THE COMPLETE
TECHNICAL PAPER PROCEEDINGS
FROM:

NCTA Technical Papers

A DIGITAL VIDEO SYSTEM FOR THE CATV INDUSTRY

by

Donald Kirk and Michael J. Paolini

ST. PETERSBURG COMMUNICATIONS CORPORATION

ABSTRACT

A digital PCM system is proposed to implement long-haul video systems for the CATV industry. A comparison is made between PCM systems and wideband FM systems in terms of repeaterability versus additional occupied signal bandwidth. This is followed by a noise analysis of coaxial cable to determine the correct PCM format and bit information rate. An eighty megabit system is selected, which uses an eight bit code in a four level-eight level-eight level pulse sequence per video sample. The selected PCM system is then evaluated for its performance on both a long-haul cable system and a long-haul microwave system. Performance calculations are made on a 500 mile cable system in terms of error rate and its related video signal to noise ratio. Repeater spacings are computed for various configurations, including the replacement of the digital regenerators with analog amplifiers. Additional performance calculations are then made on a 3000 mile microwave system, including the effect of simultaneous Rayleigh fading.

Part 1: THE APPROACH

Whenever a new modulation system is proposed for a communication service, there should be some justification made for any additional level of complexity that may occur. This is especially true for a digital video system, which at first glance appears to be an expensive, complicated, "blue sky" approach to long-haul video cable networks. Indeed, the attitude of the CATV industry today toward such a system is not too unlike the telephone industry's initial reaction to the idea of a CATV industry----expensive, technically impractical, and definitely uneconomical. Historically, these objections have applied only to the time scale of implementation, rather than being an indication of the ultimate occurence of a system.

An important point to be made is that today's distribution systems have proven in the past, and will continue tobe in the future, the correct way to distribute signals within a three to five mile radius. It is improbable that ar y fundamental changes will occur at this level of distribution in the future. A digital video system would complement today's existing equipment by supplying a high quality video signal to local (five mile) distribution centers. This configuration appears quite reasonable in view of the fact that today's high quality trunk lines are at a marginal performance level beyond a distance of approximately 30 miles, especially when economic constraints are considered. This is demonstrated in Appendix 1, which shows the relationship between cost and system length for fixed performance standards. Once this basic limitation is acknowledged, alternative solutions appear in a more meaningful light.

Adapting from microwave techniques, a solution to the long-haul video problem might be to convert the video signal into a wideband FM signal. This will yield an improvement in the detected output signal to noise ratio at a cost of additional occupied signal bandwidth. However, all communication systems that improve the detected output signal to noise ratio do so by increasing the bandwidth of the transmitted information. It remains to be determined that if the bandwidth of a signal must be increased, what is the most efficient system in terms of signal to noise improvement versus increased bandwidth.

For an FM system, the improvement in S/N over an AM system may be written as:

(1) $$\frac{S/N_{fm}}{S/N_{am}} = 3k^2$$

where $k = f_d/f_m$ and

f_d = the peak deviation of the FM signal and

f_m = the highest modulation frequency of the video signal

The bandwidth of the FM signal is:

[1] Ref 10

(2) $BW = (2k + 2) f_m$ (Carson's Rule)

Now it is convenient to define $x = BW/2f_m$ which is the normalized increase in the bandwidth occupied by the signal compared to that of an AM system. Using this definition, and solving for k in equation (2), equation (1) may be rewritten as:

(3) $$\frac{S/N_{fm}}{S/N_{am}} = 3(x-1)^2$$ which is the signal to noise improvement between the two systems defined in terms of additional bandwidth occupied by the signal.

However, for a PCM system the signal to noise improvement compared to an AM system is:

(4) $$\frac{S/N_{pcm}}{S/N_{am}} = (S_c/N_c)^{BW/2f_m} = (S_c/N_c)^x$$ which is derived in Appendix 2 where S_c/N_c is the input signal to noise ratio of the PCM carrier.

This is the signal to noise improvement defined in terms of additional bandwidth occupied and the input signal to noise of the carrier. Equations (3) and (4) permit a comparison between the signal to noise improvement of an FM and a PCM system:

(5) $$\frac{S/N_{pcm}}{S/N_{fm}} = \frac{(S_c/N_c)^{x-1}}{3(x-1)^2}$$ which is the improvement of a PCM system over a wideband FM system defined in terms of additional occupied bandwidth and the input carrier to noise ratio.

The preceeding equations, when supplied with typical system parameters, quickly show the decided advantage of using a PCM system over an FM system for a given investment in occupied bandwidth. Taking x = 3 (which is the PCM system discussed in this paper), and an input carrier to noise ratio of 60dB we have the S/N_{fm} =71 dB. If we are to permit repeatering of the signal until a detected signal to noise of 59 dB is obtained (assuming this the maximum tolerable degradation), this would leave a 12 dB margin or 2^4 maximum doubles in repeater operations which would permit a total of 16 repeaters.

However, using the same parameters for a PCM system we find that the S/N_{pcm} = $(10^6)^2$ = 120 dB which yields a margin in excess of 60 dB and would permit 2^{20} doubles or over 10^6 repeater operations. In reality, the excessive repeater capability of a PCM system can be utilized by letting the input carrier to noise ratio degrade to a lower operating level than the equivalent FM system. For example, if we assume a S_c/N_c of 40 dB for the PCM system (compared to 60 dB for the FM system), the detected signal to noise would be S/N_{pcm} = 80 dB which would be a margin of 2^7 or 128 repeater operations compared to the 16 possible repeater operations in an FM system of equivalent bandwidth operating with a 20 dB greater carrier to noise ratio.

A second method of utilizing the excessive repeaterability of a PCM system is to use analog repeater amplifiers instead of digital regenerators at a repeater site.

Analog amplifiers may be used in cascade (reamplifying the digital signal) until the S_c/N_c has degraded to 40 dB (using the previous example at the higher operating level of 60 dB S_c/N_c and cascading analog amplifiers until 20 dB of degradation has occured.) At this point a digital repeater would be used to regenerate the signal, after which it may be decoded for distribution, or followed by additional analog repeaters.

Although the preceeding advatages of a PCM system over a wideband FM system are true for all types of PCM systems, several parameters must be evaluated to determine which PCM system would permit optimum performance for video when used with cable transmission. These include the "quantizing noise" which is the intrinsic noise level that occurs when a signal is divided into discrete steps, and the consideration as to whether the system is to be used at baseband on the cable or multiplexed at higher frequencies on the cable similar to the present AM systems. Appendix 3 contains a derivation of the quantizing noise that occurs in signal quantizing, and tabulates the results in terms of the number of levels that the continuous signal is quantized. Table 1 in Appendix 3 shows that if 256 levels are used, the intrinsic quantizing noise is reduced to 59 dB below signal level. The quantizing noise of a PCM system occurs at the original coding of the signal and is independent of the number of repeater operations. In fact, nearly all normal video specifications are independent of the number of repeater operations and dependant solely on single hop performance. These include differential gain, differential phase, frequency response, square wave tilt, and video bounce. Once the video signal has been placed in a PCM format, the only signal degradation that may occur are the errors that occur in the decision process of determining the correct occurence of a pulse. These "errors" are then reflected in the detected output signal to noise ratio of the video signal. The error rate of the PCM signal is proportional to the product of the bit rate and the probability of error. Appendix 4 gives a derivation of the probability of error in terms of input carrier to noise. Figure 1 in Appendix 4 graphs the probability of error versus carrier to noise ratio for binary, quaternary, and octenary level pulses. These curves will be used later to determine the error rate of the PCM system which may then be related to final output signal to noise level.

Another fundamental consideration of the PCM system is the determination of which portion of the cable spectrum it will occupy, especially under the conditon of transmitting multiple channels. Appendix 5 contains a cost comparison between single and multicore cables. It demonstrates that for a multi-channel system, it is no more expensive to use one small cable for each channel than it is to use one large cable with frequency division multiplexing. Transmission loss requirements dictate that multichannel PCM systems be transmitted on a multicore cable with each channel occupying the baseband (lower end) portion of the cable spectrum

THE SYSTEM

Fundamental to the formulation of a PCM system is the determination of the data rate required for the system. The data rate is composed of two factors: thesampling rate f_s multiplied by the number of bits per sample (= $\log_2 N$ where N is the number of quantizing levels). The minimum sampling rate required is:

$$f_s = 2f_m \qquad \text{(Nyquist sampling rate)}$$

where f_m is the highest frequency component of the baseband information. For color video only, f_m of 4.2 mHz would be required. However, if the system is required to caray inter-carrier sound, an f_m of 4.525 mHz would be appropriate. If a pilot is to be placed in the baseband above the video information, it would not be unreasonable to consider f_m extending to 5 mHz. This would imply a sampling rate of 10 mHz. Since the bandwidth required is directly proportional to the sampling rate, it is advisable to keep the sampling rate as close to the Nyquist rate as possible.

The second factor in the data rate is the number of bits per sample, which in turn is related to the number of quantizing levels used. The determination of the required number of quantizing levels is reached by a consideration of the delivered picture "quality" . Appendix 3 gives the relationships between the number of quantizing levels and the broadband signal to noise ratio due to "quantizing" noise.

On the basis of signal to noise only, Table One in Appendix 3 would indicate that a 6 bit code (64 levels) would yield a 47 dB S/N ratio, which would be adequate for many purposes. However, the implementation of a 6 bit code, when evaluated by subjective tests, indicate a "contouring" phenomenon due to insufficient quantizing levels. Contouring transforms continuous shading into a number of discrete steps. Contouring effects are a coherent type of interference, and therefore require a higher signal to noise ratio than would ordinarily be required for random noise. This is analogous to the cross modulation requirements of a video signal being much higher than the tolerable broadband signal to noise ratio.

If random noise is introduced to reduce the contouring effects, a usable picture may be transmitted with a 6 bit code, but the system will not meet commercial broadcast standards. By using an 8 bit code, contouring effects become negligible, and the system may now be used for other purposes, such as multichannel telephone. From Table One in Appendix 3, this corresponds to a broadband signal to noise ratio of 59 dB.

With the selection of an 8 bit code (256 levels) and a 10 mHz sampling rate, the system data rate is 80 megabits/sec. For a binary system, this would imply a transmitted bandwidth of 40 mHz, with eight pulses per sample. If a quaternary (4 level) system were used, the required bandwidth would be reduced to 20 mHz, and the number of pulses per sample would be four. Appendix 6 shows a noise analysis for a digital cable transmission system. The results in Appendix 6 indicate that a satisfactory pulse format would be three pulses per sample in a 2-3-3 bits per pulse arrangement. (This means a four level pulse followed be two eight level pulses for each sample.) This format also has the advantage of reducing the required bandwidth to 15 mHz, as well as having certain advantages in the area of synchronization, due to unique transition possibilities.

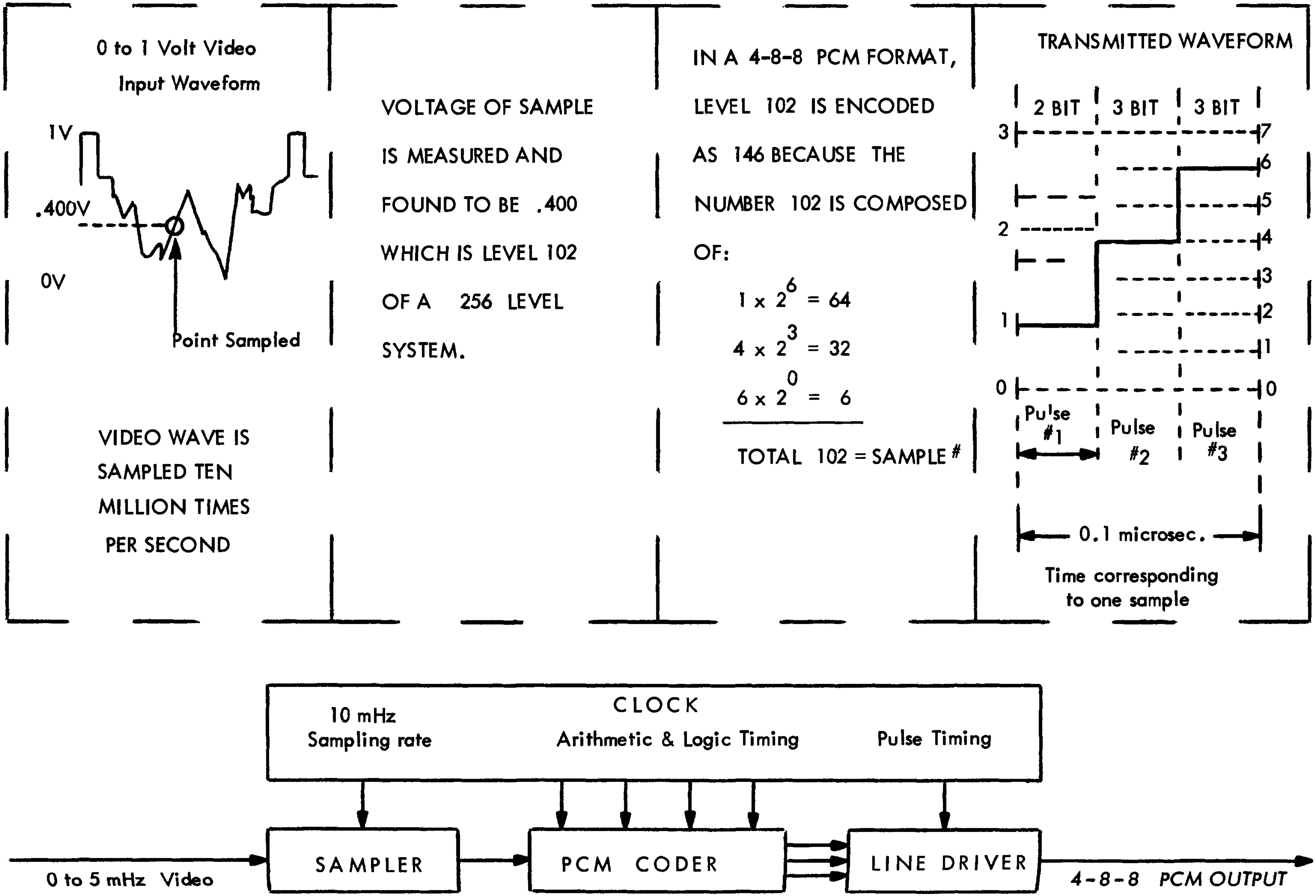
The synchronization problem in a PCM system is the determination of the correct weighting of the received pulses, that is, recognizing the sequence in which they were generated. This may be achieved by transmitting a unique code group periodically, or for a wired system, a CW clock may be passively added at the transmit end and recovered ahead of the first active device at the receive terminal. Where nonuniform coding is employed (such as the 2-3-3 bit format) synchronization may be achieved by recognizing the unequal transitions between the four and eight level pulses.

For a cable transmission system, it is desireable to employ a pulse format which does not necessitate the transmission of low frequency signals. Coaxial cables exhibit very uniform phase characteristics above frequencies of 100 kHz. Also, transformer coupling elements are more easily achieved if the number of decades of frequency range covered is held to three or four. Another advantage of not using the very low end of the cable spectrum is that hum pickup can be ignored and power for the repeater stations can be placed directly on the cable.

An eight bit code with a 2-3-3 bit format would require a pilot signal level approximately 30 dB below the video signal to ensure that the third pulse of the sample group will periodically oscillate through all possible values, even in the absence of an input video signal. By using a pilot at 4.7 mHz, it is possible to restore frequency components below 300 kHz. The pilot also permits AGC action in the receiver by guaranteeing frequent occurrence of maximum and minimum level pulses.

The preceding discussion of the PCM system is best summarized by the drawing on page 12. This drawing shows the input video signal being sampled at a rate of 10 million times per second. These output samples (which are directly proportional to the input video signal) are then quantized into 256 possible levels. Whichever of the 256 levels is selected for a sample becomes processed into a three pulse format where the first pulse can have four possible levels and the second two pulses can have eight possible levels (2-3-3 bit PCM format). The processing of a single sample is illustrated on page 12 in the drawing above the block diagram.

FUNCTIONAL DIAGRAM OF 4-8-8 PCM SYSTEM



Part 3: PERFORMANCE CALCULATIONS

Consider a single video channel to be transmitted on a 500 mile cable system, using the previously described PCM system. Assume a peak power of .1 Watt is the output at the transmitter or at any repeater site. The noise figure of the receiver and any repeater amplifier (either analog or digital) is taken to be 10 dB. From Appendix 6, the highest frequency to be transmitted down the cable is 15 mHz (F). For a video signal to noise at the end of the system of 56 dB (that is, line contributed noise equal to quantizing noise), the video signal to noise at the receiver due to error rate performance of the system must be 59 dB. The error rate for a 59 dB video signal to noise may be computed using Equation (8) of Appendix 4 or interpolated from Table Two in Appendix 4. In either method, the corresponding error rate for a 59 dB signal to noise of the video signal is 1.3×10^{-3}. These errors can be assumed to be equally contributed by each repeater of the system.

Conservatively, assume there will be 500 repeater sites in the system. (This is a pessimistic number, as the results will show that 114 repeaters are required.) This would mean that each repeater would be permitted to contribute an error rate of 2.6×10^{-6}. All of these errors will occur in the two eight level pulses (since the four level pulse is essentially operating at a higher signal to noise ratio).

The required carrier to noise ratio at the input of a regenerator to produce 2.6×10^{-6} error rate may be found by reading Figure One of Appendix 4 or interpolating Table Two of Appendix 4. The required carrier to noise ratio is 38 dB, and for a .1 Watt signal level, this defines the noise power out of a repeater to be -18 dBm. By using Equation (5) of Appendix 6, the permissible cable loss between repeater sites is found to be 61 dB at a frequency of 10 mHz. The actual distance will depend on the diameter of the cable used. For representative cable of different diameters, the results may be tabulated as follows:

Cable Diameter	Atten. dB/mile @ 10 mHz	Repeater Spacing	# Repeaters
.4 inches	13.7	4.4	114
.3 inches	18.3	3.3	150
.2 inches	27.4	2.2	225
.1 inches	54.8	1.1	450

For a cable diameter of .4 inches (.26 dB/100 ft @ 10 mHz) the repeater spacing will be 4.4 miles, which would require 114 repeaters. This shows the noise assumptions to be quite conservative, since even .1 inch cable will require less than 500 repeaters.

The preceding calculations are predicated on using digital regenerators at all repeater sites. It is interesting to investigate the possibility of adding analog amplifiers between the digital regenerators of this system. If n analog amplifiers are placed between each pair of digital regenerators, the attenuation spacing, K_d, for the digital system must be reduced by $\frac{3 \log_2 (n+1)}{n}$. The total system attenuation for N digital repeaters was:

$A_{system} = N K_d$. With analog amplifiers inserted in the system, this becomes:

$$A_{system} = N \cdot \left[(K_d - 3/n \cdot \log_2(n+1)) \cdot (n+1) \right] \quad (1)$$

For the smallest cable on the tabulation of page 14, the system loss for 500 miles was:

$A_{system} = 450 \times 61 = 2740$ dB . Inserting seven analog amplifiers between each regenerator we have: (Using Equation (1) above)

$A_{system} = 21.5 \times 10^4$ dB and the system length is 3900 miles. This indicates that the limit to this technique will be set by the difficulties of phase equalization and gain control, rather than any noise considerations.

In a similar manner, we could have taken the original 500 mile system and replaced all of the digital regenerators with analog amplifiers. For .4 inch cable, the system would have required 180 analogue amplifiers at a spacing of 2.7 miles. Again, this illustrates that thermal noise will not be a limiting constraint on the PCM system.

To allow a comparison between cable and radio systems, apply the 2-3-3 bit PCM format to a microwave transmitter. Let a combined amplitude and phase shift keying modulation system be used. (Four phase-two amplitude with no zero level). One phase position of the four level pulse may have itself identified by being transmitted at minimum rather than maximum level. This will allow phase identification for coherent detection. The signal may be filtered to produce a vestigial sideband spectrum which may be transmitted in a 20 mHz radio channel. The probability of error for such a channel (P_e) when subjected to Rayleigh fading is given approximately by:[2]

$$P_e = \frac{5}{C/N}$$ for C/N greater than 10 ~~(After Stein, Ref. 9)~~

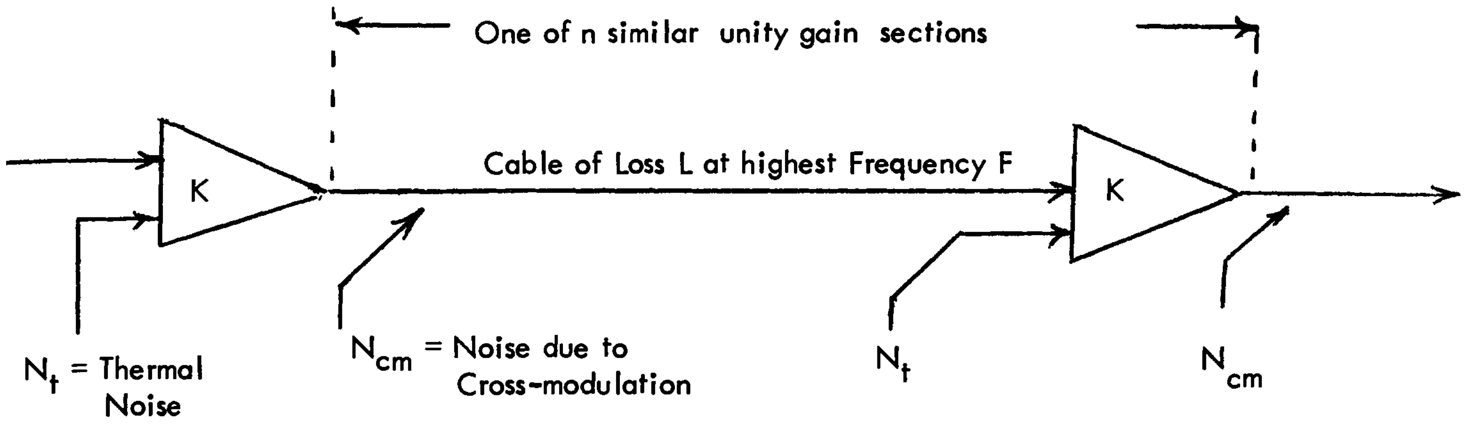
(13 dB worse than a binary DSB-PSK system)

For a 500 mile radio system consisting of 20 repeaters at a 25 mile spacing, and a system error probability requirement of 1.3×10^{-3} (as for the cable system), the individual repeater links must contribute only $P_e = 1.3 \times 10^{-3} / 20 = .65 \times 10^{-4}$.
The required carrier to noise ratio into the regenerator site is:
$C/N = 5/P_e = 7.7 \times 10^4$ or 48.6 dB .
Placing the channel at 12.5 gHz with antennas of 4 ft. diameter yields a space loss of 146.5 dB and an antenna gain (2 antennas) of 83 dB. With a 12 dB receiver noise figure the tangential sensitivity is -89 dBm and the required input carrier to noise is 48.6 dB above this or -40.4 dBm. Adding this level to the path loss minus the antenna gain yields a transmitter power requirement of +22.6 dBm which is well within the state of the art for solid state devices. The corresponding calculations for 120 repeaters (3000 mile system) yield a transmitter power requirement of +30.6 dBm which is approximately state of the art for solid state sources at 12 gHz at the present time. By increasing the antenna size, a satisfactory 3000 mile system which operated in the presence of simultaneous Rayleigh fading can be achieved.

[2] After Stein, Ref 9

APPENDIX ONE

LIMITING CONSTRAINTS ON EXISTING CABLE SYSTEMS



The preceeding drawing shows one of n unity gain sections of a cable transmission system. Let the video signals be applied in a frequency division multiplex arrangement using a modulation system comparable to conventional broadcast standards. Let the cable loss for the highest frequency channel be L and the gain of the amplifier (including a built in cable equalizer) be K where

(1) $K = 1/L$

Thermal noise of power density N_t ($N_t = 4 \times 10^{-15} f$ watts per megacycle where f is the noise factor of the amplifier input) is added at the input of each amplifier and cross modulation noise power N_{cm} (due to amplifier non-linearities) is added at the output of each amplifier. $N_{c\,n}$ may be assigned a spectral density of watts/ mHz where:

(2) $N_{cm} = MW_o^3$ where W_o is the output power per channel and M is an amplifer constant derived by subjective test. Essentially, M corrects for the fact that the well correlated third order cross modulation products cause much more picture degradation than would a similar amount of white noise. M may be written numerically by:

(3) $M = 1/BCW_t^2$ where B = the channel bandwidth in megacycles
C = S/N (power ratio) for just visible thermal noise and
W_t = Single channel output power for just visible cross modulation interference on a single amplifier test.

For a cascade of n amplifiers the equivalent noise power out of the last amplifier is:

(4) $W_n = nB(N_{cm} + KfN_t) = nB(MW_o^3 + KfN_t)$ where W_n is the total noise power out of the system. The signal to noise ratio W_o/W_n is:

(5) $S/N = W_o/nB(fKN_t + MW_o^2)$

An optimum system will operate at a level where thermal noise and cross modulation effects cause equal degradation. At this level of operation:

(6) $fKN_t = MW_o^3$ or $W_o = (fKN_t/M)^{1/3}$

Substituting this value of W_o in (5) we have

(7) $S/N = 1/2nB(Mf^2K^2N_t^2)^{1/3}$

When an amplifier noise factor and output capability M is known, the maximum number of amplifiers in cascade for a given output signal to noise ratio may be determined as

(8) $n = AK^{-2/3}$ where $A = \dfrac{1}{2B(S/N)\sqrt[3]{M(fN_T)^2}}$

The length of the whole system in terms of total system gain G is :

(9) $G = K^n = K^{AK^{-2/3}}$

To find the maximum value of G (and therfore the longest system for a given cable loss) we differentiate (9) and set $dG/dK = 0$. Hence,

(10) $dG/dK = 1 - 2/3\log_e K = 0$ or $K = e^{3/2}$ or 6.5 dB for amplifier gain.

Substituting in (8) yields the number of amplifiers for the longest system as

(11) $n_{max} = AK^{-2/3} = A/e$ and the total system gain is:

(12) $G_{max} = K^n = (e^{3/2})^{A/e} = 2.33A$ dB Where A is defined in Equation (8)

If representative numbers are placed in Equation (12) (M has a range of 20 to 200) we find the loss of the maximum cable system to be in the range of 800 to 1200 dB, depending on the particular amplifier chosen. This implies that the only method of extending the cable system is to arbitrarily increase the size of the cable (provided the best available amplifier has already been used). This is the least attractive method of increasing system length since the cost of the cable per unit length is proportional to the square of the diameter. It is convenient to define a cable figure of merit s:

(13) $s = Ld/\text{length}$ where L is the loss at Channel 13 and d is the diameter of the cable (inche and the length is measure in miles. For some of the new foam dielectric cables, s has a value of 27.5 dB/mile/inch diameter of cable. This would predict a maximum system length in the range of 30 to 40 miles if one inch diameter cable were used. However, in this development there was no provision for operating level tolerance, echo distortion noise due to vswr, and cumulative frequency response effects, all of which tend to shorten the maximum operating length.

At the point where the only alternative to increasing system length is to increase the diameter of the cable, the total cable cost of the system is:

(14) $Q = pd^2 \times (\text{system length})$ Where Q is the total cable cost and p is a constant equal to the cost of one mile of one inch diameter cable.

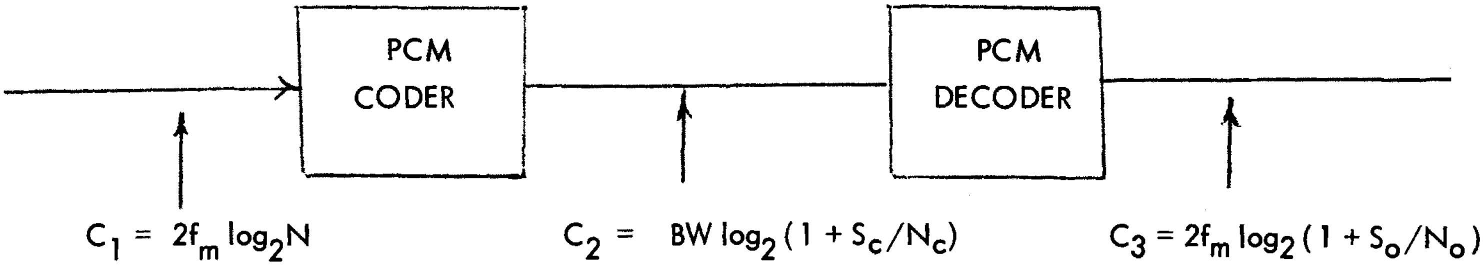
Recognizing that $d = s/L \times (\text{system length})$ and that we are operating in a range where $L = 2.33A$ (maximum system length) we have:

(15) $Q = p(s/2.33A)^2 \times (\text{system length})^3$

The results of Equation (15) may be interpreted as follows: For a short cable system (where the diameter of the cable need not be increased), the cable cost is proportional to the first power of system length. Beyond 200 to 300 dB (where cable size can be compromised by using additional amplifiers) the system designer may accept an increase in the exponent of system length (to a squared function) to make possible the use of fewer amplifiers and to provide gain margin for maintanence ease. However, in the range of 800 to 1200 dB (where the only alternative is to increase cable diameter), the cable cost starts to vary as the third power of the system length, because the limit on the quality of amplifiers has been reached and the only way on increasing system length (or level margin) is to increase the diameter of the cable used.

APPENDIX 2

DERIVATION OF SIGNAL TO NOISE IMPROVEMENT IN A PCM SYSTEM



The above diagram shows the information capacity C of a PCM system that accepts an N level quantized signal and coverts it into a PCM format for transmission, and reconverts the signal back to its original input after transmission. C_1 is the information rate expressed in terms of the highest baseband frequency f_m and the number of quantization levels N.[3] C_2 and C_3 are the information capacity of the system expressed in terms of bandwidth BW and signal to noise ratio (Shannon's Law)[4] S_c/N_c is the carrier to noise input to the decoder, and S_o/N_o is the signal to noise ratio after the decoding process.

If we postulate that there is to be no information lost in the system (i.e it will be operated at a high enough signal to noise ratio to maintain the required information capacity), the information capacity at all points in the system must be equal, and C_2 must be equal to C_3. Therefore:

(1) $$C_2 = BW \log_2(1 + S_c/N_c) = 2f_m \log_2(1 + S_o/N_o)$$

which may be written as:

(2) $$(1 + S_o/N_o) = (1 + S_c/N_c)^{BW/2f_m}$$

For signal to noise ratios much greater than one, (2) may be expressed as:

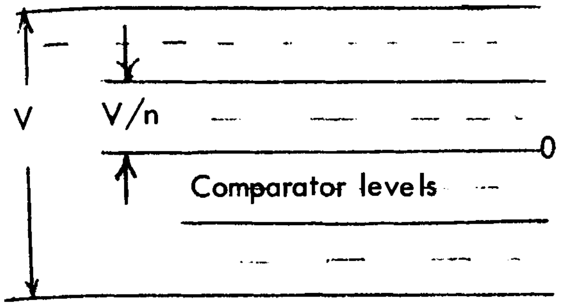
(3) $$S_o/N_o = (S_c/N_c)^{BW/2f_m}$$

3 Ref 6 Bibliography

4 Ref 11 Bibliography

APPENDIX 3

DERIVATION OF QUANTIZING NOISE IN A PCM SYSTEM



Consider a peak to peak signal V volts to be quantized into n levels. The spacing between levels is V/n volts. The quantizing comparators would be set at $\pm V/2n$, $\pm 3V/2n, \ldots (2n-1)V/2n$. Each quantized output level represents all signal values in the range of +V/2n to -V/2n about its value. The difference between the quantizing level and the true signal value is the error introduced into the system. Assuming that over a long perido of time all values of signal in the uncertainty range of +V/2n to -V/2n are equally likely, the signal may be described as $A_i + E$ where A_i is the level being transmitted, and E represents the error voltage between the actual signal and its quantized equivalent. From before, E must fall betwsen the range of +V/2n to -V/2n. We may then wri te the mean squared value of E as:

$$\bar{E}^2 = n/V \int_{-V/2n}^{V/2n} E^2 \, dE = V^2/12n^2$$

The rms value of the error is then $V/n\sqrt{12}$, and since the peak to peak signal is V, the peak to peak signal to the rms error noise (quantizing noise) is $n\sqrt{12}$ in voltage ratio or $12n^2$ for the corresponding power ratio. This is the signal to "quantizing noise" ratio, and it may be made increasingly large by arbitrarily increasing the number of levels.

TABLE ONE

QUANTIZATION S/N vs NUMBER OF LEVELS

Quantization S/N dB	# of levels, n
17	2
23	4
29	8
35	16
41	32
47	64
53	128
59	256

APPENDIX 4

ERROR RATES VERSUS VIDEO SIGNAL TO NOISE

Consider a 1V p - p video signal transmitted in a four level-eight level-eight level pulse sequence per sample at a rate of 10^7 samples per second. Let the output level of the PCM transmitter be peak power limited at a level of $P = V_o^2$. The the power for any one step in an n level pulse is: [6]

(1) $P_n \ (V_o/n-1)^2$ and the probability of error (P_e) is:

(2) $$P_e = \frac{2(n-1)N}{\sqrt{2\pi}} \cdot \exp(-V_o^2 / 8(n-1)^2 N^2)$$

where N is the rms noise voltage, and V_o/N is the signal to noise ratio in terms of p-p signal to rms noise.

Equation (2) may be used to obtain separate expressions for the probability of error for a binary, quaternary, and octenary level pulse:

(3) Binary: $P_{e_2} = 0.8(N/V_o) \cdot \exp(-0.125 V_o^2 / N^2)$

(4) Quaternary: $P_{e_4} = 7.2(N/V_o) \cdot \exp(-0.0139 V_o^2 / N^2)$

(5) Octenary: $P_{e_8} = 392(N/V_o) \cdot \exp(-0.00255 V_o^2 / N^2)$

Equations (3), (4), and (5) are presented graphically in Figure One which shows the relation ship between error rate, input signal to noise ratio.

It is useful to be able to convert readily between error rate, carrier to noise ratio, and output video signal to noise ratio. These relationships are dependent on the type of PCM format that is used. For the four level-eight level-eightlevel pulse sequence described in the text the following relationships apply:

If any of the three pulses of the code (4-8-8) are misread due to the presence of noise, there will be a noise voltage included in the decoded output. The rms value of this noise voltage will depend upon the probability that an error will occur and the noise power introduced

6 Ref 3 Bibliography

by a single error in that pulse. Since the pulses are weighted, the noise contribution of their respective errors is unequal. If normalized values of V_o = 1V p-p and Z_o = 1 ohm, we may add the noise powers contributed by each of the pulses to produce:

$$(6) \qquad W_n = (1/4)^2 \cdot P_{e4} + (1/32)^2 \cdot P_{e_8} + (1/256)^2 \cdot P_{e8}$$

where W_n is the noise power and P_{e4}, P_{e8} are the probability of error in a four level and eight level pulse respectively. To correct for second order effects, we should add the noise power contributed by 2 level errors on the eight level pulses. A 2 level error on an eight level pulse is equivalent (in probability) to a single level error on the four level pulse and has twice the weighted error of a single error that occurs on the same pulse. From this we have:

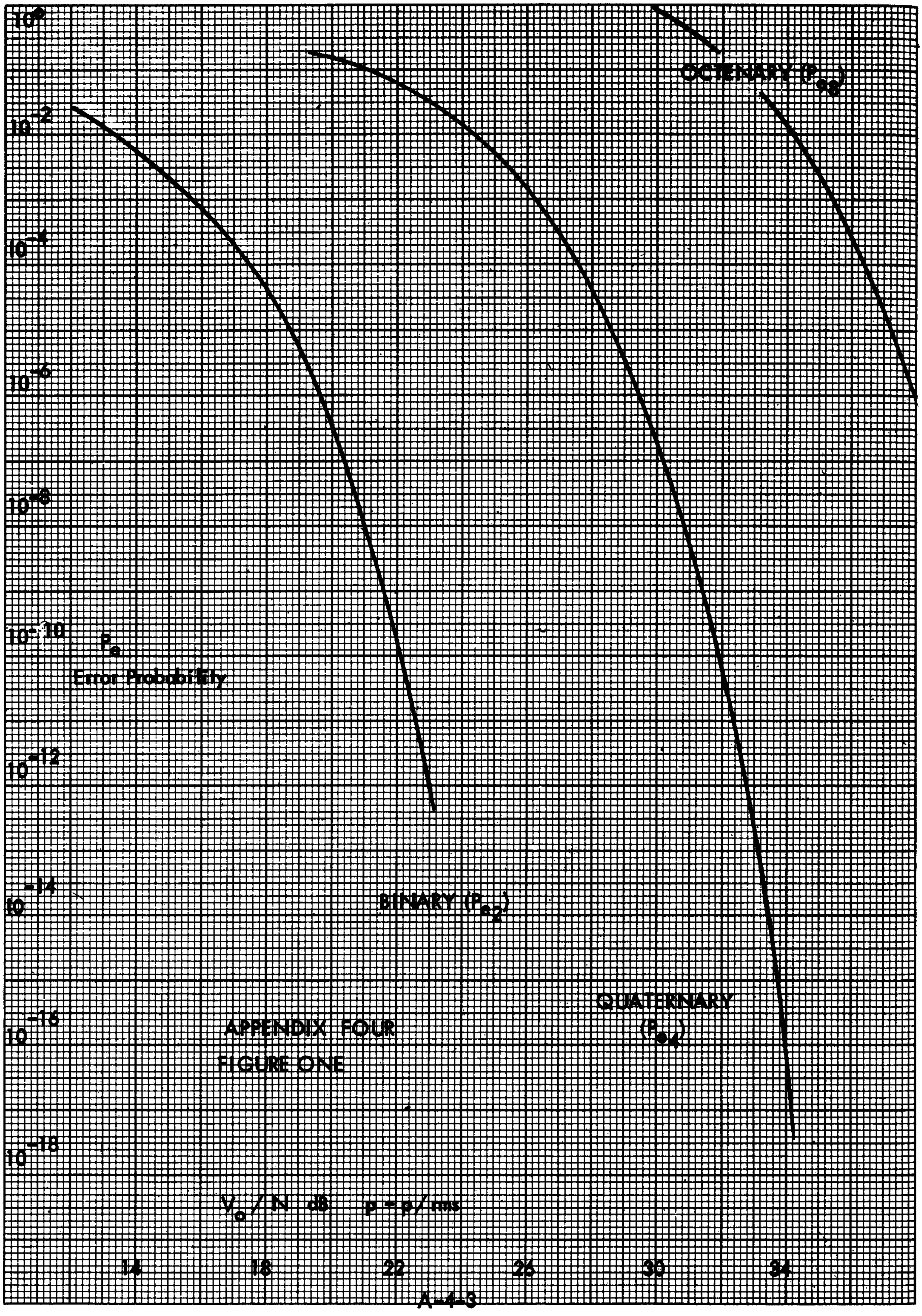
$$(7) \qquad W_n = (1/4)^2 \cdot P_{e4} + (1/32)^2 \cdot P_{e8} + (1/16)^2 \cdot P_{e8} + (1/16)^2 \cdot P_{e4} + (1/128)^2 \cdot P_{e4}$$

which upon collecting terms becomes:

$$(8) \qquad W_n = 6.64 \times 10^{-2} \cdot P_{e4} + 9.78 \times 10^{-4} \cdot P_{e8}$$

Equation (8) is useful in converting noise power into error rates directly, provided W_n is expressed directly in watts. By using equation (8) and Figure One, we are able to compute the video signal to noise in terms of input carrier to noise. This has been done, and the results are tabulated in Table Two which also includes the corresponding error rates for the four and eight level pulses.

The results of Table Two are shown graphically in Figure Two which shows the output video signal to noise ratio in terms of input carrier to noise ratio for the 4-8-8 code format. Notice that for high signal to noise ratios (carrier to noise) that the video signal to noise is limited by the intrinsic quantizing noise of the PCM system. As the carrier to noise degrades, so does the video signal to noise, first due to errors in the eight level pulses, and finally due to errors in the four level pulses.



OCTENARY (P_{e8})
10^{-2}
10^{-4}
10^{-6}
10^{-8}
10^{-10}
P_e
Error Probability
10^{-12}
10^{-14}
BINARY (P_{e2})
QUATERNARY (P_{e4})
10^{-16}
APPENDIX FOUR
FIGURE ONE
10^{-18}
V_o / N dB p = p/rms
14
18
22
26
30
34

TABLE TWO

CARRIER TO NOISE VERSUS VIDEO SIGNAL TO NOISE

C/N dB	P_{e4}	P_{e8}	Video S/N dB
23	3.25×10^{-2}	1	24.8
24	1.38×10^{-2}	1	27.2
25	5.02×10^{-3}	1	28.9
26	1.43×10^{-3}	1	29.5
27	3.01×10^{-4}	1	30.0
28	4.32×10^{-5}	1	30.0
29	3.81×10^{-6}	1	30.0
30	2.08×10^{-7}	9.68×10^{-1}	30.5
31	5.60×10^{-9}	4.08×10^{-1}	34.0
32	5.03×10^{-11}	1.75×10^{-1}	37.7
33	1.84×10^{-13}	5.62×10^{-2}	42.5
34	9.90×10^{-17}	1.29×10^{-2}	49.2
35	1.10×10^{-20}	2.23×10^{-3}	56.5
36		2.47×10^{-4}	66.0
37		1.52×10^{-5}	78.0
38		5.04×10^{-7}	93.0
39		7.47×10^{-9}	111.
40		3.29×10^{-11}	135.

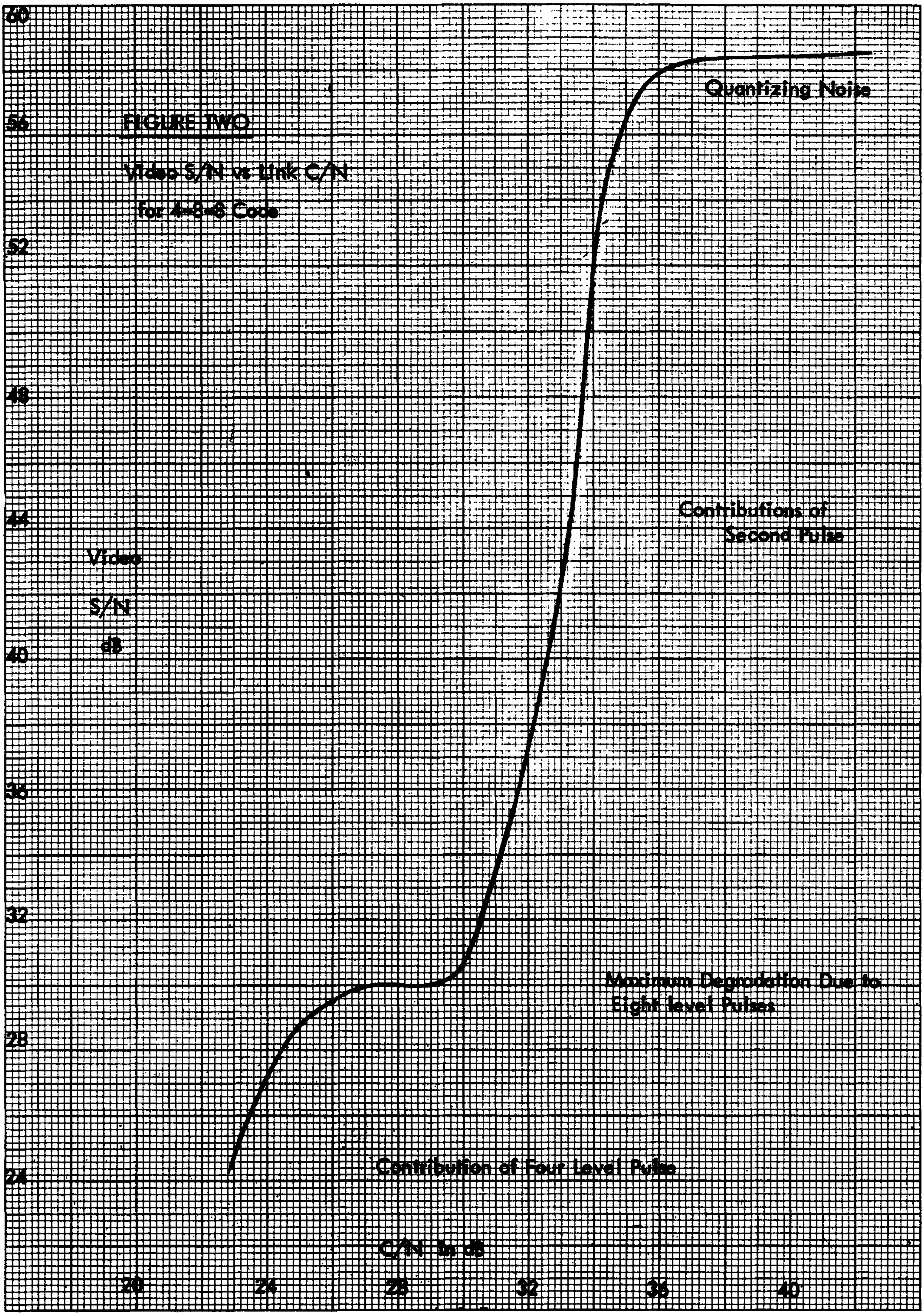


FIGURE TWO
Video S/N vs Link C/N
for 4-8-8 Code
Quantizing Noise
Contributions of
Second Pulse
Video
S/N
dB
Maximum Degradation Due to
Eight level Pulses
Contribution of Four Level Pulse
C/N in dB
60
56
52
48
44
40
36
32
28
24
20
24
28
32
36
40

APPENDIX 5

COST COMPARISON OF SINGLE VS MULTICORE CABLES

The cost of a coaxial cable may be represented by the expression

(1) $C = k_c D^m$ where
C = cost per unit length of cable
k_c = constant determined by type of cable construction
D = the diameter of the cable
m = a number approximately equal to 2 depending on the processing costs of the cable. If the cost of the cable were all material and no processing costs were involved, m would be 2 since the material content per unit length varies with the square of the cable diameter.

The attenuation of a cable where dielectric losses can be neglected is :

(2) $A = k_a R$ where
A = attenuation in nepers per unit length
k_a = a constant determined by the characteristic impedance
R = conductor resistance per unit length

The conductor resistance per unit length is:

(3) $R = \frac{k_r \sqrt{f}}{D}$ where k_r is a constant determined by the conductor material and f is the frequency in Hz.

Substituting (3) into (2) and solving for D, the cable diameter, equation (1) may be written as:

(4) $C = k_c(k_a k_r \sqrt{f} \; A^{-1})^m$

Now if N channels of bandwidth B are to be transmitted on Q cables, the highest frequency that must be transmitted is:

(5) $f = NB/Q$

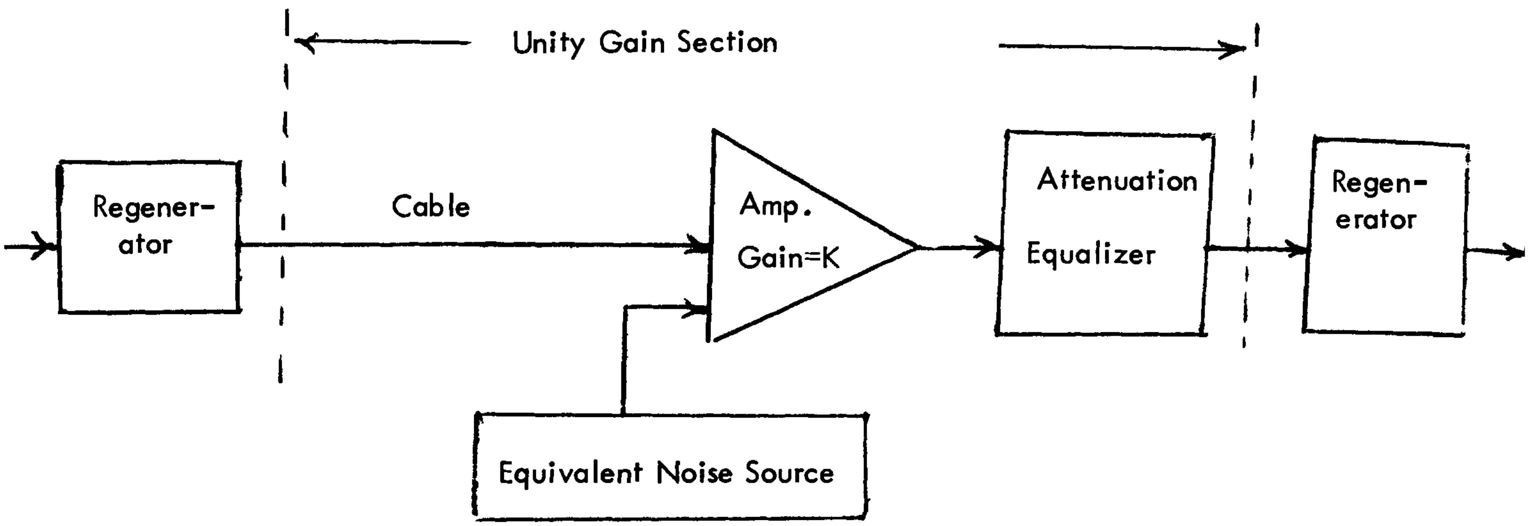
Substituting (5) into (4) and recognizing the total cost C_t is Q times C , we have for m approximately equal to 2 :

(6) $C_t = k_c NB(k_a k_r/A)^2$ which is independent of the number of cables used.

The interpretation of the above result is that for a multi-channel system, it is no more expensive to use one small cable for each channel than it is to use one large cable (whose cross-sectional area is equal to the sum of all the small cables) with frequency division multiplexing. For shorter cable runs, the cost of the multiplexing equipment would be the controlling element, whereas for long runs, the cost of single channel re-amplifying equipment becomes a consideration. In all cases, the multi-conductor cable has the reliability advantage in that a single channel failure does not interupt service on the other channels.

APPENDIX SIX

NOISE IN A DIGITAL CABLE REPEATER



The preceeding drawing shows a typical digital repeater section, with the regenerator driving the coaxial cable whose loss is given by:

(1) $A = \exp(-s\sqrt{f/f_o})$ where A is the cable attenuation and f is the frequency, and s is defined to be:

$$s = (1/4.343)(\text{loss in dB of the cable at frequency } f_a)$$

The output of the cable is combined with a white noise signal at the input of a flat amplifier. The noise source provides an amplifier input of power density $N_d = (4 \times 10^{-15})(d)$ watts per megacycle, where d is the noise factor of the amplifier. The gain of the amplifier is a factor K.

Following the amplifier, there is an equalizer for cable attenuation. Its Loss L may be described by the law:

(2) $L = \exp(s\sqrt{f/f_o}) \times \exp(-s\sqrt{f/f_o})$

For a unity gain system, we can set $K = \exp(s\sqrt{F/f_o})$ where F is the maximum frequency to which the system is to be equalized.

For a unity gain repeater section A x K x L = 1 over the frequency range from 0 to F. The noise out of the unity gain section in an incremental frequency range, df, is:

(3) $dW_n = N_d K L df = N_d \exp(s\sqrt{f/f_o}) \cdot df$ where W_n is the noise power in the section.

Therefore, W_n may be written as:

(4) $$W_n = \int_0^F \exp(s\sqrt{f/f_o}) \cdot df = (2N_d f_o/s^2)\left[\exp(s\sqrt{F/f_o}) \cdot (s\sqrt{F/f_o}) + 1\right]$$

Where the cable attenuation is high, $s\sqrt{F/f_o}$ is much greater than 1, and Equation (4) becomes:

(5) $$W_n = (2N_d\sqrt{f_o F}/s) \cdot \exp(s\sqrt{F/f_o}) = 2N_d F \cdot \frac{\exp(x)}{x}$$ where $x = s\sqrt{F/f_o}$

and $\exp(s\sqrt{F/f_o}$ is the gain of the repeater amplifier. Equation (5) gives the noise power in a digital cable repeater section. The problem remains to determine an upper bound for F interms of the information rate and the number of levels used in the PCM system.

Assume that the signal to be transmitted is a pulse of n possible levels, and the system is peak power limited to W_s watts. The power change represented by one level is then:

(6) $W_l = W_s/n-1$ and the signal to noise out of the repeater link (S/N) is:

(7) $$S/N = W_l/W_n = \frac{W_s/n-1}{2N_d F \cdot \frac{\exp(x)}{x}}$$ where S/N is the signal power for one level change divided by the total channel noise power.

For a give information rate R and a pulse repition rate of 1/T we have:

(8) $R = (1/T)\log_2 n$ and by substituting $F = 1/2T$ where F is the maximum frequency,

(9) $F = R/2\log_2 n$ which may be substituted into Equation (7) to yield:

(10) $$S/N = \frac{W_s \log_2 n}{4RN_d(n-1)} \cdot \frac{y}{\exp(y)}$$ where $y = s \cdot \sqrt{\frac{R}{2f_o \log_2 n}}$

Equation (10) permits the calculation of the repeater spacing in terms of cable loss s versus the number of levels used for fixed system parameters. For the PCM system discussed in the text, we have as parameters:

W_s = .1 Watts
R = 8×10^7 bits per second
N_d = 10^{-19} watts per cycle (14 dB noise figure)
S/N = 200 (10^{-12} error rate)
f_o = 10 mHz

By selecting f_o = 10 mHz, we will obtain our results directly in terms of 10 mHz cable loss. The substitution of the preceeding values into Equation 10 is tabulated as follows:

n	s	s in 10 mHz cable loss (dB)
2	9.77	42.4
4	13.5	58.6
8	16.1	69.9
16	18.1	78.4
32	19.7	85.3
64	20.8	90.3
128	21.7	94.2

The above tabulation, which is for a fixed error rate (10^{-12}) shows that if thermal noise were the only consideration, there would be no penalty in raising the number of levels transmitted per pulse. This means that thermal noise effects are not a consideration in determining the number of levels used in a PCM format, and that other effects such ringing, overshoot, and delay distortion should be considered.

The 4-8-8 level code selected for the 80 megabit system represents a compromise between n =4 (4-4-4-4)of the above tabulation (which would occupy a bandwidth of 20 mHz and n = 8 (which would occupy a bandwidth of 13.3 mHz.)

BIBLIOGRAPHY

1. Bennett, W.R. , "Spectra of Quantized Signals," Bell System Technical Journal, Vol. 27, 1948.

2. Bennett, W.R., Davey, James R., Data Transmission McGraw-Hill, Inc., 1965 .

3. Bell Telephone Laboratories, Transmission Systems for Communications, Third Edition, Bell Telephone Laboratories, Inc., 1964 .

4. Chang, R.W.and Freeny, S.L., "Hybrid Digital Transmission Systems", Bell System Technical Journal, Vol. 47, No. 8, 1968.

5. McMillan, B., "Communications Systems Which Minimize Coding Noise", Bell System Technical Journal, Vol. 48, No. 9, 1969.

6. Panter, Philip F., Modulation, Noise, and Spectral Analysis, McGraw-Hill, Inc., 1965.

7. Pierce, J.R., "Information Rate of a Coaxial Cable with Various Modulation Systems," Bell System Technical Journal, Vol. 45, No. 8, 1966.

8. Schreiber, W.F., and Knapp, C.F., "T.V. Bandwidth Reduction by Digital Coding," IRE Conv. Record, Part 4, 1958.

9. Schwartz, M., Bennett, W.R., Stein, S., Communication Systems and Techniques, Vol. 4, McGraw-Hill, Inc., 1966.

10. Schwartz, M., Information Transmission, Modulation,and Noise, McGraw-Hill, Inc., 1959.

11. Shannon, C.E., " A Mathematical Theory of Communication," Bell System Technical Journal, Vol. 27, 1948.

DISCUSSION

Mr. Kirk: Are there any questions?

Mr. Lowe: Mr. Lowe, Stanford Research Institute. Did you use any linear quantizing or unlinear quantizing levels for your mode?

Mr. Kirk: In what we've done here in the paper, we assume that everything was linear. This obviously--there's a reason for doing this--if you use linear quantizing now you don't have to argue with somebody when you tell them it can be used for multi-channel telephone. It's obvious that with sync pulses sticking up on top of the video, a great number of quantizing levels could easily be saved by just saying it's either black level or now it's sync because there's only two steps there and you need not worry about it. Are there other questions? Thank you.

A FREQUENCY MEASUREMENT TECHNIQUE FOR CATV

WARREN L. BRAUN, PRESIDENT
WARREN BRAUN CONSULTING ENGINEERS
HARRISONBURG, VIRGINIA

THE REQUIREMENT TO MEASURE FREQUENCIES ACCURATELY IN THE CATV FIELD IS ONLY NOW BECOMING RECOGNIZED AND RECEIVING ITS PROPER PLACE IN THE TECHNOLOGY OF THE INDUSTRY. THIS TOPIC IS BROUGHT INTO FOCUS BY EXAMINING SOME EXAMPLES OF THE REQUIRED ORDER OF MAGNITUDE FREQUENCY MEASUREMENT PRECISION REQUIRED FOR CERTAIN APPLICATIONS. FIRST, EXAMINE THE ALIGNMENT OF HEAD END PROCESSING EQUIPMENT. IN THIS CASE THE SHOULDER OF THE RESPONSE CURVE IS TO BE ACCURATELY PLACED AT A POINT BELOW (AND ABOVE) THE CARRIER, AS SHOWN IN FIGURE 1. THE RECOVERED VIDEO RESPONSE OF THE SYSTEM IS DEPENDENT UPON THE PHASE CHARACTERISTICS, AND IN TURN THIS IS DEPENDENT UPON THE AMPLITUDE CHARACTERISTICS,[1] PARTICULARLY IN THE VICINITY OF THE VISUAL CARRIER AND THE CHROMINANCE SUBCARRIER. THEREFORE, IT IS QUITE IMPORTANT THAT THE LOCATION OF THESE RESPONSE SHOULDERS BE ALIGNED WITH AN ACCURACY OF + - 25 KHZ (MINIMUM!). THE REQUIRED ACCURACY IS 25 KHZ OUT OF 200 MHZ OR APPROXIMATELY 1 PART IN 10^{-4}. IF IT IS ASSUMED THAT THE MARKER SOURCE IS A VARIABLE TUNED OSCILLATOR, ITS SCALE ACCURACY WOULD HAVE TO BE AT LEAST <u>4 TIMES THIS GOOD</u>, I.E., ITS RESETTABILITY AND SCALE ACCURACY WOULD HAVE TO BE AT LEAST 2 PARTS IN 10^{-5}, TO BE ABSOLUTELY SURE THE MARKER WAS IN THE PROPER POSITION. IT IS UNFORTUNATELY TRUE THAT MOST MARKER GENERATORS AVAILABLE TODAY SIMPLY DO NOT HAVE THIS ORDER OF ACCURACY. THE DIFFICULTY ARISES, NOT FROM THE ABILITY TO GENERATE FREQUENCIES TO THIS ACCURACY, BUT RATHER THE PROBLEM INVOLVED IN CALIBRATION. THEREFORE, IF THE ALIGNMENT OF A HEAD END SIGNAL PROCESSOR IS ATTEMPTED, ONE MUST HAVE SOME MEANS OF MEASURING THE MARKER FREQUENCY TO THIS ORDER OF ACCURACY. THE WR-99-A MARKER GENERATOR, WIDELY USED IN THE INDUSTRY, IS MARGINALLY ACCEPTABLE FOR THIS USE, IN THE HANDS OF A <u>SKILLED</u> OPERATOR, PROVIDED THAT THE MARKER GENERATOR IS ALLOWED TO WARM UP OVER A CONSIDERABLE PERIOD OF TIME (SEVERAL HOURS) PRIOR TO ALIGNMENT. EVEN THEN IT WILL BE NECESSARY TO RE-CHECK THE MARKER FREQUENCY BY THE COMPLETE CALIBRATION PROCEDURE SEVERAL TIMES DURING THE ALIGNMENT, TO BE ABSOLUTELY SURE THAT THIS FREQUENCY TOLERANCE HAS BEEN ACHIEVED.

---

[1] DIRECTLY RELATED FOR MINIMUM PHASE SHIFT NETWORKS; INDIRECTLY RELATED FOR NON-MINIMUM RELATED BIFILAR OR BRIDGED "T" NETWORKS.

A BETTER METHOD OF DETERMINING THE FREQUENCY WITH THE SAME MARKER GENERATOR WOULD BE TO UTILIZE A FREQUENCY COUNTER (AND SCALER IF NECESSARY) CONNECTED TO THE MARKER GENERATOR TO DETERMINE ITS FREQUENCY CONTINUOUSLY DURING USE. BY THIS TECHNIQUE, IT IS POSSIBLE TO MONITOR THE OUTPUT FREQUENCY TO THE ACCURACY REQUIRED FOR THIS MEASUREMENT.

A SECOND SERIES OF MEASUREMENTS AND FREQUENCY TOLERANCES INVOLVE ANOTHER CRITICAL ADJUSTMENT ON THE HEAD END PROCESSOR, THE ADJUSTMENT OF THE SOUND TRAP. SINCE THE SOUND TRAP IS IN CLOSE PROXIMITY TO CHROMINANCE SUB-CARRIER, AND SINCE ANY ATTENUATION OF THE CHROMINANCE SUB-CARRIER CREATES AN ATTENDANT PHASE SHIFT, IT IS EXTREMELY IMPORTANT THAT THE MARKER FREQUENCY FOR THIS APPLICATION BE VERY ACCURATELY IDENTIFIED. THE ORDER OF ACCURACY FOR THIS MEASUREMENT WOULD BE + - 5 KHZ OR APPROXIMATELY 5 TIMES TIGHTER TOLERANCE THAN THE PREVIOUS MEASUREMENT. IT IS MARGINALLY POSSIBLE THAT A VERY SKILLED OPERATOR COULD USE THE MARKER GENERATOR DESCRIBED PREVIOUSLY FOR THE APPLICATION, BUT IT IS DOUBTFUL THAT THIS MEASUREMENT COULD BE MADE WITH THE INTERNAL CALIBRATION MEASUREMENT WITH SUFFICIENT ACCURACY TO ASSURE THE DEGREE OF ALIGNMENT PRECISION DESIRABLE IN THIS HEAD END PROCESSING EQUIPMENT. THEREFORE, THIS SHOULD NOT BE ATTEMPTED UNLESS ONE OF TWO CALIBRATION SOURCES ARE AVAILABLE:

A) A DIRECT COMPARISON OF THE MARKER FREQUENCY TO THE ACTUAL AIR SIGNAL (SOUND CARRIER).

B) OR THE USE OF A COUNTER SCALER AS HAS BEEN PREVIOUSLY DESCRIBED.

A NATURAL QUESTION ARISES AS TO HOW ONE CAN MAINTAIN THE PRECISION OF ADJUSTMENT REQUIRED IN THE HEAD END PROCESSORS TO ASSURE THIS DEGREE OF COMPLIANCE, WHEN IN FACT, IN MOST CASES THE EQUIVALENT Q AND STABILITY OVER A LARGE TEMPERATURE RANGE SIMPLY DOES NOT PERMIT SUCH RETRACE CHARACTERISTICS. IT HAS BEEN ADEQUATELY STATED BY THE MANUFACTURERS OF HEAD END EQUIPMENT, THAT THE TEMPERATURE TOLERANCE OF HEAD END EQUIPMENT IS REALLY QUITE SMALL, AND THE CONCEPT OF ENVIRONMENTAL CONTROL FOR THE HEAD END EQUIPMENT IS NOT ONLY JUSTIFIED, BUT MANDATORY. NOT ONLY WILL IT PROVIDE FOR LONGER LIFE OF THE TECHNICAL EQUIPMENT, BUT IT WILL PERMIT A VAST DIFFERENCE BETWEEN THE ALIGNMENT CYCLES.(IF ONE REALLY CARES ABOUT THE PICTURE QUALITY AT THE HEAD END). THE MECHANISM OF THIS DETERIORATION IS QUITE SIMPLE. IF THE

HEAD END TEMPERATURES VARY OVER A PRESCRIBED TEMPERATURE LIMIT, THE EXPANSION AND CONTRACTION OF THE DEVICES USED IN THE VARIOUS RESONANT CIRCUITS EXPAND AND CONTRACT ALSO, WITH THE RETRACE CHARACTERISTICS OF THE MECHANICAL ADJUSTMENTS NOT BEING PERFECT. THIS IS THE MAJOR SOURCE OF LONG TERM SHIFTS IN ALIGNMENT (ASIDE FROM COMPONENT FAILURES OR DETERIORATION). OBVIOUSLY, IF THE TEMPERATURE CONTROL IS TIGHT (+ - 10°F) THERE WILL BE LESS NEED FOR ALIGNMENT, OR TO PUT IT ANOTHER WAY, THE ALIGNMENT CYCLES WILL BE FURTHER APART. THE NEXT EXAMPLE OF MEASUREMENT PRECISION REQUIRED IN CATV EQUIPMENT DEALS ALSO WITH HEAD END EQUIPMENT, HOWEVER, IN THIS CASE, CONSIDER THE SYSTEM WHERE THE DEMODULATION/REMODULATION IS EMPLOYED. SPLIT THIS AGAIN INTO THESE CATEGORIES:

1) THE SYSTEM WHERE THE 4.5 MHZ SOUND CARRIER IS RETAINED IN THE DEMODULATION PROCESS.

2) WHERE THE AUDIO INFORMATION IS RECOVERED AND REMODULATED. (THIS APPLIES TO LOCAL ORIGINATION ALSO).

IN THE FIRST CASE, THE TOLERANCE BETWEEN THE AURAL AND VISUAL CARRIER IS MAINTAINED AS IT WAS AT THE TELEVISION STATION, AND THERE IS NO PROBLEM MEASURING THE FREQUENCY SEPARATION AS THIS WILL NOT BE DISTURBED IN THE REMODULATION PROCESS. THE ONLY FREQUENCY TOLERANCE ATTENDANT WITH ALIGNMENT OF THIS EQUIPMENT IS THE PLACEMENT OF THE BAND EDGE OF THE BAND PASS, AND ALIGNMENT OF THE MODULATOR VESTIGAL SIDE BAND FILTER. THE FCC STANDARDS PRESCRIBE A SPECIFIC AMPLITUDE VERSUS FREQUENCY RESPONSE, AND TELEVISION RECEIVERS HAVE BEEN DESIGNED TO COMPLEMENT THESE CHARACTERISTICS, WHILE AT THE SAME TIME PROVIDING THE BEST POSSIBLE PHASE CHARACTERISTICS ASSOCIATED WITH THIS VESTIGAL ATTENUATION, THIS MEANS THAT THE CATV OPERATOR MUST DUPLICATE THIS "TRANSMITTER CURVE"[2] AS CLOSELY AS POSSIBLE TO ASSURE IDENTICAL PHASE CHARACTERISTICS[3]. THE FCC AMPLITUDE RESPONSE CURVE IS SHOWN ON FIGURE II. IT IS NECESSARY TO MAINTAIN THE SHOULDERS OF THE FREQUENCY RESPONSE TO WITHIN + - 25 KHZ, AND AGAIN, THIS REQUIRES THE SAME FREQUENCY TOLERANCE ASSOCIATED WITH THE PREVIOUSLY DESCRIBED MEASUREMENTS, AND CAN BE DONE WITH

[2] IN ADDITION, IT IS NECESSARY TO UTILIZE THE SAME PHASE PREDISTORTION CHARACTERISTICS THAT HAVE BEEN INSTALLED IN A TYPICAL TV STATION, BUT THIS IS ANOTHER MATTER FOR ANOTHER PAPER.

[3] THIS PHASE CURVE IS GIVEN IN FCC RULES 73.687 (5).

A CONVENTIONAL TYPE MARKER GENERATOR WITH SKILL AND EXPERIENCE.

ADDRESSING THE SECOND CASE OF HEAD END PROCESSING EQUIPMENT UTILIZATION, WHERE THE AUDIO HAS BEEN DEMODULATED TO BE SUBSEQUENTLY REMODULATED WITH A MODULATOR, OR WHERE AN AUDIO CARRIER IS ORIGINATED (AS WITH LOCAL ORIGINATION) THE SAME TYPE FREQUENCY MEASUREMENT IS INVOLVED, I.E., THE EXACT CARRIER FREQUENCY OF THE VISUAL CARRIER MUST BE ESTABLISHED ACCURATELY AND THE FREQUENCY SEPARATION OF THE AURAL CARRIER AND VISUAL CARRIER MUST BE DETERMINED. THE MEASUREMENT OF THE VISUAL CARRIER, CAN BE DONE WITHIN THE REQUIRED ACCURACY BY USING A MARKER GENERATOR, WHEN DONE BY A PERSON WHO HAS THE SKILL AND EXPERT KNOWLEDGE TO USE THIS DEVICE PROPERLY WITHIN ITS LIMITATIONS (AGAIN WITH A COUNTER). THE SECOND MEASUREMENT, THE INTERCARRIER DIFFERENCE FREQUENCY BETWEEN THE VISUAL CARRIER AND AURAL CARRIER MUST BE MEASURED TO A TIGHTER TOLERANCE. TYPICALLY, TELEVISION RECEIVERS WILL PERFORM SATISFACTORILY WITH A + - 5 KHZ DEVIATION OF THE AURAL CARRIER WITH VERY LITTLE SOUND DETERIORATION. AS HAS BEEN PREVIOUSLY SHOWN, THIS IS AT THE BARE LIMITS OF THE MARKER GENERATOR TECHNIQUE WHEN USED WITH A COUNTER.

ANOTHER METHOD TO ESTABLISH THE AURAL CARRIER FREQUENCY WOULD ESTABLISH THE VISUAL AND AURAL CARRIERS AND THEN MEASURE THE DIFFERENCE FREQUENCY BETWEEN THE TWO CARRIERS. THIS METHOD IS NOW WITHIN THE MEASUREMENT TOLERANCE OF A MILITARY SURPLUS METER, KNOWN AS THE "BC 221 SERIES." WITH THIS HETERODYNE TYPE DEVICE, IT IS POSSIBLE TO MEASURE THE FREQUENCY DIFFERENCE WITH SUFFICIENT ACCURACY TO BE ABLE TO SET FOR THE CORRECT INTERCARRIER FREQUENCY.

THERE ARE OTHER TECHNIQUES FOR MEASURING THE FREQUENCIES INVOLVED IN CATV TO THIS ORDER OF ACCURACY, HOWEVER, MOST OF THEM WILL BE MORE EXPENSIVE THAN THE TECHNIQUES DISCUSSED.

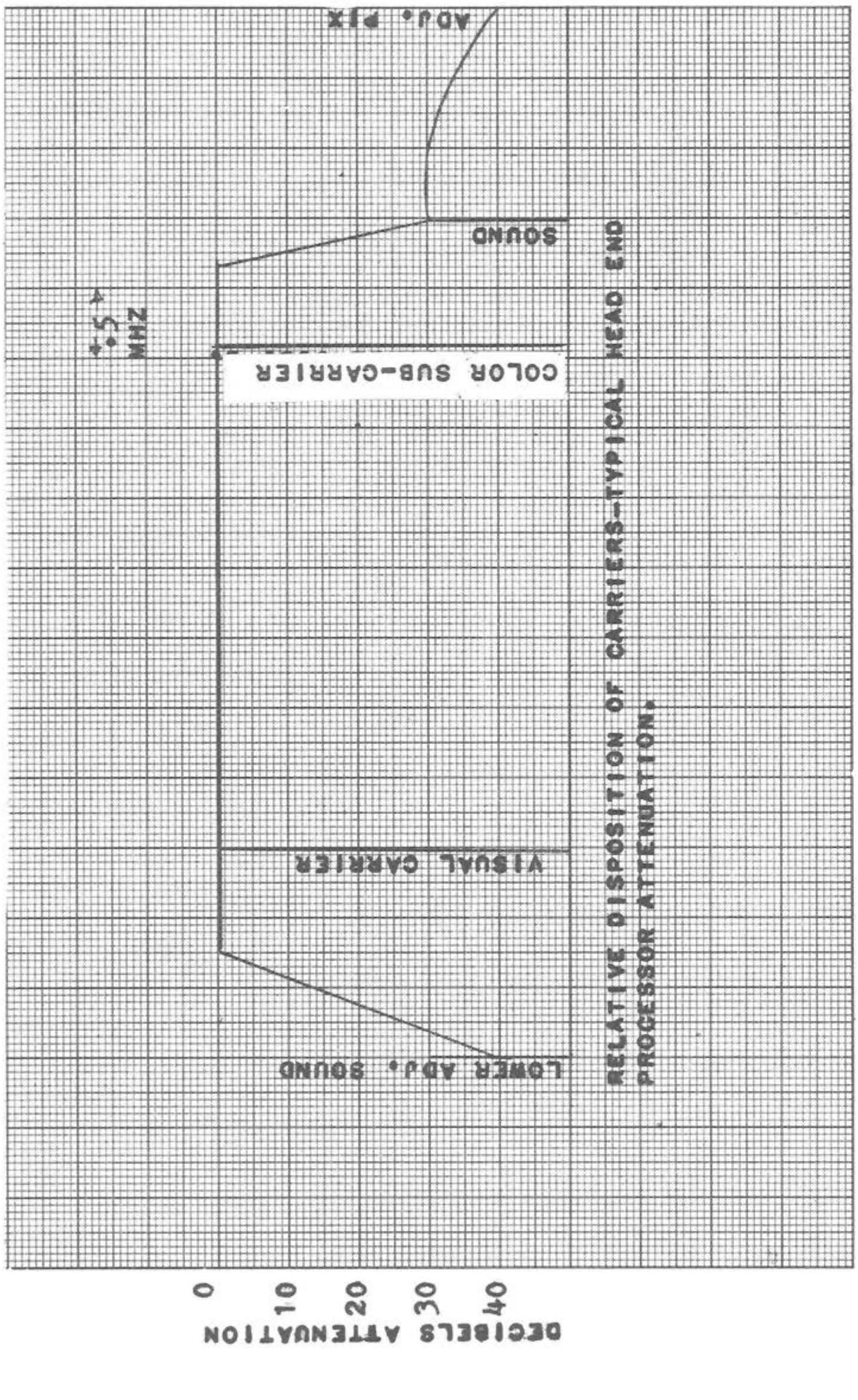
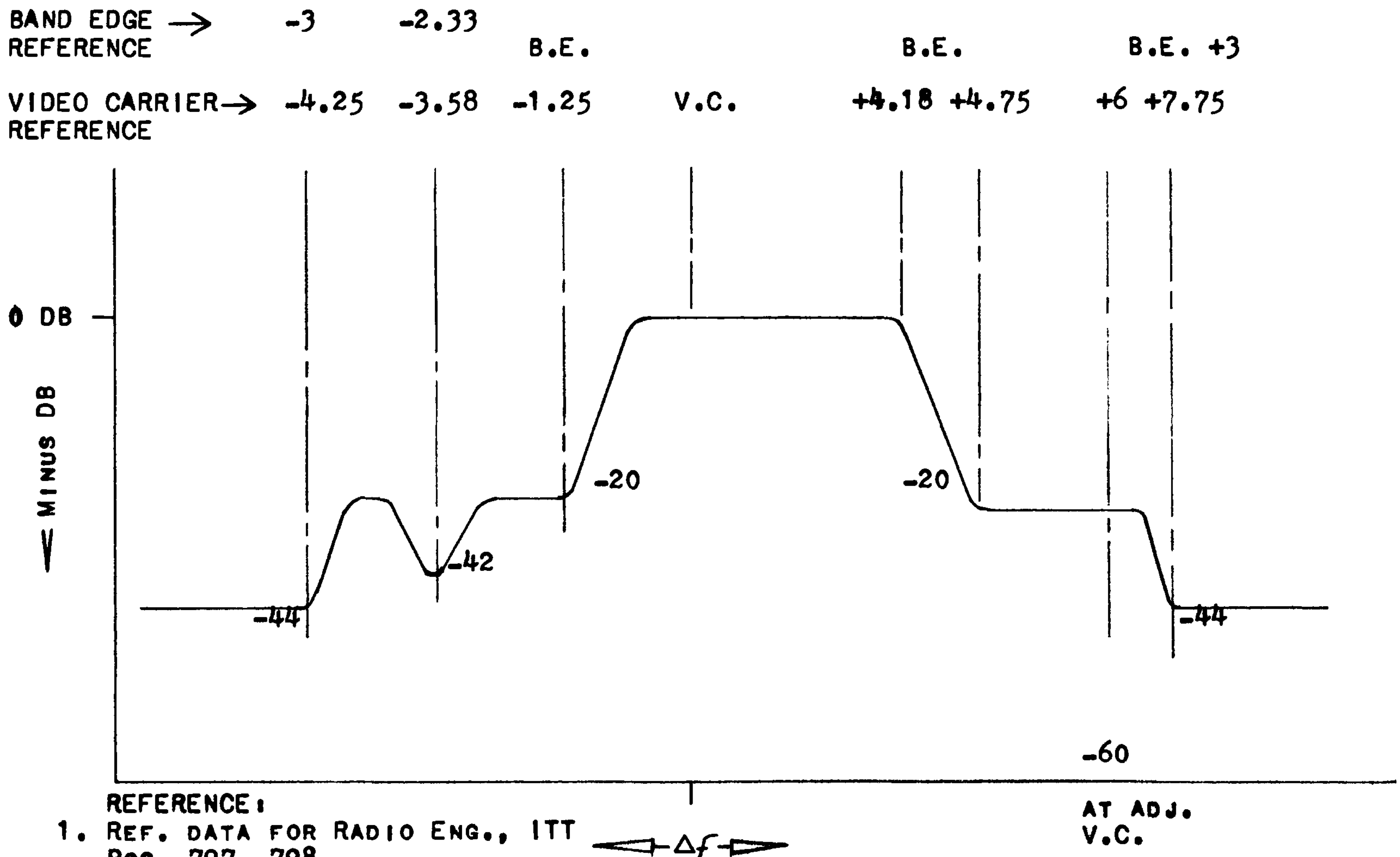

ADJ. PIX
SOUND
4.5 MHZ
COLOR SUB-CARRIER
VISUAL CARRIER
LOWER ADJ. SOUND
RELATIVE DISPOSITION OF CARRIERS-TYPICAL HEAD END PROCESSOR ATTENUATION.
0
10
20
30
40
DECIBELS ATTENUATION

FIG. 1



REFERENCE:
1. REF. DATA FOR RADIO ENG., ITT
PGS. 797, 798

FCC SPEC________

FIG. 2

A SECOND GENERATION AML

by

N. A. Begovich and L. S. Stokes
Hughes Aircraft Company
Culver City, California

INTRODUCTION

Community antenna television began as a service to isolated communities without television programming. In the ensuing years, the scope of cable TV has increased to the point where now large metropolitan areas find it expedient to use CATV to provide high equality television. Figure 1 illustrates how large metropolitan areas may be serviced from a single transmitter site using microwave links to carry television signals to individual microwave receivers located strategically within the metropolitan complex. From these receivers, cable systems deliver the television programming to individual subscribers.

In the suburban areas, located in television fringe areas or unable to receive signals from nearby metropolitan areas, microwave links can connect headend sites to the specific communities without the expense and picture quality loss of many miles of cable. Such an application is illustrated in Figure 2.

The advantage of microwave frequency links to distribute television programming to the user has been recognized for some time. The capabilities that such links must provide are tabulated in Table I which indicates that up to 20 channels per link are required to match the FCC rulemaking for local distribution systems. To meet the highest quality television standards, intermodulation distortion produced by such links must be 57 db below the video carrier. A nominal range of 10 miles for such links will meet the majority of the applications considered. Furthermore, the output VHF radio frequencies which carry the actual TV transmission, must be identical to the input frequencies if local stations are to be retransmitted on the microwave link. (For instance,

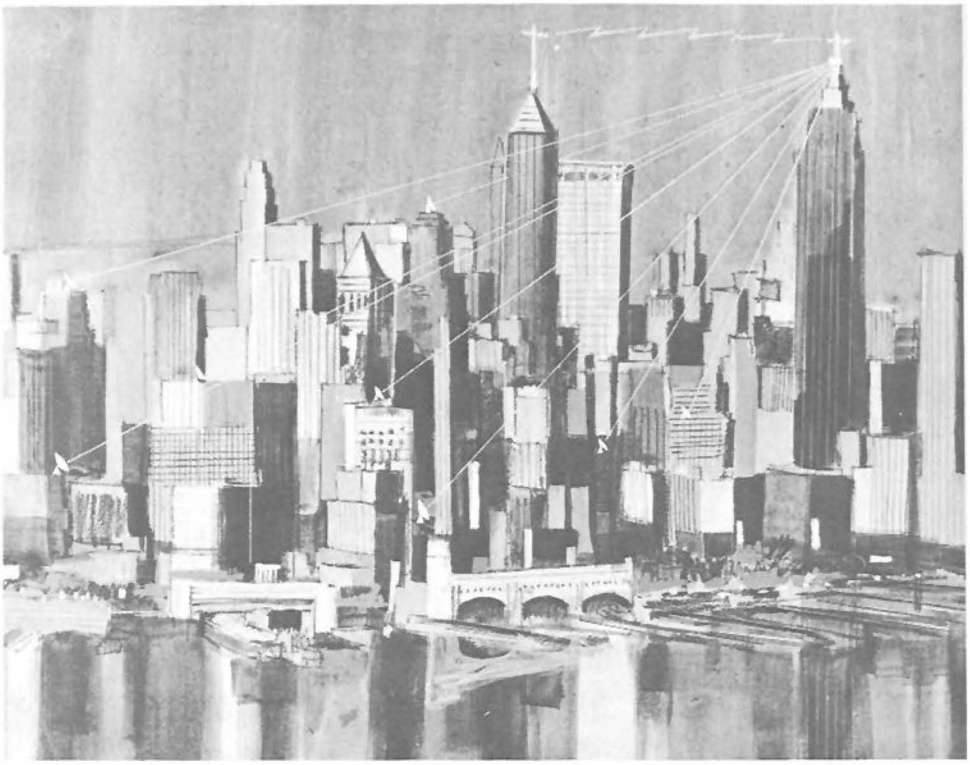


Figure 1. Artist Concept of Urban Use of AML System

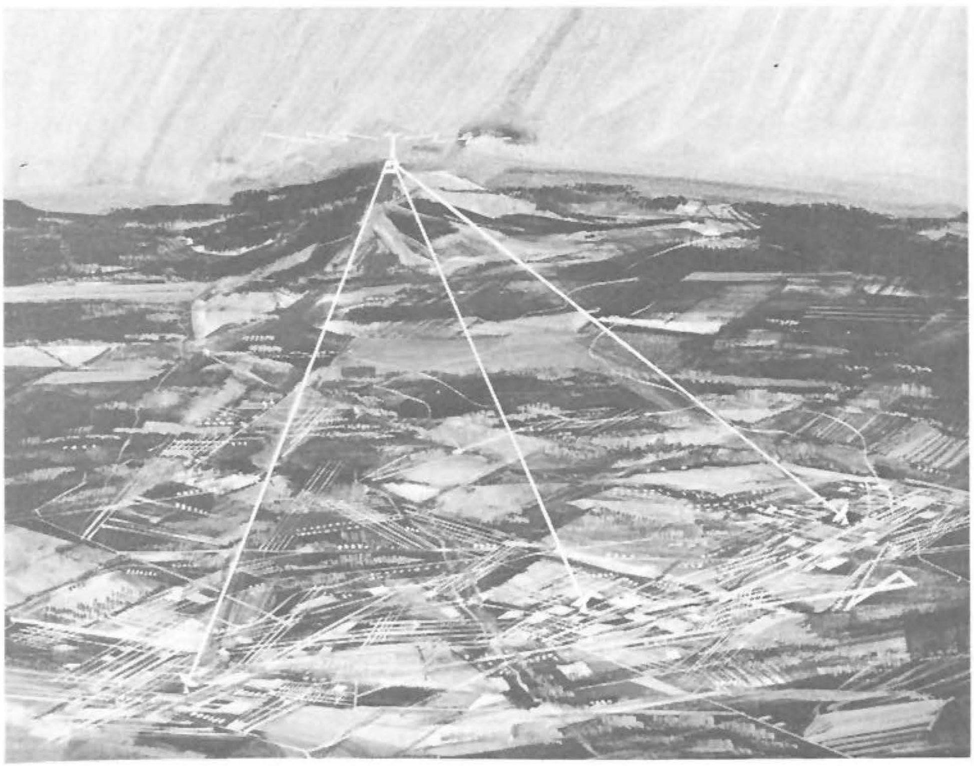


Figure 2. Artist Concept of Suburban Use of AML System

• MULTIPLE CHANNELS (UP TO 20 CHANNELS)
• LOW INTERMODULATION PRODUCTS
• RANGE OF AT LEAST TEN MILES
• OUTPUT FREQUENCIES IDENTICAL TO INPUT FREQUENCIES
• MULTIPLE RECEIVERS WITH SINGLE TRANSMITTER

Table I. Microwave Link System Goals

if Channel 2 from the microwave receiver is carried on the Channel 2 carrier, this carrier must have exactly the same frequency as a local Channel 2 transmitter. If this is not true it is possible for the Channel 2 carrier transmitted from the local VHF transmitter to interfere with the output of the microwave link and cause undesirable effects in the television picture.) It is also desirable that many receivers be used with a single transmitter since most local distribution systems have this requirement.

ECONOMIC ANALYSIS OF MICROWAVE SYSTEMS

CATV is clearly established as an economically desirable venture. The question remains whether a CATV system operating with a microwave link will maintain and improve this economic advantage. For this reason a number of cities have been analyzed to determine the overall profit and the yearly return-on-investment that would be provided by a CATV system using a microwave link. In this economic analysis (see Table II) the number of receiving sites required by each city was determined. Furthermore, the size of community and the number of homes in the community were tabulated and estimates of penetration into this community were made. Monthly rates were defined. The yearly costs to the system include the equipment amortization, the cable costs, and the effect of signal quality on income. In addition to these financial considerations, the technical considerations such as required transmitter power and the effects of rain were included in this analysis.

• NUMBER OF RECEIVING SITES
• COMMUNITY SIZE
• COMMUNITY PENETRATION
• MONTHLY RENTAL
• EQUIPMENT AMORTIZATION
• CABLE COSTS
• SIGNAL QUALITY
• RANGE
• RAIN OUTAGE

Table II. Considerations Used in Economic Analysis for Each City

The results of this economic analysis are shown in the next three figures. The first figure, Figure 3, plots yearly return-on-investment. A large number of cities had a return-on-investment of greater than 20 percent per year. A considerable number were between 30 and 40 percent a year and one had a return-on-investment as high as 92 percent per year. The actual revenue per year is indicated in Figure 4. A cumulative plot of these profit values are given in Figure 5. The indication is that the total profit from the 34 cities investigated exceeds 3 million dollars per year. Hence community antenna television systems that incorporate the convenience of microwave links are seen to maintain attractive profits.

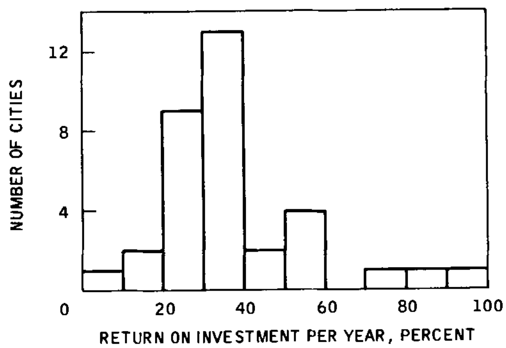


Figure 3. Distribution of Return-on-Investment

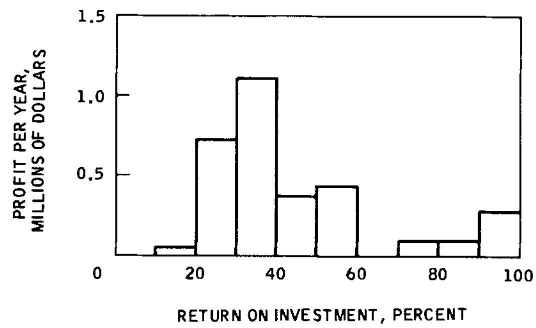


Figure 4. Distribution of Total Profits

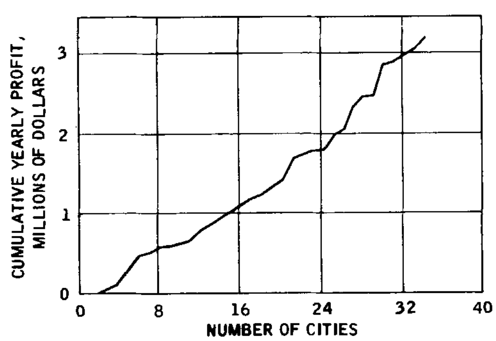
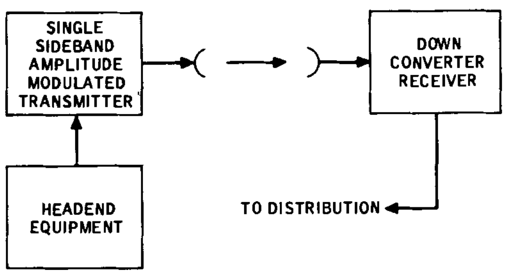


Figure 5. Cumulative Yearly Profits for 34 Cities

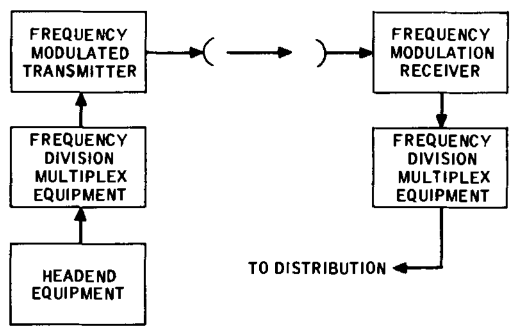
MICROWAVE SYSTEM SELECTION

Candidate Systems

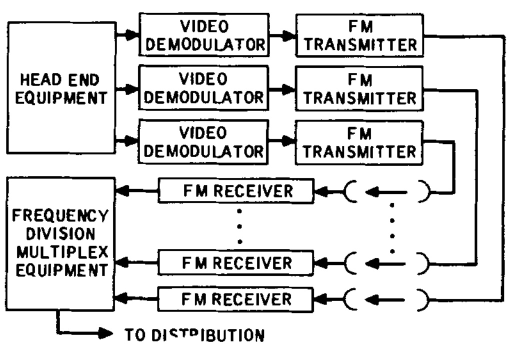
The previous analysis demonstrated that CATV systems operating with microwave links maintain a high profit position. It remains therefore to determine the best type of transmission method to use in these microwave links. Figure 6 illustrates three candidate systems. The first system is a single-carrier, amplitude-modulated link. This system accepts the output of the headend equipment directly, which in turn drives a single-sideband, amplitude-modulated transmitter. In this case the baseband VHF signals are effectively shifted to a microwave frequency for transmission. The microwave receiver detects the incoming signals and shifts them to their original spectral positions for local distribution.



a. Single Carrier Amplitude Modulation



b. Single Carrier Frequency Modulation



c. Multiple Carrier Frequency Modulation

Figure 6. Candidate Local Distribution System

The second system, Figure 6b, illustrates a single-carrier, frequency-modulated system that accomplishes the same task with increased complexity. Again the headend equipment is required to normalize the television channel signals. However, a frequency division multiplex process is required to shift these frequencies from the normal television channels to a baseband structure which begins at approximately 6 MHz. This baseband structure then frequency modulates the transmitter, is transmitted to the microwave receiver, and is demodulated to recover the input baseband structure. Again, the output of the microwave receiver must be shifted back to the VHF frequencies that are acceptable to the local user by means of a frequency division multiplexer. Thus frequency division multiplexing at both the transmitter and receiver is required. This is an additional equipment requirement to that of the single-carrier, amplitude-modulation system.

The third system, indicated in Figure 6c, is the multiple-carrier, frequency-modulation system. This system, also quite complex, requires headend equipment and demodulation equipment to detect the individual video signals from the television channels. The individual demodulated video signals then frequency modulate individual transmitters. The transmitted signals are individually received by microwave receivers to recover the video signals. The video signals must finally be remodulated upon a set of carriers corresponding to the normal channels used by a home TV receiver before the signals may go to the local distribution system.

Clearly a microwave link that is to be used in a community antenna television system must provide the required performance in a cost effective manner. From the discussion above, the single carrier amplitude modulation system requires considerably less hardware than either of the other systems. It remains then to discuss the ability of an amplitude modulated link to provide the necessary TV picture quality and number of TV channels required by industry standards.

Amplitude versus Frequency Modulation

The configurations shown in Figure 6 illustrate two basic modulation methods, amplitude and frequency modulation. A comparison of these two types of modulation (see Table III) is needed to determine their suitability 1) to maintain the desired quality for a community antenna television system and 2) to provide the desired number of TV channels. It should be noted that the quality specifications for such a system are by no means trivial. It is required that the intermodulation products be kept to extremely low levels, that the signal-to-noise ratios be maintained at high levels and that cross modulation also be kept to extremely low levels while maintaining a minimum RF occupancy.

Amplitude and frequency modulation have different characteristics relative to these several considerations. Single sideband amplitude modulation, for instance, provides a minimum RF bandwidth occupancy for a given baseband. This is extremely important in the newly reassigned CARS band where a limited bandwidth, 250 MHz, is available for all transmissions. Amplitude modulation is also more efficient than frequency modulation for a given transmitted power, when the frequency modulation index is less than 1. 4. Furthermore, amplitude modulation is not as susceptible as frequency modulation to transmission phase distortions. Disadvantages of amplitude modulation include the requirement for very linear circuits to maintain low cross modulation and third order intermodulation distortion. Amplitude modulation is also more susceptible to amplitude noise than is frequency modulation.

Table III. Comparison of Amplitude and Frequency Modulation

ADVANTAGES	DISADVANTAGES
• LESS SUSCEPTIBLE TO AMPLITUDE NOISE THAN AMPLITUDE MODULATION	• REQUIRE 4 TIMES BANDWIDTH OF SSB-AM SYSTEM TO ACHIEVE THE SAME PERFORMANCE
• CAN TRADE SIGNAL-TO-NOISE RATIO FOR BANDWIDTH	• SUSCEPTIBLE TO PHASE AND DELAY DISTORTION

a. Frequency Modulation

ADVANTAGES	DISADVANTAGES
• MINIMUM BANDWIDTH OCCUPANCY	• REQUIRES LINEAR CIRCUITS
• MORE EFFICIENT THAN FM WHEN FM MODULATION INDEX IS LESS THAN 1.4	• SUSCEPTIBLE TO AMPLITUDE NOISE
• NOT SUSCEPTIBLE TO PHASE DISTORTIONS	

b. Single Sideband Amplitude Modulation

Frequency modulation has an advantage in that it can trade signal-to-noise ratio for bandwidth. This advantage is used widely in most frequency modulation systems. However, it is difficult to apply such an advantage to the CARS band in that there is a limited bandwidth available for all transmissions. With a requirement to transmit 20 television channels, wide frequency deviations simply are not possible. In fact in the CARS band, extending from 12.7 to 12.95 GHz, the absolute maximum number of TV channels that may be transmitted with frequency modulation is 20. This requires a frequency division multiplex baseband structure of video signals starting at 0 MHz. If the baseband structure starts at 6 MHz, a suggested value, only 19 channels of television may be transmitted.

Frequency modulation has the disadvantage of requiring a wide RF bandwidth in order to have the same power efficiency as single sideband amplitude modulation. For instance, all other things being equal, the RF bandwidth for an FM system must be at least 4 times the bandwidth of a single sideband AM system for the same output signal-to-noise ratio.

Frequency modulation is also highly susceptible to phase delay distortions which produce cross-modulation and intermodulation products. The phase and delay distortion characteristics must be carefully maintained throughout the entire transmission system. This includes amplifiers, mixers and intermediate frequency amplifiers, that must maintain a very careful phase control in order to prevent undesirable distortion products. In fact, to maintain the phase distortion requirements, phase delay compensation networks must usually be added to the overall system in order to provide the necessary phase control. With careful adjustment, phase distortion can be minimized, however absolute control cannot be maintained over the phase distortion because of varying atmospheric effects such as fading, multipath and rain.

The relative costs of a single-carrier, amplitude-modulation system and a multiple carrier frequency modulation system are compared in Figure 7. A significant factor in the cost of these two systems is the requirement for multiple receivers at each receiving site for the

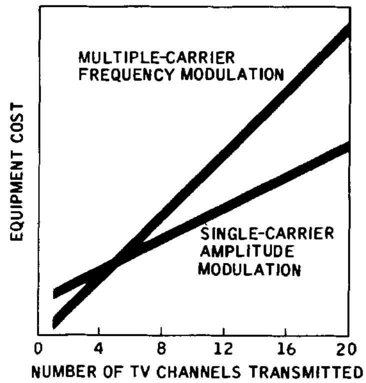


Figure 7. Local Distribution System Cost Comparison

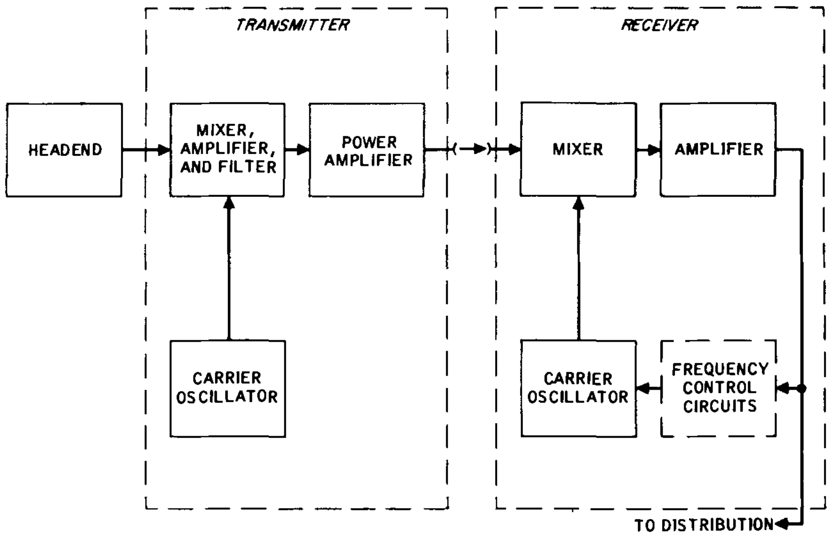
multiple-carrier, frequency-modulation system. The single-carrier, amplitude-modulation system requires only a single receiver, independent of the number of television signals being transmitted. Thus, the cost for the multiple-carrier, frequency-modulation system rises at a much faster rate, as a number of TV channels transmitted increases, as compared to the rate of increase with a single-carrier, amplitude-modulation system. For this reason, the multiple-carrier, frequency-modulation system will not be considered further.

Because of the relative simplicity of the single-sideband amplitude-modulated system over the single-carrier FM system and because of the ability of a SSB-AM system to provide 20* channels of television of the required quality, it has been chosen as the best microwave link configuration.

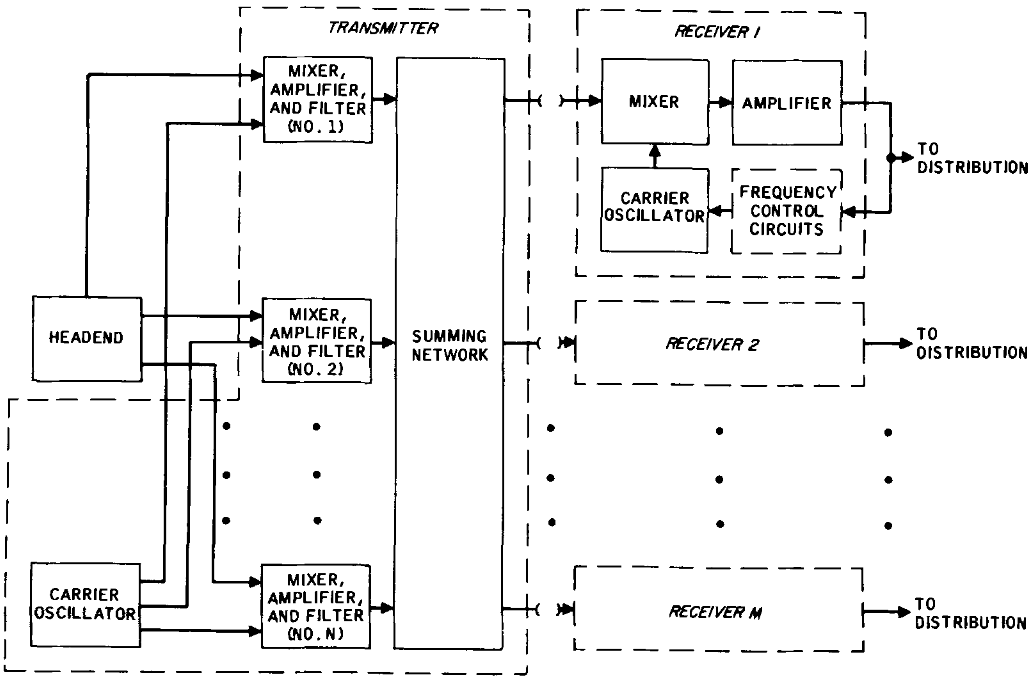
Amplitude Modulation Configurations

Two possible configurations of an Amplitude Modulation Link are given in Figure 8. Figure 8a depicts a single transmitter using a power amplifier. In this configuration, the normalized output of the headend receivers is applied directly to a mixer-amplifier-filter combination which, with the carrier local oscillator, provides the single-sideband amplitude modulation. The SSB-AM output is the same spectral configuration as the baseband input but is shifted to the desired microwave frequency. This shifted spectrum is then amplified in a linear power amplifier before transmission. The receivers consist of: 1) a heterodyne down converter to shift the microwave input spectrum to its

*The theoretical maximum is 41 channels in the present CARS band.



a. Single Transmitter Using a Power Amplifier



b. Multiple Channel Solid State Transmitter

Figure 8. Amplitude Modulation Link, Local Distribution System Configurations

original VHF spectrum, 2) an amplifier to amplify the VHF signals suitably for the local distribution system, and 3) an optional pilot tone transmitted to allow the receiver to sense the exact carrier frequency of the transmitter suppressed carrier. This allows the receiver to return the original spectra, which entered the transmitter, to exactly the same frequencies at the output of the receiver. The maintenance of input and output frequency relationships eliminates the undesirable beat notes between a locally transmitted television channel and the signal being received via the microwave/cable system.

A second SSB-AM configuration is shown in Figure 8b. Here each television signal, after it has been normalized to the proper amplitude by the suitable headend equipment, is applied to a single mixer-amplifier-filter combination. Each individual television channel is then shifted to a microwave frequency. These shifted frequencies are summed in a microwave network and are transmitted. As indicated in the figure, the number of transmitting antennas is more than one, usually 4 or 8. The receiver in Figure 8b is identical to the receiver configuration of Figure 8a and, again, merely shifts the microwave signal back to its original RF baseband spectra allowing it to be received directly by a home receiver.

The equipment indicated in the block diagram form of Figure 8a has been constructed and operated for 4 years. The actual hardware is shown in Figures 9, 10, and 11. Figure 9 is the low level mixer-amplifier-filter which performs the single-sideband function. This equipment is mounted in the rack as indicated in Figure 10 with the required power supplies and metering circuitry. The matching receiver used by this RF transmitter is shown in Figure 11, mounted on the back of the receiving antenna. The receiver is encased in a weatherproof box allowing it to be mounted externally as shown, requiring only a power supply and a coaxial cable connection.

Operation of the AML System

The AML system has been constructed and has been operating since the spring of 1966. This operation has been conducted under an experimental license and under an experimental research license as noted in Table IV. The Federal Communications Commission has granted these licenses but has not made a permanent rulemaking for the 17. 7 to 19. 7 GHz band at this time. However the FCC has authorized the operation of the AML system in New York City, in Farmington,

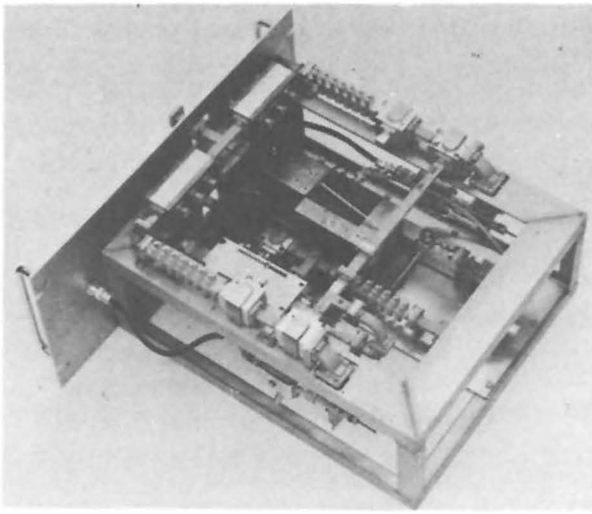


Figure 9. 18 GHz Experimental Transmitter

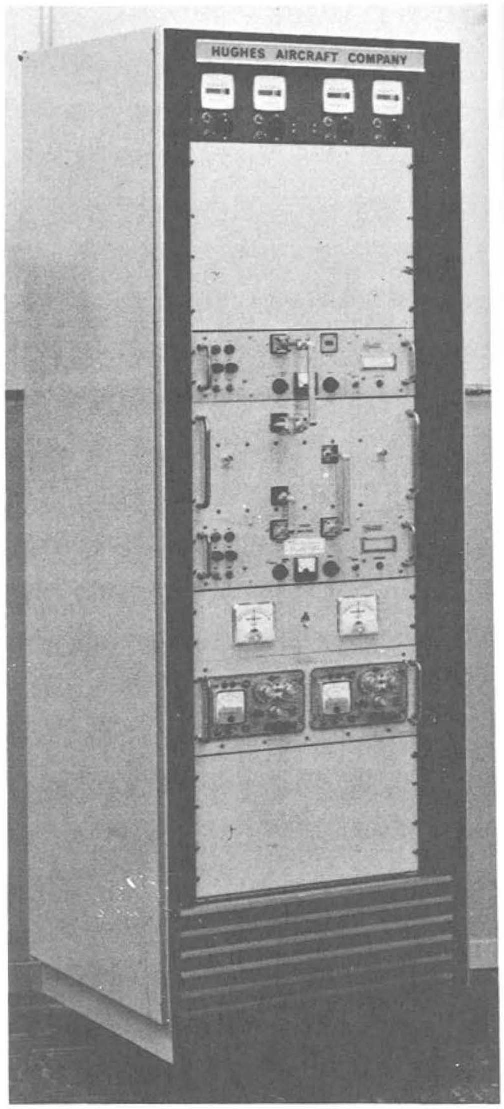


Figure 10. 18 GHz Rack-Mounted Experimental Transmitter

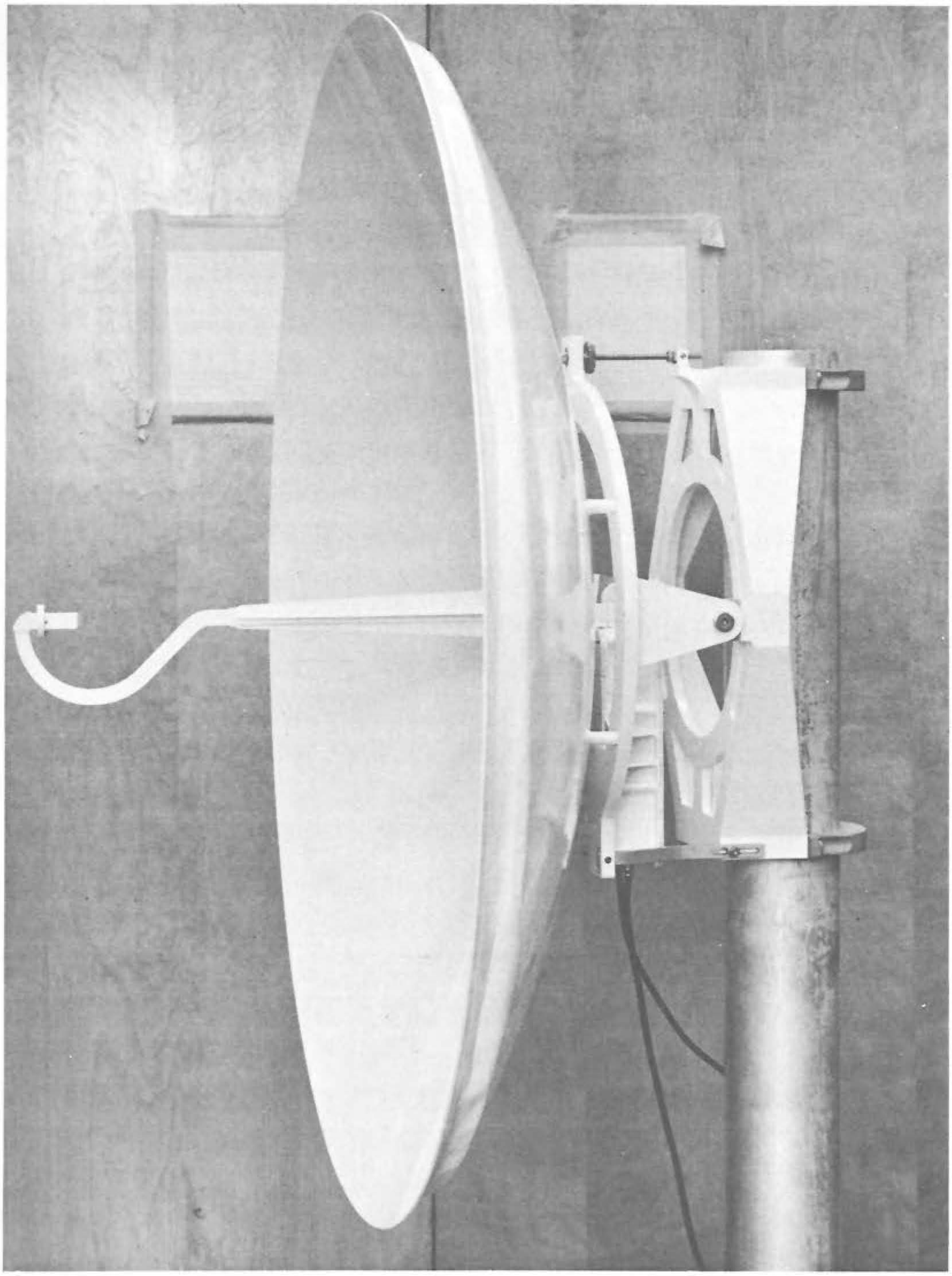


Figure 11. Prototype 18 GHz AML Receiving Antenna and Receiver

• GRANTED EXPERIMENTAL RESEARCH LICENSE
• NO PERMANENT RULEMAKING
• GRANTED TEMPORARY COMMERCIAL WAIVER
• NEW YORK CITY, NEW YORK
• FARMINGTON, NEW MEXICO
• EUGENE, OREGON

Table IV. FCC Action on 17.7–19.7 GHz Band for Local Distribution System

New Mexico, and in Eugene, Oregon. The AML has been operating in New York City since May of 1966 and has operated for a lesser time in Farmington, New Mexico.

The total operational time under the experimental license in New York City has been 33 months; the New York City operations are summarized in Table V. During this time the AML system has transmitted 12 channels of television. Operational measurements have been made on the performance of the system during this extended period of testing and considerable data relative to weather and operational procedures have been obtained. In February, 1969 a temporary commercial waiver was granted for the operation of the AML system. Since that time the AML has provided television channels to two CATV companies in the New York City area. This 16 month test used 12 channels of video transmitted over three different links. The data have confirmed the initial projections of the technical and economic feasibility of this system.

LOCAL DISTRIBUTION SYSTEM AT 12.7 TO 12.95 GHz

FCC Action

The Federal Communications Commission authorized the operation of the Amplitude Modulation Link in the 12.7 to 12.95 GHz CARS band in November, 1969 (see Table VI). This action was a permanent

EXPERIMENTAL OPERATION IN NYC • 33 MONTHS • 12 CHANNELS • OPERATIONAL MEASUREMENTS
TEMPORARY COMMERCIAL WAIVER IN NYC • JOINT USERS • 16 MONTHS (JUNE 1970) • 12 CHANNELS, 3 LINKS • CONFIRMED PROJECTIONS • TECHNICAL • ECONOMIC

Table V. 18 GHz Local Distribution Operation

• PERMANENT RULEMAKING (DOCKET 18452) GRANTED FOR AMPLITUDE MODULATED LINK, ON 7 NOV 1969
• 20 CHANNEL OPERATION
• OPERATES WITH EXISTING CARS SYSTEMS

Table VI. FCC Action on 12.7 – 12.95 GHz Band for Local Distribution System

rulemaking that allowed up to 20 TV channels for Amplitude Modulated Links. The FCC action required the compatible operation of both the AML and the existing CARS single-channel, frequency-modulated systems. Only these two systems are presently authorized to operate in this band.

With this new FCC action, the effects on the Amplitude Modulated Link were examined and the role for such systems was reconsidered. Table VII tabulates these effects. Since the frequency had been lowered from the 18 GHz band, considerable hardware simplifications resulted. In addition, a larger number of transmitter configurations became possible, in particular the use of solid state transmitters. The lower frequency also provided much better weather performance, which reduced the required link margin.

Local Distribution System Parameters

Local distribution system parameters, such as the design range, the number of receivers required per transmitter, and rain attenuation margin needed to be determined. An analysis was made of 34 different cities to determine these design parameters for a local distribution system. Figure 12 plots the number of individual links as a function of the ranges of these links. A nominal range value of 10 miles has been chosen based upon these statistics.

- FREQUENCY EFFECTS
  - HARDWARE SIMPLIFICATION
  - AVAILABILITY OF A LARGER NUMBER OF TRANSMITTER CONFIGURATIONS
  - BETTER WEATHER PERFORMANCE
- ROLE OF LOCAL DISTRIBUTION SYSTEM
  - RANGE REQUIREMENT
  - NUMBER OF RECEIVERS PER TRANSMITTER
  - APPLICATION TO INOIVIDUAL CITIES
  - BANDWIDTH CONSERVATION CRITICAL

Table VII. Effects of FCC Rulemaking at 12 GHz on AML

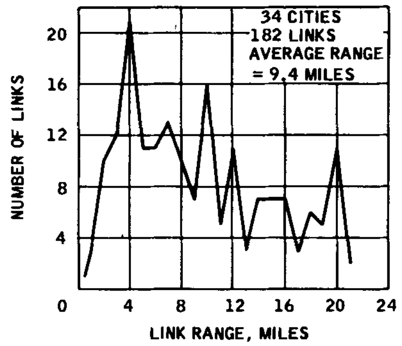


Figure 12. Local Distribution Link Range Statistics

The transmitter configuration of the AML system depends upon the number of receiver sites to be used with each transmitter. A statistical analysis of the number of receiving sites required is given in Figure 13. These data are taken for 34 different transmitting sites and include 182 receiving sites. A majority of the systems require 4 or more links from a given transmitter site. For this reason the transmitter has been designed to have either 4 or 8 outputs in order to accommodate the expected number of links at a given transmitter site.

The transmitter configuration shown in Figure 8b has been chosen for the local distribution system in the 12 GHz band. This system conveniently provides a transmitter having multiple outputs, 4 or 8, and can provide the necessary output power with solid state amplifiers.

Rain Margin at 12 GHz

The new frequency allocation reduced considerably the amount of link margin needed for rain attenuation. This is indicated in Figure 14 in which the rain attenuation is compared for 12 GHz and 18 GHz. At the nominal design range of 10 miles a 25 mm/hour rain produces 13 db of attenuation for 12 GHz and 26 db of attenuation for 18 GHz. This illustrates a marked advantage of 12 GHz over 18 GHz.

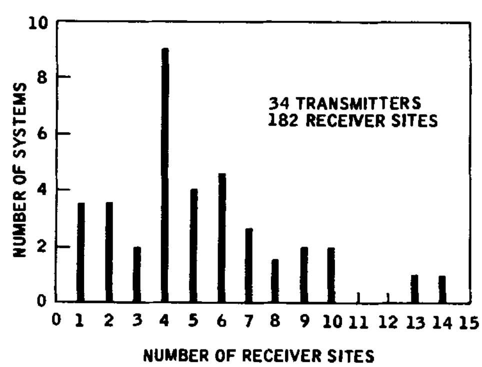


Figure 13. Number of Receivers per Transmitter Site

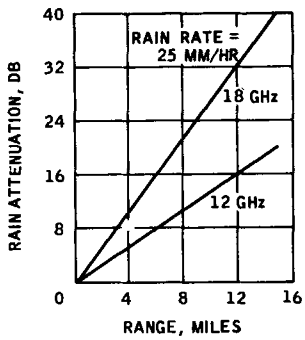


Figure 14. Comparative Rain Attenuation at 12 and 18 GHz

Rainfall data must also be correlated with the number of hours per year which exceed a given rainfall rate in order to determine the yearly performance of a link. Table VIII gives the rainfall statistics of the heaviest rainfall rate that was recorded in 5 different cities. This table indicates that the rainfall rate in Corvallis, Oregon exceeded 8.8 millimeters of rain per hour for only one hour per year while in Miami, Florida the rainfall rate exceeded 210 millimeters of rain per hour for at least one hour per year. Thus not only is the average rainfall in a given year important, but also the statistics for an hour-by-hour basis must be known to make proper attenuation estimates.

A summary of rainfall rates, given in Table VIII, is expanded in Figures 15, 16 and 17, the expected weather performance for Los Angeles, Washington, D. C. and Miami. The figures indicate the number of hours per year where the signal-to-noise ratio is expected to fall to less than a given peak video signal-to-RMS noise ratio as a function of link range. As may seen, the Los Angeles performance shows essentially no rain problem. The same is true for the expected Washington, D. C. performance. The Miami performance, with its heavy rainfall, does reduce the performance below the specified value of 45 db for about 10 hours per year if the link range exceeds 8 miles. However,

CITY	MM/HR (EXCEEDED 1 HOUR/YEAR)
CORVALLIS, OREGON	8.8
LOS ANGELES, CALIF.	10.9
ISLAND BEACH, N. J.	37.0
WASHINGTON, D. C.	50.8
MIAMI, FLORIDA	210.0

Table VIII. Rain Rate Distribution

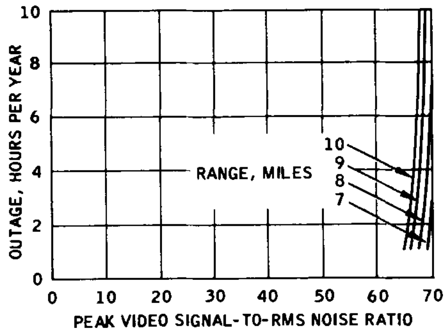


Figure 15. Expected Los Angeles Weather Performance

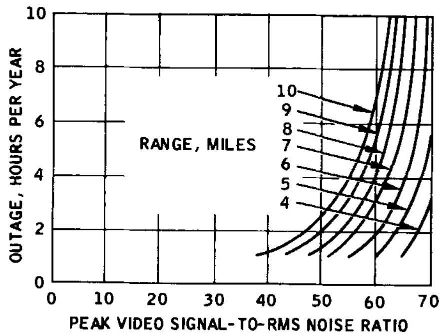


Figure 16. Expected Washington D. C. Weather Performance

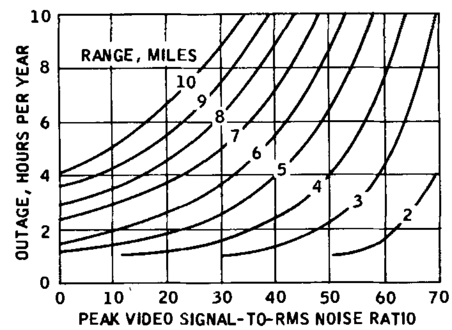


Figure 17. Expected Miami Weather Performance

the signal-to-noise ratio does not go below 35 db for ranges as large as 10 miles for more than 10 hours per year. Thus quality pictures are available even though the signal-to-noise ratio of the picture may be reduced by rain for brief periods.

To provide link margin to cope satisfactorily with rain attenuation it is possible to increase the size of both the transmitting and receiving antennas. The cost of this increase is indicated in Figure 18 which plots the relative cost of link antennas as link performance requirements increase.

12 GHz Amplitude Modulated Link

A typical link installation of a 12 GHz Amplitude Modulation Link is indicated in Figure 19. The transmitter installation, consisting of

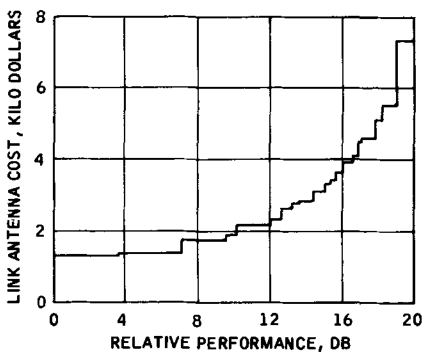


Figure 18. Antenna Cost Versus Performance

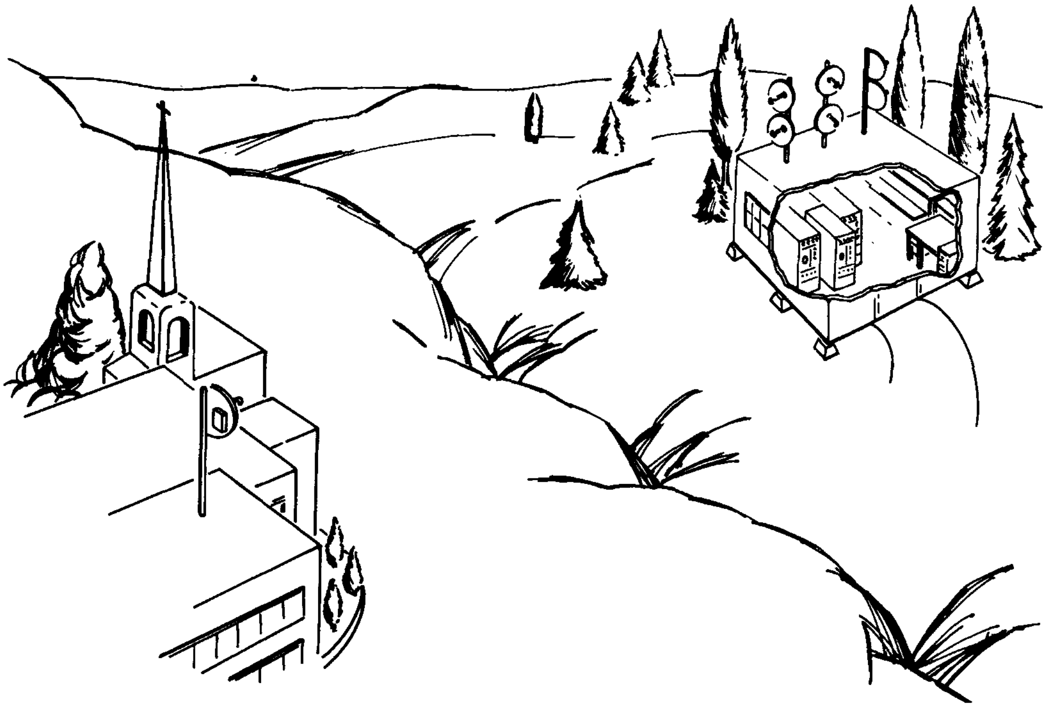


Figure 19. AML Receiver – Transmitter Installation

three modular transmitters, services receivers located in the customer communities. The receivers are then connected by cable to the distribution network.

The transmitting modules indicated in Figure 19 are each capable of transmitting 5 or 6 TV channels, depending if a pilot tone is used or not. A greater or lesser number of these transmitter modules may be used. The output of the individual transmitter modules are added before the power is divided among the several antennas that transmit to individual receivers. The present configuration of transmitting antennas uses a single transmitter to transmit the audio signals and a second antenna to transmit the video signals. All of these signals are received on a single receiving antenna.

AMPLITUDE MODULATED LINK SYSTEM PERFORMANCE

The AML system concept summarized in Table XI, has been proved both on a technical and economic basis through 4 years of field testing. The AML has provided 12 channels of simultaneous video and audio transmission while maintaining the highest technical standards. This performance has been accomplished with the use of the 18 GHz experimental band. Present FCC authorization in the 12 GHz band has provided new benefits for the AML system. . These include the ability to use a different transmitter configuration with a considerable saving in input power. Furthermore, better weather performance has considerably increased the reliability of the system and removed the need for high power amplifiers. The 12 GHz system is presently in production and is available for early delivery.

Table IX. AML

PROVEN SYSTEM CONCEPT • TECHNICAL • ECONOMIC • MULTIPLE CHANNELS
AUTHORIZED FOR 12 GHz BAND • NEW BENEFITS • SOLID STATE TRANSMITTER MODULES • BETTER WEATHER PERFORMANCE • MINIMUM PRIME POWER • IN PRODUCTION

BEST FREQUENCY ASSIGNMENTS FOR MID AND SUPER BAND CHANNELS

Michael F. Jeffers
Vice President - Engineering
Jerrold Electronics Corporation

Introduction:

During the past two decades CATV systems have carried, first, three channels of television; then five channels, then twelve channels.

For the past three years we have been advocating twenty to thirty television channels with thoughts of an almost limitless capacity for the future use of cable spectrum. As systems expanded from three to twelve channels the frequency allocation was obvious - we simply duplicated standard VHF television assignments. With the increase in the number of channels beyond twelve, serious consideration must be given to their precise frequency assignment. The purpose of this paper is to discuss some of the problems that might be encountered and to suggest ways of minimizing their effect.

The following presentation is applicable to any system carrying more than twelve channels when the standard twelve VHF channels are on the cable system at the normally prescribed broadcast frequencies, and

A. there are two levels of customer service

   1. customers receiving only the standard VHF channels - thus not requiring a converter.

   2. customers receiving the standard VHF channels plus additional channels via a converter mechanism.

B. all subscribers have a converter but the standard VHF channels are bypassed thru the device so that the television set tuner is used to tune these bypassed channels.

C. all subscribers have a converter which tunes to all programs; but the I.F. frequency of the converter is the same as the standard I.F. frequency of a television set.

D. a broad band converter is used to convert a block of Mid band or Super band programs to standard UHF or VHF channels; the twelve VHF channels on the cable being bypassed thru the converter and tuned via the TV set tuner.

E. cable systems expand to the point that TV set manufacturers build tuners which are compatible with broadcast <u>and cable systems</u>; converting all channels to the standard television set I.F. frequency.

The Problems:

A. Oscillator Leakage

An average television set has a relatively high level of oscillator leakage from its antenna terminals. As can be seen from Table I, the oscillator frequencies associated with the standard VHF channels fall into the Mid Band and Super Band spectrum of our cable systems.

VHF Channel	Oscillator Frequency(Megahertz)
2	101
3	107
4	113
5	123
6	129
7	221
8	227
9	233
10	239
11	245
12	251
13	257

TABLE I

Since none of these oscillator frequencies occur within the limits of the twelve VHF channels, this has not been a problem. However, it is obvious that oscillator leakage from TV sets tuned to channels 5 and 6 fall into the Mid Band and from sets tuned to channels 7 to 13 fall into the Super Band. Figure I depicts a typical method of subscriber connection. The following example shows how the problem occurs.

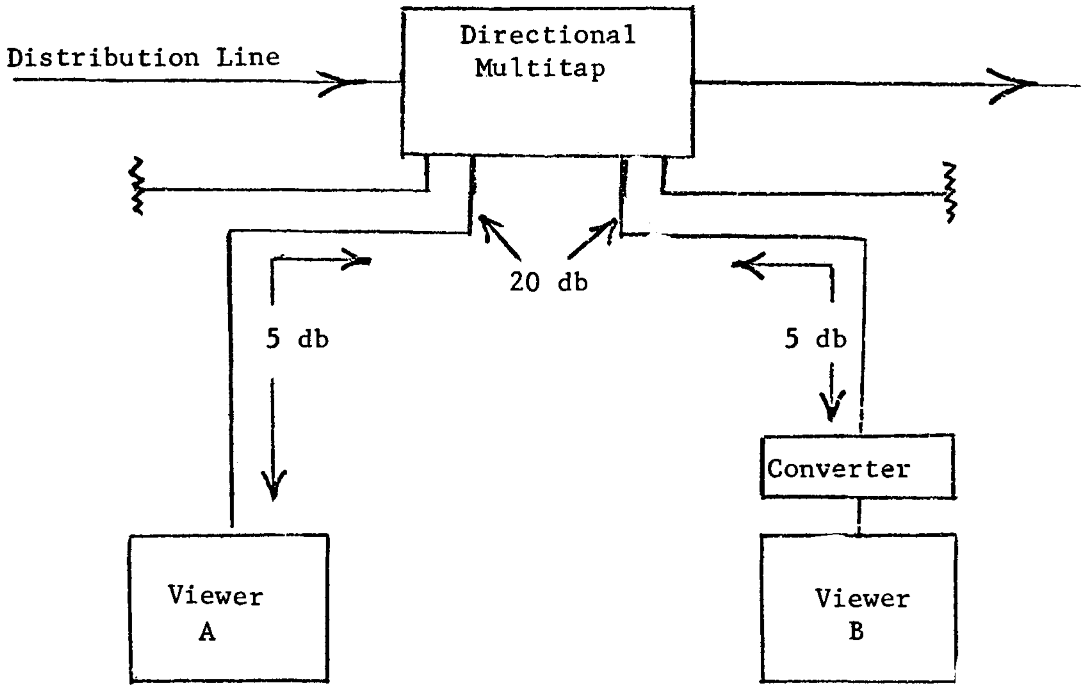


Figure I

Viewer A does not have a set converter to receive the "extra" channels. His TV set is tuned to channel 6 and the oscillator is at 129 megahertz. If Viewer B is tuned to the accepted Mid Band channel B (126 MHz to 132 MHz - video carrier at 127.25 MHz) the oscillator from Viewer A, if strong enough, will cause a beat 1.75 MHz in the channel received by Viewer B. Assume that a typical level at the input of the converter at channel B is +5dbmv and that interference to be unobjectionable must be 55db minimum below this, which is -50dbmv. Allowing 5db for each drop from the multitap and 20db isolation between customer outputs (for a total of 30db isolation); then the level of oscillator leakage from the antenna terminals of Viewer A must be -20dbmv or lower. Similarly, if Viewer A were tuned to Channel 10 and Viewer B were tuned to Channel M (234 to 240 MHz; video carrier 235.25) the oscillator from Channel 10 would appear as a 3.75 MHz beat in Channel M, which could cause a serious color beat. This second case is identical for all high VHF channel oscillators interfering as a color beat with all Super Band channels having the currently accepted frequency assignment.

Table II lists measurements of the level of oscillator leakage from the antenna terminals of several TV sets. A standard 75ohm to 300 ohm transformer was connected to the television set and readings were made with a Signal Level Meter. The tests were made while the set was receiving the desired signal at a level of 5dbmv; however, the oscillator leakage seemed to be completely independent of the input signal level and indeed, gave the same reading with the signal disconnected.

Channel	*A	B	C	D	E	F
4	-22.0	-12.0	-15.0	-12	-13.0	-24.0
5	-14.0	+ 2.0	-13.0	-11	- 8.0	-30.0
6	-11.0	+ 4.0	-13.0	- 9.5	- 7.5	-30.0
7	- 3.0	+19.0	- 4.0	+ 6	- 3.0	-18.0
8	+ 1.0	+21	- 4.0	+ 5.5	- 1.5	-14.0
9	+ 1.0	+21	- 2.0	+ 8.0	- 2.0	-10.0
10	0.0	+24.5	- 3.0	+ 9.0	- 4.0	-20.0
11	- 3.0	+20.0	- 4.0	+10.0	- 8.0	-20.0
12	- 7.0	+11.0	0.0	+ 8.0	- 3.0	-11.0
13	- 5.0	+10.0	- 1.0	+ 6.0	- 4.0	-12.0

*See Appendix for the description of the sets

TABLE II

B. Image:

The use of the spectrum commonly called Mid Band and Super Band generates another problem not encountered in twelve channel systems. The problem is "image." The image of any receiver is an interference that appears at a frequency equal to the desired signal plus or minus twice the I.F. frequency of the receiver. For the standard television receiver (or a converter mechanism whose intermediate frequency is the same as the standard TV set) the tuner oscillator is always above the incoming signal, therefore, the image will occur at a frequency equal to the incoming signal plus twice the I.F. frequency (45.75 MHz video I.F.) or at a frequency of 91.5 MHz above the incoming VHF video carriers. Table III shows the precise image frequency for all VHF video carriers.

Channel	Video Carrier (MHz)	Image Frequency (MHz)
2	55.25	146.75
3	61.25	152.75
4	67.25	158.75
5	77.25	168.75
6	83.25	174.75
7	175.25	266.75
8	181.25	272.75
9	187.25	278.75
10	193.25	284.75

Channel	Video Carrier (MHz)	Image Frequency (MHz)
11	199.25	290.75
12	205.25	296.75
13	211.25	302.75

TABLE III

An example of the image problem is as follows. Assume Viewer A is tuned to Channel 2. Table I shows that the oscillator in his TV set is at 101 MHz.

Desired Signal	TV set oscillator	-	Ch 2 Video	=	IF Video
	101.0 MHz	-	55.25 MHz	=	45.75 MHz
Undesired Image Signal	Undesired Signal	-	TV Set Osc	=	IF Video
	146.75 MHz	-	101.0 MHz	=	45.75 MHz

If a Mid Band signal were placed at 146.75 MHz, a zero beat would occur and the magnitude of the beat would be directly proportional to the attenuation of the tuner mechanism (commonly called "image rejection") to this image frequency. A 55 to 60 db minimum image rejection specification is required for all such tuners to avoid this problem. This is difficult to achieve in vacuum tube type tuners and will be far more difficult in the immediate future with the introduction of solid state tuners using varactor tuning methods. Keep in mind that there is no serious image problem for VHF broadcast TV. It is the cable industry that is using, or is about to use, the spectrum where image problems can occur, that brings this to the forefront.

Table IV shows where the accepted frequency assignments for Mid Band can cause problems.

Channel Effected	Type of Beat	Source of Interference	
2	1.5 MHz	E Video ;	145.25 MHz
3	1.5 MHz	F Video ;	151.25 MHz
3	3.0 MHz	E Sound ;	149.75 MHz
4	1.5 MHz	G Video ;	157.25 MHz
4	3.0 MHz	F Sound ;	155.75 MHz
5	-0.5 MHz	I Video ;	169.25 MHz
5	1.0 MHz	H Sound ;	167.75 MHz
*6	-0.5 MHz	7 Video ;	175.25 MHz
6	1.0 MHz	I Sound ;	173.75 MHz

TABLE IV

*Since this beat exists in cable systems today, and is not a serious problem, we can presume that the identical beat raised on Channel 5 by I video is also not objectionable. A negative beat is fundamentally less objectionable because it has the additional attenuation afforded it by its position on the slope of the IF response.

Since the image interference band for Channels 7 to 13 is 260 to 305 MHz, and although there is the intention of placing channels on the cable in this spectrum in the immediate future, these were not included in Table IV because no firm channel assignment has been in existance.

As a reference for the magnitude of the image problem in present day VHF tuners, the following information was obtained from tuner and set manufacturers:

a. Three circuit tuners: used in low cost TV sets

Image Rejection	Low Band	50db
	High Band	35db

b. Four circuit tuners: used in the better black and white sets and in color sets

Image Rejection		
	All VHF Channels	60db

Suggested Assignment to Minimize The Effect of Oscillator Leakage and Image.

Fortunately there appears to be an assignment to minimize the effects of these problems. Figure II is a graph showing the strength of an interference causing barely preceptible beat versus its location in the side band relative to the video carrier.

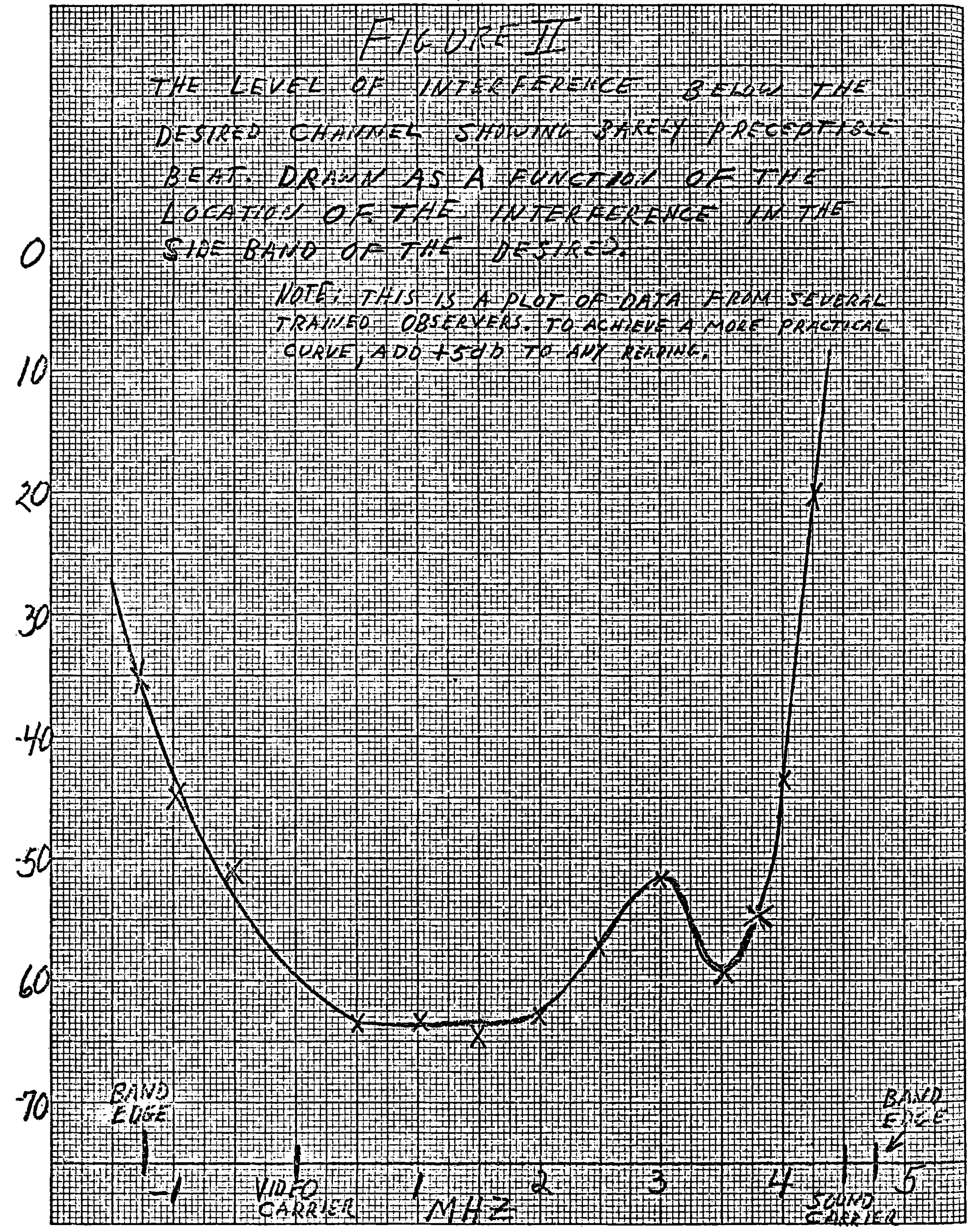


FIGURE II

THE LEVEL OF INTERFERENCE BELOW THE DESIRED CHANNEL SHOWING BARELY PRECEPTIBLE BEAT. DRAWN AS A FUNCTION OF THE LOCATION OF THE INTERFERENCE IN THE SIDE BAND OF THE DESIRED.

NOTE: THIS IS A PLOT OF DATA FROM SEVERAL TRAINED OBSERVERS. TO ACHIEVE A MORE PRACTICAL CURVE, ADD +5db TO ANY READING.

Figure II shows that an interference occuring in a desired channel at a frequency of 4.0 MHz above video carrier to the upper band edge or from about 0.5 MHz below the carrier to lower band edge can be much stronger than other interference and still be barely preceptible. This of course is in large measure due to the response shape of the I.F. amplifier of a TV set. In a black and white picture a beat occuring from 2,5 MHz to 4.0 MHz would get progressively less objectionable due to the subjective tolerance of the eye; however, in color, a beat in the region of 2.5 MHz to 4.0 MHz above video carrier begins to cause a very noticeable interference with the color carrier and a pronounced change in the appearance of the interference.

It is with reference to Figure II that I suggest the following Mid Band and Super Band assignment for consideration as a means to minimize the problems that might occur with standard television sets used on cable systems - a limitation over which we have no control at this time.

Suggested Mid Band Assignment

Channel	Lower Band Edge	Video Carrier	Sound Carrier	Upper Band Edge
A	117.0	118.25	122.75	123.0
B	123.0	124.25	128.75	129.0
C	129.0	130.25	134.75	135.0
D	135.0	136.25	140.75	141.0
*E	141.5	142.75	147.25	147.5
F	147.5	148.75	153.25	153.5
G	153.5	154.75	159.25	159.5
H	163.5	164.75	169.25	169.5

*Note: The 0.5 MHz guard band as we go from the assignment for oscillator leakage to the assignment for image.

Suggested Super Band Assignments

Channel	Lower Band Edge	Video Carrier	Sound Carrier	Upper Band Edge
I	221.0	222.25	226.75	227.0
J	227.0	228.25	232.75	233.0
K	233.0	234.25	238.75	239.0

Channel	Lower Band Edge	Video Carrier	Sound Carrier	Upper Band Edge
L	239.0	240.25	244.75	245.0
M	245.0	246.25	250.75	251.0
N	251.0	252.25	256.75	257.0
O	257.0	258.25	262.75	263.0
P	267.5	268.75	273.25	273.5
Q	273.5	274.75	279.25	279.5
R	279.5	280.75	285.25	285.5
S	285.5	286.75	291.25	291.5
T	291.5	292.75	297.25	297.5
U	297.5	298.75	303.25	303.5

Note: All higher assignments are free of oscillator leakage or image problems.

An examination of the suggested assignments listed above shows that I have placed the channels that might be effected by oscillator radiation so that this interference falls at band edge of the newly designated Mid or Super band. This would allow Viewer A to adjust his fine tuner within ± 0.5 MHz and still be within the most tolerable range for interference.

In both the Mid and Super Bands, one can see that the well planned frequency assignment of the VHF television channels by the F.C.C. avoids the possibility that any channel will be effected simultaneously by oscillator leakage while it becomes an image to another VHF channel. In the proposed assignment Channels A, B, and C are positioned to minimize interference from TV set oscillators while Channels E, F, G, and H are positioned to give minimum image effect on Channels 2, 3, 4 and 5. Similarly, in the Super Band, Channels I through N are chosen to avoid oscillator interference and Channels P through U are chosen for minimum image effect on Channels 8 through 13.

In chosing the assignment for least image effect, the following example is given -

Desired	Local Oscillator	-	Channel 3	=	I.F.
	107 MHz	-	(61.25 Vid; 65.75 Aud)	=	(45.75 Vid; 41.25 Aud)

Image	Channel F	-	Local Oscillator	=	I. F.
	(148.75 Vid; 153.25 Aud)	-	107 MHz		(41.75 Vid; 46.25 Aug)

The frequency on the system is chosen so that the image interference from the video carrier falls at 4.0 MHz or above in the IF and the sound carrier (which is usually 15db weaker than video) falls at -0.5 MHz in the channel.

Conclusion:

Cable systems are capable of carrying channels in the spectrum between VHF Channels 6 and 7 and above Channel 13. The Mid Band should give us no trouble if we chose the best possible assignment to minimize the effects of oscillator leakage and image. The data presented shows that oscillator leakage into this spectrum from the TV set is substantially lower than the leakage into the Super Band. The image rejection of tuners in the low VHF band is also better than in the high band and, in addition, the isolation between customer taps in this range is typically better than the 20db figure quoted in Figure I. The Super Band becomes more of a problem, since oscillator leakage, image rejection and customer tap isolation are all at their poorest performance level. Choosing the best frequency assignment for the Super Band is an appropriate step under any condition. If a particular customer (Viewer A) is causing interference with another (Viewer B) a special low pass filter, passing up to Channel 13 and then droppingoff rapidly, could be inserted into the offenders drop line only when such interference exists.

Appendix

Description of Television Sets Used in Table II

TV Set	Description		
A	1968 RCA	23inch	Color Set
B	1963 Zenith	23inch	Black & White
*C	1967 Zenith	23 inch	Color Set
D	1967 Motorola	17 inch	Color Set
E	1969 Zenith	23 inch	Color Set
**F	1964 Zenith	23 inch	Black & White

*Two identical sets were tested with almost identic 1 results
**This set appears to have traps at tuner input to minimize leakage. Any adjustment of fine tuner greatly worsens these readings.

DISCUSSION

Mr. Jeffers: Are there any questions? Come up to the microphone, please.

Mr. George Brownstein: I'm a consulting engineer. I've followed with quite a bit of interest the talk you just gave and there's one point which I consider quite important that I notice that you didn't cover and perhaps you have given it consideration, but I also notice that in the general trend of the industry, the industry doesn't seem to be giving it any particular emphasis. This has to do not so much with a particular frequency that you assign for the mid band frequencies, but the area of the spectrum that you assign them in. What do I mean? The area that the frequencies for the mid band being assigned to fall, generally, above the FM band and below Channel 7. That's put them practically in the middle of the aviation band and I notice that in some of your recommendations, some of your frequencies fall in the middle of the aviation band. They range from roughly 108 to 120 megacycles which is used for ILS (Instrument Landing Approach Systems).

Approximately 10 years ago, a company that I was formerly associated with, did some work and some research in this area. And we were very concerned at that time, of course, we were involved with cable systems in putting two kinds of information on the cable system both of which were video, but one was in the normal 54 to 216 band and the other was in the mid band area and the biggest thing that worried us was not that we couldn't send the signals down the cable, but suppose we did and the fault condition developed on the cable. We could just see an airplane come homing in on one of our carrier frequencies, and you can, of course, imagine what the end result of that would be.

Now I notice with a great deal of disturbance that the industry has sort of bemused itself with these additional frequencies and are doing a lot of arithmetic which makes a certain amount of sense, but what consideration has been given to the fault condition once you go into operation in this area and suppose you have been in operation for five years and a fault condition develops either in the amplifier, or the fitting, or the drop cable going into the subscriber's home? How do you answer questions like that?

Mr. Jeffers: Well, I can give you my opinion, but I am far from an expert in that area. First of all, I think you'll find the ILS band stops at 118 and although the band edge of the lost channel, I indicated, was down into that band that the carrier was up at, I believe, 118¼. Now, there is no question that that channel, I think, in certain instances, might have to be dropped. Another thing, of course, in the last 5 or 6 years with the use of solid sheath cable, the radiation from the systems is down substantially. First of all, remember the worst power level--the highest power level--in this entire CATV is of such a low magnitude that you're really in a complete open circuit probably not putting too much out to the air, although certainly something significant. I think most of the mid band would be in the aviation communications band, not in the navigation aids band. If you'll notice, also, how many millions of oscillators from TV sets are radiating into that band at 113 megahertz right now and apparently it has not been a problem. In addition, I think you'll find the modulation on the ILS is of such a nature that it would probably see through any disturbance that could come from CATV. But again, let me repeat, these are my opinions.

Ken Easton: Toronto, Canada. I'd like to make two comments. First of all, I was interested in the superposition of that curve on the Jerrold curve showing the requirements of broadcast procedure 23 which within the next 3 or 4 months is going to be a mandatory performance standard in Canada for all CATV systems, existing as well as new. I was interested to see the extent to which that curve is above the Jerrold curve by, as far as I can see, about 12 or 15 dB. I would like to make the comment that the Procedure 23 curve is, of course, a minimum performance standard and it would be expected that the design curve, the design standard, would be somewhat better than that---shall we say 12 or 15 or, at least, 10 or 12 which, I guess, would put it pretty close to your Jerrold curve. So I'm gratified to see that.

The second comment I would like to make is in support of George Brownstein's remarks concerning this question of radiation. In Canada, the Department of Communications which, of course, is the technical licensing authority of the cable systems is not, at present, prepared to approve the use of the mid band for this very reason. They are concerned and very concerned about the radiation from systems and by radiation they don't just mean radiation under normal operating condition, but as George pointed out, radiation under any, including fault conditions. There is no question that with solid sheath aluminum cable and the lower power levels which we use now compared with the older tube equipment, that the radiation from a working system in good condition is very much lower than it would have been under the old systems. But we have to concern ourselves with fault conditions and the thing that worries the Department of Communications is the fact that, in Canada, our radiation limit, at the moment, for any frequency--any frequency--including the high VHF band outside of the low VHF channels 2 to 6 is 9 microvolts per meter at 30 feet and they're very concerned about the fact that quite apart from the ILS band that George spoke of, there is a very considerable useage of land mobile within that band and many of these land mobiles have receivers having sensitivity of a half a microvolt and 9 microvolts a meter at 30 feet doesn't do them any good at all and they're looking at this whole question right now and are not prepared to license the use of this band until they have, in fact, substantially modified the radiation limits--to what, I don't know yet, but substantially better than 9 microvolts per meter.

Mr. Jeffers: By the way, again not being an expert on the radiation problem, I'll pass that, but let me emphasize that the so-called Jerrold curve, as I said, is an extremely critical curve. I would guess that a curve about 5 dB less offensive than that would be perfectly adequate. Thank you.

COPPER CLAD ALUMINUM FINE WIRE IN COAXIAL CABLE

John C. Fan, Manager, Research & Development Copper Clad Aluminum Wire
Frank A. Spexarth, Manager, Communication Industry Products

Two years ago at your annual convention in Boston, Texas Instruments presented a paper suggesting that a materials system of copper clad aluminum had significant application possibilities in the area of CATV semi-rigid trunk or distribution cables. At that time your industry had little experience, and no regular users, of this material. For that reason the paper centered on an evaluation in depth of those technical aspects of copper clad aluminum cable relevant to its suitability for CATV. It might be well to review the primary considerations of that presentation.

First, you may recall, we calculated the skin depth of copper required to transmit the range of RF signals employed in CATV, and compared it with the skin depth present on 15% copper clad aluminum cable of three common gauges. Table I summarizes these findings. Note that even at the lowest frequency, 15% copper clad aluminum in the .412 cable provides approximately 3 times the skin depth of copper needed to transmit this range of RF signals.

Next, we made a characteristic-by-characteristic comparison of copper clad aluminum and solid copper conductors, as indicated in Table II. Since this comparison would seem to indicate D.C. electrical resistance as a possible drawback to copper clad aluminum, we investigated this in considerable detail to determine what this difference in resistivity could mean in CATV applications. The conclusion, you may recall, was that use of copper clad aluminum in an actual system approximating our theoretical analysis would provide sufficient benefits in cost to more than offset the expense of additional power supplies that might be required.

The paper also explored such factors as a comparison of thermal expansion coefficients, especially as they might relate to the common problem of conductor pullout; compared weights as they might affect handling and safety; and discussed other significant characteristics of copper clad aluminum.

Today, most of these points hardly require discussion with your industry, because you have acquired considerable first-hand experience with copper clad aluminum cable. It is, as you know, receiving extensive use nationally, is readily available from important suppliers to your industry, and has been chosen by major installers for thousands of miles of CATV service.

We should, however, review briefly the experience to date on conductor pullout. At the time of our first report we suggested that the closely matched coefficients of linear expansion of a copper clad aluminum center conductor and an aluminum outer conductor should significantly reduce conductor pullout and cable buckling. As it turned out, the combination of matching coefficients of center and outer conductors only partially relieved this phenomenon. One factor remained unchanged: the coefficient of linear expansion of the dielectric material between the conductors. At least one major cable manufacturer has made significant progress recently in overcoming this final obstacle to pullout-free cable. Their solution is the application of optimum-adhesion bonding techniques between the conductors and dielectric. The field experience to date using this improved cable indicates that this approach goes a long way toward solving the pullout problem.

So much for background. Now let's examine what is happening with copper clad aluminum wire today, again as it relates to the CATV industry.

Early this year Texas Instruments announced to the cable industry the availability of copper clad aluminum in fine wire sizes as small as 40 gauge--finer than a human hair. The significance of this development to CATV is that you now can choose copper clad aluminum for drop cables, as well as for semi-rigid trunk or distribution cables.

It is particularly timely for us to re-examine drop cables today--in light of increasing subscriber demand for high quality signals and the more rigid enforcement of government regulations on signal leakage and interference. The time has come, in short, for CATV to address itself to this, the weakest link in its transmission system. First let us consider the nature of the problem with drop cables as they generally exist today.

Since drop cable is installed or used only when a subscriber requests CATV service, it has not been considered a primary installation item, as with trunk or distribution cables. Perhaps for that reason drop cable design seldom has emphasized performance. RG-59B/U (Table III) flexible cable per MIL-C-17D is used extensively as a basic drop cable. But many variations and modifications have been made on this cable, primarily to reduce its cost. Performance, in fact, has been ignored to the point that some drop cables on the market have an incompatible characteristic impedance of 88 ohms, caused from substituting foam for solid dielectric without properly adjusting the diameter ratio of inner and outer conductors. Ironically, as drop cable performance has deteriorated, the acceptance of color TV has made the need for better drop cable designs increasingly important.

What, then, are the characteristics of good drop cable design, and what are the alternatives in selecting cable for CATV systems? Table IV lists eight significant items for consideration when you are selecting drop cable. Using this as a guide, we can analyze the three types of drop cable available today, classified by the type of shielding employed: solid metallic, braid, or metallic tape.

SOLID METALLIC SHIELD

Drop cable with a solid metallic shield is similar in design to semi-rigid trunk and distribution cable, except of a smaller gauge. Without question it offers the most effective shielding, the greatest durability, and potential for long service life. On the other hand, installation is relatively difficult because it lacks flexibility and has greater weight than either braided or metallic tape cable. Principally for economic reasons, solid metallic shielded cable is used only in extreme cases of high signal density, where its exceptional shielding capability is of great value.

BRAIDED SHIELD

There are two types of drop cable employing braided shielding, one with a single braid and the other with a double braid. The principal difference between them is the significantly greater shielding provided by a double braid. A single braid cable design, RG-59, has been the standard drop cable for many years and has been specified for military applications. It received this wide acceptance because:

--It is readily available from all cable manufacturers

--Its 75 ohm impedance matches CATV system design

--Its flexibility makes it easy to handle and install

--It has good mechanical strength permitting unsupported aerial installation

--It has been performance-proven for over 30 years, including specification by the military

--Its copper clad steel center conductor can be used as a center pin for Type F connectors, thereby reducing connector costs

Its one drawback, in the single-braided form, is shielding effectiveness. On the other hand, no other design offers the flexibility and service reliability of braided cable. For that reason virtually all government-specified flexible cables are of braided design. Where required, double-braid cable can provide acceptable shielding of operating frequencies into the microwave region.

METALLIC TAPE SHIELD

The second type of drop cable in common use today is shielded by a thin metallic tape of copper or aluminum foil, frequently bonded adhesively inside a dielectric layer for mechanical strength. Tape may be applied either longitudinally or in a helical wrap. Longitudinal shielding is more effective but has poor flex-life. Helical wrapping is flexible but offers relatively poor shielding. Since the metallic tape cannot effectively conduct a power signal, a drain wire or outer braid must be added to complete cable functions. Tape-shielded cable provides excellent shielding until it is exposed to repeated flexing and alternating thermal expansion and contraction. If the elements cause even a small break at a transverse plane, severe signal leakage can result, because a signal radiator forms immediately. Cable flexing tests conducted by Texas Instruments according to standard test methods MIL-C-915 resulted in drain wire failure at 300 flex cycles and metallic tape failure after 800 cycles. (By way of comparison, braided cable has maintained satisfactory performance after more than 2,000 cycles.)

As indicated in Table V, braided cable continues to offer the best combination of drop wire characteristics. With copper clad aluminum now available for braided cable application, the question remains: can copper clad aluminum substitute for solid copper braid without compromising performance? To answer this we must examine the nature and properties of copper clad aluminum.

MAKING COPPER CLAD ALUMINUM FINE WIRE

From a material supplier's point of view, the standard copper clad aluminum material is a 5/16" redraw rod. Its solid aluminum core is covered completely and continuously with a layer of solid copper. The ratio of cross-sectional area or volume is nominally 85% Al to 15% Cu. This rod is made by a continuous metallurgical bonding process which joins these dissimilar metals at their atomic level. There is no physical separation of the metals and no bonding or adhesive elements are used in uniting the copper and aluminum layers.

As a result of this kind of bonding, copper clad aluminum rod performs and works exactly like a monometallic conductor. In fact, 5/16" copper clad aluminum rod can be drawn and in-line annealed down to 40 AWG fine wire on regular copper wire drawing machines. For cable application, 36 AWG wire is the smallest in common use today. In-line resistive annealing is essential in copper clad aluminum fine wire manufacture to assure long flex-life and more desirable spring-back qualities.

Important to the manufacture of copper clad aluminum fine wire is the fact that rod can be drawn down to fine wire sizes without any change in cladding volume or area ratio. In Figure VI-A, showing magnified cross sections of a 5/16" rod and a #36 gauge fine wire, it is obvious that cladding area is maintained uniformly and concentrically. In fact, from a 1/4" length of rod we can create 71 feet of fine wire, as shown in Figure VI-B, with no ratio change.

PROPERTIES OF COPPER CLAD ALUMINUM FINE WIRE

Since copper clad aluminum fine wire is a metallic structure of both copper and aluminum, many of its properties--such as tensile strength, spring-back and DC resistivity--reflect the basic properties of the two monometals held in bond. But some important properties such as skin effect, RF resistivity, and braid contact resistance are different enough to require analysis.

Six important RF cable properties are listed in Table VII. According to these formulas for coaxial cable, it can be seen that characteristic impedance, capacitance, reflection coefficient and VSWR will not change when a copper conductor is replaced by an equal size copper clad aluminum conductor. This is because the function formulas are independent with respect to conductor resistivity. Attenuation and Q-factor formulas do depend on resistivity of the cable conductor and its skin depth.

A major reason for using copper clad aluminum wire for CATV transmission cable is its skin effect. As mentioned in our previous paper, there is ample copper to carry television signal frequencies in all three sizes of copper clad aluminum distribution cable. RF attenuation characteristics of the cable are exactly the same as with a solid copper center conductor.

Now, what will happen in a coaxial drop cable having copper clad aluminum fine wire with a copper skin depth only 0.000256" thick? The analysis given in Tables VIII and IX shows that even if the copper thickness of copper clad aluminum fine wire is less than one skin depth at the lowest television frequency, the effect on cable attenuation is still negligible. The reasons are:

- --RF current density decreases logarithmically with respect to depth in conductor.
- --Aluminum is also a good RF conductor.
- --Braid resistivity contributes a much lower percentage of the loop resistance than center conductor resistivity.

As shown in Table IX, the theoretical attenuation increase due to AWG 34 copper clad aluminum braid shield in RG-59 B/U cable is only 2.7% at 50 MHz and 2.2% at 250 MHz. In laboratory tests, there are no significant differences in any RF property when copper clad aluminum braid shield cable is compared with solid copper braid, as shown in Table X.

BRAID CONTACT RESISTANCE

In braided cable design for better shielding, the current should be able to flow freely from one strand to another without meeting high resistance which causes current path alternation and increases cable attenuation. Therefore, braid contact resistance must be examined. An extensive test conducted in our laboratory indicates that copper clad aluminum wire braid has the lowest braid contact resistance of all cable tested, including copper, aluminum and copper clad aluminum.

--Compared to copper, annealed copper clad aluminum wire has low spring-back characteristics which produce a tighter braid with excellent electrical contact throughout the braided shield. This is an important factor why copper clad aluminum wire braided cable has shown a slightly lower attenuation and better shielding effectiveness in television frequency band during actual laboratory tests shown in Table X.

--On the other hand, when two aluminum conductors cross under low pressure a very high contact resistance results because aluminum's unavoidable oxide acts as an insulator. Aluminum also is more difficult to connect either to itself or to other materials.

SUMMARY

The cable television industry has acquired extensive experience during the past two years with copper clad aluminum semi-rigid trunk or distribution cables. It has proven itself a perfectly acceptable substitute for solid copper cable from a standpoint of performance and offers a number of desirable characteristics of its own, including the possibility of pullout-free cable. Now this metallurgically bonded materials system is available in fine wire sizes suitable for braided drop cable. Recent cable designs often ignore performance at the very time color television makes quality signals more important. A review of current designs indicates that single or double-braided cable still offers the best combination of drop wire characteristics. Detailed laboratory testing further indicates that copper clad aluminum fine wire can be substituted for solid copper in braided cable with no compromise in performance.

TABLE I

CATV FREQUENCY RANGE	USEFUL "SKIN"	CATV SIZE	"SKIN" AVAILABLE 15% CU/AL
50	.00108	.750	.00570
		.500	.00380
250	.00048	.412	.00300

TABLE II

PROPERTY	CU/AL	CU
DENSITY (#/in^3)	.132	.323
RESISTIVITY ($\frac{\text{ohms}}{\text{CMF}}$)	15.60	10.27
WEIGHT, COPPER (%)	36.8	100
T.S. Kpsi ANNEALED	17.0	35.0
Y.S. Kpsi ANNEALED	13.0	20.0
COEFF. OF THERMAL EXPANS. in/in/° C x 10^{-6}	22.0	16.8

TABLE III

RG-59 B/U COAXIAL CABLE

INNER CONDUCTOR: 0.0230" COPPER COVERED STEEL

INSULATION: SOLID TYPE A (POLYETHYLENE)
DIAMETER: 0.146" ± .004"

OUTER CONDUCTOR: SINGLE BRAID
TYPE: BARE COPPER
WIRE SIZE: 34 AWG
CARRIERS: 16
ENDS: 7
PICKS/INCH: 8.2 ± 10%

JACKET: TYPE II a (NON-CONTAMINATING PVC)
DIAMETER: 0.242" ± .004"

ENGINEERING DATA:

NOMINAL IMPEDANCE: 75 ± 3 OHMS
NOMINAL CAPACITANCE: 21.1 pf/ft
ATTENUATION: 55 mc - 2.6 db/100 ft.
83 mc - 3.2 db/100 ft.
175 mc - 4.9 db/100 ft.
211 mc - 5.4 db/100 ft.

PER MIL-C-17D

TABLE IV

BASIC DROP CABLE DESIGN REQUIREMENTS

1. ELECTRICAL TRANSMISSION QUALITY
2. LOW COST
3. SHIELDING EFFECTIVENESS
4. FLEXIBILITY
5. MECHANICAL DURABILITY
6. LONG SERVICE LIFE
7. LIGHT WEIGHT
8. AVAILABILITY

TABLE V

COMPARISON OF DROP CABLE CONSTRUCTIONS

DESIGN CHARACTERISTIC	SOLID METALLIC SHIELD	TAPE SHIELD HELICAL	TAPE SHIELD LONGITUDINAL	BRAIDED WIRE CONSTRUCTION SINGLE	BRAIDED WIRE CONSTRUCTION DOUBLE
ELECTRICAL TRANSMISSION	A	A	A	B	A
LOW COST	D	A	A	A	B
SHIELD EFFECTIVENESS	A	C	A	B	A
FLEXIBILITY	D	A	C	A	A
DURABILITY	A	B	C	A	A
LONG SERVICE LIFE	A	B	D	A	A
LIGHT WEIGHT	D	A	A	A	B
AVAILABILITY	D	C	A	A	C

KEY

A - EXCELLENT
B - GOOD
C - FAIR
D - POOR

FIGURE VI-B

ONE GRAM OF COPPER CLAD ALUMINUM WIRE

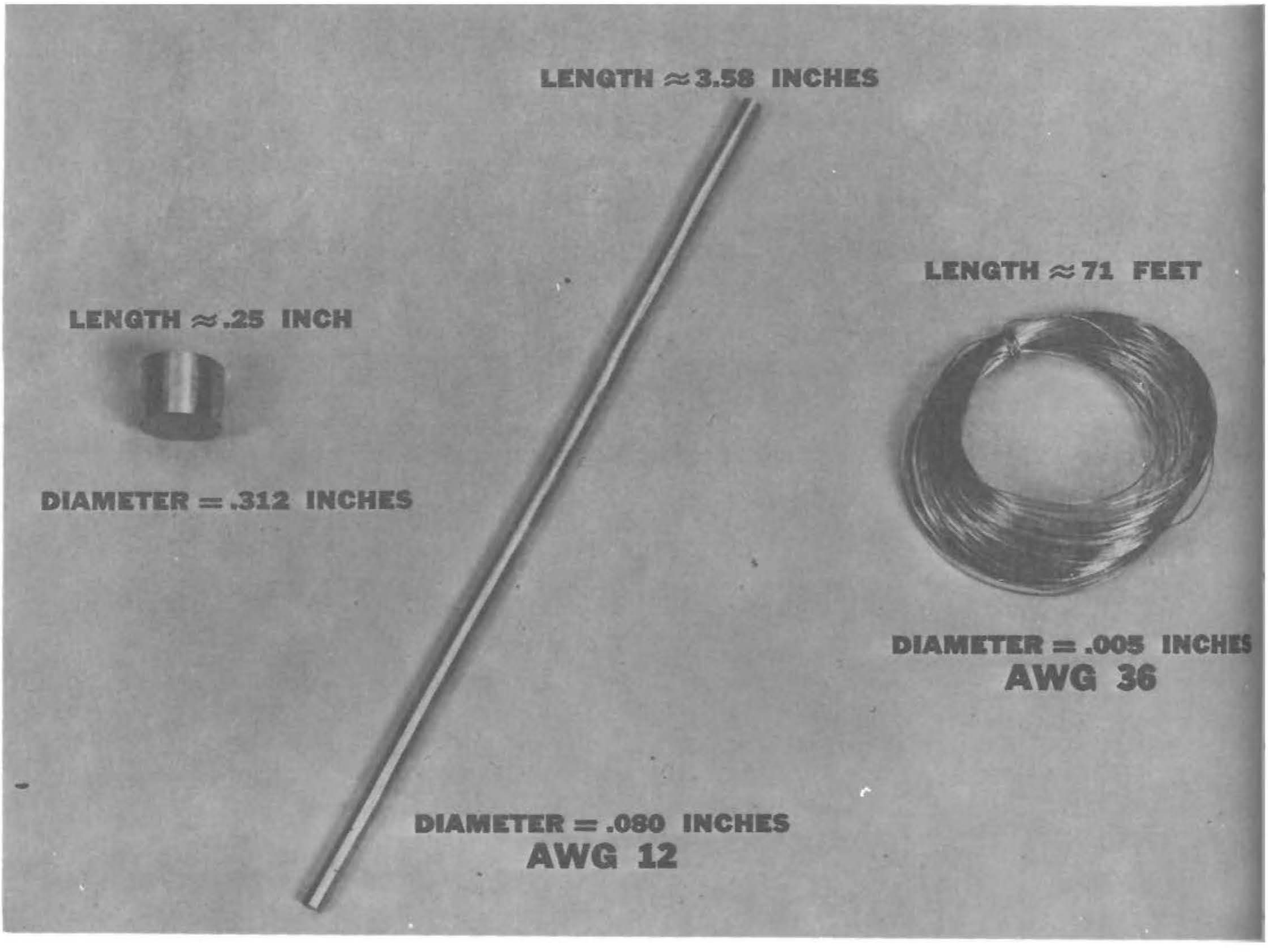
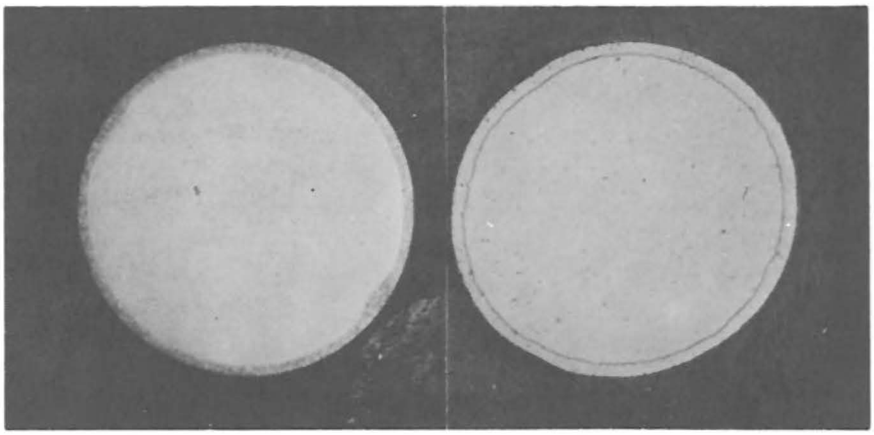


FIGURE VI-A

CROSS SECTION OF METALLOGRAPHIC PHOTO

COMPARING COPPER CLAD ALUMINUM ROD (.312")
& COPPER CLAD ALUMINUM FINE WIRE (.005")

ROD	FINE WIRE



DIA. .312 inches	DIA. .005 inches
AREA 0.07666 SQ. IN.	AREA 0.0000196 SQ. IN.

AREA RATIO = 3900

TABLE VII

COAXIAL CABLE FORMULAS

Listed here are some of the more important formulas for CATV coaxial cable.

(1) Characteristic Impedance

$$Zo = \frac{138}{\sqrt{\epsilon}} \quad \log_{10} \frac{D}{d}$$

(2) Capacitance

$$C = \frac{7.36}{\log_{10}\frac{D}{d}}$$

(3) Reflection Coefficient

$$r = \frac{Z - Zo}{Z + Zo}$$

(4) VSWR

$$VSWR = \frac{1 + r}{1 - r}$$

(5) Attenuation

$$= \frac{2.387 \times 10^{3} \sqrt{\epsilon \quad f}}{\log_{10} \quad \frac{D}{d}} \left(\frac{\sqrt{R_1}}{d} + \frac{\sqrt{R_2}}{D} \right)$$

(6) Q - factor

$$Q = \frac{\pi \; fcZ_o}{\alpha}$$

where:

Z_o = Characteristic Impedance in Ohms
C = Capacitance in Micro-Micro farads per foot
r = Reflection coefficient in Percent
α = Attenuation in decibel (DB) per 100 feet
ϵ = Dielectric constant
D = Inner diameter of outer conductor
d = Outer diameter of inner conductor
f = Frequency in cycles per second
R = Resistivity of conductor
c = Velocity of electromagnetic wave

TABLE VIII

SKIN DEPTH

$$I_z = I_o\, e^{-\frac{x}{d}}\, e^{-j(\frac{x}{d})}$$

where I_z = current function

I_o = current at surface

x = depth function

d = skin depth

If only the magnitude of current is considered, then the equation can be written as

$$I_z = I_o\, e^{-\frac{x}{d}}$$

$$\frac{I_z}{I_o} = e^{-\frac{x}{d}}$$

$$J_t = \text{Total Current} = \int_o^{\infty} I_o\, e^{-\frac{x}{d}}\, dx = I_o\, d$$

Current in outer metallic layer

$$J_1 = \int_o^{y} I_o\, e^{-\frac{y}{d}}\, dx$$

$$= -I_o\, d\, e^{-\frac{y}{d}} \Big|_o^{y}$$

$$= I_o\, d\, (1 - e^{-\frac{y}{d}})$$

THEREFORE, THE TOTAL CURRENT IN THE OUTER LAYER FOR

$$J_1 = I_o\, d\, (1 - e^{-\frac{y}{d}})$$

$y = \frac{1}{10}\, d$	$J_1 = I_o$	$(1 - e^{-0.1}) = I_o\, d\, (0.10)$
$y = 1/2\, d$	$J_1 = I_o$	$(1 - e^{-0.5}) = I_o\, d\, (0.40)$
$y = d$	$J_1 = I_o$	$(1 - e^{-1}) = I_o\, d\, (.633)$
$y = 2d$	$J_1 = I_o$	$(1 - e^{-2}) = I_o\, d\, (.865)$
$y = 3d$	$J_1 = I_o$	$(1 - e^{-3}) = I_o\, d\, (.951)$
$y = 4d$	$J_1 = I_o$	$(1 - e^{-4}) = I_o\, d\, (.982)$

TABLE IX

FREQUENCY (MHz)	CURRENT IN COPPER (%)	CURRENT IN ALUMINUM (%)	RF RESISTIVITY (RATIO TO COPPER)	RESISTIVE LOSS (BRAIDED CABLE*)
50	23	77	1.21	2.7%
250	40	60	1.17	2.2%

*NOTE: RG-59 B/U cable attenuation increases by a factor

$$A = R \frac{d}{d + D}$$

where R = Percent increase of Resistivity

d = Outer diameter of inner conductor

D = Inner diameter of outer conductor

A = .21 (.13) = 2.7% at 50 MHz

A = .17 (.13) = 2.2% at 250 MHz

For the 34 AWG copper clad aluminum braid wire, the copper skin layer has a nominal thickness of 0.000256 in. which is equivalent to approximately 0.258 at 50 MHz and 0.508 at 250 MHz. According to the above analysis the RF current was carried by the two metals in their proportions.

TABLE X

COMPARISON CHART

RG 59/U MODIFIED

	CU/AL BRAID CABLE	COPPER BRAID CABLE
Impedance	71.8 ohms	70.8 ohms
Capacitance	21.2 pf/ft	21.5 pf/ft
Velocity of Propagation	66.8%	66.7%
Conductor DCR	52.3 ohms/Kft	52.1 ohms/Kft
Shield DCR	6.25 ohms/Kft	3.98 ohms/Kft
Loop DCR	59.3 ohms/Kft	56.7 ohms/Kft
Reflection coefficient	0.178	0.170
Attenuation		
1 MHz (db/100')	.7	.5
10 MHz (db/100')	1.4	1.5
50 MHz (db/100')	2.6	3.3
100 MHz (db/100')	4.0	4.3
400 MHz (db/100')	7.9	8.6
1000 MHz (db/100')	15.6	16.0
Cable Weight Lbs/100 ft.	2.703	3.417

DISCUSSION

Mr. Frank Spexarth: Thank you. Are there questions?

Question: Not audible

Mr. Spexarth: You're referring to the trunk and distribution cables, primarily? I think on that particular question, I'd have to ask John Fan, technically, to back me up on that. You went beyond me as soon as you got into that.

Mr. John Fan: First of all, the skin depths even down to the video frequencies of 10 megacycles we test still have sufficient to carry even down to video frequency. Another thing is we use the fine wire here. Even aluminum, itself, is also a damm good RF conductor. So, basically, we really don't have to worry anything about it. We test, actually, down to 1 megacycle always.

Question: Not audible.

Mr. Fan: That is correct. It depends on the size. We cannot really describe the quality of video transmission at that range, but in the trunk end distribution cable like 412, the real sync difference we can see in the attentuating measurement is actually down to 150 KC range.

Question: Not audible.

Mr. Fan: We really get it down to a half dB or .2 dB difference of accuracy. We really didn't say that copper clad aluminum is one and half better than copper, but actually it showed a trend that it did have a slight better than copper.

Question: Not audible.

Mr. Fan: Yes. We braid this thing, actually. This was made by one of the cable manufacturers and on a standup production and they put 34 guage copper braid on and also put another roll of thin cable with 34 guage copper clad aluminum on.

Question: Not audible.

Mr. Fan: The point we tried to point out is the attentuation consists of three factors---one is resistive loss which is dissapative loss and second is leakage loss due to the braid leakage and third is reflective loss due to the cable VSWR. So if you have these three components to consider, the resistive loss seems like it starts to disappear by contribution from two other factors into the cables.

Question: Not audible.

Mr. Fan: That is correct. However, the leakage hiding the braid for the leakage loss it does show up a little bit better. Any other questions? Thank you.

DATA COMMUNICATIONS via CATV SYSTEMS

by

Thomas Chuang
Manager, Advanced Engineering
Vikoa, Inc., Hoboken, New Jersey

I. INTRODUCTION

The need for transmission of data and visual information over telephone networks has been growing very rapidly, especially in the last decade. In these telephone transmission systems the bandwidth being dealt with is usually 3 KHz for voice channels. Recently the introduction of Picturephones has created a requirement for a system with sufficient bandwidth to carry a 6.3 megabits per second signal (1) over long distances for adequate picture resolution. The wide bandwidth available in cable TV systems is attractive for the application of digital communications with the potential of higher speed transmission capability and of more narrow band channels.

This paper describes a method for implementing market survey, alarm monitoring systems, and a method for facsimile system, with relatively inexpensive and simple techniques as a step toward utilizing two-way communications in CATV. These equipments will be of great service to market research organizations, to system operators, and to any industry which has branch offices or which requires to communicate with others within the distribution of the system.

II. TWO-WAY COMMUNICATION SYSTEMS

Two-way transmission is required for data communications. The best two-way communication system is a dual cable system to handle heavy traffic and provide more channels but at the expense of high cost. However, viewed from the requirements of data channels' bandwidth and from the number of channels required per system for local origination, the proposed single cable, bi-directional system (2) shown in Figure 1 is adequate. This requires a relatively simple modification of the existing distribution system. The return direction will cover the frequency range from 5 MHz to 35 MHz. The first 6 MHz band can be reserved for data channels of narrow band, and the next 6 MHz may be allocated for data channels of wide band under time sharing basis. The rest of the channels are allotted for local origination and other public services.

III. CHANNEL MONITORING SYSTEM

One of the simple and useful applications of data communications in CATV is the channel monitoring system. The system is used for continuously sensing the channel setting of any subscriber's TV set and automatically storing or printing out its status information at a central office.

The system block diagram is shown in Figure 2. A control/display unit at a central office or headend sends out an address code data through the cable transmission line to the designated subscriber's transponder at the receiving end. Upon the receipt of a status data from the transponder, via the return path, the control unit will display the information and then send the second signal to request the second subscriber's status information. This is a sequential interrogation process which will continue until the 30th subscriber's status is recorded. The cycle will repeat again for the second group of another 30 sets until the 30th group is interrogated. The carrier frequency chosen for addressing is 73.5 MHz in the guard band and the frequency of 5 MHz is selected for the return carrier. TDM (Time Division Multiplexing) is used for the system and therefore requires synchronization (3).

A. CODING FORMAT

A clock of 60 Hz is used as indicated in Figure 3. Two clock cycles are used for two-word addresses, one for the subscriber's number and the other for sub-group number identification. Each word is represented by five binary digits to which five frequencies are assigned as shown in Figure 3. The channel setting identification code is FSK (frequency-shift-keying) modulated with five bits. Binary "one" is 3.2 KHz and "zero" is 2 KHz. Therefore the system will take 116 millisecond to scan one subscriber, and the basic system consists of 30 (sub-group) x 30 (units) = 900 subscribers. N different sets of frequencies assigned to the five bits of address code yield a system of 900 x N subscribers.

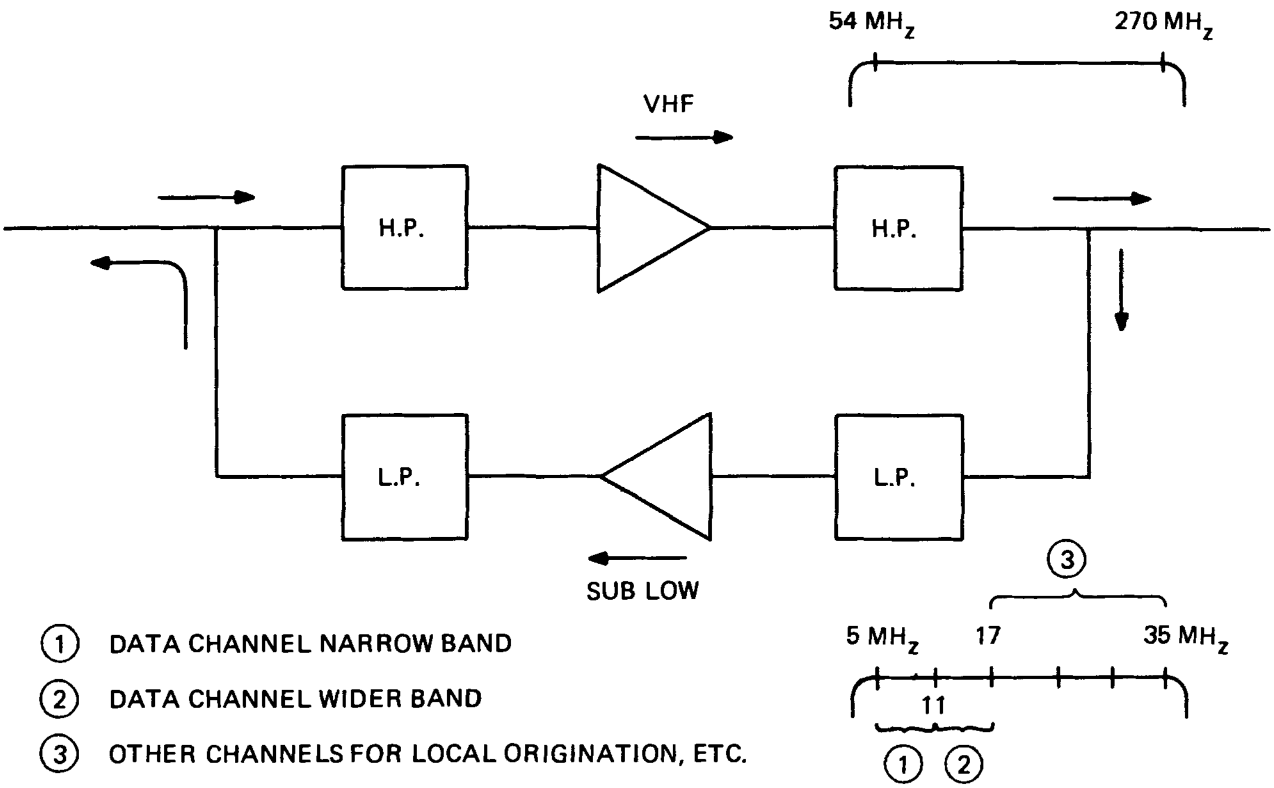


FIGURE 1. FREQUENCY DIVISION MULTIPLEX BI-DIRECTIONAL SYSTEM

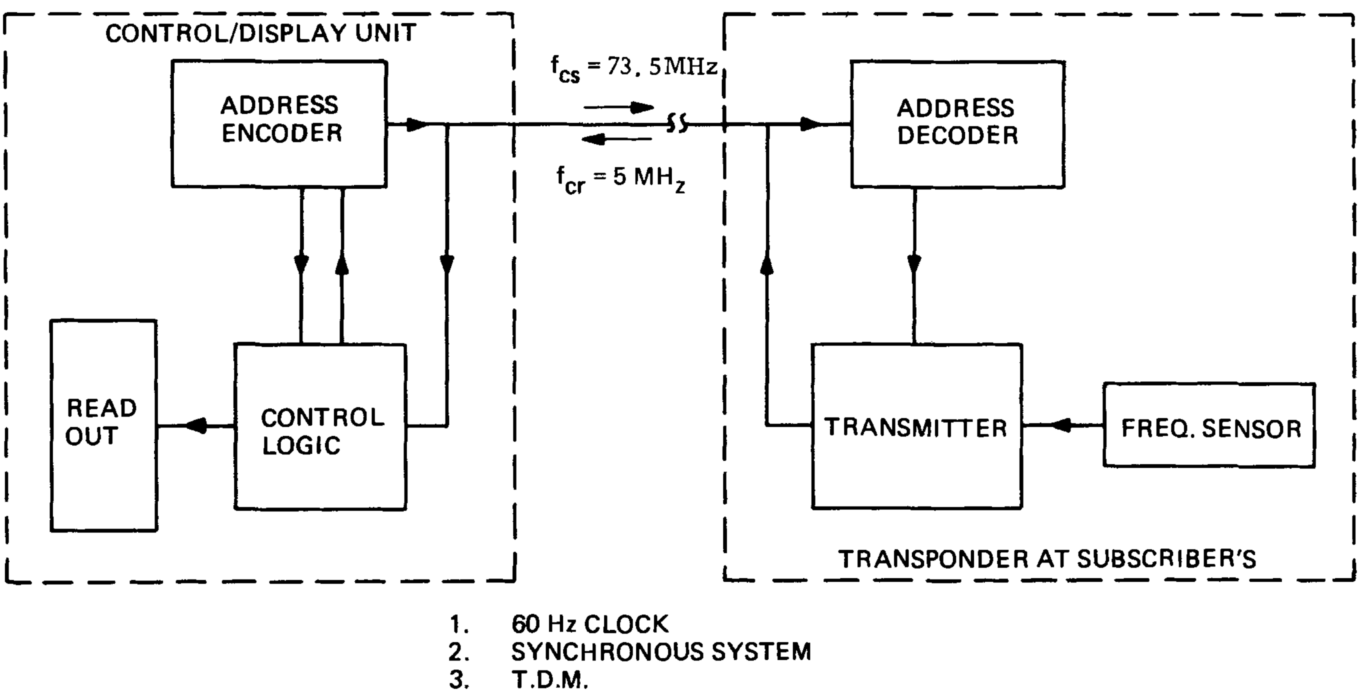
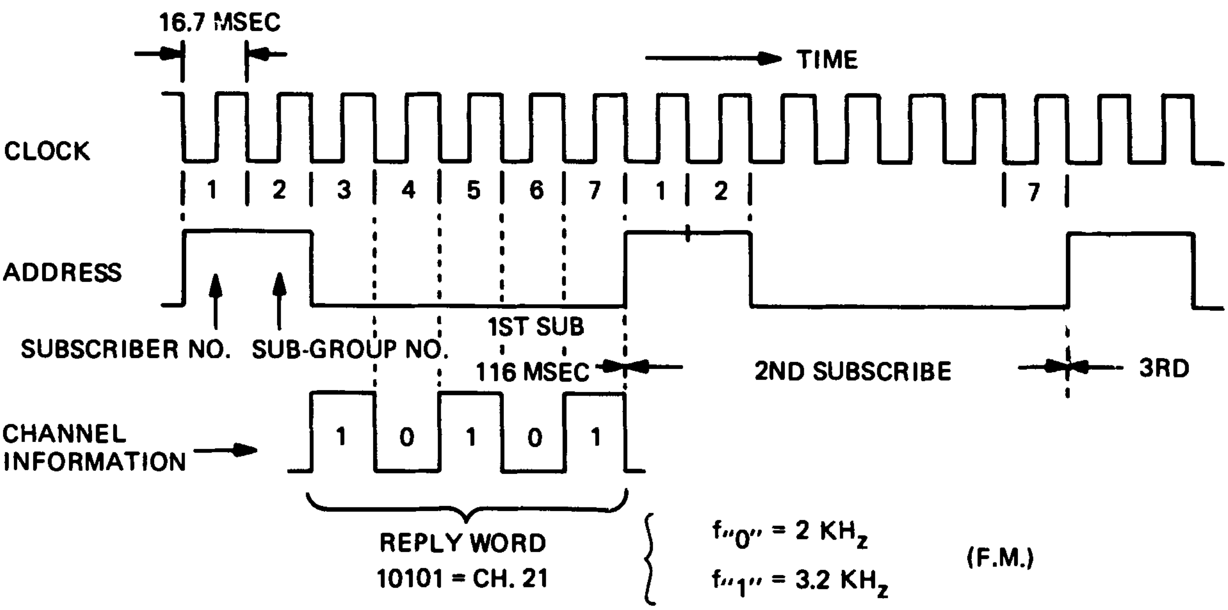


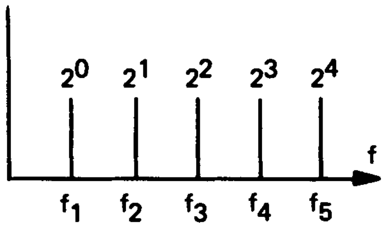
FIGURE 2. BLOCK DIAGRAM OF CHANNEL MONITORING SYSTEM



- 1ST ADDRESS CODE: (AM), f_T = 73.5 MHz

IDENTIFY

SUBSCRIBER'S NO. IN A SUB-GROUP



1ST BIT = f_1 = 600 Hz
2ND BIT = f_2 = 1.0 KHz
3RD BIT = f_3 = 1.4 KHz
4TH BIT = f_4 = 1.8 K. KHz
5TH BIT = f_5 = 2.2 KHz

- 2ND ADDRESS CODE
  TO IDENTIFY GROUP NO.

- BASIC SYSTEM CONSISTS 30(SUB-GROUP) x 30(UNIT) = 900 SUBSCRIBERS

- N DIFFERENT SETS OF FREQUENCIES ASSIGNED TO THE 5 BITS
  YIELD A SYSTEM OF 900 x N SUBSCRIBERS

- M DIFFERENT CARRIER FREQUENCY ASSIGNMENT → 900 x N x M SUBSCRIBERS

FIGURE 3. TIMING DIAGRAM

Furthermore, if M additional frequencies are assigned to the return carrier (frequency division) one will have a system of 900 x N x M subscribers.

B. ADDRESS ENCODER

Figure 4 illustrates the block diagram of the Control Unit while the encoder circuit diagram is shown in Figure 5. The ring counter counts the 60 Hz clock pulses one through eight and at every eighth count, one pulse is delivered to Counter Module 30 whose function is to perform 30 counts. The first incoming pulse which corresponds to the pulse for address code will be registered as 00001 (a code # for subscriber #1) and only f_1 is turned on. A carrier modulated with the f_1 is then transmitted. The next incoming pulse causes the register to shift into 00010 (subscriber #2) and turn on only the oscillator f_2.

After the 30th count, all registers become Logic "1" and the gate X opens and delivers one pulse. This pulse performs two functions --one, to clear the Counter so that another sequence of 30 counts will start; two, to be used for another group of 30 counts by the other Counter Module 30 (B) as shown in Figure 5. This module stores a group code so long as the acquisition of data from a group of 30 sets is still being undertaken. However, it rolls out the code data only at and during the presence of the pulse entered (gated) by the ring counter. This then forms two successive address codes for each interrogation. See also timing diagram in Figure 5b.

C. TRANSPONDER

The block and circuit diagrams of a transponder at the subscriber's set are shown in Figures 6 and 7 respectively.

The requirements on the sensor depend on the interface of a subscriber's TV set. If no converter is used, leakage from the local oscillator of a TV set may be sensed. However, with today's increasing demand of a system for more than 12 channel capability, perhaps a VHF-VHF converter will be mostly used for this purpose. The sensor depicted in Figure 7 takes the frequency from a V-V converter and feeds into a limiter for a constant amplitude and then its frequency is converted into voltage. The voltage is converted by analog-to-digital converter into 5 bits in parallel at 3.2 KHz clock rate.

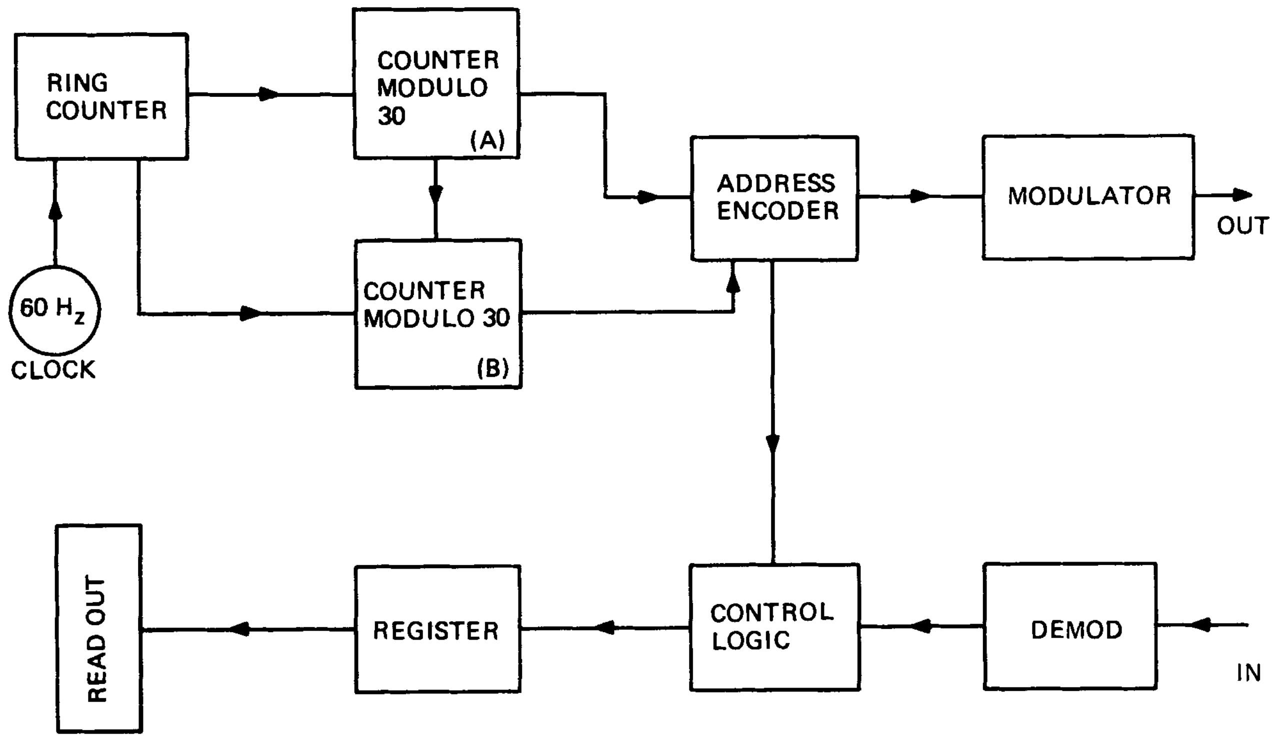


FIGURE 4. BLOCK DIAGRAM OF CONTROL UNIT

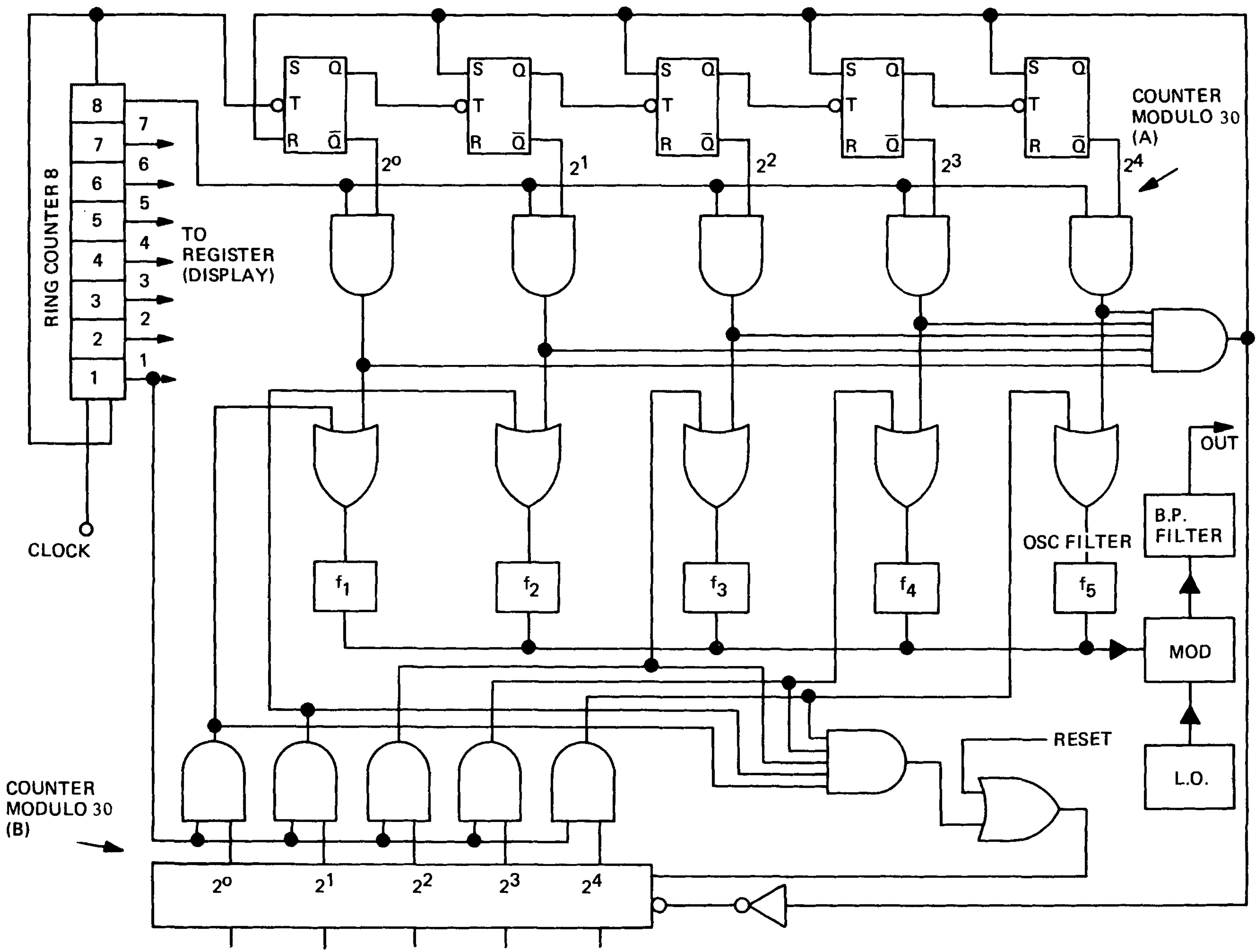


FIGURE 5A. DIAGRAM OF ADDRESS ENCODER LOGIC

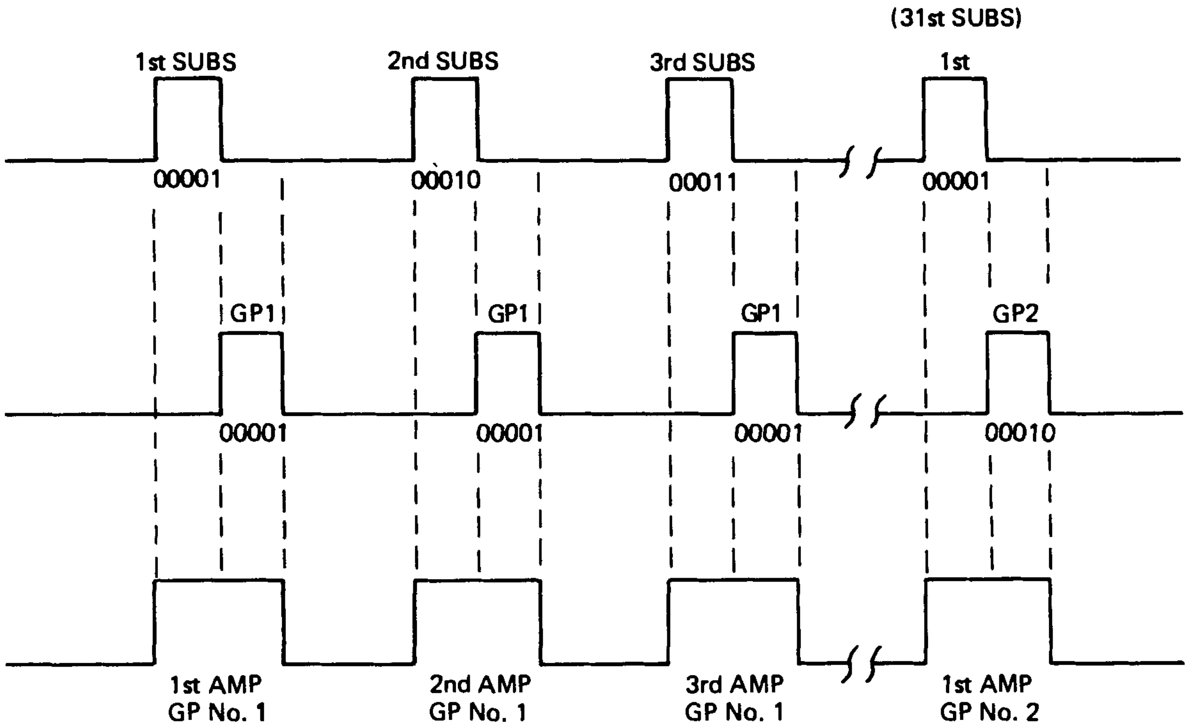


FIGURE 5B. TIMING DIAGRAM, OUTPUT OF ADDRESS ENCODER

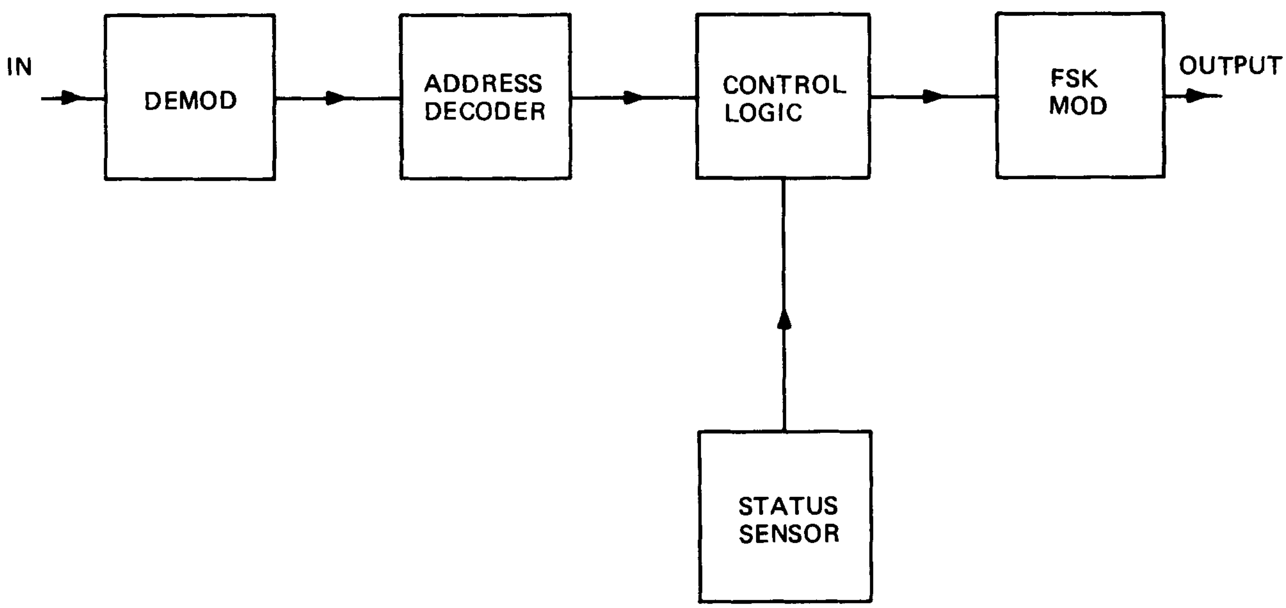


FIGURE 6. TRANSPONDER AT SUBSCRIBER'S

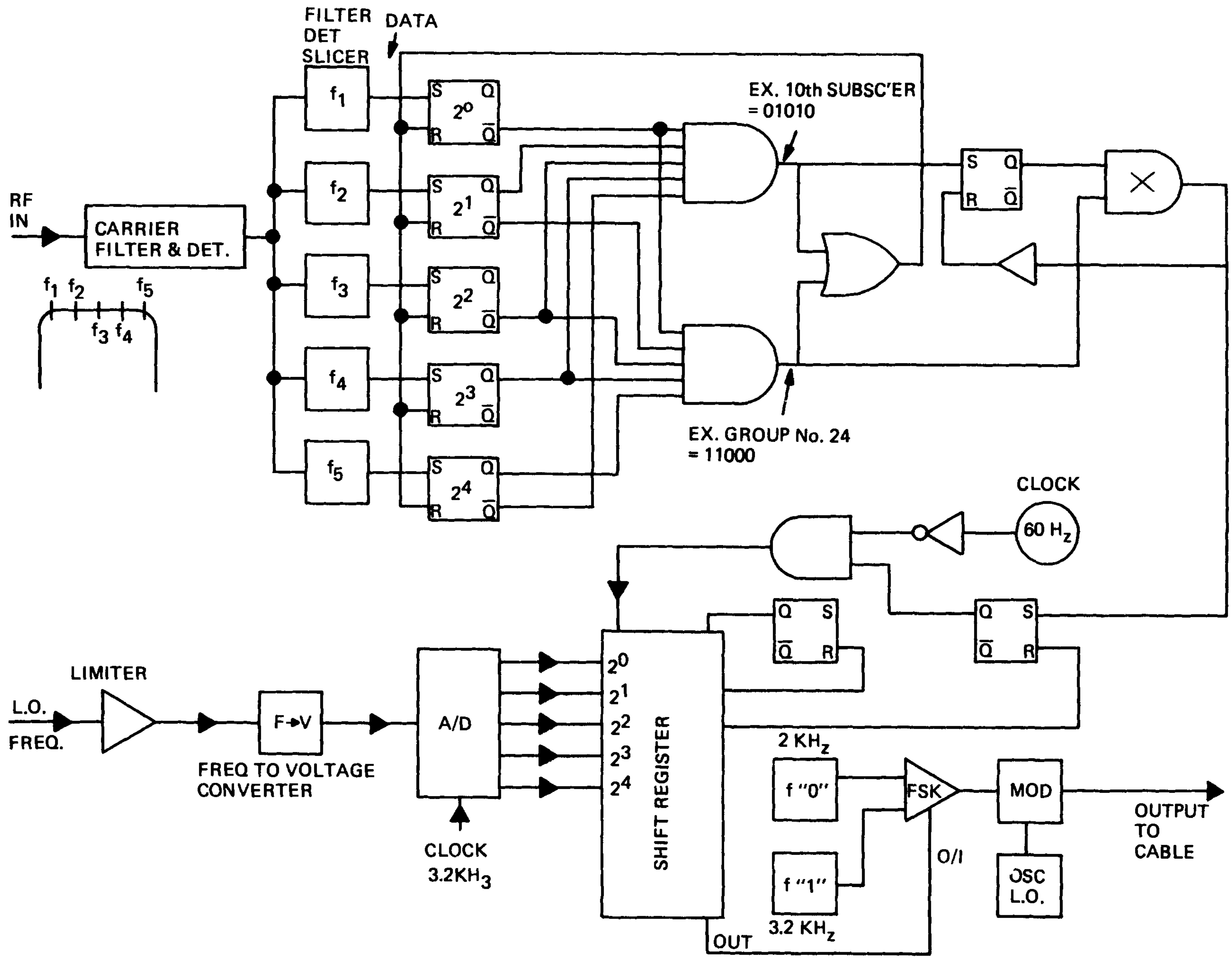


FIGURE 7. TRANSPONDER

Upon receipt of address code from the control unit, the demodulator attenuates noise and interference outside the band of interest and recovers the baseband wave. The recovered data then enters into an address decoder. If the code is in coincidence with the gate set for that code, the 60 Hz clock will be switched to the shift register to cause the five bits of channel-setting reply code to roll out serially. An integrated circuitry, gated, dual input amplifier is used for binary frequency modulation. With Logic "1" from the shift register, the input gate for 3.2 KHz signal will be open and with Logic "0", 2 KHz signal will appear at its output.

D. SYSTEM CONSIDERATIONS AND PERFORMANCE.

A synchronous system was chosen in order to reduce the complexity of the equipment, furthermore, a clock of 60 Hz is chosen to reduce the cost. Still the system can be operated to cover a distance of about 1000 miles, from a signal delay viewpoint, without getting into synchronization problems. The system performs the monitoring at the rate of 116 milliseconds per subscriber or one minute and fifty-six seconds for 1000 subscribers. By using amplitude modulation of parallel date for a reply code, this speed will be reduced by a factor of 2.3 to 1.

The bandwidth required for a transmission channel is a function of the number of subscriber sets to be monitored. The basic system of 900 subscriber monitoring requires a bandwidth of 4.5 KHz. For N=10, i.e., 9000-subscriber system, 10 sets of frequency allocation for f_1 through f_5 with 400 Hz spacing necessitates the bandwidth of system to be raised to 40 KHz which is still a very narrow band for a CATV system.

The cost of the system is largely dependent on the cost of the transponder. This is estimated to be under $30 per subscriber, based on the use of commercially available digital integrated circuits.

IV. ALARM MONITORING SYSTEM

The technique described above for channel monitoring can be directly applied for an alarm monitoring system. The alarm system is a means of detecting failures and pinpointing the failure's location in the CATV system. The need for an alarm system is justified for system maintenance for the following reasons:

A. CATV amplifiers and instruments must be in continuous operation (24 hours a day) and therefore, will occasionally fail.

B. They are usually installed outdoors and subjected to changes in temperature, weather, etc.

C. It is difficult to quickly find a failure location because the system is widely distributed.

Thus an alarm system should meet the following requirements:

A. Indication of the failure mode

B. Identification of the failure location

C. Indication of elapsed time and subsequent status change.

The channel monitoring system satisfies the above requirements with some minor modifications. Figure 8 shows an error detector in place of the aforementioned frequency sensor. The first bit of data indicates a trunk amplifier gain status, the 2nd bit indicates an amplifier tilt status, etc.

An address code identifies a status location and its timing plus a clock printer will continuously print out all events and the time of occurrence of any event.

Some features of this monitoring method over a conventional are:

A. Regardless of how many status changes occur simultaneously, the receipt of all signals is assured due to TDM technique.

B. A single channel is used.

C. The data is in a format convenient for data processing.

V. GRAPHIC DATA COMMUNICATIONS VIA CATV SYSTEMS

Thus far, systems of very low transmission speeds have been described for their particular applications. When communication of visual information between locations becomes necessary, higher data rates (for fast transmission speeds) becomes essential to reduce signal traffic and to save time.

An attractive application of visual communication on CATV systems is Facsimile. At this time this type of service is restricted

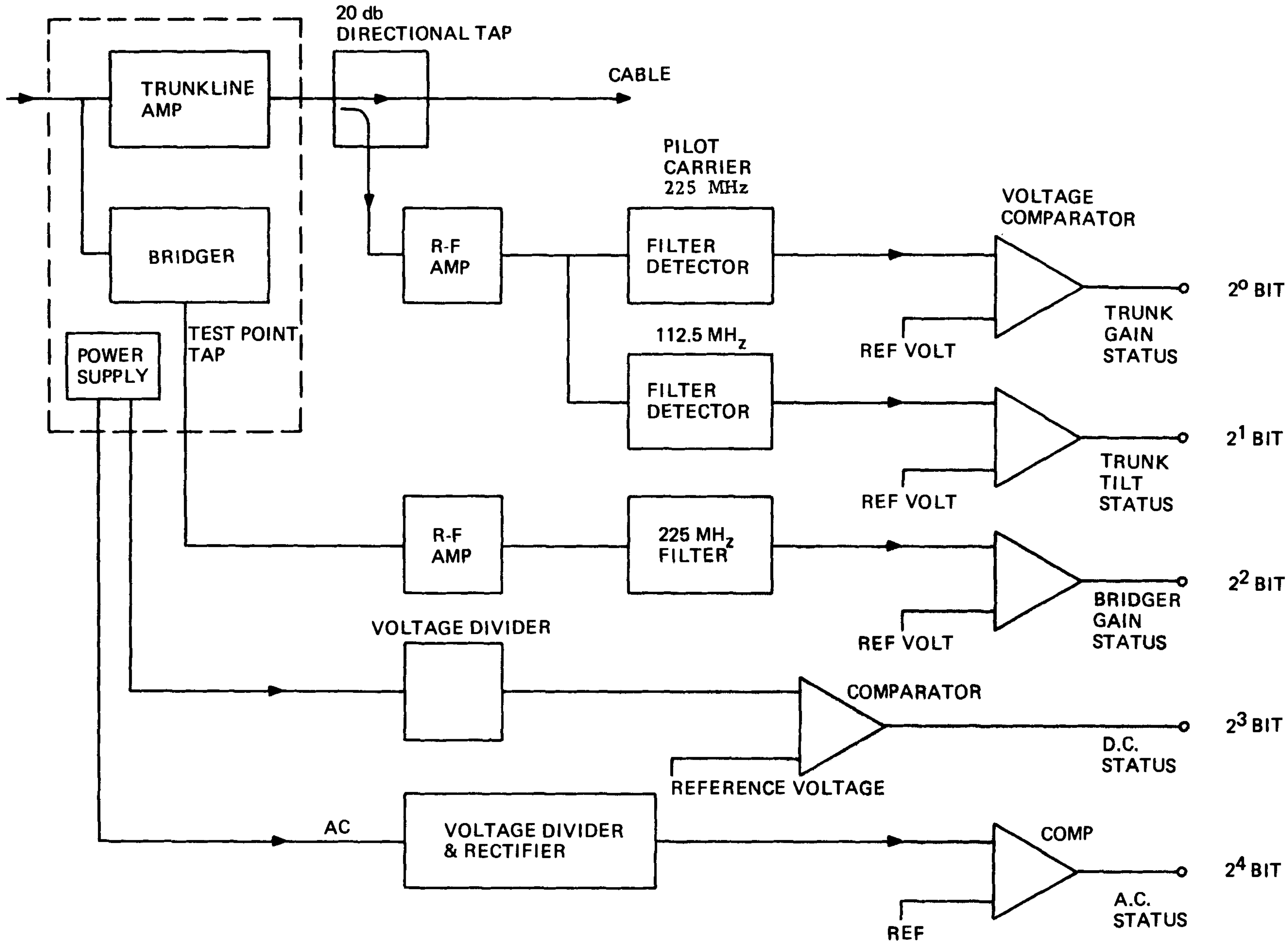


FIGURE 8. BLOCK DIAGRAM OF ERROR DETECTOR

usually for industrial or office use. However, as long as the number of installations are limited the implementation can be carried out at reasonable cost since the more sophisticated equipments which would allow TDM techniques to be utilized need not be used during this initial phase. For example a 6 MHz channel without TDM techniques could only support 20 subscribers for this type of service.

A. TRANSCEIVER EQUIPMENT

Most existing facsimile systems are designed for transmission over telephone lines, and the highest speed obtainable is 60 seconds for an average document of 8-1/2" x 11". The equipment to be described here is designed to use a flying-spot scanner and photo multiplier for image pick up and to use a CRT for display. If desired, a printing process can be added to obtain hard-copy printouts in about 10 seconds from the CRT display terminal. The basic technique used is one of "redundancy elimination" by sending video information at the two edges of an image (4) and by variable scan speed.

1. TRANSMITTER

The transmitter block diagram is shown in Figure 9. The horizontal scanning starts from left to right on the text onto which a "beam spot" from the CRT is focused. The line scanning is from top to bottom and is accomplished by means of a vertical stair-case generator. The space in time between horizontal sync pulses is analog, set by a voltage reference at the voltage comparator. When the spot travels horizontally and enters the edge of an image, a pulse of duration t is generated, i.e., a bit of information is generated to indicate the start of video. As the spot is on the verge of leaving the image, another pulse of t duration is generated to indicate the stop of video. The output is therefore a series of pulses of equal duration.

Legibility is a function of resolution. For a standard resolution of 100 scan lines per inch, one picture element = 0.01 inch. In order for a video-start-and-stop pulse t not to cause any spot positioning distortion on the CRT faceplate at the receiver, the time t should be less than the time required to scan one element. Suppose 150 KHz is used as tone signal as shown in Figure 11, thus t = 6.65 microsecond, then the bandwidth of the system = $\frac{2.2}{t}$ = 330 KHz. Total number of picture elements = 100^2 x 8 x 11-1/2 x 1/0.75 = 1,250,000 elements per page. To scan the image at a rate of 9 microsecond per element and white area at 3 microsecond per element results in:

Total transmission time per page for all black = 11.4 sec, all white = 3.9 sec, average page = 7.5 sec.

Which includes retrace 0.1 ms x 1470 lines = 1.47 sec.

2. RECEIVER

The receiver is similar to the transmitter except that a CRT is used to reproduce the exact text. It is "slaved" to the transmitter, The sync pulses are separated to drive its internal sweep generators.

The leading edges of video demodulated signals are sliced to form video pulses (Figure 10). A high persistence CRT or storage tube is used instead of using memory storage to refresh the frame. The latter may cause "jitter".

3. RESULTS AND FEATURES

a. High speed 75 kbits/sec is achieved. An average text of 8-1/2" x 11" can be transmitted in 7.5 sec. in contrast to 60 sec. of today's modern facsimile machines.

b. High resolution: 100 scan lines/in. horizontal; and 120 lines/in. vertical.

c. The latest print out equipment can be incorporated for hard copy.

B. COMMUNICATION LINK

The next immediate question is how data communications can be carried via the CATV system. A modem is built into each transceiver, to which a set of two carrier frequencies are assigned, (one for sending (f_s), the other for receiving (f_r)). A switching network is also included so that a tone signal can be transmitted to the headend for request and reply.

1. PRIVATE LINE

Referring to Figure 12, a converter with the frequency f_{12} is assigned to the two parties permanently, so that communications between the two are always at their disposal.

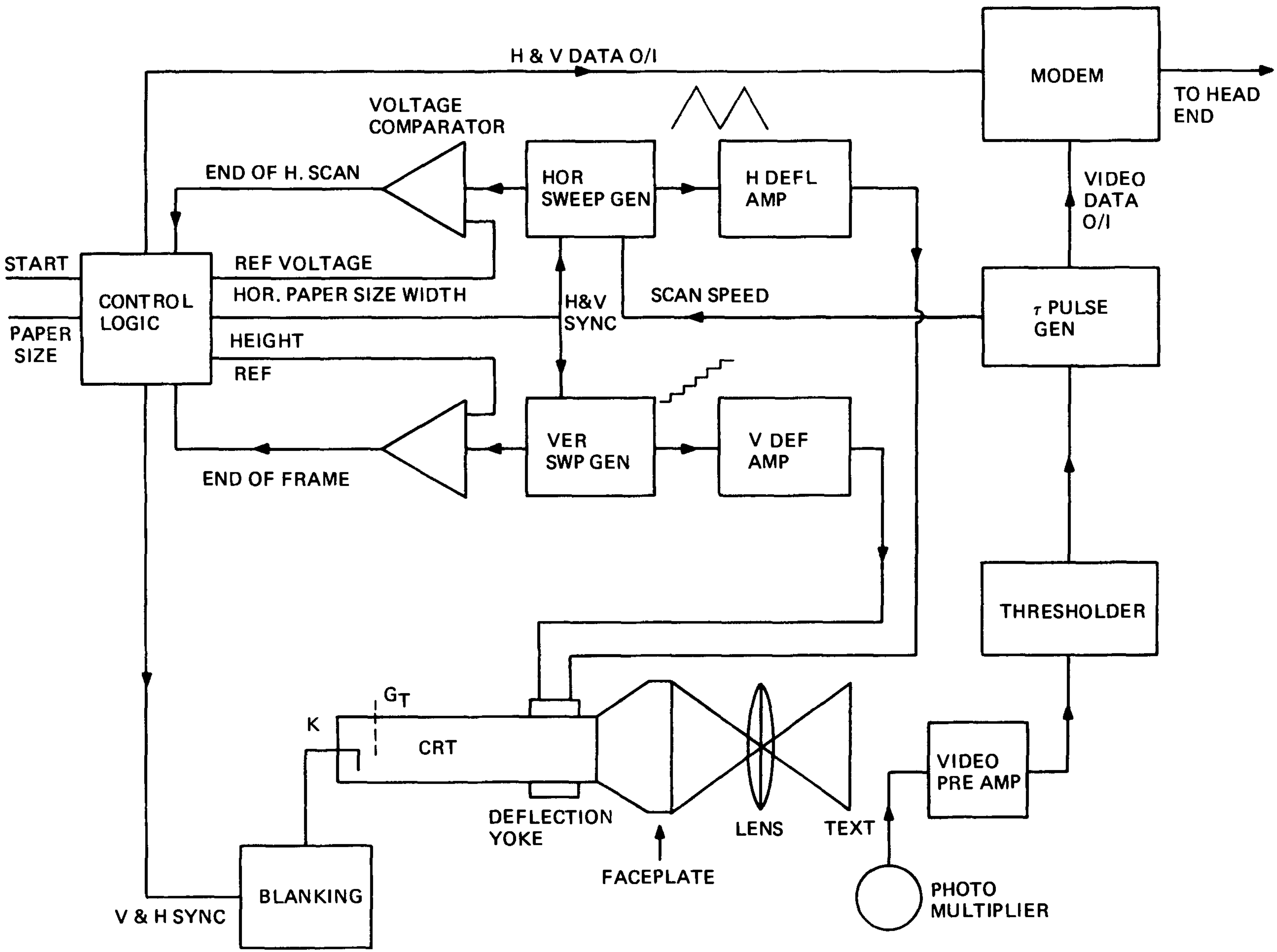


FIGURE 9. BLOCK DIAGRAM FACSIMILE TRANSMITTER

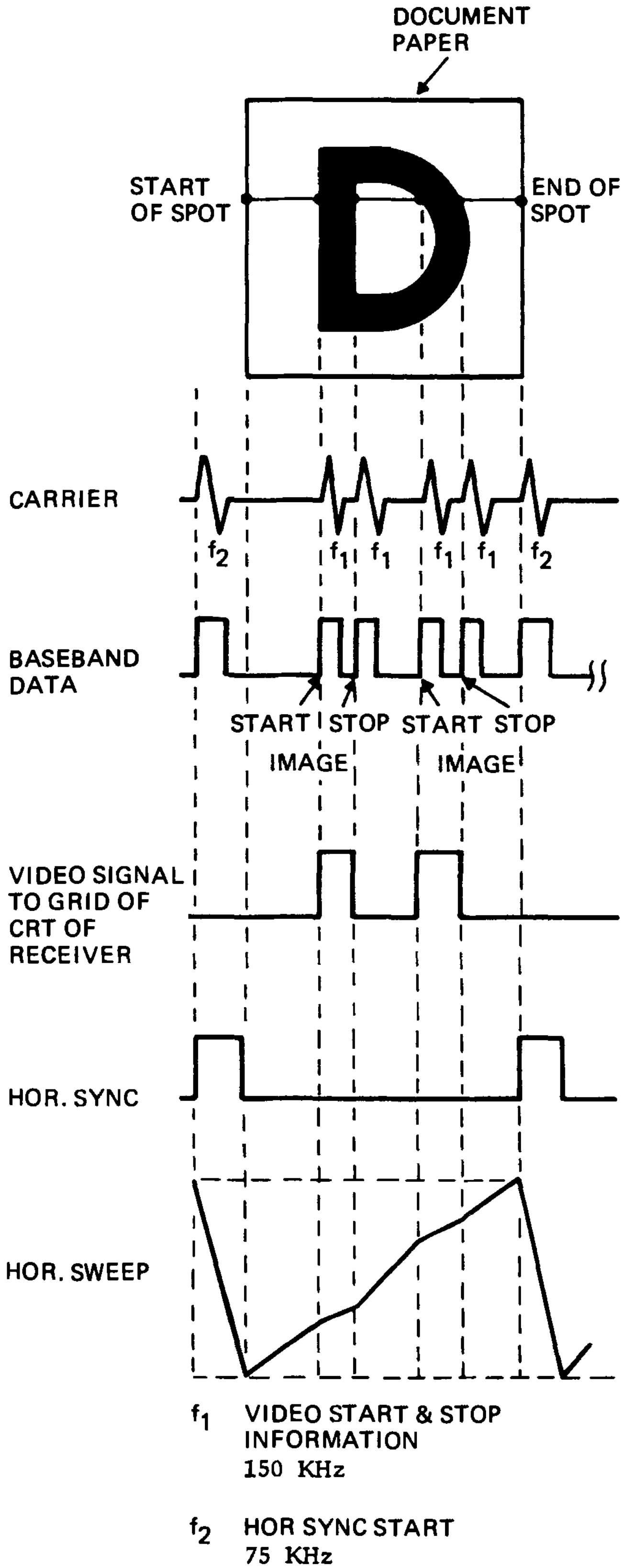


FIGURE 10. TIMING DIAGRAM TRANSCEIVER

2. PRIVATE PUBLIC LINE

As shown in Figure 13. Each subscriber can dial to any other subscriber by sending the "address code" of the party desired to the headend. If the party is free, the appropriate converter will be turned on, in the meantime, a code is returned to the addressor to activate his gate. If the party is busy, the gate remains closed.

3. PARTY LINE

Two or three subscribers can share the same set of two carrier frequencies, but each has a different code to modulate the same carrier. Additional logic is required to inhibit other parties while one of them is in the operating mode.

VI. CONCLUSION

Techniques for subscriber and equipment status monitoring systems, and a facsimile system have been discussed as a first step application of data communications on CATV and expansion into new areas of CATV services.

If the speed of facsimile transmission is not a stringent requirement, a modem can be constructed to interface the presently available equipment with CATV transmission facilities for this new service.

The headend switching unit is by no means complex at this initial phase. The same monitoring technique could be utilized in this type of switching network.

Although today's CATV systems are limited to localized distribution, with available technology, it is reasonable to expect that a national network inter-connecting local CATV systems through microwave or satellite link will be established by the end of the decade. It will then mean much more for proposed services, such as, market survey, audience response, home educational TV, transaction, video phone, facsimile mail, etc., to be available on CATV and to have direct access to each subscriber's home. Before a nationwide system is developed, however, it is natural that the first phase will be the application of data communications within local systems, spurred by the CATV industry's trend.

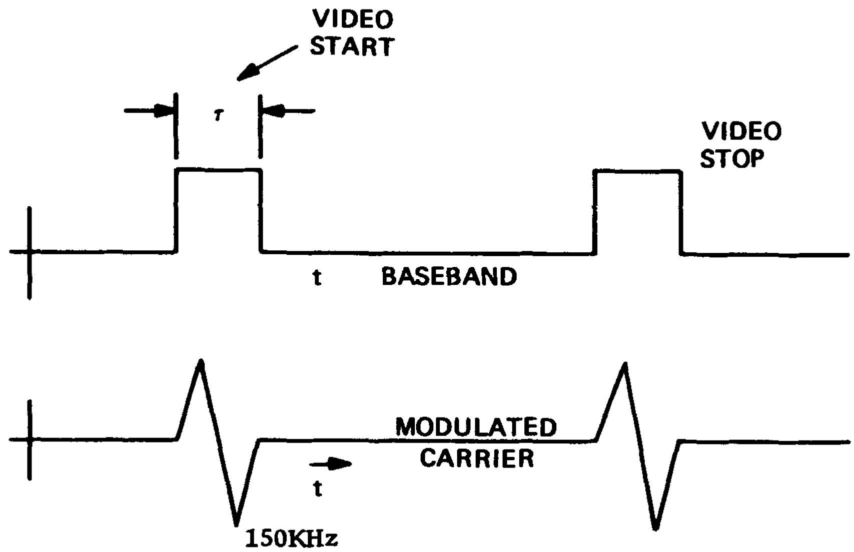


FIGURE 11. VIDEO DATA

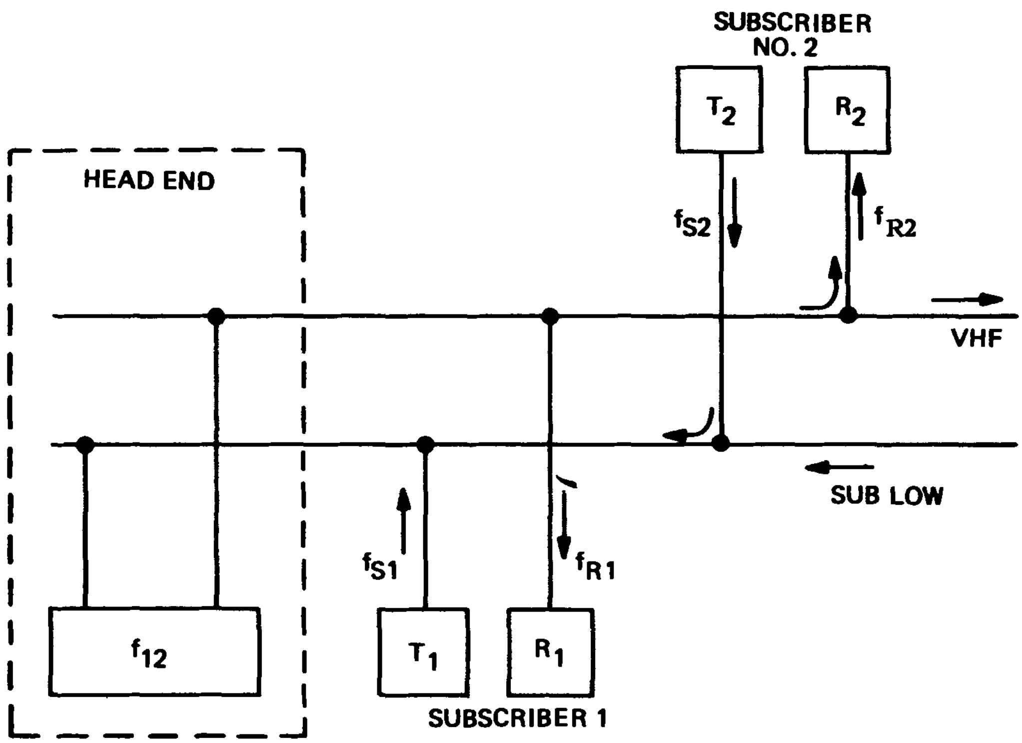
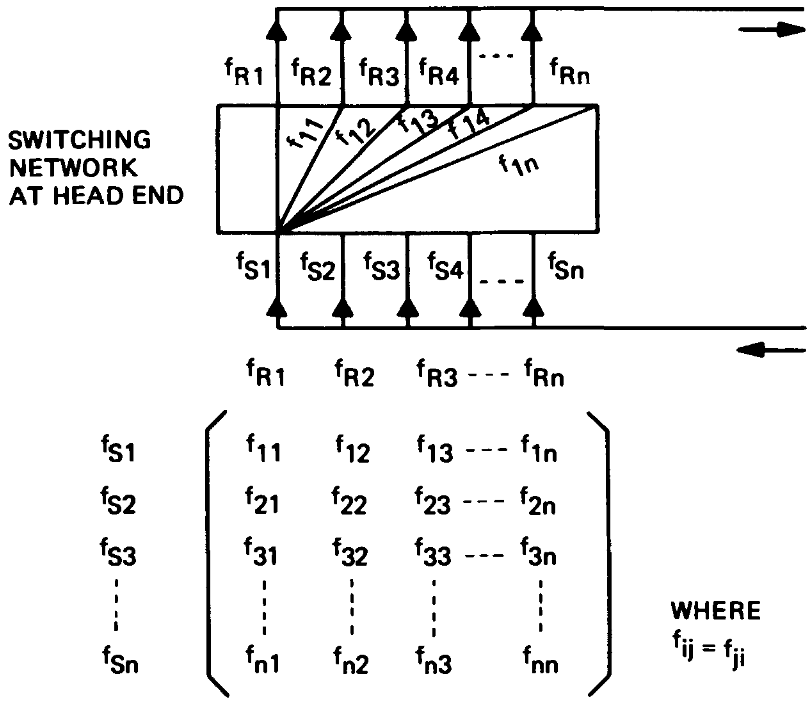


FIGURE 12. PRIVATE LEASE LINE



FOR N SUBSCRIBERS. THERE ARE N + (N-1) CONVERTERS.

FIGURE 13. PUBLIC PRIVATE LINE

ACKNOWLEDGEMENT:

Grateful appreciation is extended to M. J. Rodriguez for his advice and encouragement and to J. Straznicky for his assistance on part of this work.

REFERENCES:

1. D. W. Nast, "Picturephone Transmission", IEEE Internation Convention Digest, 1970.

2. M. J. Rodriguez, "Design Considerations, Bi-Directional CATV Systems", Vikoa, Inc. Technical Publication, 1970.

3. W. R. Bennett, J. R. Davey, "Data Transmission", pp.260-267, McGraw-Hill, 1965.

4. B. Julesz, "A method for Coding Television Signal Based on Edge Detection", Bell System Technical Journal, Vol. 38, pp. 1001-1020, July 1959.

Design of a Low Cost Color Remote Unit for CATV

By Kenneth K. Kaylor
Philips Broadcast Equipment Corp.,
Subsidiary of North American Philips Corporation

Presented at the

1970 NATIONAL CABLE TELEVISION ASSOCIATION

TECHNICAL CONFERENCE

Palmer House, Chicago, Illinois

We believe that cable television can presently fill a need in the home that the networks and the big independent TV stations have long ago forgotten even existed. Each of us wants to feel that we belong to our own local community, that we know the people, the mayor, the police chief or school principal, the minister or the Boy Scout leader. We are interested in what's going on at Little League because our kid pitches on a team. We want to know about the "sit-in" at the high school because our Number One Girl is in the senior class now.

There is no commercial advantage to the TV station for this type of programming, but the CATV operator can supply a channel for this public service at low cost and at a profit too!! Now to do it, he has to be able to run almost any type of audio-visual material available today from 16mm sound movies to picture post cards! He has to be able to interview the horseshoe champ or the beauty queen and run it all back two or three times for those who missed it the first time through!!

He has to generate almost any type of programming from almost any type of location in relatively little time and at a very low cost. One of the most appropriate ways to accommodate these goals is through the use of a self-powered remote unit (Figure #1). However, the criteria for most broadcast style mobile color television cruisers places them well out of the financing capability of even the larger CATV operations. We would like to steal a page from some of the early broadcasters and discuss with you the design of a low cost color remote unit for CATV use.

Many users will have differing requirements depending upon their location, size and climate. We will concentrate on the reasons for the selection of specific equipments and design goals so that individual companies may select those concepts which most nearly meet their needs.

Vehicle Selection:

Vehicle selection is perhaps the most critical aspect in the design of a television remote unit. This is the one subject about which even the uninitiated is an expert because of familiarity with the family car or the local installation truck. A good "rule of thumb" to use when selecting the chassis or van body is that the vehicle should cost about ten percent of the total cost of the finished unit, which includes all equipment and labor for installation. A first look at the problem usually results in a decision to procure an "off-the-shelf" truck, van, or bus because of the obvious financial savings over a "custom designed" body. However attractive, this decision does not insure the lowest total cost when operating costs, maintenance costs, and depreciation of equipment are considered over a five-year period.

Critical design factors to be considered are:

1. Sub-floor space for routing cables and wires
2. Overhead space for lighting
3. Air conditioning ducts and insulation
4. Cable storage and handling
5. Camera storage including tripods, dollys, and pan heads
6. Lighting storage
7. Audio, cue, video and intercom exit ports

8. Heat load
9. Power input and stability
10. Legal restraints
11. Load factors

When all of these factors are considered, the "custom designed" body often turns out to be the least expensive option.

System Design:

The design goals must first be established by the user: what he wants to do, how many men will staff the productions, the quality and complexity of the expected programming, budget requirements, and other pertinent data. Great economies are achieved and many uncertainties overcome if a graphic representation of the vehicle and the total system is derived before purchases start. Five schematics or wiring diagrams and two drawings will establish what is wanted.

#1 -- Video Single Line

The video single line drawing is a much simplified schematic diagram. This typical drawing (Figure 2) indicates the flow of video information throughout the system. It shows all the video sources including live cameras, TV tuners, tape recorders, special effects units, film chains, and test signals. It also shows all the switching points, distribution amplifiers, patching provisions, and monitoring requirements. When such a schematic diagram is generated on paper, it allows the designer and the owner to be sure that no vital elements have been inadvertently left out of the system.

#2 -- Audio Single Line

The audio "single line" drawing (Figure 3) establishes the flow diagram for audio functions. It also shows all audio sources, the audio levels anticipated at each point, monitoring provisions, amplifiers, patch points, filtering provisio echo, special effects, talkback, and audience reinforcement requirements. This diagram allows one to very accurately determine audio costs by indicating each point and piece of equipment in the flow diagram. Cables, switches, patches, and routing requirements are easily defined.

#3 -- Pulse Single Line

Figure 4 identifies the third requirement for successful system design by defining the pulse generation and distribution system. The same form of simplified schematic is employed to indicate the source and distribution of all pulses required to operate each piece of equipment. Since the number of distribution amplifiers are clearly shown, all guess-work is removed from cost analysis.

#4 -- Communications and Control Schematic

The Communications and Control schematic (Figure 5) shows the routing of all telephone, intercom, interphone, remote control and radio circuits which are required to operate the vehicle. Although these circuits are often considered as an afterthought, they are probably the most important aspect of the design. Since program production depends on humans, and humans depend upon communication, a breakdown in the intercom most assuredly means downtime for the show. No shortcuts should be taken in this area of construction.

#5 -- The Power Schematic

The power input, switching arrangements, auxiliary generator, lighting, air conditioning, auxiliary power, and equipment power distribution system are clearly defined in the power schematic (Figure 6) in order that all breakers may be identified and emergency power conditions remedied. Selection of power transformers, meters, and switchers to accommodate various voltages and transmission characteristics must be considered in a properly designed vehicle. Often, for the small NCTA type of vehicle, 110 volt power will be all that is utilized.

#6 -- Equipment Layout

Once the system has been fairly well defined by generation of the five basic schematics previously discussed, the designer and operator can cooperate in making an equipment layout which is much like a house plan. Both a top view (Figure 7) and a side view (Figure 8) are desired. Usually a scale drawing of the intended vehicle is prepared on the drawing board. Cutouts, also made to scale, representing each major piece of equipment, each console, each rack, and piece of furniture may then be fixed on to the vehicle drawing in several trial configurations. When a satisfactory arrangement is conceived, it is good practice to locate a large studio, barn, or loft where the vehicle floor plan can be simulated by using masking tape placed on the floor. Full scale mockup models of equipment, racks, and furniture can be fabricated from cardboard cartons and placed in the tape layout to effectively simulate the design. Some surprises are usually uncovered by this technique with respect to operator convenience and human factors. Upon final selection of the equipment configurations, a finished floor plan and elevation plan can be completed by the draftsman.

#7 -- Equipment Schedule

A specialized equipment schedule must then be prepared which lists each item to be utilized in the vehicle by type number, manufacturer, description, and price. We believe it is most appropriate to subdivide the equipment list into several subsections, including one on live cameras, one on the telecine system, one on the terminal and monitoring equipment, one on the audio equipment, one on the video tape recorder subsystem, one on the lighting subsystem, one to cover the vehicle itself with appendant cable reels, storage compartments, furniture, power system, and accessory items, one on specified test equipments, and one to cover an auxiliary power plant and trailer if desired.

#8 -- Interior Layout

Although a broadcast mobile unit is usually compartmented into three or more separate compartments involving production, engineering, audio and tape recording, the compartment of a small CATV type van will necessarily be limited due to the lack of personnel for fragmented operation. It is desirable to design the unit such that one individual can operate all of the necessary technical equipments from a control location except for strictly manual operations, such as film loading, slide loading, and tape loading of the machines.

We feel it is desirable for the vehicle to be useful for a variety of purposes Therefore, most equipments should be designed so that consoles, recorders, and test equipment can be moved into the compartment or removed from the compartment with a minimum of interference. The Norelco LDH-1 three-tube Plumbicon camera is exceptionally well suited for remote vehicle operation because it is built up from subsystem modules. A simplified version of the camera may be used as a film chain. Addition of selected items will form a complete light-weight viewfinder camera. Thus, all components, electronics modules, tubes, preamplifiers, and controls remain identical for both film and live cameras. This reduces maintenance costs and provides for simple training procedures. In addition, either one of the live cameras can be used as an independent remote unit by installing a separate plug-in color sync generator. Thus, one complete camera system can be removed from the vehicle and used for separate remote operations without interfering with the operation of the vehicle. The sync generator supplied with the camera control unit may be used to provide pulses for all other video requirements in the vehicle. A simple audio mixing system is also built up from dual modules to provide eight in-put channels switchable to sixteen in-puts with two out-puts. One half of the audio mixer can be removed from the vehicle to serve as a mixer amplifier for use with the self-contained color television camera.

The technical console is so arranged that the operator has relatively free access to the video tape recorders and the telecine system as well as to all the panels for control of live cameras, film camera, and switching equipment. All monitors, patch bays, switchers, and operational controls are located on the monitor wall in front of the console.

Telecine System

The telecine complex operates primarily in an automatic or remote control mode. Start, stop, and run facilities are located at the telecine console but are normally remoted to the technical director console. The industry has experienced considerable difficulty in maintaining optical alignment on telecine systems installed in remote vehicles. This is primarily due to vibration and shifting of the optical alignment axis. It has been found desirable to reduce the moment arms of the optical subassemblies to a minimum in order to reduce this effect. When subassemblies are mounted to the floor of the vehicle, even though they are attached to a heavy mounting base, they tend to twist and bend, thus emphasizing this drift situation. To help alleviate this problem, a unitized modular cabinet is constructed with an optical base plate attached to the top and located a very short distance from the optical center line. Each piece of telecine equipment is attached to the base plate and also attached by a mounting pin to a steel plate inserted in the side wall, thus giving two-point suspension to each component.

The telecine system may consist of a 16mm projector, a three in-put one output prism multiplexer, a turret slide projector, a precision optical periscope assembly for transferring the slides to the multiplexer, and a special housing assembly to contain the field lens and automatic light control device. The LDH-1 mobile color television camera will be mounted to the base plate by the use of a wedge adapter in order that the camera may be removed from the telecine system when it is desirable to use it as a third live camera. The system is designed so that the entire console assembly can be removed from the vehicle if desired. Storage space for film, slides, and tape is located beneath the optical base plate. A slide mounted drawer contains rewind arms and film editing equipment. Additional storage space is also provided for spare parts, microphone, and cable storage.

Video Tape Recording Subsystem

Two helical scan one-inch portable color video tape recorders are mounted on benches at the side of the tape compartment. Monitoring functions are built in over the tops of the tape machines. Remote control facilities are remoted to the switching console. The machines should be supplied with editing equipment in order that program content can be edited after field pickup. Storage for tape and supplies is provided beneath the tape recorders in a wooden storage cabinet. Additional storage space is located beneath the recorders for storage of camera cable, power cable, and microphone cable.

Power In-put Cabinet

This cabinet is found in the rear of the compartment and contains metering devices, voltage stabilization equipment, and a power transfer switch.

External and Internal Storage Concepts

A. Camera Storage

Cameras are stored in special cases provided with each camera. The cases will be mounted to clamp pins situated in the floor of the control compartment. Special shelves with springloaded hinges will be mounted on each door of the vehicle. Two doors will be adjusted so that tripods can be strapped one to either door. The rear doors will be arranged so that dollys can be strapped one to either door. Pan heads for the camera will be stored in compartments beneath the tape recorders. Microphones, field cabinet monitors, intercom boxes, and speakers, all will have divided storage space assigned to them.

B. Access Hatches

Water repellant access hatches will be provided on the sides of the vehicle for audio and video in-puts with all necessary cable connectors installed for program and test functions. Audio, video, and communcation jacks are compatible with broadcast standards.

SUMMARY

As previously stated, all users will have a different actual requirement to be concerned with when a mobile color television van is constructed. The object of this paper has been to discuss the design considerations of a small CATV color van which will help to eliminate problems for the user when he places such a van in operation. We have attempted to show several different configurations that might be considered by various CATV operators. It is believed that utilization of the design techniques offered will result in most satisfactory program origination. Complete drawings and specifications for several sizes of CATV vans are available from various manufacturers. Philips Broadcast Equipment Corporation is pleased to provide copies of sample designs for your evaluation. Please contact us at any time to arrange for your free copy.

Thank you very much for your kind attention.

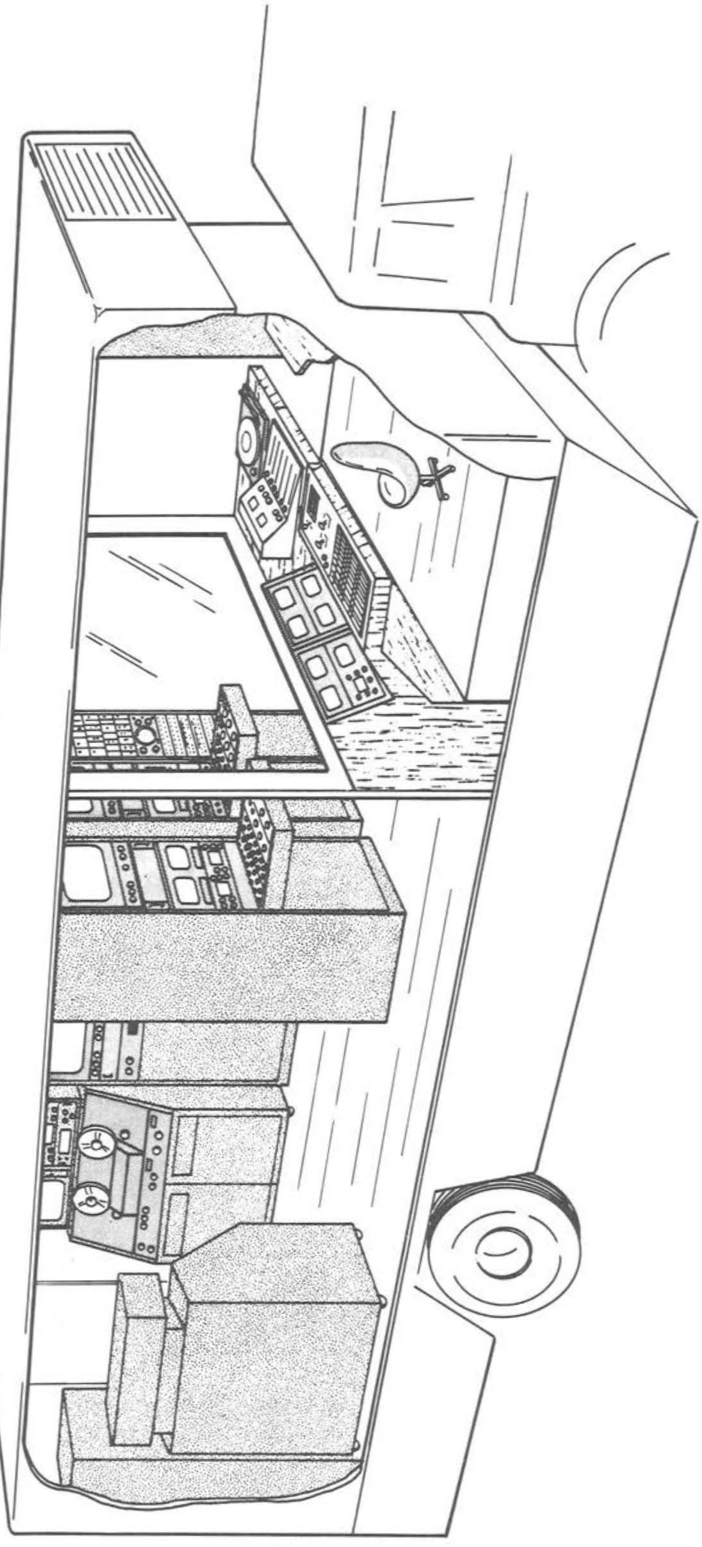


Figure. 1 24 FT MOBILE COLOR UNIT
PHILIPS BROADCAST EQUIP. CORP.

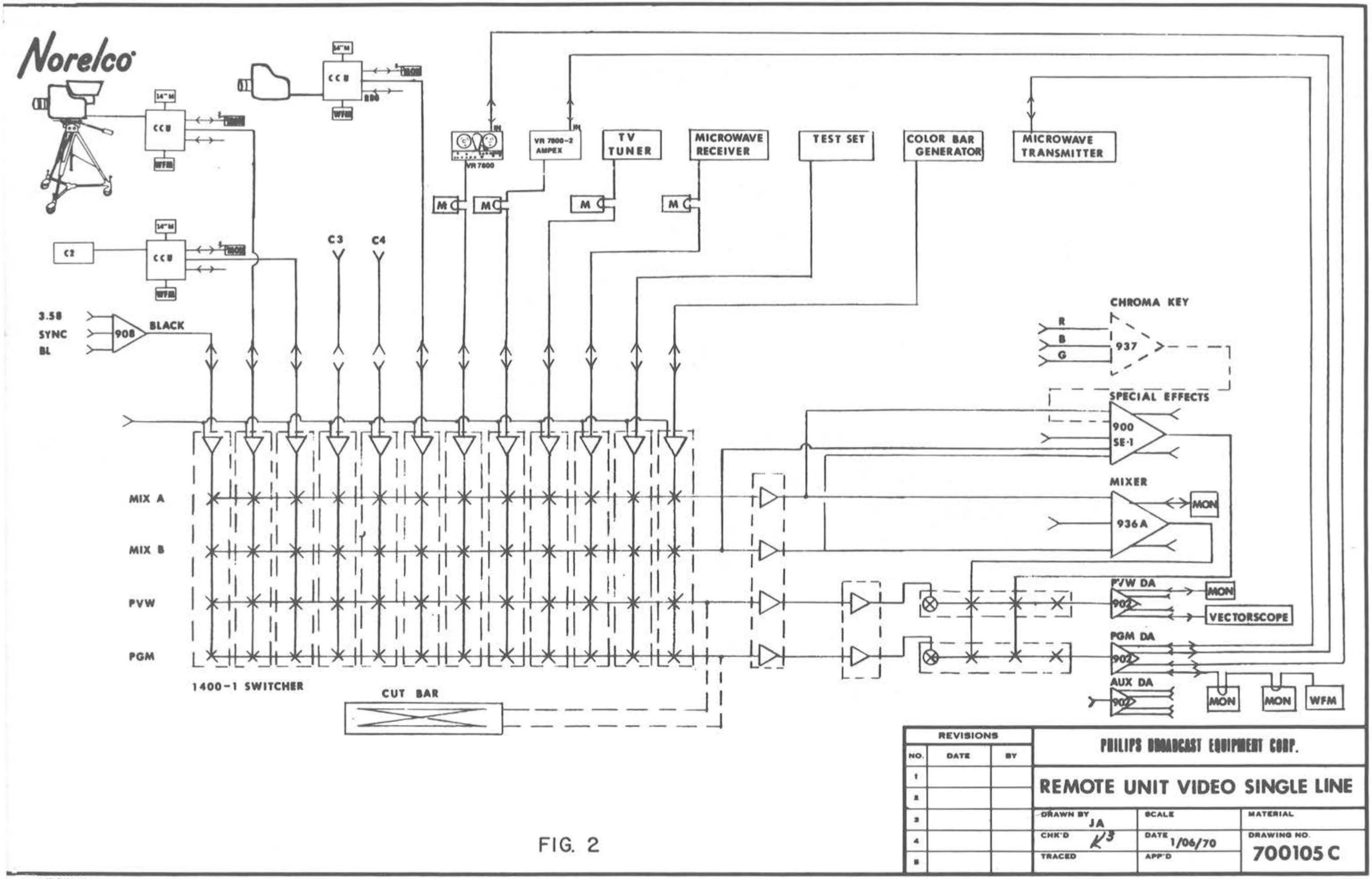


FIG. 2

1x17 (1850-64)

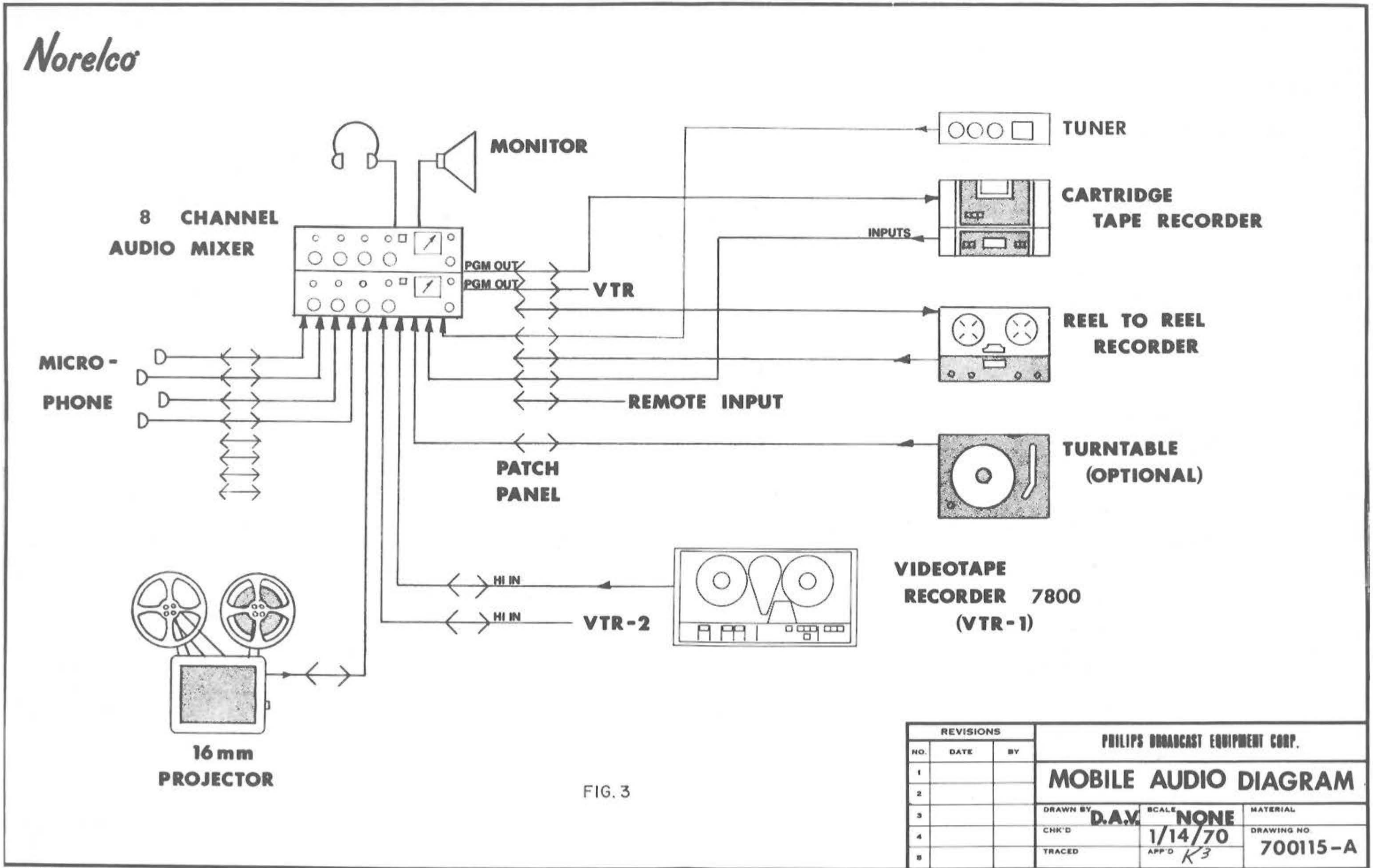

Norelco
MONITOR
8 CHANNEL
AUDIO MIXER
PGM OUT
PGM OUT
VTR
TUNER
CARTRIDGE
TAPE RECORDER
INPUTS
REEL TO REEL
RECORDER
MICRO-
PHONE
REMOTE INPUT
TURNTABLE
(OPTIONAL)
PATCH
PANEL
VIDEOTAPE
RECORDER 7800
(VTR-1)
HI IN
HI IN
VTR-2
16 mm
PROJECTOR
REVISIONS
NO.
DATE
BY
1
2
3
4
5
PHILIPS BROADCAST EQUIPMENT CORP.
MOBILE AUDIO DIAGRAM
DRAWN BY D.A.V.
SCALE NONE
MATERIAL
CHK'D
1/14/70
DRAWING NO
TRACED
APP'D K3
700115-A

FIG. 3

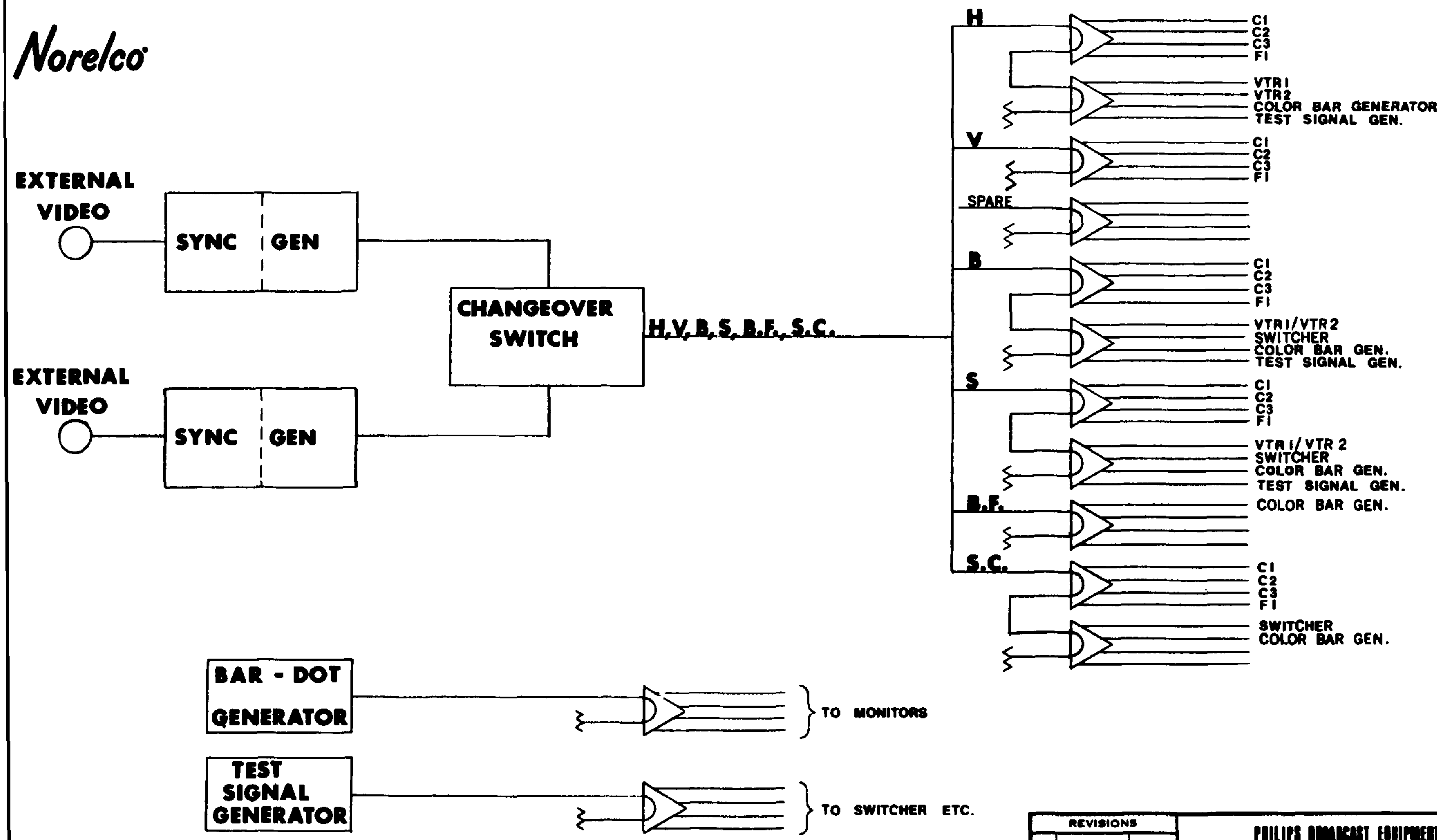


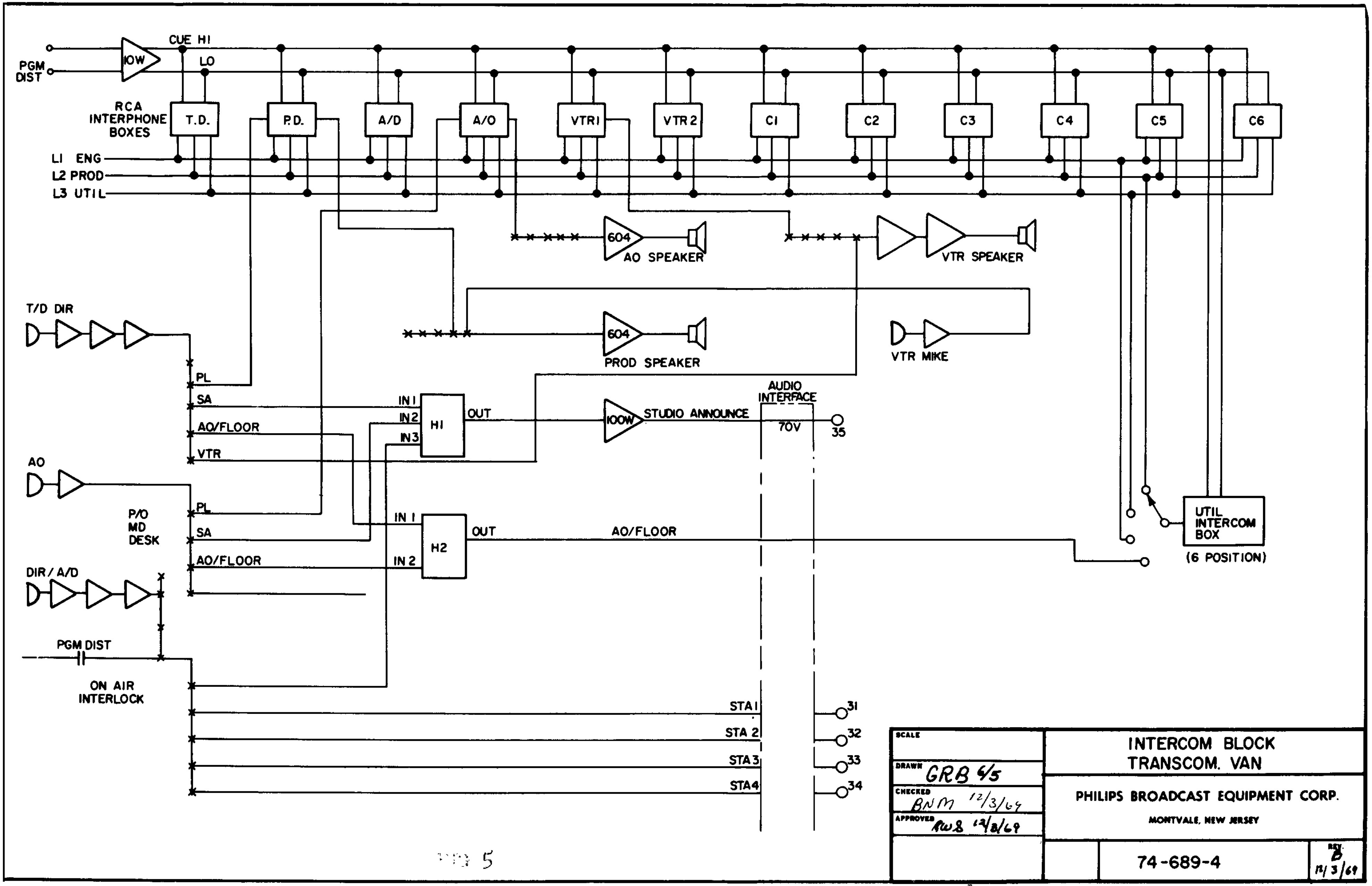
FIG. 4

REVISIONS		
NO.	DATE	BY
1		
2		
3		
4		
5		

PHILIPS BROADCAST EQUIPMENT CORP.

MOBILE PULSE DIAGRAM

DRAWN BY D.A.V.	SCALE NONE	MATERIAL	
CHK'D	1-14-70	DRAWING NO.	700114-A
TRACED	APP'D K3		



PGM DIST
10W
CUE HI
LO
RCA INTERPHONE BOXES
T.D.
P.D.
A/D
A/O
VTR1
VTR2
C1
C2
C3
C4
C5
C6
L1 ENG
L2 PROD
L3 UTIL
604
AO SPEAKER
VTR SPEAKER
T/D DIR
604
PROD SPEAKER
VTR MIKE
PL
SA
AO/FLOOR
VTR
IN1
IN2
IN3
H1
OUT
100W
STUDIO ANNOUNCE
AUDIO INTERFACE
70V
35
AO
PL
SA
AO/FLOOR
P/O MD DESK
IN 1
IN 2
H2
OUT
AO/FLOOR
UTIL INTERCOM BOX
(6 POSITION)
DIR/A/D
PGM DIST
ON AIR INTERLOCK
STA1
STA 2
STA3
STA4
31
32
33
34
SCALE
DRAWN GRB 6/5
CHECKED BNM 12/3/69
APPROVED RWS 12/8/69
INTERCOM BLOCK TRANSCOM. VAN
PHILIPS BROADCAST EQUIPMENT CORP.
MONTVALE, NEW JERSEY
74-689-4
REV. B 12/3/69

344

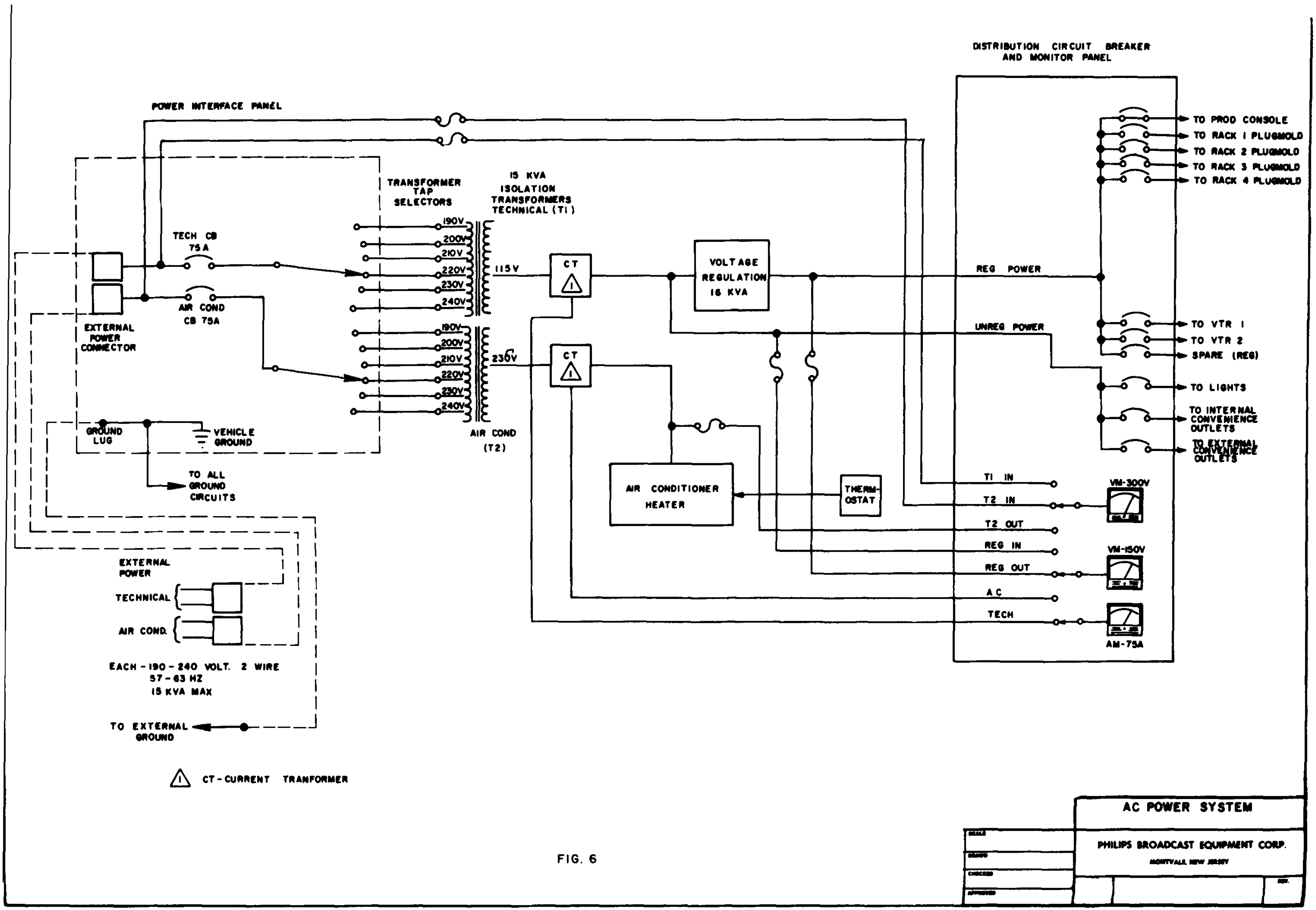

POWER INTERFACE PANEL
DISTRIBUTION CIRCUIT BREAKER AND MONITOR PANEL
TRANSFORMER TAP SELECTORS
15 KVA ISOLATION TRANSFORMERS TECHNICAL (T1)
TECH CB 75A
AIR COND CB 75A
EXTERNAL POWER CONNECTOR
190V
200V
210V
220V
230V
240V
115V
230V
AIR COND (T2)
CT
VOLTAGE REGULATION 16 KVA
REG POWER
UNREG POWER
TO PROD CONSOLE
TO RACK 1 PLUGMOLD
TO RACK 2 PLUGMOLD
TO RACK 3 PLUGMOLD
TO RACK 4 PLUGMOLD
TO VTR 1
TO VTR 2
SPARE (REG)
TO LIGHTS
TO INTERNAL CONVENIENCE OUTLETS
TO EXTERNAL CONVENIENCE OUTLETS
GROUND LUG
VEHICLE GROUND
TO ALL GROUND CIRCUITS
AIR CONDITIONER HEATER
THERM-OSTAT
T1 IN
T2 IN
T2 OUT
REG IN
REG OUT
A C
TECH
VM-300V
VM-150V
AM-75A
EXTERNAL POWER
TECHNICAL
AIR COND.
EACH - 190 - 240 VOLT. 2 WIRE
57 - 63 HZ
15 KVA MAX
TO EXTERNAL GROUND
CT - CURRENT TRANFORMER
AC POWER SYSTEM
PHILIPS BROADCAST EQUIPMENT CORP.
MONTVALE, NEW JERSEY

FIG. 6

345

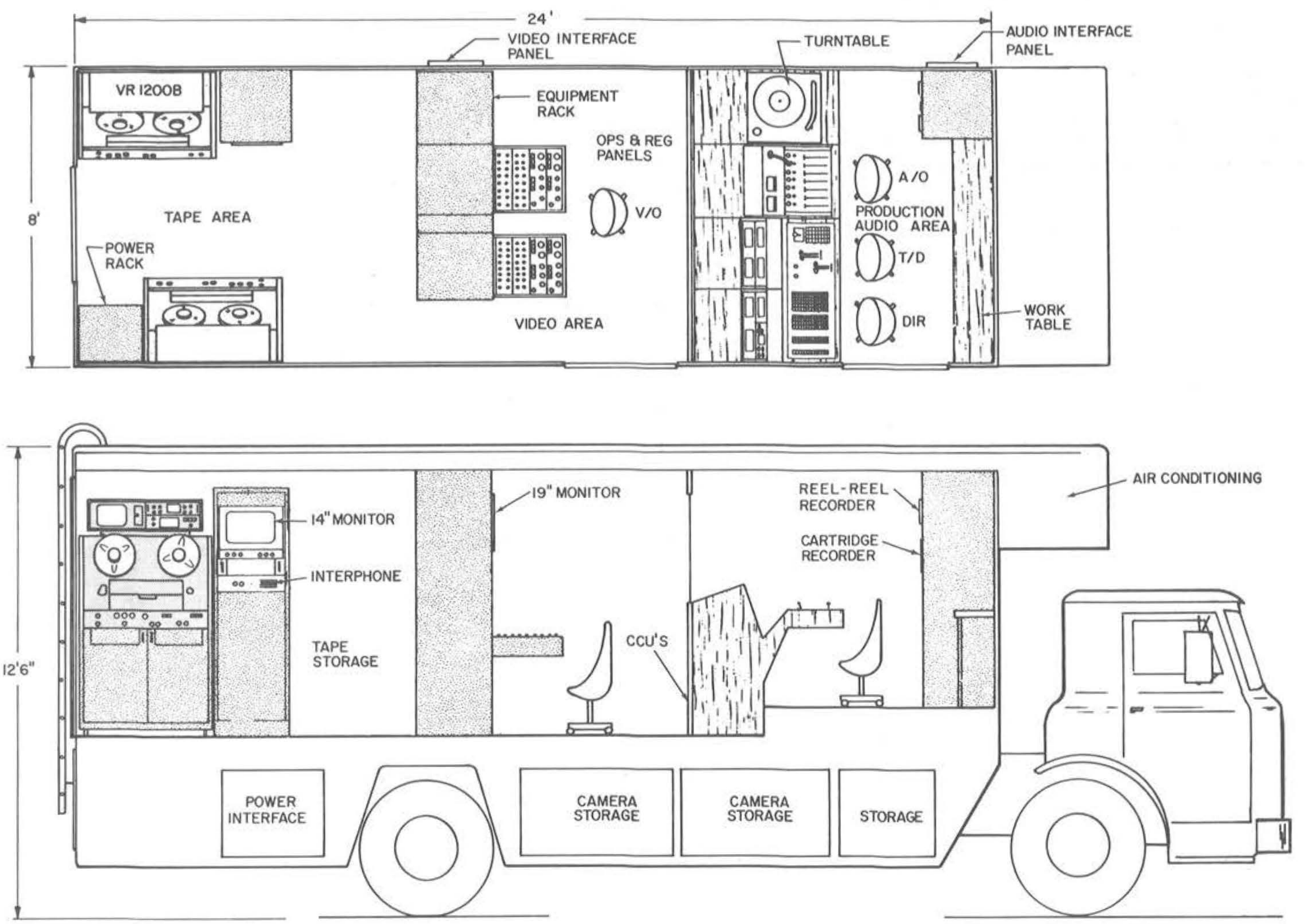


Figure 7 24 FT MOBILE COLOR UNIT
PHILIPS BROADCAST EQUIP. CORP.

Norelco

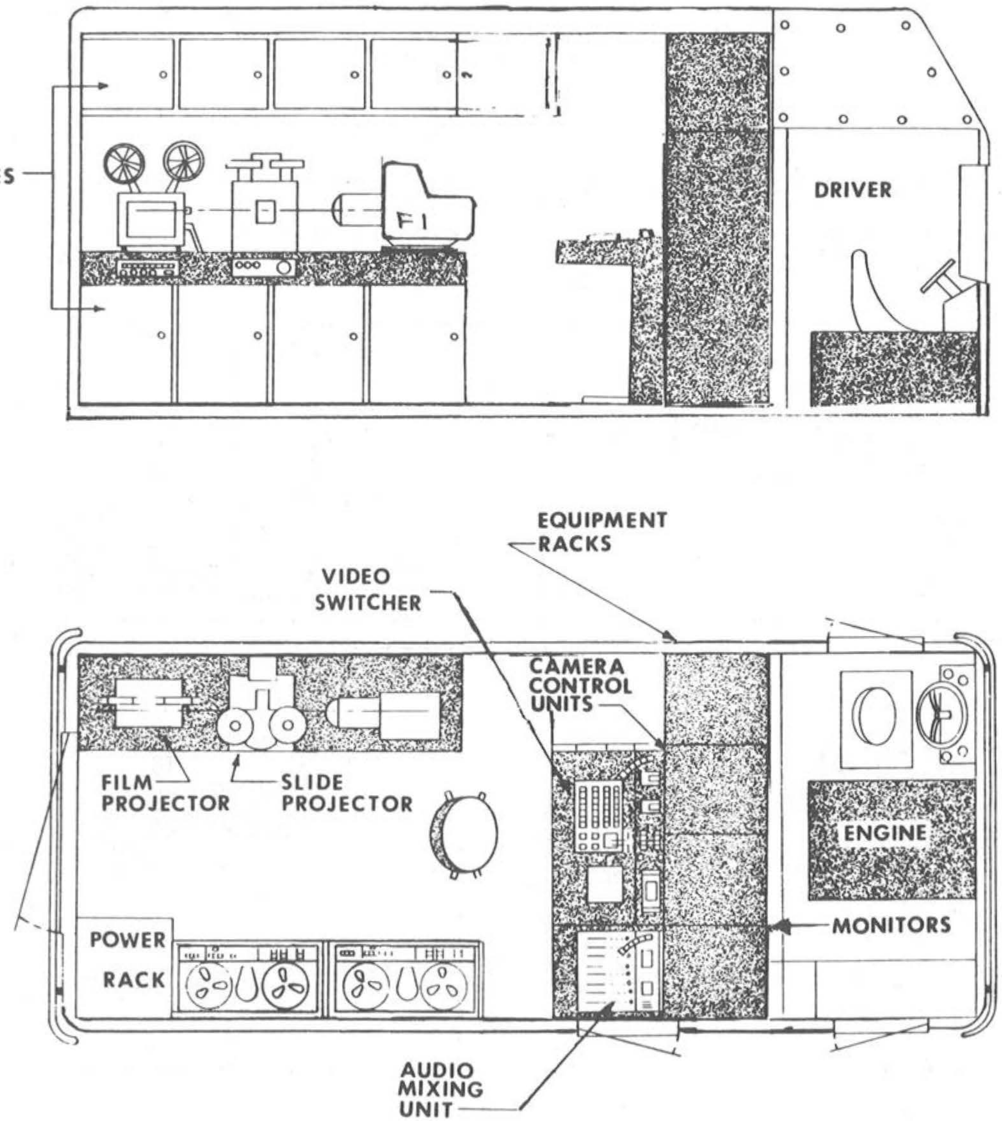


FIG. 8

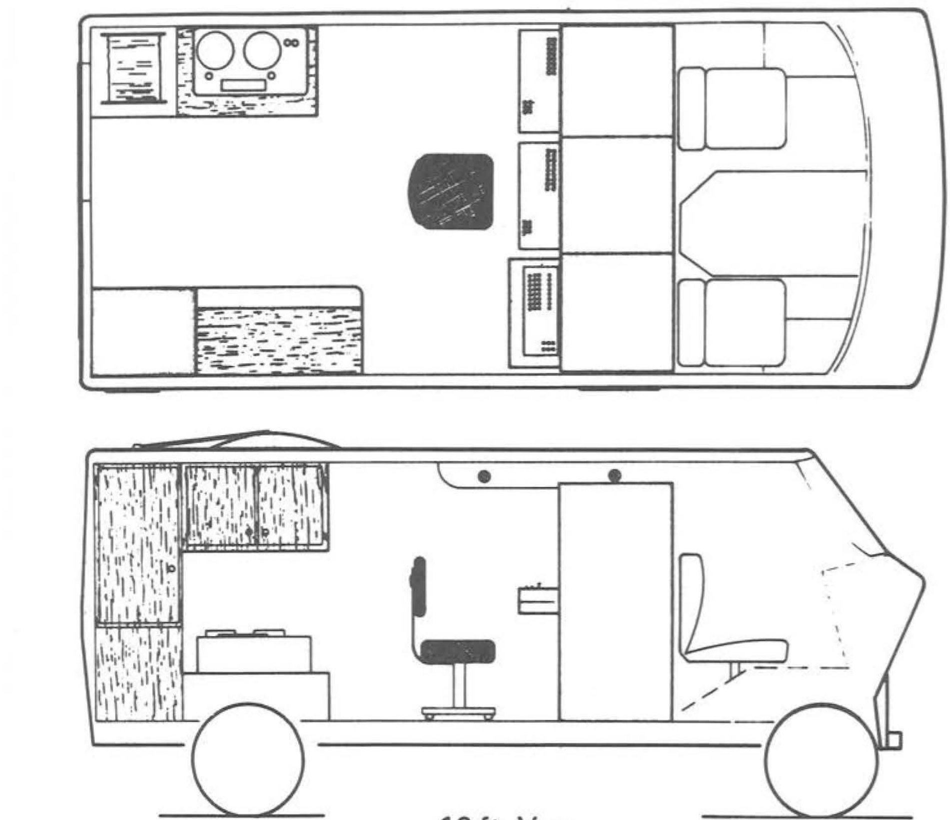
CATV MOBILE VAN

PHILIPS BROADCAST EQUIPMENT CORP.

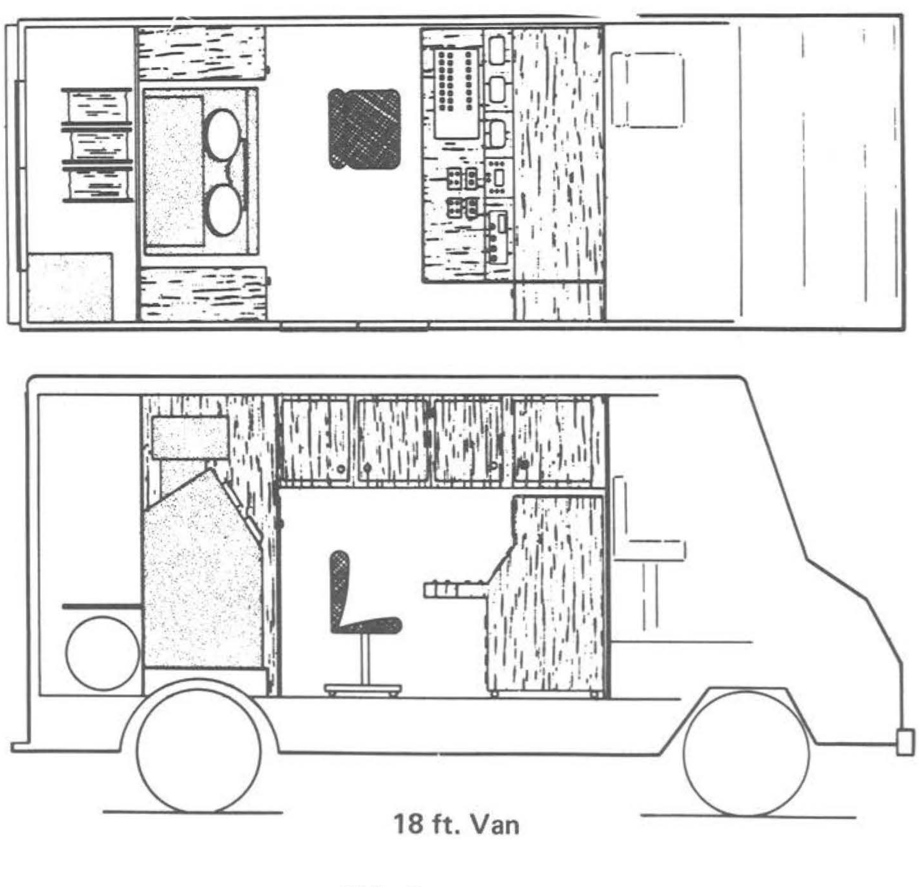
MONTVALE, NEW JERSEY

REV.

Norelco® custom



16 ft. Van



18 ft. Van

FIG. 9

DESIGN OF THE LOCAL ORIGINATION STUDIO

Larry E. Nelson
National T. V. Sales Manager
Century-Strand Inc.

Due to the recent rulings by the Federal Communications Commission, many CATV systems will now be required to provide local origination programming.

Because of these requirements, a studio facility to produce this programming must be provided. In many cases, a studio facility is already in use or is in the process of being converted from an existing facility.

In some instances, a complete studio must be designed and built from scratch. In order to either build a completely new studio or to convert an existing facility, there are many major areas which must be delt with. These areas of concern are such items as physical size, electrical power, and air conditioning from the construction stand point and fixture suspension, fixtures and dimming systems from the production standpoint.

Even though the end use of the studio is for production purposes, many engineering considerations must be studied so that the studio will truly function as an efficient production center.

The first areas of discussion will be concerned with the construction area of the studio design.

PHYSICAL REQUIREMENTS

In the design of a completely new studio facility, one of the most important single items is physical size. The physical size of the studio determines the amount of production that can be done in this studio. The smallest sized studio that will still allow a certain amount of production flexibility is approximately 20' x 30' or 600 sq. ft. A studio of this size will allow two or three individual areas to be used without having to remove one set area in order to utilize the other area. It will also allow for permanent sets that are used day after day.

When designing your studio, ceiling height is also a very important consideration. Twelve to 14 feet of ceiling height is optimum for production studios up to 2400 sq. ft. If the ceiling height is any lower than 10', it is very difficult to properly light the subjects due to the poor angles. Any height over 14 ft. would require higher wattage fixtures to give sufficient ft. candle intensity.

In most cases, the 12 ft. ceiling is very easy to obtain, but if the ceiling height is restricted to less than this certain lighting tricks can be utilized to remedy this situation.

If at all possible, the studio should have an entrance to a roadway or alley with doors sufficiently sized so that an automobile can be driven into the studio. This will allow production of automobile commercials and other revenue producing sidelines. Under no circumstances should the studio have any glass windows that would allow outside light to enter the studio.

A larger studio, of course, will provide more production capability. The average CATV origination studio is in the area of 20 X 30 feet.

ELECTRICAL POWER

A very important consideration, and sometimes a limiting factor for studio design, is electrical power. Approximately 60% of electrical power for a studio is used for the lighting. Since there is a direct relationship with power required and square footage, sufficient power must be made available to adequately light the studio. As a rule of thumb, 60W per sq. ft. of studio area is required. This is for color but should also be used in black-and-white conditions since eventually color will most likely be used. As an example, for a 20 X 30 studio, 36,000 watts of electrical power is required for lighting purposes alone. No matter how large the production area, if there is not sufficient power to properly light it, the production area is useless.

It should be noted that in all cases, such services as electrical power and air conditioning should be sized so as to take into account ultimate requirements. It is very costly to increase these services at a later date.

The most desirable AC power service for studio purposes is 3 phase 4 wire, 120/208V AC 60 cycles. This type of power service is available in most areas and is normally the most economical power service that can be provided for the studio. Since there are three phases, each carrying equal loads, each phase is only required to carry one-third of your total connected load. In the above example, the 36KW service required for the 20 X 30 studio would be 300 AMPS total. By using a 3 phase 4 wire power feed, each of the phases would only have to carry 100 AMPS. This will keep the size of feed wire and therefore, cost, to a minimum.

Another very common power feed that is available in older facilities is single phase, 3 wire service more commonly known as 120/240. This type of service uses 3 wires each of which carry ½ of the total load. In the avove example, the 36KW load will be broken down into two feeders of 150 Amps each.

It is most desirable to have a separate service provided for the studio lighting loads. In this way, there will be no fluctuation in your electrical equipment when the heavy loads of the lighting equipment are turned on.

AIR CONDITIONING

Since almost all of the electrical power used for lighting is converted to heat, sufficient air conditioning must be available to keep the studio within a reasonable temperature range that will not advdrsly effect other studio equipment or personnel. Due to the inefficiencies of lamps, either quartz, iodine or standard incandescent, all power applied must be considered as heat.

It takes approximately .14 tons of air conditioning for every 1KW (1000 Watts) of lighting. These figures may vary from one geographical location to another, but may be used generally as a starting point. For the 20 x 30 studio previously discussed, the air conditioning for the studio lighting alone would be five tons. It must be noted that this figure of .14 ton per Kilowatt is for the lighting fixtures only, and does not take into consideration any other air conditioning requirements such as electrical equipment, talent, or general illumination.

Now that the construction area has been discussed, let us proceed to the design of the actual production facilities. The first section of the design has been aimed at the engineering staff. The remainder of the discussion is concerned with the production staff.

STUDIO FIXTURES

In order to do any production work, light is required. If the production is to be done outdoors on a sunny day the light is there and free for the taking. On the other hand, since most work is done indoors, artifical lighting must be provided.

They types of lights, correctly called fixtures, as well as the number of fixtures is important. There are four basic types of illumination required. They are:

> Key Light, which is the main apparent source of illumination on the set. It is the job of this light to highlight or "key" the subject.

<u>Back Light</u>, which is used to give your subject depth and dimension by separating the subject from the background.

<u>Base/Fill Light</u>, which is used to provide an overall light level and fill the shadows cast by the key light.

<u>Set Light</u>, which is used strictly for background illumination.

For most simple one camera, one subject sets, one each of the above lights is required. Fortunately, only two different styles of fixtures are required. A 750-1000 Watt quartz fresnel is the most commonly used studio fixture for both back and key lighting purposes. On the other hand, a quartz scoop is most commonly used studio fixture for base/fill and set lighting.

<u>FIXTURE MOUNTING</u>

In order to use the lighting fixtures in the studio, some method of fixture mounting must be used. For temporary usage in small set areas, floor stands provide a very convenient method for fixture mounting. The fixtures that are mounted on these stands are then plugged into the wall circuits in the studio. This method allows the fixtures to be rapidly moved from one location to another. This method is undesirable due to the fact that these stands take up much of the available floor space on the studio floor and therefore, restrict camera movement. There is also great danger of knocking over these stand mounted fixtures.

It is more desirable to use some sort of ceiling mounted fixture support to get these fixtures off the studio floor and out of the way of the cameras and other equipment. The most desirable way to overhead mount the lighting fixtures is through the use of a pipe grid. This pipe grid is made up of sections of 1-½" I.D. black pipe and is hung from the ceiling at the desired height. As was previously mentioned, approximately 60W per square ft. are required to provide sufficient lighting for the studio. Because of this, a spacing of 4 ft. on center is used. To light the set area, the fixture is mounted to this pipe grid by the use of a fixture mounted C-clamp and is then plugged in to the electrical outlet. These electrical outlets which will be discussed in detail later, can either be ceiling mounted or can be around the perimeter of the studio on the wall. In the later case, extension cables must be provided to bring the electrical power from the floor up to the mounted fixture.

If the grid configuration is not desired, then pipes running in only one direction can be utilized. In this case, these pipes should be mounted on 4 foot centers with the pipes perpendicular to the longest studio walls. In place of pipe, uni-strut can be utilized. This is a three-sided steel channel that is mounted from the ceiling. Special mounting brackets are required to mount the fixtures to the uni-strut. The C-clamp is a much faster and more secure way to mount the fixtures to the pipe and is therefore a more desirable material.

Another type of fixture mounting hardware is the Century Strand system called MobilRail. MobilRail is extruded aluminum I-rail that is mounted to the ceiling. This system is designed so that the rail can be moved from one studio area to another from the floor. This method of fixture mounting is more expensive than the pipe grid method but its flexibility in a small studio more than justifies the additional cost.

ELECTRICAL DISTRIBUTION

Once the fixture is mounted, some method of getting electrical power to it must be designed in order to use the fixtures. If wall circuits are to be utilized, then a large number of extension cables are required. This method of power distribution is very cumbersome and sloppy and is undesirable for all but the smallest studio area. A more desirable way for power distribution is to mount the receptacles into the ceiling directly above the grid. In this way, a minimum amount of extra power cable is required.

In all cases, one 20 AMP outlet should provide for every 16 sq. ft. of studio area due to the 60 Watts/sq. foot requirements. The most desirable way to provide electrical distribution is through the use of plugging strips with pigtails. These plugging strips, usually constructed of 4" x 4" conduit, have 18 to 36 inch pigtails. The fixtures are then plugged into these pigtails for electrical power. These plugging strips are designed to be mounted directly above the pipes or hung from the ceiling. If the plugging strips are ceiling mounted, the pipes or grid can be directly mounted to strips. Through the use of 36" pigtails, one plugging strip every 8' can provide sufficient electrical circuits for the studio.

All of the circuits that are installed in the television studio should have a current rating of at least 20 Amps. In addition, one 50 Amp circuit should be provided for every ten-20 Amp circuits. These 50 Amp circuits would be used to provide power for several cyclorama lights.

The ceiling mounted circuits or plugging strip pigtails must be supplied with electrical power. This electrical power can come via circuit breaker panel or from a dimming system. In the case of a dimming system, all of the overhead and wall circuits would terminate in a cross-connection or patch panel which would then be fed in turn by a dimmer system. The dimming system capacity would depend upon studio size and requirements. Fig. 1 shows the central console for a small electronic dimming system, the Century Strand Edkotron.

BACKGROUND AND SETS

In all studio installations, it is desirable to have a variety of backgrounds. A very common background is a cyclorama curtain. This curtail is a muslin curtain that is mounted on a roller track so that it can be moved into position anywhere in the studio. This curtain can be lighted with special lights with a colored gelatin in front of them to give any desired color. In this way, the single cyclorama curtain can be made any number of colors for day to day variety. The color of the cyclorama curtain should be an off-white or beige so that full color saturation may be realized.

If background flats are to be used, they should be a fairly light neutral color so that they will not absorb too much of the lighting.

For reference purposes, two complete studio lighting packages are included. Fig. 2 is a temporary lighting package for a 10 x 15 area. This package would be designed to light a temporary set area, such as an office. The fixtures are designed for stand mounting.

Fig. 3 is a complete lighting package including electrical distribution and dimming control. This package will completely outfit the 20 x 30 foot studio.

Both packages feature quartz lights.

SUMMARY

It must be kept in mind in the design of the studio that many of the decisions made will be based on budget requirements. It should be remembered certain items cannot be economically added later and should therefore be treated as a one-time cost whereas some items such as fixtures, can be implemented at any time.

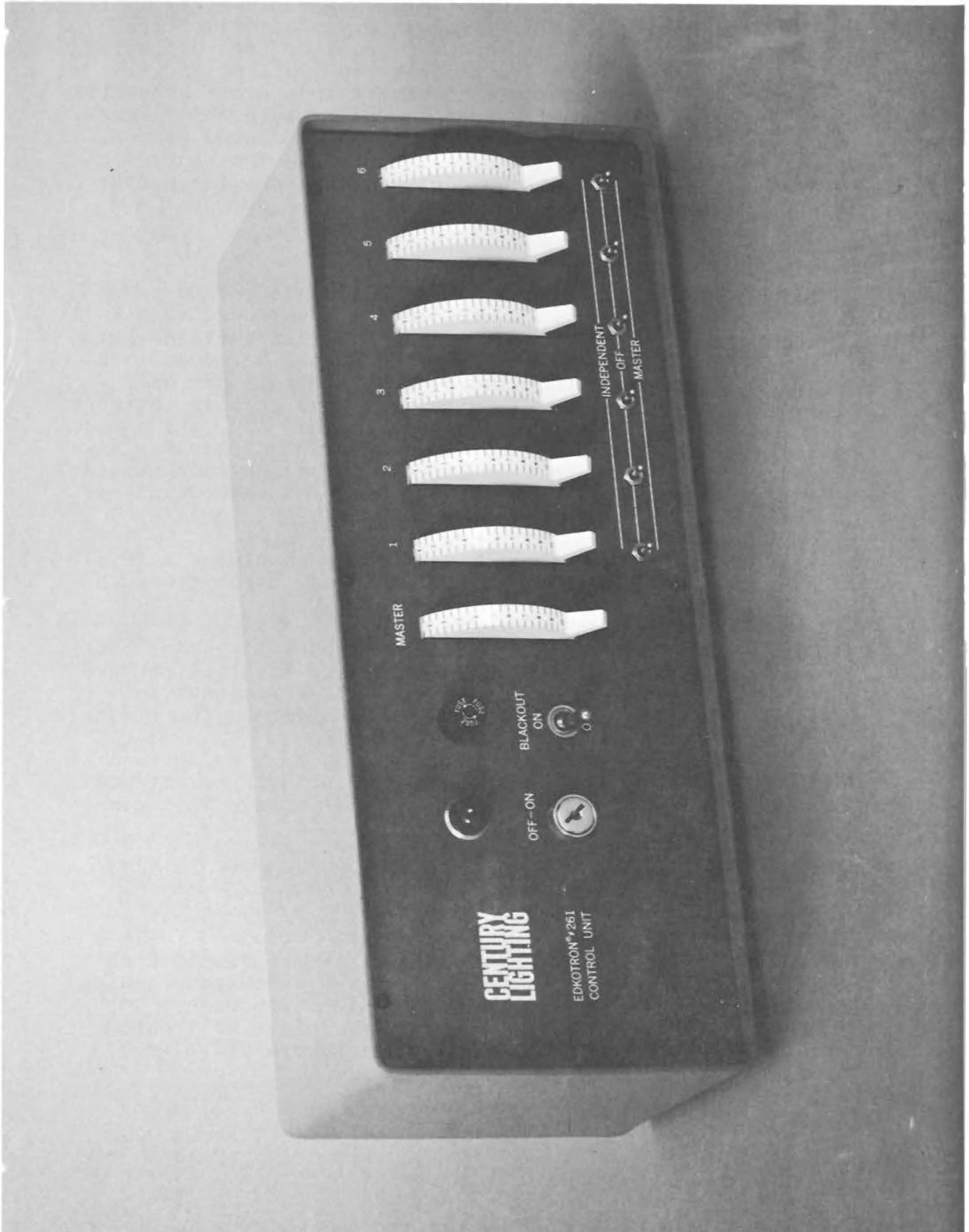

CENTURY LIGHTING
EDKOTRON® #261
CONTROL UNIT
OFF—ON
BLACKOUT
ON
MASTER
1
2
3
4
5
6
INDEPENDENT
OFF
MASTER

FIG. 2

10' x 15' "PORTA-PAC" STUDIO PACKAGE

KEY AND BACK LIGHTS

Quant.	Century-Strand Cat. No.	Description
4	33420GR	6" 750W Quartz Fresnel with stand bracket
4	13220	4-Way Barndoor
4	BTP	750W Quartz Lamp, 3200°K
4	25RCCGR	25' Extension Cable
BASE/FILL LIGHT		
4	4271GR	14" 1000W Quartz Scoop with stand bracket
4	11160	Diffusion/Gel Frame
4	EGK	1000W Quartz Lamp, Frosted, 3200°K
4	25RCCGR	25' Extension Cable
MOUNTING EQUIPMENT		
6	C691035	Channel Leg Stands
2	3256	"Grabber-Grip"
DIMMING SYSTEM (OPTIONAL)		
1	280GR	6-2kw Edkotron Dimmer Pack
1	281	Control Unit
1	283	25' Cable

FIG. 3

20' x 30' STUDIO PACKAGE

BACK AND KEY LIGHTS

10	33420GP	Century-Strand 6" 750W Quartz Fresnel
10	13220	4-Way Barndoor, Adjustable
10	11080	Diffuser/Gel Frame
10	BTP	750W Quartz Lamp, 3200°K
4	14410	7' Pantograph
4	10RCCGP	10' Extention Cable
4	3413GP	Century-Strand 8" 1000/2000W Quartz Fresnel
4	13230	4-Way Barndoor, Adjustable
4	11100	Diffuser/Gel Frame
4	BVV	1000W Quartz Lamp, 3200°K
2	14410	7' Pantograph
2	10RCCGP	10' Extension Cable

FIG. 3 (cont.)

Quant.	Century-Strand Cat. No.	Description
BASE AND FILL LIGHTS		
4	4271GP	Century-Strand 14" 1000W Quartz Scoop
8	4291GP	Century-Strand 14" 1000W Quartz Scoop, Focusing
12	11160	Diffuser/Gel Frames
12	EGK	1000W Frosted Quartz Lamp, 3200°K
6	14410	7' Pantograph
6	10RCCGP	10' Extension Cable
EFFECTS LIGHT		
1	2324GP	Century-Strand 750W Quartz Pattern Projector with Pattern Holder, Pattern Set and Gel Holder
1	EGF	750W Quartz Lamp, 3100°K
6	MS1000GP	Century-Strand 1000W Quartz Set Lights
6	FHM	1000W Quartz Lamp, Frosted
SPECIAL EQUIPMENT		
1	SGD	Spun Glass Diffusion Material
3	14340	Adjustable Stands, 5' - 8'
3	25RCCGP	25' Extension Cable
DISTRIBUTION EQUIPMENT		
6	6312-5-2P	12' Plugging Strip with 5-18" Pigtails terminated in 3 Pin Grounding Connectors
4	64620-20GP	Wall Box with 2-18" Pigtails terminated in 3 pin grounding connectors
CONTROL EQUIPMENT		
1	C70003	Dimmer Rack containing: 6 - 6KW Century-Strand Dimmers 1 - 100 Amp 3 pole Main Circuit Breaker Hanging Cord Patch Panel with: 38-20 Amp Load Cords and Breakers 42-50A Saf-T-Jacks, 6 per Dimmer & 6 Jacks/1-50A Hot Circuit

FIG 3 (cont.)

Remote Control Consolette
Containing:
6 - Controllers (1/Dimmer)
6 - Master-Off-Independent
  Switches
1 - Master Control
1 - System Key Switch
1 - System Blackout Switch
1 - 25' Control Cable

Norelco

DISCUSSION

Mr. Robert Loos: Thank you, Larry. Do we have any questions?

Question: I have a question. What happens to the color temperature as you vary the lighting?

Mr. Nelson: The color temperature varies, of course, with your applied voltage. The lower the voltage the lower the color temperature, but I think you'll find that in television operations such as your big networks, the television color cameras aren't really as sensitive to color temperature as they were a few years ago. There are certain lights that should not be dimmed such as your key light and your film base light, but your background lights and your back lights are commonly dimmed sometimes with desired effects. Dimmers in studio useage now is more for production than strictly for light control. In other words, a man who wants to start off with a silhouette opening his news show can pan in close to a newsman sitting just back lit and as this man starts talking and you bring up the spotlights and all of a sudden the man is there in perspective--or he wants his business man off and his film chain on--this is what your dimmers are used for--production for church and schools rather than actual intensity controls. That's why you put focus handles on fixtures.

"DESIGNING 18 CHANNELS OF LOCAL DISTRIBUTION SERVICE USING FILTERED PULSE WIDTH MODULATION"

by Dr. Joseph Vogelman
Fellow of the IEEE and VP of Electronic Research,
Chromalloy American Corporation

and Ira Kamen,
Senior Member IEEE and President of Laser Link Corporation

I. Introduction

The Federal Communications Commission, in supporting the dockets for the "Local Distribution Service" (LDS), has awakened a slumbering billion-dollar giant for cable operators in urban and rural areas. This new air-link, CATV service, provides a communications highway of electromagnetic energy which can accommodate a vast number of services, to fill an array of social needs. LDS will not only be an entertainment and advertising service, important as they are to society, but it will also provide information, education, community awareness, and political expression on an economical basis. LDS, as authorized, not only solves the urban problem, but provides the vehicle for taking existing systems in rural and suburban areas, and allows them to extend their services economically to contiguous small communities, jumping highways, crossing rivers, surmounting terrain barriers, and doing so for a fraction of the cost of cables. To understand one of the techniques proposed for this service, this paper is a full disclosure as to how the Quasi-Laser Link System functions.

II. General Considerations

The Quasi-Laser Link System, incorporating the proprietary technique of Filtered Pulse Width Modulation (FPWM) provides a flexible, economic, and workable means of transmitting multi-channel TV signals in a local distribution system as well as over longer haul systems requiring several hops. The FPWM technique makes possible improvements over systems using other modulation means. It does not have the limitations of Amplitude Modulation (AM) which requires extreme system linearity and lacks noise immunity.

For AM systems, these factors require the use of high power devices in order to obtain moderate linear power outputs, and also require wider physical separation between systems causing poor use of the frequency spectrum - space geometry product, and place economic hardships on users requring more than single TV channel transmission.

The Quasi-Laser Link System does its signal processing at low levels and then generates the high level transmitted signal in stages not requiring amplitude linearity so that they can be used at maximum power output and maximum efficiency. It is a system which can be used anywhere in the high frequency domain. The performance when transmitting many channels of TV information in a single transmitter has been demonstrated. System simplicity is best in the higher frequency ranges (millimeter and above) where the possibility of allocating more transmission bandwidth exists; however, much of the advantage of the FPWM technique can still be maintained in the currently allocated bands. Because the FPWM system is able to operate at lower signal-to-noise ratios than less noise immune systems, it makes possible the use of longer point-to-point transmissions or the use of multi-hops.

III. General Technical Considerations

The FPWM transmits as pulse width modulation the composite information content of a number of frequency multiplexed television channels, normally within NTSC standards (Fig. l). The individual channels from locally generated sources or off the air, are combined and down-converted to the spectral region starting at approximately 6 MHz and extending upwards in frequency in accordance with a number of channels being multiplexed.

Two systems of down conversion have been operated in demonstration links. In one case, the composite of channels 7 - 13 are translated to the frequency band 6 - 48 MHz, with channels 2 - 6 remaining approximately in their present location. In the second approach, each channel is translated on an individual basis to a location appropriate to it. To all intents and purposes, this is identical to the technique employed in sub-band transmission in present CATV cable installations. This combination of frequency multiplexed channels results in an output signal voltage whose instantaneous phase and amplitude is the vector sum of the individual channel signals and contains frequency components equal to the highest frequency in the baseband. This instantaneous voltage is fed to modulator circuitry such as to generate a train of pulses whose width is directly related to the instantaneous voltage.

In one configuration of this system, the pulse width varies from one nanosecond corresponding to the least positive voltage to 0.95 nanoseconds corresponding to the highest possible voltage for the case where three channels of television are in the baseband. For five channels, the pulse width varies from one nanosecond to .9 nanoseconds and for twelve channels from one nanosecond to .84 nanoseconds.

The average pulse spacing is equal to the average pulse width such that the average power in the pulse train closely approximates one half of the peak power. The width of the resultant train of pulses has a one-to-one relationship with the instantaneous voltages. These pulses are then limited to insure they are all of equal amplitude and are used to switch the output of a carrier determining oscillator from the "On" to "Off" stage in accordance with the rise and fall of the pulses. The resultant signal has a spectrum consisting of a carrier, the first upper-side band, the first lower-side band, a second upper-side band, a second lower-side band, etc. The spectrum is then filtered by an appropriate microwave filter structure, such as to pass only the first upper-side band.

For operation in the 12.7 to 12.95 GHz band, the carrier oscillator is at a frequency such that the upper first sideband falls within the authorized band and is centered at the assigned frequency. The signal is further amplified to the required transmitter output level by means of a traveling wave tube of a standard variety.

The radiated spectrum after filtering and amplification consists of a frequency varying signal which results from the pulse-width variation of the first modulation step. This spectrum has the general form and distribution associated with frequency modulation, and transmission of this spectrum is treated in a manner identical with the transmission of a frequency modulated carrier.

Two types of receivers have been used in this system (Fig. 2). In the first type, the received signal is amplified with a tunnel diode amplifier followed by a traveling wave tube, both of which are operating in the transmitter frequency band, with a filter limited bandwidth equal to that of the transmitter output spectrum. A solid state circuit network is then utilized to obtain an instantaneous voltage corresponding to the pulse width of the train of pulses present in the IF. The signal which results is identical to that which was used in the transmitter to produce the original train of pulses. The output signal is the instantaneous vector sum of the frequency multiplexed television band such that a standard VHF television receiver is capable of separating out the signals, and displaying them in a manner identical to that used to receive transmissions along CATV cables.

The second type of receiver consists of a balanced mixer and local oscillator at the input to immediately convert the incoming signal from the transmitter to an intermediate frequency. In this configuration, the receiver noise figure is approximately 6 dB higher than that which has been achieved with a tunnel diode input.

In this system, the signal-to-noise ratio presented at the input to the television receiver is a function of the signal-to-noise carrier, and the variation in the pulse width is used to create the transmitted spectrum. For a given carrier-to-noise ratio, the wider the deviations of the pulse widths, the larger the signal-to-noise ratio presented at the television receiver for each of the television channels. Typical tests links have shown signal-to-noise ratios at the input to the television receiver in excess of 48 dB for carrier-to-noise ratios of 34 dB for 12-channel operation.

IV. Theoretical Considerations

Since FPWM is a spectrum, which is the generalized form of an FM wave, after the first upperside band is filtered out, and amplified for transmission, the resultant signal may be analyzed, as if itwere, in effect, FM. The FM formulations derived in the literature can be applied with analyzing this form of transmission.

It is possible to get a better signal-to-noise ratio in an FPWM system than in an AM system with the same transmitter power. To achieve this advantage, however, it is necessary to use a wider range of pulse widths to represent the instantaneous voltages being transmitted. This, of course, requires more bandwidths. The improvement in signal-to-noise ratio, by using wider bandwidths, is obtained by increasing the modulation index:

$$M = \frac{\Delta p}{f_b\,[4p^2 - (\Delta p)^2]} \qquad (1)$$

Where

p = mean pulse width in microseconds

Δp = change in pulse width (max.-min.)

f_b = highest baseband frequency in MHz

With this definition, it is possible to now use formulas associated with frequency modulation to analyze the performance of FPWM. In the following analysis, ΔX is used to define the spectral distribution within the first sideband that is given by the relationship:

$$\Delta X = \frac{\Delta p}{4p^2 - (\Delta p)^2} \qquad (2)$$

ΔX corresponds to the peak frequency deviation in ordinary FM.

A. VIDEO SIGNAL-TO-NOISE RATIO

1. Amplitude Modulation

Amplitude Modulation (AM) may be used to transmit video information. A typical video waveform is shown in Fig. 3 with the vertical axis calibrated in terms of carrier amplitude. The video signal is clamped so that the transmitter power is at its maximum power P on the sync tips. The peak white level is then adjusted to 0.143 of the peak voltage.

The video is usually transmitted using vestigial sideband modulation (VSB). Such a system has the same noise performance as a double sideband (DSB) system having the same signal power and noise spectral density, providing the proper VSB filter is chosen. For this situation, the input signal-to-noise ratio in the channel bandwidth will be smaller for the DSB system since its noise bandwidth is larger. The differences of the several possibilities are summarized in Table I. These signal-to-noise ratios are for white noise in the appropriate bandwidth.

2. Frequency Modulation

Frequency Modulation (FM) is frequently used in microwave satellite or point-to-point television transmission systems. The video signal-to-noise ratio for FM is derived in terms of the RF carrier-to-noise ratio.

For an FM system operation above threshold with sinusoidal modulation:

$$(S/N)_{out} = \frac{3P\ (\Delta X)^2}{2\ N_o\ (f_m)^3} \qquad (3)$$

Where

$(S/N)_{out}$ = output signal-to-noise ratio

P = received signal power

N_o = (single-sided)noise power spectral density

ΔX = peak pulse width deviation spectrum

f_m = video bandwidth

Note that Eq. (3) does not depend explicitly on the total bandwidth W. However, W must be large enough to pass the FM signal undistorted. Also, $\frac{P}{N_o W}$ must be large enough so that operation is above threshold. Eq. (3) will now be expressed in terms appropriate to TV transmission.

The video signal power is defined as the peak-to-peak video squared:

$$S_{p-p} = (V_{p-p})^2 \tag{4}$$

This is done because the peak levels are defined by system constraints while averages are dependent on picture content.

Assuming sinusoidal modulation

$$V_{p-p} = 2\sqrt{2}\ V_{rms} \tag{5}$$

then

$$S_{p-p} = 8V_{rms}^2 = 8S \tag{6}$$

therefore

$$\left(\frac{S_{p-p}}{N}\right)_{out} = 12\ \frac{P\ (\Delta X)^2}{N_o f_m^3} \tag{7}$$

Furthermore, the video signal is clamped so that $f = f_c + \Delta X$ on sync tips, and the deviation is adjusted so that $f = f_c - \Delta X$ for peak white level. Then the peak-to-peak frequency deviation F_{p-p} is:

$$F_{p-p} = 2\ \Delta X \tag{8}$$

The minimum bandwidth required at RF (or IF) is then:

$$W = 2\ (\Delta X + f_m) = F_{p-p} + 2\ f_m \tag{9}$$

$$= 2\ f_m\ (1 + M)$$

where M = modulation index

The input signal-to-noise ratio is then:

$$(S/N)_{in} = \frac{P}{N_o W} = \frac{P}{N_o(F_{p-p} + 2\ f_m)} \tag{10}$$

Substituting (8) and (10) in (7) we get:

$$\left(\frac{S_{p-p}}{N}\right)_{out} = 24\ M^3\ (S/N)_{in} \frac{(F_{p-p})^3}{f_m} \left(1 + \frac{2f_m}{F_{p-p}}\right) \tag{11}$$

$$\left(\frac{S_{p-p}}{N}\right)_{out} = 24M^3\ (S/N)_{in}\ \frac{(1+1)}{M}\quad (2\text{-}10) \tag{12}$$

If the input signal-to-noise (SNR) is referred to the information (video) bandwidth rather than the RF bandwidth:

$$(S/N)_{in\,(\text{video bandwidth})} = (M+1)\ (S/N)_{in} \qquad (13)$$

then

$$\left(\frac{S_{p-p}}{N}_{out}\right) = 24\ M^2\ (S/N)_{in\,(\text{video bandwidth})} \qquad (14)$$

The output signal-to-noise ratios are for peak-to-peak video waveforms and unweighted "triangular" noise spectrum. They differ by a factor of eight from standard results because of the different definition of signal power.

3. Sample Calculation

FM

Let us calculate the "worst case" (in the highest channel) unweighted output signal-to-noise ratio of a 12 channel FM multiplexed television link operating at 12 GHz, with a bandwidth of 250 MHz and a transmitter power of $\frac{10}{12}$ watts per channel:

N_o = $-194\ \frac{\text{dBW}}{\text{Hz}}$ (10 dB Receiver Noise Figure)

W = 250 MHz

ΔX = 47 MHz

f_m = 78 MHz

P_t = $\frac{10}{12}$ W/channel x 12 channels = 10 W

G_t = 35 dB transmitting antenna gain

G_r = 35 dB receiving antenna gain

L_p = 145 dB, path loss

where

$$P_r = P_t + G_t - L_p + G_r \text{ (in dB)}$$

$$P_r = +10 \text{ dBW} + 35 \text{ dB} - 145 + 35 \text{ dB}$$

$$P_r = -65 \text{ dBW}$$

$$N_o = -194 \frac{\text{dBW}}{\text{Hz}} + \frac{84 \text{ dB}}{(250 \text{ MHz}} = -110 \text{ dBW}$$

$$M = \frac{47}{78} = 0.6$$

$$\left(\frac{S_{p-p}}{N}\right)_{\substack{out\\ unweighted}} = 24M^2 \quad \frac{(S)}{(N)} \text{ In}$$

$$\left(\frac{S_{p-p}}{N}\right)_{out} = 24 \ (0.6)^2 \ \frac{(S)}{(N)} \text{ in} = 8.7 \ \frac{(S)}{(N)} \text{ in}$$

$$\left(\frac{S_{p-p}}{N}\right)_{out} = S_{in} - N_{in} + 10 \log 8.7$$

$$= -65 \text{ dBW} + 110 \text{ dBW} + 9.4 \text{ dB}$$

$$= 54.4 \text{ dB}$$

If a 30 dB fade occurred the SNR would still be 25 dB -- quite an acceptable level for many observers.

Now let us calculate the performance of an AM system with <u>single</u> channel capability (6GHz bandwidth).

AM (VSB)

$$N_o = -194 \frac{\text{dBW}}{\text{Hz}}$$

$$W = 6 \text{ MHz}$$

$$P_t = \frac{10}{12} \text{ w/channel x 1 channel} \triangleq 1\text{W}$$

$$P_r = P_t + G_t - L + G_r \text{ (in dB)}$$

$$= +0 \text{ dBW} + 35 - 145 + 35 = -75 \text{ dBW}$$

$$N_o = -194 + \frac{68 \text{ dB}}{6 \text{ MHz}} = -126 \text{ dBW}$$

$$\left(\frac{S}{N}\right)_{\substack{out \\ unweighted}} = -75 + 126 = 51 \text{ dB}$$

IV. Theoretical Conclusions

Using the special definition of SNR for video modulated FPWM systems, even a modest modulation index (0.6) has been shown to produce a 3 dB advantage over a vestigial sideband single channel AM system. Additionally, we have calculated the unweighted (triangular) signal-to-noise ratio which puts the FPWM system at a disadvantage since most of the FPWM noise contribution is at high frequencies. Subjectively high frequency noise is much more tolerable to the average observer so that an additional 4 to 10 dB can be claimed for FPWM, depending on which weighting networks are used in making the comparison. Since AM noise is flat, the weighting network has little effect on system SNR.

We believe we have demonstrated that a 12 channel multiplexed FPWM system with an M of 0.6 can match the performance of a single channel AM system. The obvious equipment simplicity of a multiplex arrangement as opposed to a channel-by-channel transmission method cannot be overlooked. Furthermore, the entire question of amplitude linearity in point-to-point microwave radio systems is so overriding that AM has not been seriously considered heretofore even where it might appear to provide improved performance or reduced bandwidth.

In this type of system, the cross-modulation resulting from the link is a function of the symmetry of the modulation and the demodulation circuitry. If the translation from instantaneous voltage to pulse width is the identical conjugate of the translation from pulse width to instantaneous voltage, the cross modulation at the receiver will be identical to that achieved by combining the television channels in the frequency multiplexing circuitry. This has been shown to be the situation for the case where the television channels are translated into the sub-band, and then translated back at the receiving end. For these channels, the cross-modulation measured, when the down-converter is connected directly to the up-converter, is identical to that measured when the intervening air link is inserted into the system. The cross modulation was measured using the NCTA standard method of square wave modulated carriers. For channels 2 - 6, the cross-modulation was less than -60 dB, which was the limit of the measuring equipment.

The FPWM system must be measured for its cross-modulation as a total link, including the "On-Off" switching which is used to drive the carrier oscillator. As a result, the ability to measure the cross-modulation is limited by the maximum signal-to-noise ratio, which can be achieved in the overall system.

With available components, this limit is 60 dB. As a result, the square waves which must be detected on the unmodulated channel carrier are less than the residual noise level, and can therefore not be measured. Specialized instrumentation to probe into the noise to measure the residual cross-modulation is currently under investigation.

For multi-hop systems, required increase in signal level or antenna gain is a function of the number of hops in the identical manner to that of a standard FM network. Since no modulation or demodulation is required in the intermediate link of the network, the cross-modulation remains unchanged for a multi-hop link.

Pre-emphasis of the frequency multiplexed baseband signal can be used to equalize the signal-to-noise ratios at the television receiver. In one 12-channel configuration, the equalization was adjusted such that a signal-to-noise ratio of 54 dB was achieved at both ends of the baseband, such that even the highest up-converted signals in the normal VHF television band (channel 13) all met this signal-to-noise ratio level.

V. System Description

A. Quasi-Laser Link Transmitter (Fig. 1)

The Quasi-Laser Link transmitter has two principalsub-systems. They are a signal compiler, and a pulse-width modulator.

1. Signal Compiler

The function of the signal compiler is to accept VHF, UHF or internally generated video signals and frequency translate them into sub-band channels starting between 5 and 6 MHz. If it is desired to transmit for example 10 TV channels, then the modulation signal will occupy the frequency range from 6 to 66 MHz. The output of the signal compiler is the sum of individual signal voltages in each channel:

$$(1)\quad v_o(t) = v_1(t) + v_2(t) \text{ ------ } + v_n(t)$$

$$(2)\quad v_o(t) = \sum_{1}^{n} v_n(t)$$

where v_n has the form:

$$(3)\quad v_n = v_{cn} \cos W_{cn} t \; v_m(t)$$

when $v_{cn} \cos W_{cn} t$ is the translated carrier in channel n and $v_m(t)$ is the corresponding composite modulating waveform including video and sound. The output of the signal compiler is applied to the pulse generating circuitry of the modulator.

2. Pulse-Width Modulator

The instantaneous pulse-width is determined by the instantaneous applied control voltage. Ideally, the characteristic is such that a change in the control voltage will cause a proportional departure from the average pulse-width. This can be expressed as:

(4) $K_V(t) = \frac{1}{\quad(t)}$ where is pulse-width

Within the deviation region where the relationship holds the output pulse-width will exactly follow the output voltage variations of the signal compiler. The calculation of the exact spectrum produced is extremely difficult because of the random nature of the signals in each channel.

The pulse-width modulated train is used to key (switch "On-Off") the carrier determining oscillator. The result of this keying is a train of microwave signals which are pulse width modulated. The spectrum which results from this processing consists of a carrier, and a pair of first sidebands. Then there are higher order sidebands which are harmonics of the first sidebands and are a result of the pulse modulation. Fortunately, all of the information contained in the modulating wave is present in either the upper or lower first sideband (they are really exact replicas of the modulating spectrum) so that only one first sideband needs to be transmitted. It is not even necessary to send the carrier wave, although this does result in equipment simplification in the receiver. The name Filtered Pulse Width Modulation comes from the ability to filter out redundant or unnecessary modulation products and still retain the essential information for faithful reproduction of the input television signals at the system output. The filtering is accomplished before and in the TWT amplifier, so that the transmitted signal is the first upper sideband.

B. Quasi-Laser Link Receiver (Fig. 2)

A typical Quasi-Laser Link receiver configuration consists of a local oscillator, mixer and IF amplifier. The received signal is fed to the first detector circuit where it is translated to an intermediate frequency by mixing it with a locally generated signal such that the transmitted signal which is single-sideband suppressed-carrier, appears in the IF band. The first detector output is a frequency modulated wave whose average frequency is centered in the IF band. The signal is amplified, limited, and then applied to a demodulator, the output of which is the multiplexed television signal formed in the signal compiler of the transmitter.

The function of the receiver's de-compiler is to filter and frequency translate these signals to their required locations in the VHF television channel allocations for transmission over cable distribution systems.

VI. Experimental Results

The experimental evaluation of the FPWM technique has been conducted under experimental authorization KB2XFL and KB2XGW issued to Chromalloy American Corporation. As was noted in an earlier report 1/ tests were performed at all the assigned frequencies, 18.5 GHz, 30.0 GHz, 39.3 GHz, and 42 GHz.

The tests confirmed the expectation that the FPWM technique is carrier frequency insensitive, performing in essentailly the same manner at all four frequencies. The major test effort was then conducted at 18.5 GHz.

The overall performance of the system, when transmitting multiple TV channels such as in the LDS application, was evaluated. Television channels 7 - 13 were down-converted in the transmitter compiler from their regularly assigned frequencies to the sub-band frequencies indicated. They were up-converted in the receiver decompiler back to their regular frequencies for reinsertion into the distribution system.

Experiments on the test link were also made using the TV channels on frequency (no down-conversion & up-conversion). When the seven New York TV channels (covering a frequency band of 54-216 MHz) were used, the pulse-width modulation occupied a bandwidth of 1 GHz.

Other tests were made to determine the output signal quality under varying combinations of multiple channel transmission. The video signal-to-noise ratio at the output of the FPWM system is the measure of system quality. The signal level is that of the carrier only, and the noise is the sum of the idle noise and intermodulation products. The experimental receiver always used a full 500 MHz bandwidth so that it could receive full spectrum transmission. For the limited bandwidth tests, an improvement in carrier-to-noise ratio (C/N) could have been achieved by reducing the receiver bandwidth to that required for the spectrum being transmitted.

The output signal-to-noise ratio is dependent upon the input signal-to-noise ratio and the length of the transm ssion path.

The FPWM technique is capable of producing a 20 dB improvement in the detected video S/N ratio over the C/N ratio at the input to the receiver. Operation was simulated for different ranges by means of suitable attenuators. The data taken for a carrier-to-noise ratio at the receiver of 34 dB (equivalent to 15 miles range) was better than 48 dB for 12-channel operation.

1/Appendix B of Petition for Reconsideration in Part in the matter of Amendment of Part 74, Subpart J, of the Commission's Rules and Regulations Relative to Community Antenna Relay Stations.

It should be noted that even at a carrier-to-noise ratio of 19 dB, all the signals are marginally acceptable or better. When the received signal is increased to produce a C/N of 34 dB, all twelve channel outputs are providing excellent picture quality. At C/N of 46 dB, even better quality pictures are feasible.

A. Other Experimental Results: The seven New York TV channels, covering a frequency band of 54 through 216 MHz were transmitted in place with good picture quality being maintained. In order to make a more stringent evaluation, five channels of video programming were combined to form the wide spectrum signal. The VHF-TV channels 2, 4, and 5 were used directly, and VHF-TV channels 9 and 11 were translated into the slots for channels 3 and 6 respectively. The detected signal-to-noise ratio at the input to the TV set for each channel is tablulated below. The signal measured was that of the carrier only. The noise level was the highest level in the channel, determined by searching the band with a field strength meter. The noise measured is the total of idle noise plus cross-modulation, and as such results in the poorest ratio in each channel.

TV Channel	S/N + x mod.
2	45.2 dB
3*	47.0 dB
4	45.0 dB
5	47.5 dB
6**	48.0 dB

*Channel 9 translated
**Channel 11 translated

The experimental results indicate a significant fade margin available insuring satisfactory operation during adverse weather conditions.

From extensive tests under varying atmospheric conditions, it was determined that the FPWM technique is not frequency sensitive, giving essentially the same performance at the four assigned frequencies. It was therefore decided to concentrate the experimental effort at 18.5 GHz. A new air link was established over a 1.5 mile line of sight path. The transmitter was located at The New York Times printing plant at 64th Street and West End Avenue, and the receiving site was in The Times Tower at 43rd Street and Broadway in New York City. This link was maintained over a period from May, 1969 through November, 1969, logging a total of 750 hours of operation.

In this instance, three channels of video information were transmitted by multiplexing VHF-TV channels 2, 4 and 5 directly. The input signals were adjusted to produce identical output signal-to-noise ratios at the input to the TV sets, and the detected signal-to-noise ratios in the three channels are shown below. The measurements were made in the same manner as described previously:

TV Channel	S/N + mod.
2	57 dB
4	57 dB
5	57 dB

The transmitter TWT amplifier had a power output of 1 Watt. Both the transmitting and receiving antennas were 3' parabolic dishes. This combination results in a range capability far in excess of the 1.5 mile path length. The effective path length was increased to 15 miles by use of attenuators. Because of the advantage of the Filtered Pulse Width Modulation technique in the presence of noise and fading, this link continually performed in a satisfactory manner. Transmission maintained during major rainfalls resulted in no degradation in the received picture quality, including September 3, 1969 when the rainfall during the 24-hour period exceeded 3.32 inches.

Other combinations of channels are shown in Table III.

Table IV shows the variation in signal-to-noise ratio as a function of time of day, and shows a comparison of the performance of the headend equipment with down-converter when fed directly to the up-converter receiving equipment, and the resultant performance a short time later through the Quasi-Laser Link System. Typical spectrums are shown in Figures 4 and 5.

While all the foregoing data reflects a system test with only 12 channels, earlier tests have been run with as many as 32 channels. Based on these results, 18 channels of CATV is compatible with all the presently allocated bandwidths for LDS.

VII. Conclusion

The system we have described has complete versatility, and removes the lid on the number of services and applications possible in a locality; all beamed to rooftop, down-converter, and then, by intra-building or inter-home cabling, into existing black and white or color television receivers.

Systems like the Quasi-Laser Link give promise to a bright new prosperous era for television service. It is for those of us within the industry with vision to harness the potential of the Local Distribution Service, and bring it to fruition.

PAPERS PUBLISHED PRIOR TO JANUARY 1, 1970

Title	Author	Publication	Date Of Publication
"Novel Modulation Means Open Wide Spectrum Air Link Communications in Unused Frequency Range"	Dr. Vogelman Ira Kamen	Laser Link Presentation to FCC, Wash. D.C.	Oct. 24, 1968
"Millimeter-Wave Communications Through the Atmosphere"	Hogg, D.C.	Science 159 No. 3810 PP 39-46	Jan. 5, 1968
"Bandwidth and Transmission Bandwidth"	Feldman, D.R.	BSTJ No. 3 PP 490-595	July, 1949

Appendix B of Petition for Reconsideration in Part in the Matter of Amendment of Part 74, Subpart J, of the Commission's Rules and Regulations Relative to Community Antenna Relay Stations. Docket No. 18838

Title	Author	Publication	Date Of Publication
"The Quasi-Laser Link CATV System:	Dr. Vogelman Ira Kamen	Electronics & Appliances Specialist	April 1968
"A New Departure in Television Air Links"	Dr. Vogelman Ira Kamen	International Broadcast Conf. IEEE London	September 13, 1968
"The Quasi-Laser Link System for CATV"	Dr. Vogelman Ira Kamen	TV Communi-cations	Nov. 1968
"A New Dimension in Television Air Links"	Dr. Vogelman Ira Kamen	IEEE Trans-actions on Broad-casting	March 1969

Title	Author	Publication	Date of Publication
Tomorrow's Television - Today"	John Minor	Pageant Magazine	Feb. 1969
The Congeneric Synergism of the Quasi-Laser Link System"	Dr. Vogelman Ira Kamen	Broadcast Journal	March/April 196
Laser Link Modulation Means for Wide Spectrum Telemetry World Wide and Urban Communications Systems"	Dr. Vogelman Ira Kamen	NTC'69 Wash. D.C.	April 23, 1969
'Cabling Without Cable"	Ira Kamen	BME Magazine	June 1969
'Quasi-Laser Link Modulation Means for Wide Spectrum Airlink CATV Services?	Dr. Vogelman	N.C.T.A. Conf. San Francisco	June 23, 1969
Expansion of Communication System Horizons by New Modulation Means"	Dr. Vogelman Ira Kamen	1969 NEC Chicago	December 8, 9, 10, 1969
'Novel Modulation Means opens Wide Spectrum Air Link Communications in unused frequency Range"	Dr. Vogelman Ira Kamen	IEEE Convention Huntsville, Ala	November 19, 20, 21 196
"Signal-to-Noise Ratio for Television Transmission"	Collins	MIT	March 1969
Transmission Systems for Communications"		Bell Telephone Labs. Chs. 17-20	1964

TABLE I

SNR for AM Television Systems

Modulation	Channel Bandwidth W	$(S/N)_{in}$	$(S/N)_{out}$	$\frac{(S/N)_{out}}{(S/N)_{in}}$
DSB	$2f_m$	$\frac{p}{2N_o f_m}$	$\frac{p}{N_o f_m}$	2
VSB	$f_m + \Delta$	$\frac{p}{N_o(f_m+\Delta)}$	$\frac{p}{N_o f_m}$	$1+\frac{\Delta}{f_m}$
SSB	f_m	$\frac{p}{N_o f_m}$	$\frac{p}{N_o f_m}$	1

where:

f_m = Video bandwidth

p = Signal power (peak RF or video power, including the white-level d.c. offset)

Δ = Bandwidth of the vestigal sideband

N_o = Single-sided noise power density

W = Channel bandwidth

TV CHANNEL	ORIGINAL FREQ. IN MHz	CONVERTED FREQ. IN MHz	S/N AT INPUT TO TV SET			
			C/N-19 db	C/N-34 db	C/N-34db equalized	C/N-46 db
2	54-60	54-60	27	43	48.5	70
3	60-66	60-66	27.5	46	48.5	61
4	66-72	66-72	26	44	49.0	58
5	76-82	76-82	23	41	54.0	69
6	82-88	82-88	23	41	53.0	53
7	174-180	5.75-11.75	30	41	46.0	50
8	180-186	11.75-17.75	23	40	46.5	50
9	186-192	17.75-23.75	23	43	47.0	53
10	192-198	23.75-29.75	25	43	45.0	50
11	198-204	29.75-35.75	20	40	46.0	52
12	204-210	35.75-41.75	29	45.5	46.5	50
13	210-216	41.75-47.75	30	41	47.0	51

TABLE II INDIVIDUAL CHANNEL OUTPUT S/N RATIO FOR DIFFERENT C/N RATIOS MEASUREMENTS FOR EACH C/N RATIO MADE AT DIFFERENT TIMES.

NUMBER OF CHANNELS TRANS.	INPUT C/N dB	TV CHAN.	FREQUENCY MHz	OUTPUT VIDEO S/N Db
2	46	5 6	76-82 82-88	56 60
3	46	2 3 4	54-60 60-66 66-72	55 55.5 57
3	46	11 12 13	29.75-35.75 35.75-41.75 41.75-47.75	49 49.5 49.5
3	46	10 11 12	23.75-29.75 29.75-35.75 35.75-41.75	49 49.5 51.5
3	46	9 10 11	17.75-23.75 23.75-29.75 29.75-35.75	52 51.5 52
5	46	2 3 4 5 6	54-60 60-66 66-72 76-82 82-88	59 56 54 54.5 56
5	46	9 10 11 12 13	17.75-23.75 23.75-29.75 29.75-35.75 35.75-41.75 41.75-47.75	47.5 46 44 44 48
6	46	2 3 4 5 6 7	54-60 60-66 66-72 76-82 82-88 174-180	50 48 45 46.5 44.5 47

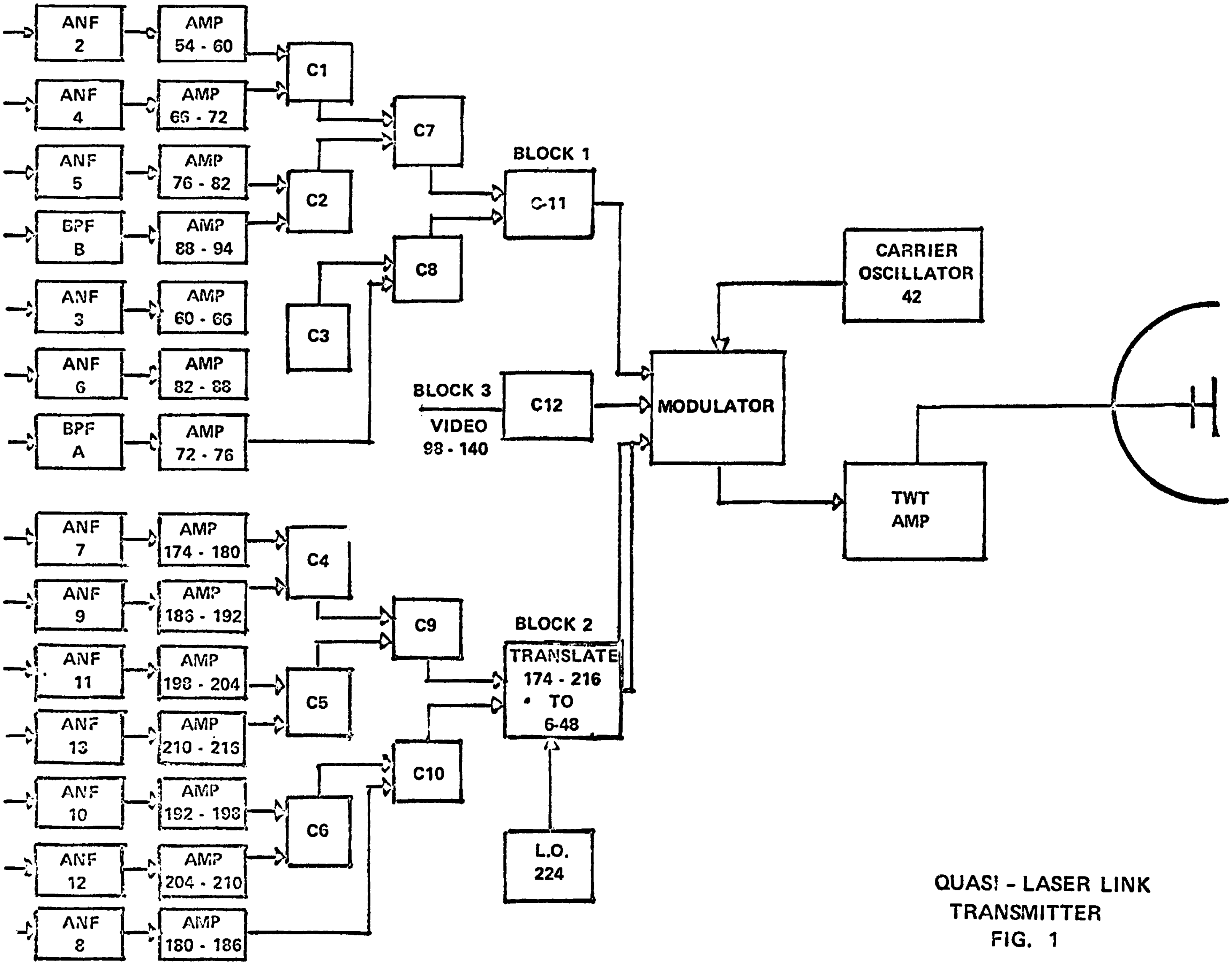
Receiver band width 500 MHz

TABLE III OUTPUT VIDEO S/N RATIO FOR VARIOUS COMPILER COMBINATIONS

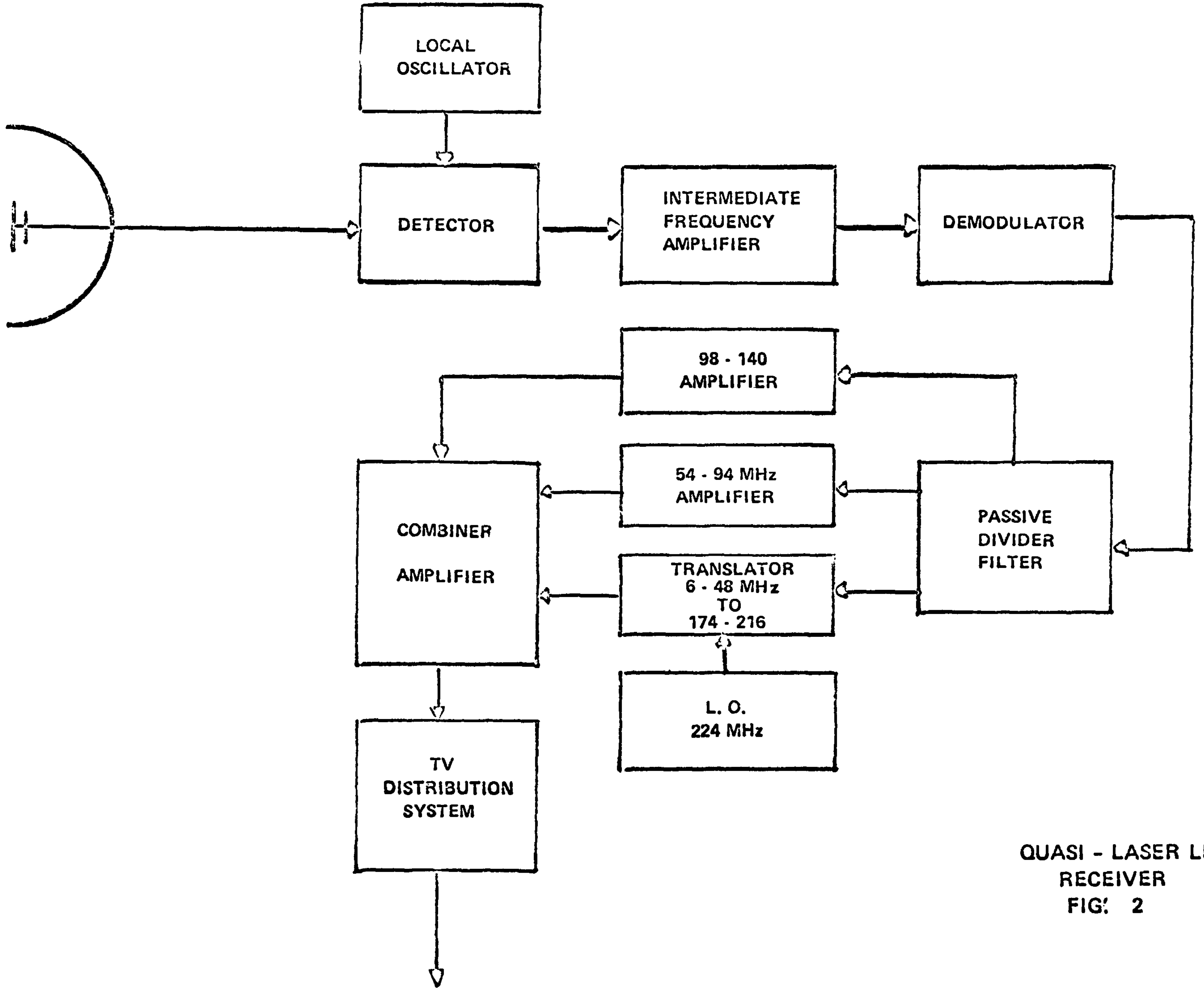
		HEADEND EQUIPMENT		QLL SYSTEM	
	CHANNEL NO.	TEST 1 S/N	TEST 2 S/N	TEST 1 S/N	TEST 2 S/N
	2	52	60	48	52
	3	53	6C	48	53
	4	61	64	50	61
	5	56	63	50.5	56
	6	54	60	53	54
	7	51	54	44	54
	8	54	55	43	54
	9	52	55	47	52
Subband	10	48	54	46	48
	11	49	52	46	49
	12	47	54.5	46	47
	13	51	55	50	51

C/N for QLL = 33db

TABLE IV COMPARISON OF HEAD END AND QLL PERFORMANCE

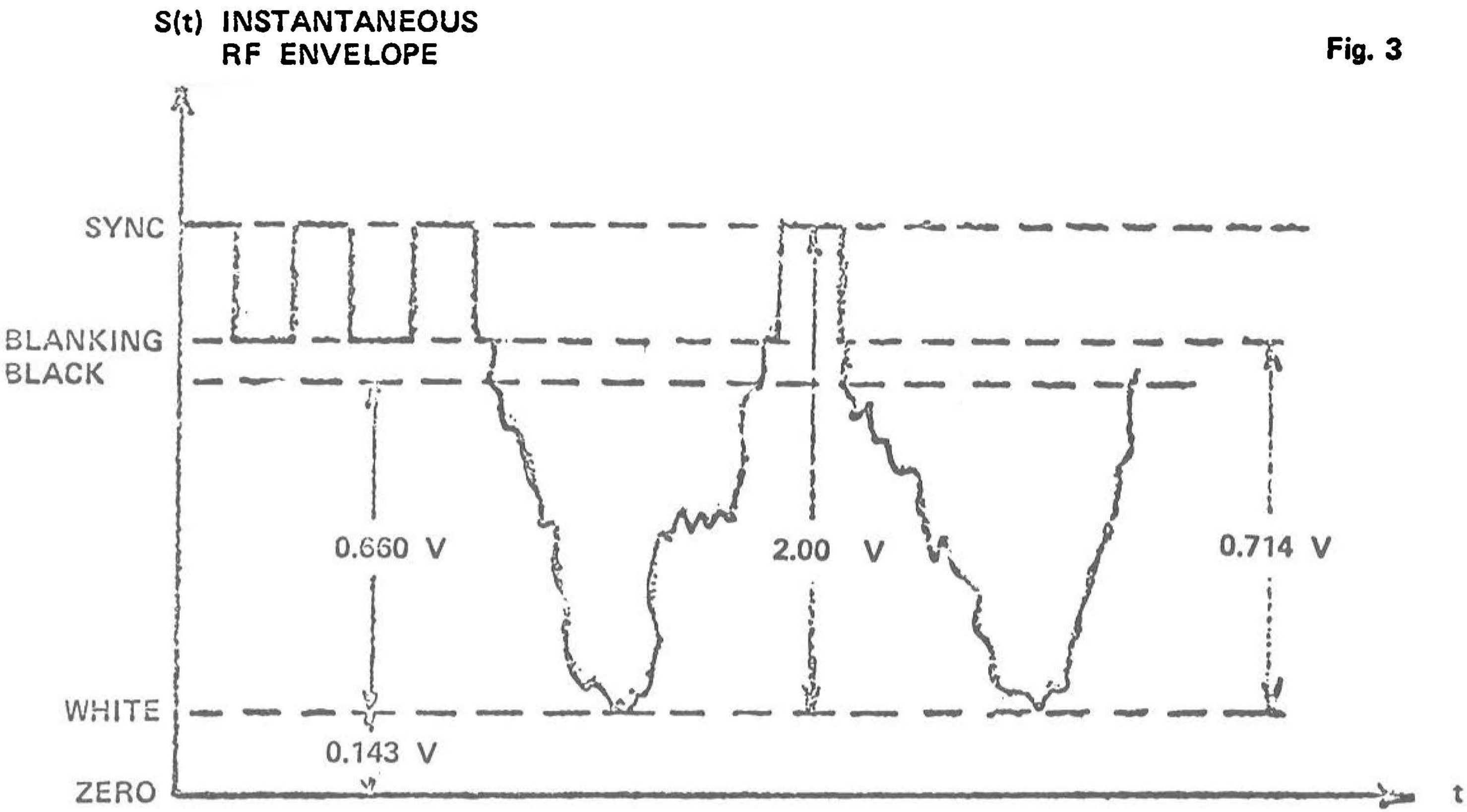


QUASI - LASER LINK
TRANSMITTER
FIG. 1



QUASI - LASER LINK
RECEIVER
FIG. 2

Fig. 3



NOTE:- Amplitudes Refer to Video Modulating Voltage and is Normally 1.0 volts Peak to Peak

FIG. 3 - Television composite signal waveform, amplitude modulation.

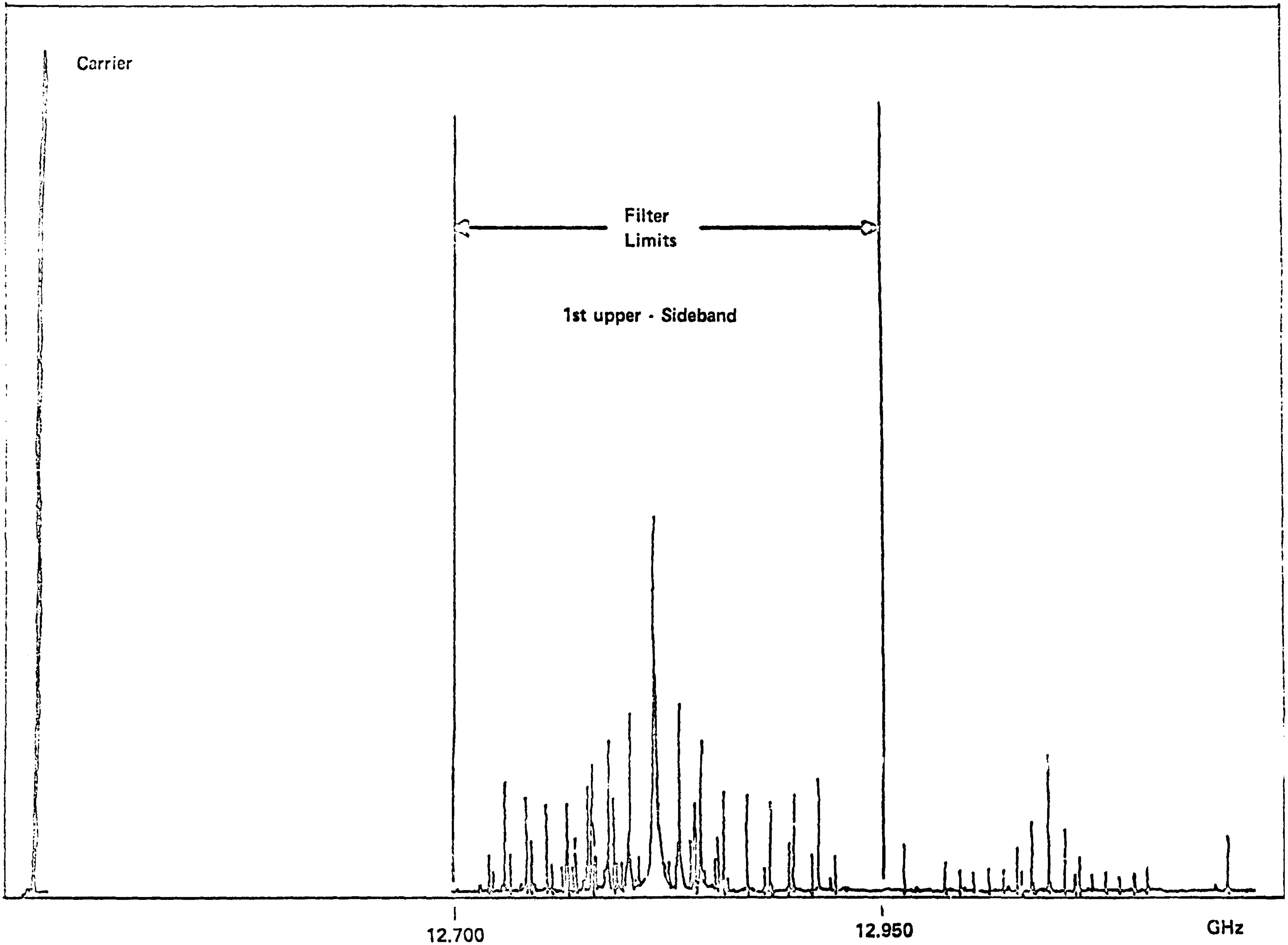


Fig 4 - Upper Half of Pulse Width Modulated Spectrum

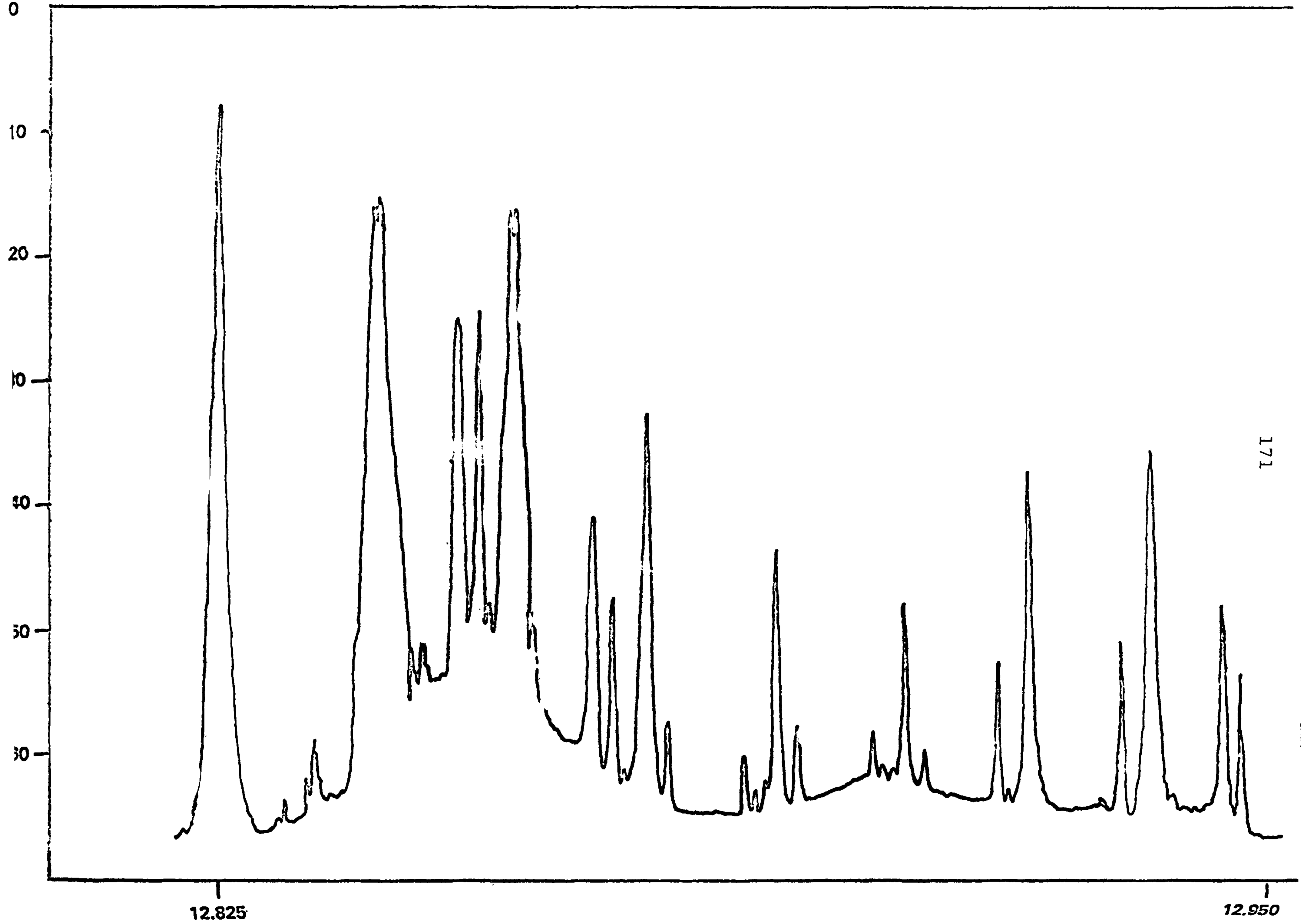


Fig. 5 - Upper Half of First Sidebond FPWM Spectrum

DISCUSSION

Dr. Vogelman: Any questions?

Mr. Hub Schlafly: Hub Schlafly, TelePrompTer. I have two questions. One is the data that you presented data that is equivalent to the proposed rulemaking at the Commission--is that equivalent to the system that has been identified in the proposed rulemaking as in terms of modulating bandwidth and RF bandwidth?

Dr. Vogelman: This all fits into 250 megahertz but is done at 18½ gigahertz.

Mr. Schlafly: So it was a 12 channel system in a RF bandwidth of 250 megahertz.

Dr. Vogelman: 250 megahertz, but at 18½ gigahertz.

Mr. Schlafly: What is the index of modulation of the multiplex (tape fadeout) signal as the Commission calls it--that first side band? Do you have a figure on that?

Dr. Vogelman: Well, I can tell you what the variation in widths are. The variations in widths run about 16%. In other words, the fattest to the thinnest pulse is 16%.

Mr. Schlafly: I understand the variations. That would be restricted in order to get the information into the 250 megacycles of CARS bandwidth.

Dr. Vogelman: Right.

Mr. Schlafly: But do you have a figure of what that modulation is or what the modulation index is because that's really the factor that determines performance, is it not, in an FM system?

Dr. Vogelman: Well, it depends on how you define this in terms, I don't know how you translate quickly from one to the other, but I would guess that if you have a base band which is 72 magahertz wide using Carson's Rule this index of modulation is about .8.

Mr. Schlafly: I applied Carson's Rule. I'm not sure that that applies on low index modulation FM systems. But if I took 114 megahertz, as you have indicated, to the Commission and FM multiplex set for 250 megahertz RF spectrum, I come out with about 0.1 index of modulation.

Dr. Vogelman: If that signal were noise, you're right. But it's not.

Mr. Schlafly: Okay, thank you.

Dial a Program—an HF Remote Selection Cable Television System

R. P. GABRIEL, MEMBER, IEEE

Abstract—A new concept in cable television systems is described in which the general form of the network is similar to the local distribution network of a telephone system. Each television set is provided with an individual connection to a program exchange where the subscriber may select, by means of a remotely operated switch, the program of his choice from the unlimited number of programs available at the exchange. The system is based on the technology of HF multipair cable television systems extensively used in Great Britain. The system is flexible in use and the facilities for two-way transmission, subscription TV, audience rating, etc., are explained.

THE need for the distribution of an ever increasing number of channels has stretched the conventional technology of broad-band VHF distribution operating with conventional television receivers to the point where either two coaxial cables must be employed or every television set must be provided with an additional tuner unit or set top converter. In these circumstances a radically different approach is worth considering. As the number of channels increases, the cost of delivering all channels to all homes and selecting the desired channel by apparatus in the home rises to the point where it breaks even with the alternative possibility of delivering only the desired channel to the home with the selection process carried out at some central "program exchange" by remote control. The break-even point will rise as the density of television sets per unit area increases and this density will be affected, not only by the average distance as the cable runs between one home and the next, but also by the number of television sets in each home. The break-even point will also be greatly affected by the extent to which the cost of a single 5-MHz channel can be reduced in comparison with the cost of a wide-band VHF connection.

The techniques which have been developed by Rediffusion International Limited in Great Britain for their HF multipair cable television systems, now serving over three-quarters of a million subscribers, offer some attractive possibilities of meeting the requirements of a remote selection system [1]. In the multipair cables, instead of each separate pair being dedicated to the distribution of a separate program, each pair is dedicated to a separate subscriber to connect him to the program exchange, as shown in Fig. 1, where he may select, by means of a telephone type dial, the program of his choice from the unlimited number which may be made available there.

Manuscript received January 29, 1970; revised February 16, 1970.
The author is with Rediffusion International Limited, London, England.

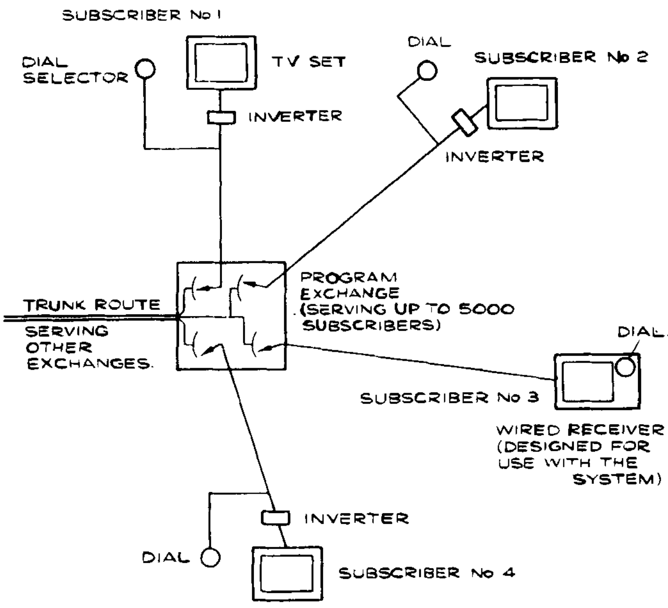


Fig. 1. HF remote selection. General arrangement.

I. The Choice of Frequency

The frequency band chosen for the distribution system must be high enough to avoid complications when demodulating the signals but otherwise should be as low as possible. Thus, all programs are transmitted on the same carrier at a frequency of 7.94 MHz with the upper sideband suppressed; and with the chrominance, therefore, falling at 4.36 MHz and the FM sound at 3.44 MHz. The fact that all programs are on the same carrier means that there can be no beats between the carrier of one program and the carrier of whatever programs happen to be carried by neighboring pairs in the cable; thus, the cross talk or "crossview" performance required is no greater than that required between two video frequency circuits, i.e., about 45 dB. Although there can be no beats between the luminance carriers, the possibility remains of beats between the chrominance carriers because National Television System Commission (NTSC) standards unfortunately allow a tolerance of ± 10 hertz in the luminance–chrominance spacing of 3.58 MHz as opposed to ± 1 hertz which is accepted in Europe. At the limits of the NTSC tolerance a higher crossview performance is required to eliminate the effects of interchrominance beating. For this reason, the upper sideband of the signal is suppressed bringing the chrominance signals

Fig. 2.

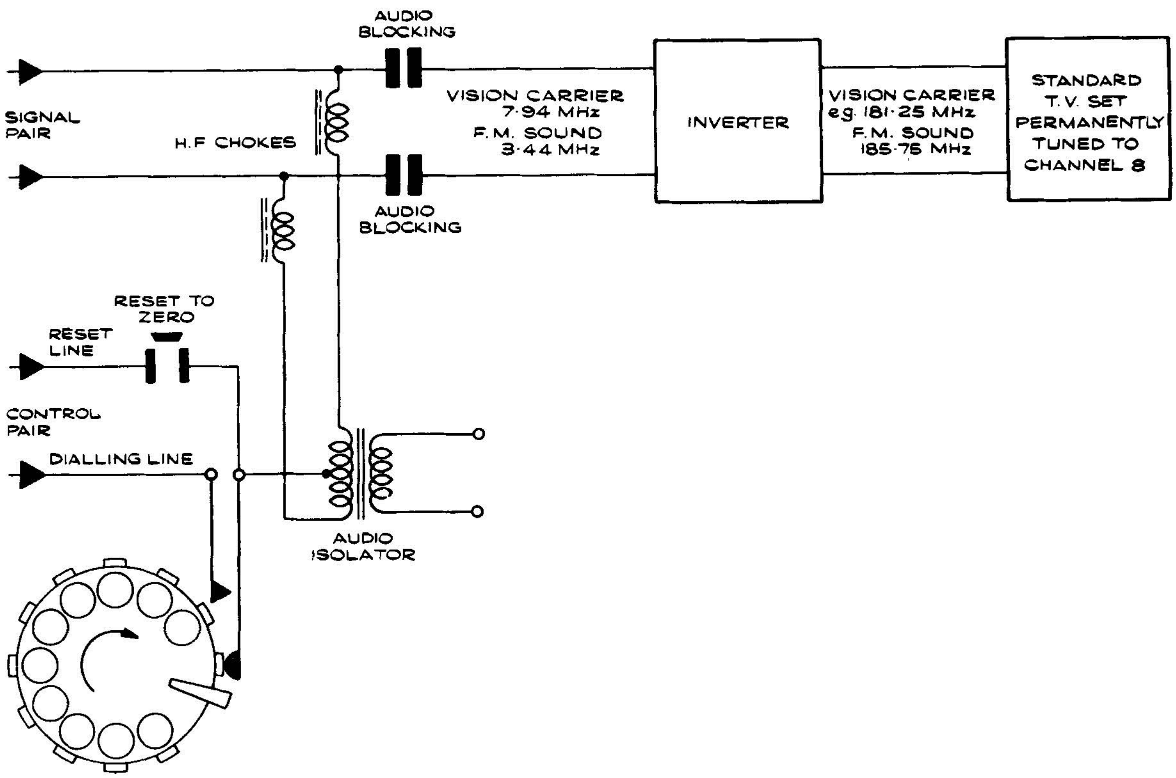


Fig. 3. HF remote selection.

to a lower frequency where the crossview performance of the cables is better.

Operation at frequencies only just above the video band has advantages in simplicity and reduced cable attenuation but introduces other difficulties which may be mitigated by careful choice of the exact frequency. For example, the second harmonic at 6.88 MHz of the FM sound carrier, which may result from nonlinearity in the repeaters or in the receiving equipment, will beat with the luminance carrier at 7.94 MHz. However, if the frequency of the luminance carrier is correctly chosen, this interference will occur at a frequency offset from the luminance carrier by a multiple of one-third of the line frequency and its visibility will be much reduced [2].

For the operation of the simplified "wired" receivers which may be used with the system, the sound is also transmitted at audio frequency.

II. Equipment in the Home

In the ordinary case, the equipment will consist of a standard television receiver connected to the cable through a small unit containing a telephone-type dial for program selection and a frequency conversion unit known as an *inverter*. The inverter unit contains a single transistor and three semiconductor diodes. It has no controls and requires no attention: its function is to change the signals on the cable to a channel in the VHF high band which is free of interference from local broadcasting stations. The television receiver is left permanently tuned to this channel—programs being selected by the dial. See Fig. 2 and Fig. 3.

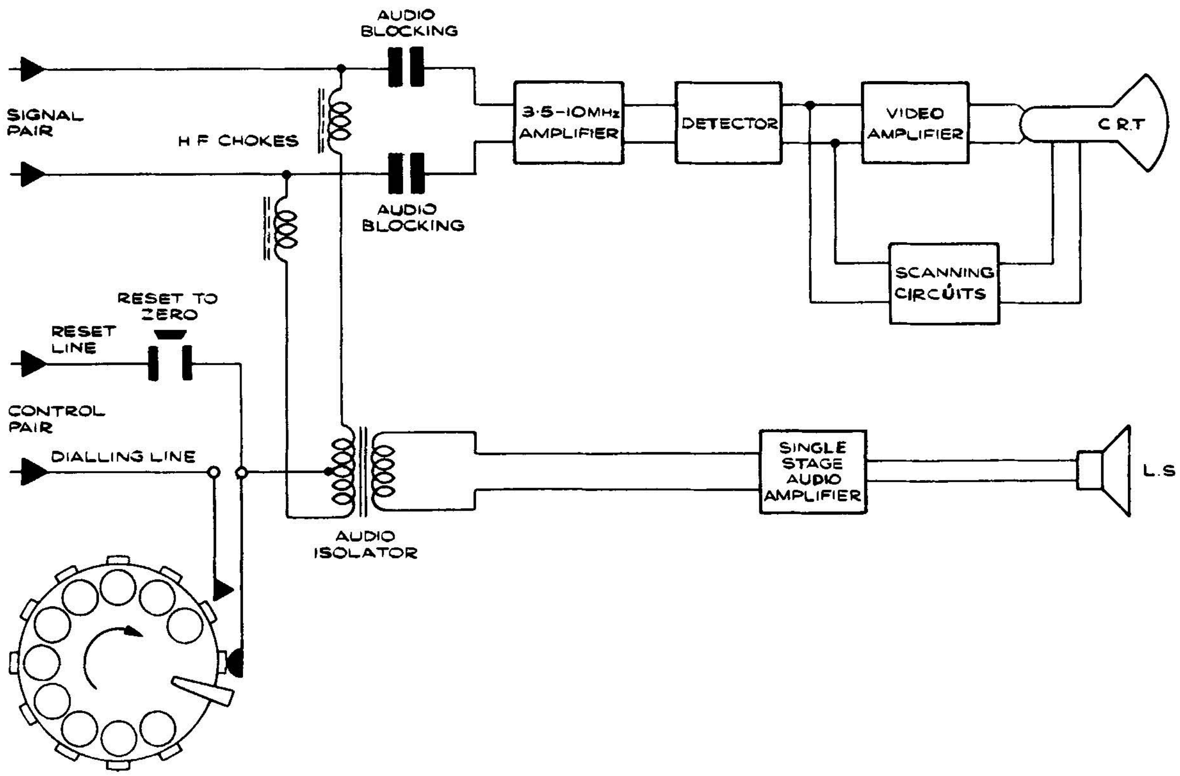


Fig. 4. HF remote selection.

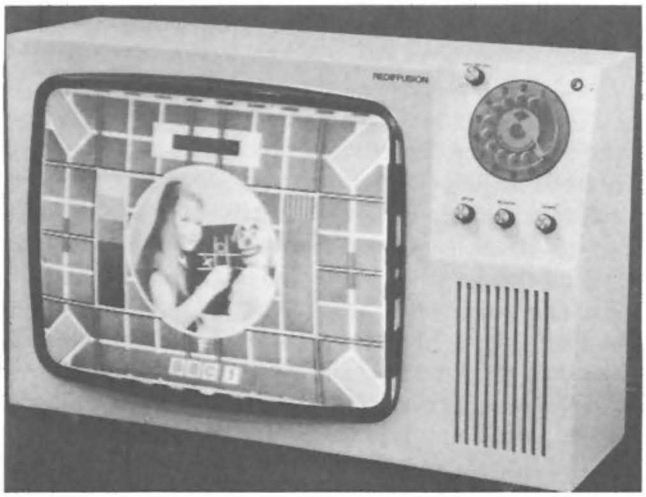


Fig. 5. Monochrome dial receiver.

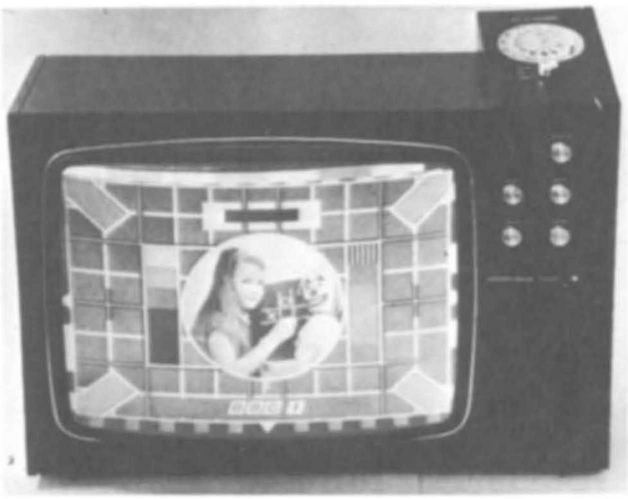


Fig. 6. Color dial receiver.

If the home or other establishment is not already equipped with television receivers, they may purchase a simplified and cheaper receiver designed for use with the system. This contains a straight amplifier for the vision signals and has no tuner. It does not respond to the FM sound signals but accepts instead the audio frequency signal from the cable which is applied through a single stage amplifier to the loudspeaker. The only controls are volume, brightness, and the program selection dial, which in this case forms an integral part of the receiver. A block schematic is shown in Fig. 4. A monochrome wired receiver with dial incorporated is shown in Fig. 5 and a color version is shown in Fig. 6.

The transmission of the sound at audio frequency to these wired receivers has several advantages; it reduces their cost, provides a very high quality signal, and enables them to accept a sound program unaccompanied by a vision signal. It also means that if, on certain programs, the FM sound is not transmitted, the sound trap in these receivers can be switched out of circuit; therefore, a very wide-band vision signal can be reproduced. Such a facility might be useful for displaying alpha-numeric information or high quality color pictures, perhaps for medical purposes.

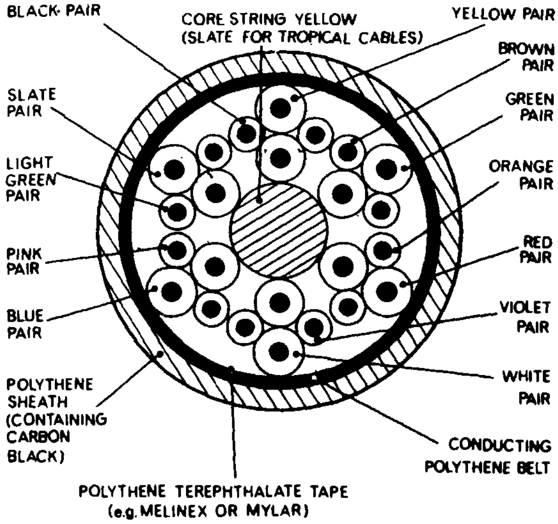


Fig. 7. Twelve-pair (6 Qwist) cable.

III. The Cables

It is not always realised that, for a given attenuation, an unscreened twisted pair will have a smaller overall diameter than a coaxial cable and will cost a great deal less. So, if a multipair cable can be made to operate in the 5-MHz to 10-MHz range with adequate isolation between one pair and its neighbors, it will represent the best engineering solution. A range of multipair cables for the Rediffusion International Limited HF system has been developed over the last 20 years and has now reached a highly developed state in which a very high degree of balance can be regularly maintained in production. A particular form of these cables is known as *Qwist* and is shown in cross section in Fig. 7 [3]. This cable, with an outside diameter of 0.41 inch, contains six larger pairs, comprising conductors of 0.018 inch, which are used to carry the vision signal. In the interstices of each of these vision pairs an additional smaller 0.016 inch pair is laid and twisted together with the vision pair. These additional pairs were originally included to provide circuits for six sound-only radio programs in the standard Rediffusion International Limited HF system but it was found that they led to a marked improvement in the crossview performance of the cable. This is due partly to an improvement in dimensional stability as the sheath is extruded over the pairs and partly to some screening effect. When these cables are used for a remote selection system this improvement in performance is very welcome since the crossview conditions are unusually severe. This is because one must design for the worst case, which arises when five subscribers served by the same cable choose the same program and the sixth chooses another. It is for this reason, and because of the simplification which is possible in the control circuits in the home and the exchange, that these extra conductors are included. Also, since they are used only during dialing, these circuits are available at all other times for other communication needs such as fire or burglar alarms or for talk-back in closed circuit systems for education and so on. Conceivably, in new housing construction, they could be used for telephone service but the saving in cable cost does not seem likely to outweigh the extra costs involved in the diversion of the normal telephone cable route, the relays required to distinguish the telephone and television system conditions, and the extra fault liability for both services.

The *Qwist* cables radiating from the exchange are terminated at junction boxes where a straight-through connection is made to six separate two-pair cables which carry the signals on to the individual television sets.

IV. Network Plan

Normally, no repeater amplifiers are used between the program exchange and the television receiver so that the length of cable is limited by the permissible attenuation and the tolerable level of crossview from the adjacent pairs in the cable which will be carrying different programs. The limit set by these two considerations is about 400 yards map distance (500 yards actual cable route) between the program exchange and the junction box with up to a further 100 yards of cable from the junction box to the television receiver. Although these lengths can be used within the limits of the performance specification, economic consideration may dictate a smaller area. As the cables approach the exchange their number builds up and, in an exchange serving say 2400 sets, one might have 100 cables leaving in one direction.

The cost per yard of this part of the network will obviously be high and it might cost less if the area were served by two exchanges instead of one. Any such saving will be offset by an increase in the cost of the trunk route serving the exchanges and the cost of this will increase with the number of programs carried. Another offsetting factor is the cost of obtaining suitable sites for the exchanges—two separate sites being likely to cost more than twice one larger site; although, in serving a series of apartment blocks, a small exchange in the basement of each block might well be the best arrangement. An unfortunate but inescapable consequence of removing the tuners from the home is that one must find space for their equivalent elsewhere.

The distance between the junction box and the receiver can be extended to about five miles if required, for example, to feed remote subscribers in country districts by the use of single channel transistorised repeater amplifiers powered over the cable and spaced every three-quarters of a mile. These repeaters and the screened two-pair cable are inexpensive and they make it possible to serve remote subscribers at a cost which would not compare too unfavorably with the cost of a telephone connection. See Fig. 8.

In general, when planning the network for a city and for the distribution of 20 to 40 programs, a reasonable compromise between the conflicting factors is obtained if the area is split up into units of about one-tenth square mile which might contain from say 300 to 3500 homes and, allowing for a second set in every other home, some 450 to 5000 television receivers. A program exchange is set up at the center of each one-tenth square mile area and the multipair cables are taken through the streets overhead or via under-

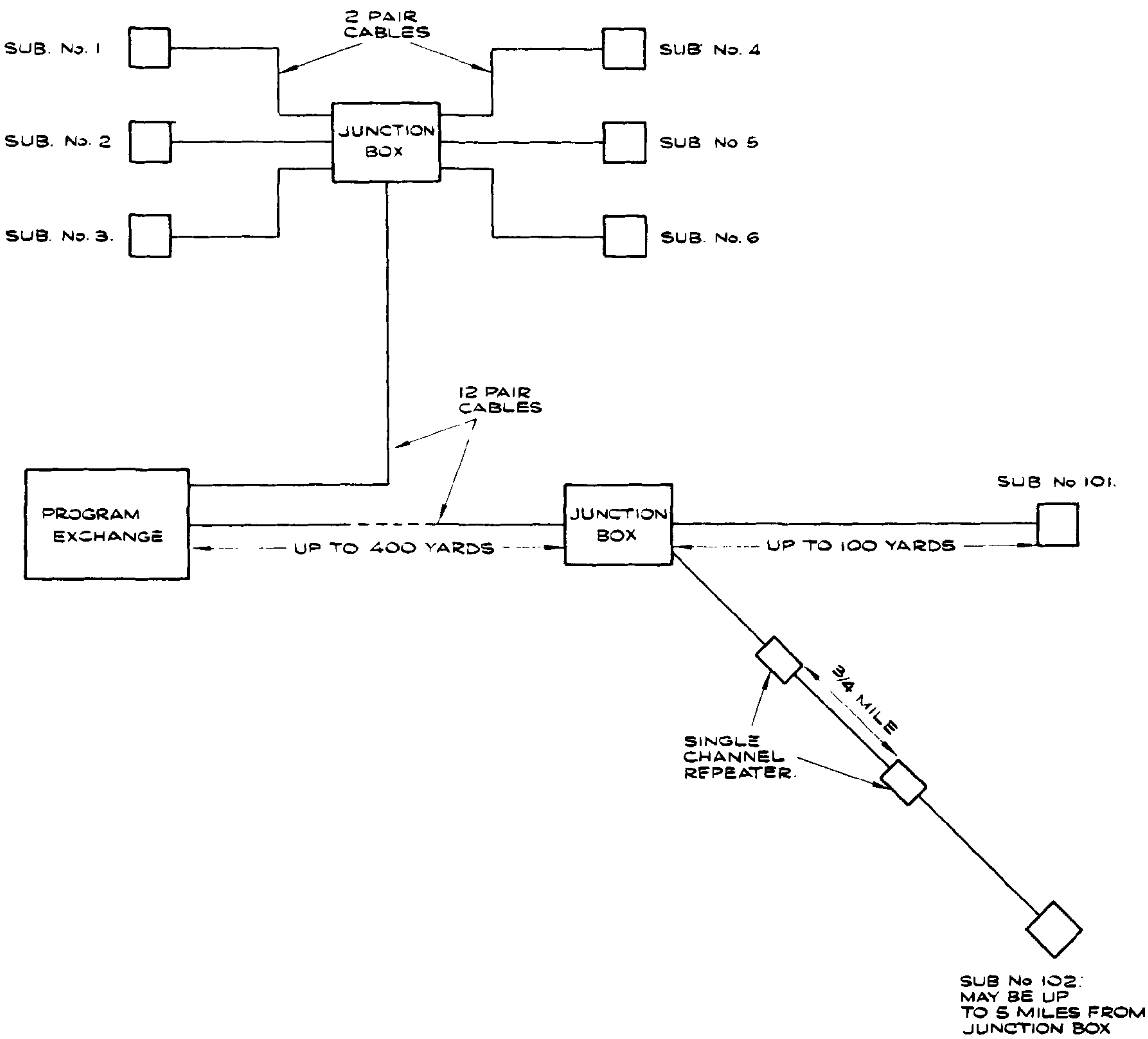


Fig. 8. HF remote selection network plan.

ground ducts of similar construction to telephone ducts. If spare telephone ducts are available they may be used without any fear of causing or suffering interference from the telephone system. The operation of the television system itself depends upon a very high degree of electrical isolation between the pairs in its cables and this ensures an even higher degree of isolation from other cables which may be nearby.

V. The Program Exchange

The program signals are delivered to the exchange by means of a primary distribution or trunk network and are amplified by single-channel repeaters and separate amplifiers for the audio signals. The audio and television signals for each program for these amplifiers are in the unbalanced mode and are combined and applied to a screened bus-bar system in the form of printed circuit conductors. Each alternate conductor is grounded and the printed circuit is backed by a continuous aluminium sheet which is also grounded. The remaining conductors, the bus-bars, carry the signals, and form transmission lines of very low impedance, 12 ohms, and distribute the signals to the subscribers' selector switches which may be plugged in and out by means of edge connectors. The bus-bar system is split up into four sections of different voltage level with the sections being connected together by step-down hybrid transformers. The most distant subscribers are served by the section carrying the highest voltage (0.75 volts) with the other sections serving groups of subscribers which are progressively nearer the exchange. The object in dividing the bus-bar exchange in this way is to ensure that the signal levels in any one Qwist cable are approximately the same so that a difference in level does not add to the crossview protection required in the cables. An outline of the circuit arrangement is shown in Fig. 9.

The selector switches make use of reed relays operated by a permanent magnet mounted on a rotating arm which is stepped round by impulses from the subscriber's dial. A prototype switch is illustrated in Fig. 10 and is shown without the cover shield which normally lies between the rotating magnet and the reeds. The main criteria in the design of the selectors is an adequate crossview performance and an ab-

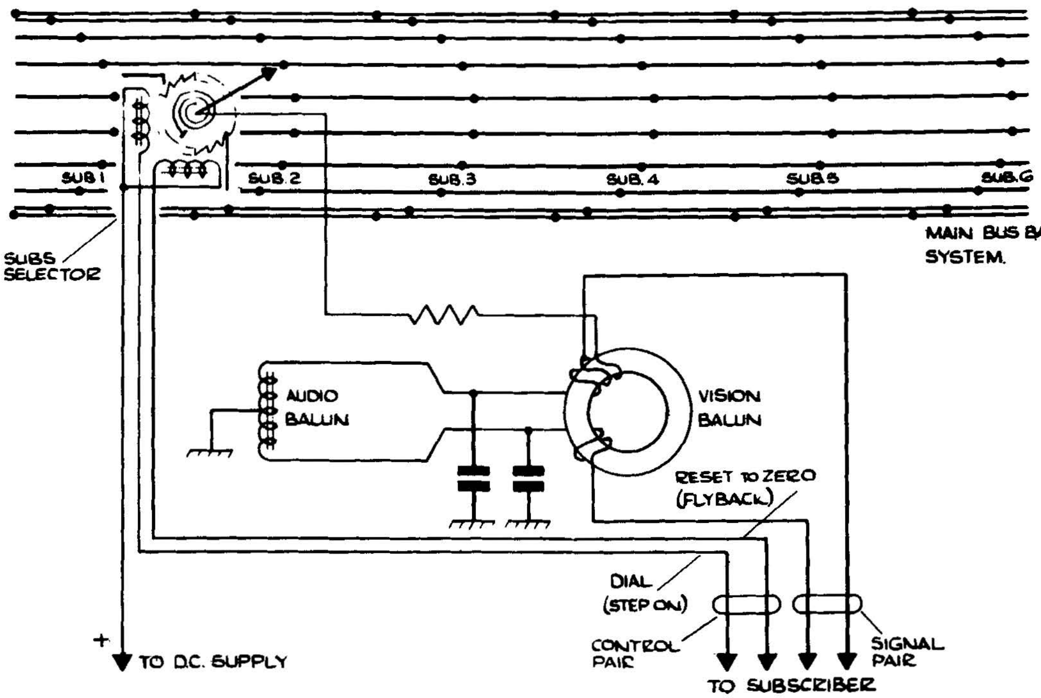


Fig. 9. HF remote selection. Subscriber program exchange (10 or more channels).

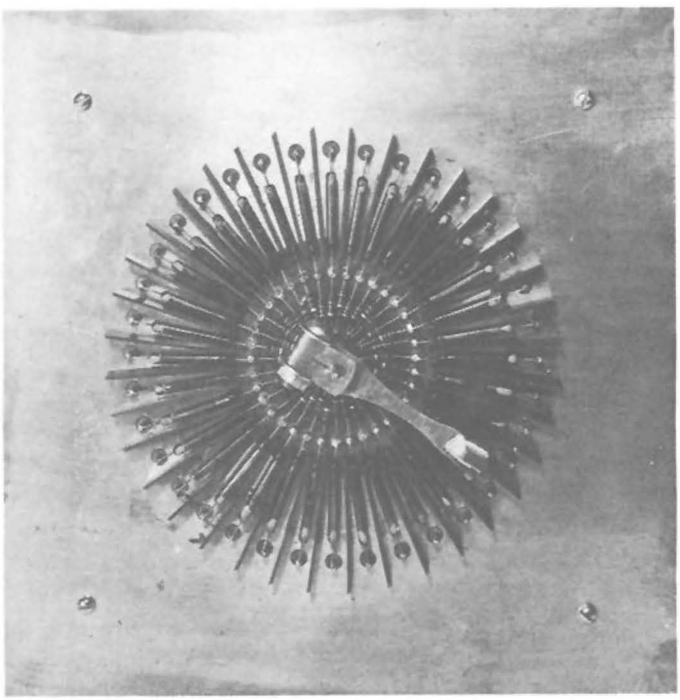


Fig. 10. Prototype selector.

solutely reliable break before make action as the magnet passes from one reed to the next. Both objects are achieved by a screen of transformer iron between one reed and its neighbor which concentrates the magnet field on the chosen reed and is effective electrostatically.

Even so, it remains of first importance that the effect of any residual coupling between neighboring circuits in the exchange should be reduced as far as possible and, for this reason, two different luminance carrier frequencies in one-third line offset relationship to each other are used for programs on alternate bus-bars in the exchange. The two frequencies chosen are 1009 times one-half the line frequency plus and minus one-sixth of the line frequency, that is, 7.935314 MHz and 7.940558 MHz. It will be seen that by choosing the frequencies in this way the condition of one-third line offset for the second harmonic of the FM sound frequency referred to in Section I of this paper is also met [2].

In order to reduce the capital cost of a program exchange, from which initially only a few subscribers may be served, and for ease of maintenance, it is desirable that the selectors should plug in and out of the bus-bar system. As reliable edge connectors are expensive, and in order to reduce the volume of the exchange equipment, the selectors are mounted in pairs on printed circuit boards as shown in Fig. 11.

The simple electromagnetic stepping mechanism is shown in Fig. 12. When the subscriber presses his reset button a catch is released and the mechanism returns to its zero position under the action of a spring. In this zero position the subscriber would, in most applications, receive a dialing directory on his screen of the programs available at that time.

Since a selector switch must be provided for each subscriber, the cost of this item is a matter of first importance in the economy of the system as a whole. It is this which has led to the adoption of an electromechanical design in which a cost of 70 cents per cross point seems well within reach

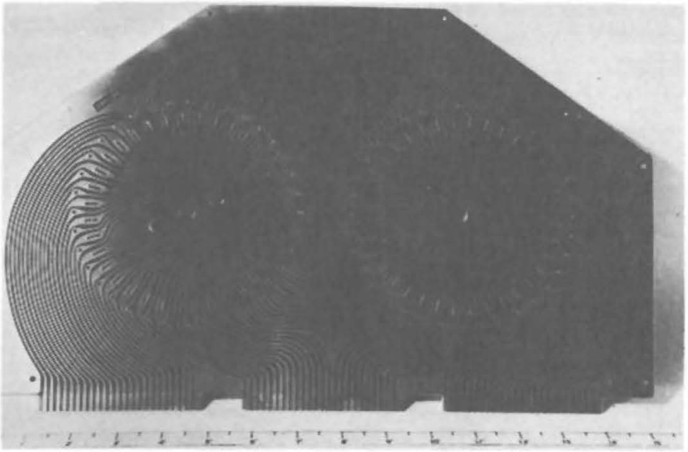


Fig. 11. Twin selector printed circuit.

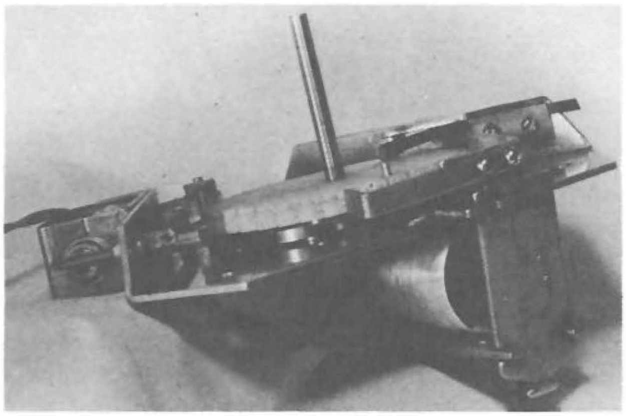


Fig. 12. Selector actuating mechanism.

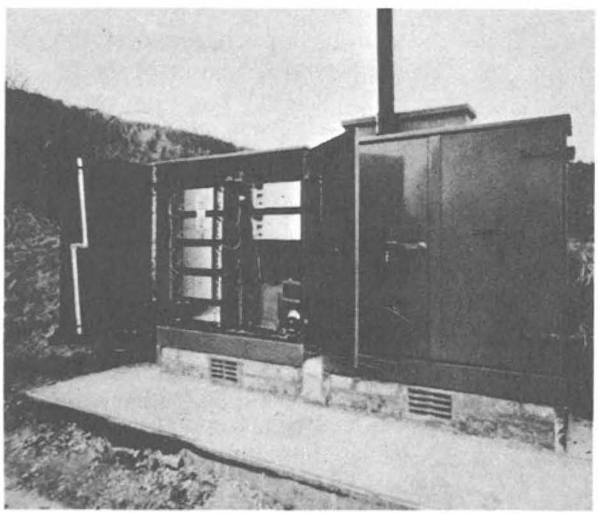


Fig. 13.

for production in reasonable volume. Although no doubt the cost of solid-state switching will fall in the course of time, it does not appear at present that such techniques can approach that figure.

The space occupied by the exchange equipment will, of course, depend on the number of television receivers served and the number of programs available to them; but for example, the equipment required for 330 receivers each having a choice of 36 programs would be contained in a volume of 8 feet by 6 feet by 2 feet. In areas of low housing density the exchange might therefore take the form of a kiosk of the type illustrated in Fig. 13. In areas of high density where the number of receivers might be about 5000, the equipment would be housed on 9-foot high racks in a room 27 feet by 15 feet.

VI. Trunk Network

The programs are delivered to the exchange by means of a primary distribution network which may use any of the conventional methods of transmitting a television signal. For example, in a city where adequate underground duct space is available the maximum simplicity and economy will be obtained by the use of a separate coaxial cable carrying the signal in the same form and on the same channel as that used for the distribution network; in this case, the equipment required in the exchange consists only of amplifiers. An inexpensive coaxial cable, having an outside diameter of 0.312 inch, will have an attenuation of less than 40 dB per mile at the operating frequency, so that a city, together with large parts of its outlying suburbs, can be covered without the use of repeaters between program exchanges. This greatly facilitates two-way transmission on the trunk routes and enables the programs from any source within reach of a program exchange to be made available throughout the network without first being carried back to a distribution center.

If the duct space is insufficient to allow the use of separate cables or if the distance to be covered is many miles, then it may become more convenient to employ a wider frequency spectrum and to carry many programs on a single coaxial cable with frequency changing equipment in the program exchanges. These techniques are, of course, very well developed and in extensive use in the United States and elsewhere for CATV.

A third alternative method is the use of multichannel microwave links between program exchanges. No doubt these three methods and other point-to-point television links that may be developed in the future would find appropriate use in a large system. Also, in a large city the program capacity of the trunk network would vary from one part to another and not all programs would need to be available to all citizens. For example, in the downtown districts many channels would be required for commercial purposes; in other parts, programs for particular ethnic groups could be limited to the areas of their concentration, thus reducing the cost of distribution. Similarly, the broadcasts of political candidates could be confined roughly to their area of concern.

VII. Two-Way Operation

In the ordinary case there is nothing except cable between the subscriber and the program exchange and the cable is therefore available to carry signals in the inward direction from the subscriber to the exchange as well as outward. The channel used for outward transmission is approximately

3.2 MHz to 9.2 MHz and the next channel above this from 9.2 MHz to 15.2 MHz may be used for inward transmission.

At the point where the subscriber's pair enters the exchange the 9.2-MHz to 15.2-MHz channel is diverted by filters to frequency changing equipment and fed to the busbar system and, if desired, also to the trunk network to serve other program exchanges. In this way the originating subscriber may, by dialing in the normal way, monitor his own program exactly as it will appear to other viewers or, if he wishes, he may continue to view some other program which might be from another subscriber with whom he was in conversation. Their connection would be similar to a videophone connection but without privacy.

VIII. Sound Radio Programs

Sound radio programs may be treated in exactly the same way as the sound accompanying a television signal; that is, they may be transmitted at audio frequencies to serve the "wired" receivers and also on an FM carrier at the usual spacing from an unmodulated luminance carrier so that they become available through the loudspeaker of an ordinary television set. Such a system is, however, rather wasteful both of television channels in the exchange and of the life of ordinary television sets. The alternative is to transfer the sound programs to a series of channels at the usual 400 Hz spacing within one of the 3.2-MHz to 9.2-MHz vision channels. In order to receive sound programs, the subscriber would dial for this particular channel and an inverter would lift the frequencies to the usual FM band for reception by a standard FM receiver. An alternative possibility is to use the control pair with a secondary and cheaper type of selector switch in the exchange with audio frequency transmission to simple amplifiers and loudspeakers in the home. This would have the disadvantage that the sound programs would be interrupted when the subscriber dialed for a vision program but would have the countervailing advantage that sound programs could be received in one part of the home and television programs in another without the necessity for two separate connections. Stereo programs could be provided by sending the L+R information at audio frequency and the L−R on the normal 19-kHz carrier for decoding in the home.

IX. Subscription Television

It will be clear from the foregoing that the selection of programs offered to particular subscribers may easily be varied and the charges made to them can be varied accordingly. Thus a basic monthly charge might provide all subscribers with all programs carrying advertising. On payment of supplementary charges subscribers could obtain access to additional programs without advertising and thus provide a source of revenue for financing such programs. A scheme of this kind would cost no extra but would not provide subscription television in the full sense that a variable price may be put on particular programs. However, only simple equipment is required to achieve this. The subscription television channels would be connected to the selector switch via reed switches which would be operated over the control pair from each subscriber upon his placing money in a coin box in his home or accepting a debit to his account. The identity of each subscriber being defined in the exchange by his incoming control pair, the recording of the money paid or due from each subscriber amounts to little more than the provision of simple counting equipment attached to each subscriber's pair.

X. Audience Rating

A sample of subscribers is selected by the usual statistical methods and an additional connection made to their control pairs so that a secondary program selection switch operates under the control of the dial selectors in the home. The subscribers in the sample are divided into their socio-economic classes and the secondary selection switches for each class are fed on each program position from a constant voltage source via a recording ammeter. The current taken by each subscriber's circuit is identical so that the total current recorded by the ammeters is a true count of the subscribers in each class connected to each program at any moment. An additional circuit guards against a subscriber being counted if his television receiver is not operating. In this way the sample may be as large as desired, the individuals comprising the sample may be changed from time to time, and, unless they are told, they have no means of knowing that they are included in the sample.

XI. Applications

In addition to ordinary cable television systems for distribution to the home, the system offers a number of advantages in other applications. For example, on a university campus it would be possible to provide every student with a television screen and a dial in his room which would give him access to a new concept of a university library in which the material would be stored in audio visual form. The dial or touch-tone-type control would also enable the student to communicate with the university computer and to receive information back either in alpha-numerical or graphical display on his television screen. Lecture rooms both in the university and possibly at schools in the neighborhood could be similarly equipped at a cost which would be a fraction of that in existing installations of this type based on video frequency distribution over coaxial cable.

Another application in which immediate access is required to large volumes of information is in the financial centers such as Wall Street and the City of London. The system is particularly suitable in these applications because of its open-ended nature with no limitation on the number of separate channels of information which may be offered to subscribers and because of the high technical quality which can be achieved if wired receivers are used exclusively for display. Other possible applications are in military control centers and large fighting ships.

References

[1] R. P. Gabriel, "Wired broadcasting in Great Britain," *IEEE Spectrum*, vol. 4, pp. 97–105, April 1967.
[2] U. S. Patent 3 290 432.
[3] U. S. Patent 3 350 647.

Mr. Lady: Thank you very much, Ralph. We do have time for a few brief questions if we have any. Would you please use the center mike for your questions?

Terry Crawford: Mr. Gabriel, my name is Terry Crawford of the Jerrold Corporation. I wonder if you could give us a brief description of how you compensate for the fact that the loading on any given trunk channel varies with the number of subscribers connected to that channel?

Mr. Gabriel: Yes, each subscriber is stood off the bus bar in the exchange by a number of decibels, but I don't know whether you noticed in the picture of the rotary selector switch each reed is connected through to the central connector by means of a resistor whose value, I'm sorry to say I've forgotten, but the specification to which we work is that the change in level between the condition under which one subscriber has selected a particular program and 100% of the subscribers have selected the same program, shall not exceed 1 dB.

Mr. John Lady: Are there any additional questions?

Question: You had mentioned cross-view talk several times, but you have never mentioned a figure of how far down this cross talk would be and secondly, coming off headends putting multi-channels on, will you have to go to one standard oscillator so everything is phase-locked in so all your carriers seem to keep this cross-talk down at a low level?

Mr. Gabriel: Yes, indeed. All the carriers for every channel are identical and phase-locked. The interference that we are concerned with, therefore, is similar to interference between 2 video signals for which a figure of 45 - 46 dB is generally accepted as adequate and that is the figure to which we work, but there are some things that you can do, in addition, to make cross-view less obvious. We use, for example, precision off-set between the carriers and the adjacent bus bar in the exchange.

Mr. Lady: Thank you very much.

Question: I was curious what the output level is if you are running out of your exchange for handling the 6,000 subscribers out of a common exchange and is there an individual pair run from the subscriber's home to that exchange for each customer? Also, how do you split off internally for your second set connections or multiple set connections on this type of situation?

Mr. Gabriel: There is a separate connection for every television set. The multiple set in the home is a problem and, as I mentioned in my opening remarks, the economy of the system is affected by this. If you want a separate set with separate program selection, it is correct to say that wanting two separate telephones, you must have two separate connections. I have forgotten your first part - -oh, the level, yes. The program exchange is divided into four separate sections. The first section delivers the---carries the highest voltage which with an output of 0.6 volts and is used for feeding the most distant subscribers. Each section then goes down to 6 dB so that the lowest section is used for feeding the nearest subscribers which will, therefore, be at a level of 6.24 dB down or 0.6 volts.

DUAL PULSED-PILOT-CARRIER ALC FOR TEMPERATURE AND COLOR STABILITY OF CATV SIGNALS

O. D. Page
P. E. Treynor

ENTRON, INC.

Introduction and Summary

Automatic level control is required in any CATV system of appreciable length, due to changes in cable attenuation with temperature. Temperature-sensing and single-and dual-CW-pilot-carrier control are discussed, and compared with pulsed or modulated pilot carrier control methods. Pulsed or modulated pilot carrier control is found to be superior, and gives much more stable output levels as amplifier circuitry is subjected to extremes of temperature. Also, Color-Beat interference between pilot carrier and video carrier is eliminated through use of a synchronized pulse generator in the head-end.

Basic Considerations

A problem which is basic to all cable transmission systems is the variation of cable attenuation with temperature. This variation with temperature is primarily due to the temperature coefficient of resistance of the cable center conductor. Most cable used in CATV systems employ a copper center conductor and an aluminum or copper outer conductor. The surface conducting area of the inner conductor is much less than that of the outer conductor; consequently most of the cable loss is due to the inner conductor resistance (current flows essentially on the surface due to skin effect). So the temperature coefficient of resistance of the inner conductor primarily determines the over-all cable attenuation variation with temperature.

For copper the loss at any temperature T (centigrade) is:

(Eq. 1) Loss at temperature $T = A_{20} \left[1 + .0019 (T-20) \right]$ (decibels)

A_{20} = cable loss at 20°C. and given frequency

In a 20-dB span of cable the attenuation will then vary .038 dB per degree C. In terms of degrees Fahrenheit, this variation is approximately .001 dB per dB per degree. Thus, for a temperature variation of 0°F. to 100°F. and a 20 dB span of cable, the attenuation change will be +2 dB. When it is considered that a trunk system can have cable spans of several hundred dB (with a 10 dB attenuation variation per 100 dB for a 100°F. temperature variation) it is obvious that a means of stabilizing levels must be found. Otherwise signals will overload amplifiers in cold weather and will drop into the noise level in hot weather.

Temperature-Sensing Level Control

The problem of level variation with temperature has been minimized in some amplifiers by employing compensating devices in the amplifier. A temperature sensitive resistance in the amplifier has been used to reduce the amplifier gain as the temperature is reduced and thereby at least partially compensates for the decrease in cable loss. There are three factors which limit the precision of this form of open loop temperature compensation. First, it is virtually impossible to insure that the temperature compensation which has been designed into the amplifier will match that of the cable with the necessary precision. Second, the average cable temperature may differ from that of the amplifier. This occurs, for example, on a hot summer day, when the CATV amplifier is shaded by a building or a tree. Third, level errors are cumulative; each amplifier does not correct for the errors in output level of the preceding amplifiers. Cascadibility of amplifiers using this type of "Automatic Level Control" is severely limited.

Composite AGC

Many amplifiers, mostly the earlier tube-type, utilized a broadband composite AM detector to sense total energy (all channels) transmitted through the amplifier. The output of this detector was used to bias amplifier tubes or drive an attenuator to change gain of the amplifier to maintain an essentially constant output. These amplifiers work reasonably well, and many are still in use today. However, precision level control, needed for very long cascades, cannot be achieved over the band because of the constant changes in energy content of the individual signals, and because certain signals are not always present. Precision control of amplifier response also presents a problem in these systems.

C-W-Pilot-Carrier Level Control With Single Detector

Another technique now widely employed is to apply stable signals to the trunk line at the head-end, and then use closed-loop control in each amplifier to hold the levels of these signals constant. Frequencies of 73.5, 116, 165, 225 MHz and others have been employed by different equipment manufacturers. Small narrow-band amplifiers (receivers) in the trunk line units provide an automatic-level-control voltage which is used to vary the gain and slope of the broadband trunkline amplifier. This control voltage is derived by rectifying the pilot carrier.

As the cable temperature drops, and the cable loss decreases, the pilot-carrier level input to the amplifier increases. This increase is amplified and rectified by the pilot-carrier amplifier. This rectified voltage is applied to a voltage-controlled attenuator in the trunk amplifier which reduces the gain of the trunk amplifier. Pilot-carrier level control acts to maintain a constant level of the pilot-carrier at the trunk amplifier output under all conditions. This method of control is therefore essentially independent of temperature differentials between the cable and the amplifier. Further, each amplifier tends to correct for any error in the output of the previous amplifier.

The loss of a given length of coaxial cable varies as the square root of frequency. A cable with 20 dB loss on channel 13 will have only 10.4 dB loss on channel 2 (20 dB $\times \sqrt{\frac{54}{211}}$ = 10.4 dB). From equation 1 it is found that for this cable length, the attenuation of channel 13 will vary by 2 dB but the attenuation of channel 2 will vary only by 1.04 dB as the temperature varies from 0°F. to 100°F. This points up a major problem in the AGC of CATV amplifiers: if a single pilot carrier is employed at midband for AGC, only the midband will be held by a <u>closed loop</u> system, the high and low frequencies can only receive a compromise (non-closed-loop) correction, which will always be subject to differences in cable and amplifiers. These small differences can be cumulative, on long cable runs. This correction, usually called "tilt-compensation" is often accomplished by applying the same control voltage also to a varicap in one or more of the amplifier stages, so that channel 13 gain is changed approximately twice as much as channel 2 gain.

When a single pilot carrier is employed for ALC in a system, it is desirable to employ the highest practical frequency. This is true for the following general reasons: (1) A high-frequency signal undergoes a greater level variation for a given temperature variation and therefore, the ALC circuits receive a greater change on which to operate. (2) A single pilot system will typically provide 1/2 dB amplifier level stability <u>at the pilot frequency</u> over the full range temperature. With only a high-frequency pilot, at the high end of the channel the pilot signal may be held constant to within 1/2 dB. With high frequency pilot control the low frequency (channel 2) gain will vary by approximately 1/2 dB for each 1 dB of ALC at the high end for correct cable compensation. If the ALC of a given amplifier is 1/2 dB due to temperature instabilities on the high end (channel 13), it is typically only 1/4 dB off on the low band. Conversely, with a low frequency pilot, at the low end of the band each 1 dB variation in gain produced by the ALC at low band must produce a 2 dB variation at high band to provide compensation of the cable. If the ALC provides the same stabilization as above, within 1/2 dB, at the low frequency pilot, the high band levels may be off by 1 dB. Thus, for a 1/2 dB pilot ALC stability, a high-frequency pilot will typically allow a maximum variation of 1/2 dB while a low-frequency pilot might allow a 1 dB maximum variation.

The most practical ALC method is to employ two pilot carriers, one at high band and one at low band. Any tilt variations caused by cable characteristics, amplifier characteristics, or passive-device characteristics will be automatically corrected by the dual-pilot-carrier amplifier, which is normally placed at every 4th to 6th station, with single-pilot-carrier (tilt compensated) amplifiers in between.

<u>Selection of Pilot Carrier Frequencies</u>

The frequencies of these carriers should be relatively close to the TV carriers but not spaced so as to interfere with the TV signals, or to allow TV signals to affect the operation of the pilot carrier and receiver.

In a conventional, standard 12 channel system the following frequencies are available: (1) below 54 MHz (channel 2); (2) between channels 4 and 5 (a band from 72 to 76 MHz is available between these channels); between the FM band and channel 7 (108 to 174 MHz); (3) above channel 13 (above 216 MHz).

In more-than-12-channel systems, the frequencies available for pilot carriers are dependent upon the system type. For mid-band systems which carry additional channels in the frequency range of 108 to 174 MHz, the following ranges are available: (1) below 54 MHz; (2) between channels 4 and 5; (3) above 216 MHz. For systems which apply additional channels above 216 MHz (1) frequencies below 54 MHz are available; (2) the band between channels 4 and 5; (3) the range between 108 and 174 MHz; (4) above the highest-frequency carrier.

The frequencies below 54 MHz and the range between channels 4 and 5 are available for a low frequency pilot carrier in all systems. For 12-channel systems and 20-channel mid-band systems a high pilot carrier above 216 MHz is optimum and for 20-channel high-band systems the high pilot should be above 270 MHz. On the basis of these considerations the following pilot frequencies are desirable:

	Standard 12-Channel System	20-Channel Midband	20-Channel High Band
Low Pilot	73.5 MHz	73.5 MHz	73.5 MHz
High Pilot	225 MHz	225 MHz	225 MHz

A low pilot of 73.5 MHz is chosen first because it provides better ALC stability than a pilot carrier below 54 MHz. Most existing amplifiers do not extend below 50 MHz in frequency response, and this low-frequency range is now being used for 2-way transmission. Further, 73.5 MHz has been widely employed by several manufacturers, and problems associated with this frequency have been solved (this will be discussed later). A pilot carrier at 225 MHz is field-proven and is again selected for systems with frequencies below 216 MHz. For high band systems with frequencies up to 270 MHz, somewhat better ALC stability is obtained if the pilot carrier is moved nearer the top of the band (between 270 and 280 MHz for example).

Pilot Carrier/Video Color Interference

The problem which has occurred in the field with a 73.5 MHz CW pilot carrier has been due to a beat product between the 73.5 MHz and the channel 5 picture carrier. This third-order beat occurs in the TV receiver tuner and is usually apparent only in color TV receivers which have poorer third order intermodulation performance. Sensitivity adjustment in the color-carrier circuit also has some effect. These receivers usually have solid state RF stages and mixers. The relationships between carriers are as follows:

(a) Channel 5 Video Carrier = 77.25 MHz

(b) Major spectral component at channel 5 video = ±N (.015750) MHz where N = any integer

(c) Third order beat = $2 F_1 - F_2$

(d) $2F_1 - F_2 = 2[77.25 \pm N(.015750)] - 73.5$

$= 81.00 \pm 2N(.015750)$

(e) Ch. 5 color carrier = 80.829545 (not-offset)

(f) For N = -5; $2F_1 - F_2 = 81.000 - 10(.015750)$ (or if pilot-carrier is off by approximately 150 kc)

$= 80.842500$

Thus the products (f above) fall within 15 kC of the color subcarrier (e). This spurious signal asynchronously aids or retards the resultant phasor formed with the color subcarrier and produces a color-bar pattern on the TV screen.

This interference problem is eliminated by using a pulsed pilot carrier at 73.5 MHz which is synchronized to the horizontal blanking pulses of channel 5 (patented). The 73.5-MHz carrier is applied for only 3 microseconds during the horizontal blanking interval of ch. 5. A block diagram of this synchronized RF pulse generator is shown in Figure 1.

Use of a pulsed pilot carrier negligibly affects the ALC circuits of the trunk amplifiers. Since the RF pilot is on for only 3 microseconds and is off for 60 microseconds, the ALC detector in the trunk amplifier requires a rapid rise time (less than 3 microseconds) and a long fall time (greater than 60 microseconds) to assure efficient operation on the pulse. This requires only a minor modification, if any, on most existing trunkline amplifiers.

Trunk-Amplifier ALC Circuit Considerations

We have seen that a pulsed pilot carrier is required at 73.5 MHz, and that a CW pilot may be employed at 225 MHz.

Consider now the conventional ALC circuitry employed in trunk amplifiers, as shown in the block diagram of Figure 2. The trunkline input is amplified by the trunk amplifier and is applied to a power splitter. The power splitter passes most of this signal to the line output but provides a low level output (typically 20 dB below the line output) to the band pass filter. This filter is typically a 2-resonator type with a band pass of approximately 1 MHz in a 73.5 MHz pilot amplifier and 2.5 MHz in a 225 MHz pilot amplifier. Insertion loss is typically 8 dB.

For a 73.5-MHz pilot the input level to the band pass amplifier is approximately 0 dB mV and because the detector requires a 1 volt input the RF gain of this bandpass amplifier must be high, typically 60 dB.

The dc output from the detector is applied to a dc amplifier and this amplifier in turn drives the ALC attenuator of the trunk amplifier to provide closed-loop gain control. For dual-pilot carrier control, an identical second loop supplies a voltage to drive a frequency-sensitive-device such as a varicap.

High RF gain in the ALC amplifier in this method is required to minimize temperature drift in the dc output of the detector. The dc voltage drop across a conducting germanium semiconductor diode is typically .3 volts but this drop decreases by approximately 2 millivolts per degree centigrade. If, for example, the detector of figure 2 receives a 0.5 volt peak RF signal, the rectified output will be only .2 volt due to the diode drop. If the temperature drops by 50°C. then the output voltage will decrease to (.2 - .002 x 50 = 0.1 volt). This in itself will cause a 6 dB ALC instability, which is intolerable and difficult to compensate. This effect is minimized by applying a large signal to the diode. Thus with 1 volt (rms) applied to the detector, a dc voltage of typically 1.1 volts dc is developed. This decreases to 1.0 volts for a 50°C. temperature decrease (1.1 - .002 x 50 = 1.0). This causes an ALC instability (of only 1 dB) which can be further reduced by compensation. The requirement of a high RF level must be met whenever a highly-stable detected dc voltage is needed. This means that a high-gain bandpass amplifier like that of Figure 2 must be incorporated in the trunk amplifier. Unless this amplifier is carefully bypassed and shielded, leakage of its signal can produce peaks in the response of the trunk amplifier. If the preselector is not carefully aligned, adjacent TV signals can also affect ALC levels. If the preselector is not temperature stable, its loss will also vary with temperature, and further affect ALC levels.

Pulsed Pilot Carrier ALC with Double Detector

We now consider an ALC system which is used with a pulsed pilot carrier. This system offers advantages of simplicity, excellent performance and flexibility. Use of this circuit requires a pulsed RF pilot carrier. Both 73.5-MHz and 225-MHz pulsed pilot carrier are easily provided. The ALC circuit of Figure 2 may be employed in the trunk amplifiers with only minor modifications for either a CW pilot or a pulse pilot carrier.

The pulsed pilot ALC system is presented in Figure 3. A portion of the trunkline output is applied through a power splitter to the ALC section. Note that this signal is applied to an isolation amplifier which precedes the bandpass filter. This isolation amplifier provides a broad band matched input impedance. When the low level output of the power splitter of the conventional ALC of Figure 2 is applied to the bandpass filter, a small ripple, on the order of .05 dB, is produced in the over-all trunk amplifier response. This dip or valley is due to the input impedance characteristic of this filter. For frequencies above and below its bandpass it is a high impedance. For frequencies within its bandpass, it is a low impedance and although hardly apparent with one amplifier, a ripple of 0.5 dB or more can occur at the pilot frequency when ten or more amplifiers are cascaded, causing the pilot-carrier level to drop off accordingly. The broadband, matched isolation amplifier of Figure 3 minimizes this problem.

Output of the bandpass filter is applied to the low-level RF detector. The output of this detector is AC-coupled to the high-gain video amplifier. The isolation amplifier has a gain of typically 15 dB; hence the signal applied to the RF detector is only 15 dBmV.

Based on the previous analysis of detector performance with temperature, it would appear that very unstable detector performance will be obtained. This is not the case because this RF detector provides an AC (pulse) output rather than DC. It has been found that a detector will deliver a very constant AC output over a wide temperature range even with low RF input levels, although the DC output will under these conditions vary widely. This is made apparent by the curves of Figure 4, which show that the volt-amp diode characteristic translates vertically as temperature is increased but the slope at any given input voltage is essentially unchanged. At a CW pilot level Vp, the DC voltage output of the detector increases by ΔV as the temperature increases from 0^{o}F. With a pulsed pilot the peak to peak pulse amplitude is virtually unchanged as temperature increases from 0^{o}F. to 100^{o}F. It is from this peak-to-peak pulse that the ALC circuit of Figure 3 operates.

The AC output from the low-level RF detector of Figure 3 is a pulse which is stable with temperature. This low-level pulse is AC coupled to a high-gain video amplifier which provides 70 dB gain. This 70 dB gain is easily obtained in a single integrated circuit package; 70 dB of video gain, in the range of 15 to 200 KHz can be obtained with smaller, less expensive circuitry than can the 60 dB of RF gain of Figure 2 (73.5 or 225 MHz). Output from the video amplifier consists of video pulses with typical peak-to-peak amplitudes of 2 volts. These high-level pulses are applied to a peak detector which provides a temperature-stable dc output voltage to the dc amplifier which in turn drives the trunk amplifier attenuator and/or varicap to provide closed-loop gain and slope control of the amplifier.

The arrangement of Figure 3 has several advantages over the conventional circuit of Figure 2 in addition to elimination of color beat in Channel 5:

(1) The unit of Figure 3 can be made physically smaller because 70 dB of video gain can be obtained in a single integrated circuit. The 60 dB of RF gain in Figure 2 requires much more space because of the inherent coupling problems at VHF, and the fact that this gain cannot be obtained in a single integrated circuit. (2) The configuration of Figure 3 is less costly than that of Figure 2. (3) The pilot carrier frequency can be easily changed. The circuitry of Figure 3 following the bandpass filter is always the same regardless of the pilot carrier frequency. The only elements of 3 which change with pilot carrier frequency are those of the bandpass filter which can be easily modified to operate at any pilot frequency. The circuit configuration of Figure 2 requires that both the bandpass filter and bandpass amplifier be modified for major pilot carrier frequency changes.

Trunk Amplifier ALC Stability

Initially, the trunk amplifier output stability will be considered for the pilot carriers. Then, stability over the total operating frequency range will be considered, since true "level control" is accomplished only when all levels remain stable.

The output level of an ALC trunk amplifier is a function of two parameters. One of these is the absolute temperature stability of the ALC system, which is essentially independent of the input signal level and is only a function of temperature. The other is the closed loop ALC gain. These relationships may be assumed linear for small signal changes. Consider the following relationship:[1]

(Eq. 2) $$\Delta E_o \text{ (dB)} = \frac{\Delta E_{in} \text{ (dB)}}{1 + G} + K \Delta T$$

where ΔE_o = dB change in output trunk level at the pilot frequency

ΔE_{in} = dB change in the trunk input level

ΔT = temperature change, °F

K = an ALC stability constant with dimensions of dB per degree F°

G = ALC gain constant

$= G_1 G_2 G_2$

where $$G_1 = \frac{\text{voltage change at final detector output}}{\text{dB change in trunk amp output}} = \frac{\text{volts}}{\text{dB}}$$

G_2 = D.C. gain of amplifier following final detector

$$G_3 = \frac{\text{dB change of attenuator in trunk amplifier}}{\text{voltage change of ALC}} \qquad \frac{\text{dB}}{\text{volt}}$$

In designing a trunk amplifier ALC system the variations in output level (ΔE_o) can be made very small for given variations in amplifier input (ΔE_{in}) by making the gain constant (G) arbitrarily large. This may be done by having a high dc amplifier gain (G_2) a high RF or video gain (G_1), or a very sensitive ALC attenuator, G_3. Good design requires a balance and optimum distribution of these gains. For the level variations encountered at the input of a trunk amplifier (± 4 dB or less), the gain constant can be easily made sufficiently large so that a ± 4 dB input variation produces only ± .1 dB output variation. Thus the first term of equation 2 does not represent a major design problem, and CATV amplifiers have been available with very high gain constants to hold output variations within very small limits, at constant ambient temperatures.

If the input level to an ALC controlled trunk amplifier is held constant, and the amplifier environment is varied from typically -40°F. to +140°F., it will be found that the output level of many CATV trunk amplifiers will significantly vary. The ALC system can handle input variation (G is high) but is basically temperature unstable (K is high), and the output level will be held only as constant as the ALC gain remains constant.

To hold the trunk amplifier output constant to 1 dB requires individual amplifiers and filters to be stable to a fraction of a dB over a wide temperature range. And when the temperature coefficient of semiconductors and economically-feasible passive components are considered, these are extremely tight limits. The most practical solution is to produce a design in which the various temperature coefficients cancel each other.

If, for example, the isolation amplifier gain of Figure 3 rises by .25 dB, the bandpass filter insertion loss increases by .25 dB, the low level RF detector and video amplifier gains increase by .5 dB for a temperature change of -30 to +150°F., then the trunk level will vary by .5 dB (+.25 - .25 +.5) and with additional small compensatory elements, this variation can be further decreased.

Unfortunately, the temperature coefficients of the elements of the ALC circuits of trunk amplifiers are not all the same. Thus, in 100 IC video amplifiers, 80 may have a gain characteristic which increases an average of 1 dB as the temperature changes from -30 to +150°F.; 15 may have a gain characteristic which increases by only .5 dB over this temperature range; and 5 may have a gain characteristic which decreases by a .5 dB over this range.

The mean and variances of the statistical distribution of the temperature coefficients of the ALC system must be considered in ALC design. The over-all system noise figure and cross modulation characteristics vs. temperature are primarily determined by the mean ALC characteristics. Thus, if in a 20-amplifier system the output of all amplifiers but 1 increase by .5 dB as temperature increases, and in only one amplifier the output increases by 2 dB, the over-all system cross modulation will increase by 1 dB (2 x .5) and will be negligibly affected by a 2 dB rise in one amplifier (assuming the amplifier remains in a well-behaved X-mod range).

The final factors to be considered are the effects of temperature variation upon the insertion loss and tuning of the pilot carrier bandpass filters of Figures 2 and 3. These filters employ high-Q tuned circuits which are very lightly loaded to provide narrow bandwidth and high rejection to the TV carriers. As a result of the high Q and narrow bandwidth, small variations in L and C can detune the filters. These variations can be caused by both temperature changes and high humidity.

Humidity can be a significant problem in a sealed trunk amplifier if, for example, the amplifier is opened on a warm humid day and then closed. The air in the amplifier now contains significant water vapor. The passive tuning elements (capacitors and inductors) generate no heat. As outside temperature drops, the warm humid air in the trunk amplifier around these components deposits moisture droplets on them when the dew point is reached and these moisture droplets significantly lower Q and vary capacitance in this VHF range.

Tests performed on numerous glass, air, and ceramic trimmers demonstrated this to be a very significant problem, and the best solution was found to be to use sealed fixed capacitors and tune the circuits by varying the inductance.

The inductors employed in these filters are air wound with heavy wire and as temperature rises this wire expands slightly and this expansion increases the inductance slightly. This effect is neutralized by employing sealed tuning capacitors with a slight negative coefficient of capacitance (typically 15 parts per million per degree C.) The capacitance decreases slightly as the inductance increases, and negligible detuning occurs.

The final temperature effect is a slight decrease in Q of the filters as coil resistance increases at higher temperature. This produces a slight increase in insertion loss, which is compensated in the video amplifier.

Level Control at Frequencies Other Than Pilot-Carrier

At frequencies other than those of the pilot carriers, the output level at frequencies other than the pilot carriers is determined partially by the temperature coefficients of frequency response of the trunk amplifier. With a single 73.5-MHz pilot carrier in an ideally-tilted amplifier, and a 0.5 dB ALC stability, the stability at 216 MHz is approximately 1 dB. It was shown earlier that use of a high-frequency pilot reduces the error, which is further reduced with dual pilot control.
A stable temperature characteristic over the full band requires a temperature-stable basic trunk amplifier response in addition to a temperature-stable ALC system.

Temperature Variations in Trunk System Performance

Consider first a perfect trunk ALC system in which the output of all amplifiers is held absolutely constant. With 20 dB spacing between amplifiers the levels below are obtained (full tilt system with 20 dB (cable) spacing on channel 13 at 60°F.):

TABLE 1 - "Perfect" ALC Stability

	Output Level		Input Level	
Temperature	Ch. 2	Ch. 13	Ch. 2	Ch. 13
- 30°F	25 dBmV	+35 dBmV	+15.9 dBmV	+16.8
+ 60°F	+25	+35	+15 dBmV	+15 dBmV
+130°F	25	+35	+14.3	+13.6

TABLE 2 - Cable Characteristics

	Cable Loss	
Temperature	Ch. 2	Ch. 13
- 30°F	9.1 dB	18.2 dB
+ 60°F	10. dB	20.0 dB
+130°F	10.7	21.4 dB

From the output levels of Table 1 we can say that there will be no change in the system cross modulation as temperature is varied (based upon the practical assumption that system cross modulation (X mod) = $K_1 + 2\Delta E_o$ (dBmV). From the input levels of Table 1 we see that the system noise on channel 13 will decrease by 1.8 dB as the temperature drops from

+60°F. to -30°F. Conversely as the temperature increases from +60°F. to 130°F. the system noise rises by 1.4 dB.

The "Perfect" ALC system described by Table 1 is characterized in equation 2 by the two constants G and K. For perfect ALC, G = 00 and K = 0. We next consider a less perfect system in which G = 00 and K = -.01 dB per F$^\circ$. This system is characterized by Table 3 below.

TABLE 3 - System Levels for K = -.01, G = 00

Temperature	Output Level		Input Level	
	Ch. 2	Ch. 13	Ch. 2	Ch. 13
- 30°F.	+25.9 dBmV	+35.9 dBmV	+16.8	+17.7
+ 60°F.	+25	+35	15	15
+130°F.	+24.3	+34.3	+13.6	+12.9

From this table we see that at the low system temperature over-all cross modulation increases by 1.8 dB but noise figure drops by 1.8 dB (Ch. 13). At high temperature system cross modulation drops by 1.4 dB but system noise figure increases by 1.4 dB (at Ch. 13).

Table 4 below shows system levels for a stability constant of +.01 dB /F$^\circ$.

Table 4 - System Levels for K = +.01, G = 00

Temperature	Output Level		Input Level	
	Ch. 2	Ch. 13	Ch. 2	Ch. 13
+ 30°F	24.3	34.3	15.2	16.1
+ 60°F	25	35	15	15
+130°F	25.9	35.9	15.2	14.5

Table 5 below summarizes the performance of the systems of Tables 1, 3, and 4.

System	Change in System Noise Figure		Change in System Cross Mod.	
	Ch. 2	Ch. 13	Ch. 2	Ch. 13
K = 0	- .9/+ .7	-1.8/+1.4	0/0	0/0
K = -.01	-1.8/+1.4	-2.7/+2.1	+1.8/-1.4	+1.8/-1.4
K = +.01	- .2/- .2	-1.1/+ .5	-1.4/+1.8	-1.4/+1.8

Parameter at -30°F./Parameter at +130°F.

From Table 5 it is apparent that ALC instabilities cause variations in system cross modulation, and the variation in cross modulation is independent of the drift direction (i.e. K = +.01 or .01). However, it is seen that the system noise figure changes less over the temperature range for K = +.01, that is, if the amplifier output level increases with cable attenuation.

We conclude that the ALC system should allow minimum output level variation and the statistical mean output of a group of amplifiers should be slightly positive for increasing temperature.

Conclusion

Use of dual-pulsed-pilot-carrier ALC control in CATV transmission system amplifiers provides greatly increased stability in the system, besides eliminating interference between the pilot carrier and TV channels. Careful design, test, and selection of components, with temperature compensation, provide for output level stability at all temperatures.

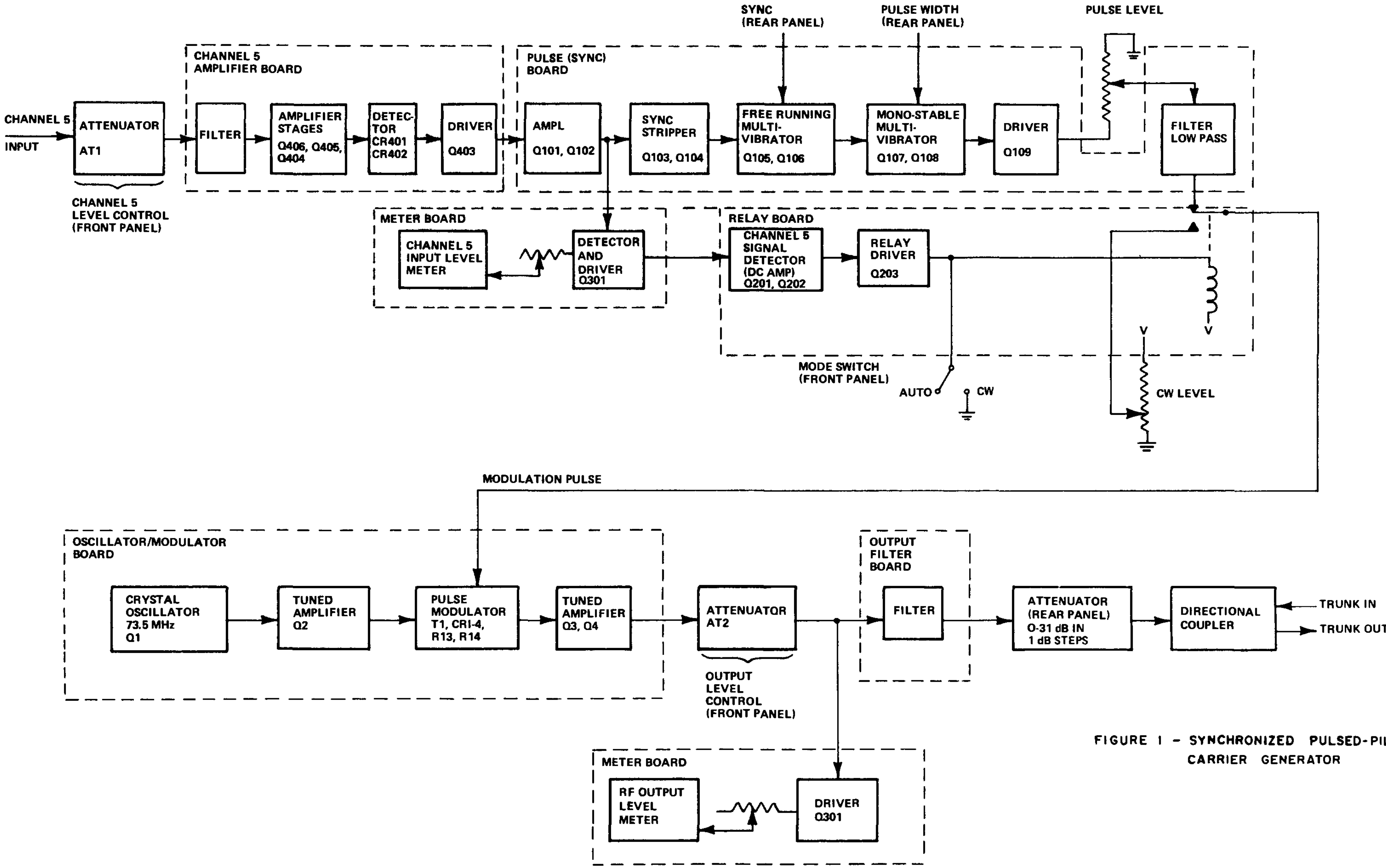


FIGURE 1 - SYNCHRONIZED PULSED-PILOT-CARRIER GENERATOR

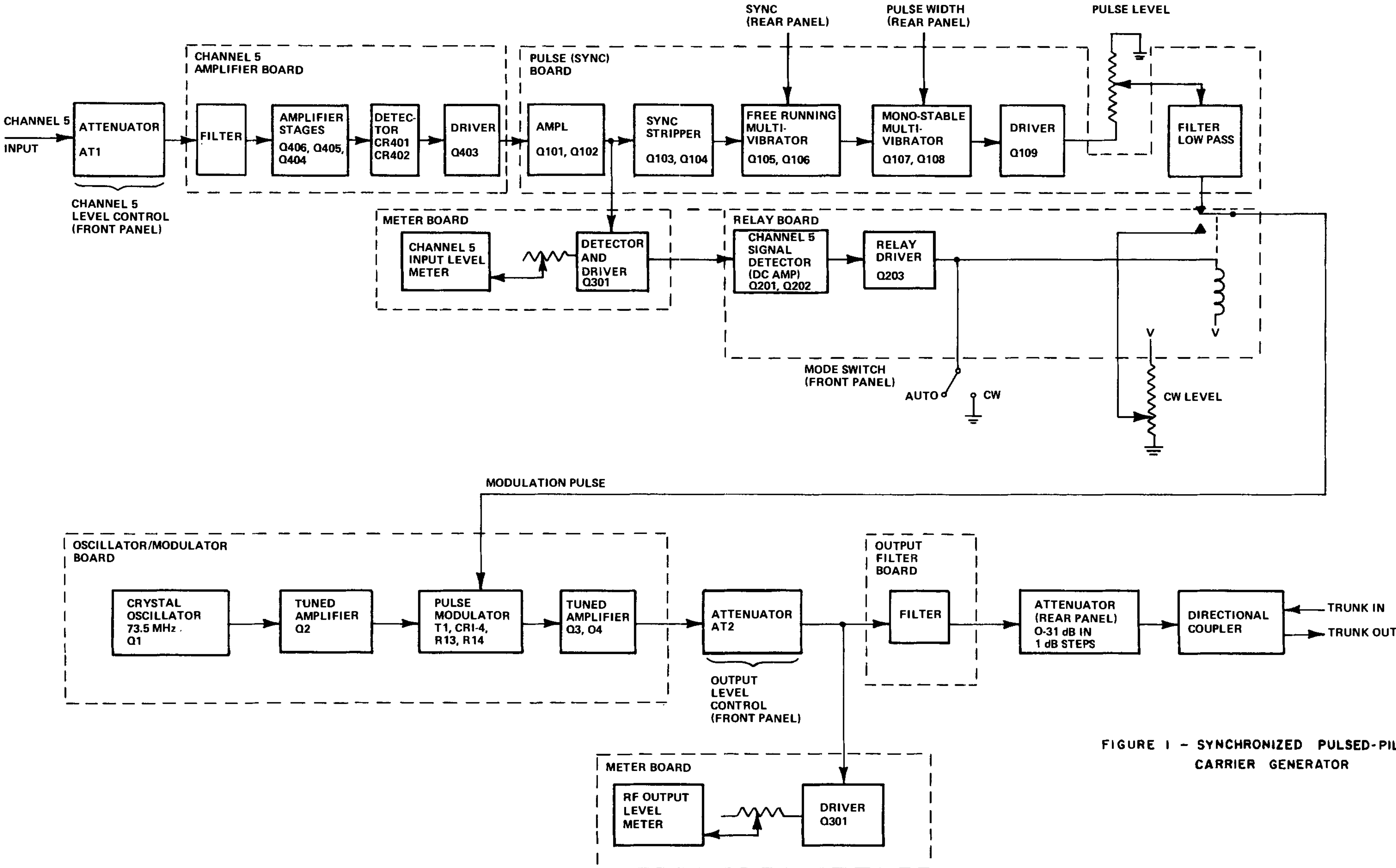


FIGURE I - SYNCHRONIZED PULSED-PILOT-CARRIER GENERATOR

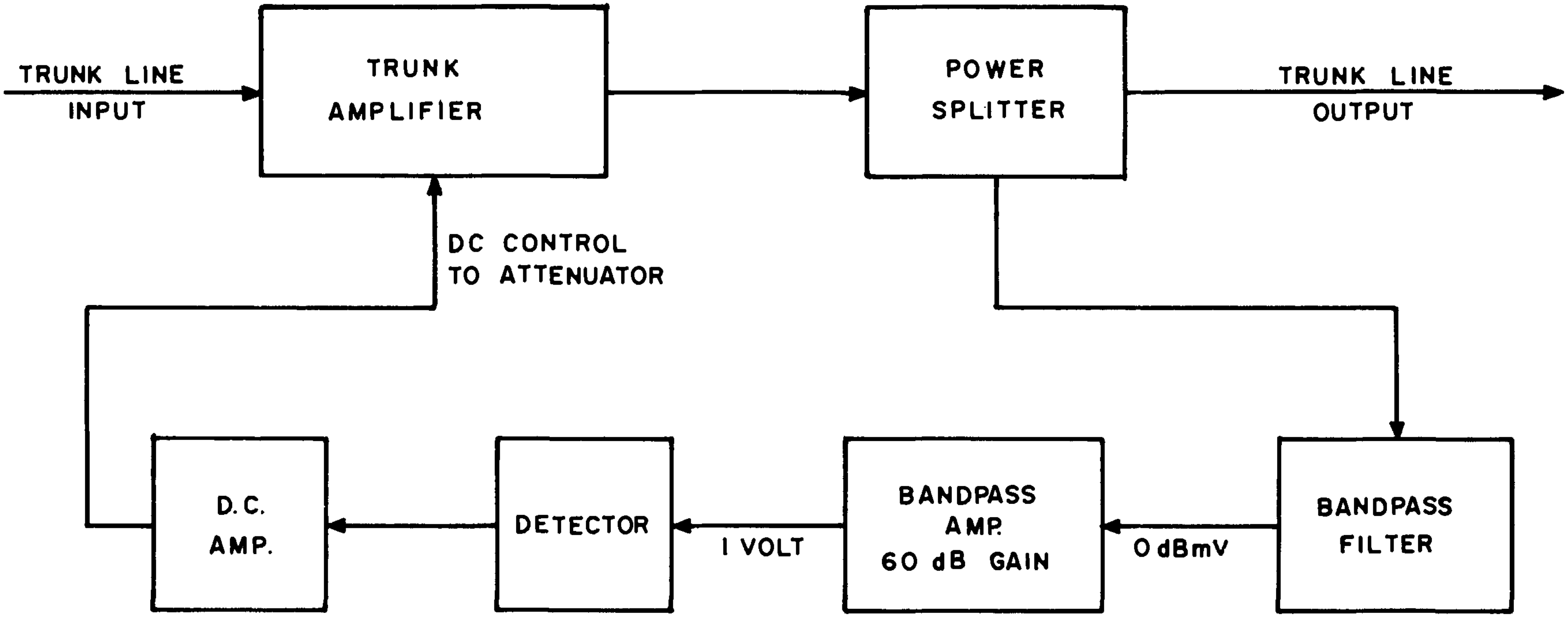


FIGURE 2 - CONVENTIONAL CW-PILOT-CARRIER TRUNK AMPLIFIER ALC

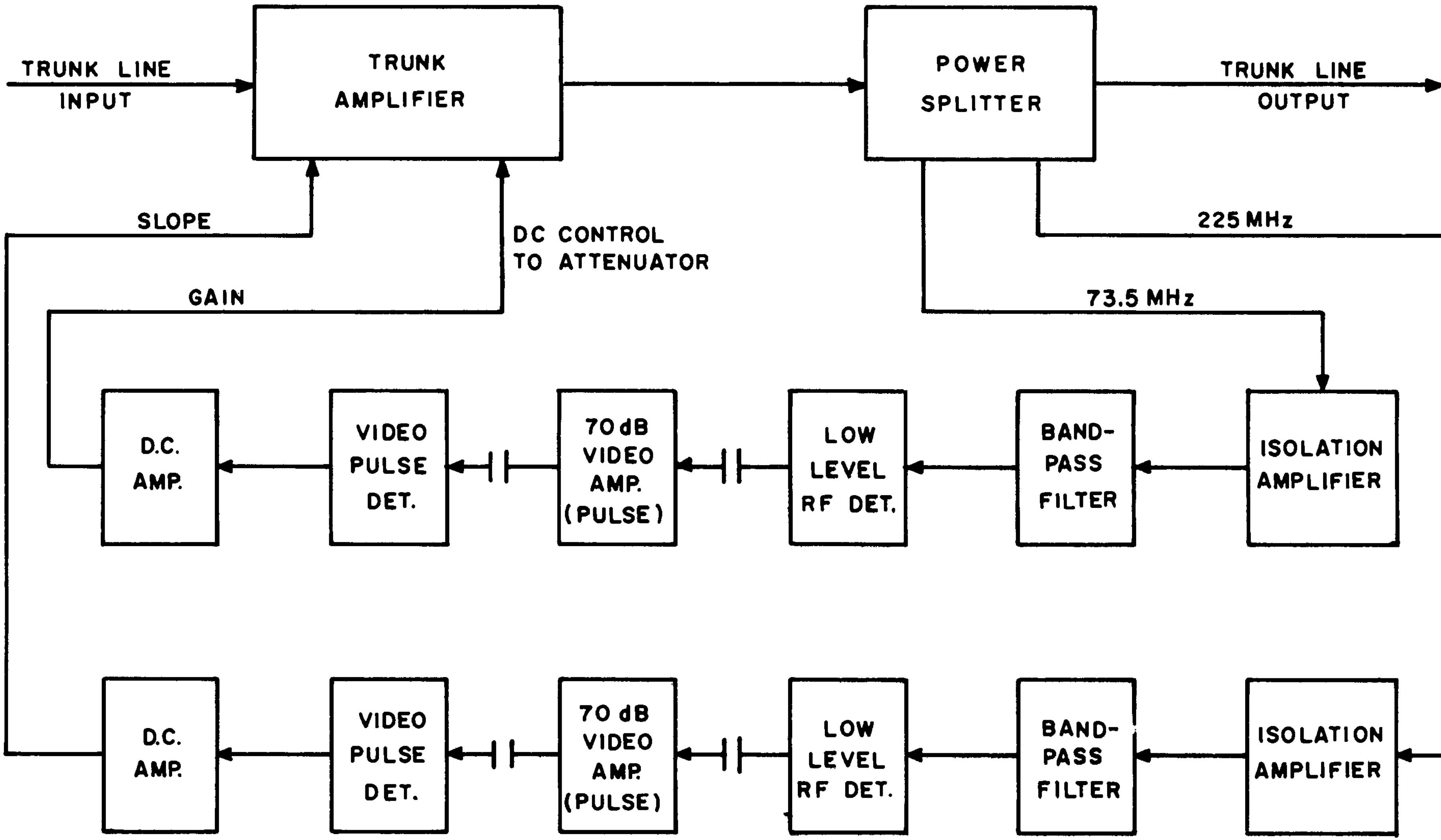


FIGURE 3 – PULSED-PILOT-CARRIER ALC

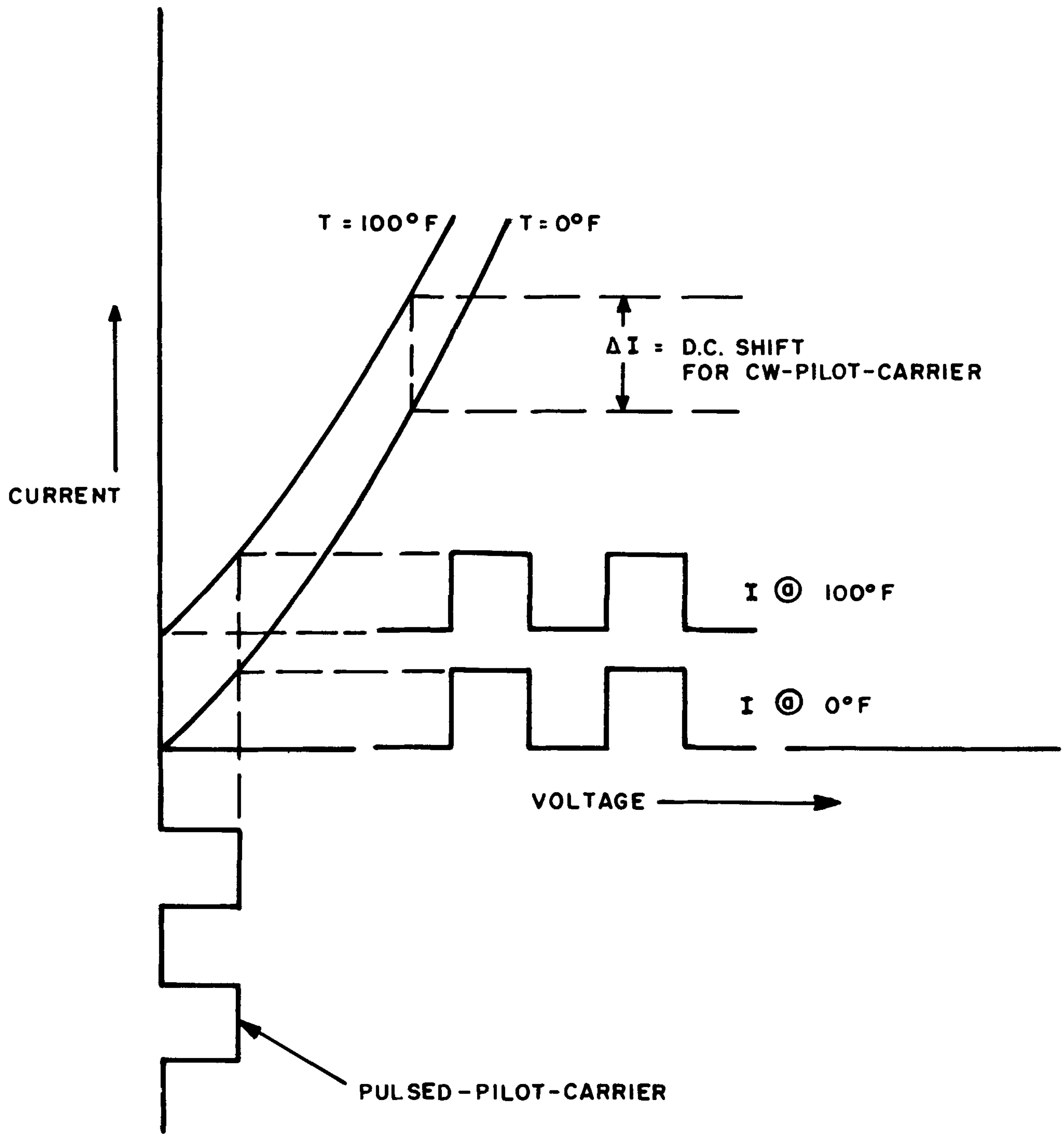


FIGURE 4 — DIODE DETECTOR CHARACTERISTICS

References:

(1) Victor, W. K., and M. H. Brockman. "The Application of Linear Servo Theory to the Design of AGC Loops". Proceedings of the IRE, Vol. 48, February, 1960, pp. 234-238.

Referring to the last equation on page 236 of this article, if we let $t \rightarrow \infty$, the steady-state expression for change in receiver (amplifier) attenuation $a^*(t)$ dbv $= \dfrac{\Delta a_{db}}{1 + \frac{1}{G}} = \dfrac{G \Delta a_{db}}{1 + G} = \dfrac{G \Delta E_{in}}{1 + G}$

(Δa_{db} in the change in input signal level $= \Delta E_{in}$)

$\Delta E_{out} = \Delta E_{in} - a^*(t)_{db}$ in dB, for small signal levels, and, substituting for $a^*(t)_{db}$

$$\Delta E_{out} = \Delta E_{in} - \frac{G \Delta E_{in}}{1 + G}$$

$$= \frac{\Delta E_{in} (1 + G - G)}{1 + G} = \frac{\Delta E_{in}}{1 + G} \quad \text{(dB)}$$

To which we add an arbitrary temperature-variable factor $K \Delta T$ to represent output level changes caused by temperature instability of the AGC Loop:

$$\Delta E_{out} = \frac{\Delta E_{in}}{1 + G} + K \Delta T \quad \text{dB}$$

EFFECTS OF NONUNIFORM COAXIAL CABLE ON CATV SIGNAL QUALITY

W. L. Roberts and F. N. Wilkenloh
Superior Continental Corporation
Continental Telephone Laboratories
Hickory, North Carolina

Variations in impedance of transmission lines are known to cause degradation of propagated signal quality. Abrupt changes in impedance can cause discrete ghosting to appear in television signals while random combinations of many small impedance changes generally contribute to the overall noise level of transmitted signals. Variations of impedance along the length of a line which recur periodically, seriously disturb only transmitted signals located in a relatively narrow band of frequencies.

The effects of echoes and noise on television picture quality have been reported in the literature over the past 20 years.[1,2,3,4] Cable impedance irregularities have been studied and mathematical relationships between discrete impedance variations and transmission properties have been published during this period.[5,6,7,8,9,10]

Direct relationships between cable impedance irregularities and actual television picture quality are quite complex since different, specific types of irregularities can result in seemingly unrelated forms of picture degradation. Echoes caused by signal reflection from a particular point along a cable can be calculated readily if the reflection coefficient and propagation characteristics of the cable are known. Experiments dealing with ghosting of this type and the effects on CATV systems were reported by Shekel.[11] Effects of randomly related echoes were studied by Ashcroft et al.[12] The direct influence of cable periodicities on picture quality does not appear to be well publicized in the literature and, indeed, may not be well known.

Presented in this paper are results of observations of a monochrome television picture transmitted over coaxial cables disturbed by impedance periodicities. Systematic impedance variations were introduced at controlled levels during the cable manufacturing process. The resonance frequency of the periodicity selected was approximately 83 MHz, a convenient frequency for use with the selected test apparatus. "Equivalent" echo ratings were assigned to the various observed conditions. Similar comparisons were made on signals transmitted through resonant networks. Application of results to CATV system requirements is discussed.

Theoretical return loss and attenuation discontinuity relationships predicted by Fuchs and Peltier[13] are experimentally confirmed as an interesting corollary to this work.

Preparation of Experimental Periodic Cables

A 15.1-inch diameter (47.8-inch circumference) sheave was trued and balanced and equipped with adjustable unbalancing weights. The sheave was installed in the coaxial cable extrusion line, wrapping the bare inner conductor around the sheave at a location prior to entering the extruder crosshead. Introduction of unbalance caused a nearly sinusoidal core diameter variation. A solid polyethylene dielectric and braided outer conductor were employed to enhance the possibility of verifying calculated impedance deviations in terms of cable dimensional variations. Expected values were calculated in accordance with Fuchs' and Peltier's predictions[14] (see Appendix). The braid assured that the core contour was "tracked" by the outer conductor. Diameter and capacitance variations were recorded during extrusion.

Six controlled periodicity levels were produced with a 0.146-inch diameter coax core (nominal $Z_o = 75$ *ohms*). In order to emphasize effects from these periodicities, return loss levels as poor as 9 dB with $\Delta\alpha/\alpha$ discontinuities of 25% were produced. Nominal sections of 40-dB loss at 83 MHz were fabricated (approximately 1300-foot lengths). For comparison to a cable with a different attenuation constant, a similar 40-dB length of 0.285-inch core diameter cable was prepared with a 30% attenuation discontinuity at 82 MHz.

As may be noted in Figures 1 and 2, experimental results agree remarkably well with calculated values for both return loss and concomitant attenuation discontinuities.

Test Method

Equipment connections are shown in Figure 3. All observations were based on viewing a 12x16-inch raster from a distance of 48 inches (in near total darkness). Experience gained through attempts to characterize the picture impairment led to use of three different viewing subjects: (1) a conventional monochrome test pattern, (2) a picture consisting of three vertical lines of approximately 1/8, 3/16, and 1/4 inch as measured on the viewing screen (about 4, 6, and 8 picture elements duration) spaced equidistantly across the field, and (3) a portrait subject providing good detail and serving as a basis for subjective echo ratings.

Two methods of assigning ratings were attempted, both intended to arrive at a number value that could be related to prior work of Mertz, Fowler and Christopher (see references 1 and 2). All comparisons were made by alternately switching the signal through an equal loss path containing a controlled positive echo delayed 3 μs from the main signal. Echo ratings are normally referred to a 10-μs (or greater) delayed ghost,[15] but results obtained using the 3-μs delay should differ by no more than one or two decibels.

Method I

An echo-free picture of the test pattern is viewed while adjusting the television monitor local oscillator frequency. After this adjustment, the vertical line pattern is selected. The test cable and the path with the controlled 3-μs echo are alternately switched while adjusting the controlled echo to the same apparent intensity as the echo caused by the test cable. (Preliminary tests with known signal levels, indicated it possible to alternately view and discern intensity differences of 1 to 3 dB, depending on the gray level.) The delay caused by the test cable is estimated by measurement of the echo displacement as observed on the monitor screen. From the intensity and delay, an equivalent echo rating for long delays is computed by reference to Figure 4 (taken from Mertz's Figure 7, reference 2). It is not likely that the slope of the tolerance-delay curve remains constant at all echo intensities, but data published for 2-μs and 12-μs delays[16] suggest that it may be constant over a range of echo levels judged *just perceptible* to *definitely objectionable* (a range of about 20 dB). For levels around *just perceptible* the bias, if any, should disappear since Figure 4 is based on this tolerance level.

Method II

After completing the set of observations in Method I, a portrait subject is viewed and an echo rating assigned by alternating between the test cable and the path with the 3-μs delayed echo—while adjusting the echo level to produce what is judged an equally objectionable deterioration of quality.

Presumably the results of Method II would be equivalent to those calculated from the data of Method I. However, Method II is wholly subjective and variances with previously established norms may be caused by differences in subject matter as well as viewer opinion.

Method I utilizes features not so subjective, but the geometric forms used may not be translatable to an equivalent level for a viewed picture such as that used in Method II. Mertz[17] showed results obtained by Doba using a picture of small solid rectangles compared with the "normal" tolerance curve. Apparently, the echo threshold level for this type of subject is lower (more sensitive) for long delays. However, at about 3-μs delay, nearly equal sensitivity is indicated. The 3-μs reference delay was selected in an attempt to minimize discrepancies between Methods I and II which might arise because of subject differences.

Qualitative Effects of Cable Periodicities

It was determined that the camera/monitor system could be tuned through a picture carrier frequency range of 80 to 90 MHz with no apparent visual degradation. It seemed reasonable then to consider varying the operating frequency through the band containing the cable discontinuity. This is illustrated in Figure 5 where the cable discontinuity is shown just above the sound carrier. The cable discontinuities studied all exhibited a "bandwidth" of less than a megahertz.

Except for the severity of ghosting, all levels of cable discontinuity exhibited similar characteristics as the frequency was shifted. It is well to note here neither sound carrier nor color subcarrier were present and effects on them would obviously have to be treated as separate cases.

As the discontinuity frequency "moves" into the upper sideband region of the composite television signal, loss of high frequency resolution is evident from inspection of the test pattern. The impairment appears quite evident as the discontinuity center frequency becomes appreciably less than 2 MHz above the picture carrier. As the discontinuity frequency approaches about 1.25 MHz above picture carrier, a distinct ghost appears, delayed by perhaps one picture element (approximately 0.1 μs). As the picture carrier is approached, the echo level and delay both increase, becoming most intense about 0.6 MHz above the carrier. When the discontinuity is at or slightly below the picture carrier, the echo becomes *differentiated*[18] and the degradation of the picture improves noticeably. Below the picture carrier the ghosting becomes negative and the intensity increases until the frequency is about 0.4 MHz below the picture carrier. Negative ghost intensity diminishes rapidly, becoming indiscernible less than 1 MHz below the carrier. With certain qualifications, the peak severity of the negative ghosting is about the same as that observed with peak positive ghosting.

The actual frequency location (with respect to picture carrier) of the various characteristic distortions is undoubtedly related to both television channel bandwidth and the discontinuity bandwidth. Still, it is believed that this same general manifestation would be observed as any discontinuity "drifted" through the television band.

Two additional comments relating to practical aspects of the picture quality: (1) It was observed that positive ghosting (with short-time delay) tended to enhance the contrast of portrait subjects and undoubtedly swayed subjective evaluations of picture quality to the point that trade-offs were made between sharpness of detail and contrast boost. (2) The negative ghosting could be practically eliminated from the monitor by slight downward tuning adjustments without suffering noticeable compromise in picture quality. When observing small discontinuities ($\Delta\alpha$ about 4 dB or less), the negative ghosting was so slight in amplitude and delay, the effect could not be measured as an echo. In fact, compared with a higher loss section or normal cable, the picture could be "matched" to a lower level input signal. Retuning was not permitted for the data reported in the following section.

Discussion of Experimental Results

Three of the six 0.146-inch coaxials with attenuation discontinuities ($\Delta\alpha$) of 2, 7, and 11 dB and one 0.285-inch coaxial with $\Delta\alpha$ of 13.5 dB were selected for detailed study. Equal intensity levels are shown in Figure 6. Delay of the echo at peak intensity levels appeared to be approximately 0.4 μs, ranging from about 0.3 μs for the 2-dB discontinuity to about 0.6 μs for the 11-dB discontinuity. It may also be significant that the peak intensity of the larger $\Delta\alpha$ occurred at a frequency nearest the picture carrier.

Figure 7 shows the intensity levels of Figure 6 redrawn to *equivalent echo rating*, allowing for time delay weighting in accordance with the slope of Figure 4. According to Figure 7 an echo rating of -40 dB would be practically achieved at a 2-dB $\Delta\alpha$ level. The subjective tests (composite results of four observers) of Figure 8 show reasonable agreement near poorest points, particularly for the larger $\Delta\alpha$ levels and indicate better than -40 dB echo rating for $\Delta\alpha$ = 2 dB and $\Delta\alpha$ = 3.5 dB. At small $\Delta\alpha$ levels, positive echoes appeared to enhance picture contrast while negative echoes simply reduced contrast and echoes as such could not be discerned.

A parallel resonant circuit was used to simulate the cable discontinuity. Results were comparable to those obtained with the test cables with regard to general characterization of the discontinuity. By switching in additional circuit loss at each frequency studied, it was varified that the $\Delta\alpha$ effects were independent of total circuit loss. Echo intensity appeared more severe than for a cable with equal $\Delta\alpha$, but this may be a consequence of the greater phase shift exhibited by the network. Phase shift determinations showed approximately 140-degree phase shift over the bandwidth displayed by the network (approximately 2 MHz). The 10-dB $\Delta\alpha$ test cable exhibited no more than 40-degree phase shift over the same frequency range.

These results imply that the effects of the cable periodicities can be interpreted in terms of amplitude and phase deviations and that compensation with lumped (inverse) networks might be feasible as a practical corrective measure in a system. This aspect was not explored in the work presently being reported.

Application to CATV Systems

A published echo rating objective for long-distance television networks is -40 dB.[19] This objective is an all-inclusive performance figure and includes corrections of equalization techniques that might be applied. If the previous results shown for a $\Delta\alpha$ of 2 or 3 dB are accepted as a CATV system objective, the following return loss levels could be permitted (assuming the discontinuity persisted over the *entire* cable system):

Loss at Highest Operating Frequency	*Permissible Return Loss for Periodicities*
100 - 150 *dB*	20 *dB*
200 - 300	23
400 - 600	26
800 - 1200	29

In practical cases, it is not likely that a single discontinuity would persist for cable systems of more than a few hundred decibel loss. The most critical region would be at the highest operating frequency, the total loss diminishing at the lower frequency bands. The return loss required at 50 MHz relative to 200 MHz would be relaxed about 3 dB. Trunk and distribution sections would likely exhibit different sets of characteristic discontinuities since the cable size is usually different, particularly if the trunk is long. Actually, variations between consecutively manufactured lengths of the same type of cable result in noticeable dispersion of discontinuities.

A discontinuity occurring at one frequency (or a limited number of frequencies) known to be peculiar to a specific cable design might be compensated with networks designed for use with that design.

It may be noted that previous studies relating to discrete discontinuities indicate critical return loss requirements (applicable only for that type of discontinuity) occurring at the low-frequency end of the operating range. Return loss requirements are thus limited at high and low frequencies because of different considerations. If results of prior work are accepted without further interpretation, return loss limits because of discrete echoes are generally more critical than limits applicable to periodicities.

It is also interesting to consider that reflections and delay times comparable to those treated in this study can be produced within the confines of a CATV subscriber's tap/drop coax/balun/TV system. Perhaps observations of "soft" pictures observed on CATV systems such as reported by Taylor[20] should be examined with this in mind.

In attempting to assess the total cable contribution to transmission nonuniformity, it would appear that a *trunk* cable with a maximum VSWR of 1.10 relative to 75 ohms (26-dB return loss) over the entire operating frequency range would ensure that the cable did not significantly degrade picture quality in runs extending to losses of 500 (perhaps 1000) dB. *Distribution* cable requirements might be relaxed somewhat since lengthy cascading is not common, but use of the same requirement (26 dB) would be desirable simply to give more margin for the trunk cable tolerance. It may be noted that 20-dB return loss periodicities in drop cable would not produce noticeable effects on picture quality, but discrete discontinuities, both in the cable and terminating devices, could easily negate all of the effort spent in delivering a flawless signal up to that point.

Acknowledgments

The authors wish to acknowledge the contributions of Mr. Paul Wilson in preparing the cable test samples and to Mr. Wilson and Mr. John Micol for their painstaking efforts in compiling test data used for this study.

References

1. P. Mertz, A. D. Fowler and H. N. Christopher, "Quality Rating of Television Images," *Proc. IRE*, vol. 38, pp. 1269-1283, November 1950.

2. P. Mertz, "Influence of Echoes on Television Transmission," *Journal of the SMPTE*, vol. 60, pp. 572-596, May 1953.

3. J. M. Barstow and H. N. Christopher, "The Measurement of Random Monochrome Video Interference," *Trans. AIEE*, pt. 1, vol. 73, pp. 735-741, January 1954.

4. J. M. Barstow and H. N. Christopher, "The Measurement of Random Video Interference to Monochrome and Color Television Pictures," *Trans. AIEE*, pt. 1, vol. 81, pp. 312-320, November 1962.

5. G. Fuchs, "Reflections in a Coaxial Cable Due to Impedance Irregularities," *Proc. IEE*, pt. IV, vol. 99, 1952.

6. W. W. H. Clarke and J. D. S. Hinchliffe, "The Evaluation of Cable Irregularities at Very High Frequencies," *Proc. IEE*, pt. IV, vol. 101, pp. 55-60, 1954.

7. G. Fuchs and Y. Peltier, "The Influence of Small Systematic Impedance Irregularities on VHF Performance of Coaxial Cables," presented at IEE Electronic Division Symposium, London, February 1964.

8. Y. Peltier, "Remarques Concernant la Propagation sur les Lignes, en Presénce D'Irrégularitiés Périodiques D'Impédance," *Cables & Transmission*, vol. 4, pp. 275-302, October 1966.

9. Y. Peltier, "Influence des Irrégularitiés Internes D'Impédance sur L'Affaiblissement des Paires Coaxiales," *Cables & Transmission*, vol. 2, pp. 139-164, April 1969.

10. H. Lubars and J. A. Olszewski, "Analysis of Structural Return Loss in CATV Coaxial Cables," presented at International Wire and Cable Symposium, December 1968.

11. J. Shekel, "CATV Reflections," *Television Horizons*, pp. 15-24, April 1962.

12. H. Ashcroft, W. W. H. Clarke and J. D. S. Hinchliffe, "Some Factors Affecting the Performance of Coaxial Cables for Permanent Television Links," *Proc. IEE*, pt. III, vol. 99, pp. 350-356, 1952.

13. Fuchs and Peltier, *op. cit.*

14. *Loc. cit.*

15. Bell Telephone Laboratories Technical Staff, *Transmission Systems for Communications*. North Carolina: Bell Telephone Laboratories, Inc., 1964, p. 383.

16. Mertz, Fowler and Christopher, *op. cit.*, p. 1275.

17. Mertz, *op. cit.*, p. 580.

18. *Ibid.*, p. 588.

19. Bell Telephone Laboratories Technical Staff, *op. cit.*, p. 384.

20. D. Kaye, "CATV TV: Slumbering Electronic Giant—A Multibillion Industry?," *Electronic Design*, vol. 20, no. 4, pp. 62-72.

21. Fuchs and Peltier, *op. cit.*

22. Peltier, *op. cit.*, pp. 275-302.

APPENDIX

Fuchs and Peltier[21] derived mathematical expressions relating return loss (VSWR) and attenuation to sinusoidal impedance irregularities occurring periodically along the cable length.

At resonance (when $\beta_o h = \pi$), the following relations apply

$$\alpha \doteq \alpha_o \sqrt{1 + u^2}$$

$$\overline{VSWR} \doteq u + \sqrt{1 + u^2}$$

$$\text{where:} \quad u = \frac{\pi}{4\alpha_o h} \quad \frac{\Delta Z}{Z_o}$$

[provided higher order terms of $\frac{\Delta Z}{Z_o}$ *can be neglected]*

α = *attenuation at resonant frequency (Nepers)*

α_o = *normal attenuation constant in the absence of the periodicity*

β_o = *normal phase constant*

h = *period of sinusoidal variation*

Z_o = *normal characteristic impedance*

$\Delta Z = Z_{max} - Z_{min}$ *difference between maximum and minimum local impedance*

For core diameter variations,

$$\Delta D = D_{max} - D_{min}$$

$$Z = \frac{60}{\sqrt{\varepsilon}} \log_e \frac{D}{d}$$

$$\Delta Z = \frac{60}{\sqrt{\varepsilon}} \log_e \frac{D_{max}}{d} - \frac{60}{\sqrt{\varepsilon}} \log_e \frac{D_{min}}{d}$$

$$\Delta Z \doteq \frac{60}{\sqrt{\varepsilon}} \log_e \frac{D_{max}}{D_{min}}$$

$$\Delta Z \doteq \frac{60 \ \Delta D}{\sqrt{\varepsilon} \ D} \qquad \text{when} \quad \Delta D \ll D$$

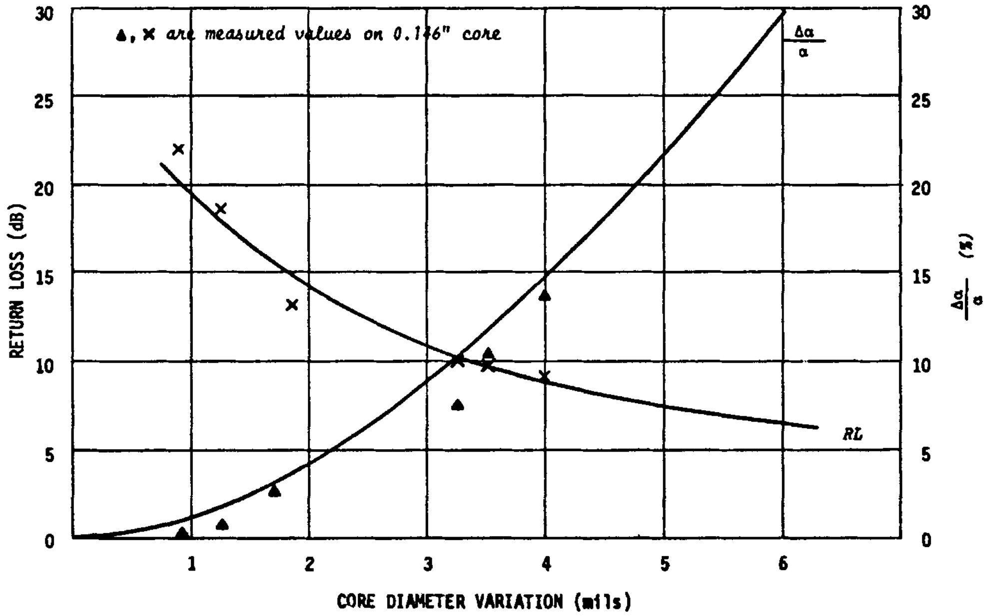

▲, × are measured values on 0.146" core
RETURN LOSS (dB)
Δα/α (%)
RL
CORE DIAMETER VARIATION (mils)

FIG. 1

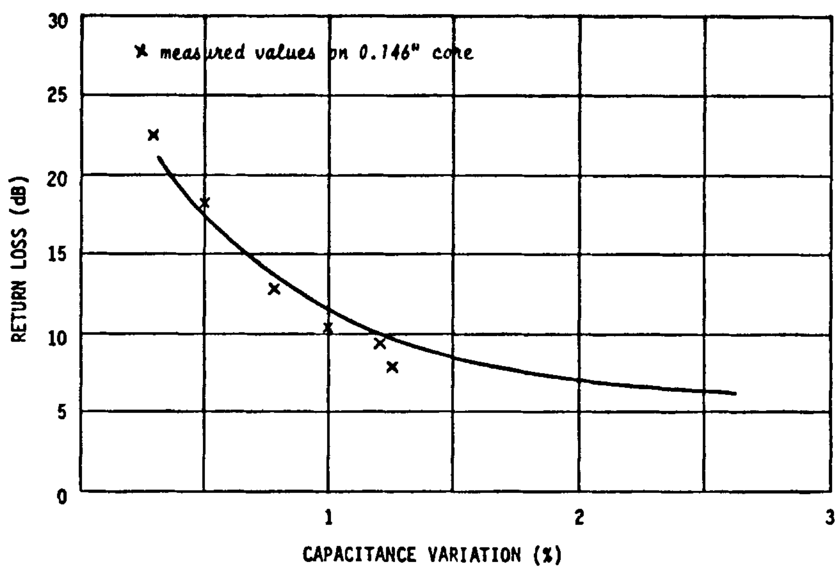

× measured values on 0.146" core
RETURN LOSS (dB)
CAPACITANCE VARIATION (%)

FIG. 2

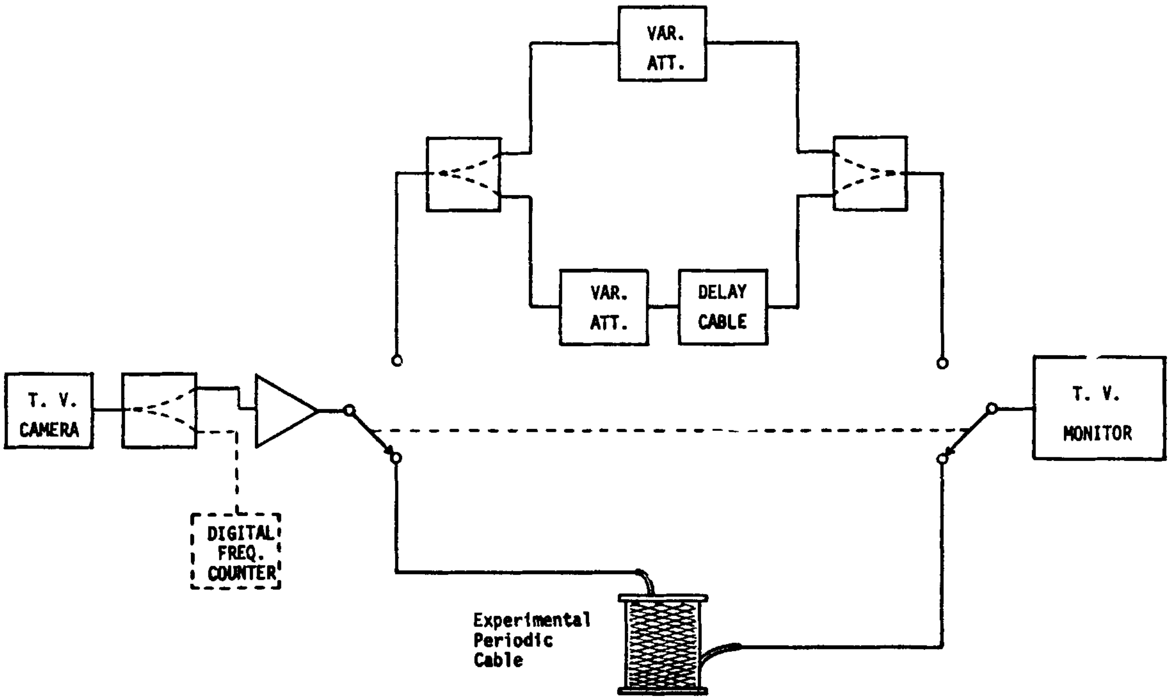

VAR.
ATT.
VAR.
ATT.
DELAY
CABLE
T. V.
CAMERA
DIGITAL
FREQ.
COUNTER
T. V.
MONITOR
Experimental
Periodic
Cable

FIG. 3

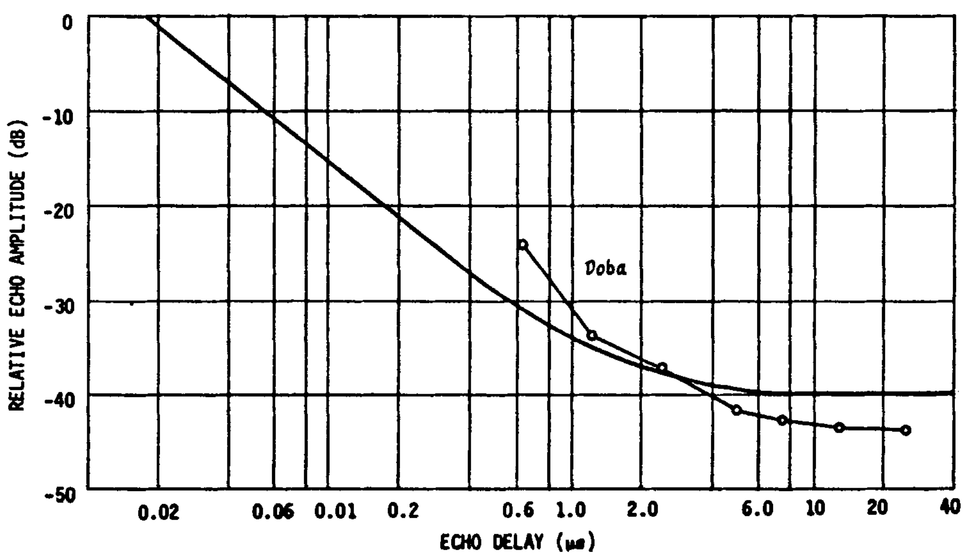

0
-10
-20
-30
-40
-50
RELATIVE ECHO AMPLITUDE (dB)
Doba
0.02
0.06
0.01
0.2
0.6
1.0
2.0
6.0
10
20
40
ECHO DELAY (μs)

FIG. 4

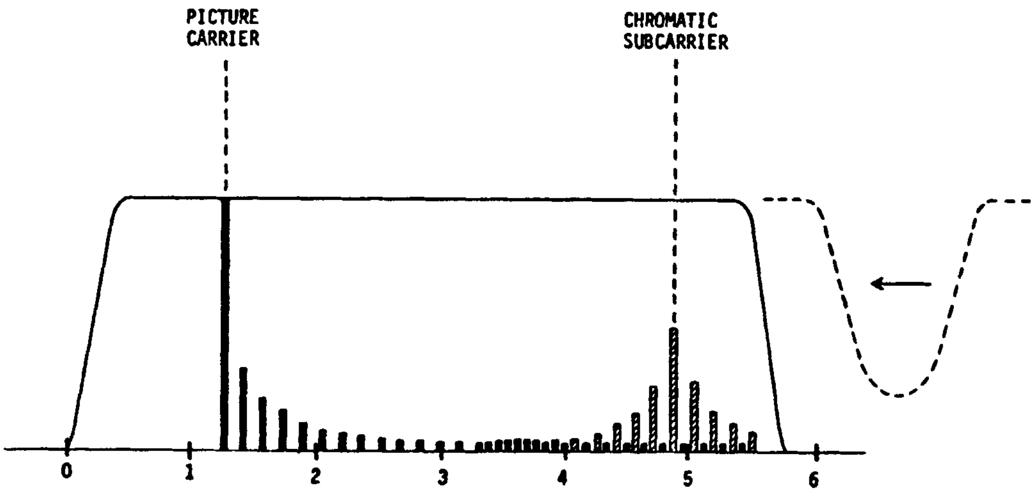

PICTURE
CARRIER
CHROMATIC
SUBCARRIER
0
1
2
3
4
5
6

FIG. 5

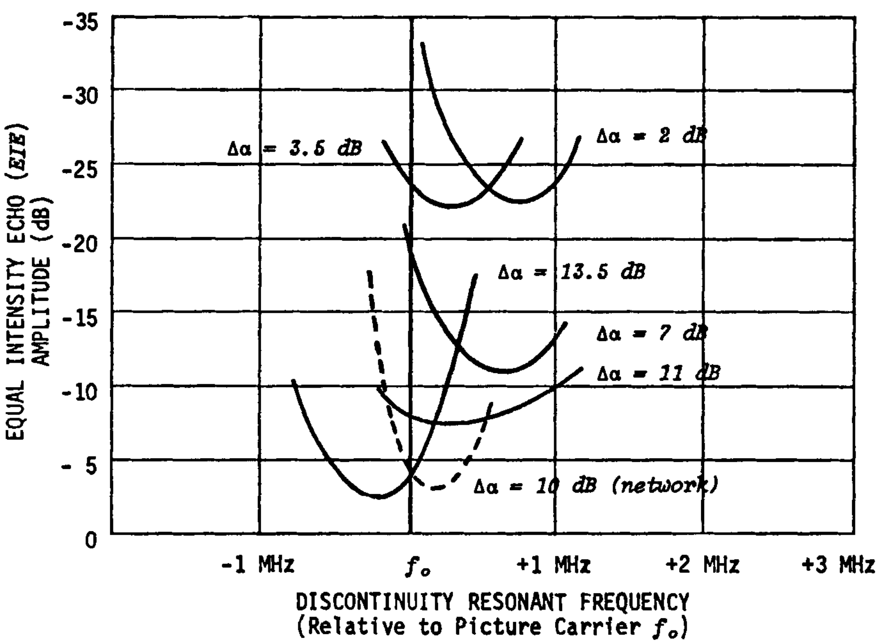
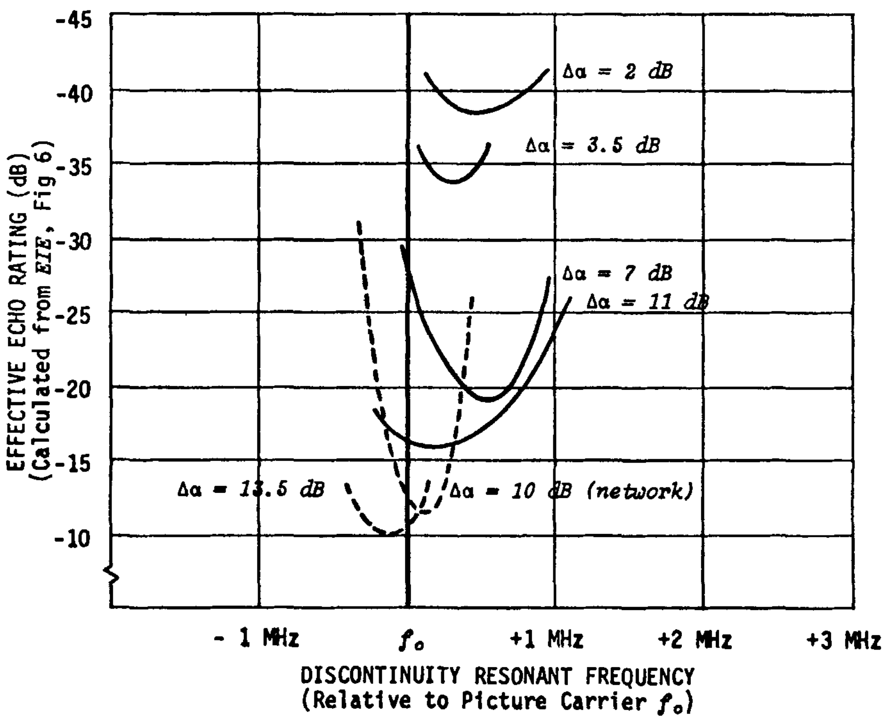
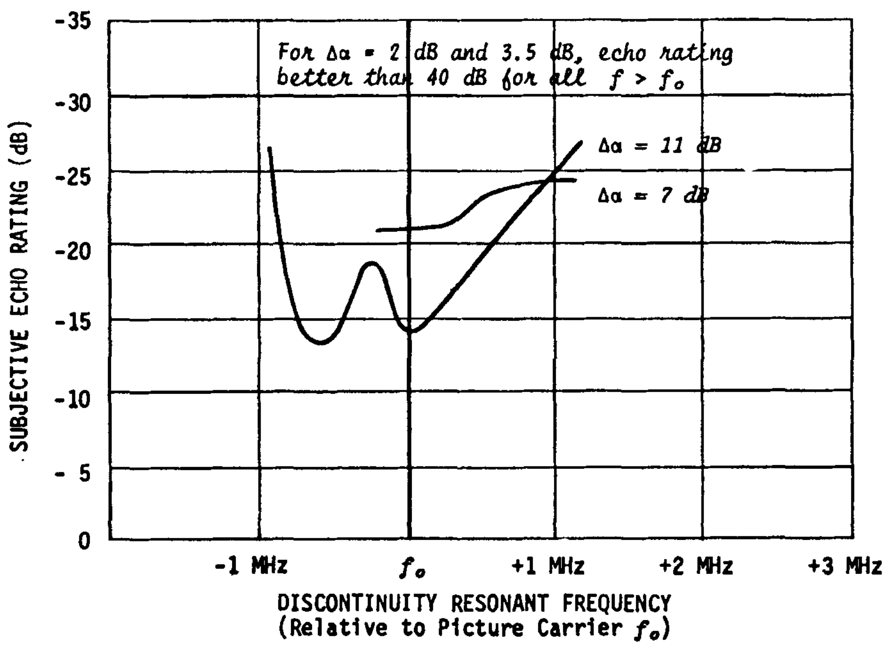

-35
-30
-25
-20
-15
-10
- 5
0
EQUAL INTENSITY ECHO (EIE)
AMPLITUDE (dB)
Δα = 3.5 dB
Δα = 2 dB
Δα = 13.5 dB
Δα = 7 dB
Δα = 11 dB
Δα = 10 dB (network)
-1 MHz
f_o
+1 MHz
+2 MHz
+3 MHz
DISCONTINUITY RESONANT FREQUENCY
(Relative to Picture Carrier f_o)

FIG. 6



-45
-40
-35
-30
-25
-20
-15
-10
EFFECTIVE ECHO RATING (dB)
(Calculated from EIE, Fig 6)
Δα = 2 dB
Δα = 3.5 dB
Δα = 7 dB
Δα = 11 dB
Δα = 13.5 dB
Δα = 10 dB (network)
- 1 MHz
f_o
+1 MHz
+2 MHz
+3 MHz
DISCONTINUITY RESONANT FREQUENCY
(Relative to Picture Carrier f_o)

FIG. 7



-35
-30
-25
-20
-15
-10
- 5
0
SUBJECTIVE ECHO RATING (dB)
For Δα = 2 dB and 3.5 dB, echo rating better than 40 dB for all $f > f_o$
Δα = 11 dB
Δα = 7 dB
-1 MHz
f_o
+1 MHz
+2 MHz
+3 MHz
DISCONTINUITY RESONANT FREQUENCY
(Relative to Picture Carrier f_o)

FIG. 8

DISCUSSION

Mr. Walter Roberts: Any questions--yes?

Question: Not audible

Mr. Roberts: These would not be weighted in any manner. I think the term "weighting", so far as I know, was used to describe at one time the tolerance that you might have as far as the inherent loss of the cable is concerned. If you looked at a particular cable at say 50 megahertz, you'd have, let's say, 1 dB loss per 100--at 200 you might have 2 dB loss and the effects of a ghost trying to weight both the delay of the ghost and the fact that the signal would be attenuated on the way down to where this ghost might occur and on the way back and in the attempt to weight all these things, you can come up with a minimum value, a worse case, and this gets more tolerant as you go up in frequency. And it gets tolerant to the tune of, what is it, 3 dB per octave or something like that--so at one time we spoke of weighting the return loss to recognize the fact that you could be more tolerant at high frequencies----a ghost or a reflection that was the result of this discreet discontinuity, but if it didn't happen to be a result of that, we don't know if the weighting is valid or not. If you look at the effect we studied here now, the return loss that results from a periodicity not a discreet discontinuity, it goes the other way. Since, in a system that's a 1000 dB long in channel 13, for example, is only 500 dB long in channel 2 for the same return loss would not develop to the half the level for the same percent. So now the upper end becomes more critical. So when I start pondering over that, I say, well, let's forget about weighting--maybe we better just have a weight that we hope will be good enough to assure us of a uniform system at any frequency here and for whatever reason that discontinuity occurs. While we may kind of be experts and be able to make certain adjustments with bridges and say---now there's a periodicity or that's not and we can. It's not a practical matter to sit and make judgment on every stinking length of cable you looked at and decide what kind of discontinuity is for us and whether or not it was serious and so on and so forth. So I think if at all possible, it makes sense to have one number if it's practical and realistic and let that number absorb all these variations and if we're talking something in a 26 or so dB level, I think it's reasonable. If we're talking 32 or 34 from the present state of manufacturing art, I think you'd better be real careful about how you specify it and whether you're worried about one type more than the other and so on and so forth. It gets pretty sticky when you really get down to those levels. Any others?

IMPROVING RELIABILITY OF CATV SYSTEMS

by

Raymond F. St. Louis

Manager of Technical Services
Blonder-Tongue Laboratories

IMPROVING RELIABILITY OF CATV SYSTEMS

by

Raymond F. St. Louis
Manager of Technical Services
Blonder-Tongue Laboratories

Much has been done by the CATV industry to assure that CATV subscribers receive high-quality TV signals and that they receive more channels than there is worth-while programming to fill. However, little has been done to assure the subscriber that he will not be entirely without TV signal due to equipment failure.

CATV equipment seemingly prefers to fail during the hours of 5 P.M. and 1 A.M. This is annoying to the subscriber and the system maintenance personnel, alike. While the subscriber waits patiently -- or impatiently -- for the programming to be restored, the system maintenance personnel must respond to these emergencies at a time that can be most undesirable.

These situations could be avoided if the CATV system was self-sustaining through the use of automatic backup equipment, enabling the system to continue operation without interruption, even though an equipment failure has occurred.

The basic design parameters of such a system are:

1. The ability to detect an equipment failure.

2. A method of automatically transferring to backup equipment upon failure of the primary equipment.

3. The ability of notifying maintenance personnel that a failure of the primary equipment has occured.

4. The ability to locate the piece of equipment which has failed.

Let us look at how these functions can be accomplished at the head-end site.

The headend equipment should be designed to prevent failures, as much as is possible with the present state-of-the-art.

Such a system is comprised of partial or full dual equipment for each channel.

This includes dual preamps, preamp P.S., dual bandpass filters, and dual-channel amplifiers capable of automatic change-over to backup equipment, and an alarm system capable of notifying maintenance personnel.

PREAMPLIFIERS

Preamplifiers should be of the single channel type with the greatest skirt selectivity available. The greater the skirt selectivity, the less electrical surge energy reaches the preamplifier's electronics from nearby lightning strikes. The preamplifier must also have maximum protection within its electronics against these surges.

Even with maximum protection against failure due to surges and solid state reliability protecting against failures due to aging, preamplifiers still can fail.

Providing a fail safe system will solve this problem.

Figure #1 shows this setup.

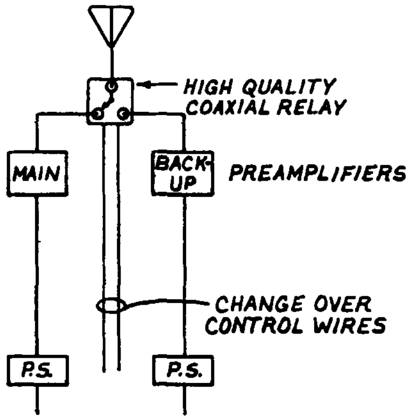


Figure #1

Two preamplifiers are fed from a common antenna. This is usually the most practical method when complex and expensive antenna arrays are used. However, when smaller antenna arrays are used for the stronger signals, a separate antenna for each preamp may be practical and greater reliability against failure can be achieved.

The antenna signal is fed to each preamplifier through a coaxial relay rather than a hybrid splitter to prevent a 3 db decrease in the signal-to-noise ratio, and to provide additional surge protection to the backup preamp in the disconnected leg of the relay.

The control voltage for this relay is supplied from the signal sensing equipment within the headend building.

Two coaxial down leads are used to provide additional backup protection in case of coax failure.

The preamplifiers each use identical broadband amplifier boards preceded by a very selective Helical resonator single-channel filter.

Stocking one spare amplifier board allows easy replacement without removing the preamp from the antenna tower.

Each preamp will be connected to its own preamp power supply. These power supplies must have excellent built-in surge protection, especially if they are of the more complex regulated type.

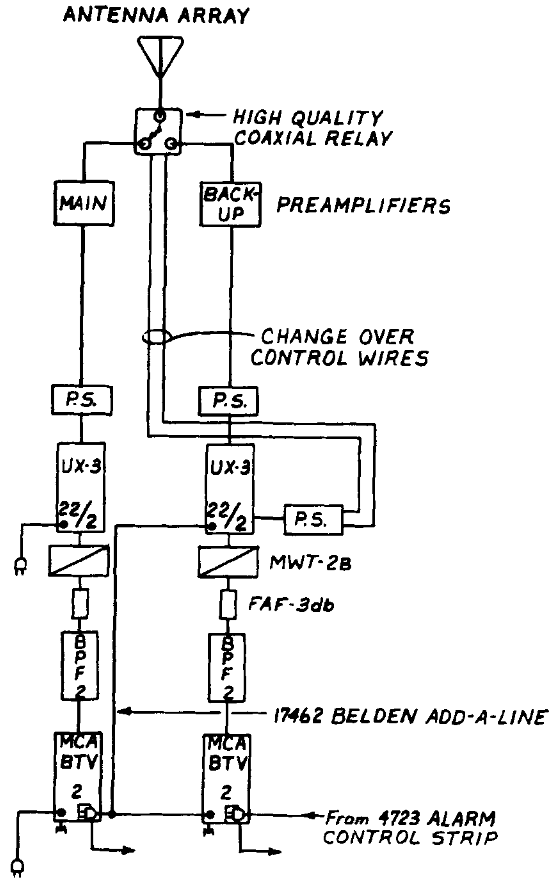


Figure #2

The preamplifiers feed either UHF converters, followed by bandpass filters, or in the case of VHF channels, VHF bandpass filters. As shown in figure #2. Both devices protect against incoming high voltage damage to the headend. The UHF converter - because very little lightning energy is concentrated at UHF frequencies, and the VHF bandpass filter - because little energy is concentrated within its narrow bandpass.

If single-channel sound traps are used to attenuate the sound carriers to the desired level, these traps must be extremely stable to prevent interference with the color sub-carrier. The ideal trap for this use is the adjustable notch depth type. The adjustable notch type differs from the standard types in that the sound carrier level is controlled not by varying the frequency of the trap, but by adjusting the coupling of the trap sections, which in turn increases or decreases the trap depth.

The heart of this improved reliability headend is the MCA-BTV signal processors. This device filters, amplifies, and provides the high AGC range for great variations in input signal level. The high gain AGC DC amplifier makes possible the method of automatically switching to backup equipment if the primary equipment should fail.

The MCA-BTV senses its own output signal and performs two functions if it, its preamplifier, or converter fails.

1. Switches AC power to its AC convenience outlet, which has been wired for "signal on - AC power off," to which is connected the backup MCA-BTV and a small DC power supply which powers the antenna preamp coaxial relay which we discussed previously.

2. Cuts off its R.F. output to prevent noise from being fed into the system by the primary MCA-BTV which is now running at full gain because it has no output signal to provide AGC voltage.

The backup MCA-BTV being turned on upon failure of the main MCA-BTV, instantly replaces the main channel, providing the subscriber with uninterrupted programming. When the backup MCA-BTV "sees" a signal at its output, it knows that the channel has not gone off-the-air, but that the main amplifier, converter, or preamplifier has failed, and switches AC power to its AC receptacle which has been wired for "signal on - AC on" (opposite from that of the primary MCA-BTV.) The Model 4723 Alarm Control Strip, which is plugged into the backup MCA-BTV convenience outlet, uses this voltage to energize a 4 form C relay within the Alarm Control Strip. This relay closure provides a 15 VAC output to a local alarm light and removes AC power from an outlet to which a Model 4722 Alarm Carrier Generator is connected. The Alarm Carrier Generator normally transmitts a 108.625 MHz carrier through the CATV distribution system, except during the time of equipment failure, at which time its AC power is disconnected. In addition, the Alarm Control Strip provides a 1 form C contact closure which can be used to activate a telephone alarm dialer or auxiliary equipment.

The 108.625 MHz Alarm Carrier Generator's signal can be received at any point in the CATV distribution system at which a Model 4624 Alarm Carrier Receiver is installed. Typical locations for these receivers are in the CATV office or in the homes of maintenance personnel.

When the Alarm Carrier Generator is "on," the Alarm Carrier Receiver receives this signal and is silenced.

If headend equipment should fail and revert to backup equipment, or if an amplifier in the distribution between the Alarm Generator and the Alarm Receiver should fail, the Alarm Receiver will generate an audible and visual alarm to notify maintenance personnel immediately that a problem has occurred.

Viewing the TV signals at the Alarm Receiver location will quickly determine whether a distribution amplifier has failed, or the problem lies in the headend.

Aside from the automatic backup and alarm function, this improved reliability headend is designed to cut maintenance to a bare minimum, by using a common, easily replaced P.C. board for all headend amplifiers, the alarm carrier generator, alarm receivers, the pilot carrier, and the carrier substitution generators.

Figure #3 shows the same system using separate aural and visual MCA-BTV's and an Aural/Visual Carrier Separator.

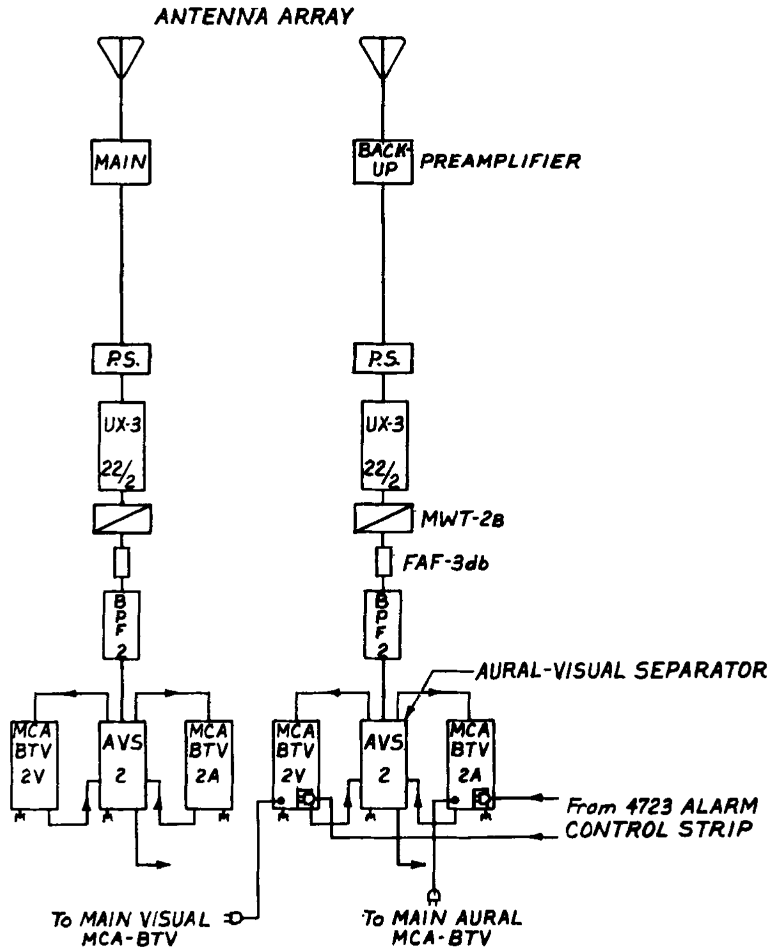


Figure #3

Other Alarm Capabilities

The features of the improved reliability headend also lend themselves to other types of alarms, such as temperature alarm, intrusion alarm, fire alarm, low AC line voltage alarm, etc., simply by plugging a low voltage controlled, 117 VAC relay (ALCO FRE-103) into the Alarm Control Strip.

Each relay takes up the place of one set of MCA-BTV amplifiers (Main and Backup.) The relay is plugged directly into a convenience outlet on the strip and the Alarm Control Strip sensing cable plugs directly into the relay.

As this relay can be connected for normally "on," or normally "off," it can be controlled by shorting or opening its low voltage output with thermostats, door switches, photoelectric cells, or AC line voltage meter relays.

The entire headend is further protected by the addition of a Model 4694 Overvoltage Protector (O.V.P.) in the AC line.

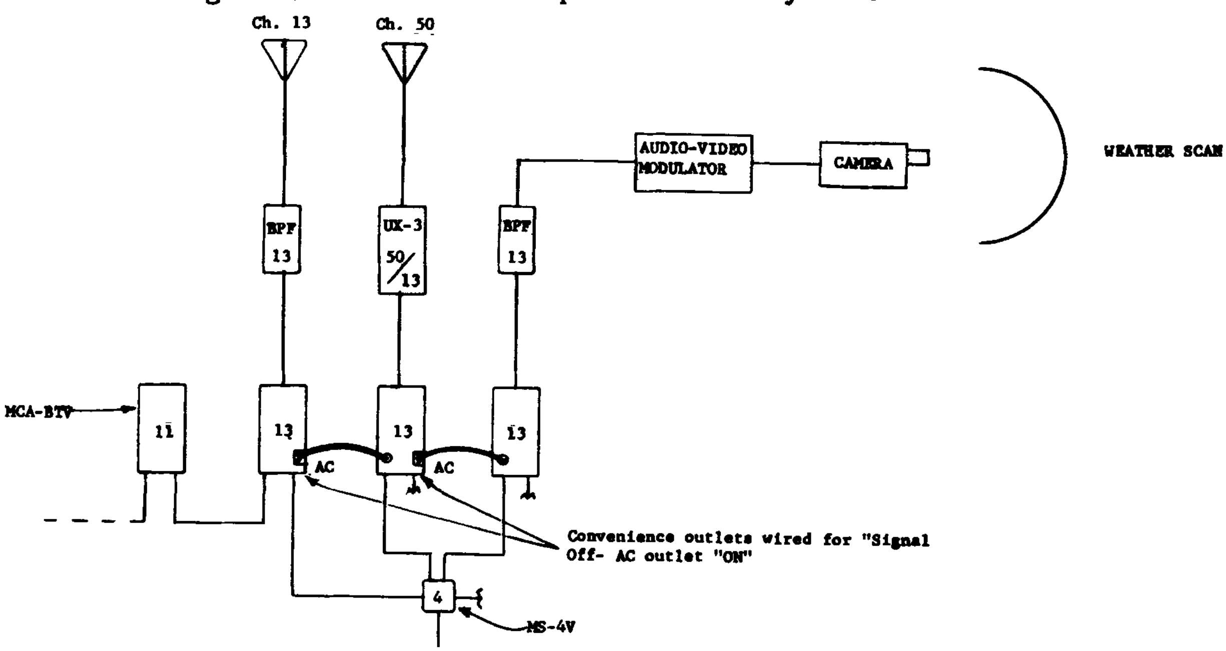
The O.V.P. protects the entire headend high voltage spikes, and more importantly, against AC line overvoltages, which can be deadly to solid-state equipment, even though the equipment is internally protected against short duration high voltage spikes.

The O.V.P. protects against the AC line voltage excesses which often accompany restoration of power following a power failure. The O.V.P. shuts off all AC power to the headend when its input voltage exceeds 135 VAC.

Having a channel amplifier which acts as an R.F. controlled relay, lends itself to many other uses, such as:

1. Channel substitution upon loss of the R.F. signal. This system can be applied to TV channels with erratic air times.

   Figure #4 shows the setup for such a system.



When channel 13 leaves the air, ch. 50 will replace it. If ch. 50 is not on the air, ch. 13 program will be provided by the weather scan.

The primary channel normally provides signal to the CATV system. If this primary MCA-BTV looses signal, it cuts off its R.F. output and switches AC power to its AC convenience outlet. The secondary MCA-BTV is then turned on and its programming replaces that of the primary MCA-BTV. If the secondary MCA-BTV "sees" no signal at its output, it will turn its output off and provide power to a third MCA-BTV which will provide a second alternate program.

Programming will again revert to the primary amplifier when that amplifier again receives signal.

2. Figure #5 shows an automatic backup of an existing headend. One MCA-BTV is used as a signal sensing relay and a second MCA-BTV is used as the backup channel processor.

3. Figure #6 shows a typical diversity reception using MCA-BTV's.

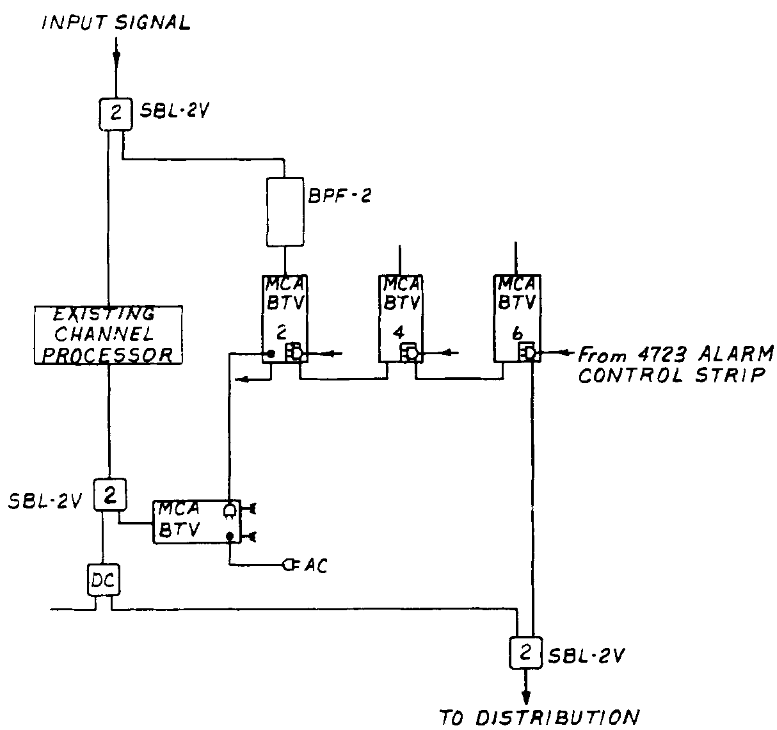


Figure #5

Automatic Backup Of An Existing Headend

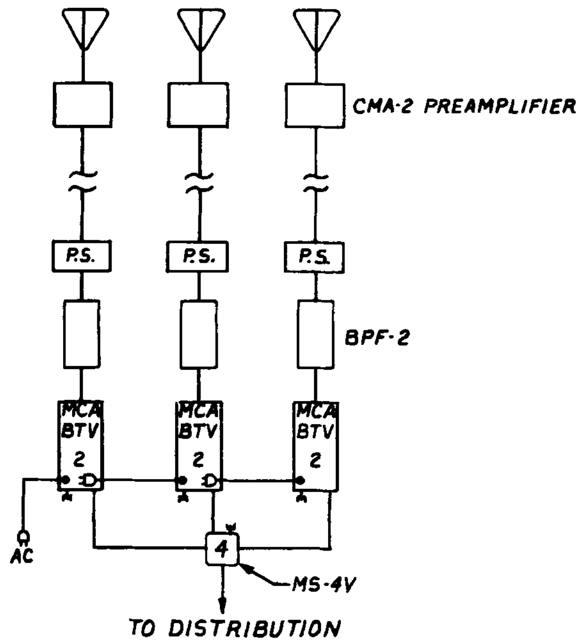


Figure #6

Typical Diversity Reception

NATIONAL CABLE TELEVISION ASSOCIATION

ANNUAL CONVENTION

CHICAGO, ILLINOIS

JUNE, 1970

PHASE LOCK APPLICATIONS IN CATV SYSTEMS

I. Switzer, P. Eng.,
Chief Engineer,
Maclean-Hunter Cable TV Limited,
27 Fasken Drive,
Rexdale, Ontario.

PHASE LOCK APPLICATIONS IN CATV SYSTEMS

Phase locking techniques have been used for some time in a variety of communications applications. The term may be generally applied to locking the phase of one signal source to another. Such sources will then operate at the same frequency and their relative phase will remain constant or nearly so. The principle application in communications has been suppressed carrier communications systems, both wire and radio, and in the transmission of colour television. Colour television subcarriers are operated in a suppressed carrier mode. Each individual colour receiver has a local subcarrier oscillator which must be phase locked to the original colour subcarrier in order to permit the colour information to be demodulated. Similarly, other suppressed carrier communications systems require a locally generated carrier to permit the transmitted information to be demodulated. The local carrier is often generated by phase locking to a pilot signal of some kind.

Figure 1 shows a basic phase lock loop. A phase comparator compares the reference signal and the local carrier oscillator. The local carrier oscillator is a voltage controlled type whose output frequency can be controlled by a DC control voltage. The comparator produces an error signal which is used to control the voltage controlled oscillator (VCO). With a suitable error signal amplifier and VCO characteristics the feedback loop will keep the local carrier oscillator tightly locked in phase with the incoming reference signal. Phase lock loops of this kind have been studied in great detail and complete analyses of a number of forms of this basic phase lock loop are to be found in the electronic engineering literature.

CATV APPLICATIONS

One of the earliest phase locking applications in CATV is the Model 220 Phase Lock generator produced by Phasecom Engineering, of Anaheim, California. This phase lock generator is employed to produce an unmodulated carrier which is phase locked to a broadcast television channel. In areas of moderate to strong local signal, it is common practice to convert the local channel to a different channel for distribution. The local channel is often left vacant, but many CATV systems use this channel within the cable system for distribution of news or weather services or other closed circuit originations. Unless the modulator used for this origination is phase locked to the air signal there will usually be annoying "co-channel" beats in subscribers' receivers caused by the beat between the direct pick-up of the air channel and the cable channel. These two channels will generally be slightly different in frequency and will cause an annoying beat at interfering signal levels which are not strong enough to cause annoying or even visible "ghost" effects. Generally speaking, the "co-channel beat" is visible before the ghost. If the CCTV modulator is phase locked to the air signal there will be "co-channel beat". If the interference is strong enough there may still be a ghost visible on some subscribers' receivers. In exceptional cases there may be no co-channel beat but there may be interference to the sound carrier. It is not feasible to lock FM sound carriers.

Figure 2 shows the basic block diagram of the Phasecom Engineering Model 220 Phase Lock Generator. A tuned RF amplifier selects the desired off-air channel. The associated sound carrier and any adjacent carriers are rejected at this point. A phase comparator (balanced diode bridge) compares the incoming air signal and the signal from the internal VCXO (voltage controlled crystal oscillator) and produces an error signal which is amplified and then used to control the VCXO.

The purpose of the phase lock generator is to provide a local carrier which is phase locked to the air signal. This carrier then becomes the carrier for the CCTV modulator, replacing the modulator's internal crystal controlled carrier. The VCXO used is a precision crystal oscillator in a temperature controlled oven. It is controlled over a very narrow range--about plus or minus 2 KHz, slightly more than the frequency drift that TV stations are permitted by the FCC. Television broadcast stations must operate on assigned frequency with only 1 KHz ($\pm$) permitted. Restricting the "pull-in range" of the VCXO simplified its design and also prevents it from locking onto unwanted modulation sidebands. When no air signal is present to provide a reference lock, the VCXO will still be quite close to the desired output frequency.

Other techniques might be used to produce a local carrier which is an unmodulated form of the air carrier. Hard limiting might have been used to "strip" the modulation from the air carrier. Most limiters tend to convert the AM modulation to a small amount of phase modulation. This would be undesirable. A limiting technique would also have the disadvantage of not providing a carrier when the air channel was not broadcasting, or if the air channel was overmodulated.

Carriers could also be effectively phase locked by the use of very precise oscillators for both the air transmitter and the CCTV modulator. The use of high precision crystal oscillators or rubidium or caesium beam oscillators is becoming more common in broadcast stations but is far from universal. A similar high precision oscillator in the CCTV modulator would keep the locally generated carrier substantially in phase with the air signal for long periods of time. Such oscillators are extremely expensive ranging from $5,000 to $15,000, depending on the degree of precision required.

Phase lock generators, which Phasecom Engineering has produced so far have been for VHF use. Because the frequency lock-in range is quite small, they must be ordered with broadcast station offset in mind. One recently installed by our Company at North Bay, Ontario operates on channel 10- i.e. 193.240 MHz instead of the nominal channel 10 visual frequency of 193.250 MHz. There has been some consideration of using the phase lock generator at IF frequency (45.75 MHz). This would permit its use in IF type modulators and would also permit the construction of an IF phase locking generator that could then be applied to a number of different situations. Phase locking at IF would require precision down conversion since the "pull-in range" of the VCXO is so narrow. A down converter from channel 13 to IF normally operates with an L.O. at 257 MHz. A stability and precision of better than 1 KHz at this frequency requires an L.O. with about .0005% accuracy.

This would be an expensive L.O.--possible, but a bit expensive. Such a phase lock system would operate an IF modulator as in figure 3. In cases where IF type modulators are used, the scheme in figure 4 is recommended. In this case, the up-converter L.O. is not as critical, and the phase lock generator operates at the same frequency as the air channel visual transmitter.

Phase lock techniques may also be used to reduce the risk of interference to direct local reception caused by signal leakage from the cable system. In the North Bay installation previously mentioned, the local air signal is not strong enough to cause serious beat and ghost problems and the system distributes sound programms only (no visual modulation) on its locally generator channel 10. The reason for the phase lock installation was reduction of the risk of c-channel interference, arising from system radiation interfering with direct local reception by non-subscribers. The system experienced a number of cases in which a subscriber and non-subscriber would have receivers back-to-back, separated only by a thin partition. A small amount of radiation from the subscriber's service drop and matching transformer leads would cause interference to the non-subscriber receiving the local station on rabbit ears on the other side of the partition. It was a case of marginal radiation causing interference to marginal direct reception. The phase lock system has completely cured this problem.

Dual cable systems would seem to be natural applications for phase locking systems. The "co-channel beat" due to excessive cross-talk between cables is visible and annoying before the "ghost" becomes visible. Phase locking the channels in adjacent cables will eliminate the "co-channel beat" type of interference. Channels of the same "number" in each cable should be phase locked to each other. One channel would be designated as the reference channel and the same channel number in the other cable phase locked to it. For example channels in the "A" cable would be designated as the reference channels and the "B" cable channels phase locked to the "A" channels. Channel 2B would be locked to 2A, 3B to 3A, etc. In some instances both the A and B channels might be locked to an air channel.

Figure 5 shows one of several schemes that permit locking channels in adjacent cables of a dual cable system. In the example shown, 9A has been derived from a channel 9 transmitter at a moderate distance. Channel 9B is derived from a UHF station, channel 50. Heterodyne processors are shown in both channels for signal processing. The phase detector compares 9B to 9A and controls a VCXO which is incorporated into the channel 50 - 9 UHF converter. This VCXO varies the phase of the 9B signal to keep it locked to 9A. The phase lock loop can be expanded to include a number of different signal processing elements.

— — — — —

Phase lock techniques can be used to eliminate third order intermodulation (triple beat) problems in CATV systems. Visual carriers in CATV systems are only nominally on specified channels. The actual visual carrier frequencies can be significantly different from the "true" visual carrier frequencies. Differences arise from the + and - 10 KHz offsets assigned to television broadcast channels, the + and - 1 KHz frequency tolerance allowed television broadcast channels, and the frequency errors arising from low tolerance oscillators in CATV channel conversion and channel originating equipment. The "triple beats" which arise as distortion products in CATV have frequencies which are the sums and differences of either three different visual carriers (aural carriers are neglected in this discussion because of their relatively low level) or the second harmonic of one carrier and the fundamental of a second carrier. If all the visual carriers in a contiguous set of carriers were separated by exactly the same frequency, the "triple beat" products would all be "zero" and would cause only very minor changes in level rather than visible interference beats. This technique is applicable to contiguous groups of channels, e.g. 2,3,4, or 7,8,9,10,11,12,13. It would also be applicable in cases where additional mid-band or superband channels are added to the high band group. These channel visual carrier frequencies could all be locked to a 6 MHz master oscillator which would set the frequency spacing between them. It is not important that the spacing be exactly 6 MHz, only that the spacing be exactly the same for all the channels.

A number of techniques could be used to derive such a locked set of channels. One scheme is illustrated in figure 6. A master 6 MHz oscillator sets the spacing between channels. A comb generator generates multiples of the master 6 MHz oscillator. These are separated and amplified and mixed with the channel 7 visual oscillator (175.25 MHz). The resulting sum frequencies are filtered and amplified to become a master set of reference carriers which are all separated by the 6 MHz master oscillator frequency. This set of reference carriers can be used as the carrier oscillators for originating modulators or they can be used as references to which specially modified heterodyne processors could be locked. The modified heterodyne processors would use a VCXO as one of the local oscillators permitting the processor output frequency to be locked to the appropriate master reference carrier.

The value of such a system is not yet known, although its cost would be substantial (but probably well within the means of larger CATV systems). Triple beats are usually a lesser problem than cross modulation interference. It is probable that they accumulate in a system at a slower rate than cross modulation products. It is likely that a system that has cross modulation products held to a tolerable level will also have triple beats under control, but there may be cases in which triple beat is still a serious problem. Still, it would be nice to eliminate the triple beats completely.

\- - - - - - - - - -

Phase lock techniques can be used to generate the local carrier required in synchronous detector systems. A television demodulator marketed by Scientific Atlanta uses phase lock to develop a local carrier for synchronous detection.

The S-A demodulator uses an elaborate system to sample the television carrier during sync pulses when the carrier is at full amplitude, in a manner analogous to the phase lock system in colour television receivers which phase locks the locally generated colour subcarrier to the colour burst transmitted on each synchronizing pulse.

- - - - - - - - - - -

Suppressed carrier techniques would considerably reduce the loading-induced distortion products in CATV systems. Suppressed carrier might be used for transportation trunks with carrier reinsertion before distribution. Alternately, suppressed carrier might be used throughout a system trunk system with carrier reinserted at bridging locations feeding the less critical parts of the distribution system. It might be possible to include carrier reinsertion in back-of-the-set or set-top multi-channel converters. Phase lock techniques locking to a carrier "burst" might be used or a pilot tone system might be considered.

- - - - - - - - - - -

The author wishes to acknowledge the assistance of Mr. Arie Zimmerman, of Phasecom Engineering, for the information on the Phasecom Model 220 Phase Lock Generator and for the provision of practical phase lock hardware to test some of the applications which have been discussed.

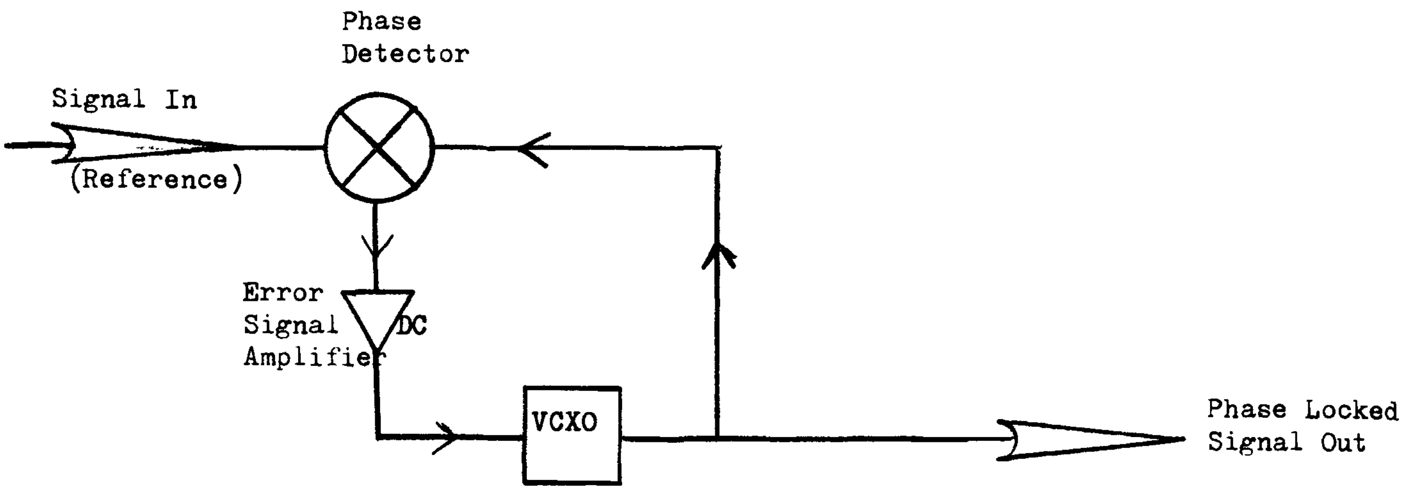


Figure 1

Basic Phase Lock Loop

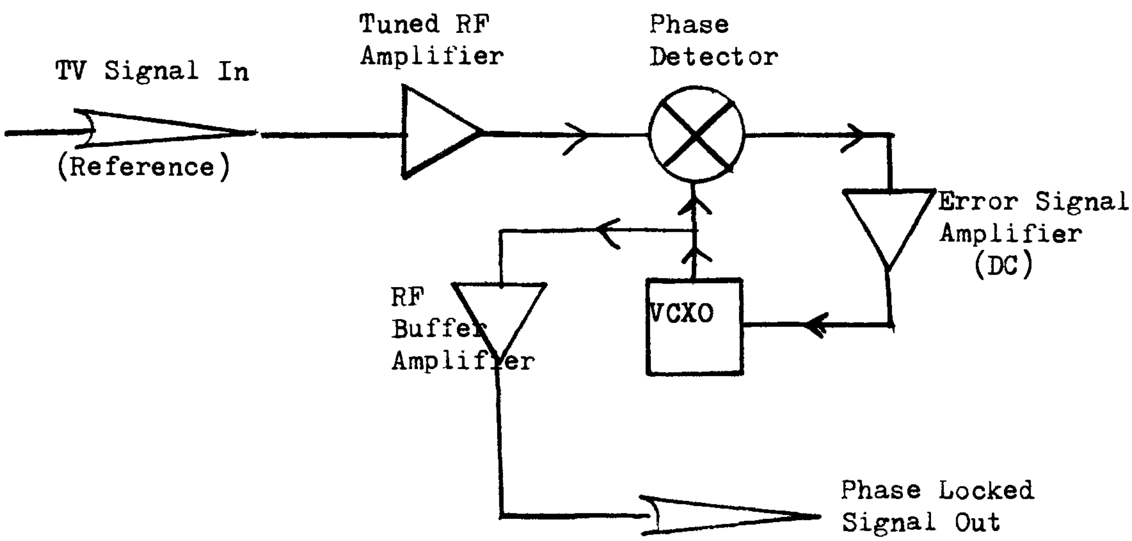


Figure 2

Basic Block Diagram

Phasecom Model 220 Phase Lock Generator

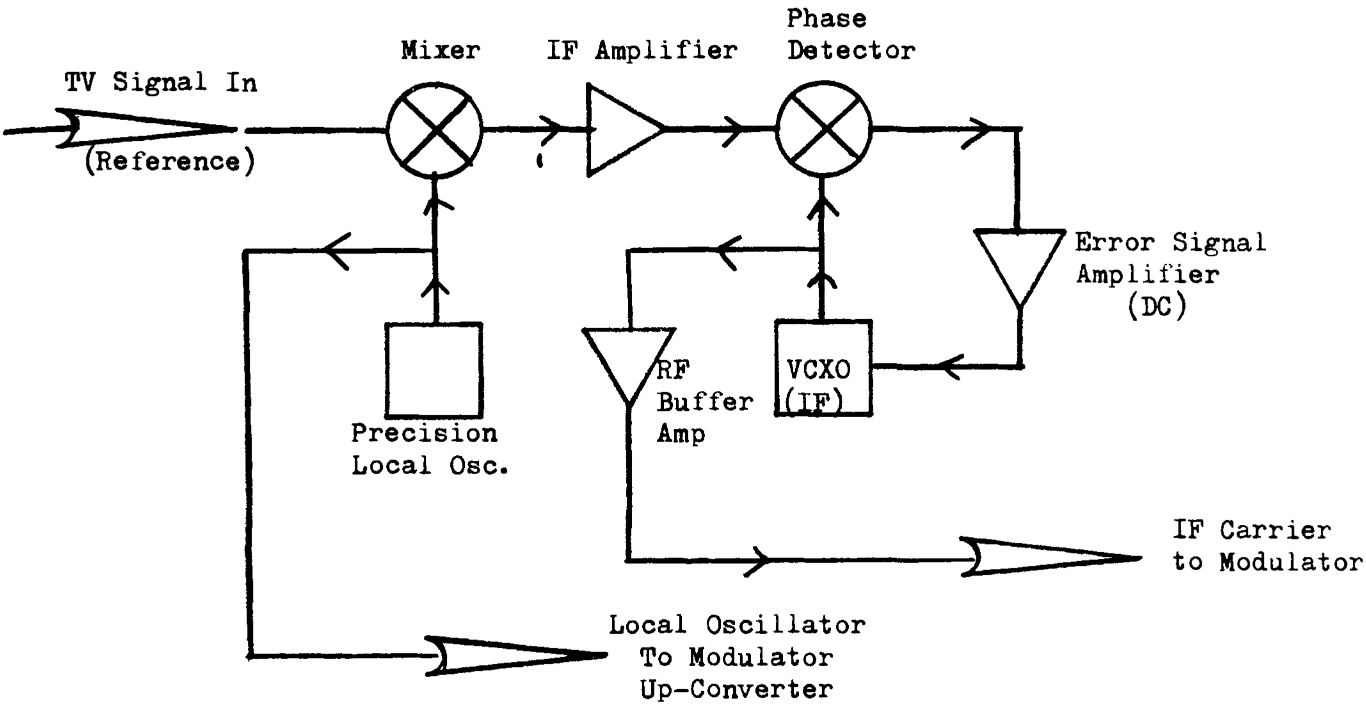


Figure 3

Phase Lock Operating at IF Frequency

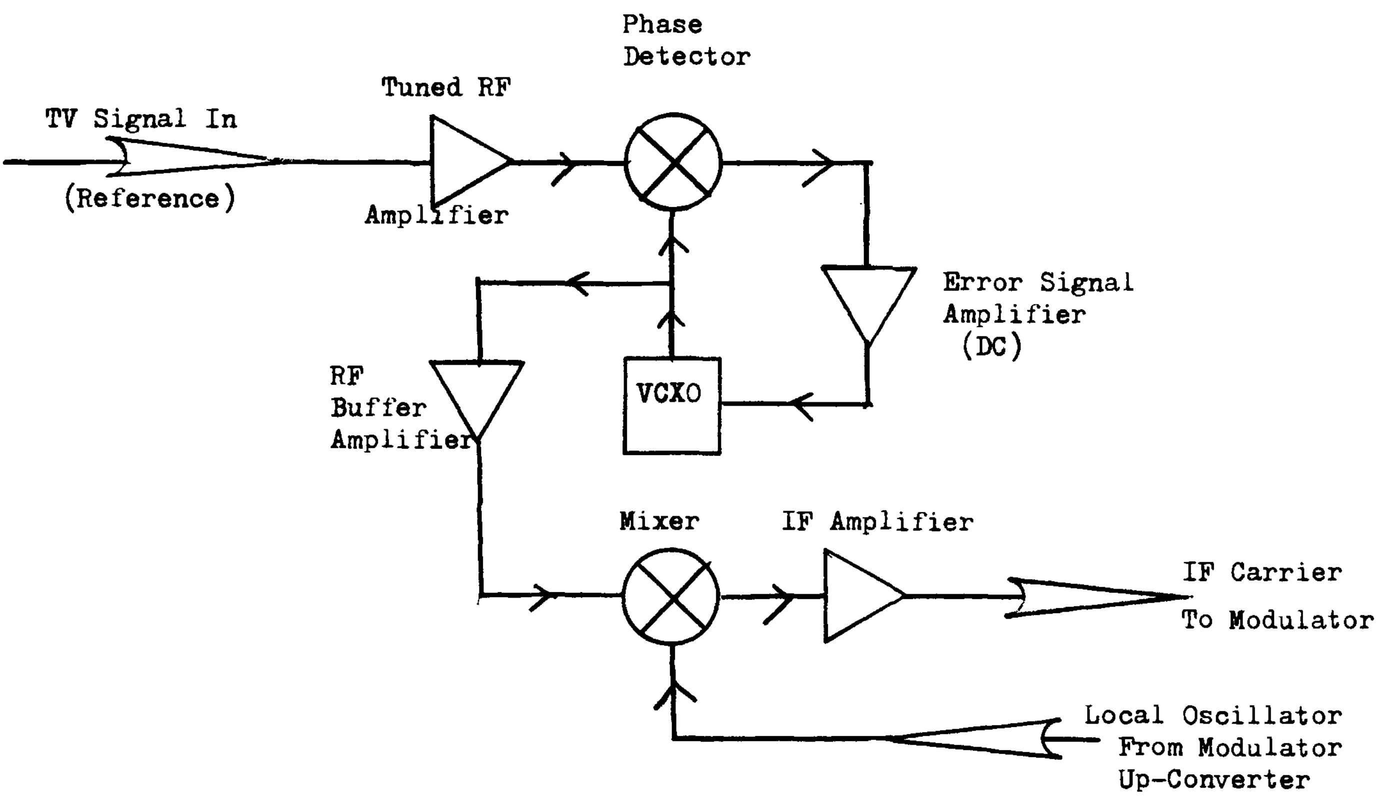


Figure 4

Phase Lock System for IF Modulator

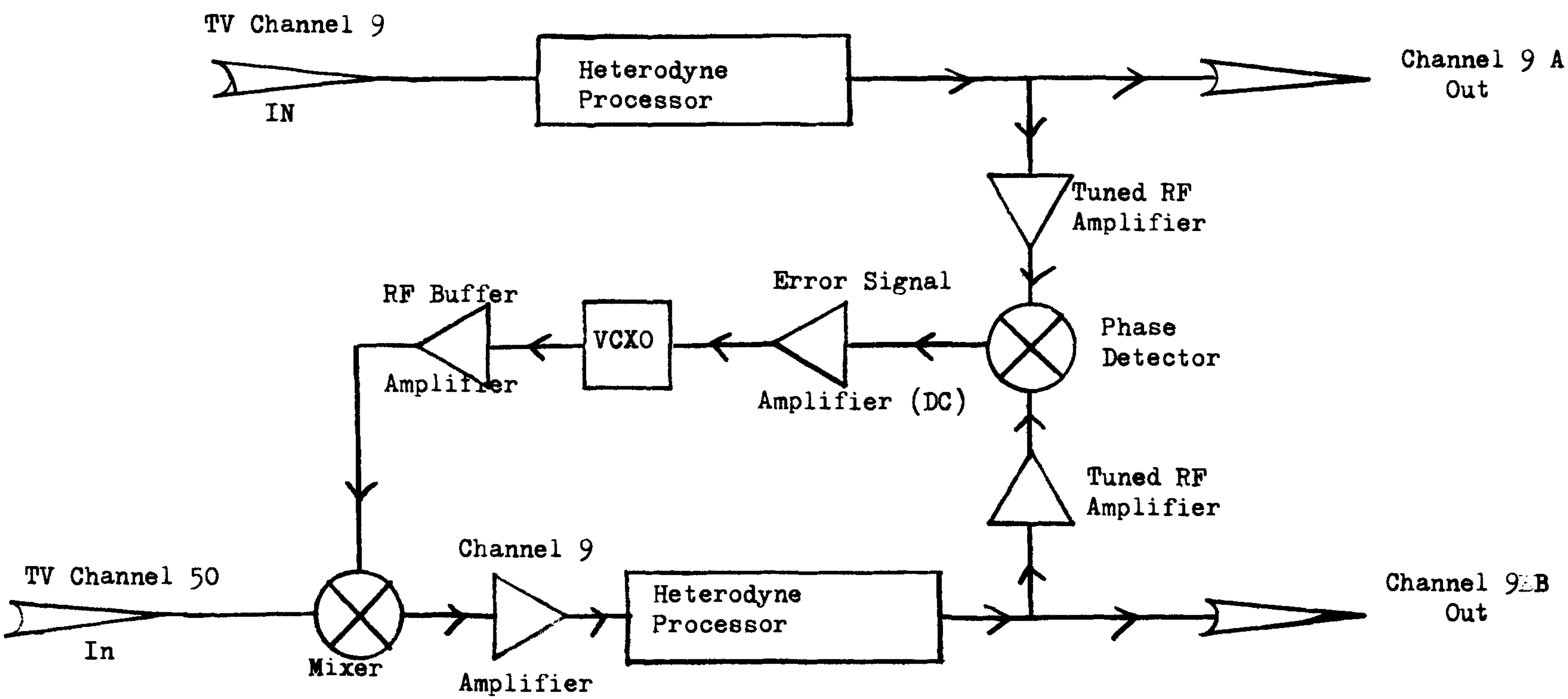


Figure 5
Dual Cable Phase Lock Application

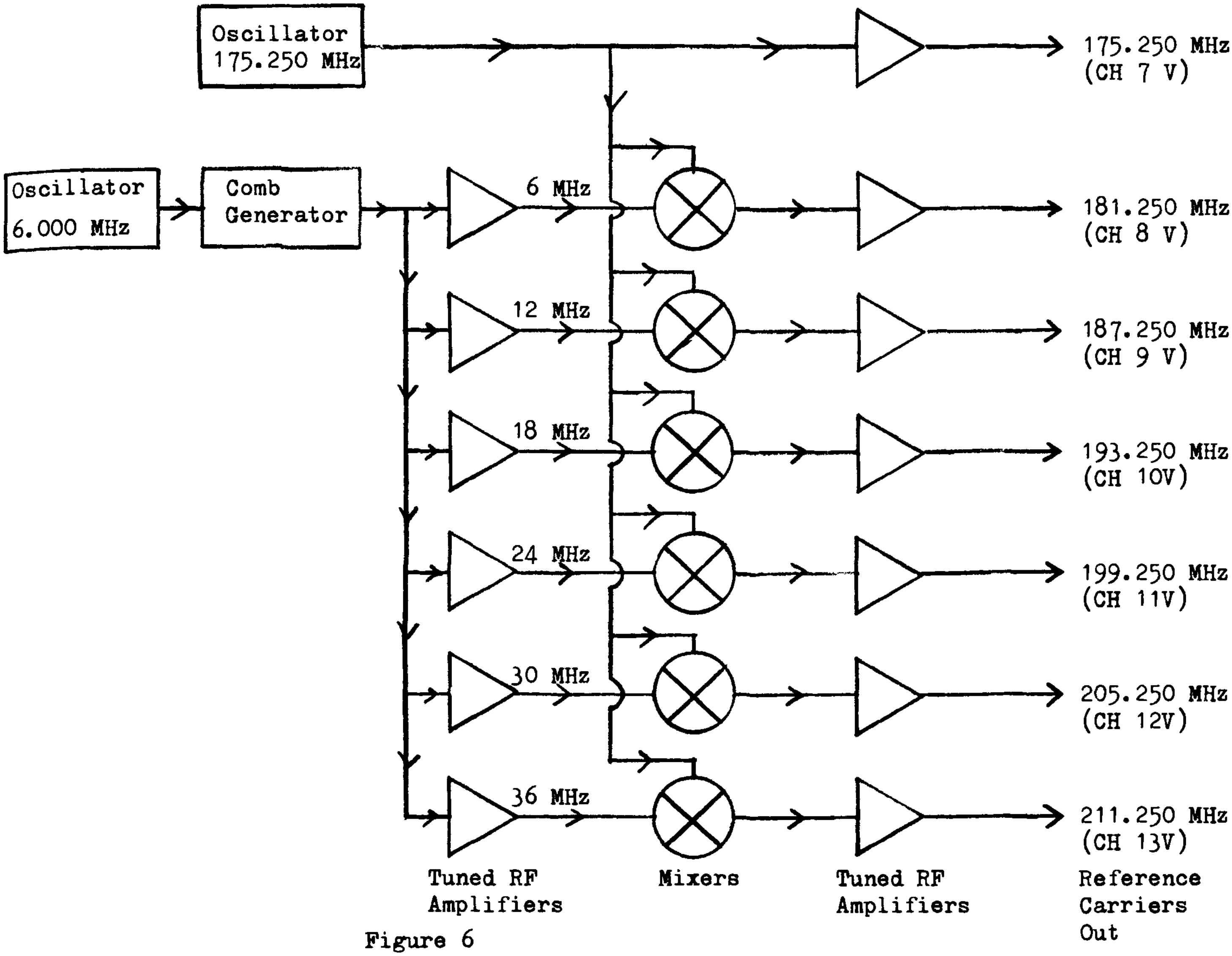


Figure 6

Block Diagram - Visual Reference Carriers locked to Common 6 MHz Spacing

PRESSURIZATION AND THE USE OF AIR DIELECTRIC CABLES

by K. R. Brubaker, General Manager
Altoona Video Corporation
Altoona, Pennsylvania
A subsidiary of Cypress Communications, Inc.

Since the use of pressurized air dielectric cables has never become popular for use in cable television systems people are usually amazed when I tell them I have used it for eight years and my experience with it has convinced me to plan a system rebuild which includes the use of air dielectric cable for trunk line.

All the figures and experiences which I will use are derived from the system located in Altoona, Pennsylvania. The trunk line is one half inch Spirofil which was installed in 1962. There are four major branches on the trunk line, the first branch occurring about seven miles from the antenna site. We have in use about 70 miles of cable with the longest single run being 17.6 miles. The entire system is pressurized from one point located at our engineering offices eight miles from the antenna site. Incidentally, we use air dielectric cables as down leads from all our antennas and we pressurize these lines from the system. The antennas are located about 800 feet from our head end building so they cannot be seen from the building in order to prevent any possible re-radiation pick-up. I realize that cable with higher losses may be advisable for use as down leads because of reflections from poor match at the receiving ends and we pad the lines as heavily as possible to improve match and use the gas filled cable primarily because of the indication we have if any structural damage, such as stray bullets in hunting season, should occur.

We use certified dry, oil pumped, nitrogen to pressurize the cable and use a pressure of ten pounds as standard. Several ounces above atmospheric pressure is all that is really needed to keep out moisture but the ten pounds gives us a better method for monitoring the performance of the pressure system. Another advantage of the higher pressure is that the cable is storing a larger volume of gas and in the event of a break in the system it takes much longer for the entire system to lose pressure. The system is broken up into 9 areas. The input of each of these grids is furnished with a gauge and shut off valve. Gas pressure and consumption are recorded at the engineering office at 8:30 A.M. and 4:30 P.M. daily and any difficulties in the system become apparent from these measurements. The employee who normally maintains the gas system, from experience, has learned what he should expect from temperature variations and has become quite expert at evaluating the variables of pressure, consumption and weather conditions.

In the event a problem occurs all the input valves to the networks are closed and the gauges monitored at three hour intervals until it can be determined which area is going flat. As soon as this is determined, usually a matter of a few hours, as much of the system is reactivated as possible. Naturally any area beyond the point of trouble will have to maintain its own pressure as a closed system until gas can be fed through the area of trouble. This presents no problem as most of the areas will lose only several pounds over a two week period due to the minute leaks which are always present in a system this size. We have had areas we have tested for thirty days and lost less than a pound in this period.

In order to pinpoint the trouble spot in a given section the section is broken up at each trunk amplifier. Each trunk segment is then pressurized to sixty pounds, a gauge installed, and sealed off. Very shortly the faulty segment can be spotted. Then a tank and regulator is installed at that amplifier location and the sixty pounds pressure maintained. In nearly every case merely walking along the trunk in this section will allow you to hear the leak if your ears do not cut off at high frequencies. We own equipment for soaping the cable in a very scientific manner with a tank, rollers and a maze of gadgets, also a supersonic detector, both of which we use in extremely stubborn cases but for day to day operation, sixty pounds of pressure and a man with good ears is much faster and a lot less bother. We were concerned in the early days about using high pressure on the cable but to date we have never been able to trace any trouble to the use of high pressure in leak detecting.

In a gas system it is necessary to by-pass the gas around every amplifier location. We use bulkhead fittings with a gas barrier and a hole in the side with a female three eighths pipe thread. The jumpers consist of Imperial Hydraulic fittings and Imperial Eastman Poly-Flo quarter inch tubing. This tubing weathers very well and requires replacement about every four years. In a solid state system I intend to block the gas at the input and output fitting of each trunk amplifier and then install a tee fitting in the by-pass tubing. I will then feed gas into the amplifier housing with a small shut off at the housing so that it may be opened without depressurizing the system. When the housing is closed the valve is opened to allow gas into the housing before the bolts are tightened to assure a nitrogen atmosphere in the housing. We are presently engaged in engineering tests and design for the necessary gas tight fittings to be used on the amplifier housings.

We have made no decision in regards to fittings splices, etc. on a rebuild which we are planning. We presently use the Entron SSU Splice Box and ER-851 flare fittings. When it comes to containing gas I am from the old school and a flare fitting when properly made with flaring tools in good condition has proved, for us at least, to be the most reliable. Great care must be used in making a flare using aluminum. Too much pressure on the flaring tool will decrease the wall thickness of the tubing to the point where it will break off under slight vibration.

Many people have expressed concern about the center conductor in an air dielectric cable having more freedom to move and as a consequence have been worried about pull outs during cold weather. This factor is really no worse than in any cable. We were very fussy when our installation was made to be sure that all pole bends were uniform and of such a radius to exert pressure on the center conductor and immobilize it. We developed a bending tool which was used at all pole bends and at every splice. As a result in eight years of operation we have had two pull outs, both occurring in below zero weather in an area where construction crews knew it would be difficult to inspect and were careless in making the proper bends. At splices we use a complete letter S and insist that the fitting cross the strand at ninety degrees.

When our system was installed structural return loss measurements were strictly for the better equipped laboratories, so we really do not know it's condition when new. We are fairly certain it was good because we have made quite a few summation sweeps including over fifteen miles of the cable and the results came out to plus or minus one half Db overall. We have had occasion to make return loss measurements on this cable after repairs in the past several years and it is not unusual to come up with 35 to 40 Db over the band of 50 to 220 megahertz. I mentioned repairs to cable. This may cause some raised eyebrows. However, we have had cases where someone put a gaff through the cable in wet weather and it became literally filled with water at a low spot. This has a very simple solution. Merely break the cable at the lowest splice you can find and install a fitting temporarily with a hole in the outer wall. At the highest available spot where gas can be introduced install a tank of nitrogen and purge the cable using the entire tank of nitrogen with the pressure regulator set at about 25 or 30 pounds. At a cost of about $6.50 for nitrogen and several hours of one man's time you have, in effect, replaced several spans of cable. The effect of water in the cable is of course the same as any other cable temporarily, high losses and severe reflections but as soon as it is purged the characteristics immediately are restored to their former values. We have never been able to measure any long term detrimental effects from water.

I am sure most of you are interested in how much it costs in terms of man hours to maintain a gas filled trunk line. In

Altoona we have accurate records of the performance of all our employees over a seven year span. Each man's performance is rated in units of work produced per hour every month so from these records it is a simple matter to extract the man hours spent on maintaining the gas filled trunk system. In fact this number is included on every monthly report I get from the chief engineer. In the past 24 months, 1,011.5 man hours were charged to gas maintenance. This averages to 42.1 man hours per month. The highest month was 76 man hours and the lowest 7.5. At $3.50 per hour this comes to $147.00 per month, or $2.10 per mile per month. Our annual cost for dry nitrogen is $311.69 or $25.97 per month or 37 cents per mile per month. This gives us a total of $2.47 per mile per month for materials and labor to maintain 70 miles of gas filled trunk line which is 8 years old. Even if your going rate for labor is double ours it is still not an item of any great consequence.

In summary I would like to list what I consider to be the principal advantages of a pressurized air dielectric system:

1. RF ground continuity of the outer conductor is assured. We lose our gas pressure long before the ground connection is poor enough to cause flashing on the trunk. In other words we locate and repair the trouble before the customer knows any trouble exists.

2. Attenuation does not change with age.

3. When water does enter the cable it can be purged and restored to like new condition with only momentary interruptions in service.

4. One advantage which is sometimes overlooked is that the attenuation versus frequency curve of air dielectric cables is flatter than most foam cables which makes for less change in equalization required for temperature change.

5. The lower loss of air dielectric cable will sometimes allow the use of the next smaller size cable and this offsets the extra cost for the air dielectric; however, the introduction of the new breed of foam cables has cancelled out this factor.

If I were to have to decide what I consider to be my chief reason for recommending air dielectric cable to you it would have to be continuity of service to the customer. We all know that troubles are going to crop up in any trunk line. The difficulties are much the same regardless of the type of cable whether it be gas filled, foam, or solid dielectric, the difference is the method of locating the problem. To site a "for instance", I am sure many of you have spent hours trying to locate a faulty splice which only gave trouble when the wind blew, usually after 10:00 P.M., when an important night football game was tied in the fourth quarter. We now have good reliable methods of locating this kind of trouble electronically and I am by no means knocking the Time Domain Reflectometer as a valuable tool. What I am saying is this, in a vast majority of cases this kind of trouble will become very evident in a properly maintained gas filled system long before the electrical integrity is destroyed plus the fact that the gas does not wait for the wind to blow to escape, ten pounds pressure takes care of this chore very nicely. In addition, electrical tests usually involve interruption of service to be properly done, which usually means night work (after 3:00 A.M. with present TV station schedules). Gas tests can be done during normal working hours with service maintained, except for the interruption for repair which can be scheduled and is usually of very short duration. I cannot remember a single customer complaint which could be traced to a trunk line problem in the past four or five years in Altoona, yet we have had many problems which we were aware of and as I said before, one of our employees spent over 1,000 man hours on these problems in the past 2 years. We feel that $2.47 per month per mile of trunk is money well spent in providing our customers with uninterrupted service.

DISCUSSION

Mr. Ken Brubaker: Thank you very much. Any questions?

Question: What about pressurization in feeder lines?

Mr. Brubaker: We don't use--I really don't think that gas filled cable has too much application in feeder lines. The advantage that you get from the gas filled cable, I believe, in a case of distribution would be offset by the amount of maintenance required on a system of that size if all your distribution lines were---remember, in a distribution line, especially in today's systems, where you have directional couplers or any kind of tap device, you can have a break in the cable at almost every customer point so that I believe you would be taking a chance of having to contend with a fitting which must be gas type at every pole. In other words, in a trunk line we're talking now 3000 feet--something like that, in that general order--between splices where in a distribution line, you're talking about a splice at every pole. And I question whether that would prove to be practical.

Question: When you start to pressurize your amplifier cases, do you still put 60 pounds on this.........

Mr. Brubaker: I think maybe I may have been misunderstood. This 60 pounds is only used in the event there is trouble. The normal pressure in a cable is 10 lbs. per sq. inch. This is what we maintain as a norm. When we experience trouble in a given section and break it up and seal it off--and this is just between trunk amplifiers--then we'll pressurize that small section to 60 lbs. We wouldn't normally, I don't think expect any amplifier or any cable system to be able to maintain this kind of pressure as a general rule. Ten lbs. per square inch is the pressure which we normally use and several ounces is really all that is required, but, as I said, the cable stores more gas under 10 lbs. and it does give you a better indication on the guage and so on. Sixty pounds is only used for testing.

Question: About how far does your pressure take?

Mr. Brubaker: We only pressurize at one point. We have 70 miles of cable all pressurized from one point. I'll tell you--the big, and this is a real headache, and there's a gentleman in the process right now near Erie somewhere who called me the other day--who is in the process of installing a gas filled system. This is the real headache--your first go around to once get everything gas tight. Once this is achieved, maintaining it is no problem. But to get it gas tight the first time, I have to confess it's murder. But, like I say, in the eight years experience that we've had with it, this was the only bad thing that we ran up against and actually the company that installed the system for us gave up and I took one of our men and spent about three months, myself, with this man getting the thing finally gas tight and once it was gas tight, it was no problem keeping it that way.

Question: One other question--do you cover the amplifier with a piece of copper tubing?

Mr. Brubaker: Plastic tubing--right. This plastic tubing has a life of somewhere over four years.

SIGNAL LEVEL METER CALIBRATION TECHNIQUES

By Fred J. Schulz, V.P.CATV Eng.
Sterling Communications, Inc.
43 W. 61st St., New York, N.Y.10023

June 1970

1. Introduction

The need to measure signal levels is an ever present one in any CATV system. The need to very accurately measure the absolute value of levels is a necessity in multichannel long cascade systems. In smaller systems it might have been adequate to correlate the readings of the various used meters irrespective of absolute calibration. This approach can no longer be defended.

The Signal Level Meter (SLM) is a basic tool of our trade, like the level and plumb are for a mason. Periodic check of SLM's is an absolute necessity and must be done very carefully and with reliable standards since the performance of the whole CATV system depends on proper level settings.

2. The nature of the TV signal

The output of a TV transmitter (picture information only) is as shown below:

Fig.1

The RF carrier strength varies with the modulation and is greatest during sync peak (interval A). We measure signal strength during sync interval, since the RF level is then steady. This requires a meter responding to the sync peak. A good SLM has therefore a true peak detector.

3. The basic SLM calibration setup

To calibrate a SLM one feeds a RF signal of known value into the meter. Signal generators used for this purpose have a sinusoidal output waveform. It is universally accepted practice to state sinusoidal waveform amplitudes as RMS (root mean square) values. The RMS value is the effective value, the value used when one computes power as the product of AC voltage times AC current. AC meters, unless otherwise stated are always calibrated in RMS value of a sinewave. The peak or crest of a sinusoidal waveform is approximately 40% higher than its RMS value as shown below:

Fig.2

(cont'd)

The SLM responds to the crest of the sinusoidal calibration signal, but it was agreed to calibrate it in RMS value as shown below:

Fig.3

If we, for example, feed a 40 dBmV (100 millivolt) signal from a generator into a SLM and calibrate at this value, a TV waveform with a 141 mV peak value will now read 40 dBmV. This is why we say that SLM's are calibrated in RMS value at sync peak.

4. Available generators/calibrators

SLM manufacturers usually use generators designed for 50 ohm loads. Such generators cost in the neighborhood of $1,500. These generators are suitable for SLM calibration, if properly matched to 75 ohm SLM's, but one must remember that they are not absolute voltage standards either. Laboratory generators also have often a lot of features which are of little use to a CATV operator such as accurately calibrated modulation capabilities; very accurate frequency calibration,etc.

The most important and desirable characteristics of a generator for SLM calibration are:

a) 75 ohm output impedance

b) good sinewave output

c) output level indicator

d) good attenuator

e) stable output over long periods of time

f) moderate output level

g) reasonable frequency accuracy

h) a stable output with temperature change

Several "low cost" generators (SLM calibrators) meeting at least some of the above specifications are available, namely:

1) Delta - model FSM-C4

2) St.Petersburg Communications - model C-524

3) Measurement Corporation - model 950

(cont'd)

One high priced unit featuring 15 separate calibrated oscillators is available from Video Instruments at $1,285.

A new type of calibrator from JFD will be discussed later. We will examine the block diagrams of the 3 above mentioned calibrators.

Delta Model FSM-C4 ($112.50)

Fig.4

This unit is very simple, it has a single transistor oscillator, the output of which is detected and indicated on a meter and its amplitude can be set by adjusting the oscillator voltage. Isolation to the output and 75 ohm back match are provided.

The calibration output voltage for each channel is marked on the front panel and is typically 15 dBmV. To obtain other values one must use an external attenuator. Frequency range is from 54-250 MHz continuous and the dial is fairly well calibrated.

St.Petersburg Comm. Model C-524 ($149.-)

Fig.5

This unit consists of a RF oscillator, a peak to peak detector with associated AGC feedback circuit to keep the output constant, a 15.75 KHz modulation oscillator, a 3 step attenuator and a regulated 117V power supply. The unit is 75 ohm backmatched. Output values from 40 dBmV maximum to -10 dBmV can be obtained in 10 dB steps. Frequency range is from 50-240 MHz in two bands with approximate dial calibration.

Measurements Corp. Model 950 ($380.-)
(Also available from Vikoa as their model 3913)

Fig.6

This product is a scaled down laboratory signal generator (Measurements Model 80). It consists of a Nuvistor oscillator, a bolometer type RF signal level detector with feedback circuitry to keep the output constant, a balance indicator and a precision 75 ohm backmatched piston attenuator. The power supply is for 117V and supplies regulated DC to all circuits including the

(cont'd)

Nuvistor tube filament. The frequency range is from 54 to 250 MHz in 2 bands with hand calibrated frequency markings within ± 0.5%. The output level can be adjusted continously from 40 dBmV to -70 dBmV with markings every 2dB.

5. Calibration techniques

It makes sense to perform the prime calibration of the SLM's at the most critical signal level to be measured in the CATV plant. The output of the trunk amplifiers is the most critical level. Since levels are measured through test points these must be taken into account. Most trunk amplifiers run at approximately 30 dBmV and use either 20dB or 30dB test points, so calibration at 10dBmV or 0dBmV is indicated.

The calibration setup is simply a length of coax cable connecting the calibrator and the SLM. The cable must be of good quality checked to 26-30 dB return loss and 75 ± 1 ohm. It is also recommended that the same relatively short (approx. 18") cable be used all the time to eliminate a further variable. A good match on the whole setup is important since standing waves can change the calibration considerably.

It is recommended that one uses a separate calibration record sheet for each SLM. A suggested sample form is shown in the figure below:

Fig.7

The calibration procedure consists of several parts:

a) Channel calibration performed at each picture and all pilot carrier frequencies; recording either compensator settings or actual readings for meters without compensators. In either case one can then from the record sheet, prepare a compensator setting table or a correction chart suitable to keep with the SLM (large self-adhesive labels are good for this purpose). The record sheet is kept in the shop it is an inventory and performance record. Deviations from prior calibrations should be noted and if more than a dB or so, close examination of the instrument (or the calibrator) is indicated.

b) One should check the accuracy of the step attenuators against a good standard. This check is recommended at Channel 2 and Channel 13.

(cont'd)

c) Scale calibration check is done against an external variable attenuator. Set SLM for full scale reading at one channel, adjust a 0-10dB attenuator dB by dB and watch for correlation of reading on the meter scale.

d) It is desirable to check the SLM input match occasionally.

e) Most meters are calibrated at a shop temperature of 70-80° F. When setting levels in the field the instrument may be at a much lower or much higher temperature. Good SLM's are temperature compensated, but not every unit is given a temperature test at the factory. The compensation can also become faulty in the field. Check the compensation once a year by placing the SLM in a cold or hot area (after calibrating it in the shop at 70-80° F). Let the SLM stabilize for approx. one hour; readjust voltage and repeak before making the new reading (calibrator must of course stay in the shop at a steady temperature).

There are a few additional guide points on SLM calibration:

a) Calibrate trunk meters at least every 4 weeks.

b) Calibrate installer meters approximately every 8 weeks.

c) Put sticker on SLM showing last calibration date and next calibration date. Meters should be recalled by the office for recalibration, don't rely on the technician to bring back meters.

d) Keep SLM's physically clean.

e) Check SLM's after factory overhaul; the unit may have been damaged in transit.

f) Replace sticky meter movements promptly.

g) Check calibrator every 6 months.

6. Calibration of generators/calibrators

All SLM producers use signal generators to calibrate SLM's and calibrators. The generators in turn, have to be calibrated themselves and for that purpose a variety of instruments are

(con't)

used. The most often used device is a bolometer, it measures power. Power can be measured quite accurately and it is a desirable quantity to measure particularly at microwave frequencies, since it is not dependent on match (VSWR) as voltage measurements are.

A bolometer is, in effect, a low power wire resistor, which changes resistance as it is heated by the to be measured RF signal and this resistance change is used in a bridge circuit with a meter, in turn calibrated in power (watts, milliwatts). To calibrate the bolometer itself one can use DC which can be measured very accurately. To measure voltages above approximately 100 mV the bolometer is the preferred method. Commercial units are however built with an impedance of 50 ohms, requiring matching for 75 ohm use.

Another method of calibrating signal generators is the use of a micropotentiometer, which contains a wire resistor heated by the to be measured RF signal flowing through it, the generated heat in turn is transferred to a thermocouple, which generates a voltage suitable to drive a DC millivoltmeter. In series with the RF path is an accurately measured low value resistor. Again, one can calibrate this setup by applying DC to it. The micropotentiometer is the preferred method at the National Bureau of Standards to calibrate Field Strength Meters, since low value RF signals can accurately be produced (see bibliography #6).

Both of the described methods are wideband and measure power. The voltage in turn is computed from the indicated power and the known instrument impedance in case of the bolometer and from the known current and resistance in the micropotentiometer method. Since SLM's measure peak voltage and above methods are based on power we must use a reasonably low distortion sinewave signal source for our calibration or additional errors will result.

The availability of a micropotentiometer together with the necessary accurate DC standards and a good stable Rf source allows one to check calibrators with accuracies tracable to the National Bureau of Standards. Such a setup is costly and it is warned that one cannot merely buy the gear, plug it together and expect NBS accuracy. A good deal of knowhow and care in making the measurements is needed. Micropotentiometers are available from sources like Ballantine (Boonton, N.J.) and Filmohm (L.I.City, N.Y.).

Another means of checking calibrator accuracy is to own two good calibrators and checking them against each other at periodic intervals and returning them both to the factory when they no longer agree with each other within specifications.

(cont'd)

Some commercial instrument calibration laboratories are equipped to check calibrators.

It is also possible to use so called RF milli or micro voltmeters, but these instruments are not very accurate and in turn must be calibrated themselves with better instruments.

7. Suggested equipment packages

	Estimated attainable overall accuracy:
A. Minimum - Delta or St.Petersburg calibrator. $115.-/$150.-	± 2 dB
B. Better - Measurements Corporation calibrator. $380.-	± 1.5 dB
C. Reassurance - Two of the above calibrators, one used as an "in-house" standard, the other used to check SLM's	
D. Best - Above calibrator, plus micro-potentiometer, RF generator voltmeter etc., to facilitate checking of calibrator. Approx. $2,000.-	± 1 dB

E. Available, but not favored:

Use of standard signal generators such as:

HP 608
Measurements Corp. Model 80 (discont.)
GR model 1021 (discont.)

Reasons: a) Generators themselves need calibration

b) Units are 50 ohm requiring matching and computations for calibration of 75 ohm SLM's

c) Old used generators may be defective and not accurate.

d) Output voltmeters read voltage into load on some units, behind 50 ohm on others, requiring computations.

F. Not acceptable - Election of one meter as a standard.

(cont'd)

8. Other methods of calibrating SLM's

Precision field strength meters ($3,000.- price range) feature built-in calibrators. These calibrators are of the impulse generating type and require no tuning over very wide frequency bands, they, in essence, provide signals at all frequencies simultaneously. The calibrators are available as separate units from Stoddard and Singer, but are of 50 ohm design.

Similarly noise may be used to calibrate SLM's. A calibrator using this principle is now available from JFD. After initial setup against a built-in single frequency generator no tuning on the generator is needed, the SLM hooked up to the generator can then be tuned through it full range giving a reading all along. The SLM reading depends on its bandwidth since both noise and impulse spectrums contain energy throughout their wide frequency ranges. The figures below illustrate this.

Fig.8 Fig.9

9. Effect of calibrator harmonics

Calibrators may have various degrees of harmonics in their output, this is not desirable since the calibrators are checked with power meters reading energy content of the fundamental and the harmonics. The SLM in turn, is a tuned voltmeter and reads the fundamental component only. Computations indicate that the error is not very large even for relatively large harmonic levels.

	Error on SLM
One harmonic 10 dB below fundamental	0.4dB
" " 15 dB " "	0.1dB
" " 20 dB " "	0.05dB
Two harmonics each 10 dB below fundamental	0.75dB
" " " 15 dB " "	0.25dB
" " " 20dB " "	0.1dB

10. Bibliography

1. TV Communications, Dec.1969, Jan.1970.
   "SLM Calibration" by I. Switzer

2. TV Communications, Feb.1970
   "Use of FSM" by Robert D. Bilodeau

3. TV Communications, May 1967
   "Characteristics of FSM's" by F.J.Schulz

4. Cablecasting, Feb.1970
   "Calibrating FMS's" by Larry Roeshot

5. Jerrold Technical Handbook for CATV systems

6. National Bureau of Standards Tech. Note 370
   "Calibration of FSM's" by H.E.Taggart & J.L.Workman

7. Proceedings of the IEEE, June 1967
   "Radio Measurement Methods and Standards"

8. Hewlett-Packard Application Note 64
   "Microwave Power Measurements"

9. Hewlett-Packard Application Note 60
   "Which AC Voltmeter?"

10. Transactions of AIEE, May 1953
    "Accurate RF Microvoltages" by Myron C. Selby

11. IEEE Convention Record 1965
    "Measurement of low RF Levels with Micropotentiometers"
    by K. M. Klarer

12. IRE Tech. Committee, 62 IRE 20 TRI
    Measuring Sinewave unbalanced RF voltage

13. IEEE, No.302, August 1969
    Standards for Measuring Field Strength

14. NAVSHIPS 94180, July 1962
    "The Radio Frequency Interference Meter"

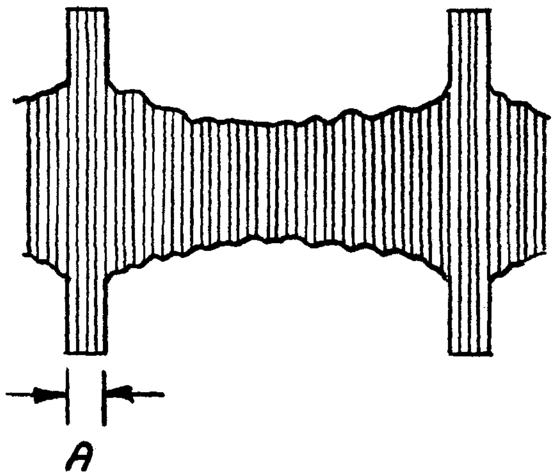


Fig. 1

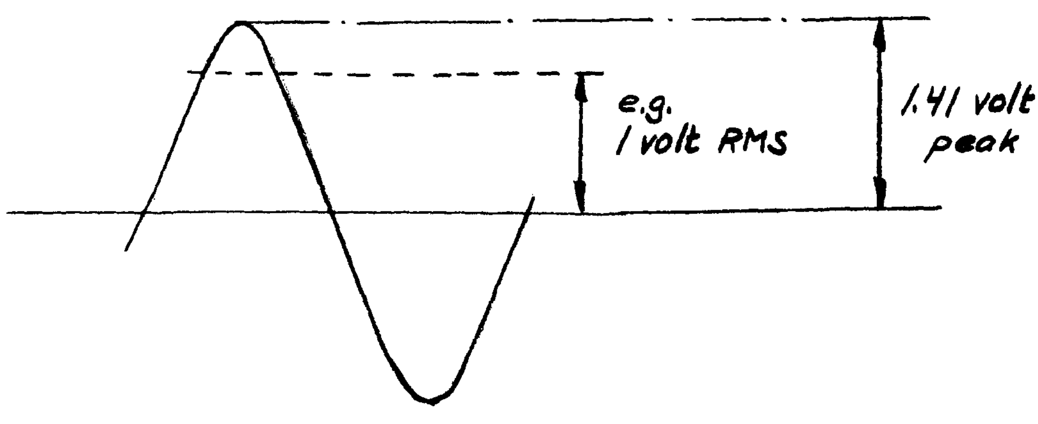


Fig. 2

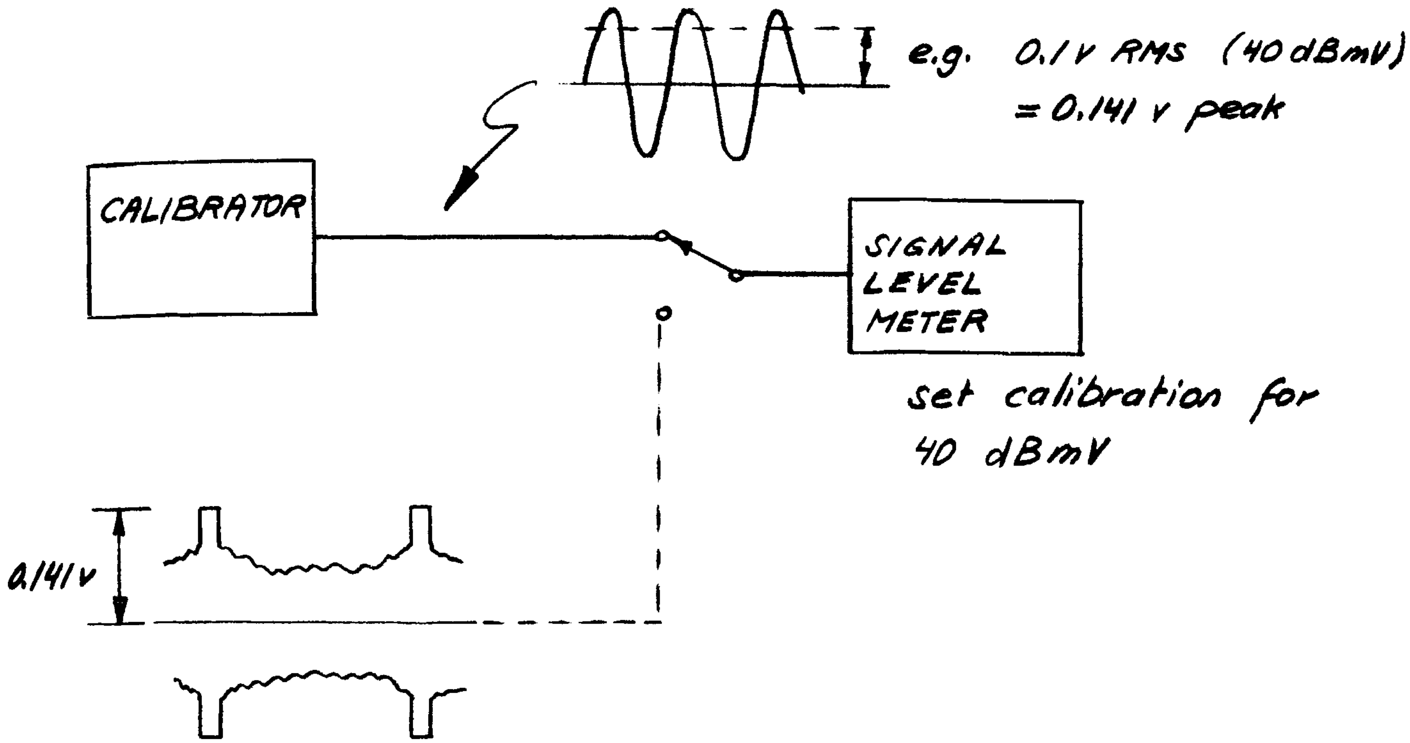

e.g. 0.1 v RMS (40 dBmV)
= 0.141 v peak
CALIBRATOR
SIGNAL
LEVEL
METER
set calibration for
40 dBmV
0.141 v

Fig. 3

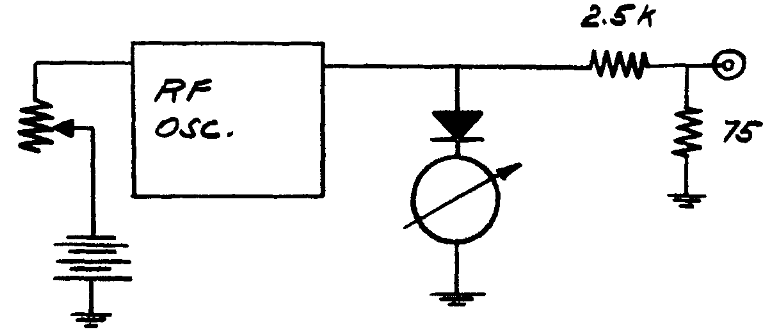

2.5 k
RF
OSC.
75

Fig. 4

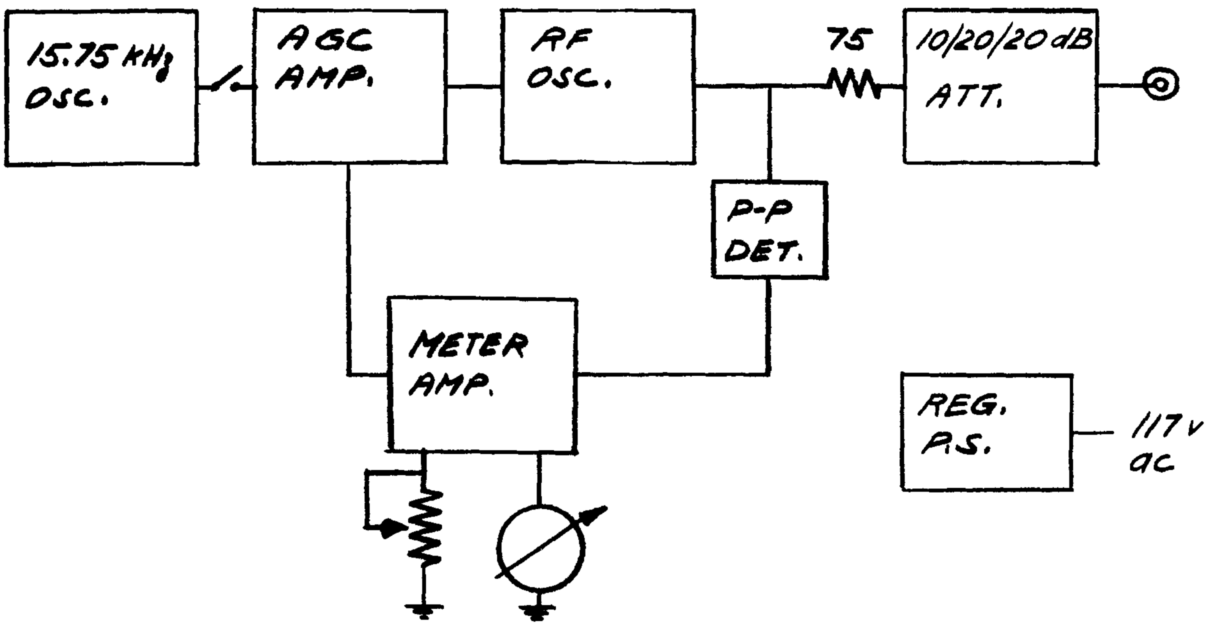

15.75 kHz
OSC.
AGC
AMP.
RF
OSC.
75
10/20/20 dB
ATT.
P-P
DET.
METER
AMP.
REG.
P.S.
117 v
ac

Fig. 5

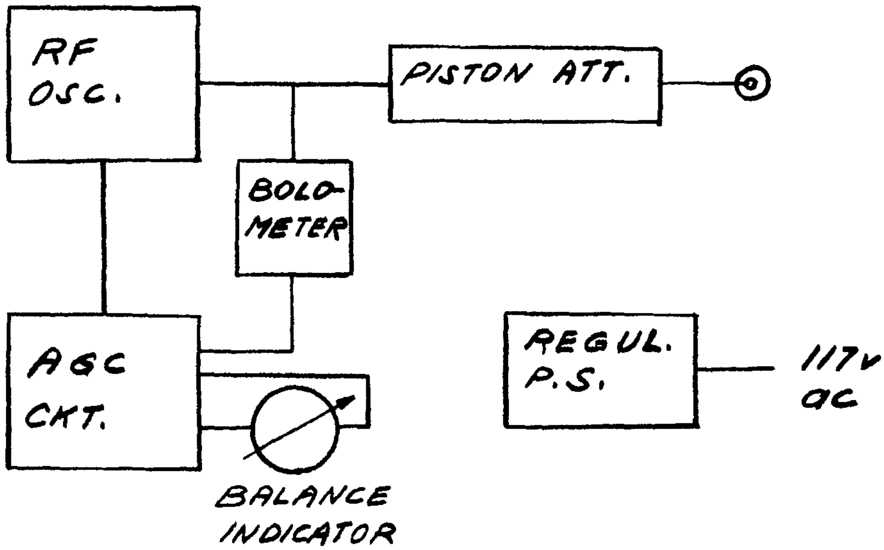
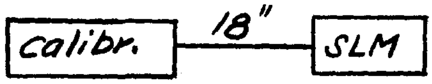

RF
OSC.
PISTON ATT.
BOLO-
METER
AGC
CKT.
BALANCE
INDICATOR
REGUL.
P.S.
117v
ac

Fig. 6

SLM CALIBRATION RECORD

Model: ------ Ser. No. ------

1. Channel calibration

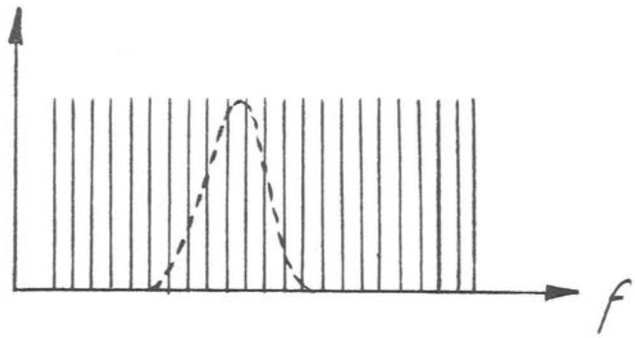


set meter and calibrator for ------ dBmV

channel	CALIBRATION DATE AND COMPENSATOR SETTING ORIG. -/-/-	-/-/-	-/-/-	-/-/-	-/-/-
2 PIX					
3 PIX					
4 PIX					
⋮					

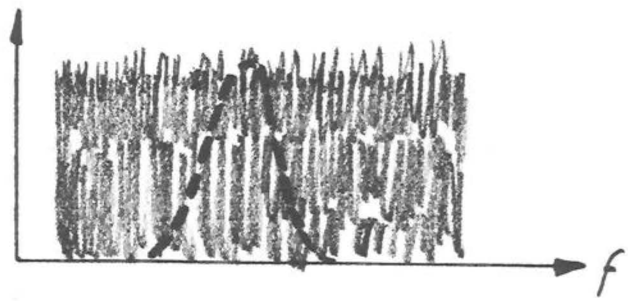
2. Attenuator check

10dB 2P					
10 dB 13P					
20dB 2P					
20 dB 13P					
20dB 2P					
20dB 13P					



IMPULSE GENERATOR SPECTRUM

Fig. 8



NOISE GENERATOR SPECTRUM

Fig. 9

SPATIAL-FREQUENCY ENCODING TECHNIQUES APPLIED TO A ONE-TUBE COLOR TELEVISION CAMERA

Albert Macovski and Earle D. Jones

ABSTRACT

An anlysis and experimental results are presented of a new color camera technique using a single camera tube. To avoid the difficult registration problem of conventional color cameras, the color information is encoded as the amplitude of two diffraction grating patterns. This is accomplished by imaging the colored scene onto a pair of colored diffraction gratings. The recovery of the color information is done by band-pass filtering of each channel followed by envelope detection. The three color signals are derived using linear combinations of the two decoded carriers along with the average signal which represents a weighted sum of all three colors.

The performance of the camera is determined by resolution, beats, crosstalk, and noise considerations. Many of these considerations are aided by proper design of the grating filter from the viewpoint of the two-dimensional spatial frequency spectrum of the encoded image. These considerations lead to the conclusion that the optimum configuration uses two gratings which are equal in periodicity and are at different angles. The final system uses a vertical red-absorbing grating and a 45° blue-absorbing grating, each having the same periodicity.

*Stanford Research Institute, Menlo Park, California 94025

INTRODUCTION

For the past few years Stanford Research Institute, under the support of RCA, Inc., has studied techniques by which color information can be encoded with black-and-white processes. The first experiments demonstrated the feasibility of recording color information onto black-and-white film through the use of a spatial-frequency carrier technique. Later a similar approach was used in conjunction with a single vidicon television camera to create a color television signal. This paper reports the results of the television camera portion of that research project.

Color television cameras today have limited acceptance because they are expensive and difficult to maintain. This is primarily because the output signal is derived from three separate camera tubes. This problem necessitates rugged and accurate construction to provide proper optical registration, as well as stable, accurately controlled circuits to ensure tracking of the three scanning beams. A number of attempts have been made to minimize these difficult problems; for example, a four-tube camera was developed for the purpose of making the broad-band luminance signal immune to registry errors by deriving it from a single camera tube rather than from the sum of the red, green, and blue cameras.[1] The colored edges caused by misregistration of these three tubes, however, continues to be a

1. H. N. Koznowski and S. L. Bondell, "Recent Developments in Color Television Camera Equipment," IEEE Trans. on Broadcasting, Vol. BC-9, pp. 31-36, February 1963.

problem. Another approach uses field-sequential color cameras employing moving color filters in the optical path of the camera.[2] Field-sequential cameras require accurate field storage devices, such as magnetic discs, to provide a compatible parallel readout of the serial red, green, and blue information. Even when done properly, the system suffers from color breakup for objects that move and therefore occupy different positions during successive field scans. Some field-sequential systems have been operated at field rates higher than standard to avoid color breakup. These systems, however, require expensive scan converters and usually suffer a resolution loss.

Another class of television cameras exists, including the one described here, where the various colors are encoded on spatial-frequency carriers or grating structures. This class represents the most successful approach to the registration problem to date. Two subclasses can be described. One subclass of these employs an optical filter consisting of alternating vertical lines of color filter material. The chrominance information modulates a subcarrier in phase and amplitude. The color signal is then extracted by sampling the video signal at the appropriate time for each color. The principal difficulty with this system is the severe requirement on scan linearity, since no reference exists for the sampling signal. A similar approach has been used employing two registered camera tubes where one tube supplied the luminance signal and the other contained a coarse structure of alternating red and blue lines.[3] This coarse line structure was adequate, since only relatively narrow-band chrominance information was derived from this tube. The tube with the grating structures supplied the red and blue signals; the green signal was derived by subtracting the red and blue signals from the luminance signal output of the

---

2. C. G. Lloyd, "Chromacoder Colorcasting," IRE Trans. on Broadcast Transmission Systems, Vol. PGBTS-1, March 1955.
3. P. S. Carnt and G. B. Townsend, Color Television, London: Iliffe Books Ltd., 1961, p. 72.

other camera tube. The scan-linearity requirement is somewhat relaxed by the wide color stripes, but the camera is still far from ideal because of the remaining scan-linearity problem and the need to register the two tubes.

The other subclass of systems employs frequency multiplexing, where the various color components appear at different frequencies and can be conveniently separated by electrical filtering.[4,5] These systems are quite immune to the normal changes in scan linearity, since the resulting frequency changes are not large enough to significantly alter the selectivity characteristics of the filters. The difficulty with these stems from crowding of the frequency spectrum. The various color encoding carriers are created by colored gratings in an image plane that cause the color information to modulate a carrier. These carrier frequencies, in addition to any beats between them, must not fall within the luminance and chrominance pass bands. If all of these grating structures consist of vertical lines, the problem of beats is difficult to avoid. The resultant grating frequencies when scanned should be outside the luminance pass band. Any attempt at meeting these requirements would require a camera tube with ultrahigh resolution, far beyond that required for the conventional broadcast standards.

This paper describes a novel system that makes efficient use of the two-dimensional resolution capability of a camera tube and avoids the problem of beat visibility.[6] Thus a conventional vidicon camera tube achieves sufficiently high performance that the limiting component in the overall system is still the color receiver.

---

4. U.S. Patent No. 2,733,291, R. D. Kell, January 31, 1956.
5. "Electronics Abroad," Electronics, Vol. 40, pp. 235-236, January 1967.
6. U.S. Patent No. 3,378,633, A. Macovski, April 16, 1968.

SPATIAL FREQUENCY SPECTRUM

We first examine the two-dimensional spatial-frequency spectrum of the camera to see where the additional color information can be added. In television systems, assuming ideal camera devices and reproducers, the resolution in the horizontal direction is limited by the bandwidth, and that in the vertical direction is limited by the number of scanning lines. If we restrict our discussion to U.S. standards, it is convenient to express both vertical and horizontal spatial frequencies in terms of equivalent electrical frequency in megahertz. For example, the maximum resolvable horizontal spatial frequency f_x, given a bandwidth of f MHz, is given by

$$f_x = \frac{f}{v} ,$$

where v is the horizontal scan velocity in unit distance per μs. Using this same notation, any spatial frequency in any direction can be expressed as an equivalent electrical frequency divided by v. This notation is convenient in that it is independent of the camera-tube dimensions. The maximum resolvable vertical spatial frequency is given by

$$f_{y\ max} = \frac{N}{2L} ,$$

where N is the number of active scanning lines and L is the raster height. This can be expressed in terms of equivalent electrical frequency by multiplying the above expression by v, the scanning velocity. For the U.S. standards

$$N = 485, \text{ and } L = (\tfrac{3}{4})53.0\ \mu s \cdot v .$$

Then

$$f_{y\ max} = \frac{6.1\ \text{MHz}}{v} \quad .$$

The equivalent electrical frequency is 6.1 MHz.

Figure 1 shows one quadrant of a two-dimensional spatial-frequency spectrum with the axes dimensioned in equivalent electrical frequency (MHz). The spatial-frequency capability of the vidicon camera is shown as a quarter-circle of radius 6 MHz. The U.S television standards appear as a square with sides of 4 MHz. (The real equivalent frequency of 6.1 MHz derived earlier must be multiplied by the Kell factor to obtain the nominal vertical resolution.) The average color-receiver capability is shown as a rectangle with f_x approximately 3 MHz and f_y established by the standards at 4 MHz. These numbers are only approximate and are used to show that a significant portion of the vidicon capability is not being used. Further, this type of diagram shows where the available resolution capability lines in the spatial-frequency spectrum. Third, it allows one to compute directly the spatial and temporal frequency of beat patterns that appear in the picture when stripe or grating structures are used.

ENCODING OPTICS

In color television, two narrow-band sources of chrominance information must be added to the wideband luminance signal in order to fully define the desired image. In the camera described here, color information is encoded as amplitude modulation onto spatial-frequency carriers using colored gratings. Figure 2 shows the simplest embodiment of such a system. Lens L_1 images the scene to be televised onto a filter plane. The relay lens, L_2, images the filter plane onto the vidicon photocathode. The filter is a colored grating structure comprising alternate vertical stripes of cyan filter separated by transparent stripes. Disposed at some convenient angle to the cyan stripes is a grating of alternate yellow and transparent stripes. Assuming ideal filters, the cyan material passes blue and green light and blocks only the red portion of the spectrum. Thus the cyan grating is a grating only to red light. Similarly, the yellow grating acts only on the blue component of the incoming light, passing the red and green components undisturbed.

As the vidicon is scanned, two carriers are generated, because the red component of the light causes a vertical grating pattern on the vidicon faceplate and the blue component causes a diagonal grating pattern. The two gratings must be used in such a fashion that the resulting grating signals, as well as the beat between the two signals, will have acceptably low visibility in the luminance pass band. In addition, the grating spatial frequencies must be within the camera resolution limits so that the grating signals can be resolved with adequate signal-to-noise ratio. As shown in Figure 1, the cyan (red encoding) grating is placed in a convenient portion of the spectrum on the f_x axis. This point appears on the f_x axis, since the x-component of its spatial frequency is 5.0 MHz and the y-component is zero (vertical lines). The yellow (blue-encoding) grating is placed diagonally at an angle of approximately 45°. It has the same spatial frequency as the red-encoding grating. Thus both gratings have the same spatial frequencies located at the edge of the camera resolution limit and create electrical frequencies that are outside the luminance pass band of the receiver. Although the beat between the two carriers occurs at the relatively low electrical frequency of 1.5 MHz, it has a spatial frequency of 3.8 MHz. The spatial frequency and angle of the resulting beat pattern can be determined from Figure 1. The distance between the two points corresponds to the spatial frequency of the beat, and the direction of the line connecting the two points indicates the axis of the resulting beat. Although the electrical frequency of the beat is well within the luminance pass band, its visibility is low because of the high spatial frequency. With ideal cyan and yellow gratings and a linear camera, no beat will be visible. Since ideal yellow and cyan filters do not absorb in the same part of the spectrum, there is no color that will interact with both gratings, creating a product action and thereby producing beats. With practical filters, however, each filter has some absorption in its pass band and some attenuation in its stop band so that a beat is generated. In addition, the nonlinearity of vidicon cameras will create a beat wherever both red and blue signals are present simultaneously. However, the cameras that have been built show that the relatively small beat amplitude and its high spatial frequency combine to make the beat essentially invisible.

The relay lens system shown in Figure 2 represents the most straightforward arrangement of lens, filter plane, and vidicon. Alternative methods exist. For example, the encoding filter could be placed inside the vidicon envelope, directly against the photoconductor. It could also be placed outside the camera, with a fiber optics faceplate used to place the filter into optical contact with the vidicon photocathode. The relay lens has the advantage that a simple, unmodified vidicon can be used. The relay lens arrangement also allows one to employ a large, coarse filter structure, since the relay lens can be used to demagnify the filter. A field lens would be included in the optical design to achieve higher light efficiency. The disadvantages of the relay lens approach are the added cost and size of the lens itself.

DECODING ELECTRONICS

When the two-color encoding gratings are disposed as described above, the two resulting carrier frequencies are in the ratio of $\sqrt{2}$ to one. The frequencies chosen were approximately 5.0 and 3.5 MHz. The spectrum of the vidicon signal and the filters used to separate the various components are shown in Figure 3. A system block diagram is shown in Figure 4.

The bandwidths around each color carrier provide 0.5 MHz chrominance bandwidth, which represents typical home-instrument performance. Larger bandwidths can be used but give greater overlap in the spectrum and increased crosstalk between the various channels. The most serious crosstalk in the system occurs when high-frequency luminance signals appear in the lower-frequency (3.5 MHz) color channel. This crosstalk appears as a blue edge occurring where high-frequency luminance signals are present, such as sharp luminance transitions. The solution to this problem, as would be expected, is to make the recovered 3.5-MHz color signal large compared to the luminance crosstalk. The amplitude of the color signal is increased by using an efficient grating having good absorption in the desired region. High-frequency luminance information can be reduced by using astigmatic optics (a weak cylindrical lens) that limit the high-frequency response in the horizontal direction only. The models that have been built show that crosstalk can be reduced to a negligible level. After the various channels are separated, the high-frequency luminance response can be restored using horizontal aperture compensation.[7]

LINEARITY CONSIDERATIONS

The linearity of this system has been studied both analytically and experimentally. We first consider the linearity problem caused by the nonlinear transfer characteristic of a vidicon. Assuming a constant gamma,

7. G. M. Glasford, Fundamental of Television Engineering, New York: McGraw-Hill, Inc., 1955, pp. 484, 495-497.

the transfer characteristic of a vidicon is given by

$$i = I^{\gamma} \quad ,$$

where i is the output current and I the incident light intensity. For a given color, the average light intensity on the photocathode will be A . Let the peak intensity of the fundamental component of the red encoding grating be P_r and that of the blue encoding grating be P_b . The resulting signal current is thus given by

$$i = (A + P_r \cos\omega_r t + P_b \cos\omega_b t)^{\gamma} \quad ,$$

where ω_r and ω_b are the carrier frequencies due to the red-encoding and blue-encoding gratings. This expression is expanded below in a series including terms up to the third order. The average value of the expression represents the luminance components, and the fundamental amplitudes of $\cos\omega_r t$ and $\cos\omega_b t$ represent the envelope-detected red and blue amplitudes. Harmonics and products of these frequencies are not important, since they do not contribute to the output. The resulting signal current is given by

$$\begin{aligned} i = {} & A^{\gamma}\left\{1 + \frac{\gamma(\gamma-1)}{4}\left[\left(\frac{P_r}{A}\right)^2 + \left(\frac{P_b}{A}\right)^2\right]\right\} \\ & + A^{\gamma}\left\{\gamma\frac{P_r}{A} + \frac{\gamma(\gamma-1)(\gamma-2)}{24}\left[3\left(\frac{P_r}{A}\right)^3 + 6\frac{P_r}{A}\left(\frac{P_b}{A}\right)^2\right]\right\}\cos\omega_r t \\ & + A^{\gamma}\left\{\gamma\frac{P_b}{A} + \frac{\gamma(\gamma-1)(\gamma-2)}{24}\left[3\left(\frac{P_b}{A}\right)^3 + 6\frac{P_b}{A}\left(\frac{P_r}{A}\right)^2\right]\right\}\cos\omega_b t \quad . \end{aligned}$$

The first term represents the Y , or luminance, output; the peak values of the second and third terms represent the red and blue outputs, respectively. Note that the relative amplitudes of these components (which determine the generated color) depend only on the <u>ratio</u> of the various peak-to-average values of the intensities, and not on their absolute intensity. Thus each color will track properly over the dynamic range. For example, if the balance is set up properly on any step of a neutral grey scale, it will remain balanced over the entire grey scale, since the relative amplitudes of luminance, red, and blue, (and therefore green) will be in the same ratio for all values of intensity.

Even though the ratio of the signals are not affected by the non-linear vidicon characteristic, the derived signals R-Y and B-Y can be adversely affected. As shown in Figure 4, the color and luminance signals are combined to form (R-Y) and (B-Y) . The relative amplitudes of R and Y are adjusted to be equal for a white signal where $R = B = G = Y = 1$. In order to make (R-Y) zero for this condition, using a gamma of 0.5, the red signal must be made 3.28 larger than that of the Y signal. The magnitude of (R-Y) is given by

$$(R\text{-}Y) = A^{\frac{1}{2}}\left\{3.28\left[\frac{P_r}{A} + \frac{1}{32}\left\{3\left(\frac{P_r}{A}\right)^3 + 6\left(\frac{P_r}{A}\right)\left(\frac{P_b}{A}\right)^2\right\}\right] - 1 + \frac{1}{16}\left[\left(\frac{P_r}{A}\right)^2 + \left(\frac{P_b}{A}\right)^2\right]\right\}$$

This equation shows the nonlinear transfer characteristic between (R-Y) and $\left(\frac{P_r}{A}\right)$. Colors with relatively high values of $\left(\frac{P_r}{A}\right)$ (red and magenta) generate significantly higher color difference output than negative (R-Y) signals (green and cyan) that have low values of $\left(\frac{P_r}{A}\right)$. For example, using the constant luminance ratios, a unity-amplitude red signal has $A = 0.3$, $P_r = 0.3$ and $P_b = 0$ resulting in an (R-Y) output of 1.45. For a normalized cyan signal $A = 0.7$, $P_r = 0$ and $P_b = 0.1$ the (R-Y) output in this case is -0.834, showing the asymmetry of the system. The principal effect of this nonlinearity is to produce somewhat overly saturated reds and blues and somewhat desaturated greens. A simple gamma corrector on the color difference signals will correct this problem without hurting the color balance. Alternatively the entire signal, $i = I\gamma$, can be gamma corrected.

The color balance can also be adversely affected by nonlinearity in the envelope detector. The threshold effect common to envelope detection produces a reduced gain at low input levels. This results in reduced red and blue output in low-light areas, with a resulting shift to the green in these darker regions. The solution to this problem is to linearize the envelope detector. Two linearizing methods have been used successfully. First, one can ensure that the signals are large compared to the diode threshold. A second method uses a feedback type of envelope detector where the effective signal level is increased in the crossover region of the detector.

PERFORMANCE CHARACTERISTICS

Assuming idealized color-encoding filters, the spectral transmission averaged over the filter area is 0.5 in the red and blue portions of the spectrum and 1.0 in the green (see Figure 5). Together with the color characteristic of the vidicon these ratios produce a luminance signal having essentially the same characteristics as a constant luminance source.

The signal-to-noise ratios of the envelope-detected signals are comparable to that of the luminance signal. Although the peak value of the carriers is 10-20 percent of the full dynamic range of the camera, the color bandwidth is only 10-20 percent of the luminance bandwidth, thus providing comparable noise conditions. It is essential, however, that a number of steps be taken to ensure adequate noise performance. The camera preamplifier should be a low-noise FET preceded by a tuned filter to boost the response in the vicinity of the color carriers. Care should be taken in both light and electron optics to ensure that the grating structures are well resolved over the entire field. A vidicon that proved suitable for this task was the RCA 8507.

A highlight overload causes clipping of the color carriers. The resulting reduction in the reproduced red and blue signals is interpreted as green. To avoid this condition, an automatic target control system can be used that detects the peak video signal and thus prevents overload. Since a system of this type is poor for live-camera use, where a single specular reflection can darken an entire scene, automatic target systems that work on the average signal are normally used. To allow an overload to take place without color distortion, a circuit was used that disables the color-difference signals when an overload occurs, resulting in a clipped white highlight rather than a green one.

EXPERIMENTAL RESULTS

The luminance and color-difference signals were fed into a conventional color receiver used as a monitor. The reproduced picture was of home-entertainment quality, being limited only by the receiver bandwidth. The recently announced low-cost RCA color camera series, including the PK701, PK730, and the PFS710 film series, are based on the techniques described in this paper. These cameras were demonstrated at the CATV Conference in San Francisco in 1969.

CONCLUSIONS

A color television camera can be constructed using a single camera tube, thus eliminating registration problems and reducing costs. By making efficient use of the two-dimensional spatial frequency spectrum of the camera, the resolution requirements are little more than that of a black and white camera. Although the initial applications are in the area of industrial and educational areas, the basic concept should find its way into studio broadcast use.

List of Illustrations

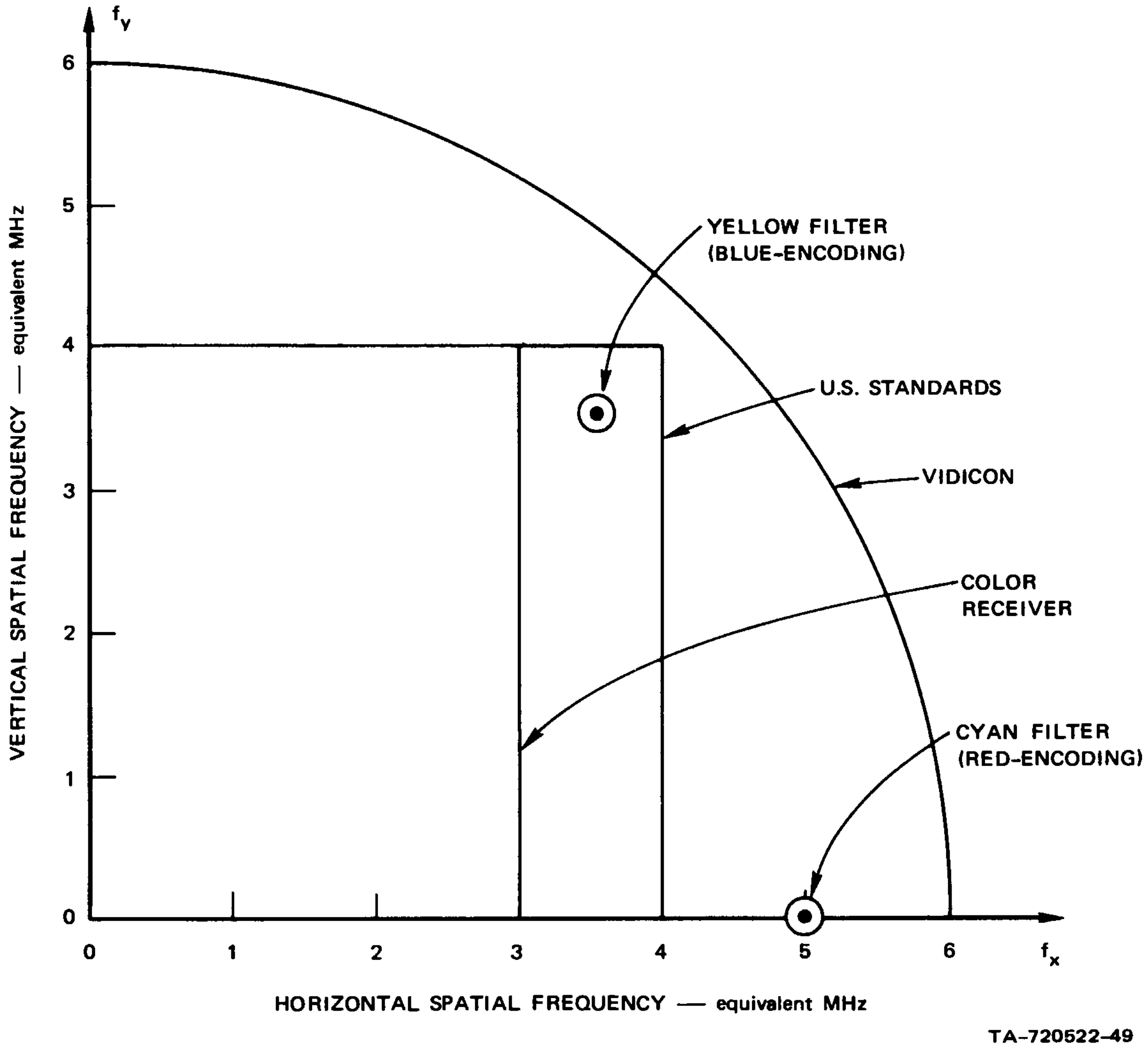


FIGURE 1 SPATIAL-FREQUENCY DIAGRAM

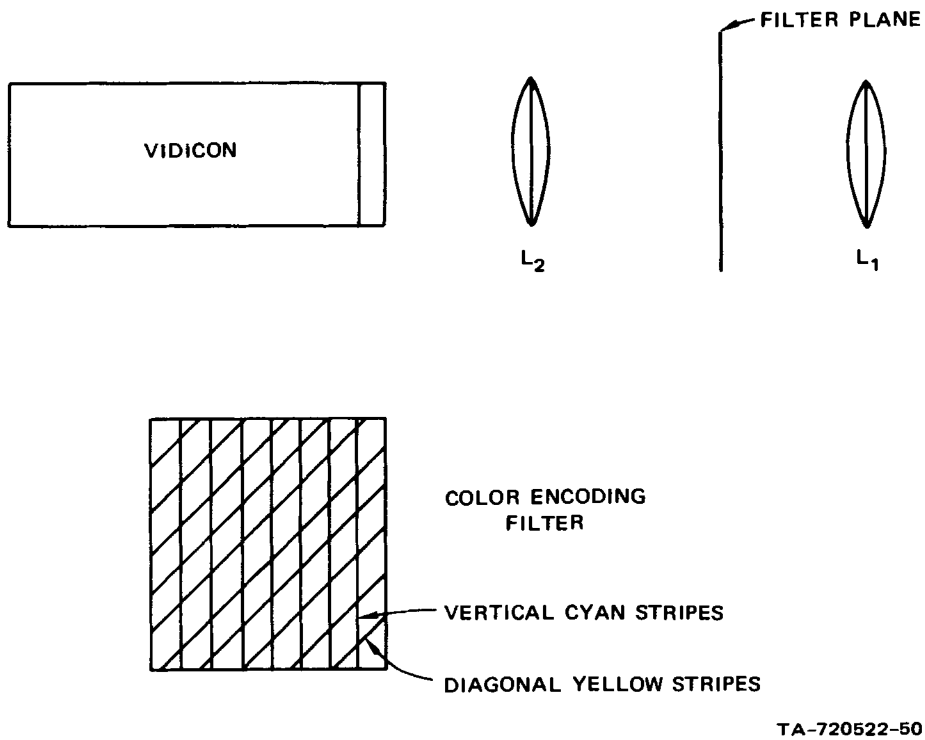
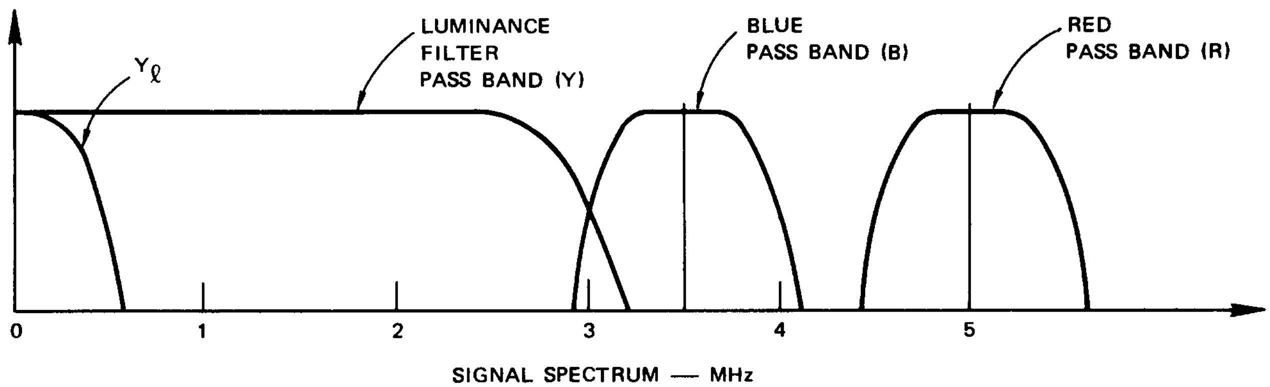


FIGURE 2 SINGLE-VIDICON RELAY-LENS TV CAMERA



TA-720522-51

FIGURE 3 VIDICON SIGNAL SPECTRUM

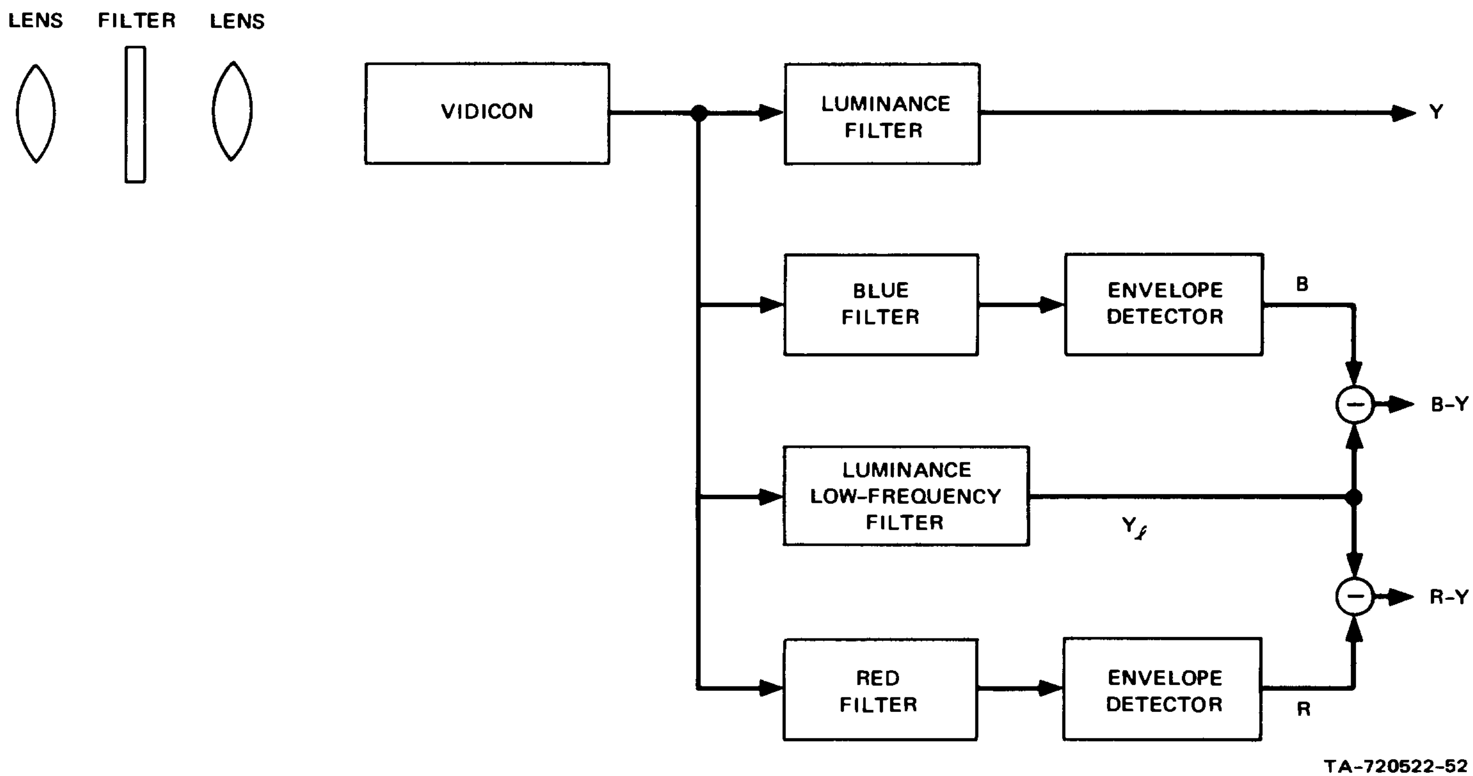


FIGURE 4 SYSTEM BLOCK DIAGRAM

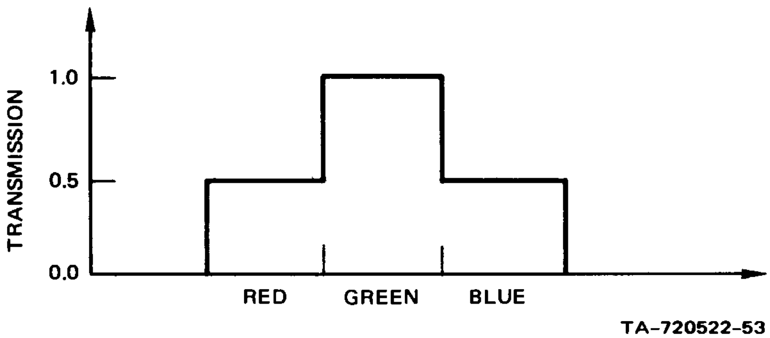


FIGURE 5 AVERAGE SPECTRAL TRANSMISSION OF COLOR-ENCODING FILTER

DISCUSSION:

Mr. Earle Jones: Thank you.

Mr. Greg Liptak: Thank you, Earle. One quick question that I might have is how does this perform under low light level conditions?

Mr. Jones: I'd have to answer--I would suggest that you talk to RCA, but I'll tell you what I think. The color television process inherently costs you something like 40% less than 1 F stop, that is if you can gather all the light in the optical system. They're evidently doing a pretty good job because I asked them down there yesterday and they said it would cost you about 1 F stop. So they use an 8507 Vidicon. The concept I heard here this morning where they take the color filter out and use it at low light level for black and white is new to me. I never heard of that before, but there is certainly no reason why you can't do it. As a matter of fact, someone here said that in a matter of an hour, you should be able to go back to color. Since there is absolutely no synchronization what-so-ever, you can move that filter any way you want to and get the same picture as long as you don't rotate it too much. You can rotate it slightly and there's no synchronous detection going on at all. I would say more like 30 seconds to go from back and white back to color.

Mr. Liptak: Very good--very pleased to hear you say that we're going to have a $2,000 to $3,000 color camera pretty quick.

Mr. Jones: Well, I didn't say that quick.

NATIONAL CABLE TELEVISION ASSOCIATION

ANNUAL CONVENTION

"TECHNICAL EYE-OPENER"

"CABLE AMPLIFIER POWERING METHODS"

June 10, 1970
8:00 A.M.
Munroe Room, Palmer House Hotel,
Chicago, Illinois

I. Switzer, P. Eng.,
Chief Engineer,
Maclean-Hunter Cable TV Limited
27 Fasken Drive,
Rexdale, Ontario, Ontario.

Reprinted with permission of:
National Cable Television Institute
3022 N. W. Expressway
Oklahoma City, Oklahoma 73112

CATV ADVANCED TECHNICIAN

CABLE POWERING

NATIONAL CABLE TELEVISION INSTITUTE

Table of Contents

CABLE POWERING

CABLE POWERING

INTRODUCTION

NOTES

As television signals travel on a cable, they are reduced in power because of the attenuation characteristics of the cable. A channel 13 signal will lose 1/10 of its power in every 100 feet of typical trunk cable (3/4-inch foam-type). At the end of 1,000 feet of cable, the channel 13 signal will have only 10% of its original power, and at the end of 2,000 feet, the signal will have retained only 1% of its original power. If the signal were not reamplified at suitable intervals, it would virtually disappear. In order to reamplify the signal, amplifiers are placed at regular intervals along the cable. These amplifiers periodically bring the signal back to its original power level and make up for the signal power losses in the cable.

A typical trunkline amplifier receives signal at its input at a +12 dBmv level and amplifies the signal to a +32 dBmv level. Considering the amount of RF (radio frequency) power involved, the amplifier has magnified the signal power 100 times. An output level of +32 dBmv represents an RF power of 21.3 microwatts, as computed in the following equation:

$$P = \frac{V^2}{R} \qquad \text{(from Ohm's Law)}$$

$V = +32$ dBmv
$= 40 \times 10^{-3}$ volts
$R = 75\Omega$

$$P = \frac{40 \times 10^{-3} \times 40 \times 10^{-3}}{75} = 21.3 \times 10^{-6} \text{ watts}$$
$$= \underline{21.3 \text{ microwatts}}$$

If 21.3 microwatts of RF power were produced on each of twelve channels, the total RF power output from a typical trunk amplifier would only be 12 x 21.3 microwatts = 256 microwatts — which is only 1/4 of a thousandth of a watt. More energy must be supplied by an amplifier from some kind of power supply. The function of the amplifier is to take some convenient form of electrical power, usually direct current or low frequency AC power, and convert it into RF power. The RF power output must, of course, be an amplified version of the RF power input.

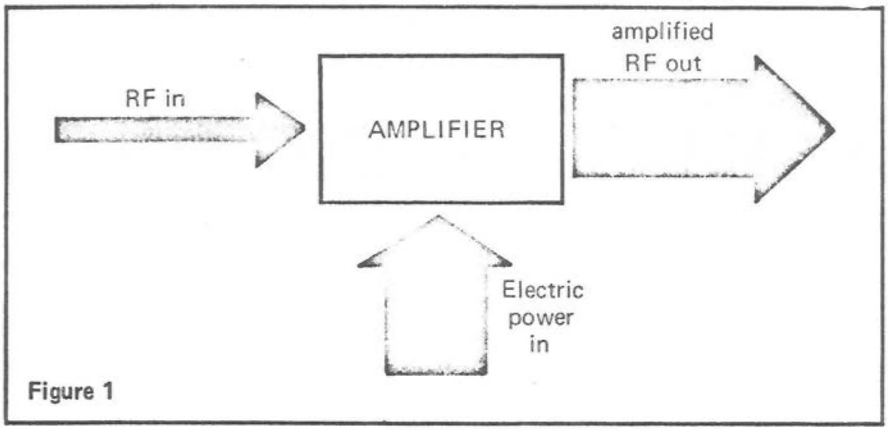


Figure 1

If the amplifier were 100% efficient, that is, if it converted all of the imput electrical power into RF output, a CATV amplifier could operate for many years on the power of a small flashlight battery. CATV amplifiers are, however, *very inefficient* in power conversion. The electrical power input is usually many times more than the useful RF power output.

The first amplifiers used in the CATV industry were constructed with vacuum tubes. A typical tube amplifier, comparable in performance to the present-day transistorized trunk amplifier, used twelve small tubes. Each tube required power of at least 1.10 watts (6.3 volts AC at 0.175 amp), just to heat the filament. Since CATV amplifiers operate in class A mode, they draw substantial plate and screen currents continuously. Therefore, the total power consumption of such an amplifier was approximately 60 watts. Tube

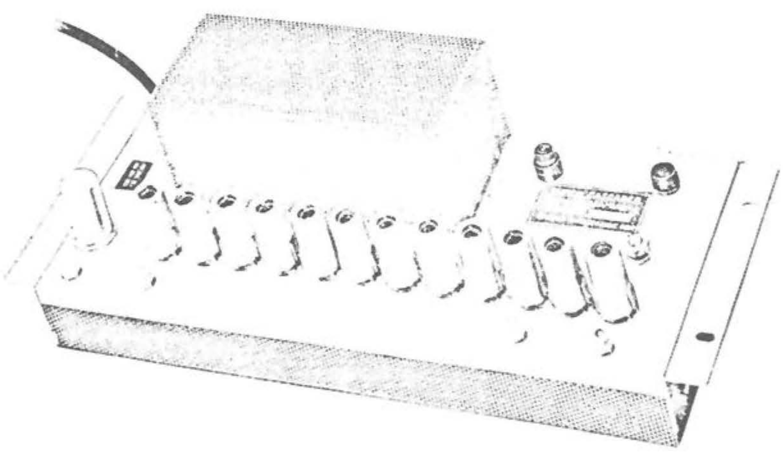


Figure 2 Typical Tube-Type Trunk Amplifier

amplifiers cannot be efficiently powered except by direct connection to local power utility services. Such amplifiers were installed in

NOTES

26

pole-mounted cabinets and were connected to the utility power system. A conventional 115-volt AC power source was required at each amplifier location.

The introduction of transistors into CATV amplifiers meant that the input electrical power requirement could be considerably reduced. Although transistors have no filament power requirement, their *plate current* (or the transistor equivalent of tube-plate current) requirement is still substantial. Transistors operated in linear, class A mode for CATV amplifiers use more current and lower voltages than tubes in similar applications.

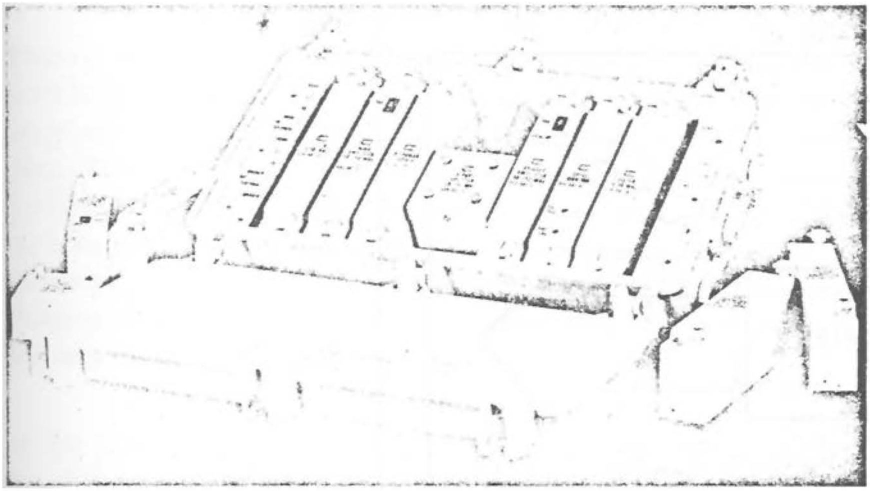


Figure 3 Transistorized Trunk Amplifier (Cover Open)

The amplifier section of a modern, transistorized CATV trunk amplifier might use 20 volts DC at 200 milliamps. This is a power consumption of only 20 volts x (200/1000 amp) = 4 watts. This electrical power requirement is substantially less than the 60 watts which a comparable tube-type amplifier required, but is still too much to be supplied by a conveniently sized battery. A set of fifteen dry cells, size D, will power such an amplifier for only seventeen hours. Obviously, the electric power requirements for the CATV amplifiers must still come from the utility power lines.

Although a CATV amplifier might require only 4 watts of power (20 volts at 200 milliamps) to operate the amplifier itself, the overall power requirement for the amplifier is usually significantly higher. The DC voltage for the transistors must be well-regulated; amplifier gain and distortion characteristics depend critically on holding the internal DC voltage to a constant value. DC regulation requires a complex voltage-regulator circuit, which consumes additional power, inside each amplifier. Many regulator circuits waste as much electrical power — dissipating it as heat — as the entire amplifier section uses. CATV amplifiers receive their electrical power as AC. This AC power must be rectified, filtered and then regulated before it can be used in the amplifier sections. The actual electrical power input for the CATV trunk amplifier is usually 15 watts (30 volts at 500 milliamps of AC power).

Since only 250 microwatts of RF energy is developed by an amplifier and fed into the output cable, virtually all of the 15 watts of electrical power used by the amplifier finally ends up as heat. This heat is carried away from the amplifier by the surrounding air and, to some extent, by heat conduction along the cables and messenger strand attached to the amplifier. Also, some heat is dissipated as radiant energy. Therefore, many amplifiers feel warm to the touch since the temperature inside the amplifier modules is high. Silicon transistors are used almost universally in CATV amplifiers because they are capable of withstanding high internal temperatures. However, the generated heat must still be carried away, or the transistors will burn out.

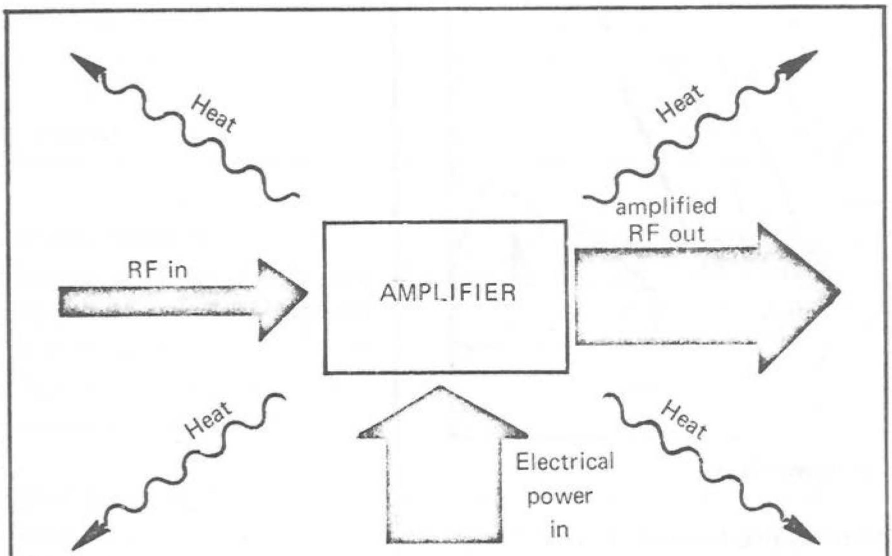


Figure 4

I. CABLE POWERING CATV AMPLIFIERS

CATV operators have long been attracted to the concept of utilizing the coaxial cable itself to carry the electrical power used by the amplifiers. The high power consumption of tube-type amplifiers made cable powering impractical except in small, line-extender amplifiers. However, because transistorized amplifiers require considerably less electrical power, cable powering is now used extensively in CATV.

A cable powering system feeding an amplifier can be considered to have the circuit shown in figure 5. This diagram of the circuit is a simplification of the actual characteristics of a cable power system but is suitable for studying the elementary characteristics of the

NOTES

28

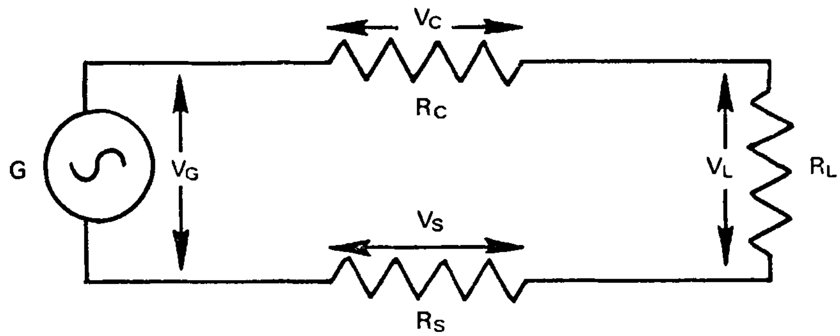


Figure 5

system. An AC power generator is shown since AC power is universally used for cable powering of CATV amplifiers. The use of DC power would have a great many advantages, but past experience with DC indicates that it causes serious corrosion problems, such as electrolysis effects at points where moisture gets into the cable system. The common use of dissimilar metals, such as copper and aluminum, in cables and connectors aggravates this corrosion problem. The use of AC power overcomes most of these problems, although electrolytic corrosion is still possible.

In the circuit diagram (figure 5) the AC power source (usually a step-down transformer from the 115-volt AC utility service) has a voltage V_G. The resistance R_C and the voltage drop V_C across the resistance R_C represent the resistance of the cable center conductor and the voltage drop in it. The resistance R_S and the voltage drop V_S across the resistance R_S represent the resistance and the voltage drop of the cable sheath or outer conductor. The resistance R_L and the voltage drop V_L across the resistance R_L represent the resistance and the voltage of the amplifier itself. This representation is a simplification since the power source G may have significant internal reactance; the sheath, or outer conductor, may be paralleled by a variety of low resistance ground paths which significantly reduce the effect of the sheath resistance; and the amplifier load is only approximately represented by an equivalent resistance (the amplifier is usually a non-resistive load, often having reactive components and not obeying Ohm's law with respect to current and voltage). The effects of these simplifications are not serious in an elementary consideration of cable powering and can be considered in detail in more advanced design studies.

Some typical values for R_C and R_S may be taken from cable specification sheets. Many specification sheets lump together the resistance of the center conductor and the resistance of the sheath, calling the combined value the *loop resistance* of the cable. In an elementary study of cable powering, it is convenient to combine these values so that the circuit uses the symbol R_C to stand for the loop resistance of the cable and retains R_L for the equivalent load resistance of the amplifier (figure 6).

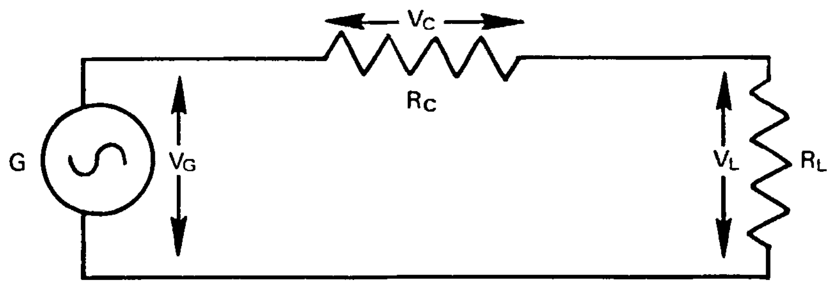


Figure 6

The following resistance values are typical of commonly used, aluminum-sheathed coaxial cables in CATV. These are values of DC or low-frequency AC resistance at a normal temperature (68° F.).

Nominal Resistance
(Ohms per 1000 feet)

Cable Type	Outer Conductor	Inner Conductor	Loop
0.412 inch	0.43 ohm	1.59 ohm	2.03 ohm
0.500 (1/2) inch	0.35 ohm	1.08 ohm	1.43 ohm
0.750 (3/4) inch	0.17 ohm	0.45 ohm	0.62 ohm

Remember, these resistance values are for a typical aluminum-sheath, copper-center-conductor, foam-dielectric cable. Center conductor resistance for copper-clad aluminum center conductors is somewhat higher, and consequently, loop resistance is also higher. Cables which have an air or foam dielectric with a higher air content than the commonly used foamed polyethylene dielectric generally have a center conductor with a greater diameter for the same size outer conductor, and consequently, have a lower center conductor resistance and a lower loop resistance.

Consider a situation in which a trunk amplifier is operating at the end of 2,000 feet of 3/4-inch cable. Assume that the amplifier requires 30 volts of 500 mA (milliamperes). Do not actually calculate the equivalent load resistance R_L representing the amplifier. The assumed amplifier specification sets the load voltage V_L at 30 volts and sets the current in the circuit at 500 mA, or 0.5 amp. The loop resistance R_C is 1.24 ohms, based on a loop resistance of 0.62 ohms per 1,000 feet, taken from the table. The flowing loop current — 500 mA, or 0.5 amp — is the current drawn by the amplifier. The voltage in R_C is derived from Ohm's law: $V = IR = 0.5 \times 1.24 = 0.62$ volt. Thus, the power supply voltage has to be 30.62 volts in order to overcome the voltage drop in the cable loop resistance and to maintain 30 volts across the amplifier load. The resulting circuit will look like figure 7 with the actual values inserted.

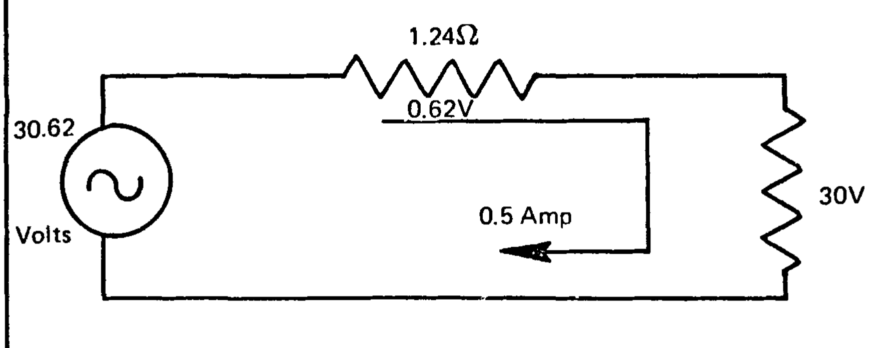


Figure 7

In actual practice the power source stays constant, and a drop in voltage at the amplifier occurs, due to the IR voltage drops in the cable loop resistance. To illustrate this more practical situation, figure 8 shows a redrawn circuit and a recalculation of the voltages.

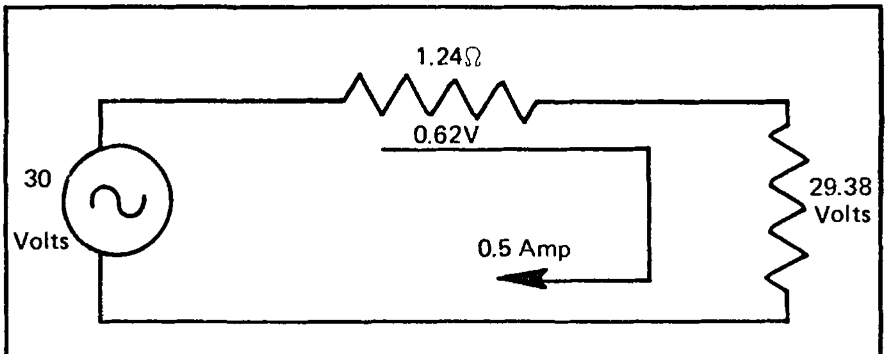


Figure 8

Most CATV amplifiers draw an almost constant current over their normal range of AC operating voltages. This is one of the less desirable characteristics of most CATV amplifiers and represents a departure from Ohm's law, which states that current flow is directly proportional to applied voltage (I = E/R). This "constant current" type of behavior is due to the operation of the *series regulator circuitry*, commonly employed by CATV amplifiers.

The amplifier works well with only 29.38 volts instead of 30 volts. In fact, it could continue to work properly with only 22 volts. This wide range of AC-voltage input is handled by the regulating circuitry inside the amplifier. However, there is a minimum AC voltage at which the amplifier will operate properly. More practical examples illustrate that this minimum operating voltage is a serious limiting factor in the design and operation of cable powering systems.

The preceding example (figure 8) represents the case of a CATV trunk amplifier powered remotely through 2,000 feet of 3/4-inch

NOTES

trunk cable. This example of cable powering represents a useful and valuable application of cable powering principles. Presumably, one amplifier could be powered at the power feed point, and another amplifier could be powered a short distance away in the opposite

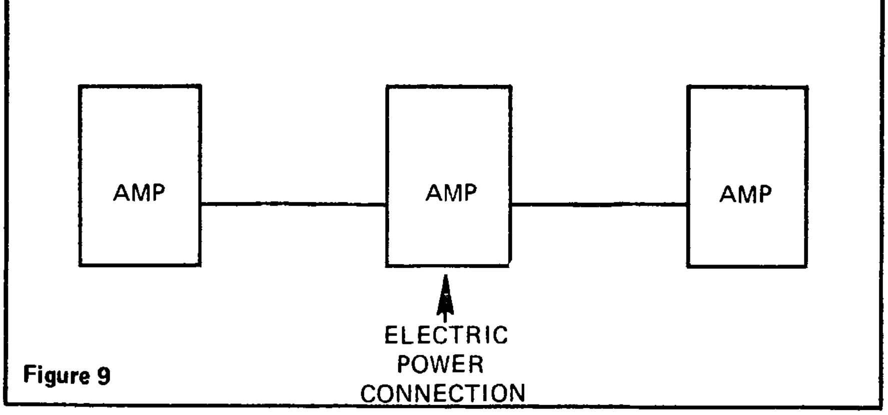


Figure 9

direction. Thus, three amplifiers could be powered from one utility power connection point, an advantage which would save the cost of additional circuit breakers, power meters, etc., and permit greater flexibility in system layout, since amplifiers can be located with very little regard for access to electrical utility power.

Consider a situation in which several amplifiers are powered from the same power feed point. Adding a second amplifier modifies the circuit diagram. Since most CATV amplifiers do not actually behave like resistors, their resistance symbol on the cable power circuit diagram is replaced with a triangular amplifier symbol, which represents the current consumed by an amplifier.

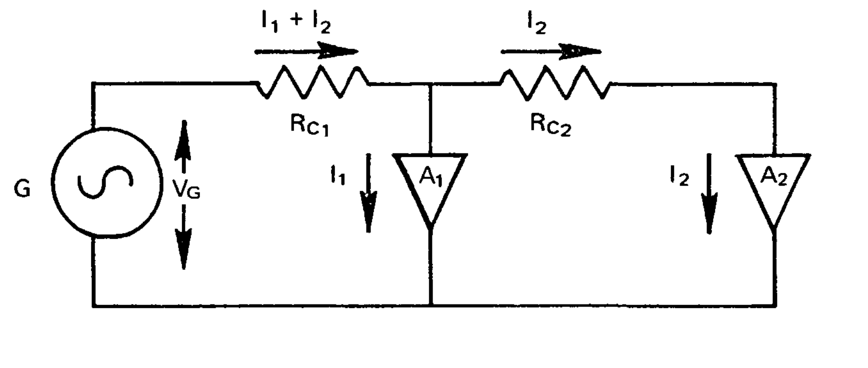


Figure 10

In figure 10, A_1 and A_2 are two amplifiers; I_1 is the current drawn by the first amplifier, and I_2 is the current drawn by the second amplifier; R_{C1} is the loop resistance of the first cable section, and R_{C2} is the loop resistance of the second cable section. Note that *the first cable section carries the current drawn by both amplifiers.* This is a very important feature of cable power systems. Cable sections closer to the power feed point carry more current than those sections further away. For a numerical example assume that both amplifiers

NOTES

32

draw similar currents, 0.5 amp at a nominal 30 volts. Assume that cable sections are 2,000 feet of 3/4-inch cable, as in the previous example. Now, fill in the resistances and currents, and calculate the voltage drops and voltage values at each amplifier, as shown in figure

NOTES

33

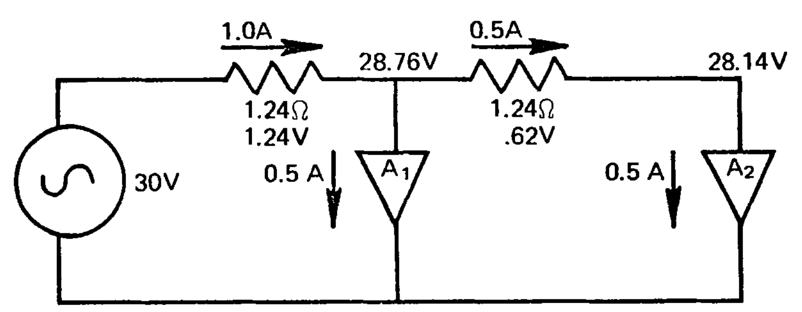


Figure 11

11. Note that the first cable section is carrying 1.0 amp, the current for both amplifiers. The second section carries only the current for the second amplifier, 0.5 amp. The voltage drop in each cable section has been calculated by Ohm's law (E = IR); thus, the voltage drop is the current times the resistance. With 30 volts available at the power feed point, the first amplifier has 28.76 volts, and the second has 28.14 volts. Add a third amplifier, using a similar diagram and calculation, and show the currents, voltage drops and amplifier voltages (see figure 12).

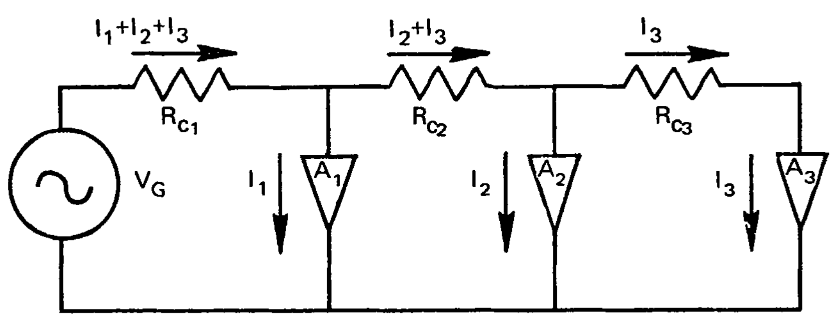


Figure 12 A

Assume:

$V_G = 30V$

$I_1 = I_2 = I_3 = 0.5A$

$R_{C_1} = R_{C_2} = R_{C_3} = 1.24\Omega$

Voltage drop in first cable section $R_{C_1} = (0.5 + 0.5 + 0.5) \times 1.24$

$= 1.5 \times 1.24 = \underline{1.86V}$

Voltage drop in second cable section $R_{C_2} = (0.5 + 0.5) \times 1.24$

$= 1.0 \times 1.24 = \underline{1.24V}$

Voltage drop in thira cable section $R_{C_3} = 0.5 \times 1.24 = \underline{0.62V}$

Voltage at first amplifier = 30.00 − 1.86 = 28.14V

Voltage at second amplifier = 28.14 − 1.24 = 26.90V

Voltage at third amplifier = 26.90 − 0.62 = 26.28V

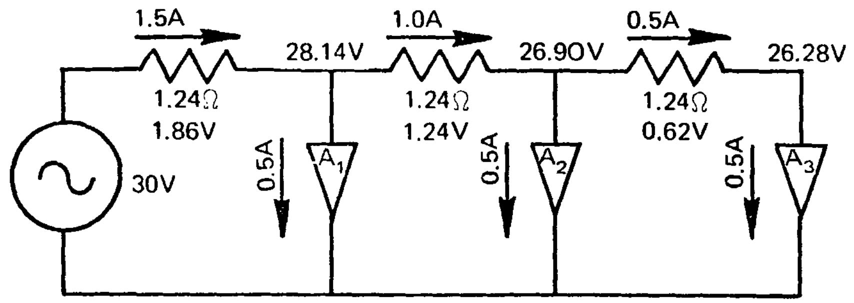
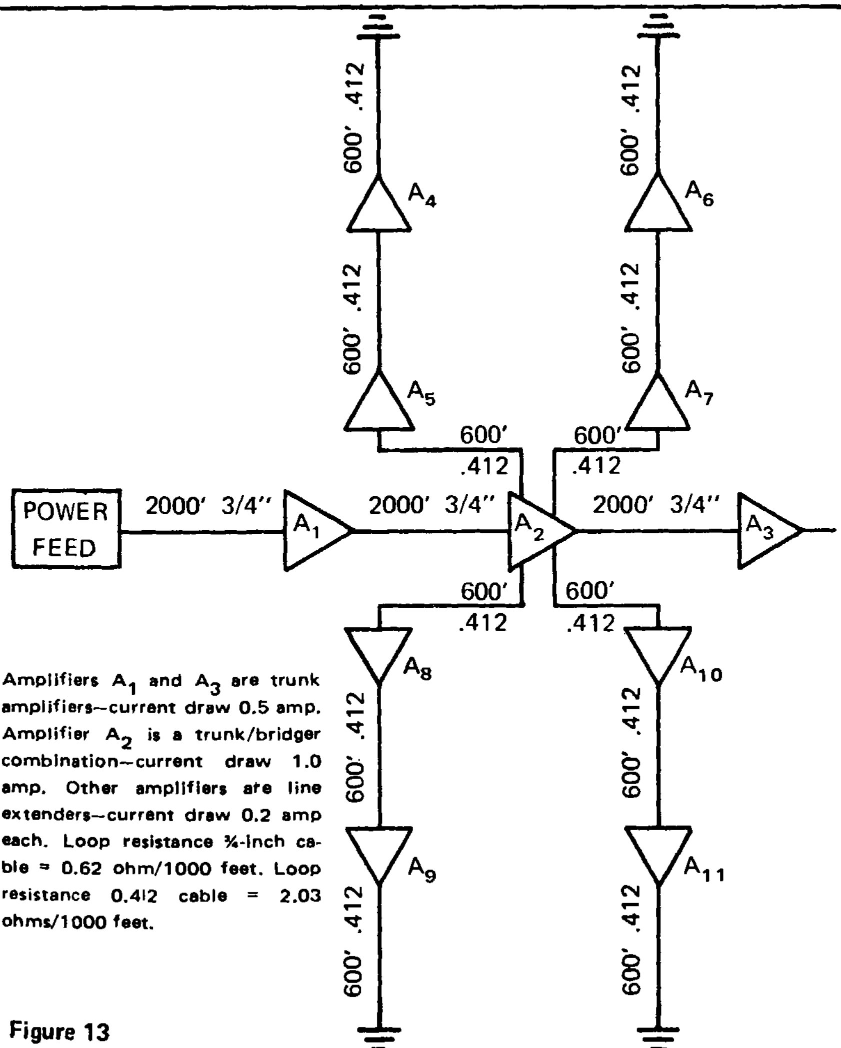


Figure 12 B

Now try a more complicated situation involving several types of amplifiers and several types of cables, including feeder lines with line extender amplifiers. To simplifiy the diagram, do not show the *return line* from each amplifier, but use a *ground symbol* to show a



Amplifiers A_1 and A_3 are trunk amplifiers—current draw 0.5 amp. Amplifier A_2 is a trunk/bridger combination—current draw 1.0 amp. Other amplifiers are line extenders—current draw 0.2 amp each. Loop resistance ¾-inch cable = 0.62 ohm/1000 feet. Loop resistance 0.412 cable = 2.03 ohms/1000 feet.

Figure 13

NOTES

common return. Incorporate a bridging amplifier which draws 1.0 amp, and line extender amplifiers which draw 0.2 amp and which are fed through 0.412 cable. The cable power voltages are not too difficult to calculate if care is taken to draw a circuit diagram for the system and to clearly mark the cable loop resistances, total currents and voltage drops as they are calculated. Cable lengths and amplifier currents have been assigned convenient "round numbers" for ease of calculation in figure 13.

The system layout can be redrawn as a schematic power diagram. Cable sections which do not carry current can be omitted, since they are of no concern in cable powering (figure 14). Note that the current drawn by each amplifier has been shown, and that the currents are then added in those cable sections which carry current drawn by all eleven amplifiers — a total of 3.6 amps. This section usually has the greatest voltage drop. Thus, it is a good practice to put power into the system at a point where the current can flow through larger cables; i.e., in 3/4-inch trunk cable in this example. The voltage distributions would have been vastly different if the power feed point had been at one of the line extenders.

NOTES

35

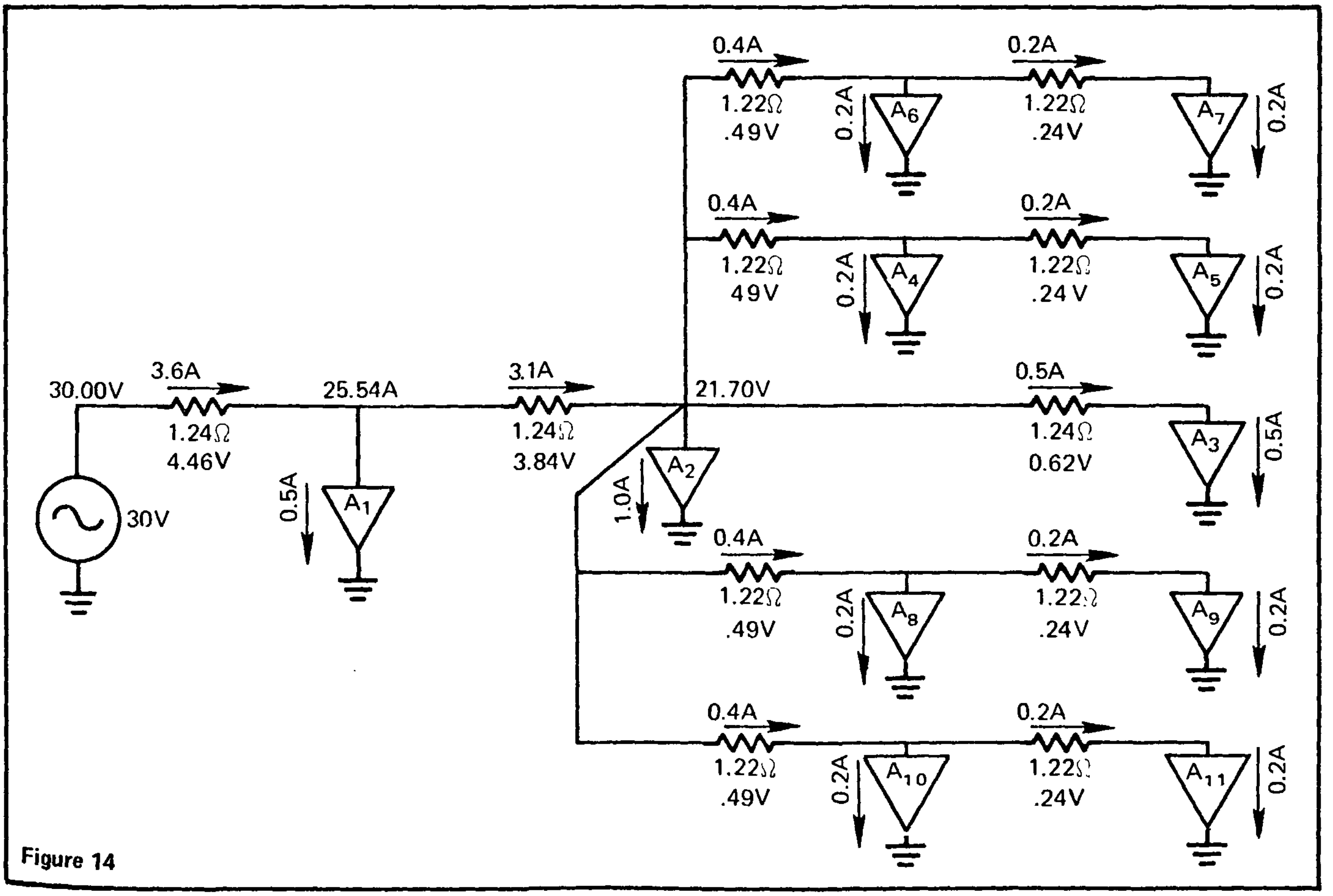


Figure 14

The technique for power calculation is simple and straightforward:

1. Prepare a detailed power flow schematic diagram showing all the amplifiers and cables which carry power.

2. Mark in the cable section loop resistances and amplifier current draws.

3. Calculate the current flow in each cable section, making sure to sum up the currents of all the amplifiers fed through the cable *for each cable section.* Start from the furthest amplifiers and work back toward the power feed point.

4. Calculate the voltage drop in each cable section, and mark it on the diagram.

5. Now, start at the power feed point and calculate the voltage at the end of each cable section by subtracting the voltage drop from the voltage at the input. Note that power often flows in a direction opposite to signal flow.

6. Check to see that each amplifier has an adequate operating voltage. Minimum operating voltage varies among manufacturers and amplifier types. An amplifier designed for nominal 30-volt operation often performs adequately on a minimum voltage of 22 volts.

If any amplifiers appear to be *starved* for voltage, some rearrangement is required. Low voltages will cause hum bars in the television picture and other operating problems. Thus, certain amplifiers may have to be transferred to another power feed point, taking care to avoid overloading power distribution facilities. Some power reserve must be allowed so that additional line extenders, active taps, amplified splitters, etc., can be accommodated at a future time without overloading the cable power system.

Proceeding through the calculations for the power flow schematic diagram (figure 14), you will discover that the system will not work. The cable power distribution system has been overloaded. There are only 21.7 volts available at amplifier A_2; and if the calculation had been continued past this point, there would have been even less power available at the inputs to the line extenders. The remedy is obvious. There is a voltage drop of 4.46 volts in the first cable section. The power is reduced to 25.5 volts and has not even reached the point of connection of the heavy power load consisting of the trunk/bridger and the eight line extenders. If the power feed point were moved to amplifier A_1, the voltage drop in the first cable section could be saved. The cable section between the power feed point and the first amplifier would then carry power in a reverse direction to a portion of the system extending in that direction (and nor shown on these diagrams). Thus, in all of these examples, it may be assumed that an additional system is being powered on the other side of the power feed point, and possibly, another amplifier and section, directly at the feed point. A single power feed point will often send power in two, three or sometimes four directions. The type of system shown in figure 14 starves itself for voltage before it starts to run very high currents. The current out of the power feed point is only 3.6 amps. Most power feed points are capable of supplying 10 to 12 amps of power at a nominal 30 volts. In order to use the capabilities of a given power feed point efficiently, it must be placed so that it can feed power in more than one direction.

NOTES

37

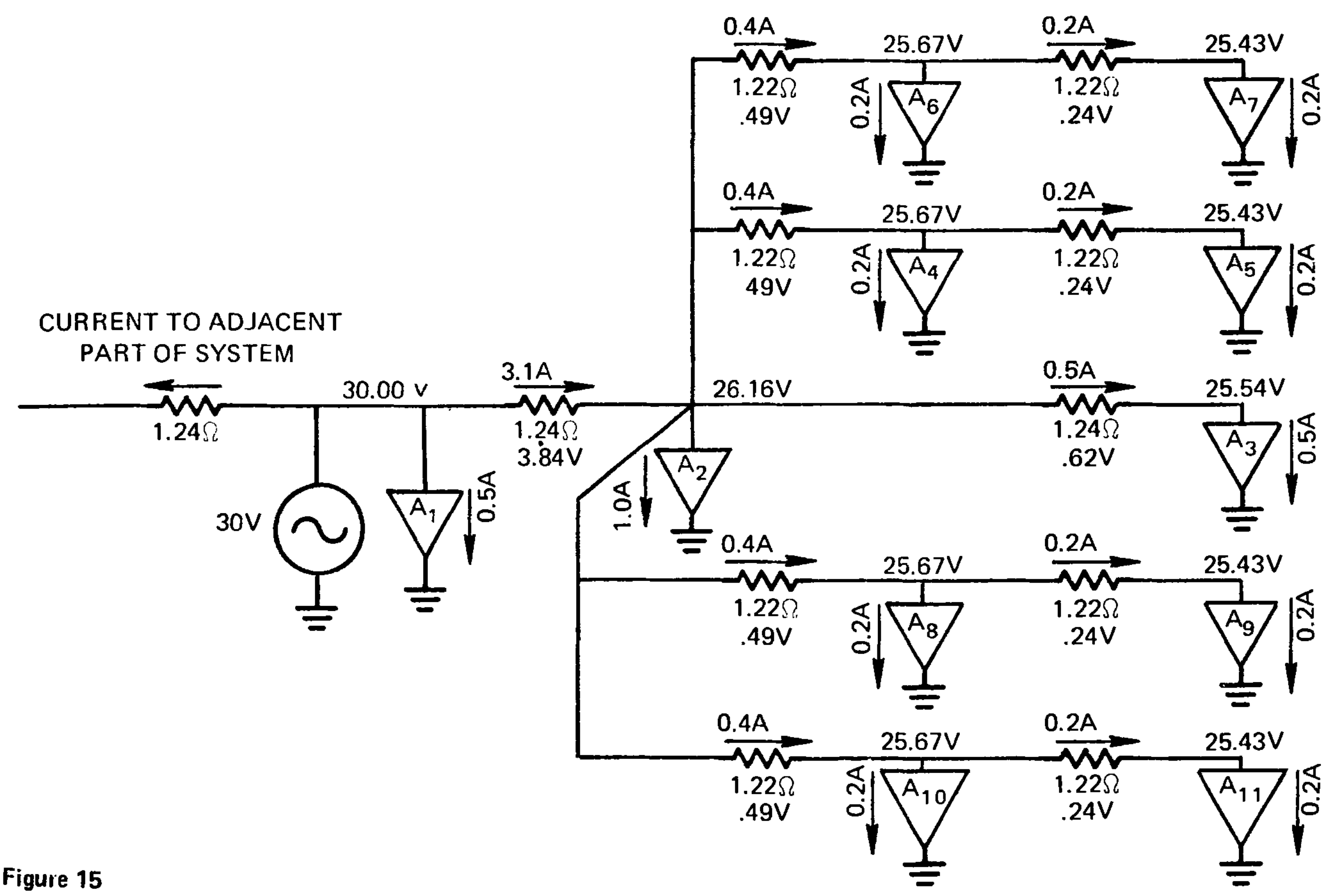


Figure 15

In figure 15 the power feed point has been relocated, as suggested by the results of calculation. The voltages of all amplifiers are now reasonable, and there is some reserve for additional power loading. Amplifier A_1 is now located directly at, or fairly near the power feed point. The A_1 current drain does not affect voltage drops in the cable power system but must be added to the total current being drawn from the power feed point and taken into account when considering the total load that can be placed on this power feed point.

Thus far, we have studied the flow of electric power in CATV systems without regard to the details involved in the mixing and separation of power and RF energy, and how amplifier power supply design affects cable powering.

II. DIPLEXING

38

Diplexing occurs when a combination of low-pass and high-pass filters form from suitable inductances and capacitances in a simple arrangement which permits mixing (or separation) of low-frequency AC power and high-frequency RF energy.

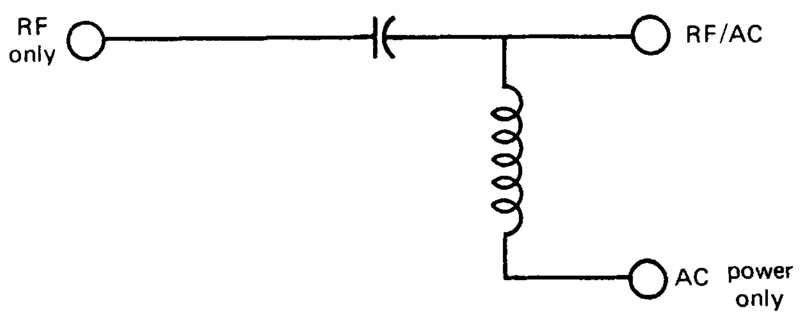


Figure 16

The capacitor is chosen (usually 0.01 μF) to present a low impedance to RF, but to present a high impedance to 60 Hz power. A 0.01 μF capacitor has a reactance of only 3.2 ohms at 5 MHz, the lowest frequency that is ever expected to be handled in the RF section of a CATV system. At 50 MHz, which is a more likely lower frequency limit for CATV RF frequencies, the reactance of a 0.01 μF capacitor is only 0.32 ohms. The 0.01 μF capacitor has a reactance of approximately 270,000 ohms at 60 Hz power line frequency. An inductance of 1 millihenry has a reactance of approximately 30,000 ohms at 5MHz, but only 0.37 ohms at power line frequency (60Hz). The reactance of this inductance rises to 300,000 ohms at the more common RF frequency of 50 MHz.

The diplexing arrangement (figure 17) can now be shown with typical values for the capacitor and inductor, with the reactance at 50 MHz and 60 Hz. The mixing and separation of RF and power can be more easily understood by following the behavior of the diplexing network at power and RF frequencies. Power will obviously follow the path of least resistance from the PWR to the RF/PWR terminals through the 0.37-ohm reactance of the inductor (RF choke). Power is kept from the RF terminal by the 270,000-ohm reactance of the capacitor. RF also follows the path of least resistance, passing easily through the 0.32 ohm of the capacitor, but is blocked from the power terminal by the 300,000 ohms represented by the inductance. Thus, the diplexer effectively separates or combines RF and power.

Actual diplexer circuitry may differ slightly from the basic arrangement discussed here. The capacitor can be omitted, allowing RF and power to appear on both the *through ports*. Two inductors may be used with small capacitors to ground, so that no RF can possibly

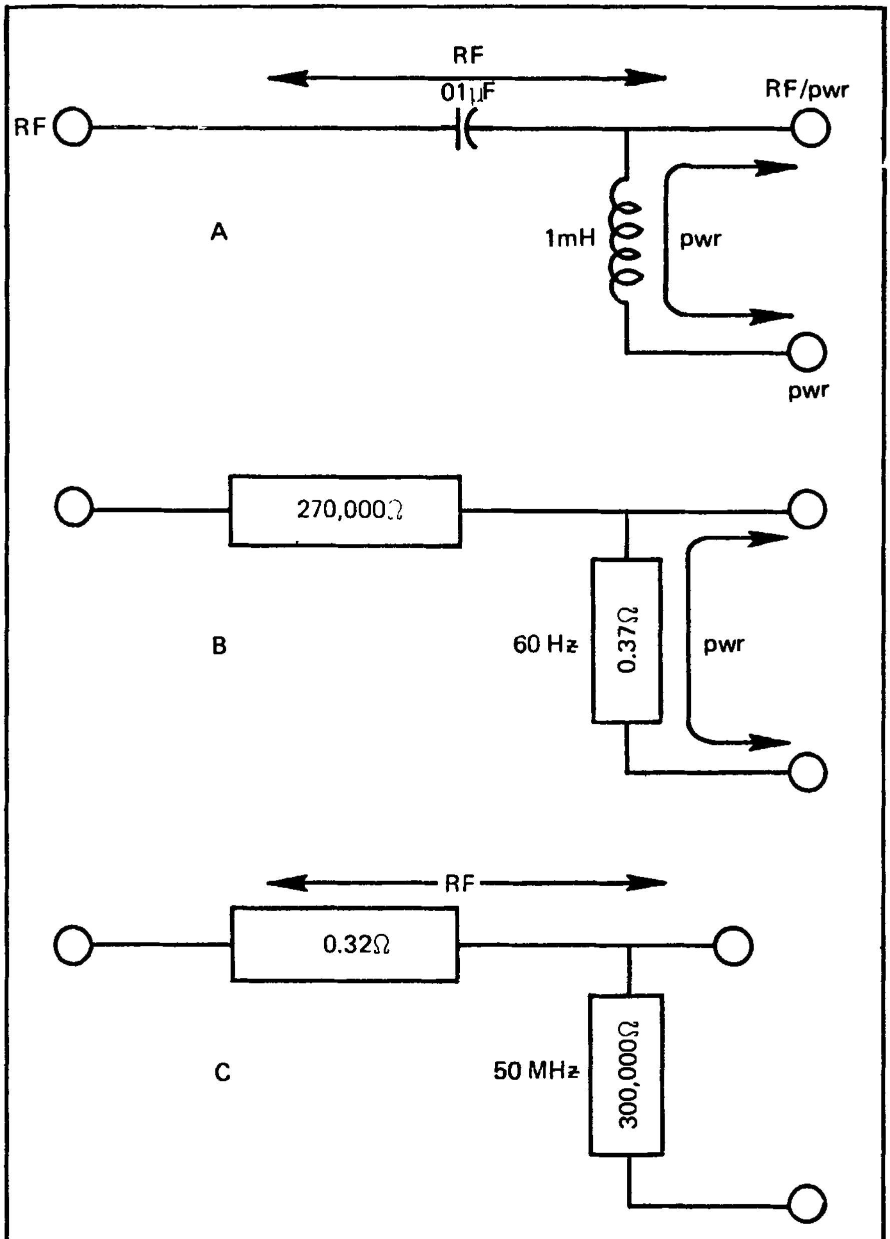


Figure 17

reach the power terminal. Such a combination may be used to permit separation of power from input and output connectors.

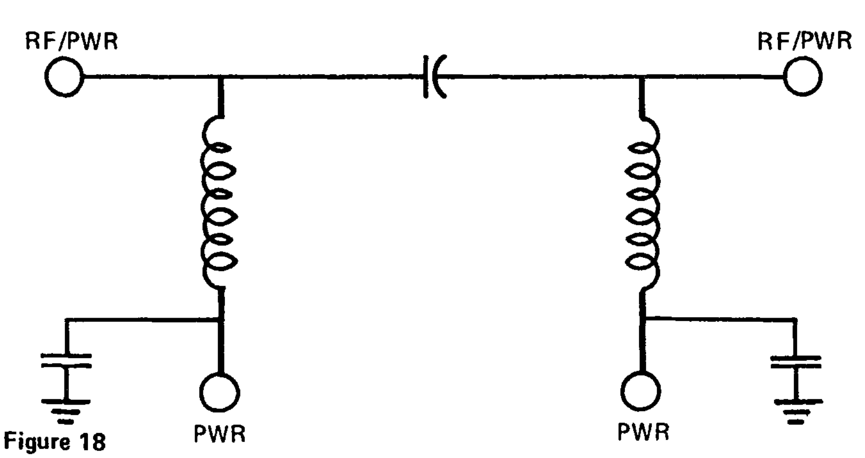


Figure 18

This arrangement permits greater flexibility in powering and is often used to mix power into a cable system. The two PWR connections may be joined together or fed separately.

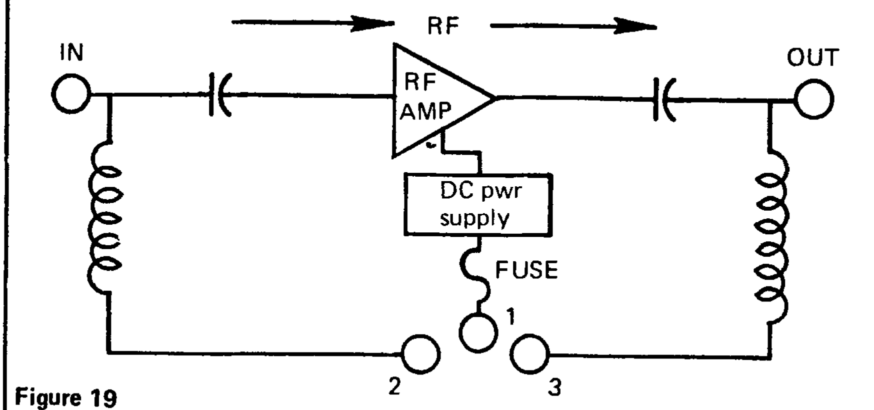


Figure 19

The separation of power and RF at an amplifier is usually accomplished as shown in figure 19. The capacitors (usually 0.01μF) keep power from entering the RF amplifier circuitry. Power is directed through the inductors to terminals 2 and 3. A system of links or jumpers is usually provided so that the flow of power into and through the amplifier can be controlled. If terminals 1, 2 and 3 are connected, the power will flow through the amplifier housing in either direction and will energize the internal power supply through the fuse. If only terminals 1 and 2 are connected, power will flow from the input connector to the internal power supply. If only terminals 1 and 3 are connected, power will flow from the output connector to the internal power supply. Some amplifiers provide an additional connector which can be linked to terminals 1, 2 and 3 to permit power to be added from an external source at the amplifier location. The amplifier then serves to mix power into the cable system. Bypass capacitors are frequently added in the "power only" sections of the circuitry to effectively short out any RF that may have found its way through the inductors.

III. AMPLIFIER POWER SUPPLIES

The purpose of cable power is to provide the energy for the RF amplifier. The amplifier power supply must change the AC cable power to DC of the proper polarity and provide adequate filtering and regulation. Some amplifiers require DC voltages of both polarities. Most require voltages of only one polarity, either positive or negative with respect to ground, depending on the details of circuit design and the types of transistors used. The DC supply for the amplifier must be well-regulated, since amplifier gain will change if the DC supply voltage changes. The DC supply must also be well-filtered; thus, for practical purposes, the DC is pure. A typical well-filtered CATV amplifier power supply will have less than 5 millivolts (RMS) ripple in a 20-volt DC supply.

CATV power supply regulation is considered as a subtopic of amplifier design detail. Various types of regulators are used, but they all require a basic source of unregulated, or sometimes partially regulated, DC which comes from the rectifier and first filter section. Both the rectifier and filter sections affect the operation of the cable power system.

Figure 20 Block Diagram of the Power Supply

Early CATV amplifiers used a *half-wave* rectifier system, feeding a *capacitor input* filter section. This is the simplest kind of rectifier-filter system. The rectifier diodes are silicon, with suitable current and voltage handling capabilities. The filter capacitor is a large electrolytic type. In a 30-volt AC system, this capacitor might typically be a 250 μF, 50-volt type. Note that the AC input and the

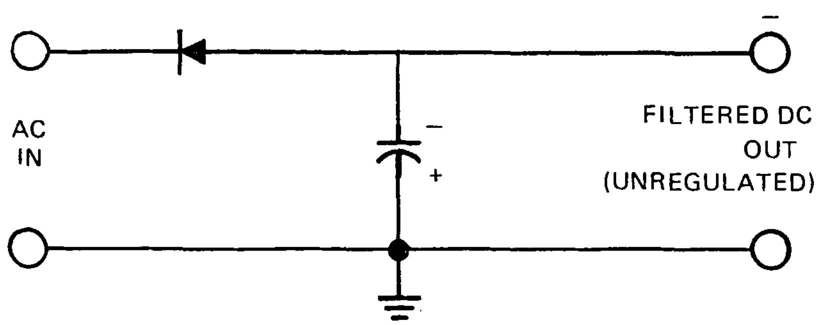


Figure 21

DC input have a common ground. The coaxial cable network is *single ended* with respect to ground; i.e., one of the circuit conductors is always at ground potential, and the amplifier itself has RF input and output grounded on one side. The normal power supply, therefore, requires that AC input and DC output have a common ground.

The action of a simple, half-wave, capacitor-input power supply can be best understood by reference to actual oscilloscope displays of current and voltage waveforms in such a system.

The student may wish to experiment with these waveforms. The photographs were taken with the aid of a double-beam oscilloscope, capable of displaying two waveforms simultaneously on a common time base. A small, low-cost *electronic switch* (e.g. Eico model 488) permits a similar display on a single-beam oscilloscope. If none of these aids are available, useful information can be obtained with only a single-beam oscilloscope of moderate capability. The arrangement shown in figure 22 was used to observe voltage and current waveforms.

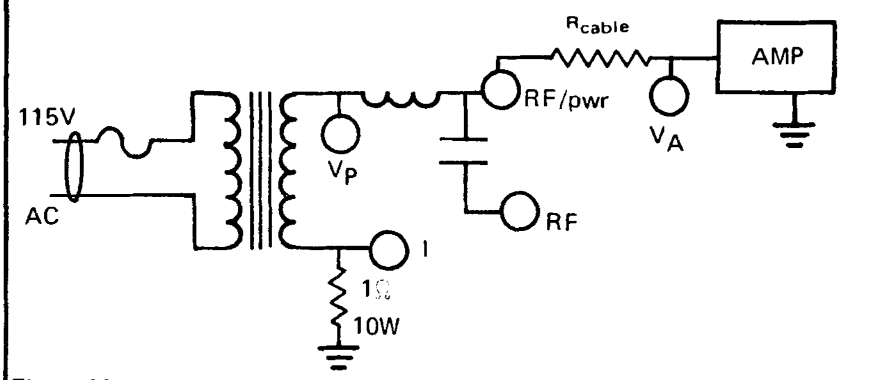


Figure 22

A conventional cable power transformer has been modified to permit observation of current waveforms. The *low* end of the transformer secondary is connected to ground through a 1-ohm resistor. The voltage appearing across this resistor represents the cable power current. The voltage from the test point shown can be displayed on an oscilloscope as representing the cable power current. If a calibrated oscilloscope is used, the peak current can be determined, since the voltage is developed across a known resistance — 1 ohm in this case. A resistance is inserted in series with the main power feed to the amplifier being examined. Since these demonstrations are usually made without the presence of RF, this resistor can be a wire-wound power type. A value of 10 or 20 ohms is convenient. Although this value is greater than cable resistances in actual systems, it permits the effects of cable resistance to be more easily demonstrated by exaggeration. The two voltage test points, Vp and V_A, permit observation of voltage at both the power supply and the amplifier. Do not try to observe current waveforms across the resistance representing the cable. Both ends of this resistance are at a voltage above ground. Connecting this resistor to an oscilloscope will short out the cable power supply and blow the fuses. Using a current sensing resistor, as shown in the secondary ground return, permits observation of current at a point which has one side grounded. Current can also be observed by using a clip-on type of ammeter or current transformer in the power lead, but this type of current observation is difficult to calibrate.

The scope photograph (figure 23) shows the voltage from a cable power feed transformer, rated at a nominal 30 volts AC. Vertical calibration is 20 volts per division, and the voltage wave is seen to have negative and positive peaks of 40 volts with respect to ground. If the AC voltage were precisely 30 volts RMS (root mean square), we would expect peaks which are 1.414 times the RMS voltage, or 42.4 volts in this case.

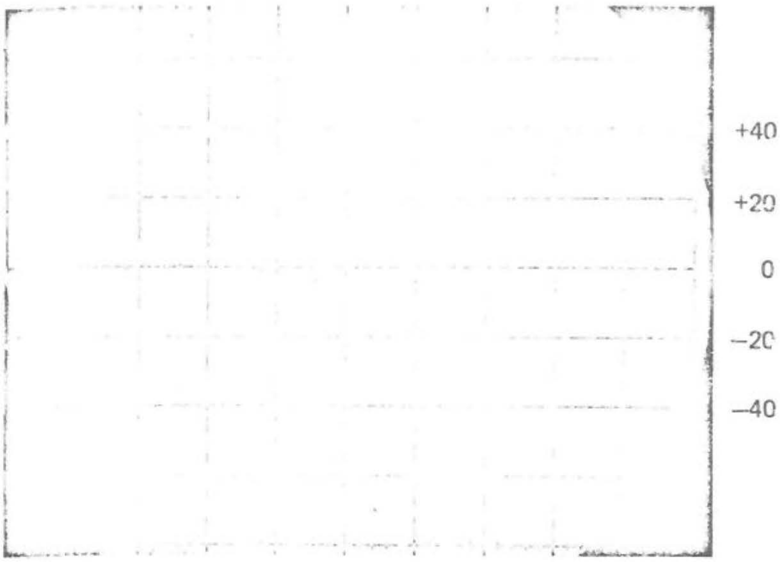


Figure 23

It is the nature of capacitor input filter systems that current flows into the filter capacitor in short pulses. AC voltage is always present, but current flows in the cable for only short periods of time, during which the instantaneous voltage exceeds the capacitor voltage. For the remaining time no current, and hence, no energy, flows in the cable. In the case of half-wave powering, only half of each cycle is used to power the amplifier. In the example shown (figure 24) only the negative half cycle of AC voltage is used by the power supply. Current and waveform are shown in this photograph. The power

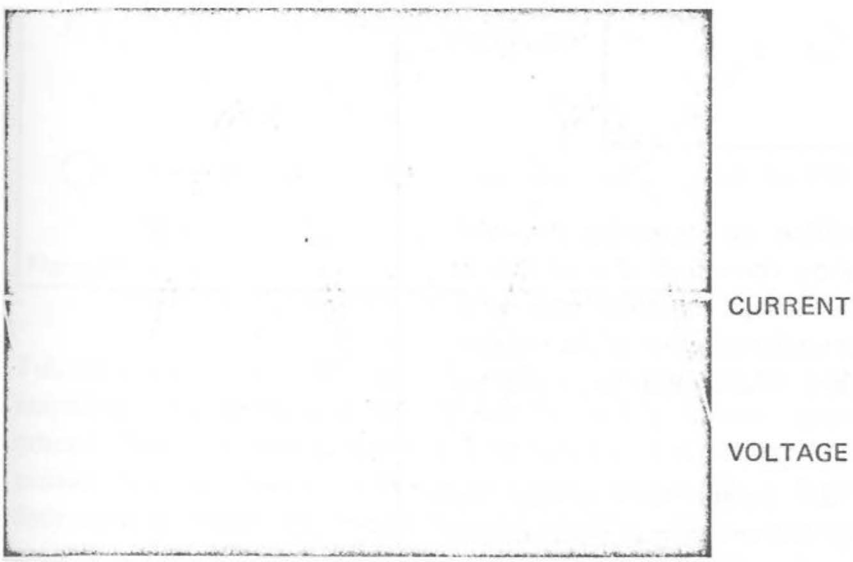
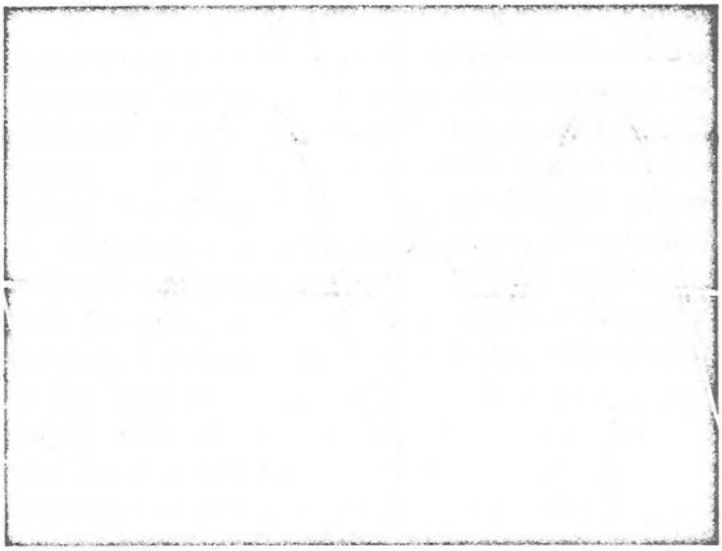


Figure 24

supply is a negative, half-wave, capacitor input supply in a small, cable-powered *active tap*. Note that current flows only during the negative half cycle, and also that the current waveform is a narrow *pulse*. Since voltage drop in the cable can occur only when current flows, the negative half cycle shows some observed distortion at the amplifier because of voltage drop in the cable during the current "pulse". The positive half cycle is not affected, since no current flows during the positive half cycle. The current flow shows positive polarity because of the position of the current-monitoring resistor in the circuit.

Measurement of the current waveform, in this case, shows that the current flows for only 22% of the available time. Since all the energy that the amplifier requires must be conveyed in only 22% of the available time, the current peak is high. Measurement of this display shows that the current reaches a peak value of 1.0 amp. Yet, the average current for this particular amplifier is approximately 0.15 amp. The actual instantaneous voltage drop in a cable is the instantaneous product of voltage and current. Voltage drop is proportional to the actual instantaneous value of the current and only occurs when current flows. In half-wave power systems, one-half of each cycle does no work at all and passes directly through the entire system without suffering any voltage drop.



CURRENT

VOLTAGE

Figure 25

The half-wave demonstration is modified by increasing the cable resistance substantially in order to show the effect of a number of half-wave amplifiers drawing current through practical cable resistances. The photograph (figure 25) shows the *clipping* of the negative half cycle by the half-wave current flow. The unused half cycle has not been affected.

In this particular demonstration, the power feed system has benefited from the clipping. The peak current flow is reduced, and the current flows for a longer period of time (almost double the time of the previous example). Since the current pulse is spread over a longer period of time, it does not have to reach the higher peak values of the narrower current pulse.

Voltage waveforms in a cable power system tend to be clipped, as shown in this demonstration When the peak voltage value is reduced to a point below which it will no longer keep the filter capacitor charged to its required value, the cable power is no longer usable.

Amplifiers, at this point in the system, will start to show excessive hum in the television picture, and the regulators in the power supplies will not function properly.

An ordinary AC voltmeter, used to check cable power in such a case, will not indicate the true situation. Most AC voltmeters are rectifier-type instruments which indicate average voltage. The average voltage in the clipped example is still high because of the presence of the unused, positive half cycles. If the voltmeter is a half-wave type, it might be reading either the positive or negative half cycle, depending on the way the leads are connected to the circuit. In such a case different readings are made; neither of which is correct.

IV. FULL-WAVE POWER

It is obviously desirable to use both halves of the AC power cycle. If energy can be transferred during both half cycles, the individual current peaks will be smaller, and the voltage drop along the cable will be less serious.

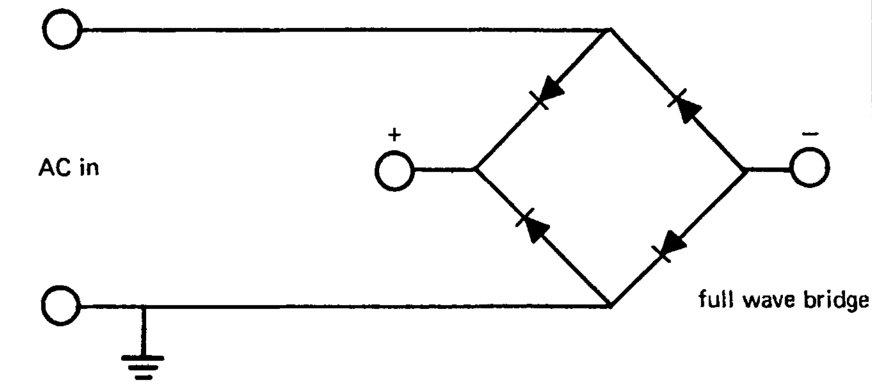


Figure 26

Full-wave power supplies are difficult to implement in CATV amplifiers. The problem is that AC and DC usually have a common ground. The full-wave bridge is a convenient and efficient way to provide full-wave rectification. This configuration in figure 26 uses four rectifier diodes, but the DC output cannot be tied to the same ground as the AC input. The filter and regulator circuitry must *float* with respect to ground. Some CATV amplifiers do just that; i.e., they float their filter and regulator and arrange the RF circuitry so that signal and power circuitry can float with respect to each other.

Some CATV amplifier designers avoid the problem by using two separate half-wave supplies of opposite polarity and by designing the amplifier to use power of both polarities. By equalizing the positive and negative power requirements, the current flow in the cable is kept balanced. This approach is expensive, since it requires two complete power supplies — one for each polarity.

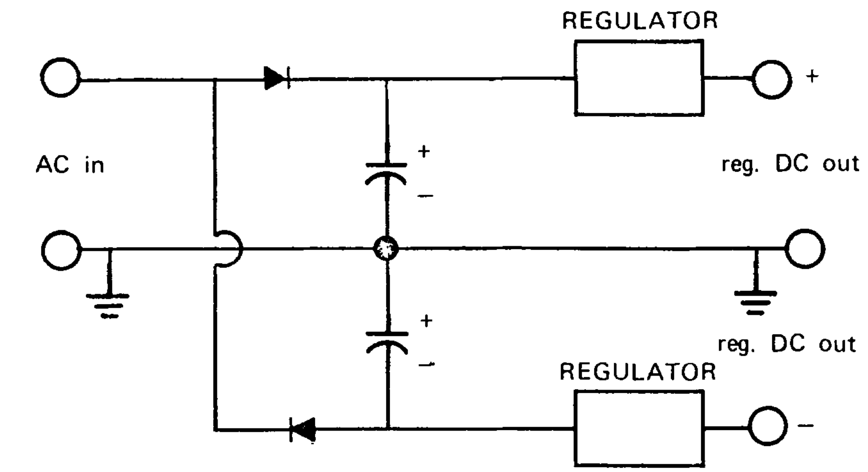


Figure 27

The most common approach is to use a power transformer, which permits use of either center-tapped, full-wave rectifiers or full-wave, bridge-type rectifiers. The transformer provides the isolation required to allow the DC and AC to have a common ground. Power transformers often have primary taps to permit adjustment for the position of the amplifier in the line. Amplifiers further away from the power feed use lower voltage taps.

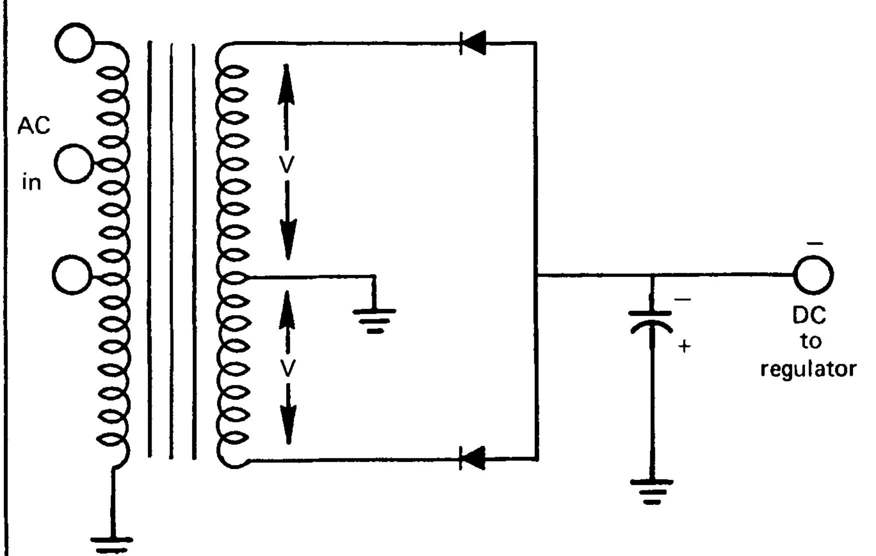


Figure 28 Full-Wave, Center Tap Transformer—Negative Polarity

The full-wave bridge is more efficient because the transformer secondary is working on each half cycle. In the center tap type of operation, only half of the secondary winding carries current at a time. Diode and capacitor polarities can be reversed if positive output polarity is required.

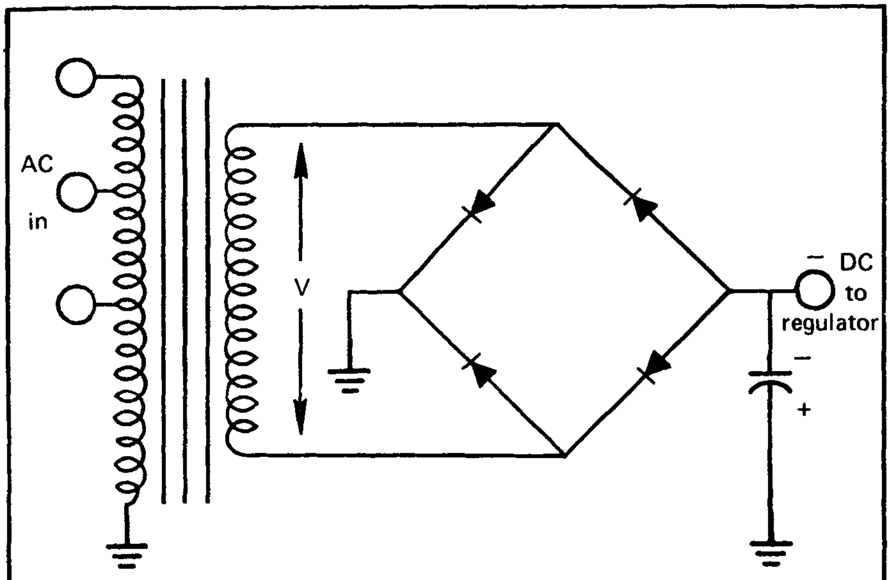


Figure 29 Transformer–Full-Wave Bridge–Negative Polarity

Voltage double circuitry can be used. This is desirable because it does not need a transformer, but output voltage is rather high because of the doubler effect. Voltage double circuitry presents no problem when *switching mode regulators* are used, since these regulators efficiently reduce high DC voltages to regulated lower voltages.

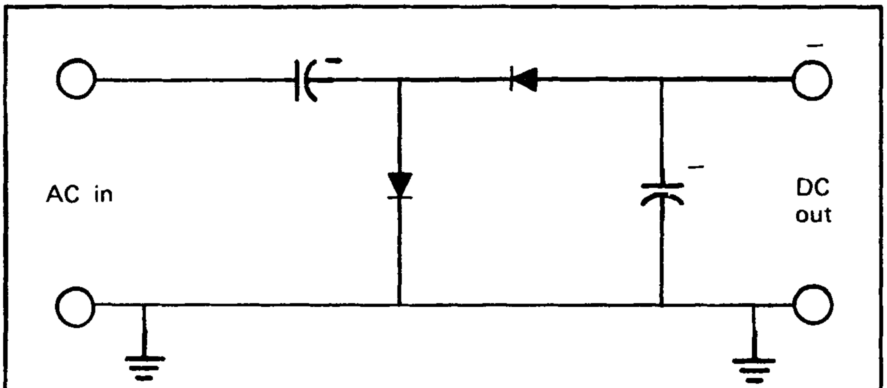


Figure 30 Full-Wave Doubler–Negative Polarity

Transformer-type supplies are now most commonly used, at least in larger CATV amplifiers. They permit simple, full-wave rectifier circuitry, and the ability to use tapped primary windings permits more efficient cable powering. Excessive rectifier output voltage is dissipated in the regulator as heat (except in switching-type regulators). Availability of primary transformer taps permits the technician to adjust the rectifier output voltage to an optimum value and also allows the regulator a reasonable operating range without excessive power wastage through heat dissipation. Transformers also provide for operation over a wide range of cable voltages. Some systems operate with nominal 60-volt AC power instead of the usual 30-volt AC power. The use of transformers permits easy change from 60-to 30-volt operation, simply by alternating transformer taps.

Full-wave operation of CATV amplifier power supplies has the advantages of permitting more efficient use of the cable for carrying power and guaranteeing the maintenance of symmetrical cable power waveform. If positive and negative cycles become different in voltage, there might be a tendency for corrosion to develop at splices and connectors.

Direct current would be the most efficient way to carry power on the cable, power would flow smoothly and continuously. Amplifiers would not need rectifiers and filters but would still require good voltage regulation. It would be easy to provide stand-by power in case of utility company power failure — a simple storage battery would be sufficient. However, DC powering is not used because of the great danger of corrosion at connectors and fittings, due to electrolytic effects. DC in the presence of moisture can act as an *electroplating current* and can cause serious damage to CATV plant.

V. WAVEFORMS IN FULL-WAVE POWER SYSTEMS

VOLTAGE

CURRENT

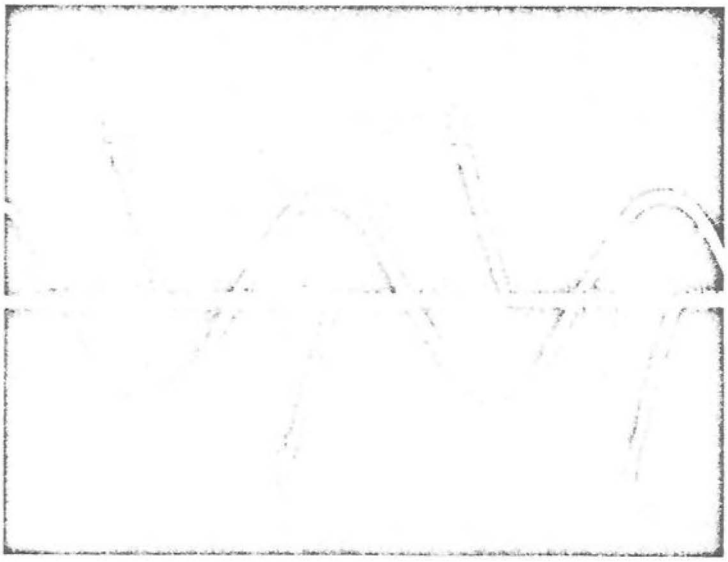


Figure 31

Figure 31 shows the voltage and current waveforms of a small CATV amplifier which has a full-wave, bridge-type rectifier system. This amplifier does not have a transformer, and its internal DC voltages are not tied to ground. Figure 31 represents a situation that occurs when the amplifier is very close to the power supply, and there is practically no cable resistance to limit the initial surge of charging current which flows into the filter capacitor at each half cycle. This sharp current pulse is not typical of current flow in full-wave systems but was observed here because the amplifier itself had no internal resistance to slow down or limit the charging current pulse. The peak current, in this case, was determined to be 0.50 amp, as measured on the oscilloscope. Compare this with the 1.0-amp current peak observed for a smaller, half-wave type amplifier. By using both power half cycles, the power supply is more efficient, and individual current pulses are smaller since there are twice as many of them. Since there is no significant cable resistance involved, there is no change in the voltage waveform.

When a resistor was used to simulate the effect of cable resistance, the waveforms changed somewhat. Note that the current pulse is smaller but wider, and that the voltage waveform is distorted by the effect of voltage drop in the cable resistance during the period of current flow. The waveforms are still symmetrical; i.e., positive and negative half cycles are the same. However, the current peak has been reduced to 0.26 amp.

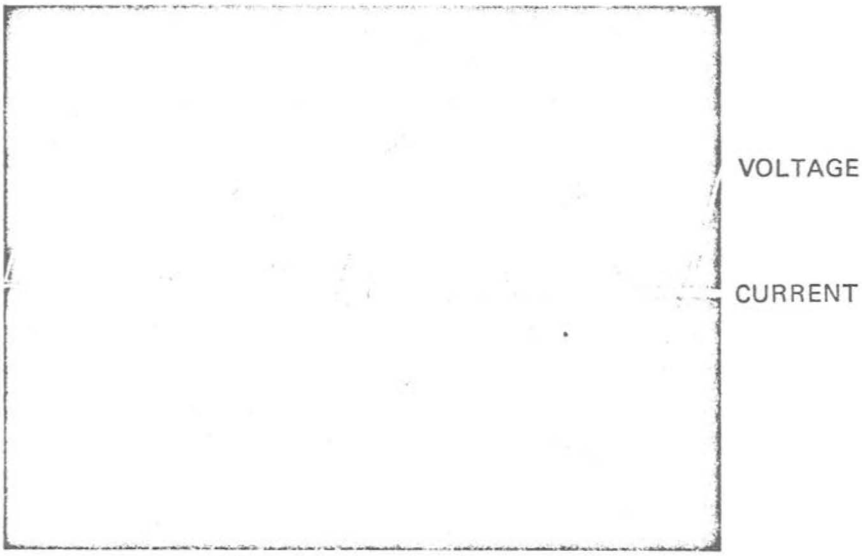


Figure 32

When a transformer is used in a CATV amplifier, the current waveform shows two components. One component is the current which flows through the rectifier diodes into the amplifier and supplies energy to the amplifier. The other component is *out-of-*

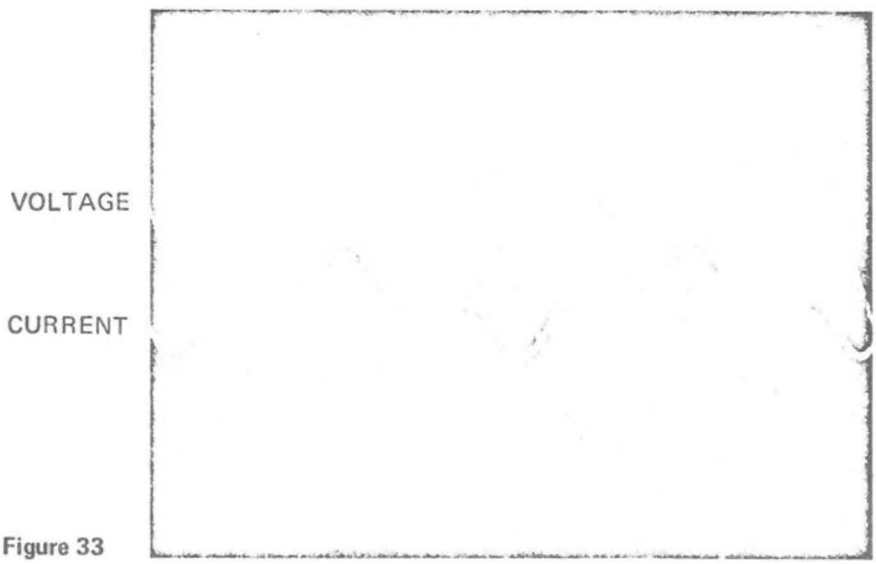


Figure 33

phase current in the inductance of the transformer itself. This current component is out-of-phase with the voltage and, therefore, does not represent any power transfer. Two photographs of current and voltage waveforms are presented — the first (figure 33), without cable resistance; and the second (figure 34), with cable resistance — to show the effect of cable voltage drop on the voltage waveform. This particular amplifier is a medium-sized CATV amplifier, and the current peak is approximately 0.7 amp.

VOLTAGE

CURRENT

Figure 34

The following tracing (figure 35) from the current and voltage form for a transformer-type power supply, more clearly shows the out-of-phase current, due to the transformer inductance. The out-of-phase current has been shaded in.

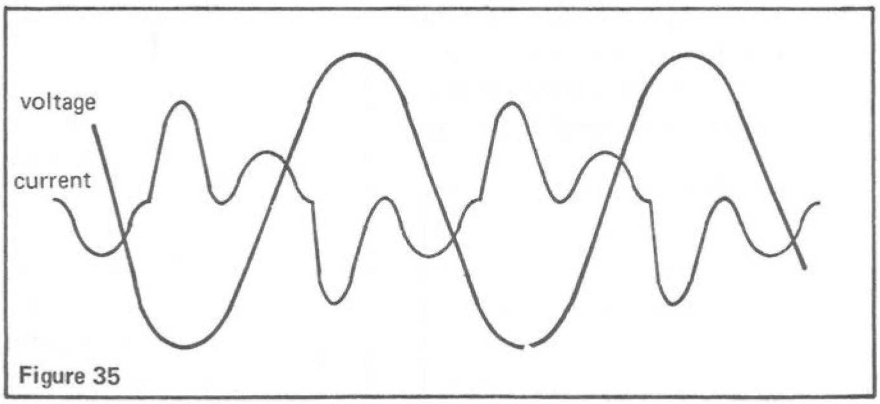


Figure 35

VI. HUM AND VOLTAGE REGULATION IN CATV AMPLIFIERS

Hum and voltage regulation are closely related in CATV amplifiers. The first filter capacitor in an amplifier power supply is usually not

NOTES

big enough to provide the required amount of hum filtering. Consider an amplifier whose internal power supply must provide 20 volts DC at 200 milliamps. A 250 μF filter capacitor is typical in such a power supply. Reference to design charts, available for detailed study of capacitor input filter systems, shows that such a filter will still have 5% ripple, which is intolerable in a CATV amplifier. This 5% ripple is further reduced to a more acceptable 0.1% level by the action of the regulator section. The regulator holds the DC voltage to a desired level and reduces hum by treating it as a form of voltage irregularity that must be removed from the DC supply. The regulator follows the individual ripple cycles and smooths them out, just as it smooths out voltage irregularities from other causes.

The detailed study of voltage regulators in CATV amplifiers is beyond the scope of this lesson, but the student will benefit from understanding the simplest and most common type of regulator, the series-controlled type. Figure 36 is a simplified schematic diagram of

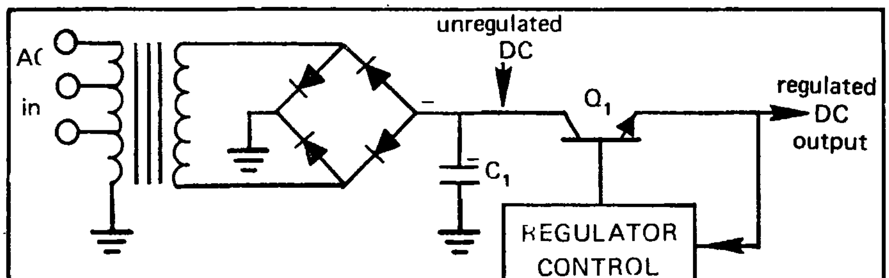


Figure 36

part of a series-regulated power supply. This diagram shows a transformer feeding a full-wave, bridge-type rectifier and a large filter capacitor. The capacitor might typically be 250 μF with a 50- or 75-volt rating. Q_1 is a large, power-type transistor which regulates the flow of current from the filter capacitor to the amplifier. Its base is controlled by circuitry, which senses the output voltage, compares it with a standard voltage (usually a zener diode), and develops an *error voltage* which controls the base of the series-control transistor Q_1. The difference between the *raw* DC at the collector of Q_1 and the regulated DC at the emitter, is dissipated as heat in the transistor. A lesser amount of heat is also generated in the transistors and circuitry in the regulator control section. The control transistor Q_1 must have some *working voltage* between its collector and emitter in order to function properly. If the raw DC at the collector falls too low, the regulator will not effectively filter out the residual hum; thus, hum bars will appear in the television pictures passing through this amplifier. If the raw DC voltage at the collector is too great, the power dissipation capability of the control transistor will be exceeded and, thus, will burn out.

The most common problem is that the voltage at the collector of Q_1

falls too low because of excessive voltage drops in the cable system, usually caused by too many amplifiers or cable runs which are too long. This low voltage of Q_1 nearly always appears as hum bars in the television picture. The situation is difficult to check with AC voltmeters, since we have seen that ordinary AC voltmeters can be misleading because of the way in which voltage waveforms change in the cable power system. The most reliable check of the performance of the cable power system is to check the voltage available at the collector of the regulator transistor. In many amplifiers this transistor is accessible for voltage measurements, and the collector is nearly always the outer covering of the transistor. A voltage check (DC) at this point is most meaningful. Many manufacturers provide instructions for making this voltage check. The voltage of some amplifiers is read with respect to ground. In other types (floating power supplies), the voltage is read with respect to a B— (B minus) terminal (not ground). Most manufacturers recommend a minimum of 6 or 7 volts between the collector and emitter of the control transistor. This voltage can be read by direct measurement between these transistor terminals, if they are accessible, or by measurement of the collector voltage to the power supply ground and comparison to the regulated output voltage. High voltages across the transistor are permissible, but they represent wasted power, since the voltage difference is dissipated in the transistor. The manufacturer's recommendation should be observed in this respect. If the manufacturer has not provided recommendations or instructions for voltage check, a procedure should be set up by the system's chief technician.

Remember, the DC at the collector of the control transistor is important; not the AC voltage that a meter might show on the cable test point. When in doubt, check the DC at the regulator section, which shows the peak DC that the available AC is capable of developing in the amplifier.

If the power supply in the amplifier is not accessible for checking DC voltage, the technician may wish to build a *power supply simulator*, using the same components that are used in the amplifier itself, but without the regulator, since the technician is interested in checking the DC available to the regulator. When an amplifier uses a transformer with primary taps, a full-wave bridge, and a filter capacitor, the circuitry might look like figure 37. This circuit (which

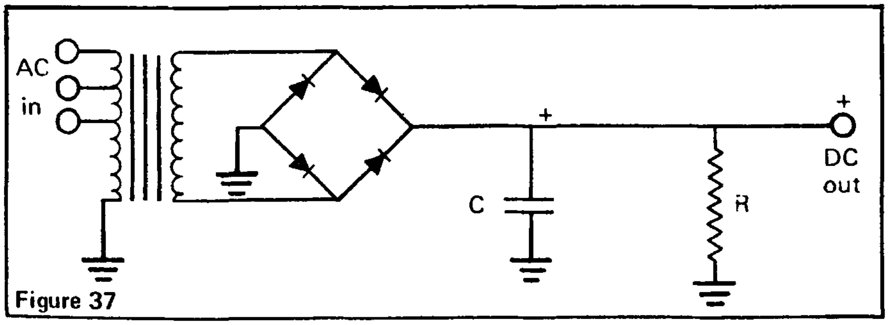


Figure 37

happens to use +DC internally) is intended to duplicate the circuitry and components in an actual amplifier. A DC meter reads the DC voltage available when the AC input is connected by test jumpers to the available AC. "C" can actually be much smaller than the filter capacitor in the amplifier, since there is practically no load on this test power supply. "R" is intended only as a *bleeder* and can be as much as 100,000 ohms. The meter that is used will carry at least a 20,000-ohm load. "C" could then be approximately 1 μF, or even 0.25 μF. Such a *test jig* would be used in cases where the amplifier power supply is not easily accessible for voltage measurement.

The importance of proper regulator operation in CATV amplifiers cannot be overemphasized. Some measurements and waveform

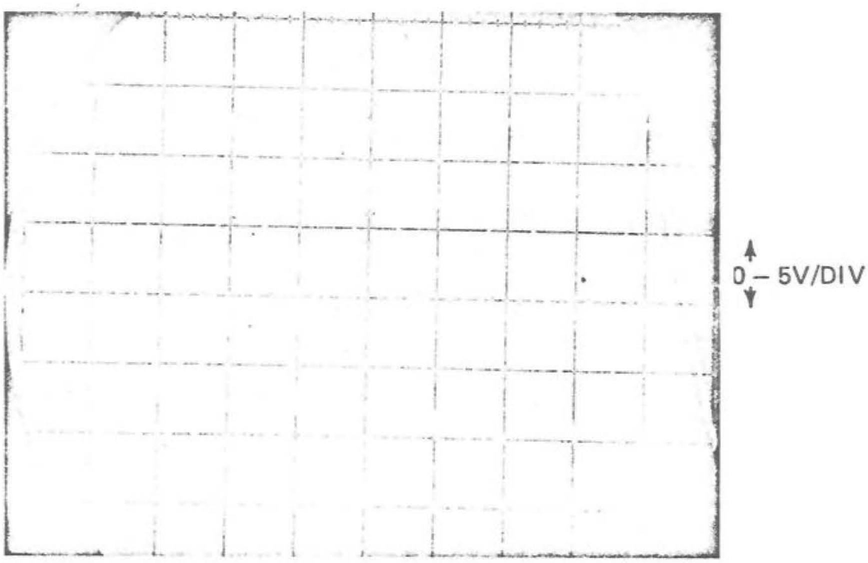


Figure 38

photographs from an actual CATV amplifier will help the student understand the importance of maintaining proper voltages in a cable power system. These measurements were made in a medium-sized

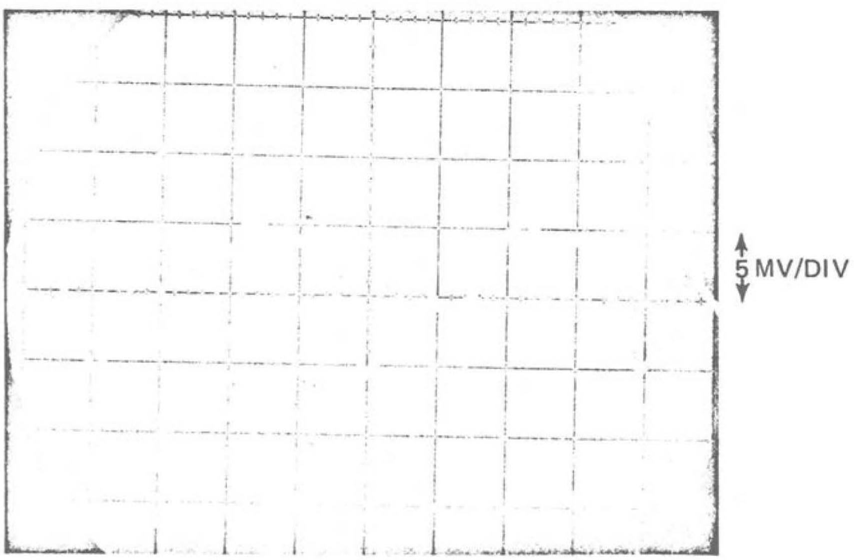


Figure 39

CATV amplifier having a transformer and full-wave bridge rectifier, as shown in figure 37. The capacitor used is 440 μF, 75 volts. Under *normal* operating conditions the voltage on the filter capacitor was found to be 27 volts DC, and the ripple at that point was measured as 0.95 V (RMS) on a true RMS voltmeter. The ripple at the filter capacitor was observed on a calibrated oscilloscope (figure 38). The DC output of the regulator section was measured as 19 volts DC, indicating 8 volts across the regulator transistor. The ripple (hum) in the regulated output voltage was measured as 2.7 millivolts (RMS) and was observed on the same calibrated oscilloscope (figure 39).

These figures demonstrate how effective the regulator is in reducing ripple to acceptable levels. RMS ripple has been reduced by a factor of approximately 300, and peak-to-peak ripple, by about the same ratio (see figure 40).

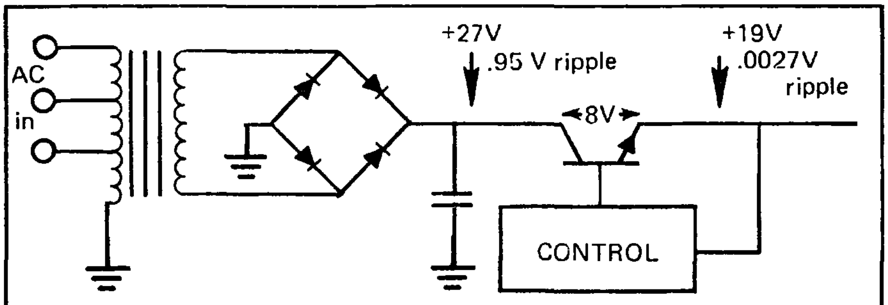


Figure 40 Ripple and regulator demonstration example.

VII. SURGE PROTECTION

CATV systems must be protected from the damaging effects of voltage surges in the cable. These surges are usually introduced through the power feed connection to the utility company power system, but may occasionally come from other sources. The voltage regulating transformers, which are often used to power CATV systems, provide some measure of protection. Additional protection is usually provided through the use of gas discharge tubes. In their simplest form, these tubes may be small neon lamps, which short out surges to ground when the surge reaches the firing voltage of the lamp (90 volts). Special surge-protection diodes are now more commonly used because of their greater surge-handling capability. These diodes have very low shunt capacitance and can be used directly across a coaxial cable in cases where their 2 pF capacitance does not affect circuit performance. Other surge-protection devices are sometimes used. These operate by shorting the surge to ground before it can harm the equipment.

NOTES

VIII. POWER CIRCUIT PROTECTION

Cable power circuits are always protected by a series of fuses and/or circuit breakers. These protective devices must be used in a logical

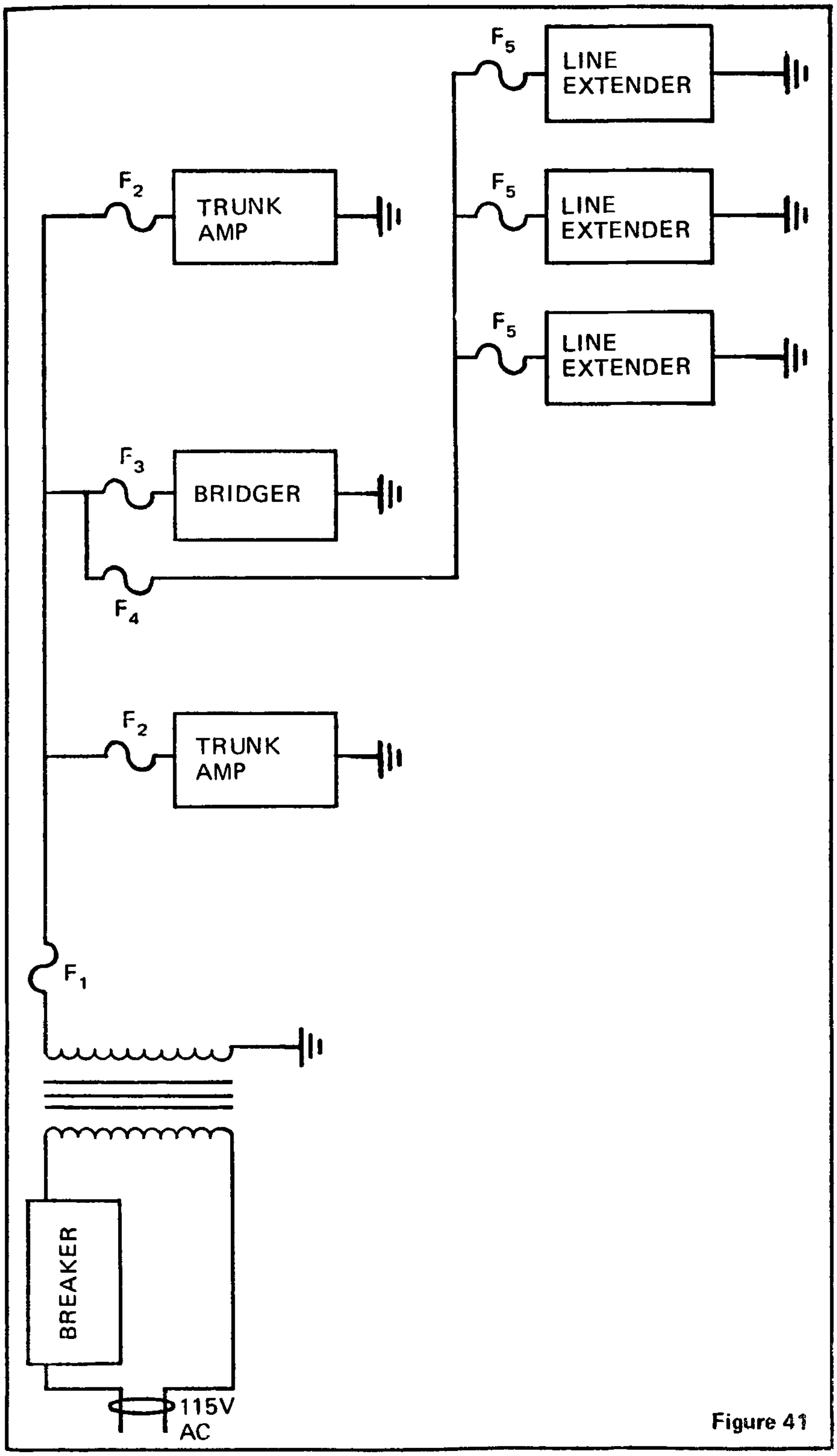


Figure 41

manner in order for them to do their job properly. Generally speaking, fuse or circuit breaker values increase along the power circuitry from the most distant equipment back toward the power source. This gradual increase prevents a minor problem on a sub-distribution line from blowing fuses at critical points in the system, thus disrupting service to large portions of the system when it should have affected only a small part of the system.

Consider a simplified schematic diagram of a portion of a cable power layout (figure 41). The main fuse F_1 must be capable of carrying all the current to be drawn from the power feed. This fuse is backed up by a circuit breaker or a fuse on the primary side of the cable power transformer. The primary breaker protects the utility company system against any kind of short circuit in the cable system or in the power feed transformer. Fuses F_2 and F_3 protect individual main line amplifiers. If any of these amplifiers develop a short, the fuse within the amplifier will blow. Obviously, these fuses must be lower in value than F_1; otherwise, a short in the furthest trunk amplifier would blow F_1, the main fuse, before it blows its own amplifier fuse. Everything connected to that power feed would then be off. If only F_2 blew, then at least part of the system would still be working.

F_4 is also considered to be a *main* fuse, since it protects a whole distribution line with several line extenders. The fuses in each line extender (F_5) must be lower in value than F_4; otherwise, a short in the furthest line extender would blow the main distribution line fuse and then blow the rest of the fuses on the entire line.

Thus, we see that the size of the fuses must be carefully controlled and designed with regard to the nature of the device or system being protected. Fuses must be graded in size in order to provide a logical chain of protection to the system.

Tracing of short circuits in a cable power system can be difficult and time consuming. If the fusing has been logically designed, knowing the position of the blown fuse in the system can be helpful. If the main fuse F_1 has blown, it is unlikely that the short is in an individual amplifier or on one of the distribution legs, since these are all protected with fuses smaller than the main fuse F_1. Similarly, if one of the distribution line fuses (F_4) has blown, it is unlikely to be one of the line extenders, since these are individually protected by fuses smaller than F_4. If the fault had been in one of the line extenders, its own fuse should have blown before the next major fuse preceding it. This emphasizes the importance of proper planning of fusing, and the importance of replacing fuses with the proper values when they need replacement.

A helpful device for locating short circuits is a small light bulb attached to a dummy fuse. A 32-volt, 10-watt (or even lower wattage) light bulb can be wired into a fuse plug so that it takes the place of the fuse when it is plugged in. The resistance of the lamp limits the short circuit current to a safe value. The lamp remains lit until the short circuit is cleared.

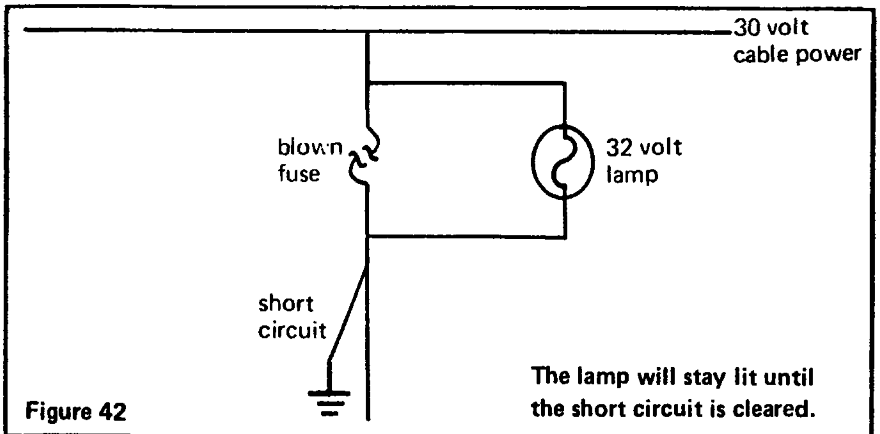


Figure 42

Indicating fuse holders that use this same principle are available. A small lamp of suitable voltage rating is wired across the fuse. When the fuse blows, the lamp lights up, since the short completes the circuit for it. The lamp goes out when the short circuit is cleared. It is not necessary to replace every fuse holder with this indicator type, but a special fuse holder can be devised with a small lamp that provides this test feature when desired.

THE CATV MODULATOR

By

Alex B. Best
Commercial Communications
Scientific-Atlanta, Inc.

Prepared for Presentation to
Nineteenth Annual Convention
National Cable Television Association
Chicago, Illinois
June 19-22, 1970

THE CATV MODULATOR

Alex B. Best, Senior Engineer
Commercial Communications
Scientific-Atlanta, Inc.

The purpose of the modulator in CATV systems is to accept a source of video and audio or 4.5 MHz FM signal and convert these to a standard television signal. This source of information may be a camera, a demodulator, or microwave relay output. Because the CATV modulator performs the same function as a television transmitter, its specifications must conform very closely to those of a transmitter.

The block diagram of a CATV modulator is shown in Figure 1. The video enters the unit through the "video phase equalizer" module. The function of this module is to predistort the phase versus frequency characteristics of the video to conform to the standard FCC predistortion requirements required for color transmission. After leaving this module the video enters the video amplifier module through a second phase equalizer section. This second equalizer, which is an integral part of the video amplifier, predistorts the delay characteristics of the video to compensate for delay errors introduced by the vestigial sideband filter and other filters in the modulator itself.

The "video amplifier" amplifies and clamps the video signal. The synchronizing pulses are clamped to a reference level in order to ensure that the sync tip level applied to the "video modulator" does not vary with the average brightness content of the signal. The "video modulator" module accepts the video signal and modulates this information on a 45.75 MHz IF carrier. Through the use of a "Cowan bridge" type modulator differential gain and differential phase are minimized.

The output of the video modulator contains both the upper and lower sidebands. The "vestigial sideband filter" module, which is at the IF frequency, passes the lower sideband and rejects all but a small portion of the upper sideband in order to conform to the standard TV channel bandwidth. The vestigial sideband filter module also has provisions to accept a combined video and sound IF signal from an external source. When signals from an external source are being used, the 45.75 MHz oscillator is automatically disabled. The output of the vestigial sideband filter goes to the output converter for conversion to the desired channel. Additionally, a sample of the vestigial sideband filter output is available at the rear of the chassis. This output can be used to provide IF signals to a heterodyne signal processor such as the Scientific-Atlanta Model 6100 to replace absent off-air signals.

The Model 6300 modulator provides two audio options. The "A" option accepts balanced audio and converts this signal to a 4.5 MHz FM subcarrier. The importance of maintaining the audio subcarrier frequency at 4.5 MHz ±1 kc is discussed in a later section of this paper. The "S" option accepts a direct 4.5 MHz FM subcarrier input or a 4.5 MHz FM subcarrier and video combined input. One of the two 4.5 MHz FM signals mentioned above is then applied to a balanced modulator along with a sample of the 45.75 MHz oscillator. The output of this modulator is fed to a filter network tuned to 41.25 MHz which rejects the unwanted sideband. A portion of this sound IF signal is also available at the rear of the modulator chassis. The output of the filter network, which incorporates a front panel level control, is then fed to the output converter. In the Model 6300 CATV Modulator, the sound IF and video IF are upconverted in separate converters and then combined at the desired output frequency in a passive device. The reason for this approach to up conversion is discussed in detail later.

To be able to accurately set both FM deviation and percent AM video modulation, a sample of the 4.5 MHz FM subcarrier signal and the video IF signal enter the meter circuit module. Depending on the position of a front panel switch, the meter on the front panel will either indicate FM deviation in kc or AM modulation in percent.

If the output signal is to be phase locked to an off-air signal, an optional input converter and phase detector are required. The 45.75 MHz IF signal which is normally crystal controlled is then phase locked to the converted off-air signal. By using the same local oscillator signal for both down and up conversion, phase lock is achieved.

The remainder of this paper will discuss the areas of particular interest in the design of a CATV modulator.

Output Converter

The function of the output converter is to convert the video, sound, and color IF signals from the standard IF frequency centered at 44 MHz to any desired output channel. There are various techniques for achieving this desired frequency conversion; however, the most common way is to sum the picture, sound, and color signals, along with a local oscillator signal, and apply them to a nonlinear device. As an example, let us consider the situation shown in Figure 2. For the standard IF frequency centered at 44 MHz,

$$f_v = \frac{\omega_v}{2\pi} = 45.75 \text{ MHz}, \quad f_s = \frac{\omega_s}{2\pi} = 41.25 \text{ MHz and}$$

$$f_{lo} = \frac{\omega_{lo}}{2\pi} \text{ is the oscillator frequency required for}$$

the desired output channel. In our example, we will ignore the color subcarrier signal to simplify the problem and discuss the consequences of this simplification later.

The transfer function of any nonlinear device can be written as a power series expansion about its operating point (1). In the above example, i_o, the output current, can be represented by the equation

$$i_o = a_o e_{in} + a_1 e_{in}^2 + a_2 e_{in}^3 + a_3 e_{in}^4 + \ldots\ldots + a_{n-1} e_{in}^n$$

where a_0, a_1, a_2,....a_{n-1} are constants determined by the particular nonlinear device and its operating point. If now in the above equation we will substitute for e_{in} the sum of $A_{1o}\cos\omega_{1o}t$, $A_v\cos\omega_v t$, and $A_s\cos\omega_s t$ and expand each term, we will find frequency components in the output current i_0 which are equal to the three input frequencies of all three input signals and their harmonics, plus a dc term. In the above example we will expand the input signal only out to the fourth order, (it is assumed that the fifth and higher order terms are much smaller in magnitude than the lower order terms if the mixer is biased properly.) and consider only those frequency components within 6 MHz of the upper and lower edges of the desired output channels. In most output converters the filter network following the mixer is adequate to reject any frequency components which are outside these limts. The spectrum of the output voltage e_0, is shown in Figure 3. Here we have ignored two other terms whose frequencies are the same as the desired outputs. They are cross modulation terms and are extremely small compared to the desired outputs. Here $a_1 A_{1o} A_v \cos(\omega_{1o} - \omega_v)t$ is the desired output picture carrier and $a_1 A_{1o} A_s \cos(\omega_{1o} - \omega_s)t$ is the desired output sound carrier. From the coefficients of these two terms we can see that they are directly proportional to both a_1 and their respective IF levels. Since the desired output signal comes from a term in the power series expansion with the coefficient a_1, that is, the square law term, the mixer or nonlinear device should be biased so as to maximize a_1. The other two terms, one 4.5 MHz below the desired picture carrier and the other 4.5 MHz above the sound carrier, are spurious outputs and if large enough will create beats in adjacent channels of a multi-channel system. Upon examining the coefficients of these spurious outputs we conclude that they are generated from the fourth order term in the power series expansion. It is unfortunate that in most nonlinear devices the bias point which maximizes a_1 also maximizes a_3. Let us pause to consider here a few minutes the importance of the coefficients of these four output signals since the results have a practical application in most up-converters in existence today. For a particular video and sound IF input level, the desired output levels will be proportional to both a_1 and their respective input levels, A_v and A_s whereas the lower frequency spurious output will be proportional to a_3, A_s, and A_v^2 and the higher frequency spurious output proportional to a_s, A_v, and A_s^2. If we assume the normal amplitude difference between the picture and sound carrier of 15 dB, we find that the lower frequency spurious output will be 15 dB higher in amplitude than the other spurious output. We also conclude that since this higher amplitude spurious output is proportional to A_v^2 and A_s, for every dB that we raise the amplitude of A_v and A_s, the relative amplitude difference between the desired output and this undesired output will decrease one dB. For this reason most manufacturers recommend the use of a good bandpass filter on the output of their up-converts if the output level is run above a certain level.

From the above conclusions, we see that if we wish to up-convert the sum of the video and sound IF signals without generating any spurious outputs, we should consider two important points. Number one is the nonlinear device selection and number two is the level of the signals at the mixer input. If, however, we now assume that we have been able to up-convert the sum of the video and sound IF signals to the desired output frequency without excessive amplitudes on the undesired outputs, we must still amplify these desired outputs to their final output level without generating more spurious outputs. With this in mind, let us now consider the situation shown in Figure 4.

If we follow the same steps as we did in the mixer, we will find the generation of the same two spurious output signals we did before; however, in the output amplifier they occur due to the third order term in the power series expansion. The spectrum of the output E_o is shown in Figure 5. Here again we have ignored two cross modulation terms whose levels are extremely small compared to the desired output. Fortunately, the bias point which maximizes A_o, that is gain, in the output amplifier tends to minimize A_2 and therefore the amplification of the output signal without generating spurious outputs is not quite as difficult as up-converting them unless of course the signals get excessive in amplitude.

In all of the above discussions we have ignored the color sub-carrier signal in order to simplify the calculations. The justification for doing this is that in a normal CATV television signal the color-subcarrier signal is typically 10 dB smaller than the 15 dB down sound carrier and therefore the largest spurious output generated by this color signal is at least 10 dB smaller than the largest one generated by the sound. Since these color generated spurious signals lie very close to the ones generated by the sound signal, their relative amplitudes will change very little due to the filter characteristics.

The conclusion to be drawn from the above example is that it is extremely difficult to sum the sound, color, and video IF signals and up-convert them in a common mixer to the desired output channel without generating spurious outputs. Even if we up-convert the sound and video plus color in a separate mixer and then add them before amplification, we would still run the risk of generating spurious outputs, especially if large output levels are desired.

The most efficient output converter, i.e., one which minimizes the generation of spurious outputs, would be one which up-converts the video with color IF signal and sound IF signal in a separate mixer, amplifies them in separate amplifiers, and then combines them in a passive combiner. It is this up-conversion scheme which is used in the Scientific-Atlanta modulator.

Audio to 4.5 MHz Converter

According to the FCC specifications on a color television transmitter, the 4.5 MHz frequency spacing between the video and sound carrier must be held to a frequency tolerance of ± 1 kHz (2). Before we consider a method by which this accuracy can be achieved in a CATV modulator, let's consider the problems which could and do arise in present day CATV systems because of errors in this frequency separation. Shown in Figure 6 is a typical color television receiver response up to the video detector. In the course of color television field tests, it has been found that receivers need about 50 dB of on-channel sound attenuation in the IF channel in order to prevent the beat between the sound carrier signal and color subcarrier signal from appearing in the picture (3). Because the color sidebands extend up to within 320 kHz of the sound carrier, extremely sharp skirted notch filters are used to suppress this on-channel sound signal.

Now let's assume that we are manually fine tuning a color television receiver to Channel 8 which has as its input the output of the modulator with no adjacent channels present. To correctly fine tune this set we would adjust its local oscillator to place the on-channel sound signal in the deepest portion of its trap. If now our video-sound frequency spacing were incorrect due to the audio to 4.5 MHz converter, the consequences would be two-fold: first, if the frequency error was in a direction such as to reduce the spacing between video and sound carrier, it would be difficult to eliminate the 920 kHz beat between the sound and color signals and still obtain a sufficiently saturated color picture. Secondly, since all sets manufactured in the United States today use the principle of intercarrier sound detection, a sufficient error in this 4.5 MHz signal could cause a distortion in the detected audio signal.

In a typical CATV system we very seldom have the single channel situation discussed above, but in more cases than not have the situation shown in Figure 7. Here each channel is bordered on each side by another channel which is of equal magnitude.

Let us now assume that we are going to manually fine tune a color television set to Channel 8 which has present at its input the composite signal shown in Figure 7. In order to produce a picture on the screen which is free from beats, our problem is two-fold. Number one, we must adjust the local oscillator of the set so as to place the on-channel sound of Channel 8 in its sound notch to eliminate the 920 kHz beat mentioned above; and second, we must place the adjacent channel sound carrier, that is, the Channel 7 sound carrier, in the adjacent channel sound trap to eliminate the 1.5 MHz beat which could occur between the Channel 8 video carrier and the Channel 7 sound carrier. If now we assume that the traps are properly tuned in the television receiver, our main concern would be with the frequency separation accuracy of Channel 7 and Channel 8's sound carriers. Of course there are several sources of inaccuracies in the separation of these two carriers. Not only is there the possible error due to an audio to 4.5 MHz converter which is off frequency, but offsets to minimize co-channel, crystal tolerances in UHF converters and input and output converters are all possible sources of error.

Although the audio to 4.5 MHz converter in a modulator is only one of many possible sources of frequency error, it could, so to speak, be "the straw that breaks the camel's back" and mean the difference between a good quality channel or one which is extremely difficult to fine tune to eliminate the beats mentioned above.

The simplest and most common way to control the accuracy of the 4.5 MHz FM signal generated in a modulator is to apply this FM signal to a 4.5 MHz discriminator, filter out the audio components, amplify the resulting dc error voltage, and apply this dc voltage to the frequency determining element of the 4.5 MHz oscillator. Such a system is shown in Figure 8.

There are several possible sources of error in such a system. First, the output frequency is determined almost entirely by the zero crossing of the discriminator and to hold an accuracy of ± 1 kHZ on 4.5 MHz would require a zero crossing which is accurate to ± .022%. This is extremely difficult to maintain with conventional tuned circuits. Another problem with the system shown in Figure 3 is that a typical 4.5 MHz discriminator is very insensitive to a 1 kHz frequency variation and therefore would have to be followed by a high gain dc amplifier to reduce the error to below 1 kHz.

A better scheme, and the one which is used in the Scientific-Atlanta modulator, is the one shown in Figure 9. Here accurate center frequency control is obtained by comparing the output of the 4.5 MHz oscillator with an accurate standard, in our case a 4.2 MHz crystal oscillator, and applying the low frequency difference signal to low frequency discriminator whose stability has a much smaller percent effect on the 4.5 MHz output signal. By using a discriminator whose zero crossing is at 300 kHz, an accuracy of only ± .33% is now required in order to maintain the output 4.5 MHz signal to within ± 1 k. This type of stability can easily be obtained with conventional tuned circuits. Also, this lower frequency discriminator offers a considerable improvement in sensitivity over one at 4.5 MHz. In actual practice, this particular design held the 4.5 MHz FM signal to within ± 500 cycles from -20° F to + 120° F.

Envelope Delay

To produce a faithful image at the television receiver kinescope, all of the frequency components which make up the video signal must have an equal time delay from their source of origination, such as a camera to the display device, in our case, the subscriber's home receiver kinescope. The majority of any variation in time or envelope delay is concentrated in the IF and video amplifier sections of the receiver and the vestigial sideband filter of the transmitter. The pictorial results of this delay distortion are three-fold: first, the variation in envelope delay between the high and low video frequency components causes preshoots and undershoots in the rapid transitions between luminance levels of the picture. This can lead to excessive ringing and smears in the picture. Second, a variation in envelope delay over the range occupied by the color signals, that is, the I and Q color difference signals, causes cross talk which results in color errors at the transitions of the color portion of the picture. Third, a variation in envelope delay between the color portion and luminance portion of the complete signal will upset the time coincidence of these signals at the home receiver and, if severe enough, will produce the so-called "funny paper" effect.

To eliminate this source of distortion in the Scientific-Atlanta modulator, a video delay equalizer is inserted at the video input to the modulator. This equalizer not only corrects for the delay errors generated within the modulator itself, but also corrects for the errors generated by the "average home receiver." To this end, the envelope delay of the "average receiver" has been determined and the inverse of this delay curve is specified by the FCC to be the delay required for a color television transmitter (4). Shown in Figure 10 is the required envelope delay distortion and its tolerances. The FCC specifications reads: A sine wave, introduced at those terminals of the transmitter which are normally fed the color picture signal, shall produce a radiated signal having an envelope delay, relative to the average envelope delay between 0.05 and 0.20 MHz, of zero microseconds up to a frequency of 3.0 MHz; and then linearly decreasing to 4.18 MHz so as to be equal to - 0.17 microseconds at 3.58 MHz. The tolerance on the envelope delay shall be ± 0.05 microseconds at 3.58 MHz. The tolerance shall increase linearly to ± 0.1 microsecond, down to 2.1 MHz, and remain at ± 0.1 microsecond down to 0.2 MHz. The tolerance shall also increase linearly to ± 0.1 microsecond at 4.18 MHz (5).

The phase equalizer in the Scientific-Atlanta modulator is comprised of two sections. One section is built in front of the video amplifier and clamper circuit and is an integral part of the video modulator module. This phaze equalizer compensates for the delay errors generated within the modulator itself. The second section of the phase equalizer compensates for the delay errors generated by the average home receiver and is an optional plug-in module through which the video signal passes before going to the video modulator module.

Differential Gain and Phase

Differential gain is defined as a change in the level of the 3.58 MHz color subcarrier as the level of the luminance signal on which it rides is varied from blanking to white. Differential gain is normally expressed in dB or percent. Differential phase, normally expressed in degrees, is defined as a change in the phase of the 3.58 MHz color subcarrier as the luminance level is varied from blanking to white. The test signal most commonly used to measure both differential gain and phase is shown in Figure 11. Here, the 3.58 MHz signal is riding on either a ramp or stairstep luminance level with the 3.58 MHz signal beginning at the blanking level and ending at a luminance level corresponding to white level of 12½% of sync tips.

The pictorial effects of differential gain and phase are related to the broad area color portions of the picture. Errors in differential phase causes incorrect hues in the picture and errors in differential gain affects the color saturation of the picture. It is interesting to note that the pictorial effects of differential gain and phase are entirely different from those of envelope delay which show up as errors in the transitional regions of the picture.

In the CATV modulator the common cause of differential gain and phase is non-linearities which occur in the video amplifier and video modulator sections. Through careful design of these areas of the modulator, differential gain and phase can be kept below 1 dB and $\pm 1^{o}$ respectively at 87.5% modulation.

Phase Locking Capability

As channel space becomes more and more crowded in CATV systems, it is inevitable that there will be some programs carried on the cable system which occupy the same channel as a local transmitter. If the channel carried on the cable has originated from a modulator whose output frequency is crystal controlled, but not phase locked, there will be a slight but unavoidable frequency difference between the cable signal and the local transmitter signal. The result of this frequency difference, due to either stray pickup in the TV receiver or leakage into the cable system itself, is a co-channel interference beat. As a general rule, this interference beat is visible all the way to the Grade B contour and beyond (6). A block diagram of a modulator which has the capability of being phase locked to an off air signal is shown in Figure 1. As shown in the block diagram, a sample of the signal to which the output of the modulator is to be phase locked is converted down to the standard IF frequency in a crystal controlled oscillator. This IF picture carrier is then fed to a phase detector along with a sample of the output of the 45.75 MHz varactor turned oscillator. The output of the phase detector is a dc voltage which is proportional to phase difference between the two IF signals. This dc control voltage is then applied to the varactor diode in the 45.75 MHz oscillator. This 45.75 MHz phase locked signal is then used as the IF picture carrier in the modulator. To complete the phase locking scheme, the oscillator for the output converter must have identically the same frequency as that of the input converter and therefore a sample of the input converter oscillator is fed to the output converter to be used for up conversion. In the absence of an input signal, the 45.75 MHz oscillator becomes crystal controlled to insure against frequency drift which could cause interference in adjacent channels.

Through the technique of phase locking, this co-channel beat can be eliminated completely. The stray pickup problem will now cause a leading ghost in the received picture, but this ghosting is generally restricted to some sets within the Grade A contour (6).

Modulation Meter

Now that local origination is becoming mandatory in many CATV systems, there will be an ever increasing need for the CATV operator to be able to quickly and accurately monitor and adjust both the percent AM depth of modulation on the video carrier and the deviation in kc on the FM sound carrier. At the television broadcast stations, these are set to be 87.5% and ±25 kc respectively. In CATV modulators it is important that these modulation standards be generally adhered to, or problems can arise. If the percent AM modulation on the video carrier reaches 100% because of variations in video levels at the input to the modulator or because the video modulation control is improperly adjusted, two problems occur: first, the "light" areas of the picture merge into a "saturated white" appearance; second, because the principle of intercarrier sound used in today's television receivers require both picture and sound carriers to be present, overmodulation of the picture carrier causes buzzing in the audio. On the other extreme, modulation percentages less than 87.5 % only cause a loss of contrast at the home receiver. For this reason, it is generally recommended that the percent AM modulation on the video carrier in CATV modulators be set at approximately 80%.

As mentioned above, the deviation on the sound carrier is normally set for ± 25 kc. If this deviation is set lower there will be a loss of volume at the home receiver. On the other hand, if this deviation is set too high, not only will the volume at the set be comparatively high, but the resulting increase in the bandwith of the sound signal [B.W. $\approx$ 2(frequency deviation + highest modulating frequency)] will make it more difficult to fine tune some sets to eliminate the 920 kc beat between the sound and color signals.

In the Scientific-Atlanta modulator, the percent AM modulation on the video carrier and deviation on the FM sound carrier can be accurately adjusted and monitored, all from front panel controls. This is accomplished in the following manner. A sample of the 4.5 MHz FM subcarrier is fed to a calibrated linear discriminator. The audio out of this discriminator is then peak detected and fed to a front panel meter which is calibrated to read FM deviation directly in kilocycles. Also, a sample of the 45.75 MHz modulated IF video carrier is fed to a detector circuit which removes the video modulation present on this carrier. The amplitude of this video signal is compared in a peak detector to that of the unmodulated IF carrier. The resulting dc voltage which is a measure of the percent

AM modulation, is then applied to the front panel meter, which is calibrated to read AM modulation directly in percent.

Conclusion

In the past few years modulators have been used primarily for converting a black and white video signal to a standard televion signal. The source of this signal was usually a camera on a time and weather channel or possibly the output of a camera used for local origination. In either case it was typically a situation where a low quality modulator would not limit the quality of the signal being delivered to the customer's receiver. This is not the case today. With local origination becoming mandatory in many situations, and with color cameras becoming available with reasonable price tags, the quality of the modulator must approach that of a broadcast transmitter.

This paper has discussed several unique design approaches for a CATV modulator. Through the use of these features in the design of a modulator, some of the technical problems which could compromise the resulting quality of the customer received picture have been overcome. By incorporating these features in a modern solid state CATV modulator, it not longer will be the "weak link" in the chain of equipment from the video origination point to the customer's kinescope.

(1) Frederick E. Terman, "Electronic and Radio Engineering", Fourth Edition, pp. 204-206.

(2) FCC Rules and Regulations 73.668.

(3) John E. Allen, "Beat Between Sound Carrier and Color Signal Components in a Television Receiver", Presented at AIEE Convention, November 4, 1953.

(4) G. L. Fredenall, "Delay Equalization in Color Television", Proceedings of the IRE, January 1954, pp. 258-262.

(5) FCC Rules and Regulations 73,687.

(6) A. S. Taylor, "On-Channel Carriage of Local TV Stations on CATV", IEEE Transactions on Broadcasting, December 1969, pp. 102-104.

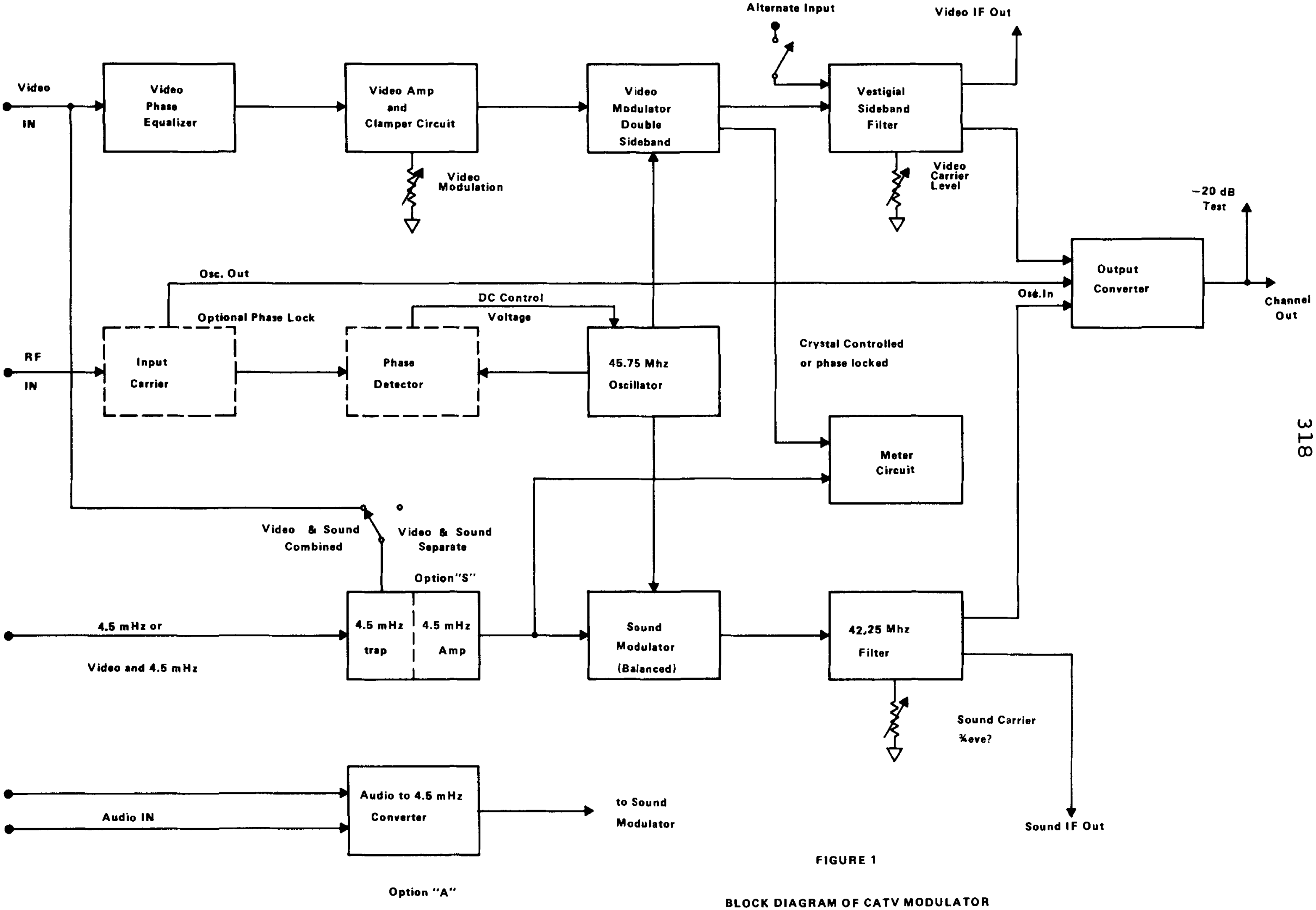
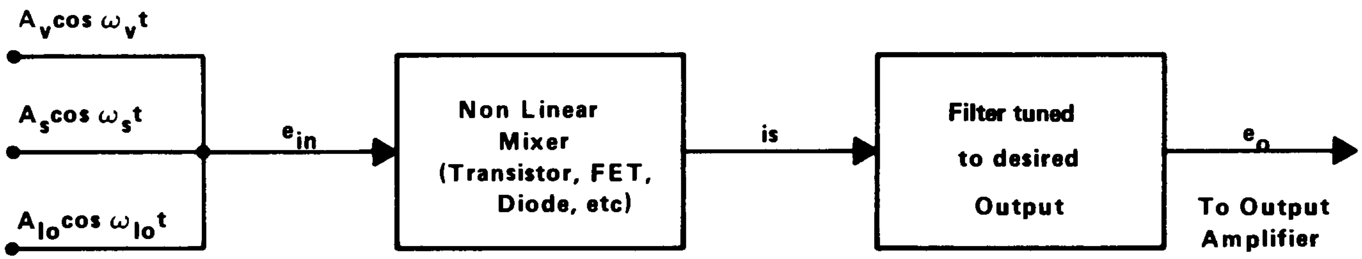


FIGURE 1

BLOCK DIAGRAM OF CATV MODULATOR

OUTPUT CONVERTER



Where $A_v \cos \omega_v t$ is the video IF signal

$A_s \cos \omega_s t$ is the sound IF signal

$A_{lo} \cos \omega_{lo} t$ is the local oscillator signal

FIGURE 2

COMMON UPCONVERSION TECHNIQUES

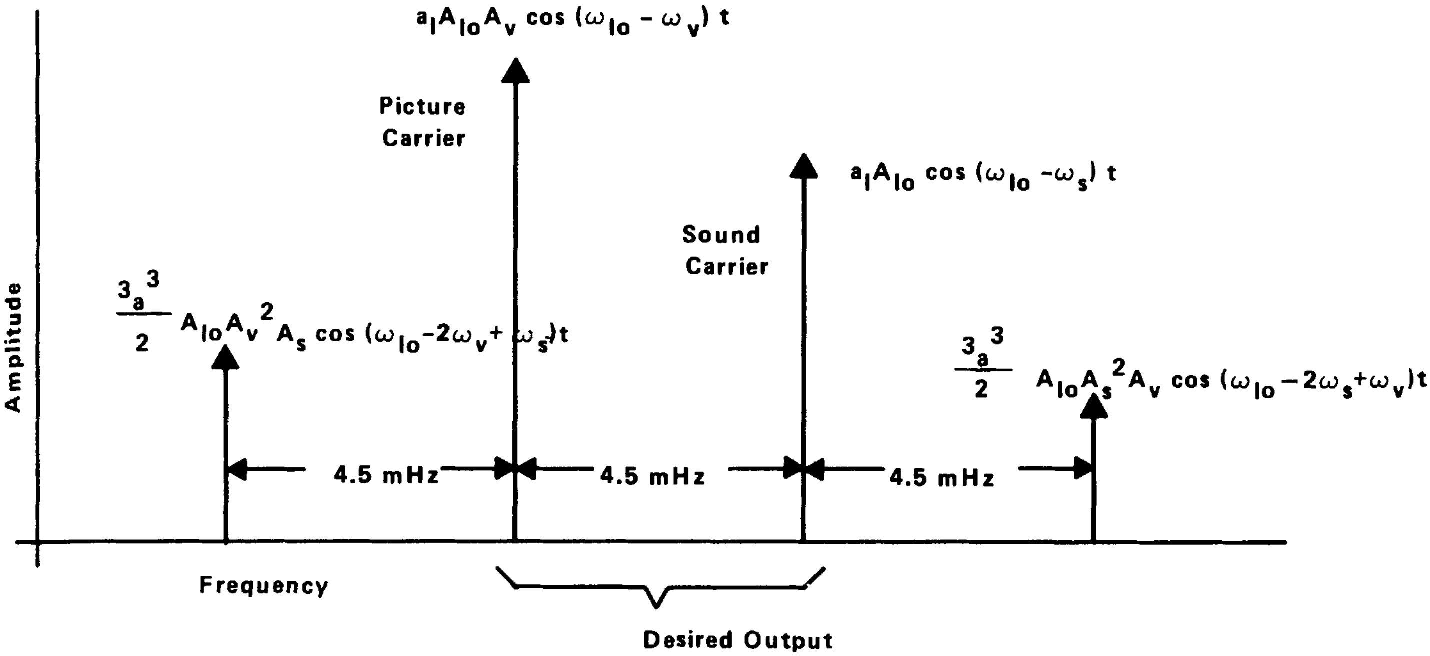
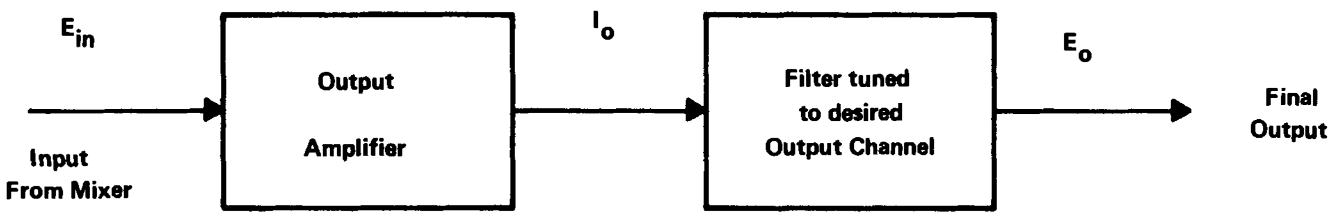


FIGURE 3

SPECTRUM OF MIXER OUTPUT e_o



$$E_{in} = a_l A_{lo} A_v \cos(\omega_{lo} - \omega_v)t + a_l A_{lo} A_s \cos(\omega_{lo} - \omega_s)t$$

and

$$I_o = A_o E_{in} + A_2 E_{in}^2 + A_2 E_{in}^3 + \ldots + A_{n-1} E_{in}^N$$

Where $f_{lo} - f_v = \dfrac{\omega_{lo} - \omega_v}{2\pi}$ = desired output picture carrier

$f_{lo} - f_s = \dfrac{\omega_{lo} - \omega_s}{2\pi}$ = desired output sound carrier

FIGURE 4

OUTPUT AMPLIFIER

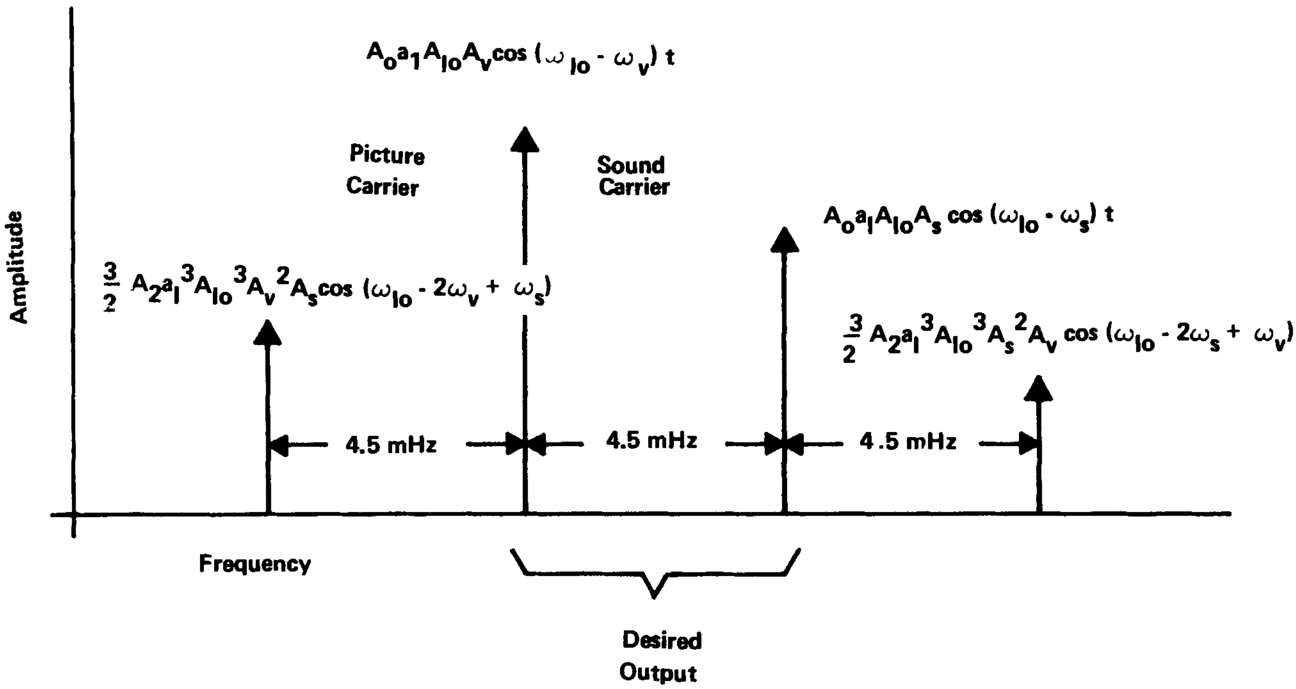


FIGURE 5

SPECTRUM OF AMPLIFIER OUTPUT E_0

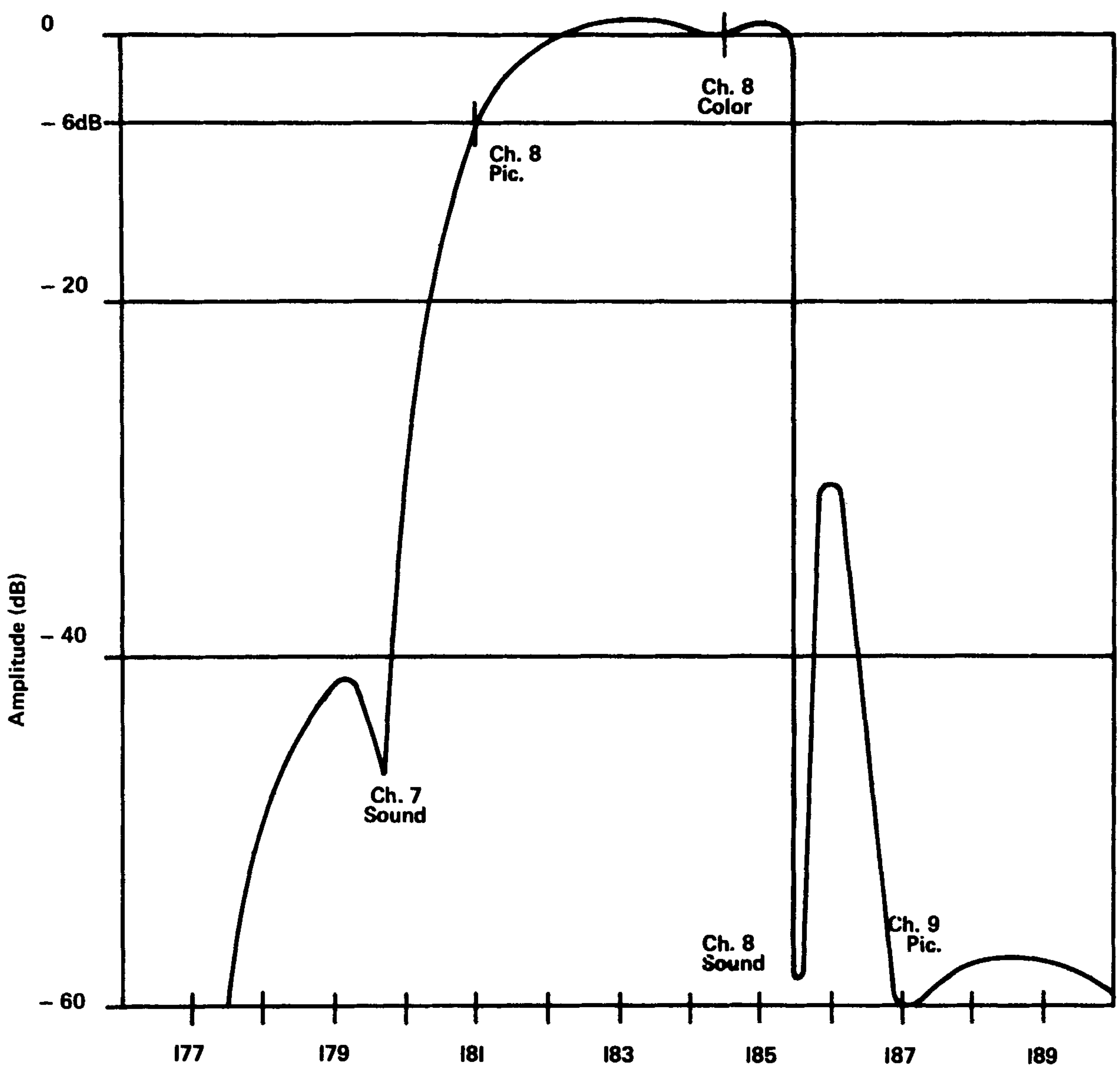


FIGURE 6

CH. 8 RECEIVER REPONSE

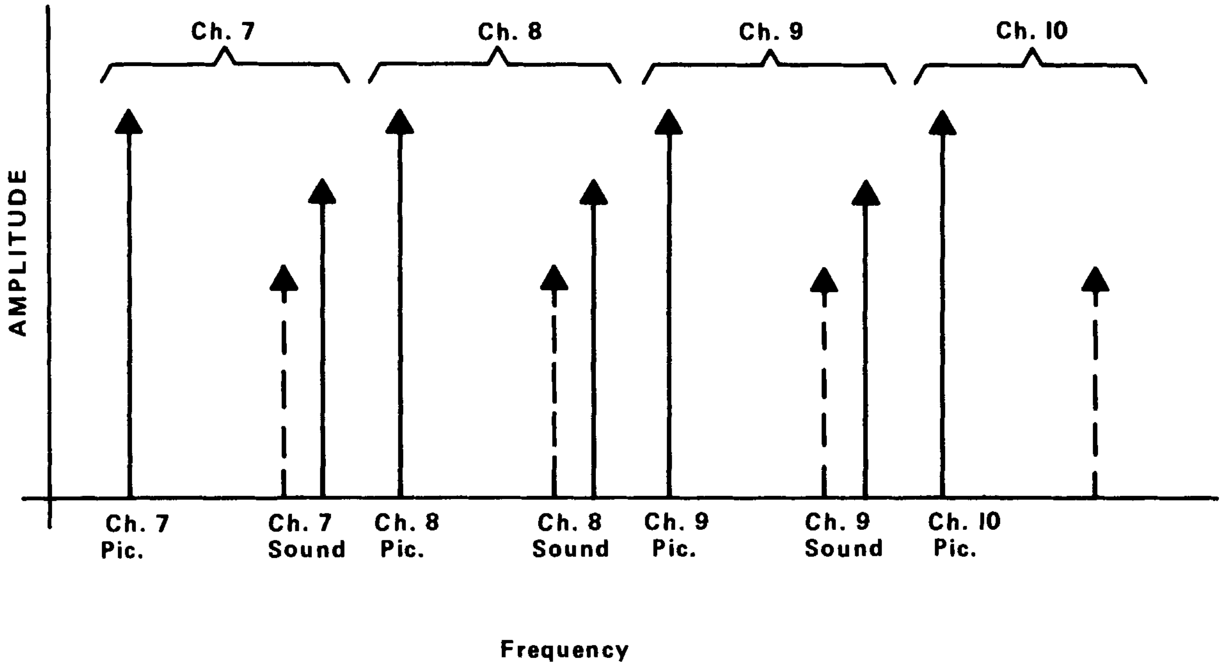


FIGURE 7

SIGNALS AT TELEVISION RECEIVER

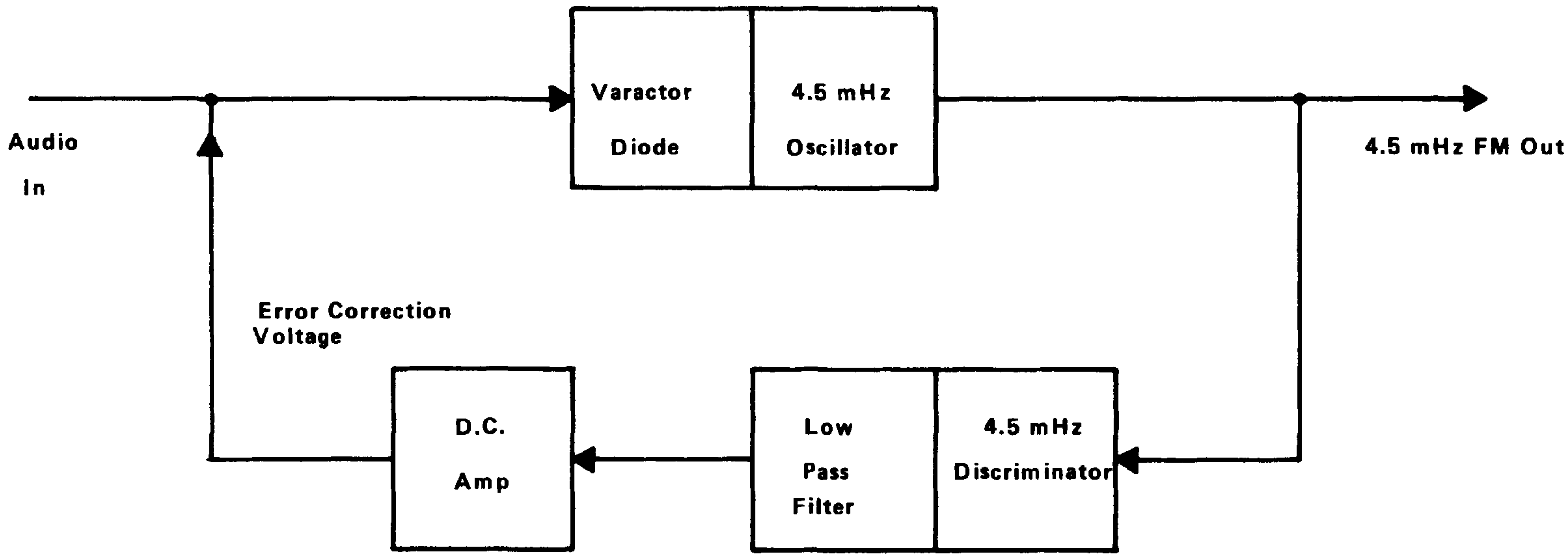


FIGURE 8

4.5 mHz SUBCARRIER CENTER FREQUENCY CONTROL

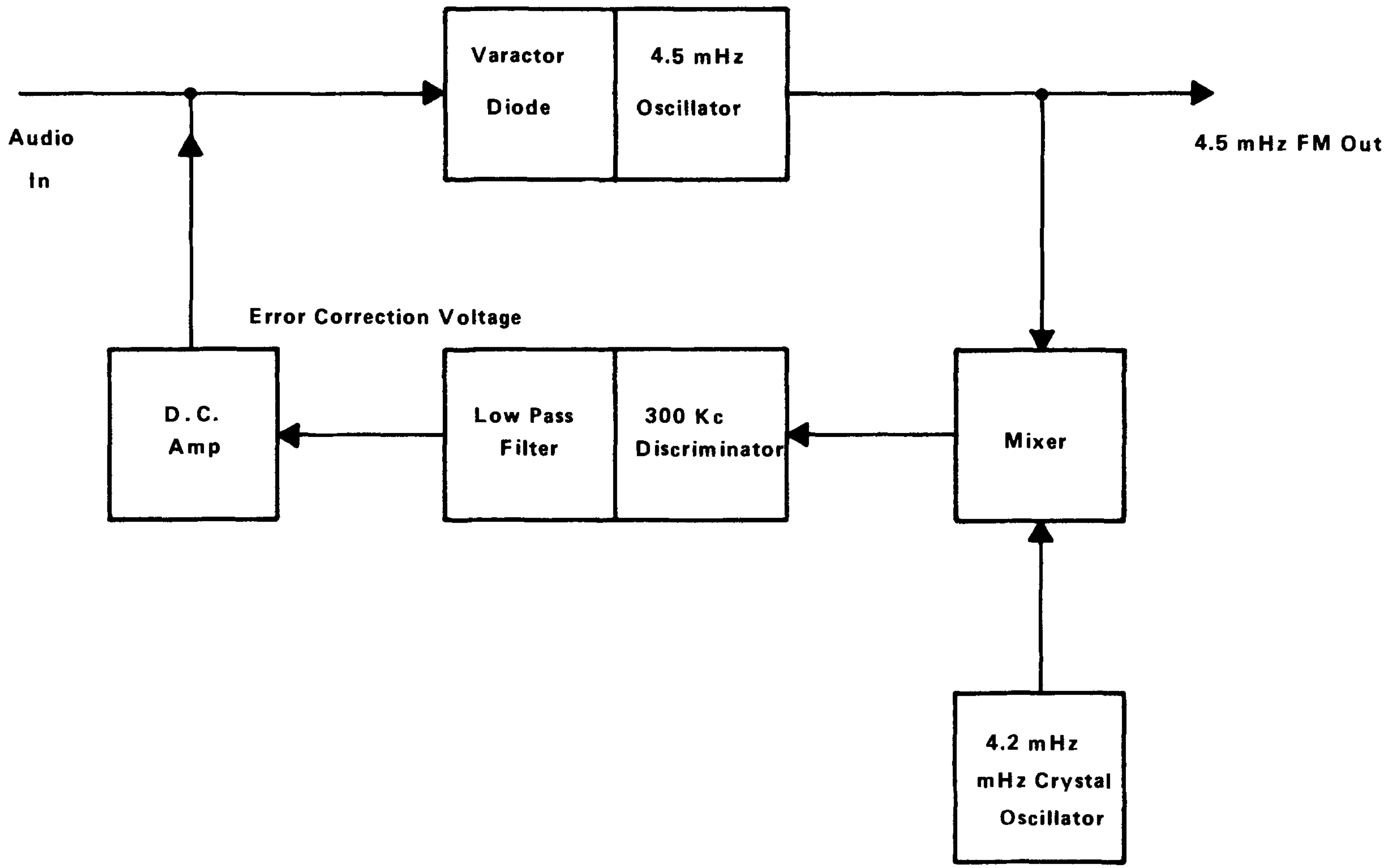


FIGURE 9

IMPROVED 4.5 mHz SUBCARRIER CENTER FREQUENCY CONTROL

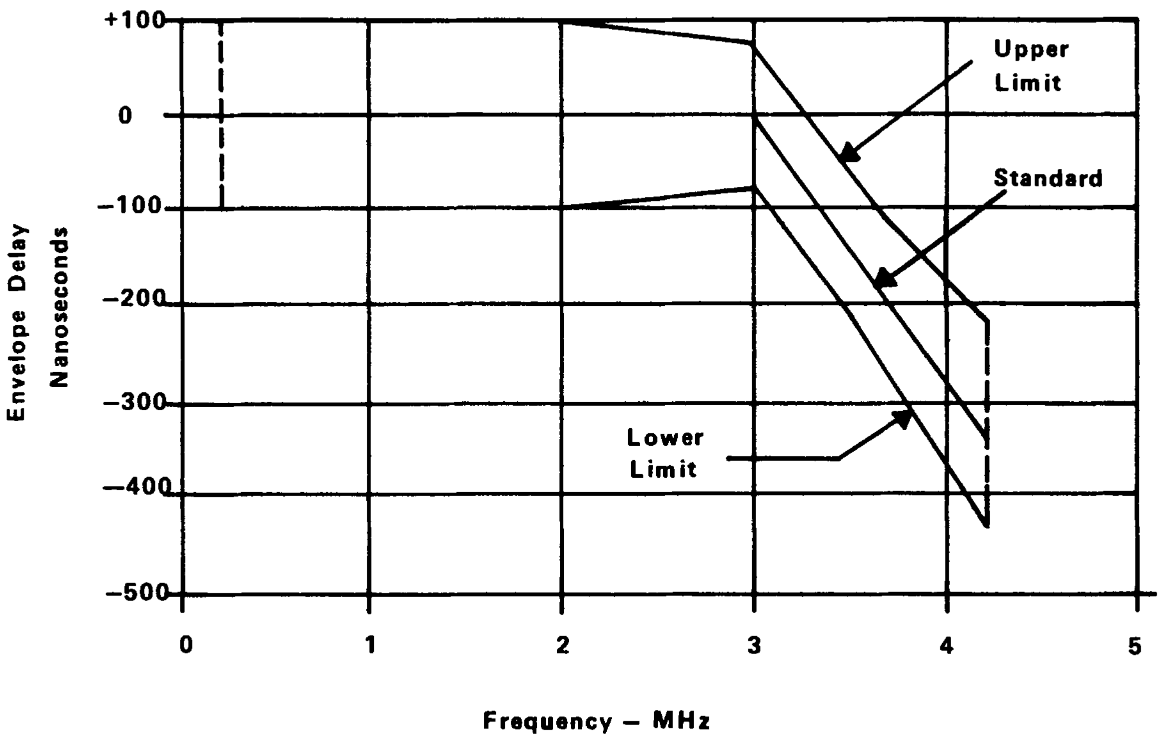
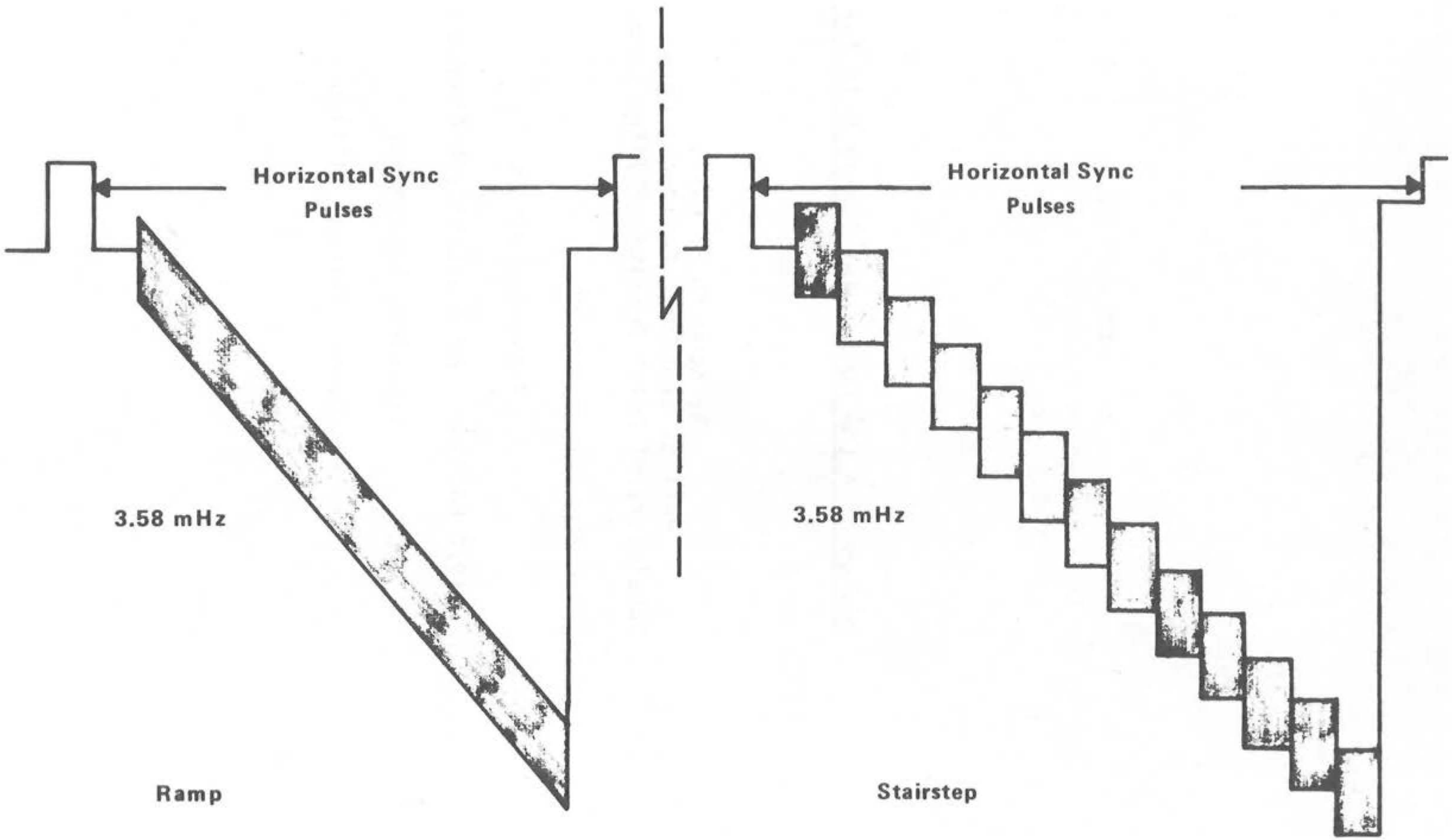


FIGURE 10

REQUIRED ENVELOPE DELAY PRE-DISTORTION AND TOLERANCES

FIGURE 11

DIFFERENTIAL GAIN AND PHASE TEST SIGNALS



"The Feasibility of Low-Cost FM Cablecasting"

by Richard L. Doering, Director
Community Music Programming/Radio 95
Box 1334, Riverton, Wyoming 82501

Abstract

The rapid growth of the FM radio over the past decade induced a young Wyoming professor to determine the feasibility of an automated high-quality low-cost FM radio station exclusively serving subscribers of the local cable TV system.

On November 2, 1968, a $3600 system designed by Mr. Doering began transmitting a wide selection of musical programs in stereo to Riverton cable subscribers. It has been operating 15 hours a day ever since, at 95.0 MHz FM.

Doering makes tapes at home from his own 2,000-album record library and sells advertising time to local sponsors at $25 a day (88 cents a minute). RADIO 95 has produced quality music programs at a mere fraction of the cost of operating an FCC-licensed on-the-air station.

The terms of Doering's lease with Community Television of Wyoming (the parent company), the type of equipment he uses, the programming material he has found most successful, audience surveys, copyright clearance, advertising and publicity methods, and other practical suggestions should be of more than passing interest.

A whole new field of broadcasting may open up to hi-fi hobbyists and cable systems as a result of Mr. Doering's unique experiences with RADIO 95.

THE FEASIBILITY OF LOW-COST FM CABLECASTING

by Richard L. Doering*, Director
Community Music Programming/Radio 95
Box 1334, Riverton, Wyoming 82501

1. Cable Radio: Wave of the Future?

Radio as a mass medium is not dying out. In fact, it may just be coming to life. Sky-high transmitters may disappear from the American landscape. But replacing them will be scores of FM modulators, owned and operated by independent programmers, "transmitting" over local CATV systems.

Anyone with a couple thousand dollars to gamble can lease an FM channel and set up his own cable radio station. RADIO 95 has shown that it can be done—quite successfully in fact. FM cablecasting is capable of revolutionizing the entire radio industry.

A fully automated cable radio station may consist of nothing more than one auto-reverse tape recorder and a simple $25 FM oscillator. Or it may utilize a $3,000 stereo generator and exciter built to FCC specifications.

Either way, the operating expenses are a mere fraction of what they would be for an on-the-air broadcast station. Programming for RADIO 95 costs only $4 a day (or 27 cents an hour). The sale of advertising time to local sponsors adequately covers this expense.

Use of the cable system to carry the oscillated RF signal makes a high-power transmitter unnecessary. In areas where more than 75 per cent of the population subscribes to CATV, cable radio can reach nearly as many people as an on-the-air station.

---

*Mr. Doering (pronounced "Deering") worked at radio station WSLN while attending Ohio Wesleyan University, where he received his B.A. degree in 1960. Two years later he was awarded a Master of Arts degree from Columbia University.

In 1967 he earned a Bachelor of Arts in Education from Kent State University. He has also studied broadcasting at the Cleveland Institute of Electronics.

While employed as an Instructor in Journalism and Industrial Psychology at Central Wyoming College in Riverton, Wyo., Mr. Doering established the CATV radio station which he describes in this paper.

This fall, in addition to being a self-employed music programmer, Mr. Doering will be authoring a book and studying for the Doctor of Philosophy degree at the University of California.

Cable FM is the perfect answer for communities too small to support an on-the-air station of their own. Ideally, every community—no matter how large or small—should have access to a dozen different music channels, two continuous news channels (local and national), and channels to air political debates, public board meetings, current issues, radio dramas, lectures, and educational subjects.

Programmers could specialize in gathering and editing one or two specific kinds of material. Copies of master program tapes could be distributed to CATV radio stations all over the country. Multi-channel FM modulators could be built to carry 10 or 15 taped programs simultaneously.

No longer would radio audiences be at the mercy of a handful of dominant stations outdoing each other to capture the "mind of the masses." No longer would good programs be scrapped because of their limited audience appeal.

Cable radio could satisfy a much wider range of listening tastes at nominal cost. The development of cable radio should be strongly encouraged by industry and government working together for the common purpose of better service to the radio public.

2. Purposes of Cable FM

Widespread FM cablecasting could successfully serve the following purposes:

(1) give listeners a broader selection of stations and programs to choose from
(2) upgrade current programming in general
(3) promote local business, educational, political and cultural interests

(4) improve radio service in small- and middle-market areas
(5) advance worthy causes such as cancer research, scholarship funding, religion in life, etc.
(6) revive public interest in good music and in the literary and dramatic arts

(7) communicate more effectively with alienated segments of our society
(8) bring radio closer to the individual, enabling him to explore social issues more intelligently, more responsibly, and more thoroughly with other people
(9) encourage more experimentation in low-cost high-quality communications technology

Our present communications system fails as a medium of meaningful exchange of thought (a) because of unnecessarily restrictive broadcast and licensing standards and (b) because of excessively high equipment and operating costs.

It is my belief that the FCC has discouraged public participation in radio by making it the province of big business. The principle of "local service" should be restored. Radio can be a very effective tool for interpersonal communication whenever people feel directly involved. Through CATV this involvement by everyone is possible.

3. Cable Radio and the Community

Riverton, Wyoming, is a fast-growing mining community of 10,000. RADIO 95 has been cablecasting in stereo over the Riverton CATV system since November, 1968, at 95.0 MHz (FM).

CATV serves 75 per cent of the community with six imported TV signals, plus the local weather channel. No FM signals are imported, although recorded music is simulcast on the Riverton TV weather channel and also at 92.0 MHz FM.

The tape deck for "Radio 92" is a twelve-year-old Ampex, and the monaural modulator is a Jerrold. The 10½-inch music tapes are supplied by a Los Angeles firm. All equipment for "Radio 92" is owned by the cable company.

RADIO 95, by contrast, produces its own music tapes and cablecasts in stereo. The CATV company owns no interest in RADIO 95, which draws its support exclusively from the sale of advertising time to local sponsors. Spot ads and announcements are programmed at 15- or 30-minute intervals on all RADIO 95 tapes.

Two AM stations are received in Riverton: KVOW (1000 watts) and KOVE (5000 watts, Lander). The nearest on-the-air FM station is in Casper, 120 miles eastward.

4. Transmitting and Receiving Equipment

Cablecasting is simply a matter of technology. Savings in high-priced transmitting equipment are passed along to our listeners in the form of better programming, since we have found that radio towers <u>are</u> dispensable.

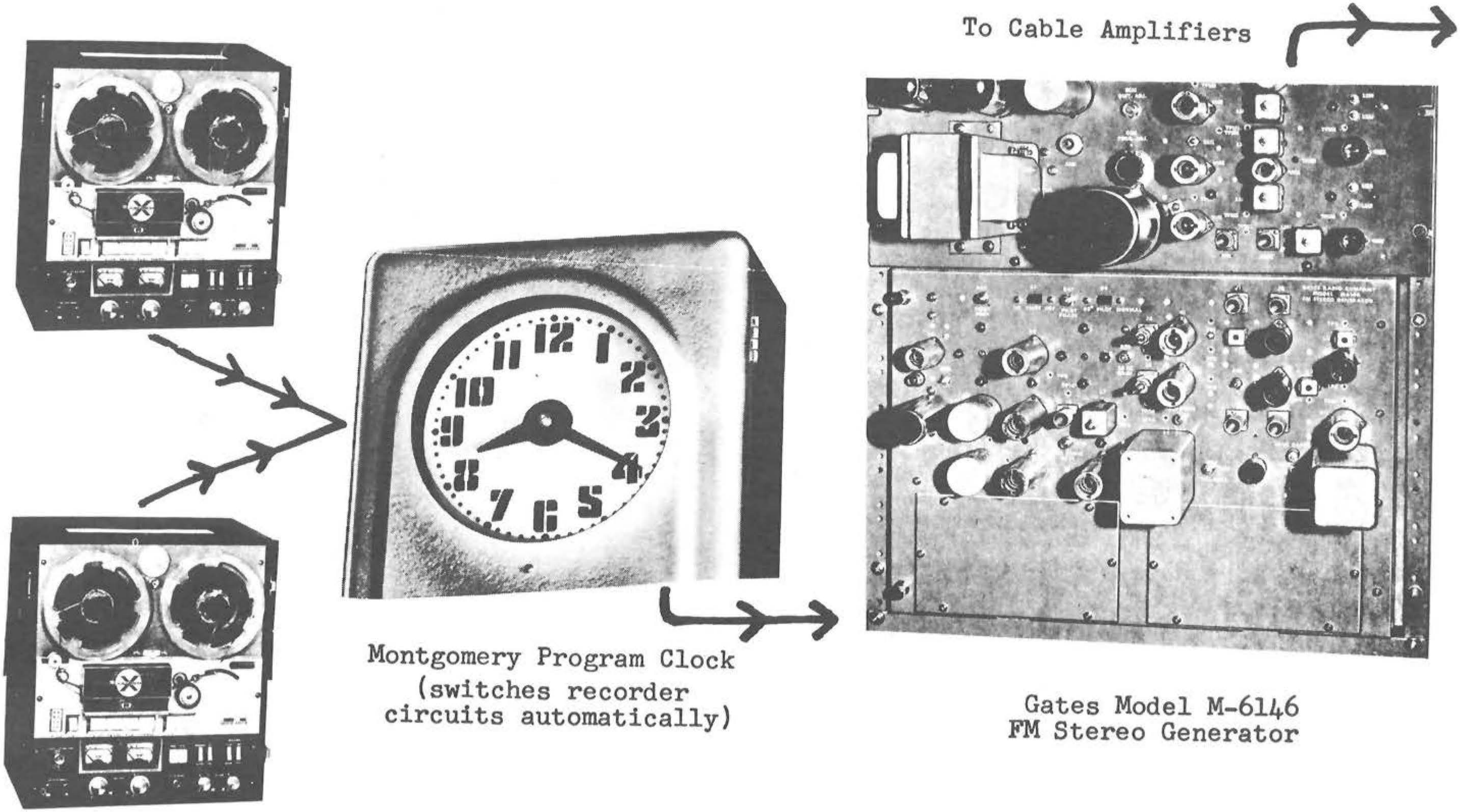
Total value of our system at RADIO 95 is estimated at $3,660. The system consists of two auto-reverse Roberts 400X tape recorders, a Montgomery program clock, a Gates 6146 stereo generator and 6095 FM exciter (see Figure 1).

We selected Roberts decks because of their superb fidelity at slow speeds (30-17,000 Hz at 3-3/4 IPS). The Gates transmitter is tuned to the FCC pre-emphasis curve and assures us broadcast-quality emissions. All transmitting equipment is rack mounted and occupies 6½ cubic feet of space.

Nearly all of our programming is taped. One 7-inch 2400-foot reel plays nearly four hours at 3-3/4 IPS. Live programming is

CABLE RADIO TRANSMITTING EQUIPMENT

Figure 1



CABLE RADIO TRANSMITTING EQUIPMENT

Figure 1

made possible simply by feeding a microphone line into the Record Amplifier of the 400X. The tape amplifiers deliver adequate direct input to the Gates equipment.

On the receiving end, listeners use radios ranging from pocket transistors to floor-model FM stereo consoles. The basic requirement is to tap off the cable TV antenna, which carries FM and TV signals together.

Many subscribers tap off their cable antenna by running 300-ohm twin lead-in wire from their TV set to the external antenna terminals of their FM radio. Telescoping FM antennas deliver satisfactory reception if placed within an inch or two of the cable antenna.

Subscribers who report unsatisfactory color TV reception ask the CATV company to install a separate FM tap for a one-time $5 charge. Black-and-white TV reception does not appear to be adversely affected by "do-it-yourself" FM hookups, which are by far the most popular and the least expensive.

Biggest bugaboos for cable radio stations are FM receivers with antennas built into the power cord. Manufacturers should be discouraged from making FM radios with no provision for an optional external antenna.

5. Channel Lease Agreement

There are three basic phases of cable radio:

(1) Programming, which involves the production of radio tapes

(2) Signal Modulation, the phase concerned with the conversion of taped program material into transmittable radio frequencies (RF)

(3) Signal Carriage, the normal CATV function of relaying RF from the headend to its destination in the listener's home

In its 1,270-word agreement with Community Television of Wyoming, Inc., RADIO 95 has assumed full responsibility for functions (1) and (2) above. The CATV system performs only function (3).

To distinguish phase (1) from phase (2) we have used the name COMMUNITY MUSIC PROGRAMMING (CMP) for the former and RADIO 95 for the latter. To date, CMP has been producing tapes only for RADIO 95. However, we are applying for copyright privileges to make tapes for other cable radio stations that want to adopt our format or use our music.

Table 1. RADIO 95 TRANSMITTING EQUIPMENT

Item	Current Market Value	Cost New (orig)	Year
1. 22-tube (10-watt) Gates M-6095 exciter, tuned and tested, 95.0 MHz	$1,000	$1,475	1967
2. 11-tube FM Stereo Generator, tuned/tested	800	1,695	1967
3. Roberts 400X Tape Recorder #1	700	800	1969
4. Roberts 400X Tape Recorder #2	700	800	1969
5. Montgomery Program Control Clock (specially designed)	300	350	1969
6. 100% Replacement Tube Set	70		
7. Matching Transformers	35		
8. Connecting Cables	30		
9. Incidental Supplies and Schedules	25		
Total Value, Transmitting Equipment	$3,660		

Here is how the three phases of cable radio may be taken on independently and yet function together:

1. The programmer (CMP) can supply:
(a) pre-recorded full-fidelity stereo tapes at $48 apiece (plus postage and copyright fees)'
(b) program scripts (no extra charge) showing the location and timing of each taped selection; the recording artist's name; and the locations where advertising spots of various lengths can be conveniently dubbed in (see Figure 2)
(c) recommended play schedules

2. The channel lessee (RADIO 95 or another):
(a) provides his own transmitting equipment (see Table 1 above)
(b) writes and sells advertising for his local sponsors
(c) dubs in advertising spots on tapes supplied by CMP
(d) may keep all profits earned from the sale of advertising

3. The CATV system:
(a) provides space for the lessee's transmitter
(b) provides electricity to operate the transmitter
(c) includes the lessee's channel in advertised listings
(d) carries the lessee's signal without charge to the lessee

Here is a précis of the channel lease agreement between CMP and Community Television of Wyoming (CATV). The agreement was drawn up by the CATV firm's legal counsel from a list of recommendations submitted by CMP:

A. Community Music Programming (CMP):

1) has the right to all gross income derived from the sale of advertising time

2) must provide its own casualty insurance for all equipment on CATV premises

3) assumes responsibility for the reproduction and transmission of copyrighted material

4) recognizes the regulatory powers of the FCC (or any other duly constituted public authority)

5) agrees to conform to general standards of conduct required of broadcasters by the NAB and the FCC

6) shall produce a radio signal compatible in quality with that produced by an FCC-licensed commercial FM station

7) shall maintain a complete program log of all broadcasts, available upon request to CATV

8) shall not broadcast signals over the air

B. Community Television of Wyoming, Inc. (CATV):

1) will carry CMP's FM signals over the Riverton cable system for the free use and enjoyment of CATV subscribers

2) will provide space to house CMP's transmitting equipment

3) shall not be party to a joint venture or partnership with CMP

4) reserves the right to request or demand that CMP cease operations over CATV's cable at any time, saving CATV harmless for any injury or damages CMP may sustain as a result

5) shall not regulate the content of CMP's programs

6. Programming Equipment, Schedules and Content

Equipment used in making tapes is located off CATV premises. Tapes are delivered to the headend for transmission only. This is our current inventory of programming equipment:

a) 2,115-album record library (book value)	$10,194.30
b) 1 Scott 382-B stereo amplifier/tuner	300.00
c) 2 Roberts 1740X tape decks	600.00
d) 1 Garrard SLX-3 turntable	100.00
e) 1 Pickering XV15/750E cartridge	60.00
f) 1 Shure 550S microphone	55.00
g) 1 Superex stereo headphone set	60.00
Total Value, Programming Equipment	$11,369.30

We have a stock of 51 master tapes. Only the twelve most recent of these are in current use. At 3:30 p.m. each day a new 3-hour 45-minute tape automatically switches into play. The tape runs two complete cycles between 3:30 and 11 p.m. The station is off the air from 11 p.m. until 8 a.m. the next morning. At 8 a.m. the same tape resumes play, and completes two more 3-hr 45-min cycles before a new tape comes on at 3:30.

Figure 2. SAMPLE TAPE SCRIPT

30 min.: Pick of the Pops (I)
Baby It's Cold Outside - Steve Lawrence & Edyie Gormé
Blue Tail Fly - Johnny Mann
Live for Life - Tony Bennett
Tiny Bubbles - Connie Francis
Moon River - Steve Lawrence

:38 FNB Bank now pays 5%

A Man and a Woman - Ed Sullivan Singers
Shadow of your Smile - Jerry Vale
Georgy Girl - Matt Monro
Paper Doll - Mills Brothers
Alfie - Barbra Streisand

:38 BRT Rug cleaning discount

30 min.: Best of Country/Western
Mountain Dew - Nashville Brass
Sweetheart of the Year - Ray Price
I Love You Because - Don Gibson & Dottie West
Orange Blossom Special - J. Cash
Born to Be with You - Sonny James

:60 RLM Wood paneling for home

Heartaches by the Number - Floyd Cramer
I'm Falling Too - Skeeter Davis
Walkin' in Loveland - Eddy Arnold
Goodbye Old Shep - Red Foley
Oh Lonesome Me - Chet Atkins
Gentle on my Mind - Glen Campbell

:22 ACC Cancer march

45 min.: Easy Listening

Instrumental music by Mantovani, André Kostelanetz, Percy Faith, Peter Nero, Tony Mottola, Guy Lombardo, Wayne King, Lawrence Welk, and so forth

30 min.: Pick of the Pops (II)
Hey There - Sammy Davis Jr.
Swingin' on a Star - Anita Kerr
Thank Heaven for Little Girls - Jack Jones
Real Live Girl - Robert Goulet

:38 TRG Polyoptic lamps

(continued - next column)

Try a Little Tenderness - Frank Sinatra
Help Yourself - Tom Jones
Valley of the Dolls - Andy Williams
Raindrops Keep Fallin' on my Head - Bill Black Combo
Buttons and Bows - Ray Conniff

:38 MRG Margate's Grand opening

15 min.: Special Program

Dixieland, honky-tonk, Gay 90's, barbershop, Beatles, religious, Andy Williams, Victor Herbert, Stephen Foster; organ, moog, polka music; Songs of the Trail; Music from Around the World (Hawaii, Germany, Brazil, Switzerland, etc.)

:60 LFS Lounge chair close-out

15 min.: Country and Western
Dream - Everly Brothers
Stand By Your Man - Tammy Wynette
Welcome to my World - Jim Reeves
Cotton Fields - Charlie Pride
Tumblin' Tumbleweed - Living Voices

:22 RJC Congrats to Best Teacher

30 min.: Best of Jazz
String of Pearls - Ted Heath
Up Up & Away - Brass Ring
Make Me a Present of You - Nancy Wilson
On Green Dolphin Street - Johnny Lytle
Corcovado - Charlie Byrd
Wade in the Water - Ramsey Lewis

:38 GBL New Chevvys have arrived

Up in Erroll's Room - Erroll Garner
Fool on the Hill - Sergio Mendes
Gone with the Wind - Dave Brubeck
Bumpin' - Wes Montgomery

:38 JWL Jade pins on sale

30 min.: Best of the Classics
Tales from Vienna Woods - Boston Pops
Humoresque - Isaac Stern
Old Folks at Home - Mormon Tabern. Ch.
Andante Cantabile - Philadelphia Or.

Here is a sample schedule to illustrate:

	Day #1	Day #2	Day #3
8:00 a.m.	Tape #42 27th cycle	Tape #48 3rd cycle	Tape #37 39th cycle
11:45 a.m.	Tape #42 28th cycle	Tape #48 4th cycle	Tape #37 40th cycle
3:30 p.m.	Tape #48 1st cycle	Tape #37 37th cycle	Tape #46 13th cycle
7:15 p.m.	Tape #48 2nd cycle	Tape #37 38th cycle	Tape #46 14th cycle
11:00 p.m.	Sign Off	Sign Off	Sign Off

Each tape runs 48 cycles over a six-month period before it is discarded. A master schedule varies the number of days between playings of the same tape. New tapes are played 10 to 12 days apart; old tapes, 23-25 days. The average (mean) interval is 17.2 days.

Since RADIO 95 is the only commercial FM station in our area, we feel committed to satisfy all of our listeners' musical tastes — except hard rock, which is readily available on local AM stations. Each of our tapes contains the following musical programs (order will rotate):

Pick of the Pops (vocal)	30 min.
Country and Western	30 min.
Easy Listening (instrumental)	45 min.
Pick of the Pops (vocal)	30 min.
Special Program	15 min.
Country and Western	15 min.
The Best of Jazz	30 min.
The Best of the Classics	30 min.
	3 hour 45 minute total

A balance is maintained between vocals and instrumentals. Since a number of offices, banks and stores play RADIO 95 for background music, we avoid pieces that could grate on somebody's ears. We like catchy, familiar tunes that everybody can hum along with.

Fourteen minutes out of every fifteen is uninterrupted music. We stick pretty much to the "Greatest Hits" of the greatest artists in each category. In order not to duplicate the efforts of AM stations, RADIO 95 has no wire news service.

Stereo versions of hits from the 1890's-1940's link the present with America's great musical past. The over-30 crowd, severely alienated by contemporary radio most everywhere today, unilaterally endorses RADIO 95.

The 2,000 albums in our record library have been hand-picked to represent "the world's most enjoyable music" (again see Figure 2).

7. Copyright Clearance

Before beginning our cablecasts in late 1968, we went to great pains in order to obtain official copyright clearance from the three major agencies--ASCAP, BMI and SESAC. We were told that there was no contract form for a cable radio station, since the copyright issue had yet to be settled by the courts.

We submitted a 14-page outline of our proposed cable operation to each of the three agencies. BMI was the only one that responded to our request: we could "go ahead and cablecast without fear of copyright infringement," they told us.

When the copyright issue is eventually settled, we expect to be contacted about signing some sort of agreement. Something on the order of the standard radio station contract would not seem unreasonable. However, any proposal must consider the fact that we are an ultra-low-budget operation, with gross annual receipts under $2,000. Excessive or unreasonable minimum copyright fees could easily sink an overladen ship.

If we ever market any of our tapes commercially, the standard copyright fees for that type of enterprise will of course apply.

RADIO 95 has done everything possible to promote the music industry. Our Special Programs often highlight a particular artist or label. On our regular programs we announce the names of all artists. We continually remind listeners to patronize local music dealers. We urge everyone to buy records and tapes. We are willing to hunt up any record in our library whose number a dealer cannot find in ordering for a customer.

Last year Riverton's largest music store reported sharply higher sales than the year before. The owner attributed this increase largely to RADIO 95's role in promoting music locally.

8. Operating Expenses (monthly)

New tapes (2)		$100.
Blank tape	20.00	
Records, Labor	80.00	
Equipment maintenance...		30.
Newspaper advertising...		20.
Directory advertising...		10.
Radio promotions........		10.
Total Estimated Monthly Operating Expenses:		$170.

Note: These figures do not take into account the cost of soliciting advertising or preparing ad copy. These expenses may vary widely.

9. Advertising

Advertising is the only source of financial support for RADIO 95. Our tapes are programmed with "slots" for ads to be dubbed in at quarter-hour intervals. During a full broadcast day (15 hours) we can carry a maximum of 30 minutes of paid advertising. Spots run in 15, 30, and 60 second lengths.

We sell our time for 75-85 per cent less than our AM competitors. Even at this modest rate we can support a full day's programming with only 12 per cent as much total advertising as an on-the-air station. A 30-minute package of ads costs our sponsor $25 — approximately 44 cents per half minute.

Advertising slots which have not been sold are filled with public service announcements provided by the Advertising Council, the Cancer Society, March of Dimes, etc.

If all of our advertising slots were filled with paid ads, we would take in $25 a day. Our operating expenses, you recall, have been estimated at only $4-5 a day, exclusive of ad writing and production.

Billing is a simple process of adding up the total amount of advertising time from the monthly program logs. We make no extra charge for ads that we run over the contracted minimum. Advertisers always get a little more time than they expect.

Until a cable radio station captures a sizeable audience, the most frequent sponsors will probably be large community-minded organizations like banks, airlines, utilities, etc.

10. Publicity

Fortunately, Riverton is small enough so that publicizing our station is neither difficult nor expensive. When we first began cablecasting, we ran half-page display ads in the local daily newspaper. Radio-TV stores absorbed some of this cost by acting as co-sponsors, since they stood to benefit by selling FM radios.

Twice a week we run a 3-inch column announcing our program highlights (see Figure 3).

In order to build our audience to its present size, we have used direct mail, telephone, and door-to-door methods. Local merchants have cooperated in giving our listeners free merchandise or discount vouchers in exchange for radio advertising time. Many listeners have won these prizes by phoning the station in response to spot give-away contests ("third caller wins...").

A cable radio station cannot achieve an official audience rating unless pollsters (1) acknowledge the station's existence and (2) list it in their area roster. RADIO 95 is not recognized as a broadcast station by the editors of Broadcasting Yearbook, which lists only licensed on-the-air stations.

RADIO 95 PROGRAM NOTES Published in The RIVERTON RANGER
(Figure 3)

ON RADIO 95

Music from Norway

"Songs from the Norwegian Fjords" are among the selections to be heard tomorrow afternoon in a special program of "Music From Around The World."

Norway's favorite vocal ensemble, the Meloditersitten" sings three authentic ballads.

"Per Spelmann" is the song of a farmer who trades his only cow for a fiddle.

The French portion features selections from the original movie soundtrack of "The Umbrellas of Cherbourg."

Student drinking songs highlight the German music portion.

Records for the program were loaned to Radio 95 by Lee F. Olson, Bill Reichert and Mel Moen of Riverton.

Radio 95 may be heard daily from 8 a.m. to 11 p.m. on any FM radio with cable TV antenna attached. — Adv.

ON RADIO 95

Country Music Bows-In

"The Best of Country and Western Music" will soon be a regular program on Riverton's FM stereo music station, Radio 95.

Popular, jazz and classical music have been programmed on a regular basis since December.

Listeners seem to like Country and Western almost as much as Instrumental and Semi-Classical, a recent survey showed.

The changeover to country music will be gradual. Eddy Arnold, Jim Reeves, and Glen Campbell will be among the top featured country artists.

Each musical program is 30 minutes long. Programs rotate throughout the broadcast day from 8 a.m. to 11 p.m.

Radio 95 can be heard on any FM radio with a cable TV antenna attached.—Adv.

ON RADIO 95

Gilbert and Sullivan Show

Saturday morning will be the time for stereo bugs to adust their sets as Radio 95 presents "A Demonstration of Stereo," with channel balancing, phasing and frequency response tests.

Remember the Gilbert and Sullivan community concert? Tomorrow afternoon's program features highlights from five G&S operettas.

Like old-fashioned barbershop ballads? "Music from the Good Old Days," starring the 1966 International Barbershop Chorus winners, runs tomorrow.

Sunday Tchaikovsky's world-famous Symphony No. 6 ("Pathetique") will be played by the Philadelphia Orchestra.

It's all on FM Radio 95. For details phone 856-9500.—ADV.

ON RADIO 95

Jazz Harp and Jazz Mass

The artistry of Dorothy Ashby, queen of the jazz harp, may be heard throughout the day tomorrow on Radio 95.

Miss Ashby, a native of Detroit, was the first harpist ever to be chosen for the Down Beat International Jazz Critics Poll.

She achieves a rare interplay between bass and drums in "Essence of Sapphire," which sounds incredibly realistic on larger stereo sets due to the technical excellence of the recording.

The Roman Catholic Church recently approved jazz liturgy, and Joe Masters' rendition of the "Credo" from the "Jazz Mass" is a rare musical delight.

That's what's happening tomorrow on Radio 95, FM stereo on the Riverton cable. — ADV.

ON RADIO 95

FM Radios For $12.99?

Just what does it cost to hook up to Radio 95? Ten downtown stores have the answer.

The lowest - priced FM radios in town are at Woolworth's, where $12.99 buys a 12-transistor AM-FM battery model.

To hook your radio up to a cable TV antenna, wire from Ace Hardware is the best buy at 3 cents a foot.

Table model AM-FM radios begin at $15.95 at Pioneer Drug, $25.95 at Stylcraft (RCA), and $21.95 at Electronic-Music Center (Admiral), where antenna couplers are now only $2.75.

Floor model combos are the best value at Modern Appliance Center, while Gambles sells tape player combos for $189.95, another good buy.

Now is the time to hook up to Radio 95. Values are tops. — ADV.

RADIO 95 cannot claim the largest share of the audience in Riverton, but we do know that our format appeals strongly to that portion with the highest per capita buying power (businessmen, teachers, professional people, etc.). If a man owns a high-priced FM stereo receiver, he'll use it — as long as there's something to listen to.

We take our listeners quite seriously. We urge them to call anytime and tell us how they like the music. Surveys are taken periodically to help us evaluate our programming. The compliments consistently outweigh the complaints by a wide margin.

Studies of ours show that well over 90 per cent of Riverton's residents have heard about RADIO 95. Apparently more of them would listen if they knew more about (1) what is meant by an "FM" radio and (2) how to make the radio-to-cable antenna connection.

Our contribution to the CATV business here is evident, also. We have kept a number of potential disconnects on cable during the TV-impoverished summer months. Many new CATV subscribers are downtown stores that want RADIO 95 for background music.

RADIO 95's strides during the past 18 months in a small Rocky Mountain community forecast a dynamic future for cable radio elsewhere throughout the nation.

THE USE OF INTEGRATED CIRCUITS IN CABLE TELEVISION

Larry F. Roeshot
The National Cable Television Center
The Pennsylvania State University
University Park, Pennsylvania

Cable television systems began with single channel amplifiers operating in a low VHF band. As the demand for cable television grew, along with technology, amplifiers were designed to cover more than two octaves and performed several additional functions, including automatic slope and gain controls.

The use of integrated circuits to satisfy some of these requirements has been limited in modern cable television amplifiers. This paper examines some of the present-day and future uses of IC's in cable television equipment in view of cost, performance, and reliability.

The development of cable television system techniques--such as the method of ALC control, tilt mode, operating level, gain, etc., together with the time consuming proof-of-feasibility for each approach--has been too dynamic to warrant any costly IC designs.

This limitation has been particularly true at VHF frequencies with physically large components which are not compatible with thin film IC's. In addition, the operating levels normally encountered in cable television amplifier output stages are beyond the present state of the art for IC's.

To determine the feasibility of thick films at VHF, a thick film broadband amplifier is shown. Using this technique, components can be trimmed to precise tolerances on low cost replaceable modules.

With the flexibility of thick film hybrid integrated circuits, utilization of these devices in cable television equipment can be expected within the next few years.

APPLICATION OF INTEGRATED CIRCUITS IN CABLE TELEVISION SYSTEMS

INTRODUCTION

The cable television industry has had a fascinating history. Cable television technology was not the product of a large electronic R and D effort or government sponsored research program. Rather, it grew out of a need and countless novel approaches, many of which produced surprising results. Fifteen years ago, few people took cable television seriously. After a 25% annual growth over the past decade together with the EIA,[1] FCC, and Justice Department favorable attitudes during 1969, cable television is becoming a major communications media. As a significant electronics industry considerable equipment refinements can be expected. Along with increased demand and competition, new functional and reliability requirements will introduce new technology including the use of integrated circuits.

It is the purpose of this discussion to examine some of cable television circuit requirements in view of hybrid integrated circuit technology, and to project some future uses of IC's for both RF amplification and low frequency control functions.

BROADBAND AMPLIFIERS

Most of the amplifiers used in early cable television systems were "broadband" in the sense that they covered a 6 MHZ bandwidth. They served to amplify the signal at the same frequency--just enough to overcome cable losses. Early systems had very few amplifiers in cascade. With nominal requirements the amplifiers did not require significantly greater dynamic range than that of the receiver front end.

As systems developed, it became necessary to carry several channels requiring broadband amplifiers. Amplifiers in the low TV band were built to accommodate up to three non-adjacent channels. Distant stations, many of which were on adjacent channels, in addition to high band channels converted for the low band system, soon filled all the available channels on the cable. This introduced a new problem for the system--that of adjacent channel interference. Television receivers, not designed for adjacent channel reception, did not have adequate selectivity to reject the next lower channel sound carrier. Reducing the sound carrier at the strip amplifiers or converters by 15 to 20 DB corrected the problem. Thus the major effort continued to be developing means to carry more channels over greater distances.

---

[1]FCC Docket 18397, Part V, Industrial Electronics Division, Electronic Industries Association, Washington, D.C., October 1969.

In order to achieve wide bandwidths several types of circuitry were used among which was distributed amplification.[2] The technique increased the gain bandwidth product of the tubes in addition to increasing the amplifier reliability. Since tubes in the distributed amplifiers were operated effectively in parallel rather than cascade, the failure of one tube merely resulted in reduced gain and not complete failure.

Transistor distributed amplifiers could not realize the full potential of the technique. The loading effect of the transistors on the input transmission line, limited designs to three transistor stages for the VHF band.[3] Although 3 DB more output power was obtained by this method, it was not enough to handle the power levels normally encountered in cable systems. Operation over +48 DBMV (about 1 MW) drove the output stages into the non-linear region which resulted in distortion. The effects of temperature on gain variations added to the design problem.

In spite of the initial shortcomings, transistor amplifiers had the potential to reduce the current requirements for cable powered amplifiers and to reduce the physical size to permit mounting on the messenger cable. The possibilities of transistor amplifiers for cable television encouraged development.

A variation of the distributed amplifier resulted in the distributed pair configuration[4] see Figure 1. The dual amplifying sections eliminated the output line reverse termination which resulted in higher output power than that of a conventional distributed amplifier. The circuit is essentially two stages in parallel, except that the input impedance remained nearly constant, about 85 ohms, over the entire VHF range.

Feedback was used around several stages to obtain gain over the 50 to 500 MHZ range. From the amplifier shown in Figure 2 it can be observed that no tuning capacitors were used. Tuning was accomplished by adjusting the air wound coils. The four-stage unit amplifiers provide 24 DB gain with 1 DB ripple and was stable from -20°c to +70°c.

---

[2] Ginzton, Hewlett, Jasberg, Noe, "Distributed Amplification," Proceedings of the IRE, Vol. 36, pp. 956-969, August 1948.

[3] L. F. Roeshot, "Transistor Distributed Amplifiers," Masters Thesis, The Pennsylvania State University, February 1960.

[4] L. F. Roeshot, "Distributed Pair Amplification," Electronic Design News, January 1963.

The gain bandwidth product of available transistors required circuit novelty to achieve reasonable gain over two octaves. Both the distributed amplifier and the distributed pair achieve broadband gain up to the f_{max} rating. As transistor gain bandwidth products increased, circuits were simplified. Maximum gain in the simpler circuits was traded for ease of alignment, repeatability and tolerance for transistor parameter variations.

Present day cable television amplifiers use variations of the circuit shown in Figure 3. In this circuit the collector-to-base feedback compensates for flat frequency response. The output power using this technique is greater at high frequencies than at low frequencies because of the mismatch.

In normal operation, amplifiers in a cable system are operated well below their maximum output levels. As amplifiers are cascaded, the third order distortion builds up in proportion to the number of amplifiers (10 log N).

System cross modulation is required to be at least -46 DB below the signal. For a single amplifier then, the cross modulation must be significantly lower. Cross modulation is reduced by using emitter degeneration. The negative feedback is made to vary with frequency by using several sections as shown in Figure 3. Using this scheme, greater emitter degeneration is obtained at low frequencies, where cross modulation improvement is needed most. The low Q adjustment of the resistor-capacitor combination is predictable and aids in obtaining positive slope response.

THE USE OF INTEGRATED CIRCUITS

Integrated circuits are not used in CATV amplifier designs except for ALC (Automatic Level Control) DC amplifiers. Integrated circuits for RF frequencies have been slow in development for a number of reasons. Standards are not yet firm. The many philosophical approaches in design, including the method of gain and slope control, operating level, and non-standard band usage has resulted in a number of design changes and field modifications. The power level used in broadband cable television amplifiers requires special design considerations. RF integrated circuits need closer tolerances than 10 and 20%, common in monolithic logic circuits. Typically, a cable television amplifier must control gain variations to less than $\pm$.25 DB in an outdoor environment.

The VHF frequency range has been a transition range for integrated circuits. At low frequencies, monolithic IC's are designed as resistance coupled circuits. In the microwave region, short wave lengths permit the use of strip line techniques for circuit functions. At VHF, neither technique is efficient. The lack of suitable broadband inductors has especially limited the use of IC's at VHF.

THICK FILM HYBRID IC'S

Thick film hybrid IC's have found wide acceptance where external inductors or other precision components must be added to the circuit. Costs are comparable to those of printed circuits, and in some instances can be less. RF circuits on PC boards often require additional components to compensate for stray effects. By using smaller geometries on a low loss substrate stray effects can be reduced which results in simpler circuits.

With present thick film technology, resistors can be screened and fired to an accuracy of about ±10 per cent. Abrasion timing can be used for ±1 per cent, or even ±.1 per cent accuracy. The temperature co-efficient of screened resistors ranges from 200 PPM/°c to 50 PPM/°c. Both resistors and capacitors can be screened or discrete components can be mixed in thick film circuits.

Discrete capacitators, available in a wide range of values, temperature co-efficients, and physical configurations, lend considerable versatility to the VHF hybrid circuits.[5]

Screened as well as thin film spiral inductors have been limited to values of a few tenths of a microhenry. For larger values, the distributed capacitance of spiral inductors results in self resonant frequencies of about 100 MHZ. Several manufacturers produce microminiature coil inductors suitable for attachment to thick film circuits. Typical inductors or .82 microhenry are .175 inches long with self resonant frequencies of 300 MHZ. Slightly larger sizes are adjustable over a two to one inductance range.

The behavior of an inductor over a broad frequency range is largely dependent on the type of fabrication. Self resonant frequency (SRF) is reached when the inductance resonates with the distributed capacitance. For spiral inductors the distributed capacitance is emperically determined and depends on the line width, the number of turns and the method of processing. For thin film spirals, this capacitance is typically .25 pf for four turns. For inductances greater that 100 uh, (needed for VHF), inductance increases about 50 uh per turn while the distributed capacitance remains relatively constant. The increased inductance lowers the SRF which degrades the high frequency performance of the coil. The equivalent inductance of the coil is a function of frequency is given by:

$$L_{equiv.} = \frac{L}{(1-K)^2}$$

where

$$K = \frac{f}{SRF}$$

[5]D. W. Hamer, "Reduced Titanate Capacitor Chip for Thick Film Hybrid IC's," International Hybrid Microelectronics Symposium, pp. 256-264, October 1968.

and the equivalent series resistance is

$$R_{equiv.} = \frac{R}{(1-K^2)^2}$$

For a spiral inductor with a SRF of 300 MHZ the inductance doubles at the high end of the VHF TV band, while the resistance increases four times. These effects are desirable in a circuit when used for shunt compensation. However, when uniform inductance is necessary over a broad band, inductors with resonant frequencies of at least four times higher than the upper frequency should be chosen.

The economies of monolithic IC and thin film circuits are best realized when the size of the circuit is reduced as much as possible. For VHF frequencies, inductors limit the reduction in size. The inductance per unit area of microminiature coils is more than ten times that of spiral inductors.

A BROADBAND THICK FILM AMPLIFIER

A broadband thick film circuit has been fabricated as shown in Figure 4. The schematic of the two-stage circuit is shown in Figure 5. Emitter degeneration was used on the first stage, while the second stage consists of collector-to-base feedback along with a tapped emitter resistor. The proper tap is selected during the bonding process of micro-soldering operation. Additional resistor adjustments are made provided by heat treating, abrasive trimming or localized heating. All of the components except the transistors and feedback inductor were screened components.

The capacitors, with values of about 470 pf, require a large amount of real estate and could be more efficiently replaced by multi-layer chip capacitors. However, the main purpose of the experiment was to screen as many components as possible. In an effort to reduce pinhole shorts the dielectric was screened twice. The resulting thickness decreases the capacitance per unit area.

A computer was used to draw the screen patterns. The four patterns, 1) first metalization, 2) dielectric, 3) resistors, and 4) second metalization, were drawn separately for independent modification. Circuit revisions are accomplished by specifying a coordinate change on a data card. After debugging the program the total drawing time was 48 minutes for the four patterns.

The frequency response of the amplifier is flat up to about 200 MHZ. Beyond this frequency, the response is highly dependent on the ground plane. These results were sufficient to evaluate the preliminary design, since the final package will largely influence the high frequency response. The cost for the basic thick film circuit without testing or packaging has been estimated to be under three dollars.

Hybrid IC modules permit amplifiers size reduction limited by the connectors. Figure 6 compares models of a conventional bridging amplifier and an IC modular unit.

A layout model for a trunk amplifier is shown in Figure 7. The four rectangular plug-in modules depict the equalizer, amplifier, power amplifier and power supply. By substituting the power amplifier module with a matching network, the unit becomes a low power line amplifier. Amplifier functions are illustrated in Figure 8.

A 7.5 watt broadband stud mounted amplifier has been developed by RCA.[6] The frequency range of 265 to 400 MHZ is just beyond the required range, but it represents the type of throw-away high power module needed for cable television applications.

FUTURE USE OF HYBRID CIRCUITS

Several television receiver manufacturers have adopted thick film technology. Oak Manufacturing has developed a TV tuner with a plug-in module using thick films.[7] This method was selected over a printed circuit or a monolithic approach on the basis of cost, performance and circuit optimization. Zenith has developed a hybrid thick film chroma demodulator.[8] RCA has announced that by 1973 color sets will be made up of seven modules.[9] The development of thick film technology for the amplification and processing of television signals is certain to have an influence on the cable television industry.

Cable television systems of the future, more clearly described as a BCN (Broadband Communication Network)[10] will perform a number of other services in addition to that of television. Many of these services will require bi-directional communications.

---

[6] W. E. Poole, "UHF Integrated Power Amplifiers," IEEE International Solid State Circuits Conference, Philadelphia, Pa., February 1969.

[7] K. S. Williams, "A Plug-In Thick Film Hybrid Circuit Module," Hybrid Microelectronics Symposium, pp. 497-505, October 1968.

[8] C. M. Engel, et. al, "A Hybrid Thick Film Chroma Demodulator and Color Difference Amplifier," Hybrid Microelectronics Symposium, pp. 487-496 October 1968.

[9] R. Sarnoff, "Bob Sarnoff Runs a New Game," Business Week, p. 89, January 1970.

[10] EIA, Op. Cit., Reference 1.

The RF spectrum used on the cable system has been proposed to be divided as shown in Table 1.[11] The simplified functional diagram of the amplifying unit is shown in Figure 8.

Proposed Frequency Plan

Frequency Allocation	Use	Channels TV
60 kHz to 5 MHz	Wideband Data & Trunk Carrier Two-Way	
5.5 to 48 MHz	Customer to Distribution Point One-Way	6 to 7
54 to 88 MHz	Distribution Point to Customers' CATV/ITV--One-Way	5
88 to 108 MHz	Distribution Point to Customers' FM plus	50 FM Channels
Music--One-Way		
112 to 170 MHz	Distribution Point to Customers' Additional TV or Other Private Line Services--One-Way	At least 2 TV Channels
174 to 216 MHz	Distribution Point to Customers' CATV/ITV--One-Way	7
220 to 270 MHz	Distribution Point to Customers' Additional TV or Other Private Line Service--One-Way	8

In future systems, the circuitry required for data conversion, telemetery, coding and performance monitoring will far overshadow the amplifier circuits. Functional modules provide the greatest flexibility for modification on field repair.

---

[11] J. O. Norback, "Modern Concepts for Cable Transmission Systems," Communication News, February 1970.

CONCLUSION

Cable television amplifiers evolved within an industry having an uncertain future. The practicality of cascading many broadband amplifiers out of doors, while preserving picture fidelity had to be proven. Having solved this technical problem, cable system designs in the future can be expected to grow more complicated with increased reliability demands.

Replacement modules have become standard in the industry. The next consideration is to make wider use of integrated circuits within the modules. An attempt was made here to describe some of the advantages and limitations of using hybrid circuits for VHF broadband amplifiers. Some details of an experimental thick film VHF amplifier were given together with a mock up to illustrate adaptation for cable television. More development is needed to include other circuit functions within the modules.

In light of the reliability, cost, and flexibility, thick film hybrids appear to be the most suitable type of integrated circuit for cable television amplifiers.

LIST OF FIGURES

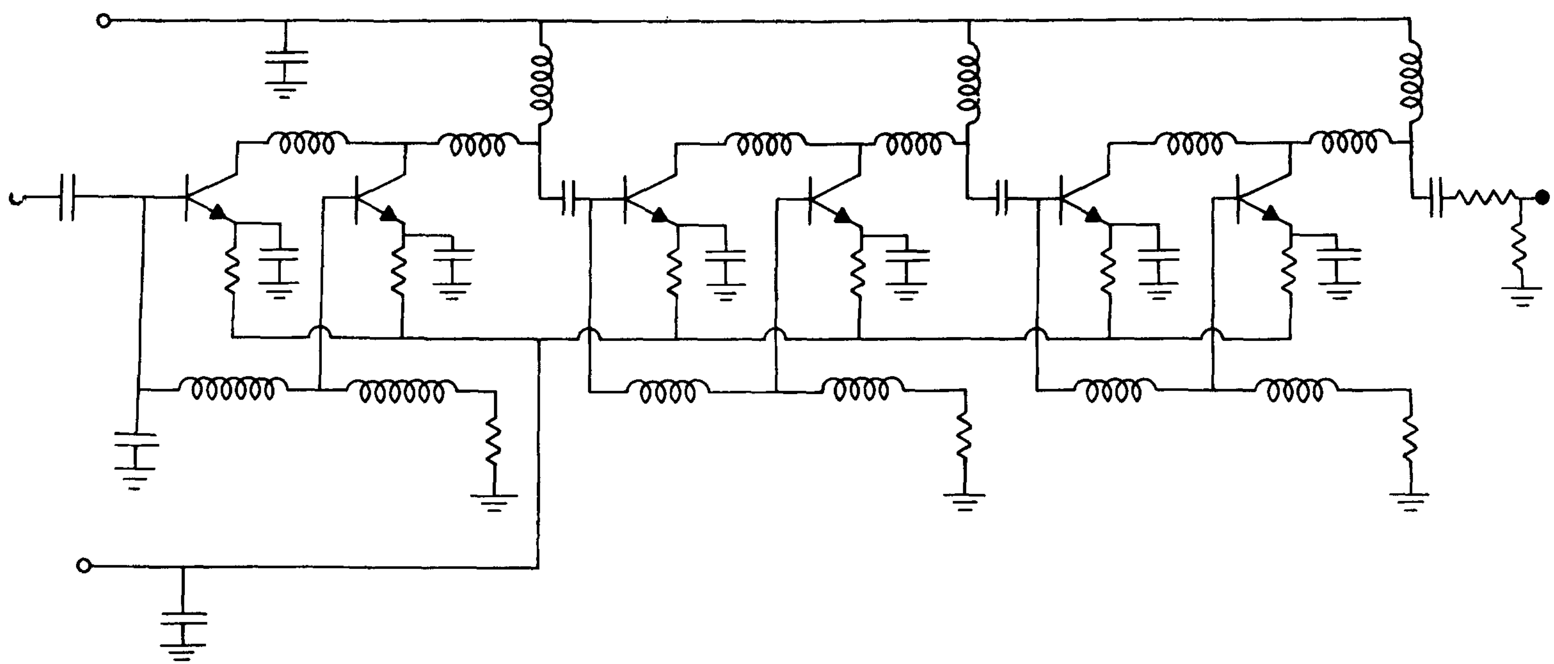


FIGURE 1. Schematic Diagram of a Distributed Pair Amplifier.

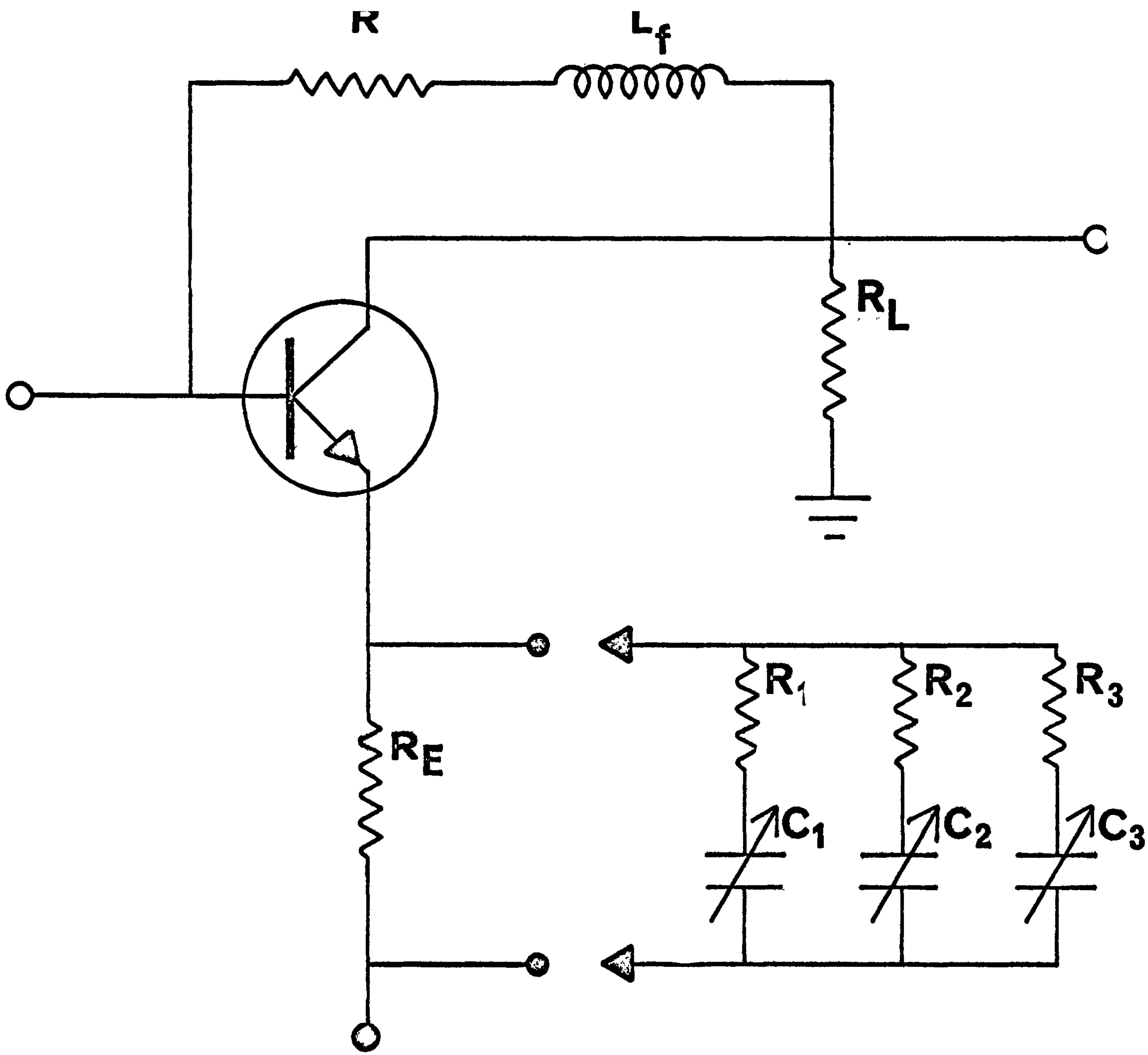


FIGURE 3. Basic Circuit for Obtaining Broadband Response in Cable Television Amplifiers.

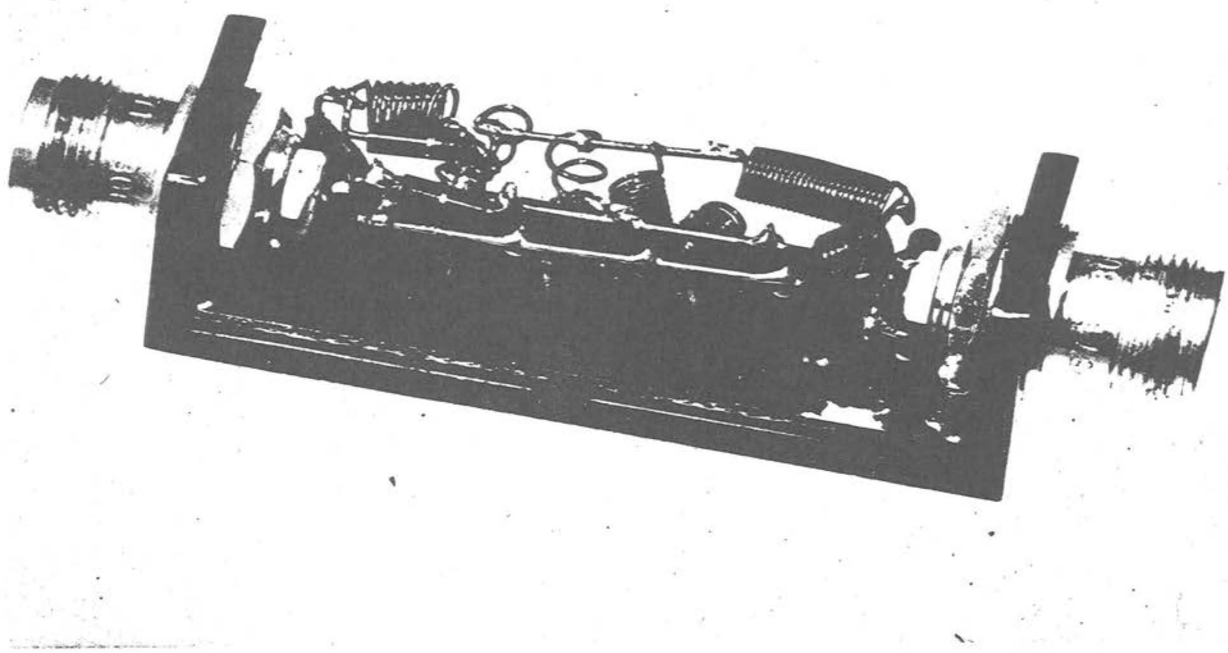


FIGURE 2. A Miniature Amplifier Covering the 50 MHZ to 500 MHZ Frequency Range.

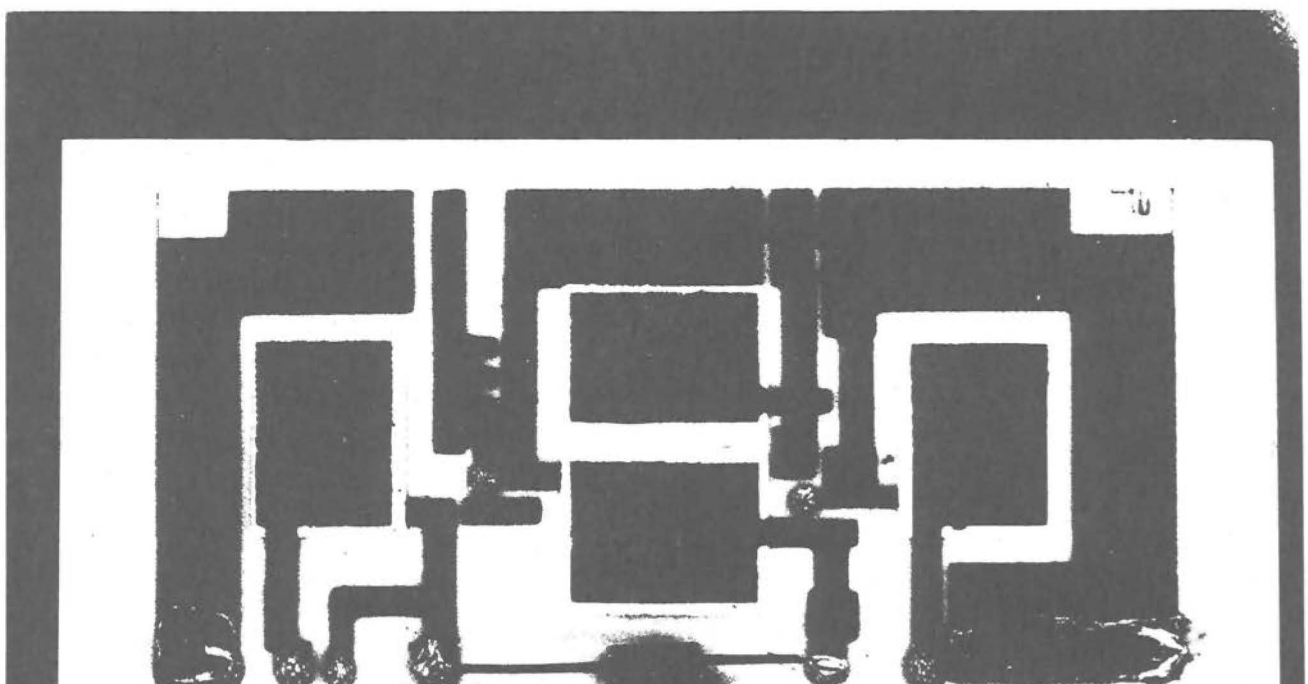


FIGURE 4 A Thick Film Broadband Amplifier

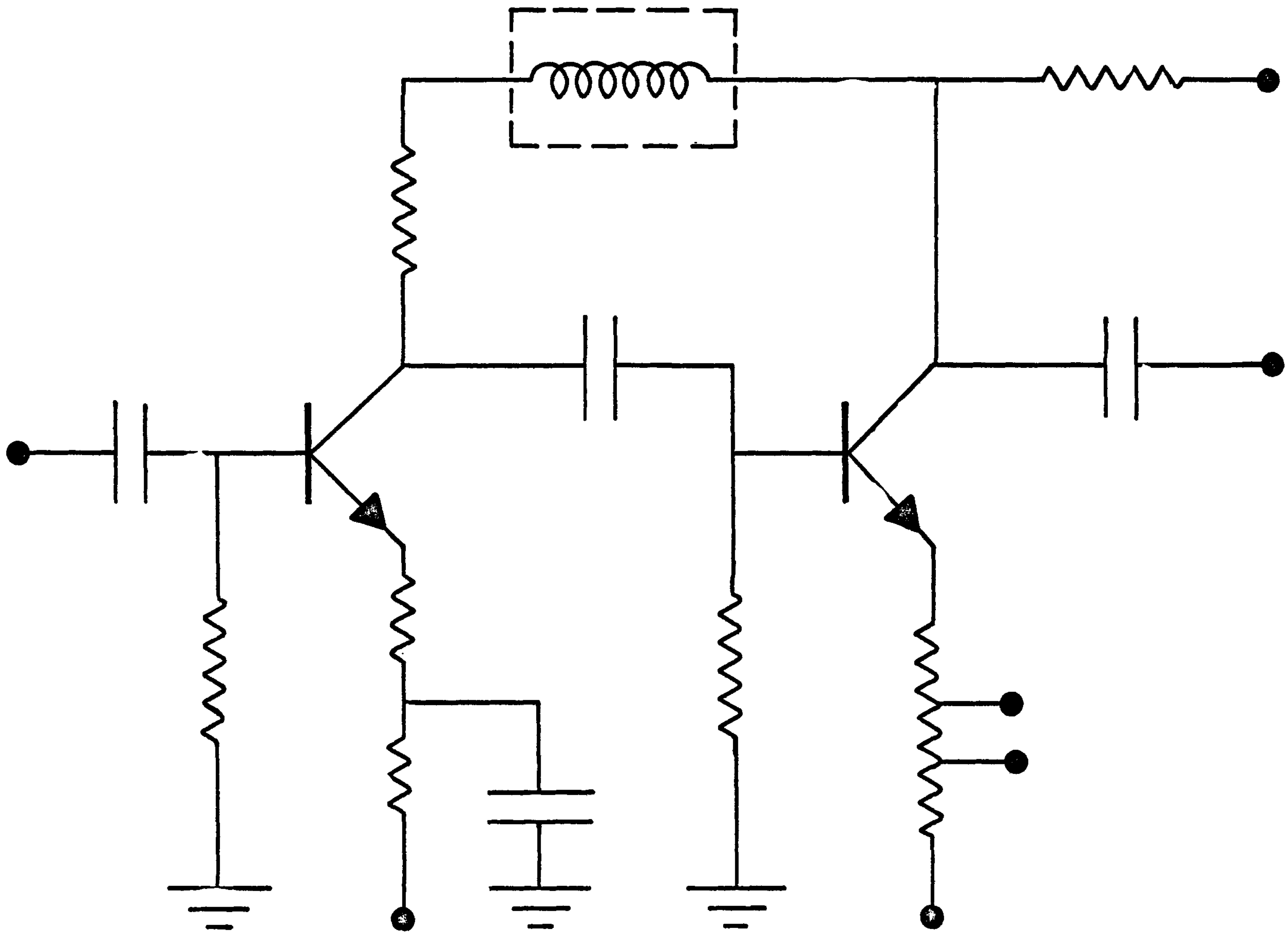


FIGURE 5. Schematic Diagram of the Thick Film Amplifier.

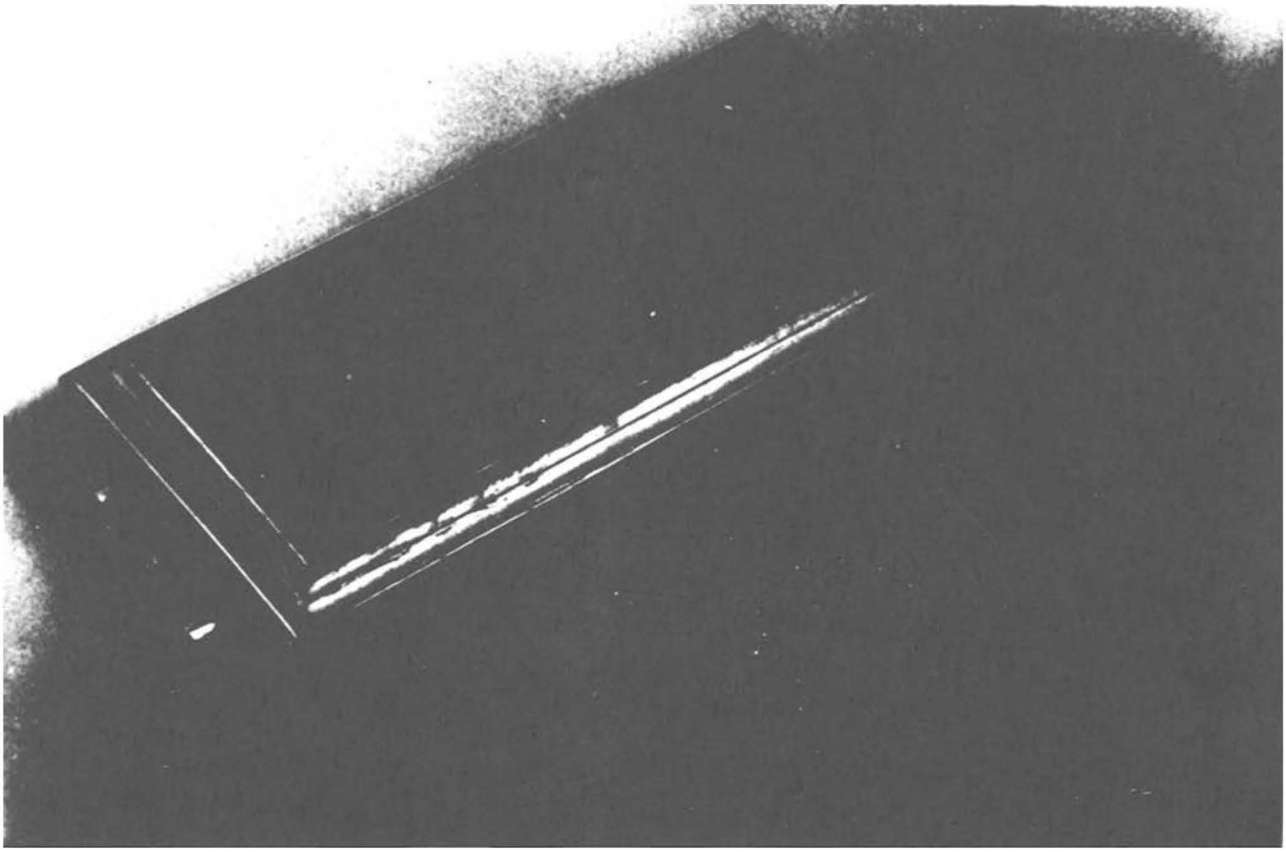


FIGURE 6. Size Comparison Between Conventional and Modular Distribution Amplifiers.

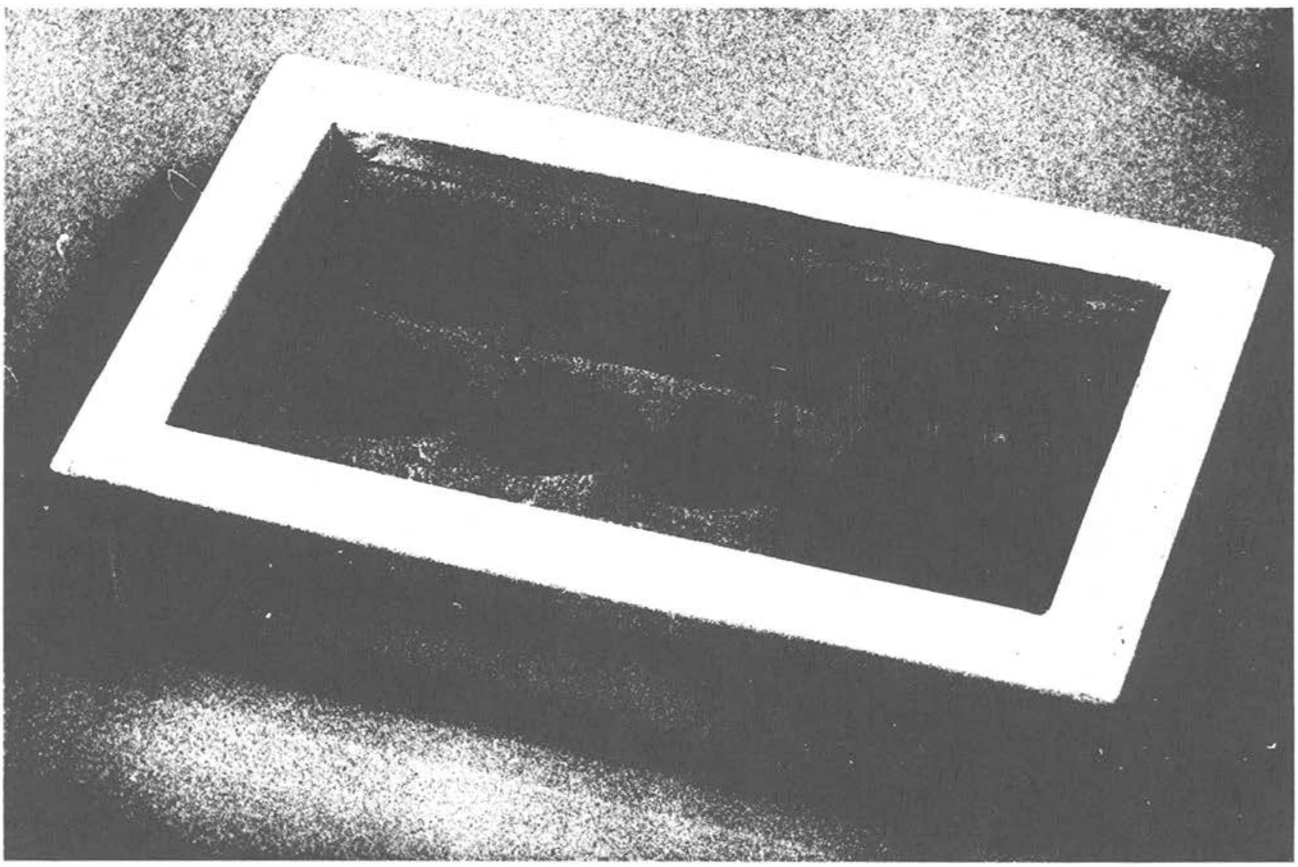


FIGURE 7. A Modular Distribution Amplifier.

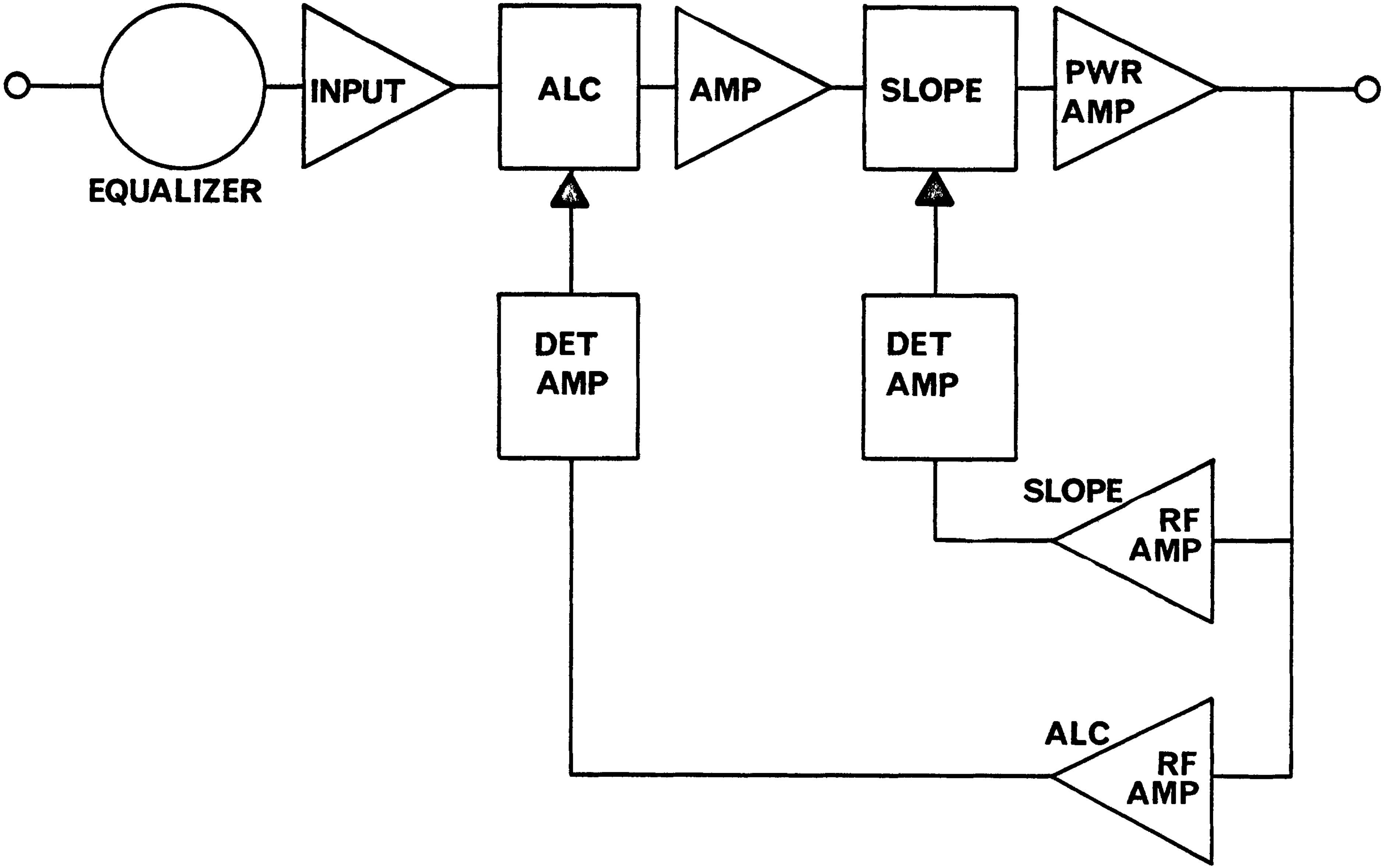


FIGURE 8. Functional Diagram of a Typical Trunk Amplifier.

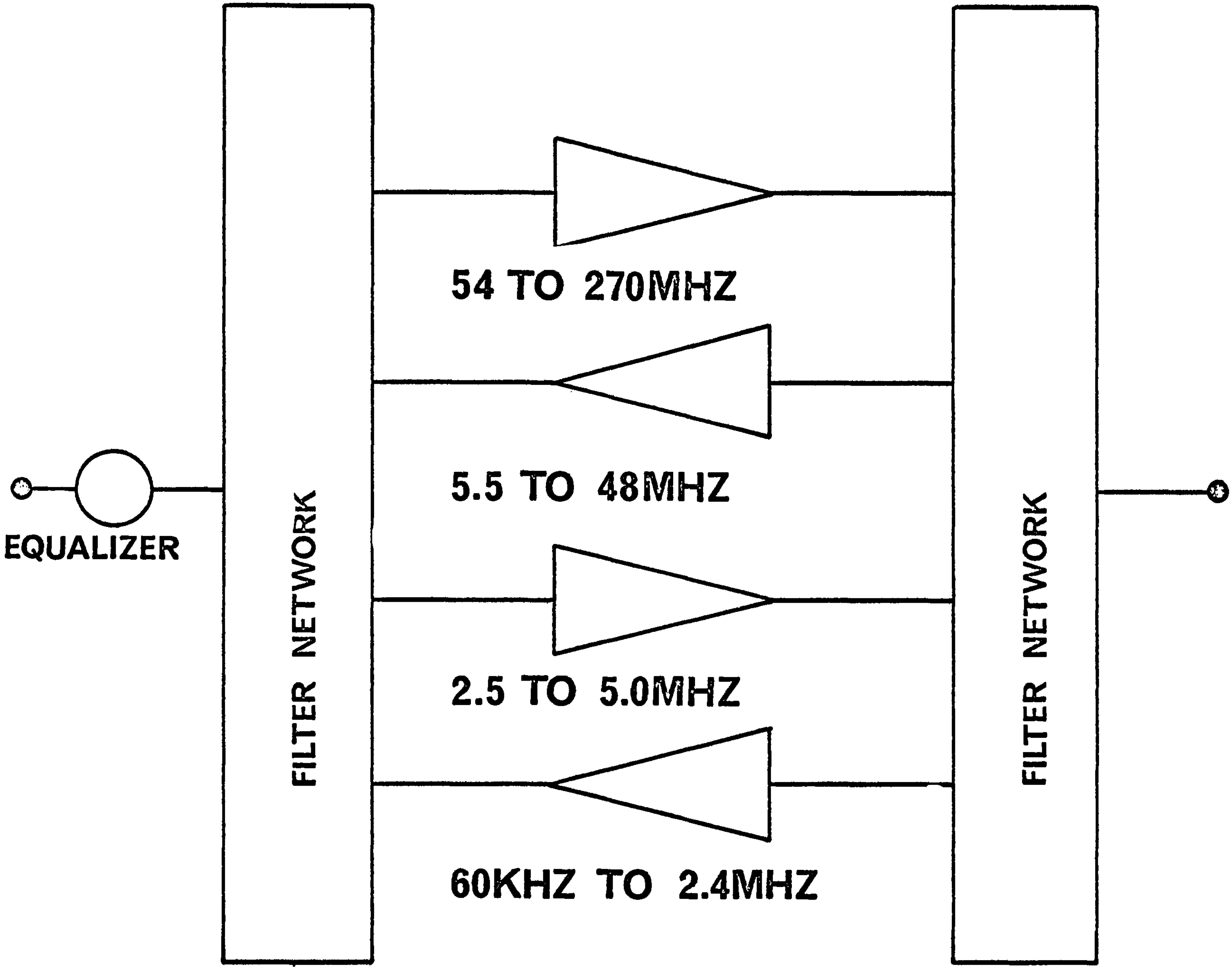


FIGURE 9. Amplifier Requirements for Proposed Frequency Plan.

TRADE-OFFS IN CABLECASTING

By Kenneth D. Lawson
TeleMation, Inc.

For the
1970 NCTA Convention

When one is attempting to decide upon what type of closed circuit television system to install, he comes to many forks in the road. "Should I go to color or black and white?" "Should my equipment be mobile or permanent?" "Should I use a 16mm film camera or a video camera on remote programs?" Most times these questions are asked in search of trade-offs which can be made when faced with problems of both capital and operating budget.

If there were no money or manpower limitations you would have enough equipment and personnel to go either way depending only on the dictates of program requirements.

This paper has been prepared to analyze two of the most common, but most difficult trade-off decisions which occur in our industry. The first is "color vs. black and white," and the second is "permanent vs. mobile studio equipment."

Color vs. Monochrome.

Other things being equal, a color television picture is far more stimulating than black and white. For this point there is no contest. But even in the face of this logic, and acknowledging the growing proportion of color home TV receivers, black and white cameras will, for the foreseeable future, still be the recommended method of televising a significant amount of locally originated cable programs.

This should not be construed as a negative attitude toward color equipment by TeleMation or me, for TeleMation has literally pioneered color television for cablecasting with eighteen studios either installed or in the process of being installed in cable systems. The point is that if you decide to rely solely on color, you will have to trade-off some types of programs which are done best, or only, in locations away from the studio.

Low cost vidicon color cameras reproduce color accurately only under controlled lighting of relatively high intensity. This type of lighting is not always available in the typical cablecasting schedule. It is not likely, for example, that the city council will consent to work under the amount of lights required to make a technically good color picture. Without these lights, pictures will tend to be off-color, ghosty, and noisy. The same electronic problem occurs in many evening sports events and social functions.

Another problem with color cameras is the large amount of paraphenalia needed to drag into the field if a smooth two camera production is to be accomplished. Color cablecasting is simply more complicated, and requires more technical skill than monochrome work. The weight and bulk of video control equipment is greater; and the amount of AC power available in many buildings is not adequate for portable lighting.

If for no other reasons than the relative simplicity and lower cost of equipment and operation, black and white productions will remain a large part of the cablecasting scene.

It is my opinion that, in general for most communities, Mrs. Jones will give up color for black and white if it is the only way she will be able to see her Bobby on TV crunched at the goal line on Friday evening.

It must be remembered, however, that it takes a fairly sophisticated black and white camera to do these marginal light productions. The pick-up tube must be at least of the separate mesh vidicon variety, and preferably a Plumbicon* (or lead oxide) or image orticon type. Separate mesh vidicon viewfinder cameras range from $1,000 to $2,000 each; Plumbicon* or lead oxide tube cameras are about $3,000, and image orthicons begin at $5,000.

Because of the growing number of low cost color cablecasting studio starts, it must mean that the trade-off in program types is considered an acceptable price to pay by some operators. For example, Tony Acone is a professional CATV program director for Coachella Valley TV in Palm Desert, California who has insisted on beginning cablecasting operations with a complete color studio system. Tony says: "In our resort and professional type of market area, people are extremely color minded. Color TV sets are the rule rather than the exception; and imported and local TV stations are color.

"Our only cablecasting limitation so far has been basketball and night football. Even with two full console bays of equipment, we take our color cameras from the studio into the field. Everything is carefully planned so that remotes are not unreasonably difficult. For example, tomorrow we will do the school dedication at 9:30 a.m., the high school baseball game at 3:00, and we will be ready for our evening show in the studio at 7:30. Slightly over 30% of our shows are in the field -- all in color."

Other cable operators avoid the program trade-off problem by having a simple, mobile monochrome system to supplement the color production center in the studio.

One multi-system operator is placing a sophisticated broadcast type, two-camera Plumbicon* color mobile van within a designated region to circulate through nearly a dozen owned cable systems. Special color programs can be staged in these communities in or out of the studio. A technically trained crew can operate efficiently on an annual basis with programs planned ahead of time. Advertising can be sold in advance of the productions which will be of a quality and nature to pull even the most jaded local businessman into the picture.

Each of the local cable systems, in this case, have their own, more modest monochrome systems for daily programming in black and white.

Because of the specific advantages of color, monochrome systems should be purchased with future convertability to color in mind. This is especially true for switchers, sync generators, film multiplexing systems, and videotape recorders. The monochrome cameras should be rugged enough to live out this evolution, so that they can be used for those "remotes" when color finally arrives in the more modest studio.

The future is likely to see the Plumbicon* (or lead oxide) type of pick-up tube used in color cameras at prices not too far different than that for existing vidicon versions. One color camera operates at extremely low light levels but costs four times more than existing low cost cameras. These types of cameras will solve some of the low light problems, but will still require more production skill and equipment logistics than black and white.

<u>Permanent vs. Mobile Cablecasting Equipment</u>.

The degree to which closed circuit television equipment is portable must be measured in three ways: (1) cameras, (2) the video and audio control center, and (3) videotape recorders.

A simplified definition of a permanent studio facility is one in which the control room facilities for camera and audio switching are not movable from the control room for producing programs outside of the studio.

Let us assume that you have a cablecasting system with two studio viewfinder cameras, a film chain, two videotape recorders, a MESSAGE CHANNEL™ and a WEATHER CHANNEL™. Your control equipment must provide a program switcher with capacity for at least six and preferably eight to twelve inputs. The extra inputs are for expansion of additional video inputs, such as special-set cameras, special effects, character generator, etc.

The control room will also need an adequate number of monitors -- at least four: a waveform monitor; a common synchronizing generator driving most, if not all, cameras; an intercom system between the technical director operating the switcher and the cameramen. It is also desirable to have remote controls placed in the control console to set up cameras and operate projectors and video recorders.

This much control equipment usually requires two consoles (three, if color) mounted together requiring floor space of approximately 4' by 4', and weighing approximately 450 to 500 pounds. A certain part of the capacity of each major equipment component in the console would be needed to produce a two camera show in the field, but to take the console into the field would obviously shut down other programming in the studio, such as films and tapes; and the problem of moving this amount of equipment is substantial. Since remote cablecasting is extremely important, the problem of portability must be solved.

Typical solutions to this problem are as follows:

(1) Rely upon single camera productions in the field without need for switching. Or

(2) Place all equipment in a van which becomes the control room, and which is connected to the studio by video, audio, film and VTR remote control, and power cables. Or

(3) Reduce video and audio control equipment to that which can be housed in one or two portable cases, and supplement it with a "bare-bones" standby switching monitoring system to be used when the portable console is taken into the field. Or

(4) Develop dual control facilities where the complete studio control room console is left in tact, and a separate, more modest mobile control unit is always available for use outside the studio. This mobile control unit can either be fixed in a van, or mounted in a portable case for use in or out of the van. The dual system is the best solution technically, but, of course, it is the most expensive in capital investment. Measuring portal to portal program time and costs, it may not be the most expensive to operate.

Single Camera Operation.

The obvious advantage of the single camera system connected directly to a videotape recorder is maximum simplicity of operation and transportation. The camera should have the following features to fully capitalize upon this technique:

(1) The camera should have an internal synchronizing generator which can be switched on when separated from the common sync system of the studio. Sync should be of the same quality as the studio system, preferably EIA or quasi EIA standard.

(2) The camera should have a lens which has a wide range of zoom focus, preferably 15mm to 150mm. Distances between camera and subject will vary more widely in the field than in the studio--especially with only one camera.

(3) The camera should have a pick-up tube which has superior low light performance, since studio lighting will not always be present

The Mobile Van.

To maintain that multi-camera "broadcast" operating format in the field, a mobile van is the most effective equipment mode. It permits a complete console arrangement; and wear and tear on equipment from moving it between studio and field is minimized by its permanent mounting in the van.

There are trade-offs in production methods which should be recognized, however, when the van is used as a control room for both studio and field. Let's look at some of these.

Film projection equipment is rarely, if ever, needed in the field. The optics of film chains are so critical that performance is enhanced if the system is securely mounted to a concrete or similar footing. A complete 3-way optical multiplexer with 2 16mm projectors and 1 slide projector is virtually impossible to fit into a van.

The film chain and at least one videotape recorder should be left in the studio when the van is away on location, so that continuous program capability is available.

The problem of splitting up your program production equipment between the control console in the van -- and the film equipment in the studio -- will occasionally cause problems for one man if he is running films with live programming.

Since the control room operator has no way to observe the studio set except by the two camera monitors placed in front of him in the van, he has to rely on comeramen to have more skill in some programs, and camera shots may require more pre-planning.

I would suggest that your two videotape recorders be of the portable type instead of rack-mounted in this situation. You are going to have some down-time on recorders, and the studio and van recorders should be completely interchangable without remounting in racks.

Storage cabinets in the van should be built with form-fitting shock padding so that cameras can be transported with lenses and tripod head attached. This will permit set up and tear-down time and wear to be minimized between studio and field.

One problem which can come up when all control equipment is fixed in the van is using cameras in locations where the van cannot be conveniently parked. For example, public buildings where council chambers are located are often built so that cables for cameras, audio, and power have to be laid across sidewalks, up stairs, down halls, and in aisles. Athletic arenas or gymnasiums can present the same problems. This is where solution number three, the portable control case, is needed.

The PORTA-STUDIO™.

The PORTA-STUDIO™ -- or portable video and audio control case -- gives maximum mobility. The trade-off is usually in terms of less switching capacity, less sophisticated remote controls of cameras and audio sources, and abbreviated test and monitoring equipment.

The PORTA-STUDIO™ also requires a certain amount of "standby" switching and monitoring in the studio, so that film, videotape, and automatic programming can be carried on while programs are being produced in the field. Again, all videotape recorders must be portable.

The program audio mixing console should remain in the studio where all microphone, VTR, film, audio tape, turntable and other inputs remain permanently connected. The PORTA-STUDIO™ needs only a simple microphone mixer with say, four mic inputs and one line level input capacity.

Dual Facilities for Video Control.

Cable systems which can afford dual studio and mobile facilities are very likely to develop color for the studio and black and white for the field. The type of mobile black and white mobile system which is developed -- van or PORTA-STUDIO™ -- will have to be determined by the effect of the trade-offs on your planned cablecasting schedule and quality requirements.

When planning your studio and remote cablecasting equipment, I hope that the trade-offs described in this paper will give you a helpful frame of reference for your investigations.

Thank you.

DISCUSSION

Mr. Ken Lawson: Thank you.

Mr. Greg Liptak: I think Ken has put down in a logical and organized fashion some of the problems that we're all dealing with on a day to day basis, particularly when you're talking about the capital investment required. We would all, I think, no matter the size of the system we're building, like to have a full color operation, but then we are in the position of are we going to have to forget about doing the local high school football and basketball games and some of these other things that really would be probably the major source of revenue for us at the start. I think it would be valuable at this point to hear from you as to your experiences with regard to color. We have in two of our systems, color cameras. They are low cost variety. We have a single camera in each system. At the start, I guess, we've had them in operation over 6 months now, we have determined exactly what Ken has told us---that we cannot use this low cost color camera in a remote situation under low light level conditions. It just doesn't work and yet we are able to provide beautiful studio pictures and under daylight conditions, excellent pictures. And it has, by the way, had a favorable marketing effect on the selling of advertising time on our local channels.

I would like to know what type of camera the fellow was using which he described in California?

Mr. Lawson: It was an IVC 3 tube camera.

Question: Was it a Plumbicon?

Mr. Lawson: No, it was a low cost Vidicon version. He's done a few things with it. He has put a 10 to 1 zoom lens on it which means he can go out and during the dedication of a building, for example, he can get off the sidelines and zoom in on the grandstand. (Tape fadeout)

Question: I was wondering--I was talking to someone from RCA yesterday on their color camera and they think if you take the color filter out of their camera which puts it into black and white. They think it will work very satisfactorily at low levels. Does anyone agree with this?

Answer: I would say there is no reason why it couldn't.

Mr. Liptak: We surely need a color camera that will operate under low light levels. If manufacturers could achieve that, our problems would be greatly lessened.

Question: Not audible.

Mr. Lawson: They claim it will work on monochrome with low light levels if you take the filter out of it and it's a matter of an hour's job to put it back in and realign the camera.

Mr. Liptak: That's part of the problem that we found--we really have to, in dealing with some of this equipment, have to have highly skilled people and that's why we like the particular variety of color camera that we're using because it's built like Collins Radio equipment--there's not too many buttons anybody can foul up. I would say there are two or three buttons you can do something with.

There is an equipment manufacturer at this show who supposedly has a low light level color camera and they say it will operate at 25 ft. candles.

Question: Of the three types of tubes optionally available for your cameras, could you give us a rundown how much light level is required and the difference between light levels?

Mr. Lawson: Just very generally because what is a good picture is different than what people think. Let's just forget about trying to produce a studio type of picture with backlighting and all the effects. We're going to talk about an electronically good picture--one that doesn't have too much noise in it and one that doesn't have too much smear in it. The 7735A type of Vidicon tube is the standard tube used most normally. To get a really good picture, you need 125 ft. candles. If you get down below 100 ft. candles, I think you begin to pick up noise and you begin to pick up a certain amount of lag and this would differ from camera to camera, but 75 ft. candles would sure be a cut-off, I would think, in terms of doing a presentable thing subject to a standard Vidicon tube. An A541 down to that point has very little change in lag at all in noise. I wouldn't worry about ft. candles at all, but if you got down to 50 ft. candles, I think you might begin to lose a little bit. Remember again, this also depends on the type of lens you have. Let's assume you have a normal 10.1 production zoom lens, you probably can get down to 35 ft. candles with a Plumicon tube and still have an acceptable picture electronic-wise. You may not have all the detail you want. See a 10.1 zoom lens has an F stop of about 2.8. This is the very important limiting factor on what's going into that tube. (Tape fadeout)

TV STUDIO ORIGINATION: THE PROFESSIONAL APPROACH

By Matt P. Spinello

Manager, Teleproductions

AMPEX VIDEO INSTITUTE

Before the

19th Annual Convention

NATIONAL CABLE TELEVISION ASSOCIATION

Chicago, Illinois

June 10, 1970

Matt Spinello

They have been calling some of our industry's televised staging effects "slap stick", our equipment "toys", and our attempts at television production, "Mickey Mouse!" A current user of video recording equipment has taken the statement literally by labeling his recorders MM-1 and MM-2. To season the wound, a printed source of information states that present day software within the industry leaves much to be desired; that material is much too boring; and that producers of videotape programming should consult first with educators before developing packages that are designed to entertain, educate, communicate.

"Quick and dirty" is the common reference they make to the method employed by most producers of videotape programming. They are so confident that cablecasters and manufacturers are oblivious to the "need to communicate" that one of them has predicted that a new method of displaying audio/visual material through a picture tube will wipe out CATV in the next decade.

Who "they" are is important if we find justification in their claims. It is equally important that they recognize the fact that CATV is here to stay, that cablecasters have established far-reaching goals, and that manufacturers of communications equipment are equally aware of the need to communicate and are constantly improving the state of the art.

We attempted to contact one of the publications that had referred to this industry's efforts as cartoonish, requesting equal space or time to explain what was being done within the industry to promote the concept of a professional approach to TV production origination. Our correspondence remains unanswered. We went a step further. We learned that Mickey Mouse was the most famous movie star of all time; that he made more successful motion pictures than any other star; and that his has been the most requested autograph of any figure displayed across the screens and television tubes of the world. Proven facts!

In answer to the accusation that our industry is producing boring softwear: we agree, with the exception that only one side of the picture is presented in the remark. Not everyone is producing sleep-generating material in this field, and much is being done to correct the problems that exist. As to the statement that CATV will be eliminated from the scene within the next ten years by a new product, we accept that remark as an ordinary challenge we expect will be met head-on, simply by nature of the fact that there is a definite need and place for film in this field, for slides and still photographs; for videotape and live camera presentations; and for cablecasters in the field to utilize the tools of today and the improvements of tomorrow.

As Americans, we're great for establishing statistics: some true, some false, some with tongue-in-cheek. Like Rowan (or was it Martin?) who stated on one of their programs last year: "by the time we are laid to rest we will have spent nineteen years of our lives watching television, four of them adjusting the horizontal hold control!" Or by the nameless statistician who claims there are 100,000 helical videotape recorders in use throughout the world in one-inch and half-inch formats. Using that figure, we can estimate that approximately 500,000 individuals are physically operating those machines. But from that total, fewer than 5,000 operators have been trained in TV production basics: less than 1.25% of VTR users. There is where the problem lies. More frightening, the majority of these operators have each been charged with a total job classification of writer/producer/director/camera operator/stage director/lighting and audio technician/video switcher; and sometimes talent.

We like to feel we're deeply involved in promoting a solution to the problem. The statements and statistics of the types presented here established the basis for an extensive lecture tour last year for members of the Ampex Video Institute. Four staff members of the AVI's Teleproduction Workshops (three with a broadcast background totaling 50 years) visited 104 cities presenting 94 road show work shops, training 4,000 students and seminar attendees, completing 16 speaking engagements, 17 remote videotape productions, and participation in 19 conventions and meetings. Two staff members established an Ampex Video Institute in Kowloon, Hong Kong; another in Stuttgart, Germany, and presented seminar segments in London, England. These projects were completed in addition to a regular calendar schedule of CCTV Workshops maintained in Elk

Grove, Illinois, plus the videotaping of 47 studio assignments for the year. Similar activity by staff members is expected to double this year.

Our findings, analyzed through travel reports made by AVI personnel, as well as on-site critiques of individual system's equipment, personnel and operation, justified our belief that the majority of problems exist through a lack of knowledge and experience with the equipment employed within the individual facilities. Through misconception, VTR operators were quick to blame equipment for lost program segments, poor quality reproduction, and equipment malfunction. Upon inspection of equipment and prerecorded segments by our road show teams, we found the majority of trouble areas had been created by what has been commonly expressed throughout the industry in slang form as "cockpit error", operational problems by the uninitiated, inexperienced, uninformed operator.

As a result of our findings, the Ampex Video Institute's road show team is totally involved in promoting concept as it relates to television production procedures. The program was developed to aid the potential user of video recording equipment by first establishing his needs through a basic understanding of the capabilities of a variety of video recording systems, as well as their limitations. The plan also serves current users. Through a series of road show seminars, consultation services, two- and five-day teleproduction workshops, the AVI team is devoted full time to serving on an international basis, those in need of guidance in the selection and proper operation of video equipment, and the duplication and distribution of recorded material.

Seminar attendees and workshop participants last year learned how a most involved program can be recorded professionally with a minimum of knowledge and equipment by applying that "basic understanding" technique to the total capabilities of their present equipment. They learned how their material, properly preplanned and committed to tape, could result in recorded segments that compete with, and meet broadcast production standards.

We do not mold network producer/director types through our brief presentations. Our capsule courses have been referred to as "crash courses", which we admit is a logical assumption. The material presented in a five-day workshop is a condensation of approximately four weeks of seminar and live studio participation. "Crammed" or "crashed", the sessions have proven that the "need to know" exists, and that studio involvement with the material covered in classroom lectures arms the participant with the understanding and experience he needs to return to his facility and produce the type of recorded package which will hold his audience's attention, convey a message, communicate!

Training is important; capable personnel to properly man selected systems are important. Engineers should be assigned technical positions, not a mis-matched combination that includes producing, directing, lighting, staging. We would not consider for a moment asking a producer/director to wire a control room, repair a video recorder or design a new camera circuit. Let us not -- in all fairness to the man, the facility and the end result -- expect an engineering type to reverse his position simply by assignment. Having the man trained in the basics of TV production as a supplemental add-on to his technical skills should definitely present a different picture (no comedy intended); an individual doubly trained and experienced should also realize a sizeably heftier pay check than one commonly issued to an individual assigned to the ordinary, everyday process of a single job function.

They (our challengers), should have realized a yesterday long ago that cablecasters are not interested in program packages that entertained televiewers in the late 40's and throughout the 50's; especially those packages that should have been sealed away in some sort of time capsule in the very early 60's. We must bestow multiple kudos upon cablecasters who refuse to be contracted into worm and torn emulsionized packages of yesterday.

We must not be shocked however, if in the process of producing our own packages, "live" or on videotape, a lack of TV production knowledge within our own facility generates a visual monster that might enhance one of yesterday's oldies on another channel. A dim lit, out of focus ball game, shot with one camera in a constant state of "zoomerama" may activate the automatic channel-switching device within a veteran televiewer's nerve center. And let us not mistake who the veterans are. If we have heard the statement before, let us mentally retain it, have it etched on the face plates of our video consoles, and flashing on and off at the entrance to our TV studios: "The youngster entering kindergarten has already witnessed 4,000 hours of broadcast television". That makes him an expert! He may know nothing of the technical aspects of tilting up a camera, trimming focus, panning left or right to properly compose a picture. But he is aware of a certain level of quality and content in a visual presentation fed through his TV monitor/receiver. As he ages, he obviously becomes more selective, until finally he joins us as armchair veterans.

The Ampex Video Institute maintains a no-charge referral service to help place AVI alumni in search of teleproduction jobs. We invite cablecasters in need of experienced personnel to contact us. We also invite cablecasters who are interested in having personnel trained to more proficient depths, to inquire of our teleproduction workshops, our consultation services, and our technical training courses; obviously without obligation. Let us know where we can help, regardless of the origin of, or the label on your present teleproduction equipment.

As for "they": our peers, critics, experts (?), may they continue to refer to our efforts as "Mickey Mouse!" We should be so famous, so successful!

ISBN 0-940272-01-6; 0-940272-08-3; 0-940272-10-5; 0-940272-11-3; 0-940272-12-1; 0-940272-14-8; 0-940272-15-6; 0-940272-16-4; 0-940272-18-0; 0-940272-19-9; 0-940272-20-2; 0-940272-21-0; 0-940272-22-9; 0-940272-23-7; 0-940272-24-5; 0-940272-25-3; 0-940272-26-1; 0-940272-27-X; 0-940272-28-8; 0-940272-29-6; 0-940272-32-6; 0-940272-33-4; 0-940272-34-2; 0-940272-35-0; 0-940272-36-9; 0-940272-37-7; 0-940272-38-5; 0-940272-39-3; 0-940272-40-7; 0-940272-41-5; 0-940272-42-3; 0-940272-43-1; 0-940272-44-X; 0-940272-45-8; 0-940272-46-6; 0-940272-47-4; 0-940272-48-2; 0-940272-49-0; 0-940272-50-4; 0-940272-51-2; 0-940272-52-0; 0-940272-53-9; 0-940272-54-7