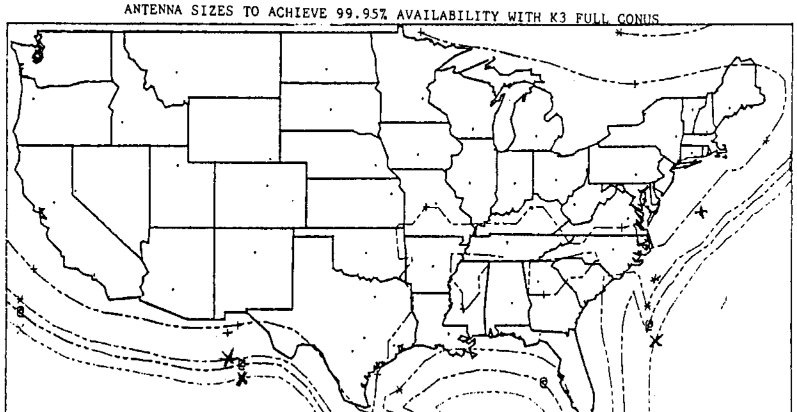
CONTOURS FOR 60 WATT TRANSPONDERS (PENDING FCC APPROVAL). 47 WATT TRANSPONDERS' CONTOURS ARE REDUCED BY ABOUT 1 dB FROM THOSE SHOWN.

+ 1.8 METERS @ 3.7 METERS

* 3.0 METERS X 5.5 METERS

FIGURE 9

ANTENNA SIZES TO ACHIEVE 99.95% AVAILABILITY WITH K3 FULL CONUS



CONTOURS FOR 60 WATT TRANSPONDERS (PENDING FCC APPROVAL). 47 WATT TRANSPONDERS' CONTOURS' ARE REDUCED BY ABOUT 1 dB FROM THOSE SHOWN.

+ 1.8 METERS @ 3.7 METERS

* 3.0 METERS X 5.5 METERS

LAB TEST METHODS FOR COMPARISON OF CORROSION PROTECTION PRODUCTS

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RAYCHEM CORPORATION

ABSTRACT

Bare metal connectors can be susceptible to corrosion especially in humid and saline environments. Various methods of cold-applied corrosion protection are available - silicone grease, paint-on coatings, mastics, self-fusing tape, and gel-based sealant strip. Some simple lab tests can be used to evaluate their comparative corrosion performance, and their respective effectiveness compared with no protection.

The paper describes 1)test sample preparation, including sources for materials. Samples range from thin metal films, metal discs, and wrapped metal bars, to connectors installed on short cable lengths. 2)how to calculate a corrosion rate for carefully prepared samples. 3)corrosion rates and other data for several methods of protection in various environments. These environments are : ASTM artificial seawater at 60degC, 5% sodium chloride solution, ASTM salt fog, "CASS" accelerated salt fog, and weatherometer (combined UV and water spray) exposure.

Other factors relating to product lifetime should also be considered when selecting a corrosion protection product. This paper focuses on corrosion performance as a key factor. All of the corrosion environments described require relatively long-term testing. The fastest method is 60degC seawater - which can be set up at low cost, using simple samples, to give a good comparison between various corrosion protection products.

INTRODUCTION

Cable television systems are becoming more complex, extensive utilities. System lifetimes are crucial not only to reduce maintenance and enhance cash flow, but also to maintain a pleased, growing customer base. Interactive systems will place still higher demands on reducing downtime. Since cable television is a serial system, reliability is a function of the weakest link. Corrosion is often the cause of that weak link. In aerial structures, salt fog and salt spray environments in coastal areas can rapidly deteriorate unprotected metals. Ground boxes and manhole systems face additional problems with ground water and, in northern areas, saltwater runoff from snow covered roads.

The predominance of aluminum in these systems can complicate corrosion problems. Aluminum is high on the galvanic series, therefore it sacrificially corrodes when electrically coupled to metals lower in the series such as steel. Inadvertant couples in ground boxes and manholes can result in catastrophic corrosion.

The corrosion process is complex. It can involve faster corrosion on hot metal such as amplifiers, and crevice corrosion in stressed areas such as coaxial cable bends or coupling threads. Acid rain will also affect corrosion rates and temperature cycling complicates the analysis of corrosion problems. A black body on a pole in Pittsburgh, for example, can reach 140°F during a sunny fall day and drop to 30°F during a cool night 12 hours later.

Corrosion, like fire, requires three conditions to happen. First, two metals or the same metal at different energy levels, must be present. This always occurs because of slight differences in stress level during manufacture for example. Second, the two metal areas must be electrically connected. In a serial system, they always will be. Finally, there must be electrolyte present. Water or water vapor in the environment is an electrolyte—especially if any salt is present. By cutting off one of the three conditions, corrosion can be stopped. In a serial system, the only practical method is to eliminate electrolyte—by coating to reduce the presence of water (or humidity) at the metal surface.

Finding the "best" corrosion prevention product involves evaluating its ability to withstand temperature extremes of the environment, ease of installation in a variety of conditions, applicability to a variety of equipment, the need to re-enter for maintenance, and the key ability to prevent water reaching metal structures.

A simple lab test is needed to compare corrosion prevention products; preferably a test to take into account the above conditions, in a reasonable time period. Several test methods are available ranging from visual observation to quantitative measurements.

TEST METHODS

Various test samples and standard environments were tried. Samples included

a) new distribuition connectors installed between sections of half inch jacket coax cable

b) copper mirrors-a thin layer of vacuum deposited copper on a glass slide

c) 120 grit finish 1" diameter 1010 steel discs, 0.06" thick

d) smooth barstock 1018 steel mandrels, 0.38" diameter, 5" long

Distribution connectors were evaluated visually to determine whether the corrosion protection had allowed water to corrode the connector (which over time would cause electronic noise and eventually metal failure).

Copper mirrors yielded a pass/fail rating after exposure—they either corroded away or were intact. Both the steel discs and steel mandrels were used to give a quantitative measure of corrosion rate in terms of mils of material gone on average over one year—mpy, or mils per year. Mild steel was used because of its susceptibility to water and a faster corrosion rate than aluminum. Enough samples were used to allow samples to be removed from the test at various time intervals to measure corrosion rates vs. time. For quantitative work, careful sample preparation was required. First, the metal was cleaned:

1. degrease in isopropyl alcohol
2. wash with detergent such as MICRO-SOAP
3. rinse in distilled water; dip in dilute hydrochloric acid (about 20 seconds)
4. rinse; rinse in acetone
5. dry in a dessicator

The samples were not touched with bare hands, which could leave corrosion sites, and were accurately weighed when completely dry. After covering with the corrosion protection and exposing the sample to a corrosive environment, each sample was rinsed in cool tapwater and blotted dry before removing the corrosion protection. The metal was gently cleaned with a nylon scourer to remove any surface corrosion products. Note: the presence of any pitting should be recorded in this kind of test, since it invalidates the corrosion rate calculation and can itself be a severe corrosion problem. In these tests the corrosion was a uniform layer. After gentle abrasion the samples were dipped in dilute hydrochloric acid until signs of rust had just disappeared.

This was followed by rinsing in water, then acetone, and thorough air-drying. Accurate weighing enabled the weight loss to be calculated. The equation used to calculate the corrosion rate was:

$$\text{corrosion rate (mpy)} = \frac{534 \times W}{D \times A \times T}$$

where W = weight loss of the sample, in milligrams

D = density of the metal used, in gm/cc

A = surface area of the sample, in square inches

T = time in the corrosive environment, in hours

Methods of applying the corrosion protection obviously varied. For tape-like products, mandrels were wrapped and discs were "sandwiched" between two flat layers with or without overlap seams. These sandwiches were securely clamped around the edges of the tape product to ensure that only the tape, or the overlap seam, was tested. The clamp was a pair of plastic rings somewhat larger than the diameter of the steel disc, bolted together with PVC nuts and bolts. The same kind of sandwich procedure was used for copper mirror samples. A variety of test environments was used. Standard ASTM D2565 weatherometer testing involved intense ultraviolet exposure plus alternating wet and dry cycles to simulate outdoor conditions. Salt fog tests using a 5% NaCl mist were done in accordance with ASTM B117.. Accelerated salt spray (CASS) used 5% NaCl mist with copper chloride and acetic acid to accelerate corrosion, in accordance with ASTM B368. To simulate manhole conditions, saltpool immersion testing was done at 30°C, in a four foot depth of circulating 5% NaCl solution.

For small lab samples, a 5% solution of ASTM artificial seawater at 60°C was a good corrosion test. The seawater solution and samples were contained in wide-mouth polyethylene jars with a loose cover to reduce evaporation; the jars were maintained at 60°C in a simple heated water bath.

For all connector testing, thermal cycling was run before the sample was corrosion tested. Thermal cycling consisted of 20 cycles between -40°F and +140°F to simulate environmental conditioning. Three cycles a day, with three hours at each temperature extreme, were run. This is sometimes referred to as "Bell" cycling since it originated in the phone system.

RESULTS

The corrosion rates over time for various environments are shown in Figure 1, for both unprotected samples and ones protected with silicone gel tape. Immersion in 60°C saltwater gave the fastest quantitative results, with reproducible corrosion rates after 1000 hours. Salt fog and weatherometer tests were useful, but required more expensive test facilities. The samples referred to in the graph were mild steel discs and mandrels, which gave the same rates. Copper mirrors yielded a semi-quantitative result—the time that corrosion failure was delayed was indicative of how much the corrosion process had been slowed down—see Table 1.

Several different corrosion protection products were compared in the lab. For example see Figure 2 which gives corrosion rates for various products wrapped on mild steel mandrels and immersed in 60°C seawater solution.

Simple exposure to a corrosion environment indicates how good a seal is obtained when first applied. For outdoor applications a more realistic evaluation includes thermal cycling. This was found to be very important when comparing corrosion protection methods--mastic tapes, self-fusing rubber tapes, paint-on compound over vinyl tape, silicone grease and silicone gel tapes.

During thermal cycling the coated-over vinyl tape contracted, allowing subsequent water ingress and corrosion. The installation process was also time consuming, although re-entry was reasonably quick. Self-fusing hydrocarbon rubber tapes split during thermal cycling. Since they did not bond to the metal substrate they allowed water ingress under the tape, visible as a spiral corrosion path along the aluminum to the connector. Badly corroded areas were noted under the splits; re-entry was easier than mastics. The mastic tapes were easy to apply and seemed to give good installations. Samples were placed very carefully in the thermal chamber because of the tendency of mastics to get very sticky when warm. After corrosion testing—and after much difficulty removing the mastic—all five samples showed water ingress and local corrosion underneath.

Silicone grease, while messy to apply, did somewhat better. Apart from corrosion on sharp edges where it was difficult to ensure coverage, the connectors and cables were mostly dry and shiny. There was a tendency for the grease to crack during thermal cycling--which could lead to longer term problems. Of course the grease was also very susceptible to removal and damage, and was messy to re-enter.

Gel tapes were easy to apply and remove. When applied correctly they gave excellent performance through thermal cycling and corrosion testing. Due to the softness of the material it was easily scuffed. However, any corrosion was limited to the area of metal exposed; water did not travel under the tape to other connector areas.

Table 2 is a summary of tests in the three fastest corrosion environments--CASS accelerated salt fog, standard salt fog, and saltpool immersion. Unprotected samples showed severe corrosion—in places right through the aluminum sheath—after 900 hours in saltpool or salt fog, or after 275 hours in CASS salt fog. Corrosion of the taps looked similar to the degree of corrosion seen in coastal cable TV systems after typically 6 months to 2 years unprotected.

CONCLUSIONS

Bell temperature cycling following by saltpool immersion is a good method for cable television engineers to try out corrosion protection products, on standard connectors. Applying and removing the products also indicates the ease of use in the field. This test does take about 1000 hours (6 weeks) and requires a temperature cycle chamber plus a saltwater tank. An accelerated test is CASS salt fog, which will give useful results in as little as 275 hours (12 days) after thermal cycling. However, it requires a special chamber to provide a carefully controlled salt fog.

At present there are no cable TV industry standards for corrosion protection products, and we do not claim to know all of the requirements for such applications. It is very clear that temperature cycling is a crucial test. Without it products may give good lab results and fail during field use. The test can be run automatically with a suitable environmental chamber, or it could be approximated using a simple oven and home freezer, changing the samples between the two manually three times a day.

For small lab samples a simple well-controlled corrosion test can be done using mild steel mandrels, and a 60°C seawater or saltwater solution corrosion test. This requires only simple lab facilities—a water bath, accurate scales (for corrosion rates) and simple chemicals for the cleaning procedure.

With more expensive complex cable television systems, corrosion effects on lifetime can be crippling. In this paper we have concentrated on simple corrosion testing as a key test. We have not evaluated signal quality after corrosion. Also in comparing product performance the specifying engineer should consider evaluation of such things as long-term effects of heat, cold, ultraviolet exposure, etc. By using a series of simple tests, the engineer can easily decide on the best corrosion prevention product for their system.

SOURCES OF MATERIALS

1. Copper mirrors (0.25" x 1.0" x 0.06" glass substrate):
   Evaporated Metal Films, Inc.
   701 Spencer Road
   Ithaca, NY 14850
   (607) 272-3320

2. 1010 steel discs (1" diam., 0.06 thick 120 grit finish):
   Metal Samples, Inc.
   P.O. Box 8
   Munford, Alabama 36268
   (205) 358-4202

3. Artificial seawater compound (ASTM D1141):
   Lake Products Co., Inc.
   P.O. Box 498
   Ballwin, MO 63011
   (314-536-1600

Table 1
Time to onset of corrosion of copper mirrors

environment	time for unprotected mirrors	time for mirrors protected by silicone gel strip
weatherometer	200 hr	3200 hr
60°C seawater	24 hr	3000 hr

Table 2
CORROSION PREVENTION PRODUCT TESTING RESULTS

Corrosion Prevention Product	Salt Fog Time (hrs)	Salt Fog No. of Samples	Results	Saltpool Immersion Time (hrs)	Saltpool Immersion No. of Samples	Results
Mastic Tape	1000*	1	Poor. Sample wet & corroded. Difficult to remove.	930	4	Poor. Sample wet & corroded. Difficult to remove.
Vinyl Tape with Paint-On Compound	1000*	1	Poor. Vinyl pulled open during cycling. Corrosion	910	2	Poor. Vinyl opened up. Connectors wet & corroded.
Silicone Grease	300**	2	Good. Grease had cracked. Corrosion only at sharp edges Difficult to apply evenly.	910	2	Fair. Some cracking & corrosion at cable jacket.
Self-Fusing EPR Tapes	275**	8	Poor. Split during cycling. Water ingress and corrosion under tape.	930	11	Poor. Split during cycling. Corrosion at cable jackets with water spiraling to connector.
Silicone Gel Tapes	275**	16	Excellent. No corrosion. One sample slightly scuffed with no corrosion at damage	930	16	Good to Excellent. Corrosion where product scuffed off; but corrosion limited to scuffed area.
	670**	4	Excellent. No corrosion			

*Standard Salt Fog
**CASS Accelerated Salt Fog

FIG.1–CORROSION RATES vs. TIME, VARIOUS ENVIRONMENTS

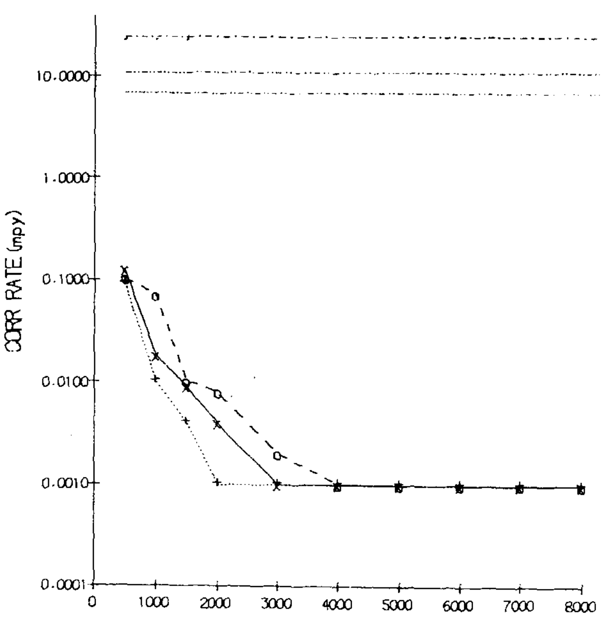
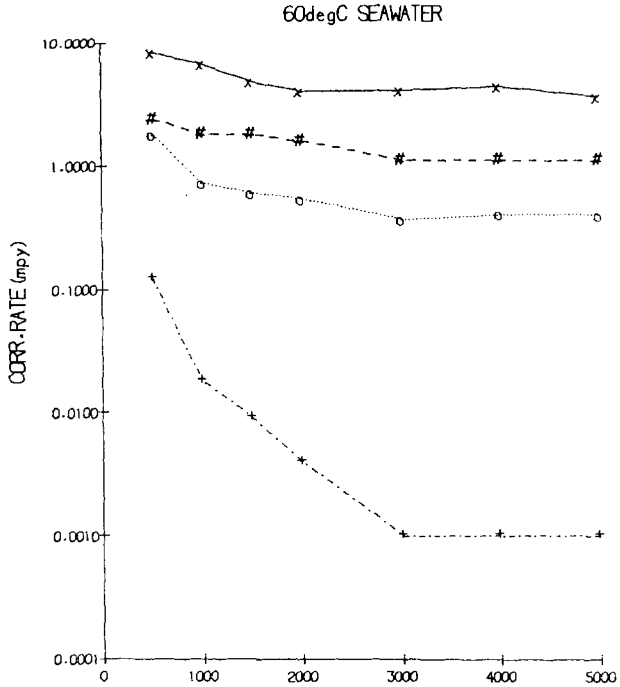


FIG.2 – CORROSION RATES FOR VARIOUS PRODUCTS
60degC SEAWATER



Launching An ANI Passing Impulse PPV System

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Centel Cable Television Co.
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Vito Brugliera

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Zenith Cable Products Division
Zenith Electronics Corp.

ABSTRACT

Pay-per-view is currently viewed as an important new source of revenue for cable systems. There are several technological approaches to the problem of order taking, among them is the use of Automatic Number Identification (ANI) based technology. ANI information: called and calling telephone numbers, identifying event and subscriber, are passed to the cable system headend where a system controller processes the impulse transactions and authorizes subscriber addressable converters in real-time without additional in-home hardware using rotary or touch-tone telephones.

INTRODUCTION

Impulse Pay-per-view (IPPV) is expected to be a new source of revenue for cable systems. However, there are a variety of transaction technologies available to accomplish impulse order taking. The dilemma facing managers and engineers is, how can this market be tapped without major capital outlay and increased operating expense? The solution selected for Centel's Traverse City system was launching an ANI passing IPPV system.

Traverse City, Michigan, was the site of an ANI IPPV trial for Centel Cable Television Company, launched May 1, 1986. The system uses Zenith Z-TAC addressable converters and CableData as a billing system. ANI information for 5600 Phonevision subscribers was supplied by Michigan Bell through the use of Science Dynamics, Multi-Access Cable Billing System (MACBS).

The criteria for selecting the order taking technology, as well as the experience of dealing with other variables in launching an ANI based IPPV system in Traverse City, Michigan, will be discussed. Among the factors discussed are:

- Network capacity and holding times
- Peak ordering
- Prefix considerations
- Third party ANI equipment
- Telco involvement
- System description
- Order and transaction sequences

The considerations before launching the ANI passing IPPV system and the experience gained through operation will also be discussed. Market research and customer reaction to the IPPV service will also be presented.

SYSTEM OVERVIEW

Automatic Number Identification (ANI) has been in use in telephone systems for quite some time. 800 and 900 prefix telephone numbers are a ready example. 800 numbers provide users free calls by charging the called number, which is likely located in another state. When such a number is called, the telephone network performs the following:

- recognizes the call as special and routes it outside the local central office switch through another part of the network for completion

- captures the called and calling numbers for billing purposes at a later time

Zenith, while working on its Z-VIEW two-way IPPV cable system, had developed a new system controller and software that permitted real time capture of two-way orders at 150 per second. Z-VIEW did this by detecting within the upstream data packet the ID of the in-home transmitter and cross referencing that to the subscriber addressable converter ID. After credit checks, the addressable converter was authorized.

The unique element was that the PPV subscriber database was contained in active memory in the system controller, and not on disc. This avoided any disc operating time and allowed near instantaneous look-up, processing and authorization. In a typical cable system the system controller/encoder authorization is driven by the billing computer - a process that can be measured in tens of seconds or longer. This was avoided by having the PPV database in active memory within the system controller, updated and tracking the billing computer database through normal edit commands. However, during the IPPV event ordering period, the transactions are in active memory and are recorded on disc by a background program within the software. After the event, the transactions are uploaded to the billing computer at a convenient time. This is a store and forward at the headend PPV system.

This order taking technology was translated to an ANI passing IPPV system, Phonevision, by recognizing that:

- the called number could be used for event identification
- the calling number could be used to identify the subscriber and converter ID
- called and calling numbers are provided by ANI
- ANI passing would be outside the local central office switch

The last item is crucial, because it means that the telephone newtwork would not be subject to the peak loading problems associated with other PPV systems using the telephone as the return path

In the Centel Traverse City cable system, PPV events are chosen by tuning to the barker channel. An event is selected and the corresponding toll-free telephone number is called, either a touch tone or rotary dial phone can be used since the ANI technology is not touchtone dependent. The IPPV subscriber call is routed through the local Electronics Switching System (ESS) to the Multi-Access Cable Billing System (MACBS). The system architecture is shown in Figure 1.

Optionally, there may be a Digital Multiplex System (DMS) used as a tandem and concentrator to link several ESS's or switches into one MACBS.

The MACBS takes the ANI information from the local ESS, translates it into a ASCII data packet. A pre-recorded digitally encoded voice response is heard, thanking the customer and requesting they hang up. Then, the packet is sent to the appropriate headend, with the called number indicating the event, and the calling number identifying the subscriber. The MACBS installed at Traverse City can send calls to four different headends or cable systems.

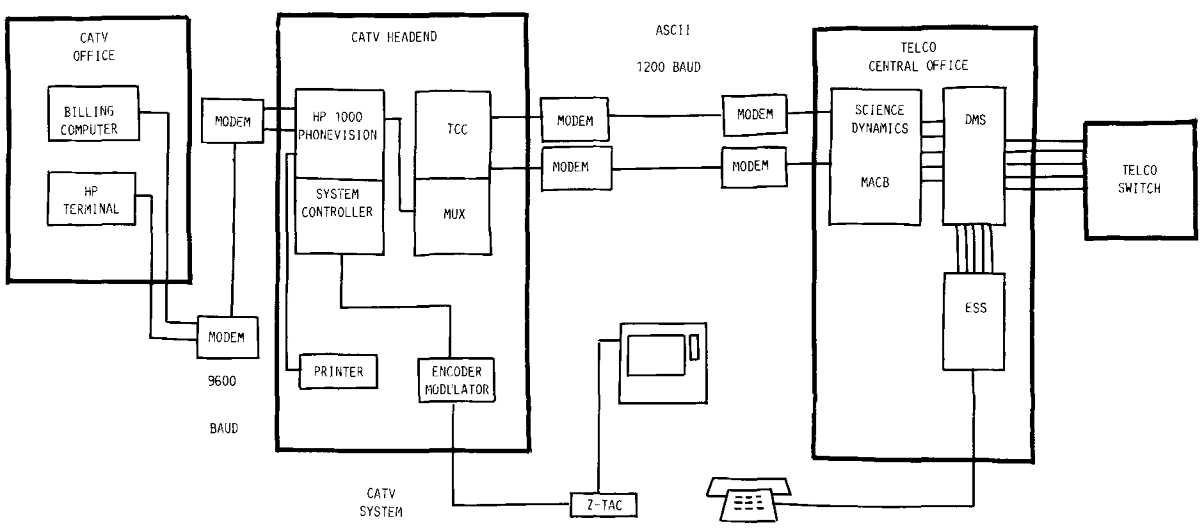

FIGURE 1. SYSTEM DIAGRAM

In the headend, the data is routed from the modems to the Telephone Communication Controllers (TCC's). The TCC verifies the protocal and holds the data in a buffer until the Multiplex (MUX) selects one of up to 16 TCC's. The Mux buffers and transfers the data into the HP1000 host computer.

The HP host computer searches the data base, previously downloaded from the billing system, using the calling number to identify the subscriber. If the subscriber has an addressable converter and is authorized to receive PPV, the HP acts as a standard system controller and authorizes the converter for the requested event by sending a downstream authorization through the channel encoder.

Upon completion of the event, all converters are mass de-authorized. Billing information remains in the HP's hard disk until purged, following a successful upload to the billing system. The storage of the transactions for post event processing is a positive feature for the cable operator. It permits IPPV even when the billing computer is down. The billing computer can be down for a variety of reasons:

- early morning down time for report generation
- modem or line failure
- billing system hardware failure

The last two are infrequent, but would represent revenue loss if the IPPV system was tied directly. The first item is fairly routine, and means that incremental revenue can be had from the night owls in your system.

TELEPHONE SYSTEM CONSIDERATIONS

The potential of an ANI based IPPV system is dependent on the percentage of your customer base the telephone company can deliver. To determine the percentage, the telephone company needs the number of subscribers by prefix, or NNX, you wish served. From the prefixes, the telephone company can determine which switches are involved.

Switch Considerations

A network could then be designed to link the switch under consideration into one or more MACBS. The number of ANI sending trunks used to link each switch to the MACBS is dependent on the subscriber base served, the expected call volume, and the holding time per trunk. Newer switches likes a 1ESS, 1AESS, 5ESS and DMS100 can be modified to pass the needed ANI to the MACBS by entering new translations.

The 2ESS and 2BESS switches require a software change called an Office Data Assembler (ODA) run. Due to the cost of an ODA run, this may delay the additions of those switches until the next scheduled software update. Older electro-mechanical switches may require physical rewiring. The Number Five Cross-Bar is capable of passing the ANI if strapped properly. Step-by-step switches may or may not be capable, if rewired, depending upon their ability to pass the ANI. All of these switches need ANI passing trunks to send the ANI to the MACBS, the number of trunks is dependent on the expected call volume.

The local telephone company staff determines what equipment changes are needed. Only after a timetable has been given showing when each prefix is available, can the true potential of an ANI system be explored.

Holding Time per Trunk

The amount of time that a customer remains on the line during a transaction is the holding time per trunk. Over the first ten months of this trial, the average holding time has been 16.5 seconds. The order message used by the MACBS in this trial was, "Your order has been accepted. Thank you. Please hang up." Shortening the message to just "Thank You", tightening up the switching time and automatically disconnecting the customer will shorten the holding time to approximately ten seconds per transaction.

At ten seconds per trunk, for a 24 trunk system, the maximum throughput per minute would be 144. The Science Dynamics MACBS has a capability of 128 trunk, at ten seconds per trunk 768 calls per minute throughput is possible.

It was expected that the majority of the order entry demand would fall in the last 5 minutes before the start of the event. However, the demand was much more evenly spread out over the 45 minute order entry time allowed before th3 start of the event. See Figure 2.

Ending the allowable order entry time at the start of the event resulted in a 18% rejection rate for calls placed after the start of the event. Allowing orders to be accepted up to 10 minutes into the start of the event reduced the call reject rate to 2%. This change in the allowable order entry time was recently introduced and is not accurately reflected in Figure 2.

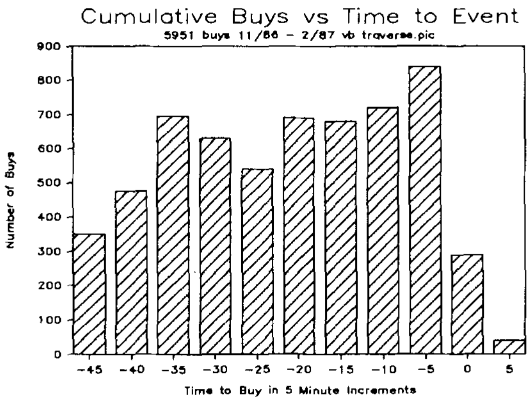

FIGURE 2. BUYS vs. TIME FOR IPPV EVENT

Database Accuracy

Once the telephone company has made the ANI information available to the cable system, some important links need to be made. First, the phone used to place the order must be linked to the subscriber, then the subscriber to their addressable converter. With the customer being identified by the phone used to place an order, the success of an ANI system is dependent on the accuracy of the data base. The particular home phone number being used as the identifier in a multi-line house or work phone number is very important, since an impulse pay-per-view order will be placed where the television is located in the home. Thus, the selection of the correct phone number among possible alternatives is very important. Having the correct customer home phone number on file to order IPPV needs to be clearly communicated.

This was accomplished through the barker channel, bill stuffers, and direct mail pieces. Also, verification of those subscribers with a listed phone number was done through the telephone company directory assistance department. Additionally, all customer contact with the CSR's includes phone number verification.

With the present software, only one converter per account can receive PPV. This converter is designated on CableData as the primary outlet and should be the only converter with its PPV flag at YES at a particular account. The link between a customer's phone number and their primary converter can be created only if the PPV flag is at YES. Therefore all converters entering the data base through the converter inventory program go on the customer's account with the PPV flag at YES. It is the responsibility of the CSR's to turn all but the primary outlet converters PPV flags to NO when starting new accounts or exchanging converters.

It's necessary to communicate to the subscribers that only one converter in their home is capable of receiving PPV. So if they order a PPV event and it does not come on the outlet they are watching, to check all the outlets. Customers with multiple converters should be informed as to which of their converters will be listed as their primary converter.

Periodically, a refresh is done to match the controller data base with the billing system. This refresh maintains the link between the customer's home phone number and a PPV capable converter. It is important that only one converter has its PPV flag at YES. In the event that more than one converter's PPV flag on an account is at YES, the phone number will be linked to the lowest converter number, which may or may not be the primary outlet. In this case, the customer may be rendered PPV incapable.

Equipment Utilization

There has been concern expressed about the peak ordering transient load, particulary by store and forward advocates, that may adversely impact portions of the local telephone network. Part of this concern may be caused by their use of an abnormally short order entry window: five minutes, prior to start of the event. Actual experience at Traverse City indicates that given a sufficiently wide order entry window, 45 minutes before and during the event, that peak loads capable of impacting the telephone network are never encountered.

Michigan Bell and Science Dynamics designed the ANI passing system to handle up to four cable headends. There are two banks of 12 trunks (24) and two banks of eight Multi-Frequency Receivers (MFR) (16). The loading of the MFR has also been a source of concern for some. Data was taken for four events with starting time and orders entered:

1:00 am	17
7:00 pm	144
9:00 pm	693
11:00 pm	22

Of the 16 MFR's available, the usage was: two MFR's handle 70 % of the incoming calls; an additional two MFR, 19.6 %; two more MFR beyond that, 6.8 % and three more MFR handle the remaining 1.9%. The number of cumulative MFR's required to handle total peak calls as a percentage are:

MFR	%
2	70.7 %
4	91.3 %
6	98.1 %
8	99.9 %
9	100.0 %

This demonstrates that half of the 16 MFR are not being used, in fact, two of the 16 MFR handle 70% of the calls. The concerns about peak loading are unwarranted.

The system was very conservatively designed and has quite a bit of expansion capability. The Michigan Bell and Science Dynamics portions of the ANI passing system worked flawlessy. Expansion to other cable systems within the calling area is under consideration.

MARKET RESEARCH

The participants in the Traverse City, Michigan IPPV test, Centel and Ameritech (Michigan Bell), commissioned Ad Factors Marketing Research Inc. to do a consumer opinion study of the PPV programming, Centel Cinema, and the ANI order entry system. A random sample of cable subscribers, who had access to the ANI order entry system in Traverse City, were selected. Telephone interviews were conducted between August 21 and September 9, 1986. The sample contained 638 subscribers:

- 247 who had not tried IPPV
- 145 one time users
- 246 multiple users

Of these:

- 65% non-users were aware of the service
- 42% of these had heard of the service via commercial insertion
- 93% of the users expressed satisfaction with ANI

The most likely IPPV user was profiled as younger, having a larger household and a premium channel subscriber. Relatively few negative comments about Centel Cinema were mentioned, the most common was about programming.

CONCLUSION

The ANI passing IPPV system trial in Traverse City, Michigan was an unqualified success. ANI order entry has been proven to be both customer friendly and technically reliable. This has been demonstrated through customer reaction and increasing buy rates. Cable customers enjoy the benefit of choosing and viewing individual events conveniently; the cable system enjoys new incremental revenue without large upfront capital expenditure, and the local telephone company initiates a new service. Perhaps as important, it was clearly demonstrated that the telephone and cable companies can work together efficiently for a common objective.

ACKNOWLEDGEMENT

We would like to especially thank the many people who have assisted in making this trial possible: Gene Walding, Tom Roach and Frank Flowers of Centel, and Ed Rzepka, Bill Coury and Russ Spranger of Michigan Bell.

MECHANICS OF AERIAL CATV PLANT

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ABSTRACT

Understanding the mechanics of aerial cable installation is essential to the cable engineer for proper plant design and for maximum cable plant reliability and longevity. A key part of that understanding is the calculation of sags and tensions. Although basic sag and tension equations are available, little is available on calculating sag and tension with changing temperature and load.

This paper extends the basic sag and tension equations to address these effects, and gives the ability to solve integrally supported (Figure 8) cable, tighly lashed cable, and unequal elevation problems. The equations are applied to several basic tension and clearance problems and further, to analyze expansion loop life, tight lashing, cable buckling, and center conductor pullouts. Obscure cable failure modes caused by wind gusting, solar heating and radiative cooling are discussed.

INTRODUCTION

The calculation of sag and tension can be quite useful to the cable television engineer. The calculations are used to select the appropriate size support strand for a given application or to determine if clearance requirements are met. They serve as an aid to the cable designers who determine the mechanical stress that the cable must be capable of withstanding or to evaluate such problems as expansion loop cracks and center conductor pullouts. The following is a discussion of sag and tension calculations which develops the simple case of a single wire suspended between two supports at equal elevation and works up through multiple elements, which takes into consideration each of the cable's components, with various loads and temperatures at unequal elevations.

BASIC EQUATIONS: WIRE, EQUAL EVALUATIONS

This first section covers the basic equations that apply. The sections that follow describe the application of these concepts. Figure 1 shows a single span of cable suspended between two fixed supports at equal elevations.

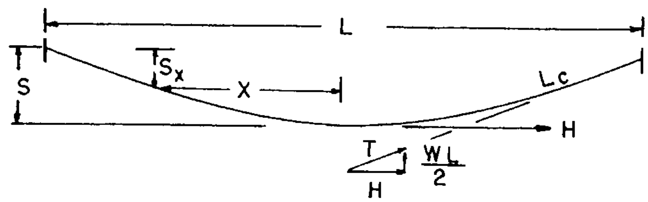

FIGURE 1.

$$H = WL^2/8S \qquad (1)$$

$$L_c = L + 8S^2/3L \qquad (2)$$

Where:

- H = Horizontal component of tension (lb)
- W = Linear weight of the wire (lb/ft)
- L = Distance between supports (ft)
- S = Sag (ft)
- L_c = Actual length of wire (ft)

These equations[1] are a parabolic approximation of the actual form that a flexible wire assumes, which is a catenary. The equation for a catenary is:

$$S = (H/W)\ [\cosh (WL/2H)-1] \qquad (3)$$

$$\approx WL^2/8H$$

The difference between the catenary and the parabola is that the parabola is approximately ½% smaller if the sag is about 6% of the span[2]. Generally, the sag is less than 2% for most cable television applications.

Another useful equation can be used to find the sag at any point along the span:

$$S_x = S(1-4x^2/L^2) \qquad (4)$$

UNEQUAL ELEVATIONS

The wire will assume the catenary form (or the parabolic approximation of this form) regardless of where the supports are located. The supports simply apply an equal and opposite vertical and horizontal force on the cable. Thus, the supports can be anywhere on the curve. Of course, after the load or temperature or both change, the sag changes. The supports are still on a parabola, but the parabola is different.

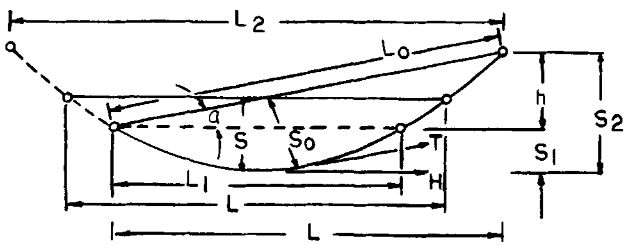

FIGURE 2

$$S = S_o = WL^2\cos\alpha/8T$$

$$= WL^2/8T\cos\alpha = WL^2/8H \qquad (5)$$

$$L_1 = L-2hH/WL \qquad (6)$$

$$L_2 = L+2hH/WL \qquad (7)$$

$$S_1 = WL_1^{\,2}/8H \qquad (8)$$

$$S_2 = WL_2^{\,2}/8H \qquad (9)$$

$$L_c = L+(4/3)[(S_1^{\,2}/L_1) + (S_2^{\,2}/L_2)] \qquad (10)$$

Where:

T = Tension in the direction of the cable (lb)

With these equations[1] (with Eqs. 6 and 7 slightly modified from the referenced source) the initial sag or tension can be calculated. Of course, one or the other must be known and span length and cable weight must also be given. They can be used, for instance, to calculate clearance over roads and sidewalks, and to determine specific size strand to be used under worst case loading. Note, however, that there is no mention of the properties of the materials such as their thermal coefficient of linear expansion or their elastic modulus. Indeed, these equations are correct but only for the initial conditions.

TEMPERATURE AND LOAD CHANGES

There are other conditions which encompass a range of temperatures and loads. For example, what if the amount of installed sag is fixed at some maximum value, and then the span is subjected to a different temperature and a load which may exceed the strand's strength? Or, if the sag and tension is known for the worst case load, what would the stringing sag and tension be at a different temperature and load? To solve these types of problems, further knowledge of the materials that are used is needed and the above equations must be adapted.

Suppose the wire is installed at some initial temperature (T_o) and the cable has a linear weight (W_o). Then the temperature drops to some final temperature (T_f) and the linear weight of the cable changes to some final weight (W_f) because, perhaps, ice has accumulated and a strong wind is blowing. What happens to the sag and tension?

As the temperature drops, the cable tries to get shorter because of its expansion coefficient, but because the span length is fixed the sag must get smaller, thus increasing the tension. The tension also affects the length of the cable due to its elastic modulus. So, as the tension increases, the length of the cable increases thus minimizing the effect. On one hand the cable tries to get shorter because of temperature, but it cannot get as short as it would like because the tension is increasing. Finally, the increased load causes the tension to increase, but because the cable is elastic it elongates in its elastic region thus increasing the sag and reducing the tension. All of this occurs simultaneously, and a balance or equilibrium is continually being maintained.

To determine the final sag and tension a mathematical description of these events will be given.

Unstressed Length

The most important quantity in the process of determining the final sag and tension is the cable's initial unstressed length at the initial temperature. The cable's unstressed length is the length of the cable if it had no tension or stress on it. It should not be confused with L_c, the actual cable length.

The unstressed cable length is necessary to know because, as additional tension is applied to the cable, its stressed length changes as a function of its unstressed length (and also its elastic modulus). To determine the appropriate amount of expansion and contraction due to temperature changes, again the unstressed length must be used (along with its thermal coefficient of linear expansion).

The stressed length of the cable (L_c) can be determined from Eq. 2 or 10. From this the unstressed length can be determined. Assuming that the material is elastic and follows Hooke's Law, its strain is proportional to the stress applied by a factor called the elastic modulus.

$$E = \sigma/e \qquad (11)$$

$$= FL/A\,\Delta L \qquad (12)$$

Where:
E = Modulus of elasticity (psi)
e = Strain = $\Delta L/L$ (dimensionless)
σ = Stress = F/A (psi)
ΔL = Change in length as the result stress (ft)
F = Force (lb)
A = Cross sectional area perpendicular to the force applied (sq in)

$$L_c = L+\Delta L \qquad (13)$$

$$= L+FL/AE \qquad (14)$$

$$L_c = L(1+F/AE) \quad (15)$$

It is assumed that the tension is the same along the entire length, but, because the wire has a finite weight per unit length, the ends of the wire near the support have more tension. The horizontal component of the tension is the same everywhere in the wire. The vertical component of the tension is zero at the lowest point since there is zero cable weight. The vertical component of tension at the support is the weight of the cable between the support and the lowest point.

The unstressed length, then is:

$$L_{uo} = L_c/[1+(H/AE)] \quad (16)$$

It should be noted that this only applies for stresses within the elastic limit of the material. Also, the elastic modulus of the material is not necessarily constant with temperature; therefore, a knowledge of the characteristics of the materials is important.

TEMPERATURE CHANGES

Once the unstressed length has been determined at the initial temperature, the unstressed length can be determined at other temperatures from the following:

$$L_{uf} = L_{uo}\,[1+a(T_f-T_o)] \quad (17)$$

Where:
L_{uf} = Final unstressed length at T_f (ft)
L_{uo} = Original unstressed length at T_o (ft)
a = Thermal coefficient of linear expansion (in/in F)
T_f = Final temperature (F)
T_o = Original temperature (F)

It should be noted that the thermal coefficient is not necessarily constant and again it is necessary to be familiar with the properties of the materials.

Most materials have positive thermal coefficients of linear expansion which cause them to get longer when the temperature increases. When the temperature decreases, they get shorter. (In some cases, materials are specially selected because they have negative expansion coefficients so that when used in conjunction with other materials the overall change in length is minimized or matched to some other component such as in fiber optic transmission lines.)

There are several factors that affect temperature. The temperature of the cable or wire is of course affected by the air temperature. The air temperature changes on a daily basis (diurnal) as well as yearly basis (annual). If electrical current is carried through the wire, a temperature rise will also occur, but, generally speaking, this rise is small for CATV cable. Another factor is radiation. During the day the cable is heated from the sun. At night the cable radiates its heat toward clear skies. The wind tends to minimize this affect.

Measurements made in Connecticut on the longest day of the year when the sun's rays are most direct indicate that a temperature rise of about 45F above ambient can be expected on black jacketed cable and about 24F rise above ambient for unjacketed aluminum sheathed coaxial cable. At night, in the same area, the temperature of black jacketed cable was about 8F below ambient and 4F below ambient for unjacketed aluminum cable.

LOAD CALCULATIONS

Before the final sag and tension can be calculated, it is necessary to evaluate the span of wire in terms of load. Initially we assumed that the wire's weight was the only load applied. Under worst case conditions, the wire may have ice formed around its circumference and at the same time a strong wind blowing on it. The total load on the wire is the resultant of all the vertical and horizontal loads. (The horizontal load here is perpendicular to both the vertical and the wire itself.)

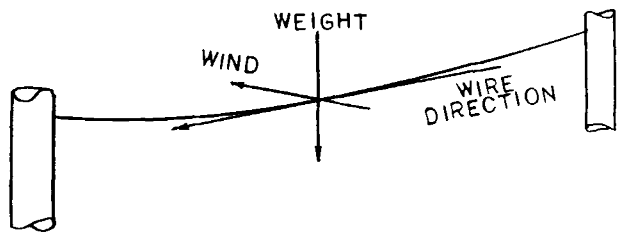

FIGURE 3

The total resultant load of the wire is the vector sum of all the horizontal and vertical components:

$$W_f = [(\Sigma W_V)^2 + (\Sigma W_H)^2]^{1/2} \quad (18)$$

Where:
W_f = Final linear cable weight (lb/ft)
ΣW_V = Sum of all vertical weights (lb/ft)
ΣW_H = Sum of all horizontal forces (lb/ft)

As an example, the vertical weight components may include: the weight of the support strand, the weight of all the cables, the weight of the lashing wire and possibly a cylinder of ice around the group. The horizontal force is only attributed to wind loading. In some cases a weight constant (e.g. in heavy loading districts, 0.3 lb/ft) is added to the resultant final weight for safety.

Ice Loading

The additional weight caused by ice build-up on the cable is usually calculated based on a hollow cylinder having an inside diameter equal to the outside diameter of the bundle of the cables and the support strand and with a given thickness which is usually 0.25" or 0.5" depending on which loading district is used. Using 57 lb/cu ft as the weight of ice[3], the linear ice weight around the cable can be calculated from:

$$W_{ice} = 1.244(D_i t + t^2) \tag{19}$$

Where:
W_{ice} = Linear ice weight (lb/ft)
D_i = Diameter over cable bundle (in)
t = Thickness of the ice (in)

Wind Loading

The wind loading on a circular surface can be calculated from[3]:

$$P = 0.00256V^2 \tag{20}$$

Where:
P = Horizontal wind pressure (lb/sq ft)
V = Wind velocity (mph)

In order to use this equation, the pressure must be converted to a load or force on the projected surface area of the bundle that faces the wind. Assuming that the wind is perpendicular to the cable and we wish the results to be in terms of lb/ft:

$$F_{wind} = PD_o/12 \tag{21}$$

Where:
F_{wind} = Wind loading (lb/ft)
P = Wind pressure (lb/sq ft)
D_o = Diameter of cable bundle (in.) (including ice if appropriate)

It should also be pointed out that horizontal loading results in sags which are not vertically directed. The horizontal and vertical components of the final sag can be resolved since they are proportional to the horizontal and vertical loads described in Eq. 18. i.e.

$$S_f = (S_v^2 + S_H^2)^{1/2} \tag{22}$$

So, for example the actual vertical component of the sag would be:

$$S_v = S_f(\Sigma W_v)^2/[(\Sigma W_v)^2+(\Sigma W_H)^2] \tag{23}$$

FINAL SAG AND TENSION

From Eq. 17 the final unstressed length of the wire can be determined based on the final temperature. From Eq. 15 we can determine the actual length of the wire if its tension is known:

$$L_{cf} = L_{uf}\ (1+H_f/AE) \tag{24}$$

Where:
L_{cf} = Final stressed length of the wire (ft)
L_{uf} = Final unstressed length of the wire (ft)
H_f = Final horizontal tension (lb)
A = Cross sectional area of the wire (sq in)
E = Elastic modulus, tensile (psi)

If the supports are at equal elevations, the final length of the wire is also (from Eq. 2):

$$L_{cf} = L+8S_f^2/3L \tag{25}$$

Where:
S_f = Final sag (ft)

This is actually a special case of supports at unequal elevations. The following progression puts Eq. 10 into the same form as Eq. 25. Consider:

$$L_c = L+(4/3)[S_1^2/L_1) + (S_2^2/L_2)] \tag{26}$$

From Eqs. 5, 6, and 7, S_1 and S_2 can be set in terms of S as:

$$S_1 = SL_1^2/L^2 \tag{27}$$

$$S_2 = SL_2^2/L^2 \tag{28}$$

So Eq. 26 can be transformed to:

$$L_c = L+(4S^2/3L^4)(L_1^3+L_2^3) \tag{29}$$

From Eqs. 8 and 9, L_1 and L_2 can be stated in terms of S and h and L as:

$$L_1 = L(1-h/4S) \tag{30}$$
$$L_2 = L(1+h/4S) \tag{31}$$

After a bit of work we find:

$$L_1^3+L_2^3 = L^3(2+3h^2/8S^2), \tag{32}$$

which can be plugged into Eq. 29 to yield:

$$L_{cf} = L+(8S_f^2/3L) + (h^2/2L). \tag{33}$$

Notice when h=0, Eq. 33 reduces to Eq.25, the equal elevation equation.

Back to the problem at hand, we were in search of another equation to set equal to Eq. 24; our search is over with Eq. 33. Actually Eq. 24 would be better in terms of S_f, like Eq. 33. This is easily done via Eq. 1.

$$L_{cf} = L_{uf}[1+(W_f L^2/8S_f AE)] \tag{34}$$

Setting Eq. 33 and 34 equal to each other yields:

$$L+(8S_f^2/3L) + (h^2/2L) - L_{uf} - (L_{uf}W_f L^2/8S_f AE) = 0 \tag{35}$$

Which can be put into the form of:

$$S_f^3+S_f(3L/8)(L+h^2/2L-L_{uf}) - (3L/8)(L_{uf}W_f L^2/8AE) = 0 \tag{36}$$

Notice that every variable in Eq. 36 is defined except the final sag (S_f). This equation takes into account a new temperature and a new load. S_f can be found via the solution of the cubic equation. Eq. 36 can be rewritten as:

$$S_f^3 + aS_f + b = 0 \quad (37)$$

Where:

$$a = 3[L^2 + (h^2/2L) - LL_{uf}]/8 \quad (38)$$

$$b = -3W_f L^3 L_{uf}/64AE \quad (39)$$

The solution of this particular form of cubic equation is as follows[4]:

$$\text{If} \quad (a/3)^3 + (-b/2)^2 \geq 0 \quad (40)$$

Then

$$S_f = \left\{(-b/2) + [(a/3)^3 + (-b/2)^2]^{1/2}\right\}^{1/3} + \left\{(-b/2) - [(a/3)^3 + (-b/2)^2]^{1/2}\right\}^{1/3} \quad (41)$$

Note, if the above condition is met and since -b/2 is always positive, the result will always be a real root.

$$\text{If} \quad (a/3)^3 + (-b/2)^2 < 0 \quad (42)$$

Then

$$S_f = 2(a/3)^{1/2}\cos\left\{(1/3)\cos^{-1}[(-b/2)/(a/3)^{3/2}]\right\} \quad (43)$$

Note, if Eq. 42 is true and since (-b/2) is always positive then (a/3) is certainly negative and again a real root will be obtained.

Once the final sag is found the final tension is easily found from:

$$H_f = W_f L^2/8S_f \quad (44)$$

CLEARANCE CALCULATIONS

Sometimes when checking to assure that the proper clearances are met (e.g. during make ready), it is necessary to know the elevation of the wire at points other than at the lowest point. For equal elevations Eq. 4 can be used. The sag at any point is given with respect to the lowest point which, for problems of equal elevation, is always exactly half-way between the supports.

For problems of unequal elevation, it is not particularly helpful to know what the sag is with respect to the position of the lowest point sag because not only is the lowest point sag not exactly half-way between the two supports, it may not be between the two supports at all!

Equation 4, however, can be adapted.

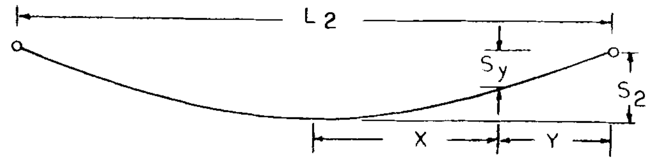

FIGURE 4

From the above figure (which is simply Fig. 2 without some of the details, but includes some new information about the sag at any point) it can be seen that x can be stated in terms of y, which is the horizontal distance from the highest support as:

$$x = (L_2/2) - y \quad (45)$$

Where:
- x = Distance from the lowest point (ft)
- y = Distance from the highest point (ft)
- L_2 = Span length (ft)
- S_2 = Sag at lowest point (ft)

Substituting Eq. 45 into Eq. 4 we obtain:

$$S_y = S_2\left\{1 - 4[(L_2/2) - y]^2/L_2^2\right\} \quad (46)$$

Where:
- S_y = Sag at y from the highest support (ft)

Equation 46 can also be used for equal elevation problems. Notice from Eq. 9 that L_2 is not fixed. L_2 will be different from its original value depending upon the tension (H). Therefore, it is necessary to determine both the final L_2 and S_2 if clearance is to be calculated.

Creep

It should be noted that materials, when exposed to stress even if below their elastic limit for a long period of time, tend to deform and take a permanent set. This slow acting deformation is called creep. For clearance calculations that require exacting accuracy, factors such as creep must be taken into consideration and creep data on the material used to support the cable should be obtained. Although it is beyond the scope of this paper to provide such information, two data points are given for discussion. The creep rate on extra high strength galvanized steel wire and strand is about 0.08% at 70% of the wire's rated breaking load. This much creep can occur in 24 hours, but very little additional creep takes place over the next 1000 hours of exposure, perhaps a total of 0.09%.

To illustrate the implications of this "small" number, consider a 100 ft span with 1 ft of sag. Suppose during the course of time the strand is exposed to such stress that would cause 0.09% creep. The implication can be calculated from Eq. 2. The resulting sag would be 2.07 ft instead of returning to 1 ft as might be expected. Although the strand should not be exposed to such stresses, the purpose of this illustration is to show the limitations of the calculations and highlight the need for caution and good engineering judgement.

SPECIAL CONFIGURATIONS

Multiple Elements

In some cases the wire may be comprised of several different materials with different cross sections. Cable with integral messengers ("Figure 8") drop cable are some examples. Each material has its own expansion coefficient and elastic modulus. When suspended, each element takes a different portion of the total load. The amount of the load that any particular cable element takes is a function of its elastic modulus and area as compared to the total resultant modulus of all the materials and the total area of all the materials.

The resultant modulus can be found by multiplying the modulus of each material by the ratio of the material's area to the total area:

$$E_r = (A_1/A_T)E_1 + (A_2/A_T)E_2 + (A_3/A_T)E_3 + \ldots + (A_n/A_T)E_n \qquad (47)$$

Where:
- E_r = Resultant elastic modulus (psi)
- A_T = Total cross sectional area of the wire (sq in)
- A_1 = Cross sectional area of element 1 (sq in)
- A_2 = Cross sectional area of element 2 (sq in)
- A_3 = Cross sectional area of element 3 (sq in)
- A_n = Cross sectional area of element n (sq in)
- E_1 = Elastic modulus of element 1 (psi)
- E_2 = Elastic modulus of element 2 (psi)
- E_3 = Elastic modulus of element 3 (psi)
- E_n = Elastic modulus of element n (psi)

These values (i.e., E_r and A_T) along with the following value for expansion coefficient can be directly substituted into the above equations for sag and tension wherever E, A, and a occur.

The resultant coefficient of linear expansion (a_r) is calculated as follows:

$$a_r = (A_1E_1a_1 + A_2E_2a_2 + \ldots + A_nE_na_n)/(A_1E_1 + A_2E_2 + \ldots + A_nE_n) \qquad (48)$$

By using these values for E_r, A_r, and a_r the final sag and final tension can be determined at the new temperature with the new load.

Because the above equations which include elastic modulus only work when the materials are stressed within their elastic region, it is important to make sure that each element's elastic limit is not exceeded.

The unstressed length of all the elements is the same only at the original temperature. It is assumed that any stress built into the cable is small compared to the installed stress. Once the temperature changes, each element expands and contracts at its own rate. Also, each element has its own elastic modulus and cross sectional area. Unless the elements are bonded together in some fashion they may move independently of one another.

In order to determine the stress on any particular element, the unstressed final length must first be determined. It can be calculated, based on the change in temperature and the expansion coefficient of the element in question, as follows:

$$L_{ufn} = L_{uon}[1+a(T_f-T_o)] \qquad (49)$$

Where:
- L_{ufn} = Final unstressed length of any particular element n (ft)
- L_{uon} = Original unstressed length of any particular element n (ft)

Note: L_{uon} is the same for all elements and exactly the same as that given in Eq. 16.

The final cable length (under stress) is the same for all elements. It can be determined from either Eq. 2 or Eq. 10. The strain that the cable element is under is then:

$$e_n = (L_{cf} - L_{ufn})/L_{ufn} \qquad (50)$$

the stress is the modulus times the strain:

$$\sigma_n = E_n(L_{cf}-L_{ufn})/L_{ufn} \qquad (51)$$

And the tension on element n is:

$$H_n = A_nE_n(L_{cf}-L_{ufn})/L_{ufn} \qquad (52)$$

Where:
- H_n = Tension on element n (lb)
- A_n = Cross sectional area of element n (sq in)
- E_n = Elastic modulus of element n (psi)
- L_{cf} = Final cable length (ft)
- L_{ufn} = Unstress final length of element n (ft)

The stress in Eq. 51 should be tested to assure that the elastic limit not be exceeded. Aside from exceeding the elastic limit of any particular element, one additional caution should be noted. Generally, "Figure 8" cable used in CTV applications is for distribution purposes and so taps are installed. To install a tap the cable must be cut and separated from the messenger wire. Cutting the cable relieves the stress in the cable at the point where it is cut. This tension, however, does not necessarily disappear. A significant portion, if not all of it, is diverted to the steel messenger wire. So, wherever taps are installed in messenger cable, you must assure that the total tension does not exceed the safe limits of the messenger if it were alone.

Cable Tightly Lashed

In cases where cable is lashed to the strand so tight as to restrict cable from moving independently of the steel strand support, a similar analysis can be performed. This condition can occur even if the lashing is not restrictive: by having no expansion loops.

We will call this third case "cable restrained." In this case, the steel strand is installed and loaded with the cable. The cable is then lashed so tight that cable movement is restricted. The resultant condition is that the steel strand is under stress and the cable has zero stress.

After the temperature, load or both change, the load is not distributed like it was in the case of "Figure 8" cable. They differ in that there is no original cable tension; so the original unstressed cable length is the same as the actual cable length (L_{co}). The original unstressed steel length is slightly shorter.

It will be assumed that the cable and steel support strand are bound together, either by tight lashing or lack of expansion loops. This may not be the case because some slight differential movement between the cable and support strand can be expected, but it will allow us to evaluate the extreme conditions; the actual[5] lying somewhere between this and the first case (i.e., the steel taking the full load at all temperatures and under loading conditions).

The first step in solving this problem is to find the unstressed length of the composite assembly. It is certainly not the unstressed strand length because the two components, the steel and the cable, are bound together. Nor is it the unstressed cable length. The unstressed length of the composite is somewhere between the two.

When the composite assembly is unstressed, an interesting condition occurs; the steel is under tension and the cable under compression.

Again, as before, the actual length of the assembly initially is found from Eq. 2.

$$L_c = L+8S^2/3L \tag{52}$$

The unstressed length can be calculated from:

$$L_u = L_c/[1+(H/A_T E_r)] \tag{53}$$

E_r is found in the same way as for Figure 8 cable, Eq. 47. The tension, H, in Eq. 53 is the tension on the steel support alone. The original unstressed length can be used to find the final unstressed length at the final temperature from Eq. 17 except that the resultant expansion coefficient (a_r) should be used in place of a. Finally, the final sag can be calculated from Eq. 41 or 43. The total tension on the assembly can then be found from Eq. 44.

The original tension on the components was easily determined; the steel had all the tension, the cable had none. At final state, that is, after the load, temperature or both change, the tension on each component may not be quite so easy to determine. We will divide the tensions into two groups: namely the tension on the steel and the tension on the other components.

To analyze the final tension on the steel, we must first determine its initial unstressed length. It is:

$$L_{uo\ steel} = L_{co}/[1+(H_o/A_{steel}E_{steel})] \tag{54}$$

Changing the temperature altered this length to:

$$L_{uf\ steel} = L_{uo\ steel}[1+a_{steel}(T_f-T_o)] \tag{55}$$

The stress on the steel is then:

$$\sigma_{steel} = E_{steel}(L_{cf}-L_{uf\ steel})/L_{uf\ steel} \tag{56}$$

The tension on the steel is:

$$H_{f\ steel} = A_{steel}\sigma_{steel} \tag{57}$$

The stress and tension on the other elements can be found in a similar manner except that the original unstressed cable length equals the original actual cable length.

$$L_{co} = L_{cuo} \tag{58}$$

The change in length of any particular element other than the steel can be determined from:

$$L_{ufn} = L_{co}[1+a_n(T_f-T_o)] \tag{59}$$

Where the subscript "n" refers to one of the cable elements.

The stress on that particular element is then:

$$\sigma_n = E_n(L_{cf}-L_{ufn})/L_{ufn} \tag{60}$$

And the tension on that element is:

$$H_{fn} = A_n\sigma_n \tag{61}$$

Unfortunately, it is difficult to first assume full load at the lowest temperature for restrained cable. For example, you may want the tension on the strand to be 60% of its breaking strength at full load at the lowest temperature. The tension on the other components can be anywhere between zero and the material's elastic limit, depending on what the original temperature was.

In the other two cases, the original temperature and load have no restrictions. But in this case, cable restrained, the original conditions must be at the temperature at which the cable was lashed to the strand. Of course, if enough information is known about the final conditions, the initial conditions can be determined.

In review, three basic configurations have been discussed with regard to sag and tension. The first was a simple case of a single wire. The second considered multiple elements which can be applied to "Figure 8" type cable or drop wire. The last, "cable restrained", addressed what happens when the cable has zero tension initially but supports some of the load as conditions change. Each configuration was considered after the temperature, load or both changed.

APPLICATIONS

Strand Tension Under Worst Case Load

To illustrate the application of these equations a few examples will be given. Consider, first, a 125 foot span having a 500 and 750 size jacketed cables supported by a 1/4" EHS strand. Assume that the cable has expansion loops and is free to move independently of the steel support strand so that there is no tension on the cable under any condition. The cable was installed at 60F and has a 1.5% sag (1.875 ft). The cable is installed in a heavy loading district with extreme wind loading of 21 lb/sq ft. The question is, what is the tension on the strand under worst case conditions?

From Table 1 and 3, the total weight of the strand, plus the 500J, plus the 750J, plus the double lashing is 0.439 lb/ft. The maximum tension on the strand must not exceed 60% of its rated break strength under worse case expected loading. The initial strand tension and cable length at 60F can be calculated from Eq. 1 and 2.

$$H = (0.439\ \text{lb/ft})\ (125\ \text{ft})^2/8(1.875\ \text{ft}) = 457.3\ \text{lb}$$

$$L_c = 125\ \text{ft} + 8(1.875\ \text{ft})^2/3(125\ \text{ft}) = 125.0750\ \text{ft}$$

The unstressed length of the steel support strand is found from Eq. 16 and the elastic modulus of the strand from Table 2 and 3.

$$L_u = (125.0750)/[1+457.3/(0.035185)\ (28\text{X}10^6)] = 125.01697\text{ft}$$

Now the unstressed length at any other temperature can be determined. From Table 4, the temperature for a heavy loading district is 0F. From Eq. 17 and the expansion coefficient of steel in Table 2, the unstressed length of the steel at 0F is:

$$L_{uf} = (125.01697)[1+7.2\text{x}10^{-6}(0-60)] = 124.96296\ \text{ft}$$

Assuming that the cables are stacked one on top of the other and from the dimensions in Table 1 and 3, the width of the cable plus the strand is 1.620 in. From Eq. 19, 0.5 in thick ice over the cable weighs 1.319 lb/ft. So the total vertical weight (i.e., ice plus the cable) is 1.757 lb/ft. See Fig. 5.

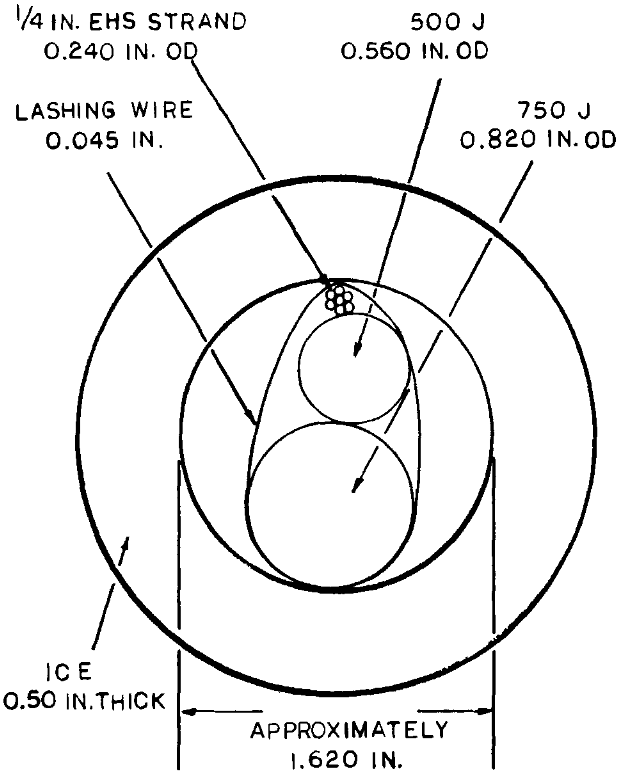

FIGURE 5

The horizontal load from the wind is calculated from Eq. 22. The projected diameter over the ice is 2.620 in (i.e. 2x0.5in + 1.620in). The wind pressure for a heavy loading district, from Table 4, is 4 lb/sq ft. The loading is then:

$$F_{wind} = (4)\ (2.620)/12 = 0.873\ \text{lb/ft}$$

The vector sum of the horizontal and vertical loads calculated from Eq. 18 is:

$$W = [(1.757)^2 + (0.873)^2]^{1/2} = 1.963\ \text{lb/ft}$$

The weight adder for a heavy loading district is 0.30 lb/ft and results in a final weight of:

$$W_f = 1.963 + 0.30 = 2.263\ \text{lb/ft}$$

To solve for the final sag in Eq. 37, a and b are first found from Eq. 38 and 39. The difference in support elevations, h, is zero so:

$$a = 3[(125)^2-(125)\ (124.96296)]/8 = 1.73608875$$

$$b = -3(2.263)(125)^3(124.96296)/64(0.035185)(28\text{x}10^6) = -26.27575$$

$$(a/3)^3 = 0.19380$$

$$(-b/2)^2 = 172.604$$

Clearly, Eq. 40 is true and the final sag (which is not vertically directed) can be found from Eq. 41 and the final tension from Eq. 1.

$$S_f = 2.779 \text{ ft}$$

$$H = (2.263)125)^2/8(2.779)$$
$$= 1590 \text{ lb}$$

This is the tension (horizontally directed) in the strand under heavy loading conditions. This is well within the 3990 lb maximum limit.

One more check is necessary and that is under extreme wind loading at 60F. From Figure A2 the wind loading is 21 lb/sq ft. The resulting wind load from Eq. 22 is:

$$F_{wind} = (21)\ (1.620)/12$$
$$= 2.835 \text{ lb/ft}$$

By adding the horizontal and vertical load components, the final load is 2.869 lb/ft. Using the unstressed length at 60F:

$$(a/3)^3 = -0.0186452$$

$$(-b/2)^2 = 277.746$$

The final sag and tension is found in the same way as before.

$$S_f = 3.30 \text{ ft}$$
$$H = 1698 \text{ lb}$$

So the strand tension under extreme wind loading is well under the 3990 lb limit.

The above example was carried out in several steps:

1. Calculate the initial tension from Eq. 1.
2. Calculate the initial cable length from Eq. 2.
3. Calculate the initial unstressed cable length from Eq. 16.
4. Calculate the final unstressed cable length from Eq. 17.
5. Calculate the final load from the loading tables and Eqs. 18 and 19.
6. Calculate the final sag from either Eq. 41 or 43.
7. Calculate the final tension from Eq. 44.

For unequal elevations problems, L_c is calculated from Eq. 10 and h is used in Eq. 38. These equations can be loaded into a computer or programmable calculator for easy calculations.

Clearance Over A Traffic Light

Consider, next, a situation where you wish to keep a 30" clearance, under all expected conditions, over a traffic light. If the steel support is installed so that it is loaded to about 5% of its rated strength will the proper clearances be met? Figure 6 shows a sketch of the span.

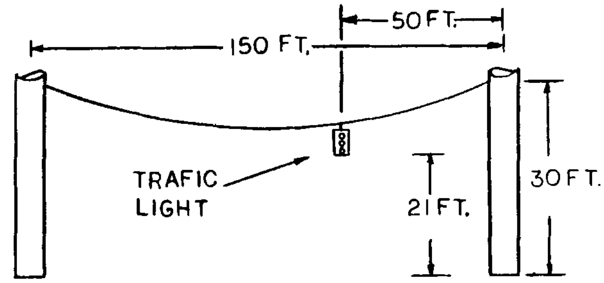

FIGURE 6

The following information is available.

SUPPORT STRAND:
1/4" EHS
$E = 28\text{x}10^6$ psi
$A = 0.035185$ sq in
$a = 7.2\text{x}10^{-6}$ in/in F

CABLE:
1 - 750J
1 - 500J
Double Lashed
Diameter = 1.620 in
Weight = 0.439 lb/ft

The cable is installed in a heavy loading district with 16 lb/sq ft extreme wind.

The initial (loaded) conditions from Eq. 1, 2, 16 and 46 are:

$H = 330$ lb $L_u = 150.198554$ ft

$S = 3.74$ ft $S_y = 3.33$ ft at 50 ft from the support.

Assuming that the worst case is at 32F with 0.5 in ice, no wind and a 0.3 lb/ft weight constant, the following results are obtained:

$W_f = 2.0576$ lb $\quad L_{uf} = 150.168274$ ft
$H_f = 1280$ lb $\quad S_y = 4.02$ ft

$S_f = 4.52$ ft

Assuming Heavy Load Conditions:

$W_f = 2.263$
$H_f = 1431$
$S_f = 4.45$

Assuming 120F with no load:

$H_f = 0.439$ lb/ft
$H_f = 296$ lb
$S_f = 4.17$ ft

For heavy loading 1431/1b is about 22% of the strands rated strength of 6650 lb. The expected creep is approximately 0.015%. Doubling it for margin, 0.03%,

$L_{cf} = 150.363381$ ft

This is a first order approximation of the increased sag over time due to creep. The sag in the middle of the span would be about 4.783 ft instead of 4.52 ft. As long as the attachments are: 21 + 2.5 + 5 = 28.5 ft high, there should not be any clearance problems.

Tension on Composite Materials

Consider a 150 ft span at 60F with 0.5% sag (0.75 ft). In this example let us assume that a single 750 non-jacketed cable is either tightly lashed to the strand or has no expansion loops so that in can not move independently of the 1/4" EHS support strand.

L = 150 ft
S = 0.75 ft
W = 0.299 lb/ft

COMPONENT	MAT	A (sq in)	E (psi)	a (in/in F)
STRAND	STEEL	0.03585	28×10^6	7.2×10^{-6}
SHIELD	AL	0.08075	10×10^6	12.7×10^{-6}
DIEL.	T4+FOAM	0.33965	42×10^3	48×10^{-6}
CENTER	CuCladAl	0.02138	10.6×10^6	12.2×10^{-6}
RESULTANT		0.476965	4.26×10^6	10.4×10^{-6}

The resultant is calculated from Eqs. 47 & 48. The condition at 60F are:

$S = 0.75$ ft $\quad H_{Steel} = 1.121$
$H_{Total} = 1.121$ lb $\quad H_{Alum} = 0$ lb

Assume the temperature drops to - 40F.

$S_f = 0.27$
$H_{Total} = 3{,}129$ lb

	TENSION	Eq. #	STRESS
H_{Steel} =	1,775 lb	(Eq. 57)	50,448 psi
H_{Shield} =	980 lb	(Eq. 61)	12,136 psi
H_{Diel} =	111 lb	(Eq. 61)	327 psi
H_{Center} =	264 lb	(Eq. 61)	12,348 psi

The stresses on the shield and center conductor have exceeded their elastic limits. This translates to excessive stress on the conductors. This span will be prone to center conductor pullout problems and stress fatigue.

Now assume instead that the cable was installed on a cold day, say at 40F, and during the summer the cable temperature rose to 120F. The intitial sag and tension are the same as before; the final conditions are:

$S_f = 2.14$ ft $\quad H_{Steel} = 1023$ lb
$H_{Total} = 393$ lb $\quad H_{Shield} = -435$ lb
$H_{Diel} = -82$ lb
$H_{Cen} = -113$ lb

Notice that the cable is under compressive forces.

The force (F) required to cause the aluminum to buckle under compressive forces is:

$$F = \pi^2 EI/X^2$$

Where:
I = Moment of inertia
X = Unsupported distance

The moment of inertia, I, for 750 cable is approximately 0.005 in^4 so the maximum unsupported distance that the cable can be exposed to without buckling is about 34in. Although the lashing wire spacing is much less than 34 in, it is quite conceivable that, at the pole, the cable will be unsupported for over 34 in and cable buckling is imminent.

Long Spans-Cable Movement and Expansion Loops

Consider a long span, approximately 200 ft of 750 cable, with 1.5% sag at 60F. Suppose that the cable has expansion loops and the cable can move freely as the temperature changes. Consider then that the temperature changes from 50F at night to 110F during the day. What is the differential cable movement? This much temperature change is not unreasonable to assume since solar heating and radiative cooling can account for a 28F temperature change on unjacketed cable even if the ambient temperature does not change.

60F	50F	110F
L = 200		
S = 2 ft	1.89 ft	2.60 ft
H = 748 lb	791 lb	575 lb
L_c = 200.05333ft	200.04768ft	200.09026ft
L_{Cable}=200.05333ft	200.02633ft	200.18832ft
	- 0.02135ft	+0.09806ft
	- 0.256 in	+1.177 in

1.43 in TOTAL

Expansion Loop Life

The life of an expansion loop can vary significantly depending on a number of factors: the surface finish of the aluminum, the type of loop, the depth of the loop, the size of the cable, the wall thickness of the aluminum outer conductor, whether it's jacketed or not, and the excursion distance per cycle. Without going into expansion loops to any great extent, the following expansion loop lifes are typical of semiflexible cable with measured 6 in depth and total 1 in excursion (i.e., + 0.5in and -0.5 in from the neutral point). The expansion loops in 0.500 in cables were formed with a LEMCO G120 and the 1.000 in with a G240.

CYCLES TO OUTER CONDUCTOR
FRACTURE (1" EXCURSION)

TYPE	UNJACKETED	JACKETED
0.500 in	29.9K	51.8K
0.750 in	18.0K	22.5K
1.000 in	8.0K	17.9K

The life of other cable sizes can be roughly approximated by interpolation. If each cycle is equivelent to one day, then the life of 0.750 in unjacketed cable would be about 49 years. Measurements show that as you double the excursion distance, the life of the loop drops by a factor of 10 for jacketed cables and by a factor of about 20 for unjacketed cables. The depth of the loop is extremely important. For a 0.500 in unjacketed cable with a 3 1/2in depth instead of 6in, the aluminum fractures at 1.5K cycles instead of 29.9K cycles.

For the specific case given above where the loop is exposed to 1.4 in excursions, the life of the 750 cable loop is about 8 years. Therefore, even with modest temperature changes, a single expansion loop may not be adaquate for long spans.

Wind Gusting

Aside from length changes due to temperature changes, changes can also occur due to load changes. Although load changes due to ice build-up are infrequent, perhaps a few times a year, load changes due to wind can occur quite frequently especially if gusting is considered.

The frequency and amplitude distribution of wind gusting is complicated and can vary significantly from one region to the next. In general, gusting is far more severe in urban areas with buildings than in flat, level country. Although there is some good data available on wind gusting[6], the next example simplifies the frequency and amplitude distribution of gusting to a single data set.

Consider a 150 ft span of cable, one 500J and one 750J cable, suspended between supports with 1.5% sag at 60 F in a town with buildings. Consider next that the distribution of wind gusts for a one year period can be represented by 1000 gusts from 20 mph to 60 mph. The following table can be generated:

Wind Speed (mph)	Wind Force (lb/sq ft)	Cable Length (ft)
0	0	150.0900
20	1.024	150.0927
60	9.216	150.1827

So, the change in length that the expansion loop must accommodate is 1.08 in. Assuming that there are 365 one inch temperature cycles per year and 1000 inch gust cycles per year, then the life of the loop will decrease by a factor of 0.27 as compared to its life due to temperature changes. For the 0.750 in cable mentioned above, the life of the loop should be degraded from about 49 years to about 13 years.

CONCLUSION

The basic sag and tension equations have been expanded to account for changes in temperature and load. Special configurations, such as tight lashing and "Figure 8" type cable with composite materials, were also addressed. The application of these equations can help the cable engineer respond to the local utility about maximum tension and proper clearance of aerial plant. The equations can also be applied to help understand various failure modes such as center conductor pullouts and premature expansion loop cracking.

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APPENDIX

TABLE 1
CABLE PROPERTIES

Unjacketed Trunk and Feeder

Type	Weight (lb/ft)	Dia. (in)	E_r (psi $x10^6$)	a_r (in/inF $x10^{-6}$)	A_r (sq in)
412	0.058	0.412	2.79	13.3	0.1333
500	0.078	0.500	2.44	13.5	0.1964
625	0.121	0.625	2.42	13.5	0.3068
750	0.171	0.750	2.37	13.5	0.4418
875	0.225	0.875	2.26	13.5	0.6013
1000	0.325	1.000	2.62	13.4	0.7854

Jacketed Trunk and Feeder

Type	Weight (lb/ft)	Dia. (in)	E_r (psi $x10^6$)	a_r (in/inF $x10^{-6}$)	A_r (sq in)
412J	0.074	0.470	2.16	13.6	0.1735
500J	0.098	0.560	1.95	13.7	0.2463
625J	0.146	0.685	2.02	13.7	0.3685
750J	0.206	0.820	1.99	13.7	0.5281
875J	0.266	0.945	1.76	13.9	0.7776
1000J	0.377	1.080	2.26	13.5	0.9161
TX565	0.106	0.625	1.77	14.0	0.3068
TX840	0.212	0.910	1.69	14.1	0.6504
TX1160	0.396	1.250	1.92	13.9	1.2282

Messengered Feeder (Figure 8)

Type	Weight (lb/ft)	Dia. (in)	E_r (psi $x10^6$)	a_r (in/inF $x10^{-6}$)	A_r (sq in)	MSGR OD (in)
412JMS	0.121	0.830	2.87	11.1	0.2207	0.109
500JMS	0.145	0.930	2.53	11.5	0.2936	0.109
625JMS	0.252	1.120	3.13	10.9	0.4272	0.186
TX565JMS	0.183	1.020	3.24	10.5	0.3657	0.186

Drop Cable and Lashing Wire

	Weight (lb/ft)	Dia. (in)	E_r (psi $x10^6$)	a_r (in/inF $x10^{-6}$)	A_r (sq in)	MSGR OD (in)
59 Quad						
Single	0.027	0.262	1.17	10.8	0.0425	
Msgr	0.039	0.417	2.03	8.89	0.0526	0.051
6 Quad						
Single	0.034	0.297	1.22	10.6	0.0556	
Msgr	0.046	0.452	1.90	9.03	0.0658	0.051
611 Quad						
Single	0.045	0.350	1.21	10.4	0.0798	
Msgr	0.066	0.527	2.24	8.66	0.0941	0.072
11 Quad						
Single	0.071	0.434	1.06	10.2	0.1267	
Msgr 1	0.100	0.657	1.91	8.59	0.1494	0.083
Msgr 2	0.096	0.658	2.60	8.44	0.1532	0.109
Lashing	0.007	0.045	---	---	---	0.109

Note: The messenger used on the 412 and 500 cable is a solid 0.109" EHS steel wire, a stranded 3/16" EHS steel wire is used on the 625 and TX565.

E_r, a_r and A_r are the resultant elastic moduli, expansion coefficients and areas respectively.

TABLE 2

MECHANICAL PROPERTIES OF SELECTED MATERIALS		
Material	Elastic Modulus (psi $x10^6$)	Expansion Coefficient (in/inF $x10^{-6}$)
Steel	28-30	7.2
Copper	16	9.2
Aluminum	10	12.7
Cu Clad Al (10%byVol)	10.6	12.2
CuClad Steel (23%Cond)	24-27	7.2
Tape/Braid	4.2-6.3	11-13
Foam Diel (T4+)	0.042	78
Jacket Mat.		
LDPE	0.028	72
LLDPE	0.042	72
MDPE	0.063	83
HDPE	0.123	67
PVC	0.0011	40

TABLE 3
PROPERTIES OF
SELECTED MESSENGER WIRES
AND EHS STEEL STRANDS[7]

Type	Stranding	Dia.	WT (lb/ft)	Area (sq in)	Break Strength x 1000lb
0.051	solid	0.051	0.007	0.002043	0.200
0.072	solid	0.072	0.014	0.004072	0.365
0.083	solid	0.083	0.018	0.005411	0.485
0.109	solid	0.109	0.031	0.009331	1.800
1/8	7x0.041	0.123	0.032	0.009241	1.83
3/16	7x0.062	0.186	0.073	0.021133	3.99
7/32	7x0.072	0.216	0.098	0.028500	5.40
1/4	7x0.080	0.240	0.121	0.035185	6.65
9/32	7x0.093	0.279	0.164	0.047550	8.95
5/16	7x0.104	0.312	0.205	0.059464	11.20
3/8	7x0.120	0.360	0.273	0.079168	15.40
7/16	7x0.145	0.435	0.399	0.115590	20.80
1/2	7x0.165	0.495	0.516	0.149677	26.90

TABLE 4
NESC LOADING TABLE[3]

	Heavy	Medium	Light	Extreme Wind
Radial Ice (in)	0.5	0.25	0.0	0.0
Wind (lb/sq ft)	4	4	9	See Fig A2
Temperature(F)	0	+15	+30	+60
Wt. Adder(lb/ft)	0.30	0.20	0.05	0.0

TABLE 5
NESC CLEARANCE REQUIREMENTS[3]
60 F, No wind

Roads, alleys and other land subject to truck traffic or farm vehicles	18
Residential driveways, no truck traffic	10
Pedestrians only	8
Railroad crossings	27
Water areas sailboats prohibited (unless otherwise specified)	15

If the spans are longer than:
175 ft. in Heavy Loading Districts,
250 ft. in Medium Loading Districts, or
350 ft. in Light Loading Districts; then
additional clearance should be added with one exception. Add 0.10 ft. (0.15 ft. for railroad crossings in heavy and medium loading districts) to the clearance for every 10 ft. of span for spans longer than the ones listed above in all loading districts. Additional clearance is not necessary if the difference between the initial sag (at 60°F with no wind) and final sag at either 32°F with radial ice (from Table 4) and no wind, or 120°F with no wind (and no ice) is greater than the added clearance.

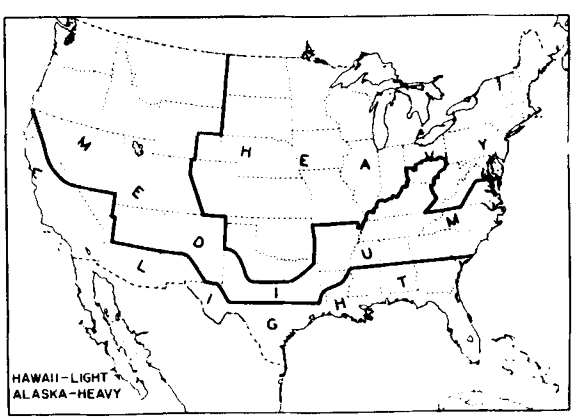

FIGURE A1
NESC LOADING DISTRICT MAP

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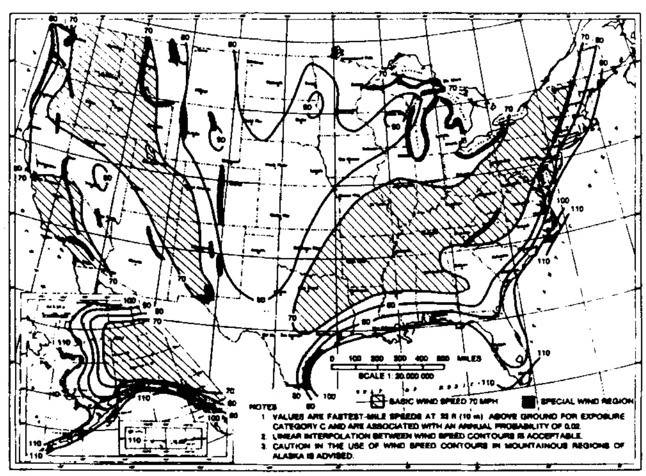

FIGURE A2
NESC EXTREME WIND LOADING MAP

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MTS COMPATIBILITY IN ENCRYPTED BASEBAND SCRAMBLING SYSTEMS

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INTRODUCTION

The advent of the MTS stereo standard has created new revenue opportunities for the MSO. MSO's may now offer premium programming with the added attraction of high-quality stereo sound. The problem of how to insure the security of the signal assumes new importance, however, and in some ways the problem is more difficult than it may appear.

Digitally encrypted audio provides the ultimate in signal security, and modern digital technology coupled with careful system design can provide two high-quality channels, utilization of which opens the possibility of a highly secure stereo signal for premium programming.

This paper will discuss the issues involved in the adaptation of a digitally encrypted baseband audio system to meet these needs, and will further discuss the technical requirements, using Oak's "SIGMA" line of converters as an example.

Realistically, it is imperative that the ultimate interface to the subscriber is in the MTS format, such that his stereo TV set can properly process and deliver the two channels of information. This format creates a signal whose bandwidth extends to at least 46 KHz, and whose amplitude and phase characteristics throughout that band are critical to proper performance.

The easiest solution would seem to be simply to directly encrypt the MTS signal, deliver it to the decoder, decrypt it, and send it on to the TV. Unfortunately, the wide bandwidth and severe phase accuracy requirements of the format make this extremely difficult and consequently quite expensive to do properly.

Moreover, the MSO is faced with the problem of non-MTS-formatted source material, such as direct left/right channel feeds from laser disc players, VCR's, and satellite receivers.

The details of the MTS requirements are easier to describe than to deal with (reference Figure 1): As with conventional FM Broadcast Stereo, the left and right channel audio signals are first combined to form two new channels, the "sum" channel (Left plus Right) and the "stereo difference" channel (Left minus Right). The sum channel is the normal monaural signal, and needs no special handling. The stereo difference channel, on the other hand, requires very special handling indeed. It is first processed through a complex and elegant compression system, then modulated to form a double-sideband suppressed carrier (DSSC) subcarrier and added to the monaural sum channel. Finally, to assist in demodulation of the difference subcarrier, a pilot tone is added to the sum/difference ensemble. The MTS baseband signal spectrum is shown in Figure 2.

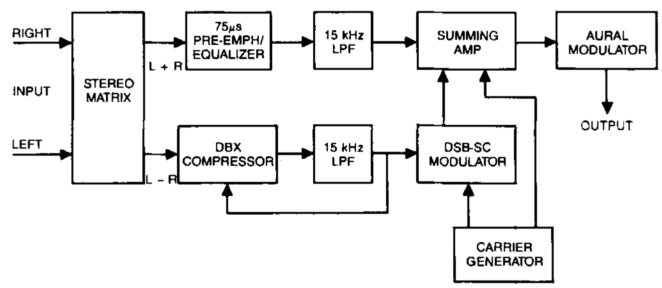

FIGURE 1. BASIC MTS ENCODER

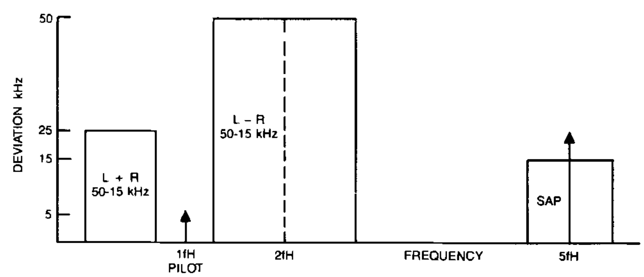

FIGURE 2. MTS BASEBAND SPECTRUM

In the receiver (reference Figure 3), the demodulated stereo difference subcarrier is expanded, dematrixed back to simple Left and Right channels, and finally amplified to drive the speakers.

In order to maintain stereo separation of 20dB or better, the amplitude and phase of the compressed stereo difference signal must match and track those of the uncompressed monaural sum signal to within +/-1 dB and +/-6 degrees. In addition, to insure accurate demodulation of the DSSC difference subcarrier (so as to enable the gain accuracy mentioned above), the pilot tone phase must be maintained within +/-3 degrees of that of the (suppressed) subcarrier.

Finally, if it is not to seriously impact the system equipment investment budget, all of the above must be achieved within severe cost constraints.

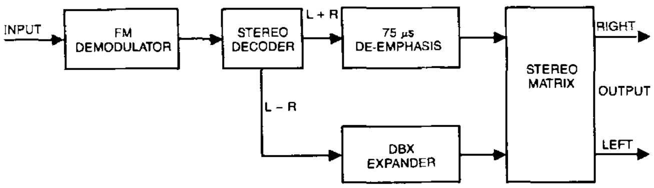

FIGURE 3. BASIC MTS RECEIVER

SYSTEM CONSIDERATIONS

The final system must meet two major requirements: it must be MSO-friendly, providing the requisite signal security and still permitting a maximum of installation flexibility; and it must be subscriber-friendly, offering convenient installation and operation.

Meeting the MSO's needs is straightforward, if not easy. The system must be easy to install, require a minimum of adjustment and provide a maximum of stability, permit a random mix of stereo TV's and hifi-only systems, and be fully compatible with the existing services. On the other hand, while the system should not be unduly expensive, it still must provide the MSO with the signal security he needs to protect his premium programming.

Dealing with the subscriber's problem is more complicated. The average home TV/VCR/Cable adapter complex is already far too cluttered and cumbersome, both physically, electrically, and operationally. Adding a stereo "sidecar" or "hotplate" would make an already clumsy situation totally unworkable. The complications of audio signal paths, control priorities, quantities of audio input/output ports, switching networks, etc., presage significant confusion and consequent dissatisfaction on the part of the subscriber.

The ideal is a system in which the installer has only to unbox and connect the cable decoder appropriate to the subscriber's TV (and add, if necessary, an audio cable from the decoder to the subscriber's hi-fi system), after which the MSO merely authorizes the services for which the subscriber has paid.

APPROACH
(Reference Figures 4 and 5)

Given that the MTS format itself occupies too much bandwidth to permit direct digitization and encryption, some ingenuity must be employed. The basic approach used here is to transmit encrypted, digital stereo, and, in the converter, convert the decrypted stereo into MTS format for remodulation and output to the stereo TV.

There are two fundamental difficulties that must be overcome. The first of these is economic: conversion or encoding into MTS requires DBX compression of the stereo difference signal, in a process that requires extreme precision and stability, and consequently is relatively expensive. The second is also essentially economic, but more subtle than it might appear: The pilot tone and the stereo difference subcarrier must be phase-locked

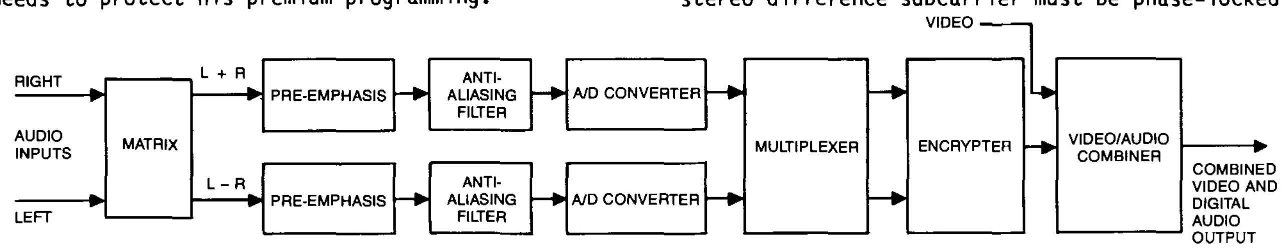

FIGURE 4. ENCODER BEFORE UPGRADE

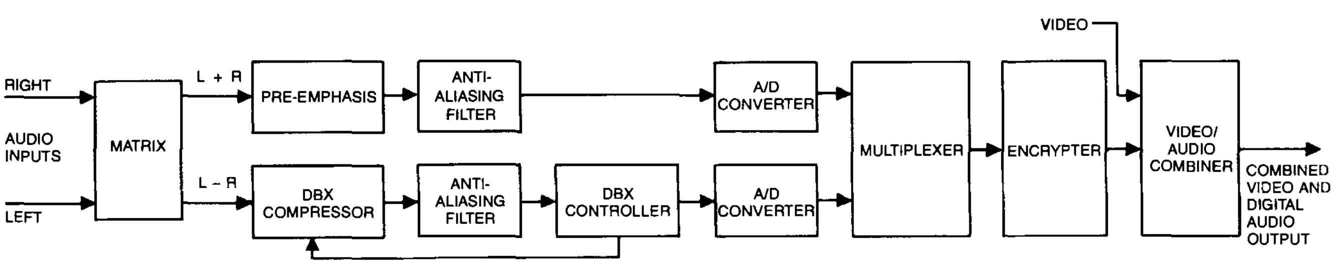

FIGURE 5. ENCODER UPGRADED FOR MTS COMPATIBILITY

to the horizontal scan rate, and to each other within +/-3 degrees; in addition, the pilot tone, at least, must be a relatively clean sine wave.

In the interests of preserving maximum stereo separation, given that the accuracy of the home stereo TV is unknown, it behooves us to make the cable portion of the stereo system as accurate as possible. Unfortunately, this aggravates the economics even further.

Ironically, with a digitally-encoded audio transmission system, and unlike the ordinary stereo-difference subcarrier approach, there is no noise penalty to the second channel. This means that the DBX compression is required simply to match the MTS format of non-encrypted stereo channels.

The answer is to do the DBX compression in the headend (reference Figure 5). There, it only needs to be done once, rather than loading the cost of each decoder; moreover, in the context of headend costs, the DBX compression is relatively inexpensive. The scheme, then, is to matrix the left and right channel audio inputs into sum and difference signals and DBX compress the difference signal; then the normal sum (monaural) and the compressed stereo difference signal are digitized, encrypted, and transmitted.

This approach yields a pair of signals which, while not yet completely in the MTS format, are nonetheless halfway there, and have, as yet, incurred no bandwidth penalties. Moreover, the most expensive portion of the MTS formatting has been accomplished with minimum cost impact to the system. After decryption, it only remains to modulate the compressed difference signal onto the subcarrier, add a suitably accurate and stable pilot tone, and sum the composite result onto the monaural sum signal.

THE DECODER
(Reference Figures 6 and 7)

The real challenge is in the decoder. Double sideband suppressed carrier modulation of the decrypted stereo difference signal is not at all difficult. Generating a clean, sinusoidal pilot tone that is accurate in amplitude and stable in phase with respect to the generated stereo difference subcarrier, however, can be expensive.

In principle, the problem appears simple: one merely phase-locks a pair of sine-wave oscillators, one at horizontal rate and the other at twice horizontal rate, to the incoming horizontal synchronizaton signal, such that the two sine waves have the proper phase relation to each other. The H-rate sine wave is then the pilot tone, and the 2H-rate sine wave is the carrier for the stereo difference signal. Unfortunately, this method is both complex and expensive. Voltage controlled sine-wave oscillator circuits with accurate, stable output amplitudes are not simple to make. Digitally synthesized sine waves are stable and accurate but require a great deal of circuitry to produce.

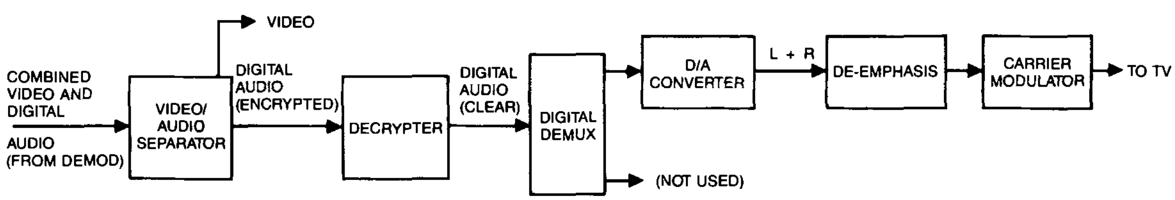

FIGURE 6. DECODER BEFORE UPGRADE. (SIGMA 1)

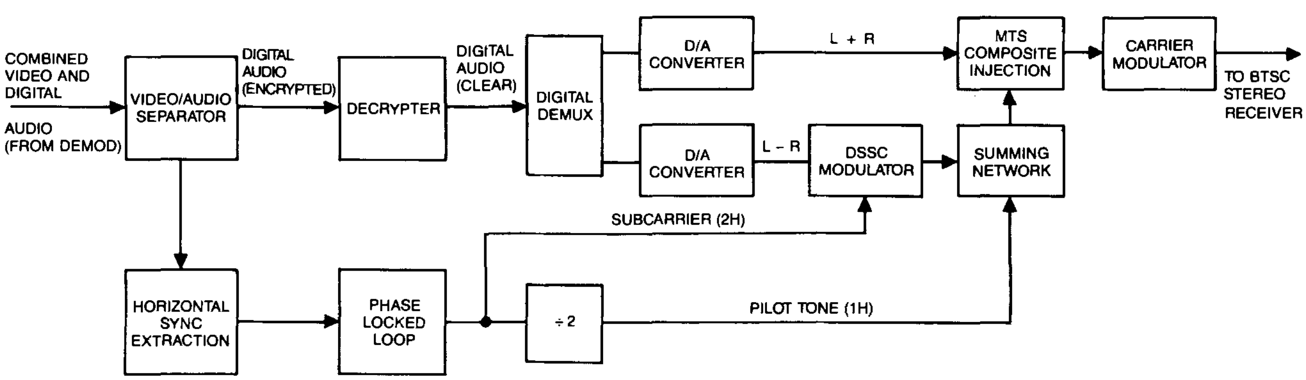

FIGURE 7. DECODER UPGRADED FOR MTS-COMPATIBLE ENCRYPTED STEREO. (SIGMA 3D)

One alternative is to generate the timing digitally and filter the resultant square waves into sinusoidal form. This also presents difficulties: the principle frequency component to be rejected is the third harmonic, which is only about 1.5 octaves above the fundamental. The filter required is large, complex, difficult to tune, and expensive to make stable; and since the desired sine wave is necessarily at the edge of the passband, the filter generates a great deal of phase shift and thus a great deal of phase uncertainty -- not an attractive proposition for a system with an allowable phase error of only 3 degrees.

The decision was made to utilize the advantages of digital timing to solve the sine-wave problem in an unusual way. An analog integrator, dating back to the days of analog computers, has the unique property of a fixed, stable, 90-degree phase shift, unaffected by component tolerances or drift. One integrator thus converts a digitally-generated square wave into an accurate triangle wave, with a fixed 90-degree phase shift. A second, identical integrator converts the triangle wave into a parabolic approximation of a sine wave, with an additional 90 degrees of phase shift. The result is a sinusoidal wave which is accurately and stably 180 degrees out of phase with the digitally generated input square wave -- that is, it is inverted.

The double mathematical integration can actually be accomplished even more simply. The smallest circuit realized to date consists of one transistor, three capacitors, and four resistors; it produces a sine wave in which the third harmonic is 30 dB below the fundamental, and whose phase shift with respect to the input is within specifications.

The double-sideband suppressed-carrier modulation of the stereo difference signal is done with a simple double balanced modulator, and the previously formed pilot tone is resistively summed into its output. The resultant composite signal is then injected into the normal (monaural) audio path at a point where its presence or absence won't otherwise affect the monaural signal, and the entire stereo difference/pilot circuit is simply switched on or off to accomodate the mode of the input program.

The decoder's remodulator thus supplies a normal Channel 3/Channel 4 MTS stereo signal to the subscriber's stereo TV, regardless of the type of input channel -- standard or encrypted premium.

SYSTEM FLEXIBILITY CONSIDERATIONS

Temporarily at least, provision must be made for subscribers who do not yet own a stereo TV, yet still who want to receive premium stereo programming. Moreover, it must be done in a way which maintains the security of the premium programming, without complicating the subscriber's life.

The subscriber without a stereo TV must use his hi-fi system to hear stereo. Existing systems may offer stereo simulcasts in the FM band, but this is cumbersome for the subscriber and expensive for the MSO (redundant equipment). What is needed is a baseband audio Left and Right channel output from the decoder that the subscriber can feed directly into his hi-fi, without further fuss. Ironically, the problems are just the reverse of the system described above. A baseband stereo encryption system can deliver Left and Right audio signals directly, but for standard MTS channels, a stereo decoder must be provided. Fortunately, incorporation of consumer-grade MTS decoder circuitry into the cable decoder is not prohibitively expensive, and the system has the added advantage that the DBX expander required for MTS decoding may also be used to expand the DBX compressed stereo difference signal generated by the encryption system.

EIA MULTIPORT

The EIA has recently completed its "EIA Multiport" standard (IS-15), which provides a baseband interface between the TV receiver and the descrambler. This interface feeds demodulated, baseband scrambled video to the decoder's descrambling circuitry, and accepts the baseband, descrambled video output for reinsertion into the TV's circuitry. The process minimizes cost and reliability problems by avoiding the existing redundancy of RF circuitry: at present, all baseband descramblers must contain both a complete TV tuner and a complete TV remodulator -- circuitry which is already present and operating in the TV receiver itself.

The decoder supplied for Multiport-compatible TV's is as simple as those described above. The decrypted sum channel is simply de-emphasized; the decrypted difference channel must be DBX expanded. The recovered sum and difference signals are then dematrixed to form the desired Left and Right channels which are fed back to the TV. Access to the input of a stereo TV's own DBX expander would, of course, further simplify and cost-reduce the system; it is to be hoped that this can be included at a later date.

CONCLUSION

The system described here achieves total stereo/nonstereo TV compatibility, and requires only that the subscriber specify which type(s) of TV he will be using. Subscribers with stereo TV's use cable decoders to convert the decrypted stereo audio into MTS format, and thus supply a standard MTS Channel 3/4 RF output to the stereo TV; those with non-stereo sets use cable decoders containing an MTS decoder, and which feed the resultant Left and Right audio into their hi-fi systems.

ACKNOWLEDGMENT

Credit and thanks are due to Mr. Graham Stubbs, formerly Vice President, Science and Technology for Oak Communications Inc., for the original concept of DBX compression in the headend.

MULTI-CONTROL REMOTE TRANSMITTER

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ABSTRACT

The undesirability of requiring a viewer to juggle several transmitters in attempting to control his decoder, VCR, and TV receiver has long been recognized. Early attempts to alleviate this problem developed transmitters which could control two units provided the units were from the same manufacturer and the user could remember to operate a slide switch or shift key to select which device was to be controlled. More recently, various manufacturers have developed units which can control VCR's, television receivers, and decoders from more than one manufacturer. One technique used for this is to "train" the transmitter to generate the proper IR format by exposing it to the IR codes that it is to replicate. A second approach is to program algorithms for generating different IR formats in firmware and then provide some means of telling the device which algorithm to branch to when it is to transmit, say, "Channel Up" for a decoder or "Pause" for VCR. The shift key or slide switch inconvenience could be eliminated by providing separate keys for all functions, but this would make for a very crowded keyboard. An improvement on this is to retain the slide switch but provide multiple keys for such functions as "On/Off", thus minimizing the need for operating the switch.

INTRODUCTION

The history of remote control transmitters is replete with example of man's continuing attempt to make television-viewing effortless, both physically and mentally. The earliest transmitters had few functions, typically power, channel change, volume change and possibly mute. For other adjustments, made less frequently, the viewer was expected to walk over to controls on the receiver. With the onset of crystal-controlled digital IC's and resultant finer resolution of signals, it became possible to include more controls on the transmitter. The forced requirement that UHF tuning capability be added to all receivers made more channels available. This made direct access highly desirable, adding ten or so keys. More and more buttons appeared, adding conveniences such as "Flashback" and signal source selection along with exotic features such as "Zoom" and telephone dialing.

The popularity of accessories such as video-cassette recorders (VCR's) and cable converter/decoders dictated that they, too, should be controlled from the easy chair. The early days saw the user sitting in that chair operating, possibly, three transmitters. With enough practice, the true genius could make a decision to control one of the units, reach for the correct transmitter, pick the proper key for accomplishing the control, press that key, assimilate the feedback that confirmed a command had been sent, received, interpreted, and responded to, and return the transmitter to its proper position, all without missing a beat in munching the turkey drumstick in his free hand. The average person had more of a problem. Simplification was clearly needed.

One phase in simplifying things was the inclusion of controls for TV and VCR in one transmitter. A purchaser of a TV and VCR from a single manufacturer could control both by the simple expedient of manipulating a selector switch, either a slide switch or a push button. The logical extension of the technique, adding a cable unit by incorporating a three-way selector, was slow in coming. This was probably due to the scarcity of manufacturers who have all three types of devices in their product line. In fact, the approach was leapfrogged by various manufacturers who realized that the three-way selector could be used to select a control set for a Brand X TV, a Brand Y VCR, and a Brand Z decoder. Different ways to realize this type of device will be examined in the following paragraphs.

DESIGN CONSIDERATIONS

Anyone involved with the design of remote control transmitters is familiar with such design aspects as keyboard scan, format generation, IR diode currents and the like. To develop a transmitter that controls TV, VCR, and decoder, possibly from three different manufacturers, one must consider four new elements:

1. How does the unit know which device it is to control?
2. How does it know which IR format to generate to control that particular device?
3. How is this information defined?
4. How was knowledge of the various IR formats put into the unit?

SELECTING THE DEVICE

The brute force approach to design of the multi-control transmitter is to provide separate sets of keys for each unit to be controlled. Because the TV, the VCR, and the decoder all probably respond to a direct channel number entry, the design thus starts with thirty digit keys. Add six more for up/down scan and four for volume up/down and it quickly becomes obvious that the final key count is in the 50-70 neighborhood. The operator of such a unit might be required to pass a test and be licensed! Technically, however, determining which device is to be controlled is a relatively simple task. Using a large number of keyboard scan lines (minimum of seventeen for 65-70 switches), the circuit couples any command key with the unit it controls as part of the scan-decoding process.

The next task, clearly, is to reduce the number of keys. The first step is to determine which keys are duplicated in two or three groups. All of the keys mentioned in the preceding paragraph send commands that are valid for at least two devices, and there are several other such keys as well. By controlling a 3-state circuit in some way, one can specify which of three devices the command is meant for. The circuit scans the keyboard to detect that, say, the "Channel Up" key is being depressed and then interrogates the 3-state circuit to determine that the command is to be sent to the cable decoder. With this knowledge, the circuit selects the proper IR data format to command the decoder.

To reduce the number of keys even further, a command used only in the VCR mode, such as "Fast Forward", can share a key with a command used only in the cable decoder, such as "Data Enter". Again, the circuit does a scan to determine which key is depressed and then interrogates the 3-state circuit. The resultant information allows the circuit to select the proper IR format.

There are several forms that may be taken by the 3-state device. The most straightforward is a 3-position slide switch, shown in Figure 1 (a). Alternately, the 3-state circuit can be comprised of flip-flops. The circuit can be forced directly to the proper state for controlling a specific device by use of the set/clear inputs, as shown in Figure 1 (b), or it may be clocked sequentially through the three states, as illustrated in the circuit of Figure 1 (c).

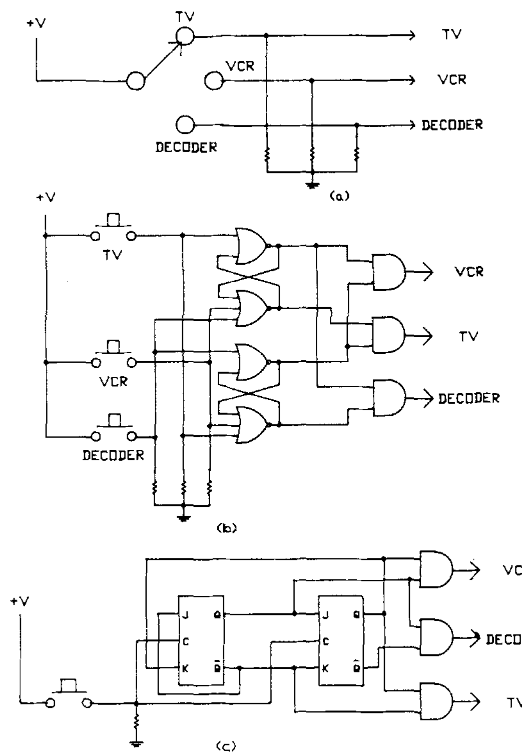

FIGURE 1 THREE-STATE CIRCUITS

The user does not wish to experiment to ascertain the state of his control determination circuit. The 3-position slide switch should have the positions clearly designated. For the circuits of Figures 1 (b) and 1 (c), some sort of readout or indicator lights must be provided, all timed out, of course, to conserve battery power. For keys which have two or more different functions, it is advisable to give the user additional help. This may take the form of color coding (green ≡ TV, etc.), backlit legends, or masked legends.

A compromise between the two key-reduction techniques is to have certain keys such as the digit and channel-scan keys selectable while duplicating the on/off keys. The theory is that in the normal lashup, the TV will have the proper channel selected once and, if it has last-channel memory, will rarely be changed. On the other hand, power will be turned on for more than one of the units the majority of time. The compromise allows minimum switching of the selector circuit, decreasing user confusion and increasing the life of the 3-position switch, if it is used as the selector.

One final effort at key reduction is to generate multiple commands from a single key closure. As an example, it may be safe to transmit a command to turn the decoder power on any time the TV power is turned on, and likewise for power off. The rationale is that in a lashup with a decoder, the decoder is on if the TV is on, while in a lashup without a decoder, no harm is done in transmitting an "On/Off" command for a decoder. Another possible multiple function would be to tune the TV to the decoder output channel automatically any time TV power is turned on. The would-be designer will conceive of others. A word of caution is in order: should two devices get out-of-sync as a result of different IR receiver sensitivities or different physical locations, the result can be confusing and, ultimately, frustrating for the user.

DETERMINING THE IR FORMAT

Once the unit has decoded a keypush and determined that it must tell the decoder to scan channels in an upward direction, it now has to generate the proper format for the IR stream. The unit needs two pieces of information. The first is the actual digital word that is to be transmitted. The designer of the device to be controlled has assigned a digital word to each command. An example might be as shown in Figure 2.

00000	DIGIT 1
00001	DIGIT 2
.	.
.	.
10100	CHANNEL UP
10101	CHANNEL DOWN
.	.
.	.
11111	MUTE

FIGURE 2 POSSIBLE FUNCTION ASSIGNMENTS.

The assignment can vary from manufacturer to manufacturer, even between those using the same off-the-shelf integrated circuit in their transmitters.

In our example, the transmitter ascertains from a lookup table that it must send the digital word 10100. The remaining information needed is how to modulate the IR stream to make 0's and 1's. This format is also likely to be contained in a lookup table. The end result is that the circuit sends out the proper waveform to turn the IR diode on and off in a properly-timed sequence recognizable by the receiving device (a cable decoder in our example) as a "Channel Up" command.

DATA DEFINITION

There are two techniques for storing the information needed to generate the IR stream. For a sampling technique, the IR stream to be replicated is sampled at a number of points and the value of the waveform at each of the points is stored in random-access-memory (RAM). The second method, an algorithmic approach, relies on program steps in firmwave to tell the unit how to compose the IR signal.

THE SAMPLING METHOD

A crude example of the sampling technique is shown in Figure 3 (a). The uppermost illustration is the waveform to be replicated. The sample points and the values stored are also shown. To replicate the waveform, the data can be shifted out of RAM serially using the same clock that samples the data. The end result is shown in the last line. Note that it has reasonable fidelity to the original waveform; in this case that is only a happy coincidence. The accuracy of the reproduction is a function of the phase. In Figure 3 (b), the phase between the waveform to be replicated and the sampling clock is shifted slightly. Because the two are asynchronous, this is an entirely reasonable possibility. Note that in this case, with the same starting waveform and sampling rate, there is considerable distortion in the replicated waveform.

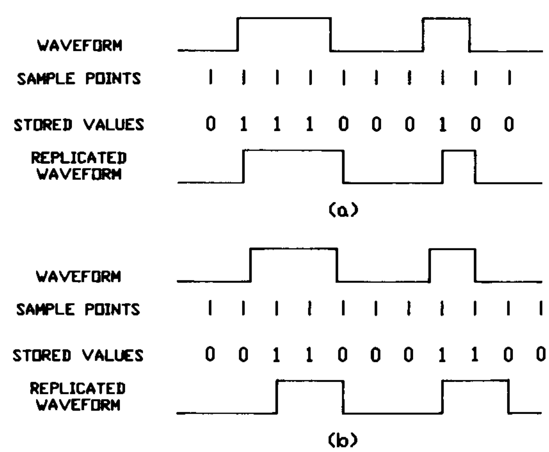

FIGURE 3 WAVEFORMS, SAMPLES, AND REPLICATED WAVEFORMS

To minimize the distortion, the samples must be taken more often. If, say, 100 samples had been taken rather than the 10 in our example, the reproduced waveform would have been truer to the original regardless of relative phases. For high-fidelity reproduction, many samples must be taken, with one bit of memory required for each sample. Consequently, the sampling circuits are very memory-intensive.

ALGORITHMIC APPROACH

To understand this approach, consider the previous example. Sampling the waveform at 100 points required 100 bits of memory. With the algorithmic approach, one can use firmwave to instruct the box to:

"Increment a counter with the system clock. Generate a signal which goes high at W-count, low at X-count, high at Y-count, and low again at Z-count."

Storage of W, X, Y, and Z requires seven bits each, a saving of 100 - 28 or 72 bits. The savings become more impressive if one considers 1000 samples. Here, each count requires storage of 10 bits, a saving of 1000 40 or 960 bits.

COMPARISON

A sampling unit has circuitry that is more memory-intensive; consequently, it will tend to be more expensive. It is also the only device that can be called "universal" with reasonable accuracy. It can be trained to replicate any IR waveform for which its clock rate is high enough to give adequate resolution. The algorithmic unit is limited to only the devices for which it has stored algorithms.

FORMAT INPUT

The last consideration of the design concerns how the information is stored in the device. In the case of a sampling unit, the unit must have exposure to the waveforms it is to reproduce. A typical way to do this is to butt the unit to be replicated against the unit to be trained and to push a function key on the replicated unit while holding down a corresponding key on the unit being trained. An IR receiver in the unit being trained produces a signal to be sampled while decoding of the key being pressed determines where the data samples will be stored.

An algorithmic unit can also be trained. Again, by butting two units together, the unit being trained can circulate through its various IR-producing algorithms to determine if there is a match. An alternative procedure is to put the transmitter into a learning mode, point it at the device to be controlled, and push a function key. The transmitter cycles through its stored algorithms, generating IR outputs from each, until the controlled device responds. At this point, the key is released, storing the data showing which is the proper algorithm for controlling the device.

Still another technique also uses stored algorithms. Rather than undergo a training cycle, however, the unit is configured through a series of switches accessible within the battery compartment. The dedicated switches for the TV receiver can be set to a code which tells the microcomputer which TV receiver is to be controlled, and similarly for the VCR and the cable decoder.

CONCLUSION

The preceding paragraphs have supplied an overview of different techniques that have been employed in the generation of multi-function transmitters. The fact that different manufacturers have taken different approaches should not imply that any is showing better judgement than any other. Rather, it demonstrates that different vendors have different assessments of the relative importance of the different features as each tries to be the one most successful in judging what the market really wants.

MUTUAL EFFECTS OF BTSC STEREO AND RF SYNC SUPPRESSION SCRAMBLING

KENNETH D. LEFFINGWELL

WEGENER COMMUNICATIONS

ABSTRACT

This paper deals with the interaction of BTSC stereo and RF scrambling systems. Both gated and sine-wave sync suppressed systems are examined. The results of separation measurements at various points in the system using both commercial and consumer stereo decoders are presented. Also presented is a subjective evaluation of both video and audio quality variances due to interaction.

INTRODUCTION

In response to the growing anxiety over BTSC Stereo vs cable television scrambling systems, Wegener Communications has been conducting a series of tests over the past months to determine the effects of stereo on a scrambled channel and vice versa. The results of these tests are the subject of this paper.

The video source for this testing was a satellite feed while the audio source changed depending on whether subjective or objective testing was being conducted. Stereo program audio was used for subjective testing; an audio signal generator provided tones for separation tests. Separation was measured by terminating the input to one channel into 600 ohms while injecting audio tones into the other channel and comparing the outputs on an oscilloscope.

A measurement base was established by measuring separation of the composite baseband output of the Wegener Model 1791-02 BTSC Encoder with a Belar BTSC Reference Monitor which was verified to follow the BTSC standard. Measurements were then made at 4.5 MHz using a precision Wegener 4.5 MHz demodulator. Finally, a CATV modulator and converter were added and full system data was taken.

TEST SPECIFICS

Set-up

Initial benchmark testing was conducted to establish a basis of comparison. These tests were performed at baseband - back to back from the composite audio output of the Wegener BTSC encoder to the input of the Belar reference decoder. Separation measurements indicated an average of 38dB between 50Hz and 12KHz. Greatest separation was 49dB at 3KHz while minimum separation measured was 32dB at 12KHz. Next, measurements were made at the RF output of the BTSC encoder. A precision Wegener subcarrier demodulator was used to demodulate the 4.5 Mhz output of the BTSC encoder. The demodulated BTSC encoded stereo signal was input to the Belar reference decoder to make the measurements. Average separation of 33dB was measured between 50Hz and 12KHz with greatest separation of 38dB at 2KHz and minimum separation of 29dB at 12KHz. Finally, the subcarrier output of the BTSC encoder was used as the input to an Scientific-Atlanta 6350 modulator with channel 26 output modules installed. The modulator output was attenuated to present a level of 0dBmv to a non-descrambling 36 channel cable TV converter. The connecter's output was looped through a Recoton consumer-grade channel 3 stereo decoder and a Scientific-Atlanta 6250 channel 3 demodulator before being terminated at the 75 ohm RF input of a TV/Monitor. The subcarrier output of the channel three demodulator was routed to the input of the 4.5 MHz subcarrier demodulator whose output, in turn, fed the Belar reference decoder. The video output of the channel three demodulator was terminated at the video input of the TV/Monitor. Figure 1 depicts a block diagram of the system. Average separation through the entire RF chain measured 33dB over the 50Hz to 12Khz band. Maximum separation of 36dB occurred at 3, 9, and 10KHz with minimum separation of 31dB at 50Hz and 2KHz as measured with the Belar reference decoder. The consumer decoder averaged 21dB in the same band with a maximum of 26dB at 9 and 10KHz and a minimum separation of 10dB at 12KHz. Figure 2 is a graph of separation measurements observed at the various points in the system.

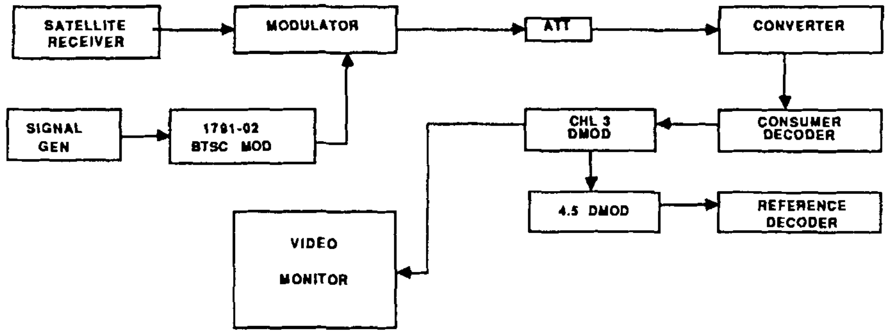

FIGURE 1

SYSTEM BLOCK DIAGRAM - NO SCRAMBLING

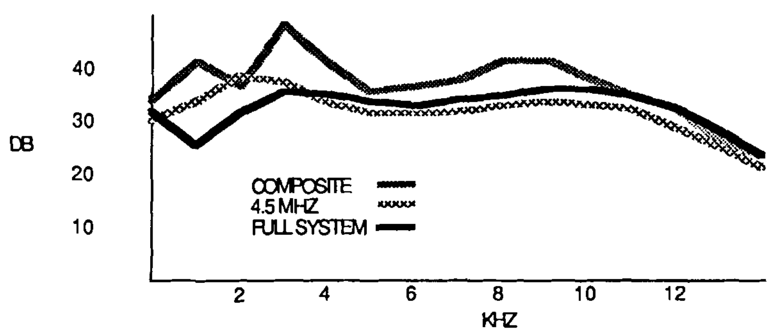

FIGURE 2

NON-SCRAMBLED SYSTEM SEPARATION MEASUREMENTS

Sine Wave Sync Suppression

An Oak Mark V Encoder and Model M35B converter/descrambler were used in this porion of the test. After installing the the encoder/decoder as in figure 3, separation measurements were made. In both the scrambled and non-scrambled modes, separation averaged 23dB in the 50Hz to 12KHz range from the consumer decoder and 25dB from the reference decoder. From the data observed, it was apparent from this data that sine wave sync suppression has very little effect on separation. As was expected with this system however, there were some audio components perceptable in the video. To determine how much of this interaction was due to BTSC, a switch was installed on the rear panel of the BTSC encoder which could select either a stereo or mono output. By switching between stereo and mono audio while observing the video, a qualitative evaluation of BTSC-related video degradation was made. Worst-case degradation occurred when video and the 4.5MHz audio subcarrier were combined at the video input of the modulator and the audio source was a constant tone. The audio interaction in this configuration was plainly visible when observing the effect using program video demodulated by a consumer grade television receiver. In comparison, when program audio was used and the video applied to the modulator separately from the 4.5MHz audio subcarrier the interaction was barely perceptible when observed using a flat field video signal and viewed on a video monitor. This may imply that a system utilizing this scrambling system would be advised to separate the video from the audio subcarrier if at all possible.

Gated Sync Suppression Scrambling

Both Pioneer and Scientific-Atlanta scrambling equipment were tested. As data taken were similar, and to avoid product- specific performance, the results are presented as an average of the two.

Again, scramblers were installed as in figure 3. Measurements were made in both scrambled and non-scrambled modes. In the non- scrambled mode separation averaged 33dB in the 50Hz to 12KHZ range on the reference decoder and 20dB using the consumer decoder. In the scrambled mode separation averaged 22dB with the reference decoder and 19dB with the consumer decoder. Please refer to figure 4 where the measurements are graphed. The data clearly indicates that gated sync suppression scrambling does degrade separation. However, according to Philys Kurtz, in her article "Maximum Separation" in the September issue of Television Broadcast, a delivery of 18dB of separation to the home is the broadcast industry target with a typical figure in the low 20's. Therefore, even though separation is dimminished somewhat by this scrambling technique, the separation is still in a range to compete effectivley with "off-air" delivery and present acceptable stereo to the subscriber.

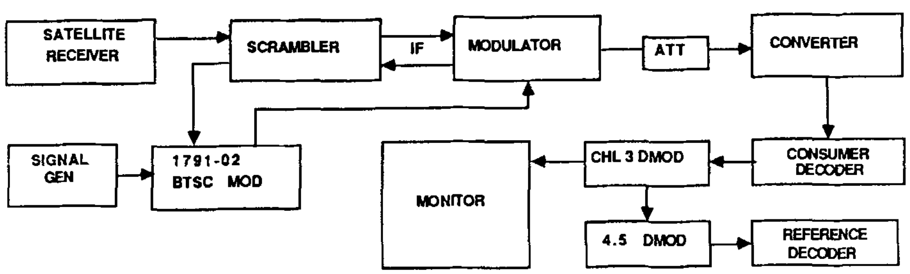

FIGURE 3

SYSTEM BLOCK DIAGRAM - WITH SCRAMBLER

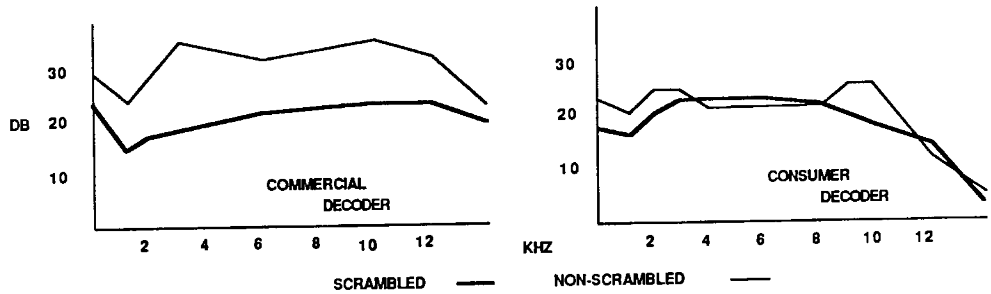

FIGURE 4

SEPARATION MEASUREMENTS IN SCRAMBLED AND NON-SCRAMBLED MODES

Video quality is not perceptibly impaired when BTSC stereo is introduced into the gated sync suppressed system. Pictures were stable - no descrambler "break-up" was observed. And, video quality did not perceptbly change when the audio programming was alternated between stereo and mono.

The most obvious detriment in a gated sync suppression system is sync-buzz. However, this phenomenon is more rightly associated with the scrambling system rather than BTSC stereo. The observed data did not significantly change when alternating between stereo and mono. There was a significant difference observed in sync- buzz level when non-scrambled mode was compared to scrambled mode. This measurement was made by setting a reference tone level at the output of the stereo decoder. The input of the encoder was terminated and the noise floor measured. this test was repeated at 1KHz, 5KHz, and 10KHZ in both scrambled and non- scrambled modes. Intuitively it would seem that the lower frequencies would be most affected. And, the data indicated this to the the case. At 1KHz in the non-scrambled mode, signal to noise measured 45 dB while in scrambled mode signal to noise dropped to 37 dB implying an 8dB noise contribution due to scrambling. At 5KHz and 10KHz signal to noise in and out of scrambling mode measured within 2dB.

CONCLUSION

After consideration of the observed data it is concluded that BTSC stereo present's no insurmountable obstacles to implementation. Inherent constraints of the scrambling techniques examined here brought to light concerns particular to each. In the case of sine wave sync suppression the major concern is video quality, which can be assured through careful treatment of the video/audio subcarrier relationship. It is recommended that the BTSC encoded audio be introduced to the modulator as a 4.5 MHz subcarrier separate from the video. On the other hand with gated sync suppression the primary concern is noise on the audio, some of which is introduced by direct modulation of the audio carrier by video. However, the major contributor in this case is the descrambling pulse. Here depth of video modulation, scrambler set-up and audio input levels are the critical parameters. All three of these will have an effect the apparent noise contribution.

Finally, a couple of general statements about BTSC. One, leave as many adjustments as possible to the factory (in particular audio deveation which is critical to separation). Two, in order to reduce interaction keep co-processing of video and audio to a minimum (ie., don't mix the audio subcarrier with the video until it's necessary). Keeping these two generalities in mind, the following order of preference can be stated for the BTSC encoder/CATV modulator interconnect.

1. 4.5MHz audio subcarrier separate from video
2. Video and 4.5MHz combined
3. Encoded baseband directly to the modulator audio inputs

BTSC stereo is compatible with the scrambling techniques tested and easily implemented in a cable television system. Neither is absolutly transparent; however both will co-exist very well. Cable systems employing these scrambling systems should have no problem enhancing the services offered with the addition of BTSC stereo.

PERCOL: A POLYURETHANE SYSTEM TO SIMPLIFY LAYING OF BURIED CABLE

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ABSTRACT

A liquid polyurethane system developed originally for rapid repair of bomb-damaged Air Force runways may simplify, expedite, and substantially reduce the cost of laying buried cable in urban streets and across highways, according to field tests conducted in a cooperative program between Prime Cable of Austin, Texas, Simpson & Sons, Las Vegas cable contractors, and ARNCO of South Gate, California, developer of the urethane system. Testing has demonstrated that highway traffic can be resumed in 30 minutes after coaxial cable is encapsulated in a (three-inch-wide by six inches deep) trench sawed into the pavement. When optimized, the system could represent a benefit of as much as $100,000,000 to the cable industry.

INTRODUCTION

The cost of installing underground coaxial cable may exceed the cost of the cable itself by an order of magnitude. Depending on specification details, a typical figure for television cable per se, including associated electronics, might average $5,000 per mile. Although stringing on existing poles can run as little as one or two thousand dollars a mile, make-ready costs to provide clearance room on old inner-city aerial routes -- often already overburdened -- can run $10,000 to $15,000 per mile before the cable is hung, and another $4,000 to $12,000 in labor afterward.

Similarly, at two or three dollars per lineal foot, the labor expended to bury cable in soft dirt in the suburbs can be as cheap as $10,000 per mile; but if the cable must be laid under existing urban pavement, this figure can escalate to $50,000 or $60,000 per mile for the labor component alone.

According to recent statistics compiled by Cable TV Technology, about 25,000 new miles of cable will be installed during 1987, and another 25,000 miles is slated for rebuild.[1] Resulting capital costs for the nation's cable companies are becoming astronomical.

Some cable companies preparing to serve new communities (many of which prohibit utility poles) have managed to circumvent such costs by working with the land developers, placing cable underground early on (at a total cost of perhaps $15,000 per mile), before streets are graded and paved -- thereby, however, making their investment as speculative as that of the real estate developers who bet their money on the success of the anticipated community. Others in this situation find themselves faced with an impractical time-restricted "no-cut" stipulation that may obviate excavation for cable work beneath streets and sidewalks for years. A cable company in Nevada recently thought they had solved the problem by leasing space in telephone cable conduit, only to be crowded out as community telephone service expanded -- not an uncommon occurrence.

In the downtown areas of major cities there is simply no choice. Streets must be barricaded as the pavement is excavated and traffic can be disrupted for weeks at a time. Common practice involves a ditch eight to twelve inches wide and 18 inches deep -- completely through the pavement (which is usually asphalt, and sometimes asphalt over brick). After the cable is laid, a concrete slurry is poured into the ditch, taking from three days to a week to cure, during which traffic is interrupted. Usually weather conditions dictate a temporary cold patch until it is possible to return and finish the job with a hot T-cap, disrupting traffic once again.

Both Boston and Chicago have recently been through this unhappy, expensive process -- as similar work commences in Philadelphia, Baltimore, Detroit, Washington, D.C., and the boroughs of New York. In Philadelphia alone, some 3000 miles of cable will be laid. In fairness, a number of cable companies are active in pioneering new technologies; but for the most part, antique sewer line and waterworks techniques prevail. Clearly, a better process is needed.

THE PERCOL SYSTEM: BACKGROUND

Prime Cable of Austin, Texas, had been investigating several new technologies when it became managing partner of Community Cable TV of Las Vegas (the Nevada company mentioned above). Prime Cable encouraged its contractor, Simpson & Sons, to carry out field tests on PERCOL, a polyurethane system that Simpson was investigating at the time. PERCOL had been originally developed for the U.S. Air Force by ARNCO, a California company which cooperated in the venture.

The original PERCOL system was developed to meet the Air Force objective of being able to resume flight operations from a bomb-damaged runway within four hours after bombardment, under all weather conditions, including heavy rainfall. The method for repairing bomb craters had been to backfill the crater with debris, followed by a layer of compacted select fill -- after which panels of heavy aluminum matting were installed, interconnected, and anchored to adjoining sound pavement. As Air Base Survivability Reasearch Engineer, Capt. Daniel J. Pierre explained, "...the matting, although structurally sound, is labor-intensive, creates a bump in the pavement that can damage fighter aircraft, does not perform well with large cargo aircraft, and interferes with tailhook barrier engagements by fighter aircraft."[2]

Developed by chemist Ransome Wyman (president of ARNCO and co-author of this paper), PERCOL is a two-component polyurethane system now available in several combinations of physical property criteria, each tailored to the requirements of the paving materials with which it is to be used. In all cases, equal volumes of low-viscosity, low molecular weight, coreactive liquids are pumped from a pair of drums through a static mixer utilizing simple equipment, as pictured in Figure 1.

Fig. 1 - Simple equipment is used to pump liquid polyurethane "A" and "B" components from a pair of drums through a static mixer and nozzle.

Fig. 2 - Air Force technician floods a debris-filled bomb crater with reactive PERCOL mixture, which gels in two minutes and cures in 30 minutes.

Table 1 - INITIAL PHYSICAL PROPERTIES OF PERCOL VS. CONCRETE[2]

Property (psi)	Percol Polymer Concrete (psi)	Portland Cement Concrete (psi)
Flexural Strength	1300	500
Tensile Strength	500	425
Modulus of Elasticity	1.62×10^6	4.065×10^6

AIR FORCE EXPERIENCE

For the Air Force application, the PERCOL chemistry was adjusted to behave hydrophobically, to permit its application during rainfall. The reactive mixture was delivered through a nozzle, flooding the crater and percolating through compacted backfill debris as shown in Figure 2. Catalyzed so as to "gel" in approximately two minutes, the PERCOL hardened and cured to a rigid solid mass in approximately 30 minutes, bonded to the concrete and sufficiently strong to support aircraft take-offs and landings.

Physical properties of ARNCO's original (rigid) PERCOL formulation, after 30 minutes were compared with adjacent cured concrete by the Air Force, and reported as shown in Table 1, above.

It should be noted that the above properties for Portland cement concrete were measured after 28 days; those for Percol R after 30 minutes. Percol R properties may be expected to achieve their maximum values after a week or more of stabilization, depending upon ambient temperatures.

The report also concluded that "Field tests of the polyurethane polymer concrete demonstrated the resolution of problems relating to warpage, expansion, placement technique, and flammability."[2]

CALTRANS TESTING

The performance of PERCOL as an "instant runway repair" system at Tyndall Air Force Base in Florida and elsewhere led to interest in its application as a highway patching medium for repairing chuck holes, bonding alligatored pavement, etc. -- as, indeed, was suggested in the above Air Force report.

Comprehensive testing of PERCOL by the California Department of Transportation (Caltrans) began in January, 1985 under the direction of Senior Materials Engineer Leo Ferroni, P.E. Adhesion to asphalt concrete was extensively investigated both in the laboratory and in the field, with installation temperatures ranging from 22°F to 35°F, at which temperatures it was observed that "the center of the patch set up in five minutes, but the outer edges did not set until 10 minutes," after which a pickup truck was parked on the patch and driven back and forth without visible damage.[3]

Results of laboratory testing, as reported by Caltrans, are shown in Table 2, below.

Table 2 - BOND STRENGTH OF PERCOL TO ASPHALT CONCRETE[3]

California Tentative Test	24 Hours	30 Days
3-Point Modulus of Rupture on 3" X 3" X 9" block samples	2665 psi	1565 psi
Bond Stress to SSDPCC, 1-Point Modulus of Rupture	515 psi	310 psi
Bond Stress to Dry PCC 2-Point Modulus of Rupture	1250 psi	775 psi

Fig. 3 - Crater (near center line) on U.S. 97 repaired with PERCOL and rock aggregate by California Department of Transportation.

Fig. 4 - Caltrans PERCOL test on Highway 118, which carries average daily traffic load of 75,000 vehicles in each direction.

It was noted in the above report that only one of the samples broke at the bond line. These results led to additional Caltrans field tests a few weeks later in northern California near the Oregon border. As reported in the trade magazine, **Highway and Heavy** Construction:

"U.S. 97, a favorite shortcut for truckers off I-5, is two lanes of asphalt concrete, puckered by alligator cracks, ruts, potholes, and depressions, subjected to freeze-thaw conditions for nearly six months each year. Though only moderately travelled, (approximately 2100 vehicles per day), half of its load is truck traffic, most with five or more axles.

"In late January, maintenance personnel poured samples of polyurethane cement over 5/16-in rock in two ragged potholes and a deep rut and flooded more resin into a four-in. deep, 10 X 3-in. depression filled with the same aggregate...Heavy traffic was rolling over the patches 20 minutes later. The outside temperature was 35 deg. F."

One of these patched craters may be seen near the center line of the road in Figure 3. During the ensuing nine months, the patches were exposed to heavy rains, snowstorms, and continual freeze-thaw cycles that occasionally exceeded a 40°F temperature change in a single day--from a low of -5°F.[3] [4]

A year later observers reported the patches to be "still intact, while the asphalt around them continues to spall."[4]

Figure 4 shows another Caltrans test location in which a large area of scaling concrete (caused by reactive aggregate) was repaired on State Highway 118 in the Simi Valley northwest of Los Angeles -- a six-lane thoroughfare carrying an average daily traffic load of 75,000 vehicles in each direction. Here, The crumbling surface of the road (on which temperatures of 125°F have been recorded) was flooded with PERCOL

Fig. 5 - Core sample from above repaired highway pavement shows good bonding at surface and PERCOL penetration up to 2 1/2 inches deep.

to test the product's ability to bond the fractured area together, and to determine if the polyurethane would penetrate sufficiently to seal the cracks and prevent further deterioration. After squeegeeing the liquid material over the area, and before gelation, the road crew broadcast sand to insure a non-skid surface.

Core samples taken from this area six months later (see Figure 5) showed good bonding at the surface, and demonstrated that the polymer had penetrated as deep as 2 1/2 inches as it percolated through the substrate.[4]

REFORMULATION

PERCOL's chemistry involves an addition reaction of an isocyanate group to the hydroxy group in the polyol component. Because of its controllable reactivity, the system can be tailored to provide a variety of physical properties to optimize compatibility with the paving materials with which it is used.

Retrospective analysis of the results of the above Caltrans field tests suggested that for highway service a more flexible form of PERCOL was desirable -- one that could accommodate movement between adjacent slabs of pavement and would also be more compatible with asphalt concrete. To this end, the original PERCOL was reformulated.

Now designated as PERCOL FL (formerly PercoFlex), properties of this elastomeric form of PERCOL are:

Table 3 - PERCOL FL PHYSICAL PROPERTIES

Durometer (Shore A)	95
Tensile Strength	2200 psi
Elongation	175%
Tear Strength (C)	235 pli
Specific Gravity	1.1

As with all PERCOL formulations, PERCOL FL reacts with low exotherm, which results in performance advantages. Since the reactive mixture does not heat up significantly, PERCOL avoids the residual stresses upon cooling which are characteristic of other more-exothermic polymer systems. The addition of elastomeric flexibility coupled with high strength also provide relief of stresses which might otherwise be induced by differences in coefficient of thermal expansion between the polymer and parent pavement.

PERCOL's low-volatility isocyanates obviate any need for protective clothing or other special safety precautions to maintain OSHA air quality standards.

NON-MILITARY AIRCRAFT RUNWAYS

The first real-world applications of PERCOL FL have been on aircraft runways. PERCOL-repaired areas of civilian aircraft runways are currently being evaluated at major international airports. Figure 6 shows a typical section of fractured concrete on an airport taxiway, which has been jack-hammered back to solid concrete. In Figure 7, the cavity has been filled with coarse, dry aggregate, which was then flooded with PERCOL, as seen in Figure 8. Again, the wet polymer was coated with sand to restore surface friction. Figure 9 shows the repaired taxiway supporting a 1200-ton aircraft.

Figure 10 illustrates a squeegeeing technique employed to apply PERCOL to a model airplane runway which had severely deteriorated. After sand was broadcast (Figure 11), the result was an exceptionally smooth, non-skid, aesthetically neutral surface. Figures 12 and 13 compare the condition of the runway before and after repair.

FIELD TESTS WITH CABLE

When some of the above references came to the attention of Simpson & Sons, cable contractors in Las Vegas, Nevada, they invited ARNCO to participtate in a series of pragmatic field tests to evaluate adaptation of PERCOL technology to cable-laying operations. Several demonstrations were arranged during 1986 for cable industry representatives.

The technique employed was a modification of one that has been utilized for installing electrical power conduits for aircraft runway lights: laying the power cable in a shallow trench in the pavement and covering it with a polymer concrete (frequently expoxy in this application).

Figure 14 shows the 36-inch carbide-tipped rock saw, built by J.I. Case, which was used in Simpson's Las Vegas demonstrations. For cable installation in the asphalt surface, the operator rapidly cuts a groove approximately 2 3/4 inches wide by six inches

Fig. 6 - Fractured concrete pot-hole on aircraft runway which has been jack-hammered back to solid concrete.

wide by six inches deep, as shown in Figure 15. After the cable is laid in the shallow trench, asphalt debris from the saw cut is swept back into the groove (Figure 16). If the debris volume is insufficient to fill the trench or is otherwise unsuitable, ARNCO recommends dry, washed pea gravel or recycled asphalt; cold mix has also been successfully used by Simpson.

Utilizing the truck-mounted equipment pictured in Figure 17, a

Fig. 8 - Aggregate filled cavity is flooded with reactive PERCOL polymer.

Fig. 7 - The cavity is filled with coarse, dry aggregate.

double-barrelled positive displacement pump delivers equal volumes of liquid "A" and "B" PERCOL components from a pair of drums through parallel hoses, to the dispensing wand, where they are intimately blended in an integral 21-element motionless mixer before flowing out of the discharge nozzle into the trench, as shown in Figure 18. Percolating through the aggregate, the reactive mixture gels in about two minutes and is capable of supporting traffic 30 minutes later. Sand, broadcast over the surface before gelation of the poymer (Figure 19), restores both surface friction and surface aethetics, making

Fig. 9 - A 1200-ton aircraft taxis over the PERCOL-repaired runway.

Fig. 10 - PERCOL reactive liquid is squeegeed over the surface of a deteriorated model airplane runway.

Fig. 11 - Broadcasting sand over the reactive PERCOL mixture before gelation.

the cut for the installed cable essentially invisible.

Figure 20 is a cross section of PERCOL-encapsulated coaxial cable in asphalt pavement, illustrating bonding to the adjacent asphalt. As has been pointed out by Simpson, such encapsulation provides abrasion protection to the aluminum cable jacket under dynamic traffic loading. Figure 21 shows how cables can be "double-decked" by staggering the process.

Early Simpson field tests revealed moisture sensitivity of the reaction, which was later overcome by (1) adjusting the catalysis of the system, (2) drying the aggregate, and (3) selecting aggregate materials that did not interfere with the reaction. Limestone, for example, should be avoided not only because of its contained water of crystallization, but because of its alkalinity. Both recycled asphalt and cold mix proved to be quite compatible with PERCOL.

ECONOMICS

While the above process represents a substantial reduction in labor and installation time compared with present underground cabling techniques, it may be expected to yield a polymer concrete that is of the order of 50 percent polyurethane. To further improve the economics of the system, work is underway in an effort to substantially decrease polymer content.

Fig. 12 - Showing condition of model airplane runway before repair.

Fig. 13 - Model airplane runway after repair, with bonded-sand surface.

Fig. 14 - 36-inch carbide-tipped Case rock saw used to create 2 3/4 X 6 inch trench in pavement.

Fig. 15 - Trench sawed in asphalt pavement in which cable will be laid.

Fig. 16 - Asphalt pavement debris from saw cut is swept back into trench, supplemented by washed, dry pea gravel, cold mix, or recycled asphalt.

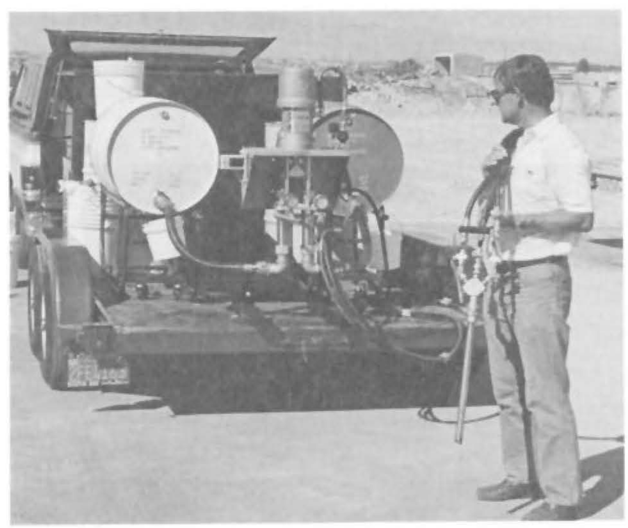

Fig. 17 - Truck-mounted drums containing "A" and "B" components of PERCOL polymer. Equal volumes of coreactant liquids, pumped by a double-barrelled positive-displacement pump, intimately mix in a 21-element motionless mixer in the handle of the operator's nozzle.

Fig. 18 - Reactive PERCOL mixture flows out of nozzle into trench, percolating through the aggregate and encapsulating the cable.

Fig. 19 - Before 2-minute gelation time expires, sand is broadcast over wet surface to restore friction and appearance.

Premixing of polymer concrete with multiple grades of screened aggregate tailored to create minimum voids between particles is an avenue that has been explored. But use of conventional mixing equipment would probably entail lengthening the short pot life of the reactive PERCOL mixture, which is intrinsic to the labor saving and speed of installation offered by the process.

Figure 22 is a polymer concrete machine developed by Respecta, a German firm (with American representation in Chicago), for Dural International Corporation, construction polymer specialists in Deer Park, LI, New York, that appears to offer considerable promise. This equipment meters liquid-polymer components and dry fillers into a mix barrel -- the long cylinder in the

Fig. 20 - Cross-section of PERCOL-encapsulated cable in asphalt pavement, showing intimate bonding of the materials.

Fig. 21 - Showing how cables can be double-decked by repeating the PERCOL process in two stages.

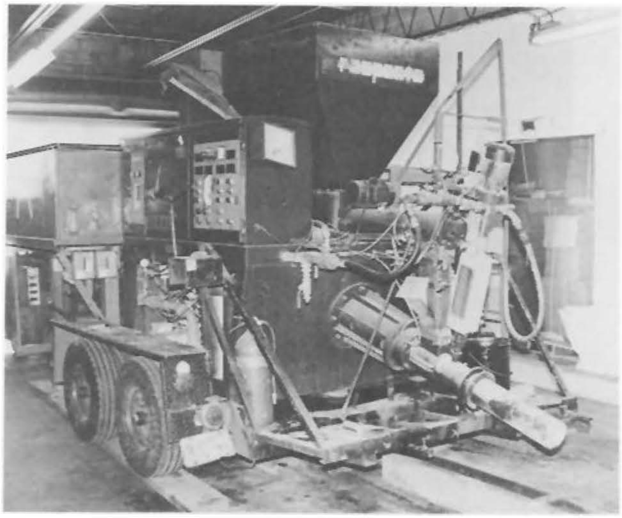

Fig. 22 - Respecta polymer concrete machine developed for Dural Corp., which will continuously meter and mix reactive PERCOL polymer with aggregate and deliver mixture into trench.

foreground, which contains a mechanical mixer. The machine is capable of delivering a continuous stream of reactive polymer concrete effluent into the grooved pavement as the machine is drawn along stradling the ditch, at a productivity rate of up to 8 cubic yards per hour. (Higher productivity rates are available, according to Repecta; depending on output volume and options selected, these machines sell for $100,000 to $200,000.)

Using its Respecta equipment, Dural has developed a five-grade aggregate mixture that contains 15 to 20 percent PERCOL polymer, yielding an estimated total material cost in the range of $3 to $4 per foot for a 3 by 6-inch cable trench. Further reductions in polymer content are possible. After optimization, it is anticipated that the equipment could move along at perhaps 1000 feet per hour, easily laying more than a mile of cable per day.

In a recent issue of **Communications Technology**, Dan Pike, Vice President of Engineering for Prime Cable in Austin, Texas, suggests that, when optimized, the process offers savings "as much as 30 to 60 percent in labor alone," compared to conventional trench installation.[5] Indeed, at an estimated cost reduction of $2,000 per mile after optimization -- and with 50,000 miles of cable currently being installed annually -- the process could represent a benefit of $100,000,000 per year to the cable industry.

PERFORMANCE PROPERTIES

For application with asphalt pavements, PERCOL FL, the somewhat flexible, elastomeric form of the polymer has been the PERCOL material of choice. Room-temperature physical properties of PERCOL FL are listed in Table 3, above. To evaluate the characteristics of this elastomer under a variety of temperature conditions, several series of tests were run in the cold rooms of the Arctic Engineering Department of the University of Alaska, Anchorage, as well as in the facilities of the Akron Rubber Development Laboratory.

After soaking in the University cold room over night at -40°F, although PERCOL FL became quite stiff at 40° below zero (having a Shore A Durometer of 100) it was possible to bend thin (1/16-in.) slabs through 180° at this temperature without fracture. In the same cold room the elastomer was evaluated for brittleness. Struck repeatedly with a heavy hammer on a steel anvil, bars of the polymer (12 in. by 2 in., by 3/4-inch thick) proved impossible to break. Severe hammer blows produced a slight flow at the unfractured surface -- a faint depression that disappeared upon rewarming to 75°F.[6]

At -40°F, both the pure elastomer and aggregates bound together by PERCOL FL exhibited great elasticity, bouncing the hammer back more than a foot, suggesting characteristics reminiscent of an ivory billiard ball.

To measure and evaluate relevant physical properties throughout the temperature spectrum, additional laboratory slabs and buttons were prepared and tested at the Akron Rubber Development Laboratory, with the results shown in Table 4 on the following page.

These data suggest that PERCOL FL maintains physical compatibility with asphalt despite changes in temperature; that is, it decreases in modulus as the temperature increases -- reducing any tendency to separate at the bond line. Moreover, the elastomer would appear to remain stronger (tensile strength is 500 psi even at 120°F) and more flexible than adjacent asphalt concrete pavement at all temperatures.

The flexibility of PERCOL FL is illustrated in Figure 23, which shows the results of an experiment performed by Simpson & Sons. Samples of asphalt pavement and PERCOL concrete approxi-

Table 4 - EFFECT OF TEMPERATURE ON PERCOL FL PROPERTIES[7]

Temperature	Durometer*	Elongation	Tension Modulus
6°F	100/100	15%	N/A
37	100/100	140	N/A
60	100/95	175	1920 psi
75	93/80	180	1075
85	82/70	190	568
100	68/60	165	315
120	60/57	155	267
140	56/56	110	265

* First value indicates initial, instantaneous Durometer reading; second is the value to which reading decays after five seconds.

mating four-inch cubes were respectively compressed in a 60,000-lb. press. While the bituminous concrete crumbled and remained in its compressed condition, the PERCOL sample (with encapsulated cable) returned essentially to its original dimensions.

In addition to PERCOL FL, half a dozen chemical variations of the PERCOL system have been released by ARNCO with physical properties tailored for specific commercial applications that are beyond the scope of this paper.

CURRENT WORK

Field testing of PERCOL is currently in progress under the direction of street authorities and highway departments in California, Arizona, Ohio, Iowa, Kentucky, and Alaska, in which state the polymer system is being utilized in such disparate applications as the containment of asbestos-contaminated soil in an environmentally-sensitive area, and the construction of a roadway comprised entirely of PERCOL and volcanic pumice. A research program at Rutgers University in New Brunswick, New Jersey, is also evaluating the product. A bridge deck over the bay at San Diego has just been completed; and some 5000 pot holes on the Santa Ana Freeway in Los Angeles will be repaired with PERCOL systems in the coming months.

SUMMARY

1. As a means of simplifying the laying of buried cable, particularly in urban paved surfaces, PERCOL polyurethane concrete technology offers significant economic advantages over existing common practice. At an estimated cost reduction of $2,000 per mile when optimized, the process represents a potential benefit of $100,000,000 per year to the industry.

2. The techniques described not only have the potential of reducing labor costs, but also avoid the repeated community disruption caused by prolonged traffic rerouting. Traffic may be resumed in as little as thirty minutes after the cable is encapsulated in PERCOL polymer concrete.

3. PERCOL may obviate rework of cut pavement occasioned by temporary measures presently employed in inclement weather during installation. Similarly, if it becomes necessary to take up the cable, the shallower cut required should reduce the cost of re-excavation.

4. PERCOL-impregnated aggregate adheres well to asphalt concrete and performs well as a compatible load-bearing element.

5. Both the physical properties and the cost-performance benefits of PERCOL polyurethane systems compare favorably with other polymer concrete systems.

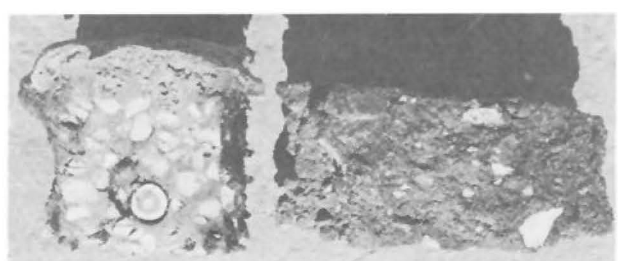

Fig. 23 - After 60,000 lb. squeeze, PERCOL concrete, left, has returned essentially to original shape; asphalt sample, right, remains compressed.

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PERFORMANCE HISTORY IN TWO-WAY CABLE PLANTS UTILIZING A PSK COMMUNICATION SYSTEM

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ABSTRACT

Pay-Per-View video programming has become a new source of revenue for cable operators. To achieve maximum potential, a true, low cost two-way impulse delivery system is needed. Two-way cable systems are the most economical method to supply this need. However, minimal reverse cable plant maintainance results in significant ingress and impulse noise that can hinder two-way data communication. The communication system must be rugged and reliable to work in such an environment. A low cost, coherent PSK system has been developed and field tested. Results indicate that field performance correlates within 1-2 db of white noise tests done in the lab. Also, a safety "margin" of greater than 24 db has been observed in the field.

INTRODUCTION

With the advent of satellite delivery Pay-Per-View (PPV) programming, the need for a low cost, true impulse delivery system now exists. Two-way cable represents the least expensive method of providing this service. However, the typical cable system's reverse plant has significant noise and ingress problems. This is the result of normal forward plant maintainance and minimal reverse plant maintainance. Upgrading these reverse plants to video quality levels is prohibitively expensive. However, a low cost data communication system has been designed to operate reliably in such a system and has been successfully field proven.

TYPICAL TWO-WAY REVERSE PLANT CHARACTERISTICS

A brief description of return plant characteristics is necessary before going into two-way cable data communication performance. A more detailed description of these characteristics can be found in Ref. 1.

There are five major degradation characteristics of a return plant: white noise, ingress, common mode distortion, amplifier nonlinearities and impulse noise.

White noise is the thermal noise generated by a 75 ohm termination. In a two-way cable plant, the headend is the focal point for all signal sources, including white noise from all the terminations and return amplifiers. All the noise is "funnelled" into the headend. This "noise funnelling" increases the thermal noise measured in the return plant at the headend. This noise can be found by using equation 1 assumming a 0 db system.

$$\text{WN floor (dbmv)} = -59 + 10\log(BW/4\text{MHz}) + nf + 10\log(N) \quad (1)$$

where

- -59 = noise generated by a 75 ohm resistor in a 4MHz bandwidth, in dbmv
- BW = bandwidth other than 4MHz
- nf = amplifier noise figure (assuming the same for all amplifiers).
- N = total number of amplifiers in the cable system

Ingress is defined as unwanted external signals entering the cable plant at weak points in the system with the most common weak points being drops and faulty connectors. Common ingress sources are amateur radio operators, citizens band operators, local AM braodcasts and local and international shortwave. Figure 1 is a plot of reverse plant ingress of an existing cable plant. The cable plant has 31,000 subscribers and 2312 amplifiers. The solid bars indicate shortwave bands; the crosshatched bar citizens band; the diagonal bars amateur radio bands; and the dotted bar AM broadcasts. The white noise floor using equation 1 and 100KHz BW is -31 dbmv.

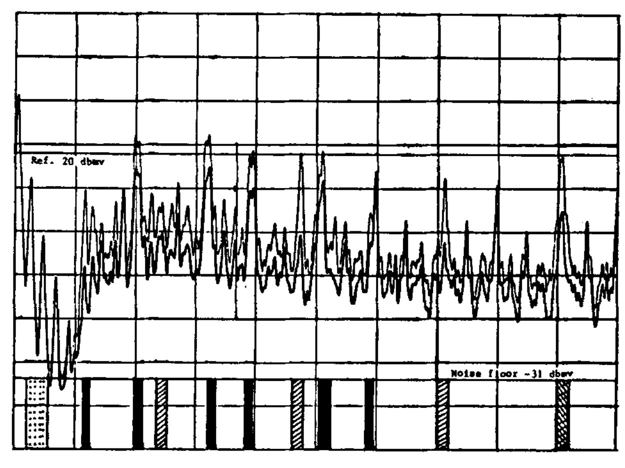


Fig 1 Reverse plant ingress of an existing cable system that has 31,000 subscribers and 2312 amplifiers.

Common mode distortion is the result of nonlinearities in the cable plant that are not due to active devices but rather to passive devices. The nonlinear function is generated by corrosion in connectors, and results in a weak diode effect that creates distortion products. The forward plant contains many video channels separated by 6 Mhz. The nonlinear effect of these "diodes" causes sum and difference frequencies; the sum frequencies fall into the forward plant's spectrum while the difference frequencies fall within the reverse plant's spectrum. Therefore, the reverse plant will have these products at 6, 12, 18, 24, and 30 Mhz.

Amplifier nonlinearities, although not very common, occur when marginally stable amplifiers are misterminated resulting in transmission line oscillations.

Impulse noise is by far the most dominant peak source of noise in a return plant and it is caused by the presence of high voltage lines at close proximity to the cable plant. Two kinds of impulse noise are caused by these high tension lines; corona noise and gap noise. These sources are primarily found on 300KV or higher lines and are random in nature. Weather conditions play a major role in the intensity of these noise sources with humidity being the prime factor. Corona noise is created by the ionization of the air surrounding high voltage lines while discharge or gap noise occurs when insulators break down creating a large spark or discharge. This occurs semi-periodically and takes on a sin x/x frequency distribution. Figure 2 shows a plot of impulse noise in the return plant at the headend of another cable plant.

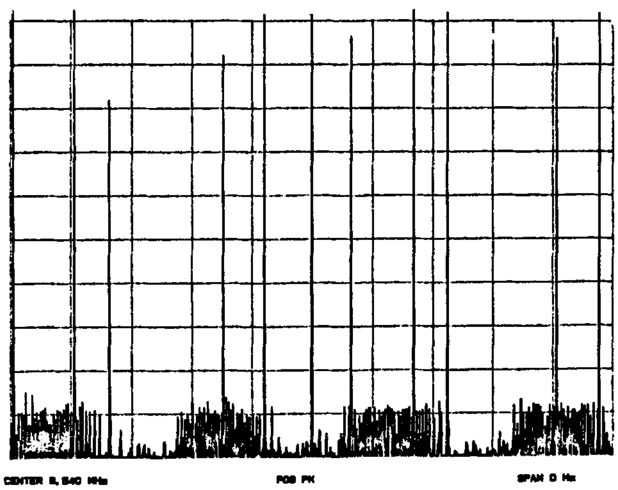


Fig 2 Impulse noise in the reverse plant of an existing cable system.

TWO-WAY CABLE SYSTEM REQUIREMENTS FOR DATA COMMUNICATION

In light of the previous description of typical cable plants, the first major requirement for a data communication system is good noise performance. Noise performance is important in any communication system. However, in the case of a two-way cable plant, the conditions can be severe due to the "noise funnelling." Not only is white noise funnelled into the headend but ingress as well. Therefore, the data communication system must be extremely rugged and have good noise/interference performance.

An advantage of two-way data communication on a cable system is that typical operation allows for data throughput to be less than 100% and still provide satisfactory performance. Return messages not received or received in error can be repeated automatically. This is true for either contention or polling systems.

A second major factor is the cost of the equipment in the system. Obviously, the thousands of home terminal units (i.e. two-way, addressable decoders) connected to the cable system must be simple and inexpensive while the headend equipment may be more complex and costly.

The reverse plant's spectrum is also important. Typically, the 5-30 Mhz return plant spectrum is divided into 4 video channels, T7,T8,T9,and T10. However, rarely more than two of these channels are ever used in the same system. This leaves sufficient room for data communication channels, although this available bandwidth should not be wasted through indiscriminate use of the spectrum. Therefore, minimal bandwidth utilization for data signals is desired.

Power levels of the signals on a return cable plant are required to be within certain ranges in order to not overload or "crossmod" the return amplifiers. This is especially important when there is video information being transmitted through the reverse plant to the headend. Typically, data signal levels into the input of a reverse plant line extender are 20-22 dbmv, in order to keep them 20 db down from any video signal carriers (40 dbmv).

DATA COMMUNICATION TECHNIQUES FOR TWO-WAY CABLE SYSTEMS

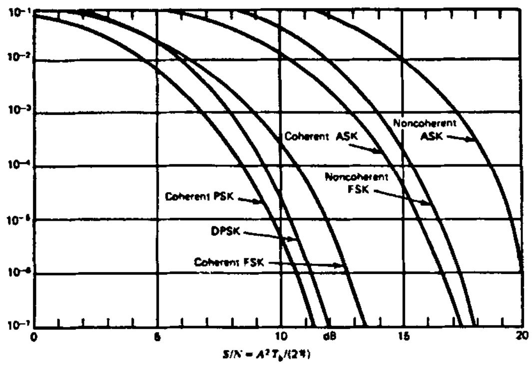
Various techniques exist for digital data communication on a cable medium. They are:

1. Coherent PSK
2. Noncoherent differential PSK
3. Coherent FSK
4. Noncoherent FSK
5. Coherent ASK
6. Noncoherent ASK

Each of these techniques has advantages and disadvantages. In choosing which one to use, they must all be weighed against the particular requirements of a two-way cable system. Parameters such as signal power level, signal bandwidth, noise performance, and cost will determine which technique best fits the need of the cable environment.

These six data communication techniques can be broken down into three categories and rated according to the requirements of a two-way cable system. Fig 3 (obtained from Ref 2) illustrates the comparison of these techniques. From this graph and the corresponding table, one can quickly draw some reasonable conclusions.

(a)



(b)

DIGITAL MODULATION METHOD	COST TRANSMITTER	COST RECEIVER	BANDWIDTH	S/N PERFORMANCE FOR 10E-4 BIT ERROR RATE
COHERENT PSK	LOW	HIGH	Fbit	8.45 db
NONCOHERENT DPSK	MED	MED	Fbit	9.30 db
COHERENT FSK	MED	HIGH	2*dF + Fbit	10.6 db
NONCOHERENT FSK	LOW	LOW	2*dF + Fbit	15.3 db
COHERENT ASK	LOW	HIGH	Fbit	14.5 db
NONCOHERENT ASK	LOW	MED	Fbit	18.3 db

Fig 3 (a) Probability of error curves for binary digital modulation schemes. (Note that the average signal power for ASK schemes is (A*A)/4 whereas it is (A*A)/2 for the others). (b) Table comparison of binary digital modulation schemes. (2*df=F2-F1 is the frequency shift and Fbit is the baseband data bit rate). Obtained from Ref 2.

First, coherent methods of detection allow for better noise performance at the expense of higher cost in the receiver. However, in two-way cable applications, the receiver is in the headend and cost is not as critical as it is for the transmitter in the home terminal units. Therefore, in terms of noise performance, coherent PSK has a great advantage over the other techniques.

Secondly, the bandwidth requirement of FSK techniques is larger, due to nonlinear modulation, than PSK or ASK techniques that employ linear modulation. Not only does this take up more of the spectrum, but also allows for more ingress to fall within the signal bandwidth and degrade data throughput. Also, the phase linearity of the bandwidth-determining filter is more critical in FSK. Another aspect is that impulse noise, which is a prime source of ingress in a cable system, has an adverse affect on FSK systems since they deal with the derivative of the phase function rather than with the phase function directly, as in PSK. Therefore, the cost savings of noncoherent FSK are cancelled by its larger bandwidth requirements and poor performance in both white and impulse noise.

Thirdly, in ASK techniques, signal level variation requires either the data decision levels to be varied or automatic gain control to be employed. This adds to the cost and complexity of the system and still gives relatively poor noise performance. In two-way cable systems, not only does the gain of the reverse plant vary with time, but the signal level from all the home terminal units will not be identical. On the other hand, PSK and FSK techniques allow for signal limiting since the data is contained in the carrier phase rather than in the amplitude. Consequently, gain or level variation is inconsequential for these techniques.

Therefore, it can be concluded from the previous three points that the best technique for two-way cable data communication is coherent PSK since it offers the lowest bandwidth and the best noise performance. The cost of the home terminal unit's transmitter is low and the additional cost that is in the headend receiver is not prohibitive. Also, the superior performance in noisy cable environments allows for less effort and cost in the return plant's maintainance.

NEW PSK DATA COMMUNICATION SYSTEM

The two-way cable communication system used in the field tests employed binary coherent PSK techniques. The home terminal unit consisted of an addressable converter/decoder for reception of the forward plant's video and control signals and a PSK transmitter for transmission of digital information on the reverse plant. At the cable headend, a coherent PSK receiver was connected to the reverse plant's cable, which often consisted of multiple trunk lines. A computer was connected to the headend receiver for collection and statistical analysis of the received digital data. The computer also controlled the forward plant's data channel and video encoders. The cable systems themselves were typical two-way cable designs with trunk, bridger, and line extender amplifiers. A block diagram of the cable system test is shown in Fig 4.

The home terminal unit's transmitter is shown in Fig 5a. It can be seen that it is quite simple and straight forward, consisting of a crystal oscillator and a multiplier. This meets the requirement of low cost for the high volume home terminal unit. Another advantage of the PSK transmitter is that its linear modulation allows the signal bandwidth to be controlled by baseband filtering rather than by RF filtering.

The headend's receiver is shown in Fig 5b. The receiver is more complex and sophisticated since it must employ coherent detection of the PSK signal. This means that the carrier must be regenerated from the incoming signal while it is being synchronously detected. In addition to amplification and narrow band filtering on the front end, the data packet must be detected and recognized in order to determine the beginning of the data as well as remove the phase ambiguity that exists with all PSK systems. Packet detection is accomplished through digital correlation thus allowing accuracy and good noise performance.

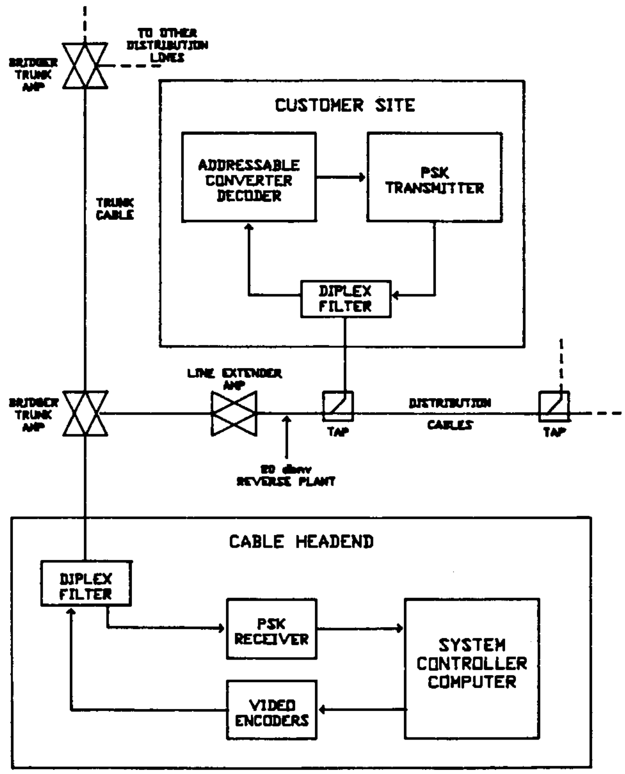


Fig 4 Block diagram of a PSK data communication system connected in a typical cable plant.

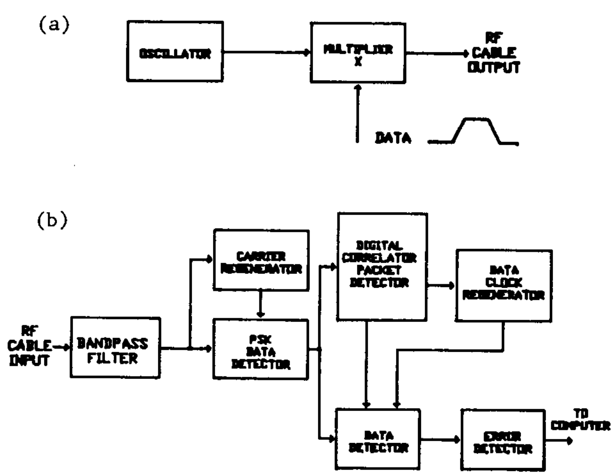


Fig 5 PSK data communication system used in field test.

Error protection is employed through the use of a cyclic redundant code (CRC) which flags all data packet messages that contain errors. No error correction is attempted since two-way cable communication systems are designed around repeating messages when errors occur. Finally, a computer is used to analyze the data, providing the statistical information needed to characterize the data communication system.

Additional parameters determine the communication system's performance. The carrier frequencies used were selected carefully by avoiding the known ingress frequencies and searching for "gaps" in the ingress spectrum. Two carrier frequencies were used in the field test; 5.5 Mhz and 11.0 Mhz. These two frequencies fit well into the ingress spectrum and allow optimum performance even when a minimally maintained reverse plant has severe ingress problems.

The data bit rate was selected to be 45 Kbits/sec. Since PSK is a linear modulation technique, the bandwidth of the channel is equal to the bit rate. This determined the bandwidth required in the reverse plant's spectrum and in the headend receiver. A double-sided bandwidth of 60 Khz was selected for the headend receiver. Regarding the bit rate, the trade-off was between faster throughput and lower bandwidth. Lower bandwidth not only has the advantage of conserving frequency spectrum but also of minimizing the chances of ingress falling within the data signal bandwidth.

Signal levels on the reverse plant are of importance to the cable operator. When video sources are being transmitted along with the data signals, then data carriers should be small in order to avoid "crossmod" problems due to non-linearities in the return amplifiers. However, if data carriers are too small, their performance in heavy noise or ingress will be degraded. These field tests were typically run with data carriers between 20-22 dbmv, which is 20 db below any video carriers on the reverse plant. It should be noted that when no return video services are offered in a particular cable system, these levels can be run higher since "crossmod" products 56 db down are not required for data communication as in video transmission.

FIELD TEST RESULTS

The two-way cable data communication system described above was taken to four different cable sites for system testing. Data signal levels from the home terminal unit were adjusted so that levels into a line extender on the reverse cable plant were 20 dbmv. Data packets were continually sent at a predetermined rate in order for the computer to calculate the statistical throughput of the system. Various conditions were simulated during the tests. In particular, the level of the data carriers was lowered in various increments until the overall system throughput was reduced to 90% (i.e. 90% of the data packets were received without error). Operation of the PSK communication system with this reduced data carrier level then provided a performance "margin" for the cable system based upon actual system noise/ingress levels.

Test site #1 had about 15,000 subscribers on the two-way plant, which consisted of three trunks. An FSK security system was already in service.

Fig 6 shows the entire reverse plant spectrum, while Fig 7 shows an expanded view of the two data channels. Notice that the C/N around the two carriers is about 50 db while the C/N at the high end of the reverse plant's spectrum is 60 db. The presence of impulse noise can be seen in both data channels with greater amounts in the 5.5 Mhz channel. The overall noise/ingress is small since the average of the ingress peaks are greater than 30 db below nominal data carrier levels. This is an indication of a well-maintained system which is necessary when FSK data carriers are present or when video signals need to be transmitted back to the headend.

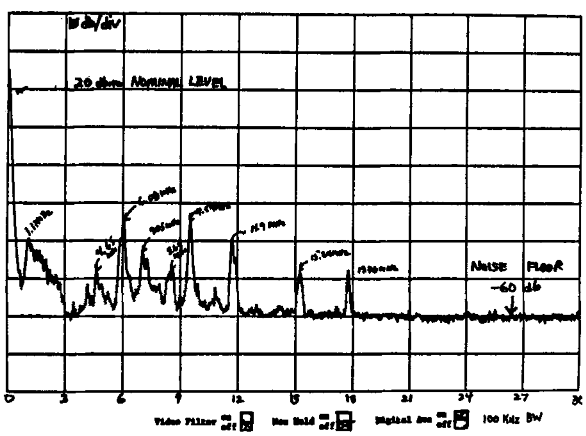


Fig 6 Reverse plant spectrum at Site #1.

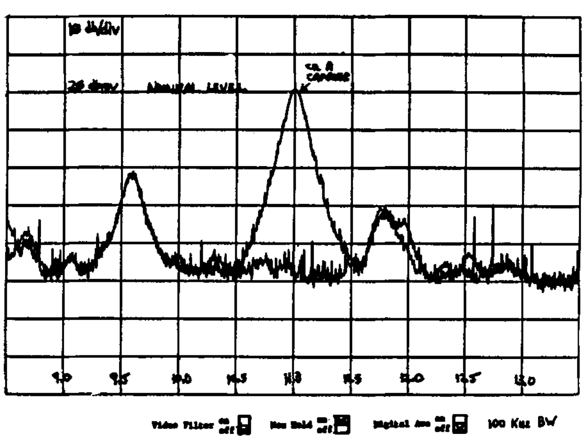


Fig 7 (a) Expanded view of 11 Mhz channel of Site #1.

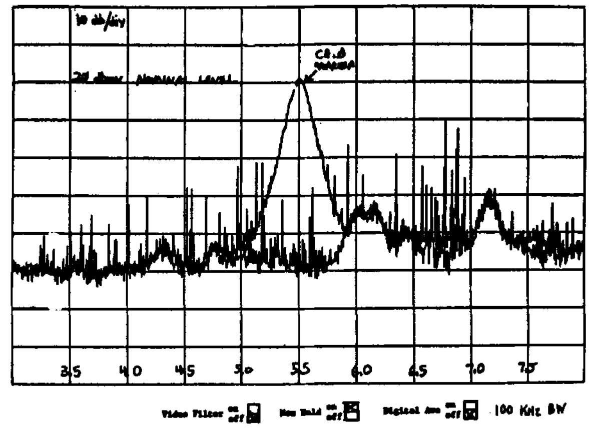


Fig 7 (b) Expanded view of 5.5 Mhz channel of Site #1.

The ingress that is present is the usual AM broadcast, amateur, and international shortwave bands. Also, impulse noise was present. The throughput under these conditions was 100%. Under these plant conditions, the data carriers were then lowered to 30 db below nominal level and Fig 8 illustrates that the throughput was over 97%.

Ch	Sent	Rec'd	Bad	Good	Rec'd /Sent	Good /Sent	Good/ Rec'd	Signal	Noise	Car+Noise /Noise
				6:38:59 pm						
A	1000	1000	0	1000	100.000	100.000	100.000	201.38	1.02	45.91
B	1000	1000	0	1000	100.000	100.000	100.000	202.09	1.05	45.68
				6:40:15 pm						
A	1000	1003	0	1003	100.300	100.300	100.000	201.27	1.03	45.81
B	1000	1002	0	1002	100.200	100.200	100.000	201.97	1.05	45.72
TOTALS										
A	2400	2343	0	2343	97.625	97.625	100.000	199.87	1.49	42.54
B	2400	2343	0	2343	97.625	97.625	100.000	200.39	1.50	42.50

Fig 8 Throughput results at Site #1 with carrier levels 30 db below nominal.

An overnight run with the data carriers reduced to 46 db below their nominal levels was also performed. Fig 9, which displays the reverse plant's spectrum, indicates that the impulse noise has increased substantially, but more so at the lower frequencies. The results, shown in Fig 10, indicate that 94% of the total data packets were received without error. However, the 5.5 Mhz channel throughput was reduced to 88.5% while the 11 Mhz channel was still greater than 99%. This difference is due to the greater impulse noise in the 5.5 Mhz data channel.

After further testing, it was determined that a throughput of 90% could be maintained with the data carriers reduced by 47 db. Therefore, the "margin" of this particular cable system was approximately 47 db.

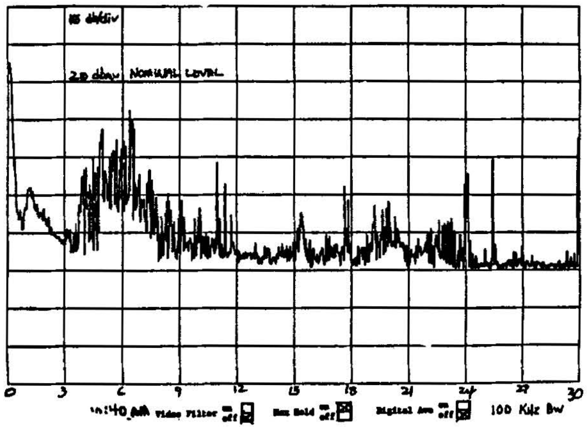


Fig 9 Reverse plant spectrum, with impulse noise, at Site #1.

Ch	Sent	Rec'd	Bad	Good	Rec'd /Sent	Good /Sent	Good/ Rec'd	Signal	Noise	Car+Noise /Noise
			1:29:40 am							
A	1000	999	1	998	99.900	99.800	99.900	36.35	5.37	16.61
B	1000	989	36	953	98.900	95.300	96.360	38.16	6.72	15.09
			2:28:08 am							
A	1000	1000	5	995	100.000	99.500	99.500	36.34	5.54	16.34
B	1000	982	59	923	98.200	92.300	93.992	38.72	7.30	14.49
			3:27:52 am							
A	1000	997	6	991	99.700	99.100	99.398	35.98	5.65	16.07
B	1000	984	37	947	98.400	94.700	96.240	38.18	6.70	15.12
			4:28:51 am							
A	1000	999	1	998	99.900	99.800	99.900	35.83	5.21	16.75
B	1000	987	26	961	98.700	96.100	97.366	38.09	6.33	15.59
			5:28:35 am							
A	1000	999	1	998	99.900	99.800	99.900	35.66	5.33	16.51
B	1000	983	61	922	98.300	92.200	93.794	37.78	6.77	14.94
			6:30:47 am							
A	1000	999	1	998	99.900	99.800	99.900	35.60	5.11	16.86
B	1000	986	39	947	98.600	94.700	96.045	37.82	6.89	14.78
			7:29:17 am							
A	1000	999	1	998	99.900	99.800	99.900	35.83	5.96	15.57
B	1000	953	143	810	95.300	81.000	84.995	38.15	7.73	13.87
			8:29:01 am							
A	1000	999	1	998	99.900	99.800	99.900	35.75	5.63	16.05
B	1000	944	175	769	94.400	76.900	81.462	38.10	8.48	13.05
			9:29:59 am							
A	1000	999	0	999	99.900	99.900	100.000	35.92	5.65	16.06
B	1000	948	116	832	94.800	83.200	87.764	38.30	8.22	13.36
			10:28:30 am							
A	1000	999	1	998	99.900	99.800	99.900	36.11	5.65	16.11
B	1000	931	146	785	93.100	78.500	84.318	38.31	7.98	13.62
			TOTALS							
A	442800	443049	469	442580	100.056	99.950	99.894	35.97	5.60	16.15
B	442800	429147	37188	391959	96.917	88.518	91.334	38.14	7.33	14.33

Fig 10 Throughput results at Site #1 with carrier levels 46 db below nominal.

Test site #2 had approximately 15,000 subscribers on the two-way plant. Four trunks, carrying all the plant's signals, were combined together in the headend. Fig 11 illustrates the two data carriers, some security data carriers (24 Mhz) and the reverse plant's noise/ingress. The ingress can be seen to come up to within 20 db of the nominal carrier levels.

The throughput under these conditions was 100% since the carrier-to-noise (C/N) in the 60 Khz bandwidth was better than 35 db despite the heavy ingress. Fig 12 illustrates this relatively good C/N at 5.5 Mhz and 11 Mhz while the common mode distortion at 6 Mhz and 12 Mhz is noticeably large. Fig 13 shows the reverse plant with the data carriers reduced by 20 db. However, the data communication system still had an average throughput of 97% at this reduced level.

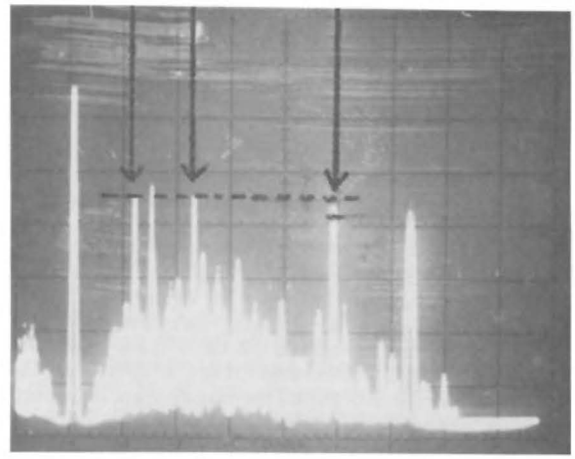
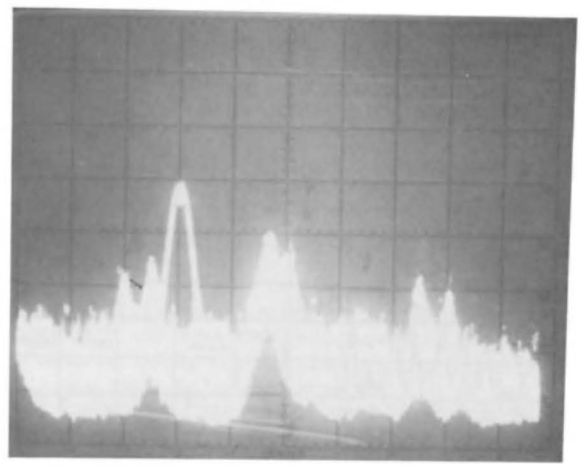
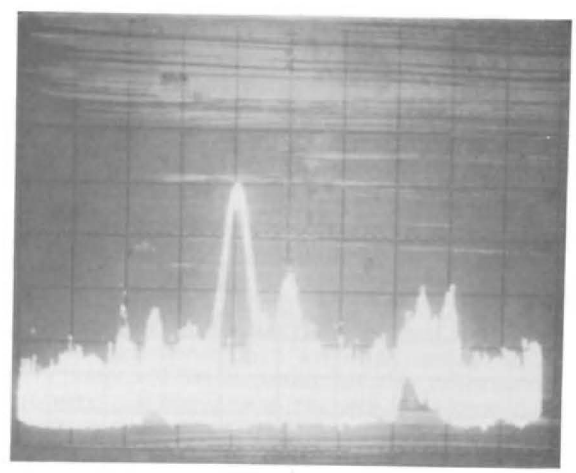


Fig 11 Reverse plant spectrum at Site #2.



(a)



(b)

Fig 12 (a) Expanded view of 11 Mhz channel at Site #2. (b) Expanded view of 5.5 Mhz channel at Site #2.

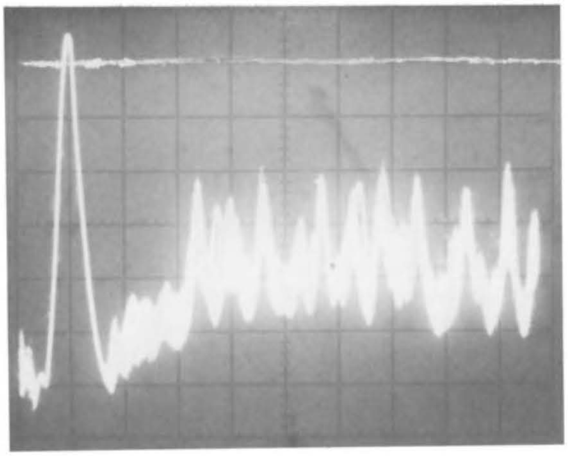


Fig 13 Reverse plant spectrum at Site #2 with carrier levels 20 db below nominal.

It was further determined that a throughput of 90% could be maintained with data carriers reduced by 30 db. Therefore, a "margin" of 30 db was given to this system.

Test site #3, a relatively new system, had about 3,000 subscribers on the two-way plant. The reverse plant was balanced and initially had low levels of ingress, 25-30 db below nominal carrier levels. However, some corona impulse noise was noticed on the plant. Under these conditions, it was found that 100% of the data packets were received without error.

An overnight test was performed with the data carriers reduced to 30 db below their nominal levels. Fig 14 illustrates the reverse plant's spectrum for the evening and next morning. Note how the noise/ingress changed greatly during the night (by as much as 30 db) yet the throughput, shown in Fig 15, was 99%, even with carriers reduced by 30 db. Throughput remained constant through the night even though the noise/ingress varied greatly. This is due to the C/N in the two 60 Khz data channels changing very little (less than 2 db).

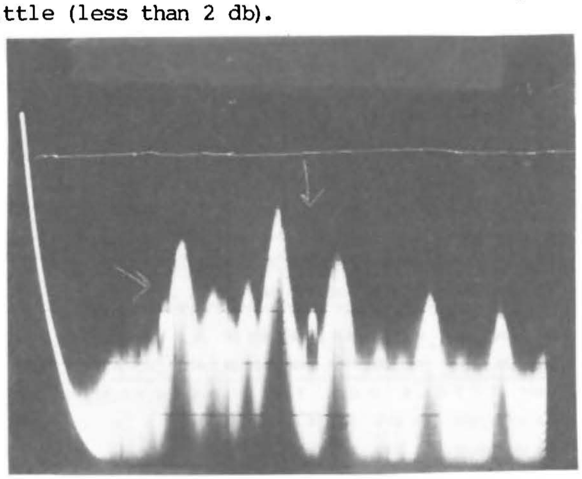


Fig 14 (a) Reverse plant spectrum at night at Site #3.

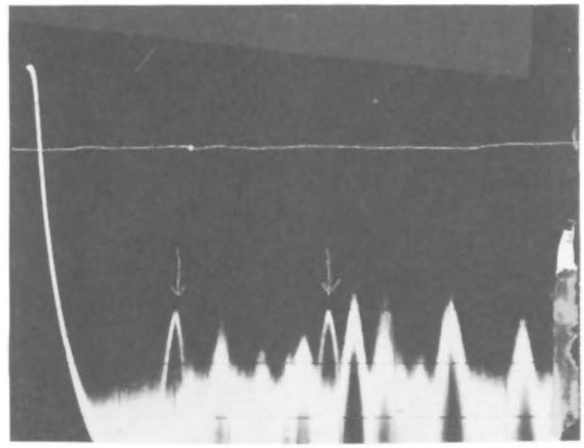
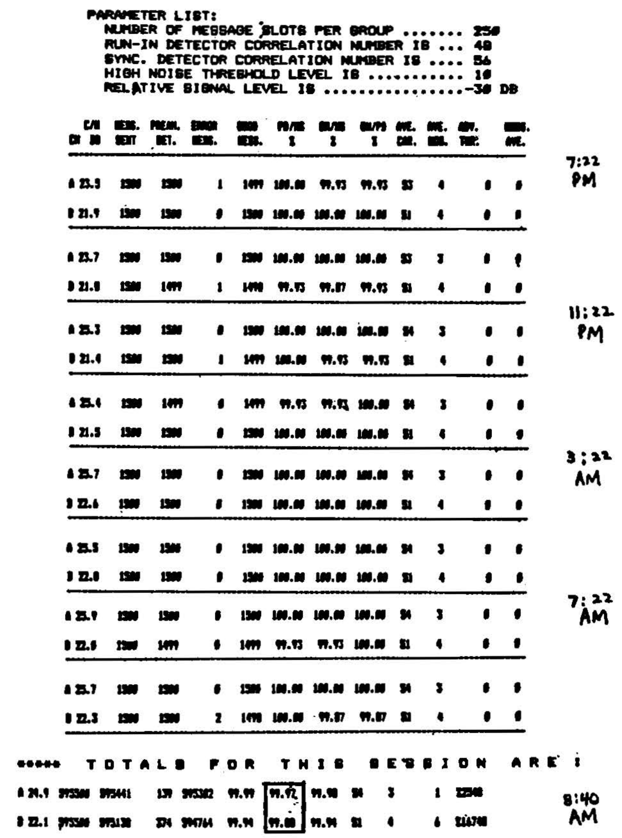


Fig 14 (b) Reverse plant spectrum next day at Site #3.



PARAMETER LIST:
NUMBER OF MESSAGE SLOTS PER GROUP 250
RUN-IN DETECTOR CORRELATION NUMBER IS ... 48
SYNC. DETECTOR CORRELATION NUMBER IS 56
HIGH NOISE THRESHOLD LEVEL IS 10
RELATIVE SIGNAL LEVEL IS ....-30 DB

***** TOTALS FOR THIS SESSION ARE :

Fig 15 Throughput results at Site #3 with carriers 30 db below nominal.

A second overnight run with the data carriers 43 db below nominal produced a throughput of 97%. Further tests indicated that a "margin" of 46 db (for 90% throughput) existed in this two-way cable system.

Test site #4 was about seven years old and had about 30,000 subscribers on its two-way plant with five trunks feeding the headend. The reverse plant had been maintained for 1200 security boxes. The plant had significant ingress as seen in Fig 16 where the peak of the ingress came within 15 db of the nominal data carrier levels. Also, some noticeable CB interference was observed. Under these conditions, the throughput of the PSK data communication system was found to be 100%.

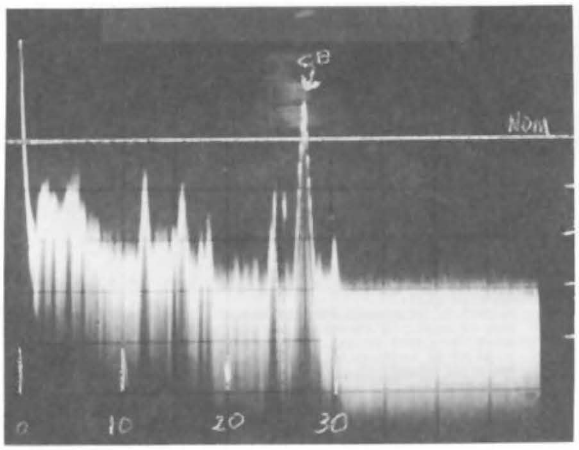


Fig 16 Reverse plant spectrum at Site #4.

A test was performed with the data carriers 26 db below their nominal levels. Fig 17 shows that an average of 74% of the data packets were received without error. However, it should be noted that the 5.5 Mhz channel throughput is significantly less than the 11 Mhz channel throughput. Fig 18 reveals that the reverse plant's spectrum has a "comb" structure at the low frequency end (4-8 Mhz) and that the 5.5 Mhz channel did indeed have a lower C/N ratio than the 11 Mhz channel. It was subsequently found that a trunk amplifier had been oscillating in the form of a pulse regenerative oscillator.

PARAMETER LIST:
NUMBER OF MESSAGE SLOTS PER GROUP 250
RUN-IN DETECTOR CORRELATION NUMBER IS ... 48
SYNC. DETECTOR CORRELATION NUMBER IS 56
HIGH NOISE THRESHOLD LEVEL IS 50
RELATIVE SIGNAL LEVEL IS ....-26 DB

CH	C+N /N	MESS. SENT	PREAM. DET.	ERROR MESS.	GOOD MESS.	PD/MS %	GM/MS %	GM/PD %	AVE. C+N	AVE. NOI.	ADV. THR.	[illegible] AVE.
A	9.3	2300	2413	94	2319	96.52	92.76	96.10	196	66	7	85
B	6.3	2300	1790	570	1300	78.32	55.52	70.89	190	96	0	116
***** TOTALS FOR THIS SESSION ARE :												
A	9.3	2300	2413	94	2319	96.52	$\boxed{92.76}$	96.10	196	66	7	85
B	6.3	2300	1790	570	1300	78.32	$\boxed{55.52}$	70.89	190	96	0	116

Fig 17 Throughput results at Site #4 with carriers 26 db below nominal.

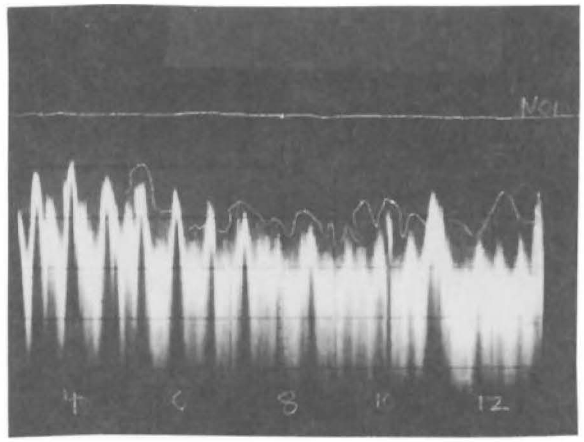


Fig 18 Reverse plant spectrum at Site #4 with carriers 26 db below nominal. Note "comb" characteristics in low frequency spectrum due to trunk amplifier oscillations.

An overnight run with the data carriers 20 db below nominal was performed with results in Fig 19 showing a throughput of about 90%. This system had a significant amount of other ingress and common mode distortion that changed over time yet the C/N within the two 60 Khz data channels changed only slightly thus allowing the data communication system to perform very well.

CH	C+N /N	MESS. SENT	PREAM. DET.	ERROR MESS.	GOOD MESS.	PD/MS %	GM/MS %	GM/PD %	AVE. C+N	AVE. NOI.	ADV. THR.	[illegible] AVE.	
A	14.6	2300	2300	29	2271	100.00	98.04	98.04	140	26	0	0	7:05 PM
B	12.9	2300	2478	127	2351	99.12	94.04	94.87	176	40	0	0	
A	13.6	2300	2493	34	2459	99.72	98.36	98.64	135	38	0	0	
B	11.7	2300	2275	109	2166	91.00	84.24	91.76	153	40	0	0	8:10 PM
A	12.7	2300	2479	35	2444	99.16	97.76	98.59	116	27	0	0	
B	11.1	2300	2345	202	2143	93.00	85.72	91.39	139	39	0	0	
A	13.6	2300	2400	22	2466	99.52	98.64	99.12	133	30	0	0	
B	11.6	2300	2387	174	2213	95.40	85.52	92.71	140	39	0	0	9:12 PM
A	14.9	2300	2381	17	2404	100.04	99.36	99.32	141	25	4	0	
B	11.3	2300	2411	119	2292	96.44	91.60	98.06	139	43	0	0	
A	12.6	2300	2447	49	2398	97.80	95.92	98.00	134	32	0	0	
B	11.4	2300	2410	127	2283	96.40	91.32	94.73	140	40	0	0	10:17 PM
A	17.2	2300	2387	19	2408	100.28	99.52	99.24	144	30	0	0	
B	11.0	2300	2440	36	2404	97.60	96.16	98.52	175	45	3	12	
A	14.5	2300	2404	17	2487	100.16	99.48	99.32	144	22	0	0	
B	11.8	2300	2430	32	2406	97.52	96.24	98.69	180	46	2	31	11:15 PM
A	13.3	2300	2451	69	2382	98.04	95.30	97.18	136	33	0	0	
B	10.4	2300	2322	172	2150	92.88	86.00	92.59	135	30	0	0	
A	14.3	2300	2497	36	2471	99.88	98.84	98.96	119	22	0	0	
B	10.3	2300	2359	74	2285	94.36	91.40	96.86	130	41	0	0	
***** TOTALS FOR THIS SESSION ARE :													
A	14.0	197305	195350	3394	191956	98.91	$\boxed{97.19}$	98.26	139	26	0	0	
B	11.2	197500	184787	14471	170836	92.52	$\boxed{86.09}$	92.06	146	40	37	84	

Fig 19 Throughput results at Site #4 with carrier levels 20 db below nominal.

After the oscillating trunk amplifier was replaced, a 4 db improvement in system performance was observed. The "margin" for this test site was then determined to be 24 db.

CONCLUSIONS

The results of all the field tests were analyzed and compiled. Two basic conclusions can be made.

First, the field tests demonstrated that the performance of the PSK data communication system worked well in the field and correlated very well against the white noise tests performed in the lab. Fig 20 is a graph of the system performance (throughput) at the four test sites over a variety of C/N ratios. It can be seen that there is a 1-2 db degradation in the throughput at the test sites compared to the white noise test. The reason for this is that there is impulse noise in the field that has large peak power that can contaminate data but a low average power that does not show up in the C/N measurements. Therefore, the curve is shifted to the left of the white noise curve. This also explains why the throughput of the 5.5 Mhz channel at Site #1 is noticeably lower; the signal 46 db below nominal was extremely small compared to the impulse noise present. Despite the existance of impulse noise, outstanding performance is still obtained.

Second, there is plenty of safety "margin" in two-way cable plants for operation of PSK two-way data communication systems. The chart in Fig 21 reveals the various "margins" in the four test sites described above. Factors that affect the margin are the amount of maintainance provided, the number of subscriber taps, the condition that the system is in, how well it was originally built, and the proximity of the ingress signals. Even in those systems where noise/ingress is severe, a 24 db margin is obtainable. Often, the margin is much greater.

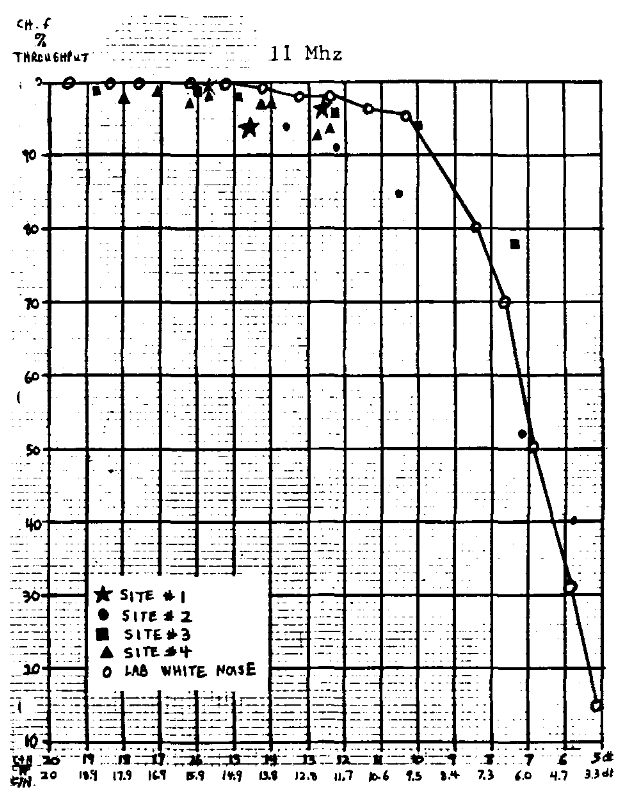


Fig 20 (a) Graph of 11 Mhz channel overall throughput versus C/N for all four test sites and white noise lab test.

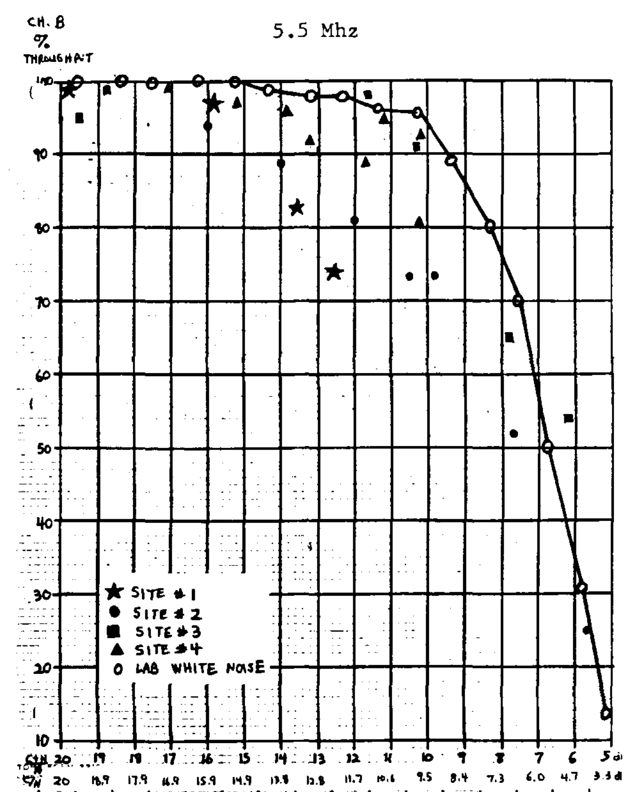


Fig 20 (b) Graph of 5.5 Mhz channel overall throughput versus C/N for all four test sites and white noise lab test.

By judicious selection of the PSK data communication parameters and carrier frequencies, large "margins" of operation are obtainable, even in cable plants with minimal two-way maintainance and significant noise/ingress.

This system has proven, in its largest installation to date, to be rugged and reliable. In three years of continuous operation in a 4500 mile two-way cable plant, the system has been shown to be technically and economically successful.

REFERENCES

1. Citta, R. and Mutzabaugh, D., "TWO-WAY CABLE PLANT CHARACTERISTICS", NCTA, 33RD Annual Convention-Technical Papers, June 3-6,1984.

2 Shanmugam, K. Sam, "DIGITAL AND ANALOG COMMUNICATION SYSTEMS", John Wiley & Sons, Inc., 1979.

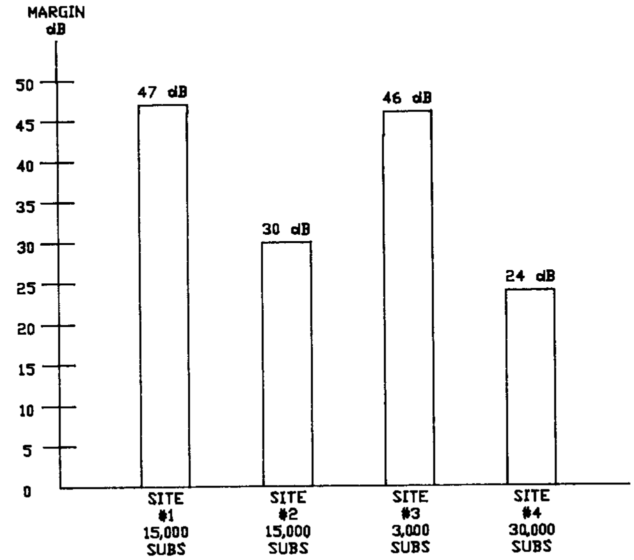


Fig 21 Chart of four field test site margins.

PRACTICAL CONSIDERATIONS FOR BTSC STEREO IN THE CATV PLANT

David A. Sedacca

SCIENTIFIC-ATLANTA

ABSTRACT

Presented is a summary of significant experience with BTSC stereo, with elaboration on areas where operators must make decisions critical to system performance. The acquisition of audio source material in the headend is possible by several means. Formats include various satellite subcarrier systems, scrambled satellite transmissions, and locally supplied audio feeds. Operators often need automatic switching for redundancy or commercial message insertion. Several methods are available for the transportation of stereo from the earth station to the headend. Some cable operators use FM links or AM links to connect various distribution systems, which may pose special problems when stereo is used. Audio input levels to the encoder must be set. Apparent loudness, peak deviation, and average levels are explained. The role of audio signal processing is discussed. Manufacturers provide multiple options for interfacing the stereo encoder with the headend modulator. Sources of error and performance degradation are cited, with techniques of minimizing them. CATV scrambling systems can affect separation and noise performance. Subscriber equipment can cause the most degradation to the BTSC stereo signal. Performance characteristics for various consumer decoders are described. There are several useful techniques for system evaluation. Complete checks require interruption of service, though basic operation can be verified while on line.

I. INTRODUCTION

BTSC stereo is appearing on more and more cable systems. The question of "when to go stereo" is giving way to the question of how to best make it work.

This paper is a distillation of experience gained in the first year of BTSC stereo on cable. The discussion is divided into eleven topics, ordered to roughly correspond to the signal path through the plant. Extra focus is placed on the characteristics of audio systems because audio is a new concern for many cable operators. Familiarity with the headend is assumed in the discussions. Grateful acknowledgement is made to the dozens of cable operators and technicians who have installed stereo in their systems and brought to light many of the problems and solutions described here.

II. BRINGING LEFT- AND RIGHT- CHANNEL AUDIO INTO THE HEADEND

Most of the services with which cable operators use BTSC stereo are satellite delivered.

If satellite transmissions are scrambled with the Videocipher system, audio information is sent in an encrypted digital format. The headend descrambler provides three audio outputs. These are "Left," "Right," and "Mono." [1] See Figure 1.

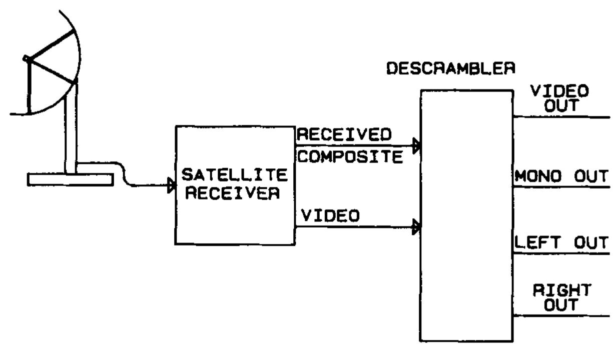


FIGURE 1: RECOVERING AUDIO FROM A SCRAMBLED SATELLITE TRANSMISSION

If satellite transmissions are in the clear, program audio is found on subcarriers transmitted with the video. These subcarriers are located between 5.0 MHz and 8.5 MHz. Usually two or three video related subcarriers are sent with each satellite delivered service. Each subcarrier is frequency modulated with one audio channel. See Figure 2.

The satellite programmer may provide "left," "right," and "mono" on three subcarriers. This is called the "discrete" channel format. When the discrete format is used, programmers often use special noise reduction with the audio signals for the left and right channels. The subcarrier can then be frequency modulated with less deviation, and bandwidth is conserved. The subcarrier demodulator at the headend has signal processing to complement the noise reduction system. See Figure 3a.

Instead of using the discrete format, programmers may use the "matrixed" format. One subcarrier sends "mono" and another subcarrier sends a "difference" channel. With the matrixed format, the left and right channels are recovered by properly combining the "mono" and "difference" signals. The "de-matrixing" circuit is usually located with the subcarrier demodulator circuits. Many receivers contain all this circuitry and provide left and right channel audio signals at their output. Most matrixed systems do not use any noise reduction other than a standard pre-emphasis/de-emphasis. See Figure 3b.

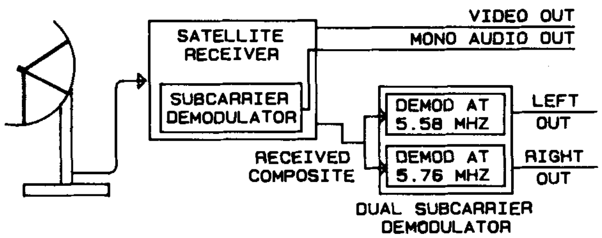


FIGURE 2: RECOVERING AUDIO FROM A SATELLITE CHANNEL WITH AURAL SUBCARRIERS

BTSC stereo may be used with locally originated program material from public access facilities or broadcast studios. A pair of FM links or leased phone lines bring audio into the headend. See Figure 4. Highest fidelity results when the left and right channels travel over identical paths. If their paths are not the same, fidelity on mono receivers may suffer, and stereo imaging may be affected.

III. CARRYING LOCAL BROADCASTERS IN BTSC STEREO

Local broadcasters can be put on the cable system in several ways. A signal processor can receive an off-air signal and place it on the cable system. The processor provides control of aural carrier level and rejects undesired adjacent signals. Many cable systems use signai processors to successfully carry local stereo broadcasts. See Figure 5. However, not all processors perform equally well. The path through which the aural carrier travels is critical. This path must be sufficiently wide to pass the sidebands that BTSC stereo creates. The amplitude and delay responses must be reasonably flat. If the aural carrier path through the processor is not wide enough, a BTSC signal will suffer increased distortion. If the path is not symmetric around the carrier frequency, or amplitude response is not flat, there may be excessive FM to AM conversion. AM on the sound carrier is recognized by some set top terminals as a cue to descramble. When a

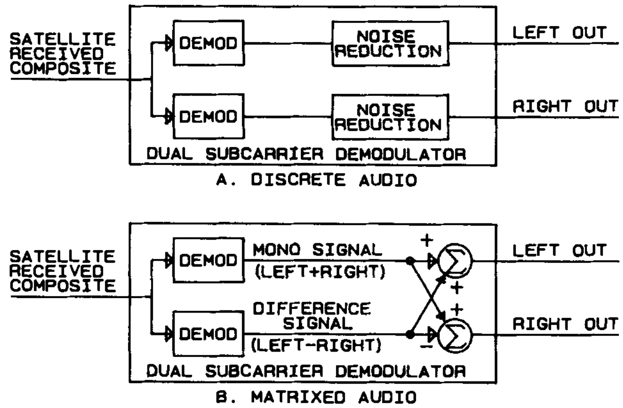


FIGURE 3: "MATRIXED" AND "DISCRETE" AURAL SUBCARRIER DEMODULATION

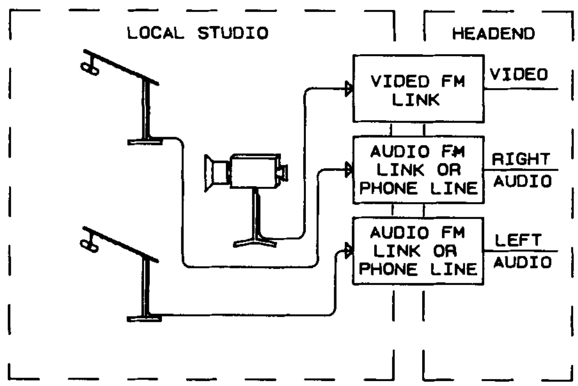


FIGURE 4: LOCAL STUDIO FEED

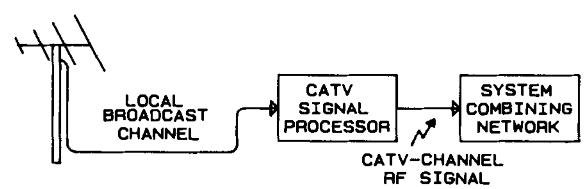


FIGURE 5: USING A SIGNAL PROCESSOR FOR A LOCAL STATION

local broadcaster uses the BTSC Second Audio Program (SAP) service or Pro channel, the carrier receives even wider deviation, and FM to AM conversion becomes more critical. Most processors provide adequate performance. Consult the manufacturer if there is any doubt.

Some cable operators place local broadcasters on their systems after demodulating the off-air signal. Baseband video and audio from the demodulator are routed through patch bays or switchers for control purposes. A modulator then places the signal on the system.

See Figure 6. Using such a demod-remod process is not recommended if the carriage of BTSC stereo is desired. The demodulator and modulator may not have sufficient audio frequency response to pass the entire BTSC signal. Typically the BTSC pilot tone will make it through, but the difference channel information that is required for true stereo performance is lost. Even if the demodulator and modulator can pass the entire BTSC baseband composite signal, stereo separation will severely suffer if deviation sensitivity of the modulator does not perfectly match the output level of the demodulator.

The demod-remod process will pass BTSC stereo if the 4.5 MHz aural carrier output from the demodulator is used instead of the baseband audio output. The aural carrier is sent to the modulator without ever being demodulated. This method preserves the critical BTSC parameter of aural carrier deviation setting. See Figure. 7.

For highest fidelity, some broadcasters provide special feeds of video and audio to nearby headends. In this case, Figure 4 again applies. Baseband left- and right-channel audio signals are brought to the headend.

IV. AUDIO LINE IMPEDANCE AND BALANCED VS. UNBALANCED LINES

Various devices in the headend will call for balanced or unbalanced lines, and for high impedance or low impedance loads.

Unbalanced Audio Sources

An audio source is said to be unbalanced if its output is presented on one terminal referenced to ground. The signal voltage on this terminal swings above and below ground.

An unbalanced source may have an "active" output or it may be "transformer-coupled." If a device has an active output, the final amplifier is wired directly to the output terminals. If the device is transformer-coupled, the output amplifier drives the primary side of a transformer. Two leads from the secondary side of the transformer provide the output signal. One of the leads is connected to ground. The other lead then has the signal voltage on it. The signal voltage swings above and below ground. See Figure 8.

Balanced Audio Sources

The output of a balanced audio source appears across two non-grounded terminals. Additionally, a third terminal at ground potential is usually available. The voltage on one terminal swings above and below ground as the signal varies. The voltage on the other terminal swings above and below ground in equal amounts, but in the opposite direction. Thus the voltages on the two terminals are "balanced," as shown in Figure 9.

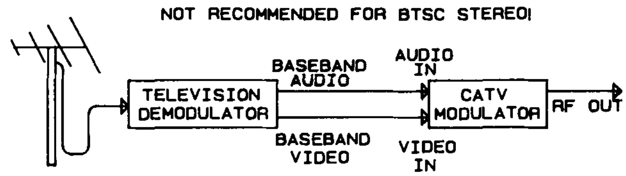


FIGURE 6: DEMOD-REMOD FOR A LOCAL STATION

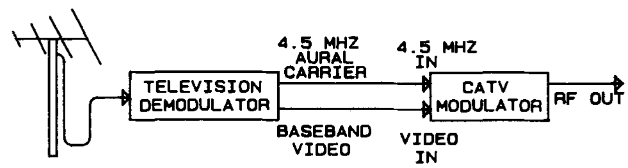


FIGURE 7: DEMOD-REMOD WITH 4.5 MHZ AURAL CARRIER

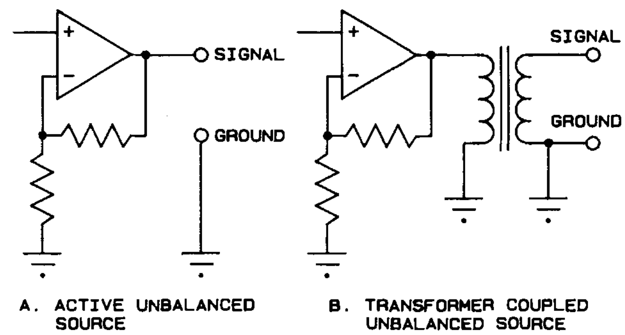


FIGURE 8: UNBALANCED SOURCES

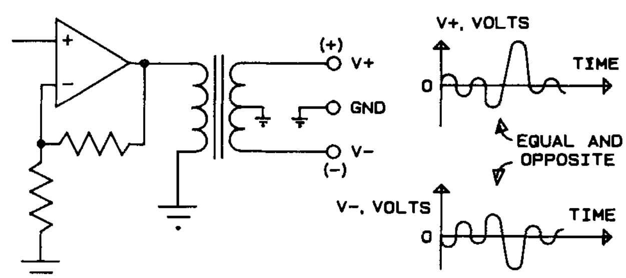


FIGURE 9: TRANSFORMER COUPLED BALANCED SOURCE

Different methods are available to manufacturers for providing balanced outputs. The output can be transformer-coupled, as in Figure 9. The audio output amplifier in the source drives the primary side of a transformer. The secondary side of the transformer has a center tap. The audio signal appears across the end leads. If the center tap is connected to ground, the DC voltage of the secondary is held at zero. The end leads of the transformer now provide a true balanced output: the AC voltages on these terminals with respect to ground are equal and opposite. If the center tap is not connected to anything or there is no center tap, the output winding is said to be "floating" and not balanced.

An "active balanced" output does not use a transformer. Instead, transistor amplifiers or integrated circuit op amps are used to create the output signals directly. See Figure 10. One amplifier drives one output terminal. Another amplifier produces an inverted version of the same signal and drives the other output terminal. A true balanced output again results: the signal terminals contain equal and opposite voltage waveforms.

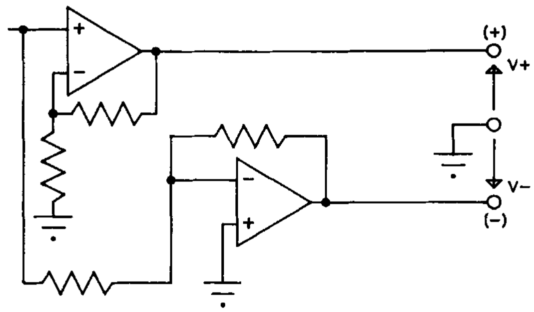


FIGURE 1Ø: ACTIVE BALANCED SOURCE

Balanced sources can usually be identified by the markings near the terminals or by reading the manufacturer's data sheet. If the output terminals are labeled (+) and (-) this is probably a balanced source. Whenever you use a balanced source, be sure that the (+) and (-) output signal lines are never shorted to ground. This may damage the audio source and will disturb operating levels.

If necessary you can usually convert a balanced source to an unbalanced one. For equipment with active balanced outputs, simply connect the signal wire to the (+) output and the ground wire to the source ground. Do not make any connection to the (-) output terminal. On equipment with transformer coupled outputs, disconnect the center tap from ground. Connect the (-) terminal to ground instead. The (+) terminal is now the signal line. See Figure 11.

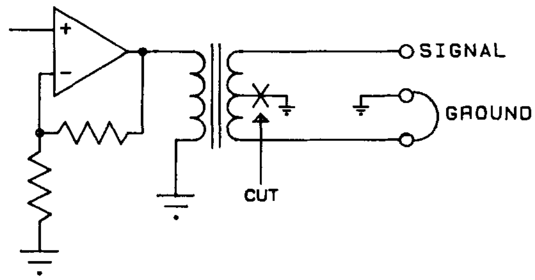


FIGURE 11: CHANGING A TRANSFORMER COUPLED BALANCED SOURCE TO AN UNBALANCED SOURCE

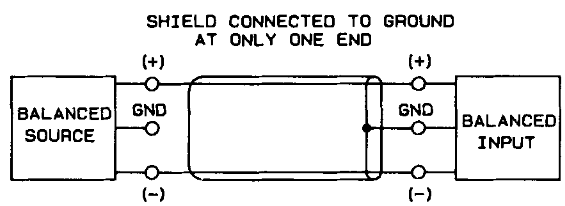
Audio Inputs

When an audio input is driven by a signal line, that input is said to "present a load" to the line. The load may be balanced or unbalanced; high impedance ("bridging") or low impedance.

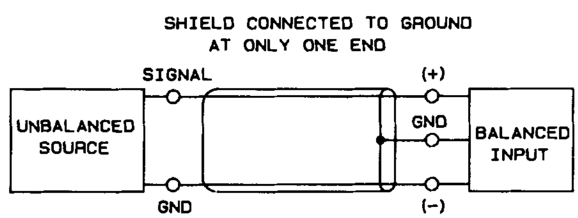
A balanced input has two signal input terminals in addition to a third terminal for ground. The ground terminal is used only to connect shielding on audio cables. The balanced input operates by amplifying the difference between the voltage signals that are applied across the two input terminals. When driven by a balanced source, the desired signal appears as a difference-mode voltage across the input terminals.

Noise may be picked up on the audio wires, often as hum or radio interference. Usually this noise is induced equally on the (+) and (-) wires. Such common-mode signals are rejected by the balanced input.

Balanced inputs may be driven by unbalanced sources. The ground wire from the unbalanced source should be connected to the (-) input terminal. The signal line from the unbalanced source should be connected to the (+) input terminal. Hum and noise rejection may no longer be as effective as with a balanced system. See Figure 12.



A. FROM A BALANCED SOURCE



B. FROM AN UNBALANCED SOURCE

FIGURE 12: CONNECTING TO A BALANCED INPUT

An unbalanced input has one signal input terminal. It is referenced to a ground terminal. The advantage of an unbalanced audio system is that fewer wires must be used for signal routing and patching. The disadvantage is that hum may easily be caused by imperfect grounding of either the audio or AC power system. An unbalanced input should not be driven from a balanced source.

Transformer coupled inputs can be driven from balanced or unbalanced sources. They provide common-mode noise rejection and can break ground loops that cause hum.

The standard source, line, and load impedance for professional audio equipment is 600 ohms. However, many products have input impedances higher than 600 ohms. These "hi-z," or "bridging" inputs are usually 10 kilohms to 50 kilohms. If audio cables are short, several hi-z loads may be driven by one 600 ohm source. It is recommended that the load furthest from the source present a terminating impedance of 600 ohms.

In most headend applications it is permissible to leave audio lines unterminated. But for lines longer than several dozen feet, termination preserves frequency response flatness and reduces induced noise. When lines are terminated, be sure that they are terminated only once. Depending on how low the source impedance is, signal amplitude will be affected by changing the load impedance. The lower the source impedance is, the heavier load (lower load impedance) it may drive. Most sources will not perform well if they must drive more than one 600 ohm load.

Connections to the (+) and (-) terminals on sources and loads should be consistent. If the (+) and (-) connections are reversed somewhere, the left and right channels will end up out of phase with each other at the receiver. Monaural receivers will have no sound output. Mono compatibility is assured when the left and right channels are wired in phase with each other.

V. SIGNAL SWITCHING FOR LOCAL MESSAGE INSERTION

Figure 13 shows one system that detects cue tones and uses a video tape player to insert messages. The tone detector listens for control signals provided by the programmer. These control signals precede pauses into which local messages are inserted. On some services a separate subcarrier is transmitted over the satellite to carry the control signal. On others, the "mono" program audio channel has control tones.

When tones are detected, the switcher selects video from the tape deck and sends it to the modulator. At the same time, a relay contact closure commands the stereo encoder to use its alternate audio inputs. When the taped message is over, program video is switched back in and the encoder is commanded to accept "main" audio.

VI. TRANSPORTING STEREO FROM A REMOTE EARTH STATION TO THE HEADEND

Separate FM links can be used to deliver video and audio to the headend. The satellite receiving system contains all descrambling or demodulating circuits to produce baseband video, left audio, and right audio outputs. Coaxial cable connects the earth station to the headend. An FM modulator uses a 14 MHz wide channel to transmit the video signal to the headend. Two separate FM modulators use channels about 200 kHz wide to transmit the left and right audio signals. [2] Three FM receivers in the headend provide baseband video, left, and right audio outputs. The stereo encoder and modulator create the BTSC television signal. See Figure 14.

Figure 15 shows a method that is not recommended. A BTSC stereo encoder is located at the earth station. Its output is a modulated

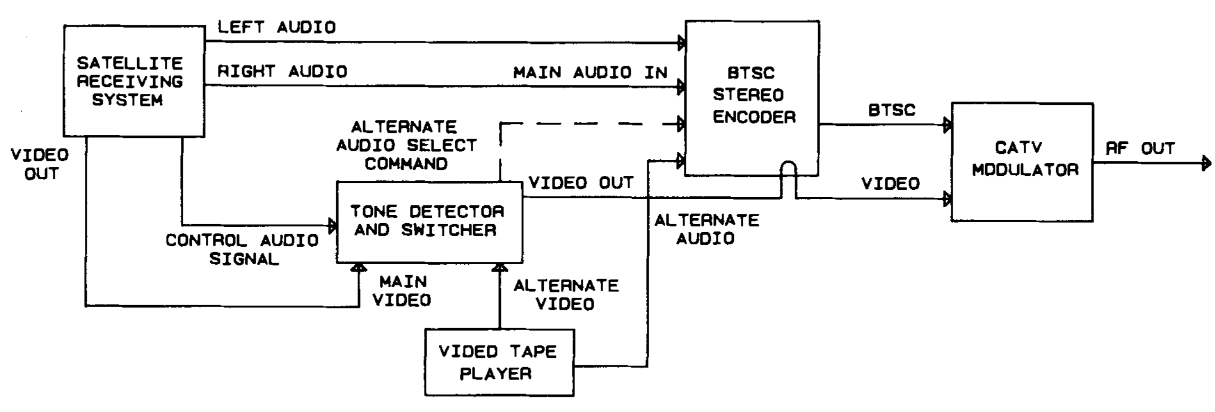


FIGURE 13: SIGNAL SWITCHING FOR LOCAL MESSAGE INSERTION

4.5 MHz aural subcarrier. Baseband video combined with the 4.5 MHz aural carrier is sent over a video FM link. This scheme may result in two serious problems: visible intermodulation beats and noisy audio. The visible beat between the aural carrier and color carrier is likely because the FM link's channel is only wide enough for quality transmission of video. When the subcarrier at 4.5 MHz is added, FM sidebands farther away from the carrier frequency become more important. But distant sidebands are attenuated in bandpass filters in the transmitter and receiver. The receiver produces an output that shows intermodulation distortion because important FM sidebands were cut out. [3] The audio is noisier because of the nature of frequency modulation systems: higher modulating frequencies suffer more noise at the receiver. The 4.5 MHz carrier at the output of the video FM receiver is surrounded by much more noise than it was at the input. The BTSC stereo format itself involves frequency modulation, with special noise reduction for its higher modulating frequencies. The degradation of carrier-to-noise ratio that the 4.5 MHz signal suffers in the FM link will combine with other system noise on the way to the subscriber. Intolerable noise performance may result.

Figure 16 illustrates the use of a composite video FM microwave link. Two FM subcarriers are modulated with the left and right audio signals. These subcarriers are combined with the baseband video to create a composite signal. An FM transmitter uses a 22 MHz wide channel in a microwave radio band. A receiving system in the headend produces baseband video, left, and right audio outputs. Because this system uses a wider channel than the simpler video FM link, it is capable of carrying subcarriers above the video with less intermodulation distortion. In addition, these subcarriers employ wider deviation or special noise reduction to make up for degraded carrier-to-noise ratios at the microwave FM receiver output. High quality audio and video transmission is achieved.

VII. HEADEND TO HUB-SITE TRANSPORTATION

When a hub-site is fed over a properly operating AM microwave link, BTSC stereo signals are not adversely affected. See Figure 17. However, in complicated hub-sites the received signal may not be routed directly to the trunk amp. Instead, signal processors or demodulator-modulator pairs may be used to create flexibility in channel assignment, scrambling, and control. The quality of received stereo depends critically on the quality of the path through which the aural carrier travels. Though any single block of the cascade may pass BTSC stereo, the combined effects of several processors, demodulators, and modulators will add up. In a complicated distribution system performance figures for stereo are more easily measured than predicted.

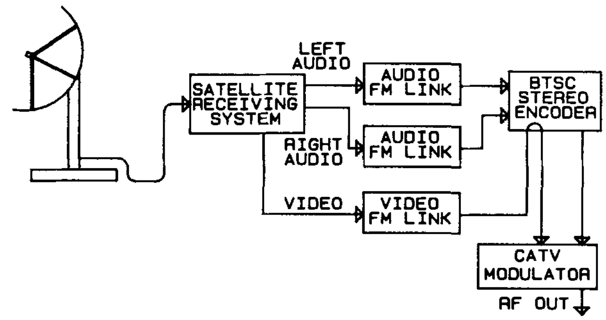


FIGURE 14: EARTH STATION TO HEADEND: TRANSPORTING LEFT, RIGHT, AND VIDEO ON SEPARATE FM LINKS

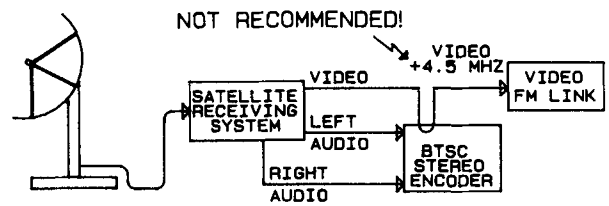


FIGURE 15: EARTH STATION TO HEADEND: TRANSPORTING VIDEO +4.5 MHZ

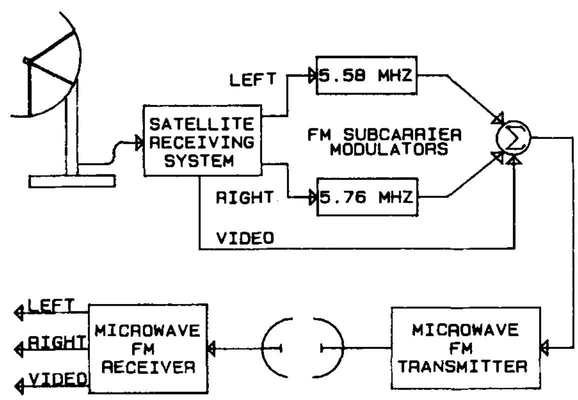


FIGURE 16: EARTH STATION TO HEADEND: COMPOSITE VIDEO ON FM MICROWAVE LINK

Some straightforward rules help preserve stereo separation:

* Do not carry baseband BTSC composite multiplexed signals more than the few feet needed between stereo encoder and television modulator.

* Once the BTSC-modulated aural subcarrier has been created, do not demodulate it to baseband and remodulate it again. Leave it in the form of an FM carrier and let the subscriber's equipment demodulate it.

* Highest fidelity comes from transporting discrete video and left- and right-channel audio, and using a BTSC encoder and television modulator at each hub site.

* Do not add the 4.5 MHz aural subcarrier to an FM link designed to carry video only.

VIII. AUDIO LEVELS

Some Terms Used In Discussions of Audio Processing

"Loudness level" is a subjective property of audio program material. Loudness is not easily measured. Instruments are available that attempt to simulate the response of human hearing and give a numerical measure of loudness. "Ideal" measurements are not achieved because of the complexity of ordinary sounds, and differences among listeners. Human perception of loudness is a non-linear, frequency-dependent phenomenon. Statistical data has been collected to determine how people judge loudness of single tone signals. The "Fletcher-Munson" curves resulted from early studies, and later researchers have produced other similar curves. [4]

"Peak level" describes the maximum excursion of signal voltage above or below zero Volts. Some waveforms will produce positive peaks of different value from the negative peaks. The human voice is one of these. Peak levels are easily measured.

"Average level" describes the average of the magnitude of the signal waveform over some time interval. This may be measured by full-wave rectifying the waveform and averaging it with a low pass filter. In such a circuit the averaging interval depends on the charge and discharge time constants of the storage network. The root-mean-square (rms) value of the signal may be measured to get a different kind of average. In this case the signal is multiplied by itself, which results in a wave that is greater than or equal to zero. The waveform is integrated over a time interval. A logarithmic amplifier may be used to provide an output proportional to a decibel reference. Otherwise a circuit may be used to simulate the square-root function and provide an output linearly proportional to rms level.

The "dynamic range" of a signal or channel is the ratio of the peak signal level that is possible to the lowest acceptable signal above noise. The "signal-to-noise" is the ratio of the present signal power to the noise in the channel. "Headroom" is the difference between the amplitude of the biggest undistorted sine that can be carried by the channel, and the nominal, average, or "operating" signal level.

Loudness control is achieved by various methods. The most common is compression. "Compression" implies gain control too slow to limit peaks, yet fast enough to substantially reduce the amplitude differences between passages with small and large average energies. Broadcast and studio equipment commonly employs frequency selective compression, which allows many aesthetic improvements over ordinary compression. Most important among these is the avoidance of loudness modulation of one frequency band by energy in a different band. Unfortunately it also allows opertors to drastically change the dynamics of recordings. A "loudness controller" contains circuitry to evaluate the spectral content of a signal. Program material is split into various frequency bands, compressed, and possibly clipped. Filters for each band get rid of harmonics caused by clipping. The bands are recombined and sent to the output.

Phase shift networks are used to make assymmetric peaks symmetric. This allows a broadcaster to fit a higher average energy within the same peak levels. It is widely

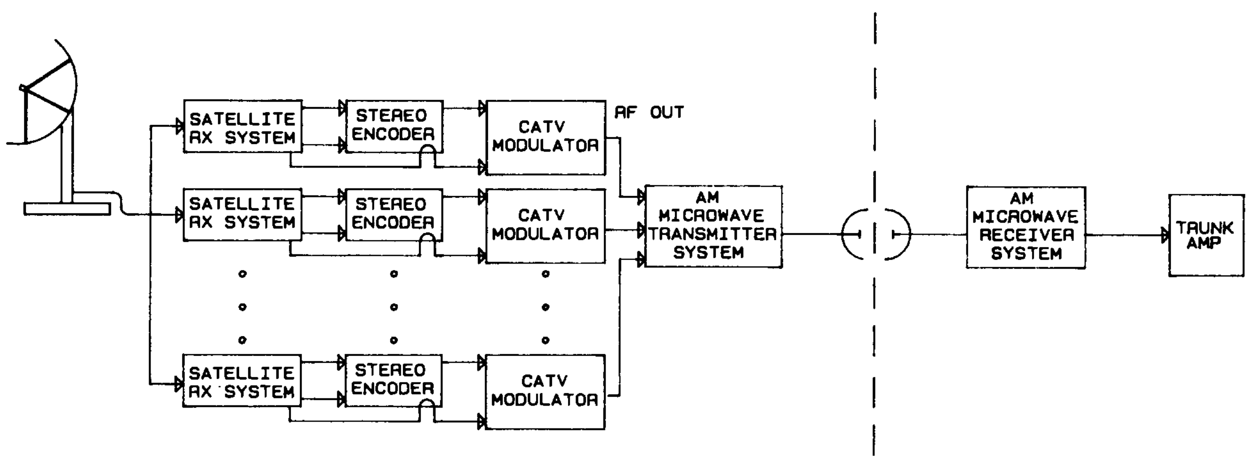


FIGURE 17: HEADEND TO HUB-SITE: SEVERAL PROGRAMS ON AM MICROWAVE LINK

accepted that most phase delays are not audible, as long as the left and right channels are treated equally. Fixed phase shift networks have been used in broadcasting for over 25 years.

Loudness Variations

Loudness levels among television channels vary greatly. Several things contribute to this.

* Some audio program material is "dense" or loud because of the program content: gunfights on horseback; chords played on an electric guitar. Other material is quiet or sparse; soft conversations or background sound effects.

* Modulation levels on some channels are set higher than on others.

* The original recording or transmission levels for programs vary.

* Some channels are compressed and limited somewhere in the transmission path. This may raise their loudness level compared with unprocessed channels without changing peak modulation.

* Some programs are compressed and limited during original production.

Audio Levels and Headend Practice

When adding BTSC stereo to a system, care must be employed when setting the left and right input levels on the stereo encoder. Most satellite-delivered programs are provided with audio that has not been compressed and limited to meet the needs of CATV transmission. Some programs may have quiet soundtracks with few loud passages, and others may have moderate to loud sound all the way through. Either situation can occur even though the peak levels indicated on the stereo encoder's meters appear the same. This is because different program producers use different audio processing when creating their soundtracks. If a program has been compressed and limited, it will have a higher perceived loudness for the same peak voltage levels. FM broadcasters take advantage of this fact: different FM stations have different loudness due to the audio processing they may or may not use, but they all have the same peak deviation of their carriers. Many people regard the over-use of these techniques as a degradation of program fidelity.

With mono program material, cable operators set levels to achieve nominally equal perceived loudness from channel to channel, and to avoid too much flashing of the "over deviation" light on the modulator. This is an effective but very imprecise method. Some stereo encoders have true peak reading meters on the front panel. The tendency is now to set levels to avoid peaks (keep the meters out of the red). This may result in a perceived loudness on stereo channels that is less than the loudness on mono channels. This difference will be heightened on satellite-delivered programming that has not been compressed and limited. If loudness levels are a problem, the operator has several options.

* The left and right input levels on the stereo encoder can be turned up some. This will drive the level meters into the red more often, which will mean that the aural carrier is being deviated more widely than called for in the FCC specs. Broadcasters are not allowed to do this, and usually don't. However, cable operators commonly overdeviate their aural carriers. Remember that excessive level can increase distortion, especially if the stereo encoder has clipping circuits to protect against overdeviation. Excessive overdeviation will cause audible distortion in recievers and visible interference on the TV screen.

* A more expensive option is to install a stereo compressor/limiter before the input to the stereo encoder. This way the stereo programming can have its peak-to-loudness ratio brought more in line with what local broadcasters have. The operator may view this as too much trouble and expense. Setting up most stereo processors is a non-trivial problem that requires the subjective judgement of the individual at the controls.

* Another option is to leave the levels set correctly and send the subscriber sound with its original characteristics. Wider dynamic range results, but the operator may have to deal with complaints about inconsistent loudness levels.

IX. ENCODER-MODULATOR INTERFACE

The stereo encoder may be connected to the modulator in many different ways. Figures 18, 19, and 20 are generalized block diagrams that show most of the configurations available.

Stereo Encoder Inputs and Outputs

Figure 18 shows the inputs and outputs of a "generic" BTSC stereo encoder. The left and right audio signals are supplied by satellite receiving equipment or local origination sources. Program video is looped through the stereo encoder so that the pilot tone may be locked to the horizontal scan frequency.

The stereo encoder provides several different kinds of output. The BTSC multiplexed baseband composite signal consists of the sum channel from 50 Hz to 15 kHz, the pilot tone at

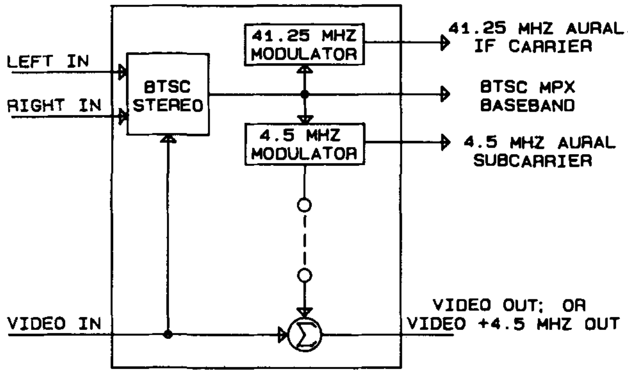


FIGURE 18: BTSC STEREO ENCODER INPUTS AND OUTPUTS

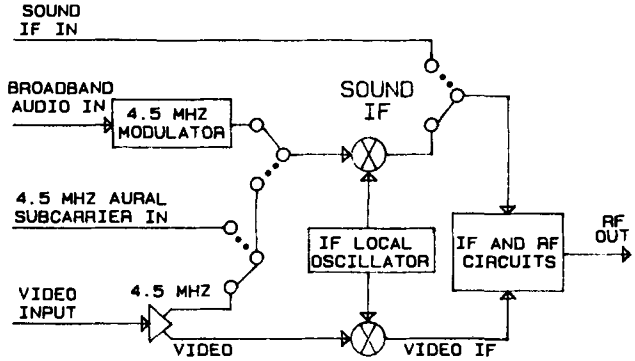


FIGURE 19: TELEVISION MODULATOR INPUTS AND OUTPUTS

15.734 kHz, and the difference channel which has been multiplexed to the 30 kHz-wide band centered at 31.468 kHz. If the Second Audio Program (SAP) channel is used, additional energy is present in a 20 kHz-wide band around 78.670 kHz. [5] A 4.5 MHz aural subcarrier modulator accepts the BTSC composite signal as its input. It frequency modulates a 4.5 MHz carrier and makes it available as an output. Some encoders allow the 4.5 MHz carrier to be combined with the video baseband signal. A 41.25 MHz modulator is also available for some stereo encoders. It produces a frequency modulated carrier at the standard sound IF.

Television Modulator Inputs and Outputs

Figure 19 shows various inputs and outputs for a "generic" CATV modulator. The video input circuit on some modulators has a splitter. This allows the introduction of video only or video plus the modulated 4.5 MHz aural subcarrier. Some modulators have a separate input for a modulated 4.5 MHz aural subcarrier. A simple modification can usually create this input on modulators that do not already have it.

The audio input on older modulators is not capable of accepting the BTSC multiplexed composite signal. Manufacturers offer upgrades and new modulators that do accept this type of input. Modulators configured for traditional monaural broadcasts include a 75 microsecond pre-emphasis circuit that is standard for noise reduction. When a modulator accepts the BTSC baseband input, its pre-emphasis must be defeated. The stereo encoder now performs the pre-emphasis. The modulator must also be able to deviate the sound carrier with frequencies up to 100 kHz.

CATV modulators usually provide some way to access the sound IF and video IF circuitry. For certain modulators, the aural carrier path does not pass BTSC stereo. In these cases the sound IF input may be the only way to successfully inject a BTSC stereo signal.

Connecting Stereo Encoders to Television Modulators

Figure 20 shows four connection schemes for BTSC stereo systems. Scheme 1 uses the BTSC multiplexed composite baseband output of the stereo encoder. The modulator must be able to accept a broadband input at its audio terminals, with no 75 microsecond pre-emphasis engaged in the modulator. This configuration requires that the deviation sensitivity of the modulator be precisely set. If deviation sensitivity is misadjusted, stereo separation degrades. Calibration tones in the stereo encoder and an accurate deviation indicator in the modulator will greatly simplify the set-up process. Specific instructions can be obtained from manufacturers supporting the interface at baseband. When correctly installed, the interface at baseband provides excellent BTSC stereo performance.

In Scheme 2 the modulated 4.5 MHz aural subcarrier comes from the stereo encoder. The television modulator has a separate input to accept the 4.5 MHz carrier. This set-up minimizes the adjustments that the operator must make upon installation. The deviation sensitivity of the 4.5 MHz modulator is calibrated at the factory, not in the field. For some modulators, a modification must be performed to create a separate input for the 4.5 MHz carrier.

Scheme 3 uses the modulated sound carrier at the intermediate frequency of 41.25 MHz. This system is best when using modulators that cannot be modified to accept either baseband input or a modulated 4.5 MHz carrier. Some modulators may be limited in their ability to pass the BTSC signal because of various filters in their signal processing chain. The 41.25 MHz modulator bypasses most of this processing. It is important that the manufacturer of a 41.25 MHz modulator provides sufficient frequency accuracy and low phase noise to maintain a quality signal.

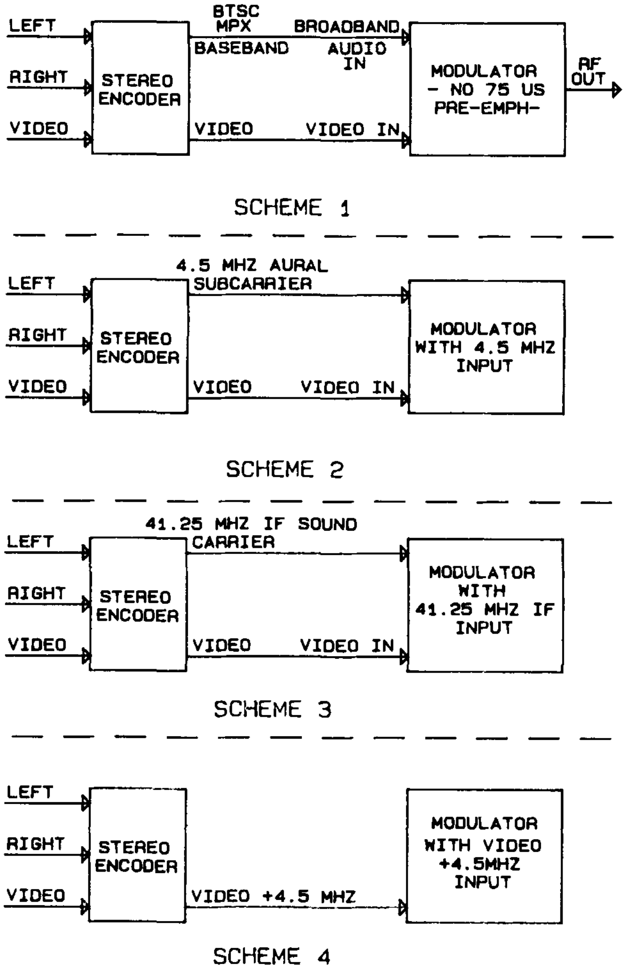


FIGURE 2Ø: CONNECTING BTSC STEREO ENCODERS TO TELEVISION MODULATORS

Scheme 4 sends video plus the modulated 4.5 MHz aural subcarrier over one coaxial cable to the video input of the modulator. This scheme is not optimum. In the modulator, the 4.5 MHz subcarrier is split from the video. Once they are there, removing video artifacts from the aural carrier band is not possible. Therefore it is best to lowpass filter the video to eliminate energy above 4.2 MHz before it is combined with the 4.5 MHz subcarrier. Video artifacts may be amplified by the aural carrier limiter stages of the modulator. The result is increased buzz in the audio. Typically the audio signal-to-noise ratio may suffer a 2 to 3 dB decrease, depending on video content. However, many operators use Scheme 4 to take advantage of existing patch bay systems in their headend. Performance for mono or stereo can be improved by avoiding video overmodulation. Additionally, the 4.5 MHz signal from the stereo encoder to the modulator may be increased above the 0.1 Volt level that is standard. This provides more immunity from video noise. Distortion is not likely to be created unless the modulator's limiter circuits are overloaded.

Because of the many connection combinations possible for BTSC stereo, it is best to consult the manufacturer of the encoder and modulator while determining the choice for a headend. Factors such as cost, switching requirements, routing requirements, redundancy, and service make it impossible to say that one way is best for all cases.

X. STEREO AND SCRAMBLING

CATV scrambling systems affect television audio if the aural carrier is used to transmit descrambling information. Television audio modulates the frequency of the aural carrier. Descrambling information modulates the amplitude of the aural carrier.

RF Scrambling Systems

RF scrambling systems disguise the television signal's synchronization information. The amplitude of the television carrier is modulated up and down by the scrambler. The amplitude of the aural carrier is modulated with coded pulses or sine waves. The subscriber's set top terminal receives the AM information riding on the aural carrier. From this it determines how to recover the synchronization information. The set top changes the amplitude of the television carrier up and down as necessary to restore its original shape. In the set top, the video carrier and aural carrier travel through the same path. When the video carrier is restored to its correct shape, the aural carrier is amplitude modulated with this correction too. This leaves quite a bit of amplitude modulation on the aural carrier.

An ideal FM detector ignores all AM on the carrier and correctly reproduces the original audio program. However, the demodulator in the subscriber's television receiver may be sensitive to AM. The descrambling information will then be heard in the audio output. Scrambling interference is heard around harmonics of 30 Hz, 60 Hz and/or 15734 Hz, depending on the scrambling scheme. It also has energy at twice the horizontal scan rate, which is 31468 Hz. These frequencies are for the most part out of the range of what most loudspeakers will reproduce, or out of the hearing range. Thus scrambling buzz has not been a big concern with monaural television audio.

When the subscriber is taking advantage of BTSC stereo, scrambling interference is more important. Scrambling interference has two effects: audible buzz and degraded stereo separation. Audible buzz is worse than that for mono systems for two reasons. The most obvious reason is that now the subscriber is listening to the audio more closely, and with better loudspeakers. The second reason is that the scrambling interference includes energy centered at twice the horizontal scan rate. This is right where BTSC stereo has its "difference" channel. If interference energy is present in the difference channel, it is translated into noise in the left and right audio outputs. BTSC stereo separation depends on the phase of the pilot tone at 15.734 kHz, the horizontal sweep frequency. One of the interference components from scrambling may appear here. If the interference component is large enough, it can add an apparent phase shift to the pilot tone. The stereo decoder will then produce incorrect difference channel information. Stereo separation also depends on the noise reduction system used for the difference channel. Buzz added to the demodulated difference channel may cause the noise reduction to mistrack, creating slight amplitude errors in the difference channel. These amplitude errors are translated to separation errors when the left and right channels are recreated. Scrambling systems that change the video carrier by 10 dB create more interference than 6 dB systems.

The degradations described here are not necessarily very severe. It must be emphasized that these effects are due mostly to AM detection in receivers, and that most receivers with stereo decoders do not suffer very much at all from this. As of this writing many systems are up and running in BTSC stereo with various RF scrambling schemes.

Baseband Scrambling Systems

Baseband scrambling systems, like RF scrambling systems, disguise the television signal's synchronization information. The baseband converter demodulates the video signal down to baseband from its RF carrier. While the video is at baseband, the converter restores the original synchronization information to it. At the same time, the FM sound carrier is demodulated to recover the audio baseband signal. While the audio is at baseband in the converter, the subscriber may use a volume control to adjust its level. The converter then remodulates video and audio onto RF carriers in the NTSC broadcast format. The subscriber's television set is tuned to this RF channel.

Baseband converters may introduce a slight amount of buzz into the BTSC stereo signal, but they always affect stereo separation. Buzz may occur as the video and audio are remodulated onto RF carriers in the converter's output circuits. Video harmonics that were created during envelope detection may spill into the bandwidth reserved for the aural carrier. The subscriber's TV will produce a little more buzz because of these harmonics. However, a quality baseband converter will not add any significant buzz to the audio signal.

Baseband converters affect stereo separation because of the volume control that is available to the subscriber. This volume control (and the demod-remod process in the converter) allows the user to change the deviation of the aural carrier. This changes the amplitude of the pilot tone that reaches the BTSC stereo TV, and more importantly, changes the absolute amplitude of the stereo difference signal. Proper operation of the BTSC system depends on the absolute amplitude of the carrier deviation. As the user changes the position of the volume control, the noise reduction circuit in the BTSC stereo decoder mistracks. Stereo separation is reduced. At the extremes of the volume control range, the stereo pilot may no longer be detected by the subscriber's receiver.

Reduction of stereo separation is tolerable in varying degrees to different people. It has been suggested that acceptable stereo signals are delivered for a useful range of baseband converter volume control levels. [6] The audio from a baseband converter no longer conforms to the BTSC standard. Still, it may well satisfy many subscribers' desire for stereo.

XI. CONSUMER DECODER PERFORMANCE

Consumer decoders can be evaluated on the basis of the stereo separation they can provide, the frequency response of their left and right channel outputs, and their susceptibility to buzz under different audio and video modulation conditions.

Many cable operators intend to test various consumer decoders so that they can provide subscribers with advice about stereo equipment. When testing BTSC stereo equipment, it is extremely important that laboratory quality test instruments be used. This especially applies to stereo encoders or decoders that are used as references for the measurements. Inaccurate or imprecise BTSC test equipment can make the stereo separation of devices under test look much better or much worse than it really is.

Consumer decoders are limited in stereo separation by the precision required to implement the noise reduction system that is used for the difference channel. Filters in the decoders also affect separation. Errors in the noise reduction tracking translate to degraded stereo separation. Stereo separation for some decoders is as good as 25 or 30 decibels, or as poor as 12 decibels. These variations are explained by the newness of BTSC stereo. Improvements occur as manufacturers learn about the unique demands of the format.

Limits on frequency response come from the filter requirements of a multiplex system. Complicated low pass filters prevent the sum and difference channels from corrupting each other in the decoding process. To extend the frequency response of these filters to 15 kHz while preserving separation calls for more expensive designs. As of this writing in February 1987, the left and right channel frequency response of consumer decoders extends to about 13 kHz. Broadcasters provide BTSC stereo with content up to 15 kHz, and some CATV quality BTSC stereo generators also meet this specification. It is expected that consumer decoders with frequency response to 15 kHz will be available as more experience with the format is accumulated.

Buzz performance varies widely among decoder models. It is likely that for the home audience video-related buzz and interference from some types of scrambling will be more of an issue than stereo separation. Manufacturers of decoders and television receivers must use quality receiver design and alignment to minimize interference. Headend operators can prevent severe buzz problems by maintaining correct video modulation depth.

XII. MEASURING BTSC STEREO ON THE CATV SYSTEM

Critical parameters of BTSC stereo performance include signal-to-noise ratio, signal-to-buzz ratio, frequency response, stereo separation, and relative phase between the left and right signals.

On-Line Checks

There are several on-line checks of stereo performance that you can do without interrupting service. These on-line checks provide qualitative indications of performance, and allow you to subjectively evaluate various parameters. Use these procedures to help troubleshoot system failures and recognize major changes in stereo performance.

To perform the on-line checks you need a BTSC stereo decoder, a pair of headphones, and an oscilloscope that can produce an X versus Y display. The BTSC stereo decoder may be of consumer quality. The better the decoder is, the more useful the measurements will be. The decoder must have a "mono/stereo" control. If you are working in a loud environment, headphones with some isolation from outside noises are recommended.

Signal-to-noise may be evaluated by listening to the signal through the headphones. Place the stereo decoder in the "mono" mode. Listen to the dynamics of the program audio. Observe program video. Buzz in the audio may be video dependent. Listen for changes in noise as scenes change. Adjust video modulation depth for possible improvements. Avoid video overmodulation. Note that some noise will originate in the program source material.

Buzz is most easily analyzed during quiet passages or passages with very good stereo separation. Listen to the stereo decoder while switching the decoder between its "mono" and "stereo" modes. If buzz is audible only when listening in the stereo mode, the noise is in the difference channel. Difference mode buzz may be caused by some scrambling systems or by interference from video.

Some frequency response problems will be audible in the headphones. Listen to the stereo decoder in mono mode or to a conventional monaural television receiver. If the audio sounds very tinny and the high frequencies are distorted, the aural carrier modulator may be adding pre-emphasis to the BTSC composite signal. This can only occur in a system where the stereo encoder provides baseband BTSC composite audio to the television modulator. The aural carrier modulator should be set to provide a flat frequency response, not the pre-emphasized response used in monaural transmissions. Pre-emphasis is already provided by the BTSC stereo encoder. Alternatively, check the audio input connections to the stereo encoder. Improper grounding or incorrect terminations may cause frequency response degradations or hum.

The left and right channel should be in phase with each other. Listen to the stereo decoder. Switch it between the mono and stereo modes. The apparent loudness between the modes should not change very much. If the loudness drops severely when listening in the mono mode, the left and right channels may be out of phase. Check the audio path at or before the stereo encoder inputs.

Stereo separation can be checked with the oscilloscope. Use the oscilloscope in the X-Y display mode. Connect the left channel output of the decoder to the "Y" input of the oscilloscope. Connect the right channel output of the decoder to the "X" input of the oscilloscope. Set the input sensitivities to be equal. The scope will now display a "Lissajous" pattern. When the signals on the left and right channels are the same, as during a mono transmission, the dot will move to create a diagonal line on the scope screen. The line will be angled 45 degrees from horizontal, travelling from the bottom left of the screen to the top right. See Figure 21a.

If the left and right channels have equal but opposite signals, they are said to be "out of phase." The display will now show a diagonal line angled 135 degrees from horizontal, travelling from the top left of the screen to the bottom right. Mono receivers will produce little audio output under this condition. See Figure 21d.

During stereo broadcasts, the left and right channels will often have unequal signals. The display will show a "scribbly" pattern, usually tilted toward the axis of in-phase mono programs. See Figure 21e. If the program is supplied in "synthesized" stereo, the lissajous pattern will usually be circular. When the left and right channels are out of phase during a stereo broadcast, the "scribbly" pattern will be tilted toward the axis of the out-of-phase mono programs. Mono receivers will produce little audio output.

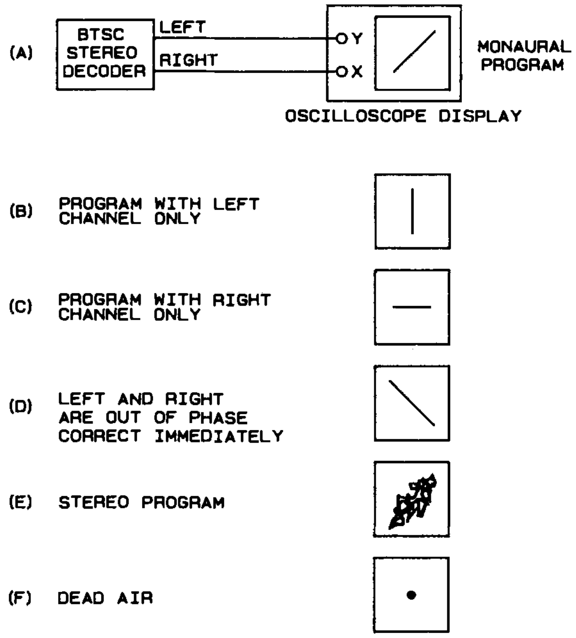


FIGURE 21: LISSAJOUS PATTERNS ON AN OSCILLOSCOPE

A different check for stereo separation in the BTSC system may be performed. It disturbs on-line performance but provides some confidence. Listen to the stereo decoder, or look at the X-Y oscilloscope display. Disconnect the right channel audio input from the BTSC encoder in the headend. The right channel audio should become greatly attenuated as heard in the headphones, while the left channel remains largely unaffected. The oscilloscope display should resemble Figure 21b.

BTSC Proof-Of-Performance Measurements

A complete proof-of-performance requires a precision BTSC stereo decoder, driven by an accurate aural carrier demodulator. A low-distortion audio oscillator and distortion analyzer are necessary. [7]

The proof of performance verifies frequency response, distortion, stereo separation, subchannel crosstalk, deviation calibration, noise performance, and pilot amplitude. A comprehensive proof usually includes tests of the stereo encoder by itself, followed by tests through the entire signal chain. Individual procedures depend on the test instruments being used. In most cases precision stereo decoders are accompanied by detailed instructions for their use in this application.

XIII. CONCLUSION

As cable operators install BTSC stereo on their systems, it is important to have access to experience at other plants. Success in implementing stereo depends on an understanding of its performance limits and where they originate, and knowledge of measurement and troubleshooting techniques.

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PROGRESS IN CONSUMER ELECTRONICS STANDARDS

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INTRODUCTION

The consumer electronics interface with cable has been a growing problem for the past decade. The principal cause of the difficulties is that the two industries developed independently without much communication. Initially, that wasn't a big deal since both cable systems and television receivers were rather simple from a technological perspective. However, as technology opened more and more options, incompatible choices were made in the two industries. Neither industry took a homogeneous stand. The various television receiver manufacturers made different choices and the large number of cable systems chose different methods of implementation. The cable subscriber, who is simultaneously a consumer electronics product buyer became caught in the middle.

Five years ago, the Electronics Industries Association, EIA, and the NCTA, Engineering Committee formed a task force to tackle these kinds of problems. The EIA/NCTA Joint Engineering Committee was born. The committee itself tackles some problems and delegates others to subgroups of specialists.

There are two primary purposes for the committee's existence. The first, and most important, is to serve as a forum for the exchange of technical information between the two industries. Mutual education will result in design choices which are more likely to satisfy the cable subscriber and the consumer electronics customer. The second purpose is to create technical standards which codify the requirements for compatibility between cable and consumer electronics technology.

There are three levels of standards. The most mature standards are in the "RS" series. RS means "recommended standard". Perhaps the most familiar example of this series is the RS232 standard used with computers and data communications devices. A step along the way to "RS" is "IS" standing for "interim standard". An IS standard is issued on a trial basis for a year or two for manufactures to attempt designs in order to more fully understand the consequences of the standard's details. After the trial period, the standard is amended to include learning from the past year and voted upon by the EIA for promotion to RS status. The least mature phase in the development of a standard is the "Recommended Practice". It is intended to indicate a direction for manufacturers to choose in an area where there may be many reasonable approaches but industry interest does not support the development of a full standard. Recommended practices are not as thoroughly debated or tested as IS or RS series documents.

It is important to realize that these standards are voluntary. Neither the NCTA nor the EIA have enforcement powers. Adherence to the standards depends on the good faith of the companies involved.

IS-6: CHANNELIZATION STANDARD

The channelization standard is a case study of the process of standards creation. Engineers from the two industries met and educated each other on the various methods used to allocated frequencies in the cable spectrum to channel designations. A debate ensued over the pros and cons of the various methods. Of course, individuals wished to preserve the methods they used in the past to minimize changes required of them. Far sighted participants tried to accommodate future needs. After much debate a compromise approach was found. Some questions were deferred until more experience was gathered. The interim standard was issued in May of 1983. Manufacturers then evolved their product designs towards compliance with IS-6.

In late 1986, the committee took up the issue of finalizing IS-6 into a proposal to the EIA for promotion to RS status. The principal issues remaining were the channelization of the FM band, the order for expanding channel capacity, and the method of counting channel capacity. The FM issue centers on the fact that receivers generally have traps (frequency selective filters) in the FM band to prevent interference with Channel 6 reception when strong FM signals exist in the reception area. This trapping practice is essential for off-air performance and therefore cable operators must use channels in the FM band accordingly. TV receiver manufacturers will likely strive to develop switchable filters for future product. While not technically practical at present, the need has been highlighted and the consumer electronics industry is now aware.

The order in which channels are added when capacity is expanded and a fair method of indicating capacity to the consumer have been agreed upon.

Before the channelization standard, cable companies used numbers and letters to designate channels in a variety of ways. There simply were a number of equally logical ways of doing this and no mechanism to coordinate between those making the choices. A serious consequence of this situation is that it became impossible for consumer electronics product manufacturers to make receivers which complied to multiple channelization methods. Now, with IS-6, cable practice and consumer electronics design can converge over time to the benefit of the subscriber.

IS-15: DECODER INTERFACE STANDARD

Perhaps the standard which has the most potential to solve consumer electronics interface problems is the IS-15 Decoder Interface Standard which is also known as the EIA Multiport. The standard is embodied as a 20 pin plug on the back of a television receiver or VCR which accepts a set-back descrambler. It has been adopted and endorsed at the IS level by both the EIA and the NCTA.

The principal advantage of the Multiport is that it makes a truly cable ready receiver possible in a scrambled environment. Because descrambling is accomplished after the receiver's tuner, the consumer electronics product can be directly connected to cable. The subscriber regains use of his remote control. In the case of a VCR, the timer again becomes useful. It can again control channel selection and turn the VCR on and off. An important secondary advantage is a significant reduction in cost to the cable operator. Set-back descramblers will be 40% to 60% the cost of set-top units. It becomes practical to provide two units, one for the TV and another for the VCR. For the first time, it's possible to watch one scrambled channel while recording a different scrambled channel at an affordable price.

A practical limitation of the EIA Multiport is that it requires the subscriber to purchase a new Multiport equipped TV receiver or VCR. This won't happen overnight. Unfortunately, TV's last too long. The typical life is twelve to fourteen years. New receivers are bought every seven years with the old unit put in the basement or donated to one of the kids who grew up with it. Significant penetration will take time. However, the subscriber who feels he needs a solution can contribute to it buy making a purchase. Even that was unavailable just a few short years ago.

The situation is dramatically different with VCR's. Since they wear out, VCR's are replaced every three or four years by heavy users. The rotating heads are a critical mechanical element in an otherwise electronic system. They clog and wear causing expensive repair bills. In many cases, the cost of repair rivals the cost of a new unit. Since the purchase of a VCR more than doubles the trouble with the consumer electronics interface, it is particularly appealing to find that the EIA Multiport can bring relief when taken as an option on a new VCR.

First TV's and VCR's with Multiport are expected on the market this year. Descrambler vendors have promised evaluation samples in the beginning of the third quarter of this year with volume delivery a few months later. The NCTA Engineering Committee has formed a subcommittee under Joe Van Loan of Viacom to promote and stimulate the Multiport among the MSO's. While there is still a lot of work to be done, progress on this important standard is heartening.

IS-23: RF-CABLE INTERFACE STANDARD

An analysis of the requirements for true cable compatibility yields two requirements: 1) the TV or VCR must be able to be connected directly to the cable without a converter or descrambler ahead of it, and 2) the internal circuits of the TV or VCR must not pick up off-air signals directly. This direct pick up problem causes ghosted images and, in HRC systems, annoying diagonal bars in the picture. There are only two ways to avoid direct pick up. First the subscriber can avoid living close to a TV transmitter. Secondly, he can own a TV or VCR with adequate internal shielding. IS-23 is intended to set technical standards that make the second option realistic. Additionally, IS-23 deals with signal levels, connector types, and the allowable level of signals back fed into the cable. The standard went up for vote at the end of 1986. TV manufacturers found its direct pick up requirement difficult to achieve. They've asked for further clarification and compelling evidence of the need for such a severe standard. The committee went on hold while this issue received further investigation. The committee will resume its deliberations in the third quarter of this year.

CONCLUSION

A lot of progress has been made. Communications between the cable and the the consumer electronics industries has increased and improved by several orders of magnitude. The result will be greater satisfaction with cable service as enjoyed through subscriber owned consumer electronics products. However, we must have realistic expectations. There are about 200 million television receivers in American homes that were designed before standards were accomplished. It will take time for these to be replaced with more compatible models. But that eventual goal would have never been attainable had it not been for the work of the EIA/NCTA Joint Engineering Committee and its subgroups.

PROPERTIES AND SYSTEMS CALCULATIONS OF OPTICAL SUPERTRUNKS FOR MULTICHANNEL TV TRANSMISSION, USING ANALOG INTENSITY MODULATION, SINGLE MODE FIBERS AND HIGH DEVIATION FM

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ABSTRACT

Optical supertrunks are being installed in an increasing number. The technology is mature and the prices are getting attractive. It is shown here how transmission bandwidth, transmission quality and optical loss budget can be used as planning elements for the determination of system parameters as modulation, frequency plans, and drive levels in an analog fiber optic transmission link.

I. INTRODUCTION

TV multichannel transmission has been tried in the past decade mostly with mediocre success, especially when multimode fiber was used. Now we have learned to use the right fiber, the right modulation, and the right frequency plan to avoid noise and intermodualtion problems so that we can transmit channel numbers and video signal qualities that a few years ago were unthinkable.

II. THE TRANSMISSION BANDWIDTH OF SYSTEMS USING SINGLE MODE FIBERS

A: The transmission bandwidth of the laser/fiber combination

The transmission bandwidth of the laser/fiber combination can be derived from the systems response to a Gaussian pulse, using the Fourier transform [1]. A good approximation for the total rise time is [2]:

$$t_{rise,tot} = \left(t_{laser}^2 + t_{fiber}^2\right)^{1/2} \qquad (1)$$

In a single mode fiber intermodal dispersion is nonexistent. Among the intramodal dispersions chromatic (or material) dispersion is predominant [3]. Chromatic dispersion is a function of the dispersion property of the fiber used, the fiber length and the spectral width of the laser. The rise time of a pulse propagating thru a fiber is therefore

$$t_{fiber} = D_c \times \Delta\lambda \times L \qquad (2)$$

with D_c : Chromatic dispersion of the fiber (ps/nm–km)
$\Delta\lambda$: Spectral width of the laser diode (nm)
L : Length of the fiber (km)

The electrical 3 dB bandwidth can now be calculated to be [1]:

$$B_{3\,dB} = 0.375/t_{rise,tot}$$
$$= 0.375/\left[(D_c \times \Delta\lambda \times L)^2 + t_{rise,laser}^2\right]^{1/2} \qquad (3)$$

The presently most often used wavelength is 1300 nm where the fiber loss is low, typically 0.5 dB or even less and dispersion is zero using low loss oxide-glass fibers. For the even lower loss wavelength 1500 nm two kinds of fibers are available or in development: Dispersion shifted or dispersion flattened [4]. Most fibers that are installed today are dispersion nulled at 1300 nm. Using a laser at that wavelength the bandwidth in (3) becomes independent of the fiber length. Typical rise times of InGaAsP lasers are around 0.3 ns so that the transmission bandwidth of this laser/fiber combination is above 1 GHz. Figure 1 shows the chromatic dispersion coefficient of a typical single mode fiber:

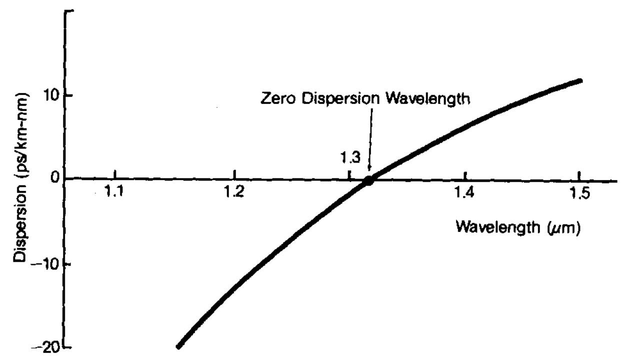

Figure 1: Chromatic dispersion coefficient of a typical single mode fiber

Using (in a WDM application with 1300 nm) a typical 1550 nm InGaAsP laser diode with a spectral width of 4 nm and a chromatic dispersion of 13 ps/km-nm, the resulting bandwidth of a 30 km link can be calculated using (3) to be only 240 MHz. Recently DFB laser diodes with a spectral width of only 0.1 nm have become available. Chromatic dispersion can then be neglected for the above mentioned fiber for lengths up to about 70 km.

B: The receiver bandwidth

The receiver rise time can be included in equation (3). Typical receiver bandwidths are 600..700 MHz using Ge APD's that are easily available and 1 GHz or higher using faster InGaAs APD's or PIN detectors.

III. CHANNEL SPACING

A: Channel spacing, deviation, and video bandwidth

The bandwidth of a frequency modulated signal is approximately

$$B = 2\,(D_p + f_m) \tag{4}$$

with D_p: Peak deviation

f_m : Modulation frequency (4.2 MHz for NTSC)

Deviation is often expressed as the deviation of the non-emphasized part of the video signal. It is called "Sync tip to peak white" (STPW) Deviation. If the video signal is pre-emphasized in accordance to CCIR Rec. 405 (525 lines) equation (4) becomes

$$B = 2\,(D_{STPW} / 1.64 + f_m) \tag{5}$$

As pointed out earlier, transmission bandwidths over 1 GHz with fiber lengths, that are limited mostly by the minimum optical receive power, are feasible today. The number of channels per fiber is a trade off versus video quality. Reasonable video specs (EIA-250-B long haul) can be achieved with channel numbers up to approximately 24 channels per fiber. Having a 30% guard band between adjacent channels, deviations below 8 MHz STPW are not optimum. A deviation of 800 kHz STPW as used in older coaxial FM equipment is useless on fiber.

B: Channel spacing as required by the adjacent channel protection ratio

The adjacent channel protection ratio is expressed here as the video signal to periodic noise ratio SPR_{video} :

$$SPR_{video} = 20 \log (100\ IRE / V_{interference,pp}) \tag{6}$$

This definition of periodic noise is used in the EIA standard EIA-250-B [5] as well as the CCIR[6]. Figure 2 shows the measured SPR_{video} (using a Rode & Schwarz UPSF2 noise meter) as a function of the frequency separation of a wanted channel and its adjacent. Both carriers are frequency modulated with 8 MHz STPW. No transmit filter is used in this test. The receiver filter has a bandwidth of 29 MHz. The video modulation of the wanted channel is "flat field", the one of the interfering adjacent is "pulse and bar" with a 30 IRE subcarrier at 6.2 MHz. A live video signal gives slightly better results.

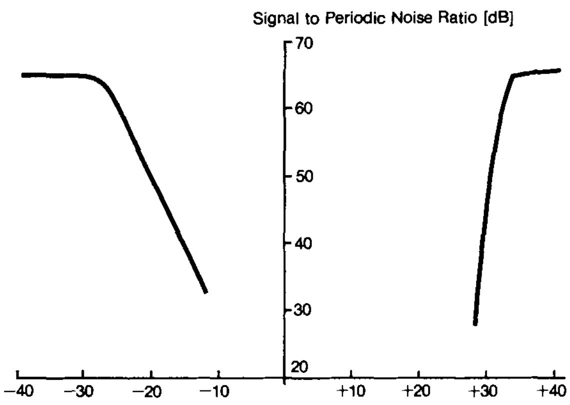

Figure 2: Adjacent channel protection ratio for 8 MHz STPW deviation

Figure 2 shows that without transmit filtering a channel spacing greater than 36 MHz is feasible, with or without using a subcarrier, when the STPW deviation is 8 MHz and the receiver filter bandwidth is 28 MHz. For other deviations or filter bandwidths this test has to be repeated.

IV. INTERMODULATION DISTORTIONS

A: Second order distortion and frequency planning

In a laser diode second order distortions are predominant. A way to get around these intermodulation products is using only an octave of the available frequency spectrum [7]. This is useful, when vestigial sideband signals are transmitted. The situation is totally different, when wideband FM is used instead. Figure 3 shows the protection ratio when an unwanted signal with 16 MHz deviation is tuned thru a wanted one with a deviation of 8 MHz. As in the adjacent channel protection ratio the video modulation of the wanted signal is "flat field", the one of the unwanted signal is "pulse and bar" with a 30 IRE subcarrier at 6.2 MHz.

Here the protection ratio is the ratio of wanted to interfering signal power in dB which is necessary to achieve a video signal to periodic noise ratio of 60 dB. It is possible to extrapolate between protection ratios at center frequency for different deviations than 8 MHz STPW using 20 log (D1/D2) as a correction factor.

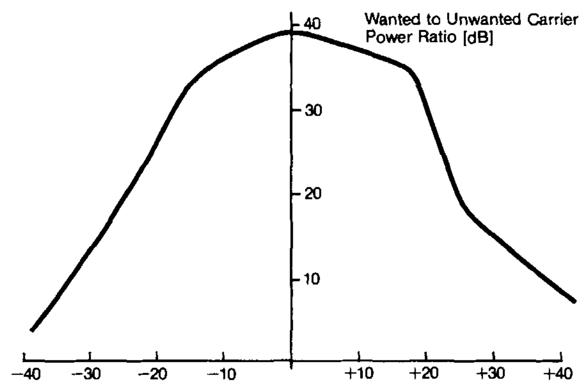

Figure 3: Protection ratio for a 16 MHz deviation signal interfering one with 8 MHz deviation for 60 dB signal to periodic noise ratio

A second order efficient frequency plan, suggested by Synchronous and used and described by [9], has the channel frequencies

$$f_{ch} = f_s/2 + n \times f_s \tag{7}$$

with f_{ch}: channel frequency [MHz]

f_s : channel spacing [MHz]

n : channel number (integer).

All second order products f_{IM2} have the form

$$f_{IM2} = m \times f_s \qquad \text{(m: integer)} \tag{8}$$

which means that they fall between channel frequencies.

From figure 2 can be concluded that with a channel spacing of 36 MHz or more second order products can be as high as −12 dBc. The efficiency of this frequency plan is therefore 27 dB, when compared to the situation where the products fall on the center frequency. Using (7) allows to ignore second order products because now third order products become predominant.

B. Third order distortion

The weakly nonlinear region and the clipping region

Intermodulation analysis is performed for the weakly nonlinear region of a nonlinear device using a power series approach [9] to determine the relative power of intermodulation products. A useful parameter of a nonlinear device is its third order intercept point [10). If it is a constant then the power series approach is justified. We have measured the third order intercept points of several InGaAsP single mode lasers with a wavelength of 1300 nm. Their average third order two tone input intercept point power was 22 dBm at 100 MHz, the standard deviation of 5 samples was 4.3 dBm. It decreases with frequency between 100 and 800 MHz by 7 dBm.

When a laser diode is instantaneously driven near its threshold then the laws of the weakly nonlinear region are not valid anymore. Typical laser diodes reach this clipping region with a two tone power of 0.5 mW per tone. Typical measured total input powers into the InGaAsP lasers we use with a 12 channel signal for optimum performance varies from 2 to 5 dBm. This indicates that the laser diode is not used in its weakly nonlinear region and that classical composite triple beat ratio calculations can not be used anymore.

The laser intensity modulation as a function of the number of channels

We found empirically, that using a depth of intensity modulation m of 0.8 the drive level has to be decreased by

$$\Delta L_{RF} = K(N) \times \log(N) \quad \text{(dB)} \tag{9}$$

with ΔL_{RF} : Decrease in RF drive level
$K(N)$: 18 for $N< 8$,
16 for $12<N<18$,
15 for $N<28$
N : Number of channels.

The RF drive level is therefore so adjusted that the RF signal peaks reach into the clipping region of the laser. Here the composite triple beats increase by more than 3 dB per dB signal increase until carrier to beat power ratio is around 40 dB, where it affects the video SNR rapidly.

V. CARRIER TO NOISE AND VIDEO SIGNAL TO NOISE

A: Carrier to noise ratio as a function of received optical power and depth of intensity modulation per carrier

Using an APD receiver followed by a transimpedance amplifier the carrier to noise ratio CNR can be approximated by

$$CNR = \frac{(m_i \times R \times P_r \times M)^2/2}{(RIN \times R^2 \times P_r^2 \times M \times B) + 2q\,(R \times P_r + I_d)BM^{2+x} + (4KTB/R_f)F} \tag{10}$$

with
CNR : Carrier to noise ratio (4.2 MHz noise bandwidth)
m_i : Modulation index of the i-th carrier
P_r : Received optical power (W)
R : Responsitivity of the APD (A/W)
M : Multiplication factor of the APD
x : Excess noise factor of the APD
I_d : Dark current of the APD (A)
RIN : Relative intensity noise of the laser diode (W/Hz)
B : Noise bandwidth (4.2 MHz)
q : 1.6E-19
k : 1.4E-23
T : Temperature (K)
R_f : approx. the transimpedance
F : Noise figure of the transimpedance amp.

The three terms in the denominator of equation (10) indicate, that an optical transmission system can be laser limited (the RIN term predominates), quantum noise limited (the seocnd term predominates) or receiver limited [11].

In short optical links (total loss less than 15 dBm) the use of a PIN detector (M=1 in (10)) is adequate. The biggest problem in short links is the laser sensitivity to reflections that are produced by connectors, splices and the detector itself [12]. This problem is harder with single mode than with multimode fibers. The reflection coefficient at the laser itself is a function of the reflection coefficient of the connector etc. and of the coupling efficiency between laser and fiber [13]. For that reason laser diodes are available with a low coupling efficiency and therefore a lower output power (−6..−10 dBm). We observed with optical losses below 10 dB and efficiently coupled lasers RIN degradations of over 10 dB. This would make a low loss link with a potential to transmit 8 channels with more than 67 dB video SNR unpredictable. Recently substantial improvements in connector design have been reported [14].

In medium length optical links, where quantum noise predominates, the choice of the right APD is critical. Using a InGaAs APD instead of a Ge APD can increase CNR by as much as 6 dB in this region.

At high optical losses the receiver noise is the limiting factor. For a given optical receive power P_r an optimum APD gain can be calculated:

$$M_{opt} = \left[\frac{4kTF / R_f}{q\,(R \times P_r + I_d)\,x}\right]^{1/(2+x)} \tag{11}$$

Because R_f is indirectly proportional to the receiver bandwidth, a careful tradeoff has to be made between the optical loss budget and the transmission bandwidth when optical losses are high.

B: Video SNR as a function of the CNR of the optical receiver

As described above a good prediction of CNR is possible by using m=0.8 in (10) and by reducing the resulting CNR in accordance with (9). The video SNR is a linear function of CNR above the FM detector threshold. Various CNR/ video SNR relations exist, depending on EIA or CCIR weighting, 100 or 140 IRE reference etc. [15]. We have verified using the Rode & Schwarz noise meter for video SNR, and the Tektronix spectrum analyzer 7L13 as well as the power meter HP 435 B for CNR the equation

$$SNR_{video} = CNR_{4.2} + 12 + 20 \log (D_{STPW}) \quad (12)$$

with SNR_{video}: CCIR weighted Video SNR, referenced to 100 IRE

$CNR_{4.2}$: CNR with 4.2 MHz noise bandwidth

D_{STPW} : Sync tip to peak white deviation ($=D_{pp} / 3.27$)

C. Performance of 8, 12, and 24 channel transmission systems

Systems with 8 and 12 channels on one fiber have been installed in the past [8]. With 12 channels we measured typically video SNR's better than 63 dB for optical losses that are less than 24 dB. To find the limit of our present optical fiber transmission system we run a test transmitting 24 channels with 4 MHz deviation (STPW) and an optical loss of 16 dB. Equation (10) and (12) predicted a video SNR of approximately 57 dB. Reading the CNR in figure 4 (Spectrum analyzer correction factor 2.5 dB) to be around 34 dB verifies the predicted video SNR. This experiment shows that very high channel densities on one fiber can be realized using high deviation FM if the signal quality does not have to be very high.

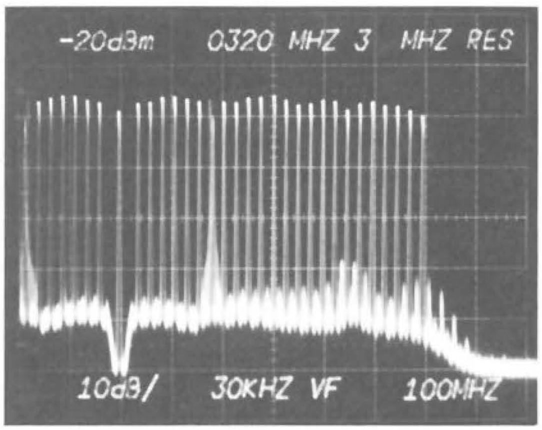

Figure 4: 24 channels over 16 dB of optical loss

VI. THE TRANSMISSION OF BTSC STEREO OVER FIBER

The transmission of stereo audio signals over fiber can be done using subcarriers or carriers at discrete frequencies. Discrete audio carriers are normally preferred because there is no interaction between audio and video. The EIA-250-B standard prescribes methods of measurement for audio SNR (video flat field with 50 IRE) and periodic noise with audio on and video off that do not allow to catch these interactions. Systems with 70 dB specified audio SNR have shown over 20 dB lower SNR with certain video waveforms. Synchronous has designed a high deviation audio transmission family. 4.5 MHz intercarriers are multiplied and transmitted with a 4 times higher deviation than the original. Stereo separation as well as audio SNR has been proven to be unaffected by this transmission scheme in field tests that Synchronous run with Gill Cable in San Jose.

VII. THE TRANSMISSION OF SCRAMBLED SIGNALS OVER FIBER

Frequency modulators are usually phaselocked so that the average output frequency is at the nominal center frequency. They are therefore acting as highpass filters with a cutoff frequency equal to the natural frequency of the PLL. DC can only be restored at the receiver with a sync driven clamp. Scrambling systems that use baseband scrambling have therefore to provide a clamp pulse that can be transmitted on a subcarrier to restore DC of the scrambled video signal at the receiver. More difficult is the transmission of RF-scrambled signals. The method of transmitting a VSB signal at a video frequency [16] has shown to be hard to realize. Linearity problems of the FM modem prohibit using a VSB carrier around 1.5 MHz. An inverted spectrum needs substantially higher deviations to get marginal video SNR's.

VIII. CONCLUSIONS

The transmission of multichannel TV up to 24 channels per fiber over optical links has reached a high degree of maturity. The user can take a choice between video signal transmission quality, number of transmitted channels and optical loss budget by selecting the right components at the transmit and receive end. Using the given equations, frequency plans and protection ratios, the planning of an optical trunk is straightforward.

Transmit filtering is not necessary when high deviation FM is used.

Using subcarriers affects the channel spacing very little, but discrete audio carriers avoid any interaction between audio and video.

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REPORT ON THE COAXIAL PORTION OF THE EIA HOMEBUS STANDARD EFFORT

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ABSTRACT

The Electronics Industry Association (EIA) has been working towards the development of a consumer electronics bus (CEBus) standard for several years. During the past year, significant forward progress has been made in a number of bus related areas, and among these is the coaxial portion of the Wired Bus (WIbus), more commonly referred to as the CXbus. This paper reports on the status of this bus topology proposal by presenting results taken on a system model, and also explains some of the thought processes that have been used to arrive at the currently proposed system topology. A possible realization of CATV interconnection is also detailed.

DISCLAIMER

The information contained in this paper is based on work in progress towards an EIA standard. It represents the consensus opinion of the members of the Wired Bus subcommittee and may not fully represent the final version of the standard. The author is not an official spokesman for the EIA.

INTRODUCTION

Consumer electronics manufacturers have shown consumers the advent of a number of sophisticated products (primarily audio and video) over the past several years. These products have produced an increased awareness of the need for the integration of similar type devices into a coordinated system. Manufacturers have attempted to provide interconnection compatibility within their own product lines and have been successful to a certain extent. When manufacturer lines are crossed however, it becomes obvious that the interaction issues are far from resolved. The VCR/CATV dilemma is but one example of this growing problem familiar to the CATV operator. The EIA Homebus standard will provide industry wide "rules of the game" that will sort out how these interrelationships between devices are established. Products that meet these rules will bear the Homebus logo and will be designed for easy "plug and play" compatibility. The standard will apply to all segments of consumer products, but for the CATV operator the integration of the home entertainment center will be the predominant issue. The author hopes that the information in this paper will provide the CATV system operator with the ability to understand the direction of the standard development, and to begin considering the future implications that this standard contains. The intent is not to consider the relative merits of the practical uses of the Homebus standard, but to share technical information on what the author believes will closely represent the CXbus standard.

LOGICAL DEVELOPMENT OF TOPOLOGY

The current topology was arrived at through a series of conceptual iterations which included star, multilevel star, tree and branch and the proposed resemblance of tapped trunk. In addition, it was known that video transport would be required for what was to be called In Home Generated Video (IHGV). The IHGV was to be in a controlled portion of the spectrum. A source would not be allowed to transmit signal unless permission was obtained to do so via the control channel link. This IHGV was to be provided from sources within the home, such as a video camera or VCR, and distributed to receive points in the home for a variety of uses. The actual frequency allocation of this IHGV was a greatly debated issue and was a driving force in determining the level of complexity of the upstream cable. There were a number of system related factors that were considered important to the successful development of the CXbus standard. The main ones were:

Minimum system requirement
Ease of system extension
Cost complexity tradeoffs
User friendliness (plug and play capability)

In addition there were many topological issues related to system performance such as:

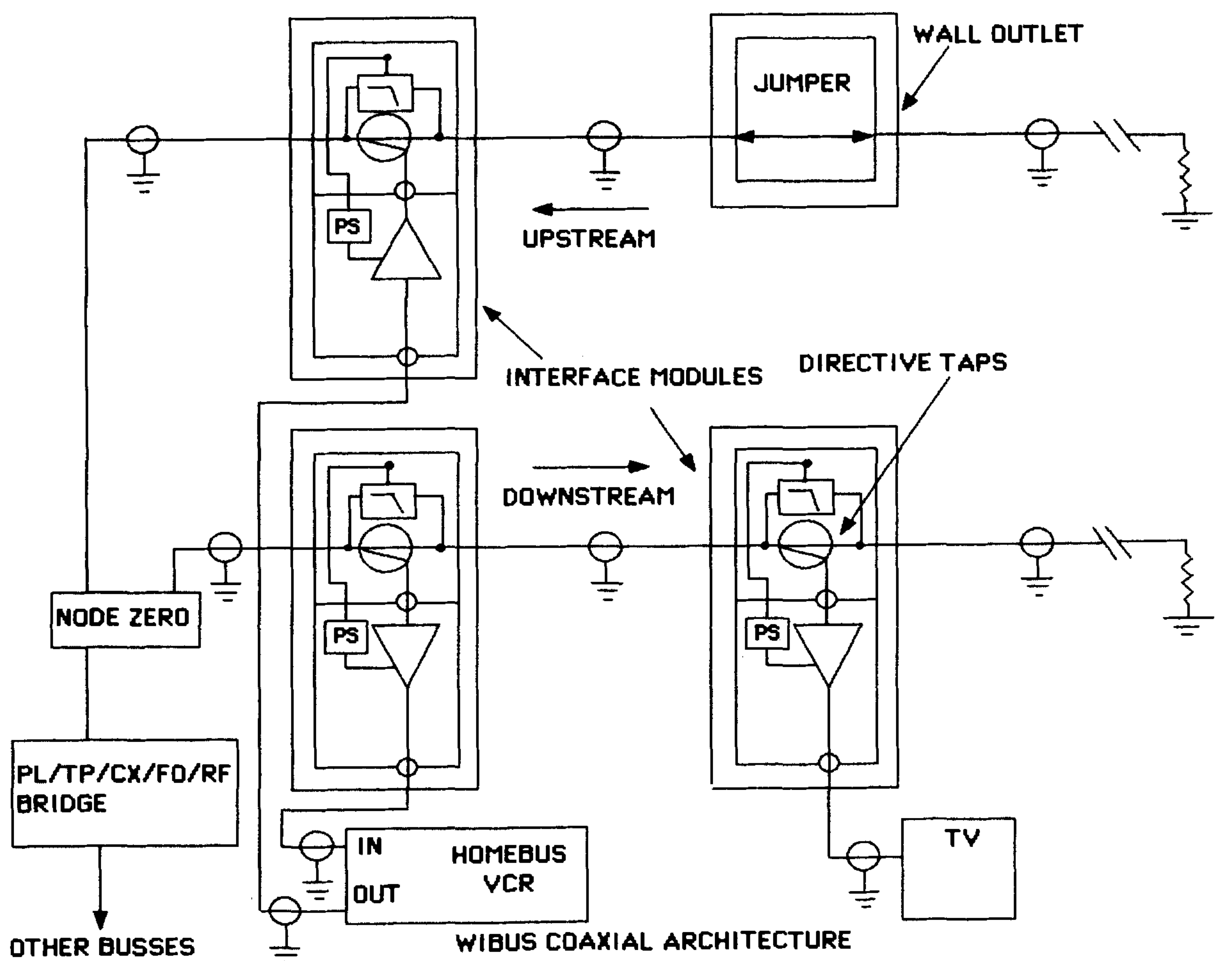
Cable footage efficiency
Practical system length
Technical performance
Powering
Required number of outlets

Through the dedicated efforts of the committee members, these and other issues were traded off against each other and the system described herein evolved.

GENERAL SYSTEM DESCRIPTION

Physical Media

The proposed topology (Figure 1) is dual cable with one cable used for downstream and the other used for upstream transmission. The separate cables exclude the need for a conventional two way distribution system, reducing the complexity of system implementation since diplex filters and two way gain capability are not required. The recognized tradeoff for this was the addition of more cable footage to the system, but the consensus was that this tradeoff was a reasonable one. The CXbus itself is capable of standing alone as an independent WIbus system. The minimum system can consist of a single outlet, with the appropriate interface module and the Node Zero realization. The system will have a defined

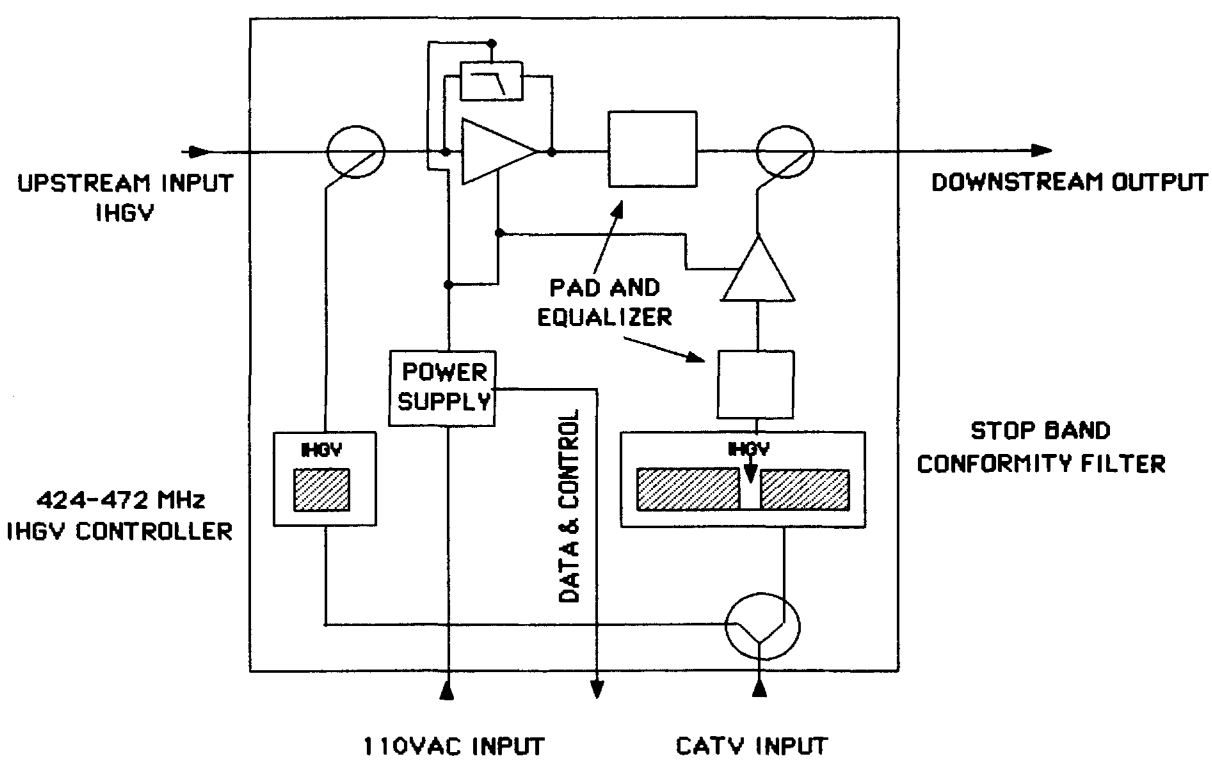


WIBUS COAXIAL ARCHITECTURE
Figure 1

operation to 550MHz with less stringent definition to 806 MHz, this upper portion of the spectrum used primarily to accommodate carrying off air UHF TV signals. The downstream and upstream cables will have the same technical definition, but will of course have different directions of signal flow. Connection to the RF portion of the spectrum of either cable is done through a low loss directional tap which is contained in the interface module. Amplification is also provided for acceptable signal level output or injection depending on which cable is being accessed. In conjunction with this RF coupling, an additional path is provided for low frequency (<1MHz) signal passing and for AC or DC powering as applicable. This low frequency path will be used for control and data applications and will be directly linkable to other bus structures. The Node Zero realization that might be applicable for a minimum system which has been interfaced to an existing CATV system is shown in Figure 2.

Frequency Allocation

As previously discussed, the frequency of operation for this system is from 0 Hz to 806 MHz. Within this wide range of operation there are a number of frequency allocation bands available for use. Figure 3 details the current frequency allocation proposal for the CXbus. Current thinking has the IHGV located in the 424 to 472 MHz range for NTSC type signals and another band from approximately 10 to 54 MHz for other wideband applications. This higher IHGV frequency allocation has some implications for CATV systems that are 450 MHz in bandwidth or higher and will be connected to a CEbus system. The IHGV portion of the spectrum will have to be "unallocated" from the CATV input since this is a reserved and controlled frequency segment. A stop band conformity filter would perform that function. In addition, an IHGV controller would provide access to those CATV channels that reside in the IHGV



NODE ZERO IMPLEMENTATION
Figure 2

spectrum. Figure 2 shows the implementation implication caused by this IHGV allocation.

SYSTEM TEST BED

Cable Interconnection

The reported system test bed consists of a 20 outlet system with a total cable footage of 312 feet. This system was constructed using RG-59 drop cable with a nominal loss characteristic of 8.35dB/100 ft at 550MHz. The cable length between each outlet and the total footage is shown in Table 1.

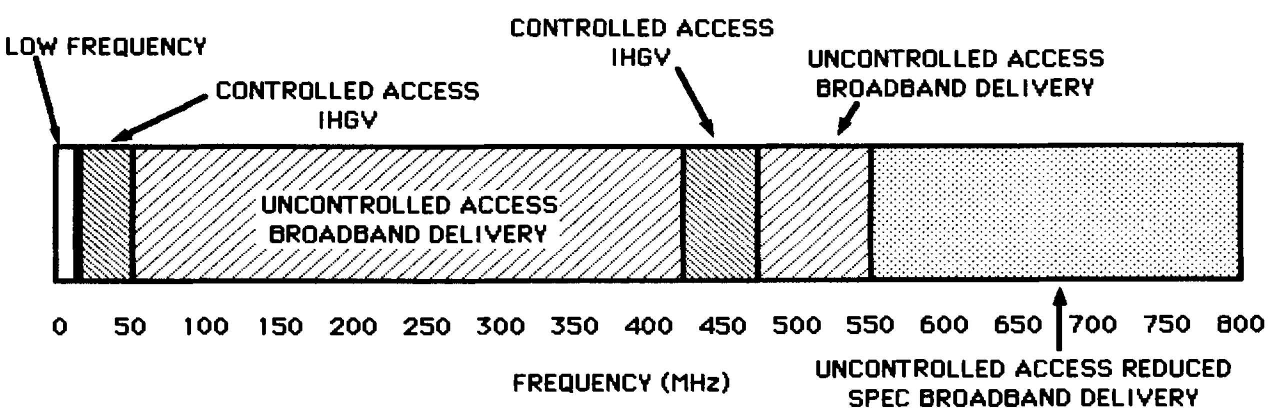
Outlet Number	Additional Footage	Total Footage
1	0	0
2	14	14
3	03	17
4	10	27
5	27	54
6	02	56
7	38	94
8	02	96
9	12	108
10	12	120
11	15	135
12	24	159
13	14	173
14	31	204
15	02	206
16	23	229
17	32	261
18	12	273
19	07	280
20	32	312

Table 1

These cable lengths were arrived at from an actual floor plan of a sample house which put one outlet on practically every interior wall. This particular house contains approximately 3000 feet^2 of living area and is in no way intended to represent a "standard" or "average" house. This does not limit the usefulness of the model however, because the dimensional relationships between interior walls is relatively constant over the vast majority of American homes.

Outlet and Module Interconnection

The outlets themselves into which the interface modules connect were constructed out of FR4 PCB and had provisions for two interface module connections, one for the downstream and one for the upstream cable. The outlets were sized to fit within a standard 110VAC electrical outlet box. Connection to the outlet PCB was made with conventional "F" connectors. Connections to the interface module were made with the custom coaxial connector used in the SA 6501/6502 distribution amplifier for diplex filter connection. This connector was used strictly for the sake of convenience. Five functional interface modules were built and used in the testing. These were four port devices with provisions for a main system input and output, a directively tapped (-17dB) RF output and a nondirective low frequency tap. Figure 1 shows an additional gain stage used in the interface module, but in this particular test system the gain stage was not an integral part of the module so that various types of gain stages could be evaluated at a later time. Plug-in jumpers were used in outlets that did not contain an interface module.



PROPOSED FREQUENCY ALLOCATION
Figure 3

The purpose of the system model was to determine the limitations of the proposed topology and to collect data on the actual performance of such a system. In addition, the model will be used to formulate and answer questions regarding complete system integration.

System Data

After construction, the 20 outlet system was swept with jumpers only to determine the losses associated with the cable and the outlet PCBs alone. This information is presented in Figure 4 as a function of four frequencies and represents the minimum system loss possible due to cable and jumpered outlets alone. The five available interface modules were added to the system and the same measurements made. These results are shown in Figure 5. The incremental loss for a single added interface module is shown in Figure 6. Figure 7 details the system loss for a given cable footage with only jumpers and outlet PCBs installed.

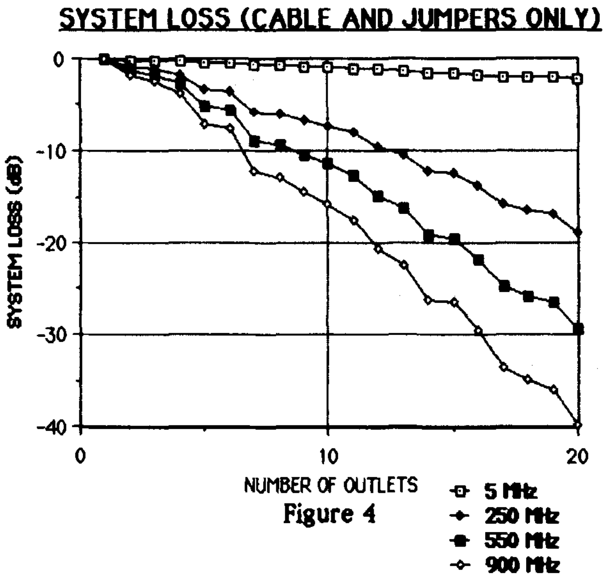


Figure 4

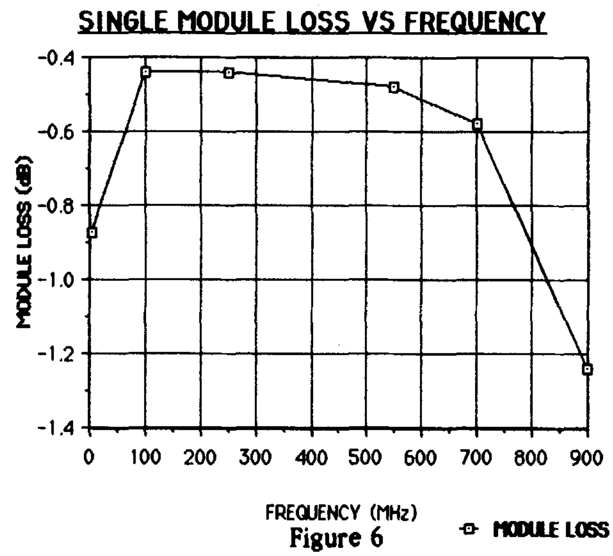


Figure 6

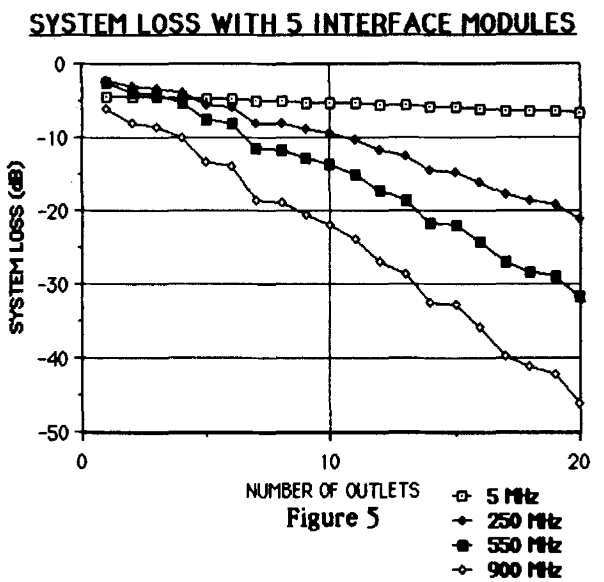


Figure 5

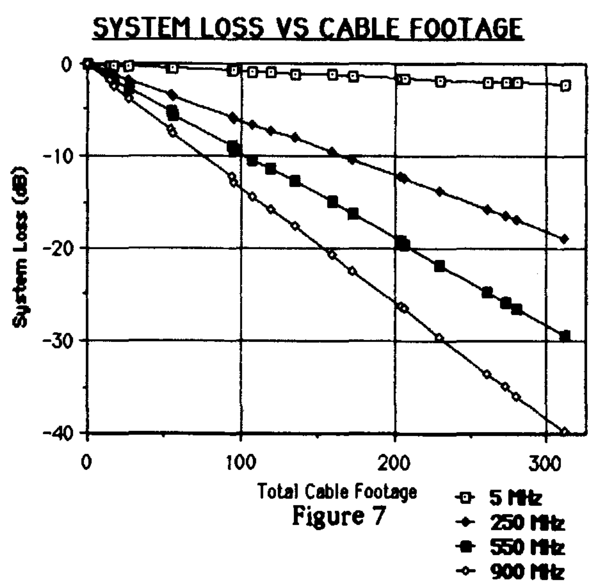


Figure 7

SYSTEM LIMITATIONS BASED ON DATA

General

As stated earlier, the intent of this system analysis was to establish more concrete guidelines for the final implementation of the CXbus structure. The perspective that one has towards the real use of this system is a great determinant in establishing the technical parameters for the operation of the system. When the CXbus structure is examined as a totally closed system, the normal specifications that the CATV industry puts on video distribution systems should not be difficult to meet because there will generally be only two amplifiers in cascade in conjunction with a manageable system loss. When the practical reality of the situation is considered however, it becomes extremely difficult to imagine this system not being connected to a CATV system in the majority of cases, and it will certainly be connected to an off air antenna in the remaining ones.

System Length

The driving issue that determines the system length is the maximum signal level that will be possible to put on this system. Based on Carrier to Noise calculations, it appears that the minimum system level should be in the range of +10 dBmV in order for the interface module to have satisfactory noise performance. This is based primarily on the noise figure of monolithic devices available for the interface module amplifier. It therefore follows that the loss allowable by the system will be:

X dBmV -10 dBmV=Y dB

where X is the maximum system level and Y is the allowable loss before an extension (line extension) of the system occurs. If one assumes that Y will be on the order of 15 dB, a general guideline for the system might be an eight outlet system with 135 feet of cable capable of having five interface modules active at a given time. This was calculated using Figure 6 to determine the loss due to the five interface modules and using Figure 7 to calculate the loss due to the cable footage and outlet PCBs. The combinations of system realizations is large once the available loss is known and one of the major challenges of the final system implementation will be to provide an acceptable means of signal level management throughout all of the possible combinations. This maximum system level issue is being addressed from a regulatory standpoint at this time.

CONCLUSIONS

The data presented represents the status of the CXbus system topology. There are a great deal of issues yet to be addressed and these will be resolved as further work is done on the standard. Some of these issues will be generated by committee input and some will come about as a result of more in depth evaluations on the system topology as it further evolves and begins to more completely unveil the related intrabus connection issues. Work to date has focussed on the broader issues relating to the system concept. Now begins the task of complete technical definition.

With regards to CATV systems in particular, this system concept moves more towards an off premise implementation of cable systems. The implications that the emerging Homebus standard has for this industry will become more apparent as other portions of the standard are firmed up, however now is the time to begin consideration and comment on the possible effects of the topology presented herein.

ACKNOWLEDGEMENT

The author would like to thank the members of the EIA Homebus committee for their inputs. A special thanks goes to Mr. Charles Bedgood of Scientific-Atlanta who designed, constructed and evaluated the system model.

SATELLITE DELIVERED TAG CHANGE SYSTEM

Andrew Ferraro

REQUEST TELEVISION

ABSTRACT

Since the launch of regularly-scheduled satellite PPV on November 27, 1985, many cable systems have discovered that their in-place hardware presents obstacles to full participation in this potentially profitable revenue stream. Specifically, their addressable hardware does not allow them to show multiple events in a 24-hour period.

The problem is this: Billing systems rely on a tag or address to identify a program service so the customer can be billed properly. With a monthly pay service, this is not a problem. In fact, tag levels were initially designed to work with monthly pay services.

It's not so simple with PPV where each event must be considered a different service, and the tag or address must be changed for each. Otherwise, the customer would only be billed for one view but actually would be able to see the complete days' programming.

MULTIPLE EVENT SCHEDULE

TIME	TITLE	TIME	TITLE
9:00AM	TOP GUN	7:00PM	THE FLY
11:00AM	TOUGH GUYS	9:00PM	TOP GUN
1:00PM	THE FLY	11:00PM	TOUGH GUYS
3:00PM	LEGAL EAGLES	1:00AM	LEGAL EAGLES
5:00PM	TOUGH GUYS	3:00AM	TOP GUN

FIGURE 1

Since Request Television has ten showings a day (Fig. 1) of as many as four different movies, it is imperative that our affiliates, or for that matter any cable operator doing PPV, be able to bill their subscribers individually for each and every showing of a movie or event that a subscriber watches.

The challenge was clear: Simply change the tags before each event. The solution, however, was a little more difficult.

Before the satellite delivered tag switching system was developed, some eager operators positioned an employee in the headend to manually change their tags for each event. But it was costly keeping employees on, day and night, to manually switch the tag levels. Showing less than a full schedule of movies was not the solution either because that resulted in fewer movies sold to subscribers.

Other operators preset different encoders to different tag levels and then switched them in line with a clock-controlled video switcher for each event. But that wasn't cost-efficient either. Each encoder cost up to $2,000. There was the cost of the video switcher as well; and in the end, this approach too, was limited to only a few events.

A third route a system could take was to update to a new controller. However, it is hard to justify a $40,000 price tag just to bring PPV into the market when the older controller is already doing every other aspect of its job.

We found a better solution for all these operators. The answer was a separate system to do the switching; one that would be completely transparent to the cable operation.

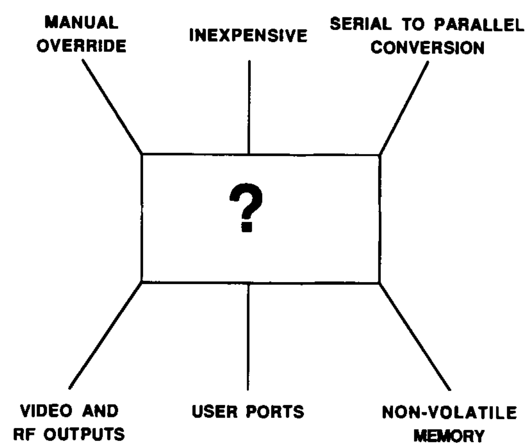


FIGURE 2

What components would be needed? (Fig. 2) The first requirement would be an intelligent device in each cable headend. The device would need the ability to accept a serial data string and convert it to a parallel output. It would also need a non-volatile memory and be inexpensive and reliable. A logical solution was the Commodore 64 Computer.

The Commodore 64 Computer comes equipped with the conversions that are required, plus an array of user ports. It's inexpensive, reliable, and best of all, should it fail, a quick trip to a local department store and a check for less than $200 would put the system back into operation.

With the aid of the game cartridge port on the computer, a prom could be programmed with all the information and look-up charts needed as well as a self-boot program for outage problems, all without fear of accidental erasure.

A means of communicating with each site would be needed. Telephone lines would be too costly, and a subcarrier on the satellite feed would not only be an additional expense but would take time to implement since each receive site would need a demodulator.

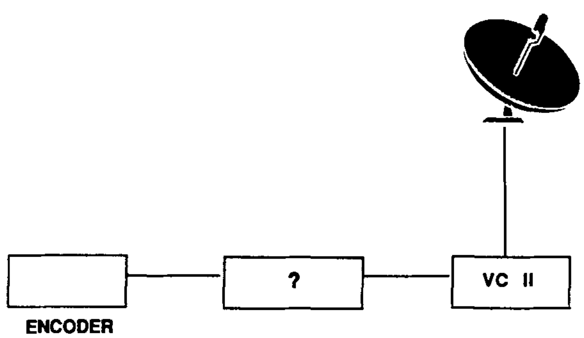


FIGURE 3

The videocipher scrambling system (Fig. 3), equipped with its data channel, provided an inexpensive path to each and every headend.

After testing, the data channel proved to be transparent under many adverse conditions, such as a high-noise ratio and terrestrial interference. A simple RS 422 to TTL converter was built and communication with the receive site was established.

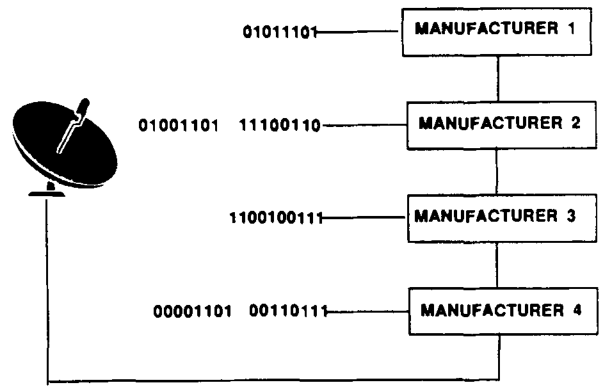


FIGURE 4

The Satellite Delivered Tag Change System was being designed to be compatible with encoders of different manufacturers (Fig. 4), and a simple means of communication was needed. Since each encoder uses a different means of changing tags, it would be impossible to send the individual set-up codes to each headend with 100 percent reliability. Instead, all of the set-up information was stored within the receive site program. In this way, only a generic signal (Fig. 5) needed to be sent. The generic signal would first identify a receive site and then instruct it to execute one stored tag.

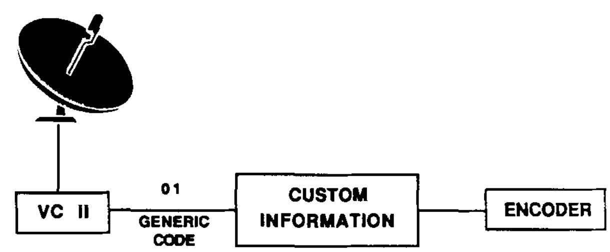


FIGURE 5

The set-up information stored represents the binary code for each tag and is arranged as a look-up table (Fig. 6). The look-up table is arranged in order of use, with the first code being assigned to the first event, regardless of time of day.

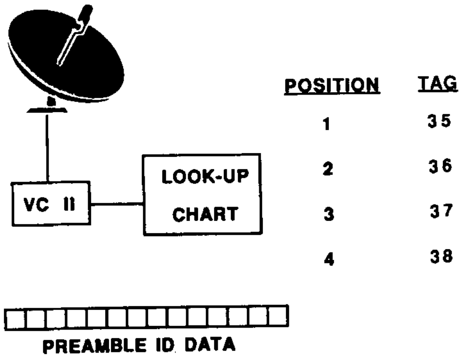


FIGURE 6

With this arrangement the host computer would need only send four sets of data: First, a preamble so the receive site would recognize this as incoming data; second, ID codes so that each system would recognize particular commands; third, a set-up command; and the fourth and final step would have all systems execute the set-up commands simultaneously. This final execution command, being separate of the set-up procedure, will allow the universe of sites to grow and still have them execute each command simultaneously. This also will allow for expansion into other controlling areas.

The functional procedure is to send the set-up code two minutes prior to the top of the hour, and to send the execution code at the top of the hour prior to the start of an event.

With this all firmly in place, a beta test was conducted. A Torrance, California system was chosen as the test site. On December 15, 1986, the equipment was installed and placed on-line.

Some shortcomings in the program were discovered which would not allow it to send the commands on time. The program was revised and a re-send of the last data option was added.

Another problem that appeared at the receive site was the accumulation of noise. The computer would collect noise as though it were data and store it until the buffer was full and then bomb out. The receive site program was revised so it would ignore all but recognizable data.

With these revisions in place, the Satellite Delivered Tag Change System was complete.

CONCLUSION

The Satellite Delivered Tag Change System allows a cable operator to enter the PPV business, operationally, without huge hardware upgrades and without additional manpower. It allows him to carry a full schedule of movies each day and to bill his subscribers according to which event they watch.

And perhaps most important of all, the Satellite Delivered Tag Change System operates independently and is transparent to the cable operator.

ACKNOWLEDGEMENTS

Many people have contributed to this project since its inception. The author wishes to acknowledge the contributions of Mr. James Schmeiser, Mr. David Rodriguez, Mr. Paul Swedberg, and the cable operators who have worked with us with a spirit of cooperation.

Security Considerations for Impulse Pay-per-View Systems

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ABSTRACT

Renewed interest in Pay-per-View applications has produced a variety of technologies for implementing Impulse Pay-per-View. Among these technologies is use of store and forward methods with addressable home terminals. There are obvious advantages to store and forward because it allows instant self-authorization of the PPV subscriber addressable home terminal, while permitting slow data communication and retrieval of the PPV buy data, thus avoiding some transaction handling problems with conventional system controllers and billing systems.

INTRODUCTION

The obvious advantages of store and forward approaches also create the opportunity for serious compromises of the scrambling and security of PPV addressable home terminals because:

- addressable home terminal is self authorizing
- valuable transaction data is stored on premise for long periods of time
- communication links between the storage element and the system headend could be vulnerable

To maintain the inherent security of the addressable home terminal under this environment it is necessary to have:

- true dynamic scrambling with constant and reliable data paths
- dynamic encryption of downstream communication
- secure non-volatile storage that is:
  1. unreadable
  2. unalterable
  3. inaccesable
- dynamic encryption of upstream communication

A description of an addressable home terminal system utilizing these criteria will be given.

SECURITY

Each succeeding generation of CATV technology has increased the level of security available to protect cable services from unauthorized reception. The parade of technology goes from mid-band tuning to sync suppression and addressability. As the perceived value of cable service has increased, so has the sophistication of the pirates. The weakest link of each new technology is discovered and exploited. This attack can be very sophisticated or just the result of brute "megatinkering" hours.

Early non-addressable decoders were soon victims of cloned or counterfeit PROM's. The advent of addressable decoders using sync suppression merely changed the battlefield. Sophisticated systems were compromised by a variety of means. More often than not, the box rather than the "system", was the focus of attack. Electronic sophistication fell to physical attack, filling the pirate pipeline with tales of "blue" and "orange" wires. However, even designers learn and the weakest link is less and less obvious. The latest generation of decoders, which have true dynamic scrambling, requiring a constant stream of data from the headend to operate properly, have not been immune from attack.

Impulse pay-per-view is emerging as a new revenue source for cable. The perceived value of this service makes it a target for security compromise. Since some of the PPV technologies utilize self-authorizing decoders, consideration of the entire system security is required if history is not to repeat itself.

History

The basic function of security is to deny unauthorized reception by the subscriber of certain program material. Authorization assumes payment by the subscriber and revenue for the cable operator and programmer. If we consider the history of cable, we find various approaches to security were used.

Early cable operators denied unauthorized reception by translating cable channels to a portion of the spectrum, midband and superband, not tuned by television receivers. A converter was required to translate those channels to frequencies tuned by the television receiver tuner. The introduction of cable compatible television tuning systems made another approach necessary.

The next step in security was the use of traps mounted between the pole and drop to alter the signal. The trap, positive or negative, had to be physically present or removed for proper reception and operation of the television receiver. Traps aged, drifted and in some instances, aided by subscribers or entrepreneural technicians, they ceased to function. Traps, now being rediscovered as "consumer friendly", became burdensome as pay services proliferated and subscriber churn kept trucks rolling.

A more sophisticated approach to security was sync suppression. At first with non-addressable converters, and later with addressable converters. Authorization was provided by a PROM, which could be reconfigured easily. A box changeout was still necessary until addressability came in. The advent of addressability allowed the subscriber's in-home terminal to have authorization levels changed electronically from the headend without the necessity of a truck roll and box changeout.

Sophisticated as they have been, addressable systems have been defeated, most often by attack upon the physical converter itself, and not the "system". A common problem has been that the signal can be interrogated to derive the scrambling parameters. An example is synch information on the aural subcarrier. On the premise that both video and audio need be present, recent approaches have left the video scrambling "soft" with sync suppression, and audio encoded digitally in what are considered "hard" forms of scrambling. This approach offers more security, but at increased cost.

PAY-PER-VIEW

The cable industry has seen the perceived value of the traditional pay services decline. Along with that decline has come a virtual cessation in subscriber pay growth. This phenomena has been attributed to external forces, among them is the rapid growth of VCR ownership and competition from VCR cassettes. VCR cassettes offer VCR owners, convenience, wide selection and choice, and product with earler release windows. This is an enormous market, measured in billions of dollars.

Cable has the ability to share in the revenues of this market through pay-per-view. Addressable technology offers subscribers the opportunity to purchase a single program. Early release of program material, time and date with VCR cassette rental and sale release, adds perceived value to the programming on cable. Best of all, the subscriber need not leave his home to enjoy the event. The ability to enjoy a single event requires that:

1. the subscriber's addressable converter be authorized and deauthroized in a timely manner to coincide with the event
2. the subscriber transaction data be captured by the billing computer for subsequent payment
3. the transaction constraints allow impulse purchase of the event

Satisfying these requirements is possible through a variety of technologies. The return path for the transaction data to reach the billing computer can be either the cable system itself using two-way communication; or the ubiquitous telephone system. Either path technology, if accomplished in real-time, must also deal with dynamic peak loads caused by impulse purchases near the event start.

One of the problems in dealing with real-time is that the very act of authorizing a subscriber can require substantial time:

- transport of the subscriber transaction to the billing computer
- dynamic updating of the billing computer subscriber database record
- transmission to the headend site the signal to the system controller to authorize the subscriber

STORE AND FORWARD

One way of avoiding the technical problems is to:

- allow the addressable home terminal to self-authorize prior to the event
- deauthorize after the event
- collect the transaction data at the addressable home terminal
- system controller database update and encoder command to authorize

These times can be measured in seconds to minutes. That magnitude of time makes impulse situations impossible with conventional technologies. Also, the voice telephone network is not very receptive to peak loads possible under these situations.

Real-time operation requires new approaches to billing technology and system controller design. Successful technologies exist for handling dynamic impulse loads in real-time: either two-way contention systems or ANI passing telephone systems.

- transfer the transaction data from the home to the system headend or billing computer at a time and data rate to avoid the peak loads on the return path or the billing computer

This approach, using an autodialer and telephone return path is the conventional approach to store and forward. Another approach uses the cable plant as the upstream return path, with an RF transmitter instead of the telephone. An advantage of the RF system is that the transmission rate for the upstream data can be faster, allowing more frequent polling.

Store and forward technology is very attractive. It allows the customer to buy programming or events on a true impulse basis, buying the event directly from the addressable home terminal. The stored information is then retrieved by polling the system and recovering it at a very slow rate.

Store and Forward Security

The home is an extremely hostile environment. The realization that value, in the stored transaction data, resides within the addressable home terminal will result in attempts to circumvent system security. To maintain a secure system, the integrity of the stored data must be maintained under adverse conditions. In addition, the decoder must self-authorize only under certain controlled conditions.

Data Integrity

Transaction data is stored in electronic memory, which must be secure. There must be no external access to this memory, and there must not be any way to alter or otherwise subvert the writing or reading of this data.

Encoding data and then placing it in a discrete external EPROM is to court disaster. Despite the mathematical odds against "breaking" the code or algorithm, we must remember that there will be thousands of boxes out there collecting thousands of "megatinkering" hours of assault by amateurs and professionals. The law of large numbers is against you no matter how infinitesmal the odds are.

In the system under consideration the data is stored in non-volatile memory within the controller IC. The controller is a proprietary application specific VLSI, and not available as a standard component. Placing the memory within the controller IC, requiring that essential dynamic data also be stored there, and that access is possible only with encrypted code, increases the security of the data. Only properly encrypted commands, which are in the in-band RF signal, can alter or read the data, making unauthorized attempts to alter the data unlikely.

Data recovery is achieved by polling the addressable converter and transmitting a message to the headend. The return path can be RF, via the cable return path, or the telephone network. This message is dynamically encrypted. Each polling of the box uses a different encryption key, generating a different message, even with the same data stored. This is absolutely necessary for a telephone return path where messages can be easily recorded and played back repeatedly for analysis. The downstream polling message must also be encrypted to maintain the security of the upstream encryption key.

Multi-Level Decoder Control

Positive control of the addressable converter requires that constant data is needed to properly decode the dynamic scrambling. This program related data is encrypted using session keys which are changed periodically. Each authorized addressable converter is individually given a new session key periodically. This provides positive control over stolen, unknown and non-pay addressable converters. Also, non-responding addressable converters where the data channel may have been subverted. The change of session key causes all of these addressable converters to stop decoding.

PPV Authorization Control

There are two levels of authorization possible in the system under consideration. There is one configuration for normal tiering, such as basic or a premium pay. For IPPV, the addressable converter and program must carry a special IPPV authorization level. Thus for proper IPPV authorization the addressable converter requires:

- IPPV program tag
- IPPV authorization
- proper session key
- correct data reception

Exceeding credit limits may deny a subscriber further access to IPPV service, but still provide normal pay services.

Additional Security Considerations

There are advantages to developing a proprietary application specific VLSI, not the least being reliability and cost reduction because of parts count reduction. The more concentrated the functions allocated to the VLSI, the higher the security of the system. The more there is in the VLSI, the fewer "blue" and "orange" wires are available. The VLSI incorporating these criteria is shown in Figure 1. The non-volatile memory is in the lower right hand corner of the chip.

The controller IC was designed to perform the following functions:

- data reception
- decryption
- decoding control
- upstream data generation
- dynamic encryption
- non-volatile memory
- decoder ID

The concentration of all these functions in one IC leaves no "hooks" to grab. The ID or serial number of the decoder is inserted in non-volatile memory at the time of addressable converter manufacture, at which time internal device gates are opened, thereby denying access to the ID resident portion of memory. As a consequence the ID is unchangeable and cannot be read or altered. Similarly, the algorithms for encryption and decryption are located in inaccessable portions of the chip. User accessable portions of the IC exist, providing functions such as tuning system control, IR remote control and favorite channel scan memory.

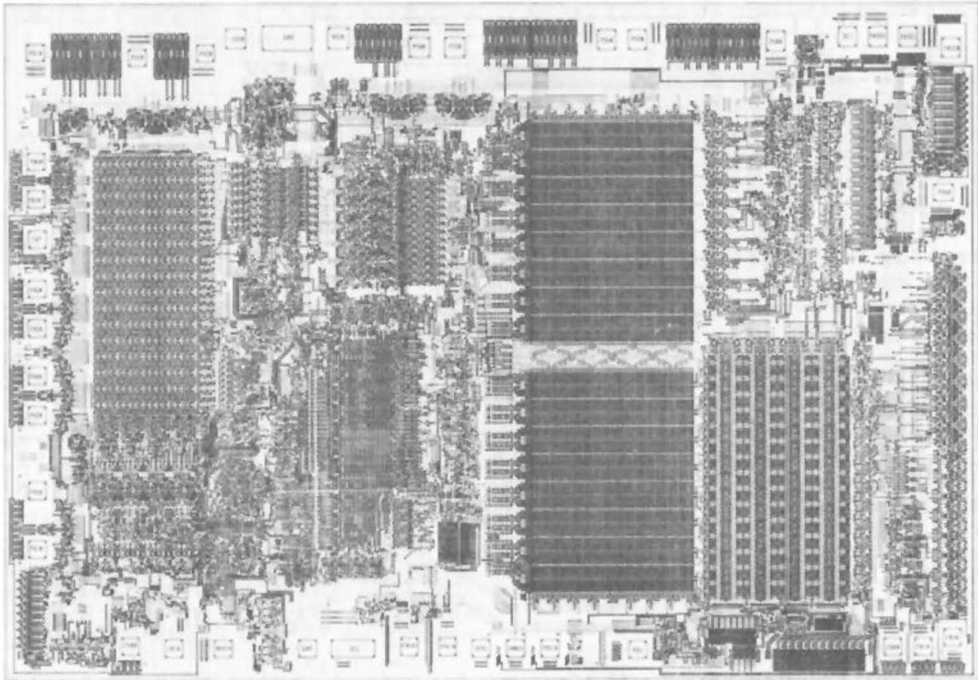

Figure 1. Custom CATV Microtuner with NV RAM

IMPLEMENTATION

An addressable converter using these security criteria for store and forward IPPV applications has been implemented. The addressable converter, by means of an external serial buss, is capable of store and forward IPPV applications, using either the telephone or upstream RF as the return path. The RF implementation, PM-Pulse, has a self contained RF transmitter. Polling occurs at up to 100,000 converters per hour. Polling at such high rates allows a typical system to be polled several times during an event, thereby assuring capture of the event data prior to completion of the programming material. Such fast polling means there is likely to be no data of value in the addressable converter. Verification of the subscriber transaction occurs because of time stamping of data at the headend.

The system polls constantly, and each upstream response from a converter supplies the following information:

- box status
- channel tuned
- authorization bit map.

Thus the box is constantly being interrogated and any unauthorized changes can be quickly noted.

CONCLUSION

Store and forward IPPV systems can be made very secure if the system architecture and hardware implementation are carefully executed. The essentials are:

- dynamic encryption
- dynamic scrambling
- constant and reliable data paths
- secure non-volatile storage
- concentration of functions in proprietary VLSI

The resulting implementation is not only secure, but capable of multiple IPPV approaches. The parts count reduction results in an extremely cost effective and reliable product. A version of this product, without tuner, provides a decoder only function using an existing "plain-jane" converter. Using ANI passing for IPPV provides an extremely low cost approach to PPV.

SELECTIVE ELECTRONIC HOME SHOPPING

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ABSTRACT

Given the extraordinary financial success recently enjoyed by televised home shopping, the viability of the electronic medium as a vehicle for the display of merchandise intended for sale now seems assured.

To date, however, the philosophy of network shopping has been to rely on conventional television distribution methods rather than to use cable technology to advantage. This has resulted in two (2) significant shortcomings:

1) Viewer satisfaction is limited to that very narrow demographic group with both the time and inclination to endure a serial product display, the viewer remaining poised to interact on a "target of opportunity" basis, and

2) Cable's unique technology is nowhere apparent. The viewer therefore perceives no enhanced value to cable TV subscribership.

OVERVIEW

This paper will describe a cable-unique, interactive, electronic home shopping service offering the subscriber full random access, for view or purchase, to potentially over 50,000 products, each displayed in full NTSC video/audio over a 6 MHZ CATV channel.

No home terminal devices beyond basic CATV and a touchtone telephone are required.

EHS Concept

Electronic home shopping (EHS) as a subscriber controlled program service can best be visualized if viewed from **Telaction's** perspective. That is, the television receiver becomes the visual equivalent of a shopping mall. Replete with the full functionality of a shopping mall, i.e., random access to a cross section of stores. The ability to enter those stores and examine a cross section of products again, through random access. And finally, to optionally purchase products, store products for future purchase or simply examine products at leisure without the necessity of visiting a conventional brick and mortar store, free of the intervention of a sales person. When and if human intervention is desired the shopper has at his/her command a simple bridge to customer service representatives from any of the stores participating in the "electronic mall".

All of the capability described above is accessible through simple, usually single, touchtone strokes to cable subscribers at systems affiliated with **Telaction**. The system - a three year, forty million dollar development effort-exists in hardware and is poised for technical test followed by market introduction in the Chicago area during the next several months.

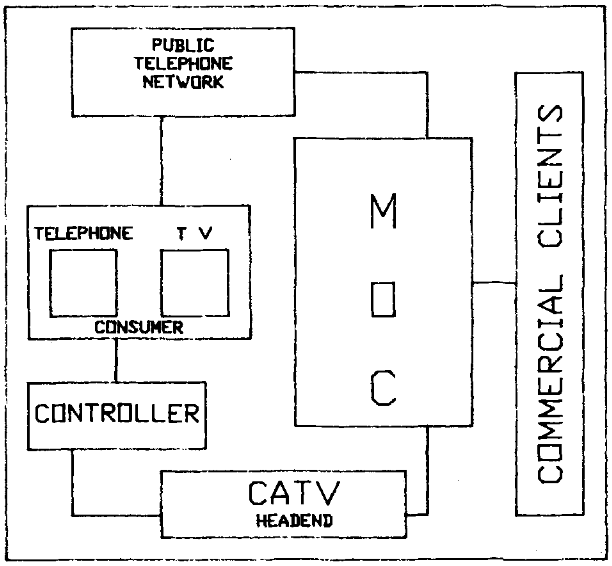
Technology nothwithstanding, the implications of so powerful a service are, of course, enormous.

Four "C's"

Operationally, the system is a hybrid composed of four (4) interactive entities; what **Telaction** refers to as the mandatory "four C's".

1. Consumer
2. Cablesystem
3. Communications
4. Clients

Amalgamation of these components in even a simplest form of life, single node net requires a hardware intensive, software driven co-operative interconnection between private residence, public utility, public switch, common carrier, cable system, and Tel**action** operating center.



(Figure 1)

Prior to describing system architecture however, its fundamental precepts must be understood. Only then, from a perspective of functionality, can the extensive hardware complement be rationalized.

Conceptually, the system will operate as follows:

The cable subscriber, tuning to the Tel**action** channel, will observe a "welcome screen". The screen, a still frame graphic, will advise him in lower one-third (1/3) script ... "to begin shopping dial the phone number".

Following these instructions the consumer removes the phone from its cradle, dials as instructed and within two (2) seconds of closure a menu will appear on screen with its associated prompt. This menu will consist of product categories. Following the prompt, now reduced to single touchtone keystrokes, the consumer may "navigate" the catalogue inventory of some thirty national and international stores. Each displayed in full NTSC video with aural accompaniment. This activity may now continue until such time as the consumer chooses to terminate the interaction by purchase, storage, customer service, feed forward, or simply hanging-up the telephone.

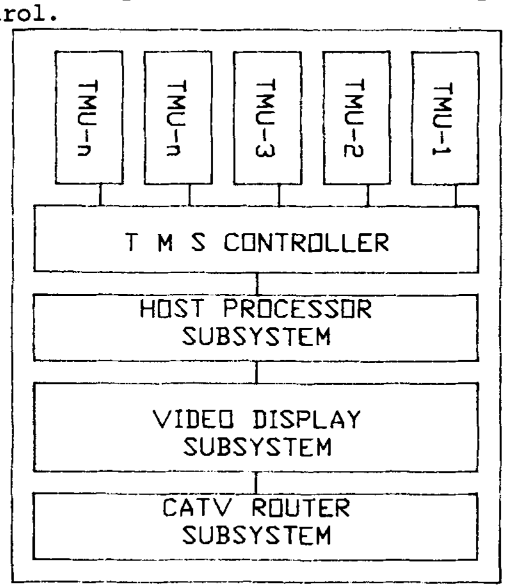
Simply stated, the viewer need not be subjected to the rigors of computer operation or any sort of intrusive hardware. Interaction is via telephone and television. Specifically cable television.

ARCHITECTURE

Metropolitan Operating Center

This level of interactivity, albeit with a far narrower range of products and services, has been achieved to date only through use of home personal computer or point of purchase (mall) kiosk devices. In order to deliver such functionality while limiting user hardware requirements to CATV and phone, transaction processing and control must be emulated externally. This is accomplished on a regional scale through a facility known as a Tel**action** Metropolitan Operating Center.

When an interactivity is initiated i.e., a subscriber dials into the Tel**action** Network via the 800 or local toll the call will be routed to the most local metropolitan operating center. Incoming traffic will process via ATT conversant telephone management system, (TMS). The conversant telephone management system is characterized by voice recognition. Voice grade circuits will direct incoming calls at the Chicago metropolitan operating center from the telephone management system, to a host computer system for data entry and control.



(Figure 2 (MOC)

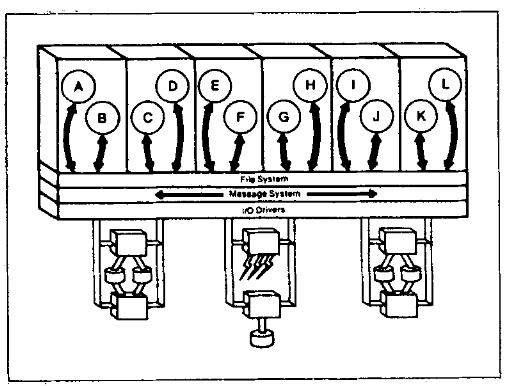
The host computer systems, in the interest of reliability, are Tandem Systems TXP and VLX multiple processor mainframes, characterized by full redundant hardware running the Tandem non-stop operating system.

While essentially transaction process computers the VLX and TXP systems are sufficiently fast to respond to the level of activity generated by electronic home shopping. Of greatest importance however, is the fundamental architecture of Tandem systems - complete hardware duplicity, each computer is a dual system, each running duplicate programs and interconnected in a failure deferral hierarchy.

Simply stated, should a host computer experience a catastrophic failure in hardware, the operating system ("non-stop") will sense the failure, shift output to the operable duplicate and continue running valid data, virtually undisturbed.

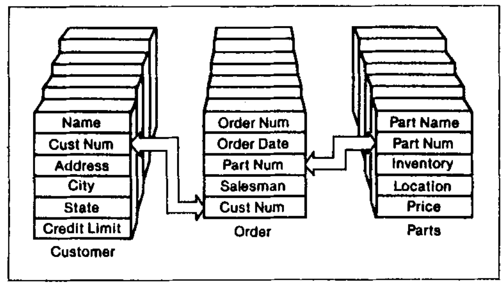
In addition to reliability, the Tandem system architecture is eminently expandable. Following market introduction the Tel**action** Network is expected to expand into sixty regions nationwide. This, according to a rather aggressive schedule will place extraordinary demands on MOC data processing facilities.

A system architecture characterized by its inherent ability to balance processor loading (Figure 3) is essential to orderly growth of the network. The Tandem system is capable of "bolt on" expansion to accommodate 4,000 processors.



(Figure 3)

Additionally, a relational rather than hierarchical data base foundation will allow those changes to existing programs inherent to developmental projects. As the network grows a rigid hierarchical construct would become increasingly resistant to change and too easily obsolete.



(Figure 4)

Following data base query as to the validity and identification of the incoming transaction, the host computer outputs a command to the video display subsystem.

The video display subsystem consists of a video display unit, (VDU), an audio distribution unit, (ADU), intelligent controllers; VBI, address inserter, and a nxl matrix switch, interconnected via high speed LAN.

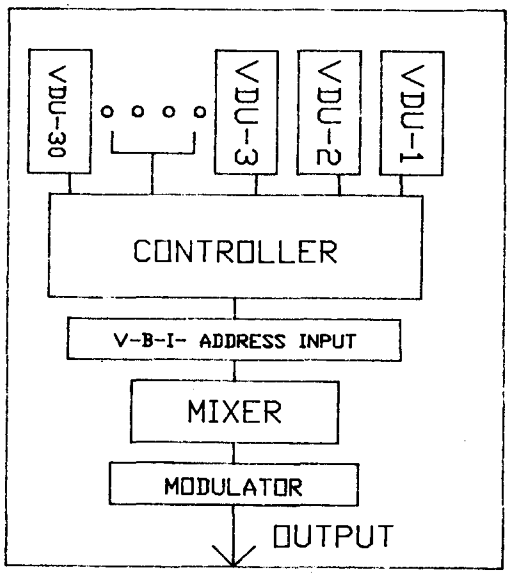
Video access is random read only as well as multiple write to accommodate dynamic data such as graphics or product change.

The entire EHS system is characterized by the output of the VDS.

The output, a series of television frames, are electronically conventional NTSC, 30 frames per second.

The frames, however, are concatenated i.e.: each is an individually fetched and addressed slide, bearing no continuity to the previous or subsequent frame.

This system of distribution allows for a very large transfer of information without stressing CATV system bandwidth or linearity.



(Figure 5)

Additional detail relating to the video display subsystem is confidential at the present time, however, it is apparent that in order to provide program audio a time compression technique is employed. The audio, once recovered, processed and addressed, is transmitted in synchronous parallel to its associated serial video frame. Audio accompaniment is limited to 40 seconds per frame. In practice considerably less time has been required.

CATV Router Subsystem

The cable television (CATV) router subsystem receives audio and video signals from the video display subsystem (VDS) and converts them into signals which can be transmitted to the cable television plant. Under control of a host software subsystem (router controller), video images from the VDS are concatenated at 30 frames per second, a frame grabber address and an audio RF reception frequency are inserted into the vertical blanking interval (VBI) associated with each frame, the video signal is modulated onto a given channel, the audio signals are converted to a given bandwidth, and the resultant audio and video signals are combined and sent to the cable television head end.

Components include a micro-computer controller (scheduler), a VBI switch, a VBI address inserter, a modulator (for the video signal), a block upconvertor (for audio), and a combiner.

The resulting combined video and audio signals are transmitted from the regional MOC system to a cable television head end facility via coaxial cable, microwave, or other wideband transmission facility.

NETWORK

A network comprised of six geographic regions is planned. The time period for completion covering 1987 through 1996, with the major activity in decline by late 1991. Network hierarchy will be as follows:

LOCAL M.O.C.
REGIONAL M.O.C.
HEADQUARTER M.O.C.

Local M.O.C.

Local M.O.C.'s would house a complete presentation system as well as CATV routing subsystems sufficient to serve their respective markets. These subsystems would comprise microwave and copper or fibre interconnects into local CATV systems.

Additional communication facilities would be necessary to provide:

a. Gateway or data base dynamics between local commercial clients and Telaction. These transactions could be relegated to voice grade Telco facilities between the Communicating entities.

b. Transaction data base dynamics between local and regional M.O.C.'s.

These transactions could be carried out on a dedicated basis via T-1 carrier.

c. Presentation data base dynamics from Headquarter M.O.C., all video presentation system modifications would be input at the local M.O.C., via T-1 carrier. These might include pricing, text or full video updates or future VDS subsystem.

d. Customer service bridge of all customer service inquiries. These would be bridged to the regional M.O.C., via dedicated voice grade (56KBPS) circuits.

e. System diagnostics, this data would be communicated to the regional M.O.C., via a dedicated T-1 "order wire" circuit. A complimentary command channel would return from the associated regional or Control Headquarter M.O.C.

Regional M.O.C.

These facilities would include all the facilities of a local M.O.C., as well as:

a. Communication facilities for transmission to Headquarters of data aggregated from local and regional transactions.

b. Diagnostic duplex channel to/from local M.O.C.'s.

c. Customer Service Telephone management system and voice communications to Headquarters.

Headquarter M.O.C.

This facility will include all of the subsystems of a local M.O.C., as well as:

a. Customer Service TMS.

b. Complete aggregation of data base inputs from local from local through regional transactions.

c. Communication links to/from commercial clients.

Communications network subsystems sufficient to support this level of information transfer would, of necessity require:

1. Interfacility communications _.e., between M.O.C.'s including data voice, and video carried out via a satellite multiplex single channel per carrier (SCPC) scheme.

Duplex R/T facilities at each communicating entity (Trans, Rec. Ant. Modems, etc.).

Protected (non-pre-emptible) satellite service across at least a single 36 HZ transponder.

2. CATV communications links-microwave, fibre and copper interconnects between M.O.C.'s and potentially 5,000 CATV systems.

3. Commercial client data communications in the form of switched telco service from local entities to dedicated T-1 SCPC over Tel**action's** transponder for national accounts.

CATV Distribution

CATV distribution closes the loop from consumer telephone to consumer display.

The incoming audio and video signals are received and demodulated at the CATV headend facility. Re-modulation for distribution is via Tel**action** supplied and modified CATV channel modulators.

Video modulation is conventional NTSC in signal characteristic, however, of concatenated frame content and devoid of aural subcarrier.

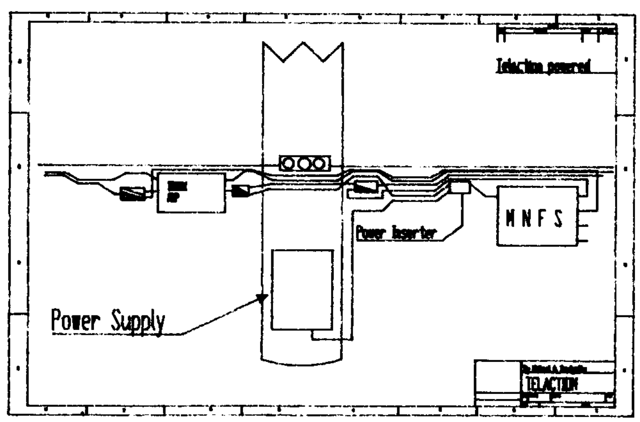
Audio distribution is via single channel per carrier, multiplexed into the CATV trunk, and may be carried on any trunk frequency irrespective of converter capability.

Since trunk distribution will route the signals through the CATV plant, selectivity ana conversion to subscriber usable format must take place externally.

These functions are accomplished in an original engineering device known as the multi-node frame store unit (FSU), designed and built by Cableshare, Canada.

Frame Store Units

Frame store units are designed for installation at off premise aerial or underground locations generally co-located with similar population to existing bridger amplifiers.



(Figure 6)

In its simplest form the frame store system consists of a combination of circuit components that 1) recognizes a video frame addressed to it, stores that frame for subsequent transmission as a video signal to the subscriber addresses it serves, and 2) receives and transmits the audio signal related to the video image being transmitted. The circuit components include a dual audio receiver, a video demodulator, an analog/digital converter, a digital frame store, a digital/analog converter, a channel modulator, and a controller which reads and interprets the addressing information in the incoming vertical blanking interval and activates the other components as required.

In its multi-node configuration the MNFS system (Figure 7), will accept an input signal, comprised of NTSC video and frequency divided multiple access (FDMA) audio.

By virtue of address the system will detect, A/D convert, discriminate, store, D/A convert, remodulate and amplify up to four specific video and associated audio "frames". Stored signals will appear simultaneously at up to four parallel feeder maker outputs of the MNFS.

Output video levels are continuously variable over a ± 20 db range, up to a maximum of +59 dbmv continuous.

In its four node configuration, the functions described comprise a component allotment of thirteen single side printed circuit boards housed in thirteen plug in modules plus transformer less power supply, line filter and receive I/O modules. The entire assembly resides in a 26" x 10 1/2" x 12" cast housing. The external appearance being similar to that of a conventional feed forward CATV trunk amplifier and should exhibit similar environmental immunity and RF radiation properties.

When a subscriber initiates a transaction call to the M.O.C., the requested frames are routed to and through the CATV trunk system. All frames appear at the input ports of all FSU's, and are selected by address for display by the FSU associated with the initiating subscriber.

Audio is routed via frequency division multiple access technique throughout the trunk system as well. The appropriate carriers are selected for feeder distribution by address and modulated on subcarrier 4.5 MHZ above video by the targeted FSU.

Simply stated, in the CATV trunk line a channel of concatenated frames as well as narrow channel of FDMA audio are routed. At the feeder level, those frames (A&V) requested by homes served by each feeder are selected by the FSU and routed only to that feeder.

Each interaction taken by the subscriber results in an additional VDS subsystem output. Each will be discriminated by that subscribers FSU for display. Tel**action's** design target for response time is two (2) seconds, from key stroke to video display at the drop.

FSU's outputs are phase locked to the input carrier and will thus introduce no instability impact as regards harmonically or incrementally related carrier systems.

Each FSU is capable of providing up to four distinct output frequencies.

Contention

Contention is a major design concern in any technology where numerous homes passed vie for one channel.

Tel**action's** design target is 86% availability during peak use and virtually unrestricted availability during low use periods. This level of contention is rationalized as follows:

IF:

$$A \times B \times C \times D = E$$

WHERE:

A = Feeder population in homes passed (national average)

B = CATV system penetration in percent basic subscribers

C = Tel**action** subscriber penetration in percent of basic subscribers

D = Probable concurrent user percent system wide

E = Contention ratio

THEN:

100 X 50% X 33% X 0.5% = .08:1

CATV Implications

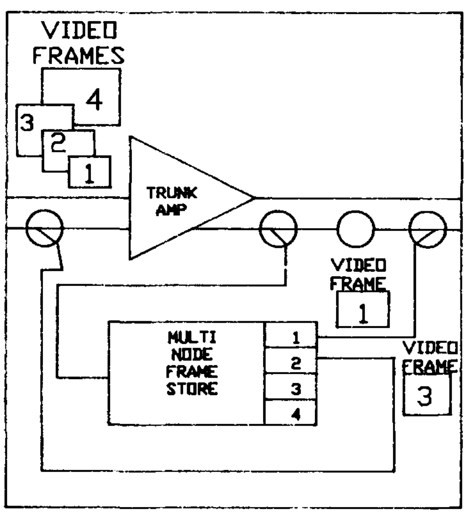
The EHS is the first service in which a series of devices need actually be installed in off premise physical plant. Consequently, the ramifications are extensive.

It is Telaction's stated intent to minimize the impact to the financial operational and signal quality performance of affiliated systems during EHS installation. To that end Telaction will assume all cost associated with installation of EHS hardware. At each affiliates option, installation will be carried out by contract organizations. Project management will be the largest CATV hardware producers to insure system integrity is maintained throughout this critical phase.

At each affiliates option all components of the EHS system residing on CATV plant are subjected to rigorous performance testing and verification on site.

Alignment and proof of performance will be carried out by Telaction prior to launch.

All subassemblies of CATV resident equipment are designed to be field maintainable. Telaction will provide on site spares to an adequate level based upon projected MTBF. Failed unit repair depots are to be located at each regional M.O.C. It is important to note that each component of the Telaction system is designed to limit failed effects to the Telaction channel through both active and passive isolation.



(Figure 7)

CONCLUSION

The Chicago market introduction of the electronic home shopping concept is intended to exercise this dramatic new application of the CATV medium.

The extensive software and hardware complement, the vast majority of which is original design, will not be without fault, therefore, Telaction intends to exhaust every resource in a pre-market series of technical tests to minimize those contingencies, while improving the fundamental system for future expansion. However, as with each pioneering endeavor; earth stations, 400 MHZ, addressability and so on, the quality of service has driven the CATV industry to endure.

Some 150 new program offerings have appeared in the CATV arena during the most recent 10 years. Of these only some 40 survive. Those few of enduring quality remain as valuable, cable apparent offerings, HBO, CNN, Discovery, etc.

And while the CATV industry has long since passed into the mainstream of establishment American industry that same innovative spirit that accepted such drastic departures from convention as HBO's, 1970's earth station concept is apparent today, and will reflect in the bottom line.

Telaction intends to be the next cable unique, program innovation.

STEREO TELEVISION AND CABLE : ONE YEAR LATER

Karl W. Poirier

TRIPLE CROWN ELECTRONICS INC.

ABSTRACT

At the 1986 N.C.T.A. Convention, the item on everyone's mind was BTSC Stereo Television. As this was a new development in cable television, much attention was focused on the technology of the standard. In particular, the questions concerned- How does it work? Can it be implemented? How do we test? and, Can the standards be met? These questions and others revolved around the encoder systems, and the generation of the signal. Subcommittees and working groups were formed to discuss issues such as proper alignment, test procedures, and specifications. This year saw much of this effort discontinued, possibly with the impressions that the obstacles had been overcome and little further concern was necessary. In reality the situation is quite different.

INTRODUCTION

The introduction of BTSC stereo generation to CATV systems has, since its inception, been faced with two situations unique to the industry.

(a) The CATV system typically involves more complex signal handling than does the basic broadcast to TV transmission.(Fig.1)

(b) The average CATV operator possesses less test equipment, particularly audio test equipment than does the average broadcaster.

These two situations were the thrust of most of the research and investigation performed up to now. We have addressed, and for the most part, solved the problems of dealing with Baseband Converters, Modulator Bandwidth, level calibration, system degradation and other fairly obvious difficulties.

We now face an entirely new set of challenges, as we attempt to implement television stereo in the average cable television system.

Source Signals

Step one of any implementation is to recover the program material from the delivery medium. This is, for most applications, a satellite delivery system. Up to now, this recovery has been a simple matter of either a single high level subcarrier, or the output of a digital descrambler system. In many cases,this signal has been designed to facilitate MONO recovery and is therefore either a DESCRETE MONO source, or a MATRIXED 'M' signal (L + R). In other cases, the signal is duplicated as a high level mono accompanied by a DESCRETE LEFT and RIGHT low level pair. In rare cases, the signal may be a high level MULTIPLEXED STEREO format. Whatever the format, the operator is probably looking for a source signal which has not been previously employed, and possibly incorrectly identified : (Feeding the L and R inputs of stereo encoder with 'M' and 'D' sources can produce some rather interesting results).

Due to the lack of both stereo programming and stereo awareness, it is practically impossible to distinguish by ear between a DESCRETE RIGHT (or LEFT) and a MATRIXED L + R (M) signal. (The aspect of stereo program context will be further examined later in this paper).

A further complication may arise as program sources which are equipped with stereo encoders today change to encrypted or digital transmission formats at a later date.

Head End Modulators

As it was realised early on that most CATV modulators would not have the baseband modulation capability required for BTSC, most ENCODER manufacturers have provided an internal 4.5 MHz modulator feature. This feature, which allows the encoder manufacturer to provide the properly deviated aural signal appears to solve the problem, however, the implementation is less than simple. Most CATV modulators which accommodate external 4.5 MHz, are designed to accept 4.5 MHz/video combined on a single input port.(Fig.2)

This brings about the somewhat pointless situation of being required to multiplex the 4.5 MHz on to the video, to be fed to the modulator, where it is immediately separated and processed separately! Not only is this a cumbersome situation, but some modulators can develop intermodulation problems when operated in this manner. The ideal situation is a modulator with separate video and 4.5 MHz inputs. This option is currently being made available by most major manufacturers, but those CATV systems employing lower grade and offshore equipment may face a difficult situation.

Terrestrial Encryption Interface

When the addition of stereo is considered, it most often involves a premium service, which is invariably protected for distribution. We are now adding a BTSC encoder not only to a receiver/modulator system, but to a receiver/descrambler/scrambler/modulator system. A particular case in point would involve a baseband scrambling system, and a video/4.5 input modulator. In this situation

- if the video is looped through the encoder before encryption, how is the 4.5 MHz added to the video after encryption but before the modulator.

  but

  if the video is encrypted prior to being fed to the encoder loop/combiner, there will be no sync to allow pilot phase lock and therefore no stereo.

To more clearly illustrate this situation, we can examine one actual case where this problem has arisen. The Rogers Toronto System employs a downtown head end hub system with baseband scrambling. Due to T.I. problems, the C BAND signals are received at the North edge of town and delivered to the hub head end via FML microwave as a baseband to baseband link to allow encryption at the hub site.

The question now becomes:

(a) Place the stereo encoder at the hub head end and attempt to deliver the DESCRETE left and right on separate subcarriers via microwave?

(b) Place the stereo encoder at the remote antenna site and attempt to transport BTSC stereo via separate microwave subcarrier?

(c) Place the stereo encoder and the baseband scrambler at the remote site?(this would involve backhaul of the encryption data).

This, and many similar situations are beginning to become apparent as implementation progresses.

In this particular case, the following method was selected as the most workable solution: The audio is BTSC encoded at the remote site and presented as a video/4.5 combined signal to the FM microwave link. At the downtown head end, the microwave is demodulated to video/4.5 combined, and diplexed to separate video and 4.5 sources. The video is encrypted and delivered to the modulator as separate video and 4.5 signals for modulation to channel.(Fig.4)

Home Installation

It is now quite apparent : only the fully stereo television receiver, with internal drive for auxiliary external speakers will provide a satisfactory solution to the customer. It is probably best, at this point, to quote an actual case example.

As the proud owner of a new large screen combination cable ready TV/Monitor with "Stereo Adapter" feature, I proceeded with utmost confidence to add an auxiliary stereo decoder. Thus armed with a multi-input (baseband, 4.5 MHz, Ch.3) decoder, and a set of auxiliary powered speakers, I undertook the 15 minute task of "just hooking this stuff up" to get stereo television. I had deliberately chosen a decoder which included a synthisizer in order to improve both my cable as well as my baseband VCR and Satellite receiver sound. Several frustrating hours later, I had achieved the following insight.

(a) The "Stereo Adapter" port on my television consisted of an amplified IF SPLITTER providing an IF sample, with a pin diode output switch(possibly a crude form of mute system!)

(b) When switched to video/audio input such as satellite, the "stereo adapter" port continued to deliver the sound from the last tuned cable channel!

(c) The amplified speakers required separate volume adjustment at every program change, which results in the final conclusion being arrived at by myself as well as other interested household members: (i.e. Take that mess down to your workshop and play with it there!)

What does, in fact, become apparent, is that with a few audiophile exceptions, most listeners will probably convert to stereo when it becomes an interesting extra in their new television receiver. Certainly many people will install decoders to be adjusted separately for a program of interest, but general "no second thought" use will only only occur with the true stereo television receiver.

Stereo Programming(or lack of):

The first observation made by a customer with stereo television that of two distinct audio formats being employed by the broadcaster. In one format, programms with little or nothing to offer in the way of stereo but transmitting in stereo: i.e. stereo transmission of a single talking head!

The second format attempts to employ every possible mechanism to exploit the fact that they are in stereo. One popular Friday night program has the soundtrack so overloaded in order to exploit stereo that it is decidedly annoying to watch(hear). (Gone forever the silent pause!).

CONCLUSION

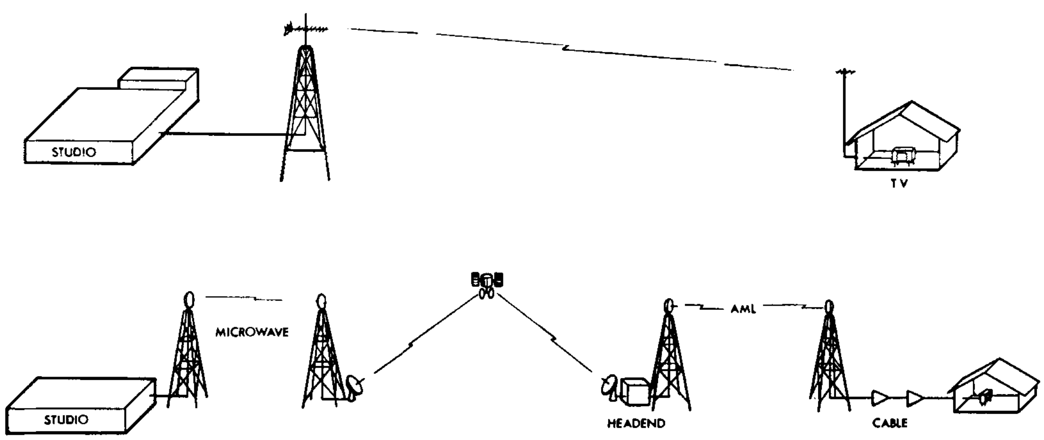
At this stage in the implementation process, it is wise to take stock of several obvious points:

(1) Those people involved with stereo over the last few years: the subcommittee, the manufacturers, the experts, etc. while knowing the field quite well, are not those who will be doing the majority of the implementation. It is our responsibility to provide as much education as possible to the system operator on whom the actual implementation burden will fall.

(2) The first stereo subscribers, the audiophiles, will be the most difficult to please, and this will, of course occur at the lowest point of the implementation learning curve.

(3) The conversion to stereo is not inexpensive but if a question of time vs capital arises, it is best for most operators to wait until they can afford to do it properly rather than to undertake half-way measurers.

(4) The list of surprises is not yet complete.



MODERN LINK COMPLEXITY FIG. 1

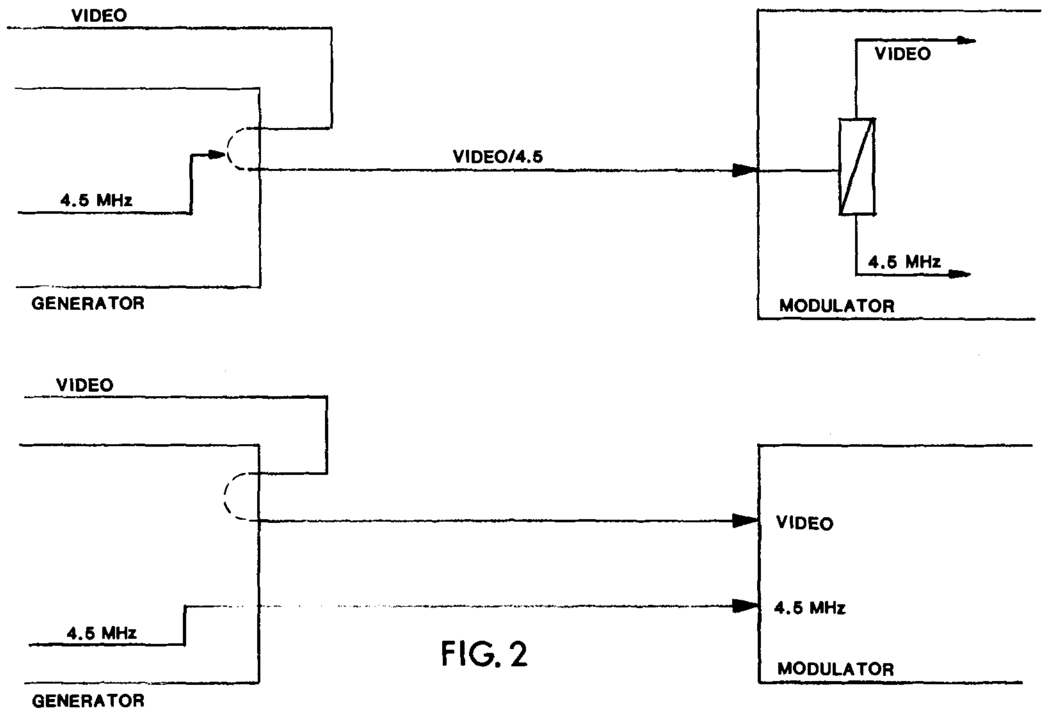

VIDEO
4.5 MHz
GENERATOR
VIDEO/4.5
VIDEO
4.5 MHz
MODULATOR
VIDEO
4.5 MHz
GENERATOR
VIDEO
4.5 MHz
MODULATOR

FIG. 2

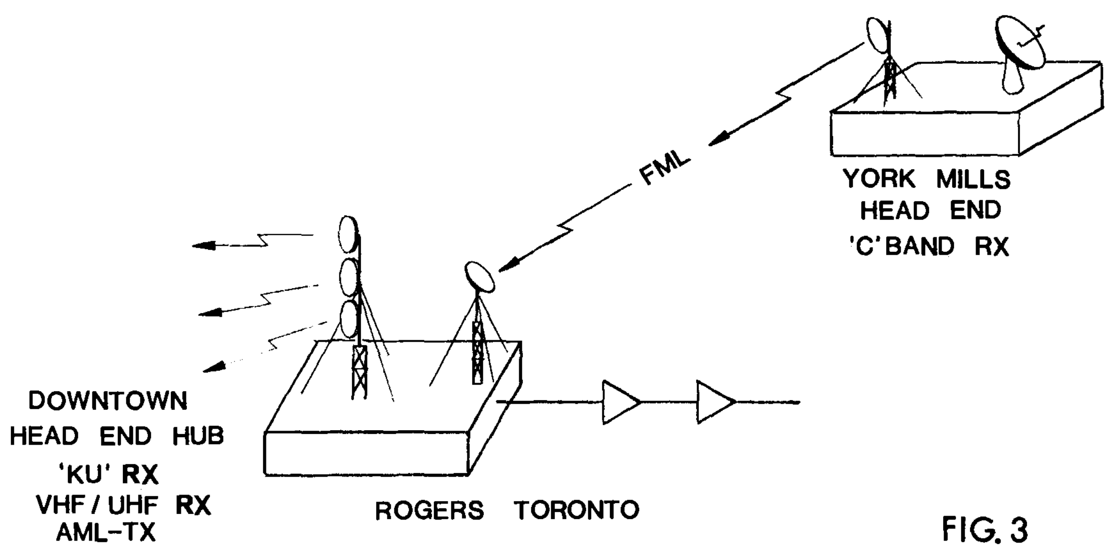

FML
YORK MILLS
HEAD END
'C' BAND RX
DOWNTOWN
HEAD END HUB
'KU' RX
VHF / UHF RX
AML-TX
ROGERS TORONTO

FIG. 3

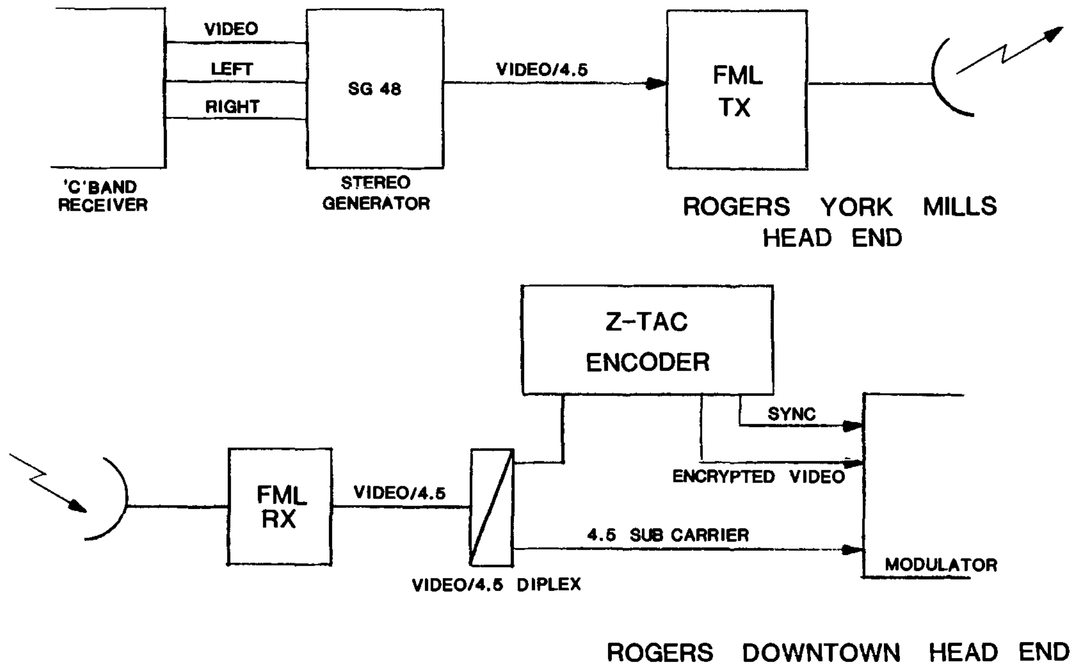

VIDEO
LEFT
RIGHT
SG 48
VIDEO/4.5
FML
TX
'C'BAND
RECEIVER
STEREO
GENERATOR
ROGERS YORK MILLS
HEAD END
Z-TAC
ENCODER
SYNC
ENCRYPTED VIDEO
FML
RX
VIDEO/4.5
4.5 SUB CARRIER
MODULATOR
VIDEO/4.5 DIPLEX
ROGERS DOWNTOWN HEAD END

FIG. 4

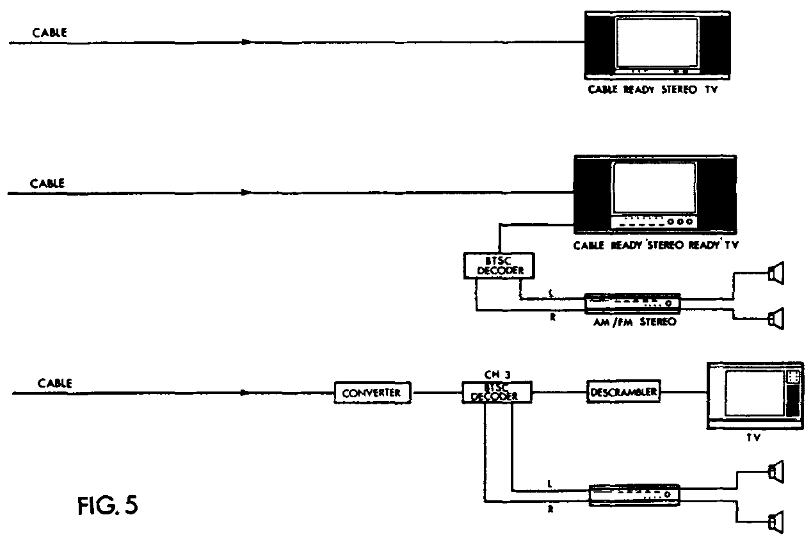

CABLE
CABLE READY STEREO TV
CABLE
CABLE READY 'STEREO READY' TV
BTSC
DECODER
L
R
AM/FM STEREO
CABLE
CONVERTER
CH 3
BTSC
DECODER
DESCRAMBLER
TV
L
R

FIG. 5

AUTHORS' VIEWS

QUESTION: What do you feel is the major unresolved problem with the implementation of BTSC on cable?

KARL POIRIER: Triple Crown Electronics Inc.

"Technical Instruction: This is possibly the most complex and foreign technology ever applied to the average CATV system. It embraces new technology, new equipment, new test procedures and new attitude. It is the responsibility of the 'experts' to pass this knowledge as quickly and effectively as possible."

WILLIAM (BILL) ARNOLD: Warner Cable, Texas

"In the decision to proceed with implementation of BTSC stereo, we anticipated a degree of subscriber education would be necessary; however, many of the subscriber comments led us to a more detailed analysis of the product delivered by the various services. Overall, the stereo product received from the music and premium services is 'acceptable', but our experience has indicated a need for the services to be more aware of the quality of their product; i.e. left/right balance, background 'garbage' and general tonal quality."

DAVID A. SEDACCA: Scientific-Atlanta, Inc.

"Headend operators must become familiar with BTSC stereo in the headend. They must learn the details that dictate how BTSC equipment may connect with their headend; then they must decide which options and signal paths will satisfy their needs for redundancy, switching, and control."

JOSEPH S. VITTORIO: General Instrument

"We feel that the major problem with BTSC today is that the measurement of BTSC signal performance in a cable TV environment has been minimal to non-existent in many cases. We feel this can be attributed to two basic problems:

1. Broadcast test equipment is too expensive for the average cable system to invest in ($35K for demodulation and $5K to $50K for instrumentation).

2. Lack of education and experience. Audio has always been something which came along with the video for free, and was generally ignored as long as it was listenable.

Stereo TV audio, however, requires care in handling and measurement. We encourage the SCTE, in conjunction with the major equipment suppliers, to develop a training program on BTSC signal handling and measurement."

TECHNICAL CONSIDERATIONS WHEN IMPLEMENTING PAY PER VIEW

Larry N. Lehman

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ABSTRACT

Many cable television system operators are currently considering implementing and offering pay per view programming. If they have an existing addressable system that decision seems to be an easy one on the surface. This paper will relate the experience of one such operator, Cencom Cable Associates, and will discuss the many things that were learned through that experience. A list of questions will be developed that need to be asked and answers obtained for before pay per view programming is implemented.

BACKGROUND

Cencom Cable Associates, Inc. is an MSO headquartered in St. Louis County Missouri. CCA was founded in 1982 and by December 1984 had acquired systems in five states with around 40,000 subscribers. The next year however saw CCA grow from 40,000 subscribers to 150,000 subscribers with the acquisition of three systems in St. Louis County. The first system purchased was from Warner Amex. This system utilized the Qube two-way interactive technology. The second system was purchased from Group W and is a one-way addressable system using Zenith Z-Tac converters. The third system was acquired from Storer and is a one-way addressable dual-cable system utilizing Tocom converters. The customer service functions for all these systems were combined to one location. The billing systems and control of all three addressable controllers were installed on one Cable Data Tandem computer.

The Qube system was offering three channels of PPV programming at the time of the acquisition. The other two systems however were not offering pay per view programming. Also, in 1985 CCA reached agreement with the St. Louis Cardinals baseball team to create the St. Louis Cardinals Cable Network. The Cardinals Cable Network offers 50 home games to cable television subscribers in the St. Louis market on either a season subscription basis or as a single game pay per view purchase. The first game offered was the 1986 home game opener, April 8, 1986. In addition Wrestlemania II was set for April 7, 1986.

The decision was made to offer three channels of pay per view programming on the two one way addressable systems also, and because of the dates mentioned above, it was to begin by the first week in April. Thus begins the experience of implementing pay per view on one-way addressable systems and the resulting list of questions that should be asked and answered before implementing pay per view.

QUESTIONS TO BE ASKED AND ANSWERED

How will the orders be taken?

To obtain the correct answer to this question could be the subject of another paper itself. For our purposes here we need to determine the impact that various order taking methods may have upon the technical operations. The first method to be considered is when orders are only taken by CSR's. This has the least impact because fewer orders can be taken and the last minute authorization rush is minimized. PPV channels will be authorized as would any pay service and this order taking method causes no problems in itself. The next method is the use of an Automatic Response Unit. With the use of a touch tone telephone, a subscriber can place a pay per view order without human interface. The affect of this device is that the ARU must interface with either the billing system or the controller directly and many more orders can be placed at the last minute. The effect of these items will be discussed later. Another telephone order method being trialed is utilizing the Automatic Number Identification capability of the telephone company in your area. Under this method we would need our controller to be capable of receiving and utilizing the data sent by the telephone network to authorize a subscriber for a pay per view event. The last method being trialed is referred to as store and forward. With this method a device is added to the subscribers converter that allows impulse or immediate ordering and reception of a pay per view event. The device descrambles the event and records the transaction. The transaction data is sent to the cable headend via phone lines at a later date. Here again, our controller must be capable of receiving and processing the data. Also the add-on device must be plugged into a telephone jack at the subscriber location. We need to make arrangements to install these jacks if one is not available. In all of these methods it can be seen that we need to know the capability of our controller. Thus the second question....

What is the capability of the controller?

Many of the controllers currently in place with one way addressable systems need to be upgraded to handle pay per view authorizations. Many times this is a software upgrade, but it may also require a hardware upgrade (increase in memory capacity). You need to determine the speed with which the controller can send authorizations, the number of authorization codes it can store and its ability to store different authorization codes for the same channel. You need to determine the method and order of authorization commands the controller uses to authorize the converters. If it is unable to send the data initially, will those commands take precedent over other commands in the global commands that follow. If the controller does not have these capabilities, then the billing system may need to perform some of these functions. Thus the next question....

What is the interface between the controller and the billing system?

In order for pay per view programming to be provided there must be on line access to customer records and control of subscriber converters. Some of the order taking methods mentioned above work directly with the controller. If the controller is capable of this, the controller would download the billing information to the billing system at a later date. However, in most situations the focal point is the billing system. The ordering data goes to the billing system which then sends the command to authorize a subscribers converter to receive a particular pay per view event. You need to determine the software capacity and capability of sending commands for multiple events on the same channel. The speed of these authorizations is also critical and the slowest link in the authorization chain needs to be determined. It is very important to understand the authorization sequence and the requirements and limits of each step from the taking of the order, the acceptance of that order placement data by the controller, the recording of the order in the billing system and descrambling of that event by the converter. This leads us then to the next question....

What are the capabilities and limits of the converter?

If you are fortunate enough to have in place addressable converters with real time and memory capabilities some potential problems will not exist. The converter must descramble then rescramble a channel around each authorized pay per view event. If the converter has real time clock and memory capacity, the beginning and ending times of an event can be downloaded to the converter when the order is placed. If this capacity is not available then the billing system and/or controller software must be set to send authorization and deauthorization commands to the converter before the event and at the end of the event.

You need to understand how data is received by the converter. Does the converter need to be on a particular encoded channel or on any encoded channel and will it receive the data when it is off. If the converter is not on a data channel when the initial order is sent, how long will it take the controller to resend the data so it is received by the converter when it is on a data channel. This needs to be understood so that subscribers can be notified to have their converters on a particular channel when an event is authorized or to tune in for a period of time before an event starts.

It is important to find out if there are any problems currently with authorizing pay services. If it takes several "hits" to authorize a channel now, the problem will become a major problem when the "hits" are sent automatically and when they are sent out often. Many converters require particular signal levels before data is accepted. The levels of the encoders and modulators should be carefully checked. Another potential problem results from the fact that many times the channels used for pay per view programming have not been encoded channels before. This means that the converters may not have been checked out for these particular channels and there may be some converters that will not take a "hit" on those channels. You need to be aware of this and if possible put pay per view programming on channels that have been checked out.

SUMMARY

If the above questions are asked and answered before implementing pay per view, many potential problems can be avoided. Experience tells me however that problems will occur that have not been anticipated. If you have answered the above questions in detail though, you will have a thorough understanding of how your addressable system operates and what its capabilities and limits are regarding implementation of pay per view programming. The key is to plan ahead and involve as many people as possible in the discussions. One last word of advice, if at all possible, implement pay per view with normal programming and not major events such as Wrestlemania II and the St. Louis Cardinals season opener. Events that generate the largest order volumes are not the ones under which you want to find out what questions you should have asked and answered ahead of time.

The Development of a New Super-Tough Cable Jacket

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SYNOPSIS

The harsh environment to which underground cables are exposed requires the best protection available. Presently the number of economical options available to the system designer is limited.

This paper introduces a new jacket design especially designed to protect sensitive coaxial cables from damage due to impact, abrasion and mechanical fatigue. Design development and testing of the new cable vs conventional cable designs are discussed within the paper.

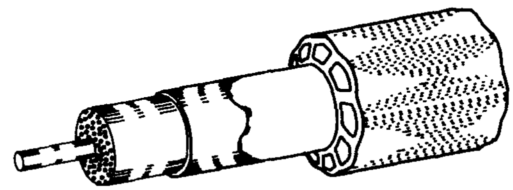
INTRODUCTION

For many years the cable industry has relied on telephone invention and research as a basis for its design. This reliance includes the use of steel armored cables for improved enviromental protection during installation and the life of the cable. Steel armored cables provide good crush resistance, rodent resistance and low frequency shielding. Unfortunately, steel armored cables are very expensive and somewhat over engineered for certain uses in the CATV applications. Many CATV cables require crush and corrosion resistance but often do not require rodent protection or low frequency shielding afforded by steel tape.

Presently the CATV operator has three expensive options available to him if he requires additional cable protection; armored cables, cable in conduit, or trench and backfill. A new cable jacket design offers an alternative to these options.

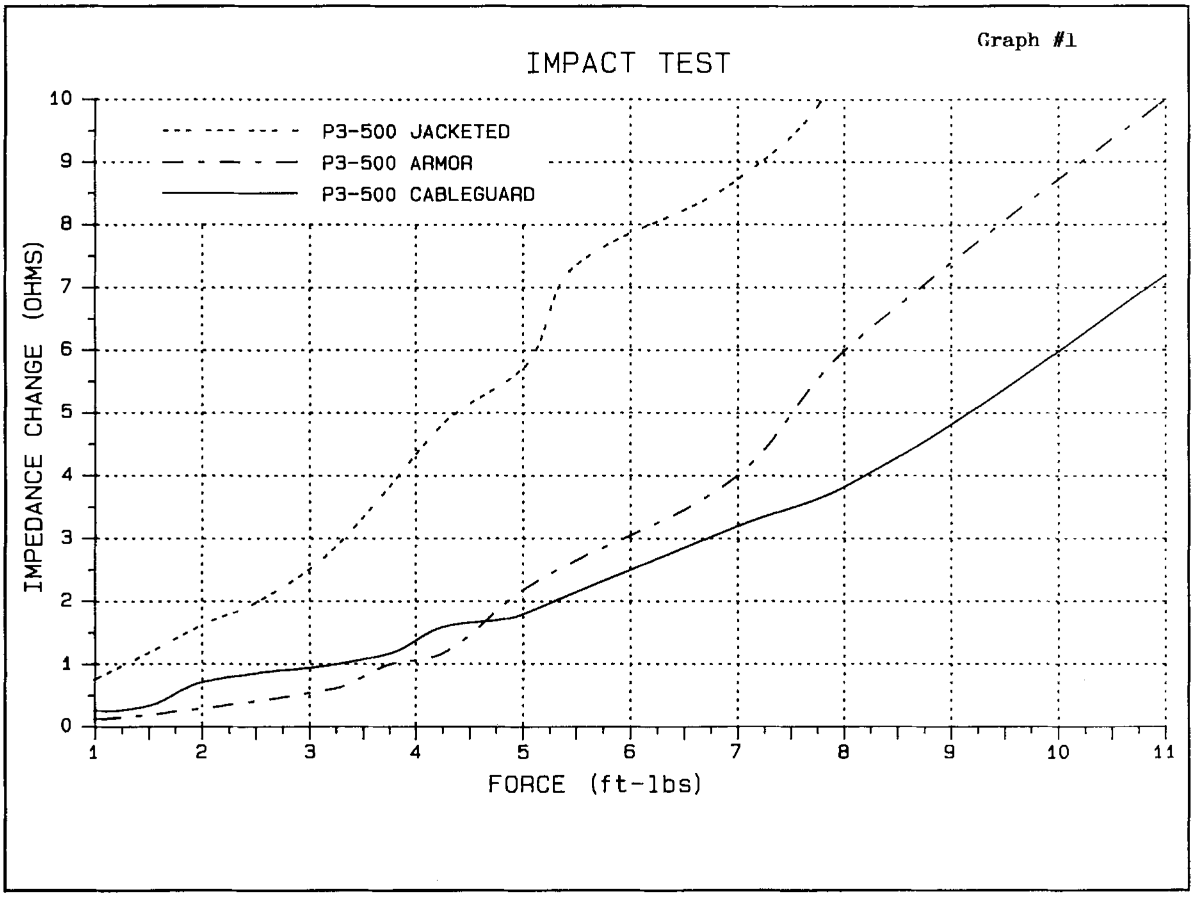
It is possible to provide a level of crush and impact resistance nearly equal to armored cables but at a significantly lower cost. The new jacket incorporates air cells surrounded by layers of Linear Medium Density Polyethylene (See Drawing #1). The air cells act as a cushion to dampen the forces applied directly to the cable conductors. The ridges that make up the cells provide an effective barrier reducing cut through and impact damage much the same as the flutes provide with corrugated packaging.

CABLEGUARD™ Drawing #1



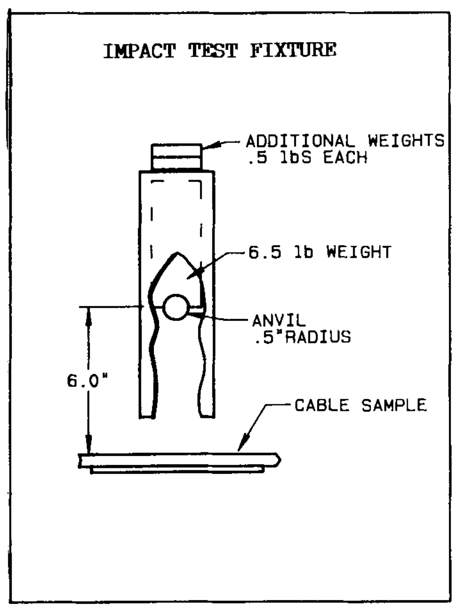
JACKET DESIGN

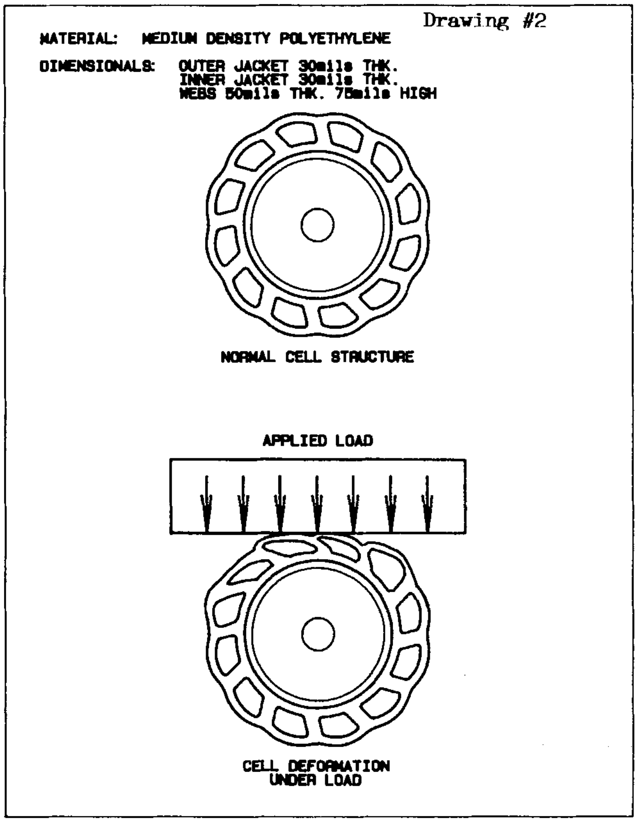
The new jacket design, called Cableguard™ was first conceived in early 1986 but only after numerous product designs and process modifications was the product available for commercial evaluation in January 1987. Upon close examination of the jacket design one sees the utilization of arches that provide support to the outer jacket. By modifying the classic arch, setting the arch at an angle one can modify the direction and magnitude of the forces directly applied to the inner foundation of the jacket. By connecting the arches together and tieing the bases of each arch together the load or impact can be more evenly distributed over a larger area. With conventional jackets and armored cables the ability to withstand impact is primarily affected by the rigidity of the individual components of the cable. With conventional jacket, once the cylindrical polyethylene jacket is deflected the majority of the force of impact is directly transmitted to the outer conductor of the cable. With armored cables the rigidity of the jacket and armor provides an effective barrier of impact up to a point but once this impact force is exceeded all additional impact force is directly transmitted to the outer sheath of the cable. Graph #1 displays the effect of impact on various designs of cable jackets.

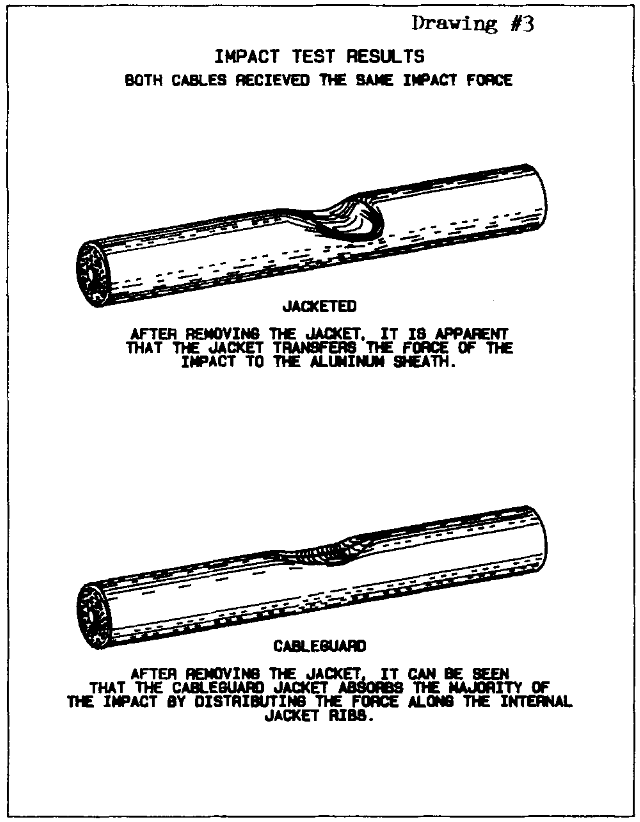


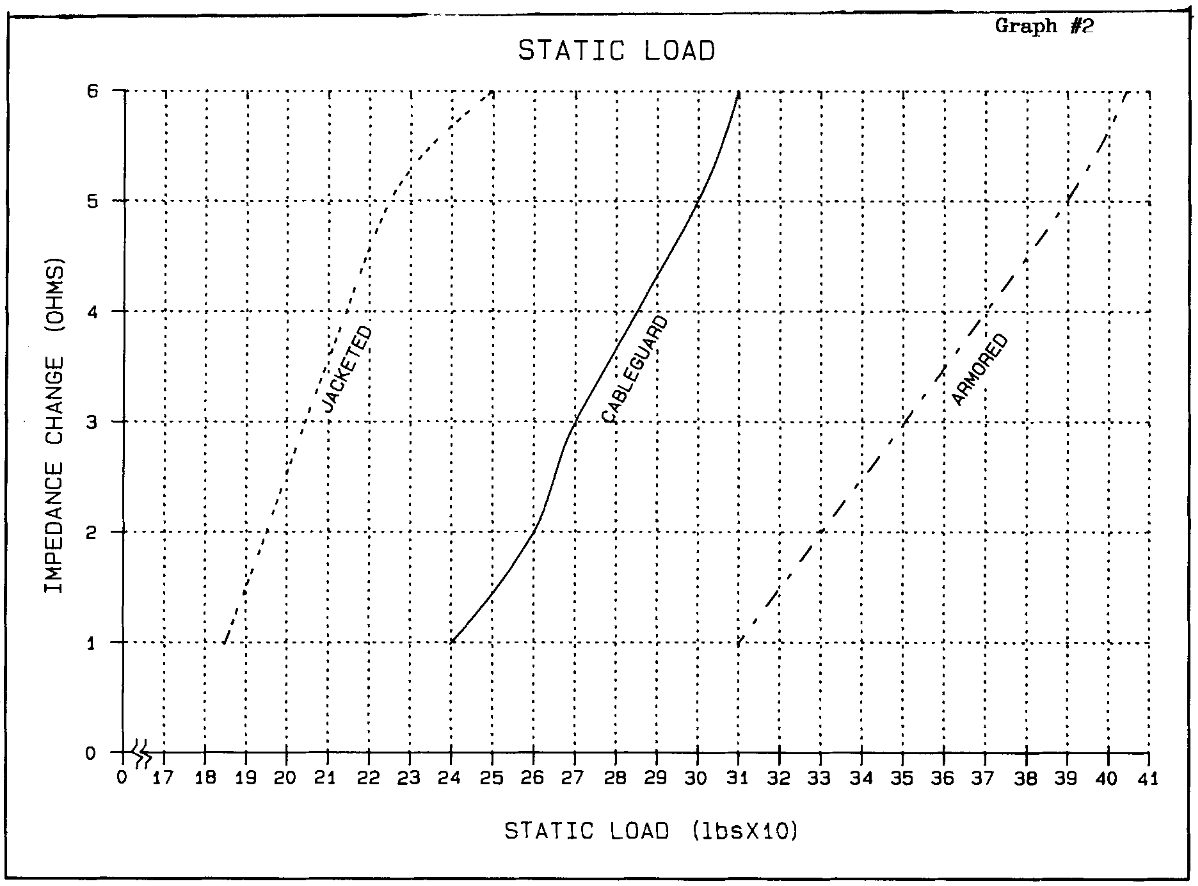
As can be seen the conventional cable jacket provides limited impact protection. The armored cable provides a much higher level of protection but as higher forces are applied the % change in impedance vs impact in foot pounds increases significantly. At low impact the Cableguard ™ jacket behaves much like armored but with high impact forces the slope of the curve is more gradual. This difference can be justified by understanding how the individual cells are deflected. As the impact force increases the more the arches tend to fold or collapse absorbing and distributing the force not only around the cable but perpendicular to the point of impact. Finite Elemental Analysis reveals for a given impact applied to the Cableguard ™ sample up to 65% of the force applied to the outer jacket is discipated before reaching the outer conductor of the cable. The analysis also reveals for conventional jacketed cable 95% of the impact applied to the jacket is transmitted to the outer conductor of the cable (See Drawings #2 and #3)

While Cableguard ™ was designed for impact resistance, armored cables still out perform Cableguard ™ in static compression test. Graph #2 compares the performance of various cable designs under static load.




Drawing #2
MATERIAL: MEDIUM DENSITY POLYETHYLENE
DIMENSIONALS: OUTER JACKET 30mils THK.
INNER JACKET 30mils THK.
WEBS 50mils THK. 75mils HIGH
NORMAL CELL STRUCTURE
APPLIED LOAD
CELL DEFORMATION UNDER LOAD


Drawing #3
IMPACT TEST RESULTS
BOTH CABLES RECIEVED THE SAME IMPACT FORCE
JACKETED
AFTER REMOVING THE JACKET, IT IS APPARENT THAT THE JACKET TRANSFERS THE FORCE OF THE IMPACT TO THE ALUMINUM SHEATH.
CABLEGUARD
AFTER REMOVING THE JACKET, IT CAN BE SEEN THAT THE CABLEGUARD JACKET ABSORBS THE MAJORITY OF THE IMPACT BY DISTRIBUTING THE FORCE ALONG THE INTERNAL JACKET RIBS.


Graph #2
STATIC LOAD
IMPEDANCE CHANGE (OHMS)
JACKETED
CABLEGUARD
ARMORED
STATIC LOAD (lbsX10)

PERFORMANCE COMPARISION:

	500 Standard Jacket (Medium Density PE)	500 Armored Jacket (2 Medium Density PE Jacket Plus .006 Steel Tape)	500 Cableguard™ (2 Medium Density PE Jackets With Air Cells)
Impact Test Impedance Change @ 5 ft/lbs.	5.7 ohms	2.2 ohms	1.8 ohms
Impact Test Impedance Change @ 10 ft/lbs.	14.0 ohms	9.3 ohms	6.0 ohms
Minimum Bend Radius	8.0"	10.5"	8.0"
Direct Burial Approved	YES	NO	YES
Cut Through (lbs. required to cut through to jacket (drop test)	37 lbs.	160 lbs	110 lbs.
Prep Time (jacket removal with tool)	1 minute	5 minutes	1 minute
Relative Cost (armored equals 100%)	56%	100%	75.%
Typical Diameter	.600	.730	.750
Compression (static crush test 3 ohm change)	205 lbs.	350 lbs.	270 lbs.

STANDARD JACKET

Advantages:

1. Good corrosion protection
2. Ease of connectorization (jacket removal)
3. Low cost
4. Product can be plowed or trenched in the ground
5. Light weight
6. Small bend radius

Disadvantages:

1. Poor cut through resistance
2. Low impact resistance
3. Does not improve crush or deformation resistance (susceptible to deformation in plow during installation)

ARMORED JACKET

Advantages:

1. Good corrosion protection
2. Excellent cut through resistance
3. Good crush and deformation resistance

Disadvantages:

1. High cost
2. Difficult to remove jacket for connectorization
3. Requires trench installation technique (often can not be plowed)
4. Heavy product to handle
5. Larger bend radius

CABLEGUARD™ JACKET

Advantages:

1. Good cut through protection
2. Excellent crush resistance
3. Moderate cost
4. Ease of connectorization (jacket removal)
5. Light weight
6. Small bend radius
7. Product can be plowed

Disadvantages:

1. Require jacket stripping tool
2. Air cells must be sealed at shrink boot
3. Diameter much larger than standard jacket

CONCLUSION

New jacket designs for coaxial cables will provide an added level of protection to the cable. While Cableguard™ does not perform as well as armored cables the various advantages and disadvantages of each should be considered. Armored cables should always be used for locations with rodent problems or severe abrasive components in the soil. Cableguard™ type jackets should be considered where fear of abrasion, mild cuts or crush resistance is a factor.

Hopefully, the evolution of jacket materials and jacket designs will allow simplication of installation and improve the service life for coaxial cables.

ACKNOWLEDGEMENTS

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THE FACTS ABOUT Ku BAND SATELLITE TRANSMISSION FOR CABLE TELEVISION

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ABSTRACT

This article treats the subject of Ku Band satellite TV transmissions from the viewpoint of the ground segment or the receiving earth station. It is, therefore, intended as a tutuorial for cable television engineers who are planning or will plan to receive the programming available from Ku Band satellites. Uplink and satellite performance parameters are discussed and included only to the extent that they affect the overall performance and serve as a benchmark for design of the receiving earth station.

Ku Band transmission has its pluses and minuses in comparison to the more familiar C Band transmission. The advantages and disadvantages are discussed without bias. The main thrust of this paper is; once given a Ku Band transmission, and a range of parameters on the signal from the satellite, how does one go about accommodating this signal to provide quality programming to subscribers?

The primary advantage of Ku Band is that it does not share the same frequencies on a co-equal basis with terrestrial microwave systems. It is, therefore, essentially a clear channel. The only terrestrial users are Local Television Transmission Services (LTTS) which operate on a secondary basis. This means that the earth station can be located anywhere as long as there is line of sight and clearance to the satellite orbital arc. The FCC will not even accept application for license in this band for receiving stations. This does not mean that a receiving station is totally free from interference. The receiving station must have sufficient discrimination against other Ku Band satellites in the orbital arc.

The primary disadvantage of Ku Band is that it is subject to weather related fading to a greater extent than C Band. The fading is explained and it will be shown that the reliability or availability of a signal of whatever quality desired can be controlled by design of the receiving earth station, usually in a cost-effective manner.

In addition, the article will discuss certain precautions to be taken in installing the hardware used in the receiving earth stations.

WHAT IS Ku BAND?

Ku Band is an old radar designation for frequencies in the range of 12 to 15 GHz. In the world of communication satellites, it denotes the frequency of two classes of satellite service. There is a Broadcast Satellite Service (BSS) which operates from 17.3 to 17.8 GHz in the uplink and 12.2 to 12.7 GHz in the downlink. There are no satellites currently flying utilizing the BSS band but several construction permits have been issued (and revoked) by the FCC. The second class of service is the Fixed Satellite Service (FSS). The FSS utilizes 14.0 to 14.5 GHz in the uplink, and 11.7 to 12.2 GHz in the downlink. The BSS band is regulated by the Mass Media Bureau of the FCC and the FSS band is regulated by the Common Carrier Bureau of the FCC. This paper deals with the FSS, Ku Band and in particular the downlink since cable operations will, for the most part, be on the receiving end of the satellite transmission.

The band 11.7 to 12.2 GHz is very nearly three times the frequency band 3.7 to 4.2 GHz of C Band. Therefore, the wavelength is 1/3 that of C Band. A parabolic antenna at Ku Band will have a beamwidth about 1/3 of the beamwidth of a C Band antenna with the same diameter. Since the gain of an antenna is proportional to the square of the frequency, a Ku Band antenna will have about 10 dB more gain than a C Band antenna with the same diameter. A 3 meter (10 ft.) diameter antenna at Ku Band is electrically equivalent to a 9 meter (30 ft.) diamter antenna at C Band.

THE CLEAR CHANNEL

No FCC License for The Earth Station at Ku Band

The frequency band 11.7 to 12.2 GHz is set aside by the FCC primarily for satellite downlink. There are a very limited number of licenses (about 50) issued to common carriers for use in the Local Television Transmission Servive (LTTS) which operate in this same frequency band on a secondary or noninterfering basis. Since the possiblity of interference is so small, the FCC does not license receive only earth stations in this band. The FCC will not even accept applications for license of receive earth stations. It would be prudent when planning a Ku Band earth station in the 11.7 to 12.2 GHz band to notify the common carrier LTTS licensee in your area. The geographic coordinates of the station and the antenna used should be given in the notification.

This "clear" channel of operation provides a great deal of freedom to anyone planning a Ku Band earth station. Any site that can support the antenna with a clear unobstructed view of the geostationary arc can be used. It need not be co-located with a C Band station although it could be.

Satellite Power

Since there are no primary terrestrial transmitters to interfere with satellite receiving earth stations, there are no primary terrestrial receivers to be victims of interference from a satellite. Ku Band satellites can, therefore, concentrate more power on the earth than C Band ones. Early Ku Band satellites have about 20 watt transmitters on board. The RCA K1 and K2 satellites have 45 watt transmitters. The RCA K3 and K4 to be launced in 1989 or 1990 will have 60 watt transmitters. The BSS satellites will have either 100 watt or 230 watt transmitters.

Interference From Adjacent Satellites

Although the Ku Band satellite service does not share frequencies on a co-equal basis with terrestrial service, there are a multiplicity of satellites in the orbital arc operating at the same frequencies. A receiving station must depend on the discrimination of its antenna to protect it from interference from other satellites. Furthermore, there does not currently exist a "standard" frequency and polarization plan at Ku Band. The consequence of this non-standarization is that polarization isolation cannot be counted on in planning an earth station.

From a practical standpoint, it would be prudent for the earth station planner to purchase an antenna meeting the FCC sidelobe envelope requirements of Para. 25.209 of the FCC Rules and Regulations. An antenna of 2.4 meters diameter or greater meeting these requirements should provide adequate protection from adjacent satellite interference.

PRECIPITATION

Ku Band transmission is subject to deeper fades and more frequent fades than C Band transmission. This is because the shorter wavelength (2.5 centimeters vs 7.5 centimeters) is affected more than the longer wavelength. Satellite links need not provide the same level of protection (fade margin) as terrestrial links, since only a small portion of the path passes through the earth's atmosphere. However, unlike C Band service, effects of precipitation must be included in the planning of a Ku Band service.

The effects of precipitation have been thoroughly studied, and solid engineering prediction models for rain fades exist which have been experimentally verified in a large number of cases. Rain fades are dependent on rain rate and the length of the path through rain. In the case of satellite transmission, the path length depends on how much atmosphere the signal passes through, or the elevation angle of the antenna. It follows, therefore, that rain fades will depend on local conditions and the location of the earth station with respect to the satellite.

Effects of Snow

The effects of snow in the atmosphere are not nearly so bad as rain regardless of the rate of snowfall. However, an uneven accumulation of snow in the dish can cause a serious de-focusing effect which will reduce the antenna gain and increase the sidelobes. It can also cause mispointing effects, further reducing the gain on the mechanical boresight axis. In this regard, the effects are not much different than at C Band, especially for dry snow. The effects of wet snow are far more degrading than dry snow. It is obvious that a dry snow condition will always degrade to a wet snow condition when the inevitable thaw occurs, if the accumulation is allowed to stay in the dish.

The net effect for those cable operators whose philosophy it has been to sweep the snow out of antennas when pictures degrade, is to be more vigilant. A good rule of thumb is to be alert to potential problems if the weather prediction is for more than two inches of snow. The alternative, for reliability, is to equip the antenna with an automatic snow removal heating system at some additional expense.

Effects of Rain

Several models exist to predict attenuation due to rain[1]. The Crane Global Attenuation Model[2] is recommended for use in planning satellite systems because it is quantitative and has been proven accurate. The Global Model relates point rain rate (for which much statistical data exists) to attenuation at various frequencies. The Global Model correlates well with measured attenuation data[3]. Figures 1 and 2 show the rain rate climate regions worldwide and for the Continental United States (CONUS) in particular. Figures 3 through 9 provides data for all of the rain rate regions in the CONUS. In the following section, it will be shown how these charts can be used in the design of earth stations in the Ku Band. For those who wish to "fine tune" a reliability prediction, use reference 2, if rain rate data exists for the site planned for your earth station.

LINK ANALYSIS

The link analysis and examples given in this paper apply to the RCA K1 satellite at 85° west longitude. The modulation prameters used in this satellite for transmission to cable systems has been optimized to provide a reasonable compromise between clear weather signal to noise ratio, and fade margin. In the transmissions from this satellite, the deviation of the main carrier by video is 9.2 MHz, and if subcarrier is used for audio, the deviation of the main carrier by the subcarrier is 2 MHz. The receiver to accommodate these signals should have a bandwidth of approximately 24 MHz.

For an FM receiver operating above threshold, the video signal-to-noise ratio (peak luminance signal to rms noise) can be expressed by:

$$(S/N)_V = 6\ m^2\ (\frac{B}{fm})\quad (C/N)_{PD}\ (PW) \ldots\ldots\ldots\ldots (1)$$

where,

$(C/N)_{PD}$ is predetection carrier-to-noise ratio:

m is modulation index $(\frac{\Delta F}{fm})$

ΔF is deviation of main carrier by video

B is predetection bandwidth

fm is highest modulation frequency (4.2 MHz for NTSC)

(pw) is combined pre-emphasis and weighting improvement factor (12.8 dB for U.S. and CCIR standards)

Substituting appropriate values for ΔF and B, this equation reduces to:

$$(S/N)_V = [(C/N)_{PD} + 35]\ (db) \ldots\ldots\ldots\ldots (2)$$

The predetection carrier-to-noise ratio is:

$$(C/N)_{PD} = (C/N)_U \oplus (C/N)_D \oplus (C/I)_T \ldots\ldots\ldots (3)$$

where,

$\oplus$ denotes power summation:

$(C/I)_T$ is total carrier-to-interference power ratio

$(C/N)_U$ is Uplink C/N

$(C/N)_D$ is Downlink C/N

Clear Weather $(C/N)_D$ is:

$$(C/N)_D = [(EIRP)_S + G_R - T_R - A_D - (k+B)]\text{dB} \ldots (4)$$

where,

$(EIRP)_S$ is satellite EIRP - dBW

G_R is Receiveing antenna gain -dB

T_R is Receiving system noise temperature -dB°K

A_D is downlink clear weather space loss = 205 dB

k is Boltzmanns constant = -228.6 dBW/°K

B is bandwidth - dBHz = 73.8 dBHz

Substituting and assuming the use of an LNA or LNB with 2.5 dB noise figure, and antenna noise temperature of 40°K:

$$(C/N)_D = [(EIRP)_S + G_R - 74.8]\ \text{dB} \ldots\ldots\ldots (5)$$

In a rain fade, the carrier-to-noise is modified by adding two addition factors:

$$(C/N)_{DF} = [(EIRP)_S + G_R - \text{Fade} - 74.8]\ \text{dB} \ldots (6)$$

should be greater than 8 dB to remain above threshold.

where,

Fade = $L_R + T_e$, L_R is rain attenuation (in dB),

and

T_e is change in system noise temperature and is expressed by:

$$T_e = 10 \log \left[\frac{T_R + T_0 \frac{(L_R - 1)}{L_R}}{T_R} \right] \text{dB}$$

where,

L_R is rain attenuation (in numeric)

T_0 is ambient temperature (290°K). The sum of L_R and T_e represents the total fade in rain.

Equation (6) is the expression that a designer of a receiving earth station has to work with. He (or she) must know the satellite EIRP for the site. Satcom K1 EIRP footprint is given in figure 10. He can then determine what size antenna to use for a desired S/N, and reliability.

Since the S/N is dependent on $(C/N)_{PD}$, and equation (6) is $(C/N)_D$ in 24 MHz bandwidth, it is necessary to combine equations (3) and (6):

$$(C/N)_{PD} = [EIRP_S + G_R - (L_R + T_e + 74.8)] \oplus (C/N)_U \oplus (C/I)_T$$

A reasonable assumption for $(C/I)_T$ is 26 dB.

A reasonable assumption for $(C/N)_U$ is 26 dB.

Therefore:

$$(C/N)_{PD} = [EIRP_S + G_R - (L_R + T_e + 74.8)] \oplus 23\ \text{dB} \ldots (7)$$

DESIGN EXAMPLE

Location of earth station site: Latitude 33°N. Longitude 97°W. (near Dallas, TX)
Desired $(S/N)_V$ in clear sky: 53 dB
Desired availability: 99.95%

1. From equation (2), for a $(S/N)_V$ = 53 dB, $(C/N)_{PD}$ = 18 dB.
2. From equation (7), for a $(C/N)_{PD}$ = 18 dB, $(C/N)_D$ = 19.6 dB.

3. Station is in rain rate region D2 (Figure 2). At this point a calculation of the antenna elevation angle must be made. Use the following formula:

$$EL = -\arctan\left(\frac{\cos\theta\cos\alpha - R/D}{\sin\theta/\sin AZ}\right) \quad \text{(degrees)}$$

where,

$$AZ = 180° + \arctan\left(\frac{\tan\theta}{\sin\alpha}\right) \quad \text{(degrees)}$$

α is earth station latitude
θ is relative longitude or satellite longitude minus earth station longitude ($\theta < 90°$)
R is radius of earth (3,957 miles)
D is radius of satellite orbit (26,244 miles)

With satellite at 85° W.L., solution of this equation yields an elevation angle of 49°.

From figure 6, rain attenuation is 5.1 dB for 99.95% availability.

$$T_e = 10\log\frac{265 + 290\left(\frac{3.24 - 1}{3.24}\right)}{265} = 2.4 \text{ dB}$$

Total fade is 5.1 + 2.4 = 7.5 dB.

From figure 10, EIRP is 47 dBW.

Using equation (5), and substituting:

For $(C/N)_D$ = 19.6 dB, the antenna gain should be:

$$G_R = (C/N)_D - EIRP + 74.8$$
$$= 19.6 - 47 + 74.8 = 46.4 \text{ dB.}$$

It should be noted that this value of gain will also satisfy equation (6) for minimum faded $(C/N)_D$ of 8 dB.

$(C/N)_D$ faded = 19.6 − 7.5 = 12.1 dB.

This analysis shows that a 2.3 meter diameter antenna will satisfy the requirement of $(S/N)_V$ of 53 dB in clear weather, with greater than 99.95% availability at this site.

In this case the $(S/N)_V$ was the critical limiting design factor. In some other region where the rain rate or elevation angle is less favorable, the critical design factor for dish size might be availability.

It can be shown that a target value of 99.95% availability (4.38 hours per year) can be obtained everywhere in the CONUS (except Southern Florida) using a 3.0 meter antenna with Satcom K1. In Southern Florida a 3.0 meter antenna will provide 99.90% availability to a $(C/N)_{PD}$ of 8.0 dB.

OTHER FACTORS

There are several precautions that must be taken for a successful installation at Ku Band. Remember, the wavelength of the Ku band frequency is 1/3 that of C Band. It follows, therefore, that much tighter tolerances are required in the microwave portion of the earth station. The components have been designed for these tighter tolerances, but mishandling or poor practice at installation could destroy these tolerances and thereby cause considerable degradation. For example, do not attempt to use a C Band dish, install a Ku Band feed, and expect performance as if it were a Ku Band design. Use antennas from reliable manufacturers. The RMS deviation from a paraboloidal surface should be less than 0.025 inches.

Extreme care should be exercised in assembly of a sectionalized dish. Don't force fit anything. Mating holes should line up without force fit. Don't overtorque or undertorque bolts.

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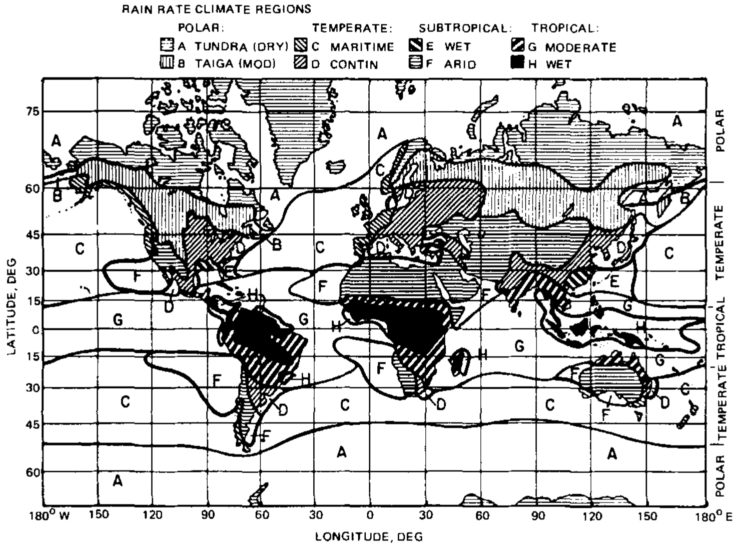


FIGURE 1. RAIN RATE CLIMATE REGIONS FOR THE GLOBAL RAIN ATTENUATION MODEL

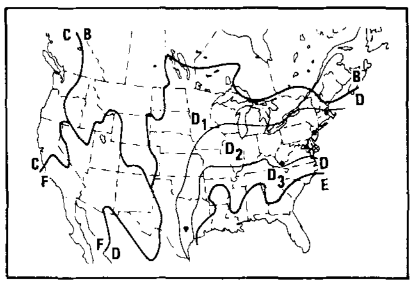


FIGURE 2. RAIN RATE CLIMATE REGIONS FOR THE CONTINENTAL U.S. SHOWING THE SUBDIVISION OF REGION D

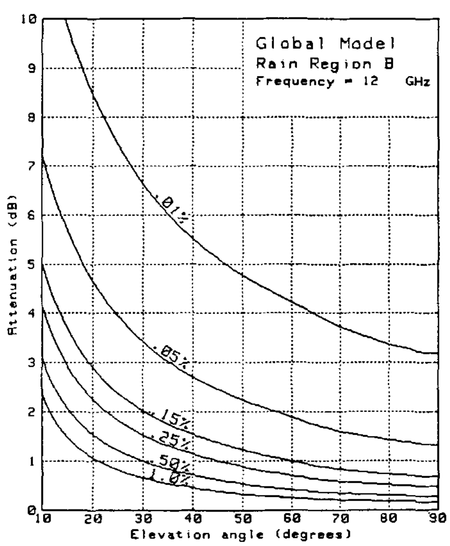


FIGURE 3. RELIABILITY DATA
RAIN RATE REGION B

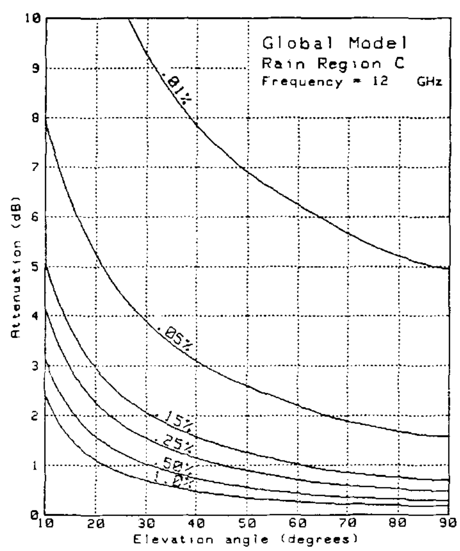


FIGURE 4. RELIABILITY DATA
RAIN RATE REGION C

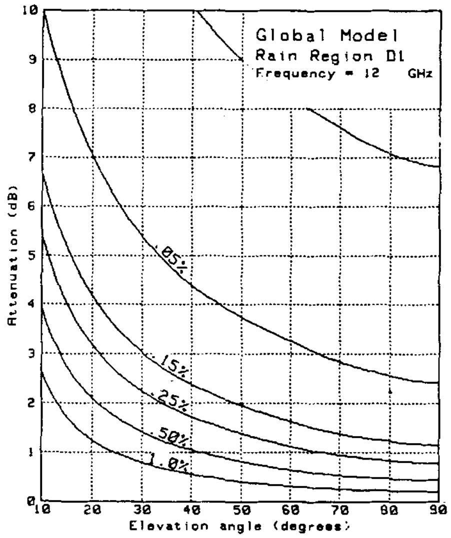


FIGURE 5. RELIABILITY DATA
RAIN RATE REGION D1

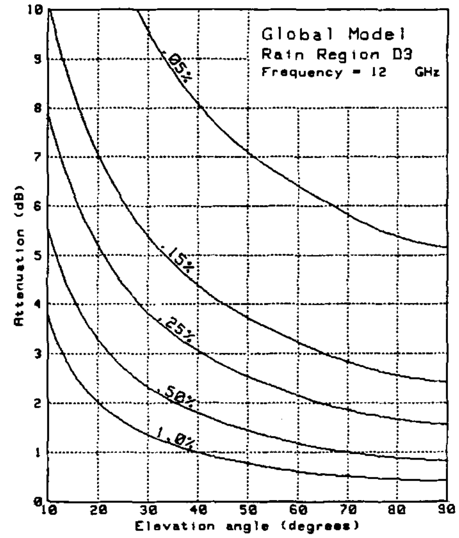


FIGURE 7. RELIABILITY DATA
RAIN RATE REGION D3

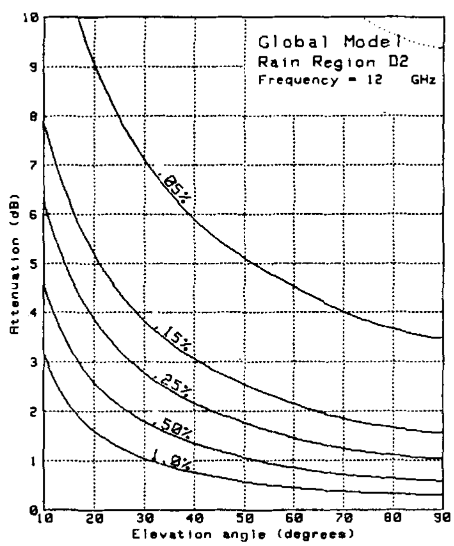


FIGURE 6. RELIABILITY DATA
RAIN RATE REGION D2

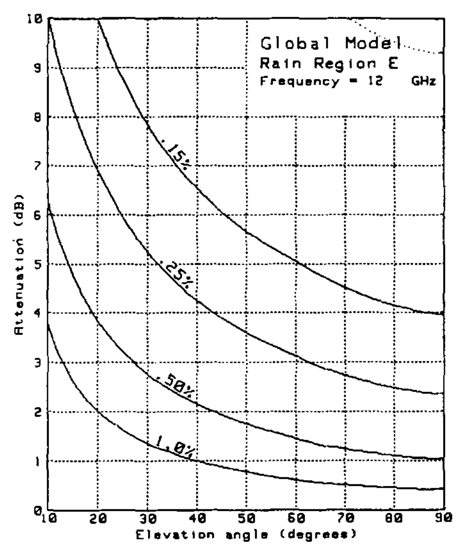


FIGURE 8. RELIABILITY DATA
RAIN RATE REGION E

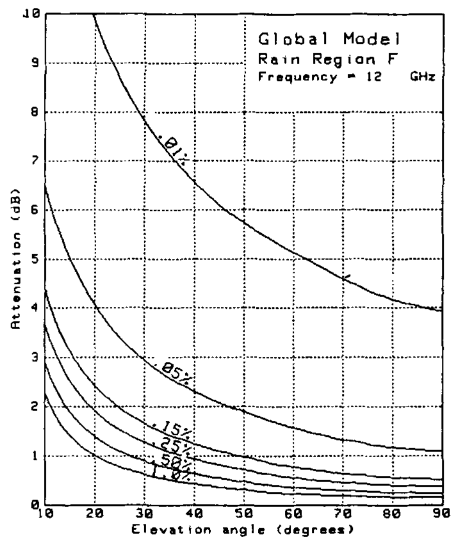
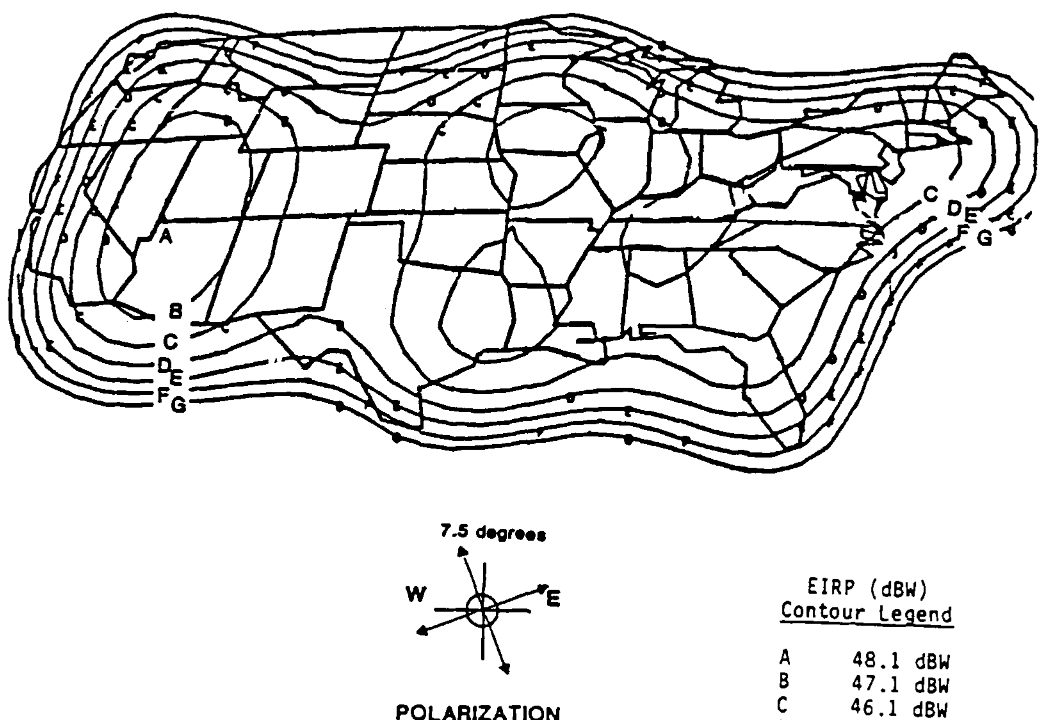


FIGURE 9. RELIABILITY DATA
RAIN RATE REGION F



EIRP (dBW) Contour Legend	
A	48.1 dBW
B	47.1 dBW
C	46.1 dBW
D	45.1 dBW
E	44.1 dBW
F	43.1 dBW
G	42.1 dBW

FIGURE 10. SATCOM K1 CONUS BEAM
EIRP CONTOURS

THE INTERACTIVE EVOLUTION

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ABSTRACT

For many years CATV systems have been viewed as ideal for the implementation of two-way interactive systems. In the late seventies and early eighties, several revolutionary systems were developed which failed primarily because they were technology driven, rather than market driven. They attempted to address a wide range of interactive applications and became too costly and too complex to operate. Presently, two-way interactive systems concentrate on specific applications, such as Pay-Per-View and Home Shopping, and are developing in a more evolutionary manner. This paper summarizes the current state of the art and explores possible approaches for evolving current embryonic interactive systems into ones with more wide spread applications. Candidate control system architectures are analyzed.

Interactive System Model

Figure 1 shows a basic interactive system model. There are three major system components to the model.

- The consumer who is requesting, receiving and paying for interactive services.
- The service component which delivers the service and receives compensation for it. In a cable based system the service component will consist of the organization providing the service and the cable system operator who provides the communications medium.
- The Customer Service and Billing component which provides the customer service, billing and payment processing functions. This component authorizes the consumer to receive services, collects data regarding the services used, bills the consumer and processes payments.

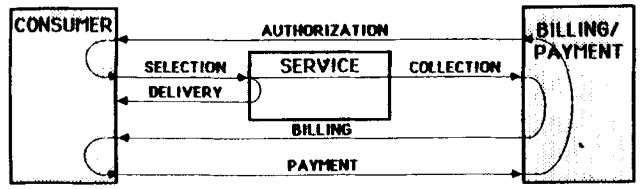

FIGURE 1: INTERACTIVE SYSTEM MODEL. LINES ARE COMMUNICATION LINKS. BLOCKS ARE SYSTEM COMPONENTS.

The communication segment can be separated into a control subsystem and a service delivery subsystem. The control subsystem consists of communications links which integrate the consumer, service and billing/payment components in the interactive system. It comprises the following.

- The authorization link which passes data allowing the consumer to use all or part of the interactive service offerings.
- The selection link, used by the consumer's equipment to request an interactive service.
- The delivery link which provides the requested service to the consumer. In today's interactive systems, the delivery link is usually a 6 MHz video channel.
- The collection link which passes data pertaining to the service a specific consumer has utilized.
- The billing link which periodically, usually on a monthly basis, bills the consumer for services utilized. The U.S. Postal System is currently used for both consumer billing and payment.
- The payment link which the consumer uses to pay his bill. Bill payment implicitly authorizes the consumer to continue to receive the interactive services he desires.

Historical Perspective of U.S. Cable Based Interactive Systems

This paper deals only with U.S. cable based interactive systems. There have been several telephone based interactive system trials, the most notable being the Knight-Ridder/AT&T Viewtron system and the Times Mirror Gateway system.

The earliest cable based interactive system to be widely installed was the Warner/Pioneer Qube system; a system which features pay-per-view programming and opinion polling. Qube uses interactivity to enhance the basic video services. Cox also introduced their Indax system, a more general purpose video and data oriented system, which went to field trial in San Diego, California and Omaha, Nebraska. Manitoba Telephone demonstrated their Omnitel system in Winnipeg which combined video, data and telephone traffic. General Instrument continued development of the Omnitel system in the United States. Other systems developed by General Instrument, Packet Technologies and World Video Library reached the prototype stage.

All of these systems, in one way or another, were technological successes and business failures. They were technology driven, rather than market driven. They all were not cost effective, primarily because of the high costs of running and maintaining the system and its data bases. Today, only Qube remains installed in some systems.

During the past few years, there has been a trend away from general purpose interactive systems to single purpose interactive systems. These systems concentrate on specific applications, such as pay-per-view and home shopping, rather than attempting to address the entire spectrum of two-way interactive applications. Even interactive game shows, such as Money Mania on the Nashville Network, are beginning to appear.

The present systems utilize a full video channel (live video) as the forward path and cable or (more likely) telephone as the return path. The degree of interactivity required is low and is accomplished by low bandwidth data communications or voice conversations. Major characteristics are:

- Video programming with a high entertainment content as the offered service.
- Quick response time, with emphasis on impulsive reactions.
- Cost effective implementation.
- Specific applications as opposed to general applications.

Interactive System Requirements

The major interactive system requirements are response time, conflict resolution, traffic handling capacity, cost effectiveness, simplicity and ability to handle multiple applications.

- Response time is the delay from requesting the service to delivering the service. From the consumer's viewpoint the response time should be as close to zero as possible. One to two seconds is acceptable for most services.
- Conflict resolution is the ability of the system to automatically deliver the desired service without the need for consumer retries due to a "busy signal".
- Traffic handling capacity is the ability of the system to accommodate the consumer population on a peak and average load basis. Bandwidth must be minimized in order to not severely decrease channel capacity for video services.
- Cost effectiveness refers to the ability of the system to be profitable for all participants - the system operator, the service provider and the equipment supplier. Cost effectiveness is a function of the value of the services offered to the consumer, the cost of offering these services, and the capital cost of the equipment involved.
- Simplicity refers to several facets of system implementation and operation. The interactive system must be reliable. It must be simple for the consumer to use, preferably from his easy chair while watching television. Most important, it must be simple for the system operator to operate. Manual operations of a highly complex nature must be eliminated. The more transparent an interactive system is to system operators, the better its chances of succeeding are.

- As we have seen, today's interactive systems focus on one application which makes their operation profitable. The ability to handle multiple applications will be important in the future if the current trends in interactive applications continue however, handling multiple applications must not overly burden system cost factors.

Candidate System Architectures

As shown in Figure 1, an interactive system can be partitioned into a communications segment, a service segment and system components. This partitioning is shown diagramatically in Figure 2. We have already covered the various interactive system components (consumer, service and billing/payment).

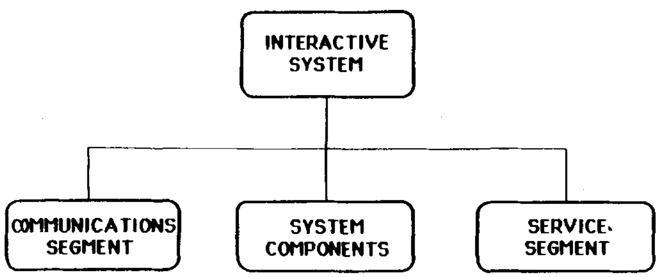

FIGURE 2: INTERACTIVE SYSTEM SEGMENTS.

The communications segment provides the links for system control and service delivery. There are various architectural alternatives for implementing the communications segment, the most common of which are shown in Figure 3.

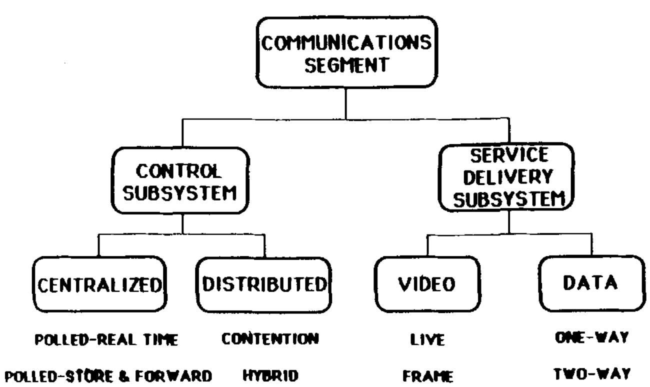

FIGURE 3: COMMUNICATIONS SEGMENT ARCHITECTURE TREE.

The purpose of the control system is to provide the link and protocols necessary to allow an orderly transfer of information between the consumer, service provider and billing/payment components. The control system can use either centralized or distributed control. Centralized control denotes a central facility which interrogates consumer equipment in order to maintain an orderly communications flow. Distributed control refers to a control system in which the consumer equipment initiates communications with central and other facilities. Distributed control requires more complex protocols and algorithms.

Referring to Figure 1, for all control subsystem architectures, the authorization, billing and payment links are very similar. For instance, authorization normally takes place over the downstream data link used for enabling addressable converters. The billing link can either be the U.S. Postal System or the downstream data link. Payment is via the U.S. Postal System.

Each control subsystem architecture will now be summarized, highlighting differences in the selection, delivery and collection links.

Polled-Real Time Control

Polled-real time control utilizes a central facility which addresses consumer equipment in predetermined sequences. Each consumer's equipment is interrogated, or polled, once every main cycle. When the consumer unit is polled, a two-way communication link is established and information is transferred from the polled unit to the central facility. The central facility performs authorization checks and delivers the service to the customer.

Response time is determined by when the request for service is made with respect to the polling cycle and the time it takes for the central facility to process the request. Delivery of service is not instantaneous, and in a large system the response time could take tens of seconds. The inherently slow response time dictates that cable, rather than telephone, be used as the return path.

Delivered service data collection is accomplished during the time that the central facility and the consumer unit are in a two-way communication mode. There is no delay in collecting service use data.

Polled Store and Forward Control

Polled store and forward control also utilizes a central facility which addresses consumers in predetermined sequences. The major difference is that the consumer equipment stores data indicating that a service has been selected and delivered. The consumer equipment is self authorizing. The polling cycle is used only to collect service use data and, therefore, does not have to be continuous, nor is it as time critical as in the real time system.

The store and forward system architecture depends on preauthorization of the consumer for the desired service set. No real time checking of consumer parameters (credit, authorization, etc.) or service availability is performed prior to delivery. Response time is almost instantaneous and there are no real limitations to system size, providing the consumer equipment has adequate memory for services requested and delivered.

The return path can either be via cable or telephone since, while telephone return is slower, there is no system requirement for fast polling cycles.

Contention Control Systems

Contention control was used in most of the distributed control architectures which were developed. Cox's Indax and General Instrument's MetroNet are examples of systems which utilized packet broadcasting with collision detection protocols, like Carrier Sense Multiple Access with Collision Detection (CSMA CD). ALOHA type systems which used a central facility to resolve common communication channel conflicts were also tested. Multiple narrow band channels are required to accommodate reasonable system sizes. The General Instrument MetroNet system, for instance, used multiple 300 KHz, 128 kbs channels which could accommodate 200 active users.

Service provider facility checks are made before a selected service is delivered. Response time is dependent on the number of active users and their average message length. In a well constructed system, delivery of the service is almost instantaneous; the critical parameter being the time a service provider requires to process the request. Collection of service use data is accomplished when the service facility and consumer equipment are in a two-way communication mode.

Contention systems require more complex protocols and algorithms for the consumer equipment. In larger cable systems, frequency agile transmitters and receivers will also be required. Consequently, consumer equipment tends to be relatively expensive.

Hybrid Distributed Systems

The hybrid distributed system uses the Public Switched Telephone Network (PSTN) for the return path (selection) communications and the cable system for service delivery. Customer selection is by means of a telephone keypad or a special keypad which interfaces with the PSTN.

Two-way communications are established over the telephone during the service selection process. The service facility performs authorization checks and delivers the service to the consumer. Data collection for delivered services is accomplished during the time when the service facility and consumer equipment are in a two-way communication mode. Response is dependent on dialing time and the service facility processing. This type of system is usually associated with video on demand, either pay-per-view or video frame dissemination.

Service Delivery Subsystems

Communications links are also required for delivery of requested services. The requested services can be video services, data services or a combination of both. Video services can either be live video (e.g., pay-per-view systems) or video frames (home shopping catalog services).

Data services can either be one-way (teletext) or two-way. One-way data services used on a pay-per-use basis could utilize a store and forward system to collect consumer use data for billing and payment processing purposes.

Two-way data systems allow data exchange between the consumer and the service facility, usually in the form of a request by the consumer and delivery by the service facility.

Today's services are usually live video with at least one one-way data application (X-PRESS). A natural evolution is the delivery of video frames to frame store equipment for catalog home shopping applications and other applications where reasonable resolution is required.

	Response Time	Conflict Resolution	Traffic Handling Capacity	Cost Effectiveness	Simplicity	Multiple Applications Capability	Communication Paths
Polled-Real Time	Relatively slow-depends on polling cycle time	Minimized-talk to one user at a time	Moderate-depends on response time requirements	Inexpensive in-home and system equipment	Simple equipment and operation	Limited	Cable downstream Cable return
Polled-Store and Forward	Fast, self authorizing	None required Self authorizing	High-non real time use data collection	Slightly more expensive in-home equipment	Simple equipment and operation	Limited-services must be on cable at all times	Cable downstream Cable or telephone return
Contention	Relatively fast-depends on number of active users per channel	Collison detection and retry algorithms	Moderate-requires multiple channels and user channel assignment	Relatively expensive in-home and system equip.	Relatively complex equipment and operation	Very flexible -data and video on demand	Cable downstream Cable return
Hybrid	Moderate-limited by telephone set up time	Retry if return path is busy	Relatively low -limited by PSTN	Inexpensive if PSTN costs ignored	Moderately complex if PSTN is considered	Moderate-data and video on demand	Cable downstream Telephone return

TABLE I: COMPARISON OF CANDIDATE CONTROL SEGMENT ARCHITECTURES

Comparison of Control Segment Architectures

Table 1 summarizes the previous discussion of candidate interactive system control architectures. Each of the candidates has strengths and weaknesses. Polled systems are relatively simple to operate and require minimum capital investment. They tend to be better suited for a limited applications set where the system is tailored to the predominent application. Most of the current cable based interactive systems are polled systems. Contention systems are relatively complex to operate, primarily because of their distributed nature, and are relatively expensive. They, however, are very flexible and can be used for a large applications set without severely compromising the efficiency of handling any specific application. Hybrid systems are attractive since they use the well established PSTN for return path communications and leverage off of an established equipment base. Traffic handling capacity is limited unless more expensive equipment complements are used. The system operator also does not have total control over the system.

CONCLUSION

Early attempts to develop general purpose, multiple application interactive systems failed, not technically but on a business basis. Today's interactive systems are tailored to specific applications and, while they are not as technically sophisticated as the early attempts, they are succeeding on a business basis. The beachhead established by pay-per-view and home shopping using polled control systems can be grown to a broader applications base using more sophisticated control techniques. Catalog shopping using full resolution video frame stores, games, stock quotes and broader pay-per-view applications appear to be the next generation of applications. As the applications set grows, distributed control systems will become more attractive - despite their added complexity. The dreams of the late seventies and early eighties are now becoming realities. The question is not whether CATV systems will support interactive services in the future. The question is whether we can evolve the current, narrowly focussed interactive services to a broader range of offerings in a cost effective and profitable manner.

THE VIDEOCIPHER II SATELLITE TELEVISION SCRAMBLING SYSTEM

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ABSTRACT

The VideoCipher II satellite television scrambling system provides secure satellite distribution of high quality video and stereo audio to both commercial distribution systems and individual homes with satellite TVRO equipment. Over 50 programming services have selected the VideoCipher II system, making it the de facto standard for entertainment program distribution. Hundreds of thousands of commercial and consumer descramblers are in operation today, providing higher quality television reception than is available with clear transmission, along with communication and control features that enhance the programming value.

Since the VideoCipher II system is designed for both CATV and DBS applications, it supports a number of important subscriber features. These include a flexible subscriber interface with an on-screen display, store-and-forward impulse pay-per-view, a parental lockout capability, and a text service for program guides, messages, and other information.

Topics in broadband modulator/demodulator design for video transmission

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1. Abstract

In a standard broadband video signal distribution system, performance degradations are caused by the system's power and bandwidth limitations. To minimize these distortions and to take advantage of the capabilities that the system offers, the best type of modulation must be selected for the application at hand .

2. Background

Traditionally,video (along with accompanying audio or data) has been used in consumer applications and entertainment.The distribution systems used to transmit these signals include over-the-air broadcast channels, CATV, satellite and lately, fiber optic links.

One of the key properties of baseband video is that most of the information it carries is concentrated in the low frequency portion of the signal spectrum. Since any distortion in this portion of the spectrum is highly visible to the human eye, the design of a high quality composite broadband video transmission system present challenges at almost every level. Any artifacts introduced by the video processing or transmission must be very low (minimum of 60 db below the signal path) or else the distortions become clearly visible.

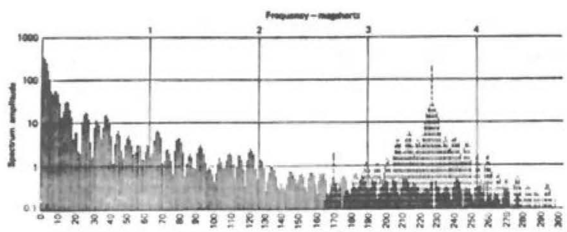


Figure 1: Composite luminance - chroma spectrum

Both the luminance and the chroma video information have periodic line spectra with a period equal to the line scanning frequency. The chroma information is modulated in quadrature on a subcarrier with frequency (approximately 3.58 Mhz) chosen to permit distortionless interlacing with the luminance line spectrum. Aural carrier, positioned 4.5 Mhz away from thr video is however only 920 Khz away from the color subcarrier.

3. Transmission channels

An NTSC baseband TV signal occupies a nominal bandwidth of 4.2 Mhz. For over the air and CATV transmission, the baseband video is VSB-AM modulated onto a carrier, while the audio baseband signals are FM-modulated onto a separate carrier 4.5 Mhz away from the video carrier frequency and at a level 15 db lower than the video carrier level. After multiplexing the video and the audio carriers, the total occupied transmission bandwidth is 6 Mhz.

a) Broadband distribution trunks

The topology of a typical CATV coaxial cable broadband distribution system includes a main trunk path length extending from several miles to several tenths of miles with amplifier boosters located every 1-2 miles. Subscribers are hooked to the main trunk via branches extending from the main trunk into the customer premises.

The key limitation of a typical CATV distribution system results from the degradation caused by the cable and the associated booster amplifiers. They limit the level and the number of the transmitted carriers as well as the achievable cable plant bandwidth. Extensive work has been performed to determine the number of carriers and their level , the noise accumulation and the second and third order intermodulation products generated as a result. Present state-of the art distribution amplifiers have a bandwidth of 550 Mhz, and can handle up to 55 high level VSB-AM signal carriers with a worst case commercial quality of 45 db C/N over the channel bandwidth.

b) Broadband distribution supertrunks

Broadband distribution supertrunk is a broadband communication link connecting typically two (sometimes more) multisource headend centers together.

For broadband supertrunk signal transportation it is imperative to be able to deliver video transparently (with virtually no system-added degradation) from geographically separated locations to distant (tens of miles away) headends.Typical CATV distribution systems having the usual number of repeater amplifiers cannot easily or inexpensively achieve this goal.

The topology of a supertrunk include path lengths extending from several miles to several tents of miles with one or more branches along the way. Because of the very high video signal performance required of a supertrunk, the number of video channels each supertrunk is required to carry is of a secondary importance only.

Coaxial cable - based supertrunks must use booster amplifiers in order to maintain the proper signal levels and CNR at all of the path sections. To guarantee, that after a cascade of booster amplifiers the required SNR at the farthest receiver location can be maintained , FM modulation is frequently selected . With the proper modulation index chosen for optimum SNR improvement versus bandwidth expansion , large improvements factors can be achieved at the expense of the number of channels which the system can support.

Fiber optic supertrunk distribution systems offer wide bandwidth with low losses and very small size and weight. Many fibers can be packed into a small space, permitting easy future expansion, excellent communication security and rapidly dropping equipment cost. Because of these properties, the system designer can select the frequency plan, the modulation method and the number of fibers offering the best and the most economical way of packing the number of required channels into a broadband supertrunking signal distribution system.

For long haul paths especially, the benefits derived from a broadband fiber optic supertrunk distribution system can be substantial - both in up front cost as well as in maintenance. Because FM modulation can offer large SNR improvement factors and the associated bandwidth expansion can easily and inexpensively be accommodated in a fiber optic link, a broadband fiber optic supertrunk can support a large number of high quality video channels inexpensively and economically.

With the rapid development of cost - effective optical sources and fiber technology, the use of fiber optic links as transmission media offers an attractive and cost-effective approach to distribution of high quality signals. It would allow the system designer to take full advantage of the wide bandwidth and of the insensitivity to EMI. Most importantly, no booster amplifiers would be required for links as long as 25 miles.

4. Modulator design for broadband trunk systems

Typically, the design objectives for a high-quality broadband distribution trunk system are set to guarantee that it will provide:

a) High quality of the delivered video and audio,
b) Reduced or no sensitivity to EMI,crosstalk etc.
c) No adverse environmental effects
d) Low equipment and installation cost
e) Low maintenance cost

A block diagram of a typical VSB-AM modulator is shown in Figure 2.

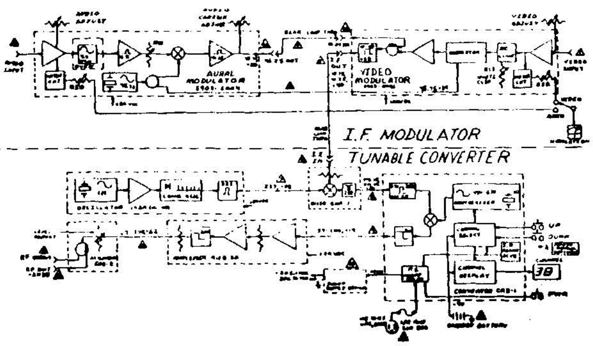
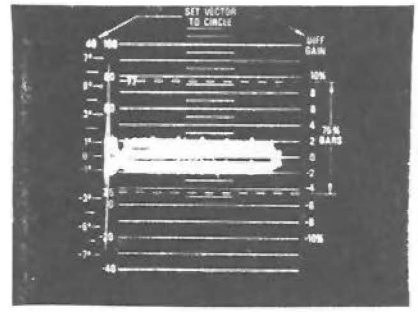


Figure 2: VSB-AM modulator block diagram

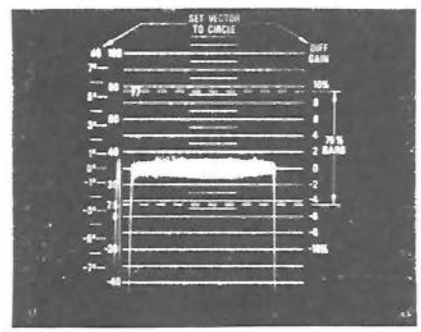
The modulator accepts baseband video and audio signals and generates the composite RF signal. Various performance specifications are imposed on the modulator to insure a high quality transmission of the baseband TV signals.

a) VSB-AM video performance with no noise present

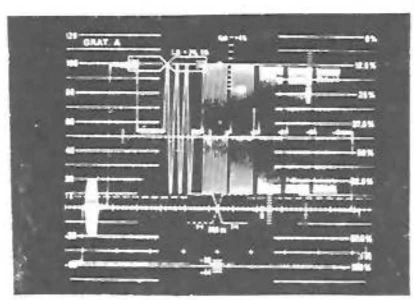
To evaluate the video performance of a system, standard test waveforms are applied to the video input and the degradations caused by the system are recorded.



Differential phase



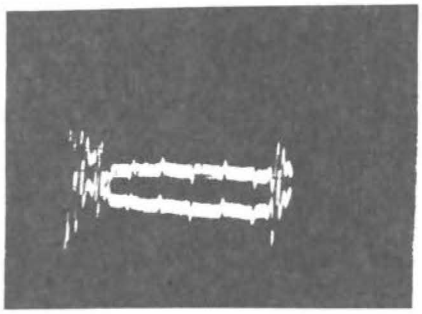
Differential gain



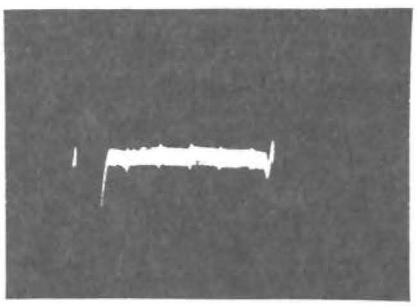
Multiburst test

Figure 3: Examples of standard video test waveforms

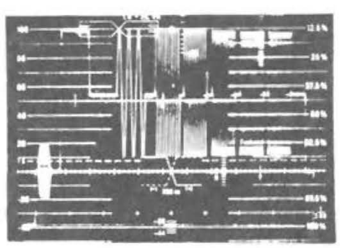
The response of a Catel CTM-20 VSB-AM video modulator to the test waveforms in Figure 3 is shown in Figure 4.



Differential phase



Differential gain



Multiburst test

Figure 4 CTM-20 VSB-AM video modulator response

A discussion and interpretation of the results reveals a wealth of information about the performance of the system.

Widely accepeted guidance standards for video transmission equipment quality are the RS-250-B standards, used extensively by the video broadcast industry.

Examples of selected short-haul (from few feet to 20 miles) RS-250-B system performance specifications are shown below:

RS-250-B Specifications

Differential gain	2.0%
Differential phase	0.5^0
Short time distortion	2.0%
Video freq. response	0.15db

The differential gain variations indicates how much the transmission gain varies with variations of the input amplitude (expressed as a percentage of maximum gain).

Differential phase expresses the maximum phase change between any two levels of the luminance signal.

The short time distortion is defined as the distortions of the time components from 0.125 to 1 microseconds and indicates the flatness of the group delay with frequency.

More tests (not shown here) are required to fully characterize the video performance. For example, the chrominance to luminance linear and nonlinear distortions are characterized by the chrominance/luminace gain inequalities, chrominance/luminance intermodulations and chroma delay relative to the luminance. Other tests such as chroma non-linear phase and gain characterize the phase distortions as a function of frequency by the modulator (more detailed explanations can be found in Reference 1).

For example, the results achieved with a broadcast quality TV modulator CTM-20 can be read from Fig. 4.

CTM-20 results

Differential gain	1%
Differential phase	0.5^0
Short time distortion	1.5%
Video freq. response	0.1db

To achieve such high performance it is imperative to design every portion of the modulator chain very carefully. Not only the frequency flatness but the return loss and match (16 db min) at each point must be well preserved, the filters frequency response and group delay must be flat, the video modulator DC response and clamp circuits must maintain the operating point to a fraction of a percent.

The frequency conversion circuits (especially output converters oscillators, mixers and IF filters must provide a flat frequency response and group delay over the conversion range, they must not introduce spurious signals and must not introduce crosstalk from the chroma or audio into the luminance channel (See Ref. 3).

b) VSB-AM performance with noise present

A signal is delivered to its final destination - a broadcast or CATV system customer - by way of transmission over the broadband trunk distribution system.

For 87.5% VSB-AM modulation, an approximate expression (Ref.5) of the (unweighted) video S/No as a function of the C/No is given by,

$$S/N_{0\,(p\text{-}p/rms)} = C/N0_{(rms/rms)} - 4.1\ \text{db} \qquad (1)$$

Adding noise weighting improvement (6.8 db-Ref.1), the weighted S/N0w becomes,

$$S/N_{0w\,(p\text{-}p/rms)} = C/N_{0(rms/rms)} + 2.7\ \text{db} \qquad (2)$$

As the VSB-AM modulated signals are transmitted, their carrier-to-noise ratio decreases with the trunk length.

Further decrease in the C/N_0 is expected at the customer premise location due to no-ideal receivers. It can be shown (Ref.6) that for a receiver with a noise figure F, gain $G_{(db)}$, received level $X_{(dbmv)}$ and input $C/Ni_{(db)}$, the output C/N_0 in 4.2 Mhz bandwidth becomes

$$C/N_{0(db)} = [G+X]_{db} - [GX/(C/N_i) + kTo(F-1)]_{db} - 10\text{Log}(4.2\text{x}10^{6}) \qquad (3)$$

For example, at -6 dbmv and a noise figure of 10 db, for input C/Ni of 43, 46 and 49 db, the output C/N_0 becomes 40.15 , 41.45 and 42.3 db - a C/N loss of 3 - 7db.
Therefore, high S/N_0 ratios for VSB-AM signals can be only be achieved with high C/N_i.

Very good insight of how this modulation method compares with others and in particular with optimum modulation methods can be demonstrated in the following paragraphs.

c) Optimum system performance

If we define the optimum communication system to be the one in which no information is lost, from the Shannon system's capacity theorem the S/N_0 is given by

$$S/N_0 = \{1 + (fm/B)(C/Ni)\}^{(B/fm)} - 1 \qquad (4)$$

in which f_m is the highest modulation frequency (4.2 Mhz for video signals) and B is the channel bandwidth.

Plots of the S/No vs C/Ni is shown in Figure 5.The curve clearly demonstrates the S/No advantages of bandwidth expansion modulation methods, since they trade power for bandwidth exponentially.

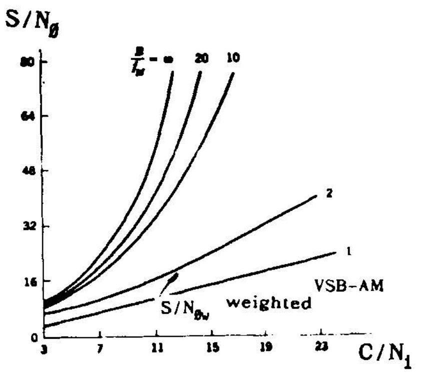


Figure 5 S/No (unweighted) of an optimum communication system

If the design objective is a broadband supertrunk system, to achieve the much higher video S/No required, FM modulation method is almost universally selected.

5. FM video for supertrunks

a) FM video performance with no noise present

To evaluate the video performance of an FM system, standard test waveforms shown in Figure 3 are applied to the input of a Catel FM modulator WFMS 3000 and the degradations caused by the system are recorded. in Figure 7.

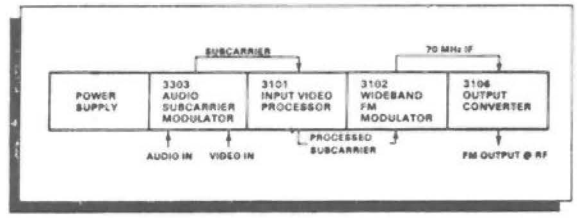
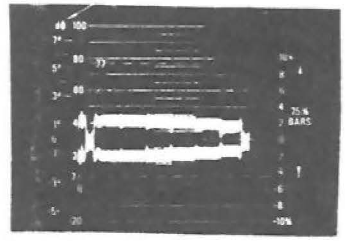
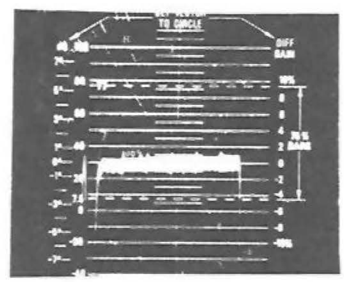


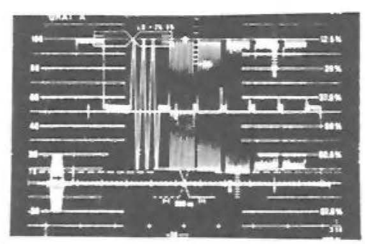
Figure 6 WFMS 3000 FM modulator



Differential phase



Differential gain



Multiburst test

Figure 7 Catel WFMS 3000 modulator response

b) FM system performance in the presence of noise

If analog FM modulation is used to transmit video, we can define the weighted output SNR_{ow} (Ref.2) of a video-modulated carrier as a function of the input (C/N)IF, the modulation index β,the IF bandwidth B_{if}, and the highest baseband video frequency F_m as follows

$$SNR_{ow} = [3\beta^2 Bif/(2Fm)] + (C/N)_{IF} + W \qquad (5)$$

Where $(C/N)_{IF}$ is measured in the IF bandwidth and

$$B_{if} = 2 \times (\Delta F + Fm) \qquad (6)$$

B_{if} is the Carson's rule bandwidth and the modulation index is

$$\beta = \Delta F/Fm \qquad (7)$$

where ΔF is the peak deviation of the video carrier and W is the video weighting improvement resulting from using preemphasis and deemphasis (3.6db with CCIR-405-1 characteristic) , CCIR noise weighting (11.5db) and P-P/RMS conversion (9db).

Note that ΔF is the peak deviation of the carrier by a **sinusoidal** signal with no preemphasis included.Other deviation definitions, used by equipment manufacturers (including Catel) include **sync tip to peak white deviation** $\Delta F_{st\text{-}pw}$. It can be shown (Ref. 4) that the two deviation definitions are related as follows:

$$\Delta F = \Delta F_{st\text{-}pw} / (2x0.3) \qquad (8)$$

If we refer the noise generated by the receiving equipment to the input, the carrier to noise ratio C/N becomes

$$(C/N)_{IF} = Pr/(kT^0_{eq} x B_{IF}) \qquad (9)$$

Where k is the Boltzman constant, T^0_{eq} is the equivalent noise temperature given by:

$$T^0_{eq} = T^0_0 x(F-1) \qquad (10)$$

in which T^0_0 is the ambient noise temperature (300^0K) and F is the noise figure of the receiver.

To estimate the theoretical achievable performance of an FM video modulation system, the SNR_{ow} will be calculated with the following assumptions:

Carrier deviations :

$\Delta F_{st\text{-}pw}$ = 4Mhz , 6Mhz

(corresponding to ΔF = 6.67Mhz, 10Mhz)

IF Bandwidth B_{IF} = 30 Mhz, 40Mhz

Worst case NF : 20 db

Worst received power : -24 dbm

Substituting in equations (5) to (10), the SNR_{ow} becomes:

For $F_{st\text{-}pw}$=4Mhz and IF bandwidth = 30Mhz,
SNR_{ow} = 72 db.

For $F_{st\text{-}pw}$=6Mhz and IF bandwidth = 40Mhz,
SNR_{ow} = 75 db.

Although not shown in the analysis, it has been demonstrated that FM can reject interference from other sources including adjacent FM channels, intermods, crossmods and any other interference not coherent with the in-channel video. This feature permits the system designer to select the modulation to cost-effectively design high performance multichannel video FM systems over broadband supertrunks

Early multichannel FM video systems using peak carrier deviation of 1.6 Mhz and RF bandwidth of 16 Mhz have been used in coaxial cable supertrunks and achieved S/N_0 improvement factors of 10 db over the C/N_i.

With the appearance of cost-effective high performance fiber optic transmitters and receivers and single mode fibers, it become economical to build high performance broadband supertrunk fiber systems. With these systems, it become possible to accomodate more channels and to achieve longer distances without repeaters.

Examples of present state-of -the art high performance wide deviation video FM systems can carry up to 16 channels, each 40 Mhz wide with multiple subcarriers at a distance to 40 km and achieve 65 db video S/No per channel.

Shown in Figure 8 is measured performance of 12 channel video FM fiber optic-based system delivered by Catel.

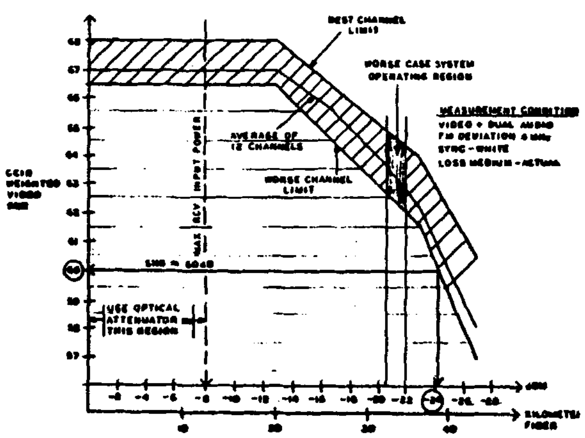


Figure 8: S/No of a 12 channel FM system over a signle mode fiber media

The chart shows that worst case of 60 db S/No per channel at 40 km distance can be achived with worst case received power of -24 dbm and dual audio channels.

6. Flexibility of video FM systems

Because of their high performance, video FM system can also be quite flexible.

For example, the system can be easily adapted to transmit data. Since multi-level PCM data is a video-like signal, PCM multiplexers can be readily interfaced (as shown in Fig.ure 9) with the FM modulator and demodulator permitting high data rate signals to be transmitted over broadband networks.

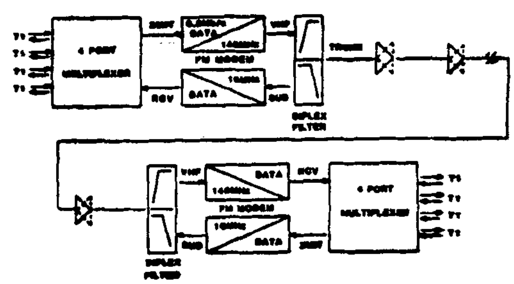


Figure 9: PCM multiplexer interfaced to an FM system

Other type of complex signals (such as BTSC - stereo audio signals) can be conviniently and efficiently carried as subcariers above the audio subcarrier signals.

Even the most complex scrambled signals can be transmitted over an FM system. Converting the scrambled RF signal into a subcarr-like signal and then modulating it permit preservation of the properties of the scrambled signal and does nor require decoding and encoding for supertrunk distribution. Since the modulation is AM/FM however, the FM advantage is lost and the C/N of the scrambled signal is only 3 db better than before downconversion and FM modulation.

7. Comparison to an optimum system

It is instructive to plot the S/No performance of an FM demodulator and compare it with the optimum system performance (shown in Fig.12).

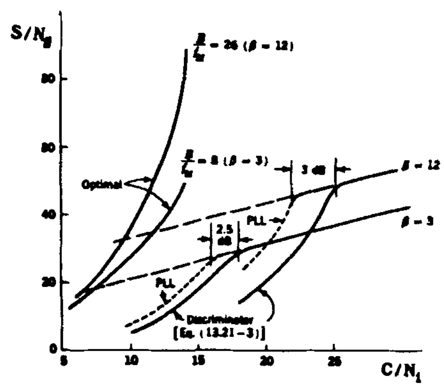


Figure 12 Comparison of FM demodulator performance to the optimum system performance

The results are quite interesting. It can be seen that although large FM improvement factors can be obtained for wide carrier deviations by the baseband video, audio and data, the FM system is quite inefficient when compared to the optimum system - especially at high C/Ni! Unfortunately however,although Shannon theorem gave us the bound on the achievable optimum system performance, it did not specify what is this system!

May be it is a digital video system? What type of digital modulation/coding scheme should we use to achieve this bound? Future research will determine this question.

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Dr. Mircho A.Davidov has received his PH.D. in 1981 from the University of Southern California and Msc and Bsc in 1976 and 1974 respectively from Tel Aviv University.

Dr. Davidov presently directs all of the engineering activities at Catel Telecommunications Inc. developing products in the areas of FM modulators and demodulators for video transmission, VSB-AM modulators and demodulators, FDM-FM fiberoptics transmitters and receivers and frequency translators for LAN data signals.

He was Director of Corporate Research at Oak Industries from 1981-1985. His responsibilities at Oak included developing systems for securing the transmission of high quality video signals. He worked for Honeywell and Brunswick Corporations between 1979-1981 and was responsible for development of RF communication systems for wireless monitoring of energy and control systems and wireless transmission of language translations. Between 1976-1979 he was a consultant to LinCom Corporation in Los Angeles, performing studies in the area of PLL, synchronization and signal processing for satellite communications.Between 1970-1976 he was an engineer for the Israeli Governmental Communication Office,designing circuits for the International Telephone,Telegraph and Telex lines. Between 1967 and 1970 he managed a military communication lab designing communication equipment for the Israeli Defence Forces.

He was an assistant professor at Cal State Northridge in 1981 and teaching assistant between 1974-1979 at Tel Aviv University and the University of Southern California. He speaks 5 languages fluently and is a member of IEEE.

Dr. Davidov has numerous publications and 7 patents pending.

USING IMPULSE TECHNOLOGY TO IMPLEMENT HOME SHOPPING

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ABSTRACT

Cable's impulse technology is a sophisticated, computer-controlled system that has established itself as a consumer-friendly ordering technique. Automated data collection from subscribers requires hardware installed in the home and a refined software package for the addressable controller. Many of the refinements in impulse technologies have been made to provide application flexibility.

While pay-per-view programming has been the driving force behind impulse technology, its application to home shopping is being explored today.

A traffic model comparing home shopping to pay-per-view will explain why the existing store-and-forward technology is capable of handling home shopping order entry. Examining the traffic under peak loading and penetratrion criteria identifies the configuration parameters for addressable computer equipment required to support impulse shopping.

Full addressable control of the subscriber impulse equipment provides the required security. Also, a subscriber code can be used to identify an authorized purchaser. Downloaded activation, credit limits and disconnect commands provide simple purchase authorization/deauthorization.

Finally, as in pay-per-view, impulse home shopping can be phased into a system. The impulse technology can be used in conjunction with phone ordering.

The real beauty is that today's impulse technology can be used for home shopping and pay-per-view, while its flexibility lends itself to other future applications.

TRAFFIC MODEL

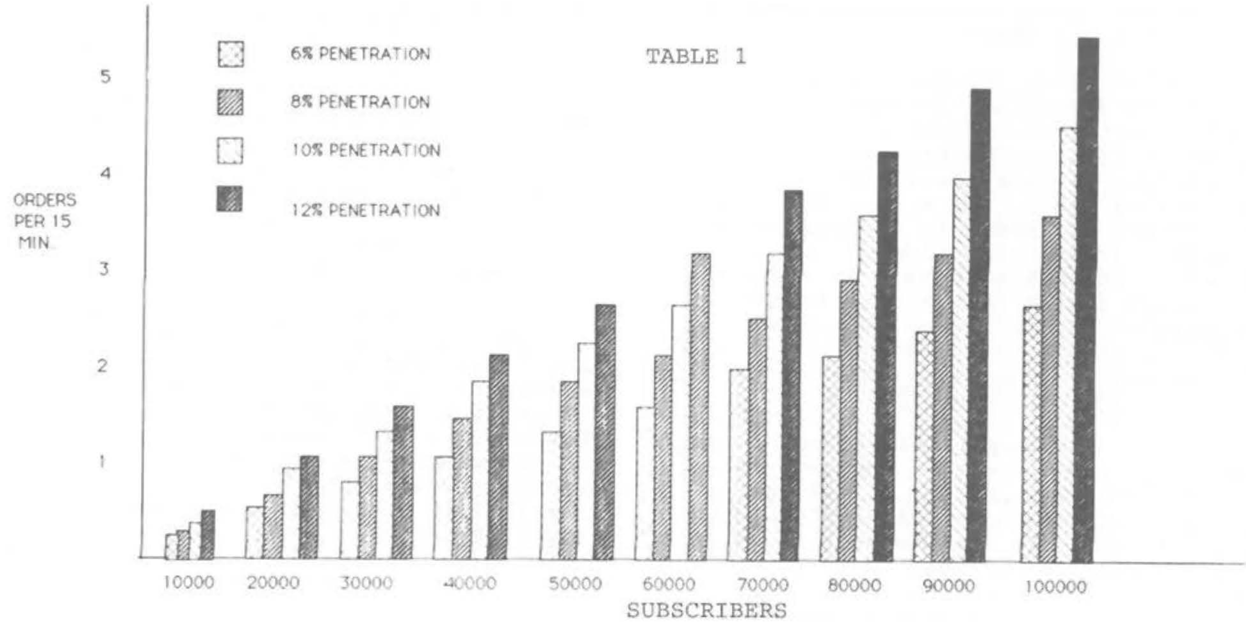
As an overview, there are currently two system configurations today of store-and-forward technology, two-way RF and a hybrid of cable and phone. In both of these systems the addressable computer is in full control of the incoming communications. For the cable return configuration, the computer controls the timing of the communications, and incoming frequency and level. The computer in the phone line return configuration controls the incoming phone line traffic through its polling sequence. The control in both systems is a factor of the software installed in the system and therefore can modified to fit a specific system.

The polling rate in each configuration is dependent on the specific system parameters and the technology in use. Two-way RF is dependent only on the return traffic on the cable system and is therefore the most efficient means of return path. Of course the main issue in a two-way system is the concern of maintaining the return path. The incoming call rate in a hybrid system is dependent on the phone lines in the community, the number of incoming lines to the computer, and the length (number of digits) of the number being dialed.

In the realm of impulse pay-per-view, the peak order load occurs during the 10 minutes before an event begins to 5 minutes into an event. As the impulse system ages and subscribers become accustomed to impulse purchasing, they order events closer to the event start. Thus, buys are truly made on "impulse".

The home shopping industry is truly based on impulsive buying. The products shown on a network are on the air for less than 15 minutes, with an average of 4 to 8 minutes. While the frequency of products on a shopping network, 10 per hour, is much greater than typical pay-per-view programming, one per two hours, impulse technology can meet this application.

To define the average order load for home shopping, the following calculations were made. Home shopping is first divided among a smaller precentage of



subscribers. Typically 6% to 12% of the entire subscriber base are shoppers, while pay-per-view is available to up to 80% of the subscribers - those with addressability. Table 1 shows the average order per 15 minutes which is derived from the following industry averages:

6% to 12% of cable subscribers have made a purchase.

The average subscriber who has purchased will make 15 buys per year.

Table 1 represents the number of subscribers in a system multiplied by the 6%, 8%, 10%, and 12% factor. This figure is then multiplied by 15 purchases per year. The total purchases per year is then divided by 365 days, 24 hours, and finally by 4. The 15 minute time period was used so we could compare it to the 15 minutes of ordering in a pay-per-view environment.

In actuality, purchases from home shopping will peak when a product is first seen on the air, peaking 5 to 12 times per hour. Pay-per-view has experienced a peak load on a single offering of 17%-20% of their subscribers with an average of 3%-5% per offering. Even if all the home shoppers in a system ordered an item at the same time (12% of subscriber base) the number of orders would still be less than the maximum load which has already been handled in pay-per-view.

The response time for home shopping purchase is dependent on the ordering method in use and its associated factors. Currently the most common ordering mechanism is human interface. The controlling factors here are the actual dialing of the phone (with pulse dialing this time is longer than tone), and the actual order taking. The software used by the shopping network allows the orders of previous buyers to be entered quickly. The software relies on the data base for shipping and credit information and requires limited information to place an order. It is common for an operator to be able to place an order in 15 to 20 seconds with a total connect time of 30 seconds. This is then added to the time for connect and disconnect.

A new technology to the shopping network is Voice Response Units (VRU). The speed of this ordering method is again dependent on the dialing method and the education of the buyer. If a subscriber is accustomed to ordering with the Voice Response Unit and they have a programmable touch tone phone, they can speed their ordering. This would require the subscriber to program in the shopping network telphone number, identification code (club number), and possibly a size or color identifier.

While this programming will reduce the ordering time from 15 to approximately 10 seconds, many of the subscribers will either not have programmable phones or will wait for the voice prompts before entering information. Therefore, VRU ordering may reduce the ordering time in some but not all cases.

Ordering with impulse technology not only offers a consumer-friendly ordering technique but also can control the response traffic. The ordering process is dependent on data speed instead of tone dialing rates. Data from the telephone return equipment is transmitted at 300 baud. Transaction time can be reduced to under 5 seconds. In a store-and-forward system though time is not critical.

PREVENTING GRID LOCK

One of the complications seen in the home shopping business is the same as that seen in pay-per-view, local telephone system grid-lock. A "run" on a single product has been known to lock up a local switching office. Not only is this dangerous (emergency calls cannot get through), but valuable orders are lost. With the current order entry system, the phone operators have no control over incoming traffic. Adding a VRU, while reducing order entry time, still does not offer control over incoming traffic. With store-and-forward ordering the addressable computer can control the incoming traffic and prevent grid lock while receiving all the orders.

From the network standpoint the economies of time offered by impulse technology will reduce the grid lock often experienced on their own local switch. Again this grid lock can result in lost orders. Impulse equipment can place orders in their local systems and the addressable controller can be used as a repeater or consolidator. This eliminates the connect and disconnect time for the network's local switching office.

Also important to the ordering process is the identification of authorized purchasers. Current systems utilize a club or identification number to confirm an authorized buyer. A store-and-forward system will support a user selected identification code or "PIN" of up to four digits. This will eliminate orders by non-authorized people (e.g. juveniles). This code can be entered or changed by the authorized buyer for additional security.

For the cable operator, increasing the services (including non-entertainment services), adds value to their transportation system. Impulse can be phased into a system and used in conjunction with phone ordering. Impulse technology can be utilized for pay-per-view and home shopping and its flexibility will lend itself to future applications.

SUMMARY

The benefits of impulse technology's ordering simplicity have been documented in the pay-per-view business. Home shopping, on impulsive business, is the next venture for cable's impulse capability. Order frequency will improve due to impulse technology, impulse-pay-per-view profitted from a two-fold improvement over phone-in ordering. Additional lift in ordering may also be felt in systems which implement impulse shopping and pay-per-view.

Finally, the cable operator will be promoting a new non-entertainment service - Impulse Home Shopping - which is not available to non-cable homes.

Impulse's value is a combination of phone traffic control, consumer-friendliness and, most important, cable exclusivity.

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