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A New Approach to Evaluating CATV System Triple Beat Performance

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As the number of channels carried by a CATV system increases much past twelve, the predominant distortion limiting performance changes from Cross Modulation to Triple Beat products. The mathematics of Triple Beat build-up and its characteristics are briefly reviewed for background. Experimental data on triple beat measurements using various techniques is then presented to illustrate the problems and limitations of current test methods as related to practical system design and operation. Based on further experimental work, a new approach to evaluating CATV system triple beat performance will be discussed, which yields a test technique whose results may be directly correlated to visible degradation of the received picture.

Triple Beat distortion in CATV systems is not something new, but rather has been with us all along. It was just not recognized as such and was generally referred to as "busy background". The recent emphasis on triple beat performance in the industry is due to the increase in the number of channels carried by modern systems, since as the number of channels carried by a CATV system increases much past twelve, the predominant distortion limiting performance changes from cross modulation to triple beat products.

Mathematically, the beats are caused by third order distortion in an amplifier, and are of the form $F_1 \pm F_2 \pm F_3$ and $2F_1 \pm F_2$ where F_1, F_2 and F_3 are discrete frequencies. Cross modulation is a special case of third order distortion where the modulation from one carrier is transferred to another carrier.

It has, therefore, been suggested from time to time that the individual triple beat level, relative to the desired carrier, be established as a more fundamental criterion of amplifier and/or system performance. The question is, what criteria should govern? If we relate to the 51 dB NCTA cross modulation visibility threshold, the corresponding triple beat threshold is given by:

$$(\text{Triple Beat})_{\text{Threshold for X-Mod}} = [\,51 + 20 \log (N-1)\,]\ \text{dB} \qquad (1)$$

where N is the number of channels.

This relationship will generally over-estimate the triple beat requirement since the measured NCTA cross modulation is rarely as bad as one would predict based on a measured triple beat and a simplified theoretical relation between triple beat and cross modulation. (1)

Another approach is based on Arnold's observation (2) that the threshold for picture degradation due to triple beat is given for $B > 50$, by $53 + 10 \log B$, where B is the number of beats falling in the "worst" channel. We will shortly show that $B = 0.034N^{2.6}$ is a fair approximation for the standard channel selections. Thus,

$$(\text{Triple Beat})_{\text{Threshold for Triple Beat Noise with CW Carriers}} = [\,38 + 26 \log N\,]\text{dB} \ (N > 15) \qquad (2)$$

(1) "The Decibel Relationship Between Amplifier Distortion Products" - K. E. Simons, Proc. IEEE, 58, P 1077 (July 1970).

(2) "Required System Triple Beat Performance" - B. Arnold, Dec. 1972.

Yet a third equation is found in the Netherland PTT CATV System Technical Requirements. This states that:

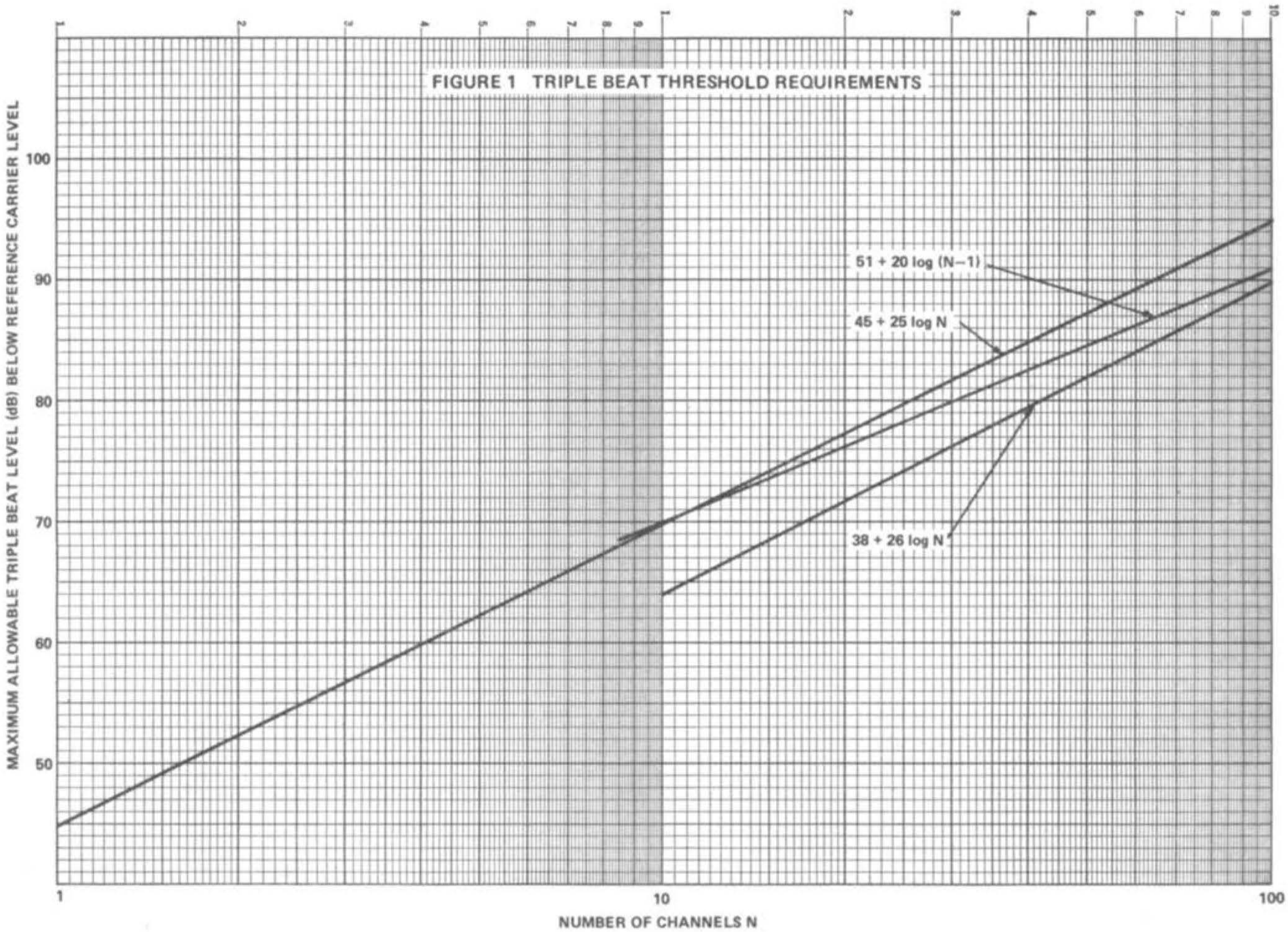
$$(\text{Intermodulation})_{\text{Level below carrier}} \leqq [45 + 25 \log N] \text{ dB} \quad (3)$$

Figure 1 is a plot comparing equations (1) through (3). Note that in order to correctly predict the cross-over between triple beat and cross modulation visibility threshold, equation (1) would have to be dropped 4 to 5 dB. It appears that equation (3) is too severe a criterion although it does have nearly the same slope as equation (2).

Returning to the relationship between number of beats and number of channels, consider the case where the channels are spaced at regular 6 MHz intervals without any gaps. Figure 2 shows that the number of beats is then approximately proportional to N to the 2.2 power ($N^{2.2}$). However, if the standard 12, 21, and 35 channel configurations including channels 5 and 6 are considered, there are fewer beats for a given number of channels(3), and the number of beats is approximately proportional to N to the 2.6 power ($N^{2.6}$). One would expect that as the number of channels becomes very large the 2.6 power curve will asymtotically approach the "no-gap" case wiping out the effects of channel 5 and 6 and the FM band.

For any specific channeling plan, of course, the exact number of beats can be determined on a computer. We have run such a program for a thirty channel system, channels 2-13 and A-R, with three pilot carriers, the results of which are shown in Table 1. For all practical purposes, the results are typical of a 33 channel system. From the Table, it may be seen that the maximum number of beats occurs in a broad

(3) " Third Order Distortion Buildup in a Multi-Channel Cascade" - R. Bell and P. Rebeles, Presented at 1973 NCTA.



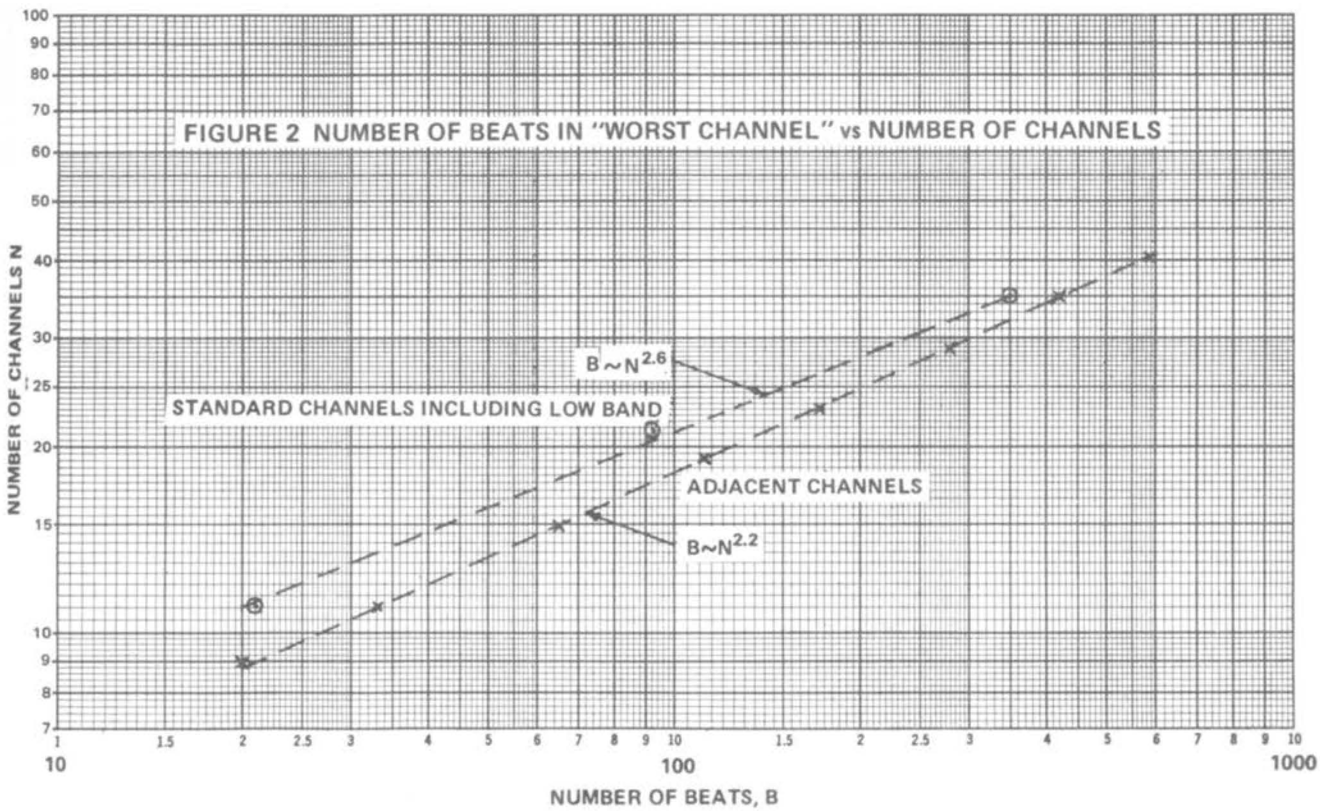


FIGURE 2 NUMBER OF BEATS IN "WORST CHANNEL" vs NUMBER OF CHANNELS

area around channels 7 through 10, and further, that approximately 88% of them occur at the carrier frequency.

We now have some conflicting theoretical number for required triple beat performance and some idea of the number of beats in the "worst case" channel. The next question, of course, is how do these numbers correlate with the visual effects observed in a multi-channel system. Unfortunately, the answer is "a little bit, but not very much."

In order to understand this it is necessary to examine some of the factors associated with three carrier triple beat measurements.

First, there are no industry standards or procedures for this type of test. Thus, it is possible for different people to make these tests using different techniques and come up with completely different answers, which makes comparing specifications or system performance impossible. A good example of this is in calibration where there are currently three popular approaches.

(a) Use of a field strength meter or other receiver to feed an audio wave analyzer, using 100% square wave modulated signals such as NCTA cross modulation as a reference. This is a fairly convenient technique, but can yield errors up to 4 dB since the reference is a double sideband signal, whereas the beat appears as a single sideband. It also requires manually scanning the wave analyzer to locate the beat, which can be very frustrating.

(b) Insertion of another CW signal set a known amount below the carrier, usually 40 dB, and offset in frequency by a small amount to establish a reference on the audio wave analyzer. This is quite accurate but very time consuming. You still have to manually scan for the beat.

(c) Direct observation on a spectrum analyzer. Dynamic range is a problem here and it is difficult to observe close in beats. There is also a question of how much of the beat is generated in the test equipment.

TABLE I
THIRD ORDER BEAT PRODUCTS

Channels

2 through 13
A through R

Pilot Carriers

109.25
271.25
301.25

CHANNEL	$F_1 \pm F_2 \pm F_3$		$2F_1 \pm F_2$		TOTAL	
	At Carrier	In Channel	At Carrier	In Channel	At Carrier	In Channel
2	135	228	10	18	145	246
3	146	237	10	15	156	252
4	154	243	11	17	165	260
5	27	253	0	17	27	270
6	27	258	0	19	27	277
A	237	297	13	16	250	313
B	247	304	12	15	259	319
C	254	308	14	19	268	327
D	262	313	12	15	274	328
E	266	314	14	17	280	331
F	272	319	13	14	285	333
G	275	320	15	17	290	337
H	280	324	13	15	293	339
I	282	325	14	16	296	341
7	287	326	13	17	300	343
8	285	326	15	16	300	342
9	286	324	14	19	300	343
10	285	323	15	18	300	341
11	284	322	14	16	298	338
12	282	318	15	18	297	336
13	278	314	14	17	292	331
J	274	308	15	20	289	328
K	270	304	14	16	284	320
L	265	297	15	19	280	316
M	260	293	14	17	274	310
N	253	287	15	20	268	307
O	247	282	14	17	261	299
P	239	274	14	18	253	292
Q	231	265	14	19	245	284
R	220	256	14	18	234	274
TOTAL	7,110	8,862	380	515	7,490	9,377

A second factor is the choice of frequencies used to make the measurement. An amplifier's triple beat performance is not constant over its operating frequency range.

A third factor is that in normal system operation, signal tilt is employed. This greatly complicates the question of what is an acceptable triple beat level, as is illustrated in Table II.

The first column in Table II is a list of the specific triple beats measured in a typical 16-amplifier cascade. Note that with channels 2, 3 and 4 a difference product in the low VHF band was measured as well as a sum product in the high VHF band. Column 2 is the measured beat level for each group in dB down from the desired carrier. Column 3 is the measured beat level in dB down when each group was set for the tilt, but not the level, that it would have in normal operation. Here we can see that this cascade changes 2 for 1 in beat level with system level quite nicely, even though the levels are substantially above normal. Column 4 shows the calculated beat level which results when the system is derated on a 2 for 1 basis to normal operating levels. It was not possible to actually measure these due to the dynamic range limitation of the test equipment. Finally, column 5 shows the beat level calculated back to a single amplifier at normal levels.

By examining this Table we can see that what appeared to be reasonably consistent performance when everything was measured at a constant level, became a large spread in numbers when reduced to an operational mode. It also points out that the effective operational triple beat performance becomes worse with increasing frequency. This is to be expected since third order distortion does increase with increasing frequency in both discrete and hybrid amplifiers.

One final factor which becomes evident from Table II is that it is generally necessary to use elevated levels in order to get a measurable single beat. In the case of a single amplifier, this may be a substantial amount, driving the amplifier into a completely unrealistic operating mode and generating even higher order distortion products. The multiplying effect of a cascade permits more realistic operating levels, but not too many people can build long enough cascades for practical equipment evaluation.

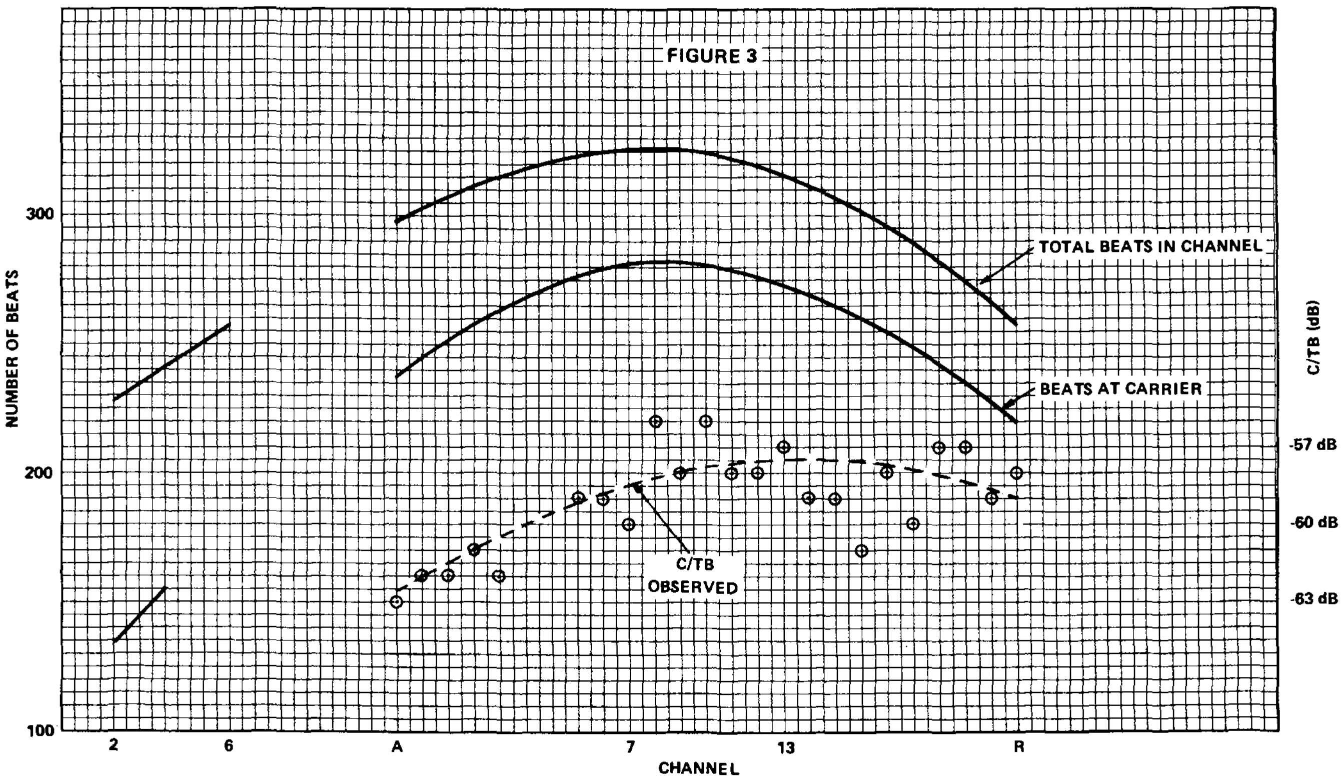
In view of the above factors, plus a lot more experimental data, we at Theta-Com concluded that three carrier triple beat numbers were for all practical purposes, of little value in system design, and that there had to be a better way to handle the problem. The approach we decided to take was very similar to that used in the early 1950's to establish the criteria for cross modulation performance. Namely, to set up an operating system, and, using a large number of observers, determine the visual threshold for triple beat, as well as the tolerance range. Then, using these conditions, devise a measurement technique which would accurately represent the visual effect.

You will recall that in Table I we saw that the number of beats varied from channel to channel and reached a maximum at approximately channels 7 through 10 for a 30 channel system. This data is shown again in Figure 3. From this we selected channel 7 for our test channel, since it is in the maximum beat area, would not require a converter ahead of the TV set, and was not a local channel.

TABLE II

MEASURED TRIPLE BEAT (16 Amplifier Cascade)

Channel	Output Levels @ +40 dBmV	Output Levels @ +40/43/45/47 dBmV	Derate to Normal Level	Single Amplifier
2 + 4-3 on 3	-60	-60	-90	-114
G + I-H on H	-58	-53	-83	-107
9 + 13-11 on 11	-56	-46	-76	-100
S + U-T on T	-54	-40	-70	-94
2 + 3 + 4 on 8	-65	-70 Below Ch. 8	-100 Below Ch. 8	-124 Below Ch. 8



A 32 amplifier cascade, with bridger and two line extenders was then set up for the test. Twenty-nine channels of off the air video from our headend were fed into the system, with channel 7 fed from a modulator with test patterns to obtain a clean, steady signal. A Sony Trinitron receiver/monitor was used as the TV set.

The procedure was then to observe the TV set and adjust the system levels as a group, until triple beat interference was just on the threshold of visibility. This was done with people who could be classed as trained observers (critical viewers), and average viewers.

A first cut measurement was then made by using a standard 600 KHz bandwidth signal strength meter to make a carrier to "Triple Beat Noise" ratio measurement by measuring the peak carrier level and then measuring the noise with channel 7 removed, but all other carriers on. The result, for trained observers, was a carrier to triple beat noise ratio of 46 dB. (The term carrier to triple beat noise ratio quickly degenerated to carrier to garbage, but that is not very technical). The carrier to thermal noise ratio, uncorrected, under the same conditions was 48 dB, so we were not very far above thermal noise, but it was measurable, and repeated very well.

This procedure was then repeated with a spectrum analyzer using various bandwidths, sweep rates and filters to obtain better noise discrimination. As would be expected, the threshold number varied quite a bit depending on the control settings selected, but again, for a given set of conditions, yielded consistent numbers with trained observers. It was also noted that all of the near carrier beats fell in the range of $\pm$ 20 KHz from the carrier, which is in line with the expected range of $\pm$ 40 KHz which could be expected from standard FCC channel assignments. Figure 4 shows the triple beat noise spectrum as displayed on a Hewlett-Packard 8554L/8552B spectrum analyzer with the following control settings:

Bandwidth	1 KHz
Sweep Width	10 KHz/Div.
Scan Time	0.1 Sec. /Div.
Video Filter	Off
Scan Mode	Internal
Scan Trigger	Automatic
Storage Mode	

For contrast, Figure 5 shows the same measurement with 20 channel loading and Figure 6 with 12 channel loading.

All 30 channels were then examined with the

FIGURE 4

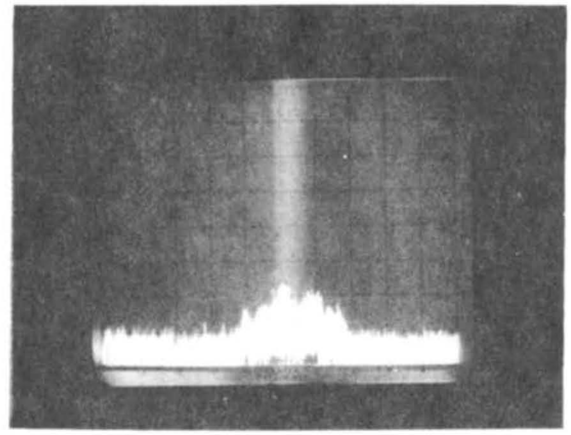


FIGURE 5

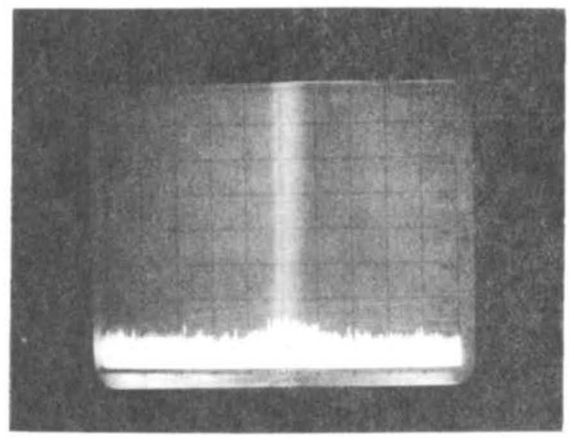


FIGURE 6

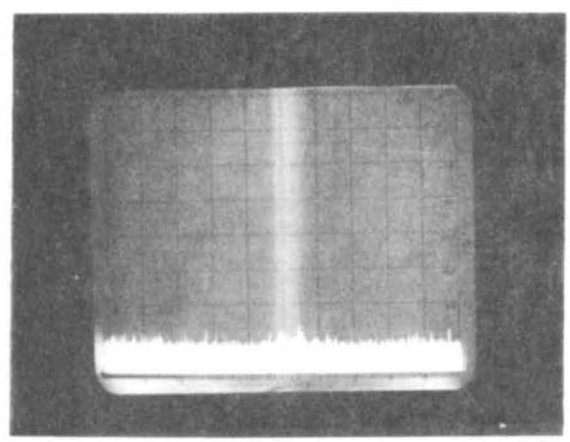


TABLE III

CARRIER TO TRIPLE BEAT NOISE RATIO

30 Channels

Channel	Condition 1		Condition 2	
	Beat (-dB)	Noise (-dB)	Beat (-dB)	Noise (-dB)
2	-	-67	-67	-69
3	-	-63	-63	-65
4	-	-64	-63	-67
5	-	-65	-68	-68
6	-	-65	-65	-68
A	-63	-67	-61	-70
B	-62	-67	-59	-70
C	-62	-67	-57	-69
D	-61	-66	-56	-68
E	-62	-66	-57	-70
F	-61	-66	-55	-70
G	-60	-66	-54	-68
H	-59	-64	-54	-67
I	-59	-65	-54	-67
7	-60	-66	-55	-69
8	-56	-65	-53	-68
9	-58	-64	-53	-66
10	-56	-64	-52	-66
11	-58	-64	-54	-66
12	-58	-63	-54	-64
13	-57	-63	-53	-65
J	-59	-64	-55	-68
K	-59	-64	-55	-67
L	-61	-66	-57	-69
M	-58	-64	-54	-66
N	-60	-64	-56	-67
O	-57	-62	-51	-64
P	-57	-61	-52	-63
Q	-59	-63	-55	-66
R	-58	-62	-54	-65

spectrum analyzer to determine if they followed the pattern predicted by the computer analysis. Two conditions were used. 1) with the system levels set for threshold on channel 7, and 2) with the system levels elevated 3 dB. These results are shown in Table III and also in Figure 3.

The plotted data shows some scatter, which may be due in large part to the visual integration required with the spectrum analyzer display. However, it does generally follow the predicted curve up to the maximum in the channel 7 to 10 region, but does not roll off as fast at higher frequencies as predicted by the computer data. Visual check of all channels also confirmed the plotted data as near as could be seen. The apparently poor values at channel 8 and 10 could not be confirmed visually, and may have been due to off the air pickup, as these are local channels.

During the observer portion of the visual tests for the threshold and tolerance, several interesting characteristics of triple beat were noted:

1. With a test pattern and a trained observer, the break between non-visible and visible beats is quite sharp, about 1 dB in signal level change.

2. With off air pictures and a trained observer, the threshold signal level is about 1 dB higher than the test pattern case, but still quite sharp.

3. With off air pictures and an average observer, the threshold signal level is about 2 dB higher in signal than for a trained observer.

4. Average viewers felt the picture was better (up to a point) with signal levels higher than a trained observer's threshold. This apparently is because these people are more tolerant of "busy background" than they are of thermal noise.

All of the above tests for threshold and measured beats were done with video modulated carriers. Since this is not practical for field testing, the use of CW carriers was investigated. Again, the procedure was to observe a test pattern on channel 7, but with 29 CW carriers on the other channels. The CW levels were then varied relative to the channel 7 level until the same visual effect was noted. The carrier to triple beat ratio was then measured with the signal strength meter and the spectrum analyzer with the result that the visual effect and measured numbers were the same when the CW carriers were 3 dB below normal carrier levels. Thus, this technique is practical for field use with relatively simple equipment, and is being used for proof-of-performance testing in several Theta-Com turnkey projects. The criteria for acceptance is a 46 dB carrier to triple beat noise ratio, with the measured carrier at normal level and the remaining CW carriers 3 dB lower in level, using a signal strength meter.

While the technique of measuring carrier to "triple beat noise" ratio with full channel loading shows great promise for determination of this important parameter, there is a great deal of work to be done before it can be considered as an industry standard. Such factors as measurement bandwidth, dynamic range, detector characteristics and readout, must all be investigated more and correlated with visual observations. Threshold limits with specific instrumentation must be determined and perhaps even special instrumentation designed.

In addition, the technique must be extended to measurement of individual pieces of equipment so that they can have meaningful specifications. And those specifications must have significance to the system designer so that he can predict final system performance.

In conclusion, we have found that:

1. Three carrier triple beat measurement has limited use in determining equipment and system performance.

2. A carrier to triple beat noise measurement can be related to actual system performance and further that the industry should proceed in this direction toward establishment of standards.

A COAXIAL CABLE CONNECTOR DESIGN TO OVERCOME COLD FLOW CHARACTERISTICS OF CABLE MATERIALS

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Coaxial cables used by the CATV industry utilize aluminum for the outer conductor and aluminum or copper for the inner conductor.

Both aluminum and copper are "soft" materials and exhibit cold-flow characteristics under compression.

Electrical connectors used on coaxial cables employ compression connections on both the inner and outer conductors. The very best electrical connectors that exhibit fine electrical characteristics when initially installed will deteriorate as soon as cold-flow occurs. Cold-flow will occur. There is no way to prevent it. How long it takes for this action to be significant depends upon the initial pressures employed in the connector design and the ambient temperatures encountered.

By the very nature of the compression connection, even a very minute increment of cold-flow results in a gap in the circuit. The obvious answer is to back the compression connection up with a spring.

How to do that and still release the cable when desired constituted the problem.

The solution was obtained with a unique clamp design and by taking advantage of the "spring" characteristics of the newer plastic materials.

In the new design every increment, however minute, is immediately compensated for by a spring back-up force. In this manner initial electrical chacteristics can be expected to be maintained for many years of severe service.

Electrical connectors, to most engineers, are a necessary evil. This is due, at least in part, to the tendency of engineers to leave selection of connectors until the very last thing. In every project connectors are taken for granted, and little thought is given to them until trouble occurs.

Engineers should consider connectors a vital part of the CATV system, and give full attention to them in the early stage of design.

JUST WHAT IS A GOOD CATV CONNECTOR?

Cost, appearance, and ease of assembly are all factors to be considered, but a good connector is one that will guarantee against signal distortion under severe usage and adverse weather conditions; one that will continue to do so for years on end.

Perhaps the single, biggest, problem encountered today with CATV connectors is not their initial performance, but deterioration of performance after months and years of usage. What matters cost, appearance, or ease of assembly if a connector doesn't do a good job?

WHY SHOULD AN APPARENTLY GOOD CONNECTOR, WHEN PROPERLY INSTALLED, DETERIORATE?

The answer lies in the materials used to fabricate the cable itself, not in the connector. What do we mean by this? Shall we blame the cable and look for better cable? No, I think not. We would do better if we accepted the cable as it is, and designed our connector to overcome, if possible, or at least live with, the apparent shortcomings of the cable.

The first step is to recognize that all cables utilize copper and/or aluminum for both the inner conductor and the outer conductor. Both are soft materials, and both are subject to large increments of cold-flow. The very nature of a connector magnifies the results of cold-flow in the cable, and this is where the problem starts.

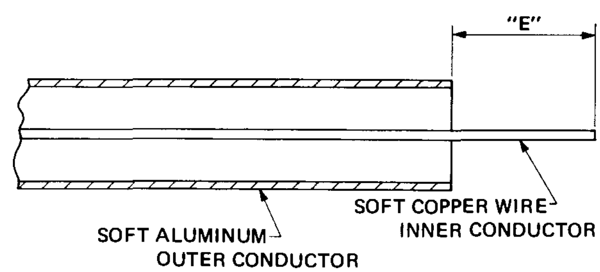

FIGURE 1

We will illustrate this with a few slides. A typical cable is illustrated in Figure #1. In this instance the inner conductor is a solid copper wire and the outer conductor an aluminum tube. Because of difference in the coefficient of expansion between these two materials and the difference in mass between the two components, movement of the center conductor relative to the outer conductor will take place with ambient temperature changes.

Referring to Figure #1, extension "E" on any particular cable is not a fixed dimension. In other words, if not properly clamped in place, the extension of the inner conductor from the outer conductor will vary according to weather conditions and other noncontrollable factors.

What is the solution? There is only one. Clamp both conductors with sufficient forces to overcome the stresses encountered.

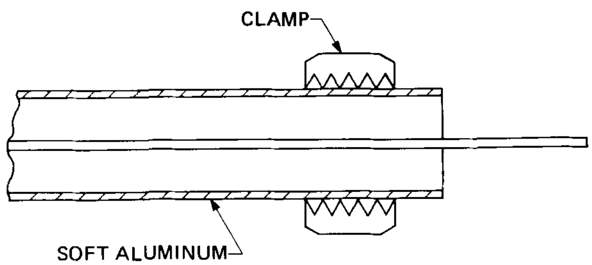

FIGURE 2

In Figure #2 we have a pictorial illustration of a connector. A cable clamp, or grip, is placed around the outer conductor or soft aluminum tube. All coaxial cable connectors are built in this manner. This is just what the doctor ordered to promote cold flow in the aluminum.

What then should we do? For one thing, we can place a stainless steel tube, or sleeve, inside the cable. This is illustrated in Figure #3. It is a big help and certainly improves the life expectancy of the installed connector. It has one serious drawback in that a special hollow mill, or tool, is required to core out the dielectric from the cable. This is a task that must be done by the installer on the job.

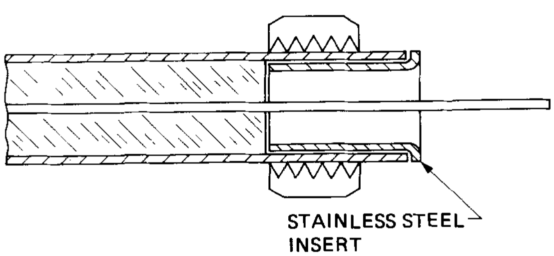

FIGURE 3

As will be seen later, the stainless sleeve becomes unnecessary in the connector design we are describing here.

Our next problem is to secure the inner conductor. This is smaller in cross section, and therefore a larger clamping force will be required to hold it securely. Our clamp must really bite into the copper, as shown in Figure #4.

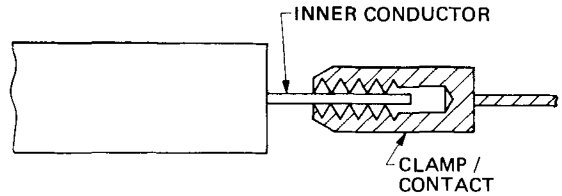

FIGURE 4

Unfortunately, these clamps are also our electrical circuits, and this is the problem. Any increment of cold-flow that occurs, either in the center conductor or the outer conductor immediately shows up as signal distortion. Sometimes this cold-flow doesn't show up for months, or even for years, but it is bound to show up sooner or later.

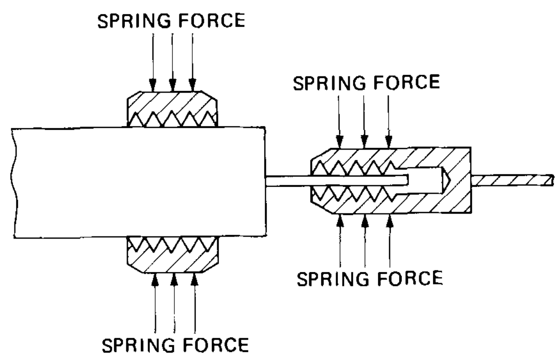

FIGURE 5

So, we again ask the question "What do we do about it?" After careful consideration of a number of ideas, we tried backing up both clamps with springs. This is illustrated in Figure #5. Since this spring force is constantly applied, any increment of cold-flow is at once compensated for. It can be seen that we now have a connector that can "live with" the inherent deficiencies of the cable.

Let us now see how this spring force is applied to the center contact. Figure #6 illustrates a partial section of a cable connector. Item "A" is a conventional brass contact with saw slots to obtain 'tines' or 'fingers' to grip the conductor. The contact is provided with a fine sharp thread to serve as teeth for gripping the copper conductor. The front of this contact is tapered on the outside to match a similar taper on the barrel (Arrow "B"). In the 'free' position the tines on the brass contact will be open and the copper conductor may enter freely. As the insulator "C" is moved forward, the taper on the insulator barrel closes the contact tines around the copper conductor.

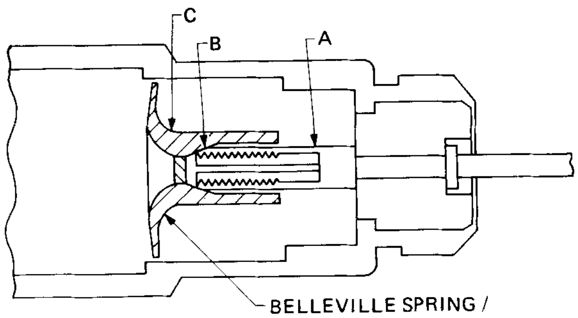

FIGURE 6

We would like to show you now just how a spring is added to the design. First we must look at one of the newer plastic materials. DELRIN is a trade name for Dupont's acetal molding compound. This material has been successfully used in spring applications of all kinds. It possesses a remarkable fatigue resistance. It is fortunate, indeed, for the connector designer to be able to find a material suitable for springs, yet which has good electrical characteristics and good moisture resistance. Best yet, this material performs remarkably well in the form of the well-known Belleville washer.

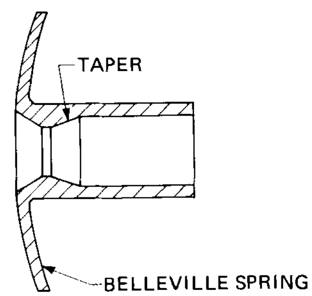

FIGURE 7

Referring to Figure #7 we see that our insulator is a modified Belleville spring, incorporating not only the spring member, but the taper barrel for a contact support. For obvious reasons this is referred to as a "mushroom".

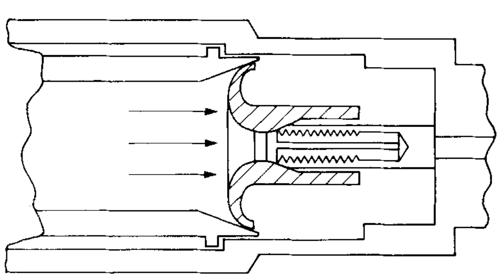

FIGURE 8

Now, in Figure #8 the spring action of this mushroom can be observed. As illustrated in a previous slide, a horizontal movement of the insulator will close the tines firmly around the copper wire conductor. A point is reached where further movement of the insulator is prevented when the teeth of the contact have firmly gripped the copper conductor. A continuing push in the outer periphery of the insulator, as illustrated in Figure #8 will force the insulator into a trueBelleville spring. Load calculations for practical purposes may be obtained by use of a conventional formula. Anyone interested in further information on this subject will find it in an article written by J. H. Crate, and published in a recent issue of Dupont Company's Engineering Design Magazine.

Figure #8 certainly makes it clear that the gripping action of the contact teeth is backed up with a continuous spring action tending to force the teeth forward into any increment of cold-flow that should occur in the copper conductor.

It will be observed that the spring action of DELRIN has been previously utilized by connector engineers, but in all cases this action opens the contact for conductor removal. In this unique design the spring action is utilized to close the contact, NOT to open it. It is easily seen that any spring action tending to open the contact would permit cold-flow increments to remain in the circuit.

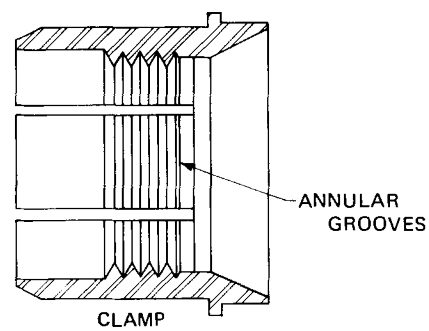

FIGURE 9

The problem has been solved so far as the center conductor is concerned. Our next problem is with the outer conductor, or aluminum tubing. We will see how the spring action was obtained here also. Figure #9 illustrates the clamp, or grip, designed for the aluminum tubing. This clamp is turned from a solid bar of 2024-T4 spring temper aluminum, and is provided with annular teeth for gripping the tubing. Milled slots in the outer wall provide tines, or fingers, for the best type of 360° grip on the tubing wall.

In Figure #10 the clamp is noted as Item "A". A sealing ring, Item "B" is illustrated with a taper cooperating with a taper on the clamp. As the sealing ring is moved forward, the tines on the clamp are forced downward to engage the teeth with the tubing.

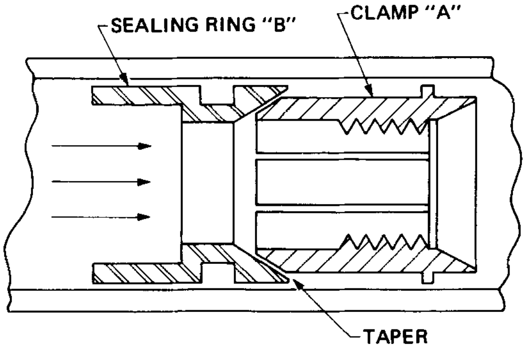

FIGURE 10

Turning to Figure #11 we see that in a manner similar to that described for the contact grip, a point is reached where the clamping action is complete. However, the cantilever action spring tine, Item "B", continues to be "loaded" as it bends in toward the tubing. Here, also, as with the copper wire, the aluminum tubing is constantly under a spring load so any increment of cold-flow in the aluminum is instantly taken up by the six cantilever springs. This provides the long life connector we are seeking.

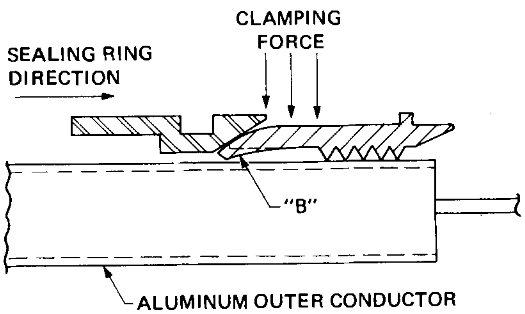

FIGURE 11

We cannot go into the calculations and details of design in this paper; There are many other features of the connector contributing to its' ability to continue year after year without deterioration in signal distortion. Therefore, we have included Figure No. 12 which illustrates a complete connector. A study of this cross-sectional drawing will reveal other reasons why this connector has been so successful in retaining its' initial qualities.

First, the center contact (a) is independent of the other components in that it is rigidly secured into the front insulator (b). The full stress load is transmitted directly to the outer shell (c).

Although the primary purpose of this paper is to show how to overcome the effects of cold-flow of the copper and aluminum conductor materials, there are other often overlooked factors that permit deterioration of performance after initial installation.

Good seals against moisture and dirt are absolutely necessary. "O" Ring seals are very satisfactory if properly lubricated when installed, and if the sealing surfaces are smooth. To properly seal the soft aluminum tubing, or outer conductor, a good rubber compression gasket is needed.

"O" Ring seals are used at (e) and (f) where smooth surfaces are encountered.

We thought you might be interested in seeing the results of some tests conducted on these connectors to be sure nothing was lost in one direction while pursuing another.

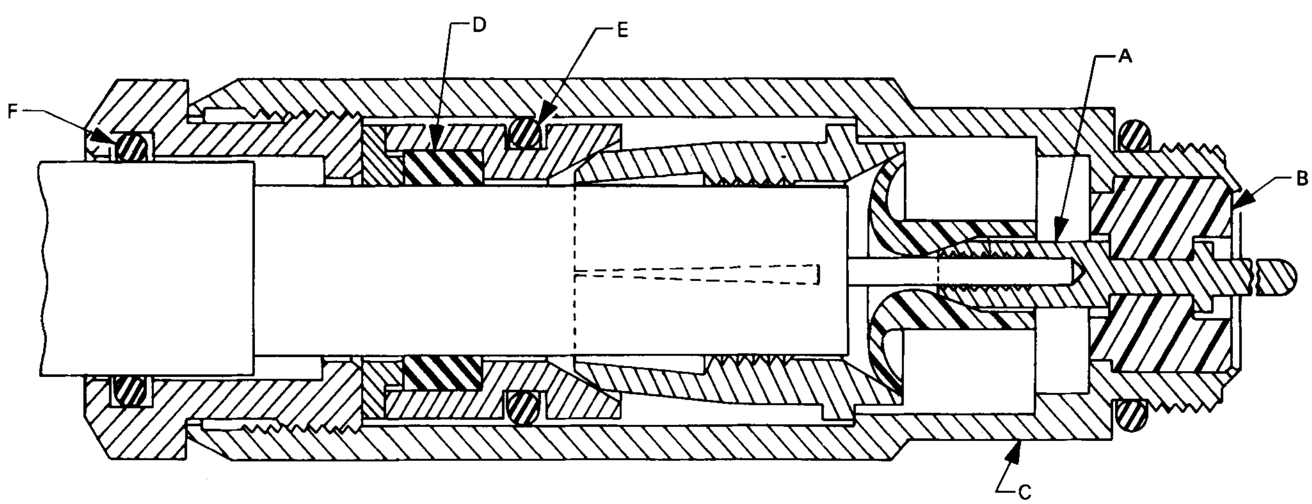

FIGURE 12

Connectors were subjected to environmental tests consisting of:

Salt Spray
Humidity
Vibration
Temperature Cycling
Air Pressure
Ozone
Thermal Shock
Conductor Pull-out

A test specification has been prepared for manufacturing quality control purposes. A copy is available to interested engineers.

Some results of these tests are:

RFI (Closed Loop Method-----	125-130 dbl
RFI (Enclosed Tube Method)--	73 db min.
Return Loss----------------	37-42 db.
Reflection Co-efficient-----	.01 max.
Impedance	75 plus or minus ½ ohm.
Pull out forces	
Center Contact	150 lbs. Min.
Outer Contact	200 lbs. Min.

SUMMATION:

Serious RFI problems and other difficulties, tending to shorten the useful life of CATV connectors have been overcome by adding spring forces to the clamping components of both the inner and outer conductors. These forces tend to eliminate the effect of cold-flow inherent to the "soft" conductor materials such as copper and aluminum.

AN ECONOMIC MODEL OF THE SELECTION PROCESS:

MICROWAVE VERSUS CONVENTIONAL TECHNIQUES IN CATV DISTRIBUTION

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ABSTRACT

When is the use of multi-channel microwave economically attractive? How can the technical trade-offs be converted into dollars? The models used in assessing the parameters of the decision process will be discussed. These include a comparison of alternative methods of signal delivery such as super trunk, separate headends, and multi-channel microwave. Further consideration of the selection process includes estimated signal quality given the alternate distribution methods and the effect of this signal quality on distribution system cost. A method of assessing the viability of the addition of another subscriber area will be discussed.

I INTRODUCTION

On the basis that a cable television system's purpose is to make money for its owners, we can narrow any decision down to an analysis of the financial impact of that decision.

The future profitability of a CATV system can depend heavily on the choice of the type of signal delivery system. Three types of systems will be considered: Super trunk, separate headends and multi-channel microwave. In planning an overall CATV system, the operator has many choices. A choice which appears attractive in the short term, for example, to serve only one community initially, can, in terms of the long term systems viability, be exactly the wrong approach. This is probably most true in small discontiguous systems with several population clusters.

A unique aspect of this analysis will be that the economic impact of the choice of signal delivery system on the distribution system cost is taken into account. Moreover, the need for flexibility of the signal delivery system in allowing the generation of new revenues, either through expansion of the distribution system(s) or the addition of other revenue generating services, is taken into account.

THE ECONOMIC MODEL

Because of the wide range of CATV system configuration possibilities the best possible way to analyze signal delivery systems is with a model which is as generalized as possible. A model, however, is constructed on the basis of a number of assumptions. At the most basic level, for example, linearity is assumed; i.e., two miles of distribution cost twice what one mile costs. The assumptions which were used in constructing this model are discussed below. Signal quality at any subscriber's tap was held constant.

Three methods of signal delivery are considered in the model: 1) Multiple headends, which are attractive where no more than three distribution areas are widely separated and premium TV, imported channels, etc. will not be used. 2) Super trunk, which is most attractive where a single run of ten miles or less serves two or three distribution areas. 3) Multi-channel microwave, which is attractive when three or more distribution areas are to be served, especially if premium TV, imported channels, or system expansion are contemplated.

In the model, only costs which are unique to a particular one of the signal delivery systems have been considered. That is to say, the cost of items which are common to all distribution systems have not been considered. Thus, the cost of subscriber taps and office-related expenses to not appear in this model.

Towards the end of a trunk cascade, line extender cascades have to be kept shorter and physical spacing between them is shorter due to the lower levels at which they are operating. It then follows that when the signal on the trunk is less degraded, then longer line extender cascades and less trunk can be used. Indeed, as the cross modulation characteristics of the signal source at the distribution center improve, up to a certain point, the signal levels on the trunk can be increased with a resultant increase in the strand distance between the bridgers and the first line extender and between each subsequent line extender. In addition, a longer cascade of line extenders can be used. Thus, in an optimized CATV distribution system design it is possible, with high quality low distortion signals, not only to improve the trunk to feeder ratio but also to improve the cost effectiveness of the feeder by decreasing the cable size and/or decreasing the quantity of electronics required. Figure 1 compares the signal quality at the center of the distribution system using various approaches to signal delivery. Figure 2 represents a distribution area fed by a 10 mile super trunk run. Figure 3 represents a distribution system serving the same area but fed from a centrally located multi-channel microwave receiver. Comparing Figures 2 and 3 the difference in distribution system cost is readily seen. It is clear that the system being fed by the higher quality signal will have a lower initial cost as well as a lower maintenance cost.

SIGNAL QUALITY OF DISTRIBUTION SYSTEM

	S/N	Cross Mod 20 Channels
Super Trunk		
5 Miles	48 dB	-73
10 Miles	45 dB	-67 dB
15 Miles	43 dB	-63 dB
Separate Headend*		
After Combining	60	-90
After Two Trunk Amps	53	-83
Microwave** Multi-Channel		
	53	-72 or
	50	-78 or
	47	-82

* Assuming such quality is available at each site.

** Theta-Com AML - S/N vs Cross mod adjustable with one pad.

Figure 1

In an optimized CATV system the cost of the distribution systems will vary with the quality of the signals available from the signal delivery system at the feed points of the distribution system. In this model it was assumed that distribution systems fed with signals with good signal-to-noise ratios and good cross modulation will cost 20 percent less than a distribution system fed with super trunked signals. It was assumed that two miles of super trunk would cause distribution systems to be 15 percent less than normal cost, four miles, 5 percent less than normal, etc. This percentage is based on Theta-Com's experience in costing distribution systems being fed by a high quality signal from a multi-channel microwave receiver as opposed to a 5 or 10 mile super trunk feeding of a distribution area. With regard to maintenance costs the model assumes a 10 percent differential between the high and low quality fed systems.

Plant maintenance costs, when viewed over a ten year period, can have as significant an effect on profitability as the initial cost of the plant. Maintenance costs, therefore, must be included in any examination of the economic impact of a specific course of action. Inasmuch as capital expenditures occur in one or two years, while the maintenance expenses are incurred over a period of time, the cost of maintenance must be brought to its present value. In the model all maintenance expenses are inflated at 4 percent a year and brought to present value by discounting them in accordance with the interest rate entered into the model.

The cost of maintaining a microwave system, whether multi-channel, or single channel has been assumed to be $3,000 per year. This estimate is based on Theta-Com's evaluation of the labor and material cost necessary to maintain AML microwave equipment. The assumption that the maintenance for one or two channels of single channel microwave equipment would be comparable is based on the fact that broadband microwave is only marginally more complex than microwave equipment which is capable of carrying only one channel.

The cost of maintaining a distribution system has been adjusted depending on the number of active devices in the system. Thus, in the model it has been assumed that an optimized CATV system being fed with high quality low distortion signals will use fewer electronics and that the maintenance will be about 10 percent less than it would ordinarily be expected to be.

Headend maintenance in the model has been assumed to be $1000 per headend/yr. Headend costs were assumed to be $26,000 plus $2,600 per channel for a headend with no origination except for a time and weather channel. Where a centrally originated program material, such as premium television, was considered it was assumed that material available at one site would be microwaved by single channel microwave to the other headends. The turnkey cost of the microwave equipment to perform this task is definable in the model by the user. It was assumed that headends would be $7,500 and two amplifiers away from the center of the distribution system.

Super trunk costs were assumed to be directly proportional to the number of miles of super trunk required to interconnect the distribution areas. Super trunk maintenance was assumed to be just as expensive as distribution system maintenance, on a per mile basis.

As a simplifying assumption land acquisition cost for separate headends, super trunk pole rental, and microwave mounting structure rental, were assumed to balance out. They are therefore not included in this analysis.

Multi-channel microwave costs used are those which Theta-Com uses to budgetarily estimate the turnkey cost of a given microwave system. The transmitter, and headend serving it, were assumed to be located independently of the distribution areas. If they were adjoining a distribution area then the multi-channel microwave would be $15,000 less. The cost for LDS using single channel transmitters and single channel receivers has not been considered, but like other factors in the model which are not specifically variable by data entry, this can be modified in the program itself.

The model is shown as Appendix A. It is written in basic and has been run several hundred times on a Wang 1200 Mini Computer. A sample input and output is shown as Figure 4. The inputs are underlined.

ADDITION OF A DISTRIBUTION AREA

After the model has analyzed the comparative cost of the three methods of signal delivery the marginal cost of a new service area can be seen by entering "yes" in response to the appropriate question. Anytime in this model that the number of years of maintenance is equal to zero, the model output will display plant cost per subscriber. The plant cost per subscriber will indicate the viability of the additional distribution area being considered. See Figure 5

CALCULATED RESULTS

In order to compare the results of this model with previous comparisons of the cost of the signal delivery systems alone, three, four and five distribution area systems with 6 total miles of distribution plant were evaluated. The model was run in order to develop a number of points that indicate the comparative advantage or disadvantage of the various signal delivery systems. Figures 6, 7 and 8 show the system cost with three, four and five distribution areas.

The assumptions used in running this system in the model were that super trunk costs $7,000 or $9,000/mile, that the "standard" distribution system costs $5,500/mile and that aerial maintenance costs $350/mile/year. Interest was assumed to be 12 percent and the single channel microwave was assumed to be $15,000 at the transmit site and $18,000 at each sub-headend.

A study of these graphs will show that multi-channel microwave can often be more cost effective than separate headends or super trunking. Multi-channel microwave is often more cost effective than any other signal delivery system, even in system configurations where microwave might ordinarily not have been considered.

ADDITIONAL CONSIDERATIONS

There are a number of factors contributing to the economic selection of a signal delivery system which have not been dealt with in the model.

1. Significant extra cost in using one signal delivery system in a given situation due to large physical obstructions to super trunking, inadequacies of off-air signals at a headend site, and a microwave path obstruction.

2. Cost of separate headends due not take into account the lower subscriber

penetration resulting from a lack of non-broadcast material available at the separate headends, or conversely, the expense of bringing more than one channel of microwave from a central site to the headends. The cost of just one imported channel at $400 per month at one site over a ten year period is $58,000, at present value.

3. Signal delivery system flexibility in the addition of new distribution areas, or expansion of existing areas.

4. Speed of installation for earliest subscriber revenues.

GENERALIZED CONCLUSIONS

Running this model several hundred times has resulted in some generalized conclusions which were not previously apparent. Multi-channel microwave is economically viable in most cases where it might be considered, providing that distribution areas being served have not been built or will be rebuilt shortly and that the system design will take advantage of the high quality low-distortion signals. One basic exception is the case where a single super trunk run of 10 miles or less in length can serve several distribution areas, provided it is assumed that no expansion in the number of distribution areas is required.

SUMMARY

The CATV operators' primary goal is to manage profitable systems. In order to generate as much profit as possible it is necessary to evaluate the signal delivery system's impact on total system cost. The model shown in Attachment A will provide a means of analyzing the cost implications of various signal delivery systems. If some of the assumptions in the model do not agree with your experience then the program can be adjusted accordingly. For the majority of the cases considered in this paper multi-channel microwave is the most economical approach.

APPENDIX A

```
10 Input "Number of Channels", C
20 Input "Interest in Percent", Il:Il=
   Il/100
30 Input "Super Trunk Cost", Sl
40 Input "Dist Cost--Conventional", Dl
50 D2=Dl*.8
60 Input "Aerial Maintenance", D3
70 S3=D3
80 Input "Super Trunk Miles" S
90 Input "Distribution Miles", D
100 Input "No Of Systems", N
110 Input "Pay TV MW TX", Pl
120 Input "Pay TV MW RX", P2
130 M9-0
140 Input "Number of Subscribers", S8
150 Input "Number of Years of Mainte-
    nance", M8
160 Print, "-Super Trunk -"
170 Print, "Basic", "Premium TV"
180 Hl=C*2600+25000:A=Hl
190 If M9>0 Then 210
200 Goto 220
210 Print "Headend", "0":P=0:A=0:Goto 250
220 Print "Headend", Hl
230 E=1000
240 Gosub 480
250 Print "He Maint", P
260 A=A+P
270 S4=Sl*S
280 Print "Super Trunk", S4
290 A=A+S4
300 E=S3*S
310 Gosub 480
320 Print "St Maint", P
330 A=A+P
340 D4=Dl*D:S5=S/N:If S5  5 Then 360
350 D4=D4+(.75+(S5x.05)):D4=Int(D4)
360 Print "Dist Syst", D4:A=A+D4
370 E=D3*D
380 Gosub 480
390 Print "Dist Maintenance", P
400 A=A+P
410 Print "Total", "0", A
420 A8=A/S8
430 A8=Int (A8)
440 If M8>0 Then 460
450 Print "$/Sub", A8, "0", A8
460 Stop
470 Goto 580
480 If M8  0 Then 500
490 P=0: Return
500 P=0
510 For T=l to 11
520 If T=M8+l Then 570
530 P=P+(E/(1+I↑T))
540 E=E*1.04
550 Next T
560 Goto 510
570 P=Int (P): Return
580 Print, "-Separate Headends-"
```

```
590 H4=H1*N
600 Print, "Basic", "Premium TV"
610 A=14
620 P4=P1+(P2*N)
630 If M9>0 Then 650
640 Goto 660
650 P4=P2
660 Print "Headends", H4, P4
670 E=N*1000
680 Gosub 480
690 H3=P:A=A+H3
700 E=3000
710 Gosub 480
720 P3=P
730 Print "He Maint", H3, P3
740 1=7,500*N: Print "Headend Runs", 1
750 E=S3*N: Gosub 480
760 Print "He Run Main", P
770 A=A+1+P
780 D4=D2*D:A=A+D4
790 Print "Dist Syst", D4
800 E=.9*D3*D
810 Gosub 480
820 A=A+P
830 Print "Dist Maint", P
840 P5=P3+P4
850 X=A+P5
860 Print "Total", A, P5, X
870 A8=A/S8:P8=P5/S8:X8=X/S8
880 A8=Int (A8):P8=Int(P8):X8=Int(X8)
890 If M8>0 Then 910
900 Print "$/Sub", A8, P8, X8
910 Stop
920 Print, "-Microwave Dist-"
930 If M9=0 Then 950
940 H1=0
950 Print "Headend, H1
960 E=1000:Gosub 480
970 If M9=0 Then 990
980 P=0
990 A=H1+P
1000 Print "He Maintenance", P:M1=C*
     2200+(N*14580)
1010 If C<7 Then 1050
1020 If C<15 Then 1060
```

```
1030 If C<23 Then 1070
1040 If C<30 Then 1080
1050 M1=M1+32605:Goto 1090
1060 M1=M1+48385:Goto 1090
1070 M1=M1+65855:Goto 1090
1080 M1=M1+78435:Goto 1090
1090 If M9>0 Then 1110
1100 Goto 1120
1110 Print "Microwave", M9, "0": M1=M9:
     Goto 1130
1120 Print "Microwave", M1, "3200"
1130 A=A+M1
1140 E=3000: Gosub 480
1150 If M9=0 Then 1170
1160 E=500:Gosub 480
1170 Print "MW Main", P
1180 A=A+P
1190 D4=D2*D
1200 Print "Dist Syst", D4
1210 E+.9*D3*D:Gosub 480
1220 Print "Dist Main", P
1230 A=A+P+D4
1240 If M9>0 Then 1260
1250 P8=3200:Goto 1270
1260 X=A:P8=0:Print "Total", A, "0",
     X:Goto 1290
1270 X=A=3200
1280 Print "Total", A, "3200", X
1290 A8=A/S8:P8=P8/S8:X8=X/S8
1300 A8=Int (A8):P8=Int (P8): X8=Int
     (X8)
1310 If M8>0 Then 1330
1320 Print "$/Sub", A8, P8, X8
1330 Stop
1340 Input "Do You Want To Add A Distri-
     bution Area", A$
1350 If A$ >"No" Then 1370
1360 Gosub 160
1370 Input "Super Trunk Miles", S
1380 Input "Distribution Miles", D
1390 Input "Number of Years of Mainte-
     nance", M8
1400 Input "Number of Additional Subs",
     S8
1410 N=1:M9=14580
1420 Gosub 160
```

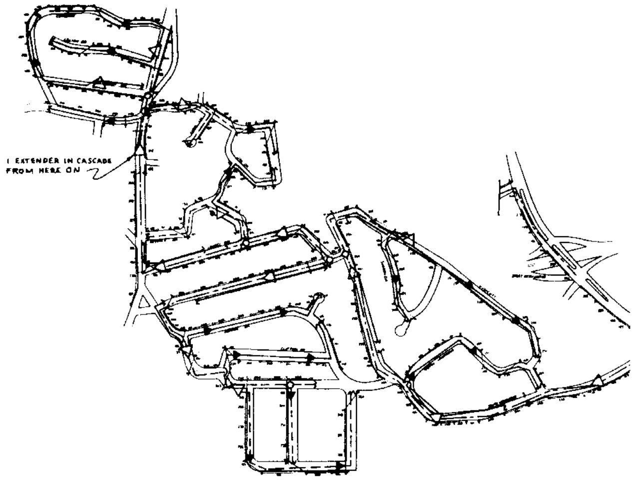


FIGURE 2

DISTRIBUTION SYSTEM FED BY 10 MILES OF SUPER TRUNK
Active Equipment: 13 Trunk Stations
18 Line Extenders

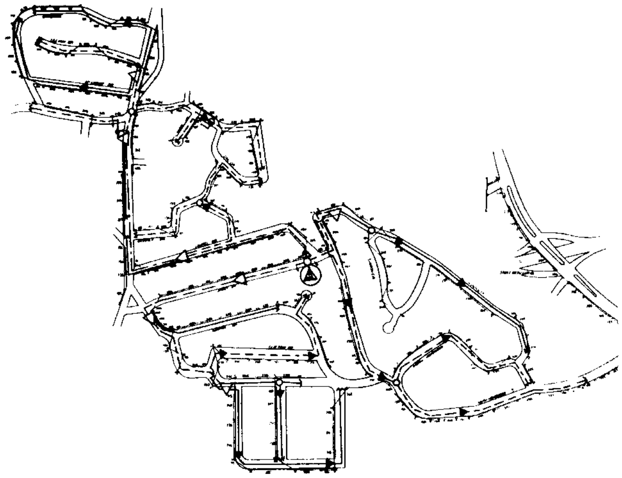


FIGURE 3

DISTRIBUTION SYSTEM CENTER FED WITH LOW DISTORTION SIGNALS
Active Equipment: 7 Trunk Stations
15 Line Extenders

FIGURE 4 SAMPLE INPUT AND OUTPUT PRINT-OUTS OF THE ECONOMIC MODEL

Run
Number of Channels? 6
Interest In Percent? 12
Super Trunk Cost? $9,000
Dist Cost--Conventional? $5,500
Aerial Maintenance? 350
Super Trunk Miles? 30
Distribution Miles? 100
No Of Systems? 4
Pay TV MW TX? 15000
Pay TV MW RX? 8000
Number Of Subscribers? 4500
Number Of Years Of Maintenance? 10

-Super Trunk-

	Basic	Premium TV	
Headend	40600		
He Maint	11882		
Super Trunk	270000		
St Maint	124761		
Dist Syst	550000		
Dist Maintenance	415872		
total	1413115	0	1413115

Stop
Continue

-Separate Headends-

	Basic	Premium TV	
Headends	162400	47000	
He Maint	47528	35646	
Headend Runs	30000		
He Run Main	16634		
Dist Syst	440000		
Dist Maint	374285		
Total	1070847	82646	1153493

Stop
Continue

-Microwave Dist-

Headend	40600		
He Maintenance	11882		
Microwave	104125	3200	
MW Main	35646		
Dist Syst	440000		
Dist Main	374285		
Total	1006538	3200	1009738

Stop

FIGURE 5 SAMPLE INPUTS AND OUTPUTS FOR AN ADDITIONAL DISTRIBUTION AREA

Continue
Do You Want To Add A Distribution Area? yes
Super Trunk Miles? 8
Distribution Miles? 20
Number Of Years of Maintenance? 0
Number Of Additional Subs? 1000

-Super Trunk-

	Basic	Premium TV	
Headend	0		
He Maint	0		
Super Trunk	72000		
St Maint	0		
Dist Syst	110000		
Dist Maintenance	0		
Total	182000	0	182000
$/Sub	182	0	182

Stop
Continue

-Separate Headends-

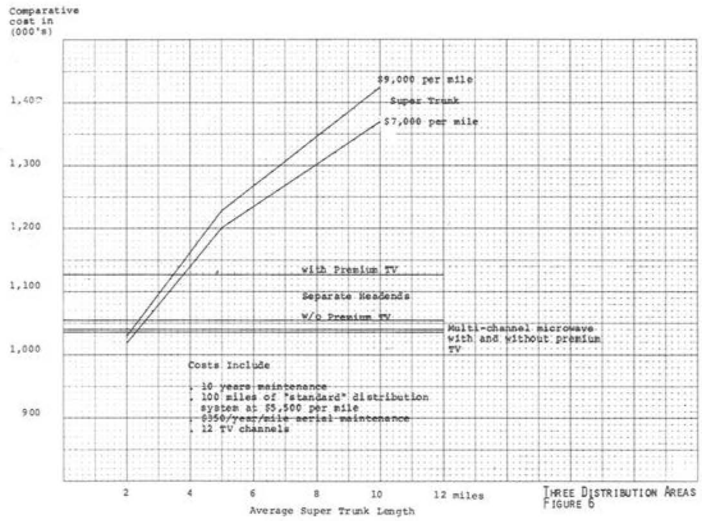
	Basic	Premium TV	
Headends	40600	8000	
He Maint	0	0	
Headend Runs	7500		
He Run Main	0		
Dist Syst	88000		
Dist Maint	0		
Total	136100	8000	144100
$/Sub	136	8	144

Stop
Continue

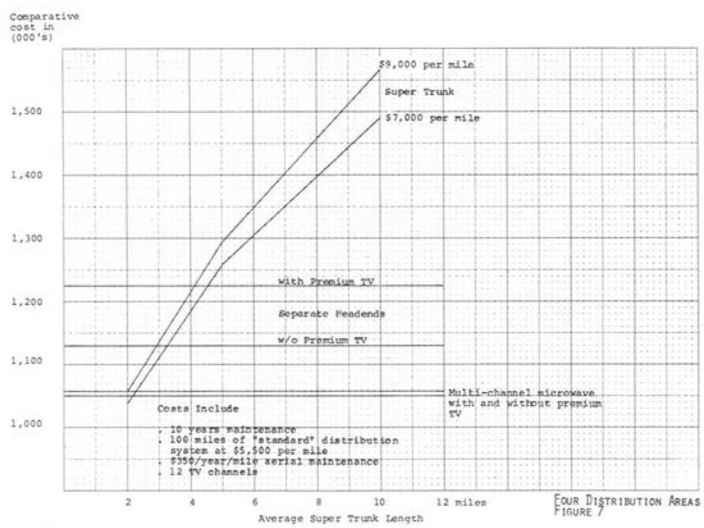
-Microwave Dist-

Headend	0		
He Maintenance	0		
Microwave	14580	0	
MW Main	0		
Dist Syst	88000		
Dist Main	0		
total	102580	0	102580
$/Sub	102	0	102

Stop



THREE DISTRIBUTION AREAS
FIGURE 6



FOUR DISTRIBUTION AREAS
FIGURE 7

AURAL SERVICES FOR THE FM BAND

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ABSTRACT—Most CATV systems carry FM signals from direct "off air" or microwave delivered sources. This paper reviews several additional services which can contribute toward the development of a total CAFM (Community Antenna FM) concept. Included is a discussion of the local origination of monaural, stereo and quadraphonic programming TV/FM simulcasts, and special services, such as, standard time, weather and international short wave rebroadcast.

Most CATV operators carry "off air" or microwave delivered FM broadcast signals on their systems. Traditionally, FM has been treated as an incidental service to television. This is understandable, but with CATV systems now reaching a high level of technical performance, and with the need to find new revenue sources, many operators are now placing greater emphasis on FM services. With 80% of all U.S. homes equipped with an FM receiver[1] and a music oriented public, the CAFM (Cable FM) concept ideally complements CATV.

Since this session deals primarily with local origination, and on the specific topic of Aural Services for the FM Band, the discussion will be limited to services in the 88-108 MHz. frequency range, although there are tempting opportunities for use of aural signals in other parts of the spectrum.

An effort will be made to acquaint you with special aural services, which are being used, or can be used, to offer new personal and community benefits as a means of attracting and maintaining subscribers on a cable system.

The objective of an effective CAFM program, as with cable television, is that of offering signals of a variety and quality not readily available to the subscriber on a direct pick up basis. The following are some examples of aural programming which can meet that objective:

An activity which is growing very rapidly is that of CAFM cablecasting by colleges, universities and other schools. A recent article[2] in the Journal of College Radio indicated that there are approximately 50 such stations now operating, with an estimate that this number will reach 75 by the end of 1974. In 1970 there were only 10 stations.

There are hundreds of campus radio stations using carrier current for on campus broadcasting. Since these stations have audio equipment, and regular operating schedules, all they require to expand their coverage to the local community is the addition of an FM modulator and the cooperation of the local cable system operator. Most of the school stations use the combined carrier current - CAFM technique.

In the IBS survey it was shown that 31% of the stations are operating in stereo. Of the college radio stations using "on air" broadcasting with 10 Watt (Class D) stations, only 5% are using stereo. Over 3/4 of the schools own the origination equipment, and all of them report that they pay rental of telephone lines when required, to the CATV "head end". It is apparent that the CAFM approach makes it economically possible to offer stereo and also, and perhaps most important, the signal-to-noise, problems with a 10 Watt transmitter have limited most stations to monaural programming.

For the cable operator who is looking for a first step beyond "off air" signal carriage, he should look to local schools where he will often find active interest, talent, and in many cases, the equipment to begin CAFM origination.

The high level of student interest in programming, originating from their school, insures many additional FM "hook ups" to the cable system.

Connection with the CATV system usually involves special telephone lines from the campus station to the FM Modulator at the "head end", although in some cases direct access is provided through the cable system. This latter approach is especially attractive for stereo origination where dual line charges may be beyond the limited school budgets.

Recent changes in telephone tariffs offer a concept of community "cable radio" services which should be considered in any CAFM aural origination program. Under these tariffs it is possible to use direct dial lines for remote radio broadcasts, so access from any location in the community can be offered, simply and inexpensively, for any "voice grade" programming.

There are many relatively inexpensive remote units being used by broadcast stations. One, which also serves as a conference telephone, can be leased from most phone companies for approximately $15 per month, plus an initial installation charge. A direct dial remote can provide access to the FM Modulator, from the local football field, little league ball park, mayor's office, church, service club meeting room or law enforcement headquarters. A number of police headquarters are now linked with other agencies using direct or telephone interface with the cable system.

This highly flexible system of community "cable radio" is available as one of the very inexpensive services which can increase subscriber interest and, as an extra benefit, perhaps reduce demands for television remotes. Just as cable television origination can serve a narrower area of interest than broadcast television, cable radio can economically reach an even smaller segment. One can envision serving special interest groups of only a few members, where cost of origination equipment would make television prohibitive.

When looking at cable FM origination it is well to consider some of the technical factors which make cable a potentially superior means of transmitting high fidelity musical programming.

Most of us have experienced the rather dramatic difference between a musical performance heard in a concert hall, with that received in the home or automobile FM receiver. This difference is especially evident when an attempt is made to record "off air" musical programs. The "flatness" or "lack of brilliance" of the sound reproduction is a problem in dynamic range, where the sound recording and transmission system cannot match the capability of the human ear to discerne differences in level.

The sound intensity range of the large symphony orchestra can be as high as 70 dB (a power ratio of 10 million to 1) and a good high fidelity amplifier has a range of 60 dB, so the problem is in the transmission medium, or more precisely, what is done to the original signal before it reaches your ears.

In an article entitled "Who's Monkeying With Your FM Signals? (3), Peter E. Sutheim of Radio Station KPFK, Los Angeles, speaks of the many signal processing steps which occur between a live performance and the human ear as "tinkering" with the sound beyond the straight amplification and transmission of the musical signal.

The reason for this "tinkering" is the economic pressure on an FM broadcaster to achieve maximum coverage, approximately (4.5 million units sold in 1973) (3) especially to the growing number of listeners using automobile FM Stereo receivers. In the conflict between fidelity and marketing, it is understandable that concessions will be made in fidelity, except in the case of non-profit broadcasters who, being less subject to economic pressure, often can offer a level of fidelity superior to their commercial counterparts. Mr. Sutheim points out that the effort to be the loudest station results in "tinkering" with gain and frequency response to achieve an increase in apparent loudness while not overmodulating the transmitter. Since it is estimated that maximum useful dynamic range for listening to music in an automobile is only 20 dB, and with much of the FM stations market on wheels - the compromise with fidelity is understandable.

Using cable for origination makes possible the transmission of music with no need for processing beyond the normal pre-emphasis to be compatible with long standing FM broadcast requirements. Compression and special limiting techniques, to achieve maximum coverage, are not necessary in FM cablecasting, so truly high fidelity transmission is possible, for the home recording enthusiast, or for the critical listener, whether of Rock or Bach.

Thus far, we have spoken only of the transmission of mono and stereo, by cable. With the high fidelity industry actively promoting quadraphonic (4 channel) sound, we are certainly faced with this kind of transmission as part of the future aural services from cable. Cable transmission, with its capability of providing good signal to noise and freedom from multi-path interference, can offer improved stereo, as compared to "off air" broadcasts. For the same reasons it is expected that quadraphonic signals, with even more critical demands on precise phase and level relationships, can be better delivered by a cable system than received "off air".

Although we are not in a position to get involved in the matrix vs discrete quadraphonic issue, our company has, for the past year or so, been demonstrating the origination of matrix (SQ) quad at various CATV conventions. We are now involved with a consumer products manufacturer and a major west coast cable system in testing CD-4 (Discrete) quadraphonic transmission by cable.

So much for cable FM origination---How about some ways in which aural services can add to the effectiveness of Cable television origination?

Not being directly involved with the problems of television origination perhaps limits our perspective, but it is difficult to understand why TV/FM simulcasts, with the audio portion of TV origination carried simultaneously as Hi-Fi mono, or stereo, along with its' conventional mode, has not been used by system operators.

Every business man agrees that for any product to be successful, it must be capable of satisfying a need. We think that TV/FM simulcasts - in CATV origination, have an opportunity of doing just that - for a high fidelity conscious audience.

With the high level of perfection of the picture quality of TV color presentations, the remaining weak spot in television is the sound. Although the television broadcast industry is looking at ways to improve sound, any change will involve much time with perhaps a whole new generation of TV receivers, and certainly CATV doesn't need any more long pay off items. In the early days of FM stereo, simulcasts, in which the AM station transmitted one channel and the FM another, proved quite popular, to this suggests an approach for cable. In this case, however, we are not suggesting splitting the channels for stereo, but rather simultaneous sound transmission on TV and any 88 to 108 MHz channel.

The reported success of the ABC Network, "Wide World of Entertainment" rock concerts, and some special classical music programming, where the TV station carries the conventional sound and picture, and a local FM station carries the stereo sound - has demonstrated the public wants this kind of programming. CATV is in a position to offer television high fidelity on a regular basis. A simple and relatively inexpensive FM modulator makes it possible to give the subscriber a service that is not regularly available from "off air" sources. Putting the local origination audio on the FM band as a monaural signal can greatly improve programs with high musical content, although stereo would, of course, be preferred. Also, using simulcast, pay television can offer greater appeal by providing "theater like" sound, instead of limiting the sound portion of a movie to the capability of the audio and acoustical system in the TV set.

In addition to local origination, we feel that "off air" television should be carried on a cable simulcast basis - to offer an option to those who want to use their high fidelity equipment to listen to the Boston Pops, Lawrence Welk, or a Broadway Show as a TV "Special". This, however, would get us off into a discussion that is not part of local origination - and possibly a commercial emphasis not appropriate for this paper.

The CAFM origination ideas we have presented involve people and equipment which can represent a substantial investment in getting started, as well as in sustaining the program. There are, however, a number of excellent aural services that can be added to your FM system that are available with a minimum of equipment, and that don't require continuing "people" involvement.

The National Oceanic and Atmospheric Administration, National Weather Service, now provides 24 hour per day weather broadcasts in most major coastal areas. This service is being rapidly expanded to other key population centers. These broadcasts are on VHF-FM frequencies which are not regularly available, without the use of special receivers. Many consumer goods manufacturers are offering the weather channel options, in new models, and this provides verification of public interest in "radio weather".

By converting the weather broadcasts to the standard 88-108 MHz FM band and carrying the signal on an unused FM channel, this extremely useful public service is made available to thousands of additional homes. Also, since the weather broadcasts include signals which may be used to alert the listener to weather emergencies the cable system can provide a means of using these signals to initiate warning procedures and to insert weather and civil emergency information.

For many years the National Bureau of Standards has broadcast time and frequency information by high frequency radio station WWV. Although the transmissions have been widely used by those who had a specialized need for such information, the use by the public has been limited because of the relatively high cost of receiving equipment and the amount of technical knowledge required to properly interprete the signals. Recently the format of the standard broadcasts has been changed with features such as precise time announcements each minute and the elimination of some of the encoded information which was confusing to the public. Like most everything else in our changing world, however, the WWV broadcasts show a new image in announcing time as "Coordinated Universal Time" instead of the Greenwich Mean Time (GMT) reference we have known through the years.

Now that watches are being sold which use quartz crystals and even digital readout, it would seem that the ultimate step in "conspicuous consumption" is that of knowing that ones precision time piece was checked daily - using a reference standard that is accurate to approximately one part in 10^{11}. Seriously, the availability of a readily available time standard has proven of subscriber interest in a number of systems, as have the weather, propagation forecasts and standard audio tones available from WWV.

To add the time broadcasts to the FM band, the cable operator needs a high frequency receiver, appropriate antenna system, and an FM modulator. Simultaneous monitoring of night time and day time frequencies, with automatic change over to the best signal would be desirable, but CATV systems, using the standard time broadcasts, are realizing satisfactory results using a manual change to the best channel, for a given distance from the transmitter, and the time of day and year. A sattelite transmission system to provide relay on a VHF frequency has been tested by the Bureau of Standards and will no doubt simplify reception problems in the future.

In spite of the tremendous educational and cultural content of international short wave broadcasts, and the long history of such services, their use has been limited to a relatively small segment of our population who have the equipment and the knowledge to cope with the complexity of high frequency signal propagation. An estimate by the publishers of World Radio/TV Handbook shows 1300 short wave broadcast transmitters currently operating world wide with 40 Million receivers in daily use. Out of an estimate of the 140, 000, 000 receivers capable of short wave reception, there are only 2, 500, 000 in the U.S., yet our diverse backgrounds would suggest a high degree of interest in foreign broadcasts. By using a high frequency receiver and proper antennas, with appropriate equipment for the re-transmission of signals, the cable operator can add a virtually unlimited variety of special interest programming for his subscribers.

Along with time and weather broadcasts, Buckeye Cablevision of Toledo (a system which deserves much credit for pioneering effort in special FM services) carry BBC - London, Radio Berlin and Radio Moscow on the cable. In this case, their hub distribution plan includes carriage of the special services via microwave, so only one receiving location is required.

Another area closely related to cable origination is that of background music services that can take advantage of the unique capability of cable.

At present, background music is monaural, using SCA or wire lines for distribution. In addition to the transmission limitations, speaker placement problems and economic considerations, the specialists in this field tell us that the whole idea of background music is to minimize involvement on the part of the listener, so stereo is not used. Since total involvement with music would seem to best satisfy the needs of the listener in the dentists, or doctors office, or in fact most waiting rooms, it would appear that at least part of the argument for mono lies in the huge investment in monaural music libraries and equipment on the part of the music service operators.

I would like to go into other areas, such as point-to-point voice and data links, remote control of surveillance cameras, transmission of slow scan television and facsimile which involve signals of an aural nature, but realize that to deal with these would be going beyond the limits assigned for this paper.

The stated purpose of this paper was to acquaint you with some of the ideas that are being currently used, and others that are not presently used but readily available as aural services in the FM band. Only time will tell the ultimate success of some of these concepts, but they do represent simple and relatively inexpensive services which bear the serious consideration of cable operators who see CAFM as part of their future.

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BENDING YOUR SYSTEM INTO SHAPE - HINGING

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Abstract - This paper discusses a new technique, called "hinging", for calculating system design levels to meet a specification while minimizing cost. The condition of a signal is treated as a normalized noise and distortion pair in a specialized coordinate space. This leads to a discussion of amplifier vectors, signal quality and system margin, which is related to tolerance. This approach is an expansion of some early work by D. Carson. (See Ref. 1)

This paper addresses the following question:

Give system and amplifier specifications, at what levels should the amplifiers be operated to minimize system cost?

It is generally accepted that the cost of a cable system is reduced by increasing the feeder-to-trunk ratio. The length of the feeder is a function of the output levels of the bridger and line-extenders, and the minimum acceptable subscriber level. For a given minimum subscriber level, the greater the output levels of the bridger and line-extenders, the longer the feeder, the better the feeder--to-trunk ratio, and the lower the cost.

Feeder systems are, therefore, normally operated at the highest possible levels within the allowable distortion specification. This is accomplished by reducing the distortion in other parts of the system, lowering their levels. These levels in turn are limitted by the carrier-to-noise specification. Because the feeder will contribute to the overall system noise, and the low-level sections of the system will still contribute distortion, care must be taken to insure that the system is still within specifications - as the feeder levels are increased. One major advantage of the approach discussed in this paper is that this problem can be handled directly.

Noise - Distortion Coordinates

The technique uses a coordinate system based on the two quantities which normally limit the operating levels of a CATV amplifier:

(1) Carrier-to-Noise
(2) Carrier-to-Distortion

These are normally specified in deciBels. (dB)

The distortion is assumed to be third order in nature, varying 2 dB for each 1 dB variation of amplifier level, and accumulate by voltage addition. It is usually cross-modulation or triple-beat, relative to the carrier level. Carrier-to-noise, as defined by the NCTA, varies 1 dB for each dB change in amplifier level and accumulates by power addition.

Unfortunately, neither of the two quantities mentioned above adds conveniently in deciBels when cascades are considered. It was therefore found more useful to consider them as inverse ratios:

(1) Noise-to-Carrier as a power ratio adds in cascade. Therefore, the output signal of a cascade of M amplifiers has a noise-to-carrier M times the noise-to-carrier of the same signal passed through only one amplifier. (Assuming a unity-gain system, identical amplifiers)

(2) Distortion-to-Carrier as a voltage ratio also adds in cascade.

It is further convenient to express these quantities in a normalized form, dividing them by their system specification. Thus, the condition of a signal is specified by the ordered pair (n,x) where,

n = Normalized noise-to-carrier ratio
x = Normalized distortion-to-carrier ratio

If $(N/C)_o$ is the noise-to-carrier of the signal,

and $(N/C)_{sp}$ is the system noise specification

then,
$$n = \frac{(N/C)_o}{(N/C)_{sp}}$$

Similarly, if $(X/C)_o$ is the distortion-to-carrier ratio for the signal and $(X/C)_{sp}$ is the specification,

$$x = \frac{(X/C)_o}{(X/C)_{sp}}$$

Figure 1 shows the representation of the condition of a signal by a point in a n-x coordinate system. Note that the dotted lines represent the boundaries of the region of permissible signals, a signal on the boundary having "used-up" either of the specifications.

At (1, 1) the signal has used-up both the noise (n = 1) and distortion (x = 1) specifications. A "clean" signal is represented by the origin (0, 0).

Quality and Margin

Consider a signal which has a normalized noise-to-carrier of 0.3 and a normalized distortion-to-carrier of 0.7. It can be represented by the point (0.3, 0.7) of Figure 2. Another signal represented by the point (0.7, 0.3) in the coordinate system might be said to be equivalently degraded. This is expressed by a

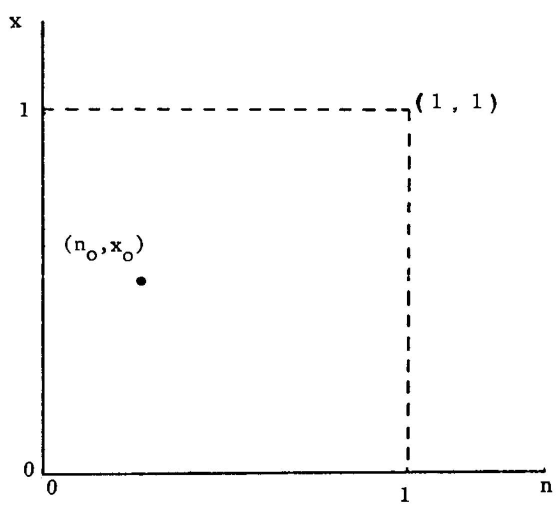

FIGURE 1
NOISE-DISTORTION
COORDINATE SPACE

n_o - normalized noise-to-carrier

x_o - normalized distortion-to-carrier

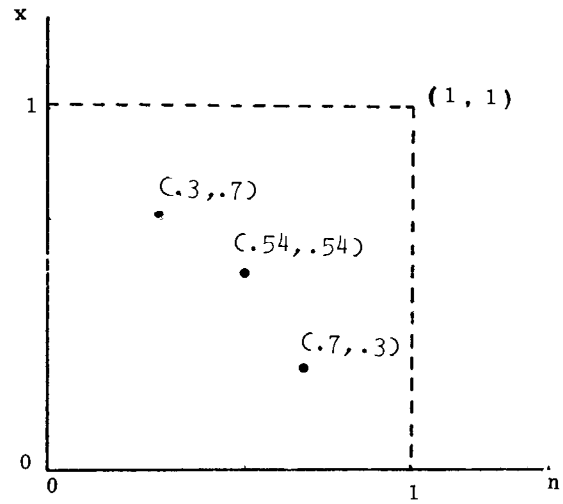

FIGURE 2
EQUAL Q (QUALITY) POINTS

mathematical quantity Q - the quality of the signal --

$$Q = \sqrt{(1-n)\ (1-x)}$$

Note that both of the points above have a Q of 0.46. Other points having a quality of 0.46 are: (.54, .54), (.1, .76), (.76, .1).

The quality of a signal has the following properties:

1. If the signal is clean,
   n = 0, x = 0, Q = 1.

2. If either of the system specifications is used up,

   n = 1 or x = 1, Q = 0.

3. Q has real values only when the signal is within specifications, and Q is always between 1 and 0.

Another term is also useful: The quality of the lowest-Q signal in a system is called the "margin" of the system and is usually expressed in percent. This signal normally occurs at the extremity of the longest cascade. Appendix I shows that this quantity is closely related to system tolerance as described in the Technical Handbook for CATV Systems by Ken Simons. (3rd Edition, Chapter VII, pg. 46)

Amplifiers, Cascades

If points q_1 and q_2 represent, respectively, the input and output signals of a given amplifier, then an arrow or vector drawn from q_1 to q_2 can represent that amplifier.

Figure 3 shows such a vector and demonstrates the effects of raising or lowering the output level of the amplifier.

(All other parameters: Gain, Noise Figure, Output Capability are, of course, assumed fixed). As the level of the signal at the output of the amplifier is varied, the vector, connecting the pair of points representing its input and output signals, changes in direction and length. The arc drawn through the points of all possible output signals is a hyperbola.

If a signal has equal n and x coordinates, i.e. $n = x$, it is said to be "midway" between the distortion and noise specifications. If such a signal is the input to an amplifier, the shortest length vector and the vector which causes the minimum change in Q, falls on a line which contains all of the "midway" points. The output signal of such an amplifier is "midway" and the amplifier is said to be operated at its "midway level".

The Handbook for CATV Systems shows that the maximum cascade can be achieved by operating an amplifier at midway level. At this level both specifications are used-up at equal rates. The shortest amplifier vector also results, and all the vectors lie on a straight line. Thus, the longest cascade from a point on the midway line is also on the midway line; and the longest cascade between two points is a straight line.

An amplifier vector has the following equivalent properties:

(1) It points from a signal of quality Q to another signal whose quality is always less than Q.

(2) Its angle with respect to the n - axis is between 0° and 90°.

(3) For a midway input signal, the shortest length amplifier vector occurs when the noise and distortion specifications are "used-up" at equal rates and the change in Q is minimized.

System Design

A cascade of amplifiers can be represented by placing their vectors heel-to-toe as in Figure 4. Amplifiers can be lumped together into a single vector by observing the input and output signal Q's and varying all the amplifier levels together. Figure 5 shows a two-vector cascade representing the trunk and feeder of a given system.

A cascade of identical amplifiers with a clean input signal and operated midway can be represented by the vectors in Figure 6. This type of system results in the highest possible margin and is optimum with respect to performance.

Rarely, if ever, is a system operated midway.

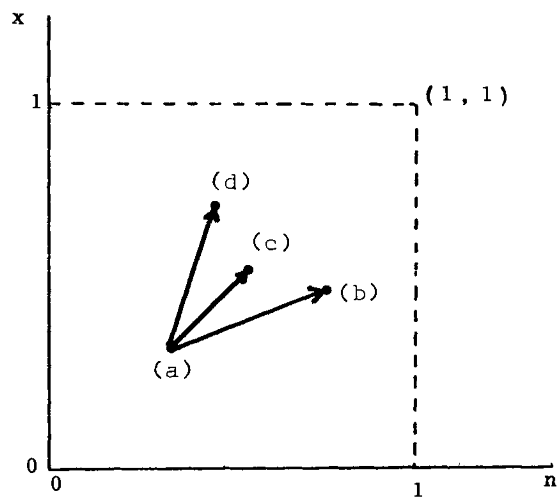

FIGURE 3
AMPLIFIER VECTORS

(a) input signal
(b) low level output signal
(c) midway output signal
(d) high level output

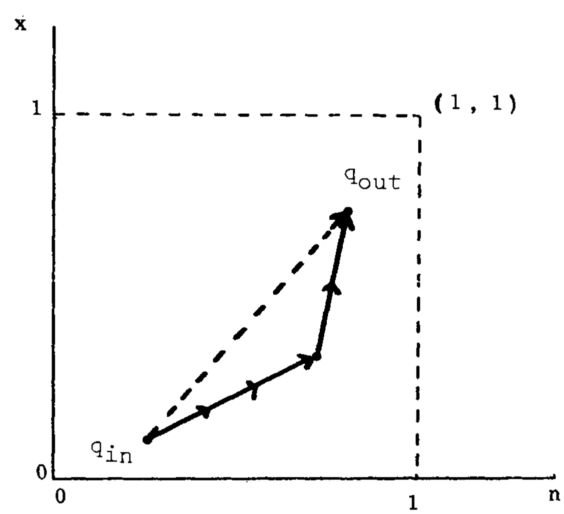

FIGURE 4
AMPLIFIER CASCADES

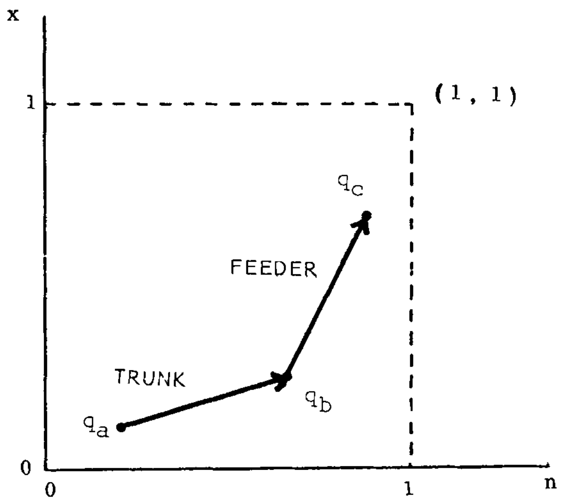

FIGURE 5
TRUNK AND FEEDER
VECTORS

q_a - signal into trunk
q_b - signal out of trunk, and into feeder
q_c - signal out of feeder

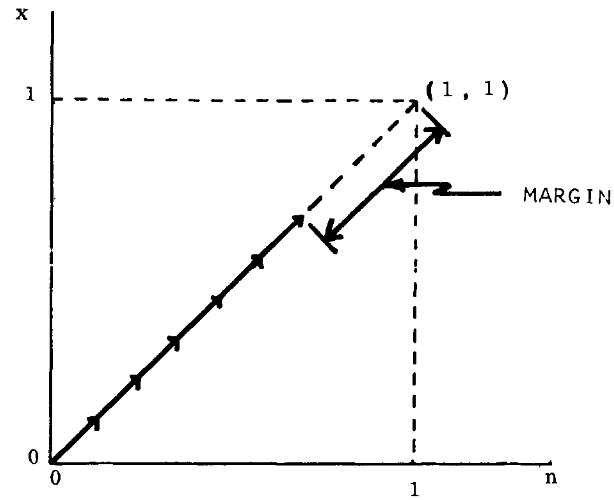

FIGURE 6
"MIDWAY" LEVEL SYSTEM

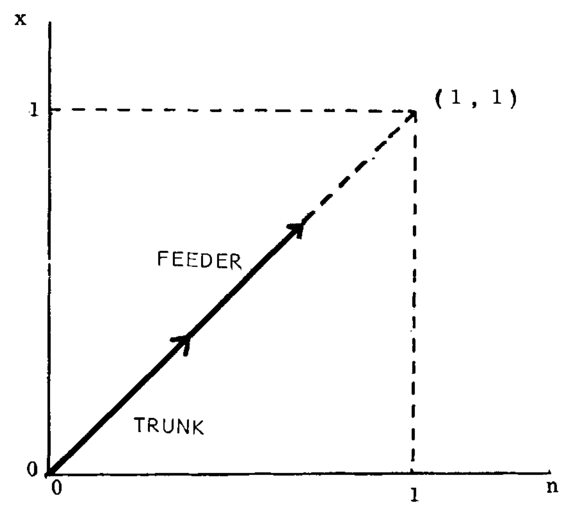

FIGURE 7A

OPTIMUM PERFORMANCE
SYSTEM
(HIGH COST)

The levels are usually very low, resulting in a low feeder-to-trunk ratio and high cost. Fortunately a cascade much shorter than the maximum cascade is usually required, and the excess margin of the cascade is "used-up" to obtain better levels.

Operated midway, the system margin is maximized in Figure 7a. This margin is reallocated to reduce cost by increasing the feeder levels and reducing the other levels in the system. This results in the diagram of Figure 7b which suggests the name "hinging".

The maximum feeder levels occur when,

(1) All of the available specifications have been used, and the margin is zero.

(2) The distortion and noise specifications are simultaneously used-up, i.e. the last vector ends at (1, 1).

(3) The hinge point is such that the trunk vector has the lowest possible level.

The system which fulfills (1) - (3) is the solution to the problem of minimum cost.

The hinging technique has been used successfully at Jerrold for one-way and two-way systems. In a one-way system the terminating bridger of the longest cascade and the line-extenders are treated as the feeder vector. The trunk amplifiers are lumped together into another vector. The two vectors are then hinged. In a two-way system the return trunk, return feeder, and any hub-to-hub trunks, are included in the trunk vector to take full advantage of all excess margin.

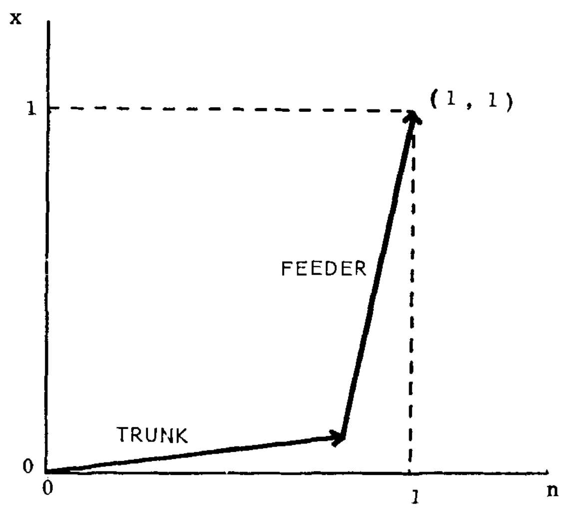

FIGURE 7B

OPTIMUM COST SYSTEM
(WITHIN SPECIFICATIONS)

Appendix I
Margin and Tolerance

The tolerance of a single amplifier, for a given set of system specifications is defined in the Handbook for CATV Systems as

$$T_1 = E_{max} - E_{min}$$

where E_{max} and E_{min} are, respectively, the maximum and minimum output levels, expressed in dBmV, allowed by the system noise and distortion specifications. In a unity-gain cascade of identical amplifiers the system tolerance is

$$T_s = T_1 - 20 \log m$$

in deciBels, where m is the length of the cascade. System tolerance is also called reserve tolerance.

Margin, as defined in this paper, can be shown to be related to system tolerance when margin is computed based on a midway system.

Given the system tolerance T_s, the midway level E_o is,

$$E_o = E_{xm} - T_s/2$$

$$T_s = 2 (E_{xm} - E_o) \qquad (1)$$

where all levels are in dBmV and tolerance is in dB. E_{xm} is the maximum output level for each amplifier such that an m amplifier cascade just meets the distortion specification.

Operating midway, the resulting distortion from each amplifier is

$$X_o = X_s + 2 (E_{xm} - E_o) \qquad (2)$$

in deciBels. X_s is the system distortion specification.

Since the distortion and noise are used up at equal rates, the normalized noise and distortion ratios are equal,

n = x (n is the normalized noise; x, the normalized distortion)

and the margin M is,

$$M = \sqrt{(1 - n)(1 - x)} = 1 - x$$

$$x = 1 - M = 10^{-X_o/20} \big/ 10^{-X_s/20}$$

where x and M are ratios and X_o and X_s are the deciBel values of signal distortion and system distortion respectively.

Therefore,

$$1 - M = 10^{(X_s - X_o)/20}$$

$$X_s - X_o = 20 \log (1 - M) \qquad (3)$$

and from (1) and (2) above,

$$-T_s = -2 (E_{xn} - E_o) = X_s - X_o \qquad (4)$$

so that, from (3) and (4),

$$T_s = -20 \log (1 - M)$$

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February 28, 1974

CABLE & SATELLITE: FRAMEWORK FOR A NEW INDUSTRY

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SUMMARY

Satellites and cable systems form a natural technological base for a nationwide interconnected broadband network. An organizational concept to bring such a network into being does not exist, but initiatives have already been taken which could bring about a new multi-service telecommunications structure based on the knowledge and experience of the present industry. Conditions in the regulatory, economic and technological areas appear to be favorable to such a development.

The concept of a cable network is enhanced by the two fundamental advantages of satellite distribution, which are insensitivity to the distance factor and simultaneous coverage of all receiving points located within the area of visibility. The viewable area from the bird is approximately one-third of the earth's surface. The antenna can be designed to cover a continent, a single time zone, a smaller area such as Puerto Rico or Hawaii by spot beam, or several such areas simultaneously.

Use of domestic satellites is proposed for TV distribution, cable system interconnection, private line service, data traffic, facsimile including electronic mail delivery and many other services not now accommodated by the point-to-point and switched message networks.

The primary reasons behind satellite usage in these modes are operational economy, high quality of signal, fewer terrestrial problems, flexibility of use and capability for expansion of services at declining costs.

Institutional arrangements in domestic satellite service are as yet undefined, and access is relatively open to both space and ground environment, as compared with the international satellite system which is, by law, carrier-configured.

The integration of high-capacity cable communications service with the fundamental advantages of satellite distribution requires a comparatively low cost but highly flexible earth receiving station, such as was first demonstrated at the 1973 NCTA Convention in Anaheim, California. This station, built by Scientific-Atlanta to specifications supplied by TelePrompter, was used to demonstrate long-haul satellite relay in the first SpaceCast^{sm} program. Although this was a point-to-point transmission and thus not unique, it was an historic first United States' use of satellite technology for a domestic TV program communication. It was also a first in the association of a cable-originated program with a satellite carrier, American Satellite Corporation, and in the reception of signal through a cable-owned earth station interconnecting with local cable systems.

The first-generation domestic satellite facility that seems to be emerging in 1974-75 consists of a carrier-owned-and-operated satellite accessed through carrier-owned-and-operated transmit and receive stations which will be located in the pattern New York - Los Angeles - Chicago - Dallas - Washington - San Francisco - Seattle - Atlanta.

Cable interface with these facilities would occur at receive-only earth stations located at or near system head-ends. It has been estimated that for the immediate future between 80 and 100 earth stations in combination with regional terrestrial microwave systems could interconnect cable systems serving 50% of the present CATV market.

The satellite frequencies to be used in the present and immediate future are those shared with terrestrial microwave in the 4 and 6 Ghz bands. Later on the 6 Ghz band may be displaced by the 14 Ghz band for transmission to the satellite, thus eliminating present interference problems. Down-link transmission, continuing at 4 Ghz, would not materially change the configuration of the proven 'bird' design, would retain the propagation advantage of this lower frequency and would protect the investment of receive station pioneers.

Cable's immediate need is for distribution of special interest programming which in turn requires the economic characteristics of satellite distribution to become attractive to advertisers. The audience for this type of programming will become a viable market through the cumulative process, or incremental build-up of small audience segments. Whether these segments are reached in larger population centers or in rural areas, the satellite sees them all at a uniform, one-time space cost. Indeed the satellite alone makes attainable the concept of a large national cumulative audience for special interest program material.

A brief review of a typical satellite carrier rate structure reveals why this is so. A one-hour satellite transmission from New York to Los Angeles, on a point-to-point basis, for a full color video channel with two 4Khz audio channels, including up-link, down-link and space segment, costs between $725 for off-peak hours and $1450 for prime - time - scheduled service.

For a point-to-many-points service, the expected CATV pattern of usage, the same charges would be made by the carrier, including a down-link charge only where the carrier itself owned the receiving station. No charge at all is made for reception at cable-owned receive-only stations, regardless of number. Thus the carrier charge for satellite service, including transmission, could be subdivided by as many receive-only stations as would be in operation, a number eventually in the thousands, for the cable industry alone. The cost of reaching cable subscribers in this way with marketable special interest programs begins to shrink to manageable proportions.

A reasonable estimate for the near future would suggest that a cable - satellite network could be constructed wherein strategic placement of eighty earth stations in combination with existing terrestrial facilities would serve cable systems reaching some 4 1/2 million subscriber homes. Satellite transmission, or up-link plus space segment costs, would be about $1,200,000 per year for full-time usage of one transponder. No charge is made in this model, for reception at cable-owned earth stations. The earth stations themselves are estimated to cost $70,000 each, but this amortizes over at least a ten year life and involves low operating costs.

With these figures to build on, and including ongoing distribution costs from the 80 earth stations, the costs to the cable system operator for participation in such a network average out to approximately 10¢ per subscriber home or about 4¢ per potential viewer per month.

The earth stations are expected to be capable of use for more than TV alone. With minimal upkeep and operational cost and a ten-year lifetime they represent a modest investment in facilities that will stimulate new subscribership as well as provide additional revenues from ancillary services.

It would seem at least reasonable to expect that the economics of this distribution method would support the special interest, cumulative audience concept as well as time division multiplexed digital services which have been advanced as cable's great promise for the future.

Will the public respond to such an alternative program service? Projections and theories have been offered, but nobody really knows. Meanwhile at least one group is proceeding to action in this field, rather than meditation. Westport Broadcasting, of Kansas City, has organized an interconnected cable-only program service to begin on May 1, 1974. The approximate dimensions of the venture show a 12-state area, some 300 cable systems belonging to various MSO's, over 800,000 subscribers and a program schedule of 10 hours a day without movies. The service has sold itself and is spreading. National advertisers have signed on, and systems can sell protected local spots as part of the arrangement.

Here is a starting point for a national cable service that will include program, sales, traffic and affiliate activities. As an outgrowth of a straight UHF operation it may show how the broadcasting industry will profit, hedge and expand in partnership with cable rather than in opposition to it. Indeed, the broadcasters, with proven expertise and facilities for program production are the most likely source of programming for cable distribution.

Should Westport install or feed a satellite transmitter, its cable program service would become available to the entire country over RCA, Western Union or Amsat space facilities. There are several letters of intent from cable systems now in the hands of earth station manufacturers awaiting this event. Equally feasible would be the use of a satellite receiver at the center of the Kansas City network for the reception of supplementary program material and other services from east or west coast origination points where transmittal facilities already exist.

In addition to this venture there is also the nucleus of an industry-wide organization which could complement and broaden the Kansas City regional network. This is the so-called Cable Satellite Access Entity, a group of 32 MSO's, independents, manufacturers and others who have commissioned Booz-Allen&Hamilton to document the opportunities of cable interconnection by satellite. The consortium, if it becomes operational, could provide the immediate vehicle for expansion of a regional network to national dimensions and multi-service marketing. It could also provide the base for an earth station - leasing venture which need not be limited to cable but could serve independent broadcasting stations as an interconnect, a concept which was recently exposed at the INTV Convention at Dallas.

Whether a cable system or other entity leases or owns an earth station, access thereto by law must be completely free of restriction. From the cable point of view, the economic advantage in owning an earth station is appealing. Linked up with the community cable delivery system, it may rapidly become the most logical and effective means for the distribution of broadband communications.

The carrier backbone satellite system contemplates that major send - receive stations will be owned and operated by the carrier which owns and operates the spacecraft. It does not seem likely, for a few years at least, that the broadcast industry or the cable industry will be interested in a satellite dedicated to a particular need or service. But both broadcaster and cablecaster have an identifiable interest in maintaining contact with their customers. In the broadcaster's case this contact might eventually be the rooftop microwave dish aimed at a direct broadcast satellite. We will see this in undeveloped countries perhaps before we see it in this country, for obvious reasons. But the cablecaster, with direct access to his subscriber, needs to control the gateway to his service which is represented by the receive station associated with his head-end. This is not only legal, but common sense as well. The incentives for provision of earth station services are as many as there are revenue-producing services which may be offered. The ownership of earth stations while not limited by law would seem logically to fall to the entity providing service to the ultimate user who is, of course, the cable subscriber.

The question might then arise, is cable system ownership of multi-channel receive-only earth stations a common carrier function, in view of FCC rules encouraging leasing of channels and non-video services which cable can provide?

The definition of the cable-system operation of a network reception point, the earth station, really turns on the larger question of the existence and function of a cable network, regardless of who provides the satellite or other means of interconnection.

A cable network may be divisible into separate and distinct functions. It is well to specify the sense in which the word network is used, as the implication of a broadcast network organization as it exists today may not apply.

A cable network could be carrying programs as a leased channel service; the programmer would be leasing channels for this purpose, and the network as the lessor organization would presumably be out leasing other channels for any purpose to whatever markets and for whatsoever services a customer desired.

However, cable systems, regionally or nationally, might wish to program themselves, in which case their activities would resemble those of any small conventional network, except that their revenue sources would include subscriber growth and possibly interconnected pay-TV channels as well as advertising. Mr. Whitehead suggests, in his recent report, that while control over facilities and program should be separated, the cable industry might be encouraged to program one or two of the many channels available. This certainly would not exclude a satellite network channel or channels or the ability of a cable operator to pick and choose available satellite programming for his own channel distribution. Nor is a cable programming network prohibited by any existing regulations. It might even be encouraged by a broad interpretation of the Commission's thrust toward cable program origination.

But a cable program network and a leased channel network are two concepts separated by a single lack - that of an organization that could do both, or either. When it becomes necessary to define cable's status, will it continue under present rules or will it be a common carrier, specialized carrier, non-carrier or some combination of all three? It will be necessary to include in the definition the function of networking on a national multi-channel basis. There will have to be an identifiable industry-wide body of some sort to which responsibility could attach. If a common carrier structure emerges, there must be someone to contact for leasing a channel for a marketing pattern that suits your purposes. If a client decides to conduct a program and sales operation on a network scale, the geographic preserves of single MSO's must give way to an organization capable of representing many - an organization which today does not exist.

That such an organization should exist is accepted by many as inevitable. Whether it should come into being as a new profit-oriented corporation or as an industry-sponsored consortium which contributes the administrative, operational and technical know-how, is an interesting question. It amounts to the imposition of a new technology on an old establishment.

Public and regulatory policy has seen to it that the expansion of cable need not be at the expense of familiar and vested interests. At the same time policy is moving toward cable expansion freed from some of the present establishment restrictions. Freedom of viewer choice and availability of low cost communications services may not only be necessary but may stimulate established facilities to greater service and expanded business. Naturally some interests feel threatned, but if cable does not organize as an industry, it poses its main threat to itself.

As soon as it is clearly recognized that an operational entity, having a specific operational mission for the industry, in contrast with the general responsibilities of a trade association, is a feasible organization, the cable industry will be in a position to deal from strength in the various readjustments that are now impending. Such an entity, foreshadowed in events that have been mentioned, would owe its cohesion to the glue that holds our form of society together, the profit motive. For the prospects of a multi-service national broadband communications network for this country are real, and differ, in an order of magnitude, from the limited structure of today. It remains for those who see these prospects to bring them to fruition. History may well record that 1974 was the year when the cable industry realized these possibilities and took the necessary steps to provide the best administrative and creative environment for them.

CABLE REPEATER STATION DESIGN
USING FIXED GAIN BLOCK PREAMPLIFIER AND POWERAMPLIFIER

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1.0 Introduction

A cable repeater trunk amplifier station design in common use today consists of a flat (or nearly flat) frequency response, fixed gain preamplifier and poweramplifier, separated from each other and from the station input and output connectors by various loss networks. Refer to the block diagram in Figure 1.

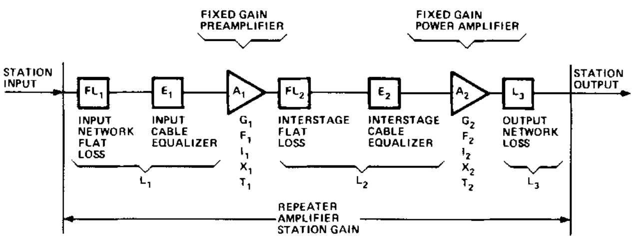


Figure 1. Repeater Station Block Diagram

The networks introducing loss provide the functions of cable equalization, gain and slope controls, bridging and AGC amplifier and power takeoff, and frequency division multiplex filters, all of which are necessary for cable repeater amplifier station performance. Apportioning the magnitude of loss in these networks in the station design should be done in a manner which, in the limit, allows the ultimate station noise figure to approach the value of the preamplifier (A_1) noise figure and the station distortion values to approach the values of the poweramplifier (A_2) distortion. Fixed gain amplifiers A_1 and A_2 will be referred to as IC (hybrid integrated circuit) amplifiers throughout the paper, although the analysis is general and applies to discrete component amplifiers as well.

The IC amplifier (A_1 and A_2) distortion (crossmodulation, second order intermodulation distortion, and triple beat) is usually specified by IC vendors for signal carriers at equal signal levels. Equal signal levels are used for convenience in measurement and also provide a convenient reference.

The signal levels at the inputs and outputs of A_1 and A_2 in the repeater amplifier are normally not flat because of the cable attenuation frequency response and the cable system block tilt many times used. The magnitude and frequency response of the input, interstage, and output loss networks L_1, L_2 and L_3 have very significant effects on the station noise and distortion performance. It is, therefore, difficult to predict and compare repeater amplifier station performance on the basis of various IC amplifier specifications based on flat level specifications. To solve this problem, a fast, simple means of predicting ultimate station performance from the known performance of IC amplifiers A_1 and A_2, system block tilts, and station equalizer and loss network values has been developed by deriving a number of normalized performance degradation factors. This paper will define and explain the use of these degradation factors, and show the advantages to be gained by their use. The degradation factors, when added to the flat level IC specifications at a given station operating level, provide the station distortion and signal-to-noise ratio specifications. A list of symbols and definitions of terms used in the paper is listed on page 13.

Three amplifier design problems which inspired the derivation and use of the degradation factors are summarized in Table I. The derivation and use of the Degradation Factors provides a simple, quick means of solving the three problems listed below and summarized in Table I.

1. Determine *station design* from fixed gain *IC amplifier specs* to *maximize station specs.*
2. Determine fixed gain *IC amplifier specs* and *station design* to *meet* required *station specs.*
3. *Determine station specs* for a *given station design* and *given fixed gain IC amplifier specs.*

Table I. Problems Solved with D-Factors

GIVEN	DETERMINE
1. IC Amplifier Specs	Station Design to Maximize Station Performance Specs
2. Station Performance Specs	IC Specs and Station Design to Meet Station Performance Specs
3. Station Design, and IC Amplifier Specs	Station Performance Specs

Ultimate station performance would be attained if the station noise figure were equal to the preamplifier noise figure F_1, and the station distortion were equal to or less than the poweramplifier distortion specifications I_2, T_2 and X_2. The D-Factor derivation assumes that any distortion contributed by the preamplifier is additive on a voltage basis (20 log). The station distortion could be less than the poweramplifier (A_2) distortion by the use of interstage distortion cancellation networks, particularly second order distortion. However, the distortion cancellation design approach is not in common use in present generation trunk station amplifier designs, and is not considered in this paper. Five D-Factor sets are derived for each particular trunk station design and level tilt. Each D-Factor is the number of dB that the trunk station performance deviates from the fixed gain IC amplifier specification measured at any flat level within the applicable dynamic range of the IC. This relationship is shown schematically in Figure 2.

The D-Factors (D_F and $D_{S/N}$) are precisely accurate at each frequency calculated for station noise figure and signal-to-noise figure. The D-Factors (D_I and D_T) are also accurate for the second order distortion product of each pair of frequencies calculated and each set of third order (triple beat) products calculated individually. The magnitude of the combination of second

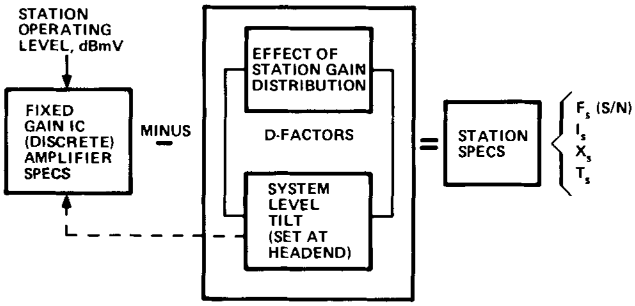


Figure 2. Relationship of IC Specs to Station Specs by D-Factors

and third order beats which occur at any single frequency for a given number of channels in the amplifier passband cannot be calculated in a practical way (Reference 6). The D-Factor (D_X) for crossmodulation in each channel calculated using the rule that "crossmodulation with n channels is [n-1] times that of a two channel measurement" is normally not accurate (Reference 9, 7) for broadband VHF amplifiers. A more precise answer for D_X can be obtained by measuring the actual fixed gain amplifiers with various block tilts and using these measured results as they apply to the calculation of D_X.

2.0 IC Amplifier Characteristics

Fixed Gain IC amplifier parameters are summarized in Table II below. The numbers used are those listed in the published TRW data sheet for the CA100 (preamplifier) and CA200 (poweramplifier) IC amplifiers.

Table II. Summary of IC Amplifier Specifications

PARAMETER	IC SPECIFICATION A_1 (preamp)	IC SPECIFICATION A_2 (poweramp)	STATION SPECIFICATION CATV MANUFACTURER
Gain (Fixed)	G_1 = 16.3 dB	G_2 = 16.3 dB	G_s = 22 dB (@ 300 MHz)
Frequency Response	Flat	Flat	Shaped to Equalize Cable Loss
Noise Figure	F_1 = 7.5 dB	F_2 = 10 dB	$F_s = F_1 + D_F$
Distortion, Flat Levels @ +50 dBmV			
Intermodulation (2nd) Order)(CH 2, 13, R)	I_1 = -68 dB	I_2 = -70 dB	$I_s = I_2 = D_I$
Crossmodulation (32 Channels)	X_1 = -52 dB	X_2 = -57 dB	$X_s = X_2 + D_X$
Triple Beat (3rd order)(CH 3 + CH 4 + CH A on 245 MHz)	T_1 = -73 dB	T_2 = -78 dB	$T_s = T_2 + D_T$

To attain bandwidth, input and output match, repeatability, gain stability and minimum cost, an IC amplifier with the devices available today must have more than one stage. At this time, it is difficult to design IC amplifiers with responses which equalize the various types of cables in use. It is for these reasons that the IC preamplifiers and poweramplifiers available today have a flat frequency response and approximately 16 dB of gain.

2.1 Fixed Gain Preamplifier

The preamplifier is designed for minimum attainable noise figure. The gain G_1 is fixed at approximately 16 dB. Preamplifier IM (I_1) distortion, crossmodulation distortion (X_1), and triple beat (T_1) is usually minimized by increasing the bias (which means more power consumption). An increase in bias tends to increase the preamplifier noise figure. Therefore, the trunk station designer should determine what preamplifier I_1, T_1 and X_1 he must have so that both power consumption and preamplifier noise figure can be minimized.

The station noise figure (F_s) design goal is to achieve an F_s which is limited only by the noise figure F_1 of the preamplifier. The station F_s is increased above F_1 by the loss networks in the station and also by the poweramplifier noise figure F_2.

2.2 Fixed Gain Poweramplifier

The station distortion design goal is to achieve station distortion (I_s, T_s and X_s) which is limited only by the distortion of the fixed gain poweramplifier (I_2, T_2 and X_2). The gain G_2 is fixed at approximately 16 dB. The IC poweramplifier is designed for minimum distortion (I_2, T_2 and X_2) and its noise figure F_2 is allowed to be higher than F_1. The minimum distortion is ultimately limited by the transistors (or active devices) used. The contribution of preamplifier distortion to the station distortion should be minimized by careful apportionment of the required loss networks in the station.

The station design challenge is to select the best compromise between conflicting requirements for the amount of loss apportioned to each loss network in the station (figure 1.0). Ultimate station noise figure should be limited by preamplifier noise figure and station distortion limited by poweramplifier distortion.

3.0 Relating Station Performance Parameters to Fixed Gain Preamplifier and Poweramplifier Parameters by D-Factors

Even though the fixed gain amplifier specifications such as those listed in Table II for A_1 and A_2 are known, the station specifications obtainable are not immediately obvious. The various loss networks required in the station, as shown in Figure 1, make it impossible to attain station specifications which are as good as the A_1 and A_2 specifications. To further complicate the prediction of the station specifications from the known specifications of A_1 and A_2, the station output level of each channel in the passband is normally not flat, but is tilted (block-tilted or linear-tilted).

A set of D-Factors[1] will be derived for a given set of station design requirements and system level tilts. The D-Factors are normalized such that they are independent of signal level. The fourth column of Table II lists the relationship between station and A_1, A_2 specifications by D-Factors. An outline of the station performance analysis leading to the D-Factors derivation and their use is shown in Figure 3. A station design example follows to show the type of results expected from the analysis. The numbers used are obtained from the analysis which follows later in the paper.

[1]Because of the various functions required in a repeater amplifier station design, the station specifications will be worse than (or Degraded from) the Fixed gain A_1 and A_2 specifications; hence, the term D-Factor.

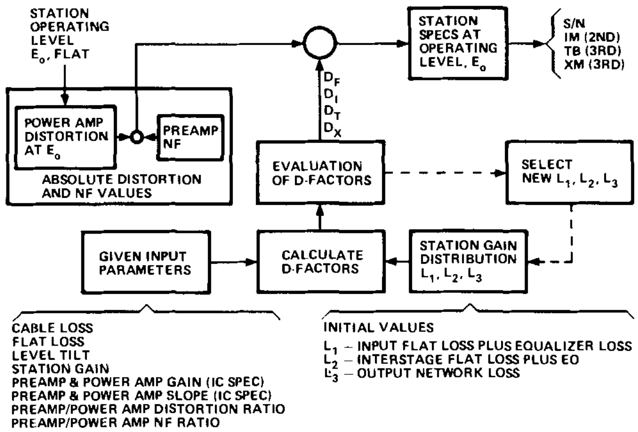


Figure 3. Station Performance Analysis Flow Diagram Using D-Factor Analysis

3.1 Design Example

GIVEN PARAMETERS

STATION	FIXED GAIN IC AMPLIFIERS (See Table II)
Cable Loss = 22 dB @ 300 MHz	$G_1 = 16.3$
G_s (Station Gain) = 22 dB @ 300 MHz	$G_2 = 16.3$
$FL_1 = -1.5$ dB	$F_2 - F_1 = 2.5$ dB
$FL_2 = -3$ dB	$I_1 - I_2 = 2$ dB
$FL_3 = -1.5$ dB	$X_1 - X_2 = 5$ dB
E_1 (input Equalizer) = 14 dB	$T_1 - T_2 = 5$ dB
Station Output Level = 31 dBmV	
Level Tilt[1] 4/2/0/0	

The D-Factors are calculated from the given parameters and the analysis procedures outlined in Figure 3 and derived in the following sections of this paper. The IC specs at +31 dBmV are summarized in the following table[2]. The D-Factors are added to the IC specifications to give the station specifications.

Note that the D-Factors are a function of frequency, so that D-Factors for a sufficient number of sets of frequencies are necessary to characterize the station performance across the passband of the amplifier. Note the difference of the station specs of column 4 versus the IC specs of column 2 in Table III.

The signal level of each channel applied to a station must be known in addition to the station noise figure F_s in order to determine the station signal-to-noise ratio (S/N). Since it is S/N and not N (noise) alone which is a measure of signal quality, an S/N degradation factor $D_{S/N}$ is also derived. The operating signal levels in a system requiring a cascade of a number of repeater stations assume considerable importance. The relationship between channel levels, which will be defined by a number of level tilt settings, effect not only station S/N, but also the station distortion. The significance of level (block) tilts will be discussed in Section 6.0. Since the D-Factor $D_{S/N}$ includes D_F, the D_F factor will be derived first, followed by the derivation of $D_{S/N}$ in Section 5.0.

Table III. Station Design Example Summary

Parameter	IC Specs (in dB) @ +31 dBmV, Flat Levels	D-FACTOR (in dB) Calculated from given Parameters Listed Above	Station Specs (in dB) @ +31 dBmV & 4/2/0/0 Block Tilt
NF	F_1 (300 MHz) = 7.5 F_1 (Chan 2) = 6.5	D_F (300 MHz) = 2.8 D_F (Chan 2) = 12.8	F_s (300 MHz) = 10.3 F_s (Chan 2) = 19.3
S/N	Input Level at 300 MHz is +9 dBmV	$D_{S/N}$ (300) = -2.7 $D_{S/N}$ (Chan 2) = -2.0	S/N (300 MHz) = +57.8 S/N (Chan 2) = +58.5
IM	I_2 (CH 2 + 13 R) = -89	D_I (CH 2 + 13 in R) = 3.4	I_s (CH 2 + 13 in R) = -85.6
TB	T_2 (CH 3 + 4 + A in Ch 0) = -116	D_T (CH 3 + 4 + A) = -3.8	T_s (CH 3 + 4 + A) = -119.8
XM	X_2 (CH 2, 32 Channels) = -95	D_x (CH 2) = 4.1	X_s (CH 2) = -90.9

4.0 Normalized Preamplifier Noise Figure Degradation, D_F

It has been previously stated that the ideal station noise figure (F_s) would be equal to the fixed gain preamplifier noise figure (F_1). In other words, D_F would be equal to zero in equation 4.1.

$$(4.1) \quad F_s = F_1 + D_F$$

The magnitude of D_F is a function of the following:

1. Input network loss (flat and sloped).
2. Interstage network loss (flat and sloped).
3. Poweramplifier to preamplifier noise figure ratio.

The input and interstage network losses are in part determined by G_1, G_2, and the station gain G_s. The magnitude of D_F will be different at maximum G_s compared to operating G_s. The noise figure F_1 of A_1 and F_2 of A_2 is normally a smooth function of frequency with today's devices, and is about 1.2 dB higher at 300 MHz compared to its value at 55 MHz. Also, L_1 and L_2 are functions of frequency, so that D_F is a function of frequency.

The noise figure at the IC preamplifier input terminals will be calculated first. The effect of input loss networks on the station noise figure will then be added to this result. Lower case symbols indicate numeric values and capital letters indicate dB values. The noise figure is independent of signal level and block tilt.

Referring to Figure 4, the expression for f_1' is

$$(4.2) \quad f_1' = f_1 + \frac{f_2' - 1}{g_1}$$

[1] Level Tilt notation defined in Section 6.0.

[2] This example assumes an IC distortion characteristic for which relative third order distortion (Crossmodulation and triple beat) decreases 2 dB for every 1 dB of level reduction and relative second order distortion decreases 1 dB for every 1 dB of level reduction. Note that the A_1 and A_2 distortion versus level characteristic must be defined in order to obtain a valid I_s, T_s and X_s.

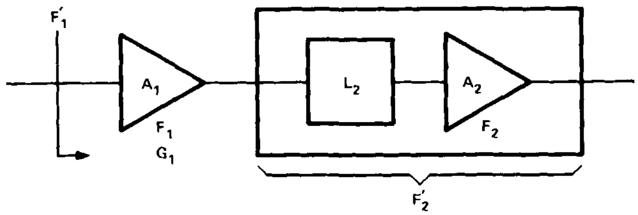


Figure 4. Block Diagram of Amplifiers Cascaded & Separated by Loss

The value of interstage loss L_2 is determined by G_s, G_1, and G_2 and the apportionment of loss between L_1, L_2 and L_3. L_2 adds to the noise figure F_2 on a dB for dB basis so that

$$(4.3) \quad F_2' = F_2 + L_2 \qquad \text{in dB}$$

The quantity l_2 is the numeric gain (loss) of the interstage network and is less than 1.0. Equation 4.2 becomes

$$(4.4) \quad f_1' = f_1 + \frac{(f_2/l_2) - 1}{g_1}$$

By factoring out f_1, the effect of f_2 and l_2 on f_1', the total noise factor, can be shown

$$(4.5) \quad f_1' = f_1 \left[1 - \frac{1}{f_1 g_1} + \frac{f_2}{f_1 g_1 l_2}\right] \qquad \text{Numeric}$$

$$(4.6) \quad f_1' = f_1 \left[\frac{f_1 g_1 - 1}{f_1 g_1} + \frac{k}{g_1 l_2}\right] \qquad \text{Numeric}$$

The ratio of f_2/f_1, denoted k, is normally greater than one, which means that the power amplifier noise factor has the effect of increasing f_1 as shown in equation 4.6. It is assumed here, without loss of generality, that F_1 versus frequency and F_2 versus frequency have the same relative values versus frequency.

Since $f_1 g_1 >> 1$ for most practical cases, equation 4.6 can be written

$$(4.7) \quad f_1' = f_1 \left[1 + \frac{k}{g_1 l_2}\right] \qquad \text{Numeric}$$

Equation 4.7 is a very important result. The interstage loss l_2 is a number less than one. The higher the interstage loss, the smaller l_2 and the larger f_1 becomes. Written in dB

$$(4.8) \quad F_1' = 10 \log f_1 + 10 \log \left[1 + \frac{k}{g_1 l_2}\right]$$

The factor D_F' is defined as the degradation of the preamplifier noise figure looking into the preamplifier (A_1) input terminals. It is a function of preamp gain, interstage loss; and the f_2/f_1 ratio.

$$(4.9) \quad D_F' = 10 \log \left[1 + \frac{k}{g_1 l_2}\right]$$

Preamplifier noise figure degradation as a function of interstage loss and A_2 to A_1 noise figure ratio is summarized in Table IV. Note that each dB of A_2 noise figure increase relative to F_1 is equivalent to a dB increase in interstage loss. Also note that the quantity l_2 is a function of frequency.

Table IV. Noise Figure Degradation for Diagram Shown in Figure 4.0 (Calculated Using G_1 = 15.5 dB)

INTERSTAGE LOSS L_2 in dB (at a Specific Frequency)	PREAMPLIFIER NOISE FIGURE DEGRADATION, D_F' ($K=F_2 - F_1$ in dB)			
	K=0	K=2	K=4	K=6
2	.17	.28	.44	.69
4	.28	.44	.69	1.06
6	.44	.69	1.06	1.59
8	.69	1.06	1.59	2.31
10	1.06	1.59	2.31	3.26
12	1.59	2.31	3.26	4.43
14	2.31	3.26	4.43	5.81
16	3.26	4.43	5.81	7.37

The noise figure degradation factor D_F' for k = 0, 2, and 4 is plotted in Figure 5 versus interstage attenuation.

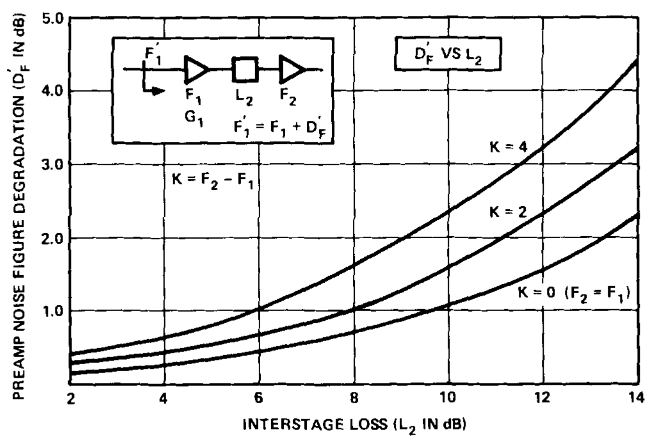


Figure 5. D_F vs L_2 (Interstage Loss) for G_1 = 15.5 dB

The station noise figure F_s is equal to F_1 in dB increased by the amount of loss of the input network (input equalizer loss plus flat loss).

$$(4.10) \quad F_s = F_1 + 10 \log \left[1 + \frac{k}{g_1 l_2}\right] + L_1 \qquad \text{in dB}$$

The normalized preamplifier noise figure degradation factor D_F, for the station, is therefore

$$(4.11) \quad D_F = 10 \log \left[1 + \frac{k}{g_1 l_2}\right] + L_1 \qquad \text{in dB}$$

The preamplifier noise figure degradation factor D_F is plotted in Figure 6 versus frequency for two different input equalizers, at operating station gain, and at maximum gain for an E_1 of 14 dB.

5.0 S/N Degradation Factor Derivation ($D_{S/N}$)

In an amplifier cascade design in which the amplifier gain at each frequency offsets the loss of the cable at each frequency – i.e., a unity gain system, the repeater station noise figure, input signal level (function of frequency), and the number of amplifiers in cascade define the cascade signal-to-noise (S/N) ratios.

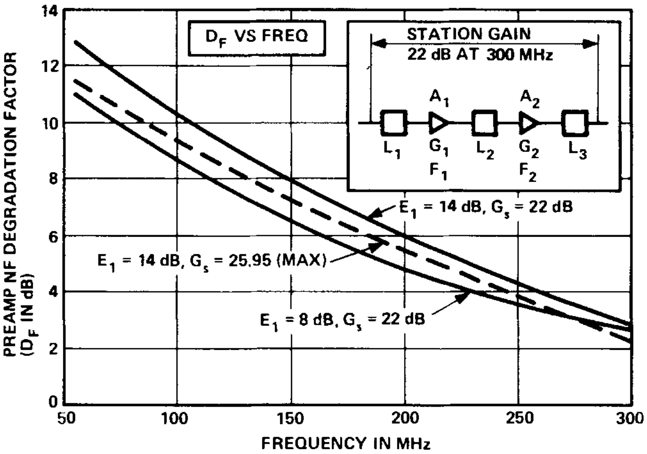


Figure 6. D_F versus Frequency using Design Example Parameters

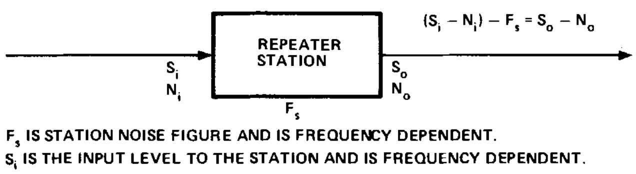


Figure 7. Station S/N

The signal to noise ratio of an N amplifier cascade, assuming identical amplifiers, is:

$$S_i - N_i = S_i - [-59 + F_s + 10 \log_{10} N] \qquad \text{in dB}$$

$$(5.1) \quad S_i - N_i = 59 + S_i - F_s - 10 \log_{10} N \qquad \text{in dB}$$

-59 dBmV is the thermal noise threshold for a 4 MHz bandwidth in a 75 ohm system at 20°C.

For convenience, the S/N of a single station is analyzed and the results presented for a single station only. The 10 log N factor can be added later for specific system cascade lengths. The single station S/N is:

$$(S_i - N_i) = 59 + S_i - F_s$$

An S/N degradation factor is derived which serves a number of purposes. It is used to calculate station S/N as a function of station gain distribution and IC amplifier (A_1) noise figure. It normalizes the noise figure F_1 versus frequency characteristics of preamplifier A_1 to the highest frequency channel value. It normalizes the input signal S_i to the highest channel frequency signal level so that system analysis for all channels uses only this one signal level. The signal level normalization accounts for the cable loss versus frequency response and for the block tilt used.

The S/N degradation factor is denoted as $D_{S/N}$ and is a function of the following:

1. frequency
2. cable loss and cable system flat loss
3. F_1 versus frequency variation
4. poweramplifier to preamplifier noise figure ratio
5. interstage loss (flat and sloped)
6. input network loss (flat and sloped)
7. output network loss
8. block tilt
9. preamplifier and poweramplifier gain
10. station gain

The single repeater station S/N equation using the S/N degradation factor has the form:

$$(5.2) \quad S_i - N_i = 59 + S_i \text{ (at 300 MHz)} - F_1 \text{ (300 MHz)} + D_{S/N} \qquad \text{in dB}$$

S_i is the input signal level in dBmV at the highest frequency, which for this analysis is 300 MHz.
F_1 is the noise figure of the IC preamplifier at 300 MHz.
$D_{S/N}$ is the signal to noise ratio degradation factor and is equal to D_F plus a level normalization factor, a noise figure normalization factor and a block tilt factor, B_F.

Figure 8 shows the general shape that input levels to a repeater station will have for flat level (solid curve) operation and block tilted level (dashed curve) operation. Reducing levels of lower frequency channels by block tilting levels lowers the S/N of these channels, but it will be shown in the sections on station distortion analysis that station distortion performance is improved. Block tilting levels provides a compromise between station S/N ratio on all channels and station distortion performance. Equation (5.3) defines $D_{S/N}$:

$$(5.3) \quad D_{S/N} = -10 \log \left[1 + \frac{k}{g_1 l_2}\right] - L_1 + \Delta A + B_F$$

Where ΔA is the difference in level in dB for each carrier frequency station input level compared to the station input level at 300 MHz.
B_F is the number of dB which must be subtracted from each channel for a given block tilt, compared to flat level operation.

The last two terms added to the expression for $D_{S/N}$, ΔA and B_F, could have been omitted since the effect they produce is taken into account in the S_i term of equation (5.1). However, by including them in the $D_{S/N}$ factor, the S_i term becomes a constant defined at one frequency. Since the $D_{S/N}$ factor contains other terms which vary with frequency, it is convenient to combine the ΔA and B_F terms with the D_F term. The term ΔA is positive and thereby increases the S/N. The term B_F will either be negative or zero at channel frequencies across the band. The cable system S/N will be highest when $D_{S/N}$ is most positive.

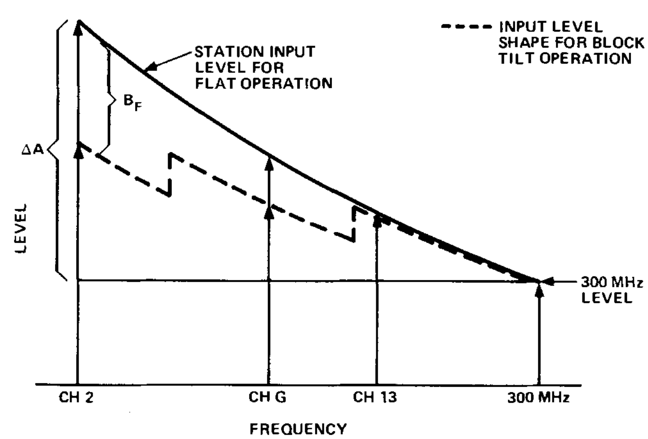


Figure 8. Station Input Level Relationships

5.1 Example of $D_{S/N}$ Application

The values of $D_{S/N}$ plotted in Figure 9 were calculated using the design parameters given in Section 3.0. The $D_{S/N}$ for flat level (0/0/0/0/) operation is plotted in Figure 9a, and a block tilt level (4/2/0/0) condition in Figure 9b. A set of $D_{S/N}$ values for input equalizer values of 8 dB and 14 dB is shown in each figure. Values of $D_{S/N}$ for Channels 2, 6, and 13 for two level tilt conditions and two input equalizer values are taken from the curves shown in Figure 9 (a & b) and listed in the table below. Also listed in the table are station S/N for an A_1 noise figure of 7.5 dB (300 MHz) and S_i (input level @ 300 MHz) of +9 dBmV. The S/N is evaluated from equation 5.2.

5.2 $S_i - N_i = 59 - F_1\ (300\ \text{MHz}) + S_i\ (300\ \text{MHz}) + D_{S/N}$

CHANNEL	Level Tilt 0/0/0/0 Input EQU=14 dB		Level Tilt 4/2/0/0 Input EQU=8 dB	
	$D_{S/N}$	S/N	$D_{S/N}$	S/N
2	1.98	62.5	-.14	60.4
6	1.38	61.9	-.83	59.7
13	1.09	59.4	-.14	60.4

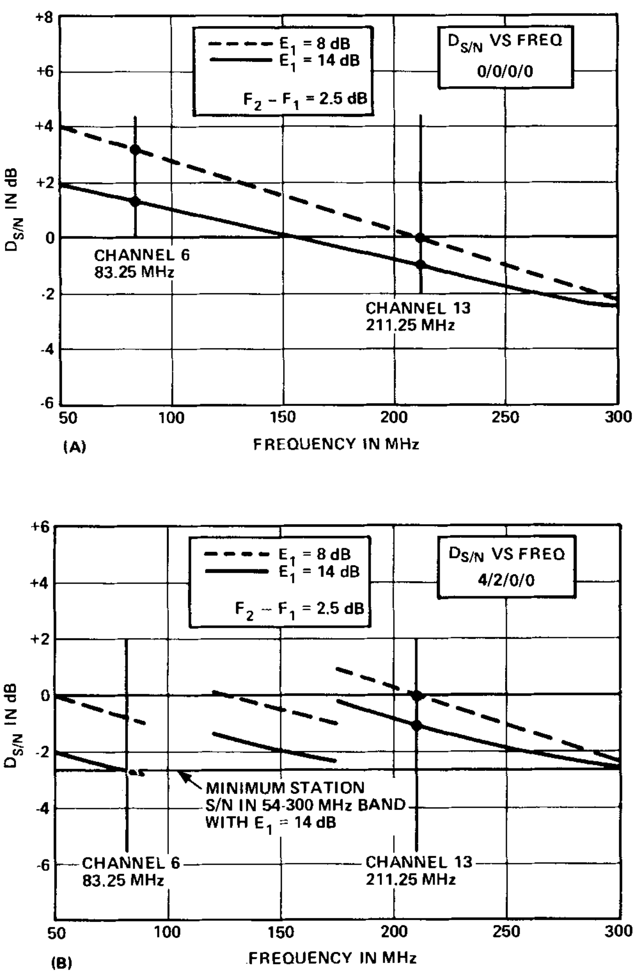


Figure 9. Comparison of $D_{S/N}$ versus Frequency For Two Level Tilts and Two Different Input Equalizers

The station design with a 14 dB input equalizer and operation with flat levels results in a better S/N (62.5 versus 60.4) at Channel 2 and worse S/N (59.4 versus 60.4) at Channel 13 than the station design with an 8 dB input equalizer and block tilted 4/2/0/0 levels. Choice of level tilt obviously has a significant effect on station S/N.

The system operator has some leeway in selecting a trunk station level tilt, but before deviating from the station manufacturer's recommended level tilt values, he must understand the consequences on the station S/N and distortion parameters.

6.0 Signal Level Relationships (Block Tilts)

The amplifier cascade design is a unity gain design. The gain of each repeater amplifier station at each frequency in its passband is set to exactly offset the loss which precedes it. Therefore, the relationship between levels at each frequency in the passband at the output of each repeater station will be identical to the level relationship set at the headend, modified only by the response flatness of the amplifier cascade. An example is shown in Figure 10.

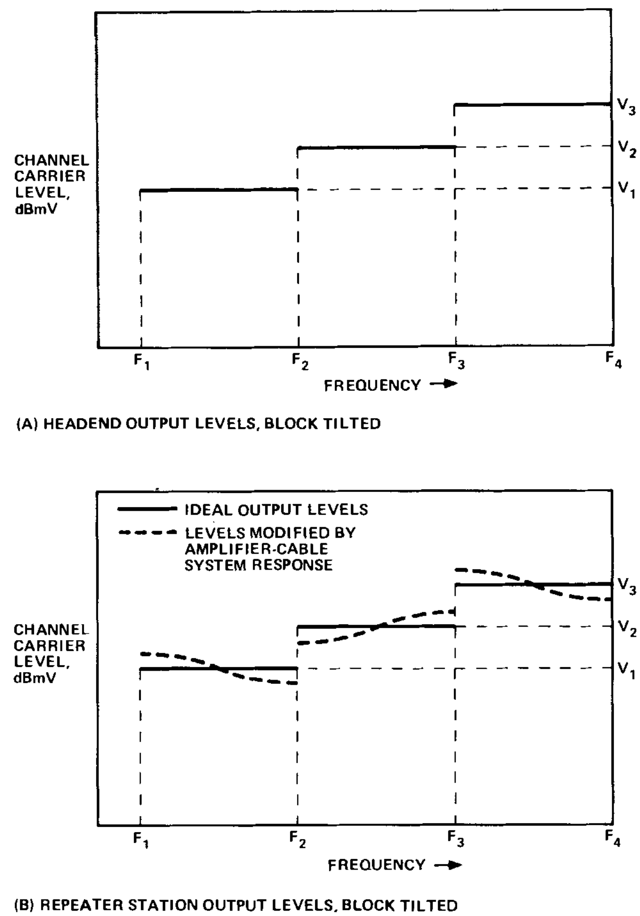


Figure 10.

The level at the repeater station output normally refers to the levels of the channel carriers in the highest frequency band block. In Figure 10, the level V_3 for channel carriers between frequencies f_3 and f_4 is referred to as the operating level of the station. The passband of the amplifier of Figure 10 is $f_4 - f_1$ and is divided into three frequency bands of blocks, $f_4 - f_3$, $f_3 - f_2$, and $f_2 - f_1$. When the level of each of the bands or blocks of carrier frequencies is set to a constant level, the levels are described as "block-tilted". Choice of the magnitude of difference in levels between frequency blocks ($V_3 - V_2$ and $V_2 - V_1$) and number of blocks is based on obtaining the result which gives the best compromise to maximize system S/N and minimize system distortion. A given repeater amplifier design may dictate the use of a specific block tilt to be used in order to optimize system performance. However, many times enough station performance margin is available so that the choice of a number of different "block tilts" can be used with a given repeater station design without sacrificing system performance.

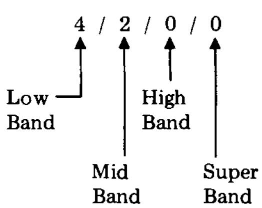
The most important reason for selecting and operating with amplifier station output levels which are *not equal* at each frequency in the passband is that the amount of signal power at the station output can be reduced. Less signal power results in many lower magnitude distortion products, although some distortion product magnitudes will increase. Offsetting the advantage of reduced level of distortion products by use of a level tilt is a decrease in signal-to-noise ratio on those channels at reduced levels. However, consider that the level of channel 2 carrier frequency is attenuated by the cable approximately 13 dB less than a carrier frequency at 300 MHz for 22 dB spacing at 300 MHz. Therefore, reducing the levels of the low frequency channels can be down without reducing the S/N of low frequency channels relative to higher frequency channels. Also, the noise figure of the active amplifying devices have somewhat lower noise figures at lower frequencies relative to higher frequencies.

6.1 Shorthand Notation for Block Tilts in Use

The frequency band from 54 to 300 MHz is commonly divided into four bands.

Low band	54-88 MHz	CH 2-6	5 channels
Mid band	120-174 MHz	CH A-I	9 channels
High band	174-216 MHz	CH 7-13	7 channels
Super band	216-300 MHz	CH J-CH W	14 channels

Using the superband block of frequencies for the reference level, the block tilt (shorthand) notation takes the form:



The high band levels are the same as the super-band levels, mid-band levels are 2 dB below the super-band levels, and low-band levels are 4 dB below super-band levels. A summary of block tilts specified for trunk stations taken from a number of manufacturers' data sheets in the past year are summarized in Table V.

Table V. Examples of Block Tilts Published in Manufacturer's Data Sheets

COMPANY	BLOCK TILT
Anaconda	4/2/0/0
Jerrold	3/3/0/0
Theta-Com	7/4/2/0
Scientific Atlanta	4/2/2/0
Magnavox	0/0/0/0 (No block tilt)

6.2 General Comments on Distortion Product Accumulation in Broadband VHF Amplifiers.

Care must be exercised in attempting to analytically predict second order intermodulation product, triple beat (third order) product, and crossmodulation product accumulation in a single channel in a broadband VHF amplifier as a function of channels and amplitudes of individual channels. The only practical way to evaluate the cumulative distortion products of an amplifier station is by measurement. However, calculation of the magnitude of single second and third (triple beat) order beat products can be done precisely, and thereby provide considerable insight into the station design.

For reasons pointed out in both references 6 and 7, broadband VHF amplifiers used in CATV repeater stations have transfer characteristics which in general, invalidate the rule that "n channel crossmodulation distortion is n-1 times the 2 channel crossmodulation value". However, because measurement on a fixed gain amplifier in common use today does follow the "n-1 rule" for crossmodulation accumulation in channel 2 for trunk station application, a B_x factor for channel 2 is analytically derived. Obtaining a B_x factor must normally be done by measurement first, followed by analysis. Table VI contains a summary of the methods of analysis for these types of distortion products.

Table VI. Amplifier Distortion Characteristics Analysis Methods

DISTORTION PRODUCT AS FUNCTION OF LEVEL (Flat, Block, Linear) TILT	METHOD OF ANALYSIS
2nd Order → 2-Channel (Single Beat) → Accumulation (n-channel)	Analytical Measurement in Each Channel
Triple Beat → 3-Channel (Single Beat) → Accumulation (n-channel)	Analytical Measurement in Each Channel
Crossmodulation → Accumulation (n-channel) → Accumulation (n-1 times 2-channel)	Measurement in Each Channel Analytical

6.3 Second Order IM (A ± B) Block Tilt Factor, B_I

A flat gain amplifier operated with a pair of equal level signals produces second order beats (A + B and A – B) which, if in the passband of the amplifier, interfere with a third signal. The ratio of second order beat to the signal for flat output levels is different than the ratio obtained when the amplifier is operated with linear or block tilted levels. The difference in second order IM ratio (in dB) resulting for tilted level operation versus flat level operation for the same amplifier is defined by a block tilt factor B_I. Values of B_I for six pairs of channel carrier frequencies and three level tilts are summarized in Table VI.

As an example, for the numbers listed in Table VII, the 2nd order IM product of Channel 13 minus (—) Channel G producing a beat which interferes with Channel 2 is 2 dB worse when operating with a 4/2/0/0 block tilt than with no block tilt (0/0/0/0) or flat levels. However, the IM ratio for the Channel 2 plus (+) Channel G beat in Channel 13 provides a 6 dB improvement for the 4/2/0/0 block tilt condition.

The magnitude of the combination of total number of second order beats occurring at a specific frequency for a specific system application and given number of channels can practically be determined by measurement only. Second order product addition depends on magnitude *and phase* of each product.

Table VII. Block Tilt Factor B_I For Six Pairs of Channel Carrier Frequencies and Three Block Tilts.

RELATIVE IM DISTORTION PRODUCT (TV CHANNELS)	BLOCK TILT FACTORS B_I IN dB		
	4/2/0/0	3/3/0/0	0/0/0/0
R – 13 (in Ch 2)	+4	+3	0
13 – G (in Ch 2)	+2	0	0
13 – 2 (in Ch G)	–2	0	0
2 + G (in Ch 13)	–6	–6	0
13 + 2 (in Ch R)	–4	–3	0
R – 2 (in Ch 13)	–4	-3	0

6.4 Third Order [Triple Beat (A ± B ± C)] Block Tilt Factor, B_T

The B_I factor is calculated considering the level difference of two frequencies and the level difference between that of the beat frequency and that of the frequency with which it interfaces. The B_T factor is calculated in a similar manner, except that the level difference of three frequencies rather than two frequencies is accounted for. A third order IM (2A ± B) factor could also be calculated, but is not included in this paper.

Triple beat products are considered to be a limiting parameter in 30+ channel systems with today's amplifiers. One complication of the triple beat problem is the sheer number of possible products for 30+ channel operation. It is therefore naive to assume that the calculation of a few triple products define an amplifier station's triple performance. However, calculation of the triple beat product amplitude for a few products for a given level tilt relative to flat level operation will indicate a trend, indicating whether triple beat magnitude reduction is possible by the use of level tilts. A more useful and detailed analysis of this problem must be reserved for another paper.

B_T factors for a few triple combinations and level tilt conditions are summarized in Table VIII.

Table VIII. Triple Beat Block Tilt Factors as Function of Block Tilt

TRIPLE BEAT PRODUCTS LISTED BY CHANNEL COMBINATIONS	TRIPLE BEAT BLOCK TILT FACTOR B_T IN dB		
	4/2/0/0	0/0/0/0	3/3/0/0
3 + 4 + A in 0	-10	0	-9
A – J + Q in 5	+2	0	0
G + S – J in 13	-2	0	-3
13 + R – J in Q	0	0	0
7 – 13 + L in G	+2	0	+3

A negative number in the table indicates a lower amplitude beat at the station output compared to flat level operation and a positive number indicates that the triple beat product for that specific three channel beat product is higher (or worse) than that resulting with flat level operation.

6.5 Crossmodulation Block Tilt Factor, B_X

The effect of tilted levels and fixed gain preamplifier crossmodulation distortion contribution for a given number of channels must be known in order to determine station XM from the given flat level XM specification of the fixed gain preamplifier and poweramplifier. Remembering the precautionary comments of Section 6.2, and applying the "n-1 XM accumulation" rule, a B_X factor for channel 2 is derived. The station level tilt factor, B_X, defines the difference between the XM produced with flat levels, compared to the XM expected with tilted levels for a given number of channels in a given channel. The B_X term is then included in the XM D-factor, D_X, for a given station design.

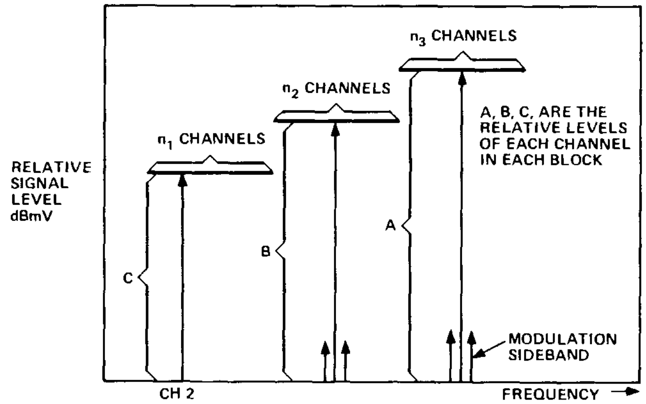


Figure 11. Block Tilted Levels Versus Frequency

The crossmodulation at the level of channel 2 for block tilted levels compared to flat levels is given by equation 6.5.1 (assuming equal modulation percentage on each modulated channel.)

$$(6.5.1)\ m_2 = \frac{(n_1 - 1) + n_2\,(b/c)^2 + n_3\,(a/c)^2}{n - 1} \qquad \text{Numeric}$$

where $n_1 + n_2 + n_3 = n$ (total number of channels)

To obtain the B_x factor for channel 2 compared to flat level operation at level A (which would be the flat level condition):

$$(6.5.2)\ B_x\ (\text{Channel 2}) = 20 \log (m_2) - (A\text{-}C)$$

As an example of the use of equation 6.5.2, consider the following:

Seven high band channels at level A, 5 dB above five (5) low band channels at level C.

$$n_1 = 5,\ n_3 = 7,\ A/C = 10 \exp (5/20) = 1.79$$

Substituting these values in (6.5.1)

$$m_2 = \frac{(5 - 1) + 7\,(1.79)^2}{(12 - 1)} = \frac{4 + 22.4}{11} = 2.4$$

Evaluating equation 6.5.2

$$B_x\ (\text{Channel 2}) = 20 \log (2.4) - 2\,(5) = -2.4 \text{ dB}$$

Operating with 12 channels, 5 of the channels 5 dB below level A, results in channel 2 XM 2.4 dB below the XM value of channel 2 when all 12 channels are at level A.

Using equations 6.5.1 and 6.5.2, the B_x factor for channel 2 is calculated and summarized in the table for a number of block tilt and number of channel conditions:

Table IX. XM Block Tilt Factor B_x
(Calculated using n-1 rule.)

	12 CHANNELS				
	LOW		HIGH		B_x CH 2
No. Chan.	5		7		
Relative Lev.	-3 dB		0 dB		-1.7 dB
	-5 dB		0 dB		-2.5
	21 CHANNELS				
	LOW	MID	HIGH		
No. Chan.	5	9	7		
Relative Lev.	-3	-3	0		-3.4
	-5	-2	0		-3.1
	35 CHANNELS				
	LOW	MID	HIGH	SUPER	
No. Chan.	5	9	7	14	
Relative Lev.	-3	-3	0	0	-1.8
	-4	-2	0	0	-1.6

7.0 Normalized Poweramplifier Second Order IM Distortion Degradation Factor, D_I

The quantity D_I is derived so that repeater station 2nd order IM distortion as a function of station gain distribution and block tilt can be easily related to the flat level IC preamplifier and poweramplifier IM distortion specifications. One D_I is calculated for each A + B and/or A − B product considered necessary to define the station IM performance across the passband.

(7.1) $I_s = I_2 + D_I$ in dB

The quantity I_s is the relative 2nd order IM for a pair of frequencies interfering with a third frequency at the station output. The quantity I_2 is the relative 2nd order IM ratio in dB for the same pair of frequencies at equal levels at the poweramplifier output.

D_I, the degradation factor, is independent of the absolute station output signal levels and is a function of the following:

1.0 Preamplifier IM distortion relative to poweramplifier distortion.
2.0 Preamplifier gain and poweramplifier gain.
3.0 Interstage network loss (flat and sloped), a function of station gain
4.0 Block tilt.

Two assumptions are made in the derivation of D_I which do not affect the result in general. First, the IM distortion specification for each pair of frequencies of the flat preamplifier is assumed to be related to the IM distortion specification for the same pair of frequencies of the flat poweramplifier by a constant number of dB. Second, the IM distortion of the preamplifier is assumed to add in phase with the IM distortion of the poweramplifier. This represents a worst case condition for the total IM distortion of the repeater station. Second order IM addition is phase and amplitude sensitive so that in the actual hardware design, some degree of cancellation will occur, either by accident or by design.

In the following analysis, lower case letters indicate numeric, and upper case letters indicate dB. The total repeater station IM distortion for a given pair of signals is the sum of the IM distortions generated by all sections of the amplifier. Assuming that the only sources of IM distortion products are the preamplifier and poweramplifier, then the station distortion is:

(7.2) $20 \log_{10} (i_1 + i_2) = 20 \log_{10}[i_2 (1 + i_1/i_2)]$

$i_1 = 10 \exp (I_1/20)$ Preamplifier IM distortion

$i_2 = 10 \exp (I_2/20)$ Poweramplifier IM distortion

The terms I_1 and I_2 represent the second order distortion ratio in dB for a given pair of frequencies as measured for the preamplifier and poweramplifier with flat output levels. Modification to the IM ratios for block tilt operation compared to flat level will be added later.

(7.3) The station second order IM distortion can be written as:

$$I_s = \underbrace{20 \log (i_2)}_{\text{Poweramplifier IM Distortion}} + \underbrace{20 \log (1 + i_1/i_2)}_{\text{IM Distortion Contributed by Preamplifier}}$$

The numeric ratio (i_1/i_2) may be solved by working with the amplifier constants in dB, since the numeric ratio can be written:

(7.4) $i_1/i_2 = 10 \exp \quad (I_1 - I_2)/20$

where I_1 and I_2 are in dB.

The difference between preamplifier and poweramplifier IM as measured with flat levels at the same level, is assumed equal to a constant D.

(7.5) $D = I_1 - I_2$ in dB (a positive number).

The quantity D is usually positive, since the preamplifier IM distortion is normally worse (less negative) than the poweramplifier IM distortion. If D = 0 dB, the preamplifier IM distortion is equal to the poweramplifier IM distortion.

Relative second order IM distortion is proportional to signal level, i.e., two signals whose levels are each changed by 1 dB produce a relative IM product change of 1 dB. As an example and to prepare for the analysis which follows, consider a poweramplifier with flat output levels preceded by a flat loss network and a flat output preamplifier, as shown in Figure 12.

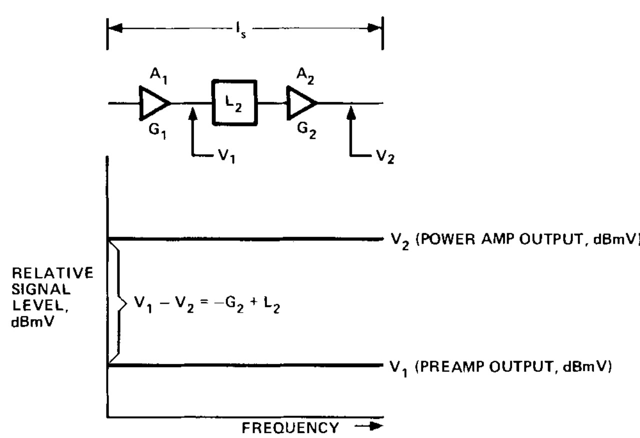


Figure 12. Flat Output Preamplifier and Poweramplifier

The ratio of preamp (A_1) to poweramp (A_2) distortion can be written:

$$(7.6)\quad (I_1 - I_2) = \underbrace{(V_1 - V_2)}_{\text{Due to level difference of } A_1 \text{ and } A_2} + \underbrace{D}_{\text{Due to IM difference of } A_1 \text{ and } A_2}$$

The preamp (A_1) output level V_1 is G_2 dB below the poweramp (A_2) output level V_2 increased by the interstage loss L_2. Equation 7.6 can then be written as:

$$(7.7)\quad (I_1 - I_2) = -G_2 + L_2 + D \qquad \text{in dB}$$

The station distortion I_s shown in Figure 12 can then be written:

$$(7.8)\quad I_s = I_2 + \underbrace{20 \log \left[1 + 10 \exp\left[(I_1 - I_2)/20\right]\right]}_{\text{Degradation of Poweramp IM Distortion Spec}}$$

In a cable repeater station, neither amplifier A_1 or A_2 may operate with flat levels. The poweramp (A_2) may operate with a level tilt, which has been set at the cable system headend, and the preamp (A_1) levels will be sloped as a result of cable equalization in the station.

It has been instructive to solve the flat level A_1, A_2, and L_2 problem described by equation (7.7). Using this equation as a building block, the final solution is obtained by the following procedure:

Step 1: Solve for the station ($I_1 - I_2$) ratio by considering the actual frequency response of the preamp A_1 output levels. The amount of cable loss, and cable equalization placed before the preamp A_1 will define this level relationship, which is sketched in Figure 13. The term L_2 in equation (7.7) must be modified to account for the sloped levels at preamp A_1 output. If the term L_2 in equation (7.7) is the interstage insertion loss at the frequency which is interfered with by the IM product, then a term ΔL must be added to the equation to account for the level differences of the frequencies at the preamp output.

Equation 7.7 now becomes:

$$(7.9)\quad (I_1 - I_2) = -G_2 + L_2 + \Delta L + D \qquad \text{in dB}$$

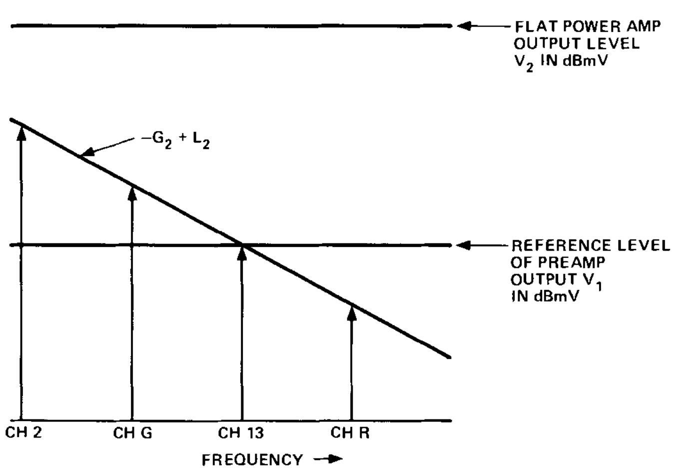


Figure 13. Sloped A_1 Output Levels When A_2 Output Levels are Flat

Step 2: Substitute the value of ($I_1 - I_2$) defined by equation (7.9) in 7.8, add the system block tilt factor B_I, and add L_3 which is any loss between A_2 output and station output terminals (connector).

The station IM ratio for a given pair of frequencies is then

$$(7.10)\quad I_s = I_2 + \underbrace{20 \log \left[1 + 10 \exp\left(\frac{(I_1 - I_2)}{20}\right)\right] + B_I + L_3}_{D_I}$$

The difference between station IM distortion and A_2 IM distortion for a given pair of frequencies is then:

$$(7.11)\quad D_I = 20 \log \left[1 + 10 \exp\left(\frac{(I_1 - I_2)}{20}\right)\right] + B_I + L_3 \quad \text{in dB}$$

where ($I_1 - I_2$) is determined by evaluating equation (7.9) for a given set of station design parameters.

Using A_1 and A_2 specifications and station design parameters given in Section 3.0, equation (7.11) was evaluated and the results plotted in Figure 14 for a flat level system and Figure 15 for a block tilt (4/2/0/0) system. The $D_{S/N}$ factor for channels 2, G, and 13 are also included in Figures 14 and 15 to show the relationship between D_I and $D_{S/N}$ as a function of input equalizer value. Specific values will be taken from these curves and used in the summary (Section 9.0).

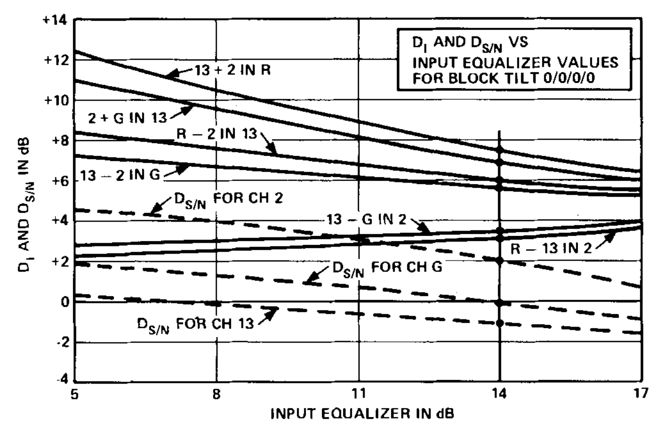


Figure 14. D_I and $D_{S/N}$ versus Input Equalizer Values for Block Tilt 0/0/0/0

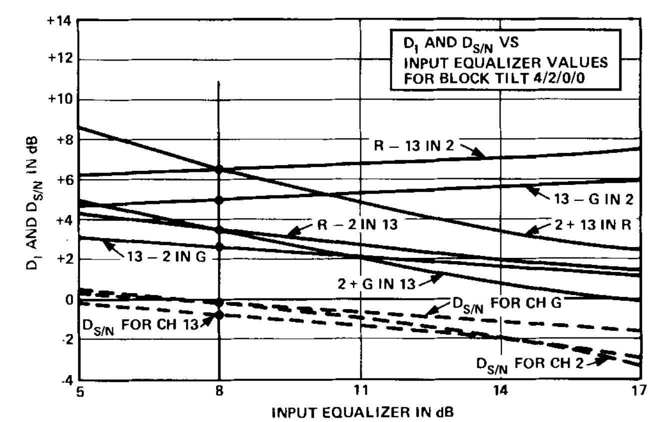


Figure 15. D_I and $D_{S/N}$ versus Input Equalizer Values for Block Tilt 4/2/0/0

8.0 Normalized Poweramplifier Triple Beat (A ± B ± C) Distortion Degradation Factor, D_T

A D_T factor for a few sets of A ± B ± C beats are calculated for the A_1 and A_2 specifications and station design parameters given in Section 3.0. The equations used for the D_T factor calculation are similar to those used for the D_I factor, with the exception that the relative levels of three frequencies causing a beat interfering with a fourth frequency are determined compared to relative levels of two frequencies causing a beat interfering with a third frequency. The station triple beat product in dB for three frequencies is:

(8.1) $T_s = T_2 + D_T$

Assuming that the only sources of triple beat distortion are A_1 and A_2:

(8.2) $20 \log (t_1 + t_2) = 20 \log [t_2 (1 + t_1/t_2)]$

$t_1 = 10 \exp (T_1/20)$ A_1 distortion

$t_2 = 10 \exp (T_2/20)$ A_2 distortion

The station triple beat distortion can be written as:

(8.3) $T_s = \underbrace{20 \log (t_2)}_{A_2\ \text{TB Distortion}} + \underbrace{20 \log (1 + t_1/t_2)}_{\text{TB Distortion Contributed by } A_1}$

The ratio of A_1 to A_2 triple beat distortion in dB is:

(8.4) $T_1 - T_2 = (2)(-G_2 + L_2) + \Delta L + T$

where: T is the difference between A_1 and A_2 triple beat product amplitude specification measured at the same flat levels

ΔL is a factor which accounts for the effect of sloped A_1 output levels on T_1

L_2 is the interstage network attenuation at the channel frequency interfered with by the triple beat frequencies

The station triple beat ratio for a set of three frequencies then becomes, using the value of $(T_1 - T_2)$ evaluated in (8.4).

(8.5) $$T_s = T_2 + 20 \log \left[1 + 10 \exp \left(\frac{(T_1 - T_2)}{20}\right)\right] + B_T + 2L_3$$

The factor B_T accounts for the effect of any level block tilt at the station output. The factor (2) appears in both equations (8.4) and (8.5) because of the assumption that the triple beat product changes 2 dB for every 1 dB of level change of the three beat producing frequencies. The D_T factor is:

(8.6) $$D_T = 20 \log \left[1 + 10 \exp \left(\frac{(T_1 - T_2)}{20}\right)\right] + B_T + 2L_3$$

Values of D_T for the beat products listed in Table VIII are calculated and listed in Table X in Section 10.

9.0 Normalized Poweramplifier Crossmodulation Distortion Degradation Factor, D_X

The D_I and D_T factors for a given set of channel frequencies are easily calculated, becoming cumbersome only by the number of sets required to describe station performance over its entire passband. These factors do not consider the accumulation effect of beats at a given frequency. The D_x factor, however, must include the accumulation of crossmodulation from each channel onto a given channel carrier to be of any use. For the reasons discussed in Section 6.2, the method of calculating n-channel XM by assuming that n-channel XM is (n-1) times two channel XM may not be accurate. However, this calculation method does result in the maximum value of XM distortion for n-channel XM accumulation, if the two channel XM is pure AM (Ref. 6). As has been previously stated, channel 2 XM for one type of IC amplifier (A_1 and A_2) in common use today, can be analyzed fairly accurately by using the n-1 accumulation rule. Just remember that the following analysis must be applied with good judgment and only after the A_1 and A_2 XM characteristics have been adequately defined by XM measurements.

The D_x factor for channel 2 is calculated using equations similar to those used for calculation of D_T (equations 8.1 thru 8.6).

(9.1) $X_s = X_2 + D_x$ (One for each channel.)

Assuming that the only sources of XM distortion are A_1 and A_2.

(9.2) $20 \log (x_1 + x_2) = 20 \log x_2 (1 + x_1/x_2)$

$x_1 = 10 \exp (X_1/20)$

$x_2 = 10 \exp (X_2/20)$

The station XM distortion can be written as:

(9.3) $X_s = 20 \log (x_2) + 20 \log (1 + x_1/x_2)$

The ratio of A_1 to A_2 XM distortion is:

(9.4) $X_1 - X_2 = 2(-G_2 + L_2) + B_x' + D_x'$

D_x' is the ratio of A_1 to A_2 XM measured at the same flat level

B_x' is a factor accounting for the sloped preamp (A_1) output levels. The magnitude of this factor varies as a function of input equalizer value and number of channels.

L_2 is the relative level of the channel for which D_x is being calculated.

A simplifying assumption is made to reduce the complexity of the B_x' factor by dividing the sloped A_1 output levels into average level flat blocks as sketched in Figure 16. Calculated

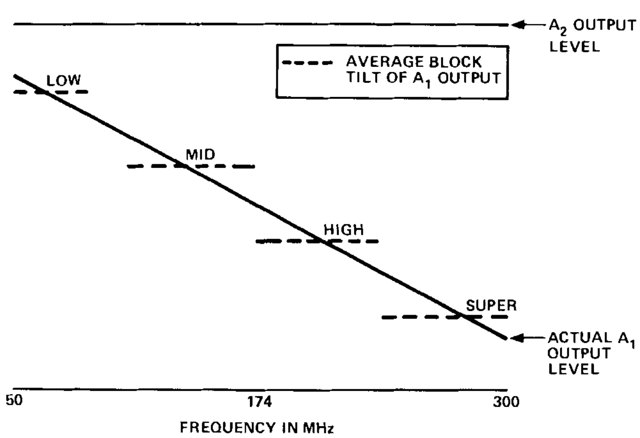


Figure 16. Approximation of Sloped A_1 Output Level by Flat Level Blocks

relative preamp output levels for two input equalizer and block tilt values and the design parameters of Section 3.0 are shown in Figure 17.

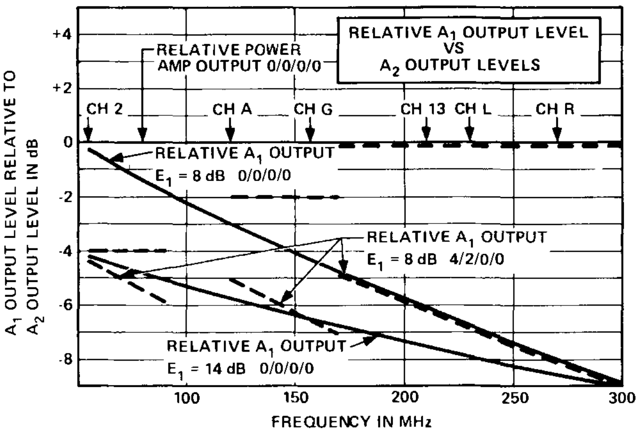


Figure 17. Relative Preamp Output Level versus Frequency

The station XM is now described by:

$$(9.5) \quad X_s = X_2 + 20 \log \left[1 + 10 \exp\left(\frac{(X_1 - X_2)}{20}\right)\right] + B_X + 2L_3$$

The B_X factor accounts for the effect of any level block tilt at the station output. The factor (2) appears in both equations (9.4) and (9.5) because of the assumption that the XM ratio changes 2 dB for every 1 dB level change of every channel. The D_X factor is:

$$(9.6) \quad D_X = 20 \log \left[1 + 10 \exp\left(\frac{X_1 - X_2}{20}\right)\right] + B_T + 2L_3$$

D_X factor is evaluated for channel 2 using design parameters in Section 3 and listed in Table X of Section 10.

10.0 SUMMARY

The D-factors calculated for two station designs, differing only by the values of input and interstage equalizers and two block tilts are summarized in Table X. The station, A_1, and A_2 parameters used in the calculations are those given in Section 3.0. Note that both station gain distribution (input, interstage and output loss networks) and system block tilt are interrelated and have a very significant effect on the station S/N and distortion.

The station specifications for the design parameters listed in Section 3 and in Table X can now be easily calculated by adding the D-factors listed in Table X to the IC amplifier specifications at the station operating level selected.

Comparison of the two columns of D-factor numbers in Table X, indicates that each design has its strengths and weaknesses. The final design selection requires more extensive analysis of the D_T factors and also inclusion of measured weighting factors in the analysis to account for the distortion product versus frequency characteristic of the IC amplifiers.

Table X. Summary of D-Factors Calculated for Parameters Given in Section 3.0

D-FACTOR PARAMETERS	INPUT EQU. 14 dB BLOCK TILT 0/0/0/0	INPUT EQU. 8 dB BLOCK TILT 4/2/0/0
D_F		
2	12.8	10.9
G	7.7	6.3
13	5.7	4.7
W	2.8	2.6
$D_{S/N}$		
2	2.0	-.1
G	-.1	-.7
13	-1.1	-.1
W	-2.7	-2.5
D_I		
R - 13 in 2	3.2	6.5
13 - G in 2	3.6	5.0
13 - 2 in G	5.7	2.6
2 + G in 13	7.0	3.4
2 + 13 in R	7.4	6.5
R - 2 in 13	6.0	3.5
D_T		
3 + 4 + A in 0	8.3	-3.8
A - J + Q in 5	4.9	7.1
G + S - J in 13	5.3	3.9
13 + R - J in Q	5.3	5.9
7 - 13 + L in G	5.2	8.7
D_X		
Chan 2	5.9	5.5

11.0 CONCLUSION

An analysis method using normalized D-factors for relating the noise and distortion parameters of a fixed gain preamplifier and poweramplifier to the cable repeater trunk station amplifier has been developed. The station design configuration and level tilt limit the noise and distortion performance obtainable from any given fixed gain IC amplifier. Use of the D-factors by the equipment manufacturer, IC manufacturer, and end user provides an aid in analyzing the relationships between IC amplifier performance specifications and trunk amplifier station performance specification. It must be noted that the final station design is determined by combining both measured and analytical results.

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SYMBOLS DEFINITIONS

A_1 Fixed gain preamplifier, could be IC or Discrete component design.

A_2 Fixed gain power amplifier, could be IC or Discrete component design.

G_1 Gain of A_1 in dB

G_2 Gain of A_2 in dB

G_s Station Gain

F_1 Noise Figure of A_1 in dB

f_1 Noise factor of A_1, numeric

F_2 Noise Figure of A_2 in dB

f_2 Noise factor of A_2, numeric

F_s Noise Figure of Station in dB

I_1 Carrier to second order intermodulation product ($A \pm B$) ratio of A_1 in dB

I_2 Carrier to second order intermodulation product ($A \pm B$) ratio of A_2 in dB

I_s Carrier to second order intermodulation product ($A \pm B$) ratio of station in dB for given I_1, I_2, station design, and block tilt

X_1 Carrier to crossmodulation distortion ratio of A_1 in dB, measured with equal channel levels for given number of channels.

X_2 Carrier to crossmodulation distortion ratio of A_2 in dB, measured with equal channel levels for given number of channels.

X_s Carrier to crossmodulation distortion ratio of station for given X_1, X_2 and station design and block tilt.

T_1 Carrier to triple beat product ($A \pm B \pm C$) ratio of A_1 in dB

T_2 Carrier to triple beat product ($A \pm B \pm C$) ratio of A_2 in dB

T_s Carrier to triple beat product ($A \pm B \pm C$) ratio of station in dB for given T_1, T_2, station design, and block tilt.

E_1 dB value of cable loss at reference frequency for which input equalizer equalizes.

E_2 dB value of cable loss at reference frequency for which interstage equalizer equalizes.

FL_1 Flat loss in dB proceding preamplifier A_1.

FL_2 Flat loss in dB in interstage network between A_1 output and A_2 input terminals.

L_3 Flat loss in dB between A_2 (poweramplifier) output and station output.

L_1 Total loss of input networks, flat loss plus equalizer loss.

L_2 Total interstage loss in dB equal to sum of interstage equalizer loss and flat loss.

D_F A_1 noise figure F_1 degradation factor for given station design $F_s = F_1 + D_F$ in dB.

$D_{S/N}$ Signal to noise ratio degradation factor for given station design. $S_i - N_i = 59 + S_i$ (ref. frq) $- F_1$ (ref. frq) $+ D_{S/N}$

$D_{S/N}$ includes D_F, block tilt factor, signal level variation and noise figure variation across frequency passband.

D_I A_2 second order IM degradation factor for given station design and I_2.

$I_s = I_2 + D_I$ (One D_I required for each $A + B$ and $A - B$ product).

D_T A_2 third order (triple beat) degradation factor for given station design and T_2.

$T_s = T_2 + D_T$ (One D_T required for each $A \pm B \pm C$ product).

D_x A_2 crossmodulation degradation factor for given station design and channel.

$X_s = X_2 + D_x$

B_I Block tilt factor relating difference, in dB, of second order IM product magnitude measured at block tilted levels compared to the magnitude measured at flat levels. Each $A \pm B$ product has a B_I.

B_T Block tilt factor relating difference, in dB, of third order (triple beat) product magnitude measured at block tilted levels compared to the magnitude measured at flat levels. Each $A \pm B \pm C$ product has a B_T.

CATV PROOF OF PERFORMANCE TESTING

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INTRODUCTION

Testing a cable system to determine compliance with the FCC standards can provide results which are truly meaningful when related to the quality of service the subscriber receives. In some instances though certain test procedures may produce results which bear little relationship to the picture quality as seen on a home television set.

Because video measurements rather than RF transmission measurements have a more direct relationship to what is seen on a television screen that approach has been taken in this paper. The more conventional methods of testing CATV distribution systems often do not include the effects of headend equipment or microwave where it is used. Baseband video measurements also enable the tests to be done on an in service basis with the system in normal operation. Invaribly in service test methods will gain wider acceptance with engineers and subscribers than will sleepless nights and interrupted programs.

These techniques require equipment which is often not a regular part of a cable system's test equipment. They are not the least expensive nor the most expensive methods to perform the FCC tests, but they do provide data which can give a real indication of the quality of service the subscriber receives. In many cases, the tests described can be done faster and with more accuracy than conventional distribution system only tests.

THE TEST POINTS

At the present time the required FCC tests to determine compliance with the standards are, with the exception of radiation tests, very closely tied to the subscriber terminal. The subscriber terminal lies at the interface between cable system owned equipment and subscriber owned equipment. In most cases this means the point to conduct tests or which to relate any other measuring points to is the output of the 75 ohm unbalanced to 300 ohm balanced transformer.

The output of the transformer is where the interface takes place. Some engineering judgement must, of course, be used in selecting the actual test point. In the test methods that follow, the actual measurements are made at the 75 ohm unbalanced impedance side of the transformer. In a converter equipped system, this point is at the output of the converter. The assumption that the transformer does little to degrade the quality of service is usually a very valid one.

FCC rules require that the tests be conducted "at no less than three widely separated points in the system, at least one of which is representative of terminals most distant from the system input in terms of distance." If the tests are viewed as an opportunity to monitor your system's performance, rather than as a burden or hardship, it is then logical to place all three test points at system extremities, preferably on different trunk lines. Each test point should be near the end of a feeder line and after all active equipment. Two drops at each test point should be available near ground level. On a tap having more than two ports, the two test point drops should be hooked up to two ports fed from the same two-way splitter within the tap. In this way, you can simulate worst case customer to customer isolation. Each test point should have twenty-five (25) feet of drop cable between the tap and the actual test point near ground level. The two drops should be installed permanently by placing the ground level test points in a waterproof junction box with the cable running down the pole in conduit. They should be terminated when not in use. Excess cable should be coiled up and fastened on the line near the pole, put in the junction box, or in the case of underground plant, put in a pedestal. Use approved construction practices in all cases.

By establishing the test points permanently, measurements will have more repeatability and comparisons can be made from year to year, or on a more frequent routine maintenance basis.

When the tests are actually conducted another one hundred (100) feet of drop cable can be added to reach the test equipment so that a worst case customer drop should result. As much as possible, try to establish the three test points at locations where a vehicle carrying test equipment can drive up and have easy access to the test point. Power for the test equipment can come from a small gasoline generator. The generator is recommended over inverters or belt driven generators for vehicles. A gasoline generator usually has better regulation, it is easier to use, and the dollar/watt ratio is better! If the generator is kept about seventy-five (75) feet or more away from the test antenna, ignition interference should be no problem during radiation measurements.

THE LOG FORMS

Sample log forms which were, in part, inspired by Tektronix CATV Proof of Performance forms are included at the end of the paper. The log forms, test procedures, and a copy of the FCC Rules and Regulations regarding CATV tests and standards which were in affect at the time of testing should be put in an appropriately labeled binder and retained at the system local office for at least five years, as required by the FCC. A summary of the FCC standards and tests now in effect is included at the end of this paper.

Keep in mind that proof of compliance with FCC standards at the three selected test points in no way relieves the system from the responsibility to maintain such standards at every subscriber terminal. Installation and service records which record signal levels and picture quality should provide evidence to the FCC that every subscriber is being properly served.

Since the installer's or technician's field strength meter or other similar instrument is the key to maintaining system quality, their calibration should be checked regularly and as a part of the FCC performance tests.

LOG FORM INFORMATION

The System Information form, LVOC 21-73, should be completed and on file prior to conducting the tests. The blank spaces in the "Cable CH" column may be used to list FM stations carried on the cable by listing their frequencies. For systems with more than twelve (12) channels, two sheets may be used. The first sheet can list channels 2-13 and the second sheet can list the other channels by striking out 2-13 and putting in the desired channel letter or number in the space below the diagonal.

The second column, "Class", refers to the class of the cable channel as defined by the FCC in Volume III, Part 76, subpart A, 76.5. Class I channels are those cable channels carrying the signals of a television broadcast station. Class II channels are all channels which carry television signals, other than Class I, which can be received by the subscriber without auxiliary decoding equipment. This generally includes local origination, news, and weather channels. Class III channels carry data other than television signals or television signals which require auxiliary decoding equipment. A channel which carried computer data signals, an FM channel or a scrambled television signal would be a Class III channel. Class IV channels are channels which carry signals from a subscriber terminal to any other point in the cable system. For your own information, it is worthwhile to make these performance tests on Class II channels in addition to the required Class I channel testing.

The column labeled "Grade" refers to the Grade A or Grade B field intensity contours of a television broadcast station as defined by the FCC in Part 73, sub-part 73.683. The pickup site location with respect to the Grade A or B contours must be determined and listed. In some cases, the pickup site may be beyond the Grade B contour and should be listed as "none". This information is available from the "TV Digest" or the TV station in question.

The comments column can contain information on programming, i.e., network, independent, news or weather, for example.

<u>Proof of Performance Data</u> form, LVOC 22-73 should be filled out by the person actually making the tests.

<u>System Test Equipment Data</u> form, LVOC 23-73, is not specifically required by the FCC, but is for informational purposes and will serve as a handy record of test equipment calibration data. Because the field strength meter, FSM, is so important to the maintenance of a system, calibration of the system's FSM's as part of the tests should be standard procedure.

List any equipment on a feeder line

(distribution amp, bridger amp, line extender, etc.) under the classification of line extenders on the Test Point Information form LVOC 24-73

A set of System Tests forms, LVOC 26-73 and 27-73 are used at each test point. Cross modulation, intermodulation and co-channel tests are not required by the FCC, but a system operator may want to conduct these tests for his own information, so space has been provided to record the results.

The visual signal level on each channel in dBmV should be listed on the 24 Hour Level Variation Test form, LVOC 28-73. Temperature in degrees Farenheit should be recorded on the bottom row.

THE TESTS

In order to make the tests more manageable, they have been broken up into headend and system tests. At the output of the headend and just before the headend signals enter the distribution system, a directional coupler should be inserted which allows sampling of all outgoing signals. Immediately ahead of this coupler on the headend side of it, another directional coupler should be inserted so that test signals may be inserted into the system. Couplers with eight (8) or twelve (12) dB taps will do the job. Once these are installed, it is a good idea to sweep test them from the sweep input point to the headend output test point. All headend tests are then conducted from the first accessible point in the headend, such as a preamplifier input, to the headend output test point. System tests either test the headend and the distribution system or they are conducted from the system sweep input point to the system extremity test points.

Because accurate results from the carrier to noise and hum or low frequency disturbances tests depend upon knowing the relationship between the video modulating signal and the magnitude of the modulated carrier it is necessary to check the depth of modulation on all channels. On channels where the cable operator has no control over the depth or percentage of modulation, i.e., where channel to channel processors or stripline amplifiers are in use, the broadcast signal will usually be close enough to 87.5% modulation that your measurement error will be insignificant. It is possible to set the modulation on your own modulators by comparison with with a broadcast signal, use of a FSM and a scope, or with a spectrum analyzer. First, check to see that the modulator has the proper composite video input, which in most cases is one (1) volt peak to peak. Using a spectrum analyzer, tune to the channel desired. Then reduce the scan to zero and the bandwidth or resolution to maximum, usually about 3 MHz. Fine tune the analyzer, which is now operating in the time domain, for maximum signal. Then switch from a log to a linear display and adjust the vertical gain until the top of the signal is at the top graticule or reference line. This line now represents the maximum carrier amplitude. The bottom graticule is then zero carrier amplitude.

On a display having eight (8) major vertical divisions, each division represents 12.5% modulation. Set the modulation so that the peak to peak signal displayed on the analyzer occupies seven of the eight major divisions. If a Vertical Interval Test Signal, VITS, is present, it will provide a convenient reference as the maximum peak to peak video signal.

FREQUENCY MEASUREMENTS

If the signal is being carried on channel and you have checked to make sure that the same oscillator is being used for both down and up conversion, then no frequency measurement is needed to determine compliance since the frequency is controlled soley by the originating broadcast station. Similarly, measurement of the 4.5 MHz. intercarrier sound is not necessary unless it is demodulated to baseband audio by the CATV system.

Aside from using a signal processor capable of stripping the modulation from a TV signal and providing a direct counter output, the best way to measure frequency on the system is to set a signal generator to the same frequency as the signal to be measured and then count the signal generator. Most any signal generator or sweep generator in the CW mode is stable enough for the short time needed to make the measurement. The system and the signal generator may be mixed through a splitter and the combination then fed into a set top converter. A spectrum analyzer connected to the output of the converter is then tuned to the output channel of the converter in order to detect the beat between the signal to be measured and the signal generator.

It is important to adjust the signal generator level using an external variable pad to the same amplitude as the carrier to be measured for a good zero beat. Record the counter reading of the generator frequency when a zero beat is obtained on the analyzer. By using a converter ahead of the analyzer, only the signal generator and the converter will have to be adjusted as you move from channel to channel. The analyzer may be left tuned to the converter output. A television set or a FSM could be used to detect the zero beat, but it is more difficult to obtain the correct results using them.

If it is desired or necessary to count the 4.5 MHz. intercarrier frequency, the counter may be connected to the 4.5 MHz. output of a demodulator. By setting the counter for a ten (10) second counting interval, it will average the FM deviation over this interval and provide an accurate count of the 4.5 MHz. aural intercarrier.

In systems using converters, it would be wise to determine that all visual carriers at the input of the converter can meet the ± 25 KHz. standard before making measurements at the output to determine if the ± 250 KHz. standard can be met even though the FCC may only require the latter.

AMPLITUDE MEASUREMENTS

Measurement of the visual and aural carriers is a straight forward process we are all familiar with. A FSM or a spectrum analyzer may be used providing their calibration has been checked. On a spectrum analyzer what is often called "vertical interval roll through" will occur. The picture carrier will increase in level when the vertical interval is displayed at the top of the picture carrier. This occurs because the television station vertical sync rate and the analyzer line sync rate are not equal or phase locked together. It is this maximum amplitude which should be measured and recorded when using the analyzer.

While the analyzer is being used for other tests, a man with a FSM can make the rounds of the test points for the twenty four (24) hour level variation checks. A check of all channels at each test point every three hours should be adequate. Pilot carrier levels and the temperature should also be recorded for your own use in analyzing the data. If you are involved in testing several systems, or wish to do these long term variation checks on a regular basis, an automatic all channel level recorder should be considered. These could be placed at each test point to record the levels over a 24 hour or longer period. One could probably be put together with a detector and chart recorder on the output of a converter which was cycled through all channels every hour and then the entire unit shut off until the next hourly reading was desired. Mounted in a weatherproof box which could be chained to a pole or pedestal, a battery operated unit could be put anywhere in the system as a recording level monitor. Such a unit might not be equipped to give absolute levels, but the FCC requirement over a 24 hour period only involves relative level changes.

Before using such a device to prove compliance with the standards, make sure it is relatively insensitive to humidity, temperature and mechanical changes.

HUM AND LOW FREQUENCY DISTURBANCES

A waveform monitor connected to the video output of a demodulator can be used for single channel tests of hum and low frequency disturbances. Measuring the hum on a pilot carrier may prove the distribution system, but it does not measure the effects of headend equipment. Inadequate low frequency response may be seen as tilting or slope in the synce tips or on the vertical blanking interval. With the video level adjusted for a peak to peak display of 140 IRE units on the waveform monitor and assuming 87.5% modulation, eight (8) IRE units of hum or low frequency disturbances correspond to five (5) percent peak to peak variation in visual signal level. This may be expressed as:

% P-P signal variation =

$$\frac{\text{P-P video variation IRE}}{\text{140 IRE}} \times 87.5$$

A poor signal from a broadcast station may cause an out of limits condition using this measurement technique, but if that is suspected, it can easily be checked by making the measurement directly off the air and by using more conventional methods to measure the distribution system only hum.

The five (5) percent peak to peak variation is actually 2.5% AM or hum modulation (-32 dB) as most commonly used in the CATV industry.

FREQUENCY RESPONSE

This is one test which must be performed on at out of service basis. Done during the early morning hours, it will cause little interruption to the customer's service.

The results of the headend and system frequency response tests may be added together for a worst case result. The headend poses no small problem with its combination of demodulators, microwave, modulators and processors. If preamplifiers are in use, they too should be tested.

Because headend processing equipment often has frequency response characteristics, similar to a television transmitter, testing from -1 MHz. to +4 MHz. with reference to the visual carrier will not allow the ± 2 dB specification to be met in many instances. In the headend, tests from -.5 MHz. to +4.2 MHz. are more realistic or if you test from -1 to +4, a note indicating that most of the response variation lies below the visual carrier frequency, if that is the case, should be included in the test data.

Preamps or other tower mounted devices can be tested by disconnecting the antenna lead and hooking up a test cable which runs down into the building. Even a long, lossy cable will have very little slope in its response over a 5 MHz. bandwidth. By putting a slow sweep on the input line and a spectrum analyzer with variable persistence on the line which would normally connect to the headend processor, a picture of the frequency response can be "painted" on the analyzer. The analyzer should be set for 1 MHz./division scan width and 30 KHz. resolution. Vertical sensitivity should be set to 1 or 2 dB per division or linear for all frequency response checks. If used in the linear mode, it will be necessary to determine how much vertical deflection on the analyzer a 1dB input change causes. The sweep should be inserted through a 10 dB pad at the normal input level for the preamp. The 10 dB pad will improve the flatness of the sweep.

A demodulator should be put in the manual gain position and set for the same operating level as when it is in AGC. Insert a sweep signal with 1 MHz. markers and use a video detector and a scope to display the frequency response at the output.

When FM microwave is in use, a video multiburst test signal should be used. Clamper amplifiers and other devices which depend upon sync references make sweep testing difficult. Vertical interval test signals will give you a clue as to your system's performance, but these signals have traveled through a lot of processing equipment which can affect frequency response before they reach the CATV system antenna. If you operate your own CARS microwave,testing will be no problem, but if you are served by a common carrier, it may be wise to ask them for annual certification of frequency response in their headend equipment and microwave. Just because you do not have direct control over this will not relieve you of your responsibility to deliver high quality service to your customers.

Modulators may be tested by connecting a slow or manual video sweep to the video input with the spectrum analyzer at the headend output test point. Again, a picture of the frequency response will be displayed on the analyzer.

Processors can be tested by putting them in the manual mode and then adjusting the gain for the same operating level as when in AGC. Slowly sweep the input with the spectrum analyzer connected to the headend output test point. The sweep input level should be the same as the normal signal input. Disable or turn down the sound IF path during this measurement or the frequency response may appear to be excessive as the sweep approaches 4.5 MHz. above the visual carrier frequency. The limiter stages in the sound path will run wide open without an input.

The distribution system frequency response may be tested by inserting a slow or manual sweep at the system sweep input point and using the spectrum analyzer at the system test points to display the frequency response. The sweep level should be set about 4 dB below the visual carriers. This will allow these carriers to be used as frequency markers on the analyzer. It may be necessary to install notch traps at the pilot carrier frequencies on the output of the sweep to avoid AGC amplifier gain or slope changes. If more than one trap is used, a 6 dB pad should be used between traps.

No more than five or six channels should be displayed on the spectrum analyzer screen at one time. In a converter equipped system, only the output channel need be displayed, as the system channel frequency response tests are done one at a time. If a converter is hooked up to the headend output test point and every channel's frequency response is tested through it, then system frequency response checks could be conducted at a point which would normally be the converter input. This can save considerable time and still be representative of overall system performance when the headend and system frequency response measurements are summed.

CARRIER TO NOISE AND COCHANNEL

Use of a video noise test set, such as the Tektronix 1430, on the output of a demodulator is a fast and simple way to make in-service checks of the overall noise performance of the headend and the distribution system. Conventional methods require terminating the system input and then measuring the noise level. Substitution of a termination for the antenna at the input of the first active device is not always easy with tower mounted preamps and must be done at night to avoid interruption of service. Video signal to noise measurements are particularly relevant to systems using microwave where the signal is present as baseband video. Local origination studios and video sources for other cablecasting channels can also be tested by a video noise test set.

On Class I channels, this technique also measures the noise contributed by the broadcaster, but the noise power of the broadcast signal with typical S/N readings of 50-55 dB is negligible when added to the noise power of the cable system where S/N readings of 37-42 dB are typical at the system extremities. If the broadcast signal to noise measured with the demod and noise set is 10 dB or greater than the CATV system signal to noise, the contribution by the broadcaster to the overall measurement made at the system extremities is less than .5 dB. The demodulator with an input of 0 to 10 dBmV will also have a negligible effect on the S/N reading. The FCC requirement and the measurement most often made in the CATV industry is carrier to noise. Assuming 87.5% modulation of the visual carrier and a noise free video modulating signal S/N may be converted to a 4 MHz. bandwidth C/N reading by adding 4 dB to the S/N reading. This 4 dB correction factor should not be confused with the 4 dB factor used for correction of noise measurements made with a FSM. The S/N to C/N conversion factor comes from the amplitude relationships of modulating signal to carrier level and video level to composite video level. The formula relating them is:

$$C/N = S/N + 20 \log ((100\,IRE/140\,IRE) \times .875)$$

Where 100/140 represents the ratio of video to composite video (video and sync) and .875 represents the depth of modulation (87.5%). The actual factor which results from the above formula is about 4.08 dB but 4 dB is accurate enough for most purposes. The normal C/N ratio is not weighted, so do not confuse weighted S/N numbers with the unweighted number used above.

Offset cochannel can be measured using a spectrum analyzer set for 5 KHz./division scan width and 3 KHz. or 300 Hz. resolution. The horizontal sync sidebands will appear at 15,750 KHz. above and below the picture carrier and if offset cochannel is present, it will appear at 10 or 20 KHz. above or below the picture carrier. Several interfering cochannel signals may be present simultaneously, but suspect your measurement if any two of them are more than 20 KHz. apart since broadcast stations are normally only assigned ± 10 KHz. offsets.

INTERMODULATION

The spectrum analyzer is one of the best methods available to detect low level beats in a cable system. Set up the analyzer for 200 or 500 KHz./division scan width and 300 or 500 KHz. resolution using variable persistence. Slowly scan the channel of interest, stopping several times to look for beats. Do not confuse the color subcarrier or regularly spaced sync sidebands with real interfering beats. Probably the most distinguishing feature of a beat is that it normally does not change in frequency or amplitude. If the CRT brightness and persistence are kept low enough, a beat "down in the noise" can be measured since it will appear as a hole or bump in the noise. Even a high quality analyzer may not detect beats which are close to the picture carrier and masked by the modulation. If, after viewing a television set, you believe such a beat is present use your RF spectrum analyzer in the time domain as a fix tuned receiver. The vertical output of the RF analyzer may then be used to feed a wave analyzer or a low frequency spectrum analyzer either of which can detect a low frequency intermodulation product. If a low frequency analyzer is not available, try removing the channel from the system and then examine, with the RF spectrum analyzer, the area around the visual carrier frequency for intermodulation products. Removing the visual carrier may also remove some of the beats, but if interference can be seen on a television it is likely that non linearity in the system will have created several intermodulation products. Take care not to drive your analyzer so hard that intermod products are created internally. If adding a three (3) dB pad to the input causes the suspected beat to drop more than three (3) dB then the beat is being created, at least in part, internally in the analyzer. Reduce the input level until this effect is no longer present.

TERMINAL ISOLATION

Display the frequency range to be tested on the analyzer. Single measurements from 54-300 MHz. will produce accurate results. Adjust the sweep generator for an output level of about twenty (20)dB above the visual carrier levels measured at the subscriber terminal under test. Connect the analyzer directly to the slowly sweeping generator and adjust the analyzer for a reference level near the top of the screen, making sure the sweep output is flat. Connect the generator to one test point and the analyzer to the other test point using the shortest drop cables possible in order to simulate a worst case condition. Now measure the sweep level on the analyzer and the difference between the measured and reference values at each channel is the terminal isolation. The visual carriers seen on the analyzer serve as convenient markers.

RADIATION MEASUREMENTS

Making radiation checks is one measurement which has no tie with the subscriber terminal. In fact, worst case radiation is more likely to occur at high level points in the system, such as amplifier locations, rather than at subscriber drops.

Three measurements of radiation would appear to satisfy the FCC annual test requirements, but to insure that the system is really in compliance, on a continual basis, will require more extensive testing.

In making the annual measurements a homemade dipole or a commercial calibrated dipole antenna should be used in conjuction with a preamp and a spectrum analyzer. The antenna should be positioned the required distance from the system components and at least ten feet above the ground and away from any metal objects. A slightly modified studio tripod used with a fiberglass extension pole or "layup stick" can be used as a support for the antenna.

If the analyzer does not have its own preamplifier, a line extender can be used. line extender or other CATV amplifier is convenient since its gain and slope controls may be adjusted to compensate for the loss in cables interconnecting the antenna, preamplifier, and spectrum analyzer.

Since radiation is generally a broadband phenomena in a cable system, testing at one or two frequencies will suffice. Select a channel which is not present as an off the air signal in the area. Connect the analyzer to the system, tune in the cable channel to be measured, reduce the scan width to 5 KHz./division and the resolution to 3 KHz. in order to minimize the internal noise displayed on the analyzer. Then, being careful not to change the center frequency, connect the analyzer to the test dipole and increase the sensitivity to maximum. Rotate the horizontal antenna about its vertical axis for maximum signal.

Measure the absolute amplitude in dBmV and use the antenna correction factor to convert to uV/M. Use of a bandpass filter ahead of the preamp is almost a necessity in urban areas where the multitude of off air signals will overload the preamplifier and the analyzer.

Because of the many possible sources of radiation in a cable system and the complex interaction of their near and far fields, it is not recommended that measurements be made at points other than the ten and hundred foot distances from the cable the FCC radiation limits are specified at. Attempts to simply relate field strength to the inverse of the distance will not always give "real life" results.

The NCTA has recently issued a bulletin on radiation measurements describing a technique devised by Ken Simons of Jerrold Electronics, using a sensitive television receiver and a dipole antenna calibrated and mounted on a vehicle to make system radiation checks at many locations by driving around the system and observing the TV set for any signal leakage from the cable. Not only will this method also meet the annual performance test requirements, but it can be used in a routine maintenance program.

CONCLUSIONS

The methods described in this paper are the results of considerable field experience and although they will yield accurate and repeatable data, so will other methods, some more complex and some less. Which methods to use is a matter which the individual system operator must determine based upon test equipment available to him and his own particular technical experience.

Some of the test techniques used in this paper are described in more detail or from a different viewpoint in the Tektronix "No Loose Ends" proof of performance program by Cliff Schrock.

ACKNOWLEDGEMENTS

I would like to thank Mr. Ron Peterson for his help in drafting the log forms, Ms. Carolyn Ningen and Ms. Debbie Cosmos for preparing the manuscript, and the many LVO Cable personnel who helped perfect the test techniques.

APPENDIX A

SUMMARY OF TECHNICAL STANDARDS

(FCC Volume III, 76.605)

Required completion dates for the first performance tests for various standards are listed below. These incremented test dates apply only to systems in operation before March 31, 1974. Systems which began operation on or after that date are required to complete their first performance tests for all standards by March 31, 1974. Testing to determine compliance with the offset co-channel standard, Section 76.605 (a) (9), and the intermodulation standard, Section 76.605 (a) (10), has been suspended and is not required for either "old" or "new" systems at this time.

Pre March 31, 1972 systems are not required to comply with most FCC standards until March 31, 1977. They are, however, required to meet the radiation standards, Section 76.605 (a) (12), since compliance has been required of all systems since March 31, 1972. Compliance with the standards for systems commencing operation after March 31, 1972 was effective March 31, 1972. Although testing for offset co-channel and intermodulation is not required, compliance with these standards is required of post March 31, 1972 systems. It may also be assumed pending further action by the FCC, that pre March 31, 1972 systems will be required to comply with these standards by March 31, 1977.

TEST	STANDARD	FIRST REQUIRED DATE OF TESTING *
1. Frequency of the visual carrier:	1.25 MHz. ± 25 KHz. above channel boundary	3-31-76
a. At output of converter:	1.25 MHz. ± 250 KHz.	3-31-76
2. Frequency of aural subcarrier:	4.5 MHz. ± 1 KHz.	3-31-76
3. Minimum visual signal level:	1 mV across 75 ohm (0 dBmV)	3-31-74

TEST	STANDARD	FIRST REQUIRED DATE OF TESTING *
4. Permissible signal level variation:	12 dB total/24 hr.period	3-31-74
a. Maximum adjacent channel variation	3 dB	3-31-74
b. Maximum signal level	12 dB	3-31-74
5. Maximum signal level:	Below threshold of degradation (overload point)	3-31-74
6. Maximum hum and low frequency disturbance level:	5%	3-31-75
7. Within channel frequency response:	± 2dB from -1 MHz. to +4 MHz.	3-31-75
8. Aural signal level:	13 to 17 dB below visual	3-31-74
9. Carrier to noise level for all signals picked up or delivered within its Grade B contour:	36 dB S/N ratio 36 dB co-channel	3-31-74 None
10. Signal to intermodulation and non-offset carrier interference:	46 dB	None
11. Subscriber terminal isolation:	18 dB	3-31-75
12. Radiation: Up to 54 MHz:	less than 15 uV/m @ 100'	3-31-74
54 to 216 MHz.:	less than 20 uV/m @ 10'	3-31-74
Above 216 MHz.:	less than 15 uV/m @ 100'	3-31-74

* Incremental testing applies only to systems in operation prior to March 31, 1974

CATV

Proof-of-Performance

LVO CABLE, INC.
P.O. BOX 3423
TULSA, OKLAHOMA
74101

EXPANDING ENTERTAINMENT AND COMMUNICATIONS BY CABLE • AN LVO CORPORATION COMPANY
LVO CABLE INC.

SYSTEM INFORMATION

August 1973 MM
LVOC 19-73

LOCATION__________ DATE__________ SYSTEM__________

MINIMUM VISUAL SIGNAL LEVEL (CLASS I CHANNELS) ______ dBmV

CABLE CH	CLASS	LOCAL, OFF-AIR, OR MW	CALL LETTERS	STATION LOCATION	CH	GRADE	COMMENTS
2							
3							
4							
5							
6							
7							
8							
9							
10							
11							
12							
13							

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PROOF-OF-PERFORMANCE DATA

LOCATION__________ DATE__________ SYSTEM__________

SENIOR ENGINEER'S STATEMENT

ENGINEER__________ TITLE__________ COMPANY__________

ADDRESS__________ SOCIAL SECURITY NO.__________

FCC LICENSE CLASS__________ LICENSE NO.__________ EXP. DATE__________

EXPERIENCE__________

EDUCATION__________

TEST EQUIPMENT

MODEL	MANUFACTURER	TYPE	SERIAL #	CALIBRATION-DATE	BY

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SYSTEM TEST EQUIPMENT DATA

LOCATION__________ DATE__________ SYSTEM__________

TEST EQUIPMENT

MODEL	MANUFACTURER	TYPE	SERIAL #	CALIBRATION-DATE	BY

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TEST POINT INFORMATION

LOCATION__________ DATE__________ SYSTEM__________

TP #_____ LOCATION DESCRIPTION__________
__________ POLE / PEDESTAL NO.__________
TRUNKLINE AMPS CASCADED TO THIS POINT_____ TYPE OF TRUNKLINE AMPS__________
LINE EXTENDERS CASCADED TO THIS POINT_____ TYPE OF LINE EXTENDERS__________
TYPE OF TAP__________ TAP SIZE __________ LENGTH OF DROP CABLE__________
TYPE OF DROP CABLE__________

TP #_____ LOCATION DESCRIPTION__________
__________ POLE / PEDESTAL NO.__________
TRUNKLINE AMPS CASCADED TO THIS POINT_____ TYPE OF TRUNKLINE AMPS__________
LINE EXTENDERS CASCADED TO THIS POINT_____ TYPE OF LINE EXTENDERS__________
TYPE OF TAP__________ TAP SIZE __________ LENGTH OF DROP CABLE__________
TYPE OF DROP CABLE__________

TP #_____ LOCATION DESCRIPTION__________
__________ POLE / PEDESTAL NO.__________
TRUNKLINE AMPS CASCADED TO THIS POINT_____ TYPE OF TRUNKLINE AMPS__________
LINE EXTENDERS CASCADED TO THIS POINT_____ TYPE OF LINE EXTENDERS__________
TYPE OF TAP__________ TAP SIZE __________ LENGTH OF DROP CABLE__________
TYPE OF DROP CABLE__________

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HEADEND TESTS

LOCATION__________ DATE__________ ENGINEER__________ SYSTEM__________
* AURAL-VISUAL CARRIER SEPARATION UNAFFECTED BY CATV SYSTEM.

CH	STANDARD FREQUENCIES		OFFSET kHz	VISUAL MHz	AURAL MHz	AURAL-VISUAL MHz	RESPONSE ± dB	REMARKS
	VISUAL — MHz	AURAL — MHz						
2	55.25	59.75						
3	61.25	65.75						
4	67.25	71.75						
5	77.25	81.75						
6	83.25	87.75						
7	175.25	179.25						
8	181.25	185.75						
9	187.25	191.75						
10	193.25	197.75						
11	199.25	203.75						
12	205.25	209.75						
13	211.25	215.75						

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SYSTEM TESTS

TEST POINT #____________ DATE____________ ENGINEER__________________ SYSTEM____________

CH	VISUAL dBmV	AURAL dBmV	P-S dB	INTERMOD dB	CROSSMOD dB	CO-CHANNEL dB	CARRIER/NOISE dB	RESPONSE ± dB	HUM %
2									
3									
4									
5									
6									
7									
8									
9									
10									
11									
12									
13									

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SYSTEM TESTS

TEST POINT #____________ DATE____________ ENGINEER__________________ SYSTEM____________

CH	RADIATION		TERMINAL ISOLATION dB	REMARKS
	μV/M	DISTANCE-FT.		
2				
3				
4				
5				
6				
7				
8				
9				
10				
11				
12				
13				

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24 HOUR LEVEL VARIATION TESTS

TEST POINT #__________ DATE__________ ENGINEER__________ SYSTEM__________

CH	TIME												MAXIMUM VARIATION dB
2													
3													
4													
5													
6													
7													
8													
9													
10													
11													
12													
13													

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CATV SYSTEM GROUNDING TECHNIQUES

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Grounding and bonding are important to CATV operators for the protection of personnel and plant from accidental power line contact and lightning strikes. The increasing complexity of CATV systems and a growing use of buried plant, with its special susceptibility to lightning damage, should stimulate operators to analyze grounding and bonding needs beyond those historically established in joint-use and utilities codes requirements. The important aspects of these special grounding techniques are described in this paper.

Standards for bonding and grounding for CATV have generally been obtained from the National Electrical Safety Code, which was the result of work between the power companies and the communications companies. Since CATV is a communications industry, many of the requirements of the code can be directly applied to CATV, and have been in past installations. In the future, when CATV pole plant use grows to telephone plant levels in urban areas, the NESC will undoubtedly be modified to include more specific joint use requirements for CATV. In the meantime, it would appear desirable for the CATV industry to give some special attention, in the design of new systems, to the grounding problems which they will encounter in both joint use and on their own supporting structures.

It is the purpose of this paper to point out several of the items to be considered and the actions to be taken not only to conform to the requirements of the NESC but to properly protect CATV employees, subscribers, and the CATV plant where joint use may not be involved.

Coaxial cables for cable television use, consist of a center conductor, an insulating medium, and a cylindrical outer conductor on the same axis as the center conductor. The outer conductor of today's cables is usually made of aluminum and is frequently the exterior of the cable, unless the cable is jacketed with a sheath of polyethylene or some other plastic, or reinforced with an outer sheath and a steel tape or tapes for strength. The electrical impedance of the coaxial conductor is determined by the diameter of the center conductor, the spacing between this conductor and the outer conductor, and the thickness and diameter of the outer conductor. This is usually 75 ohms unbalanced to ground.

The ground serve not only as a base from which potentials are measured, but also provides a protective function in the event the coaxial cable is contacted by outside potentials.

As it is virtually impossible for CATV systems to avoid exposure of facilities to lightning strikes and occasional accidental power contacts, CATV system design engineers should be as meticulous in the drafting of grounding and bonding standards as they are in establishing system operational standards.

Cooperative and coordinated efforts with other users of distribution networks, such as the telephone and power companies, will provide protection against accidental contact with power circuits if all users adhere to accepted clearance requirements and other construction practices. We should, however, by a careful analysis of the type of network to be built, its location and the nature, and the type of joint users, set grounding and bonding standards which, in many instances, could be more stringent than the minimums placed on us as part of a pole attachment agreement or by a regulatory body order.

Before attempting to derive system bonding and grounding practices for a specific CATV system, one must first be aware of the nature of construction (aerial or buried) and our "right of way" neighbors' installations.

Looking first at the aerial possibilities, we see that we might find shared pole space with both power and telephone, with power alone, with telephone alone, or not shared at all if it is our own pole. For the underground, we might be using direct buried techniques or leasing existing conduit space on a shared basis with the telephone company. All of these possibilities present different grounding and bonding requirements, and should be so treated in drafting system construction practices.

Since television cables are communications conductors, and, if we assume the same standards apply to them as apply to telephone and telegraph cables, the configuration and many of the structural details, such as clearances and strength, have already been standardized in various codes and practices. Those are a result of long years of experience and study.

The grounding of television cables in each of the situations mentioned above will be discussed on the basis of the necessity to protect, as far as possible, against foreign potentials such as fallen power wire contacts or lightning resulting in injured personnel or damaged plant.

Case 1 - Where power, telephone and CATV share space on a pole the requirements normally found in our pole attachment agreements will be sufficient for system protection. The telephone company has already established an aerial communications plant that is effectively grounded and bonded to the multi-grounded network of the power company, in accordance with the requests of the NESC. By following the bonding and grounding requirements set forth in the attachment of renewal, the CATV system also achieves a satisfactory ground. However, when jacketed cable is used for the CATV system it should be noted that a low impedance bond or tie between the strand and the coaxial outer conductor must be maintained to avoid a "system contained" difference of potential as bonds to power networks are usually connected to the strand. With unjacketed cable, the lashing wire provides a continuous bond eliminating the need for further connections.

Case 2 - Where CATV and power share space without other communications lines being present, the CATV company, even if not required by the pole attachment agreement, should attempt to follow the pertinent sections for joint use of power and communications services set forth in the U. S. Department of Commerce, National Bureau of Standards Handbook 81, Part 2, "Safety Rules for the Installation and Maintenance of Electric Supply and Communications Lines" and the "National Electrical Code" published by the National Fire Protection Association.

Special safety considerations or cautions that should be exercised in this case, particulary where relatively inexperienced people are involved:

1. Always look for open or broken vertical ground wires before climbing the pole or commencing work. Any questionables wires should be checked with a voltage tester and if an excessive potential is noted, the power company should be notified.

2. All connections required to be made to the power company's network should be made by the power company, not by the CATV company.

3. Where suitable vertical ground wires exist, the CATV company may complete the connection between the strand and the ground wire but with the first connection being made to the strand. The use of rubber gloves is also recommended whenever work is being done on a joint use pole.

4. Whenever found, all street light fixtures mounted below or if in close proximity above the CATV attachment should be bonded by the power company to a vertical ground wire to which CATV should also bond.

5. The CATV strand should be bonded to a multi-grounded neutral supply system vertical ground wire at intervals no in excess of a ¼ strand mile.

Case 3 - Where telephone and CATV

are to share pole space without electrical supply lines, the CATV company should follow as closely as possible the practices set forth by the telephone company, particularly bearing in mind the following:

1. Where strand is not continuous (i.e. a dead end) a strand to strand bond should be installed and the bond connected to a vertical ground wire.

2. The strands should be bonded at all cross-overs.

3. The strands should be bonded and connected to the vertical ground wire at all power supply locations.

4. The strands should be bonded at intervals not exceeding 10 spans.

5. All guys should be bonded to the strands where the guy connection to the pole is made on a separate bolt.

Case 4 - Where a CATV network is installed on system-owned poles not shared with other users, standard telephone procedures for bonding and grounding non-joint use systems should be followed. Several of the more important of these practices are:

1. Strand supporting aerial plant should be effectively grounded at least once in every strand mile of system and at all power supply locations.

2. Where joint use crossing poles occur with electrical supply lines and communications lines, bonds should be made between the strands and the supply line neutral, which in turn should be connected to a vertical ground wire.

3. Where joint use crossing poles occur with only communications lines, bonds should be made between strands.

4. Grounds should be installed as close as possible to all strand dead end locations.

5. Strand continuity should be maintained throughout the CATV system.

6. At head end locations, the strand should be bonded both to the equipment rack and the tower ground.

One final caution for all aerial construction concerns the safety of employees while attempting to restore service after a major disaster. When plant damage is encountered, such as broken or burned joint plant caused by high winds, ice storms, collisions, or other disasters, no attempt should be made to restore the CATV cables until a survey of power contacts has been made, the contacts and reduced clearances referred formally to the power company, the power company has cleared them, and an inspection has been made to be sure they have been adequately cleared in at least a semi-permanent manner providing adequate strength and clearance. Only then should CATV personnel start to clear their plant.

While most of the television cables have been installed aerially, the increasing restrictions on aerial utility plant in most cities will probably mean that future cable television distribution systems will be underground rather than aerial, and it is highly probable that significant percentage of existing aerial plant will be replaced by underground plant in the future.

Underground cables can be placed either in conduit, or directly buried in the ground. Unless existing conduit runs are available, and can be leased from established carriers at reasonable rates, it will be found that the direct burial of television cables in most cases will be more economical. In either case, in conduit or direct burial, the need to prevent electrolysis indicates that the underground cables should have an insulating plastic sheath. As this jacket has a tendency to insulate the cable from ground special care should be taken to achieve the lowest possible impedance to ground throughout the system. Sheath continuity should be maintained to permit the distribution of foreign potentials over as much of the CATV system as possible.

Trunk amplifiers, line extenders, and other equipment having metallic cases which are connected to the grounded outer conductor should be grounded to the same extent as the television cable, whether they are buried directly or brought above ground in a pedestal for maintenance purposes. Special care should be taken at pedestal locations by attempting to limit any possible potential difference between the CATV system and other facilities that could be contacted simultaneously by employees, subscribers, or the public. Normal practice would be for the CATV system to bond to all other facilities or metallic structures at least within ten feet of the pedestal. Junctions of aerial and underground plant require additional engineering considerations but as there are several variables that could effect the bonding and grounding techniques, it would be too lengthy to go into here. Whenever a junction of this type occurs in a CATV plant it would be advisable to either research the requirements in the NESC or to discuss the situation with a knowledgeable plant engineer.

Lightning can have serious damaging effects on buried cable. While it would appear that aerial cable might be more severely damaged by lightning than buried cable, this is generally not the case. Substantial damage may occur under certain circumstances. When lightning strikes the ground, very high potential is created at the point where it strikes. This high potential tends to equalize itself over an area the size of which is determined by the resistance of the earth (earth resistivity). In areas of high earth resistivity, such as is found on the Piedmont plateau on which Atlanta, Georgia is located, the tendency of the lightning would be to find a conductor of some sort from which it could spread in rather long distances to equalize the charge created at its earth strike point. Good bonding and grounding techniques allow dissipation of such strikes with little or no damage to the CATV cable.

In areas of high earth resistivity and high frequencies of thunderstorms, currents carried by severe lightning flashes running as high as 200,000 amperes, will be common.

Taking all these factors into consideration, it is obvious that buried cable should be jacketed for protection against electrolysis and well grounded for protection against lightning strikes, particularly on long runs, and most particularly in areas of high resistivity and frequency of storms. Also, in rural areas where the earth resistivity is high or on long dead runs between communities it would seem desirable to consider the possible installation of lightning protective wires, which can be buried above the television cable and one or two feet apart. These wires need not be very large, in fact they could be #12 wires. The two wires act to provide a conducting plane which further protects the cable from lightning. Attempts should also be made in these areas to achieve uniformity of grounds throughout the system.

No discussion of bonding and grounding requirements would be complete without some mention of subscriber drops. In cases where distribution cable and drop cable outer conductors are electrically continuous the network ground is usually sufficient, although we are all aware that braided RG-59U cable, particularly over a period of time, offers little possibility of electrical continuity. Thus all subscriber drops should be grounded as close as possible to the home entrance by means of at least a #12 wire to a cold water pipe.

In summary, as we build large and more sophisticated CATV systems we must become aware that bonding and grounding requirements are at least as important as noise, hum and cross modulation specifications as the lack of necessary bonds and grounds could lead to the physical destruction of sections of our system, damage to subscriber's property or even to serious injury or death of an employee or a subscriber. System engineers, lets make a grounding test part of our proof of performance!!

CONTROLLING POWER SYSTEM SURGES IN CABLE TELEVISION SYSTEMS

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This paper discusses the causes of commercial power system surges and presents recorded data indicating the nature, magnitude and frequency of occurrence of surges to be anticipated on 120 volt distribution circuits. The effects these surges can have on a cable system's reliability are discussed, and the various devices available for surge protection are reviewed. Particular attention is paid to a surge problem, and the corrective measures taken, in General Electric's Decatur, Illinois cable system. In Decatur, surges in excess of 1200 volts were indicated on residential, 120 volt circuits and repetition rates up to 480 surges in a 24-hour period were recorded at the output of one of the cable system's 60 volt power supplies. Finally, a modification for cable television use of the electrical protection section of REA specification PE-60 is offered. This modification provides voltage and current surge acceptance tests for cable television amplifiers and power supplies.

Three of the principal causes of surges in 120 volt power circuits are inductive load switching, lightning induced transients and direct lightning strikes to the power system. Probably every cable system has experienced both short term, self-correcting service interruptions and long term service outages caused by lightning. However, as the services carried on cable systems, particularly bidirectional cable systems, grow in complexity, the need for higher levels of system reliability and immunity to voltage transients and surges also grows. For example, one to two hour outages cannot be tolerated in a cable system carrying a burglar alarm and fire alarm service.

First, just what is the nature of the surges we encounter or can expect to encounter? What is their peak voltage, wave shape, frequency of occurrence and duration? We will omit any consideration of direct lightning strikes to the cable system since these result in large mechanical failures despite the best grounding and bonding measures.

In a general study[1] on 120 volt secondary power surges that took place over two years and at 400 locations in 20 cities, surges from 600 to 5600 volts were measured. Their wave shape was either unidirectional, with durations to 15μS, or oscillatory with one to four cycles of frequencies from .2 to 1 MHz. The wave shape varied considerably from city to city and even location to location within a given city and was dependent upon the impedence of the power system at that particular point.

Although many of these surges were lightning induced, even surges within homes caused by water pumps, fluorescent light switching and oil burner starting were from 500 to 2000 volts! Although many surges greater than 4000 volts were recorded, if it is possible to draw a "typical" lightning induced surge, it would look something like Figure 1.

Surges are present almost all the time, at almost all locations and at surprisingly high voltages. One of the most convincing pieces of evidence for this is the fact that when General Electric increased the winding insulation in their 120 volt electric clocks from a 2000 volt withstand to a 6000 volt withstand[1], motor winding failures were reduced by a factor of 100:1!

These are the surges we must live with. How? There are three main classes of surge protectors or arresters: zener diodes, metal oxide varistors and spark gaps. Each of these devices has its advantages, disadvantages and area of application. The device currently being used in cable television amplifiers is a spark gap sealed in an inert gas atmosphere - commonly called a gas tube. The gas tube has two main disadvantages: a time delay in establishing the arc and a difficulty in extinguishing the arc. The first can allow fast rise time pulses of considerably higher voltage than the tube's DC firing voltage to pass. Because of the gas tube's second difficulty, the arc in a gas tube that does not have a series limiting resistor can be maintained with a much lower voltage than that required for arc initiation. The arc may even be maintained on the normal 60 VAC powering voltage; this can lead to destruction of the gas tube, leaving it fused in a shorted condition.

The first stage in surge protection occurs in equipment selection for the construction of new cable systems. It is hoped that the surge specification included as an appendix will help in this. This is a modification of the surge portion of REA specification PE-60, "Trunk Carrier Multiplex Equipment", that we have included in our general construction specifications. Figure 2 shows the type of surges this modified test specification applies to cable television equipment.

The most common surge problem, however, is in a cable system which already exists. This problem is particularly severe in General Electric's Decatur, Illinois system. Figure 3 shows the history of power related failures in Decatur. Although these failures include burned circuit boards, broken circuit breakers, circuit breakers requiring resetting and burned out power transistors, by far the predominate failure mode was the fusing of gas tube surge protectors. The turnkey vendor for our Decatur and Anderson systems, Anaconda Electronics, placed a Biomation transient recorder at the output of one of the Decatur AC power supplies and set it to record any time the output voltage exceeded 70 volts or dropped below 50 volts. In the 30 day recording period, 750 such excursions were recorded - 480 of them in one 24 hour period! General Electric had previously measured surges on 120 volt services in Decatur in excess of 1200 volts[1].

Although we are experiencing excessive power failures in both our Anderson and Decatur systems, certain areas of our Decatur system seem to be the hardest hit. We have increased our grounding frequency in these areas and now have another transient recorder on loan from the IEEE to our Research and Development Center installed in one of these areas.

Anaconda has taken two actions in the hope of improving their system's vulnerability to surges: they have provided 230 volt surge protectors to replace the factory installed 145 volt units and have installed large capacitors (Figure 4) at the output of selected AC power supplies. Unfortunately, neither action has solved the problem or, it appears, even diminished it.

In an attempt to determine the surge attenuation of the AC supply, a sample was loaded to an output current of 6 amperes into a nonreactive load and surges were applied to both the input and output at our Research Center in Schenectady. A 2kV, 5μS surge and a 2.5kV, 70μS surge were applied across the input line and from an input line to ground. In no case was more than 30 volts of the surge left at the power supply's 60 volt output. When the 2.5kV, 70μS surge was applied across the 60 volt output of the operating, loaded supply, the gas tube contained in the supply fired, clipping the peak value of the surge to 900 volts and then recovered, restoring the 60 volt output. This testing would seem to indicate that the surges causing our problems are not entering the system through the AC supply.

At the writing of this paper, the problem is still not solved. In conjunction with the General Electric Research Center, we are now investigating induced voltages and ground currents as possible causes.

(1) Surge Voltages in Residential and Industrial Power Circuits, F. Martzloff and G. Hahn, IEEE Transactions on Power Apparatus and Systems, July/August 1970.

(2) Bibliography on Surge Voltages in AC Power Circuits Rated 600 Volts and Less, Ibid.

APPENDIX

2.5 Electrical Protection

2.5.1 General

Adequate electrical protection of distribution equipment shall be included in the design of the distribution system.

The characteristics and application of protection devices must be such that they enable the distribution equipment to withstand, without damage or excessive protector maintenance, the voltages and currents that are produced in the equipment as a result of induced or conducted lightning surges.

All power supplies will be equipped with circuit breakers and a surge protector, and all amplifier trunk ports will be equipped with easily replaceable gas discharge surge protectors. All trunk amplifier bridger ports will be equipped with self-healing overload protectors.

Compliance with Specification 2.5 will be demonstrated to a representative of General Electric Cablevision Corporation on two randomly selected samples of each of the following: a "fully loaded" trunk amplifier station, an AC power supply and a line extender amplifier.

In each demonstration, the device being tested shall pass the following tests: (1) capacitor discharge test, (2) current surge test and (3) protector response delay surge test. The capacitor discharge test and the protector response delay surge test are performed with all protectors removed; the current surge test is conducted with all protectors in place. In each case, the device will be tested while it is in normal operation.

2.5.2 Capacitor Discharge Test

The magnitude of the capacitor discharge surge shall be the maximum DC breakdown voltage of the protector used and shall have a rise rate of not less than 100 volts per microsecond and shall decay from peak voltage to one half the peak voltage in not less than 100 microseconds. The rise rate is defined as the peak voltage divided by the time required to reach the peak voltage.

Five surges shall be applied at one minute intervals to each port of the device under test. The polarity of the surge generator shall then be reversed and the procedure repeated.

AC power supplies shall have the power input terminals surge tested by surges applied between the power input terminals and between each power input terminal and ground. The number, repetition rate and polarity of the pulses shall be as above.

2.5.3 Current Surge Test

A 500 ampere peak current surge with a rise rate of not less than 100 amperes per microsecond and a peak to half value decay time of not less than 1000 microseconds shall be applied to all ports and terminals as described in 2.5.2. Three surges of each polarity shall be applied at one minute intervals.

2.5.4 Protector Response Delay Surge Test

Additional capacitor discharge tests shall be performed to determine the ability of the equipment to withstand voltage surges encountered because of the operating delay time of the protector. These surges shall be applied to the ports and terminals and in the number, repetition rate and polarity described in 2.5.2.

The magnitude of the surge is the surge striking voltage of the gas tube, and the rise shall not be less than 500 volts/microsecond. The surge striking voltage shall be determined from the protector manufacturer's data, using a rise rate of 500 volts per microsecond.

Power supply AC power line input terminals shall also be given protector response delay surge tests. The striking voltage of the protector shall be determined at a 10kV per microsecond rise rate and the rise rate of the surge shall not be less than 10kV per microsecond.

TYPICAL MEASURED SURGE

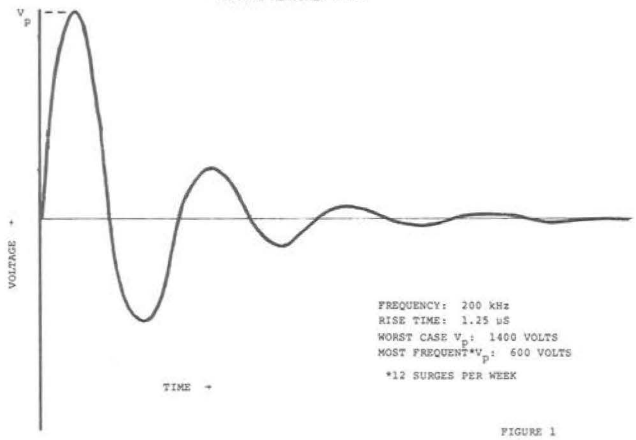


FIGURE 1

POWER RELATED FAILURES
DECATUR, ILLINOIS
(APPROXIMATELY 255 MILES)

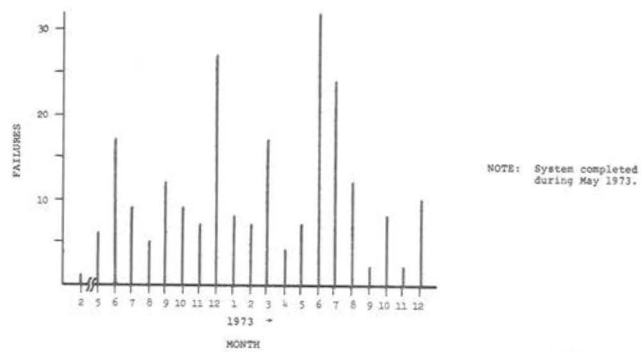
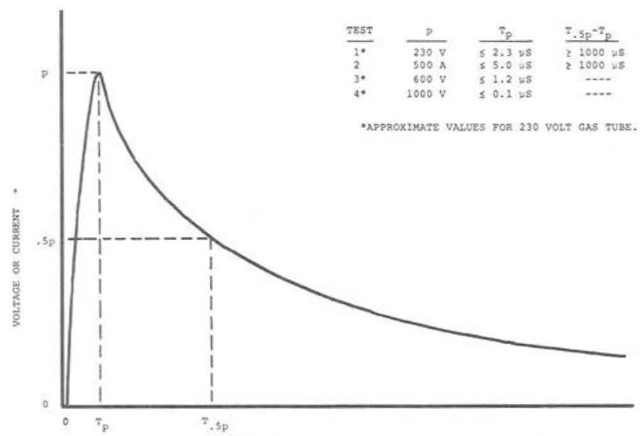


FIGURE 3

TEST SURGES



TEST	P	T_p	$T_{.5p}-T_p$
1*	230 V	≤ 2.3 μS	≥ 1000 μS
2	500 A	≤ 5.0 μS	≥ 1000 μS
3*	600 V	≤ 1.2 μS	----
4*	1000 V	≤ 0.1 μS	----

*APPROXIMATE VALUES FOR 230 VOLT GAS TUBE.

FIGURE 2

VOLTAGE SURGE SUPPRESSOR
FOR INSTALLATION IN
60 VAC POWER SUPPLY

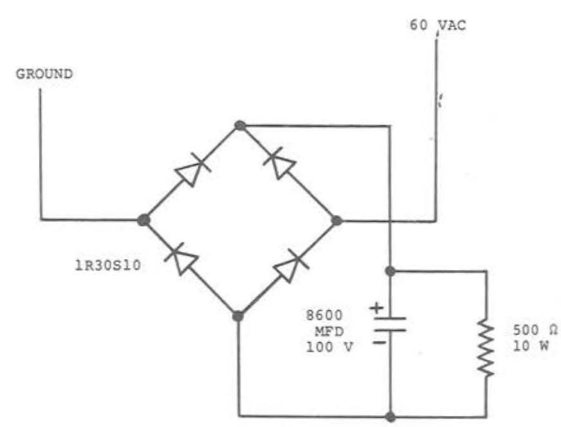


FIGURE 4

ESTIMATING CATV DEMAND PATTERNS FOR SYSTEM DEVELOPMENT

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This paper describes a methodology for estimating CATV subscriber densities suitable for use in the financial planning and engineering design of a CATV system in a city with good over-the-air broadcast TV reception.

The method provides estimates of demand as a function of time, and confidence limits on these estimates, for sections of the city formed in consideration of various characteristics of the city. The method is based on an advanced econometric model of CATV penetration developed by R.E. Park of the Rand Corporation and takes into account characteristics of local over-the-air broadcast stations, features of the proposed CATV system, demography of the community served, and growth stimulating factors.

The paper includes a brief review of the basic econometric model, a discussion of an adaptation made to it and an illustrative case study.

INTRODUCTION

Lenders to the CATV industry require cash flow projections for various stages of a system's development in order to determine the cash flow available for debt servicing, for structuring the amortization schedule of the loan and for evaluating the long term economic viability of the system. Since a CATV operator's prime source of revenue is monthly subscription fees, among the key factors in establishing positive cash flow are the system density and its penetration.[1]

It was with the investor's viewpoint in mind that this method of estimating CATV subscriber demand by sections of a proposed system as a function of time was developed. By being able to rank-order sections of the community on the basis of their estimated subscriber densities and associated construction costs it should be possible to develop an optimum engineering plan in terms of the expected number of subscribers, related costs, and revenues upon which to base a financial plan that demonstrates positive cash flows during the early years of operation and a long term economic viability.

[1]David M. Carlisle, "Financing Community Antenna TV," Thesis, the Stonier Graduate School of Banking, Rutgers University (June 1973).

DESCRIPTION OF METHODOLOGY

General Approach

Demand for CATV in a community depends on the existing TV environment, the characteristics of the proposed CATV system and the demography of the community served. For engineering and financial plans, highly disaggregate estimates of demand for CATV are desired. For example, if estimates by neighborhood or block were available, capital costs and revenues could be balanced by developing the most promising routes in the early years.

An econometric model expressing CATV penetration in terms of the TV environment, the proposed CATV system and the community served has been developed by R.E. Park of the Rand Corporation[2]. However, this model is based on CATV systems in their entirety. Furthermore, the data required by it is generally not available at the neighborhood or block level. Thus in developing an estimation method, we were faced with three major (and conflicting) considerations.

1) The engineering and financial plans should, ideally, be based on estimates of demand at a highly disaggregate level.

2) The most advanced econometric model of CATV penetration available in the literature was developed on the basis of entire CATV systems.

3) The level at which the data required by the model is readily available varies from city-wide to census tract.

In the suggested method we resolve these conflicting conditions by using Park's model (adapted to take into account special interest channels) to obtain

[2]R.E. Park, "Prospects for Cable in the 100 Largest Television Markets," The Bell Journal of Economics and Management Science 3, No. 1, 130-150 (Spring 1972).

an indication of penetration at the census tract level. Then, we combine contiguous tracts of similar subscriber densities into sections of the city which are large enough for valid application of Park's model while, at the same time, sufficiently disaggregated to be of value in developing engineering and financial plans. Estimates of the demand in each section are obtained at several stages of development by taking into account the dynamics of demand. By ranking the sections of the city according to their estimated subscriber densities, an ordering of areas for system development can be provided.

The Econometric Model

Park's model of CATV penetration is a cross-sectional model based on data from 63 cable systems located in communities whose television environment resembles that of the 100 top TV markets. That is, at least three television stations are received in the community and over-the-air reception is generally unimpaired (except possibly by distance from the transmitter). While a full description of the systems in the sample and their associated data are not available, it is reasonable to assume from the date of the study (1971) that the sample systems offered a minimum of local originated programming and no two-way services. Furthermore, the sample systems are of different ages; a maturity factor is included in the model to compensate for variations in penetrations due solely to age differences.

The estimating equation is linear in the logarithm of the exogenous variables which include income, price, color set ownership, and a set of variables which compare the service offered by the CATV system to that available over-the-air. A comparative service variable is defined for each type of channel (network, duplicate network, independent, educational, and foreign); it is expressed as the ratio of the number of CATV channels of a given type to the number of available TV stations of the same type, weighted according to the transmission mode (VHF, UHF) of the station, the percent of UHF receivers in the community and by a reception degradation factor defined in terms of the distance the community is from the station's transmitter.

In developing an econometric model*, a basic functional form is assumed along with a set of exogenous variables in an attempt to explain, in part, the observed behavior of an economic variable. Consistent observations on the dependent and independent variables are used to determine the values of the coefficients in the model which will account for a significant amount of the observed variation in the phenomenon. The model can then be used to investigate the logical consequences of the assumptions (and changes in them) or to estimate the value of the endogenous (dependent) variable in cases where it cannot be observed. The validity of the results obtained with the model is limited by the extent to which the assumptions on which it is based apply; the amount of variation unexplained by the model is reflected in the standard error associated with the estimate, or equivalently, in its confidence limits.

In using the suggested estimation method, care must be taken to stay within the domain of the basic model. It should not be applied to communities with significant TV reception problems, nor to systems whose appeal is based on innovative services (e.g., two-way services, premium TV) and no attempt should be made to obtain estimates for small homogeneous sections. In all applications the method should be validated by combining the estimates by section into an estimate for the system as a whole and comparing the result with the estimate obtained by applying the model to the entire system. If the two estimates differ by more than several percentage points the size and composition of the sections should be re-examined.

Treatment of Special Interest Channels

Certain channels offered on the proposed CATV system may be of primary interest to a special interest group within the community so that subscriber demand for the service may vary from section to section as the special interest population varies. Since we wish to rank the sections of the community according to their demand, this variation in demand due to a special interest channel was taken into account by a modification of the basic model which takes advantage of the continuity of the regression equation and is intuitively satisfactory. The modification consists of redefining the comparative service variables to include the value of a special interest channel in the functional form $p + k(1-p)$ where p is the proportion of the special interest group in the population and k is a factor representing the general interest of the remainder of the population. This allows the value of the channel to vary from 0 to 1 depending on the values of p and k. Foreign language channels and channels having a major portion of their programming with an ethnic orientation are handled in this fashion.

Data Sources

Most of the data required by this method is readily available. Demographic data at the census tract level can be obtained from the U.S. Bureau of Census. Areas by census tract, maps showing census tracts and land use, and projections of population can usually be obtained from local planning councils. The current TV Factbook provides the necessary information on the over-the-air stations received in a community.

Some substitutions and a certain amount of adjustment of the raw data are required. For instance, median household income, an exogenous variable in the basic model, is not provided at the census tract level; median family income, available by census tract, must be used as a proxy. Data on color set and UHF receiver ownership is generally available only on a citywide basis and must be assumed to apply uniformly throughout the community. Data for a census tract not wholly contained within

*E. Malinvaud, Statistical Methods of Econometrics, Rand McNally and Company, Chicago (1966) Chapter 2.

city limits must be adjusted to reflect the fraction of the tract within the city. Estimates of non-residential land use in the tracts usually have to be made from maps.

Preliminary penetration estimates are obtained for each census tract from the model. These, along with the demographic data of the tracts constitute the information required for forming sections of the city.

Forming Sections of the City

The sections of the community can be formed in consideration of various characteristics of the city, e.g., areas of poor reception, land use patterns, cabling requirements, municipal districts and socio-political boundaries. The number and composition of the sections are determined by the population and from the range of subscriber densities in the community. Sections are formed of contiguous census tracts and should contain a minimum of 7500 - 10,000 households in order to stay within the domain of the model. The constituent tracts of each section should have similar subscriber densities and the resulting sections should have significantly different subscriber densities so that a definite ranking of the sections can be made. Demand estimates for each of the sections are obtained by applying the econometric model to each section using data constructed from the data of its constituent tracts. For all but one of the variables the data for the section can be formed either as the aggregate or as the weighted average of the tract data. However, the median income of the section is not the average of the median incomes in the tracts. In order to estimate the median income for the section, the income distribution for each tract is required. If these are not available demand estimates for the section could be obtained by using the weighted average of the penetrations of the tracts.

Demand Estimates

Our method provides estimates of penetration, subscribing households and subscriber densities.

Penetration is calculated by the model as the ratio of subscribing households to households passed by the cable. We are using this parameter as a probability of subscription and as such it is most meaningful for many miles of cable, i.e., for relatively large areas.

The estimate of subscribing households is obtained as the product of penetration and households passed by the cable. In the absence of a construction schedule, estimates are based on the total number of households in the section which assumes that all homes are passed by the cable. If, however, information is available on the projected number of miles of constructed cable and the number of homes per cable mile for each section, this data along with the penetration estimate should be used in computing the estimate of total subscribers.

Subscriber densities, by which the sections are ranked, can be expressed either in terms of subscribers per square mile or subscribers per cable mile, depending on the data available. Ranking the sections on the basis of subscribers per square mile is equivalent to ranking by subscribers per cable mile provided the ratio of cable mile to square mile and the construction costs per cable mile do not vary significantly from section to section. If there are significant variations in these factors, they should be included in the estimates when determining the construction schedule.

Confidence limits for each demand estimate are provided based upon the standard error associated with the econometric model. This error is applied to the penetration estimates calculated for each section to give upper and lower limits on the penetration. These limits are then used to obtain corresponding limits on total subscribers and subscriber density.

After a CATV system is operational, the number of subscribers can be expected to change due to changes in system penetration resulting from the maturing of the system, from changes in the exogenous variables (e.g., reception quality, income, price, etc.) and from population shifts. The effect of these dyanmic factors has been included in our methodology by the use of income and household projections which take into account shifting population and income patterns within the community, and by the maturity factor in the econometric model. Thus we are able to make estimates of demand at different stages of the proposed systems operation.

Use of Demand Estimates

The primary use of the results of this method is for comparative analysis. While the actual demand estimates for the sections contain a substantial amount of uncertainty, the relative rankings of the sections should be quite reliable provided their subscriber densities differ significantly. The method also lends itself to use in a sensitivity analysis in which relative changes in demand or shifts in the ranking of the sections due to changes in the independent variables are studied.

The confidence limits associated with the demand estimates reflect the uncertainty in the estimates and should, therefore, be taken into consideration in any additional analysis based upon the numerical values. They should be used to obtain confidence limits for any further estimate derived from either the penetration or demand estimates.

A CASE STUDY

To illustrate this method consider the following situation: A CATV system is proposed for a city located in one of the 100 top TV markets with good-over-the-air reception. There is currently being received in the city a total of 6 over-the-air stations; 3 VHF network stations, 2 UHF independent stations and 1 VHF educational station. The subscription fee for the proposed cable system is $72 per year. The system will offer a total of 10 channels; 3 network channels, 5 independent channels and 2 educational channels. One of the independent channels is a Spanish language station and one of the educational channels has a major portion of its programming devoted to Black oriented programming.

Our objective is to divide the city into sections of uniform density which can be ranked for system development. For the sake of this example we will assume that construction costs per cable mile are approximately the same and that the ratio of cable mile to square mile does not vary significantly throughout the city. Therefore, estimates of subscriber density in subscribers per square mile are adequate for ranking purposes. No specific construction schedule will be included in this analysis; the estimate of subscribers and subscriber densities in each section is based on the assumption that all households are passed by the cable. Estimates of demand after two years of operation (1975) will be provided.

The 1970 Census is the most recent enumeration for the 61 census tracts in the city. At that time the total population of the city was 520,000 with a total of 180,500 households. Data for the key demographic variables in the census tracts exhibited the following ranges of values.

Population: 126-23,000

Households: 34-7118

Area (sq. mi.): 0.26-9.80

% Black Households: 0-100

% Spanish Surname Households: 0-75

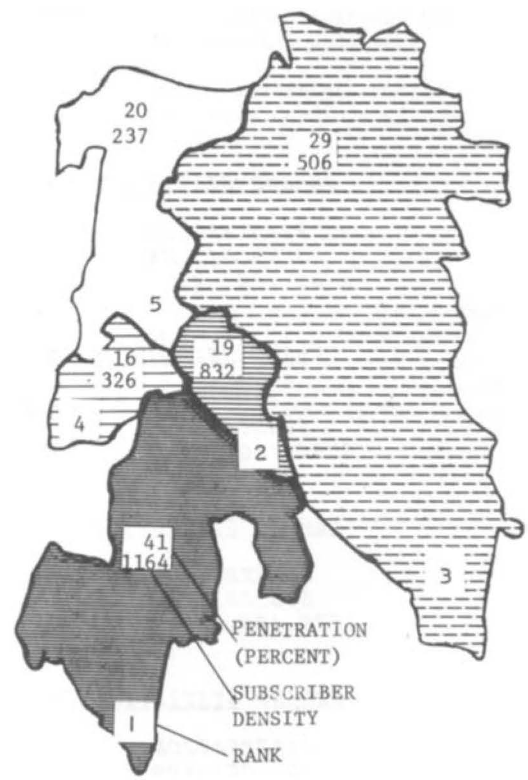
Median Income: $3,383-$29,663

Data pertaining to 1975 is developed in the following ways: Projections of households by census tract for 1975 are obtained from the local planning board. Because projections by ethnic group are not available for 1975, this data is constructed by applying the 1970 percentages to the 1975 household projections by census tract. Income projections by census tract for 1975 are developed by using a factor based on data for the city obtained from the Bureau of Labor Statistics. Recent citywide data on color set and UHF receiver ownership are assumed to apply uniformly throughout the city.

Topological data for the city indicates that reception quality of the over-the-air broadcast stations can be assumed to be uniformly good and so degradation factors of zero are used for all stations. On the basis of surveys, general interest in the two special interest channels is estimated to be 25% in the educational channel and 15% in the independent channel.

The estimated penetration obtained by applying Park's model to the city as a whole using 1975 data is 31.1% with a subscriber density of 626 subscribers per square mile. Preliminary estimates of penetration by census tracts range from a low of 9% to a high of 67% reflecting, for the most part, the income ranges and, to a lesser degree, the ethnic composition in the census tracts. Subscriber densities in the individual tracts range from 60 subscribers per square mile to 1440 per square mile. By grouping contiguous census tracts of similar subscriber densities, we can form five sections of the city with penetrations in the sections ranging from 16% to 41% and with subscriber densities of 237 to 1164 subscribers per square mile. The penetration and subscriber density of the city obtained by aggregating the section estimates are 30.9% and 621 subscribers per square mile which are in close agreement with the estimates obtained by applying the econometric model to the city as a whole. The recommended order for system development based on the ranking of the sections is shown in Figure 1. The demand estimates and their confidence limits for each section are given in Table 1.

FIGURE 1.



DEMAND ESTIMATES AND RANKS FOR

SECTIONS OF THE CITY WITH

GOOD RECEPTION

TABLE 1. *GTE LABORATORIES INC.,PROJECT 482*

CATV PENETRATION PROFILE
(BASED ON MODEL BY R.E.PARK, RAND CORP.)

ASSUMPTIONS

PERCENT UHF SETS: 95.0
PERCENT COLOR SETS: 66.0
SPANISH STATION INTEREST:0.15
BLACK CHANNEL INTEREST:0.25
BROADCAST STATIONS: 6
CATV CHANNELS:10
RECEPTION QUALITY: *GOOD*
PRICE:$72
AREA BASE: *ADJUSTED*

DEMAND STATISTICS: 1	LOWER LIMIT	EXPECTED VALUE	UPPER LIMIT
PENETRATION	17.86	40.82	47.05
SUBSCRIBERS	11340	25918	29870
SUBS./SQ. MILE	509.43	1164.33	1341.86

DEMAND STATISTICS: 2	LOWER LIMIT	EXPECTED VALUE	UPPER LIMIT
PENETRATION	9.68	18.65	25.51
SUBSCRIBERS	1654	3185	4356
SUBS./SQ. MILE	431.85	831.59	1137.33

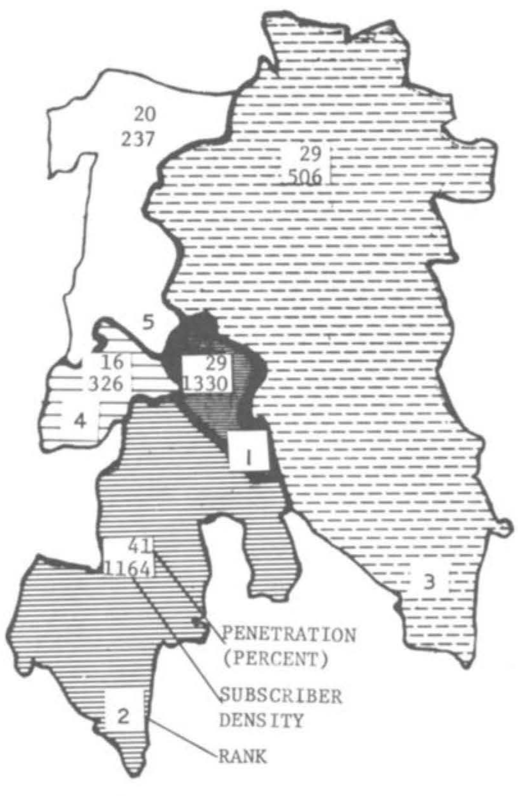
DEMAND STATISTICS: 3	LOWER LIMIT	EXPECTED VALUE	UPPER LIMIT
PENETRATION	14.00	29.40	36.88
SUBSCRIBERS	12591	26444	33165
SUBS./SQ. MILE	240.92	506.00	634.61

DEMAND STATISTICS: 4	LOWER LIMIT	EXPECTED VALUE	UPPER LIMIT
PENETRATION	8.58	16.18	22.60
SUBSCRIBERS	912	1719	2400
SUBS./SQ. MILE	173.05	326.18	455.40

DEMAND STATISTICS: 5	LOWER LIMIT	EXPECTED VALUE	UPPER LIMIT
PENETRATION	11.01	21.76	29.00
SUBSCRIBERS	1679	3318	4423
SUBS./SQ. MILE	120.61	238.36	317.74

Variations can be introduced into the problem: Suppose, for instance, that over-the-air TV reception in the community is unimpaired except in one area of the city where, due to tall buildings, the reception quality is equivalent to that experienced midway in the B contour of the stations. Assume, also, that this is an area with a high population density and relatively low median income. Should system development start in this area rather than in a higher income, less dense area where reception quality is good? Using reception degradation factors of 0.5 we find an increase of penetration of 10 percentage points resulting in a subscriber density of 1330 subscribers per square mile and a change in the ranking of the sections as shown in Figure 2.

FIGURE 2.



DEMAND ESTIMATES AND RANKS FOR

SECTIONS OF THE CITY WITH

LOCALLY IMPAIRED RECEPTION.

It should be noted that, in practice, construction costs might also be significantly different in this area and should be taken into account in the ranking.

Conditions imposed by the franchise can also be included in the analysis of a community. If, for instance, portions of the city are in different political districts (e.g., counties) and the franchise stipulates that system development must progress simultaneously in each, sections can be formed and ranked within each political district and a development schedule provided for each.

CONCLUSIONS

A flexible and systematic method for estimating CATV subscriber densities in a large city with good over-the-air TV reception has been developed. The method has the following features:

1) Provides demand estimates suitable for engineering and financial planning.

2) Incorporates local constraints and/or restrictions on system development and includes local demography.

3) Employs readily available data.

4) Is based on the most advanced model of CATV penetration currently available.

As more cable systems are implemented, and as innovative subscriber features are introduced, a new penetration model which takes into account these advancements could be devised and used in place of the econometric model presently used in the method.

HELICAL VIDEO TAPE INTERCHANGEABILITY REQUIREMENTS

K. Blair Benson
Goldmark Communications Corporation

BACKGROUND

In the initial applications for helical video tape recording and playback equipment, the necessity for recordings to be playable on any number of players in a variety of locations generally was not a serious consideration. In most cases recordings were made off-air or of local originations, either of entertainment or educational programming, for subsequent playback on the same VTR equipment used for the recording. Although there were some exceptions, such as regional or state educational organizations where video tape was used to avoid the expense of intercity transmission and the scheduling limitations of network programming release from a central location, these were in minority and thus not a major factor at the time in forcing development of technical improvements and standards dictated by interchangeability requirements.

More recently, however, with the increased use of helical color video tape in fully automated cable and hotel pay-TV origination systems, for advertising and sales promotion, and for expanded educational non-interconnected networks, all requiring quality competitive with that of television broadcasting, the need for optimum interchangeable performance has become acute. The problems have been compounded, first by the large number of organizations providing a recording service in a manner not unlike that of a film processing laboratory producing prints for distribution to theaters and other exhibitors, and second by the increasing number of manufacturing sources for equipment and tape.

Apart from technical quality, another complication which results in increased cost of duplication and distribution, is the large number of different recording formats. In fact, not only are there a variety of tape widths, but in addition, a variety of formats exist for any one width. The problem is considerably more complex than in motion-picture film where the gauges are limited to Super-8, 16mm, and 35mm; and the formats to two types of wide-screen and the traditional 3-by-4 aspect ratio.

To cite a few of the cassette or cartridge formats in use, in 1/2-in. there are the EIA-J and Philips, in 3/4 in. the U-Matic and in 1-in. the IVC and Ampex. The problem has grown to such proportions that it is no longer amenable to standardization through the work of industry technical committees. Instead, the question of the best format for each application undoubtably will be answered in the marketplace by the user. Therefore, industry activity appears likely to be the development of recommended practices for each of the different formats which are available from two or more manufacturing sources.

The following review of the interchangeability requirements, for the most part will be general in scope; in the interest of brievity, where specifics regarding parameters and tolerance are discussed, the 3/4-inch U-Matic or 1-inch IVC formats will be used as examples.

SURVEY OF INTERCHANGEABILITY FACTORS

It is the obvious conclusion from the foregoing that interchangeable performance has to be developed to a level where no operating or set-up adjustments are required in order to accommodate tapes made on different recorders or head assemblies. In order to achieve such highly desirable operational flexibility, rigorous control must be exercised over many parameters. The scope and magnitude of the design and operational problems is indicated by the following tabulation of the most significant factors to be taken into account:

Video Head
Dihedral alignment
Head-to-Tape speed
Gap azimuth alignment
Tape guiding
Recorded track dimensions

Magnetic Tape
Physical dimensions and properties
Magnetic properties
Cassette dimensions
Humidity and temperature effects

Video Signal
Carrier frequencies for sync level
Carrier frequencies for reference black level
Carrier frequencies for reference white level
Luminance and chrominance bandwidths
Color carrier signal
Pre- and post-emphasis

Control Track
Recorded level
Signal phasing to video signal
Track dimensions and placement
Gap azimuth alignment

Audio Head and Track
Recorded level
Track dimensions and placement
Gap azimuth alignment

As will be evident in the subsequent discussion, several of the items listed cannot be adjusted or corrected except during manufacture. For these parameters, the user's control is limited to a periodic check in order to determine if the equipment remains within the permissable limits of tolerance. However, many of the factors are subject to set-up or operating adjustment and must be held under careful control if interchangeable performance is to be comparable to that achieved in non-interchangeable operation.

VIDEO HEAD
One of the objectionable degradations in helical video tape playback results from timing errors in the playback signal. These can be caused by misalignment in the head assembly, or more frequently, differences between recording and playback, head-to-tape speeds.

Dihedral Alignment: If the two video heads are not precisely 180-degrees apart on the head drum, following the switch between heads the television fields from each of the heads will be displaced horizontally. This will be apparent as dual images of vertical lines at the top of the picture. The longer the receiver horizontal sync time-constant, the further down from the top of the picture the split of vertical lines will be present. Adjustment tension by the skew correction control will have no effect on this error. Normally dihedral errors are corrected by factory alignment. However, in the event it is required, an adjustment can be made in the field using an alignment tape provided by the manufacturer as a test reference.

Head-to-Tape Speed: A difference in recording and playback speeds will result in a cumulative error in signal timing. This will be evident as a horizontal shift of the video immediately following the switch between heads, and differs from dihedral errors in that both fields shift identically rather than in opposite directions. The effect is commonly known as skew. On receivers with a short horizontal sync time-constant, the effects of such time-base errors do not cause a serious degradation in picture geometry, since the shift in horizontal positions is usually limited to a few lines at the bottom of the picture after head switching and is covered by the receiver mask. However, on most receivers of American manufacture, having a long time-constant, the horizontal circuits may not recover from the abrupt change in timing of sync until a third or more down from the top of the picture. In this case, the resultant skewing is very disturbing and must be corrected if viewer satisfaction is to be achieved.

The time-base error can result from differences in radius of the video heads' travel because manufacturing tolerances and wear of the pole pieces. However, the largest errors usually are the result of tape tension adjustments and environmental effects of humidity and temperature; the latter will be covered in detail under magnetic tape factors.

The recording writing speed and playback time-base errors are adjusted by control of tape tension, the latter in part or completely depending on the type of control used. Goldmark Communications has developed an automatic braking system for tension control and accompanying skew correction which reduces the error to under a micorscond.

Other systems for timing error correction are available from Television Microtime and Consolidated Video Systems which reduce the time-base error by introducing an automatically-controlled variable delay in the playback video signal.

Tape Guiding: Improper tape guiding around the head drum can cause partial or total loss of the video signal. It also can result in improper wind of tape on the feed and take-up reels and possible jamming of tape in the player.

MAGNETIC TAPE
Magnetic Properties: In order to obtain a maximum signal-to-noise ratio, helical scan equipment normally is designed to operate with high energy magnetic tape. Two types of coatings provide this character-

istic; they are chromium dioxide and cobalt oxide. Both of these materials provide substantially higher playback signal levels than ferric oxide tape. However, in order to achieve this performance, higher record current is required. In summary, with proper record current levels, and with the more efficient ferrite video heads, an improvement of 3 or 4dB is achieved.

Humidity and Temperature Effects: Environmental effects from variations in humidity and temperature can be major factors in restricting interchangeability. Because of the long track traced by each head, a dimensional change in the tape, or in diameter of the head drum, can cause a large variation in resultant tape-to-head writing speed. The primary control of this parameter is through an adjustment of tape tension as it passes over the head drum so as to introduce a controlled amount of tape stretch. Thus, a head-to-tape reading speed can be maintained during playback which is equal to the writing speed during recording.

The range of timing difference resulting from humidity and temperature changes can be calculated, for example, for the 3/4-in. U-Matic format using the following constants:

Humidity coefficient of tape
11×10^{-6} in/in/%RH

Temperature coefficient of tape
15×10^{-6} in/in/°C

Temperature coefficient of aluminum
11.7×10^{-6} in/in/°C

Head-writing speed
404 in/sec.

Track length
6.7 in/field

The calculations of change in microseconds for each percent of relative humidity and degree fahrenheit are tabulated below:

$$\frac{11 \times 10^{-6} \times 6.7}{404 \times 10^{-6}} = 0.182 \text{ u sec/field/\%RH}$$

$$\frac{(15 - 11.7\ )\ 10^{-6} \times 6.7}{404 \times 10^{-6}} \times \frac{5}{9} = -0.2 \text{ u sec/field/\%RH}$$

It is interesting to note that temperature changes cause a timing error opposite to that from humidity. This results from the fact that the effect of a temperature change on the head drum is greater than on tape and, in addition as related to timing errors, is in the opposite direction.

The importance of environmental control is highlighted by a calculation of the timing error resulting if a recording made under conditions of 50% RH and 70° RH is played back with an ambient of 90% RH and 90°F. The humidity increase will cause an increase in recorded track length of 7.28 microseconds, whereas the increase in temperature will cause an increase in head speed which is equivalent to a reduction in track length of 4.0 microseconds. The net effect is a timing error of 3.28 microseconds. This will be apparent on the television picture as a shift of video information to the left after the switch between heads.

VIDEO SIGNAL

One of the major problems in any phase of television program transmission is the consistent maintenance of uniform video signal levels. In video-tape recording and playback these parameters are dependent not only upon the gain of the various video amplifiers, but also upon the deviation employed for the frequency modulation of the carrier signal applied to the recording heads. Second in importance to level control is the choice of pre- and post-emphasis for optimum compromise between video bandwidth and signal-to-noise ration.

Carrier Deviation: The choice of deviation frequencies is dependent upon two opposing factors: (a) The bandpass limitation of the tape and head combination, and (b) the need for a maximum signal-to-noise ratio. The signal-to-noise ratio will vary directly with the peak-to-peak magnitude of the deviation. Therefore, it is desirable to modulate the carrier over as wide a band as the system elements will permit. However, if the high-frequency cutoff of the head and tape is exceeded, a distortion of peak white signals is produced. In some types of playback circuitry this results in a failure of limiting action prior to demodulation and permits the random noise from the tape and preamplifier during playback to be amplified to full signal level by the high-gain amplifiers. The resultant noise and streaking in peak white are very objectionable in appearance. In other circuits the effect merely is a loss in peak whites similar to that produced by a peak limiter. It is apparent that the maximum frequency deviation must be limited to that which can be accommodated by any or all playback equipment.

Pre- and Post-Emphasis: Current practice in most systems is to provide a rising amplitude frequency characteristic wherein the 3 MHz response is 10 to 12 dB above the low-frequency response. In order to avoid overload distortion from the higher frequency components, the rising response characteristic is restricted normally to frequencies under 3 MHz. Appropriate post-emphasis is employed to provide a flat overall system response over the desired luminance bandwidth.

CONTROL TRACK

The control track synchronized the rotation of the heads with the tape travel. Most equipments have means to automatically phase the head rotation with the control track. However, this will not correct for errors in recorded phase of the control track relative to video tracks. Correct phase is essential for maximum signal output and signal-to-noise ratio.

CONCLUSION

If the maximum potential quality and operational flexibility in the use of helical tape equipment on an interchangeable basis are to be realized, it is essential that a vigorous control of recording and playback parameters, as well as environmental conditions, be maintained. However, these disciplines in control require that agreement be reached on recommended practices and standards among manufacturers and users. Realizing the importance of the problem, the Joint Committee on Intersociety Coordination (JCIC)* has assigned the task to the Society of Motion Picture and Television Engineers (SMPTE). The SMPTE, in turn, has instructed the Helical Recording Subcommittee of the Video Tape Committee to proceed expeditiously to develop recommendations for industry adoption.

* Representing EIA, IEEE, NAB, NCTA, and SMPTE.

IBM - TELEPROMPTER DIGITAL TRANSMISSION TEST SATELLITE & CATV

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The IBM Corporation has been conducting a series of technical experiments on the use of satellites for wideband digital communications at its Development Laboratory in Poughkeepsie, New York.

In addition, experiments have been conducted using digital techniques for transmission over wideband CATV facilities.

IBM's RF and satellite access was supplied by RCA Global Communications as a "Development Satellite Transmission Service" as Tariff FCC No. 92. The access to Telesat Canada's ANIK Satellite Transponders via this tariff and the associated leased RF Terminal deployed onto IBM property have well satisfied the RF Transmission needs.

The TelePrompTer Corporation has been touring the United States with its "Space Cast" Project, a transportable receive-only Earth Station, making test measurements of signal strength and interference of video signals from the Telesat Canada Domestic Satellite in geostationary orbit 22,300 miles over the earth's equator.

The unique opportunity to deploy a replica of the Poughkeepsie digital transmission equipment, and utilize the facilities at Rochester, Minnesota using the TelePrompTer Earth Station and their CATV system in a realistic digital transmission environment occurred the week of December 17, 1973.

This paper will describe the results of this cooperative digital Transmission Test.

Industrial Two-Way CATV Systems

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ABSTRACT

The technology of two-way CATV transmission has provided a unique solution to the internal communications problems within large scale industrial complexes. At one time the requirements for audio, video, and digital information within a plant were treated on an individual basis. Cable technology combined with minicomputers and inexpensive terminals is beginning to integrate these needs in a single cost effective system in the automotive and chemical industry.

INTRODUCTION

The progress of two-way CATV growth has been slowed due to many inter-related factors. These can best be summarized in terms of the lack of a defined market for the range of potential services which were available. Not only has it been difficult for the cable industry to successfully demonstrate the services, but it also has been equally difficult for subscribers to define the nature of the service that they would purchase. This stalemate may yet be broken under the proper conditions. However, the effort which has been expended in the development of the hardware associated with two-way cable systems has not been misspent. Many industrial or institutional complexes represent a market in which the "subscriber" has clearly drawn communication requirements which can be uniquely fulfilled through cable technology.

Industries which are spread over a significant area have been faced with an ever growing internal communication need. As plants have expanded, the need for acquiring and delivering data on a real-time basis to many locations has outgrown the conventional means of distribution. Similarly, the audio and video applications have grown in similar proportion. The central problem in all three areas is not the devices used for digital, audio, and video needs, but rather the more mundane issue of the wiring required to interconnect literally hundreds of possible communication points either with each other or to computers. The wiring problem for data usually involved the installation of twisted pairs within conduit. The audio intercom employed a separate twisted pair facility, and the distribution of video is commonly carried on individual base band video cables also in conduit. It is this maze of wiring which prevents the coherent growth of industrial communications.

However, the wide-band coaxial cable represent a consistent and unified approach to the problem, and by its very nature permits an orderly expansion of the communication network without having to rewire a plant each time. In many ways, the industrial plant looks like a small community in which there are hundreds of points effectively tied together through a coaxial cable. Therefore, it may be useful to the CATV industry to briefly review the progress to date in industrial CATV and perhaps conjecture about the role that the cable operator might play in this growing field.

INDUSTRIAL APPLICATIONS

The typical scope of two-way applications can be divided into three broad categories. The first involves the transmission of digital data from the production floor and between computers. Production information is generated by a variety of methods. The most typical is the entry of keyboard data directly by machine operators, inspectors, and foremen. This provides a real-time data base for the types of defects which are encountered and the status of production machinery. Another class of digital data arises from the acquisition of sensor outputs, such as piece count, analog process variables, alarms, badge and card readers, etc. In addition to the retrieval and processing of this information, there is a need to provide operators and management with real-time reports of the significant data. The latter is usually accomplished through CRT displays to minimize the generation of hard-copy reports.

The second category of transmission requirements concerns the distribution of video or conventional TV. This includes not only the ordinary security surveillance, but also the expansion of two-way television for in-plant training, management conferences, random access to micro-fiche material, remote inspection of production line problems, etc. Furthermore, the audio/video conferences are now being considered between distantly located plants as well as within the plant environment itself. There is also a control aspect associated with these TV applications, particularly in the form of remotely controlling the motion of the camera.

The third category of information transmission is the standard audio distribution. The first level of requirement is for the establishment of party-line conversations among production and supervisory personnel outside of the usual telephone circuitry. The audio associated with the TV program is, of course, another application. However, a more encompassing problem is the in-plant telephone network itself. As will be seen, this situation can also be accommodated on the coaxial cable.

Again, the common factor in all of these applications is the extensive wiring requirements, the large number of terminal points, and the extreme bandwidths involved from simple switch contact closures to television. However, this is precisely the transmission problem envisioned in the use of two-way services for community CATV using a broadband coaxial cable. When viewed in this light, it becomes very practical to utilize the low cost modems that were developed for interactive TV services.

INSTALLATIONS

Utilizing the above concepts, Interactive Systems, Inc. (ISI) has installed coaxial cable and the associated modems and computer hardware at the following sites.

A. Oldsmobile - Lansing, Michigan

Approximately 15,000 feet of coax was installed for the operation of a defect monitoring system. Inspectors would key in observed defects in real-time and the computer would provide CRT displays of the most frequently occurring defects. Plans are underway to interconnect another plant by cable which is two miles away and to extend the applications to other data displays and television programming.

B. Chevrolet - Detroit, Michigan

A security surveillance system was installed with 8,000 feet of coaxial cable. In addition to carrying the TV signal, the cable was also utilized to transmit all of the signals for remotely controlling the cameras. The cable was run parallel to the axle production lines, so that future data acquisition systems can be easily added, merely by tapping into the cable and avoiding the need for new wiring.

C. American Motors - Kenosha, Wisconsin

Over 15,000 feet of cable has been installed with 80 terminals again for a real-time defect monitoring application as well as television for employee programs. Consideration is also being given to extending the cable to neighboring plants to integrate the data acquisition and retrieval and to include machine status monitoring.

D. Dow Chemical - Midland, Michigan

This is probably the largest facility encompassing 40,000 feet of trunk cable over a two square mile area and interconnecting 75 buildings. The initial application involves television programming for in-plant surveillance, training, and then expanding to two-way audio/video conferences. An experimental data acquisition program is currently in progress to transmit in a digital mode, analog process information from scattered buildings to a central computer.

Plans are now being completed to install another cable facility at Corporate headquarters about two miles from the production facility. Although it will have its own head-end facility, there will be one interconnect to the initial installation. Part of the same program includes an experimental microwave facility to interconnect a distant Dow facility with the Midland complex for two-way audio/video conferences.

Several other corporations are also planning to install the ISI Videodata network for even more sophisticated applications involving computer to computer transmission.

GENERAL TECHNOLOGY

A. Cable Configuration

In almost all cases, the solid aluminum sheath cable was used for trunking and feeder lines, with double shielded RG-6 used for the short drops to terminals. Considerable cost reductions were achieved through the elimination of conduit.

Where extensive applications were envisioned within a complex, it was decided to use a single cable (multiple leg) mid-split spectrum in order to provide symmetric transmission potential from any point in the system. Unlike conventional practice, the trunk system is tapped for branches rather than using bridger amplifiers.

B. Data Transmission

The digital network is similar to those previously described in the literature on broadband cable. ISI utilizes a two computer system; one for the general polling of terminals, and the second for the actual application data processing. A generalized modem has been developed which permits a very simple interfacing to a variety of peripheral devices such as:

a. keyboards
b. printers
c. CRT displays
d. alarm monitors
e. production controllers
f. analog-to-digital and digital-to-analog transmission
g. production sensors

The data rate for these channels has been adjusted for particular customer preferences and varies from 12,000 bits/second to one megabit.

C. Audio/Video Transmission

For straight video transmission, conventional CATV modulators and demodulators are utilized with security achieved through the use of mid-band and super-band channels, computer controlled converters, and scrambling. Separate modems have been developed for party-line audio networks or for control systems which could not effectively utilize digital modems.

The use of the cable network for telephone communication within a plant has recently been developed by Collins Radio. The basic system can accommodate up to zero phones on the equivalent of one TV channel. Further, study is currently underway to integrate this process with the previous audio, video, and digital transmission network. Thus, within the not too distant future, new industrial plants may be completely cabled for their entire communication needs.

ROLE OF CABLE OPERATOR

Although the industrial and institutional area may seem foreign to the cable operator, this may actually afford new opportunities to provide services on a selective basis. There are many advantages to the operator, particularly MSO's, in entering some portion of the market.

1) The cable installation and maintenance is similar to his present activity.
2) It offers a means of learning the two-way technology in a more controlled environment.
3) In some cases, the transmission between plants could involve his own cable network, or the overlaying of a new link which is leased or owned by the customer.
4) The market is profitable.

The industrial field will undoubtedly grow more rapidly than CATV under present conditions, and this may be the time for re-evaluation.

NEW COMMUNICATIONS TECHNOLOGY IN NEW TOWN DEVELOPMENT (ROOSEVELT ISLAND)

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One of the numerous new communities across the country, Roosevelt Island provides 5000 New Town units within New York City, and is being developed by the Urban Development Corporation of New York State. This report offers a look at how we can begin to shape an integrated communications infra-structure based upon CATV technology.

Several examples of existing community based video applications are described which themselves could be suitably interfaced into local service for the Island. Other examples are mentioned that could be implemented if the resulting community determines that such uses would satisfy their programmatic needs.

Finally and most importantly, unresolved issues are highlighted. Perhaps the watchwords should be "enact no major constraints." For better or worse, this experience has been developed on the expedient principles of "ad hocracy". Time will tell.

The Roosevelt Island Development Corporation is considering the provision of a cable television network with sufficient capacity for two-way communication, and additionally use the system for burglar and fire detection. It is contemplated that the system would be used for community meetings, to communicate between the educational spaces and to provide a link to the hospitals from apartments of the handicapped and homebound.

A community information and referral center, composed of an Educational Exchange and a Communications Center is proposed to be located in the Town Center. It may not be practical to develop a full range of services, programs and offices needed for this number of residents--18,000--but the Center would make direct referrals and provide assistance to Island residents, much as in the demonstration of community information services to be demonstrated under a Federal grant by the Brooklyn Public Library at some 50 different neighborhood offices in its service area

EDUCATIONAL PROGRAM

The proposals for the community educational system prepared for RIDC in 1970 (footnote 1) outlined a number of educational programs situated in a locationally dispersed but functionally integrated system that would require a high level of communications interaction, that potentially offers a formulation of standards for adoption by other new communities These programs, which would be administered under the auspices of School Board District #2, are largely structured to respond to the growing educational needs of one-third of the population who are non-adult (6000).

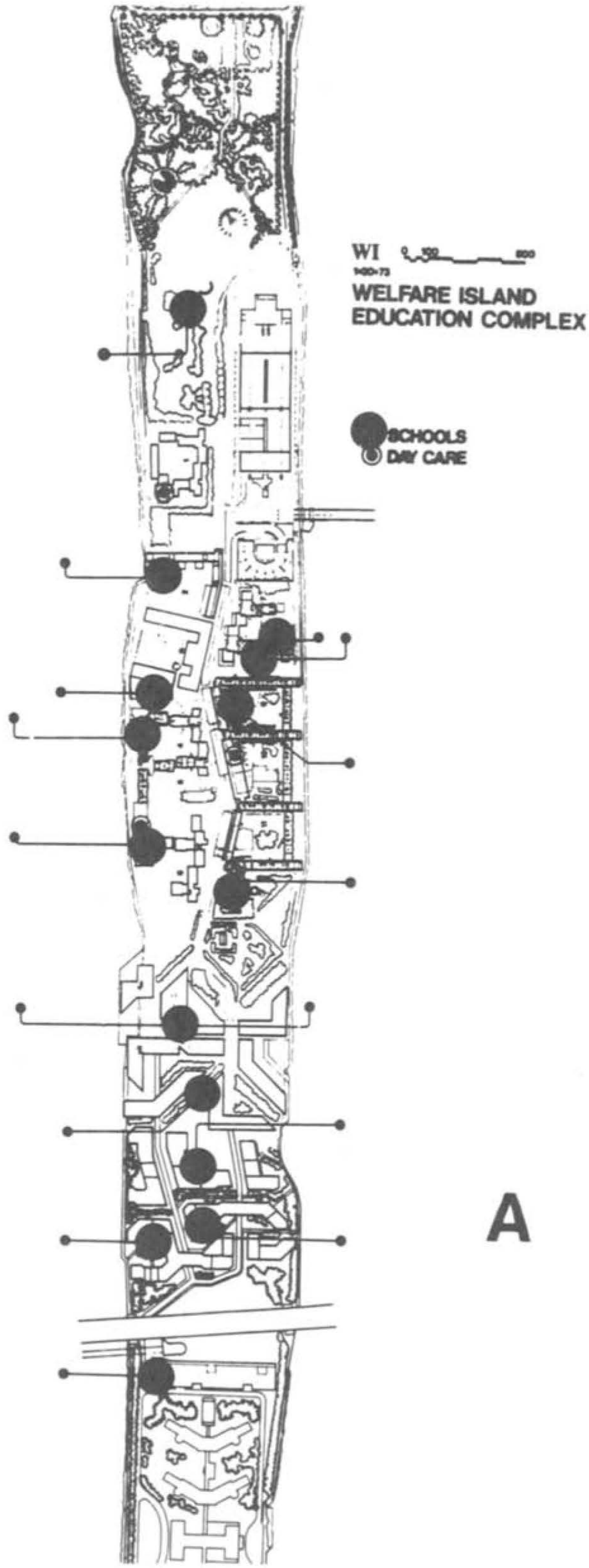
Since the idea of joint occupancy by educational and residential space has been pioneered by the Board of Education and the Educational Constructional Fund, with scattered sites woven into the residential component, intercommunications among educational spaces has become more important. On the Island, educational sites will be further dispersed among the residential and other functional locations within the community (figure A). Perhaps fifteen different locations will provide educational spaces, including the hospitals, library, police/fire station, theatres, stores and elsewhere. Certainly the cultural locations will be prime resources. The main educational facilities for older children in particular will be in seven major centers, each organized around a central concept, such as the arts, communications, economics and work, science and technology, behavioral science, environment and planning, and recreation. Two centers for high school programs will accomodate a Visual and Performing Arts program, and a Physical Science and Technology program. The former will include curricula in both the fine arts of dancing, music, drama, and film, but will also provide instruction and participation in the applied arts of the media, including film, television and video, audio, and graphics design. The Physical Science and Technology Center will share spaces for laboratory and studio use in electronics and telecommunications.

COMMUNITY CENTER

As proposed by this educational report, the nerve center for a sophisticated organism such as the community education system, must be the Communications Center. Unless communications between facilities and people are almost instantaneous, direct and accessible to all of the participants, the system will falter. In addition, the Education Exchange is to be a bank of information about the community which is available to all Island residents, and ideally would be a computerized information and learning system tying the entire school system and the community together through the common cable television and computer terminal network encompassing all of the Island and any of the residents who wish to participate. The CATV network distributes the educational "supermarket" of ideas and information among a cadre of inspired administrators, teachers and facilitators who--hopefully--will help make things happen for the participants.

The institutionalization of this function needs further definition if it is ever to take shape in the life of Roosevelt Island. But Robert Bartz finds that the technology of a "Community Exchange" (footnote 2) is easy to describe: "It links the individual, at a neighborhood center or even in the home, with public agencies and organizations which offer services, including information; it links him with sources of educational/entertainment/cultural programs, either live or upon request; and it links him or his interest group with other parties concerned with public matters (here the dialogue element becomes all-important. The links are quite generally two-way--that is one can send as well as receive. They are normally both visual and audio, and they are normally instantaneous. Overall this technology is well advanced, and it is also, for many applications close to economical. There are technical challenges which remain, but they are largely of only long-term significance.

"Community Exchanges can originate with a library, museums of the modern variety, and in other ways. These origins might or might not lead in particular situations to a specifically incorporated entity to manage the facility and its operation. The Community Exchange can be small and informal, or large and structured. Emphasis will vary widely. The economic, the training, and the health requirements of a typical community consisting largely of retired people are utterly different from the requirements of a young community which is heavily minority, and of very low income. Any one Community Exchange will reflect such differences in its genesis and growth; it must itself be an expression of the felt and perceived needs of the citizen. El Segundo, California will be unlike Dallas, which will be unlike Boston; there's a different and interesting story behind each of these communities for its promise as a setting for a Community Exchange."



A

CONSTRUCTION STANDARDS

The HUD environmental statement (footnote 3) reports that "the present Island telephone facilities are supplied by cable from Manhattan and reach the Island through conduits under the deck of the Queensborough Bridge in much the same manner as the electric power lines. The cables enter conduit risers attached to the west tower of the Bridge, which are routed into manholes at the base of the tower. The cables then run north and south to the extremities of the Island in conventional 12 duct terracotta conduit banks running under the westside of the Island. Fire alarm and other circuits are carried in cables in adjacent ducts in the same bank. Most of the present facilities will be removed.

"The proposed new system will consist of 12 duct conduit banks running under the new streets. The 4" ducts will be plastered and encased in concrete and placed on a reenforced concrete base to preserve alignment. Service ducts into building parcels will generally be 4" steel pipe. The duct system as presently designed will be capable of accomodating adequate police call and fire alarm systems over the entire Island and available for the Sterling franchisee to install a cable television system."

Subsequently, a statement of detailed recommendations was prepared by the telecommunications consulting firm of Malarkey, Taylor & Assoc (footnote 4). These recommendations centered around the principle of anticipating the real immediate and future communications needs, and of originally providing within the buildings themselves, sufficient horizontal and vertical ductwork to significantly minimize the incremental costs of enlarging capacity.

The substantive position of the report states: "The most important points, by far, are to provide adequate riser space to accommodate any conceivable future communications requirements. Riser space includes floor sleeves or installed conduit of adequate size, and the enclosed spaces should be designed to permit only authorized access to pull-boxes. We cannot overstate the importance of providing adequate floor sleeves and riser space to avoid the future necessity for breaking holes in concrete floors, or opening and repairing finished walls. The cost of unused riser space has to be so small compared to the cost of providing for it after the building is completed and occupied, that there can be no sound reason for not providing for any conceivable future communications riser requirement.

"We recommend that serious consideration be given to providing floor sleeves in addition to the minimum required, so that additional duct could be installed for unanticipated future needs. Nearly equal in importance is adequate horizontal wire-way for communications interconnection between buildings, and between each building and the Empire City duct system throughout Welfare Island and Manhattan. Although 4" interconnecting duct may seem excessive, repeated painful and costly experience indicates a high probability that its utility and value will be gratefully acknowledged in the future. It is so much less costly to install such ducts in new construction than to break holes in walls and excavate streets when the need arises, that failure to provide adequately for interconnection now must be considered imprudent, at best. Overproviding is far less costly than underproviding.

"We suggest that careful consideration be given to the probably future need for additional outlets within each apartment. At least 20% of all CATV subscribers now have two or more TV outlets per dwelling unit, merely for family convenience in viewing TV. When the predicted broadband communications services become available, we can certainly anticipate a demand for interactive terminals requiring additional outlets in the kitchen, den, student's room, etc. The demand is too ephemeral at this time to warrant additional outlets initially. But we strongly advise that a suitable plan be developed by which additional outlets can be reasonably installed in the future at most any location in the apartment."

Additional MT&A recommendations:

aa) basement connection panels.
a) one cable per riser for MATV.
b) one separate cable for each apartment for CATV.
c) suggested duplicate of the CATV system for strictly local, Roosevelt Island communication, allowing for future expansion.
d) up to six cables for each school or day care center--to provide for possible instructional uses of television, including camera origination in classrooms.
e) stores and office spaces should have the same facilities as apartments, since commercial applications of broadband communications may develop more rapidly than residential applications.

By closely scrutinizing and largely adopting these basic recommendations, the Roosevelt Island community will be able to keep pace with the rapidly developing telecommunications component of our society.

UNIVERSAL CATV WIRE-UP

However, the franchise operator, Sterling Manhattan, and the Roosevelt Island Development Corporation have not been able to negotiate an agreement for wiring the subscribing CATV apartments. RIDC wishes to provide a master antenna system (MATV) for all its residents within ductwork provided in the original construction. Sterling would rather utilize exterior cabling that can be more easily maintained and policed for nonpaying taps. However, if the MATV system is provided, the economic inducement for subscribing to the Sterling CATV is greatly diminished. If on the other hand, Sterling were to agree to RIDC's proposed monthly contract rate of $3.50 per apartment ($17,500 for all 5000 units), it would monetarily correspond to an effective 58% penetration rate--which is much higher than their market success at present, with less cost and maintenance for internally constructed ductwork, obviating nonpaying taps, and without the severe competition of a universally provided MATV connection--and with the social utility of achieving universal wire-up for the transporting of all educational and community video communications to all apartments of all income levels.

APPLICATIONS

According to our knowledge of today's practical applications of cable communication, the following examples would have immediate applications in Roosevelt Island's cable communications system:

TELEMEDICAL

Mt Sinai Hospital has been operating a two-way fully interactive video telemedical consultation service with the Wagner Pediatric Clinic in East Harlem (figure B, footnote 5) serving 1300 children with medical consultation with pediatric specialists at Mt Sinai, assisted by staff nurses at the Clinic. It is important to note that this access to medical specialization is delivering a health care service that would not otherwise be available to children at the Clinic. This is one of the half dozen or so operating systems in the US which provide both the patient and the doctor with a fu-ly interactive, social consultation relationship. While the service occupies two channels of a separate dedicated cable installed for this purpose by TelePrompTer when their regular cable television net was installed, it is used in the typical closed-circuit fashion as separate trunk line uninterrupted by subscriber taps. This dual trunk feature is implicit in most institutional network capability as a multiple trunk cable shadow net (MT&A recommends six trunks for Roosevelt Island educational uses). Undoubtedly the present coordination of the planning of educational, cultural resources, medical, and other institutions who may share their use of various telecommunications modes will have an effect on future delivery of services and their costs.

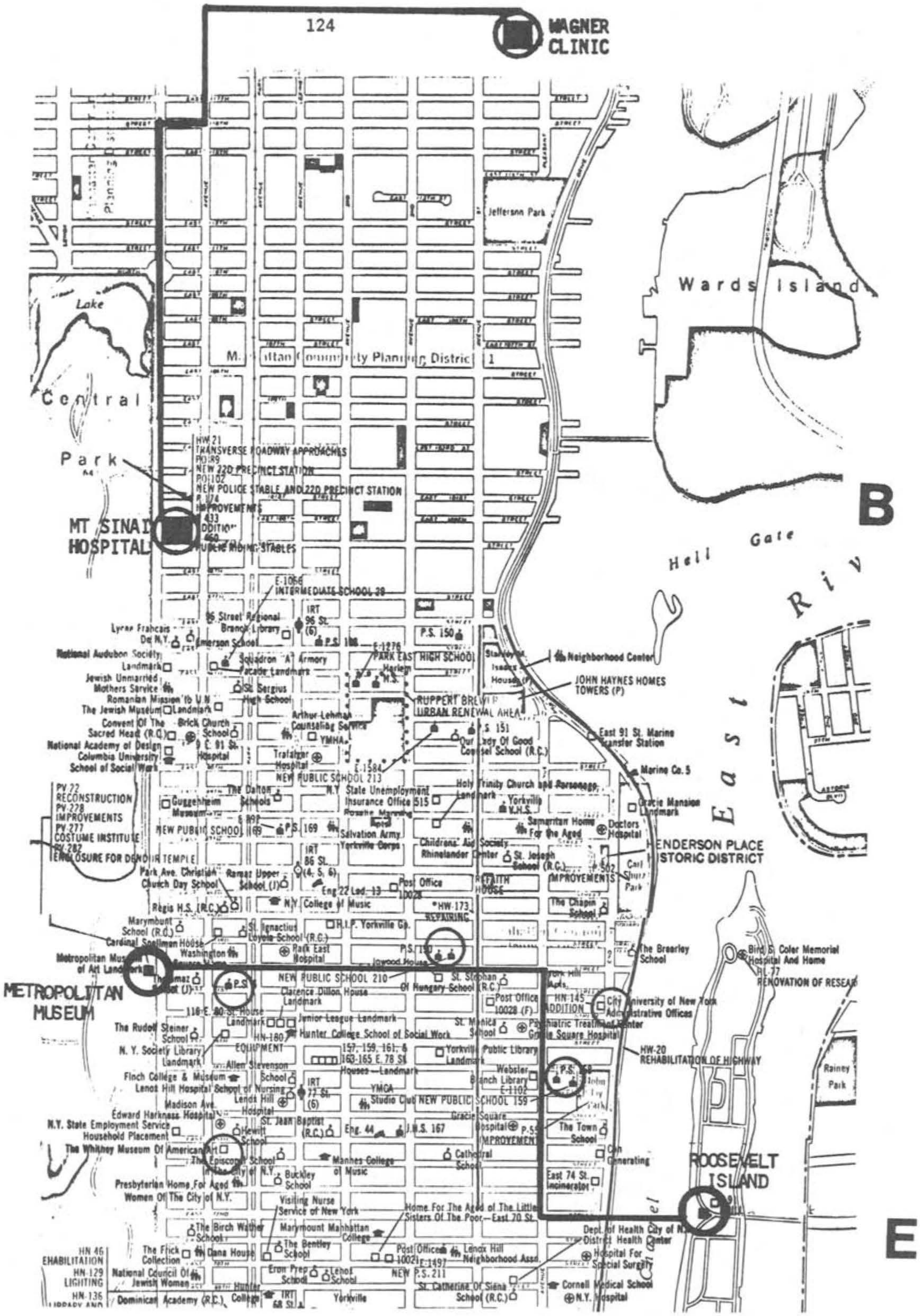
A utilization similar to the aforementioned Mt Sinai telemedical link is being proposed for another nearby area by the Northwest Queens Task Force for Health Services which would use a two-way video link between the Astoria Health Station and Elmhurst City Hospital as a demonstration for savings in time and cost on providing effective response to emergency medical needs. The Task Force reports (footnote 6) that: "...The manner in which emergency ambulances are used affects their ability to provide prompt life saving service to people in need of emergency care. ...By freeing nonurgent cases from the need to use emergency services,-the emergency medical services can then respond to emergent cases more rapidly. ...We propose to demonstrate that effective quality health and medical care can be rendered by physician assistants and nurse practitioners in a community-based environment with proper medical supportive services via telecommunications links. ...Moreover, as more experience is gained from this project, information on cost and performance will serve to develop guidelines for establishing other mini-telemedical centers in New York City and other urban communities."

This demonstration would utilize a moderately inexpensive directional microwave link prepared for that purpose between the two facilities. A more elaborate, shared utilization might become more economical if several institutions that would be passed by an alternative cable link (figure C) were to coordinate their mutual telecommunications needs. It is noteworthy that one such major institution, LaGuardia Community College (which is developing a health/media program) is preparing an architectural study (footnote 7) of the former Army Pictorial Studio under a grant by the Educational Facilities Laboratory to determine the feasibility of renovation of this significant studio complex for multi-use purposes, including community media and CATV, by the College. It is also noteworthy that Mt Sinai Hospital is the back-up teaching hospital for Elmhurst City Hospital, and comprises the Mt Sinai School of Medicine of the City University, and thus a logical CUMBIN (City University Mutual Benefit Instructional Network--an interactive television network; reference 8) participant along with LaGuardia.

HUMANISTIC USES

The two municipal affiliated hospitals to remain on Roosevelt Island, Coler and Goldwater, with their community of 2000 immobile residents with chronic disabilities, present a striking challenge to any thoughtful utilization of the enormous potential capacities of broadband communications. Specialized health programming and uses suitable for wide cable distribution include channels with sign language and subtitles for the deaf, multiple audio channels of talking-book service for the blind, two-way video correspondence for the confined, and hospital life-support monitoring systems for separate residential quarters.

A surprising and rewarding result of the Overland Park, Kansas experiment of two-way interactive video between homebound children and their school class was the emotional stimulation and motivational reenforcement shared between the


124
WAGNER CLINIC
Wards Island
Central Park
MT SINAI HOSPITAL
B
Hell Gate
East River
METROPOLITAN MUSEUM
HENDERSON PLACE HISTORIC DISTRICT
ROOSEVELT ISLAND
E

children. Thus, it is clear that the opportunity to significantly improve the quality of life through the leverage of greatly enhanced communications for immobile residents should not be foregone without an intensive assessment of the means for implementing such programs via the cable communications system of Roosevelt Island

OFFICE VIDEO CONFERENCING

Advanced applications of video conferencing have been demonstrated by the Joint Unit for Planning Research in London (British Post Office sponsored), by the New Rural Society in Connecticut (HUD sponsored), and are being operated by the Metropolitan Regional Council in the New York tri-state area (NSF sponsored; figure D; footnote 9). For $14,000 per location, a simultaneous two-way audio/video teleconferencing can be established between the 200 employees of the Urban Development Corporation central offices, if they move to Roosevelt Island as planned, and their several regional offices in the New York Metropolitan area through the established and operating microwave network of the Metropolitan Regional Council. This network provides interconnected video conferencing among some 20-30 of the 600 governmental agencies in the NY/NJ/CT tri-state area.

For the forseeable future, MRC would be able to confirm the availability of at least one to possibly several hours per day for UDC internal video conferencing use. The normal switching capabilities of the MRC operating would provide the flexibility of Regional to & from Central, Regional to & from Regional, and likewise between participating MRC cities. A future link with Albany has been considered, providing UDC with potèntial access to its additional branch offices. With more elaboration and utilizing cable links, this operation could serve as a prototypical demonstration of a switched governmental communications channel among central offices, Borough headquarters, and satellite departments. Each additional microwave location would be another $14,000 for two b/w monitors, two cameras and one mixing console. No additional cost is necessary if the office is within line-of-sight of the World Trade Center (MRC) facility. Origination and terminal equipment can be user operated.

DELIVERY OF CULTURAL RESOURCES

The New York State Education Department has proposed a $338,500 demonstration of on-demand cultural resource video programming between the Metropolitan Museum of Art and School Board District #2 (footnote 10) that would provide 10 wired origination galleries at the Met with connections to 10 classrooms in separate schools with video reception and audio feedback. Color origination capability for any particular location, including other participating institutions, would start at $15,000, including all ancillary hardware. Classrooms would use an ordinary color TV receiver, a cable system connection and some audio feedback feature. A dedicated cable connection with 10+ channel capacity might run to an additional $100,000 at most, and serve a half dozen schools in the Met's vicinity, as well as extend to the new community school facilities on Roosevelt Island (figure E).

In addition, Lincoln Center, in close proximity to the Sterling cable system headend, has an education program in the performing arts of music, dance, opera, film and drama supported by the Board of Education which reaches 300,000 students in schools annually and hosts another 30,000 at the Center. Their Workshop intensively works with an additional 270 students. Through the reach of cable video and utilizing audio feedback participation, the feasibility of serving additional remote workshops on the Island could be explored.

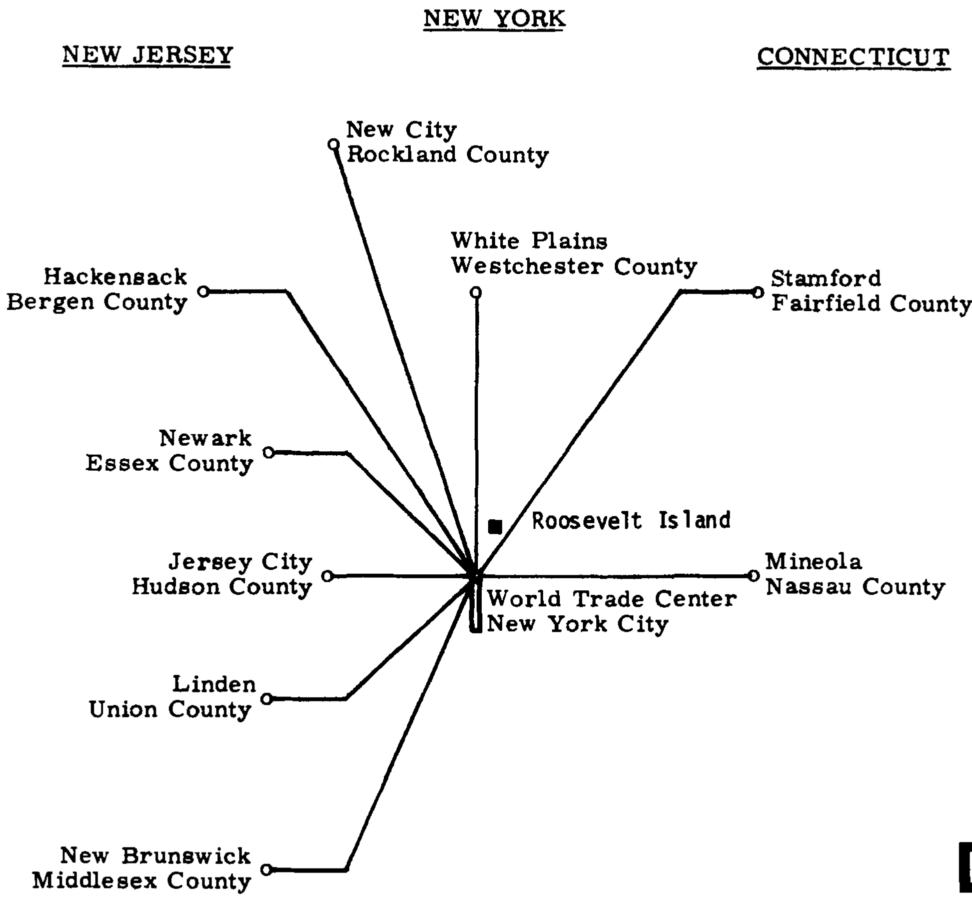
The efficiency of reaching many remote sites with their larger groups--containing more potential users--is clear. Perhaps of even greater relevance, exploratory considerations undertaken by NYS Commission on Cultural Resources indicate that the implications for economically enriching the lives of isolated individuals in hospitals (2000 on Roosevelt), prisons, and the homebound sick and elderly are obvious. However, the planning for trunk networking that would allow these user institutions to economically share headend access as well as provide additional capacity for intercommunication of resources among themselves is an important issue that is not currently being addressed. Even so, quite an array of private and public funding sources are active in this area of technology assessment, and the modest amounts of funds necessary are available, or are within marginal limits of conventional City budgetary considerations.

COMMUNITY EXCHANGE

A Community Exchange can be shaped around various components, such as an "Everything for Everybody" register, or the Island's own indigeious form of TICCIT, being used in Reston, and a CUIC: Citizens' Urban Information Center format such as being demonstrated by the Brooklyn Public Library under a Federal grant (footnote 11), and by the emerging national informations networks of NTIS, ERIC, Pandex, PLATO IV, and the NY Times--wherein the library component serves as the institutional agent for access to the information system and then economically redistributes the responses over its own cable channels, each with one video or 120 audio or 1000 digital display channel capacity, into the school or community facility or residence requesting them.

Brigitte Kenney of the Drexel University School of Library Science comments that the role of the library can become that of a community


ELMHURST HOSPITAL
LAGUARDIA COLLEGE
St. Michael's Cemetery
Sunnyside yard
Physicians Hospital
Boulevard Hospital
Astoria General Hospital
Queens Vocational H.S.
Aviation H.S.
Long Island City H.S.



D

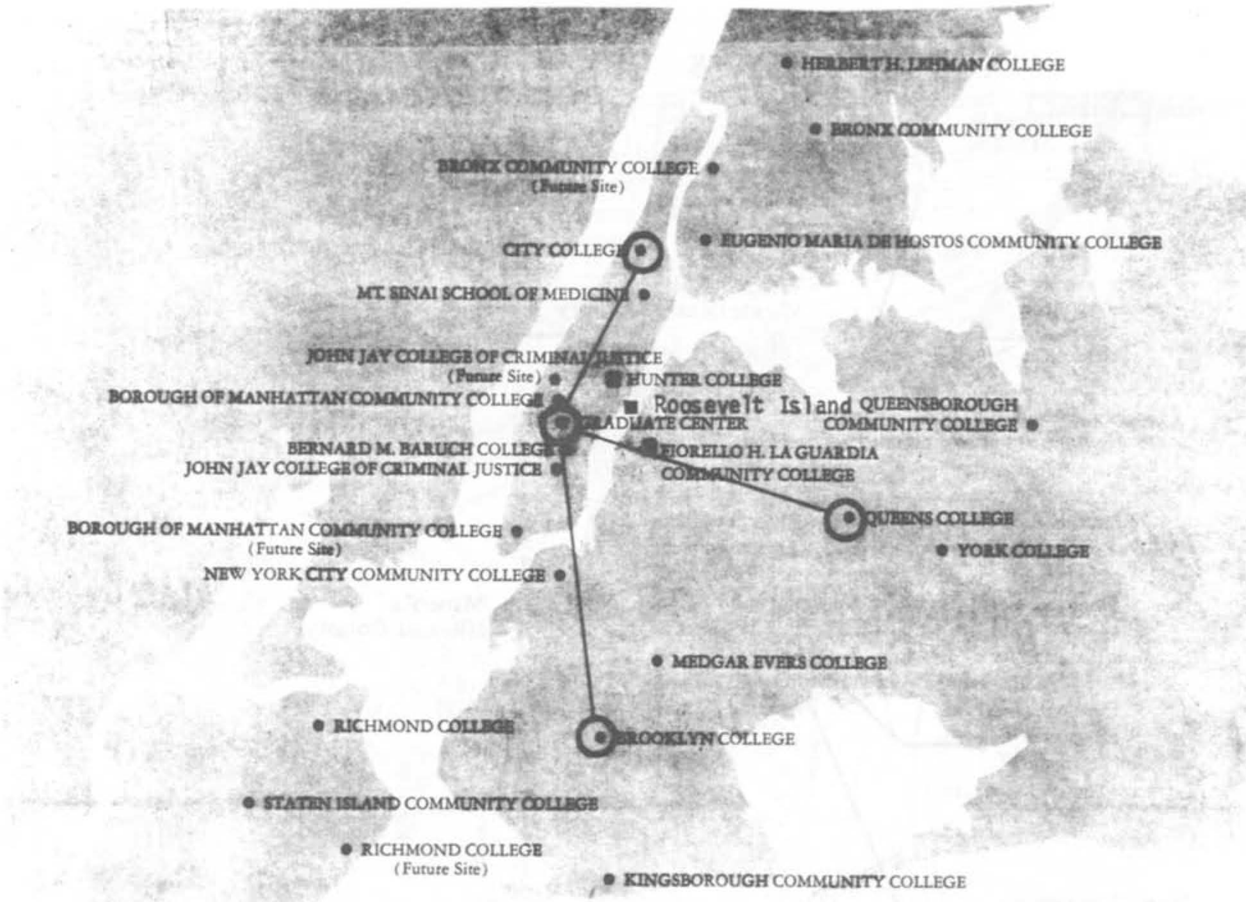
MRC-TV
Interactive Microwave Television Network
Stage One: Ten Sites

information catalyst by assuming a back-up function, an in-depth information service for other groups and agencies, which only the library can provide. "...The inevitable new structure can be envisioned as a series of specialized networks, each staffed by information mediators who know where answers to questions may be found and who can formulate both. It is a people oriented and people operated series of networks: telecommunications links, computers, and a vareity of terminals, including the home television set connected by cable to the network, are used to speed up the information delivery process and to send the information where it is needed--directly without passing through various levels of hierarchy. When this kind of information service develops, society might well feel the considerable impact of it. Information is power, and in our age of information overload, information must be effectively available to the people."

LEARNING SYSTEM

A learning system (footnote 12) presently being considered by another School District involves the use of 96 audio channels to be used by a language arts diagnostic center to screen 400 students at a time of the 26,000 in the District. It would be the first building block of a centralized instructional distribution system of supportive services for that School District's particular educational program, and which facilitates individualized review and evaluation of a student's performance. With the assistance of parallel programmed texts, this propietary system is intended to be particularly useful in determining the level of bi-lingual or non-readers.

The 96 audo channels are digitally multiplexed and modulated for a standard 6MHz video channel that can be carried over a school's internal MATV, and between schools over a properly dedicated channel of the CATV. Each of the 96 diagnostic program channels can



CUMBIN:
City University Mutual
Benefit Instructional Network

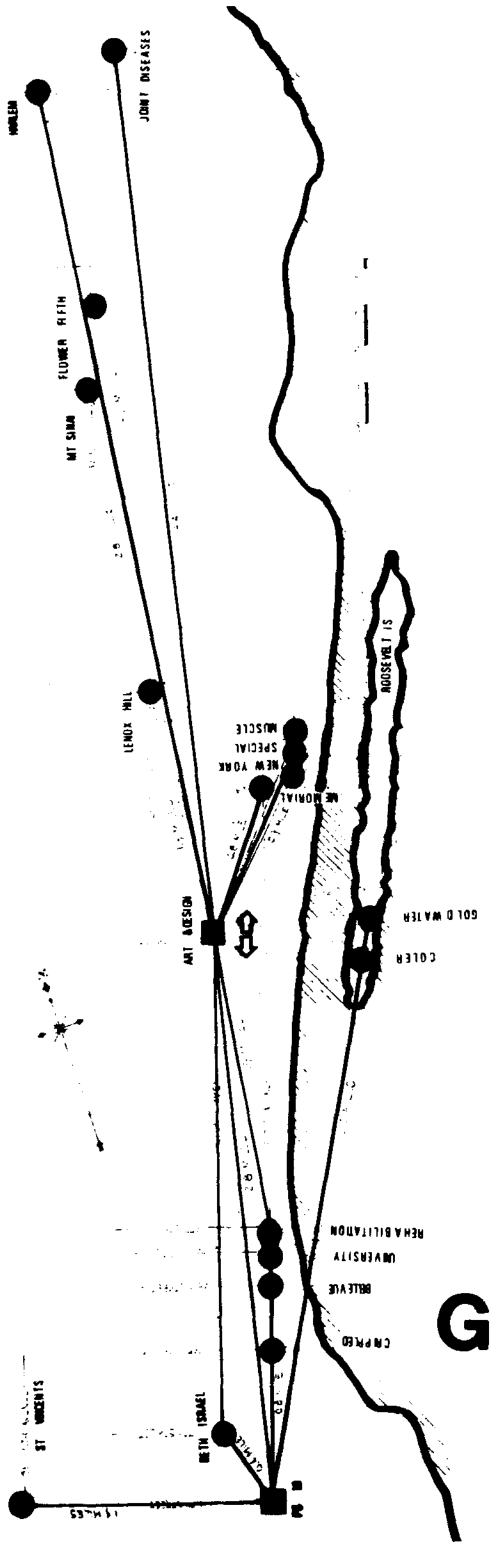
F

be selectively accessed by any remote terminal provided with a switch and earphones.

REMOTE WORK-STUDY

LaGuardia Community College, in adjacent Long Island City, will resubmit a grand application for telecommunications research (footnote 13), which seeks to evolve a model, whereby community colleges with work-study programs can participate in a consortia of educational institutions, community agencies, business and industry in applying telecommunications capability for a mutual exchange of information and resources, and deliver instruction and training for supportive and technical jobs where skilled people are in short supply. The research aims at development of a cooperative educational and human resources policy system--which is specifically addressed to the needs of urban ethnic, minority and special groups; it should also be applicable to instructional systems generally.

This model, taking advantage of the unique opportunities offered through LaGuardia's cooperative educational work-study program set in an industrial urban community, can result in applications which improve the imperfect linkage among the college's outreach to learners, feeder schools, families and employers; give the learners realistic preparation and experience for the job market; and facilitate training and placement services with feedback on the quality of work and training in progress, with the expectation that students learn more, become more career effective and be better integrated into the workworld. The model would build on alternative media approaches, with assessment of advantages and tradeoffs. It would design demonstration experiments between the College and cooperating organizations, using telecommunications channels now feasible in Northwest Queens (and Roosevelt Island), to test the objectives and to develop RFP guidelines in anticipation of 1977 broadband cable television availability in the Northwest Queens area of service.



PROPOSED:
HOSPITAL SCHOOLS
TELEVISION NETWORK

Parenthetically, a dedicated cable or directional microwave linkage between Hunter College and Roosevelt Island and LaGuardia College would give two-way access (with a new link from Hunter) to the City University's interactive video network (CUMBIN; figure F) at all three locations.

ISSUES

Generally, th educational establishment of Roosevelt Island, characterized by its scattered sites, can measurably benefit from the television feeds and the telecommunications paths for interchange, sharing, and utilization of mutually generated resources. By full utilization of these electronic pathways within the cable television system, the Community Exchange will, in large measure, be transported into every apartment and facility in the community, providing a new social plasma that will become more prevalent in our evolving society.

Specifically, in telecommunications planning there appear to be three fundamentals to be ascertained: capacity, format (media mix) and networking. However, more specifically, at the present state of considerations, the local Island issues of desired capacity for educational programmatic needs, provision of ductwork in the original construction, and universal CATV wireup are not sufficiently resolved to insure full realization of potential communication benefits.

Moreover, these items serve to further illustrate several of the generic issues in the City of the social benefit and economic considerations of universal wire-up in housing projects, and adoption of ductwork standards by HDA for new construction and renovation, and desired additional neighborhood cable capacity based upon local educational programmatic needs.

In addition, the issues of interfacing CUMBIN and other resources with the CATV metworks will include the configuration of multiple trunk shadow networking, patterns of shared utilization with other institutions, quasi-public ownership and lease-back alternatives, capital budget reallocations, and revised locational considerations.

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PHASE PHIDDLING

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HISTORICAL

The system of harmonically related carriers (HRC) used on the St. Catharines system is based on studies begun in early 1970. A paper presented at the June, 1970 convention of the National Cable Television Association, in Chicago, reported on applications of phase lock loop (PLL) technology in cable television. Maclean-Hunter had successfully used PLL techniques to overcome some "direct pick-up" problems in CATV systems. The paper reported these results and speculated on further applications of PLL in CATV. One of the applications considered was the third order intermodulation problem (triple beat) in CATV systems. The paper suggested that PLL techniques could be used to generate carrier systems with uniform spacings. Third order intermodulation products would then be "zero beat" and their visibility would be significantly reduced. Not much was known about the problem at the time and the speculative discussion in the 1970 paper erred on the conservative side in assessing the importance of third order intermodulation in CATV systems. Third order intermodulation has since been shown to be more important in CATV system picture degradation than was suspected in the 1970 paper.

By the summer of 1971 these concepts had been developed at Maclean-Hunter into a more comprehensive system. The importance of third order intermodulation was now more widely and better understood. A paper making firm proposals for coherent carrier systems for CATV was prepared and read to the IEEE Broadcast Group Symposium at Washington, D. C. in September of 1971. This paper proposed coherent carrier systems and concluded with a recommendation for a "Complete System" using harmonically related carriers. This paper, although read to the Washington symposium, was not published and was considered the draft for a more detailed paper which was presented to the National Cable Television Association at their May, 1971, convention in Chicago. At this time there was still no practical experience with coherent systems.

The company was first able to experiment with a set of harmonically related carriers in the Fall of 1972. A twenty channel set of HRC modulators had been built for use as the "base band" in an experimental wide band, multi-channel microwave system. Carriers ranged from 6 MHz to 120 MHz. Laboratory tests with this set of TV carriers confirmed expectations with respect to suppression of visibility of intermodulation beats but raised some questions about the amount of cross modulation in an HRC system. This led to the writing of an unpublished paper "SUPERPOSITION OF COHERENT CARRIERS" in the fall of 1972. This paper suggested that prudent control of the relative phase of coherent carriers at the head end would result in reduction of overall distortion in a broad band CATV system. The harmonically related carriers were considered as the solution of a Fourier analysis problem. The repetitive waveform resulting from the superposition of these harmonically related carriers could have high peak-to-peak excursions of amplitude or the peak-to-peak amplitude could be reduced even though the RMS value of the composite repetitive waveform remained constant.

An attempt was made to obtain an analytical expression for optimum phase relationships for minimum peak-to-peak amplitude. Assistance was obtained from mathematicians at Rand Corporation. No analytical solution could be found. Literature searches subsequently confirmed the probability that analytical expressions were probably too difficult. One of the Rand computers was programmed to explore optimum phase relationships and some usefull tabulations were obtained.

At this time Maclean-Hunter had formed a partnership with California interests to establish a company to design and manufacture equipment for CATV "head-end" application. Both conventional and "coherent" type products were developed and marketed.

By September 1973 about twelve HRC systems had been built and installed in CATV systems in the United States. These installations were made to facilitate PAY-TV experiments which needed mid-band transmission capacity in CATV systems which had excessive second order distortion. Use of HRC head-ends enabled use of mid-band channels without time consuming and costly system amplifier replacement. Maclean-Hunter felt that an experiment under Canadian conditions and control was desirable and consequently installed an HRC system at St. Catharines, Ontario, beginning HRC operation on August 28th, 1973.

This paper reports experience with the St. Catharines system.

EXPERIENCE

The St. Catharines system at this time is carrying twenty television channels and two pilot carriers (Table 1).

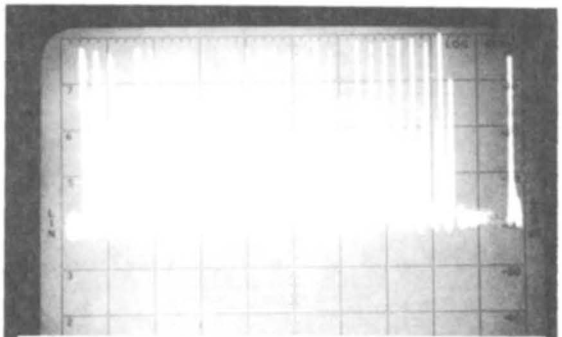


FIGURE I Spectrum Analyzer Display
Head-End 20MHz/div BW 300 KHz

The visual carriers and pilot carriers are derived by phase-locking to the selected component of a harmonic comb generated by a 6 MHz pulse generator. The pulse generator is driven by a crystal controlled oscillator.

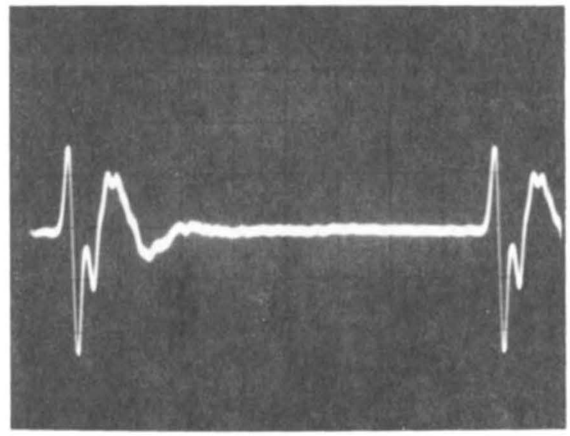


FIGURE 2

6 MHz pulse displayed on HP-183 oscilloscope with 250 MHz vertical amplifier

.02 usec/div 0.1 volts/div

TABLE 1

Channel	Visual Carrier MHz	Harmonic (N)	Nominal Deviation from Broadcast Freq.
2	54.000	9	- 1.25 MHz
3	60.000	10	- 1.25
4	66.000	11	- 1.25
PILOT	72.000	12	
5	78.000	13	+ .75
6	84.000	14	+ .75
A	120.000	20	
B	126.000	21	
C	132.000	22	
D	138.000	23	
E	144.000	24	
F	150.000	25	
G	156.000	26	
H	162.000	27	
7	174.000	29	- 1.25
8	180.000	30	- 1.25
9	186.000	31	- 1.25
10	192.000	32	- 1.25
11	198.000	33	- 1.25
12	204.000	34	- 1.25
13	210.000	35	- 1.25
PILOT	240.000	40	

\+ FM radio carriers

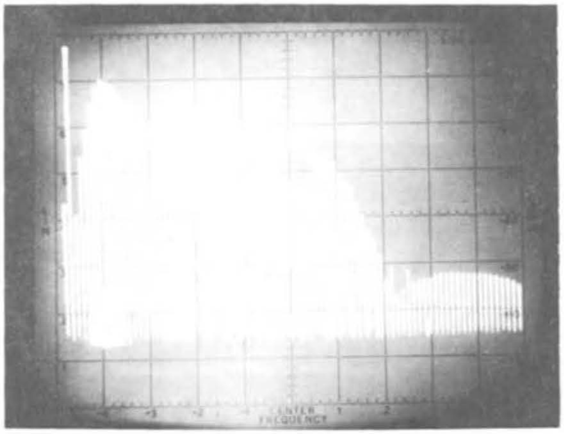


FIGURE 3

Spectrum of 6MHz pulse 50 MHz/div BW 100KHz
level ref + 42 dbmv

The "coherency" of the carriers can be observed by displaying the "synchronizer pulse", i.e. the controlling 6MHz pulse and individual system carriers on a suitable dual channel, high frequency oscilloscope. The following photographs taken on an H-P 183 oscilloscope with dual 250MHz vertical channels compare individual carriers with the synchronizing pulse. All displays were triggered by the 6MHz synchronizing pulse.

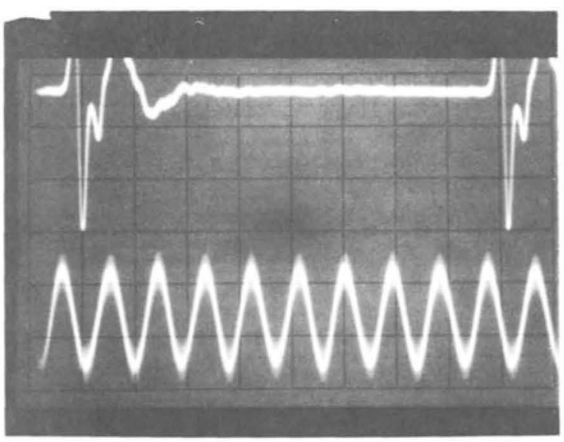


FIGURE 4

HRC Channel 2 .02 usec/div

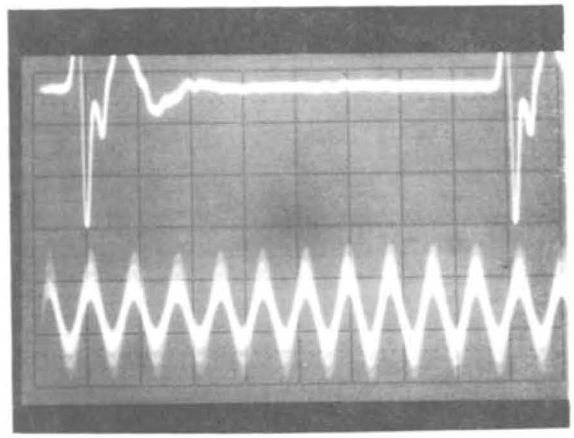


FIGURE 5

HRC Channel 3 .02 usec/div

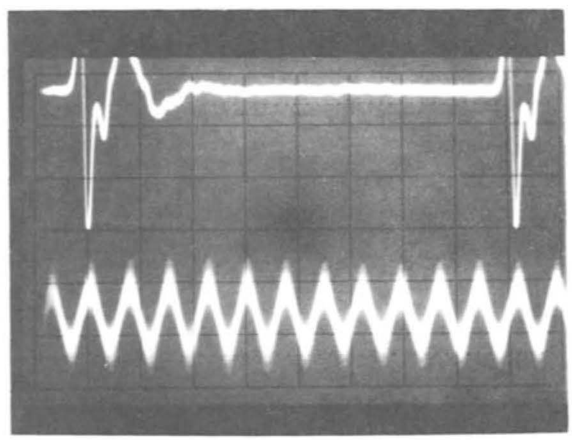


FIGURE 6

HRC Channel 4 0.02 usec/div

As expected, the display of HRC channel 2 shows nine complete cycles of RF carrier in each synchronizing pulse period. Channel 3 carrier shows 10 complete carrier cycles. Similarly channels 4, 72MHz pilot and channel F show N complete carrier cycles, N being the channel harmonic number. Amplitude "jitter" in the carrier traces is due to the amplitude modulation on the carriers.

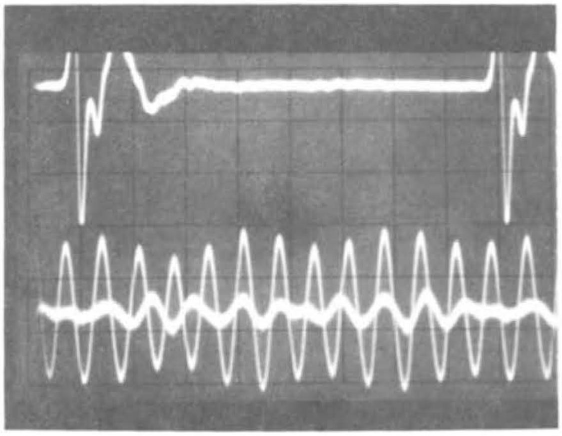


FIGURE 7

HRC Pilot 72 MHz 0.02 usec/div

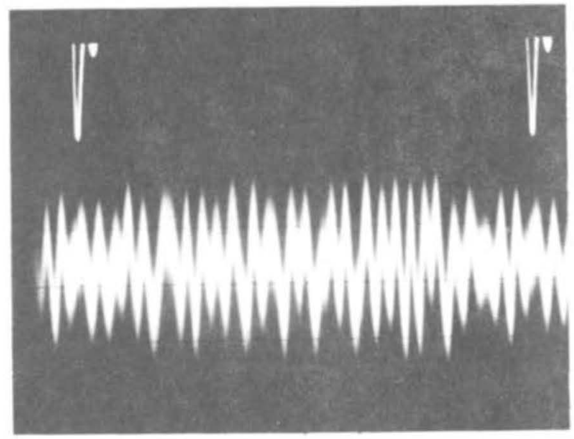


FIGURE 9

0.02 usec/div

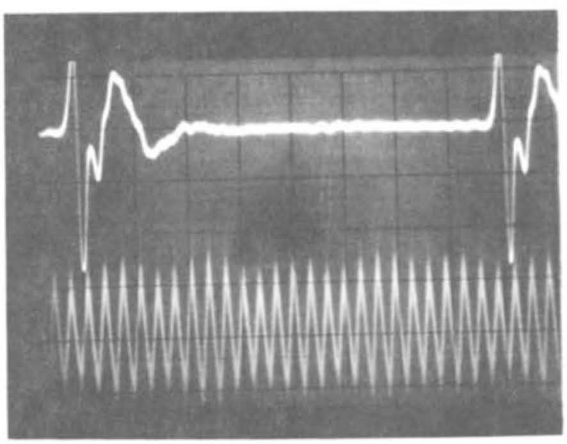


FIGURE 8

HRC Channel F 0.02 usec/div

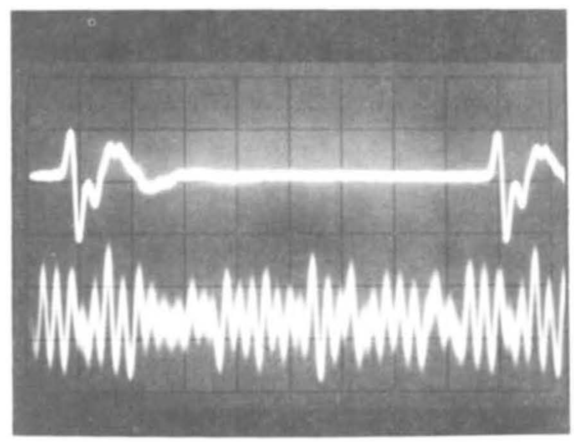


FIGURE 10

0.02 usec/div

Figures 9, 10 and 11 show the composite output of the head end as observed at the head end test point, including all twenty picture channels, two pilot carriers (coherent) and a number (approximately fifteen) FM radio channels. Figure 9 shows a nearly optimum phase relationship between visual carriers. "Peaks" and "valleys" seem evenly distributed within the period of the 6MHz synchronizing pulse. Figures 10 and 11 show less optimum phase relationships between carriers. A more detailed discussion of experiments with the relative phase of carriers will be found later in this report.

PROBLEMS

The major problem that was anticipated was that of expecting receivers to tune the new HRC channels without the use of set-top converters. It was, of course, expected that set-top converters would be used to tune the supplemental channels. Table 1 lists the carrier frequencies used. Most broadcast channels were shifted a nominal 1.25MHz lower. Channels 5 and 6 were shifted 0.75 MHz high.

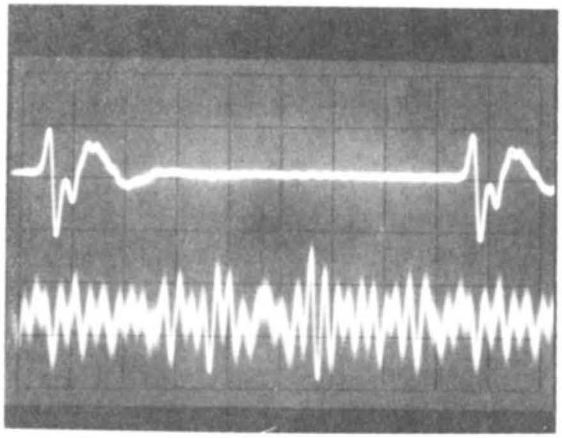


FIGURE 11

0.02 usec/div

Experience in American systems that had converted to HRC indicated that about 20% of subscribers would be unable to readjust their own fine tuners and would require assistance from the cable company. American experience also indicated that only about 1% of all receivers would lack sufficient fine tuning range to accomodate the new carriers. The changes were extensively advertised in the newspapers and in a special mailing to all subscribers with detailed explanations on the fine tuner changes required. A videotape showing a typical tuner arrangement was shown frequently on the community service channel (channel 8) on cable.

Switch-over to HRC channels on August 28th, 1973, still required a very large number of service calls. Additional staff had been brought in from other CATV systems and staff worked overtime to make service calls to help subscribers readjust their fine tuners. Set-top converters were provided on a loan basis for those subscribers whose receivers would not retune to the new channels. Most receivers had sufficient fine tuning range. Some receivers required adjustment of fine tuner range with a tuning tool through the front of the receiver. A very small number (less than 100) of receivers were readjusted in this way. It should be borne in mind that the St. Catharines system has about 14,000 subscribers.

The direct-pick-up problem was seriously underestimated. Some problem had been anticipated since St. Catharines is only about 30 miles from a maximum parameter station on channel 11+. This proximity had caused some marginal direct pick-up problems with conventional operation but the situation had not been considered serious. With HRC operation the direct pick-up was no longer coherent but was 1.26MHz into the HRC channel, i.e. 1.260MHz above the HRC-11 carrier. We have since determined that this beat has a threshold about 15 db less favourable than the coherent beat (worst case). After most subscribers receivers had been retuned for HRC channels the direct pick-up problem remained the most serious problem. A very few receivers, particularly succeptible to direct pick-up, showed some direct pick-up beat on other channels as well, notably on channels 7 and 9, both about 40 - 45 miles distant. This number was very small and has not been considered significant at all.

The initial approach to the direct pick-up problem on channel 11 has been to "unlock" HRC channel 11 and shift it slightly to take advantage of the "interlace" provided by "channel offset". The visibility of the direct pick-up beat was reduced about 10 db my moving HRC-11 so that the pick-up beat fell halfway between H sidebands. The 1.260MHz beat falls just above the 80th H sideband of the HRC-11 carrier. The HRC carrier was shifted slightly downward so that the beat fell halfway between the 80th and 81st H sidebands. This required a shift of about 6.6KHz. The adjustment was made visually for minimum visibility of beat. At first this was not completely effective since there was considerable direct leakage of channel 11 into the cable system. Sources of leakage were found and corrected and this leakage has been held to a minimum. Figure 12 shows the extent of this leakage and shows the offset of HRC-11.

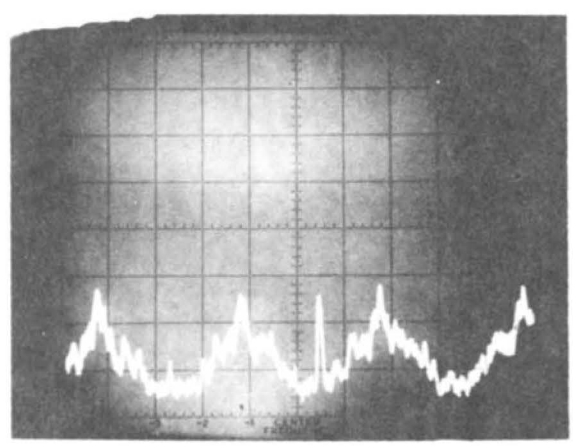


FIGURE 12

Leakage of channel 11 into system 5KHz/div BW 300Hz video filter 10Hz level reference HRC-11 carrier

This spectrum analyzer photograph was taken at the cable system office. The offset is only approximate due to drift of the "unlocked" HRC-11 carrier. The leakage into the system at this point is about 54 db below desired carrier level. HRC-11 frequency was subsequently adjusted to more optimum "mid-point".

The "unlocked" operation of HRC-11 causes a few undesired beats in the system. Figure 13 shows HRC-4 observed at the same location showing beats due to the HRC-11 offset. These beats disappear when HRC-11 is locked to the synchronizing comb.

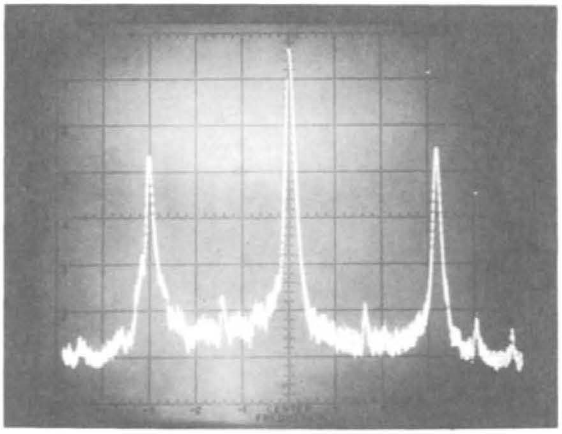


FIGURE 13

HRC-4 showing beats from "unlocked" HRC-11
5KHz/div BW 300Hz video filter 10Hz

At the time of the installation of the St. Catharines HRC system we had devised a solution for the direct pick-up problem from channel 11 which involved locking the system to the channel 11 transmitter. It was not possible to get the necessary equipment designed and constructed in time for the St. Catharines test and the risk of direct pick-up had to be accepted.

A system for relieving this direct pick-up problem has been devised and is being engineered and constructed for installation in St. Catharines within the next few months: -

Limited PLL lock-in range does not allow shifting the synchronizing pulse frequency to cause the direct pick-up beat to fall midway between H scan lines. A special phase-lock loop is being built to cause the beat to fall midway between V scan lines, as follows:-

For colour television the following "standards" apply:

Colour subcarrier	=	3.579545 MHz
H = C/455 X 2	=	15734.26374 Hz
V = H/525 X 2	=	59.94005 Hz

Midway between H sidebands would require beat to be:
$80\frac{1}{2}$ H = 1.2666082 MHz

This requires a shift of about 6KHz in HRC carriers and would require new VCXO's in all signal processors. The nearest beat frequency which falls halfway between V sidebands is:

80 H + $21\frac{1}{2}$V = 1.26002981 MHz

This requires only a 30Hz shift and will not affect the present VCXO's. A phase lock loop will control the 6MHz synchronizing oscillator so that the beat between HRC-11 and the broadcast channel 11+ will always be 1.26002981MHz with a tolerance of +- 0.1Hz. Effectiveness depends on the stability of the colour synchronizing generator at the channel 11 broadcast station but this is considered to be a good quality generator. The channel 11 transmitter has been observed, however, to drift +- 40Hz within a period of a few minutes. This suggests that a precise offset operation requires locking our master oscillator so that the beat between the two carriers (broadcast 11+ and HRC-11) is maintained at a very precise frequency relative to the line scanning frequency. It is expected that this system will allow coherent operation of HRC-11 with minimum direct pick-up problems.

Alternative Technique for Controlling "Direct Pick-Up" at St. Catharines

The technique we would have preferred to use to reduce direct pick-up at St. Catharines would be to lock the HRC-11 channel to the local broadcast channel 11. The HRC carriers need not be harmonics of precisely 6.000000MHz. The fundamental frequency may vary somewhat as long as the system carriers remain harmonics of the fundamental. Variations from 6.000000MHz merely affect the spacing between channels, most importantly the spacing between lower adjacent sound and the next higher visual carrier.

In this case it would be helpfull if the fundamental frequency was adjusted so that its 33rd harmonic (HRC-11) co-incided with the local broadcast channel 11. In fact the fundamental master oscillator would be derived by locking an oscillator to the local channel 11 and then counting down, digitally, to obtain the master fundamental frequency. Since the local channel 11 is actually channel 11+, nominally 199.260MHz, the master oscillator would be 199.260/33 = 6.038182MHz. This oscillator frequency would be floating up and down as the local channel 11+ transmitter shifts in frequency around its nominal assigned frequency. Table II lists the HRC carrier frequencies in this case and their deviations from nominal broadcast frequencies.

Note that the high band channels are quite close to nominal broadcast frequencies. Channel 11 is identical and locked to a local broadcast channel. The other high band channels are within 150KHz of nominal broadcast channels, well within fine tuning and AFT ranges on virtually all receivers. The low band channels are improved for channels 2, 3 and 4, but channel 6 may have more problems since it is shifted up in frequency toward any FM traps that may be present in the RF stage of the receiver tuner.

TABLE II

Channel	Visual Carrier MHz	Harmonic (N)	Nominal Deviation from Broadcast Freq.
2	54.344	9	- 906 KHz
3	60.382	10	- 868
4	66.420	11	- 830
5	78.496	13	+ 1246
6	84.535	14	+ 1285
A	120.764	20	
B	126.802	21	
C	132.840	22	
D	138.878	23	
E	144.916	24	
F	150.955	25	
G	156.993	26	
H	163.031	27	
I	169.069	28	
7	175.107	29	- 143
8	181.145	30	- 105
9	187.184	31	- 66
10	193.222	32	- 28
11	199.260	33	0
12	205.298	34	+ 48
13	211.336	35	+ 86
J	217.375	36	
K	223.413	37	

etc........

The spacing between lower adjacent sound and desired visual carrier becomes 38KHz greater than nominal. A broadcast situation might have a - offset channel adjacent to a + offset and allowing for the 1KHz tolerances allowed in visual and intercarriers might have this spacing at 23KHz greater than nominal 1.500MHz. This 38KHz increase is an additional 15KHz. This might reduce the effectiveness of adjacent channel traps in some receivers. Our experience is that this effect is not significant and that this not a practical obstacle to implementation of this channeling variation. Many cable systems already have adjacent channel sould spacings of greater than 1.523 MHz due to inaccuracies in their carrier frequencies on cable. These inaccuracies usually occur in UHF channels being converted to VHF channels for cable distribution. Local oscillators of 600MHz or so are quite common in UHF to VHF conversions and a 30KHz error represents only .005% and many of the local oscillators used in cable systems do not meet .005% tolerance. The new FCC standards require tighter tolerance but the fact is that systems have operated with rather slack tolerances for many years, and it may turn out that .005% tolerances are difficult to achieve in UHF conversions. We deliberately introduced a 1.538MHz adjacent spacing into a pair of the channels on the Maclean-Hunter Cable system and could find no evidence of increase in adjacent channel interference problems on these two channels during a test period of several months.

The direct pick-up problem is less serious in a system which uses converters on all channels. Such a system has no practical problem in tuning HRC channels since all the set-top converters we have seen can easily be adjusted to HRC channels instead of broadcast channels. Such a system should also be immune from direct pick-up problems and operation on either broadcast or HRC channels would require maintenance of a high degree of system integrity. In such cases there would seem to be little effect on the subscriber from system use of HRC channels, and the cable system might prefer to use a system based on a precise 6.000MHz master oscillator.

SYSTEM PERFORMANCE

Rigorous system performance measurements are still being made, but some preliminary measurements are available. These measurements were made at the system office, which is about 2/3 of the way along the maximum system extension. The transportation trunk from the head end to the beginning of distribution had been replaced with push-pull amplifier equipment (20 amplifiers in cascade). The remaining route to the office consists of nine trunk amplifiers, a bridging amplifier and four line extenders, all single ended design. The distribution amplifier within the office is push-pull design, conservatively operated.

Some systems performance measurements made on January 25th: (Cable Office, 45 Wright Street)

Noise: Figure 14 shows system noise as observed at the office. Channel 3 was chosen as a convenient test channel. The Global TV network is distributed on this channel and since its programming does not start until late afternoon, the standby carrier on this channel could be used as a test carrier.

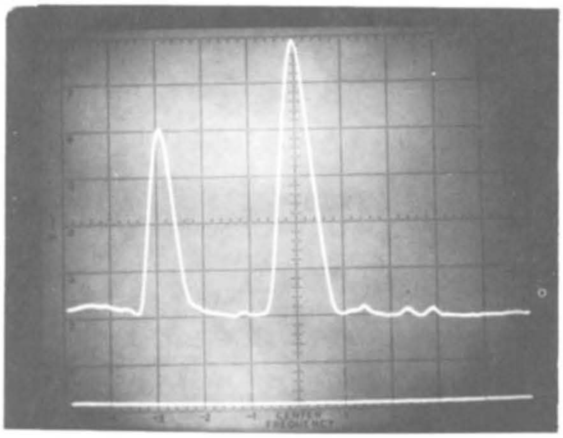


FIGURE 14

500 KHz/div BW 100KHz video filter 10Hz
Standby carrier on Channel 3 showing channel 2 sound. Double exposure shows spectrum analyzer noise level

Noise measurements were made with the spectrum analyzer. The noise bandwidth for the 100KHz IF filter had been measured using Hewlett-Packards recommended technique. Correction factor, including factor for detector and logarithmic display was calculated to be +18 db for the particular analyzer used. An IF bandwidth of 100KHz was used with a video filter of 10Hz (10,000 X) to give effective averaging of noise. Figure 15 shows the smoothness of noise display around the channel 3 standby carrier. Standby carrier had been adjusted for same peak carrier level as normally modulated carrier. This photo indicates carrier/noise ratio of 60 db displayed, correcting to 42 db for 4MHz noise bandwidth. Similar measurements at other parts of the spectrum show C/N ratios of from 44 db near channel 2 to about 41 db in mid-band to about 40 db in the high band.

Beats: Control of undesired beats is one of the main features of HRC operation. Figures 15 and 16 show the channel 3 standby carrier and the beats at this portion of the spectrum when the channel 3 standby carrier is removed from the system.

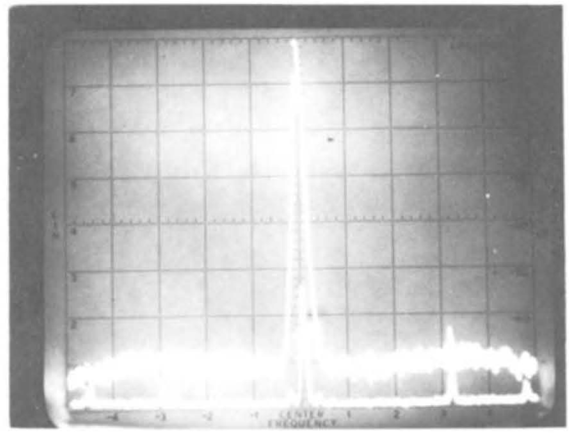


FIGURE 15

Channel 3 standby carrier (all channels coherent)
5KHz/div BW 300Hz video filter 10Hz
Double exposure showing beats present when carrier removed

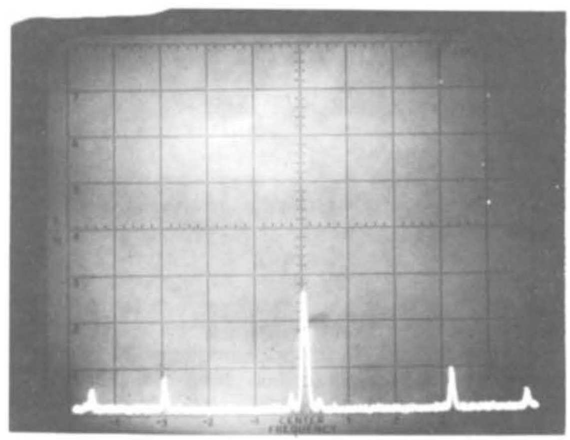


FIGURE 16

As in Figure 15 but showing beats only
(Channel 3 standby carrier removed)

The beats in Figures 15 and 16 arise from second order and third order causes. Because of the coherency of the system all the beats are coincident and only the resultant sum of all the beats is observed. Phase cancellations of beats does occur. The observed resultant beat is 54 db below desired visual level and is, of course, at visual carrier frequency. The H sidebands, which are effectively cross-modulation, can be seen only when the channel 3 carrier is removed since they are at a level just below the noise in most observations.

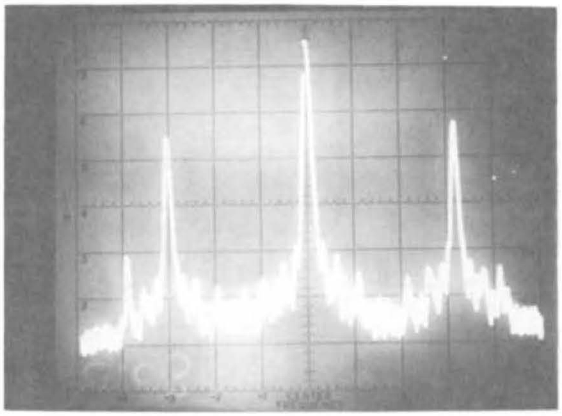


FIGURE 17

Normally modulated channel 3 (Global 6+)
5KHz/div BW 300Hz video filter 10Hz
Ref level adjusted for peak carrier showing H sidebands and -20KHz co-channel interference

Figure 17 shows that normal H sidebands are typically about -24 db relative to peak visual carrier. Observed H sidebands with carrier removed seem to be about 70 db below peak visual carrier indicating a cross modulation ratio of about 46 db. No cross modulation coult be visually observed in the pictures at this point. Figure 17 also shows an unwanted co-channel carrier (originated as channel 6-) at about 50 db below desired peak visual. This level of co-channel interference was not visible in the pictures.

Figures 18 and 19 show the beats with a different phase arrangment at the head end. Figure 20 shows still another phase arrangement. Note that the level of beat has changed due to differing phase cancellation.

The cross modulation produced H sidebands could hardly be distinguished in the channel noise when observing the standby carrier so an attempt was made to improve this observation. The first lower H sideband was selected and maximum spectrum analyzer resolving power and "signal averaging" using the oscilloscope screen storage was attempted. A tunable preselector and amplifier were used to improve the spectrum analyzer noise figure and the spectrum analyzer gain was raised 20 db from previous observations. The H sideband can just be distinguished at about -50 db on the display, indicating -70 db from peak visual carrier, or equivalent to about -46 db cross-modulation.

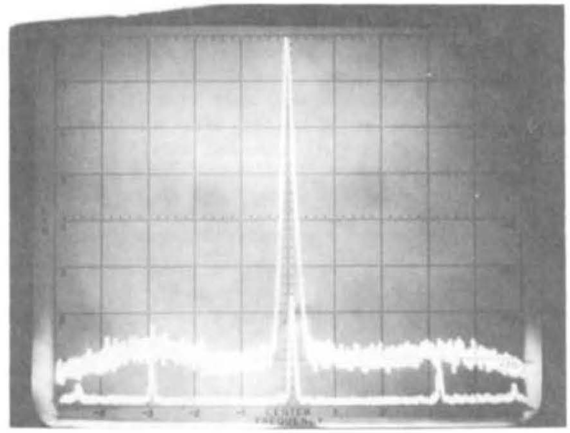


FIGURE 18

Double exposure with and without channel 3 standby
5KHz/div BW 300Hz video filter 10Hz

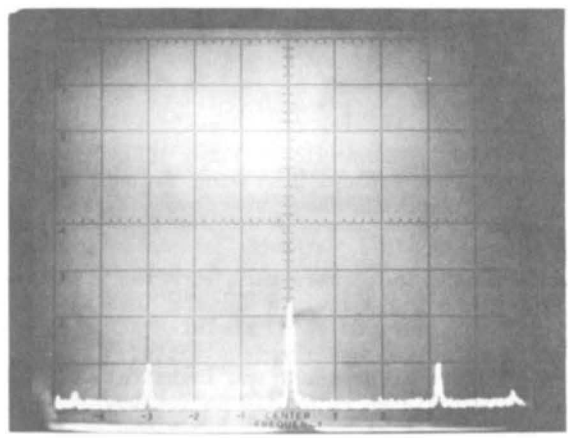


FIGURE 19

As figure 18 but single exposure - beats only

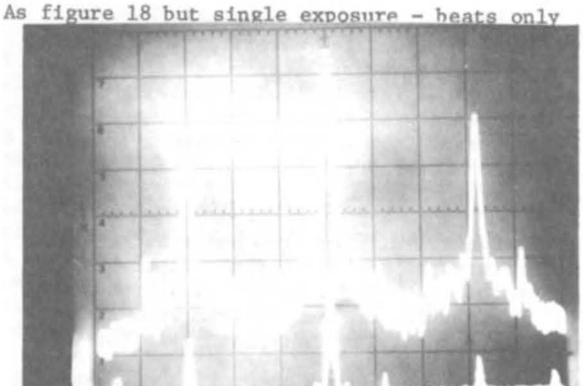


Figure 20 Double Exposure showing beats when channel 3 removed 5KHz/div BW 300Hz video filter 10Hz Carrier phases altered from Figures 18/19

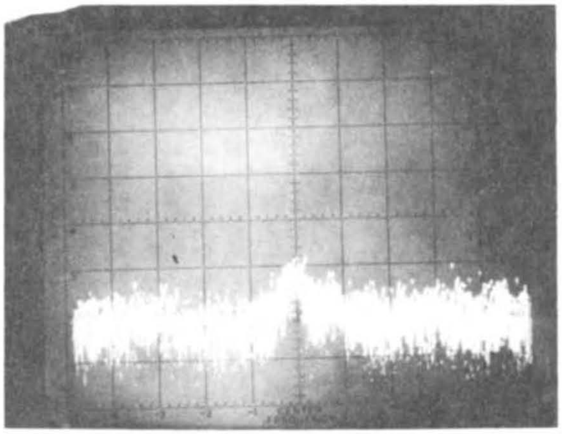


FIGURE 21

Lower H sideband channel 3 standby carrier
20Hz/div BW 10Hz video filter 10Hz
integrated by storage CRT

Intermodulation:

The systems freedom from intermodulation beats around the visual carriers was demonstrated by observing the channel 3 standby carrier with increasing resolution and decreasing dispersion. Figure 22 shows low level intermodulation products due to channel HRC-11 being non-coherent.

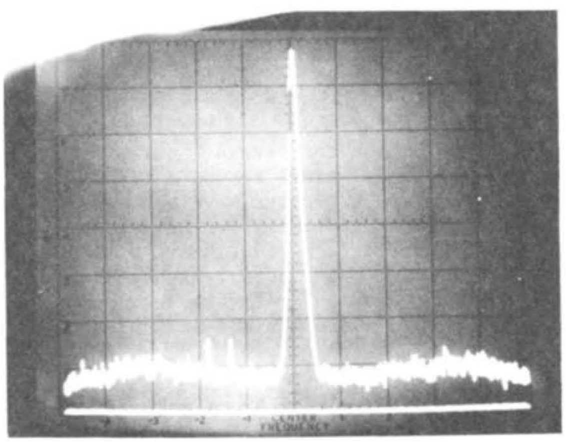


FIGURE 22

5 KHz/div BW 300Hz video filter 10Hz

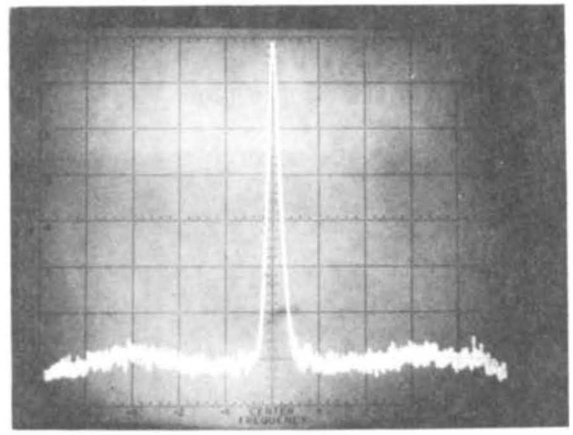


FIGURE 23

5 KHz/div BW 300Hz Video Filter 10Hz

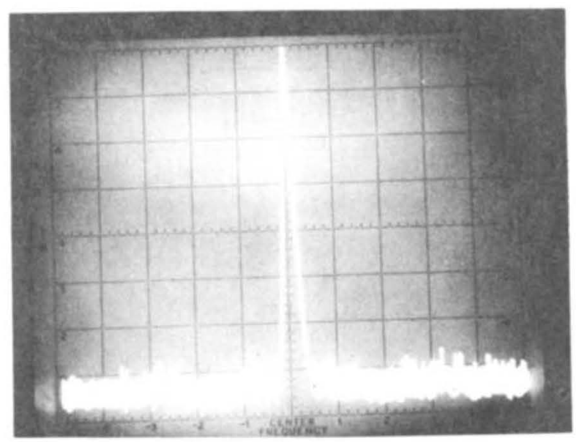


FIGURE 24

2 KHz/div BW 100Hz Video Filter 10Hz

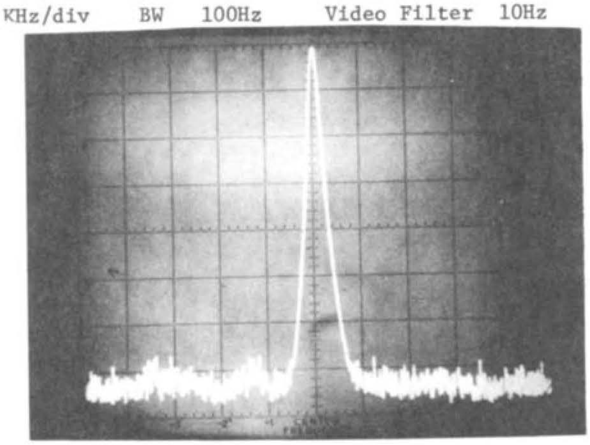


FIGURE 25

1 KHz/div BW 100 Hz Video Filter 10Hz

Locking reference (6MHz synchronizing comb) was then disconnected so that all visual carriers were non-coherent. Figure 26 shows that the channel 3 standby carrier has moved about 1.5KHz and several intermodulation beats appear.

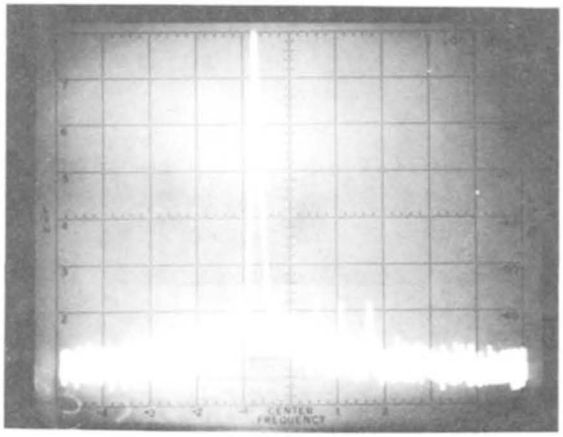


FIGURE 26

2KHz/div BW 100Hz Video Filter 10Hz

The channel 3 standby carrier was then removed and the beats examined in greater detail, Figure 27

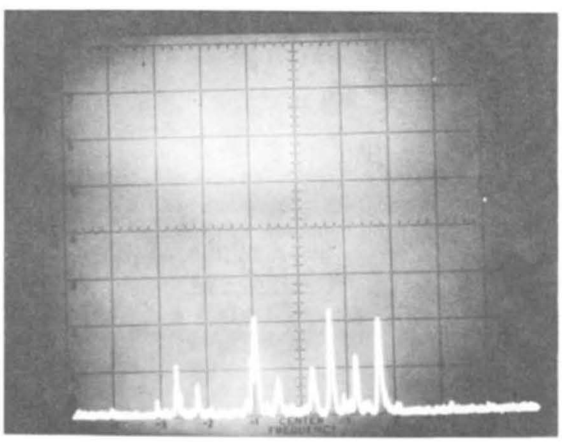


FIGURE 27

2KHz/div BW 100Hz Video Filter 10Hz

Hum Modulation:

Although not the subject of this particular discussion hum modulation was observed in terms of hum modulation sidebands on the channel 3 standby carrier. These are shown in Figure 28. Hum modulation appears to be about 1%.

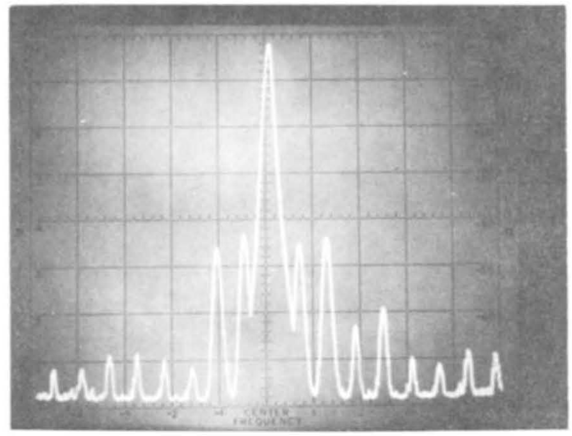


FIGURE 28

100Hz/div BW 10Hz Video Filter 10Hz

"Phase Phiddling"

A cable television system is a "very nearly linear system". The slight non-linearities give rise to intermodulation and cross modulation effects. The effect of the harmonic relationship of the carriers is to make intermodulation products zero-beat and therefore reduce the visibility of these non-linear effects. The high degree of linearity means that we can consider that the carriers in the system are "superposed", i.e. that they add algebraically in such a linear system. We can therefore consider that the simultaneous carriers can be considered as adding to give a sum waveform that can be displayed on an oscilloscope or oscillograph. In a non-coherent system the phase relationships of the carriers are quite random and the sum waveform is quite "random". There is no repetition of any particular waveform and an oscilloscope shows a "jumble". There is a high probability that voltage addition of all the carrier amplitudes could occur, although the power in the system at any point in time cannot exceed the sum of the power present in the individual carriers.

The superposition (addition) of coherent carriers can be considered to be the inverse of Fourier analysis - a sort of Fourier "synthesis". Any periodic function can be analyzed into a Fourier series - a series of sin, cos, or sin and cos terms of the fundamental and harmonics (depending on nature of symmetry of the periodic function). The physical implication is that any periodic function can be analyzed by Fourier methods into a series of harmonics with definite phase and amplitude relationships. A periodic function defined only in terms of the amplitude of its Fourier constituents is not uniquely defined. The phase relationships between the Fourier components must also be defined.

Waves which are harmonically related may be superposed (added) into a periodic waveform. The periodicity will be that of the fundamental. The exact nature of the periodic function resulting will depend on the amplitude and phase of the component waves.

The visual carriers of a cable television system may be considered to be a set of waves which are being added in a broad band distribution system. If these carriers are harmonically related they may be considered to be components of a Fourier "synthesis" which will yield a composite waveform with a period which is that of the fundamental on which the carriers are based. The resulting function will have a unique form dependant on the relative amplitudes and phases of the component carriers. The RMS value of the resulting function will depend only on the amplitudes of the component carriers and represents the addition of the power content of the component carriers. The peak amplitude of the resultant function depends critically on the phase relationships between the component carriers. Peak amplitude reached by the periodic function can be minimized by prudent selection of relative phases.

Carriers which are not coherent, i.e. not locked to a common "fundamental" frequency, will have random relative phases and will add in such a way that the sum waveform often reaches a peak which is the sum of the peak amplitudes of the component carriers. Carriers which are coherent have controlled phase relationships. Such carriers can be phased so that their additive peak amplitude is substantially less than the sum of the individual peak amplitudes.

The behaviour of an amplifier when the input is a set of coherent carriers may be described in terms of its response to the "sum" waveform. If the "sum" waveform has reduced peak amplitude excursions we can consider that the amplifier output will have reduced distortion because of reduced excursions along the amplifier transfer curve. The amplifier will always be working on a more linear part of the transfer curve rather than experiencing frequent large peak amplitude excursions along the transfer curve. We propose that the effect of this would be to allow derating of CATV amplifiers for increasing number of channels by a "power addition" law rather than by "voltage addition" law. We doubt that exact "power addition" is actually achieved but we do believe that we are achieving benefits somewhere between "power addition" and "voltage addition".

The potential benefits may be estimated from this table of "derating factors" relative to 12 channel loading, based on "power addition", "voltage addition", and a "14 log" law estimated to be about half way in between.

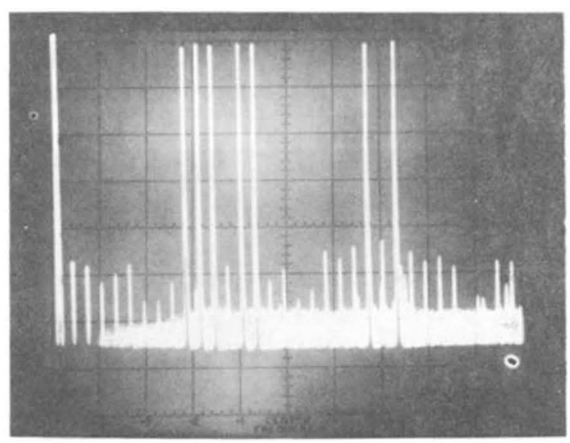
N Channels	Power Addition 10 log N/12	14 log N/12	Voltage Addition 20 log N/12
15	1.0 db	1.4 db	1.9 db
20	2.2	3.1	4.4
25	3.2	4.5	6.4
30	4.0	5.6	8.0

The effect of "phase phiddling" on distortion has been demonstrated in the laboratory and in limited field tests. The following is typical of laboratory demonstrations:-

Seven HRC carriers were generated (unmodulated) as follows:-

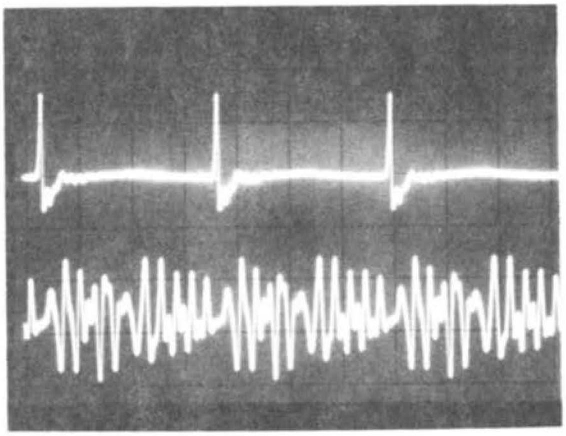
Channel	Frequency	N
HRC-2	54.000	9
-3	60.000	10
-4	66.000	11
-5	78.000	13
-6	84.000	14
-C	132.000	22
-E	144.000	24

This set of carriers was amplified by a high quality single ended MATV type amplifier to introduce a moderate level of distortion. Input signal level (flat) was adjusted so that output distortion products were typically 45 db below carriers for second order and about 55 db below carriers for third order intermodulation products.

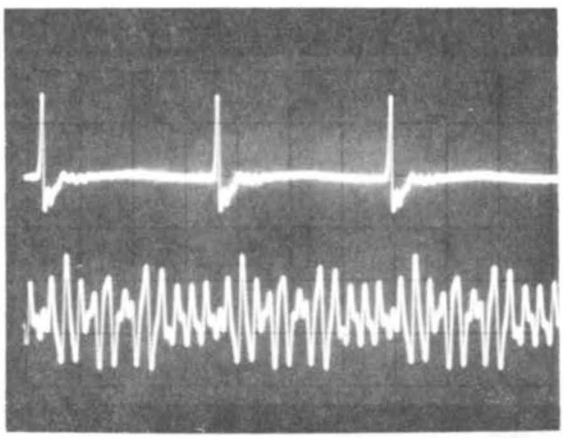


7 Channels 10 db/div 20MHz/div

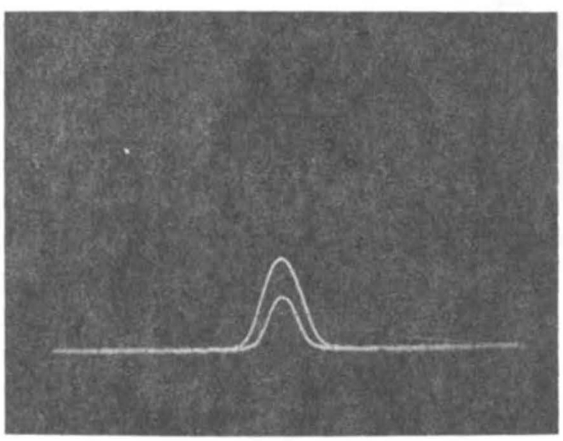
The intermodulation products at 72.000MHz were chosen for study. Channels C and 3 produce a second order product at 72.000MHz so channel C was removed to allow study of the underlying third order products.



"Sum" Channels 2,3,4,5,6,C,E .1V/div .05 usec/div



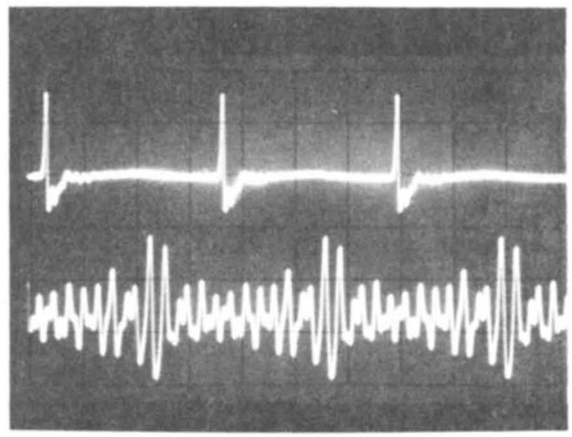
"Sum" Channels 2.3,4,5,6,E .1V/div .05 usec/div



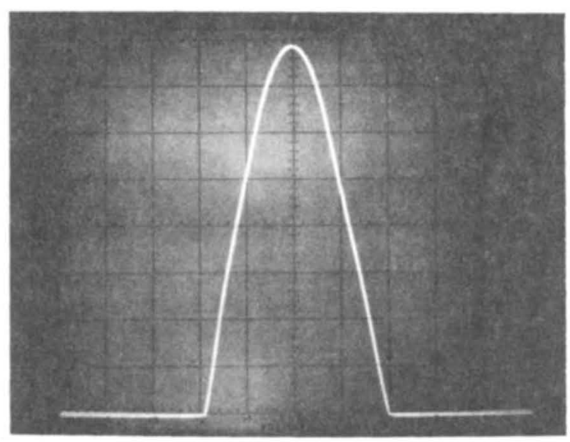
Double Exposure 20KHz/div BW 10KHz video filter 10Hz 10 db/div showing third order IM under 2nd order product

The phase of the six remaining carriers was than adjusted for maximum amplitude of the displayed third order products. We had previously found that a single distortion product (intermodulation) which is the result of only two carriers intermodulating is not sensitive to the phase of the contributing carrier. Dr. J. Shekel, of Jerrold, was kind enough to contribute an analysis to this effect and we have verified this experimentally. Removal of the single second order distortion product at 72.000MHz allowed us to observe the third order products "underneath". The observed third order product is the vector sum of seven contributing third order products.

Carrier phases were adjusted to produce maximum amplitude of the 72.000 MHz product. Corresponding sum waveform and intermod amplitude are displayed below:

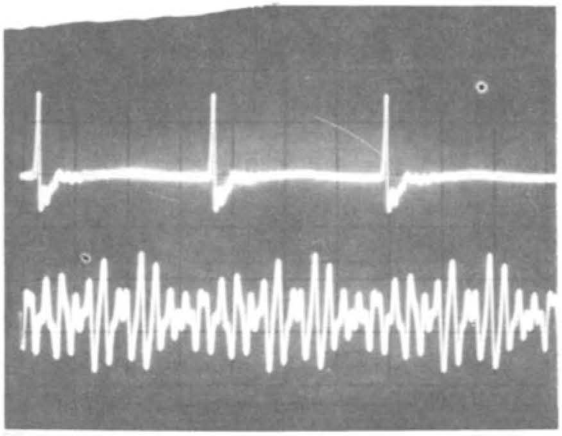


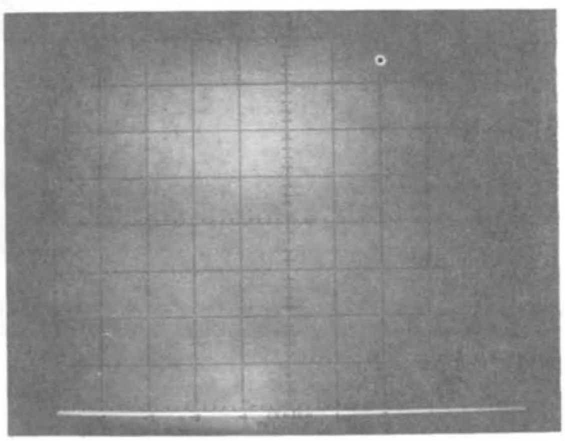
Sum waveform for maximum IM at 72MHz
0.1 v/div .05 usec/div



Amplitude of IM at 72MHz 2db/div 20KHz/div
ref - 54 dbm

Carrier phases were then adjusted for minimum IM product at 72.000 MHz. This product was reduced by more than 16 db dropping out of sight on the 2 db/division spectrum analyzer display. The sum waveform shows reduced peak to peak amplitude .





Since channel E did not contribute to the 3rd order product at 72MHz we tried removing it. Removal of E reduced the third order product by 1 db. Phase adjustments on E alone produced a similar 1 db change, whereas the phase of the other 5 channels had far greater effect on the amplitude of the IM product at 72MHz.

The intermod product at 48MHz was also observed to see whether it decreased in amplitude along with the product at 72 MHz:

	IM Product	Level	N (products)
"Worst Case"	72.000 MHz	-56 dbm	7
	48.000 MHz	-61 dbm	3
"Best Case"	72.000 MHz	-83 dbm	7
	48.000 MHz	-70 dbm	3

The 48.000MHz product decreased in amplitude but not as dramatically as the 72.000MHz product for which the adjustments were optimized.

EXPERIMENTAL PHASE ADJUSTMENTS

Some experiments with head end carrier phase adjustment were conducted in late November 1973.

A technique was developed for observing and adjusting the "relative phase" of the visual and pilot carriers at the head end. The concept of "relative phase" of carriers of differing frequency was taken to mean their relative phase at periods of the fundamental frequency from which the coherent carriers were derived. The main head end test point was designated as the visual carrier reference point and one of the "synchronizing pulse" test points was designated as another reference point. Test cables of convenient length were cut and marked to connect these test points to the dual channel high frequency oscilloscope. The oscilloscope (H-P 183 with dual channel 250MHz vertical amplifiers) was triggered by the 6MHz sync pulse. The individual channel processors were then all disconnected from the mixing networks and then reconnected one at a time for individual observation beginning with channel HRC-2. The length of the jumper cable from the channel HRC-2 processor to the mixing network was then adjusted until the first "peak" of the sync pulse coincided with a "valley" of the visual carrier. This is illustrated for the 72MHz pilot carrier in Figure 29. This was considered "zero phase". Figure 30 is an expanded version of Figure 29. In Figure 31 a half-wave length of cable has been added to the mixing jumper to show the 180 degree phase reversal case.

HRC-2 was then disconnected and HRC-3 was connected and a similar adjustment of mixing jumper made. Similarly all 21 visual carriers and the two pilot carriers were adjusted to "zero phase" with the sync pulse. Being "in phase" with the same pulse now meant that they were "in phase" with each other, meaning that there was a time during the period of the 6MHz sinc pulse when all the carriers would have co-incident peaks. This was expected to be a "worst case" condition. Figure 32 shows the composite of all these carriers in the coincident with the sync pulse. Switches were then installed in each channel mixer jumper to allow easy addition of half wave length of cable. Channels could then be quickly shifted 180 degrees in phase.

Some practical problems have arisen in detailed evaluation of effectiveness of this phase adjustment. The coincident phase adjustment made in late November was very tedious. It required adjustment of one channel at a time, while the system was out of service (early morning hours). Several problems were apparant very soon, mostly from the fact that phase stability and adjustability had not been considered in the design of the head end equipment at St. Catharines. It had been spec'd for coherency, i.e. ability to phase lock to desired harmonic comb components, but not to maintain any specified phase relationship with that comb component.

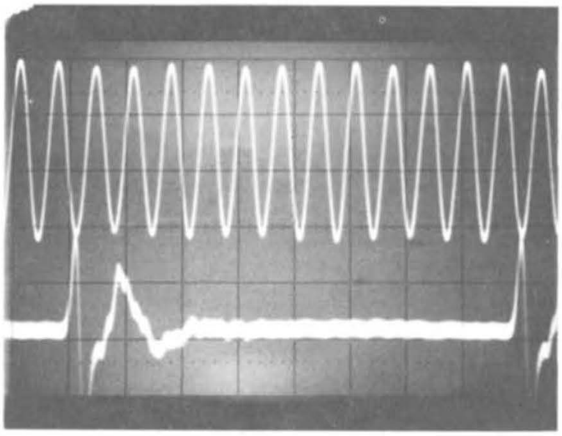


FIGURE 29 0.02 usec/div

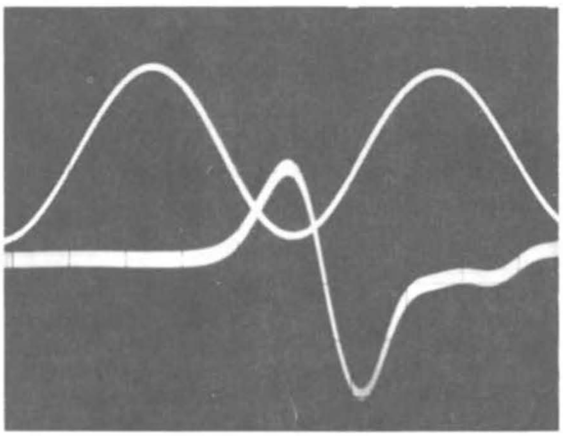


FIGURE 30 0.002 usec/div

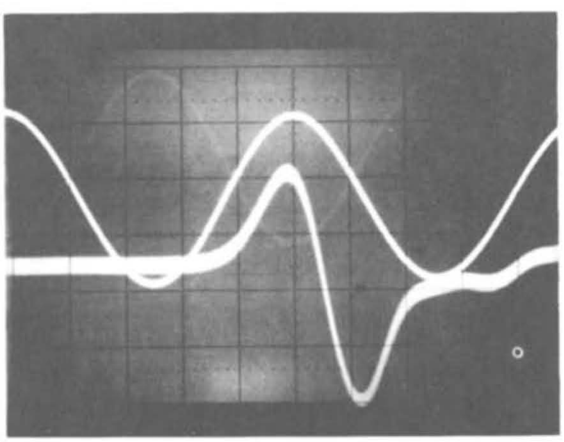


FIGURE 31 0.002 usec/div

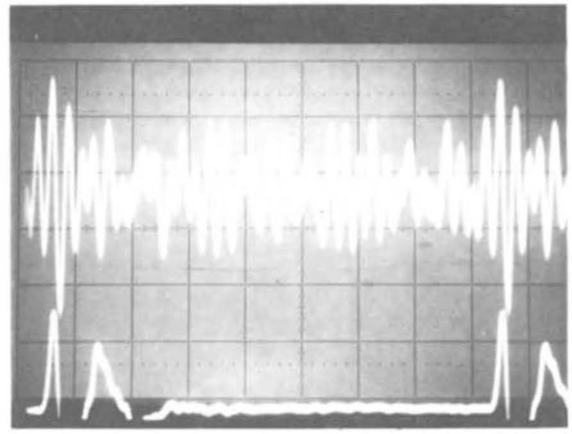


FIGURE 32 0.02 usec/div

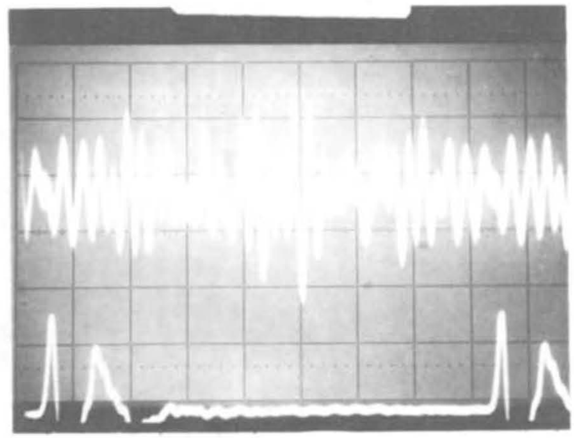


FIGURE 33

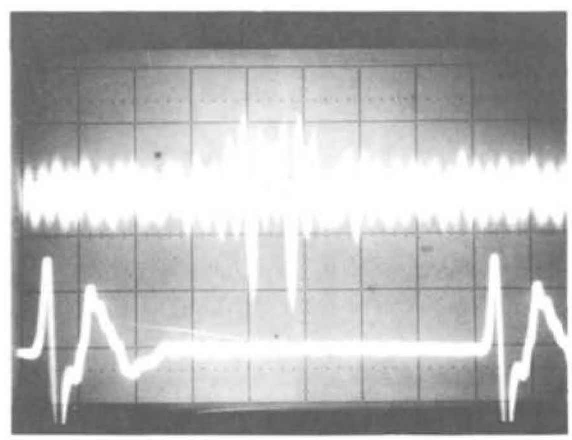


FIGURE 34 0.02 usec/div

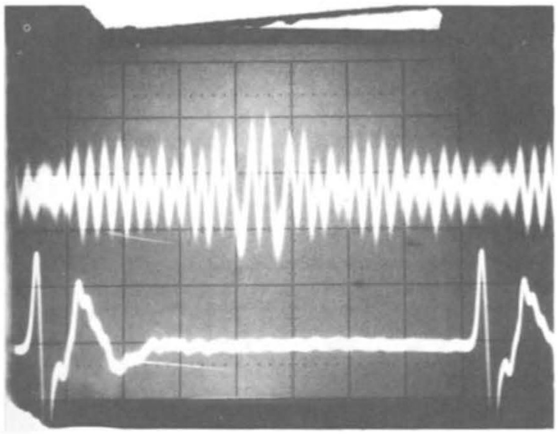


FIGURE 35   0.02 usec/div

We found that the standby carriers had different "phase" than the normal carrier. We also found that "relative phase" drifted with time. Adjustment of phase locking controls or any of the tuned circuits in the equipment also changed the relative phase. We therefore found it difficult to reproduce the coincident phase condition or any other specific relationships. Having observed the "worst case" coincident phase and photographed it we could now recognize it if it should occur again. We are getting the impression that the range of "optimum" adjustments is quite broad and that as long as the worst case "coincident phase" is avoided we are probably realizing near maximum benefits.

The new generation of coherent head end equipment for subsequent application will be designed for stability and ease of phase adjustment to overcome the experimental difficulties we have experienced with this installation.

ADDITIONAL COMMENTS

The new amplifiers installed in the transportation trunk use modulated pilot carriers. The pilots, at 72 and 240 MHz were coherent but were initially modulated at 30KHz. The 30KHz modulation was occasionaly visible as a slight beat in system pictures. Modulation was changed to 23.601 KHz (3/2 X H). This "interleaves" any visible beat that is produced. The "beat" or cross modulation from the pilot was slightly visible when ordinary cross-modulation was not visible because of the high modulation index and spectral nature of the modulation. The pilot modulation sidebands are very high in level compared with individual sidebands in normal TV modulation, see figure 36. The 23.601KHz modulating frequency has remedied this problem.

Interference potential to other services has been considered, particularly to the aircraft distress frequency, 121.5MHz. HRC carrier frequencies are closely controlled. HRC-A at 120.000MHz is 1.500MHz below the distress frequency. Relative energy density of a typical television channel is shown in figure 37, which is a spectrum analyzer display with 30KHz bandwidth but no video filtering. Sweep was slow enough to display peak energies due to picture components. Energy 1.5MHz above the carrier, at 121.5MHz is about 40 db down from the visual carrier. Radiation at 121.5MHz would be at a level about 40 db below that expected from carriers in the system. If as much as 1 volt of carrier was available for radiation in a system fault condition this would make available about +10 dbm of power for radiation at carrier frequency. At 121.5MHz the available power could be considered to be about 40 db less than this or about -30 dbm, or -60 dbw (1 microwatt). This is a very low level of energy and not likely to cause any interference to an aviation receiver system.

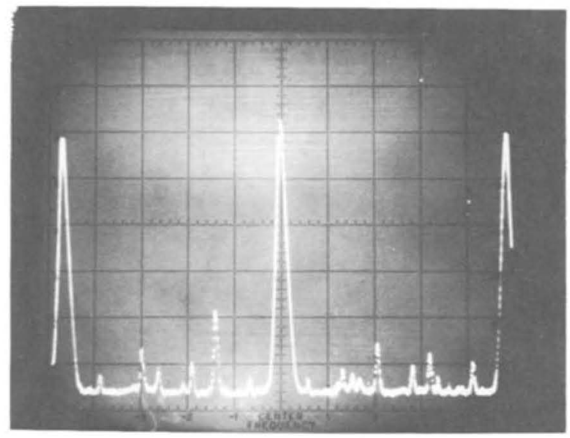


FIGURE 36 5KHz/div BW 300Hz video filter 10Hz showing modulation sidebands on 72MHz pilot carrier and beat from offset HRC-11

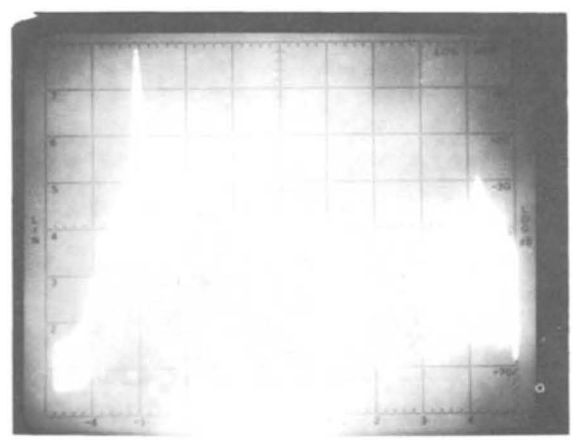


FIGURE 37   500 KHz/div BW 30KHz

SUPPLEMENTAL MEASUREMENTS

Additional oberservations were made at a subscriber's home close to the end of the system (6 Colton Street). System at this observation point has 20 push-pull trunk amplifiers, 24 single ended trunk amplifiers, 1 single ended bridging amplifier, 2 single ended line extenders. A good quality single-ended amplifier with flat band-pass characteristic was used in the subscriber's home to boost levels to overcome spectrum analyzer noise. Figure 38 shows system carrier levels at this subscriber drop. The 240MHz pilot carrier has been attenuated by the single ended equipment.

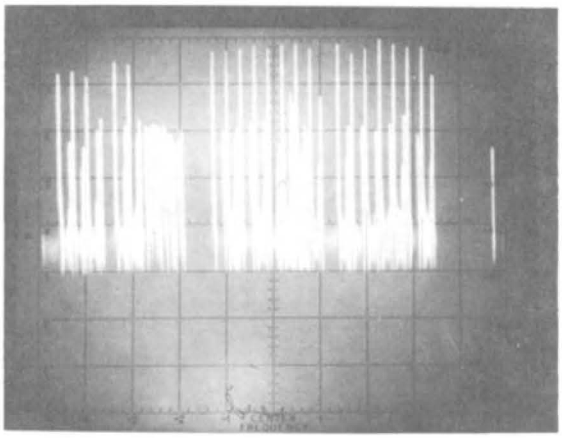


FIGURE 38 20MHz/div BW 300KHz ref level +6.5dbmv subscriber drop 6 Colton Street

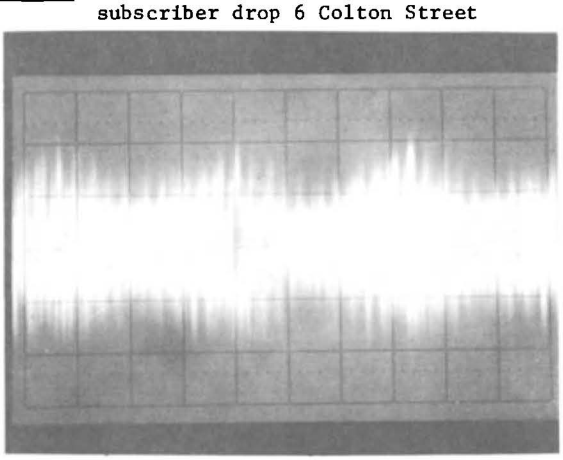


FIGURE 39 0.5 usec/div subscriber drop 6 Colton Street

Figure 39 shows the composite waveform at the subscriber terminal. This photograph is "inconclusive" because of the trouble in triggering since there is no 6MHz system reference signal readily available at this point in the system to provide reliable scope triggering. The photo does suggest that there may be undesirable co-incidence of carrier peaks.

Group delay distortion on a "broad band" basis can cause the phase relationships established at the head end to change along the system. This effect was anticipated in the preliminary technical paper and if necessary we plan to use passive, all pass, delay correction networks in the system.

Noise measurement at channel 3 was made. Figure 40 shows the carrier to noise ratio at the output of the head-end processor on HRC-3. A C/N ratio of about 52 db is indicated.

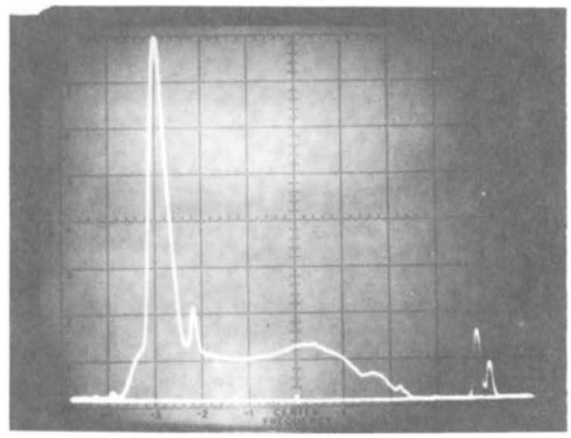


FIGURE 40 HRC-3 Standby at head-end 1MHz/div BW 100KHz video filter 10Hz showing C/N correction to 4MHz noise BW= +18db

Figure 41 shows the noise at the subscriber location. The channel 2 aural carrier has been attenuated somewhat and the channel 4 visual carrier has been severely attenuated by the preselector filter used to prevent overloading of the spectrum analyzer. The C/N ratio is indicated to be about 39 db.

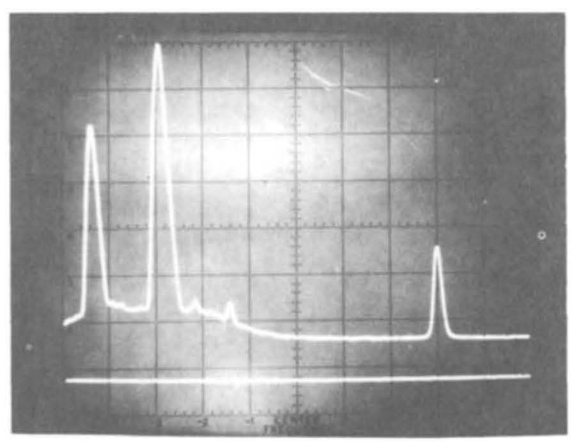


FIGURE 41 HRC-3 standby 6 Colton Street 1 MHz/div BW 100KHz video filter 10Hz correction to 4MHz noise BW = + 18db

Leakage of air channel 11+ into the system was checked. Figure 42 shows this leakage. The photo was made during the readjustment of the offset of HRC-11 for minimum beat visibility and does not represent the optimum position of the beat. This photo should be considered as representing levels only. Reference level is -17.5 dbmv, indicating the channel 11+ leakage to be about -48 dbmv or about 55 db below HRC-11 at this subscriber location. This is considered an acceptable leakage level since this signal/beat ratio would not be visible. Any visible beat would be due to direct pick-up in the receiver connected to the system at this point.

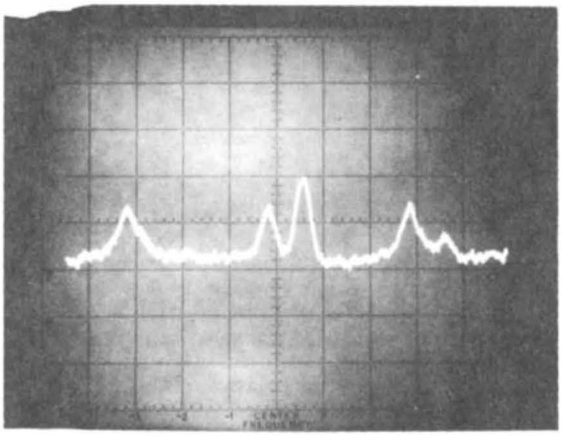


FIGURE 42 Channel 11+ leakage into system
6 Colton Street
5KHz/div BW 1KHz Video Filter 10Hz
ref level -17.5dbmv

Figure 43 shows intermodulation beats in HRC-3 standby due to HRC-11 being unlocked, i.e. not coherent. HRC-11 is being operated so that the direct pick-up beat from air channel 11+ falls halfway between H sidebands. The intermodulation beat caused in other channels then shows up as beats above and below the desired carrier. In this case the undesired beat is about 47 db below desired carrier. The beat is not visible on TV sets because of the H/2 offset. Figure 44 shows beats from other channels into the offset HRC-11 at a similar level.

We tabulated all the second order beat products that would fall into HRC-3 due to the added mid-band channels. Since all the beat products fall directly on each other the spectrum analyzer cannot distinguish any single beat. The many third order beats which are generated also fall on each other but are much lower in level than the second order beats. The beat which is seen when the carrier is removed is the resultant (vector sum) of all the beats which are generated. There are ten second order beats, all "differences" which fall on HRC-3.

Only the D - 5 and the E - 6 beats would be harmfull in a "conventional" system since the rest of the beats fall at the band edge where their effect would be minimal. Channel HRC-3 was chosen as an example because of its convenience for testing during system operation. It is not the worst case channel for second order beats. It is available because Global starts programming late in the afternoon and it is tunable on the high resolution H-P spectrum analyzer which makes possible a degree of precision not easily achieved for measurements above 110MHz.

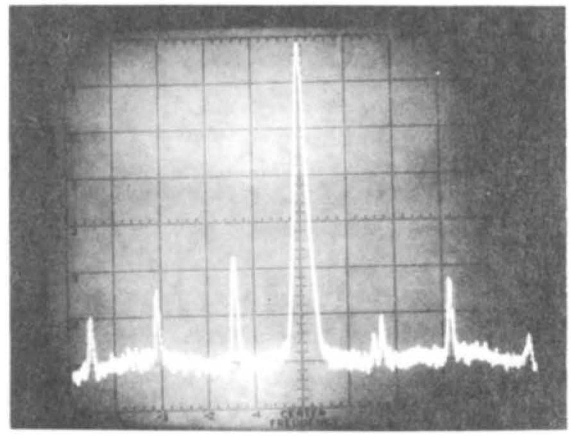


FIGURE 43 Intermod beat into HRC-3 standby due to unlocked, offset HRC-11
5KHz/div BW 300Hz Video Filter 10Hz

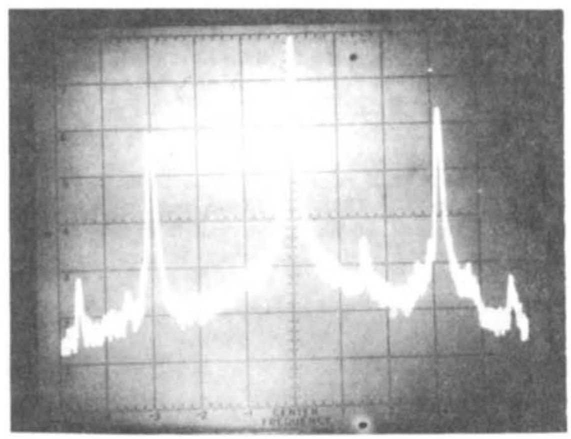


FIGURE 44 Intermod into HRC-11 due to unlocked offset operation of HRC-11
5KHz/div BW 300Hz Video Filter 10Hz

Beat Product	Freq. in "conventional" system	Freq. of video beat in conventional system
13 - F	60MHz +-20KHz	band edge
12 - E	"	"
11 - D	"	"
10 - C	"	"
9 - B	"	"
8 - A	"	"
A - 3	"	"
B - 4	"	"
C - 72MHz pilot		
(73.5)	59.75MHz +-	CH 2 sound
D - 5	62 MHz +-	0.75 MHz
E - 6	62 MHz +-	0.75 MHz

Figure 45 is a triple-exposure showing channel HRC-3 in normal picture modulation, standby, and removed. This shows the level of beats under the carrier and also gives some further noise information. Some of the noise observed in standby carrier is due to the head end processor and that the system noise, with the head end processor disconnected is a further 6 db down.

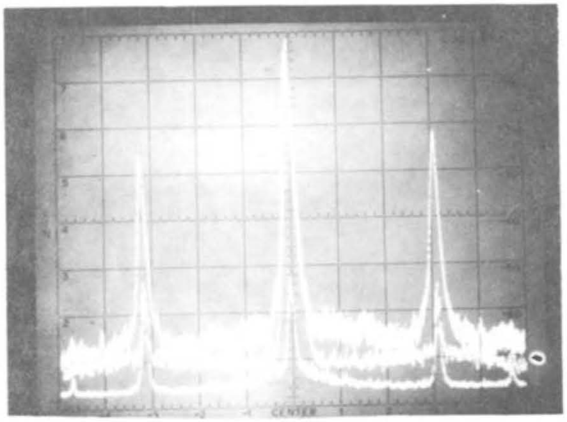


FIGURE 45 5 KHz/div BW 300Hz Video Filter 10Hz

Figure 46 shows the HRC-3 standby carrier at 6 Colton Street and figure 47 shows the intermod beats under the carrier, i.e. carrier removed. The carrier beat is about 48 db below desired vidual. Being zero beat (HRC-11 was locked in for these observations) there was no visible effect on the screen due to carrier beats. The H sidebands associated with the carrier beats are about 55 db below desired visual carrier but only about 30 db below the normal H sidebands in a normally modulated picture channel. This could be interpreted as -30db cross modulation if the observed H sidebands are considered to be undesired modulation.

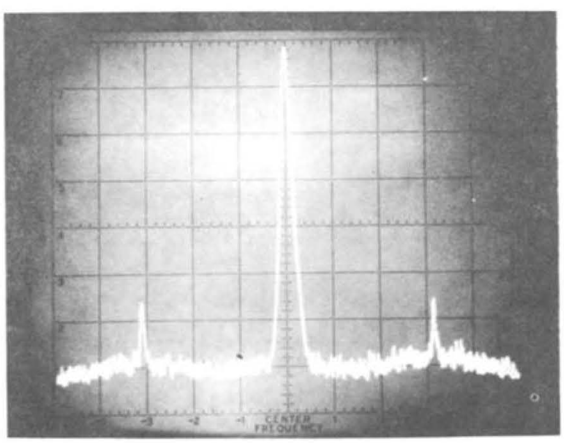


FIGURE 46 HRC-3 standby 6 Colton Street
5 KHz/div BW 300Hz Video Filter 10Hz

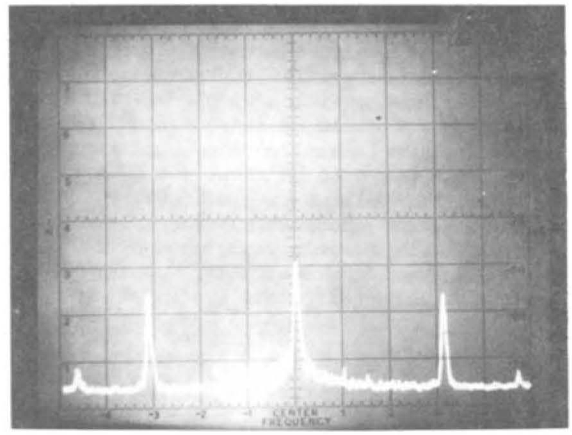


FIGURE 47 Standby carrier removed
5 KHz/div BW 300Hz Video Filter 10Hz

All carriers were then unlocked to show effect of non-coherent operation. Figures 48 and 49 show the resultant beats. The frequency spread is due to the head-end processors dropping out of phase-lock. Restoration of the reference signal then pulls them back to locked, coherent operation. Note that there are four beat products at a -48 db level and additional beats at lower levels. If these individual beats had added "in phase" on a voltage basis they might have produced a beat some 12 db higher when they are "brought together" by coherent operation. Even with power addition as a worst case the sum of these four beats might have been 6 db higher than each individual beat. Figure 47 shows that the resultant of all the carrier beats is no greater than any of the individual beats. Obviously an advantageous cancellation of beats has taken place, or at least a partial cancellation.

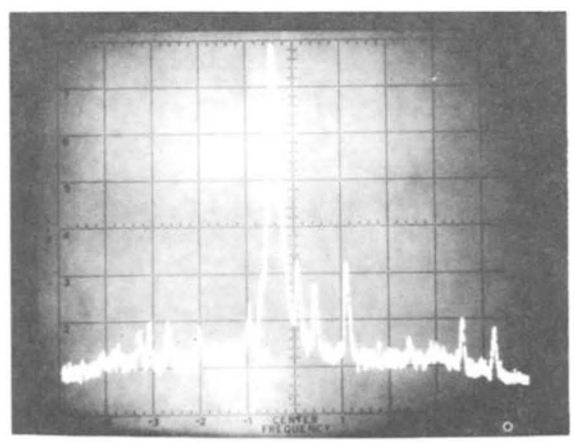


FIGURE 48 HRC-3 standby all carriers unlocked
5 KHz/div BW 300Hz Video Filter 10Hz

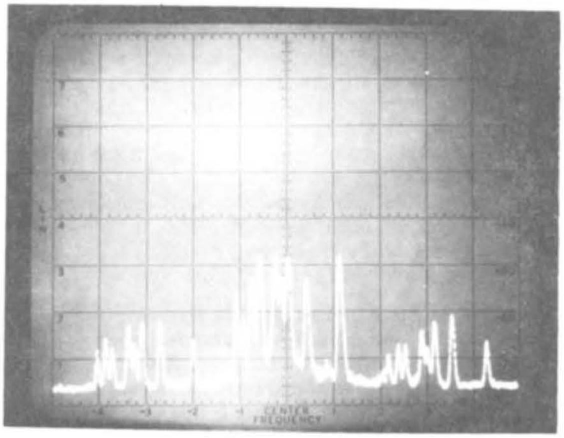


FIGURE 49 Carrier removed
All other carriers unlocked
5KHz/div BW 300Hz video filter 10Hz

The proper way to "add" beats at different frequencies as in figures 48 and 49 is not really known. These beats are spread over a 50KHz span around the carrier. We have made similar observations of "dispersion" of individual third order beats around a carrier in conventional systems. How does the eye sum the consequent video beats as seen on a TV receiver? Is the visual effect of a number of dispersed beats the same as from a single beat which is the power sum of the individual beats? One paper presented at the 1973 NCTA convention (Arnold - "REQUIRED SYSTEM TRIPLE BEAT PERFORMANCE") suggests that a "Power Law" is applicable. In a 24 channel example the individual third order intermod beat would have to be 76 db down from desired carrier so that the sum effect of the 198 triple beats falling on the worst case channel (9) would meet the visibility threshold requirements he sets out.

The "apparant" 30 db cross modulation observed in figure 45 is alarming when viewed on the spectrum analyzer, yet the pictures visually observed on a test receiver and the subscriber's receiver were judged of "good quality" with no visible cross modulation. At the date of this writing (March 1st, 1974) we have not been able to explain this apparant inconsistency.

RURAL EXTENSION TECHNIQUES AND SYSTEMS

Victor Nicholson
Cable Television Information Center
Washington, D. C.

The difficulties of extending cable TV service into low density areas have plagued cable TV operators and small town residents since the construction of the first community antenna television system in 1949. The principal problem was that the low density of rural America coupled with the high cost of constructing cable systems made it financially uninteresting for cable operators to build systems in rural areas and outside the denser portions of towns. Furthermore, in recent years when substantial amounts of risk capital began flowing into the cable industry, all of the attention was turned to the major urban areas where the existing vast television markets are located. As a result, by]972 when the new FCC Report and Order was promulgated, most of the activity and excitement centered around efforts to win franchises in suburbs and central cities located in the major urban markets.

In this talk, I would like to turn your attention away from central cities to the sizeable and as yet untapped markets which exist in rural, small town, and fringe suburban areas.

Two types of market exist in these areas. The first is the market for extension of service from existing systems in small town and suburban areas. The second is the market for new cable systems largely in rural and suburban areas.

The magnitude of the market involved is tremendous; almost 40 million households are in rural, small town, and suburban areas. Contrast this with the present approximately 7 million cable subscribers outside of central cities. The cable industry could triple in subscribers without even touching cities of 50,000 population or more.

The need for expanded telecommunication service is even more important than in urban communities because of the isolation of the people and the and the distances involved. Telecommunications can be used to deliver medical, legal, educational and other social and governmental services directly to the home. The added factor of gasoline shortage makes it even more imperative to substitute the transportation of electronic signals in place of transporting people.

This market must be served -- if not by cable TV then by other technologies. A recent study shows by OTP[1] translators to be an efficient means for delivering broadcast television to rural homes. It discusses the need for variances in federal regulations to make translators viable funding, suggested sources of funding, but I can assure you that if there continues to be a lack of TV service in rural, small town, and suburban areas, -- there will be more pressure by farm organizations, regulatory agencies, and others for alternative means of television distribution.

In this talk, I want to suggest that the so-called problems of system extension and new construction in rural, small town, and suburban areas should be viewed as opportunities by the cable industry to provide more service to more subscribers and, consequently, opportunities to make additional profits.

The Cable Television Information Center feels the main barriers to system extensions and rural area service have been economic and technical; that there is a rather substantial "information gap" in which it is difficult for small system operators to keep track of the rapidly changing equipment capabilities, construction techniques and system design. There has also been a lack of understanding of the results in savings to an operator and of extended distance of possible service due to improvement of amplifiers and cable. Keep in mind that there are already some operators today that have been serving areas with densities of 9 households per mile or less; as examples: Joe Gans in the Poconos of Pennsylvania and Less Biederman in Traverse City, Michigan. I am sure there are others but the numbers are far to few.

What are some major cost reducing factors? First there needs to be an understanding that undergrounding which can cost $12,000 to $100,000 per mile in central areas of cities can cost less than $1,000 per mile in rural areas. There are vibratory plows today that eliminate the separate actions of trenching, laying cable, covering and compacting. The equipment is available with accessories for under $15,000 but for those opera-

1. Denver Research Institute, "Braodband Communications in Rural Areas", Nov. 1973.

tors who prefer to contract out the construction, bids vary from 12 to 18 cents per foot. To this must be added about $100 per black top crossing -- but there are few of these per mile in rural areas. Another important long term saving by undergrounding is the elimination of pole rentals. The $175 per mile yearly cost is a major cost factor when there are only a few subscribers per mile.

Another saving for low density areas is in the use of system design where a single cable tapped trunk is substituted for standard separate trunk and feeder cables. The major saving is in elimination of the parallel cable and construction. Using tapped trunk design, figure 8 cable (with messenger included) is applicable and therefore the need for linemen to climb poles is reduced. With earlier construction techniques the lineman climbs to install the strand, again for the cable and again to lash them together. Using figure 8 cable the lineman need climb each pole only once.

The suggestion of tapping the trunk is contrary to past CATV practice where this technique has been rejected as a source of reflections. Conditions have changed; the coaxial cable today has a return loss of 30 dB, in contrast, for many years the manufacturers refused to release this specification: probably becasue measurements snowed cable to have less than 20 dB return loss. The amplifiers used today are terminated both at input and output -- or years that of many suppliers had a high impedance output; finally the tap-offs used today are directional with a good impedance match at in/out and thru terminals. For low density household areas, it is reasonable to use the same cable for both trunk and feeder for a saving of several hundred dollars per mile.

Another important saving in system design is applicable to long haul rural areas where several small communities are served. Typically costs of microwave between towns is added to the distribution costs making a proposed system non-viable. The center recommends the use of low-sub amplifiers with a top frequency of 108 MHz. This permits transportation of 12 TV channels plus FM over a distance of more than 60 miles -- still meeting a Signal/Noise of 43 dB and Signal/Intermodulation of 57 dB. The same transportation cable is used for distribution of signals enroute such that two levels of service are delivered. For a lower fee, the subscriber receives only the 5 VHF channels plus FM. Those also desiring the 7 high VHF channels would buy a set converter. Upon reaching a community the operator has the option of splitting the system and converting the distribution section to 12 channels and so eliminated the need for set converters.

The feasibility of such long haul cascades contradicts much application design criteria of the past where 10 to 15 miles was the maximum.

Todays amplifiers are far better than those of the past and being solid state require improved power supplies for protection of the transistors or IC's; as a side effect they have a better than 20 dB improvement in reduced hum modulation. Over the years the power handling capability has also improved -- a very important factor since a 6 dB improvement permits a doubling in amplifier cascade. Finally solid state devices tend to have a relatively flat amplitude versus frequency response across a wide range of frequencies. The amplifiers are also designed with well matched input and output circuitry, recognizing the importance of minimizing reflections.

Likewise coaxial cables are improved; the attenuation being reduced, frequency range increased, and reflection characteristics improved.

Finally accessory devices such as cable connectors, splitters and tap off devices are designed to be well matched to the 75 ohm coaxial cable and are no longer major sources of reflections. Another factor is the use of directional coupler subscriber taps that provide isolation of TV receiver discontinuities from the downstream signals.

Even for extension of service, advantage can be taken of improved amplifier capabilities to reduce costs. Today's better line extenders use similar IC circuitry as in the trunk amplifiers -- with slightly lower performance capability. Some line extenders include equalization, gain and slope control and are of push-pull circuitry to permit use of the mid and superband frequencies. These line extenders are superior in performance to trunk line amplifiers of only a few years back and for low density areas where minimum cost is very important; can supplant trunk amplifiers.

Keep in mind that for top 100 urban markets the present system design is necessary, providing extra reliability, redundancy, standby power, modulator capability and future two-way. For rural America areas where limited channels and one-way services are acceptable, there are available lower cost amplifiers of better capability than is generally recognized. As for two-way, these subscribers will presently settle for use of the telephone.

The CTIC is making a study of techniques, costs and profitability of investment for eleven different options descending from a top 100 market, two-way aerial system down to a low cost rural underground system at$2,150 per mile. We have separated them into two categories: the first category is extensions from existing systems where certain assumptions are made; (1) that there need be no additional headend cost, (2) that cable system facilities can be used for office warehousing of material, (3) that equipment and cable list prices shall be used as a base for calculations even though there may be a discount of up to 30% for amplifiers for large volume users, (4) that a markup of 25% for overhead is included in installation costs, permitting the construction to be contracted out, (5) that the cost of tree trimming and pole rearrangement in top 100 markets are assumed to be $700, (6) that the cost of pole rearrangements in areas of lower household density are assumed to be $200 per mile, and (7) that annual pole rental is $175 per mile in top 100 markets and $140 per mile in areas of lower density.

The second category is construction of new systems in currently unbuilt rural areas, where the costs of headend, office and warehouse need to be added to the distribution system. Here too lower cost headend equipment can and should be used where warranted. However for a long system with many subscribers the added cost per subscriber for improved headend equipment is not an important factor.

Here are a few graphs showing one of the Options Options (#9) covering the cost, design, equipment used, and the rate of return on investment depending on subscriber density, installation charge and monthly fee. Since the expected subscriber penetration for any area is know by the operator, he can convert subscribers per mile to household density per mile.

In conclusion the center feels that operators need no longer limit their service to areas of 40 households per mile which at 50% penetration is 20 subscribers per mile. We calculate a possible rate of return on investment of better than 15% for areas of only 7 subscribers per mile using a $25 installation fee and amortizing costs over 10 years.

In short, cable operators now have the opportunity to increase their subscribers and profits by using modern methods to build in rural, small town, and suburban areas. Central cities may have to wait for new cable services to attract subscribers. Low density areas offer millions of subscribers who are ready now.

Finally, I would like to extend my appreciation to many operators and to Cadco for providing construction techniques and costs and I hope that this information will be helpful, to such that this year the cable industry will make an all out effort to wire up the low density areas of America.

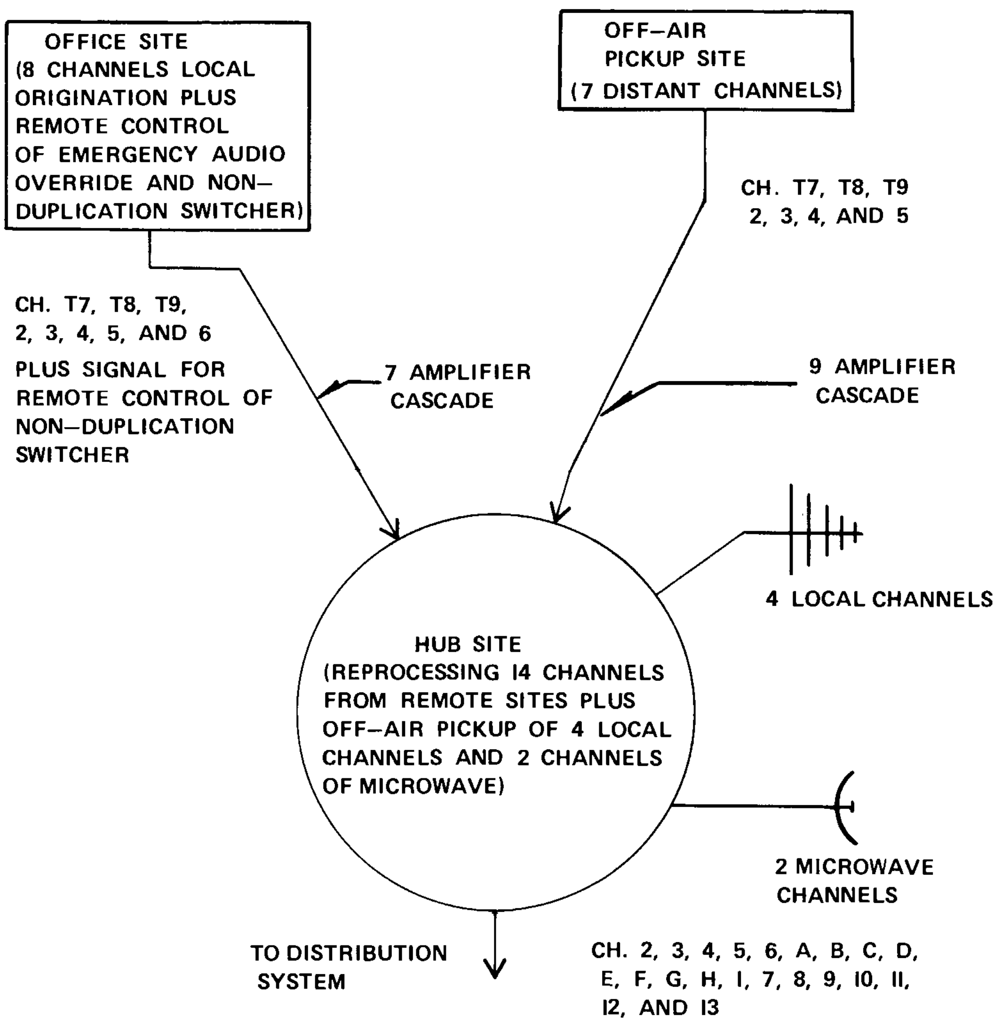
SIGNAL PROCESSING REQUIREMENTS FOR MODERN CABLE SYSTEMS

By Alex Best

Scientific-Atlanta, Inc.
Atlanta, Georgia

The movement of cable TV over the past few years from the rural areas into the large metropolitan areas has brought with it new demands on the techniques of processing signals. To a large degree these changes in processing requirements have been dictated by the fact that in metropolitan areas, the best geographical location for receiving signals off-air is separated by some distance from the most convenient office and local origination site, and in most instances both of these are located some distance from the optimum point to distribute the signals. Headend electronics capable of operating in this environment without producing excessive degradation have increased the technical performance specifications of headend equipment, especially heterodyne signal processors. To investigate these increased performance demands, consider a "typical" headend system for a metropolitan area such as the one shown in Figure 1.

FIGURE 1 TYPICAL HEADEND SYSTEM FOR A METROPOLITAN AREA



Here we have the situation discussed above, where the office and local origination site, off-air pickup site, and hub site are physically located several miles apart. The processing problem arises at the hub site, particularly with the channels from the other sites that must be reprocessed. For the sake of discussion, let us consider the channel 3 signal being fed into the hub site from the off-air pickup site. This channel was originally received off-air as a standard VHF or UHF signal and converted to channel 3 in a signal processor. After combining with the sub-low and low band channels, it is fed through a 9 amplifier dedicated distribution system to the hub site. Here the combined signal is split as many times as there are channels coming in and fed into the input of a second processor. In this discussion we will assume that the channel 3 coming from the off-air pickup site will be carried on channel 3 on the distribution system. This being the case the signals present at the input to the channel 3–3 processor at the hub site will appear as in Figure 2.

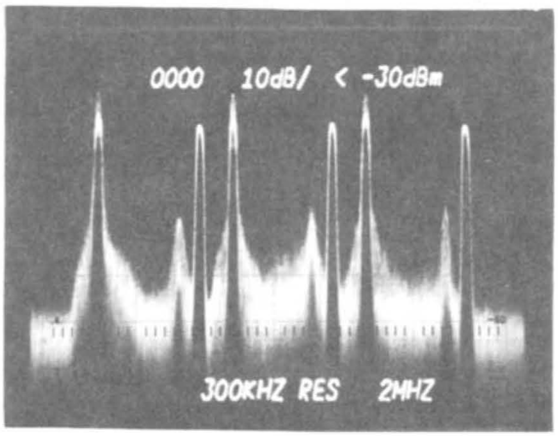


Figure 2. Ch 2, Ch 3, Ch 4 out of seven channels coming from off-air receiving site.

Here we show only three out of the seven signals that are actually present. For purposes of demonstration the sound carrier levels have been set 6 dB below the video carrier levels when in actual practice they are set approximately 15 dB below. The question here is: how much adjacent channel rejection must this processor have in order not to interfere with other channels being distributed on the cable? The worst case problem would exist if the channel 4 signal being distributed on the system is different from the channel 4 signal present at the input to the channel 3–3 processor. Under these circumstances, any of the channel 4 signal from the
through the channel 3–3 processor would fall back on channel 4, with an unavoidable frequency difference between it and the channel 4 distributed on the system. In this case 60 dB would be the minimum adjacent channel rejection allowable. In Figure 3 we show the amplitude response of a Scientific-Atlanta Model 6150 channel 3–3 processor. Here we see "notches" at the upper adjacent picture carrier frequency and lower adjacent sound carrier frequency in excess of 70 dB. We also see undesired but unavoidable "response upshoots" on the high frequency side about 20 dB down and on the low frequency side about 30 dB down. Would a processor with this

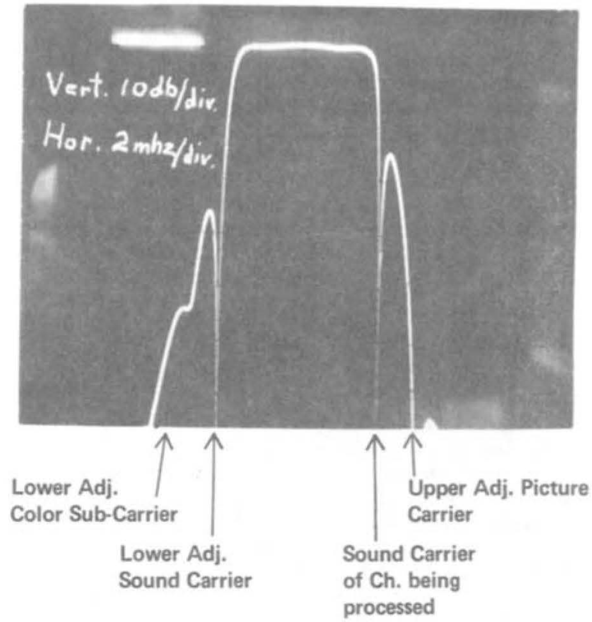


Figure 3. Amplitude Response of 6150 Signal Processor.

amplitude response be specified as having 60 dB adjacent channel rejection? Shown in Figure 4 is the output spectrum of this processor with the input spectrum shown in Figure 2. Here

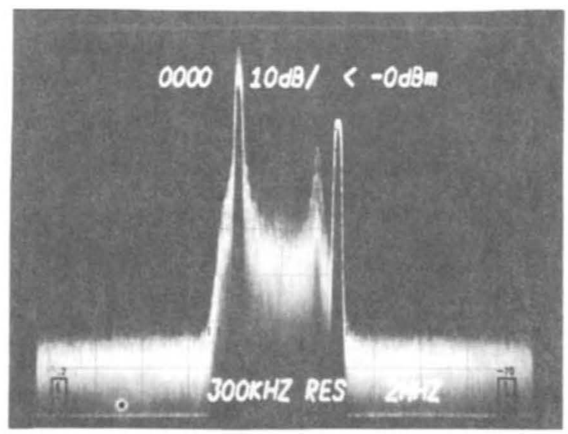


Figure 4. Output spectrum of signal processor with amplitude response shown in Figure 3 and input spectrum shown in Figure 2.

we see that there is no energy falling in the upper and lower adjacent channels greater than 60 dB below the desired video carrier level. Therefore, if we define adjacent channel rejection as the rejection to equal level adjacent channels that have "normal" energy distributions (pc = 0 dB, sc = -6 dB, cc = -17 dB), this amplitude response would have 60 dB adjacent channel rejection. The point here is that deep notches alone do not constitute adjacent channel rejection. Nor does "response upshoots" necessarily hurt you. It is a combination of the processor amplitude response and the energy distribution of a "typical" television signal that together make up adjacent channel rejection. Signal processors with 60 dB adjacent channel rejection have only recently become available in the CATV industry. Before this, processors located at hub sites required external traps/filters to achieve this degree of selectivity.

It should be pointed out that any additional increase in selectivity does not come without penalty. The price we pay is an increase in the differential time delay across the passband of the processor. Shown in Figure 5 is the amplitude and envelope delay response of the IF section of a Scientific-Atlanta 6150 processor, without delay equalization. The IF portion of the processor constitutes the majority of the selectivity and therefore produces practically all the delay. In the CATV industry envelope delay is defined as the differential time delay between the picture carrier frequency and the color sub-carrier frequency. From Figure 5 we can see this is approximately equal to 130 nanosec. for the processor with no equalization. It is also interesting to note that although the "envelope delay" is only 130 nanosec., the total differential time delay from band edge to the center frequency is approximately 500 nanosec. It should be pointed out that for "minimum phase shift" networks, into which category virtually all headend equipment falls, the amplitude response uniquely defines the delay response.[1] Therefore, regardless of how the response shown in Figure 3 is obtained, the delay would be the same.

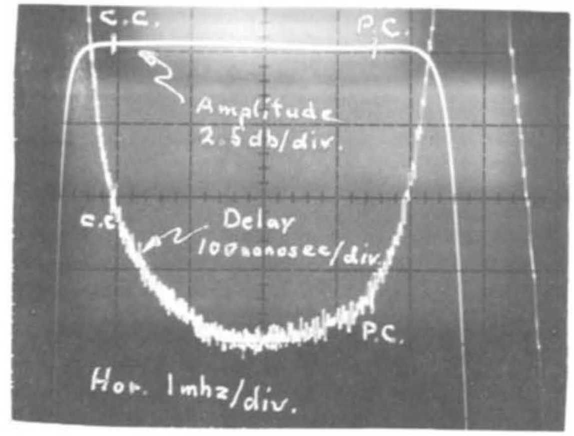


Figure 5. Amplitude and envelope delay response of IF section of 6150 processor (without delay equalization).

To determine the detrimental effects of processing signals through a device that has the delay characteristics shown in Figure 5, the test set up shown in Figure 6 was used. Here we

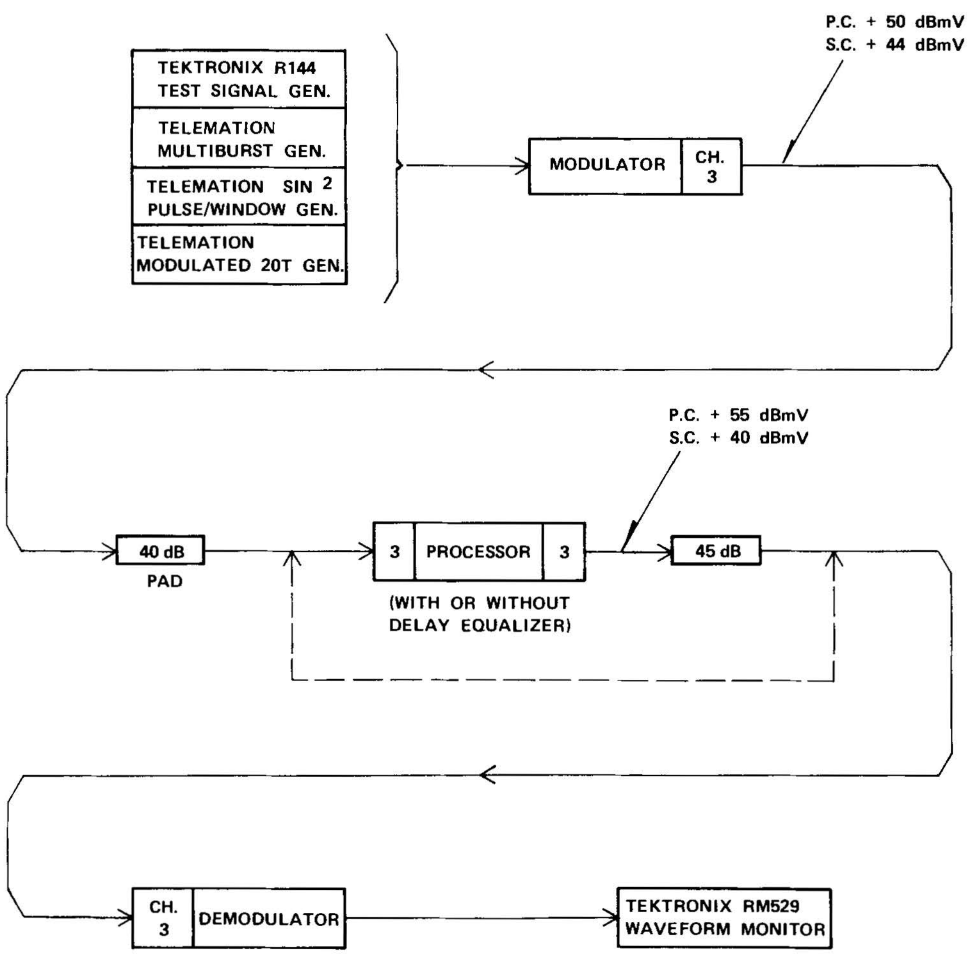
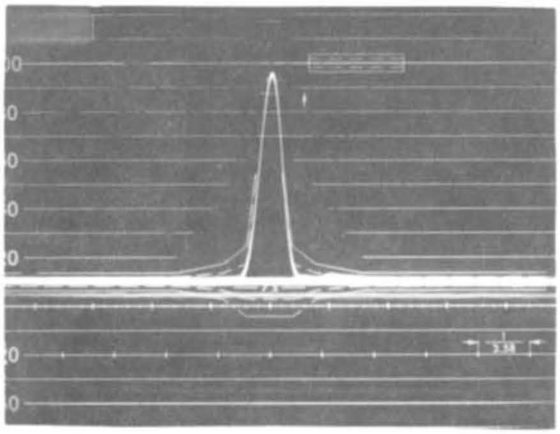
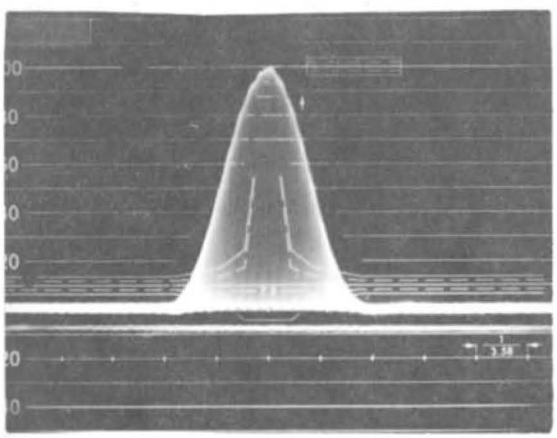


FIGURE 6 EQUIPMENT SETUP FOR WAVEFORM TESTING OF SIGNAL PROCESSOR

have a channel 3 modulator and channel 3 demodulator with the capability of either connecting them back to back or inserting the channel 3–channel 3 processor between them. Two video test signals were used in this evaluation, one being the "Modulated 20T pulse" and the other the "2T Sin² pulse." The first test signal was designed primarily to indicate the relative delay between the video carrier and the color subcarrier frequency. The "Sin² pulse" is more sensitive to the amplitude and delay in the vicinity of the response that the luminance signal energy occupies. Figure 7 shows the response of the modulator–demodulator back to back for the two test signals described above. As can be seen, the modulator–demodulator combination is very transparent, indicating very little distortion. The "Modulated 20T pulse" indicates virtually no envelope delay and the "Sin² 2T pulse" has a K factor of 2%.



K factor
≈2%
2T Sin² Pulse

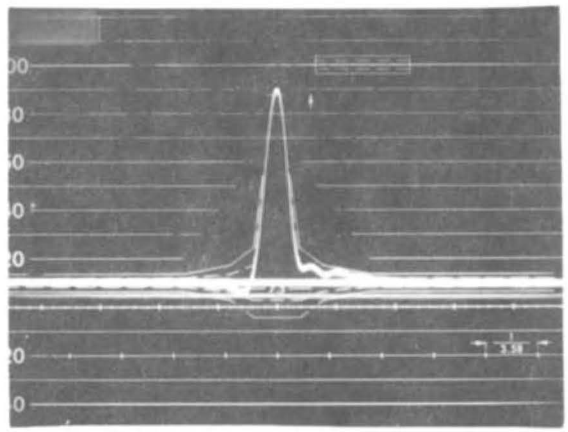


Envelope Delay
≈0
Modulated 20T Pulse

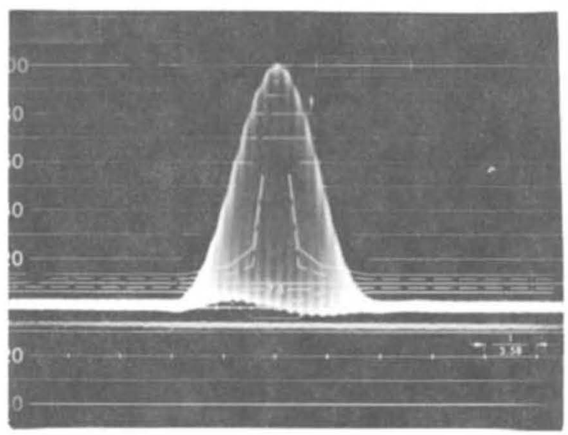
Figure 7. Video test signal response of modulator–demodulator.

Shown in Figure 8 is the result of passing these test signals through the same mod–demod, but with a 6150 processor with no delay equalization inserted between them. The distortion

off-air pickup site feeding



K factor
≈3.5%
2T Sin² Pulse



Envelope Delay
≈110 nanosec
Modulated 20T Pulse

Figure 8. Video test signal response of modulator – processor (without delay equalization) – demodulator.

passing through only one processor with no delay equalization. In headend-hub systems many signals pass through two processors. In this case the envelope delay would become approximately 220 nanosec. and the K factor would increase by lization. From the amplitude response we see that although the all-pass delay equalizer is theoretically a lossless device, it does introduce a small amplitude variation (±.3 dB typical). Actually, the delay equalizer alone produces a 5 or 6 dB dip in the response which must be amplitude equalized to achieve the desired flatness. The results of passing the multiburst test signal first through the mod-demod back to back and then through the mod-processor (with equalizer)--demod can be seen in Figure 10. As evident from this photograph, the amplitude variation

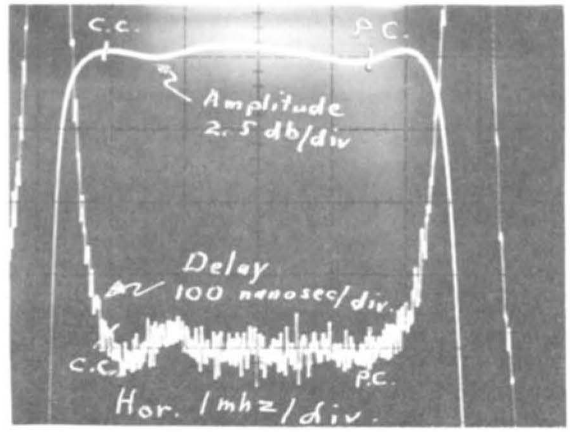
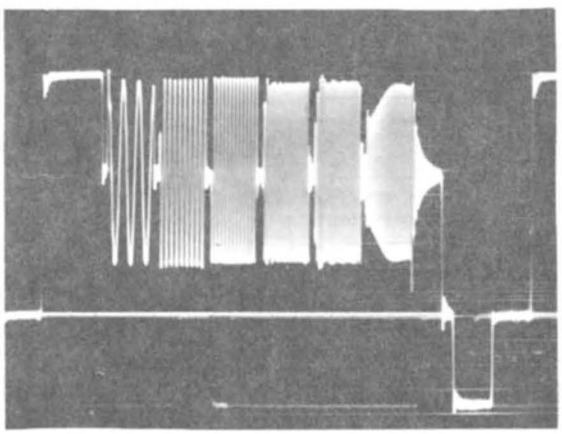
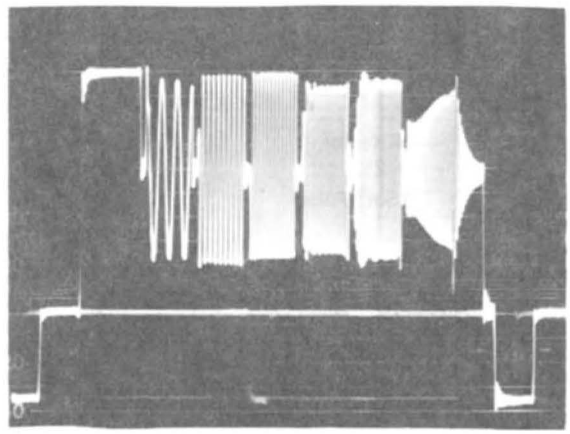


Figure 9. Amplitude and envelope delay response of IF portion of 6150 processor (with delay equalization).



Modulator—Demodulator

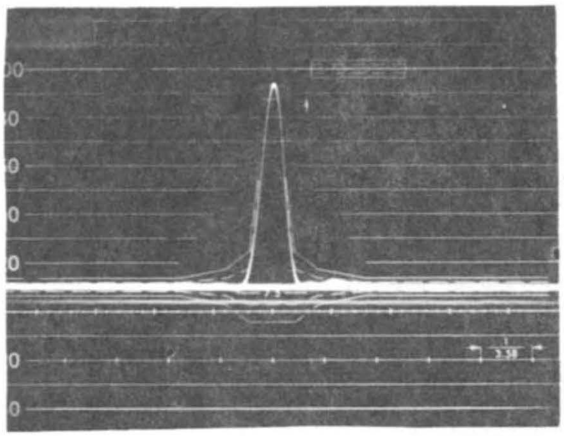


Modulator — Processor (with delay equalization) — Demodulator

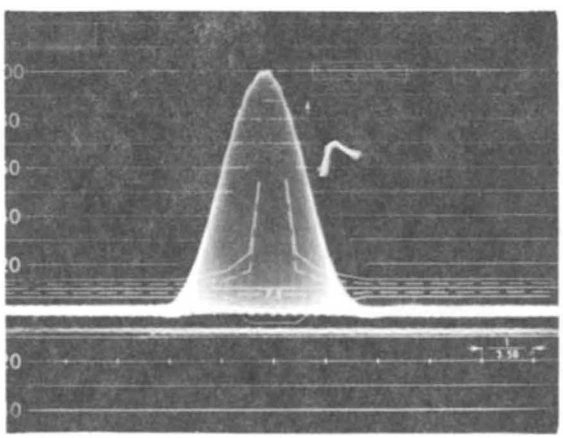
Figure 10. Multiburst response of modulator — demodulator and modulator—processor (with delay equalization) — demodulator.

is certainly tolerable. Obviously, when comparing Figure 9 and Figure 5, the most dramatic difference caused by the delay equalizer is in the flatness of the envelope delay response. From this response it can be seen that the relative delay between the video carrier frequency plus .5 MHz and the color sub-carrier frequency is flat ±20 nanosec. Also evident is that from the two frequencies mentioned above to the band edges the delay is approximately 200 nanosec. The reason for this compromise in equalized delay is two-fold: first, the increase in complexity in an equalizer to achieve a flat delay from band edge to band edge would be tremendous. Such an equalizer would prove to be prohibitive not only from an alignment standpoint but also from a physical size standpoint, not to mention the problem of amplitude equalizing the device. Second, and even more important, the additional degree of improvement that would be obtained by complete equalization is negligible, compared to the improvement already achieved. This can be seen from Figure 11. Here we see the results of sending the Modulated 20T and Sin² 2T pulse test signals through the mod-processor (with equalizer)--demod. Note that the degree of distortion to the base line of the 20T pulse is virtually nonexistent, indicating, as is already evident from the swept delay curve, that the differential envelope delay at the video carrier frequency and color sub-carrier frequency is zero. Even more important however, is the complete lack of undershoot and trailing smear that was evident when there was no processor equalization. This is due to the flat delay in the vicinity of the video carrier. Comparing Figure 10 to Figure 7, it is apparent the processor with equalization is transparent to the test signals discussed above.

Now that we have an equalized processor, with 60 dB of adjacent channel rejection, does not insure that the use of external traps and filters can be eliminated. Discussed earlier in this article is the fact that at the hub site many processors will have as inputs the desired signal plus many more, including equal level adjacents. To insure that no external traps/or filters are required at the input of each processor, it must have the capability of "handling" these signals without generating in-band intermodulation. This distortion could occur in any active stages prior to removal of the adjacent channels and once generated, no amount of adjacent channel rejection can remove it. To demonstrate this problem, I combined the output of a channel 2, channel 3, and channel 4 modulator. The carriers are not modulated to assist in looking at in-band intermodulation. The combined spectrum of these three modulators is shown in Figure 12.



K factor
≈2%
2T Sin^2 Pulse



Envelope Delay
≈0
Modulated 20T Pulse

Figure 11. Video test signal response of modulator—processor (with delay equalization) — demodulator.

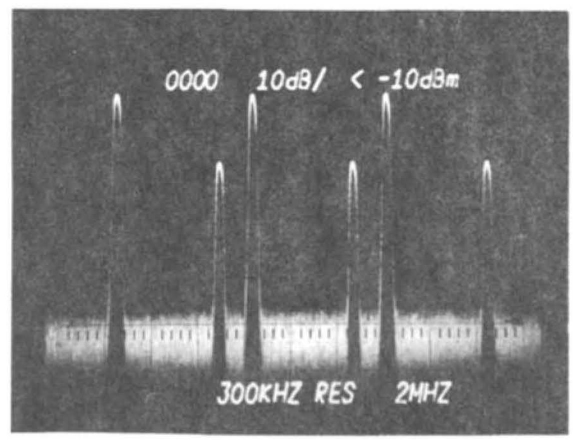
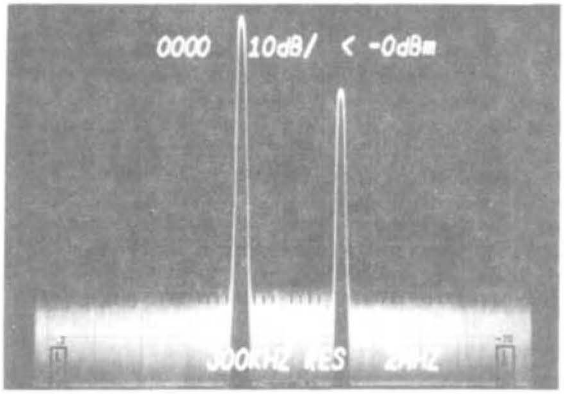
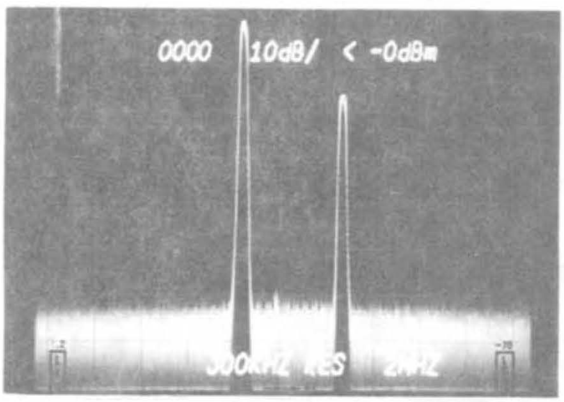


Figure 12. Combined output of Ch 2, Ch 3, and Ch 4 modulator.

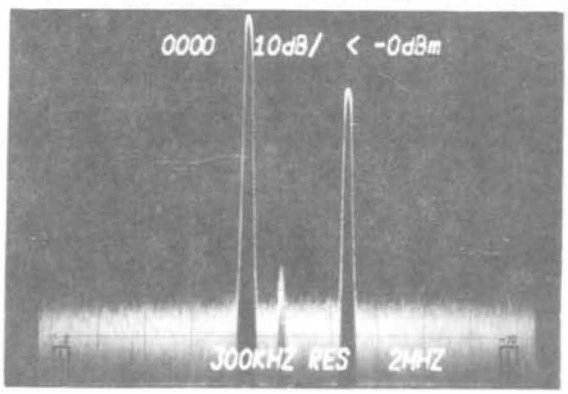
These signals were then fed through a 6150 channel 3–3 processor with the input levels set first at +10 dBmV, then at +20



Input Level
+10 dBmV



Input Level
+20 dBmV



Input Level
+30 dBmV

Figure 13. Output spectrum of 6150 Ch 3 – Ch 3. processor with equal level input signals on Ch 2, Ch 3, and Ch 4.

dBmV, and finally at +30 dBmV. Figure 13 is the output spectrum of the processor for these three levels. Note that with input levels of +30 dBmV, an in-band spurious signal is visible 1.5 MHz above the desired video carrier. Fortunately, this undesired signal is highly dependent upon the input levels and when they are set to +10 dBmV it is well below the noise level. It is a well known fact that due to the delayed AGC, very little improvement is achieved in the output signal to noise ratio by feeding the processor with more than a +10 dBmV. Because of this no loss in output signal to noise ratio must be sacrificed in order to achieve sufficient "linearity" prior to removal of the adjacent channels.

This article has discussed three important requirements for signal processors if they are to function successfully in modern day headend-hub systems. Certainly 60 dB adjacent channel rejection and the capability of handling equal level adjacent channels are a necessity if the processors at the hub site are to operate without external traps and/or filters. Delay equalization, if not a necessity, is certainly desirable, especially in cases where the signal is processed twice. This by no means exhausts the list of desirable and in many cases necessary features for signal processors. Obviously, spurious free outputs from 5 MHz to 300 MHz are required if external filters are to be eliminated. IF switching options, battery power capability, standby carrier modulation capability, phase lock capability are all desirable features. The requirements I discussed were chosen because headend-hub systems emphasized these limitations in first generation solid state signal processors.

[1] Frederick E. Terman, Electronic and Radio Engineering, Fourth Edition, Sec. 11-3, pp. 379-381.

[2] Tektronix, Nomographs for Measuring Relative Delay and Amplitude Using Modulated 20T Pulse for 625/50 Standards.

[3] T. M. Gluyas, Jr., "Television Transmitter Considerations in Color Broadcasting." RCA Review, September 1954, pp. 312-334.

STATUS MONITORING SYSTEM

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Cable Dynamics Inc.
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Leo Hoarty
Buckeye Cablevision
Toledo, Ohio

One of the most pressing problems in the CATV industry today is the lack of trained technical personnel.

With the wiring of major market systems, size has become a major consideration as it relates to system maintenance.

The major market system demands that "normal" system maintenance has now become "super maintenance" and to try and supply technical competence simply by numbers as relates to personnel required to maintain a 1,000 mile system using the same formulas as required to maintain a 100 mile system will not work.

Therefore, it was felt essential that electronics become the maintenance tool of the major system.

System status monitoring can be one of the answers to the manpower crunch.

The two items that seem to be in shortest supply in the CATV industry are money and trained technical personnel.

These two problems are magnified exponentially when you start talking about major market systems, particularly major market systems that exceed 2,600 miles of electronics and four hub locations.

A study was made of the very <u>few</u> systems that fall into this category and/or are planning to be in this area of magnitude. Based on this study, it was determined that you could anticipate a minimum of one "maintenance technician" for every 200 strand miles of plant. This technician's sole function is that of trunk line maintenance.

Using this as a (conservative) estimate, it was felt that the time had come to start using electronics instead of manpower to insure constant maintenance of CATV systems.

It seemed apparent that if we could eliminate the need for some of this (almost impossible to find) manpower and possibly save money over the long run that we would be doing our industry a good service.

CDI, at the request of one of its major clients, Buckeye Cablevision in Toledo, Ohio, (actually it was more akin to insistence) started working with several manufacturers to try and achieve a <u>positive</u> status monitoring system. Jerrold Electronics was chosen as the prime contractor to work on the proposed system.

As in many pioneering efforts, early attempts were shot full of arrows, however, it was determined that status monitoring is:

1. Technically feasible.
2. The cost could be maintained at a reasonable level.
3. That it could replace manpower.
4. That in addition to system monitoring, it could be used as an outstanding maintenance tool.

It was determined that by the use of a status monitoring system, Buckeye Cablevision could eliminate a minimum of eight maintenance technician slots in their Toledo, Ohio system. This amounted to an approximate savings in salaries alone of over $100,000 a year. In addition, it was felt that the cost savings realized in not having to equip these eight men for trunk line maintenance would save an additional $120,000.

The cost to implement the status monitoring system for Buckeye Cablevision has now been reduced to below $130 per strand mile of system, and that the total cost for status monitoring will not exceed $200,000 for the entire system. It does not take an accountant to see that the status monitoring system was not only economically feasible, but almost a necessity.

In addition to the dollar savings and manpower availability problems, the status monitoring system is an excellent diagnostic tool and offers almost instant response to any system problem, which of course, has an incalculable value to any major market system. CATV in a major market must offer outstanding picture quality as well as outstanding reliability, and a system that is self-monitoring assures the latter.

The status monitoring system that was developed by Jerrold and ourselves is a relatively simple system. Its simplicity is arrived at in that our information is not encoded.

The status monitoring transmitters are installed in selected mainline stations (usually in every third amplifier and preferably following an AGC station and at the extremity of each trunk line). These selected stations have the capability of indicating the actual levels (±.75db) of one low band and one high band carrier. We have found that we can maintain this accuracy throughout a temperature range of -40°F to +95°F.

The status monitoring transmitters themselves require 50 milliamps at 27 volts, and measure the carrier levels at 65.25 and 199.25 MHz. (The measured carriers can be changed to fit system requirements). This is accomplished by developing a DC voltage that is proportionate to the amplitude of the selected carriers. This voltage in turn controls the frequency of an audio tone. Since there are two selected downstream carriers, each has an assigned audio tone (451.6Hz and 43.6Hz). The DC voltage developed from the downstream carrier is fed into an integrated circuit voltage to frequency converter. Variations in the amplitude of the downstream carrier therefore result in the variation of the frequency of its comparison audio tone. The output of the status monitoring transmitter is a crystal controlled RF carrier generator in the 4 to 6 MHz region. The two audio tones are applied to a varicap which is part of the frequency determining network and operates in conjunction with the crystal. Thus the FM modulation system consists of a carrier and two audio tones.

The concept of using frequency modulation in the return path removes any sensitivity to amplitude variation caused by gain changes in the return path. (An extremely important feature). It is important to note as mentioned before, there is no encoding nor is there any interrogation responder relationship between the status monitoring transmitter located at the selected trunk line stations and the status monitoring receiver. The status monitoring transmitters are all "talking" simultaneously and carry back only analog information.

The status monitoring receiver consists of an FM receiver whose first mixer injection is frequency controlled by reference to a 4kHz precision carrier and a divide by N counter. (Drawing I)

The recovered audio tones from the discriminator in this frequency stepped receiver passes through a frequency to voltage converter, and these recovered voltages operate the two meters in the front panel on the status monitoring receiver. These meters indicate the discrete carrier level of the two selected downstream carriers. The system is specified to an accuracy factor of ±.75db. This accuracy is of course as good as most of the meters currently available and in general system usage.

The digital portion of the system starts from a micro processor which developes three octaves of binary coded decimal numbers. This micro has three manual inputs. The first input limits the lowest three digit number the micro will seek. The second sets the high limit and the third can be set to provide continuous readouts of any specific mainline station location. Normally the micro scans from the three digit number on the low limit to the three digit number on the high limit.

The rate of scan in the system is adjustable for a matter of operator preference and is not a function of system performance. The three octaves of binary coded decimal information developed by the clock and within the preset limits controls the operation of the divide by N counter in the receiver local oscillator loop. Under control of the clock in the divide by N counter, the receiver local oscillator scans in frequency from the low preset limit to the high preset limit in 4kHz steps. These steps are accurately controlled because of the phase lock reference to a precision 4kHz source. It is in the receiver that a specific identification to a given status monitoring transmitter is generated.

The indicator path through all stations is identical. The only way one station can be distinguished from another is by its incoming carrier frequency. As the receiver steps in 4kHz increments, the three octave digit display indicates the transmitting location being monitored.

Each status monitoring transmitter in the system has its own assigned RF carrier frequency. Because of the nature of the scanning, which is sequentially numbered, should there be a gap in the numerical assignment of stations, the receiver will scan past and will indicate an alarm as it scans the unassigned station.

There are five alarms normally displayed on the front panel of the status monitoring system. These alarms are:

1. Low band carrier level low.
2. Low band carrier level high.
3. High band carrier level low.
4. High band carrier level high.
5. No return carrier.

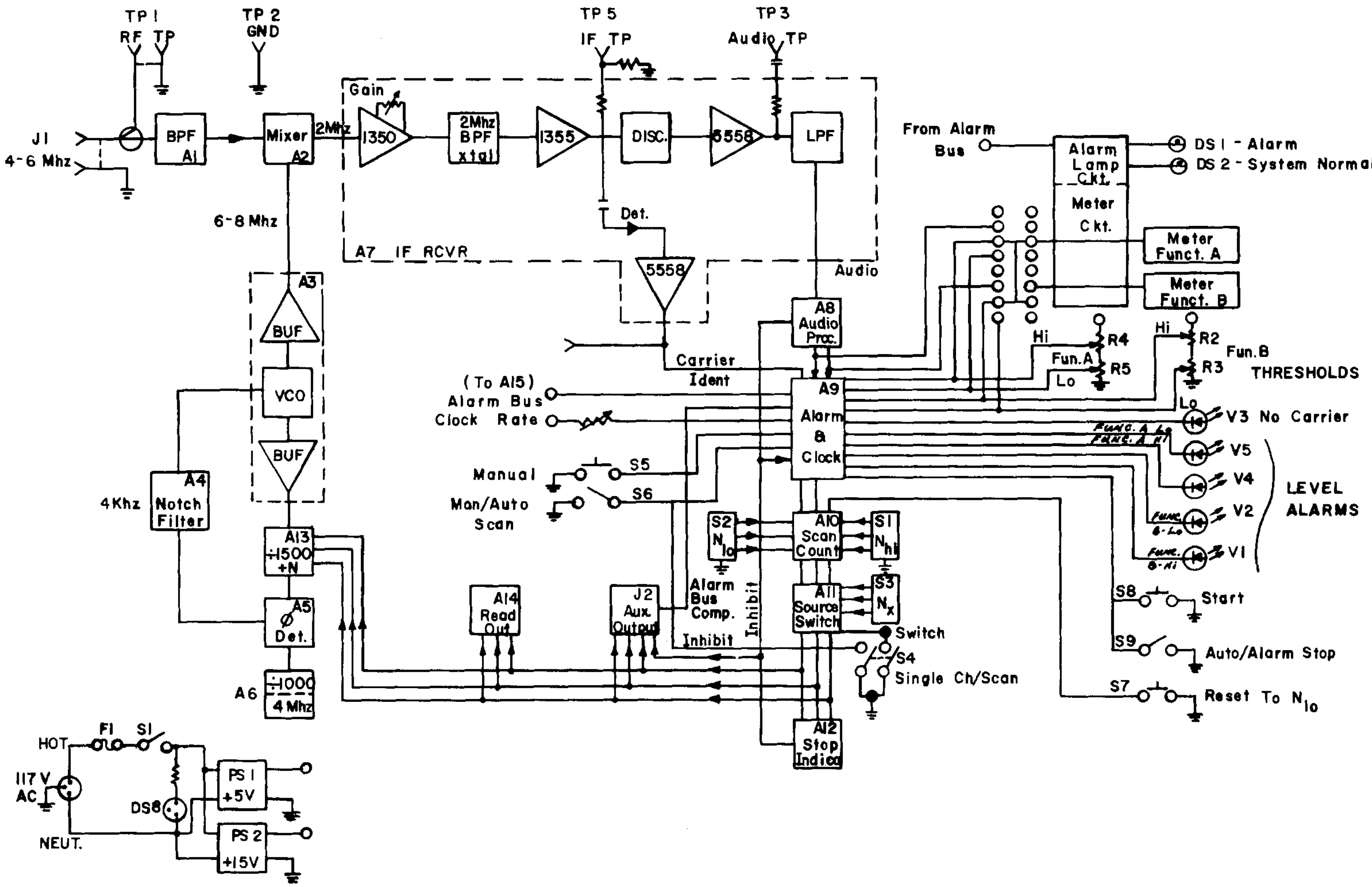
The status monitoring's receiver front panel meters indicate the absolute level of the two selected downstream carriers and their preset alarm limits. The purpose of these limits is of course to preclude arbitrary alarming at some random point in the system. The operator can adjust the low and high level points at which an alarm will be generated. The return carrier level is not measured and the status monitoring receiver merely senses its presence (or lack thereof). The purpose of this last alarm is to avoid operator problems which could be created by false downstream alarms caused by misadjustment or failure of the upstream or return path system. Of course downstream and upstream information could both fail due to mechanical cable separation and/or power pack failure in an individual amplifier. The exact location of this of course can be determined by the status monitoring receiver.

In addition to the receiver, it was felt that an active display would facilitate the use of untrained personnel as "system monitors". Therefore, CDI has developed a system that goes from the transmitter to a map display of the entire system. (It is interesting to note that the map display in a 1-600 scale covers and area in excess of 15 x 13 feet). [Drawing II]. This analog information from the transmitter is transferred to a logic board. The logic board (Drawing III) operates as binary coded decimal information center which singles out and identifies each amplifier location as the receiver it is scanning. A light emitting diode with a time delay latch will be used to indicate an out of specification transmitter location. Of course the map operates at the same scanning rate as the receiver so any trouble in the system can be located immediately as the logic board used as the interface between the receiver and the readout board is designed to lock on an amplifier that is out of its specified range.

This will immediately alert the operator that a condition out of specification is existing. He can then determine that condition immediately by the use of the status monitoring receiver. In addition to the above, CDI is currently working on a system that will give a digital and printed readout of the amplifier number and exact street location in the system. The advantages of this technique are of course obvious, and in addition, we are working on the interface that will be fed directly to a computer center that will allow permanent data storage for all system failures and variables that should be of great assistance in future design concepts.

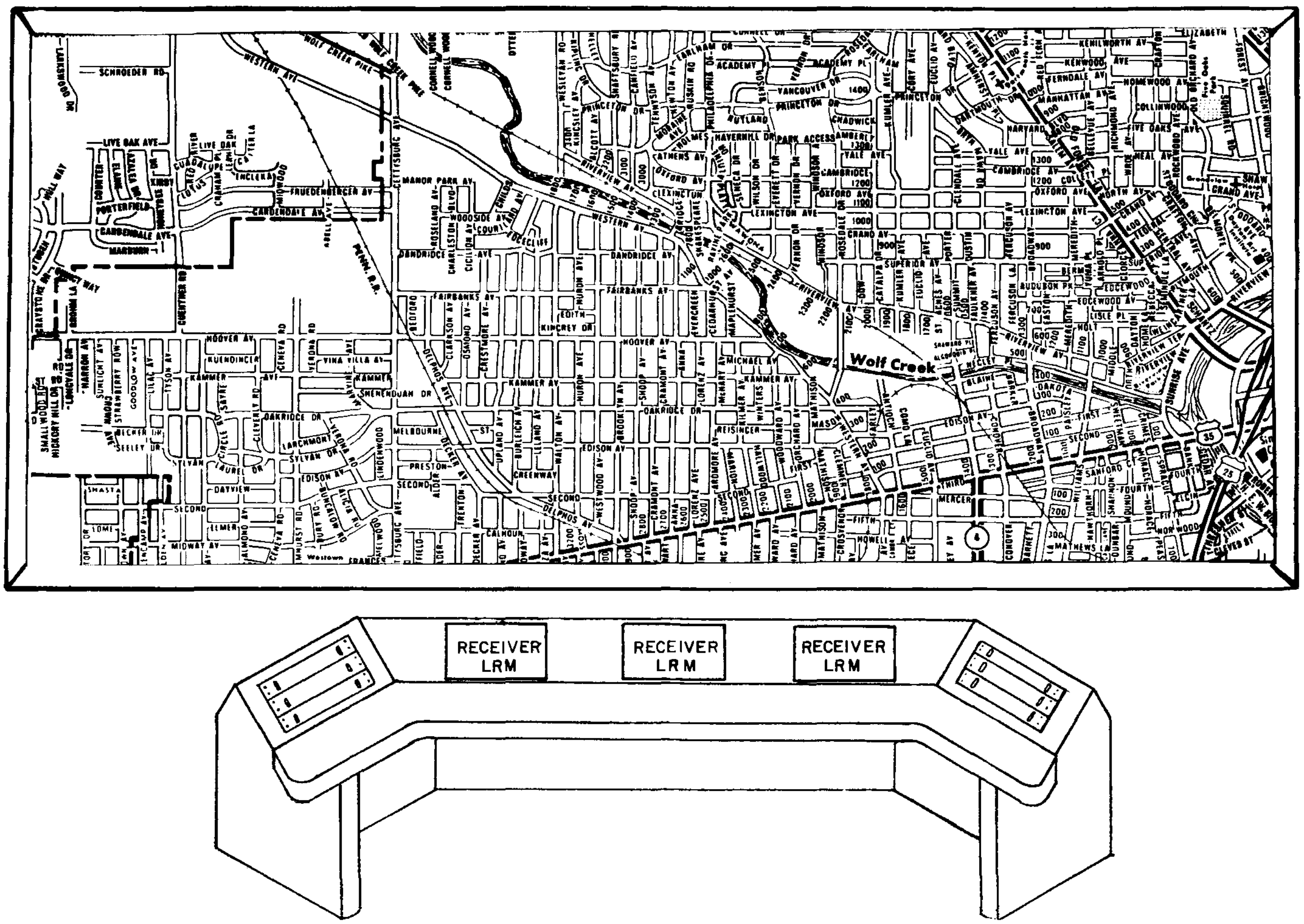
As in all systems, some items were compromised. However, we feel that this system as a technical trade off to dollar cost has very few shortcomings. The principle one is that the transmitter unit is mounted inside the standard mainline station. Since generally there are three different devices fighting for the same space in the mainline housing (downstream AGC, upstream AGC and the status monitoring transmitter), we are restricted to locating a status monitoring transmitter at approximately every third mainline station. We are currently working on an auxiliary housing arrangement for status monitoring transmitters, so that we can eliminate this problem and place the transmitters at any predetermined point in the system. This of course will allow any CATV system (with bi-direction capability) to insert and use the status monitoring system.

It is Cable Dynamics' opinion that a status monitoring system technique is an absolute necessity in major markets and as more status monitoring techniques are developed, it could become economically feasible for even the smallest system.

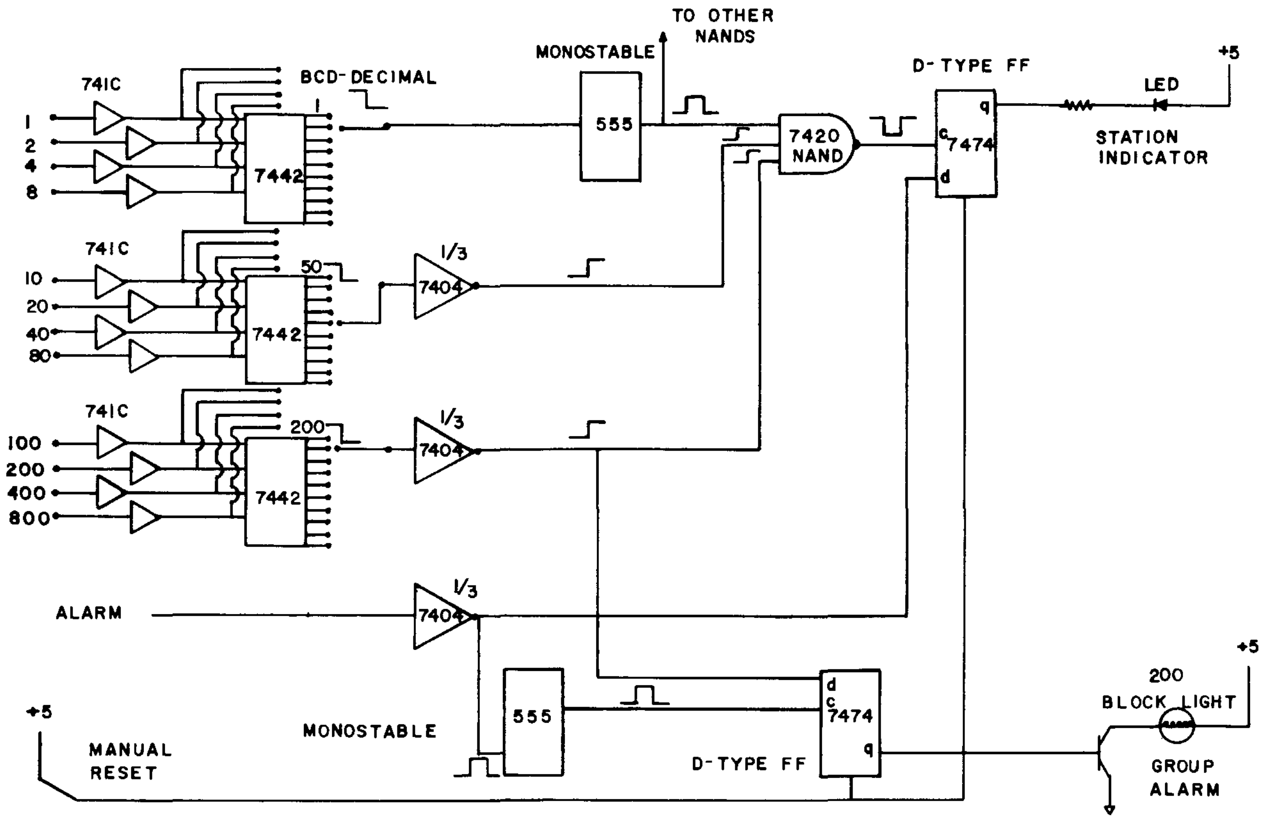


STATUS MONITOR RECEIVER

MODEL LRM



STATUS MONITORING
CONTROL PANEL



LOGIC BOARD INTERFACE

SYSTEM CONSIDERATIONS AFFECTING ANTENNA PREAMPLIFIER DESIGN

By Frank Pennypacker, P. Eng. and Emanuele DiLecce, P. Eng.

Lindsay Specialty Products Ltd.

This paper presents the results of a study conducted as part of a preamplifier design project. It was considered important that the preamplifier being designed be the best combination of performance to suit system operator requirements. These results will also be of great interest to the system operator to aid him in choosing a suitable preamplifier.

The emphasis was placed on system considerations, that is how the performance of the preamplifier affects the overall performance of the system.

Antenna preamplifiers are used to improve the signal to noise ratio of weak signals. It is hoped that an overall improvement in signal to noise ratio will result for the entire system. The antenna preamplifiers can be characterized numerically by certain measurable parameters. These are: noise figure, overload capability, gain, and match. Not as easily measured is reliability.

The characteristics of the preamplifier are not the only factors affecting a particular system performance parameter. The signal to noise ratio of the system is affected by noise received by the antenna plus the noise figure of the head-end processor, and the noise figures and number of distribution amplifiers. Beat products and cross modulation can be generated by the antenna preamplifier and also any other active device of the cable system. For the purpose of our investigation, we have assumed values that we consider typical of the rest of the system.

Before we launch into a detailed discussion of antennas, preamplifiers and noise figures, let us talk about thermal noise. Noise is generated in any resistive device, due to the random motions of the electrons caused by the heat contained in the device. This noise power is proportional to the temperature in absolute degrees, the resistance and the bandwidth. It is a common misconception that a receiving antenna generates a noise level of -59 dBmv. This corresponds to the noise that would be generated by a 75 ohm resistor at room temperature. In the case of the antenna, the noise developed across its terminals is not being generated by the antenna, but rather, is a transformation of the electro-magnetic fields being received by that antenna. The antenna is composed of metallic elements with extremely low resistance and insulators with extremely low conductance. The impedance of the antenna is not associated with the actual elements of the antenna but rather with the space to which the antenna is coupled by the electro-magnetic fields.

The noise received by an antenna is composed of a certain amount of radiation generated by the earth, which is at approximately room temperature; atmospheric noise and radiation from outer space. The amount of noise received will depend on the pattern of the antenna and the direction in which the antenna is pointed. An antenna pointed down will receive more radiation from the earth, and an antenna pointed up will receive less. The upward pointing antenna will, of course, receive more radiation from space. The closest source in space is our sun. The amount of radiation from the sun depends on the direction and time of day. In other words, the noise is greatest when the antenna is pointed more or less directly at the sun. In certain directions the radiation from outer space can be extremely low; the temperature of intergalactic space appears to be about 3 degrees kelvin. However, the center of our galaxy is an extremely strong source of energy and the antenna pointed in that direction will receive a large amount of radiation.

Figure 1 shows graphically the relationship between noise level and frequency. Notice that the lower frequencies have a greater noise then higher frequencies. Expecially note, that the noise at the frequency of channel 2, is considerably larger then the often quoted -59 dBmv. It has often been said that the low band is noisy, this is one reason why. Figure 2 shows the noise received by an antenna pointed well above the horizon throughout a 24 hour period. The peak represents a time when the antenna is pointed at the center of our galaxy. The dip represents a time when the antenna is pointed at the very cold regions of intergalactic space. The height and width of the peaks will be affected by the beamwidth of the antenna. The antenna which has a narrower beam width will cause a higher, narrower peak.

The above does not take into consideration the affects of man-made noise. This can come from such sources as electric power distribution systems and ignition noise from vehicles. This is extremely variable and unpredictable, but, we have a certain degree of control over it, and can do

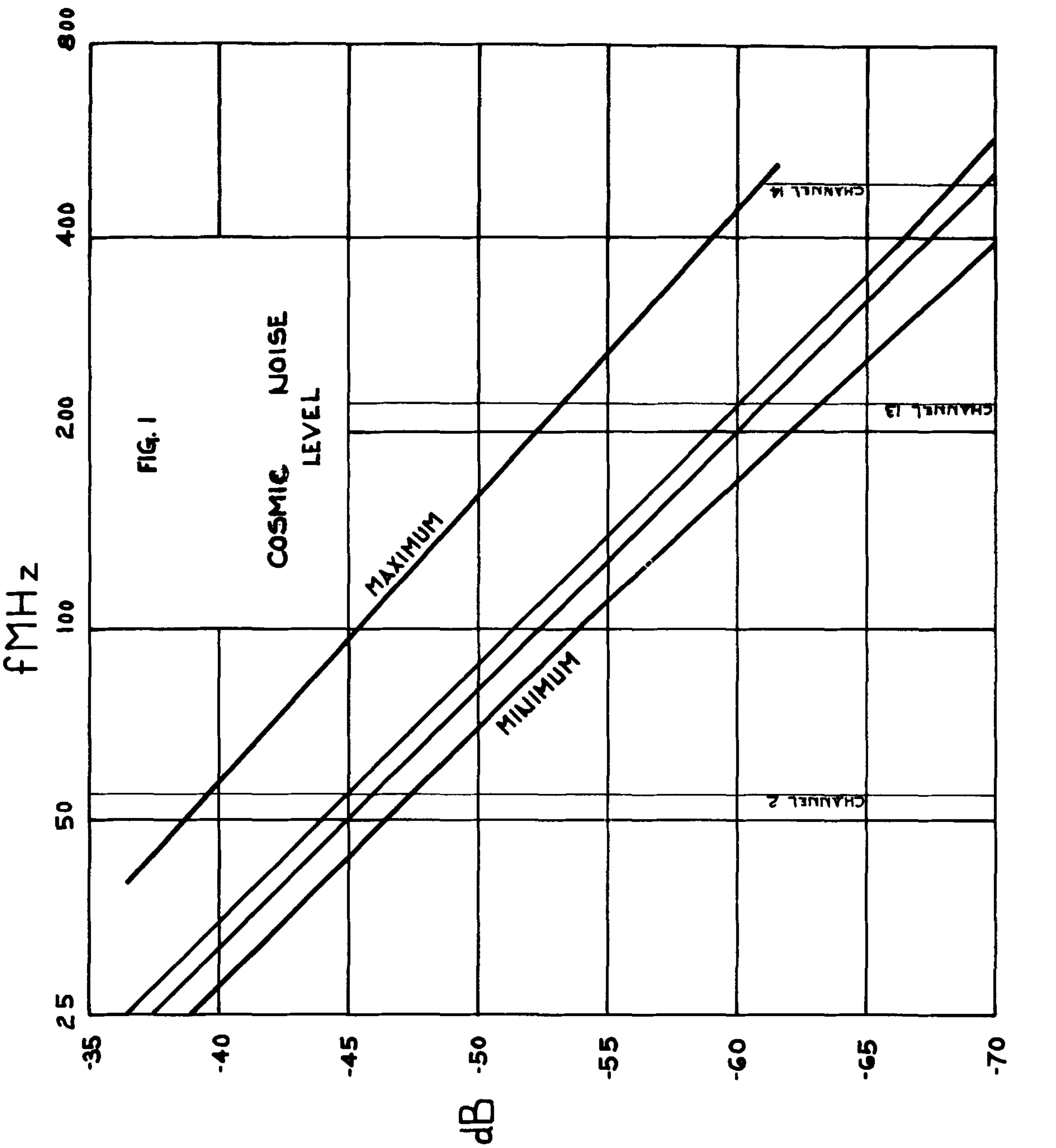


FIG. I
COSMIC NOISE LEVEL
MAXIMUM
MINIMUM
CHANNEL 14
CHANNEL 13
CHANNEL 2
fMHz
25
50
100
200
400
800
dB
-35
-40
-45
-50
-55
-60
-65
-70

something to reduce it.

As an aid to analysis and understanding, we are going to define a quantity called Delta signal to noise ratio and abbreviate it ΔSNR. This is a quantity which we will express in dB; it represents the improvement in signal to noise ratio which will be realized by using a particular kind of preamplifier. We define 0 dB ΔSNR as the signal to noise ratio which would result if the antenna were directly connected to the input of the head-end processor by a lossless cable. Thus, if the antenna is connected to the processor by an actual cable having a certain amount of loss the ΔSNR will be negative and equal to the loss of the cable. For that reason the actual improvement realized by using the preamplifier will be equal to the ΔSNR with the preamplifier minus the ΔSNR without the preamplifier. Note that ΔSNR is the improvement of signal to noise ratio which will be realized at the processor output.

By comparing ΔSNR for various situations, we can determine how much improvement in signal to noise ratio has been made by changing system parameters. We can determine how closely we have approached the maximum possible signal to noise ratio and what we will have to do to approach more closely. This can all be done with much less calculation and complexity than working with signal to noise ratio directly. After the system parameters are determined, the actual signal to noise ratio can be calculated by well known techniques.

Figure 3 shows the relationship between noise figure and ΔSNR for various levels of received antenna noise. Note that for -65 dBmv received noise, about the best that will be encountered at Channel 13, the affect of antenna preamplifier noise is quite large; much larger then one would expect, assuming that well known -59 dBmv. The change in noise figure, 1 dB to 2 dB, results in more than 2 dB change in signal to noise ratio. For -60 dBmv received noise, which is almost equal to -59 dBmv, the change is almost 1 dB for 1 dB. Let us look at the case of -45 dBmv, which is about average for Channel 2, here the change from 0 dB noise figure to 4 dB noise figure results in only 3/10 of a dB change in signal to noise ratio. This is quite an insignificant amount.

We can conclude that for low band channels, the noise figure of the preamplifier is relatively unimportant, but for high band channels it is of significant importance. This is a particularly unfortunate situation since, low noise figures are much easier to achieve at lower frequencies.

This discussion on noise figure would seem to indicate that for low band signals, the preamplifier is of no use. Let us investigate the effect of not having gain at the antenna. Figure 4 is the relationship of gain to ΔSNR, for an antenna noise level of -45 dBmv. For 5 dB loss between the antenna and the processor ΔSNR has a value of -1.2; for 4 dB noise figure and 15 dB of gain ΔSNR is +0.4. The total change is 1.6 dB. Here to simplify calculations and presentation, we have subtracted the loss between the preamplifier output and the processor from the gain of the preamplifier and called it effective gain. It appears that an effective gain of 15 dB results in a negligible degradation in ΔSNR compared to what would be realized with infinite gain. Figure 5 shows the same relationship for an antenna received noise of -55 dBmv and a processor noise figure of 7 dB. Note that a 15 dB effective gain results in only about 3/10 dB worse signal to noise ratio than would result from 30 dB of gain. We consider this 3/10 of a dB to be insignificant. Figure 6 shows the relationship for an antenna received noise of -65 dBmv.

In the case of the preamplifier with 2 dB noise figure, the difference between 15 and 30 dB of gain represents a change in signal to noise ratio of 6/10 dB. These figures show that 15 dB effective gain is sufficient for the conditions just described. These conditions are typical for reception of VHF signals. If we allow for 5 dB of cable between the antenna and the processor and the use of a hybrid splitter to run two processors off one antenna, this means that the preamplifier should have a minimum gain of 24 dB.

In the case of UHF reception, the preamplifier is the major source of noise. At these frequencies the present state of the art transistors have much higher noise figures and received antenna noise is much lower. We have assumed a received noise level of -65 dBmv and a processor noise figure of 14 dB. The results we presented in figure 7. We again see that an effective gain of 15 dB is adequate allowing for a 10 dB down lead loss we find that 25 dB is the minimum gain.

The problem of overload by high signal levels is one that has all too often been ignored by preamplifier designers. If the designer ignores the problem, the man who uses the preamp and has to suffer with it, can't. There are preamplifiers on the market with noise figures of 1 to 2 dB, which overload and produce noticeable cross modulation and beats with inputs of only 10 to 20 dBmv; in many cases these amplifiers are not satisfactory. To determine what would be satisfactory, we have made a statistical study of the maximum levels encountered in CATV head-ends. Figure 8 shows the results of this study. Here we have presented the probability of finding any particular signal level. Here are probabilities for both high band and low band. Note that there really isn't much difference. The curves indicate that a preamplifier which will not overload with 26 dBmv input will handle 95% of the head-end requirements; the preamplifier which will handle approximately 33 dBmv input without overloading will handle 99% of the requirements.

These curves were made by taking a sample of ten head-ends randomly selected throughout North America. For each head-end the maximum signal level on the high band and the maximum signal on the low band was noted. We then assumed a normal distribution curve and computed a mean and standard deviation. It is from these values that the curves were generated. Admittedly, this

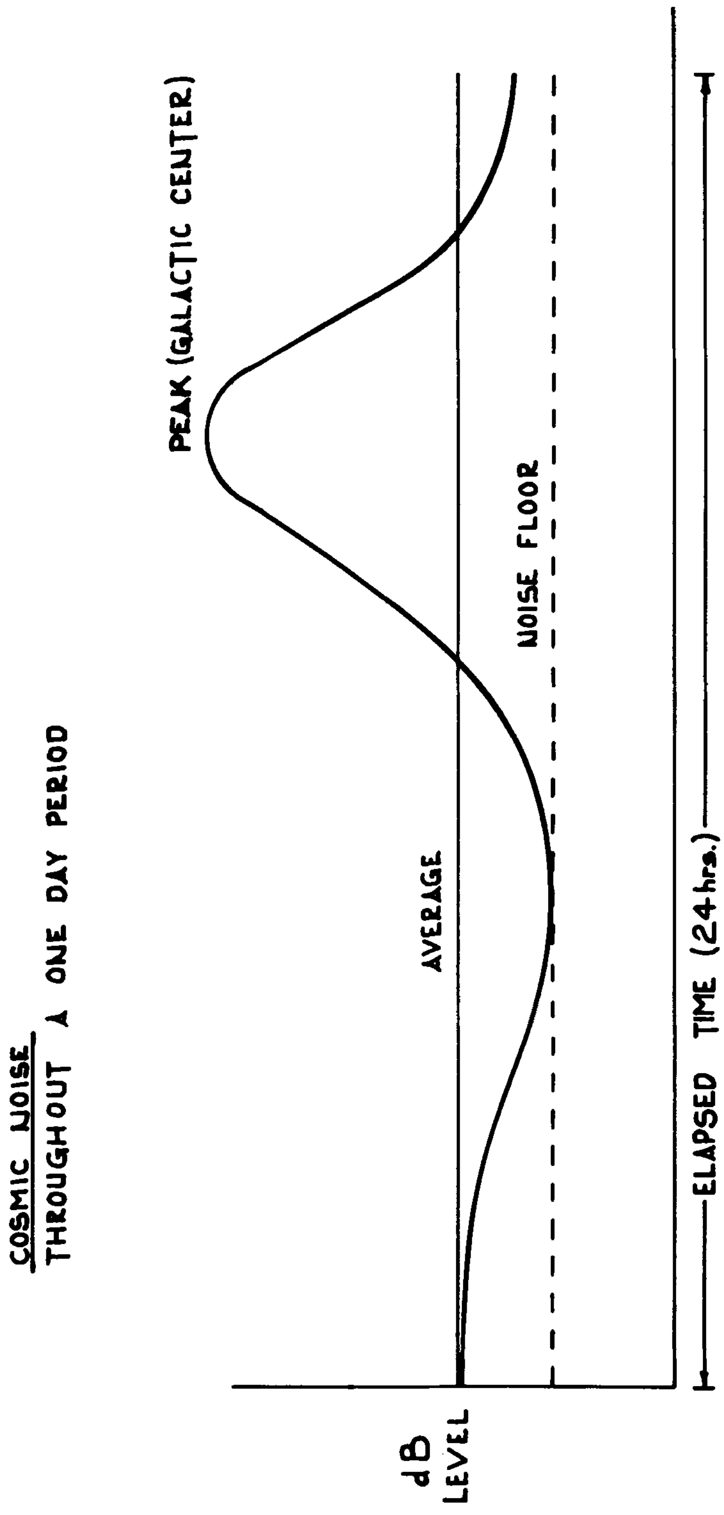

COSMIC NOISE
THROUGHOUT A ONE DAY PERIOD
PEAK (GALACTIC CENTER)
AVERAGE
NOISE FLOOR
dB
LEVEL
ELAPSED TIME (24 hrs.)

FIG. 2

is rather a small sample, but our experience in doing signal surveys of hundreds of head-ends indicates that the numbers are reasonable. In computing these curves no allowance has been made for rejection of unwanted signals by phasing antennas or other such methods. Depending on relative angles of the wanted and unwanted signals, it may not be pssible to create any rejection or the rejection might be as high as 30 dB. The curves then represent a worst case situation.

We conclude that a preamplifier that will accept +40 dBmv inputs without over-loading will handle virtually all signal conditions that would normally be encountered. In fact such a preamplifier will eliminate a considerable amount of trouble for the Antenna installers by not forcing them to carefully phase the antennas to eliminate unwanted signals. In the case of a search antenna, which is a broad band antenna used for signal surveys or as a standby, it is imperative that the preamplifier be able to accept whatever input signals are present. This, we believe, will confirm the field experience of those people who install the antennas and produce working head-ends.

There is, of course, a trade off between achievable noise figures and achievable signal handling capability. Figure 9 shows the trade off which is achievable with to-day's state of the art. As can be seen +40 dBmv input capability will involve the loss of about 2 dB of noise figure. When extremely high levels are present it is better to accept this small sacrifice in noise figure and have a bit more snow in the screen rather than have a herringbone pattern or visible cross modulation. For the low band, where the noise figure of the preamplifier is not so critical, it is our feeling that there is no merit in building an extremely low noise figure preamplifier with poor overload characteristics. If you will recall from our previous discussion, a change in 1 dB of noise figure in the preamplifier resulted in a very small fraction of a dB change in noise figure for the entire system. We have encountered numerous cases of overload which is quite a serious problem. For the high band it is questionable whether an extremely low noise preamplifier should be built. For a preamplifier with a 3½ dB noise figure the system signal to noise ratio will only be about 2 dB better than for a preamplifier with a 1½ dB noise figure. Again, we state that this 2 dB difference in signal to noise ratio will be invisible while overload distortion products will be quite visible.

For UHF channels the situation is somewhat different because of the much lower antenna noise level. In many cases there are no strong local UHF stations which would interfere with reception of more distant stations. Here, the lowest possible noise figure is called for. In other cases there are strong local stations which might interfere with distant stations. Because of the two possibilities, it appears that two different kinds of preamplifiers are required; an amplifier with high handling for cases in which there are strong local signals and a lower handling, lower noise figure amplifier for cases in which there are not strong local signals.

In the interest of thoroughness, we have also looked into the matter of the minimum acceptable return loss. Figure 10 shows time delay versus the level of reflections. As is well known, for a very short time delays the reflection can be much stronger without being visible. Superimposed on this curve showing minimum conditions, is a curve of calculated conditions for what we consider typical worst case. We have assumed a processor with a 16 dB return loss on the input and a preamplifier with 16 dB return loss on it's output; the cable is standard .412 diameter with 1.7 dB of attenuation at 216 MHz and .85 dB attenuation per hundred feet at 54 MHz. Relative velocity was assumed to be 81%. The worst case occurred with about three hundred feet of cable. Here, there was still an 11 dB margin before a visible ghost would occur. Of course, for higher frequencies the cable attenuation would be much greater and the margin much greater also. From this we can conclude that with readily attainable levels of match, there should be no ghosting problem.

Reliability is not a matter which lends itself to a similar analysis. However, one would hope that the reliability would be such that the antenna system would not have more than one failure in several years. In order that a mean time between failures of less than three years each preamplifier will need a mean time between failures of about 30 years. With proper design such a reliability level can be achieved. However, it is not easy to measure. The only practical course is to have a large number of preamplifiers in service for several years to evaluate their reliability. It should be possible to accelerate failures by such means as vibration and temperature cycling. It will not be possible to know exactly what is the exact amount of acceleration.

Preamplifiers should be designed with the maximum possible protection from lightning by incorporating both gas discharge surge protectors and diodes to protect semiconductors. Since lightning energy is mainly low frequency the preamplifiers should incorporate filters to admit only frequencies in the band of interest. Power supplies for preamplifiers should incorporate devices to prevent surges on the power line from entering the preamplifier. All components, of course, should operate at much less than their maximum ratings. It is recommended that a redundant power supply be used. Construction should be sturdy and rigid to withstand the effects of vibration, with tuning adjustments locked by some sort of adhesive to prevent detuning under vibration and temperature cycling. Although the user of the preamplifier cannot readily measure life time or mean time between failures of a preamplifier, he can examine them to determine if the above principles have been followed.

It is hoped that this paper has improved understanding on an area that has not been well understood in the past. We hope that this will result in better preamplifiers and better usage.

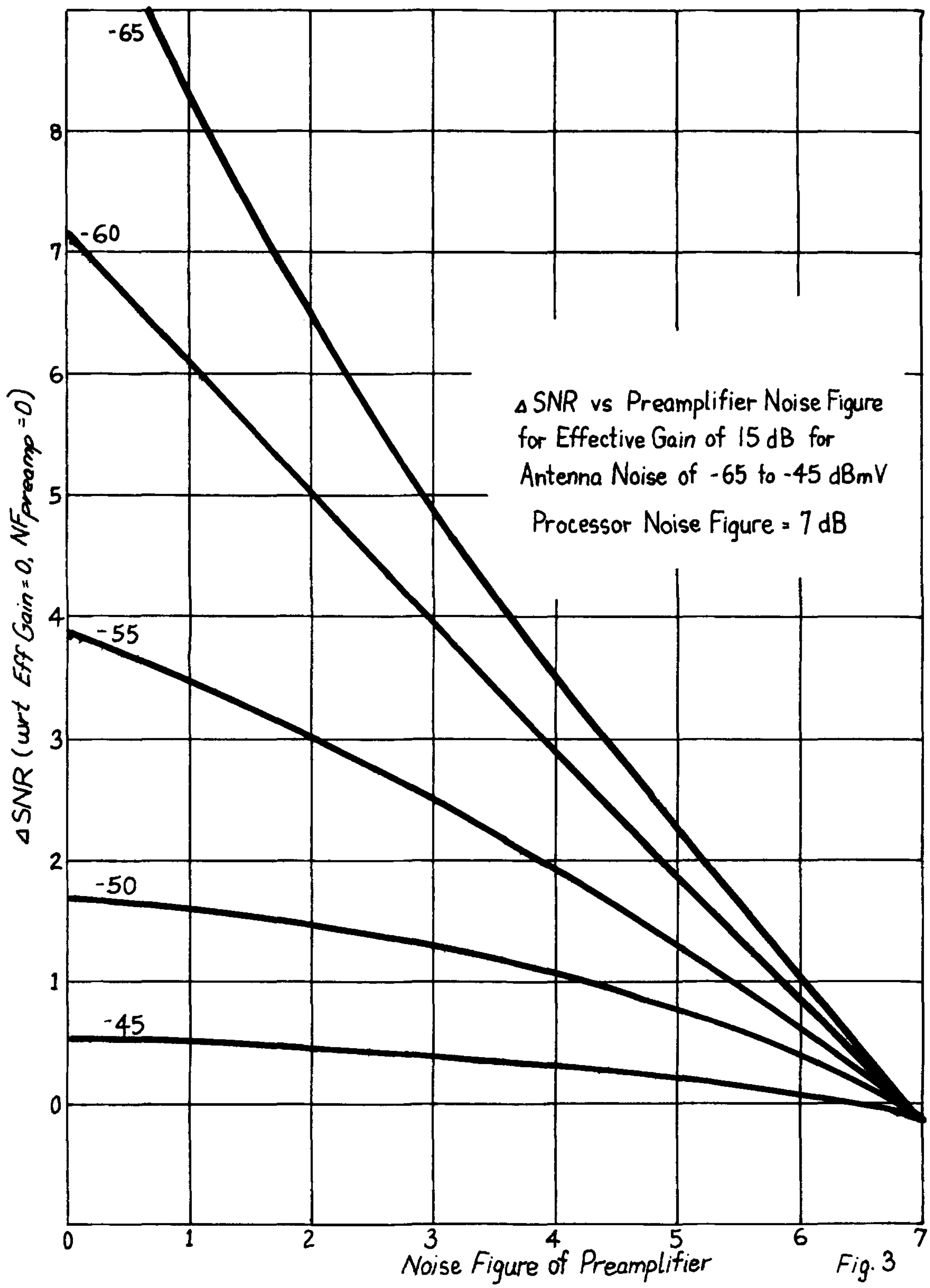

∆ SNR vs Preamplifier Noise Figure for Effective Gain of 15 dB for Antenna Noise of -65 to -45 dBmV
Processor Noise Figure = 7 dB
∆ SNR (wrt Eff Gain = 0, NF preamp = 0)
Noise Figure of Preamplifier
-65
-60
-55
-50
-45
0
1
2
3
4
5
6
7
8

Fig. 3

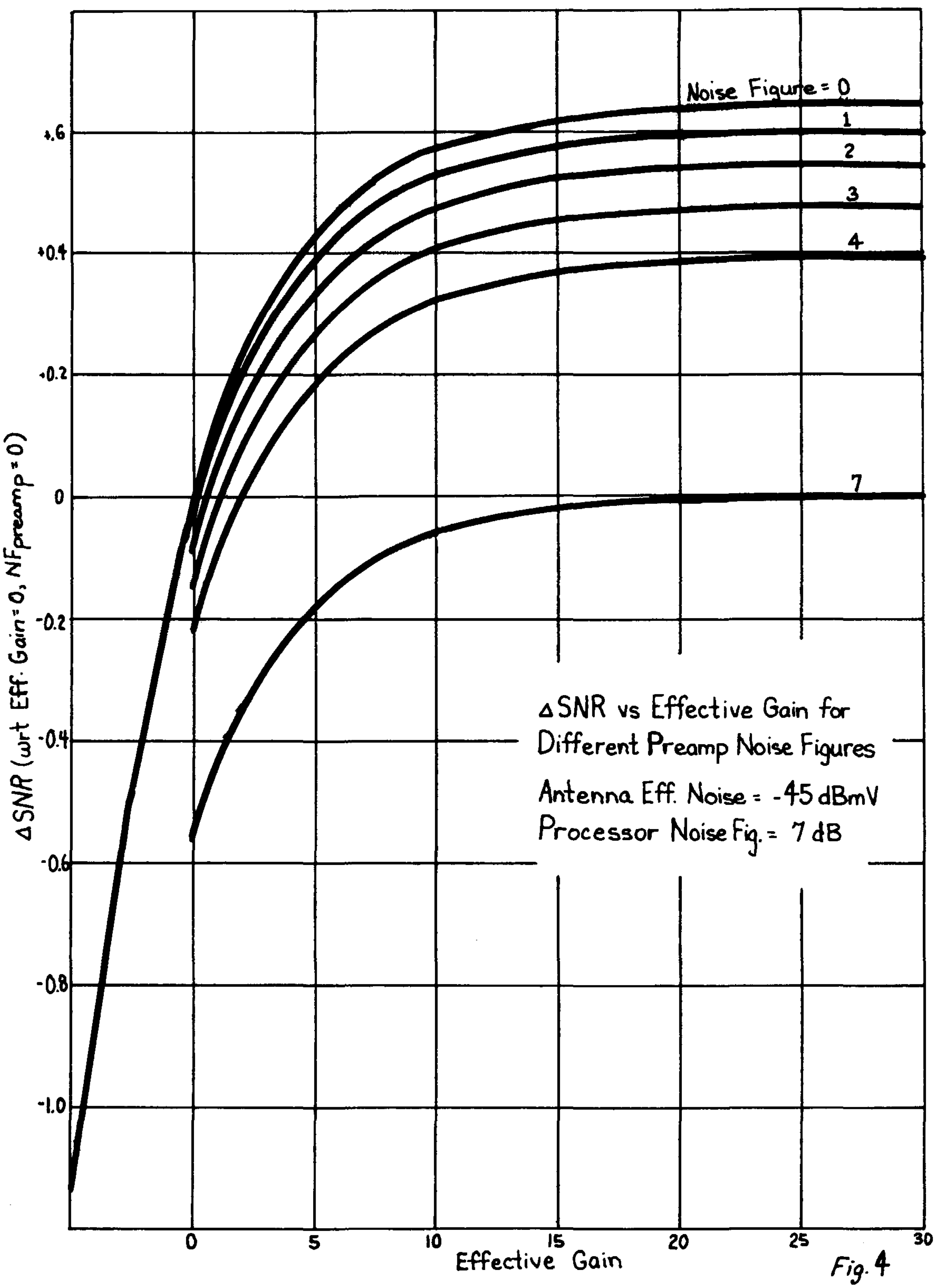
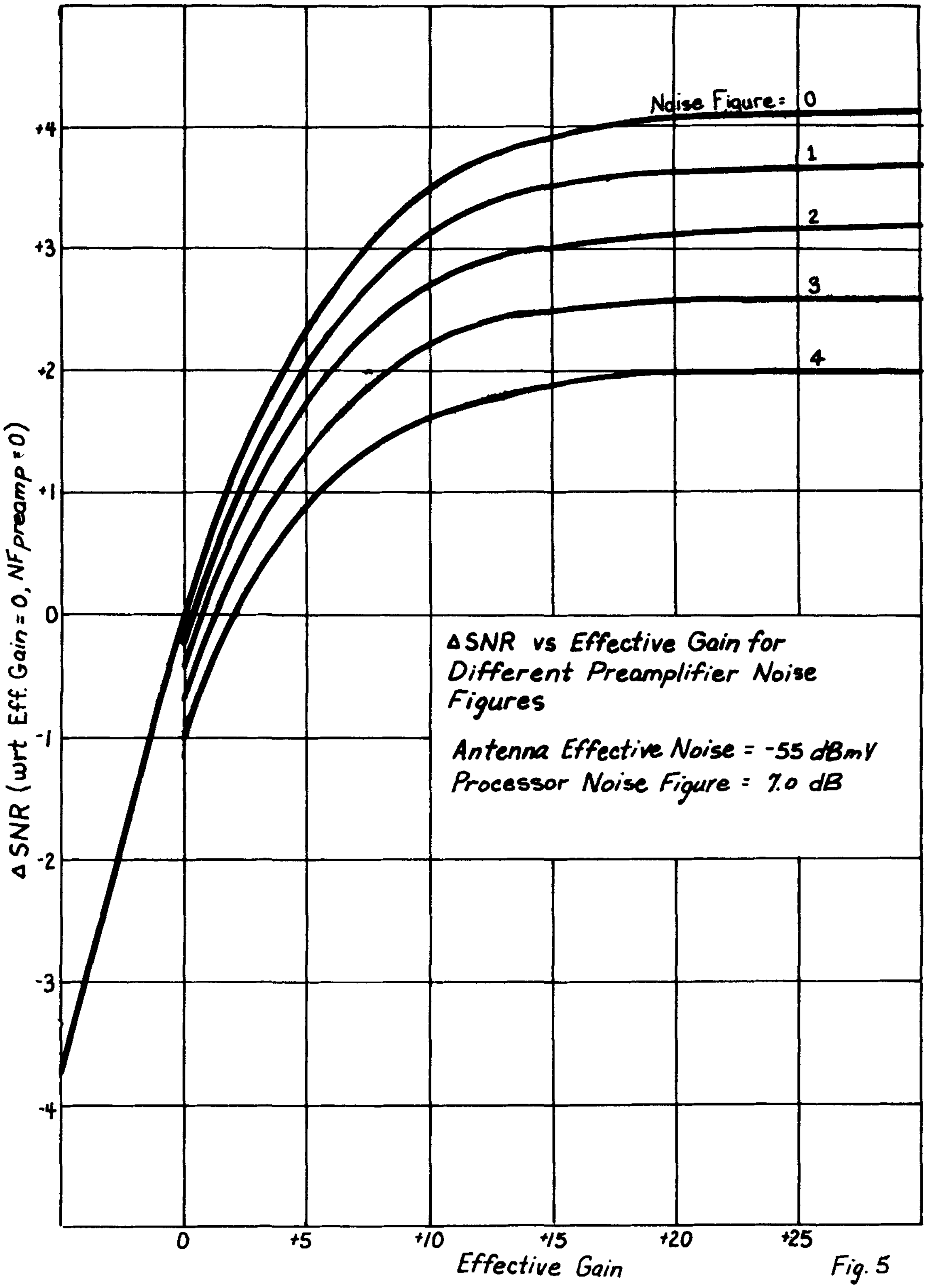

Noise Figure = 0
1
2
3
4
7
+.6
+0.4
+0.2
0
-0.2
-0.4
-0.6
-0.8
-1.0
ΔSNR (w.r.t Eff. Gain = 0, NF preamp = 0)
0
5
10
15
20
25
30
Effective Gain
ΔSNR vs Effective Gain for
Different Preamp Noise Figures
Antenna Eff. Noise = -45 dBmV
Processor Noise Fig. = 7 dB

Fig. 4


Noise Figure = 0
1
2
3
4
+4
+3
+2
+1
0
-1
-2
-3
-4
Δ SNR (wrt Eff. Gain = 0, NF preamp = 0)
Δ SNR vs Effective Gain for Different Preamplifier Noise Figures
Antenna Effective Noise = -55 dBmV
Processor Noise Figure = 7.0 dB
0
+5
+10
+15
+20
+25
Effective Gain
Fig. 5

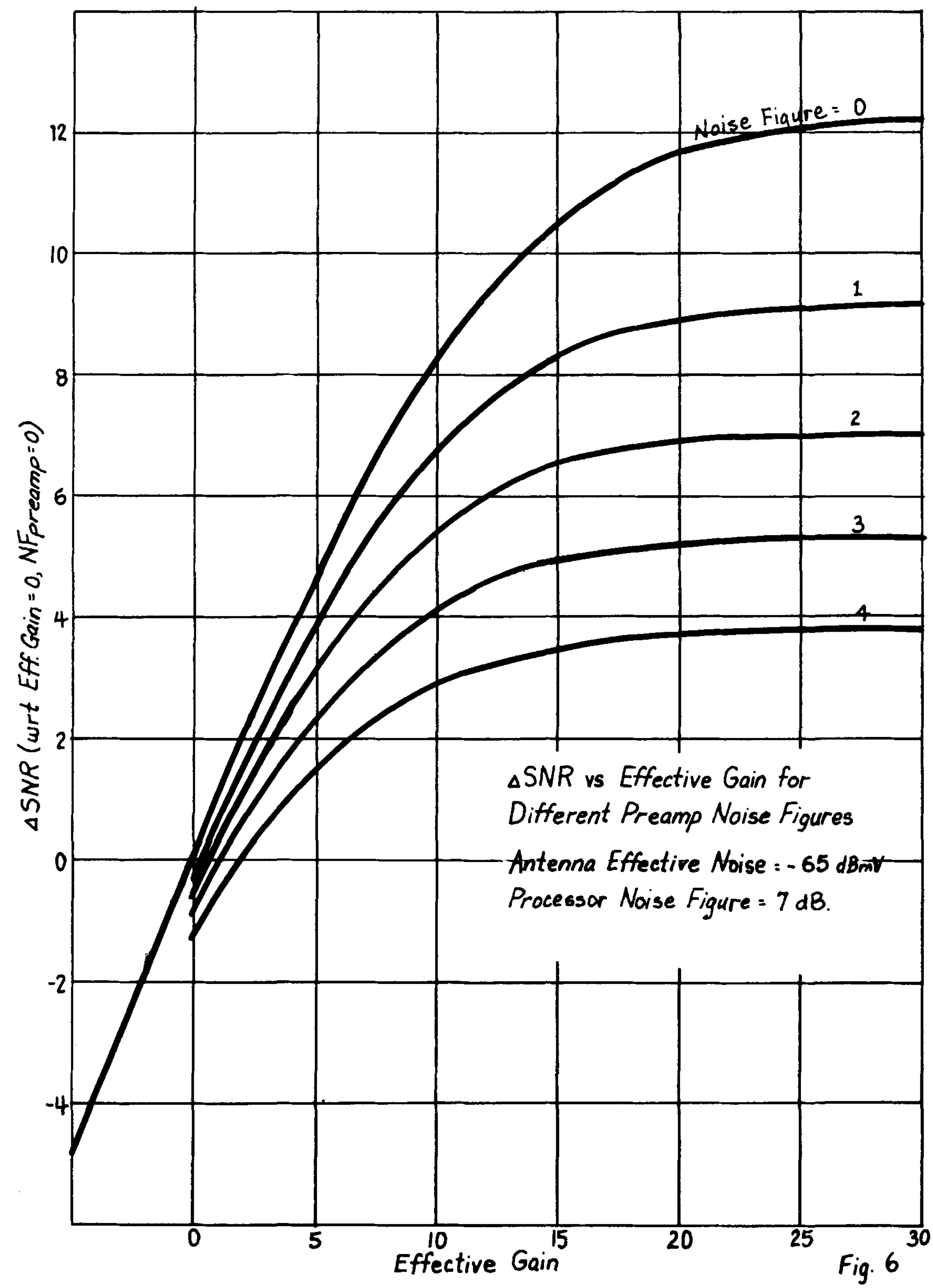
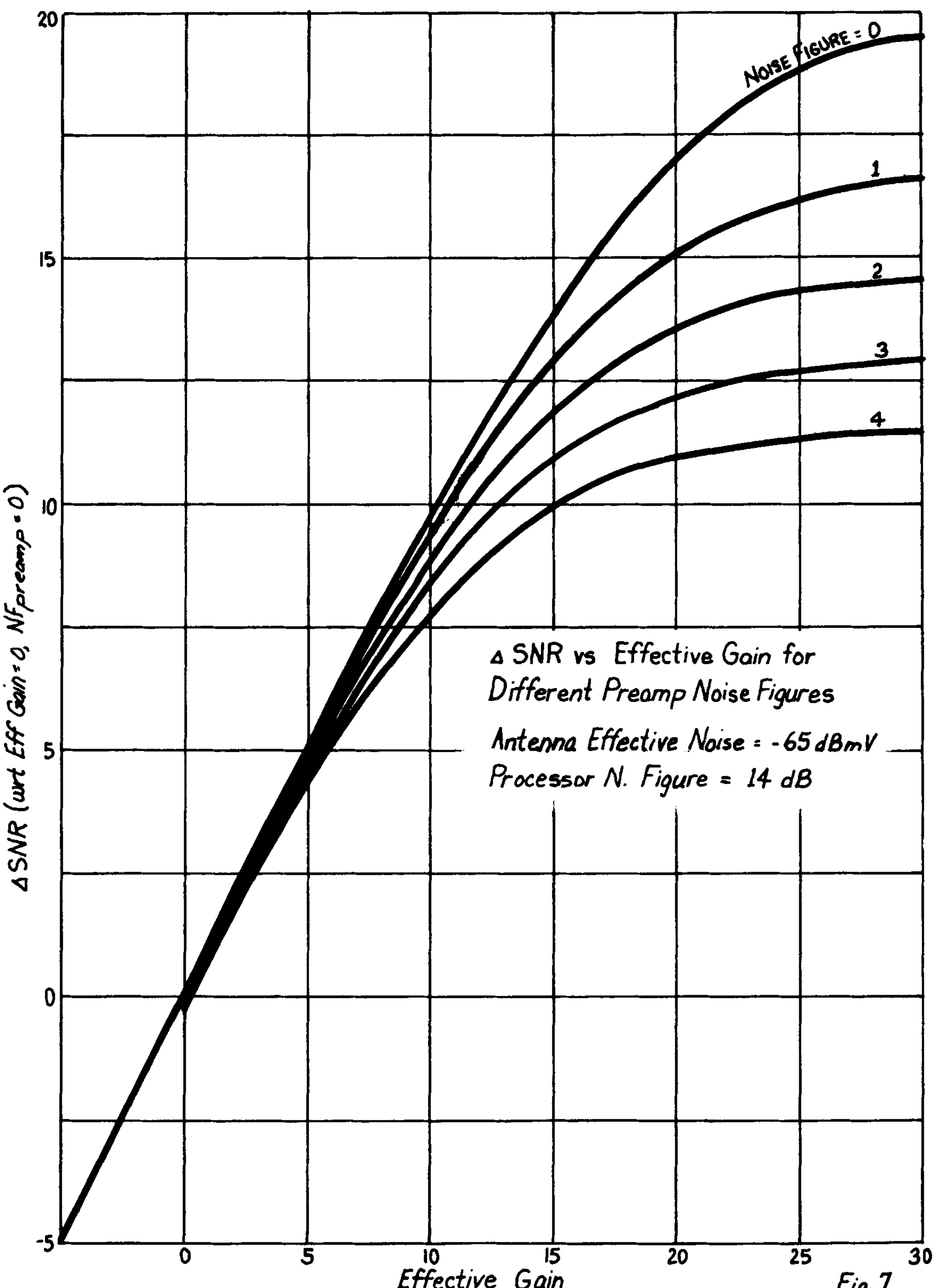
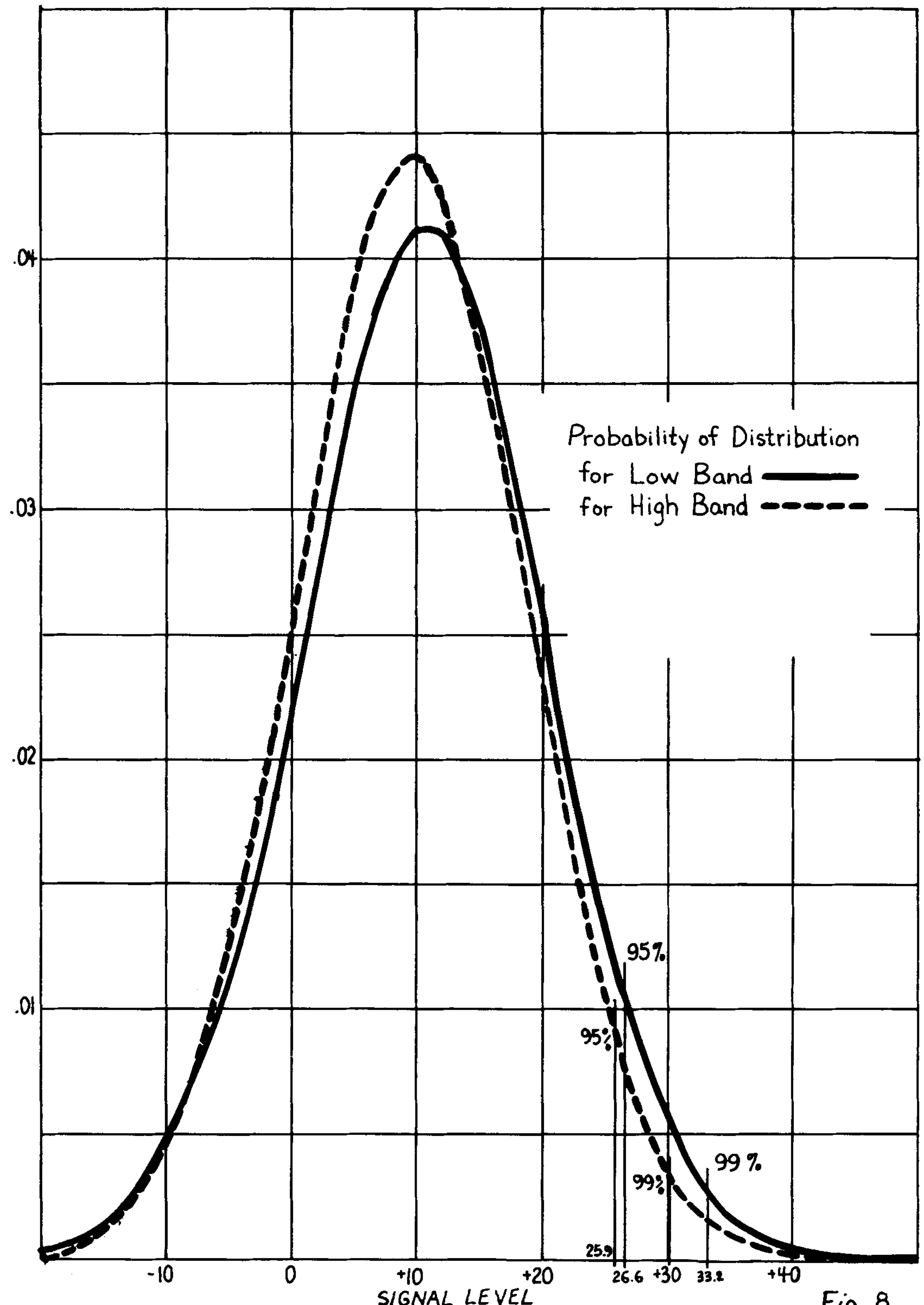

Noise Figure = 0
1
2
3
4
ΔSNR (wrt Eff. Gain = 0, NFpreamp = 0)
Effective Gain
ΔSNR vs Effective Gain for
Different Preamp Noise Figures
Antenna Effective Noise = -65 dBmV
Processor Noise Figure = 7 dB.
12
10
8
6
4
2
0
-2
-4
0
5
10
15
20
25
30

Fig. 6



20
15
10
5
0
-5
0
5
10
15
20
25
30
Effective Gain
Δ SNR (wrt Eff Gain = 0, NF preamp = 0)
Noise Figure = 0
1
2
3
4
Δ SNR vs Effective Gain for
Different Preamp Noise Figures
Antenna Effective Noise = -65 dBmV
Processor N. Figure = 14 dB

Fig. 7



.04
.03
.02
.01
Probability of Distribution
for Low Band
for High Band
95%
95%
99%
99%
25.9
-10
0
+10
+20
26.6
+30
33.2
+40
SIGNAL LEVEL

Fig. 8

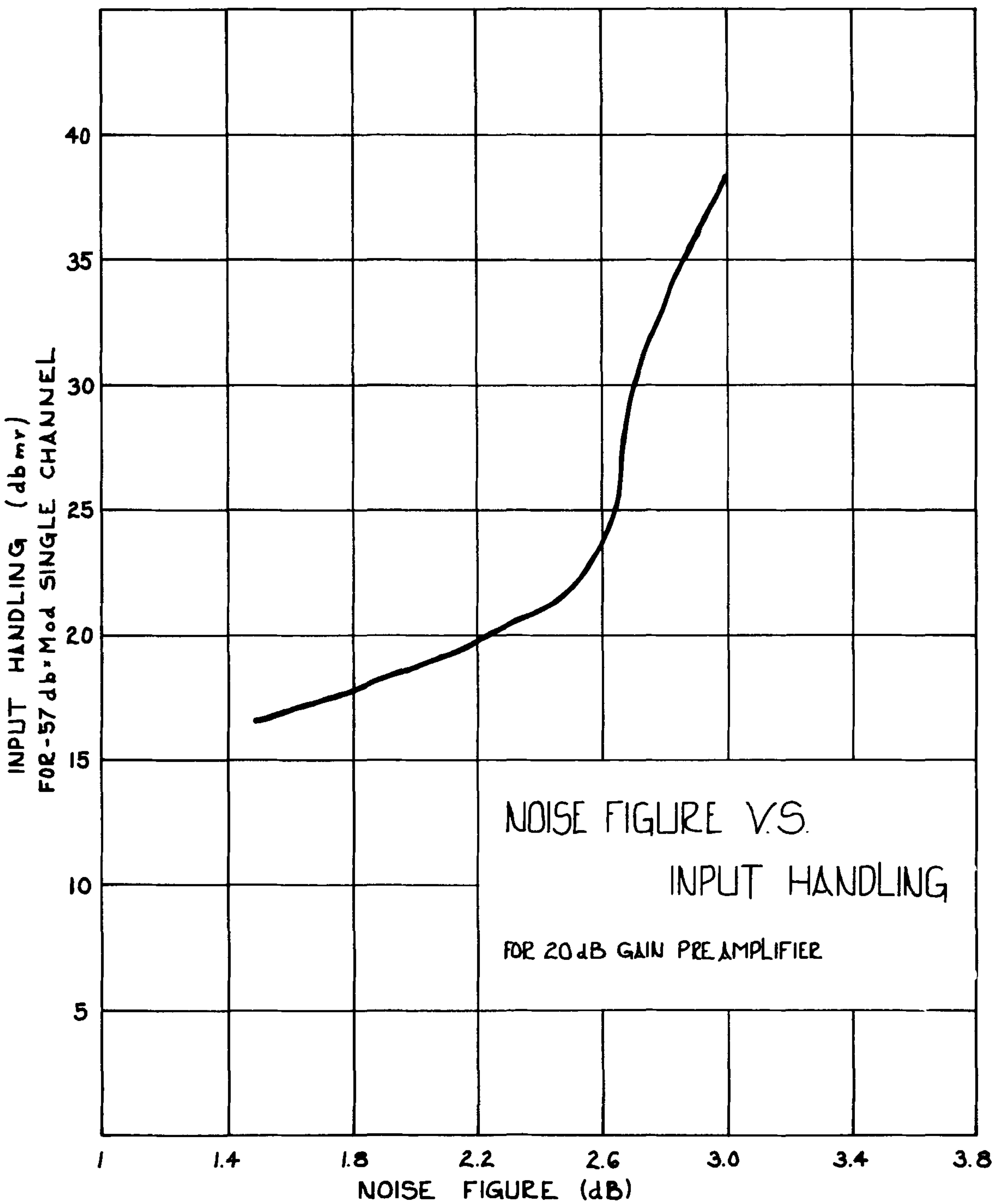

NOISE FIGURE V.S.
INPUT HANDLING
FOR 20dB GAIN PREAMPLIFIER
INPUT HANDLING (dbmv)
FOR -57 db X Mod SINGLE CHANNEL
40
35
30
25
20
15
10
5
1
1.4
1.8
2.2
2.6
3.0
3.4
3.8
NOISE FIGURE (dB)

FIG. 9

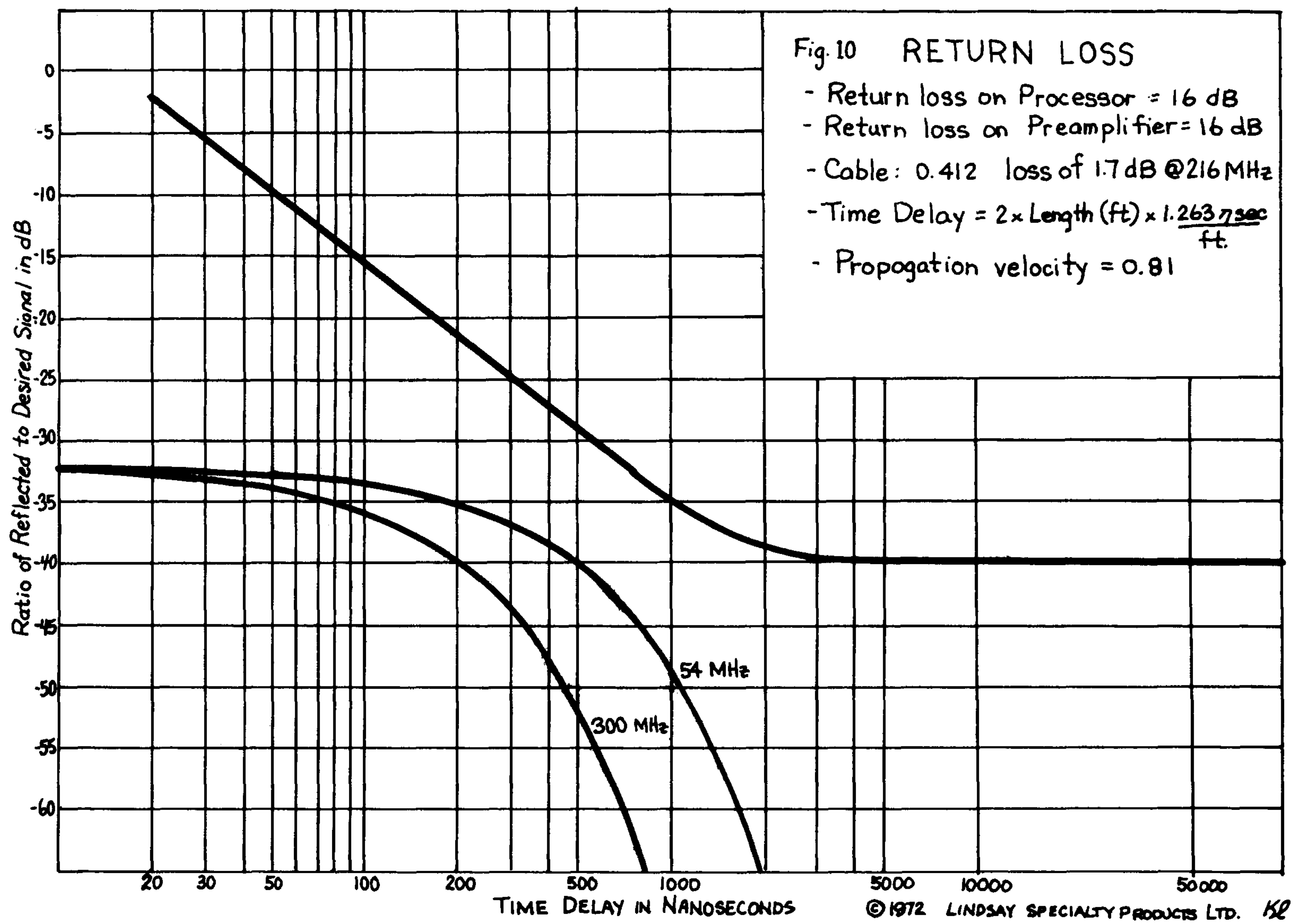


Fig. 10 RETURN LOSS
- Return loss on Processor = 16 dB
- Return loss on Preamplifier = 16 dB
- Cable: 0.412 loss of 1.7 dB @ 216 MHz
- Time Delay = 2 × Length (ft) × 1.263 ηsec/ft.
- Propogation velocity = 0.81
0
-5
-10
-15
-20
-25
-30
-35
-40
-45
-50
-55
-60
Ratio of Reflected to Desired Signal in dB
54 MHz
300 MHz
20
30
50
100
200
500
1000
5000
10000
50000
TIME DELAY IN NANOSECONDS
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PREVIEW

TECHNICAL EYE OPENER WORKSHOP

COPING WITH 1977 TECHNICAL STANDARDS

Sponsor

Society of Cable Television Engineers

National Organizer

Robert Bilodeau
Suburban Cablevision
West Orange, N. J.

Moderator/Organizer

Robert Cowart
Gill Cable
San Jose, Calif.

Panelists

Richard Covell
GTE Sylvania
El Paso, Texas

Kenneth Foster
N. Y. State Commission on
Cable Television
Albany, N.Y.

Early Monroe
FCC
Cable Television Bureau
Wash., D.C.

Keneth Simons
Jerrold Electronics
Horsham, Pa.

James Wright
CATV of Rockford
Rockford, Illinois

Technical standards mean different things to different people. The FCC's imposed standards are forever subject to modification determined, to a great extent, by the real life situations in CATV systems. Most recently, the timetable for implementing certain measurements was revised so that only minimal requirements need be met by a March 31, 1974, deadline, with provisions for more complex measurements in several tiers, and "full" compliance by 1977. This panel's charge is to relate the FCC's standards and state/local jurisdiction standards to what the public needs or wants. An attempt will also be made to translate technical standards of various jurisdictions into manpower, economic, material and hardware requirements to implement the standards as they are either required or desired.

Hopefully, some of the confusion surrounding multi-tiered standards and compliance will be cleared away.

PREVIEW

TECHNICAL EYE OPENER WORKSHOP

PRACTICAL APPLICATIONS OF FREQUENCY CHANNELING CONCEPTS

Sponsor

Society of Cable Television Engineers

National Organizer

Robert Bilodeau
Suburban Cablevision
West Orange, N. J.

Moderator/Organizer

Wayne McKinney
Texas Community Antennas, Inc.
Tyler, Texas

Panelists

Edward Chalmers
Zenith Radio Corp.
Chicago, Ill.

Malcolm Ferguson
Television Communications Corp.
Pennsauken, N.J.

Mike Jeffers
Jerrold Electronics
Horsham, Pa.

Henry B. Marron
Scientific-Atlanta
Atlanta, Ga.

Sruki Switzer
Switzer Engineering Service
Mississauga, Ontario

In the past several years, various proposals have evolved suggesting certain reassignment of television channel frequencies for carriage on a cable T.V. system. Some of these proposals have reached the available hardware stage in head-end equipment. The purpose of the exploration into the reassignment of frequencies is to determine if an arrangement other than the standard FCC broadcast assignments, in conjunction with standard midband and superband assignments, will provide an improvement in the performance of CATV amplifiers, specifically as the output capability of CATV amplifiers manifests itself in terms of second- and third-order beat products.

Three variations to date have been: the constant interval, the harmonically related constant interval, and the pivotal harmonically related constant interval. Each plan may have its own particular advantages and disadvantages and specific location considerations vis-a-vis off-air ambient signal level. In addition, a decision to employ a "non-standard" channeling plan is locked into the terminal equipment decision. The purpose of this panel is to state the pros and cons of the various techniques and relate them to practical applications. The panel represents a cross section of interested parties in this particular area of technology; i.e., receiver manufacturers, head-end equipment manufacturers, terminal equipment manufacturers and the cable operator.

PREVIEW

TECHNICAL EYE OPENER WORKSHOP

SYSTEM REQUIREMENTS FOR TWO-WAY

Sponsor

Society of Cable Television Engineers

National Organizer

Robert Bilodeau
Suburban Cablevision
West Orange, N. J.

Moderator/Organizer

Warren L. Braun, P.E.
Com-Sonics, Inc.
Harrisonburg, Va.

Panelists

James W. Stilwell
Tele-Systems Corp.
Elkins Park, Pa.

Edward Callaghan
American TV & Communications
Denver, Colorado

Nick Worth
TeleCable Corp.
Norfolk, Va.

Tim Ellers
The MITRE Corp.
McLean, Virginia

Steve McVoy
Coaxial Scientific Corp.
Sarasota, Fla.

Richard T. Callais
Theta-Cable
Inglewood, California

Each of the panelists will provide a 10-minute thumb nail sketch of their practical experience with two-way CATV systems, with application to the day-to-day system operation, together with the system design constraints that result from their findings.

Mr. Stilwell's comments will be focused on system ingress problems, with comments on remedial technology, i.e., instrumentation to locate sources of system ingress and upstream spectrum analysis. Comments will also be directed to hardware and maintenance limitations.

Mr. Callaghan's talk and comments will be oriented to the following areas of concern:

I. CATV system should be designed initially as a two-way system.

II. Unique requirements for two-way system components.

III. Reverse path design considerations.

IV. Other considerations

TeleCable operates sub-low-return CATV systems in Spartenburg, S. C. and Overland Park, Kansas. In each of these systems, extraneous signal ingress has proven to be a problem requiring special measures to reduce the ingress to manageable levels. Mr. Worth will cover system requirements and TeleCable's experience in solving these problems.

Mr. Ellers will comment on MITRE's experience with interactive signal communication, resulting in revised engineering requirements for two-way circuit signal implementation. His discussion will be focussed on system requirements viewed through the end use of the system in interactive mode.

Mr. McVoy will review their system experience with two-way responsive signaling. He will be unveiling a recently developed unique solution that solves many of the more difficult upstream system problems. Detailed information of this system will be disclosed in Mr. McVoy's presentation.

Mr. Callais will discuss the field experience and data accumulated from the SRS (Subscriber Response System) operation at Theta-Cable. Data will be presented and conclusions drawn. A technical paper will be available at this session.

TELEPRODUCTION AND AUTOMATIC PROGRAMMING SYSTEM
FOR LOW-COST CASSETTE AND OPEN-REEL VTRS

C. Robert Paulson
Television Microtime, Inc.
Bloomfield, Connecticut

The modern-day Japanese made television receiver with a Sony Trinitron picture tube is the world's finest time base corrector. Connected directly to the output of a low-cost cassette or open-reel VTR, the receiver's fast-acting lock up

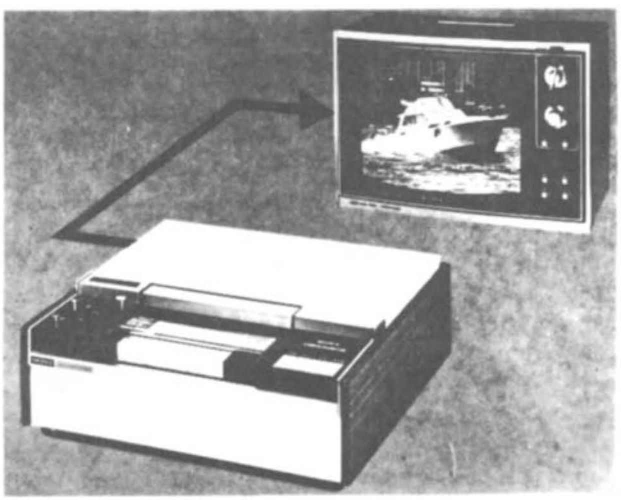

Fig. 1. World's best TBC Combination

circuits track the sync and subcarrier frequencies from the VTR, and mask head switch point skew tension errors of many microseconds. Just about the only problem the receiver can't handle is the inability of the VTR to track an improperly recorded interchanged tape.

Substitute an older, marginally maintained receiver, or even a new unit with long time constant AFC circuits designed for fringe area reception, and the picture changes. There's flagwaving at the top of the picture, perhaps vertical jumpiness, and the color may not lock up.

Precisely the same picture chaos results in a teleproduction application, when an attempt is made to mix the VTR output with a camera locked to it through even the most expensive of proc amps. Assemble or insert editing of tape segments into a second generation edited master has also been impossible, and even second generation dubs of continuous tape recordings lose all of the first generation's color freshness and pick up almost intolerable noise and jitter.

It has become fashionable in CATV circles to believe that adding a "time base corrector" in the VTR output channel--especially a revolutionary new "all-digital"

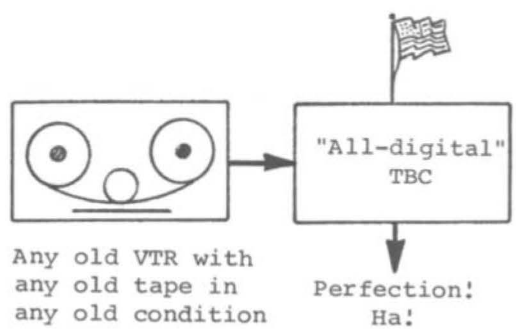

Fig. 2. No TBC offers a cure-all for bad recordings, uncorrected dubs, careless tape storage, and cheap, poorly maintained VTRs.

unit--will immediately eliminate all the problems. It is then quickly discovered that no TBC is a cure-all for all the problems that are created in the taping and dubbing activities described above. No TBC can do anything about the signal's time base, amplitude and sync/subcarrier phase problems recorded in to a tape from earlier generations. Further, getting the VTR's free-wheeling output under control and integrated with other VTRs and camera picture sources in a teleproduction or automatic programming system is a complexly related three part problem. It requires a system solution, and cannot be accomplished by adding a "standalone/wide-window/universal color box for all reasons" to the system, or any other excitingly described and uniquely heralded TBC, for that matter.

In order for any two NTSC color composite video signals to be mixed together through a switcher or special effects mixer, their transmission to and time of arrival at the mixing point must be arranged so that:

1. Their vertical sync pulse timing is coincident well within ±1/2 H line (±30 microseconds);

2. Their H sync pulse timing must be coincident within a nominal 1 microsecond, with a relative jitter at least below 100 nanoseconds, and preferably under 50 nanoseconds, peak to peak;

3. The two burst signals must be identical in frequency, with a relative phase difference, referenced to an arbitrary external subcarrier source, of substantially less than ±10 nanoseconds (±13° of subcarrier phase).

Compare the differences between a color camera composite video signal generated from a broadcast stable color sync generator and the output of a typical low cost cassette or open-reel VTR, in Subcarrier Frequency (SC), Horizontal Frequency (H), and Vertical Frequency (V):

	Broadcast Stability Requirement	Typical Low-cost VTR Output*
(SC)	3,579,545 Hz ±10 Hz "Reference 3.58 MHz"	Free-running frequency somewhere near Reference 3.58
(H)	15,734.--lines/sec. (3.58 MHz ÷ 455/2)	Free-running slipping, jittering, discontinuous signal somewhere near 15,734
(V)	59.94 fields/sec. (15,734.-- ÷ 525/2)	Stability problems like H, but 2:1 coherent, as long as the signal originated in a picture source driven by 2:1 sync

*It is testimonial to the capabilities of the modern-day color receiver that it can make an acceptable color picture even if all three sync references are independently varying and far removed from broadcast-spec frequencies.

There are a variety of MICROTIME™ 74 teleproduction systems which can be assembled from products manufactured entirely by Television Microtime, Inc., to accomplish the desired end of assembling "a teleproduction and automatic programming system for low-cost cassette and open-reel VTRs".

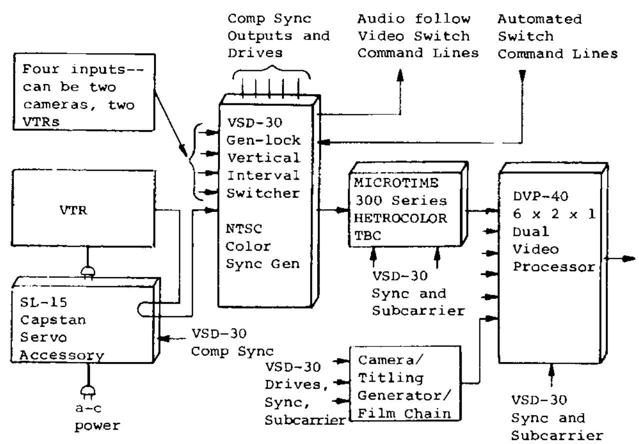

Fig. 3. HETROCOLOR '74 Teleproduction/ Automatic Programming System

TM Trademarks of TMI.

The success of the MICROTIME system design approach is based on compromise, since it is not reasonable to assume that an under $1,000, mass produced, battery-powered VTR can be induced, by either electromechanical or metaphysical techniques, to stabilize its jitters and broaden its responses to act like a $50,000 to $100,000 broadcast quad or helical scan VTR.

The three parts to the compromise solution to the three part problem are most economically achieved in a system with four coordinated-design MICROTIME products. These are the MICROTIME 300 Series HETROCOLOR™ Time Base Corrector, with a VSD-30 Video/Sync Director on its input and a DVP-40 Dual Video Processor on its output, and one or more SL-15 Capstan Servo Accessories providing vertically phase a-c driving power to the VTRs.

There are subtle differences in the sync and burst phases in the composite video output signal of this system, compared to the composite video signal at the master control output of a television broadcasting station. You won't be able to detect this difference without a $10,000 test instrumentation system, and those old unresponsive receivers in your subscribers' homes will never know what the difference is. For the record, however:

Subcarrier (burst frequency	3,579,545 ±5 Hz;
Vertical frequency and phase	59.94 fields/sec. average, ±8 H lines even with the most rubbery of cassette VTRs on line;
Horizontal frequency and phase	15,734 lines/sec. average, with H phase drift limited to ±1/2 H line at low frequency, and relative jitter between a camera and any VTR reduced to less than 35 nanoseconds peak to peak.

The functions of the MICROTIME products which accomplish this amazing transformation of the VTR output signal are:

VSD-30--generates broadcast stable NTSC color subcarrier reference, burst, and burst flag, and coherent RS-170 composite sync;

Provides five composite sync outputs to vertically phase up to five capstan servo'd VTRs to system vertical phase;

Provides additional comp sync, comp blanking, V drive, H drive, and burst flag outputs to drive stable picture sources--cameras/titling generators/switchers/special effect generators;

Makes vertical interval switches among its six inputs, transferring the RS-170 sync generator to gen-lock tracking of a non-synchronous VTR.

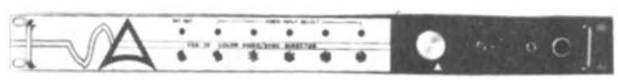

Fig. 4. The VSD-30 is an impressive multi-function unit including an NTSC Color Sync Generator, a second gen-lock RS-170 sync generator with burst flag included, a 6-input vertical interval switcher, a "coherent color" feature for dubs to quad VTRs, and 75-ohm outputs for all required drives and sync signals.

SL-15--generates a-c driving power for the VTR of the proper frequency and phase to maintain vertical phase coincidence between the VTR output and the reference sync generator, by comparing the VTR output phase continuously against the comp sync reference from that sync generator.

Fig. 5. The SL-15 corrects the tape speed and phases the playback timing of a power-line driven or capstan servo'd VTR or VDR against any external vertical sync reference.

MICROTIME 300 Series TBC--corrects the VTR output signal jitter, skew tension errors, and burst frequency/phase errors against the VSD-30 comp sync and subcarrier references, to a residual raster jitter of ±17 nanoseconds maximum against the comp sync reference, and color phase shift to less than ±4 nanoseconds relative to VSD-30 subcarrier reference.

Fig. 6. The MICROTIME 388 HETROCOLOR TBC is the most versatile in the industry, accepting any kind of color signal, from any kind of properly maintained VTR or VDR, with either RS-170 (broadcast) or RS-330 (2:1 interlace) composite sync.

DVP-40--switches the outputs of up to six TBCs or cameras or other picture sources which are driven by the VSD-30 sync generator, to either side of a 2x1 switcher/fader, where they are combined into one output.

The system output is therefore the manually selected combination of inputs from any of the studio's picture-generating sources--VTRs, cameras, titling generators, special effects generators, even the newly emerging video disc recorders (VDRs). An additional feature of this HETROCOLOR system is that <u>all the switching functions may be commanded remotely, either manually or under automatic programming system control</u>! Program audio from the VTRs automatically follows video, and any synchronous inputs (camera/titling generator/film chain) can have announce mike or recorded audio sources follow their selection.

This MICROTIME teleproduction and automatic programming system meets all the FCC specifications for broadcasting of a composite video signal, as regards the frequency and stability of the burst, vertical sync and horizontal sync. (There is still a raging debate about what is really required for sync and subcarrier phase coherency.)

A broadcast engineer would describe the output of this program origination system as "non-phased color". It is a well known fact of life in the industry that many stations broadcast non-phased color tapes from VTRs and VDRs or ancient vintage and variable capabilities as a matter of convenience, and this departure from broadcasting a signal driven by the station NTSC color sync generator is not detectable on any home receiver.

In view of this confusion within the industry, we do not take any position as to whether this MICROTIME teleproduction and program origination system "meets FCC specs for broadcasting". Its output signal is technically capable of being broadcast, processing the output signals of low cost cassette and open-reel VTRs. Its output signal is certainly "broadcast quality". The system described contains four processing amplifiers, a complete NTSC color unit in the TBC and three limited function units in the DVP-40, in addition to the broadcast-spec NTSC Color Gen-lock Sync Generator in the VSD-30.

An additional signal processing capability not important in non-broadcast program origination, but frequently important in teleproduction processing, is an intriguing "coherent color" mode of operation. In this mode, the reference subcarrier frequency to the system is derived by multiplying the gen-lock composite sync H frequency within the VSD-30 by 455/2.

The H frequency will be tracking the output of the VTR or VDR, which may be capstan servo'd, a-c power line driven, or even a battery-powered. The time base corrected output of the TBC will have coherent burst and sync, but their frequencies will be floating and off spec, depending completely on the characteristics of the source VTR/VDR.

The second generation playback of this signal will be coherent at the DEMOD output of the VTR, and may be time base corrected against broadcast-spec sync and subcarrier, if the unit is appropriately equipped. There is a limitation on this transfer process, however, associated with the amount the average frequency of H (and therefore of subcarrier) is displaced from the centerline specification of 15,734 lines/sec. A 6 Hz displacement of H results in a 1300 Hz displacement of subcarrier frequency, which equates to a 30 degree hue shift of any color across each H line, with respect to stable re-inserted burst. The broadcast VTR must have a velocity error corrector in its TBC complement to remove this unacceptable color aberration from the second generation playback. At some lesser displacement of H, the hue shift error becomes tolerable, but this is a subjective judgement.

An outstanding feature of the SL-15 Capstan Servo Accessory is its ability to alter the tape speed of either a line-locked or a capstan-servo'd VTR, so that reproduced H frequency is 15,734 lines/sec average. This means that even a battery-pack tape, recorded extremely slow because of low tape voltage, may be corrected to 15,734 lines/sec average frequency by playing it back on a capstan-servo'd VTR driven by an SL-15 unit.

An additional MICROTIME 74 TBC series is now available for broadcasters and others who require the ability to produce a broadcast-stable, phased (coherent) color tape in one step from a signal reproduced on a capstan servo'd (V locked) VTR or VDR. Called the 600 Series DIGI-MATIC™ TBCs, these units utilize an all-digital signal processing technique.

Fig. 7. The 600 Series DIGI-MATIC TBC is the newest addition to Television Microtime's rapidly expanding product family.

They work with direct and heterodyne record 1-head-per-field helical VTRs which exhibit large time base errors from inherent instabilities in the tape drive and video head drive servo systems, large irreducible changes in tape dimensions manifesting themselves as skew tension errors, and/or 1/2 H-line or greater discontinuities in sync timing caused by wrong field edits in capstan servo'd VTRs.

The DIGI-MATIC TBC design features advanced proprietary automatic circuitry designed to minimize the effects of digitizing noise and quantizing errors and heterodyne processor noise found in earlier all-digital TBCs. A unique capability included in the digitizing clock function compensates for velocity errors detected in the playback of direct-record signals. With the unit interconnected to the VSD-30 input accessory, one switch selects whichever processing mode is required by the phased or non-phased characteristics of the incoming composite video signal. All units contain a MICROTIME full-color proc amp with front panel operating controls and internal maintenance adjustments described previously.

The current MICROTIME family of time base and velocity error correctors and teleproduction accessories now includes six base product series and almost a dozen accessories. A hallmark of every MICROTIME designed product is universal adaptability and a competitive price consistent with quality and reliability. Individual designs incorporate all the currently feasible time base error correction technques, alone and in combination--EVDL (Electrically Variable Delay Line), SBDL (Switched Binary Delay Line), heterodyne color sign recovery, and all-digital, to reach these objectives.

THE MITRE INTERACTIVE TELEVISION EXPERIMENT

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The MITRE Corporation has announced plans for a demonstration of interactive television in Stockton, California, in cooperation with Educational Testing Service and Big Valley Cablevision of Stockton. The intent of this paper is to report the status and progress of the planned project to the cable television community.

The project will experiment only with public service aspects of interactive television, and will include no commercial services. The major emphasis is on education service provided by a computerized education and information system. An essential feature of the system is the use of a time-shared computer system which can provide completely individual service to about 100 users at any time. The unmodified home TV receiver is used as the computer display. The computer generates full color, still frames of alphanumeric and graphic information, under control of the home user. The user enters information requests on a small keyboard which channels to the computer via the reverse path of the bidirectional cable system. Any user is free to choose from the complete data base, which includes over 100 hours of material, at any time, independently from any other user. If more than 100 users request service at the same time, the overload users get a busy signal and are entered into a que.

The particular implementation chosen for use in Stockton is based on providing the maximum experimental flexibility, with least investment in hardware, in order to concentrate resources on content development and social evaluation. The Big Valley system is divided into six separate service hubs, each trunked to a central headend. Approximately 16 channels will be made available (of the 60 channel dual cable system) for the interactive experiment. The 16 channels will be used independently in each service area, allowing 96 users. This approach allows a dedicated channel for each active user, eliminating the need for a home located refresh device, and allowing the delivery of moving video and audio.

About 1000 homes will participate in the project, sharing the available computer ports. Each participant will be provided a keyboard and non-standard channel converter.

A computer system will be located at the headend. The computer offers access to 200,000 stored frames, 12,000 graphics images, 5 hours of random access audio, plus video tape and picture files. All services are in color.

The particular system chosen is more ambitious than most systems presently under commercial development. Our research indicates that it can be practical over a wide variety of usage and service assumptions. To further develop the system, however, we must have a much clearer idea of how people will use it, and what types of service are attractive and effective. There is no way to accurately determine these usage characteristics except to try out a variety of approaches and evaluate the results. The content approaches will be developed by a number of subcontractors and partners with demonstrated expertise in authoring and education. Educational Testing Service will provide assistance in evaluation and social experiment design. Thus, the primary mission in Stockton is to gather data about how people will actually use interactive television. In addition, Stockton will provide a major demonstration of the future services which cable may provide.

The proposed schedule calls for material authoring on the computer system installed at MITRE during 1975. The system will be moved to Stockton by 1976, and will deliver services for two years.

It must be emphasized that the project is presently in the proposal stage, and has not received full funding. The project is the result of our previous research in the interactive television field, which has been funded by the National Science Foundation since 1971. A formal proposal to support the complete project in Stockton has been submitted to the National Science Foundation in April, and a decision on it is expected in 6-9 months. A public announcement before funding approval is required because of the need to seek the support and cooperation of the Stockton community. In addition, we have discussed the project with a number of cable systems during our search for the most suitable experiment site, so that awareness of our plans are widespread.

THE PHAZAR, A NEW INTERFERENCE REJECTION DEVICE

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Steven I. Biro — Biro Associates, Princeton, N.J.

The PHAZAR, a recently introduced interference reduction system eliminates the need to know the direction of the off-the-air interference source, and in most cases makes the application of an auxiliary antenna unnecessary. A sequential manipulation of the controls will result in a 20 to 30 dB interference rejection.

In this paper the authors give a brief summary of the conventional interference rejection methods used before. Than, using block diagrams and phasor charts, the authors describe the theory and operation of the PHAZAR SYSTEM "A".

Finally, the interference rejection capabilities of the PHAZAR will be demonstrated simulating harmonic type and co-channel interference conditions.

Off-the-air interference problems, such as co-channel interference, the harmonics of local FM stations and Citizens Band transmitters, adjacent channel carriers, high voltage power line noise, reflections (ghosting) have impaired the CATV picture quality for a long time. In order to ascertain the best picture quality CATV engineers used one or more of the following approaches to protect subscribers against interference problems:

1. Better antenna site selection. To prove by an on-site survey that the location is free of interference.

2. Phased antenna-arrays, forcing an antenna radiation pattern null in the direction of interference.

3. The application of bandpass filters and traps.

While the use of filters and traps might bring satisfactory relief against strong adjacent channels, where the interfering carriers fall outside the received channel, filters and traps are useless against co-channel or harmonic type of interference if the spurious signal shows up between the video and sound carrier. (FIGURE 1).

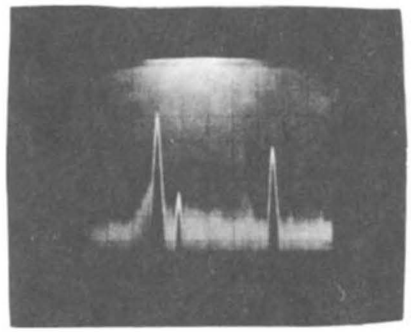


FIG. 1

This spectrum analyzer picture shows a spurious signal .75 MHz (3/4 Division) above the video carrier. No trap or filter could be applied to eliminate the interference.
Horizontal scale: 1 MHz/Division
Vertical scale: 10 dB/Division

The PHAZAR, a recently introduced interference rejection device, greatly simplifies the task of interference reduction or elimination. The advanced (SYSTEM "A") version of the PHAZAR even eliminates the need to know the direction of the interference source, or the installation of a second (auxiliary) antenna-array, thus rendering an attractive and economical solution combating off-the-air interference.

The concept of interference cancellation, employing a bucking antenna, oriented toward the interference source, has been known for a number of years in the CATV industry. External attenuator pads were used to equalize the amplitude of the interfering signals, and different length of coaxial cables were introduced to achieve out-of-phase (cancellation) conditions. However, the process was somewhat crude and tedious, the monitoring instrumentation inadequate. After many disappointing field applications the bucking antenna method slowly faded away.

The PHAZAR, a passive interference rejection device, operates basically on the bucking antenna principle. The first input port of the PHAZAR is connected to the main antenna array (FIGURE 2), receiving the desired station plus the undesired interferences. The second input port is connected

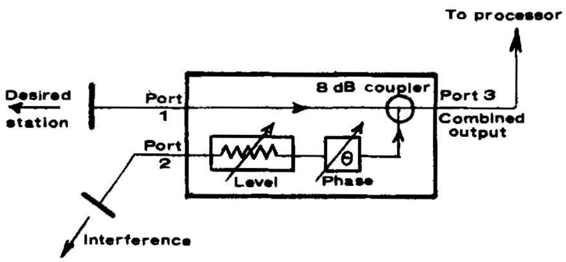


FIG. 2

Schematics of the PHAZAR concept

to the auxiliary antenna, oriented toward the interference source, picking up primarily the interference, and a fraction of the desired signal. A portion of the interference signal is than added out-of-phase to the desired signal, resulting in a more or less complete cancellation, as illustrated on the vector diagram of FIGURE 3.

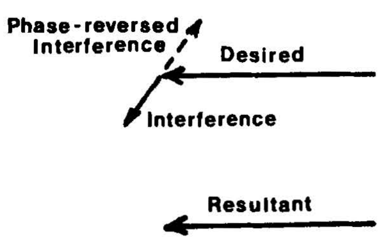


FIG. 3

Vector diagram of interference cancellation

The cancellation (phasing and shaping) process of the PHAZAR is convenient and effective because the instrument is head-end rack mounted, and has built-in attenuators and phasing controls. (FIGURE 4).

The instrument controls provide:

1. A maximum of 40 dB attenuation of the undesired signal by step-attenuators.
2. 0 to 360 degree phase variation through 10 and 90 degree step controls.
3. A continuously variable 0 to 2 dB attenuation, and a 0 to 10 degree fine tuning phaseshift vernier.

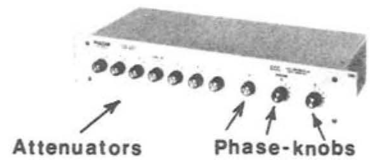


FIG. 4

Rack-mounted version of the PHAZAR

It is the fine and continuous variation of amplitude and phase which makes the PHAZAR really successful, the nulling so effective, producing a minimum of 20 dB, occasionally as high as 40 dB interference rejection.

The PHAZAR SYSTEM "A" is a more advanced version of the PHAZAR concept, using, beside the PHAZAR, a sum-difference hybrid combiner, which replaces the original 3 dB hybrid splitter (combiner) of the antenna array (FIGURE 5). If the system already has a two-ay or quad antenna-array working on the problem channel, there is no need for the installation of an auxiliary antenna.

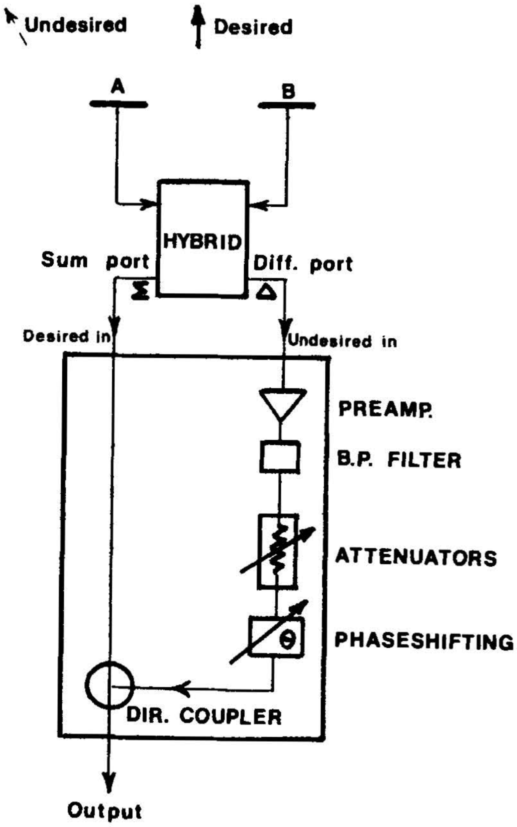


FIG. 5

Schematics of the PHAZAR SYSTEM "A"

When both antennas of a horizontally stacked antenna-array are oriented toward the desired station, the output from the hybrid sum port will be the desired signal, plus an interference signal component. The output from the difference port will not contain any desired signals (see the split lobes of FIGURE 6), only the interference signal will appear at this port with a considerable amplitude.

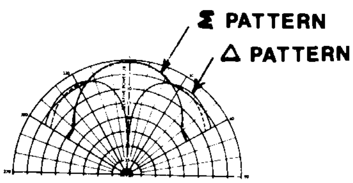


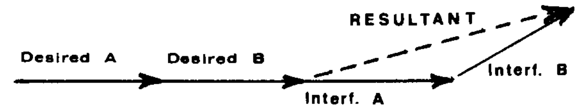
FIG. 6

The split-lobe radiation pattern shows that the desired station is received at a -25 dB level. The deep null also falls in the direction of the desired station.

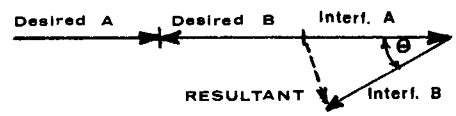
Connecting the sum port to the "DESIRED IN" port of the PHAZAR, and the difference port to the "UNDESIRED IN" port of the PHAZAR, we are ready to adjust the level and phase controls on the PHAZAR to achieve interference cancellation. There is a built-in preamplifier after the undesired-in port ascertaining an initial high interference signal whioh is than attenuated to the exact cancellation level.

The theory of operation is illustrated with the aid of the vector (phasor) diagarams of FIGURE 7. The signals received from antennas A and B are vectorially added at the sum port, and vectorially subtracted at the difference port. Assuming that antennas A and B have received the desired signals in-phase, and they also have equal gain, the two desired signals simply add at the sum port (FIGURE 7/a). The interference signal received by antenna A is arbitrarily shown as in-phase with the desired signal. The interference signal picked up by antenna B has a phase angle θ relative to the signals of antenna A. The dashed line represents the resultant interference vector.

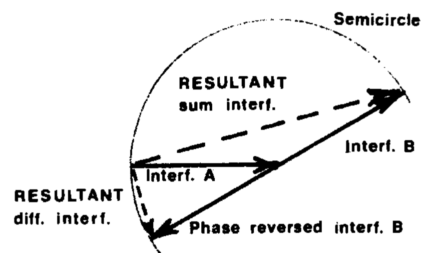
FIGURE 7/b depicts signal conditions at the difference port. The desired signals are equal but 180° out-of-phase, thus resulting in a perfect cancellation. (No desired signal at the difference port). The resultant interference signal is made up by the interference signal of Antenna A and the interference signal of Antenna B. However, the latter vector shows a reverse direction to indicate subtraction. The dashed vector of FIGURE 7/b pointing downward is the resultant of the interference signals.



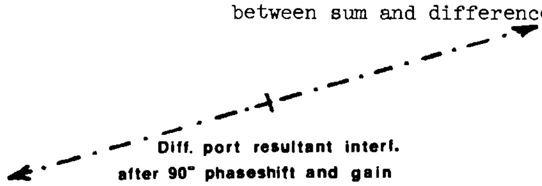
a. Hybrid sum-port output



b. Hybrid difference-port output



c. Interference signal relationships between sum and difference port



d. Resultant interference vectors after cancellation process.

FIG. 7

Phasor diagrams for the PHAZAR SYSTEM "A"

FIGURE 7/c illustrates the relationship between the sum and difference port interference vectors. By noting that interference vectors A and B are equal in length, it can be seen that these vectors describe a semicircle in which vectors A and B represent the radii of the semicircle. Second, the resultant interference vectors (dashed lines) are always at right angle relative to each other (Thales principle).

The remaining task is to rotate clockwise the difference port interference vector by 90° and make it equal amplitude with the resultant sum port interference vector. (FIGURE 7/d).

This critical amplitude shaping and phasing process is achieved by the built-in preamplifier, the step and fine tuning attenuators and phase shifters.

Since there is a fixed 90 degree phase relation between the two (sum and difference ports) resultant interference signals, the PHAZAR's phase controls can be set to a predetermined position corresponding to the 90 degree difference. Then, only the level (attenuation) controls need to be varied sequentially in order to find the setting of the best cancellation. This, in turn, substantiates our statement in the introduction that the direction of souce must not be known to achieve cancellation.

The PHAZAR and PHAZAR SYSTEM "A" are trade names of the CATV Component Company, St. James, N.Y. Patent application is currently being processed. It is the authors firm opinion that the above discussed interference rejection technique, when properly applied, will provide CATV operators with a convenient and economical tool to combat co-channel interference, ghosting, local RF and AC interference conditions, resulting in a much improved picture quality, less service calls, and hopefully, an increased demand for CATV services.

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THE RELATIONSHIP BETWEEN THE NCTA, EIA, AND CCIR DEFINITIONS OF SIGNAL-TO-NOISE RATIO

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One might think that such a common term as signal-to-noise ratio has only one definition, but this is unfortunately not the case. The definition depends upon what is meant by "signal" and what is meant by "noise" and these meanings determine how the measurement is made. To compare an NCTA S/N measurement made at VHF with an EIA or CCIR S/N measurement made at video baseband one must follow the vestigial sideband television signal through an ideal vestigial sideband demodulator. The resulting relationships show that there is only a small difference between the various definitions. Experimental results which back uo the theoretical findings are described.

A. Definitions

The following are some of the definitions commonly used to define TV signal quality.

1. NCTA : Signal - rms power of the VHF signal during the synch pulse.

   Noise - rms noise power in a 4 MHz wide VHF channel.

The measurement is necessarily made at VHF and generally with a field strength meter. Signal power is read directly from the field strength meter. Noise power is also read off the meter after the signal is removed but a correction factor of 3.9 dB, (at +5 needle reading) which accounts for the fact that the meter bandwidth is less than 4 MHz and also that the meter attempts to read noise peaks rather than rms value, is added to the reading.

Alternatively, a VHF spectrum analyzer may be used to make the measurement. The sweep speed must be sufficiently lowered, and the sweep width decreased so as to insure that one does see the true synch pulse peak. For the noise measurement, video filtering can help establish the rms value but the true noise bandwidth must also be carefully established.

2. TASO : Signal - rms power of VHF signal during the synch pulse.

   Noise - rms noise power in a 6 MHz wide VHF channel.

Note that this is exactly the NCTA definition except for the different noise bandwidth. Thus the TASO and NCTA definitions are related to each other by a simple bandwidth correction factor. On the other hand, they are distinctly different from the definitions which follow.

3. EIA : Signal - difference in voltage between the synch tip and the reference white.

   Noise - rms noise voltage (nominally between 10 KHz and 4 MHz) weighted by the curve shown in Figure 1.

The measurement is necessarily made at baseband frequencies. A wide band oscilloscope is used to measure the peak-to-peak volts at the output of the weighting network shown in Figure 2. An rms indication meter is then used to measure the noise voltage with the signal removed. Low frequency noise due to hum is excluded.

4. CCIR : Signal - difference in voltage between the blanking pulse and the reference white.

   Noise - rms noise voltage weighted by the curve shown in Figure 3.

The measurement is made as for EIA except that the signal is defined as above and the weighting network shown in Figure 4 is used.

To be more precise, this definition applies only to the CCIR Norm M television signals used in Canada and the United States. In other countries, different weighting networks are applied.

In addition, one often hears reference to an unweighted CCIR signal to noise ratio. One could view the "unweighted" case as one where the weighting network has a flat frequency response.

5. BTL : Signal - difference in voltage between the synch tip and the reference white.

Noise - rms noise voltage weighted by curve shown in Figure 3.

This is clearly a hybrid description in which signal is defined as in EIA but the CCIR noise weighting is used. Here also the "unweighted" definition exists. The unweighted and weighted ratios are simply related by the appropriate weighting factors which will be given later in this paper. The relation is such that the weighted signal to noise is always larger than the unweighted signal to noise ratio.

B. Relation Between RF and Baseband S/N

Within the definitions given in Section A of this paper, there are obviously two classes of S/N, i.e. measurements made at VHF and measurements made at video baseband. For the latter type measurements, a noise weighting is applied which attempts to take into account the variation in subjective evaluation to interference at various baseband frequencies. In that sense the latter definitions are more nearly a measure of the true quality of the TV picture delivered to the customer. Both EIA and CCIR noise weighting shows that in general, noise at high baseband frequencies is less objectionable than noise at low video baseband frequencies. The difference between the two is that the EIA applies to color TV while the CCIR is only applicable to black and white. The greatest confusion arises from imprecision in stating which baseband definition, whether weighted or unweighted, is to be compared to the NCTA definition.

In this section, we derive a general relation between the baseband and rf signal to noise ratio.[1] As a starting point, we begin with the familiar equation for a double sideband amplitude modulated wave

$$g(t) = A_c(1+mf(t)) \cos \omega_c t \qquad (1)$$

where m = modulation ratio
ω_c = carrier frequency
A_c = carrier amplitude
$f(t)$ = modulation function

The carrier envelope varies from A_c (1+m) to A_c (1-m) because $|f(t)| \le 1$. The detected peak to peak signal voltage is therefore proportional to $2mA_c$.

This signal is accompanied by noise assumed to have a uniform spectral power density, η, over the full 2B rf bandwidth of the receiver. B is the spectral width of the video modulating signal. If the noise on one side of the carrier is uncorrelated with noise on the other side of the carrier, the noise voltages from the two "side bands" add in an rms fashion upon detection. This is the case when the predominant noise is generated at rf as in most CATV systems. The ratio of the detected peak to peak signal power to the detected rms noise power is then

$$\left(\frac{S}{N}\right)_{AM} = \left(\frac{2mA_c}{2B\eta}\right)^2 \qquad (2)$$

The rms carrier power during the peak of the modulating cycle is given by

$$C_p = (1+m)^2 \frac{A_c^2}{2} \qquad (3)$$

Substituting this into (2) one obtains

$$\left(\frac{S}{N}\right)_{AM} = \left(\frac{2m}{1+m}\right)^2 \left(\frac{C_p}{\eta B}\right) \qquad (4)$$

The factor $2m/(1+m)$ represents the envelope variation relative to the peak envelope. Clearly if the definition of peak to peak signal changes, as for instance between CCIR and EIA, then we would accordingly adjust the modulation factor. The second factor in equation (4) could be the NCTA definition of carrier to noise.

In ideal vestigial sideband receivers, the signal and noise are first passed through a filter having the characteristics shown in Figure 5 and only then detected. The filter serves to just compensate for the extra low frequency ($f \le .75$ MHz) vestigial sideband which is transmitted. The detected signal output is then a nearly undistorted replica of the modulation waveform applied at the transmitter. This is achieved by adjusting the filter so that the voltage response at the carrier frequency is just $\frac{1}{2}$ of the response at frequencies above 0.75 MHz. Since the two sidebands of the signal are correlated, the voltage on either side of the carrier are added and the post detection signal voltage characteristic is independent of frequency up to the upper limit of the receiver response.

The receiver filter effectively eliminates half the sideband voltage. However, since the carrier voltage is also reduced by $\frac{1}{2}$, the envelope variation as a fraction of the peak carrier remains the same as it was immediately following the double sideband AM process in the transmitter. Thus the factor $2m/(1+m)^2$ retains its validity and meaning for vestigial sideband.

Consider now the effect on noise of the receiver filter. Because the two noise sidebands are uncorrelated, they add in rms fashion in the detection process. The resultant noise spectral density is shown in Figure 6. It is 3 dB down at zero frequency and increases quadratically to 0.75 MHz when it becomes flat. The equivalent noise power bandwidth is given by

$$B_N = \int_0^{f_m} n(f)df \qquad (5)$$

where n(f) represents the distribution of baseband noise with frequency and f_m is the maximum frequency of the receiver response.

By rewriting the second factor in equation (4) as $(2\ C_p/2B\eta)$, one can now modify this factor for the vestigial sideband case. In particular the equivalent noise bandwidth, 2B, for double sideband is replaced in vestigial sideband by B_N. Also, the carrier power is reduced by a factor of 4. Thus the second factor becomes $(\frac{1}{2}C_p/B_N\eta)$.

Consider now the possibility of noise weighting. Let the weighting filter be characterized by a frequency response w(f). Then the weighting factor is defined by

$$K_w = \frac{\int_0^{f_m} n(f)df}{\int_0^{f_m} n(f)w(f)df} \qquad (6)$$

The general equation relating baseband signal to noise with vestigial sideband rf peak rms carrier to noise is then

$$\left(\frac{S}{N}\right) = \frac{1}{2}\left(\frac{2m}{1+m}\right)^2 \left(\frac{C_p}{\eta B}\right)_{rf} \left(\frac{B_{rf}}{B_N}\right) \frac{B_N}{\int_0^{f_m} n(f)w(f)df} \qquad (7)$$

Alternatively,

$$\left(\frac{S}{N}\right) = \frac{1}{2}\left(\frac{2m}{1+m}\right)^2 \left(\frac{C_p}{\eta B}\right)_{rf} \left(\frac{B_{rf}}{B_{NW}}\right) \qquad (8)$$

Where B_{NW} is the equivalent weighted noise bandwidth given by

$$B_{NW} = \int_0^{f_m} n(f)w(f)df \qquad (9)$$

Equation (8) is the simpler form since it involves only one integral. However, it is most useful to express equation (7) in logarithmic form since weighting factors are generally given in dB and defined as in equation (6).

$$\left(\frac{S}{N}\right)_{baseband} = 10 \log \left\{\frac{1}{2}\left(\frac{2m}{1+m}\right)^2 \left(\frac{C_p}{\eta B}\right)_{rf} \left(\frac{B_{rf}}{B_N}\right) K_w\right\} dB \qquad (10)$$

C. Tabular Results

Equation (6) is a very general form for the weighting function. If n(f) has the form shown in Figure 6, the weighting applies to vestigial sideband. If n(f) is "flat", then the weighting applies to double sideband AM. If $n(f) \sim f^2$ it is because the rf noise spectrum is triangular, as in FM systems. Table I summarizes the noise weighting, in dB, for each of these cases for both CCIR and EIA weighting curves.

TABLE I. Noise Weighting (dB)

	"White" Noise (AM)	"Triangular" Noise (FM)	Vestigial Side Band Noise
EIA (color)	4.0	6.4	4.1
CCIR (monochrome)	6.1	10.2	6.7

In any one of the three types of systems, deviations from the "ideal" noise spectrum, in particular excess noise at low detected baseband frequencies, would reduce the actual improvement factor obtained from noise weighting and correspondingly result in degraded picture quality.

Table II shows the relationship between the NCTA signal to noise ratio and the various other signal to notse ratios defined in Section A.

TABLE II. S/N Relationships

$(S/N)_{TASO}$	=	$(S/N)_{NCTA}$	- 1.8 db
$(S/N)_{EIA}$	=	$(S/N)_{NCTA}$	+ 0.1 db
$(S/N)_{CCIR}$	=	$(S/N)_{NCTA}$	- 0.2 db
$(S/N)_{BTL}$	=	$(S/N)_{NCTA}$	+ 2.7 db

As an example, consider the relationship between $(S/N)_{EIA}$ and $(S/N)_{NCTA}$. The synch tip to reference white voltage is .875 of the peak signal. The NCTA bandwidth is 4 MHz and B_N obtained from integration of Figure 6 is 3.8 MHz. Rewriting equation (10), we have

$$\left(\frac{S}{N}\right)_{EIA} = 10 \log\left\{\frac{1}{2}(.875)^2 \left(\frac{S}{N}\right)_{NCTA} \left(\frac{4}{3.8}\right) K_w\right\} dB$$

$$= \left[-3.0 - 1.2 + \left(\frac{S}{N}\right)_{NCTA} + 0.2 + 4.1\right] dB$$

$$= \left(\frac{S}{N}\right)_{NCTA} + 0.1 \text{ dB}$$

Note that the relation between unweighted baseband signal to noise and the NCTA signal to noise can be readily determined from the two tables. Thus, for instance, unweighted BTL is given by

$$\left(\frac{S}{N}\right)_{BTL,\ unweighted} = \left[\left(\frac{S}{N}\right)_{NCTA} + 2.7 - 6.7\right] \text{dB}$$

$$= \left[\left(\frac{S}{N}\right)_{NCTA} - 4\right] \text{dB}$$

D. Experimental Verification

The experimental verification of the theoretical relationships obtained in section C are not so easy to come by as one might think. Aside from the normal instrument calibration problems, one is faced with the fact that vestigial sideband demodulators are only rarely a good approximation to the ideal assumed in the theory. Other factors enter in as well. For instance, the rf noise generator does have to have a fairly high output level without clipping the thermal noise peaks. This is best done by bandlimiting the noise before bringing it to full power in an output amplifier.

Nevertheless, verification of the theoretical expectations has been obtained. The most thorough experiment was recently carried out at a working session of the CTAC working group on noise.[2] This work, which was performed in February 1974, did verify, within ±1 dB experimental error, the predicted relationship between CCIR and NCTA signal to noise.

E. TASO Revisited

Although only peripherally related to the foregoing discussion, the following information may be of particular interest to CATV. The question concerns the subjective quality of television pictures as the S/N is varied over a wide range of values. The most extensive work along this line is, of course, the TASO study.

Some 14 years ago, the Television Allocations Study Organization undertook a comprehensive study of the subjective effect of random noise at various interference levels on the quality of the TV picture. The experimental program first established a set of optimum psychological definitions which were printed on the observer's scoring sheet, reproduced in Figure 7.[3] Several still scenes were viewed on good quality black and white and color 21-inch receivers. Viewing distance was between 90 and 126 inches and the average room illumination was 0.6 foot-candles. A 40 dB range of interfering noise ranging from not perceptible to completely masking was employed. Test results varied very little with the scene used. For the most extensive tests with the "Miss TASO picture" a total of 76 observers were asked to rate 20 showings of the subject with 10 different signal-to-interferance ratios each repeated twice in a random order. The results are tabulated in Figure 8.[4] Note that this data presentation is in a percentile form. For instance, at a $(S/N)_{TASO}$ of 27.5 dB, 50% of the viewers considered the picture "passable" or better, but also the most critical 10% of the viewers considered the picture "inferior". This lies 4 dB below the EIA recommendation of 33 dB[5] which is to be considered as an "outage" for microwave propagation fades.

In order to see if the TASO results could be used as a guide to the application of LDS microwave in CATV, a brief experiment was conducted at Theta-Com during one of the AML technical training seminars. The idea was to repeat the TASO type evaluation although necessarily under quite different conditions. An off-the-air television signal was processed through an AML microwave system and displayed on a 17" Sony television receiver. The $(S/N)_{NCTA}$ was varied from 50 dB down to 14 dB in 4 dB steps. This was controlled by a microwave attenuator placed between the AML transmitter and receiver. In all, 20 scenes were shown, 2 each at the same S/N, but in a completely random sequence.

The 24 students were mostly CATV technicians and engineers. They were each given a copy of the TASO Scoring Sheet and asked to evaluate the pictures on a personal rather than professional basis. The students were arranged in 4 rows of seats, the furthest being some 18 feet from the television screen. The room light was extinguished but enough illumination was available to permit the score sheets to be filled out. An A-B switch was used to switch the signal directly to the head end during the intervals when the microwave attenuation was reset.

Figure 9 summarizes the results of these tests. It is seen that the results are quite similar to the TASO results. As might be expected, the viewers in the front row were slightly more critical than those furthest from the screen. On the average, the CATV technicians and engineers were about 2 dB more critical than the TASO volunteers were back in 1960. Perhaps this is more a reflection of our rising expectations for good signal quality rather than any other factor which impacted these experiments.

REFERENCES

(1) Similar arguments are given in an unpublished memorandum by J. J. Bisaga. Other workers, notably J. J. Gibson, have followed slightly different lines but with essentially the same results.

(2) J. J. Gibson, private communication.

(3) G. L. Fredendall and W. L. Behrend, "Picture Quality Procedures for Evaluating Subjective Effects of Interference" PROC IRE 48, 1030-1034 (June 1960).

(4) C. E. Dean, "Measurements of the Subjective Effects of Interference in Television Reception" PROC IRE 48, 1035-1049 (June 1960)

(5) EIA Standard RS-250A (February 1967)

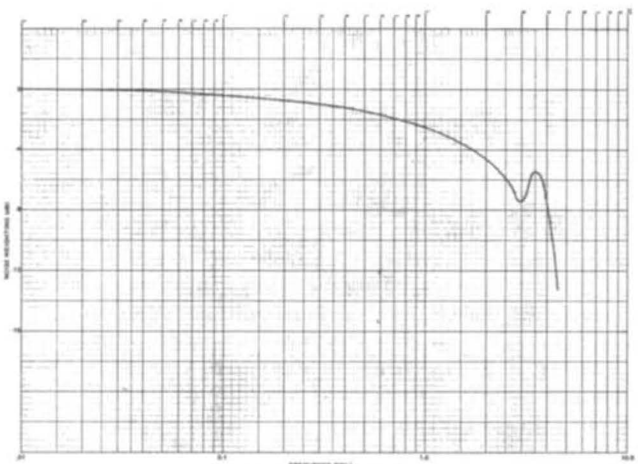

FIGURE 1. EIA NOISE WEIGHTING

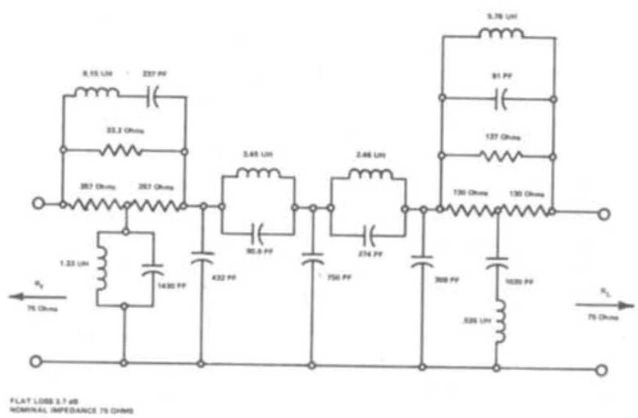

FIGURE 2. EIA NOISE WEIGHTING NETWORK

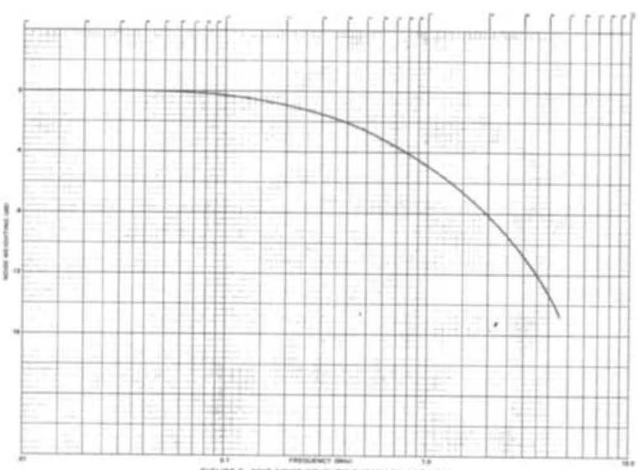

FIGURE 3. CCIR NOISE WEIGHTING (CANADA AND USA)

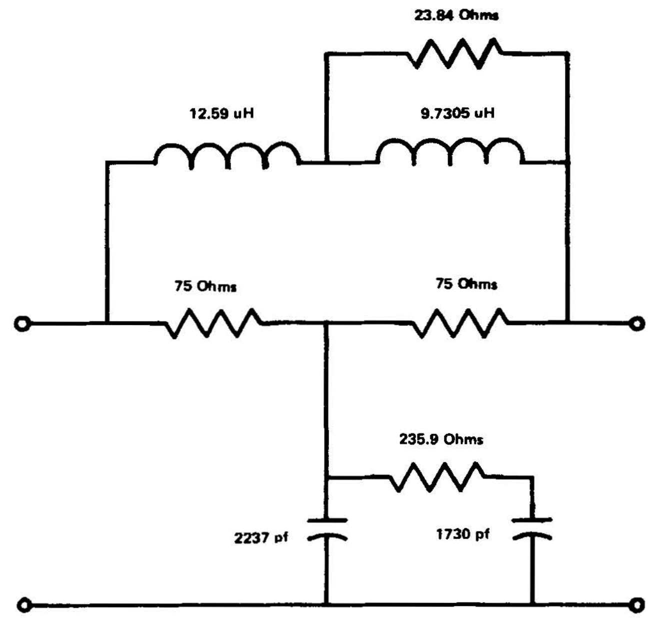

FIGURE 4. CCIR NOISE WEIGHTING NETWORK

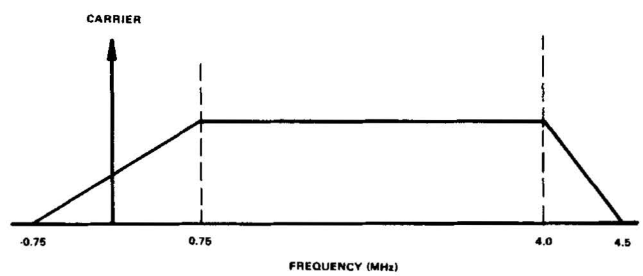

FIGURE 5. RECEIVER TRANSFER CHARACTERISTIC

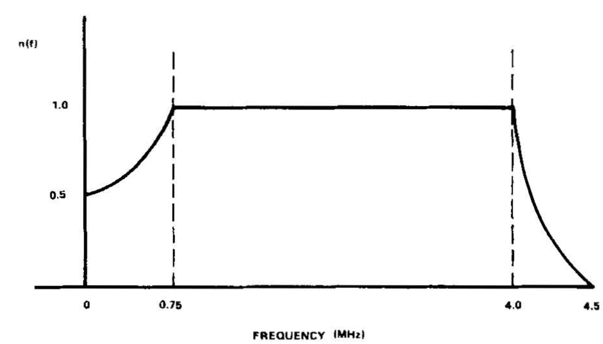

FIGURE 6. DETECTED NOISE SPECTRAL DENSITY

506

TELEVISION ALLOCATIONS STUDY ORGANIZATION

Panel 6

TEST NO ________ TV SET ________ OBSERVER ____________

EXCELLENT. The picture is of extremely high quality as good as you could desire.

0 1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16 17 18 19 20

FINE. The picture is of high quality providing enjoyable viewing. Interference is perceptible.

0 1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16 17 18 19 20

PASSABLE. The picture is of acceptable quality. Interference is not objectionable.

0 1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16 17 18 19 20

MARGINAL. The picture is poor in quality and you wish you could improve it. Interference is somewhat objectionable.

0 1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16 17 18 19 20

INFERIOR. The picture is very poor but you could watch it. Definitely objectionable interference is present.

0 1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16 17 18 19 20

UNUSABLE. The picture is so bad that you could not watch it.

0 1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16 17 18 19 20

DATE____________ TAB ______

FIGURE 7. TASO OBSERVER SCORING SHEET

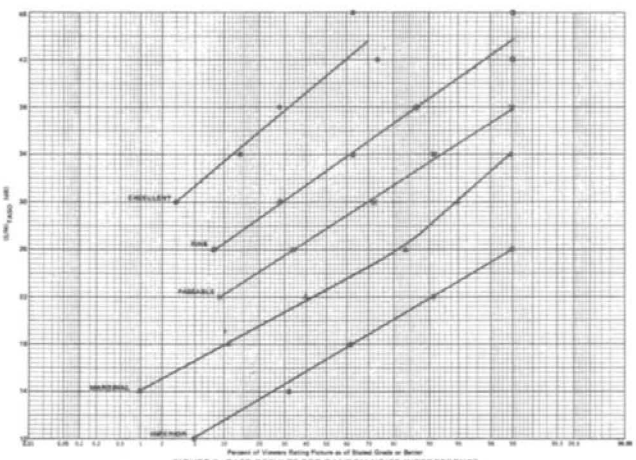

FIGURE 8. TASO RESULTS FOR RANDOM NOISE INTERFERENCE

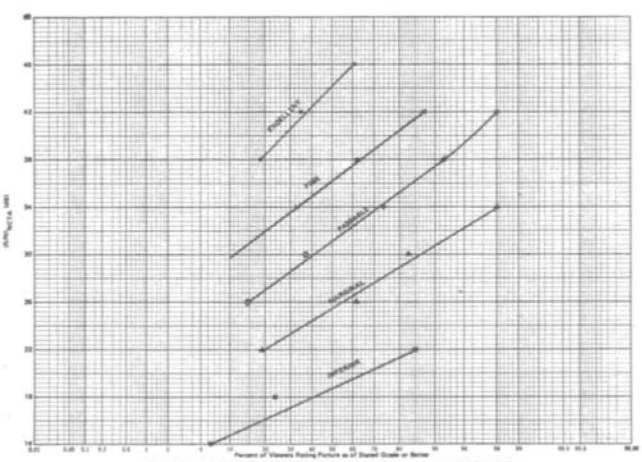

FIGURE 9. EVALUATION OF PICTURE QUALITY AT AML TRAINING SEMINAR

VALIDITY OF SUBJECTIVE TESTING

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Tektronix, Inc. Beaverton, Oregon

ABSTRACT

In view of the present FCC standards for CATV, there has been much discussion as to the validity of inexpensive test techniques; specifically subjective testing. A systematic study into test techniques and results was performed in an attempt to open up this needed field of testing. The goal was to provide the smaller or financially burdened operator with simple, inexpensive, repeatable, and valid alternatives to known, expensive instrumentation test techniques.

We chose to study three areas felt to be the most restrictive financially for the small operator; Intermodulation, Co-channel Interference, and Signal-to-Noise ratio. Using the technical standards suggested by the FCC in §76.605 (now partially suspended) we performed a comparison between an elaborate $11,000 test set, and a test set consisting of simple equipment and a TV set. The results we feel substantiate our theory, that subjective testing can be valid, and in some cases superior to other techniques.

INTRODUCTION

Subjective testing, it has been theorized, is the answer to inexpensive test techniques. After all, the basic test of a CATV system is it's ability to deliver a good picture. Cannot then, one use a TV as the test instrument?

Subjective testing in it's most pure form would involve looking at a TV picture and "quesstimating" the respective amounts of Noise, Crossmod, Intermod, Hum, etc. Many old timers refer to this form of testing as the "Calibrated Eyeball Technique".

Advocates of subjective testing are generally refering to a modified form of subjective testing. By controlling variables and providing the technician with reference distortions, subjective testing can be transformed from a "crystal ball" game into a highly valuable analysis tool. It is to this end I address the following discussion.

In spite of the discussions on subjective testing, little has been done to correlate procedures, accuracy, cost, and repeatability. My experiments are little more than a crude "stab in the dark" attempt to shed some light on subjective testing. Hopefully some direction and momentum can be developed to open up this needed field of testing.

TESTING USING DISTORTION PLATES

One modified form of subjective testing uses a series of calibrated plates (photographs) that represent various amounts of known distortions. The user simply selects the plate that most closely matches the picture on a test set and thereby ascertains the type and amount of picture impairment.

This form of testing has some definite drawbacks:

1. The plates are stationary representations of distortions that normally move or change on a TV screen.
2. The plates must be compared to picture material of various colors, hues, shades, etc.
3. TV sets vary, even among the same make and models. Sony Trinitrons, for instance, tend to indicate results 3 - 4 dB better than most other color portables.
4. Lighting of both the test plates and the TV screen is critical.
5. Multiple distortions or a combination of a distortion and low test drop strength can be difficult to interpret.

Even considering the drawbacks, I chose to go ahead and evaluate this technique. The results were suprisingly good. By controlling certain variables, the technique could yield quite useful results. In fact, in some instances of close in intermodulation distortions, the subjective technique was more sensitive in locating problems than other instrumentation techniques.

PREPARATIONS OF DISTORTION PLATES

The distortion plates are the key to the accuracy of this form of subjective testing. I chose to use a series of color plates as outlined in the following charts. I chose to limit the experiment to Intermodulation, Signal-to-Noise, and Co-Channel tests although subjective testing could be extended to include Hum, Ghosting, and Envelope Delay.

	INTERMODULATION			CO-CHANNEL		S/N
	1 KHz	100KHz	1 MHz	10KHz	20KHz	4 MHz BW
50						
45						
40						
35						
30						
25						

dB of Distortion

Figure 1 Chart showing Distortion Plate Values

To control some of the variables, the following were observed for the preparation and use of the plates:

1. Two plates were prepared for each distortion value, one using picture material, and one using color bars. A total of 72 plates were prepared.

2. The picture material included a substantial amount of white or light colors, distortions being most readily visible in the lighter areas. Waiting for light scenes on the receiver would also help to improve test accuracy.

3. The plates were taken from a standard color portable with an input of +5 dBmV, using a shadow mask tube.

4. The plates were of sufficient size to permit easy viewing (5 x 7 ").

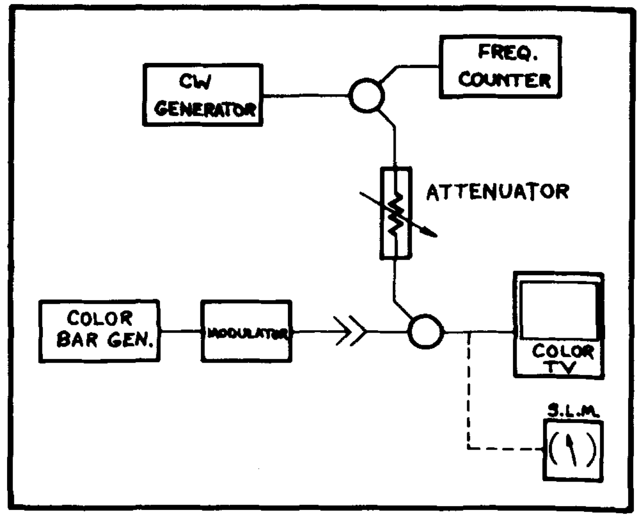


Figure 2 Setup for producing Intermod and Co-Channel Plates

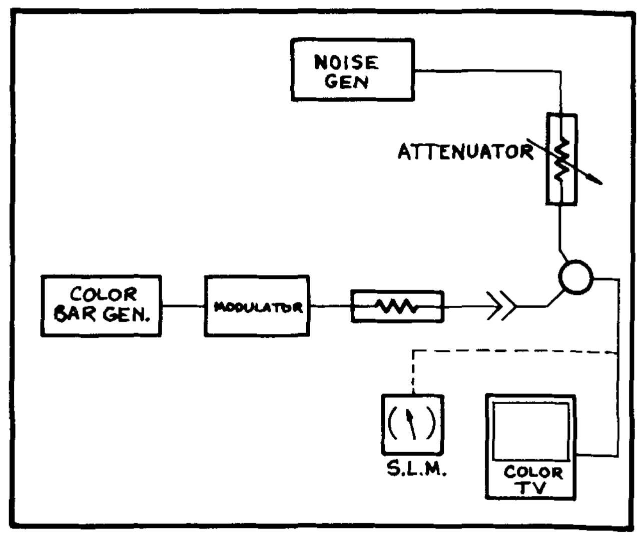


Figure 3 Setup for producing Noise Plates

ACCURACY TEST

A series of sample tests were performed using a group of random observers. The distortions were imposed over standard network program material and the observers were given as much time as necessary to analyze each distortion.

Figure 4 Using Distortion Plates

Six distortions were selected in each of six categories; and the subjects were told the type of distortion, such that the only subjective variable was the amount of distortion.

Verification of the amount of each distortion actually on the TV screen was made using a Tektronix 7L12 - 7L13 combination.

A total of eight observers used the distortion plates, and while the results are inconclusive at this time, some important observations could be drawn.

1. Most subjects could determine within ± 5 dB, the amount of an impairment.

2. Using color bar plates and test signals, the subjects could measure distortions in the higher ranges (30 to 45 dB) with high accuracy.

3. Accuracy on all distortions, especially Signal-to-Noise was spotty above 45 dB.

CONCLUSIONS

I feel that this form of testing, while having definite limitations, can be refined into a valuable evaluation tool for the CATV industry. A large scale subjective test should be setup to further refine both the distortion plate technique, and distribute the results to the industry.

TESTING USING A TV SET COMPARISON TECHNIQUE

This second technique for performing subjective testing relies upon a comparison between two live TV pictures, thereby eliminating many of the variables encountered using the distortion plate technique. The only equipment required is a TV set, an attenuator, a CW signal generator (units in the $100 price range work satisfactorily) and an A-B coax switch.

The theory of testing involves taking a clean TV picture and purposely distorting it by a known amount. Then the pictures are compared visually (subjectively) with the system pictures and an evaluation is made as to the amount of a distortion.

To control as many variables as possible, the following is suggested:

1. To eliminate differences in sets, it is recommended that an A-B switch be used with one set for the comparisons as shown in the figure.

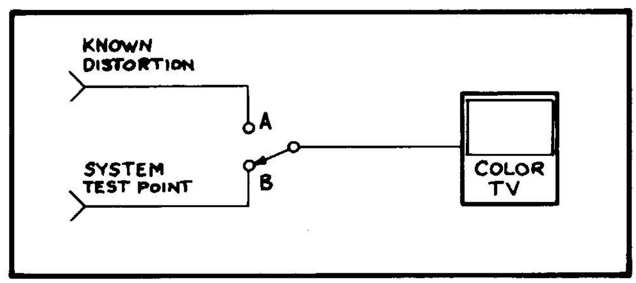


Figure 5 Comparison using A-B RF Coax Switch

2. A clean signal can be derived from a test antenna in most systems. In systems where there are no off-air signals, the output of an inexpensive rainbow-bar-dot generator can be used.

3. The "known distortion" line should have the same carrier strength as the "system test point".

4. The comparisons in most cases will be between two different pictures containing different programming. Different scenes camouflage or highlight some distortions.

The TV set comparison technique shows the highest promise of satisfying the accuracy requirements for the financially encumbered operators. Procedures were devised for Intermodulation, Co-Channel, and Signal-to-Noise (Carrier-to-Noise as defined in §76.605).

SUBJECTIVE-COMPARISON TECHNIQUE FOR INTERMODULATION

Two parameters can be obtained using this technique; the frequency of the interfering carrier, and the level of the interfering carrier relative to the picture carrier.

To obtain a rough idea of the interfering frequency, the number of bars cutting across one horizontal line can be counted as indicated in the two photographs. Then, for most TV sets, the frequency is determined:

Bars x line rate x frame rate x (1.4) correction factor = frequency

For example

$$23_{\text{Bars}} \times 525_{\text{Lines}} \times 30_{\text{Frames}} \times 1.4 = 507 \text{ KHz}$$

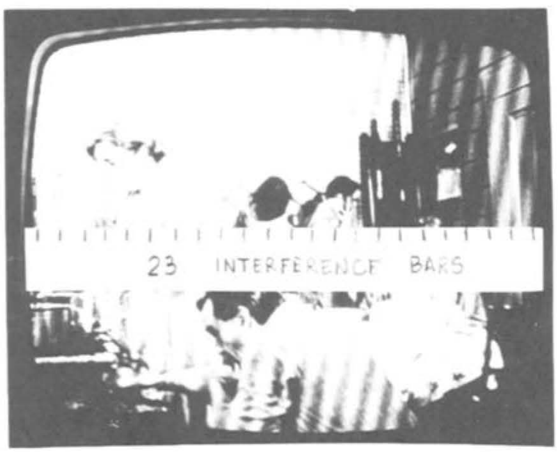


Figure 6 500 KHz Intermodulation

Figure 7 Count the Bars on one H Line in Center of Screen (1 MHz Intermod)

Generally, intermod frequency below 50 KHz will appear as multiple horizontal lines sometimes called the "venetian blind effect". Higher frequency intermods up to 3 MHz will appear diagonally, and can be counted using the formula.

On color sets, intermodulation products within 500 KHz of the color subcarrier will cause colored bar patterns. A strong intermod within 50 KHz of the color subcarrier will cause the color to drop out entirely causing a monochrome picture.

To evaluate the relative strength of a distortion, a known or calibrated distortion must be generated. Equipment is setup as illustrated in Figure 8.

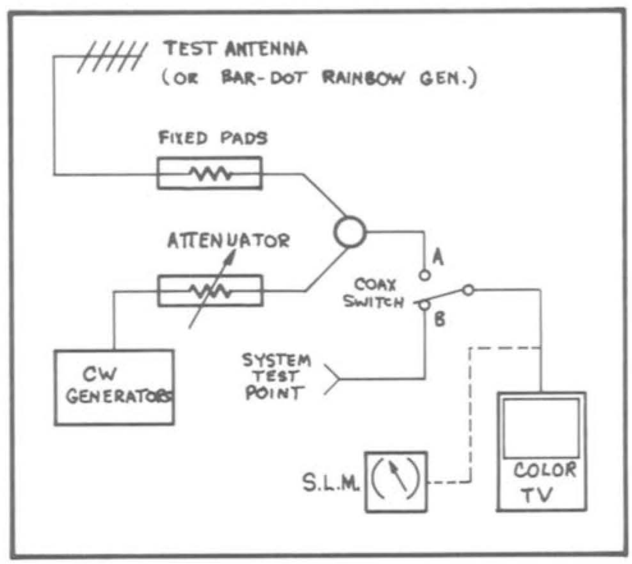


Figure 8 Setup for Subjective Intermodulation Test

The antenna is used as a source of a clean signal. If no off-air signals are available, then and inexpensive bar-dot-rainbow generator (such as a Heathkit IG-28) should be utilized. This clean signal should then be attenuated using fixed pads until it is equal in strength to the system test point. The CW generator should be adjusted to the frequency of the clean signal and with the ATTENUATOR in the thru position (no pads inserted) the generator output should be set equal to the clean signal and the system test point.

The attenuator can then be adjusted for calibrated amounts of distortion, the pads inserted equal to the dB down of the intermodulation carrier.

Starting in the B position, select channel 2 on the test set, and note any picture impairments associated with intermodulation (beat bars). Then select position A and the clean channel on the TV set, and using the attenuator and the CW generator frequency control, match the distortion pattern and intensity. This procedure should be repeated until a subjective match is produced.

The bar pattern frequency can be determined using the formula (as discussed), and the dB down of the interfering carrier is the dB of pads inserted with the variable attenuator.

SUBJECTIVE-COMPARISON TECHNIQUES FOR CO-CHANNEL

Although Co-Channel and Intermodulation are two very different distortions in origin, they appear similar in effect. For subjective testing Co-Channel can be considered to be little more than a special case of intermodulation. Specifically, Co-Channel appears as beats due to either a 10 KHz or 20 KHz intefering TV station.

The beat pattern produced by a Co-Channel situation is very distinctive. The picture will be cut by 30 to 40 almost horizontal beat bars. Once recognized, the technician will have no trouble differentiating between Co-Channel beats and Intermodulation beats.

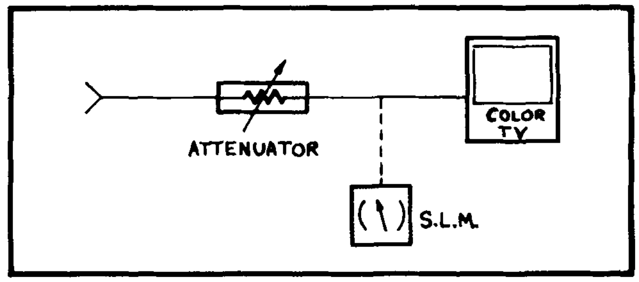
The equipment setup is the same as the Intermodulation setup (Figure 8). The CW generator should be carefully set 10 or 20 KHz from the picture carrier of the clean signal. This is easily accomplished by carefully zero beating the generator. This will be evident when the picture is cut with one vertical bar; half the screen dark, and the other half light.

Then the generator is carefully tuned until the pattern turns horizontal. This will be the 10 KHz reference. Continue tuning the generator until the pattern once again becomes vertical (two light and two dark bars), and then becomes horizontal. This will be the 20 KHz reference.

The same A-B comparison technique is followed as in the Intermodulation technique. Again, the pads inserted in the attenuator will be the dB down of the interfering Co-Channel after a satisfactory match is made between the distorted signal, and the incoming signal.

SUBJECTIVE-COMPARISON TECHNIQUE FOR SIGNAL-TO-NOISE

The Signal-to-Noise Technique actually yields a Carrier-to-Noise ratio. The distorted picture could be generated using a noise generator, however, this is not usually standard equipment for a small operator. We have instead derived a chart based upon the fact that, as signal strength is reduced, a standard TV set AGC will increase the noise on the screen in an almost perfect 1 dB correlation. This correlation is illustrated in the figure below.



C/N	55dB	50	45	40	35	30	25
SIGNAL LEVEL	6dBmV	1	-4	-9	-14	-19	-24

Figure 9 Signal Level v.s. Signal-to-Noise for a Standard Test Set

This correlation assumes many variables, however, after careful checks, it was found that the standard transistorized American TV sets vary ± 2 dB at the most from the chart. One must be careful of the following:

1. The clean signal must be free from any visible noise.
2. At some point in reducing the input to a TV set, the limit of the AGC range will be reached. Generally this will be coincident with a loss of sync lock.
3. The chart assumes a front end noise figure of 7 to 9 dB (normal for most portables).

The equipment for the actual test is setup as shown in Figure 10. The Signal Level Meter is used to monitor the distortion channel (A). The technician compares one channel at a time to the distortion channel to obtain a match. The attenuator is adjusted as necessary. Then using the S.L.M. values, the actual Signal-to-Noise ratio is derived from the chart (Figure 9).

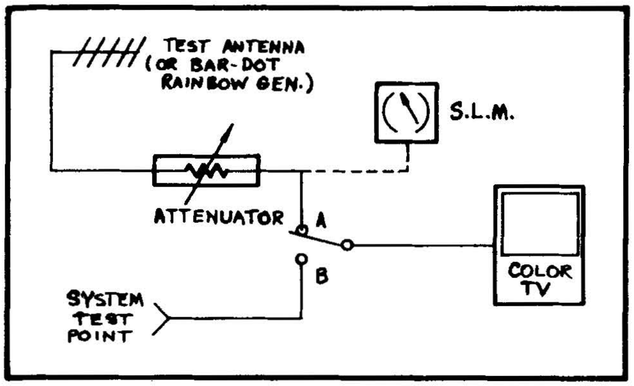


Figure 10 Equipment Setup for Subjective Signal-to-Noise (C/N) Test

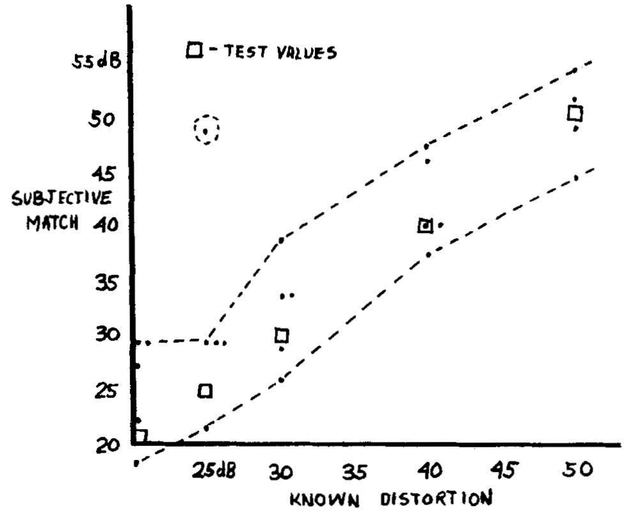


Figure 12 Validity of Subjective Intermod Test

VALIDATION OF SUBJECTIVE-COMPARISON TECHNIQUE

To get a rough idea of the accuracy one could expect from this form of testing, six observers measured the distortions on TV picutres. The distortions were also measured using a Tektronix Proof-of-Performance Package.

Standard off-air TV pictures were used. Some of the program material varied during the tests from excellent to marginal. No special precautions were observed as it was felt that the tests must represent a true, real life situation.

Figure 11 Using the A-B Comparison Technique

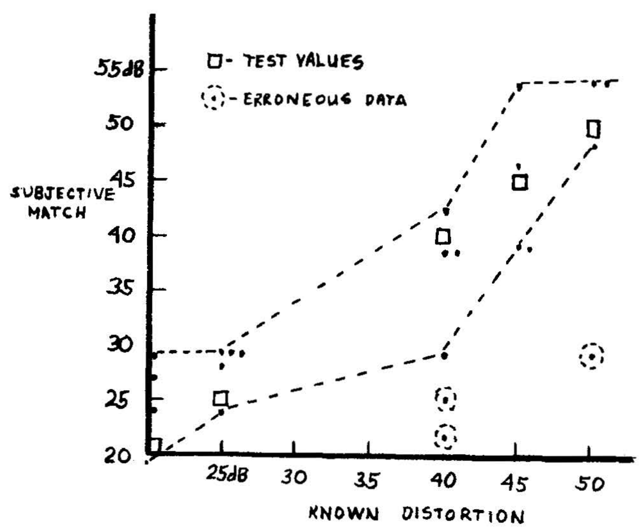


Figure 13 Validity of Subjective Co-Channel Test

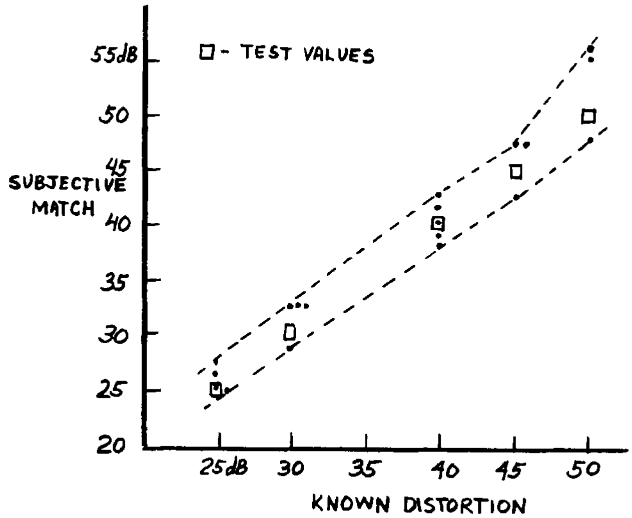


Figure 14 Validity of Subjective Signal-to-Noise Test

CONCLUSIONS

The results of this experiment were gratifying. It proved that subjective testing must be considered as a test technique. The remainder of the conclusions cover a variety of areas and are listed.

1. This experimant was only a start. Much work needs to be done in the area of subjective testing.

2. Subjective testing can be used to evaluate one and possibly two simultaneous distortions, such as intermodulation and noise. Beyond two, the test becomes quite confusing.

3. The lower ranges of each distortion are easiest to measure. CATV operators may have difficulty measuring higher distortion values (generally in excess of 50 dB) however, within the ranges recommended by the FCC as the minimum acceptable values, subjective testing can yield results within ± 5 dB or better.

4. Subjective testing is influenced by picture content. Various techniques should be explored such as B & W vs color, and raised contrast levels to make distortion more recognizable.

5. Simple training of the subjects increased accuracy tremendously. Some of the subjects could match within ± 2 dB consistently on some of the tests.

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