CATV SYSTEM ENGINEERING

THIRD EDITION

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By William A. Rheinfelder

How to Plan and Design
Modern Cable TV Plants

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Preface to the Third Edition

For this edition, the entire text was generally reviewed, in view of today's technology, and changes and additions were made to bring the content up to date. A new chapter discusses concepts for systems with more than 12 channels, including other services. Also, a number of new entries were made in the Glossary of Terms.

From a scientist's viewpoint, there are still a great many subjects related to CATV system and amplifier design which are not well understood, and only research in the future will gradually lead to a more thorough understanding. Even to this day there is no valid equivalent circuit for transistors capable of broadband operation above 100 MHz. There are several approximations, but these are not adequate to explain the vital characteristics essential to CATV system design. Fortunately, development work is progressing, and unquestionably will be perfected in the future. As always, any suggestions and comments will be welcomed by the author.

Phoenix, Arizona | W. A. Rheinfelder
December 1969

Preface to the Second Edition

In the last two years the CATV industry advanced at a far more rapid rate, as far as equipment performance is concerned, than could be anticipated on the basis of progress during the early solid-state years from 1960 to 1965. The key feature of the new breed of amplifiers is the increased overload-to-noise ratio which has come about by major improvements in circuit design by several manufacturers, particularly in the area of output capability.Only during 1965, the typical 12-channel overload level of the best amplifiers was on the order of 45 dbmv, while equipment available now from several manufactures reads in the vicinity of 60 dbmv and more when tested under identical conditions.*

This somewhat fantastic and unprecedented increase in output capability of 15 db necessitates a thorough re-examination of the entire CATV concept, keeping in mind that every 6 db increase in dynamic range allows doubling the number of amplifiers which may be operated in cascade for the same system performance. 15 db allows, for example, a cascade of amplifiers more than five times longer. Or, stated differently, 15 db extra margin allows, for the same system length, additional errors in system level of ±7.5 db without serious consequences. This is equivalent to good system performance even with much sloppier and less accurate field adjustments. Also, a 60 dbmv output capability is now higher than was ever possible with the very best vacuum-tube amplifiers. Taking full advantage of these improved equipment capabilities, a modern CATV system uses high-level distribution, improved bridging methods, and better powering systems. In the main trunk, automation and factory alignment have become widely accepted for the ultimate in reliability and freedom from maintenance.

Due to these unexpected equipment advances, it was necessary to revise and update much of this book. All graphs were extended to cover the new equipment parameters. New chapters were added to cover Amplifier Design Con-

*Differently expressed, new equipment shows a considerable reduction of distortion at system operating levels.

cepts, High Level Distribution, and Cable Powering. Other chapters, including the appendices, were expanded and re-arranged. The basic philosophy of avoiding mathematics and lengthy derivations in the main text was retained. However, the interested reader will find additional material by consulting the footnotes for bibliographical references, as well as the various appendices which are given to a mathematical treatment of some pertinent topics. New Appendix III gives details on taps, and Appendix IV contains system mathematics with calculated examples. Several errors which occurred in the drawings or text were eliminated. Lastly, an alphabetical index was added for greater usefulness.

Again, the author will be grateful for any suggestions for further improvements of the book.

W. A. Rheinfelder

Anaheim, California
March 1967

Preface to the First Edition

The CATV industry, since its beginning about 15 years ago, has come a long way. However, the explosive growth of CATV, particularly during the past few years, has increased the need for knowledge and understanding of the technical aspects, as well as improved system standards. Many presently operating CATV systems are functioning only because of the ceaseless and undaunted efforts of a few technicians who, by cut-and-try methods and with little guidance, have succeeded in making a number of uncorrelated components perform with each other. It is for these individuals, as well as for all other persons interested in CATV, that this book is written. Particularly, it is designed to serve as a technical handbook for the instruction or reference of all personnel involved in planning, installing, operating, and maintaining a CATV system.

The year 1965 represented a turning point for the CATV industry. For the first time, a theory of fully integrated systems has evolved as a result of freedom of system design brought about by the advances of new solid-state equipment. Advances in design will, in the next few years, result in a greater standardization and stabilization of system design and layout. Also, more equipment specifically designed for this industry will become available, eliminating the use of units modified or adapted by cut-and-try methods. Considerably better performance at lower cost will be achieved in such new, fully-integrated systems. But also, in older systems, worthwhile improvements will be achieved by modernization along the lines discussed in this book.

In order to keep the material easier to follow for technicians without formal schooling, mathematics has been generally avoided; however, some of the more important mathematical derivations are included in the Appendices.

The first Chapter gives a short introduction to CATV engineering and lists recommended performance standards for a high quality system. The following Chapters then treat segments of the system in logical sequence. Chapter 2 covers head ends, and Chapter 3 is devoted to amplifiers. The rather important aspect of system spacing is discussed in Chapters 4 and 5. Chapters 6 and 8 treat two important considerations in system design—the use of level diagrams, and the problems associated with reflections. Chapters 10 and 11 have to do with the maintenance and automation of CATV systems. Finally, Chapter 13 discusses new methods for achieving precision measurements on CATV components. Additional helpful material is given in the various Appendices.

In writing a first manual for a new and rapidly developing industry, the material of necessity, had to be limited. Particular emphasis was placed, however, on developing a sound understanding of the general engineering principles involved without going too much into distracting variations. Two factors will have a greater influence on future system engineering than any other—the theory of spacing, and the introduction of solid-state equipment. It is on these two points that a considerable change in thinking is required, especially for newcomers in the field of transistorized amplifiers. Although transistors have been around for quite a number of years, their impact on CATV is only recently being felt and there is no predicting what the future will bring.

The author will be thankful for all suggestions concerning this book. May it bring some technical footing to our fledgling industry and contribute to increasingly sound and prosperous engineering.

W. A. Rheinfelder

November, 1965

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Active - any circuit containing amplifying devices, such as tubes or transistors.

Antenna preamplifier - a small amplifier located in the immediate vicinity of the antenna, used to amplify extremely weak signals, thereby improving the signal-to-noise ratio of a system.

Automatic gain control (AGC) - a circuit which automatically controls the gain of an amplifier so that the output signal level is virtually constant for varying input signal levels. Sometimes referred to as automatic level (ALC) or automatic volume control (AVC).

Automatic spacing - a method whereby unavoidable errors in amplifier spacing are automatically corrected.

Automatic Temperature Control - a method whereby changes due to temperature in amplifiers or coaxial cable are automatically corrected by either a closed- or open-loop servo system.

Automatic tilt - automatic correction of changes in tilt.

Block tilt - a form of half-tilt where all low-band channels are set to the same level as Channel 2, and all high-band channels to the same level as Channel 13.

Bridging amplifier or bridger - an amplifier which is connected directly into the main trunk of a CATV system. It serves as a high-class tap, providing isolation from the main trunk and multiple high-level outputs.

Cable powering - a method of supplying power to solid-state CATV equipment by utilizing the coaxial cable to carry both signal and power simultaneously.

Closed-loop system - a servo feedback system where the residual error after correction is fed back directly into the servo system for inverse proportional control.

Coaxial cable - the most commonly used means of signal distribution, consisting of a center conductor and a cylindrical outer conductor (shield). Other types of transmission line used in CATV systems include open-wire (two-wire line) and Goubeau line.

Combining network - a passive network which permits the addition of several signals into one combined output with a high degree of isolation between individual inputs.

Countermodulation - a type of cross - modulation distortion where interfering modulation appears as out-of-phase modulation of the desired signal. "White Windshield Wiping" is typical countermodulation. May be caused by a combination of second-order distortions. See reference 20, App. VII.

Critical distance or cable length - the length of a particular cable which causes a worst-case reflection if mismatched; depends on velocity of propagation and attenuation of cable.

Cross-modulation - a form of distortion where modulation of an interfering station appears as a modulation of the desired station. Caused by third and higher odd order nonlinearities. A typical example of cross modulation is the form of overload known as "windshield wiping."

Directional coupler - a high quality tapping device providing isolation between tap and output terminals.

Distribution system - the part of a CATV system used to carry signals from the head end to subscribers' receivers. Often applied, more narrowly, to the part of a CATV system starting at the bridger amplifiers.

Dynamic range - see overload-to-noise ratio.

Emitter tuning principle - a circuit which extends the high-frequency response of transistors by using variable trimmer capacitors to neutralize detrimental emitter inductance.

Equalization - a means of modifying the frequency response of an amplifier or network, thereby resulting in a flat overall response.

Equalized Loss - any loss in CATV systems caused by coaxial cable; also, insertion loss of components designed to match cable loss characteristics.

Feeder line - the coaxial cable running between bridgers, line extenders, and taps.

Flat Loss - equal loss at all frequencies, such as caused by attenuators.

Flat Outputs - operation of a CATV system with equal levels of all TV signals at the output of each amplifier, corresponding to fully-tilted input signals.

Frequency response - the change of gain with frequency.

Full-Tilt - operation of a CATV system with maximum tilt at the output of each amplifier (flat input signals at each amplifier).

Fully integrated system - a CATV system designed to take advantage of the optimum amplifier - cable relationship for highest performance at lowest cost. Such a system is also admirably suited to the fully automated CATV system concept.

Gain - a measure of amplification, usually expressed in db.. For matched CATV components, power gain is readily determined as insertion power gain. Gain of an amplifier is often specified at the highest frequency of operation, for example, at Channel 13 for all-band equipment.

Half-Tilt - operation of a CATV system half way between full-tilt and flat-outputs operation. Compared with a FULL-TILT system, the level of Channel 2 is up at the input and output of an amplifier by one half its slope.

Head end - the electronic equipment located at the start of a cable system, usually including antennas, preamplifiers, frequency converters, demodulators, modulators, and related equipment.

High band - TV Channels 7 through 13.

House drop - the coaxial cable from line tap to the subscriber's TV set.

Hum modulation - form of distortion where the power-line frequency modulates the TV signal, causing hum bars to appear in the picture.

House drop - the coaxial cable from line tap to subscriber's TV set.

Inline package - a housing, for amplifiers or other CATV components, designed for use without jumper cables; cable connectors on the ends of the housing are in line with the coaxial cable.

Insertion loss - additional loss in a system when a device such as a directional coupler is inserted; equal to the difference in signal level between input and output of such a device.

Integrated system - a term used to denote a system in which all components, including various types of amplifiers and taps, have been designed from a well founded overall engineering concept to be fully compatible with each other. Such a system results in greater economy at improved performance through the avoidance of over specification by well engineered design center values.

Intermodulation - a form of distortion where two modulated or unmodulated carriers produce beats according to the frequency relationship $f = nf_1 \pm mf_2$, where n and m are whole numbers. Intermodulation is caused by second and higher order curvature, and is essential for the proper operation of frequency converters, mixers, modulators, and multipliers. Second-order curvature by itself does not cause distortion of the modulation envelope, but is often responsible for parasitics. The order of the intermodulation product is defined as n plus m.

Jumper cable - short length of flexible coaxial cable used in older CATV systems to connect system coaxial cable to amplifiers or other CATV components; they have no place in a modern, high quality system.

Level diagram- a graphic diagram indicating the signal level at any point in a system.

Line extender or distribution amplifier - types of amplifiers used in the feeder system.

Low band - TV Channels 2 through 6.

Main trunk - the major link from the head end to a community or connecting communities.

Noise figure - a measure of the noisiness of an amplifier. Noise factor is defined as input signal-to-noise ratio to output signal-to-noise ratio. Noise figure is noise factor expressed in db. The lowest possible value for a matched system is 3 db.

Noise modulation - also called "Noise Behind the Signal." Increase in demodulated noise level with the application of the desired carrier.

Open-loop system - a servo correction system in which the residual error is unrelated to the means of correction.

Overload-to-noise ratio - the ratio of overload-to-noise level measured at or referred to the same point in a system or amplifier, usually expressed in db, and commonly used as an amplifier figure of merit. Not to be confused with signal-to-noise ratio.

Pancake package - a flat amplifier housing designed to accommodate printed-circuit boards.

Passive - a circuit or network not using active devices such as tubes or transistors.

Reflection coefficient - ratio of reflected wave to incident wave—mathematically related to VSWR.

Return loss - reflection coefficient expressed in db.

Semiconductor - material with an electrical conductivity between conductors and insulators. Most commonly used semiconductors for transistors and diodes are germanium, silicon, and gallium arsenide.

Signal-to-noise ratio - the ratio of the signal to noise level with both measured either at the input or output of electronic equipment, usually expressed in db.

Single-ended package - a housing for electronic components having connections at one end only.

Slope - difference in amplifier gain, or change in cable attenuation, between Channels 2 and 13, in db.

Solid state - a term taken from physics, used interchangeably with the word transistorized; also includes other semiconductor elements, such as diodes. Generally refers to tubeless equipment.

Spacing - length of cable between amplifiers expressed as db loss at the highest TV channel provided for in a system, usually at Channel 13 in an all-band system; equal to amplifier gain in main trunks.

Span - distance between line extenders or distribution amplifiers; also, distance between taps.

Splitter - a network supplying a signal to a number of outputs which are individually matched and isolated from each other; usually based on hybrid coils.

System level - the level of the highest signal frequency, usually Channel 13, at the output of each amplifier. Must be carefully chosen and maintained for least distortion and noise.

System Mode - or System Tilt. The tilt at the output of each amplifier, normally set in the head end for the main trunk (or bridger for the distribution system).

Tap - any device used to obtain signal voltages from a coaxial cable. The earlier forms such as capacitive and transformer types have been replaced by directional couplers in modern systems.

Terminator - a resistive load for an open coaxial line to eliminate reflections; usually capacitively coupled to avoid shorts in cable-powered systems.

Tilt - difference in level between Channels 2 and 13 at output of amplifier in db; in fully-tilted system operation, equal to slope. Often used interchangeably with slope. See also full-tilt, half-tilt, flat outputs.

Tilt-compensation - the action of a tilt-compensated gain control, whereby tilt of amplifier equalization is simultaneously changed with the gain so as to provide the correct cable equalization for different lengths of cable; normally specified by range and tolerance.

Velocity of propagation - velocity of signal transmission. In free space, electromagnetic waves travel with the speed of light. In coaxial cables, this speed is reduced. Commonly expressed as percentage of the speed in free space.

VSWR - abbreviation for Voltage Standing Wave Ratio. Reflections present in a cable due to mismatch (faulty termination) combine with the original signal to produce voltage peaks and dips by addition and subtraction. The ratio of the peak-to-dip voltage is termed VSWR. A perfect match with zero reflections produces a VSWR of 1. For freedom from ghosting, most matches in a CATV system must have a VSWR of 1.25 or less.

Windshield wiper effect - onset of overload in multichannel CATV systems caused by cross-modulation, where the horizontal sync pulses of one or more TV channels are superimposed on the desired channel carrier. Both black and white windshield wiping are observed and are caused by different mechanisms. See also countermodulation.

CHAPTER 1

The CATV System

CATV, or Community Antenna Television, is a system designed to provide television service to outlying communities where none, or only unsatisfactory and limited off-the-air reception, is possible. The basic idea is to erect receiving antennas at a site where good quality TV signals are available. These signals are then relayed to the community and distributed to individual subscribers via cable or wire. Since many subscribers share in the same system, receiving equipment of the finest quality can be installed, resulting in superior quality TV reception. Among additional benefits are increased TV service such as many new program channels, the choice of FM background music, an information channel with automatic time and weather reports and the like.

1-1. FUNCTION AND PURPOSE OF CATV

Generally, a CATV system consists of the receiving system, or head-end equipment, and the distribution system which carries the signals to the individual homes (Fig. 1-1).

The head end includes a number of receiving antennas, usually mounted on a tower or located at a fairly high elevation in order to achieve a high signal strength. Various pieces of electronic gear amplify the incoming signals and, if necessary, convert them to different channels for better system operation and freedom from various types of interference. The channel output levels are individually controlled and finally provide a combined signal of up to twelve or more TV channels of studio quality. No effort or expense is spared at the head end, since even in a very large CATV system only one head end is required. A poorly engineered head end would be conspicuous throughout the entire system.

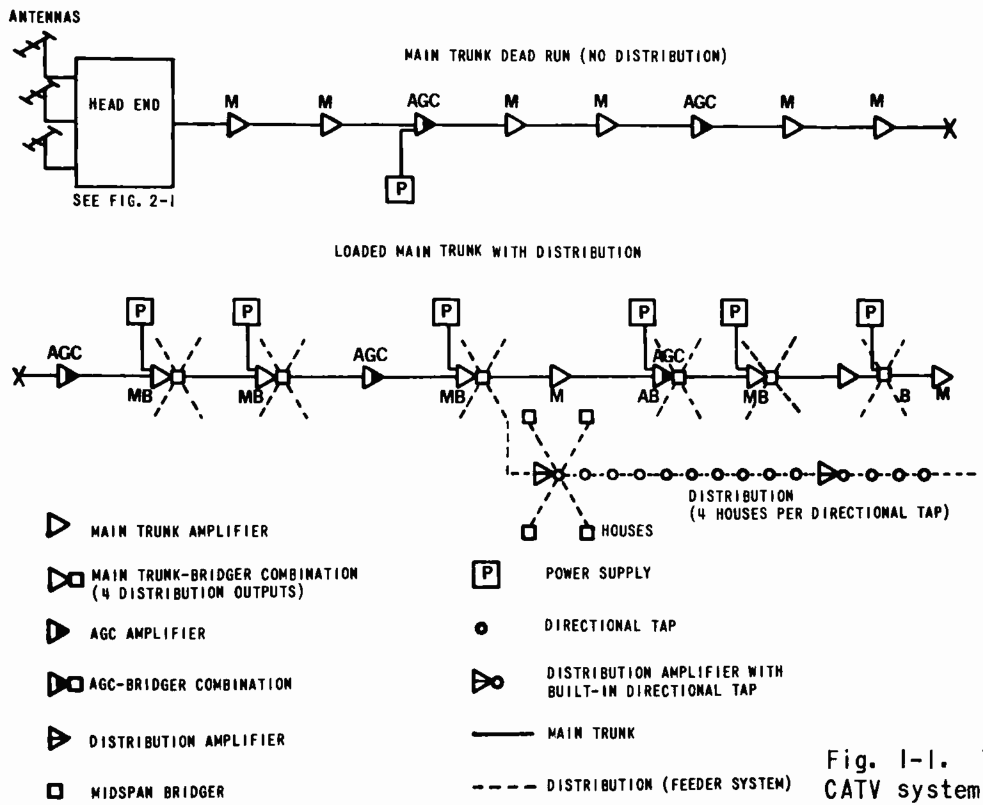


Fig. 1-1. The basic components of a CATV system are shown in this sketch.

Of much greater economic importance is the distribution system from the head end to the subscribers' homes. The main trunk line is the major link from the head-end site to the community. It consists of high quality (large diameter, low loss) coaxial cable, with repeater amplifiers (main trunk amplifiers) spaced along the cable. Also, as part of the main trunk in special level control positions, AGC amplifiers (Level Control Amplifiers) provide automatic correction for various changes in signal level due to temperature, etc. The main trunk is generally routed over the shortest possible path between the head end and various distribution points, and may even supply more than one community. The maximum system length is determined mainly by the length of the trunk which can be used without causing excessive degradation in picture quality. In recent years, great advances have been made in the development of improved amplifiers which, together with better cables, have led to substantially increased trunk length. Consequently, CATV systems are becoming a possibility now where before no practical solution existed or some other form of transmission, such as microwave, had to be used.

Normally, no taps can be used directly on the main trunk to connect feeder lines except in high-level main trunks. Therefore, a special type of amplifier, called a bridging amplifier, is used to provide isolation from the main trunk. It has several outputs, and enough gain to make up for the isolation loss and the power loss inherent in multiple outputs. This amplifier is often combined with the main trunk amplifier (mainline-bridger combination). From the bridging amplifiers, feeder lines are run along a row of houses. The feeder or distribution system is usually less critical and requires less exacting equipment. However, a recent study indicates that a high output capability is even more desirable in the distribution system than in the main trunk (see Chapter on High Level Distribution). Consequently, a new type of amplifier, called a distribution amplifier, has come into use in feeder systems, replacing the older low-level line extender. Chief advantages of the distribution amplifier are higher system economy and reliability (fewer amplifiers for a given number of subscribers). Both types of feeder amplifiers compensate for the losses in the feeder system, and each feeder line may be four to ten or more amplifiers deep. Power for the various amplifiers is supplied either individually for tube equipment, or simultaneously with the TV signal via the coaxial cable for solid-state equipment (cable powering).

Between successive distribution amplifiers or line extenders, directional taps or couplers are provided at an average spacing of 150 feet. These taps usually have four outputs each, to which individual house drops are connected. In the past, these taps were often pressure taps, a form of capacitive tap which mechanically pierced the coaxial cable and made pressure contact to outer and center conductors. Also, transformer taps, of the unmatched or the better back-matched variety, were often used. More recently, directional couplers are specified since other forms of taps are unsatisfactory as far as reflections (ghosting) are concerned. Moreover, in an integrated system design, the somewhat higher cost of directional couplers can be offset by savings in other devices. Therefore, the total system cost remains unchanged, or is perhaps even reduced, although system quality is actually higher.

In a fully integrated system, care should be taken not to overspecify or underspecify any particular component. For example, if directional couplers are used (which is a necessity in a high quality system), the built-in suppression of reflections will allow a considerable relaxation in the specifications (as far as match is concerned) of other pieces, such as line extenders. This reduced requirement reflects itself in reduced cost and improved performance of other more important amplifier parameters.

I-2. PERFORMANCE STANDARDS OF CATV SYSTEMS

The performance and quality of a CATV system is directly related to the signal-to-noise ratio of the system. In a more general sense, noise includes thermal noise ("snow") as well the effects of interference and distortion such as "windshield wiper" effect, "ghosting," and other forms of signal degradation. Since picture quality is a rather subjective measure—dependent on the observer as well as the particular TV receiver and the setting of its controls, room lighting, etc. —extensive tests* have been made by the

*See Appendix VII, References I-6, 9, 10, 11.

SMPTE and other organizations to relate observed picture quality to signal-to-noise readings, and to establish a quality standard of picture transmission. These tests indicate that a flawless television picture of studio quality results with a signal-to-noise ratio of 40 db, while still very good picture quality is achieved with a signal-to-noise ratio of 30 db (Table 1-1). It should be noted, however, that these values are the true signal-to-noise ratios of video signals with a 4-MHz bandwidth. Readings taken with a field strength meter having a different bandwidth must be correlated to these readings. Another correction would be necessary if a peak reading meter were used to read random noise, etc.

It is thus the objective of CATV system design to achieve at least the minimum acceptable signal-to-noise ratio at every connected receiver, under the varying conditions of system operation such as changes of signal strength, temperature, etc. To obtain a high system signal-to-noise ratio, it is necessary that all components in the system feature an even higher individual signal-to-noise ratio due to the cumulative build-up of noise and distortion when electronic equipment is operated in cascade. Thus, a cascade of 10 amplifiers has a 20 db lower overload-to-noise ratio than that of an individual amplifier, and 40 db lower for 100 amplifiers. (More about this in Chapter 4.) The number of amplifiers which may be operated in cascade is therefore limited, particularly with amplifiers which have a low individual overload-to-noise ratio.

However, new advances in amplifier design have largely overcome limitations to cascading. Degradation of system performance with cascading follows a logarithmic law. For example, with older amplifiers, any circuit improvement which allowed a cascade to be increased from 10 to 20 amplifiers had a significant effect on system design. With newer amplifiers, the corresponding change might be from 100 to 200 amplifiers operating in cascade. In both cases, system degradation is 6 db for doubling the number of amplifiers. It must be remembered that even in newer systems the maximum possible cascade is often not realized because of other limitations (see Chapter 5), or on purpose in order to trade system length for increased performance margins on picture quality.

Regardless of the length of the system, for best picture quality it is also essential to maintain the correct system operating level (the signal level at the output of each amplifier). As the system level is increased, the effect of noise decreases proportionately until overload occurs. At this point distortion rapidly increases. There is then an operating level where the optimum signal-to-noise or distortion ratio is achieved. Since distortion increases much more rapidly than noise, it is desirable to set the operating level slightly below this value to obtain smoother system operation with the normal, unavoidable system changes.

In summary, for a high quality system, the design and concept must be such that a video signal-to-noise or interference ratio of at least 40 db can be obtained at every subscriber's TV set. This is the only requirement of importance, and all other system specifications must be directly related to this basic standard. With this fact in mind, we shall discuss the various portions of a system, with their components and functions, in detail.

Table 1-1. Picture quality rating.

Video Signal-To-Noise Or Distortion Ratio	Rating
40 db	Flawless, studio-quality picture
35 db	Some snow or distortion visible
30 db	Acceptable, but with flaws clearly visible
25 db	Snow and distortion becoming objectionable

CHAPTER 2

Head-End Concepts

The head-end equipment, located at the start of a CATV cable system, must produce the high-quality TV signals which are essential for satisfactory system operation. Because of the unavoidable degradation of signal quality in the cable system, the requirements for the head end are exacting. For example, a high-quality, solid-state CATV amplifier has an overload-to-noise ratio of 85 db. A cascade of 30 such amplifiers produces a system overload-to-noise ratio of 55 db. Consequently, a head-end overload-to-noise ratio of only 55 db would, by itself, result in a degradation equivalent to 30 amplifiers, a situation which cannot be tolerated.

2-1. FUNCTIONS AND REQUIREMENTS OF HEAD END

Generally, head-end equipment can easily be designed to exceed the dynamic range of the repeater amplifiers. For one thing, only a single channel has to be amplified, which greatly increases the overload capacity. Therefore, we should rightfully expect the overload-to-noise ratio for the head-end equipment to exceed that of the best amplifier in the system. Noise, of course, in a more general and meaningful sense, is meant to include all types of interference. But not all spurious outputs manifest themselves as a degradation of the video dynamic range. Therefore, a 60 db suppression of spurious outputs is often satisfactory.

The large video dynamic range is achieved by filtering, AGC, and circuit concept. Some filtering must be provided in the RF stage and antenna preamplifiers, to avoid difficulties of cross-modulation with other signals, with more filtering in the IF strip. AGC is provided in both forward and backward directions to keep the output constant for normal variations in signal level. All these functions are performed

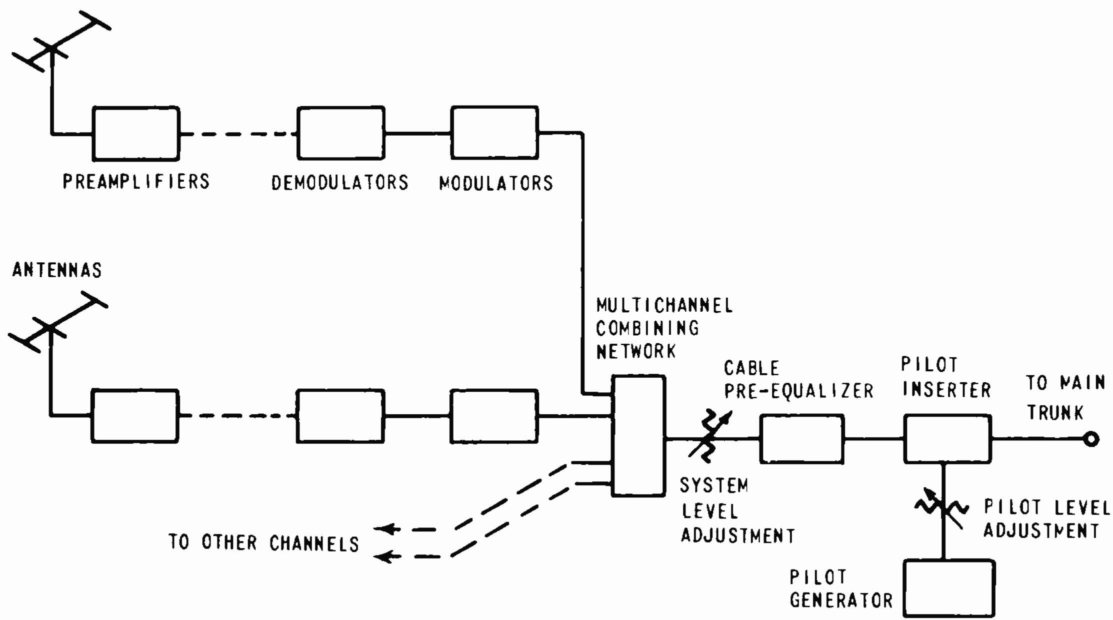


Fig. 2-1. Diagram of basic head-end equipment.

similarly in a less perfect way in a normal TV set. However, the achievement of a large dynamic range is only part of the function of the head end. Normally, frequency translation (conversion from one channel to another) is necessary in all-band CATV systems to minimize interference problems. Also, any received UHF channels are converted to VHF channels at this point. This conversion is necessary to keep losses in the system within reason. Consequently, no use of the built-in UHF capability of newer TV sets is generally made in CATV system design.

In addition to frequency conversion, the head-end equipment allows for the precise adjustment of the composite signal level for each channel, as well as a separate correction of picture- and sound-carrier levels. These adjustments are provided to permit the correct balance of the various TV channels with each other and to reduce interference problems due to the sound carrier in all-band CATV systems. The sound carrier is normally reduced below the picture carrier by about 15 db. All head-end signals are added in a multichannel combining network which provides isolation between channels and freedom from spurious outputs due to interaction. The combined output signal is then applied to a cable preequalizer, which adds the necessary frequency correction for the section of main trunk cable from the head end to the first main trunk amplifier. Any pilot carrier for AGC purposes is added after the preequalizer, just before the start of the main trunk system.

The complete head-end setup (Fig. 2-1) may include separate antenna preamplifiers, if the distance between antenna site and head-end building is excessive, in order to keep the total system noise figure at low values. Electric power for the head-end equipment is sometimes a problem, particularly on mountain-top locations, and a self-contained motor-generator power station may be required. With transistorized head-end equipment, some form of cable powering becomes a logical possibility.

A number of different concepts for head-end equipment have been developed through the years. These concepts are not equivalent, but rather evolved historically, and knowledge of the various approaches will be of value to system engineers and technicians.

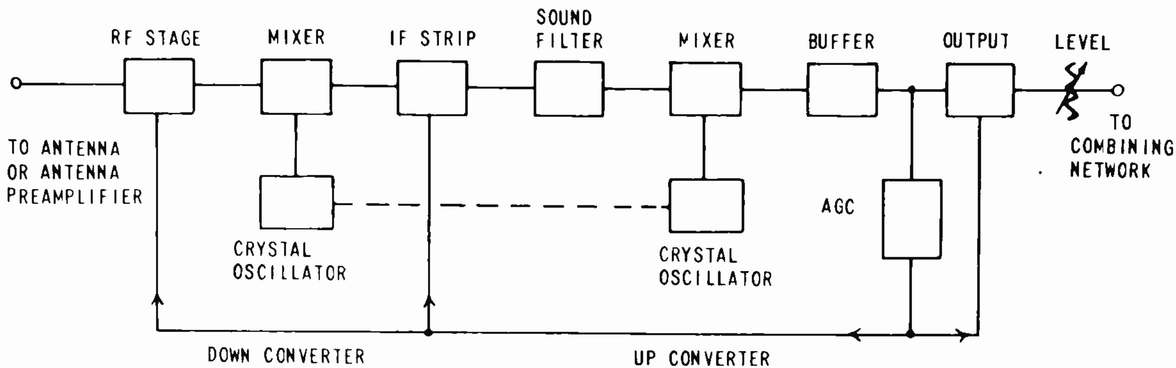


Fig. 2-2. Block diagram of a typical reconversion head end.

2-2. FREQUENCY CONVERSION

A block diagram of the frequency conversion type of head-end equipment is shown in Fig. 2-2. The incoming channel is converted down to a standard IF frequency, then reconverted up to the desired channel frequency. The adjustment of the sound carrier is performed by adjustable trap circuits in the IF amplifier. This type of head-end equipment can be perfected to a high degree and may be fully transistorized to advantage.

In order to achieve the high quality essential for CATV, cross-modulation must be kept to a minimum. This requirement necessitates the use of a push-pull mixer. Consequently, two RF stages must be used in the RF amplifier to achieve a low overall noise figure. A high-quality front end of this type would of necessity be based on circuits using field-effect transistors (FETs), because of their far superior performance in this application when compared to regular bipolar transistors or vacuum tubes. The oscillator is crystal-controlled for high stability and low maintenance. Generally, no temperature stabilization of the crystal is required for CATV work, particularly with the small heat stress of transistorized equipment. The IF amplifier has, as a main objective, the removal of interference by its inherently high skirt selectivity. Lower phase distortion results from the use of specially designed double-tuned bandpass filters. Low phase distortion and reduced group delay is important for high-quality picture reproduction, particularly for color signals.

An adjustment for the sound carrier must also be provided in the IF strip. Generally speaking, conventional traps also affect the picture to a considerable degree, and crystal filters are generally too narrow to be useful. Thus, while it is possible to develop a satisfactory circuit at some expense, the control of the sound carrier represents a weak spot in the otherwise advantageous reconversion concept.

For on-channel operation, that is where the output signal has the same frequency as the input, the second oscillator used in the up-converter must be controlled by the crystal of the down-converter to avoid intermodulation beats. This connection is indicated by a dashed line in Fig. 2-2 The up-converting mixer is followed by buffer and output stages, which are designed to reduce the unwanted mixing products and to provide the necessary output power. It is desirable to have a fairly high undistorted power level out

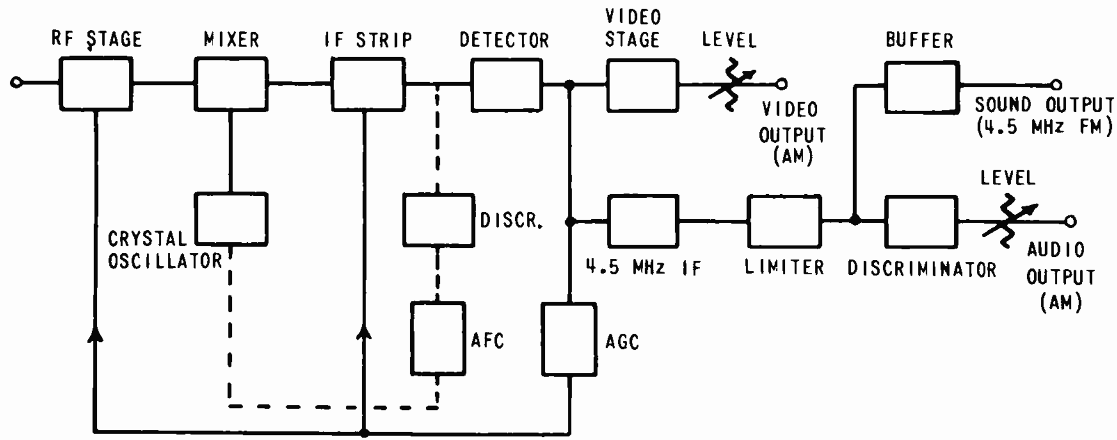


Fig. 2-3. Block diagram of a typical head-end demodulator section.

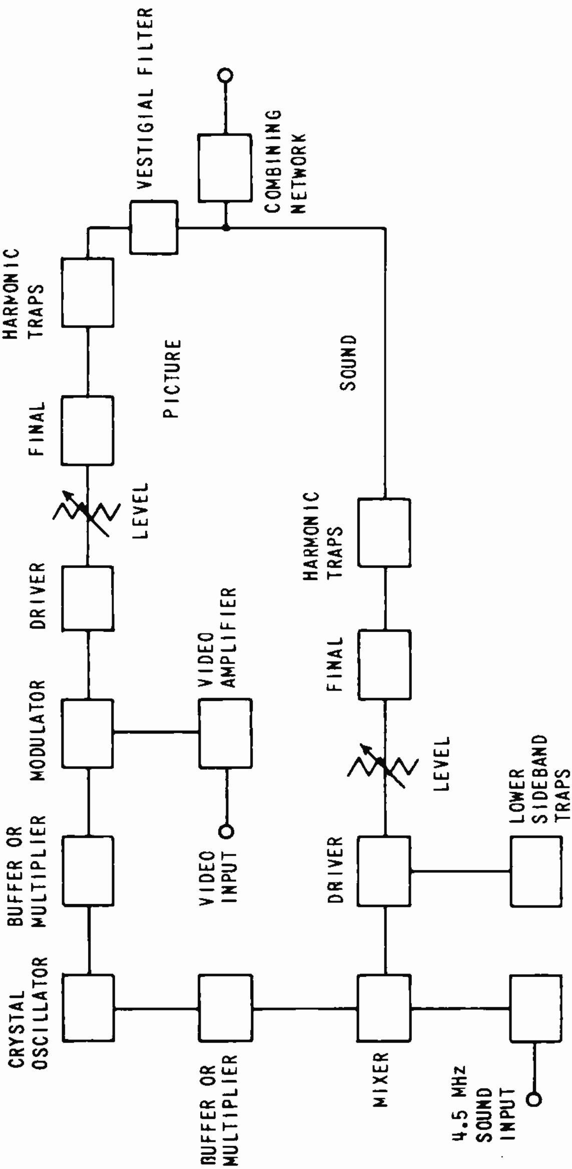


Fig. 2-4. Block diagram of a sound-and-picture modulator.

of the head-end equipment, because of the rather severe insertion loss inherent in the following combining network, where all channels are added to form the single output signal fed into the cable.

AGC is generally designed to keep the output constant for an input change of 20 db. This AGC action, while not difficult to attain, calls nonetheless for a well engineered concept with the correct amount of forward AGC. The reconversion principle has the advantages of simplicity, low cost, and the elimination of possible sources of distortion in demodulator or modulator circuits. It causes difficulties with control of the sound carrier due to simultaneous phase distortion of the picture signal. Reconversion AGC requirements are stringent and frequently not met by available equipment. The reconversion principle will not correct inferior modulation, and this concept does not lend itself to adaptation for use with microwave links or program origination.

2-3. REMODULATION

Basically the remodulation principle involves a receiver or demodulator, and a transmitter or modulator. The demodulator receives its input signal in the same general way as the reconverter; however, the IF strip is followed by a video detector and amplifier, providing a high level (about 1 v peak-to-peak) output signal at low impedance. Also, a 4.5-MHz sound amplifier with limiters is included to provide a constant, low-impedance output of the frequency-modulated 4.5-MHz sound carrier.

Generally, a demodulated audio signal is also available from a built-in discriminator. This signal may be used for monitoring purposes or special applications. A separate discriminator is necessary for automatic frequency control if the front end (tuner) is not crystal-controlled; this is shown by dashed lines in Fig. 2-3. Such a receiver then provides a studio-type video signal and a 4.5-MHz FM sound signal, both held constant by AGC and limiters. The frequency deviation of the sound signal is, of course, identical to that of the broadcast station.

The video and 4.5-MHz FM sound output signals are then used to modulate the transmitter as shown in Fig. 2-4. The picture and sound transmitters are kept separate to the very end to avoid intermodulation products. In the picture transmitter, a crystal oscillator and buffer and/or multiplier is followed by a modulator, usually of the push-pull variety. Negative feedback may be included in the modulator circuits to maintain low distortion and wide frequency response. Of particular interest for color is the magnitude of differential phase and gain produced in the modulator and the following stages of the picture transmitter. Considerable engineering effort must be spent in the circuit design of the modulator to obtain flawless picture modulation, and this possibly is one of the shortcomings of this concept. DC restoration is normally a part of the video amplifier and modulator, enabling acceptance of all kinds of video signals.

The sound portion is relatively simple. The sound carrier is obtained by mixing the picture frequency with the 4.5-MHz FM signal from the demodulator. The resulting sound carrier needs only filtering and amplification, and is ready for combination with the picture carrier. In particular, the lower sideband resulting from the mixing process must be removed by several traps and careful shielding to avoid interference on other channels.

The advantages of the remodulation concept are obvious. Picture and sound level are inherently constant and easily controlled without interaction. The quality of the video signal may be improved over that available from the station. Also, the picture modulation level may be reduced for decreased cross-modulation in the cable system. Camera signals may be directly applied to the video input for program origination, such as advertising, weather information, news, etc. The sound signal is only reconverted and not demodulated. Consequently, no degradation in sound quality occurs unless the bandwidth of the sound circuits is too narrow. The only critical factor is the picture modulator, but excellent performance is possible with a well engineered circuit. This concept is also fully adaptable to microwave.

For the remaining function of complete program origination, a separate audio modulator is needed, since this transmitter accepts only 4.5-MHz FM sound. A modulator which converts audio signals to 4.5-MHz FM sound is shown in Fig. 2-5. The audio signal is fed through a pre-emphasis network and amplifier to the reactance modulator which, in turn, controls the frequency of a 4.5-MHz oscillator. This frequency-modulated signal is amplified and brought to the output. The only problem with such a modulator is automatic frequency control (AFC). AFC may be accomplished by using a discriminator or ratio detector tuned to 4.5-MHz; however, this method is inaccurate because of the limited slope of the control function for these devices. Also, no precise frequency setting is possible due to the poor temperature stability of normal tuned circuits. These problems can be overcome by using the phase-locked AFC system shown in Fig. 2-5. The phase error between a 4.5-MHz crystal oscillator and the frequency-modulated oscillator is measured in a phase bridge, and the output is used to control the reactance modulator. Such an AFC system is far superior to other systems since it provides the basic accuracy of a crystal oscillator.

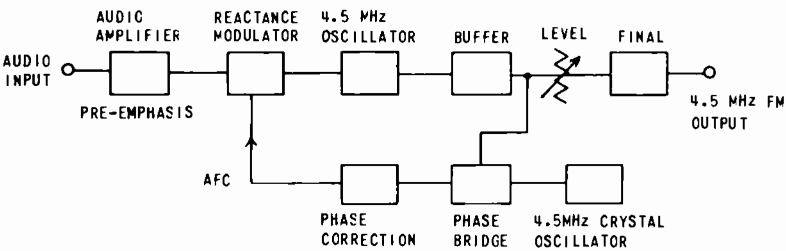


Fig. 2-5. Block diagram of a 4.5-MHz FM audio modulator.

All head-end equipment can be transistorized to advantage. Not only does solid-state equipment result in higher reliability and reduced current drain, but it also leads to performance far superior to that attainable with vacuum tubes. Transistors have a far higher transconductance than tubes, a decisive factor for this type of equipment. Also, semiconductor diodes possess far better modulation characteristics than vacuum tubes.

2-4. STRIP HEAD-END EQUIPMENT AND ANTENNA PREAMPLIFIERS

The previously discussed methods of providing TV signals for CATV systems are capable of the highest quality; however, they are somewhat more costly than individual straight-through RF amplifiers (so-called strip amplifiers) for each channel. This approach is only possible, of course, for on-channel operation—that is, where incoming TV channels can be fed directly into the cable. Anybody familiar with receiver design is aware of the problems associated with this approach, including cross and intermodulation, selectivity, stability, sound control and AGC. None of these problems can be dealt with adequately by circuit design; if attempted, a far more critical and costly circuit would result than with the methods already discussed. Therefore, it seems that this approach must be left to low-quality systems as an interim solution until better equipment is installed.

Antenna preamplifiers, while of similar design, have a different purpose and are a part of the highest quality system whenever the need arises. The main function of an

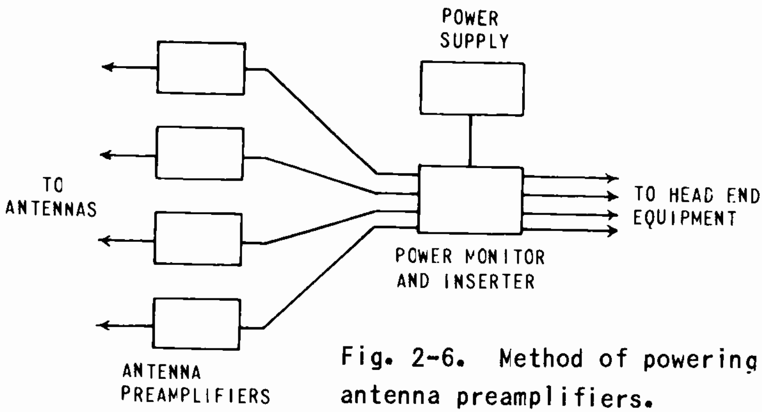


Fig. 2-6. Method of powering antenna preamplifiers.

antenna preamplifier is to keep the total system noise figure down. The need for antenna preamplifiers arises when weak signals are received from the antenna, when the following head-end equipment has a high noise figure, or when the distance from the antenna to the head-end equipment is excessive. In order to keep the system noise figure low under these conditions, it is necessary that the antenna preamplifier itself be of low-noise design and have sufficient gain to override the noise of the following equipment. These objectives are achieved with a noise figure of 5 db or less and a gain from 14 to 25 db. Higher gain is of no advantage, and usually causes additional problems due to regeneration. Since some form of interference rejection is desirable, it is common practice to trade gain for increased selectivity. Also cross-modulation and other forms of distortion and group delay may be minimized while, at the same time, achieving a flat-topped response with good skirt selectivity.

Antenna preamplifiers are mounted immediately adjacent to, or as part of, the antenna, and remotely powered with DC current from the head-end equipment site (Fig. 2-6). The remote power supply includes current monitoring to assure proper operation of each preamplifier. Antenna preamplifiers should not be used indiscriminately, but only in cases where an excessively low signal level at the input to the head-end equipment warrants their use.

As with other pieces of CATV equipment, the performance of RF preamplifiers often falls short of what is possible with modern circuit engineering practices. A substantial improvement in CATV system performance can be achieved by fully integrated system design, where the entire CATV system is based on professional engineering principles.

CHAPTER 3

CATV Amplifier Characteristics

Disregarding head-end equipment which consists of antennas, frequency translators, etc., and excepting the various passive components and the coaxial cable itself, the amplifiers used in cable TV comprise the most important investment in a CATV system and effect, as no other piece of equipment, the quality and performance of the entire system.

3-1. AMPLIFIER REQUIREMENTS FOR CATV

CATV amplifier requirements are exacting, since they must meet the more stringent characteristics of repeater equipment. A situation similar to re-recording equipment in sound recording exists; as a piece of music is re-recorded several times, all faults such as noise, flutter, distortion, etc., are multiplied, necessitating far superior quality for the electronic equipment than is ultimately required for the whole process. In CATV systems, TV signals are amplified again and again in repeater amplifiers which compensate for the normal cable losses. Since all deficiencies inherent in these repeater amplifiers are cumulative, it is clear that rather strict requirements must be met by the individual cable amplifier.

In order to fully appreciate the problem of CATV amplifier design, it is desirable to examine the essential amplifier parameters in detail. While there are several different types of amplifiers needed in a CATV system—such as main trunk, distribution, bridger amplifiers, etc.—the general characteristics are closely related and will be discussed first.

3-2. EQUALIZATION

Coaxial cable, the most frequently used transmission line, has an attenuation characteristic which is a function of frequency and varies according to an exponential law. On log-log paper, attenuation versus frequency may be plotted as an approximately straight line (Fig. 3-1). All cables show a deviation from the straight line at higher frequencies. This deviation is less pronounced and occurs at higher frequencies for better quality cables. For the TV range from

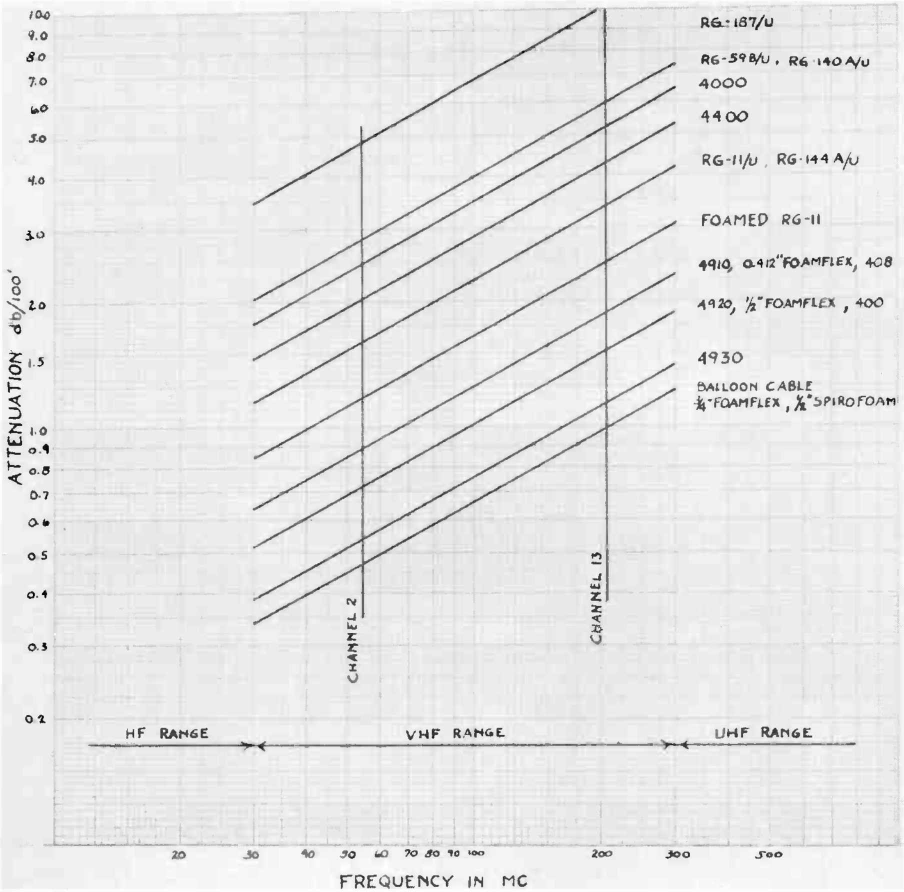


Fig. 3-1. Typical coaxial cable attenuation curves.

50 to 220 MHz, a straight-line approximation may be used. In CATV work, it is common practice to specify cable length in db attenuation at the highest channel in use (Channel 13, or 213 MHz, for modern all-band systems). Fig. 3-2 shows the attenuation characteristic of typical high-quality coaxial cable for 25 db cable length, a common spacing between repeater

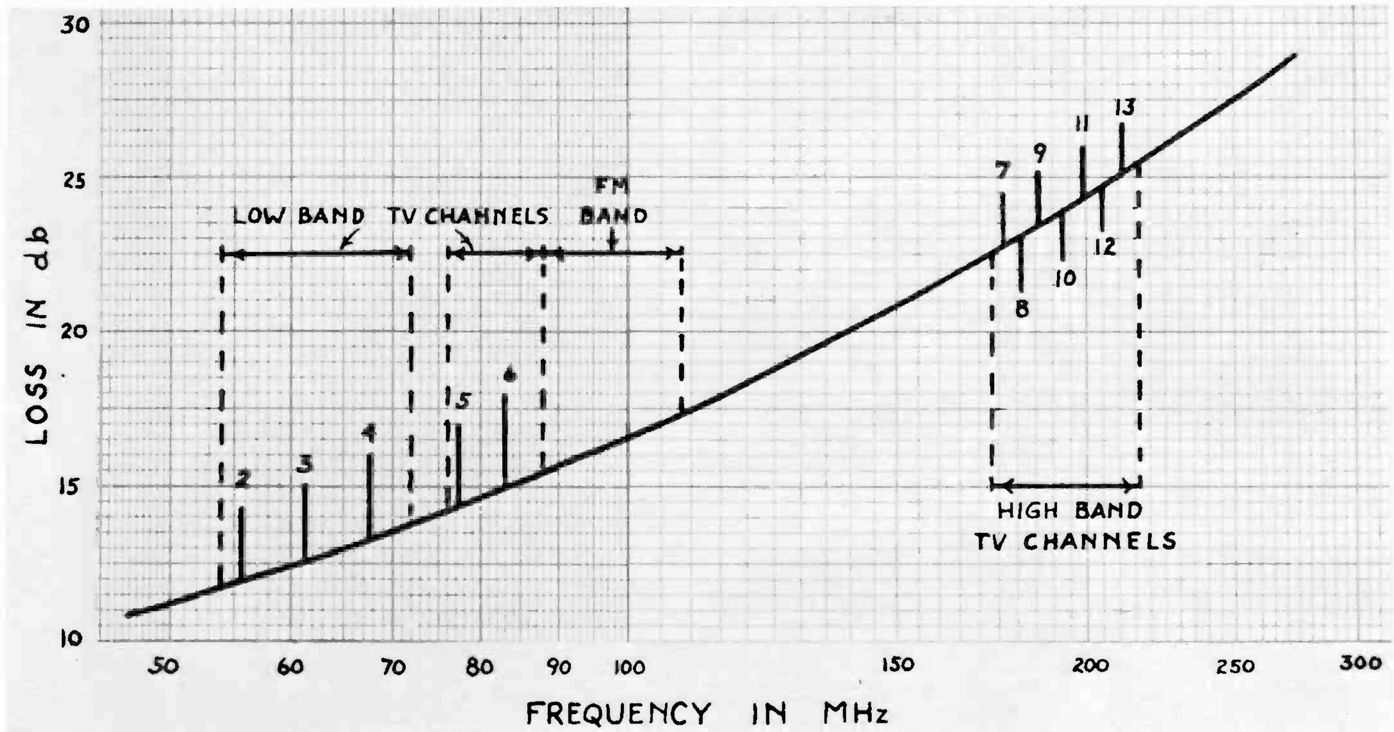


Fig. 3-2. Cable loss for 25 db spacing.

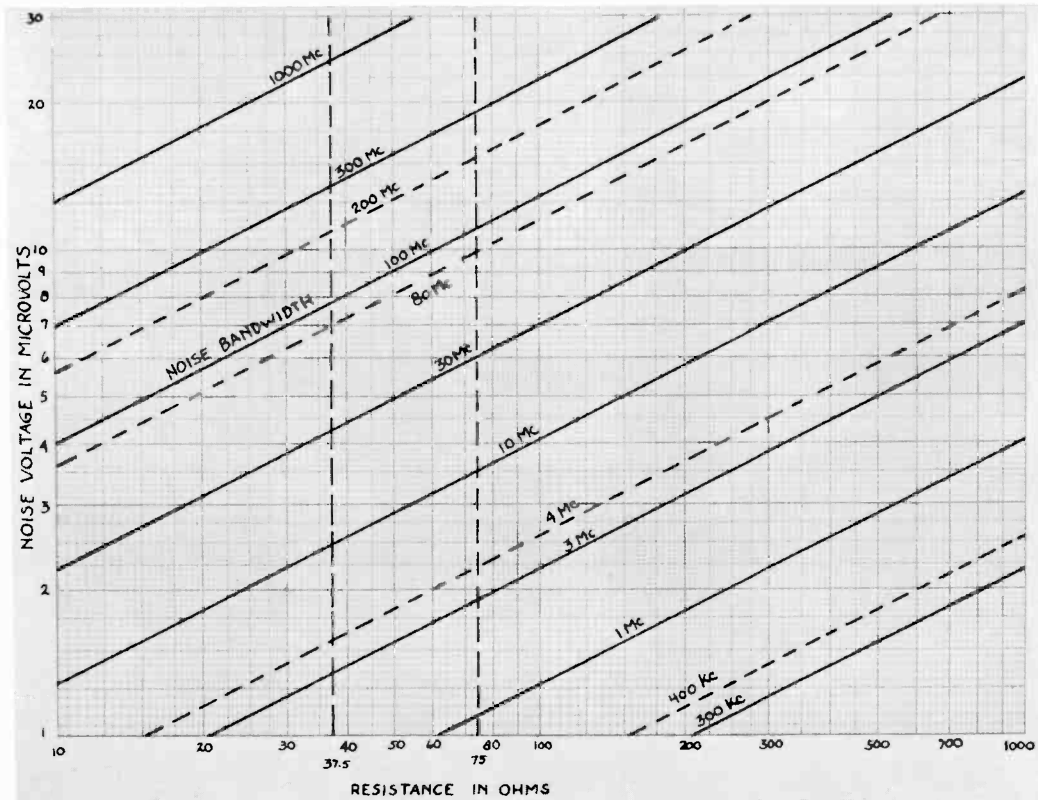


Fig. 3-3. Graphical determination of thermal noise.

amplifiers. The corresponding loss at the picture frequency of each channel is clearly indicated. It is obvious that the gain of a repeater amplifier operating at 25 db spacing must exactly match the cable attenuation at each channel; that is, the gain must be, for example, 12 db at Channel 2, while it is 25 db at Channel 13.

The accuracy with which this equalization can be achieved is today the limiting factor in cascading repeater amplifiers and in the realization of long trunklines. Previously, the limit for CATV system length was determined solely by signal-to-noise considerations. In order to achieve the proper frequency correction in cascaded systems, it is particularly important to provide for ease and precision of field adjustments by well-functioning amplifier controls and accurate test points. Another method which achieves similar results in based on full automation of all applicable functions in a CATV system (see Chapter 11).

After equalization or gain characteristic of an amplifier, next in importance are noise and overload, which determine the all-important dynamic range of the amplifier.

3-3. NOISE AND NOISE FIGURE

All electrical circuits generate noise to a varying degree. The most commonly known and unavoidable type of noise is thermal or Johnson noise, which is due to the irregular thermal motion of charge carriers such as electrons. The magnitude of this thermal noise depends on temperature, the number of charge carriers involved (which may be expressed in terms of ohmic resistance), and the electrical bandwidth over which the noise is observed.* The chart in Fig. 3-3 relates the actual root-mean-square noise voltage to resistance and bandwidth at normal room temperature. For example, a resistance of 75 ohms at a bandwidth of 4 MHz produces an RMS noise voltage of 2.21 microvolts at 75°F. This is the lowest possible noise value of an ideal resistor. Practical resistors may produce up to ten times this amount of noise.

In addition to thermal noise, several other types of noise are distinguished, such as shot noise, flicker effect, noise modulation, etc. The exact noise-producing mecha-

*For a mathematical treatment, see Appendix VII, Reference 7.

nisms in an amplifier are of no interest here and it is more convenient to speak of the total equivalent input noise level of an amplifier. A comparison of the equivalent input noise of an amplifier with the possible minimum noise level (thermal noise only) gives a measure for the noisiness of a particular amplifier. For a complete noise figure of merit, it is also necessary to consider changes in signal level which affect signal-to-noise ratio. The most basic and precise definition* for noise factor is based on the signal-to-noise degradation as the signal passes through an amplifier or circuit. For example, assume that a signal with a signal-to-noise ratio of 80 db is applied to the input of an amplifier. At the output of the amplifier, a signal-to-noise ratio of 70 db is observed. The signal-to-noise degradation in the amplifier is then 10 db, or the amplifier has a noise figure of 10 db. Noise figure is noise factor expressed in db. A circuit producing twice as much noise power than ideally possible has a noise factor of 2, or a noise figure of 3 db.

Noise factor may be directly calculated from the ratio of input signal-to-noise ratio to output signal-to-noise ratio. It is clear that every time a signal loss occurs with the noise unchanged, the noise factor also changes proportionately. For instance, the addition of a 6 db pad to the input of an amplifier automatically increases the noise figure by 6 db. Noise factor is also greatly affected by matching. The best noise factor results when the source works into a high impedance. However, this unmatched condition is not acceptable for the input circuit of CATV amplifiers, due to the severe requirements regarding reflections. With a terminated input circuit, the signal input power is reduced by 6 db, and, although the noise power due to the source is also reduced by 6 db, an equal amount of noise has been added in the termination, increasing the total noise power by 3 db. The signal-to-noise ratio under matched conditions is therefore reduced by 3 db; stating it differently: the best possible noise figure under matched conditions is 3 db. It is well to have a clear understanding of this point.

Noise sensitivity for different noise figures in matched 75-ohm systems can be readily calculated by considering that a 3 db noise figure corresponds to a matched 75-ohm system (a resistance of 37.5 ohms, using the

*Appendix VII, Reference 7.

values of Fig. 3-3). If, in addition, microvolts are converted to the more commonly used dbmv (db above 1 mv), a very important graph results—one of the essential design charts for all CATV systems and amplifier problems having to do with noise (see Fig. 3-4). For example, for the common TV bandwidth of 4 MHz, a noise figure of 10 db corresponds to an equivalent input noise level of -49 dbmv, a figure well worth memorizing.
NOTE: A CATV amplifier with a bandwidth from 50 to 220 MHz and equalized for 22 db of cable has a noise bandwidth of 80 MHz (which may be determined graphically). The 80-MHz line is also indicated in Figs. 3-3 and 3-4.

3-4. DISTORTION AND OVERLOAD

As the signal input level to an amplifier is increased, eventually a point will be reached where the amplifier ceases to behave as a linear circuit and distortion products not originally contained in the signal will appear. For simple CW input signals, the output will contain components described by the sums and differences of the integer multiples of the input signals. This distortion, called intermodulation, causes the well known beat-frequency effect and is the basis for mixers and frequency converters. It can be shown* that pure second order nonlinearities will produce ideal mixing action without affecting the modulation envelope itself—that is, without distorting the intelligence being transmitted. Therefore, intermodulation itself is normally of little concern in CATV systems, except in systems carrying more than 12 TV channels. Also, two cascaded second-order nonlinearities may produce countermodulation, a special form of cross modulation.

Third order nonlinearities produce a much more serious effect, called cross-modulation. When this form of distortion is present, the modulation of an undesired interfering signal appears as modulation of the desired signal, and it can in no way be separated from the desired modulation. In an individual TV channel, cross-modulation may cause sound interference in the picture. To avoid this effect, the sound carrier is usually operated at a much reduced level, because cross-modulation products are proportional to the square of the signal levels. With a single TV channel, distortion products then would be possible only between the various components of the video signal, and this level should be considered the ultimate

*Appendix VII, Reference 7,8.

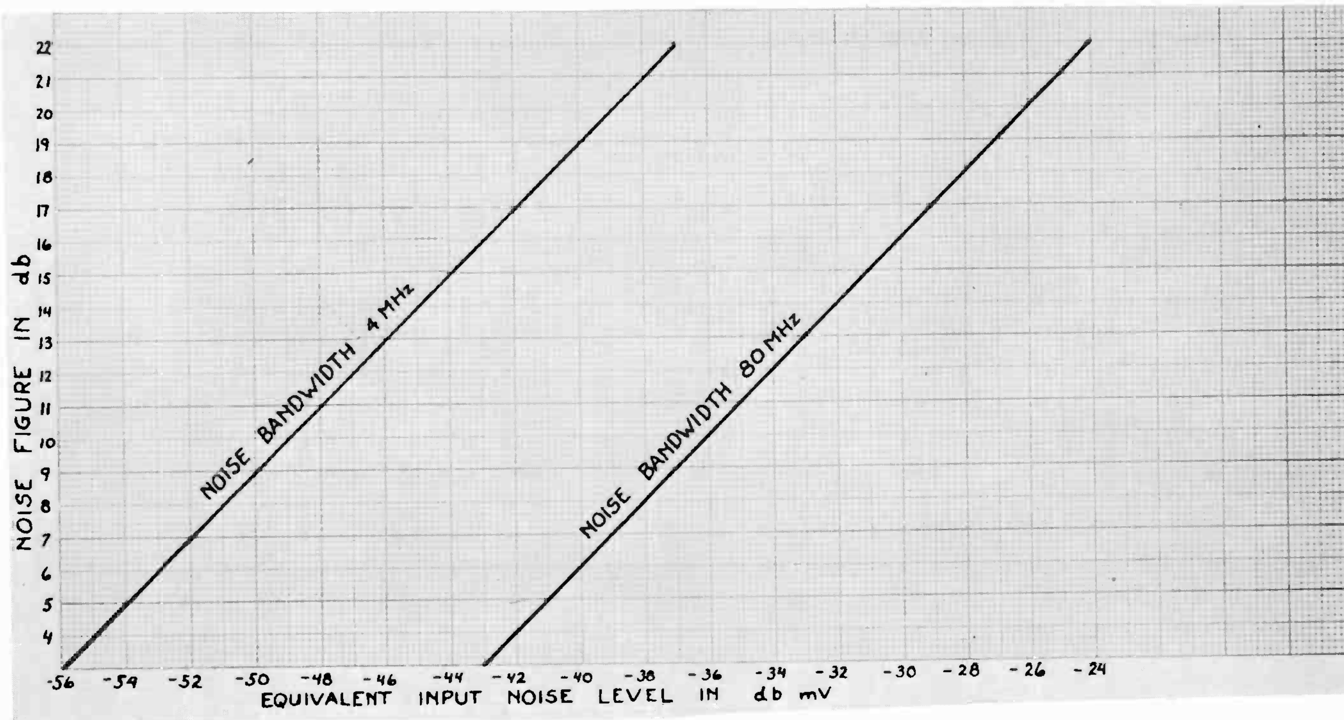


Fig. 3-4. Sensitivity chart for 75-ohm systems.

overload point of an amplifier. The single-channel overload level is often more precisely defined by compression of the sync pulse.

With multi - channel CATV systems, a much reduced overload level is found due to "windshield wiper" effect. This effect is caused by cross-modulation of the horizontal sync pulses of different broadcast stations. Consider two channels, A and B (Fig. 3-5), with a fixed 25- microsecond

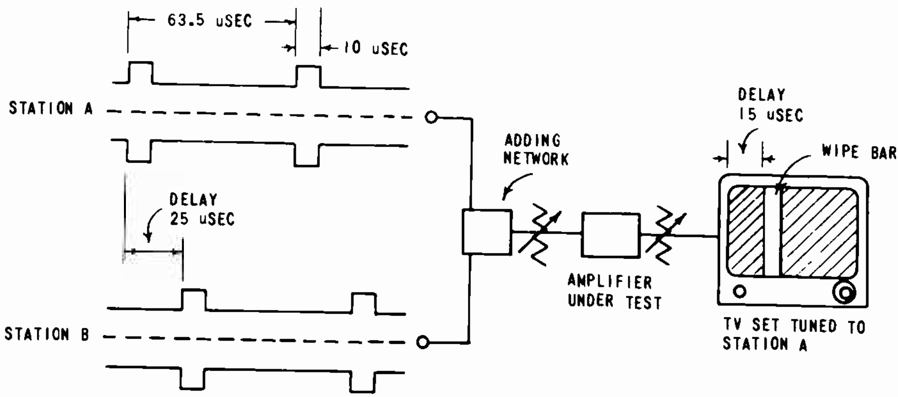


Fig. 3-5. Illustration of "windshield wiper" effect.

delay for the sync pulse of station B. If cross or counter-modulation takes place in the amplifier shown, the sync pulse of station B will be impressed as modulation on the video portion of signal A and appear as a vertical bar on a TV monitor. In reality, no fixed delay between the sync pulses of different

The "windshield wiper" effect illustrated in Fig. 3-5 is the "white wipe" commonly observed in practical systems. It is caused by countermodulation which is generally due to a different mechanism than that which is responsible for distortion measured at low levels which produces black bars.*

systems, and the maximum permissible output level of an amplifier is then called 12 - channel overload level. This amplifier overload level is the maximum permissible signal output level in all-band operation for a flawless picture.

The "windshield wiper" effect illustrated in Fig. 3 -5 is the "white wipe" commonly observed in practical systems. It is caused by countermodulation which is generally due to a different mechanism than that which is responsible for distortion measured at low levels which produces black bars.*

* See Reference 26, Appendix VII

3-5. AMPLIFIER DYNAMIC RANGE

In order to quickly evaluate different amplifiers and also to determine their best system operating gains, it is desirable to establish a figure of merit which can be directly related to simple amplifier measurements. Such an amplifier

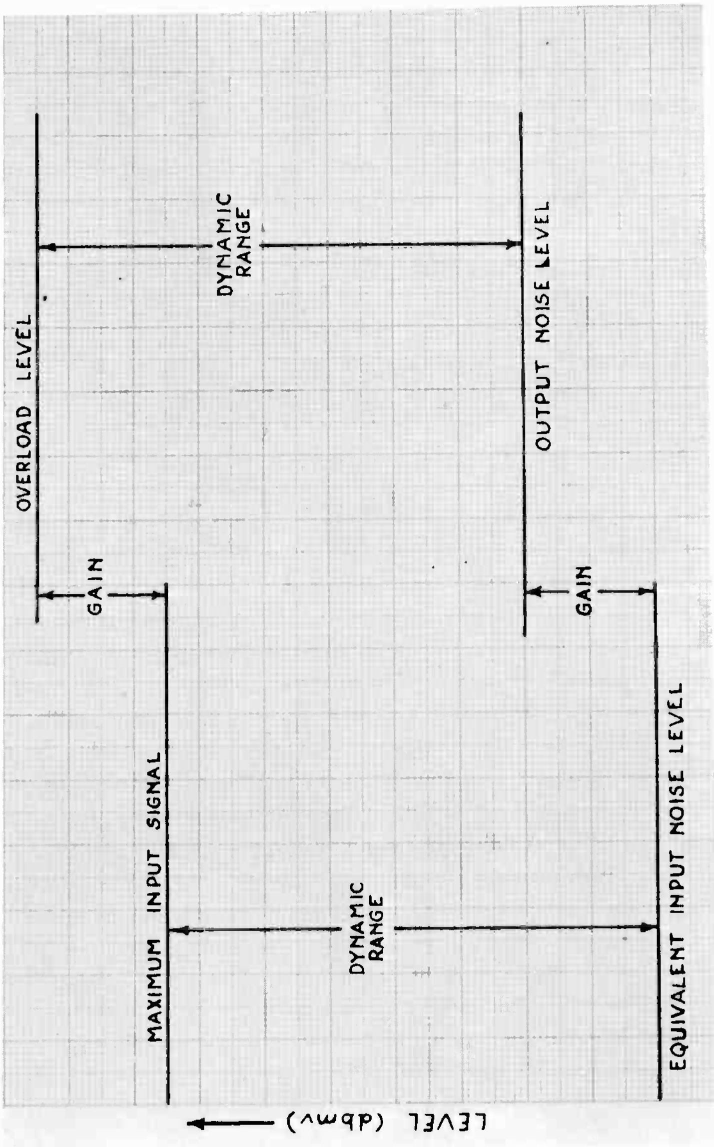


Fig. 3-6. Graphical explanation of dynamic range.

figure of merit relates of necessity to overload and noise levels only, because other characteristics such as equalization, matching, control functions, etc., have no direct theoretical limitations, and can be designed to any degree of perfection desired; also, they have no direct bearing on system performance.

Dynamic range is a term used in communication systems to indicate the range of signal levels that a particular piece of equipment or system can accommodate. The maximum permissible level is determined by distortion—that is, the onset of windshield wiping in CATV amplifiers or overload level. The minimum signal is limited by the noise level of the amplifier. Dynamic range is then identical to overload-to-noise ratio. Both overload and noise must, of course, refer to the same point in an amplifier. It will not do to compare input noise level with output overload. If input noise is used, then the maximum input signal level must be determined from the output overload level of the amplifier and its gain. Alternately, output overload level can be compared only with output noise level to determine dynamic range. Output noise level is obtained from equivalent input noise level and gain of an amplifier (Fig. 3-6).

Amplifier and system dynamic range are most important figures in CATV system design. Amplifier dynamic range is directly related to amplifier performance characteristics and is the decisive amplifier figure of merit. System dynamic range is related to amplifier dynamic range, spacing, and length of system, and reflects overall system performance (see Chapter 4).

Amplifier dynamic range is readily determined by measuring the overload level and noise figure for various channels, normally at the band edges (at Channels 2, 6, 7, and 13), and for different gain settings. Often, only the Channel-13 dynamic range is used because many times it is a worst-case figure. The detailed determination of dynamic range is as follows:

1. Align amplifier to flatness of ±0.25 db for the desired spacing (cable length) and determine gain at the desired TV channel.

2. Measure the noise figure of the amplifier under test at the desired channel. (See Chapter 13 for test procedures.)

3. Convert noise figure to equivalent input noise level by using Fig. 3 - 4 (4-MHz channel noise bandwidth).

4. Calculate output noise at the desired channel by adding the gain in db of the amplifier at that frequency.

5. Measure overload level at the desired channel (see Chapter 13).

6. Calculate amplifier dynamic range at the desired channel as output overload - to - noise ratio from Steps 4 and 5.

For example, suppose that an amplifier has been carefully aligned with 22 db of cable (at 211.25 MHz). The gain at Channel 13 is therefore 22 db (Step 1). Noise figure is now measured at Channel 13 and found to be 12 db (Step 2). From Fig. 3 - 4, equivalent input noise level is then -47 dbmv (Step 3). Output noise level is -47, +22, or -25 dbmv (Step 4). Next, it is necessary to measure output capability of the amplifier. This test should be performed under the same system tilt as used in the actual system (see Chapter 6). Let us say this amplifier is to be used in a half - tilted system and the overload level in half-tilt is 50 dbmv at Channel 13 (Step 5). The dynamic range of the amplifier is then found at 75 db.

Similarly, an amplifier with an overload level of 55 dbmv and a noise figure of 12 db, both measured when aligned for 35 db cable (35 db gain at Channel 13), is calculated as follows:

Input noise level from Fig. 3-4:	-47 dbmv (Step 3)
Output noise level with 35 db gain:	-12 dbmv (Step 4)
Dynamic Range with 55 dbmv overload level:	67 db

The second example—typical for a vacuum-tube amplifier—clearly shows how performance of an amplifier with otherwise good parameters (55 dbmv overload level, 12 db noise figure) is severely degraded by excessive gain. (More on this in Chapter 7.) Compare with the parameters and the dynamic range of the first example.

NOTE: Dynamic range or overload - to - noise ratio should not be confused with signal-to-noise ratio, which is affected directly by signal level rather than by amplifier performance.

Obviously, the maximum signal-to-noise ratio which can be achieved is equal to the dynamic range. However, the actual signal operating level in a system depends on cascading considerations which are treated in Chapter 6.

3-6. CASCADED FIGURE OF MERIT

Dynamic range is an ideal figure of merit for an individual amplifier or piece of communications equipment. It is found that as several such pieces of equipment are used in cascade, overall dynamic range is reduced (see Chapter 4).

Table 3-1. Correction for cascaded amplifier figure of merit. For 1000 db system subtract an additional 32 db.

db Gain	db Correction
10	−7.9
12	−6.4
15	−4.5
18	−2.9
20	−1.9
22	−1.2
25	0
28	+0.9
30	+1.6
35	+2.9

The longer the cascade, the greater the reduction in system dynamic range. Ultimately, the performance quality of an amplifier must be judged by the system dynamic range which can be achieved with this unit. The amplifier which results in the largest dynamic range for a given system—of, say, 500 db—would have the highest figure of merit. For an amplifier having a different gain (spacing), a small correction must be applied to its dynamic range in order to have a precise cascaded amplifier figure of merit. This correction is necessary because a different number of amplifiers, and with it different system derating, is needed for a given total system length, depending on the gain of the unit.

For example, in a 500 db system, it will take 25 amplifiers at 20 db gain, or only 20 amplifiers at 25 db gain. The derating for 25 amplifiers in cascade is somewhat greater than for 20. This correction is made readily by

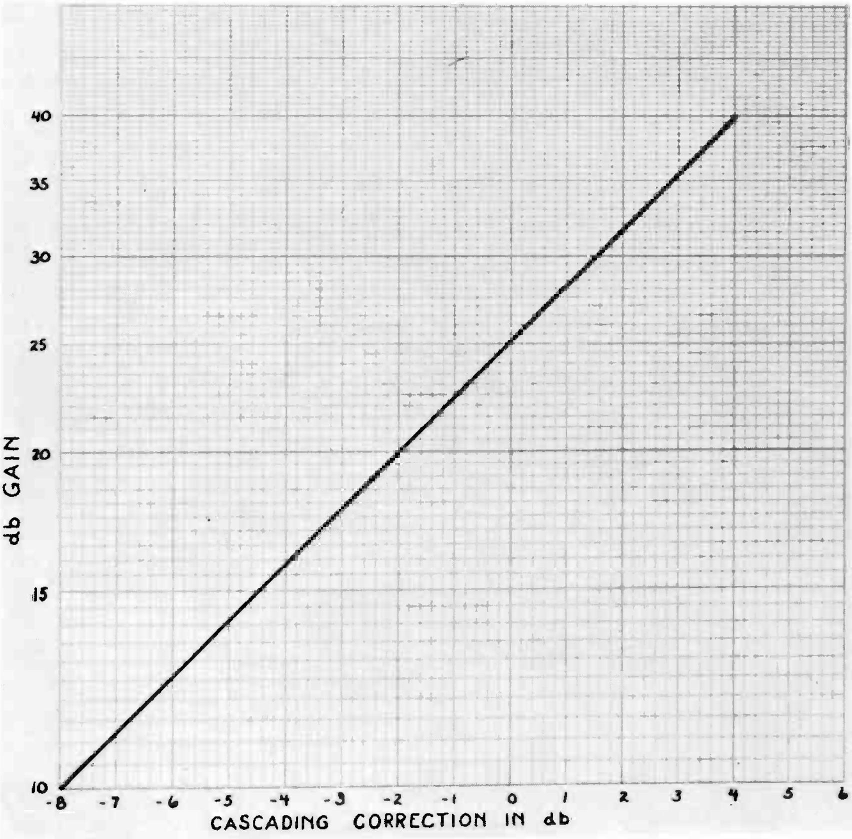


Fig. 3-7. Correction chart for cascaded amplifier figure of merit.

using Table 3-1 or Fig. 3-7. The cascaded amplifier figure of merit so obtained is correct for a CATV system of any length and represents a valuable yardstick for comparison of different CATV amplifiers.

Comparing the two amplifiers from our previous example, we find for the first amplifier with a dynamic range of 75 db and 22 db gain, a correction of -1.2 db, or a cascaded figure of merit of 73.8 db. The second amplifier with a 5 db higher overload level had a dynamic range of 67 db and 35 db gain. The cascading correction is +2.9 db and the cascaded figure of merit is 69.9 db. Therefore, the first amplifier is 3.9 db better in overall system performance than the second. Expressed differently, the first amplifier will permit a 56% longer system to be built for the same signal-to-noise ratio.

CHAPTER 4

Cascaded Amplifier Systems

After the dynamic range of an amplifier is determined, the question arises as to the combined dynamic range for a complete system made up of a cascade of amplifiers. Considering noise first, it is clear that at each repeater amplifier input the same equivalent noise power is added, assuming identical noise figures for the amplifiers, while the signal level is unchanged. Therefore, for two amplifiers the noise power is doubled or increased by 3 db; for 10 amplifiers the noise power is ten times that of an individual amplifier or 10 db higher. The increase in noise power is simply proportional to the number of amplifiers and is readily expressed in db. For amplifiers with different individual noise figures, a separate calculation must be made.* However, in most cases it is sufficient to use an average noise figure and to derate from there. This is particularly true for cascades of 10 or more amplifiers.

4-1. SYSTEM DYNAMIC RANGE

A similar derating must be made for the overload level. Mathematically, a quite different relationship results** and this must be taken into account if only a few amplifiers are operated in cascade, or for amplifiers with different individual overload levels. However, for longer cascades of 10 or more amplifiers, sufficient accuracy results by using the average overload level and derating from this value, using the same relationship as for noise. This procedure is justified because it has been demonstrated that distortion products in CATV amplifiers behave as if fully uncorrelated.

The degradation of dynamic range in a system of cascaded repeater amplifiers is clearly visible in a system de-

*Appendix I, and Reference 7, Appendix VII.

**Reference 18, Appendix VII.

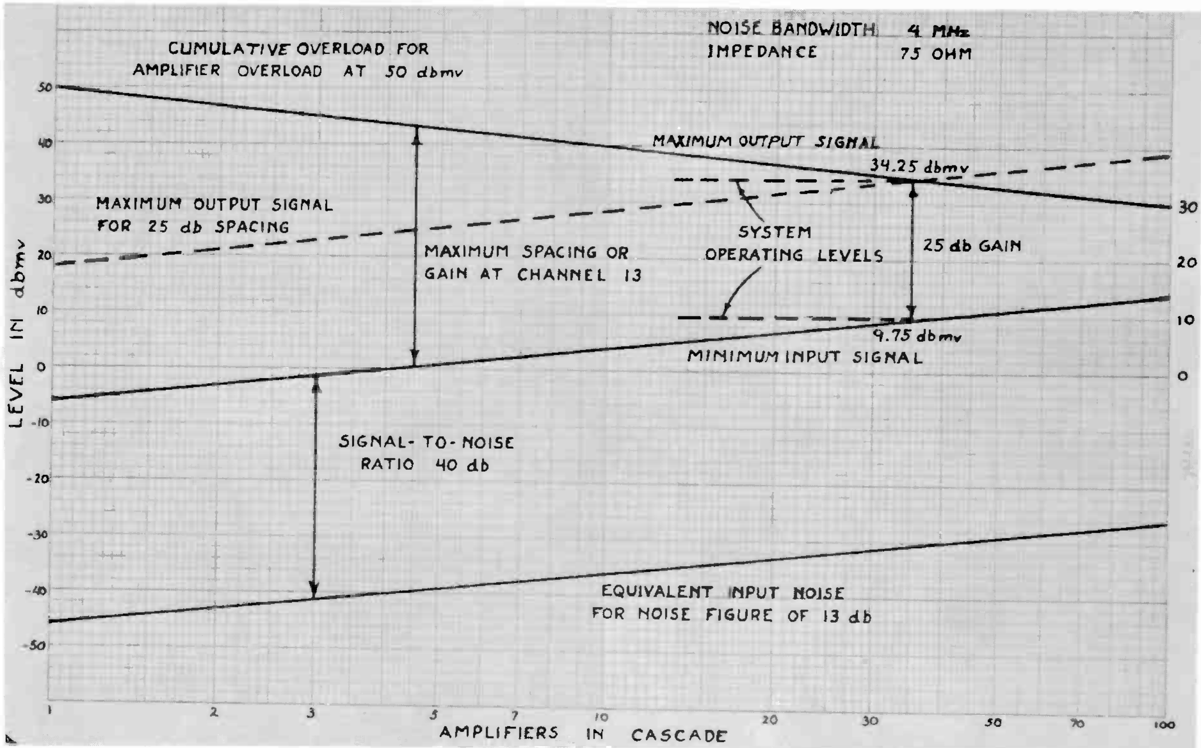


Fig. 4-1. System derating diagram for the amplifier of example.

rating diagram (Fig. 4-1). For simplicity, identical amplifiers are assumed with a noise figure of 13 db and an overload level of 50 dbmv. These values are typical for transistorized amplifiers of moderate quality.* A noise figure of 13 db corresponds to an equivalent input noise level of -46 dbmv (from Fig. 3-4). The total input noise level can then be plotted for any number of amplifiers; it increases, for example, to -26 dbmv for 100 amplifiers in cascade.

It has been determined (see Chapter 1) that for a flawless TV picture of studio quality, a video signal-to-noise ratio of 40 db is required, and that a 30 db figure would suffice for a very good picture. These figures are not to be confused with readings taken with some narrow bandwidth field strength meters. Consequently, a minimum input-signal line may be drawn 40 db above the input noise level. The maximum output signal is determined by the overload level and is derated for the system of cascaded amplifiers according to the line shown. The difference between the minimum and maximum signal lines indicates the range available for amplifier gain as well as for safety margins to allow for errors in level settings and other factors. For example, with 30 amplifiers of this type operating in cascade, the total allowable signal range would be 26 db. The amplifier gain might be, therefore, 25 db, and the safety margin 1 db—0.5 db on each end. The recommended system levels would then be 9.75 dbmv at each amplifier input and 34.25 dbmv at each amplifier output. These levels would apply at the highest frequency (highest amplifier gain) in the system—that is, at Channel 13 in an all-band CATV system.

4-2. MAXIMUM AMPLIFIER AND SYSTEM GAIN

It is clear from such a system derating diagram that as the number of amplifiers is increased, the maximum permissible amplifier gain decreases sharply. Thus, with one amplifier, a maximum gain of 56 db would be permissible, while only 16 db could be used for a cascade of 100 amplifiers (disregarding safety margins for the moment). As

*This amplifier is used here for instructional purposes. A derating diagram for CATV amplifiers of advanced design is given in Chapter 6.

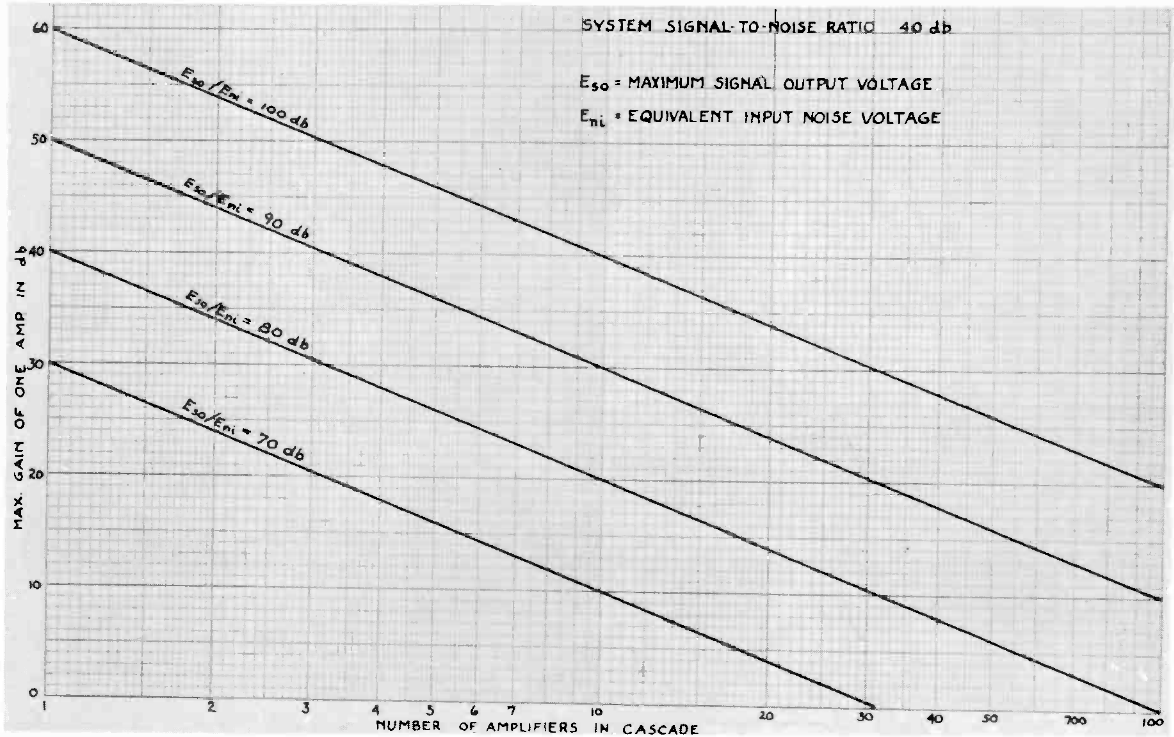


Fig. 4-2. Derivation of optimum amplifier gain.

far as the actual system length goes, with the first amplifier, the system length and gain was 56 db, while in the second case, the gain for the total system was 1600 db. In both cases, the signal-to-distortion and -noise ratio after the last amplifier is the same. From this example, it becomes clear that a single high-gain amplifier leads to far inferior performance, allowing only a very short system length, running just as close to overload and noise levels as the longer cascade of many low-gain amplifiers with a total system length of 1600 db. The obvious question is then: Is there an optimum amplifier gain which will produce the highest signal-to-noise ratio in a given system? Stated differently: Is there an optimum spacing of repeater amplifiers which results in the greatest system length for a given system signal-to-noise ratio?

4-3. OPTIMUM SPACING AND GAIN

The question of optimum spacing of repeater amplifiers is of the greatest importance for the development of a sound concept for the overall CATV system, as well as for the actual electronic design of the CATV amplifiers. At first glance, it appears that optimum gain would be different for amplifiers with different overload-to-noise levels. To eliminate gain from our previously used figure of merit, it is desirable for this discussion to use the ratio of overload level to equivalent input noise to describe the quality level of a particular amplifier. This is a valid procedure for many amplifiers because the noise figure is largely affected by the input stage, and the overload level is determined primarily by the output stage, with both values reasonably independent of the actual gain setting of the amplifier. (More on this in Chapter 7.)

Using this amplifier characteristic number and an ultimate overload-to-noise ratio of 40 db, the cascaded system degradation with an increasing number of amplifiers may be plotted (Fig. 4-2). The parameters of the straight lines are determined by the quality of the particular amplifier, with values of 100 db or higher representing very good amplifiers, when compared at the same gain settings. The maximum permissible gain for each amplifier is indicated in the left-hand scale. For example, for an amplifier with a characteristic number of 90 db, the maximum pos-

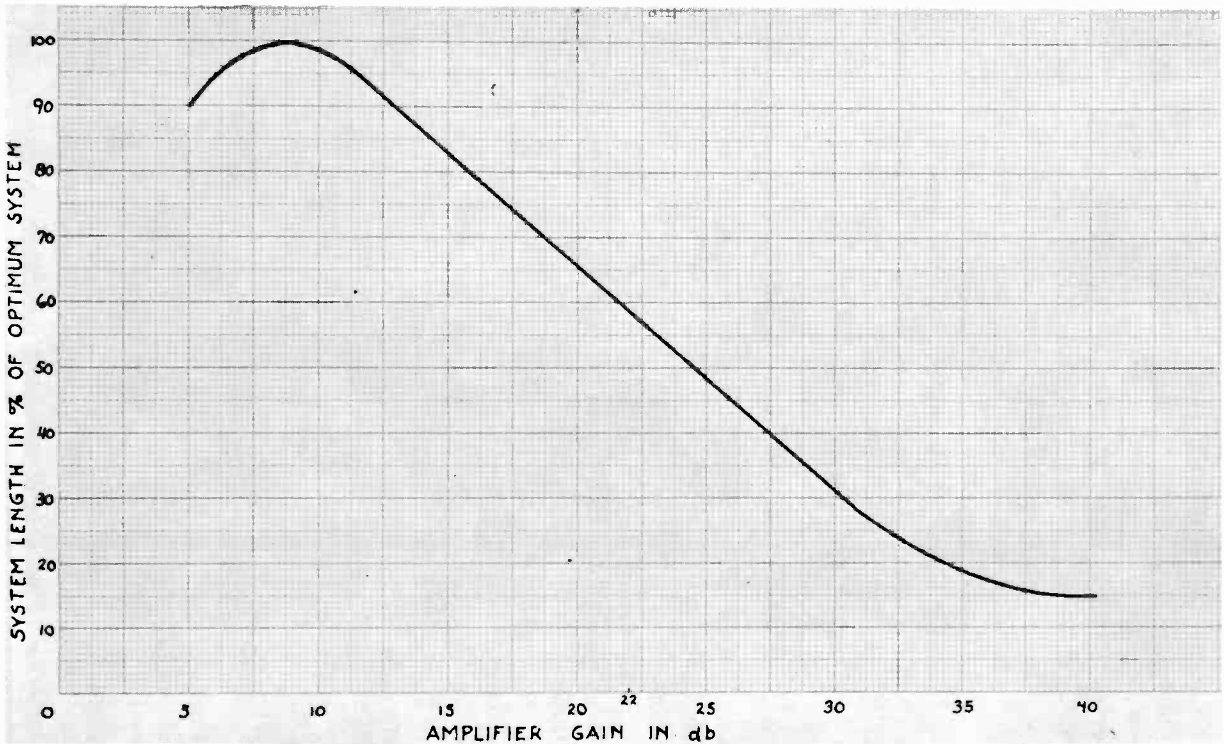


Fig. 4-3. Maximum system length as a function of amplifier gain. Constant output signal-to-noise ratio, single-stage amplifier.

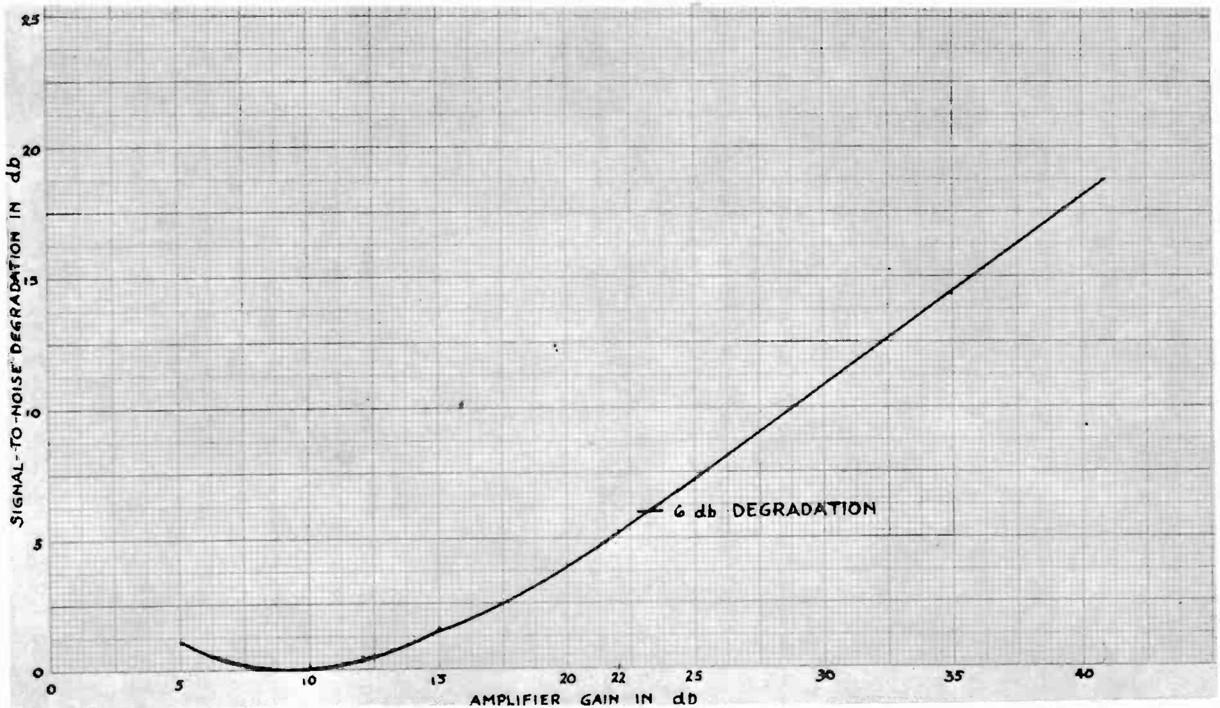


Fig. 4-4. System signal-to-noise degradation as a function of amplifier gain. Total system gain constant, single-stage amplifier.

sible gain for a cascade of 10 amplifiers is 30 db; for 30 amplifiers, 21 db; and for 100 amplifiers, 10 db (in order to have a signal-to-noise ratio of 40 db at the end of the system). The interesting point is, however, the greatly varying system lengths for the various cases, achieved at the same system quality (signal-to-noise ratio). This is best shown in Table 4-1.

This table indicates that a peak occurs somewhere between 10 and 4 db amplifier gain. Repeating this process with amplifiers of different characteristic numbers, exactly the same optimum gain is found. A mathematical analysis (Appendix II) shows this optimum gain to be 8.69 db. Since the actual noise figure, overload level, and other amplifier characteristics drop out, the result is valid for any type of repeater application, such as microwave,

Table 4-1. System lengths with amplifier characteristic number of 90 db for system signal-to-noise ratio of 40 db.

Number of Amplifiers	Maximum Gain Per Amplifier (db)	System Length (db)
1	50	50
10	30	300
30	21	630
100	10	1,000
200	4	800
300	0	0

oceanic cables, CATV, etc. This optimum gain is rather low, and there may be economical and practical reasons for a different spacing. It is therefore necessary to examine how sharply this optimum gain peaks; that is, the extent system performance would suffer if gains other than the optimum were used. It is a relatively simple matter to compute the actual possible system length, in terms of the ideally possible, for other than optimum amplifier gain (Fig. 4-3), or in terms of system signal-to-noise degradation (Fig. 4-4). These curves indicate a fairly shallow peak and no rapid degradation in performance for reasonable deviations from the optimum. As indicated in Table 4-2, it appears that a gain of 15 db causes an acceptable degradation, while spacings in excess of 25 db should be avoided.

Before coming to any hasty conclusions, it must be kept in mind that in CATV work, due to the necessary equalization required in each amplifier, it is not possible to meet the optimum gain requirement for each TV channel. A compromise must therefore be chosen. For example, by using the method of "splitting the error," the gain at Channel 2 is chosen below the optimum, and at Channel 13 above the optimum—say 9 db at Channel 2 and 19 db at Channel 13. Some compensation might then be attained by better amplifier performance at Channel 13, so that no net loss in system dynamic range results.

4-4. LIMITATIONS OF SPACING THEORY

The spacing theory discussed so far is based on certain assumptions, it it now necessary to examine these boundary

Table 4-2. System degradation for other than optimum spacings.

Spacing or Gain	System Length (%)	Signal-to-Noise Degradation (db)
8.68	100	0
10	98	0.2
15	83	1.6
20	63	4.0
25	43	7.0
30	29	11
35	18	14

conditions more closely. The results given so far—that is, an optimum gain of 8.68 db regardless of amplifier figure of merit—are valid for the case that both noise figure and overload level are independent of the actual gain setting of the amplifier. These conditions are normally met by amplifiers, since the gain control is usually located after the noise producing stages and ahead of the output or distortion producing stages. Therefore, a change in the setting of the gain control has no effect on noise figure and overload in such an amplifier.

However, when we are talking of a total amplifier gain on the order of 10 db or less, it is obvious that no clear distinction can be made between noise and distortion producing stages. With a very small individual stage gain, the noise contribution of the second stage must be considerable, thereby degrading the overall noise figure with decreased gain, and similarly, distortion products in the first stage reflect in a reduced overall overload level.

It is therefore clear that the optimum gain derived above applies strictly to single-stage amplifiers only. While multistage amplifiers normally meet the conditions set forth, this is certainly not the case when the gain drops too low. Single-stage amplifiers for CATV systems are impractical. The many conflicting requirements which must be met, such as match, noise figure, overload, equalization, gain, temperature control, etc., cannot be achieved presently in a single stage. But even if such an amplifier could be constructed, the need for a separate, high-quality regulated power supply makes this concept economically unsound.

It seems, therefore, that practical CATV amplifiers will have from two to four stages of amplification, depending on control functions such as AGC, or auxiliary characteristics such as bridging outputs. As the state of the art advances, fewer stages will be need to accomplish the needed performance standards.

The strict mathematical treatment for multistage amplifiers becomes somewhat involved. It can be shown (Appendix II) that for the 2 - stage amplifier, optimum gain is 11.0 db, assuming identical transistors or tubes in both stages (Table 4 - 3). It can also be shown that a system built with 2-stage amplifiers has a system dynamic range of 2.5 db below that attainable with single-stage amplifiers. This is certainly an acceptable degradation, considering the greater economy of the 2-stage circuit.

For more than two stages, a mathematical treatment is possible, but quite involved. Also, the determination of constants poses difficulties. For example, it is not possible to measure individual noise and overload levels for each stage to the required degree of accuracy. However, it is quite possible to compute optimum gain for a few cases such as identical stages, or higher overload and noise for the output stage, etc. In this fashion a good insight into the possible range of optimum gains is obtained, into which all practical circuits must fall. However, this analysis should only be considered a guide in amplifier and system design. After a new amplifier has been fully developed, optimum spacing for this particular amplifier can be determined quite rapidly from simple measurements, and a comparison with the theory then indicates the quality level achieved.

4-5. DETERMINATION OF OPTIMUM SPACING FROM AMPLIFIER MEASUREMENTS

For any given amplifier, it is a simple matter to determine the optimum gain setting, that is, the spacing which will result in the best system dynamic range or stated differently, a spacing which leads to the longest system for

Table 4-3. Optimum gain (spacing) for various amplifiers.

Amplifier Type	Optimum Gain	Gain for 90% Maximum System Length
One stage	8.69 db	13 db
Two identical stages	11.0 db	15 db*
2nd stage 6 db higher overload and noise	15.0 db	19 db*
Three identical stages	13.0 db*	16 db*

*Estimate

a constant system dynamic range. The measurements are made on a number of amplifiers to avoid inconsistencies due to production spread. All measurements are best combined in a table for ease of computation, such as Table 4-4.

The objective is to measure the overload-to-noise parameters at each channel to determine amplifier as well as system performance. These measurements are made over the useful gain range of the amplifier, the range over which the amplifier can be made to equalize the corresponding length of cable. In Table 4-4, 5 db steps of gain were taken for an amplifier which had a total gain range from 15 to 30 db for an equalized flatness of ±0.25 db. If so desired, closer steps may be taken, such as 3 db; however,

Table 4-4. Determination of dynamic range and optimum spacing for high-quality transistorized amplifiers.

Ch	Noise Fig. db	Input Noise dbmv	Gain db	Output Noise dbmv	Overload Level dbmv	Dynamic Range db	Cascaded Fig. of Merit db	100C db System Dynamic Range db
2	14.2	-44.8	7	-37.8	47	84.8	80.3	48.3
13	10.8	-48.2	15	-33.2	53	86.2	81.8	49.8
2	12.3	-46.7	9.3	-37	50.5	87.5	85.2	53.2
13	9.8	-49.2	20	-29.2	57.5	86.7	84.8	52.8
2	11.5	-47.5	12	-35.5	51	86.5	86.5	54.5
13	9.5	-49.5	25	-24.5	59	83.5	83.5	51.5
2	11.5	-47.5	14.5	-33	51	84.0	85.6	53.6
13	9.5	-49.5	30	-19.5	59	78.5	80.1	48.1

Note: For 1000 db system subtract 32 db from cascaded figure of merit.

5 db steps are usually sufficient. The exact optimum spacing may then be determined graphically or by interpolation. In Table 4-4, the amplifier is first aligned with 15 db of cable (at Channel 13) using all adjustments provided for the flattest possible response. A tolerance of ±0.5 db is sufficient for this measurement, although this would be inadequate for use in a system. The test is best performed by using the alignment methods described in Chapter 13. After the completion of alignment, noise figure and overload are read for Channels 2 and 13. These measurements are recorded in columns 2 and 6 of Table 4-4. Thereafter, the procedure is repeated for an alignment at 20, 25 and 30 db of cable, etc. Again, the flattest possible response must be achieved in all cases.

Now the system dynamic range is determined as follows:

1. Noise Figure is converted to equivalent input noise using Fig. 3-4 and the 4-MHz line. The value so obtained is recorded in column 3.

2. For output noise in column 5, add column 3 and 4. Subtract columns 5 and 6 to obtain amplifier dynamic range in column 7.

3. After that, the cascaded figure of merit is obtained by adding the correction factors of Table 3-1 to column 7 to obtain column 8.

Instead of the last step, it is possible to calculate the dynamic range for a system length which is an integer multiple of the gain. For example, a ready comparison at Channel 13 is possible by calculating a 300 db system as shown in Table 4-5. The second column is obtained from Column 7 of Table 4-5. The third column figures are chosen so that the total system length is 300 db in all cases. Degradation for cascade is 20 log the number of amplifiers, the standard derating factor for cascaded systems. The difference of columns 2 and 4 is recorded in column 5 as dynamic range of the 300 db system. The difference in readings in column 5 is maintained, regardless of the length of the system (300 db, 600 db, 3,000 db, etc.). For a

Table 4-5. Calculation of 300 db system at Channel 13.

Amplifier Gain db	Amplifier Dynamic Range* db	Number of Amplifiers	Degradation for Cascade db	System Dynamic Range db	Relative Change db
15	86.2	20	26.0	60.2	3
20	86.7	15	23.5	63.2	0
25	83.5	12	21.6	61.9	-1.3
30	78.5	10	20	58.5	-4.7

*Channel 13

3,000 db system, simply subtract 20 db from each value in column 5, leaving the difference between values unchanged. This difference (column 6) is, of course, the important final result, which indicates at which gain setting or spacing the best system performance is possible for the particular amplifier type. It can also be obtained directly from column 8 of Table 4-4, where correction factors were used to obtain system performances. The correlation of this method of determining optimum spacing and system performance to the amplifier dynamic range, discussed earlier, is obvious. A meaningful comparison between different amplifiers may then be made, and the optimum gain setting (spacing) may be determined for a particular amplifier as well. For example, the amplifier of Table 4-4 and 4-5 should be spaced at 20 to 25 db.

CHAPTER 5

Practical Aspects of Spacing

As we have seen, closer spacing is desirable in the interest of improved system dynamic range. However, there are a few myths concerning closer spacing which should be dis-spelled from the start.

5-1. COST AND RELIABILITY

A point frequently made is that closer spaced systems are more costly, since more amplifiers are used. This would be true if the amplifier type were the same. Actually, a closer spaced system runs lower in cost. This is so because of the considerably lower cost for the amplifier, which is, however, not reflected in a lower price with some manufacturers.

Typically, going from a spacing of 30 db down to 20 db requires 1.5 times as many amplifiers. However, an excellent amplifier designed for 20 db spacing can be manufactured at perhaps one-fourth the cost of a 30 db amplifier. The cost of amplifiers for such a closer spaced system is therefore lower by a factor of 2.7 or equal to 37% of the cost of a system spaced at 30 db.

The reasons for the much lower cost of a 20 db amplifier are tied directly to the "theory of diminishing returns" in wide-band amplifier design. As a second amplifying stage is added in a wide-band amplifier, total bandwidth is reduced. In order to achieve the original bandwidth requirement for the whole amplifier, the bandwidth of the individual stage must be widened in excess of the overall requirement. This widening must take place in trade for a reduction in gain. Adding a second stage has therefore not produced twice the gain. With each additional stage, the increase in gain for a constant overall bandwidth becomes less, until eventually a point is reached where adding another stage actually decreases the gain.* Long before this point is reached, however, extra effort must be spent for every extra db of gain. This leads to a con-

*Appendix VII, Reference 19.

siderably more complicated and costly amplifier for the higher gain version. Also, because of the disproportionately greater number of transistor stages, both noise and distortion are considerably higher.

Another important factor is reliability. One argument makes the point that with fewer amplifiers, reliability should be higher; hence, a closer spaced system would be less reliable. This argument is correct if the same amplifier type but with different gain settings were used in both systems. However, this should not be done and a different amplifier should be used instead.

Typically, a closer-spaced 20 db system would use a 2-stage amplifier and a 30 db system would use a 4-stage amplifier. Extensive studies indicate that reliability is related simply to the number of components and their individual reliability. Thus, the same reliability results if a 4-stage circuit is split into two 2-stage circuits, since the same number of components is involved. However, as far as the overall system is concerned, fewer total electronic circuits are needed in closer spacing, since only 1.5 times two stages are used in the 20 db system as compared to four stages in the 30 db system for the same overall gain. As a result, the reliability of the closer spaced system is higher.

For exactly the same reasons, current drain is less in a closer spaced system. A high-gain amplifier uses proportionately more stages than the increase in gain justifies, and therefore has an excessive current drain.

All these factors are borne out by the amplifiers available today, except for the price which is often not related to manufacturing cost. The spacing values of 30 and 20 db were used only as examples to illustrate the point; however, they closely represent the spacing values of modern systems. Optimum spacing leads to much improved system performance, as we have seen in Chapter 4, and there is no objection to the required closer spacing from the standpoint of cost, reliability, and current drain.

5-2. TRANSISTORS VS TUBES

Thus far, all of our theoretical conclusions pertaining to optimum system concepts are generally true, regardless of amplifier details related to solid-state vs. vacuum-tube circuits. However, a comparison between transistorized and tube equipment is instructive and warrants a detailed discussion.

In earlier days, vacuum-tube amplifiers were the only type available for CATV systems. Due to the cost of bringing in line power to every amplifier, spacing between amplifiers was increased to the maximum, often as high as 35 to 45 db. As we have seen, this leads to inferior system performance, which must be offset by equipment of extreme dynamic range. Indeed, tube technology permitted the design of amplifiers with overload levels of 60 dbmv, and noise figures of 15 db over the whole frequency range from 50 to 220 MHz. As calculated in Chapter 3, the input noise for such an amplifier would be -44 dbmv; output noise would be -9 dbmv, for a gain of 35 db. The dynamic range would then be 69 db. This relatively high value is needed because the excessive spacing causes a rapid degradation in a cascaded system. The degradation due to spacings other than optimum was covered in detail in Chapter 4. As we shall see, this tube system is inferior in many respects to a well designed transistor system.

With the advent of transistors, cable powering became possible. Line power is applied to the cable at few strategically located remote powering points. The obvious merit of this system is economy, because only very few power stations along the cable must be provided. With cable powering, complete freedom as to spacing and other system concepts exists. Obviously, as far as the cost of powering is concerned, it does not matter if amplifiers are spaced 15, 25, or 35 db apart.

With this concept, optimum spacing may be used. It is now possible to design amplifiers with an added degree of freedom, because obviously a lower dynamic range for the amplifier is allowable (possibly even desirable, as we shall see) if system degradation due to non-optimum spacing is held down.

This interplay of the various amplifier parameters and overall system performance is best shown in an example. A typical, older transistor amplifier has an overload level of 50 dbmv (note that this is 10 db less than for the tube amplifier) and a noise figure of 10 db (5 db less than for the tube amplifier). The equivalent input noise is then -49 dbmv and output noise with 25 db gain, is then -24 dbmv. The dynamic range for the transistor amplifier is then 74 db, while the dynamic range for the tube amplifier was only 69 db due to the higher gain of the tube amplifier.

However, since more transistor amplifiers must be cascaded, it is necessary to calculate a whole system for a valid comparison. Thus, five amplifiers at 35 db gain, or seven amplifiers at 25 db gain, make a 175 db system. Dynamic range for the 175 db system is then 55 db for the tube system and 57 db for the solid-state system. The comparative figures are given in Table 5-1. As shown, this particular solid-state amplifier beats the tube amplifier by 2 db; this difference remains the same, regardless of actual system length, as the case of the 1750 db system shows.

Actual transistor amplifier specifications today are far better than the values given in Table 5-1 and now even exceed the output capability of tube amplifiers. The important point, however, is that the tube amplifier with a 10 db higher output capability above the older transistor amplifier used in our example, still resulted in poorer overall system performance due to the greater deviation from optimum spacing. It is therefore meaningless to ask blindly for a higher overload level. <u>A higher overload level coupled with higher gain does not lead to improved system performance!</u> This is well shown in the example. It is not even desirable to have a higher overload level without limit. Generally, twice the power output capability requires twice the power dissipation (or close to it) and with it, increased current drain. While it is quite possible to design an amplifier with ten times the output capability, this generally results in a tenfold increase in current drain, which may be undesirable. Therefore, output capability must be matched to both gain and current drain to be meaningful. A truly improved amplifier achieves a higher overload level at no increase in current drain or in spacing.

These factors were not recognized when some of the first transistorized amplifiers were designed. First attempts concentrated on simply copying specifications for tube equipment, without taking advantage of the freedom allowed in the spacing of transistorized amplifiers. This, together with the poor performance of early transistors,

Table 5-1. Comparison of tube and transistor systems.

	Tube System	Transistor System	
Amplifier Overload	60 dbmv	50 dbmv	50 dbmv
Noise Figure	15 db	10 db	10 db
Gain (spacing)	35 db	25 db	17.5 db
Input Noise	-44 dbmv	-49 dbmv	-49 dbmv
Output Noise	-9 dbmv	-25 dbmv	-31.5 dbmv
Amplifier Dynamic Range	-69 db	74 db	81.5 db
Number of Amplifiers in 175 db System	5	7	10
Derating for Cascade of Amplifiers	14 db	17 db	20 db
Dynamic Range for 175 db System	55 db	57 db	61.5 db
Dynamic Range for 1750 db System	35 db	37 db	41.5 db

led to far inferior systems as compared to tubes. When advantage was taken of closer spacing with transistorized amplifiers, system dynamic range increased above the value possible with tubes even though amplifier output capability was inferior to that attainable with tubes. Today, solid - state amplifiers have achieved a dynamic range far greater than was ever possible with tubes. The improvement in system dynamic range is even more drastic now so that tubes remain only in some of the older CATV systems. In Table 5-1, a spacing of 17.5 db for the transistor amplifier is shown in the last column. This closer spacing, which is entirely possible with cable - powered equipment, results in the example in a further improvement of 4.5 db in system dynamic range. However, for practical amplifiers optimum spacing should be determined following the procedure given in Chapter 4, because overload and noise figures are a function of gain and do not remain constant as in the example of Table 5 - 1. This will be explained more fully in Chapter 7.

Besides the greatly increased system dynamic range with transistors, the following further advantages are also inherent in solid - state equipment: greater reliability, reduced current drain, reduced maintenance, hermetically-sealed packaging, no replacement cost, low-cost circuitry, small size and weight.

Reliability of transistors is evidenced by the elimination of vacuum tubes in military designs where high reliability is a prerequisite. In this connection, it should be mentioned that the transistors used in low - cost portable radios and similar equipment usually are second-rate devices with reduced reliability. By contrast, transistors designed into amplifiers for Cable TV or other industrial and military applications are premium devices. Apparent poor transistor reliability is often caused by the user of transistorized equipment. Although all transistor manufacturers have warned repeatedly against removing a transistor from a "hot" socket, this advice frequently is not heeded and transistors are pulled from sockets without first turning off the power. The result is often a damaged transistor, called a "drifter," with seemingly unchanged performance. However, the life of the transistor is now reduced to a few hundred hours instead of the several 100,000 hours which have been demonstrated so far by transistor manufacturers. A transistor damaged in this fashion is most easily recognized by a jump in leakage current from five to ten times the initial value. For greatest reliability, modern CATV equipment avoids transistor sockets altogether and transistors are soldered directly into the circuit board like all other components.

5-3. PRACTICAL LIMITATIONS TO SYSTEM LENGTH

The ultimate limit to system length depends entirely on the dynamic range which can be maintained at the end of the system. Disregarding variable effects such as temperature changes, etc. (which will be covered in Chapter 11), it seems natural to consider amplifier dynamic range, as well as system spacing, as the major criteria for the feasibility of long systems.

However, at the present state of the art, this is no longer the case; the major factor limiting system length is now flatness of equalization. There is little difficulty in constructing a system with a cascade of 100 amplifiers if flatness can be maintained. Consider, for example, Fig. 6-1, where a system of 220 amplifiers leads to a final dynamic range of 40 db. According to SMPTE standards, this represents a flawless picture. However, a flatness error of only 0.1 db per amplifier at any particular channel leads to a 10 db error for the whole system, which will seriously degrade performance of the system. This reduces the cascade of 200 amplifiers mentioned above (Fig. 6-1) to approximately 85, quite a substantial reduction for such a small error as 0.1 db. An error of 0.1 db per amplifier might be still acceptable; however, it has been demonstrated that errors up to a total of 2 db per amplifier actually exist in practical systems, although the amplifiers themselves had been aligned flat within 0.25 db at the factory.

The cause for this change in response has been traced to the use of jumper cables and other measurement errors, and it has been shown that the use of jumper cables is a major limiting factor to increased system length. Since jumper cables cannot be tolerated in modern high-quality systems, it is essential to use only specially designed amplifiers which avoid the use of jumper cables Such amplifiers must feature a light-weight, weather-proof housing with built-in cable connectors and provisions for strand mounting. A well designed amplifier of this type also keeps the internal lead lengths from connector pins to circuitry to less than 0.5 inch.

With the elimination of jumper cables the major limitation to longer systems lies then in the flatness of alignment achieved in the design of the amplifiers, and the correction of variations in the system to the necessary degree of perfection (such as for temperature, incorrect spacing, etc.). A major contributor to reduced system performance is inaccuracy or difficulty of field adjustments. This is discussed in more detail in Chapter 10. Here it suffices to examine the effect of level errors on overall system performance regardless of the cause for such errors.

Assume, for example, a level meter or test procedure is used which results in a total accuracy of ±3 db when making field adjustments. Such a tolerance will allow one channel to be set up to 6 db higher in level than some other channel. If this is done consistently, at least one channel will be 3 db into overload and at least one other channel will be 3 db into noise. Consequently, system length will have to be reduced so as to regain the lost 6 db in dynamic range. It is seen that system length is cut to one half the normal

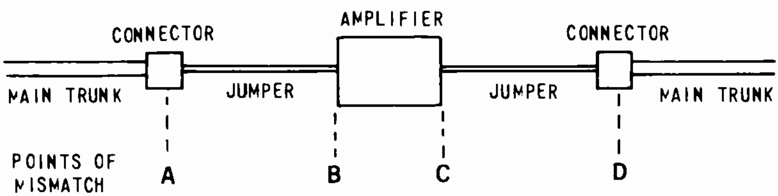


Fig. 5-1. Illustration of mismatches caused by jumper cables.

length if field adjustments cannot be performed more accurately than to a tolerance of ±3 db. Accuracy of field adjustments can be improved by better meters, back-matched test points, etc. However, even with a better accuracy of ±1 db, system length is still reduced by 20%. Due to these tolerances of instrumentation, human errors, etc., it is highly undesirable to field-adjust each amplifier, a procedure which must lead to incorrect settings and consequent system degradation. The only concept which

*This is a very common and often specified accuracy with presently available CATV test equipment.

avoids all of these pitfalls is based on a fully automatic CATV system which will be discussed in Chapter 11.

5-4. JUMPER CABLES

In order to fully appreciate the effect of jumper cables on frequency response, let us consider a typical amplifier position in the system of Fig. 5-1. Discontinuities in impe-

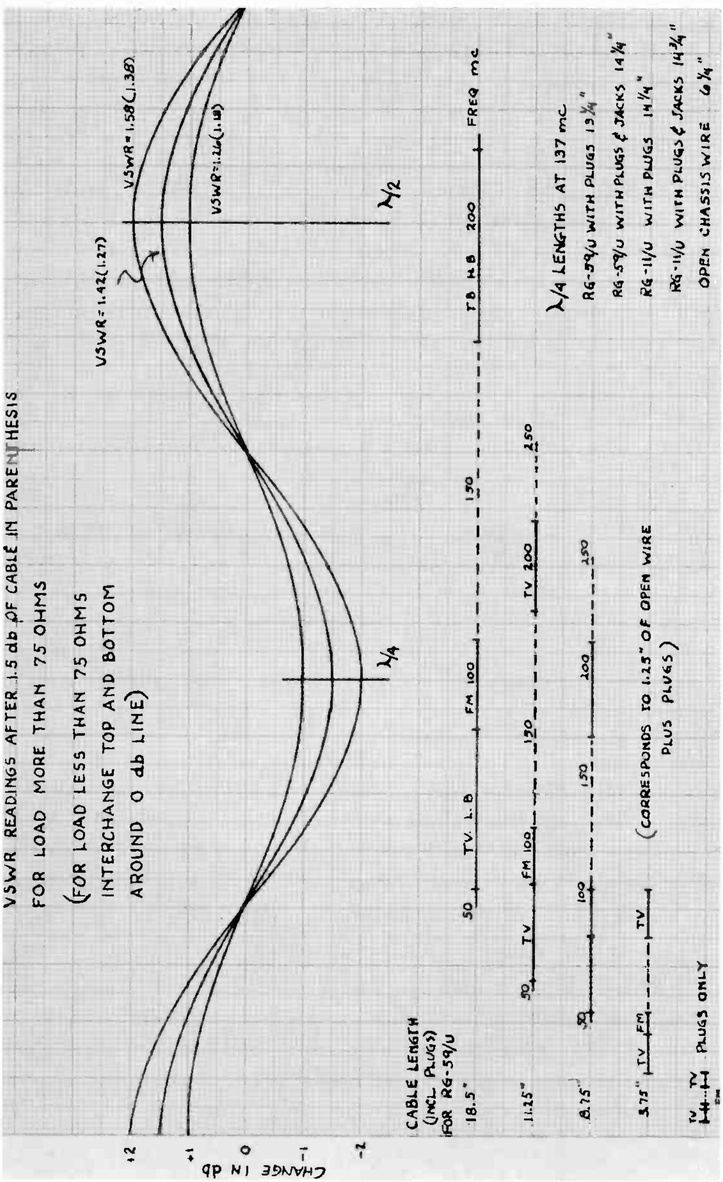


Fig. 5-2. Effect of jumper cables on frequency response.

dance are indicated by letters A to D. The first mismatch occurs between the main trunk cable and the jumper at A. Part of the mismatch is due to poorly controlled impedance of the cables, partly due to the connectors. A VSWR of 1.1 is quite common for this point, with 1.05 achieved in rare cases. The next mismatch is at the amplifier input at B. Since this VSWR is often measured with 1.5 db of cable (discussed in Chapter 13), the actual VSWR ranges from 1.31 to 1.38 db for Channels 2 and 13, respectively for an (inexactly) specified match of 1.25. It is entirely possible to design amplifiers with a lower input VSWR; however, this is done at the expense of noise figure and a compromise must be chosen.

Both of the discontinuities at A and B seriously affect the frequency response of the signal applied to the amplifier. Factory alignment with jumper cable included does not solve the problem, because the mismatch at A cannot be properly simulated; then, in turn, the influence of the mismatch at B is also affected by the system. This is so because the signal reflected at B is again reflected at A, and both mismatches directly affect the frequency response. The magnitude of the effect is evident from Fig. 5-2, where the maximum change in response is given for different lengths of jumper cable. For a measured VSWR of 1.25 (center curve), the maximum deviation is ± 1.5 db. The length of the jumper cable determines at which frequencies the effect is most severe. For example, a jumper of 18.5" of RG-59 cable produces an increase of 0.4 db in the center of the high-band channels, as compared to Channels 7 and 13, with the overall high band raised by 1.1 db. The low band is tilted, with Channel 6 at 0.9 db below Channel 2. With different lengths of jumper cable, the frequency scale is shifted as indicated in Fig. 5-2. Depending on the exact location of the mismatches and their magnitude, the reflection passes through the jumper cable two or three times, thereby causing a shift in the frequency of the first dip.

A situation similar to that at the input of the amplifier also occurs at the output, thereby leading to a cumulative error in frequency response of rather severe magnitude. These errors generally cannot be compensated by a different alignment of the amplifiers, although this is sometimes attempted. Another method, called "tuning the jumpers," where the length of various jumpers is trimmed, has also been used. Both methods are incorrect. The author has experienced a situation where amplifiers were rejected in final tests at the factory for flatness errors of 0.25 db, when the design of the amplifier required the use of jumper cables which would introduce an additional error of up to 2 db per amplifier in the system. Also, the alignment procedures often used are incorrect, with uncontrolled source and load matches, numerous jumper cables, and other deficiencies.*

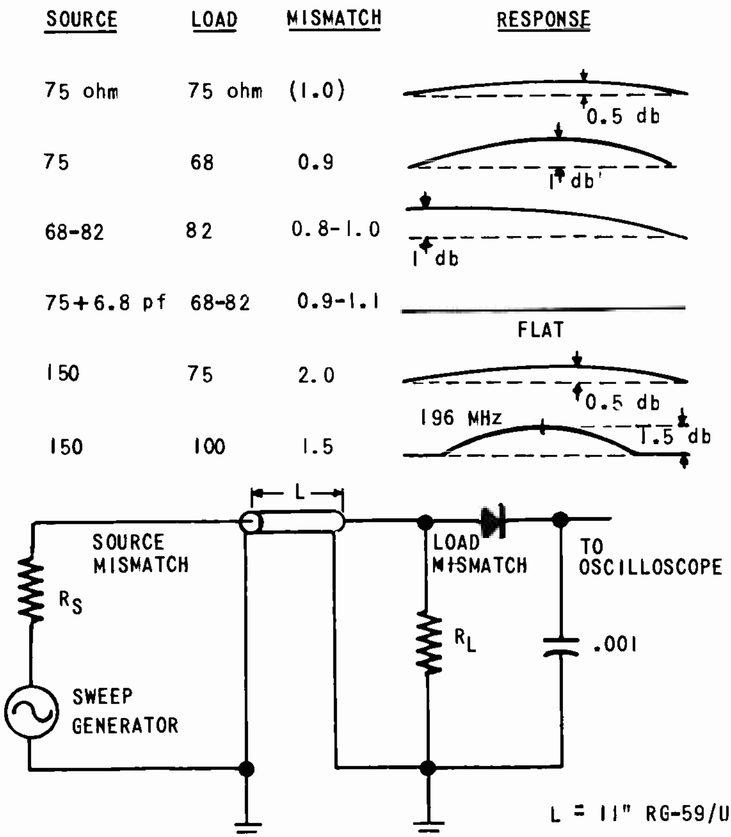


Fig. 5-3. Effect of input or output jumper cable.

It is futile to try to align to a flatness of 0.25 db under these conditions. However, such procedures are still attempted on occasion.

The theoretical curves of Fig. 5-2 are pretty well substantiated by actual response measurements under various

*For correct test methods see Chapter 13.

cal curves are obtained for a lossless jumper cable and purely resistive mismatches. Practical mismatches include capacitance and inductance as well. Two cases must be distinguished; with the point of observation at the end of the jumper cable, or at the head of the jumper. The first case applies to both jumpers of Fig. 5-1, and measurements are given in Fig. 5-3. The only time a flat response

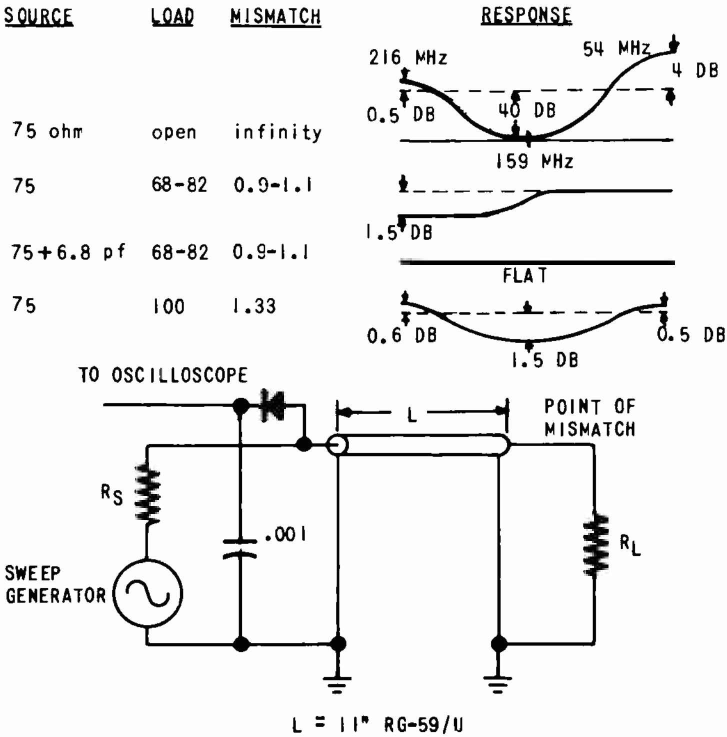


Fig. 5-4. Effect of jumper cable on taps or bridgers.

results is when the source is perfectly matched—that is, a resistive and capacitive match, since any practical, lossy cable is perfectly terminated by a resistance and capacitance only. The effect is more severe for the output jumper, since amplifier outputs are generally not too well matched to obtain the highest overload capability. As is seen, even an output match as good as 1.1 would be of little benefit.

Another effect occurring in bridger applications is depicted in Fig. 5-4. The last example in this figure corresponds directly to the theoretical curves of Fig. 5-2.

Since there is no compensation possible for jumper cables, and it is impractical to obtain VSWR's better than 1.1 (even this value would be insufficient), the use of jumpers must be restricted to lower quality systems of limited length. In the past, jumper cables could not be avoided because no suitable amplifiers were being manufactured for use without jumper cables. With well designed amplifiers available now, no modern system should be constructed with jumper cables, particularly if full advantage is to be taken of the increased cascadeability of modern amplifiers for added system length.

CHAPTER 6

System Level, Level Diagrams, and Tilt

The system level or system operating level is the signal level at the output of each amplifier at the highest signal frequency, usually Channel 13. Main-trunk level is determined mainly from cascading considerations while distribution level may be based on system efficiency (see Chapter 9 on high-level distribution systems). As we have seen, in cascading repeater amplifiers, distortion and noise increase steadily and system dynamic range is reduced as more amplifiers are operated in cascade. For a high quality CATV amplifier, a system must be derated as shown in Fig. 6-1. After 100 amplifiers, system overload is reduced to 40 dbmv, and system output noise level is at -7 dbmv. For a signal-to-noise ratio of 40 db, there is then a margin of 7 db and the system operating level may be positioned right in between the allowable maximum and minimum signal with 3.5 db margin on either side.

6-1. SYSTEM OPERATING LEVEL

For the example given in Fig. 6-1, the system level is then 36.5 dbmv. With this system level and no safety margins, this amplifier would allow cascades of 220 amplifiers to be built. Comparing this result with Fig. 4-1, where only 35 amplifiers could be cascaded, we see the following reasons for the drastic improvement: larger overload level and reduced noise figure of individual amplifiers, plus reduced spacing from 25 to 22 db. The reduced spacing implies, of course, that overload level and noise figure were obtained at this spacing. Considering this drastic improvement with the more advanced CATV amplifier, we still find a remarkably small change in system operating level, 36.5 dbmv in Fig. 6-1 versus 34.25 dbmv in Fig. 4-1. On analyzing various other factors involved it appears that main-trunk levels will generally remain in the vicinity of 35 dbmv at Channel 13

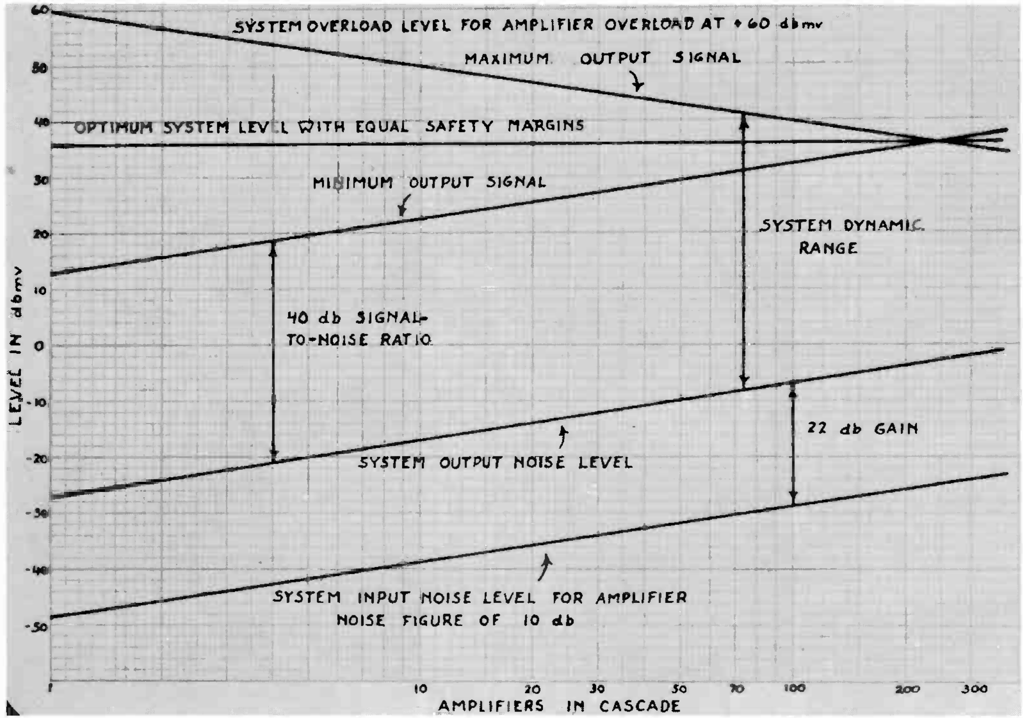


Fig. 6-1. Derating diagram for advanced CATV amplifiers.

Even in shorter systems—unlike the distribution system—there is normally no advantage in raising or lowering the main-trunk level from the optimum system level.

The optimum system level is not necessarily set at equal safety margins from overload and noise. It is found that overload is far more critical than noise. If system level exceeds the system overload level by only 2 db, due to some misadjustment, temperature change, etc., this leads to a very pronounced loss of system performance. The overload margin from excellent to totally unacceptable is usually less than 5 db.* This corresponds to a change of about 15 db with noise (see Table 1 - 1). Therefore, the safety margin from overload should be greater than the one from noise. Optimum system level is, therefore, usually chosen 2 to 3 db below the level determined from Fig. 6-1. This puts the system level at approximately + 34 dbmv.

In practical system construction, it is wise to determine the maximum and minimum system level after completion of the system. This measurement is made at the system extremity (taken from a system map) by changing the combined signal level in the head end to the onset of windshield wiping and noise.** The total safety margin is then noted and the system level is set about 2 - 3 db below the exact middle, between overload and noise. The total safety margin so measured is a figure of merit for a completed CATV system. It relates to the worst-case condition at the system extremity. However, the worst subscriber location chosen for this test must not have any other unusual characteristics, such as extra-length house drops, splitters, etc., but must be representative of normal operation at the system extremity.

6-2. DISTRIBUTION LEVEL DIAGRAMS

In order to have a better understanding of level changes in a CATV system it is advantageous to use level diagrams as an aid in system design. At a glance, it is then possible to detect any flaws in system performance and make the necessary adjustments. In developing level diagrams for

*For a true square law overload characteristic this number would be 7.5 db. However, above the overload level, distortion products increase at a higher power law.

**This test normally cannot be performed with systems not using pilotcarrier AGC, since all AGC amplifiers would have to be switched to "manual" operation and the level errors become excessive.

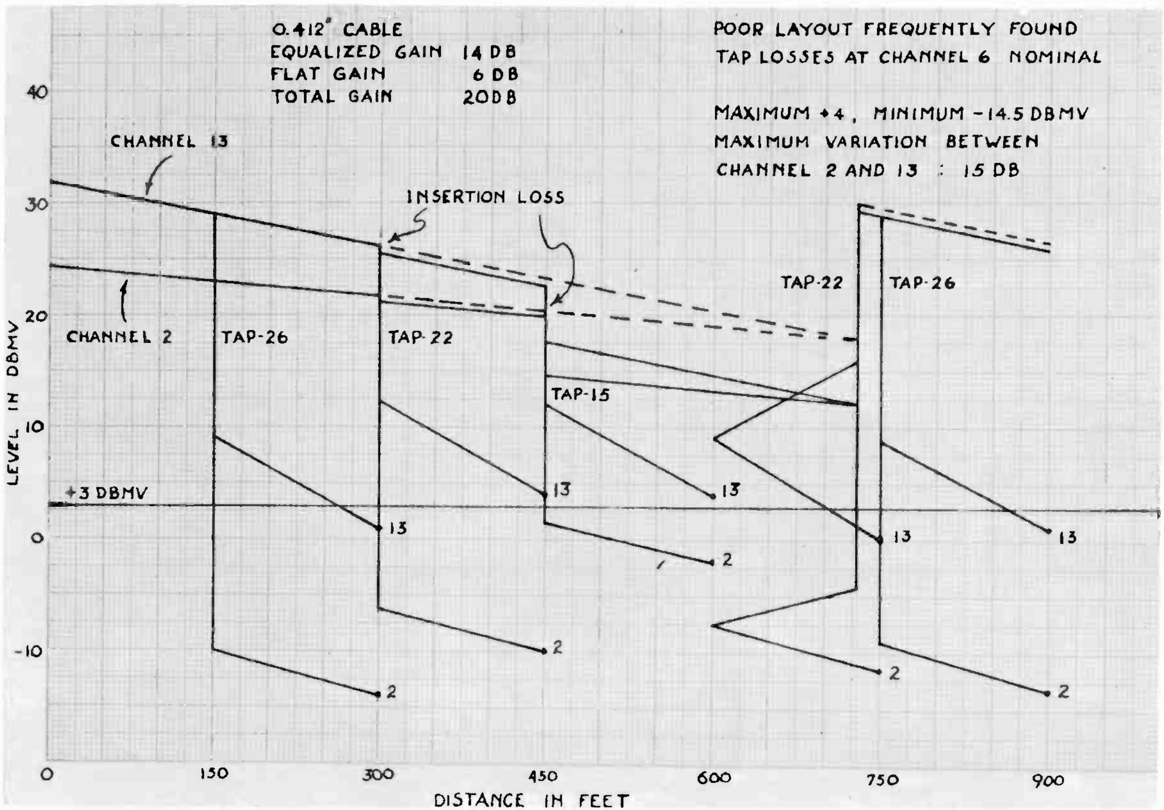


Fig. 6-2. Distribution level diagram with capacitive taps.

an overall system concept it is best to consider the distribution system first and start with the levels at the subscriber's TV set.

We already know that a noise figure of 10 db corresponds to an input noise level of -49 dbmv. Consequently, the minimum input signal to a TV set is -9 dbmv to meet SMPTE standards for flawless reception. A noise figure of 10 db is not met by most presently manufactured TV sets; therefore, a higher signal input level is desirable. On the other hand, standard TV sets are not designed for adjacent-channel reception. Thus, in all-band TV systems, only a limited input signal can be handled without distortion (windshield wiping). This point is reached for most sets with an input signal of 20 dbmv, although some sets show windshield wiping at much lower levels. With this in mind, a level of 3 dbmv (at 75 ohms) is often used as a standard for the input level, with a tolerance stated at ±7 db (an input signal level from -4 to +10 dbmv). This tolerance with a total range of 14 db is rather high and has been chosen to accommodate some of the unavoidable variations such as different lengths of house drops, etc.

In laying out a distribution system, average values should be used as design center. In such a layout, with average 75-foot lots, a four-house tap will be required every 150 feet. The length of the house drop itself may vary considerably, with 150 feet a good value to choose, to allow for aging of the usually lower quality house drop cable. Let us also keep in mind that a line extender or distribution amplifier should have two stages of amplification, no more and no less, which provides a maximum gain from 20 to 25 db.

It is instructive to first look at a distribution level diagram for actual distribution systems, commonly used in the past, as an example of poor layout design practice. These distribution systems are based on the use of capacitive pressure taps, and the level diagram is shown in Fig. 6-2. The left-hand scale gives the level in dbmv and the bottom scale distance in feet. Starting at a level of 32 dbmv, the signal drops along a sloping line until it hits the lowest point at about 730 feet, where it is amplified and brought up to a level of 30 dbmv. The sloping line represents, in this case, 0.412" aluminum cable. Two lines are shown, for Channels 2 and 13. Channel 2 starts at a level of 24 dbmv and drops to the same level as Channel 13, 12 dbmv, at the input of the next amplifier. Therefore, we are using the system operating mode called "full tilt," where the amplifier outputs are fully tilted and the inputs to the amplifiers are flat, with the same level for both Channels 2 and 13.

Let us now follow the signal path in detail. Channel 13 is applied to the cable with a level of 32 dbmv. After 150 feet, we arrive at the first 4-way capacitive tap. The signal level in the distribution cable is down to 29 dbmv at this point. A tap with a nominal tap value of 26 db (at Channel 7), is used, with an actual loss of 20 db at Channel 13, and 39 db at Channel 2. The signal then goes through the house drop (slanted line) and arrives at the TV set with a level of 1 dbmv at Channel 13. Similarly, it is found that Channel 2 has a level of -14 dbmv at this point. Clearly, there will be excessive noise on Channel 2, even if there is no noise in the system itself, because it would take a TV set with a noise figure of 6 db (!) to provide a good picture with such a low signal level.

At the 300-foot tap location, a tap with a nominal loss of 22 db was specified. The actual losses as indicated in the level diagram are -28 db at Channel 2 and -13 db at Channel 13. As the tap loss becomes less, an insertion loss is incurred in the distribution system, as indicated by the step from 26 to 25.5 dbmv at Channel 13, and 22.8 to 22.5 at Channel 2. The levels at TV sets fed from this tap are -10 and +4 dbmv respectively.

At the 450-foot position, a tap with a nominal loss of 15 db had to be used. This kind of tap has an insertion loss of about 5 db, indicated by the large step from 23 to 18 dbmv for Channel 13. Whenever the insertion loss of the tap exceeds 3 db, no further tapping of the distribution cable is possible until after the next distribution amplifier. Therefore, no tap was possible at 600 feet.

In order to supply the four houses at this location, back-feeding is used, indicated by the lines slanted to the left. This is accomplished by inserting a tap immediately following the line extender. Because of the longer run of house-drop cable, a lower value tap (22 db) is used. Thereafter, the system repeats itself and the level diagram is similar to the first section shown.

Similar layouts have been used in thousands of cases, mainly because of ignorance of proper system techniques, which include level diagrams. The faults of this layout can be seen at a glance: The variation between channels is too large for high picture quality. This tremendous variation is caused by the capacitive taps. The house-drop cable improves the spread only insignificantly. Someone might argue that if the amplifiers were run with flat output and tilted inputs, the situation would be improved. If this were done, the spread between channels on the TV sets would be reduced by 8 db; however, Channel 2 would not be brought up 8 db; rather Channel 13 would have to be dropped. With flat outputs, the overload level of the amplifiers is materially reduced.

The only conclusion which can be drawn then is that capacitive taps cannot be used. This leaves transformer taps and directional couplers, which are identical as far as the level diagram is concerned. However, as we shall see in Chapter 8, there are good reasons why only directional couplers can be used in high quality systems.

A distribution system with optimized directional couplers is shown in Fig. 6-2. Again, as the tap loss decreases, the insertion loss increases, leading to larger steps in the top slanted lines, which represent the losses in the distribution cable for both channels. Instead of feeding only 12 houses directly, as in Fig. 6-2, now 20 houses can be supplied. No back-feeding is required in any position. The maximum spread between channels at the subscriber's TV set has now been reduced from 15 db to 4 db. All this has been achieved with an amplifier gain of 17 db, instead of 20 db. However, the reduced gain requirement does not permit the addition of extra subscribers unless the distribution level is also changed. A further examination of this approach leads to a radically different distribution concept which is discussed in Chapter 9.

The distribution system shown in Fig. 6-3 can be achieved only by careful design, and it is a far cry from the cut-and-try methods which result in poor layouts such as in Fig. 6-2. Obviously, it is necessary to develop distribution level diagrams for the most common cases in order to be able to standardize on directional couplers and other details. From these results it is then possible to determine

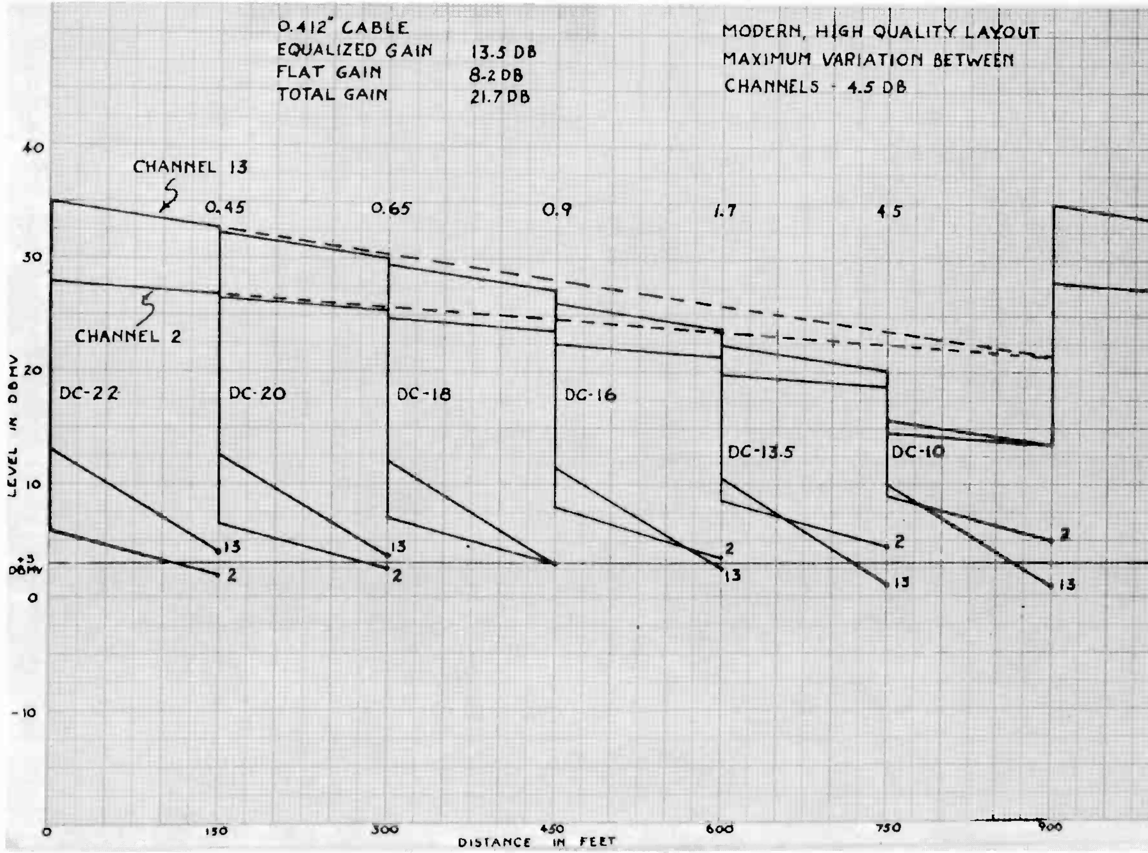


Fig. 6-3. Distribution level diagram with directional couplers.

the specifications of the ideal distribution amplifier. With such matched and standardized components, it is a rather simple matter to design high-quality integrated distribution systems. Future development will likely not affect the layout of the distribution system to any great extent, since the boundary conditions are becoming fully established.

6-3. MAIN TRUNK LEVEL DIAGRAM AND TILT MODES

A main-trunk level diagram is much simpler since no taps need be shown. For a spacing of 25 db, the level diagram looks as in Fig. 6-4. There are three modes of system

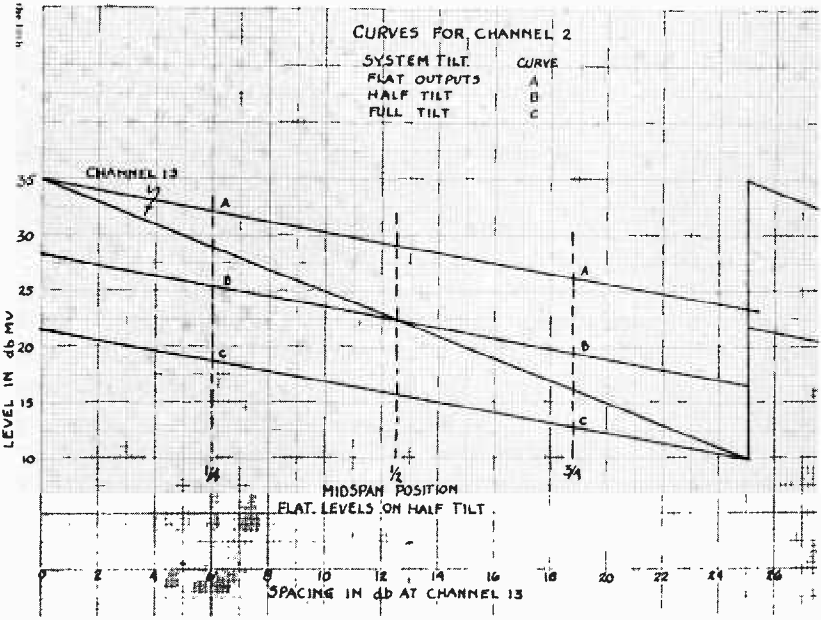


Fig. 6-4. Main-trunk level diagram.

operation: Flat Output, Half Tilt, and Full Tilt. These three different modes are set up in the head-end equipment. Flat Output refers to operation with equal signal levels on Channels 2 through 13 at the outputs of the amplifiers. Full Tilt provides maximum tilt at the output of each amplifier. In this case, the input to all amplifiers is flat. Half Tilt is between the two extremes. In the level diagram (Fig. 6-4), Channel 13 is unaffected by the various tilt modes. An amplifier output level of 35 dbmv is used. This level drops, at 25 db of spacing, to 10 dbmv at the input of the following amplifier, as indicated by the sloping parallel lines (different cable loss at Channel 2) for the three tilt modes. It can be seen at a glance that in Half-Tilt operation, equal levels are obtained in the midspan position.

There are various arguments favoring Full-Tilt operation, with flat inputs and fully tilted outputs. The noise figure generally may be made reasonably flat over both TV bands. This would indicate that flat inputs are most desirable for good signal-to-noise. In the past, the noise figure often was higher on Channel 2 than on Channel 13, due to the type of equalization used.* To preserve a good signal-to-noise ratio on all channels, it was then desirable to run a higher signal input level on Channel 2, corresponding approximately to Half Tilt. As for the output, it can be shown that distortion is lower at Full Tilt, since this decreases the level of some channels in the critical output stage. It seems, therefore, that Full Tilt is the best system mode in the main trunk. From the discussion of distribution systems, we have already seen that Full Tilt is also most desirable there. Because of this, future system design is likely to specify Full Tilt throughout. The use of Half Tilt in the distribution system was generally necessitated by the use of capacitive taps, and in the main trunk by equipment deficiencies with respect to noise on Channel 2, either by circuit design problems or the use of an external cable equalizer or attenuator (see Chapter 7). Although the use of Half Tilt allowed a reasonable compromise with bad equipment in the past, such operation is still far inferior to Full Tilt which is now possible with well designed amplifiers not using a separate cable equalizer or otherwise having an excessive noise figure on Channel 2.

*See also Chapter 7, Section 2.

CHAPTER 7

Disadvantageous Amplifier Design Concepts

Occasionally the question is asked if it would not be better to use fewer high-gain amplifiers, presumably to arrive at a lower system cost and added reliability. The answer is no. The spacing aspects of this question were already treated to some extent in Chapters 4 and 5. However, there are four reasons why high-gain amplifiers lead to poor CATV performance; these will be covered in detail here due to their importance in basic system design.

7-1. THE HIGH-GAIN AMPLIFIER

The question, "Why not a high-gain amplifier?" seems natural when we consider amplifier behavior with gain (Fig. 7-1). All amplifiers show highest output capability (least distortion) as the gain is increased. This is quite obvious when we consider that as gain is increased, eventually distortion is generated predominantly in the output stage. Hence, the overload level of such an amplifier at a high-gain setting is essentially the overload level of the output stage. As gain is reduced, eventually distortion in the input stage(s) becomes the limiting factor, leading to an ultimate proportional reduction in overload level—that is, 5 db reduction in overload level for a 5 db reduction in gain. Actual measurements of a CATV amplifier are shown in Fig. 7-1. At every gain setting, equalization was corrected for the corresponding length of cable. The general shape of the curves is typical for any CATV amplifier.

Noise figure, on the other end of the level range, is also improved at maximum gain. Clearly, at maximum gain, noise of the output stage(s) becomes less important and the total noise figure of the amplifier is essentially the noise figure of the input stage. As gain is reduced, noise from other stages is added until eventually a proportional increase in noise figure results with a reduction in gain.

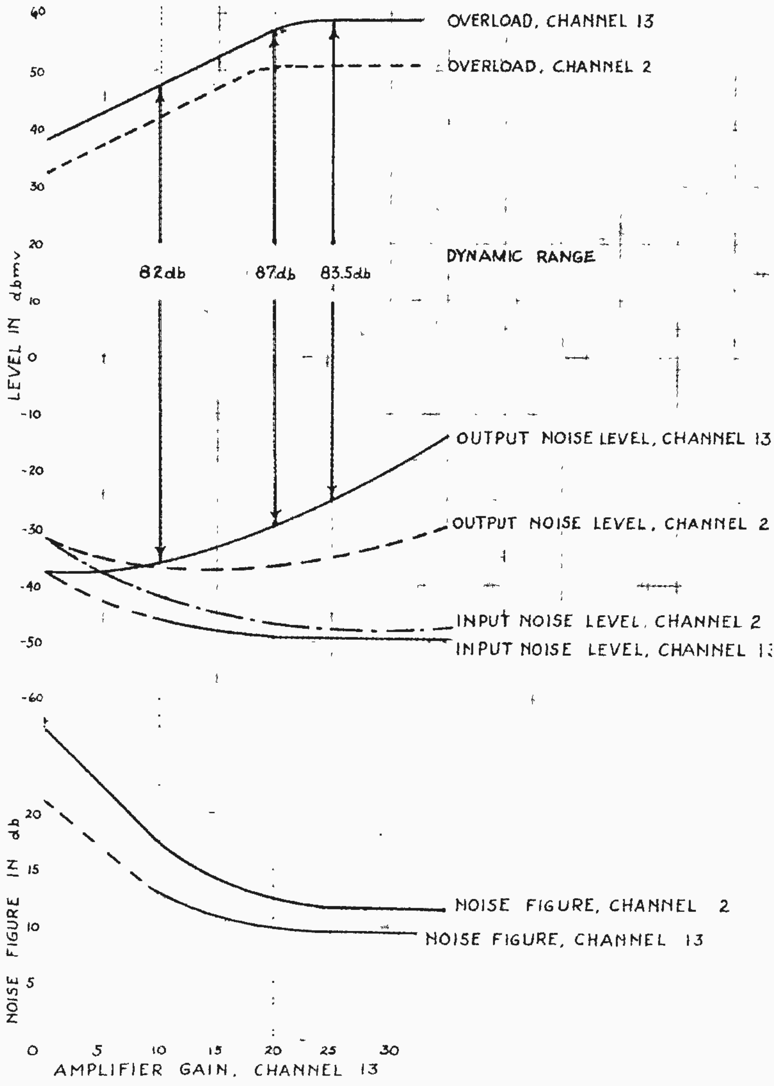


Fig. 7-1. Optimum amplifier gain chart.

Why then, if both noise figure and overload level are best with a high-gain amplifier, would it not be advantageous to use such amplifiers in CATV systems? This seemingly logical question is based on fallacious reasoning, for noise figure is a number directly related to equivalent input noise, while overload level is referred to the output of an amplifier. To calculate overload-to-noise ratio (dynamic range), numbers which are measured at the same point in a CATV system or amplifier must be used. If the objective were merely to achieve highest overload level and lowest noise figure, this would be most readily achieved at maximum gain. However, if the objective is maximum dynamic range and system signal-to-noise ratio, reduced gain must be used, for maximum amplifier gain results also in maximum output noise level, although the noise figure and its equivalent input noise level may be low. In Fig. 7-1, noise figure is converted to equivalent input noise level using the relationship of Fig. 3-4. The addition of gain at each point results in the output noise level of the amplifier, plotted at both Channels 2 and 13 in Fig. 7-1. The objective is now to choose an amplifier gain so that the distance between the overload and output noise curves becomes largest—that is, to find the peak of amplifier dynamic range. Examination of the graph shows an optimum gain near 20 db for the example. At 25 db, amplifier dynamic range is reduced by 3. 5 db, although overload is up 1 db and noise figure has dropped 0. 5 db.

As discussed in Chapter 4, maximum amplifier dynamic range must be corrected slightly if we are to have a true cascaded figure of merit for system performance. In Fig. 7-2, such a system correction of the overload curves is made by superimposing the correction of Fig. 3-7. The amplifier and system dynamic range can then be plotted directly from measurements as shown in Fig. 7-3. Here amplifier dynamic range is shown together with dynamic range of a 1000 db system (for a 2000 db system subtract 6 db, etc.). Optimum spacing equals the gain setting, resulting in optimum system dynamic range. Since Channel 13 is the more critical one for the example given, a spacing of 22 db is optimum for this amplifier. A normal level variation of ± 3 db results in less than 1 db degradation in dynamic range for this spacing. Note that optimum spacing is slightly more than the gain setting giving maximum dynamic range for an individual amplifier.

This detailed discussion explains why a high-gain amplifier falls short of meeting the gain requirement for both maximum amplifier dynamic range as well as optimum system spacing, both of which are close together. A third consideration has to do with the number of stages in an amplifier. It can be shown (Appendix II) that a two-stage amplifier results in a system degradation of 2.5 db as compared to single-stage amplifiers. A three-stage amplifier is worse again than a two-stage circuit. Consequently,

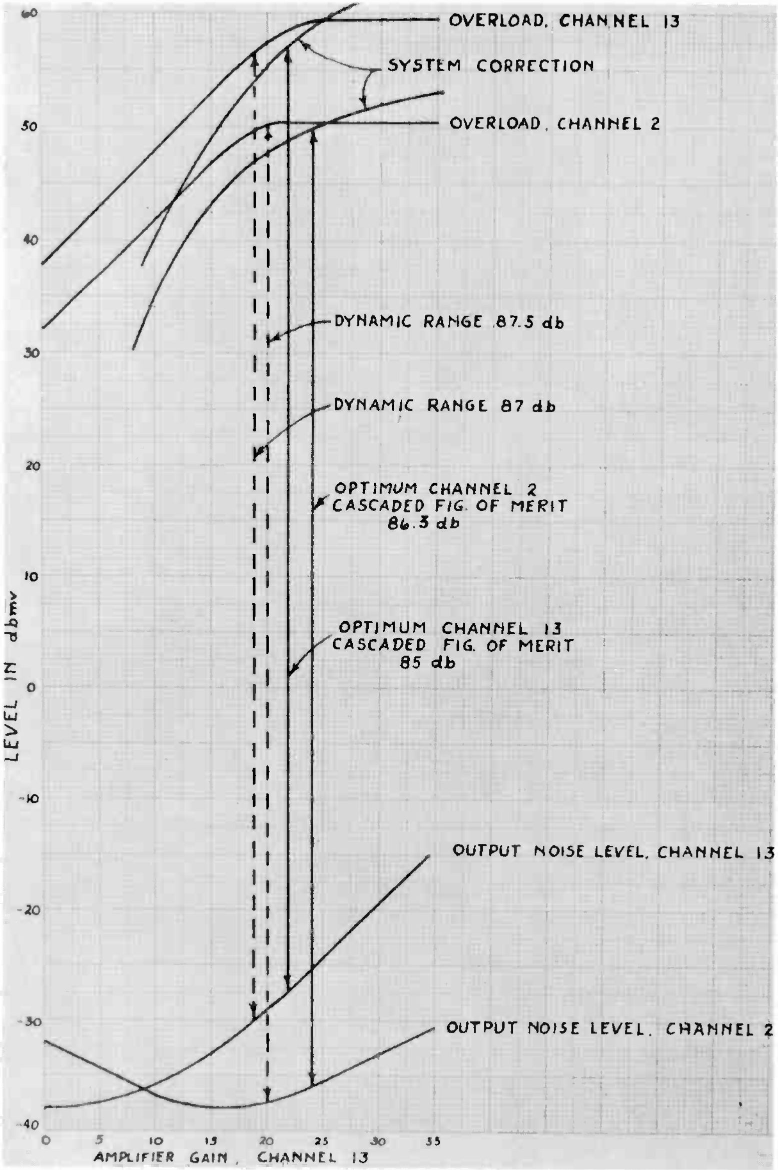


Fig. 7-2. Graphical analysis of amplifier dynamic range and cascaded figure of merit.

the amplifier with the fewest number of stages, which still meets all system specification, is superior as far as system performance is concerned. With the rapid advances in circuit design, two stages are required now where previously three or four were used. It is unlikely that amplifiers will be reduced to single stages in the near future, because of increasingly difficult circuit design. There are also economic reasons which favor a two-stage design over single-stage amplifiers.

Lastly, excessive spacing results in a greater variation between Channels 2 and 13, which creates special prob-

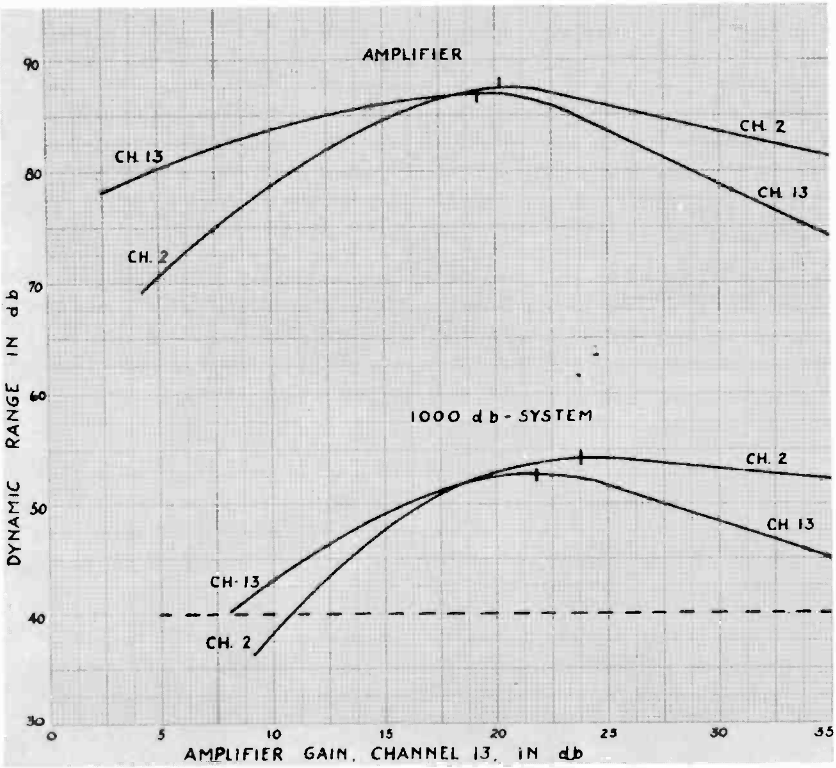


Fig. 7-3. Amplifier and system dynamic range curves.

lems in the equalization of bridger amplifiers. Also, it necessitates even more the use of midspan bridgers with their difficult equalization problems. With lower gain amplifiers, midspan bridgers can be more readily avoided. Also, bridgers can be built into the same housing with the main trunk amplifiers, thereby avoiding special equalization problems.

7-2. PASSIVE EQUALIZERS AND ATTENUATORS*

In designing amplifiers for CATV systems, two basic approaches can be taken. In one concept the amplifier is de-

*See Reference 27, Appendix VII.

signed "flat," with equal gain at all frequencies, and a separate, passive equalizer is used to compensate the cable. This separate equalizer can be built, of course, into the same housing, or even into the same circuit board; it still is basically a flat amplifier, together with a separate cable equalizer circuit. The other approach uses equalization as an integral part of the amplifier. This latter type of amplifier is far more difficult to design, but it has significant advantages as compared to the first approach.

First, let us consider the system dynamic range of both amplifiers. We assume an ideal basic amplifier, having a noise figure of 10 db, an overload level of 60 dbmv, and a gain of 25 db. One of these amplifiers is equalized and can be used as is in the CATV system. The other amplifier is flat and must be used with a separate equalizer. It does not matter in principle if this equalizer is placed before or after the amplifier; in one case the noise figure is increased, in the other case the overload level is decreased; the reduction in dynamic range is the same, however.

For this example, we place the equalizer before the flat amplifier. Such an equalizer for 25 db of cable has an insertion loss of about 1 db at Channel 13 and a loss of 14 db at Channel 2 (13 db tilt). Connecting such an equalizer ahead of any amplifier results in a proportionate increase in noise figure, since the signal level is attenuated while noise is not. Thus, the flat amplifier plus equalizer is identical in every respect to the equalized amplifier, except its noise figure is 1 db higher on Channel 13 and 14 db higher on Channel 2 (noise figures of 11 and 24 db as compared to 10 db on the equalized amplifier).

The disadvantage of the flat amplifier-equalizer concept is obvious. If any passive equalizer is to be used, it must be used after the noise-producing stages and ahead of the distortion-producing stages for best system design. Such an equalizer, after the noise-producing stages, reduces noise and signal level by the same amount at Channel 2, thereby not causing a degradation in noise figure or in the overload factor. Actually, since in CATV amplifiers stage gain must be low, noise- and distortion-producing stages cannot be fully separated, and equalization must be distributed in a certain way throughout the amplifier, which is the approach taken in the other concept—the "equalized amplifier."

A degradation of 14 db in noise figure is very serious, and allows only one fifth of the system length that would be possible with an "equalized amplifier." Since the degradation at Channel 2 was considerably worse than at Channel 13, it appears that with a different system tilt the degradation might be averaged out better, so that the total system degradation as compared to the "equalized amplifier" is reduced. In Table 7 - 1, system dynamic range is given for both the "equalized amplifier" and the flat-amplifier-equalizer combination for the three system tilt modes—Full Tilt, Half Tilt and Flat Outputs. The optimum system tilt results in an equal dynamic range at Channels 2 and 13. For any other tilt, system performance is limited by the poorer of the two channels. Thus, for the "equalized amplifier" of our example, Full Tilt is best and results in a 1000 db system dynamic range of 52 db. For the amplifier with separate equalizer, Half Tilt is best and leads to 47.75 db. Thus, the "equalized amplifier" is better by 4.25 db. In terms of system length, it permits a 63% longer system to be built. The "equalized amplifier" is better for any system tilt. When operated Half Tilt the improvement is only 1 db at Channel 13, yet it is 14 db at Channel 2.

In analyzing different system tilts, it must be remembered that for 25 db gain, the difference between Channels 2 and 13 is 13 db. Hence, in Half Tilt, Channel 2 is increased by 6.5 db, and for Flat Outputs, 13 db, with no change in Channel 13. Similarly, it can be shown that overload level is reduced by 3.25 and 6.5 db respectively at Channel 13 for Half Tilt and Flat Outputs, while overload at Channel 2 is increased by the same amounts, as compared to Full-Tilt operation.

This relationship is pictured for clarity in Fig. 7 - 4, where the 1 db insertion loss for the equalizer was neglected in order to avoid confusing lines. It is clearly seen why Half Tilt results in best system operation for the flat amplifier - equalizer combination. Also, it is evident that the "equalized amplifier" is superior under all tilt modes and operates best at Full Tilt.

While the foregoing was, in part, a theoretical analysis of the effect of various system tilts and amplifier design concept, the results of this analysis can be readily compared with the curves of Fig. 7-3, which apply to an

Table 7-1. Comparison of equalized amplifier with flat amplifier plus separate equalizer (10 db noise figure, 25 db gain, 60 dbmv overload).

		Equalized Amplifier		Ampl. W/Sep. Equalizer		
		Ch 2	Ch 13	Ch 2	Ch 13	
FULL TILT	Noise Figure	10	10	24	11	db
	Equiv. Input Noise	-49	-49	-35	-48	dbmv
	Output Noise	-37	-24	-23	-23	dbmv
	Overload	47	60	47	60	dbmv
	Dynamic Range	84	84	70	83	db
	1000-db System	52	52	38	51	db
HALF TILT	Noise Figure	(4.5)*	10	(17.5)*	11	db
	Equiv. Input Noise	-55.5	-49	-41.5	-48	dbmv
	Output Noise	-43.5	-24	-29.5	-23	dbmv
	Overload	50.25	56.75	50.25	56.75	dbmv
	Dynamic Range	93.75	80.75	79.75	79.75	db
	1000-db System	61.75	48.75	47.75	47.75	db
FLAT OUTPUTS	Noise Figure	(-3)*	10	(11)*	11	db
	Equiv. Input Noise	-62	-49	-48	-48	dbmv
	Output Noise	-50	-24	-36	-23	dbmv
	Overload	53.5	53.5	53.5	53.5	dbmv
	Dynamic Range	103.5	77.5	89.5	76.5	db
	1000-db System	71.5	45.5	57.5	44.5	db

*Equivalent noise figure. Used here for simplicity of calculation to show increase in signal-to-noise ratio as compared to Full Tilt system.

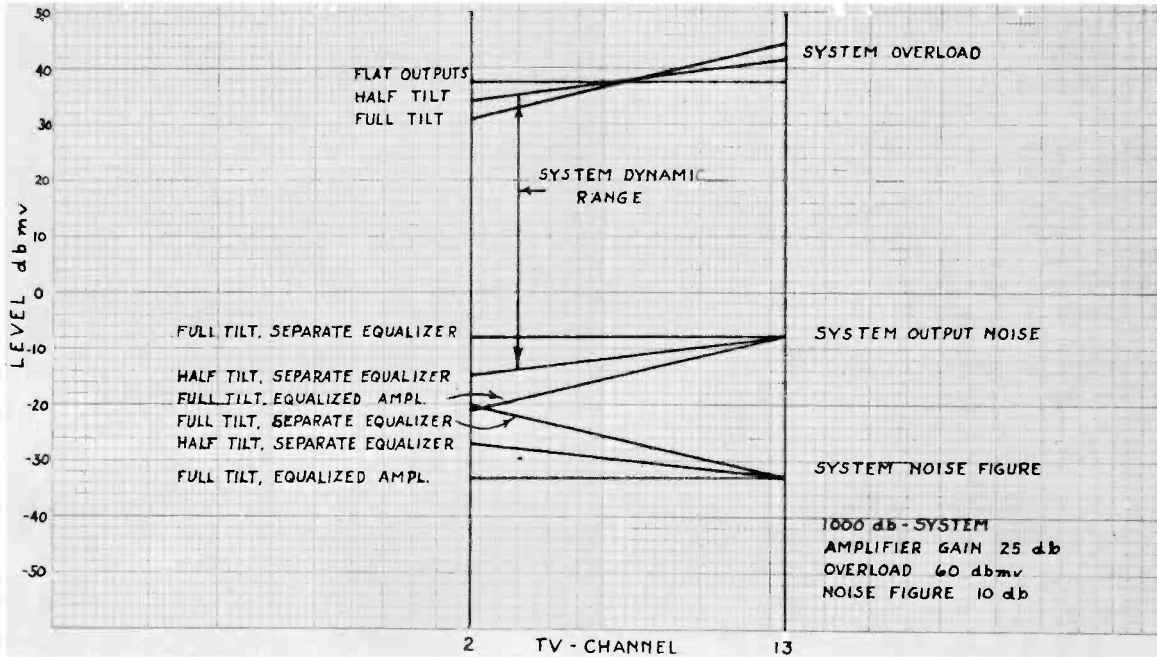


Fig. 7-4. Diagram of system tilt and dynamic range for two basic amplifier concepts.

actual "equalized amplifier" operated at Full Tilt.

In conclusion, it should be mentioned that there are other serious drawbacks to the flat amplifier-equalizer approach. For example, it does not adapt itself to the automatic spacing principle (discussed in Chapter 11), which is one of the cornerstones of fully automatic CATV systems.

Attenuators are used occasionally at the input of amplifiers to compensate for variations in flat loss. The effect on noise figure is similar to the one already described for a separate equalizer. For example, a 6 db input attenuator also increases noise figure by 6 db. This degradation in system performance can be avoided by a different compensation technique for changes in flat loss. As discussed in Chapter 10, a properly designed tilt control permits ready correction for different amounts of flat loss; this adjustment is made with no increase in noise figure, an approach which is clearly superior to the use of a switchable input attenuator in an amplifier.

CHAPTER 8

Matching and Reflections

Transmission lines must be matched (terminated in their characteristic impedance) if reflections are to be avoided. In CATV systems, mismatches with their resulting reflections manifest themselves by a power loss and the production of secondary images or ghosting. The power loss results in a change of frequency response and affects flatness of an amplifier; however, with proper alignment methods (see Chapter 13) and modern amplifier design, errors in flatness have been largely overcome, even with large mismatches.

Of a much more serious nature is ghosting when a second, weaker picture appears on the TV screen, displaced from the original picture by whatever time delay is incurred for the reflected signal. In considering such ghosts, it is obvious that no ghost will be seen if the delay of the reflected picture is short enough; the reflected picture is then simply superimposed on the direct signal. In this case, ghost visibility is limited by the resolution of a TV set and present transmission standards. As the delay time is increased, however, a decreased horizontal focus is noticed and ultimately, a clearly recognizable ghost picture results. For any such ghost picture, it is also clear that if the reflected signal is weak enough, it will not be objectionable, regardless of delay time.

In order to determine a tolerance limit for ghosting, thorough studies were made under controlled conditions,* and a curve was established as a tolerance limit for the visibility of ghosts (Fig. 8 - 1). This curve defines the maximum reflected signal level which can be tolerated in order to obtain a ghost-free picture, for any particular delay time. This curve was obtained from measurements with

*References 20, 21, Appendix VII.

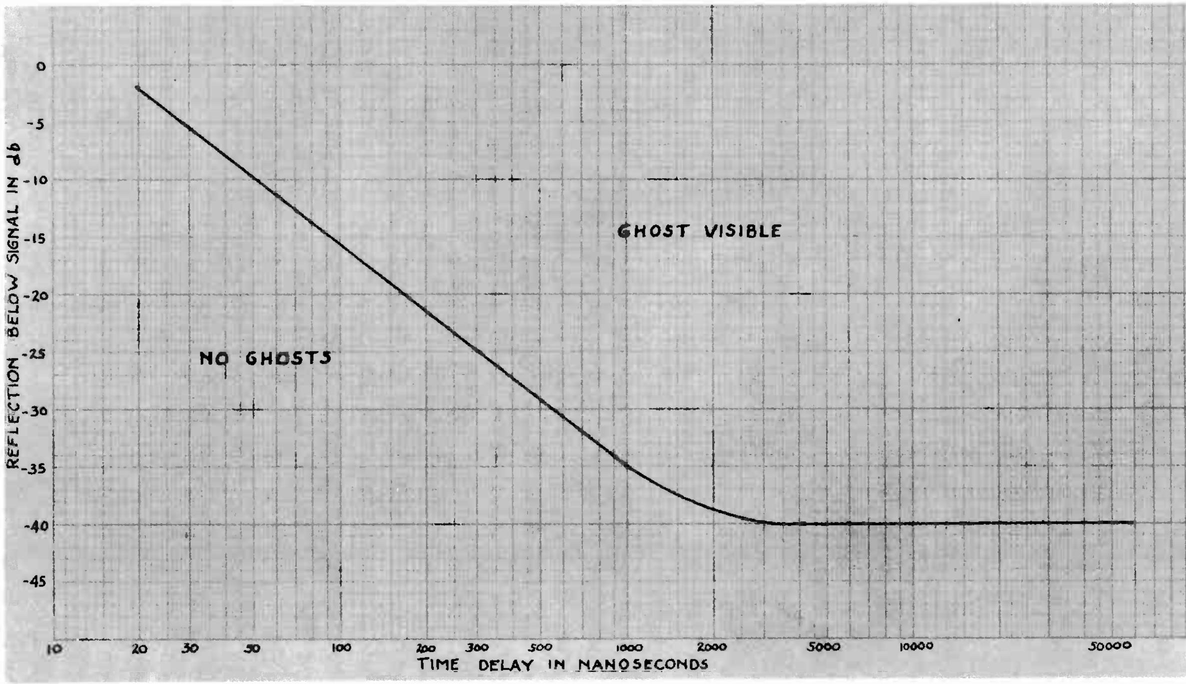


Fig. 8-1. Curve showing perceptibility of ghosts.

studio equipment under ideal conditions. It is, therefore, likely that a somewhat reduced quality level would still be acceptable for average quality home receivers. However, for a flawless, high-quality picture, this curve should be aimed for under all circumstances.

8-1. THE CRITICAL CABLE LENGTH

Delay time and loss are intimately related to a particular transmission line. Consider the generation of a typi-

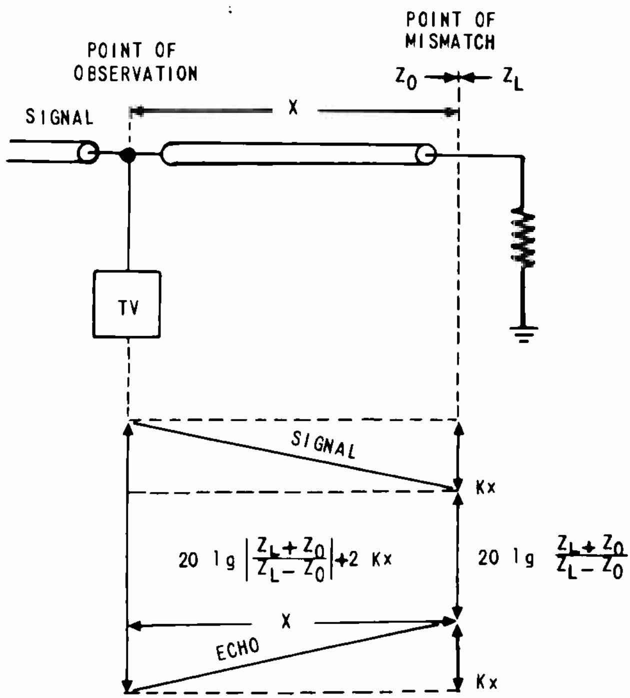


Fig. 8-2. Diagram showing echo generation.

cal reflection (Fig. 8-2). A TV set is located a distance x away from a point of mismatch. The magnitude of the reflected signal at the point of observation depends on the loss in cable x and the severity of mismatch. For a mismatch M, the reflected signal is 20 log (1 + M) / (1 - M) db below the direct signal.* In addition, the signal and its reflection passes through cable x in both directions; therefore, an additional loss of 2 kx results, where k is a constant.

*See Appendix IV for complete mathematical formulae.

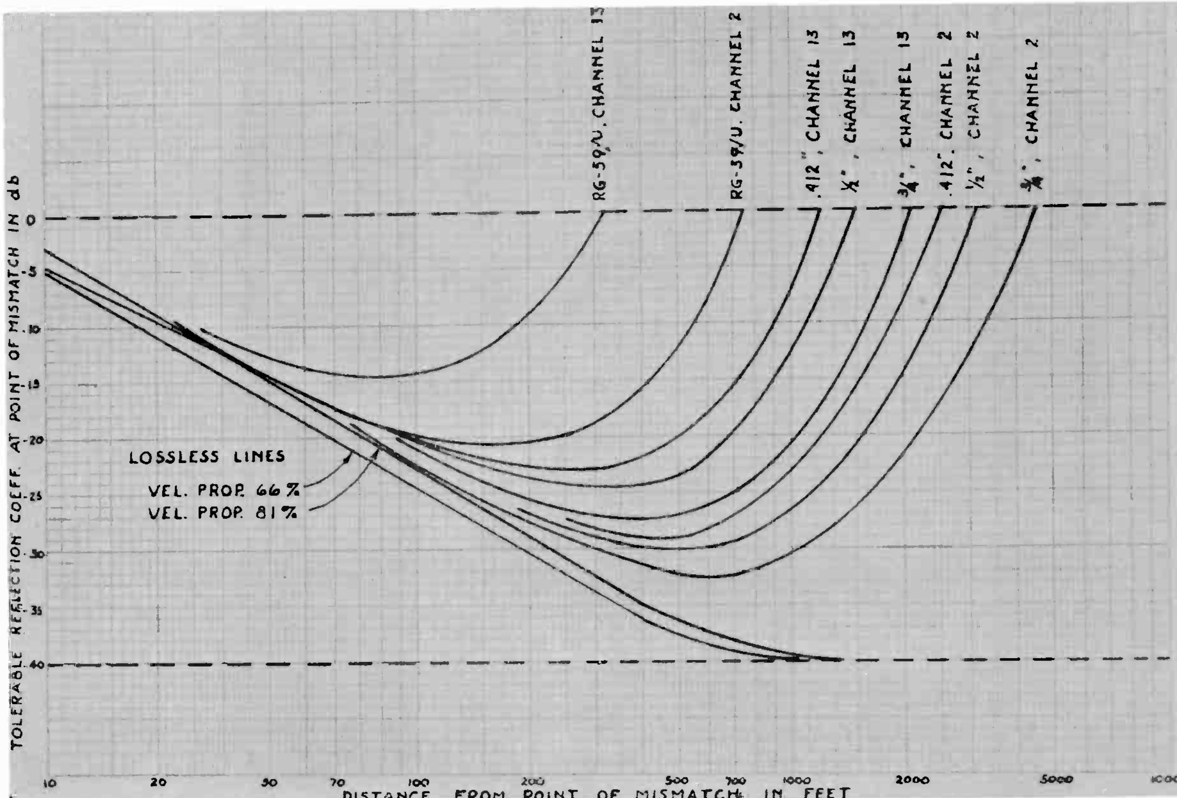


Fig. 8-3. Graph for determination of critical cable lengths.

This relationship is indicated in the level diagram of Fig. 8-2. The time delay, in turn, depends on the propagation constant and the length of cable x. It is then possible to replot the original "perceptibility of ghosts" curve (Fig. 8 - 1) for various cables and to use, instead of delay time in nanoseconds, length of cable in feet (Fig. 8-3). Instead of specifying the reflection below the signal level at the point of observation, the maximum amount of mismatch (return loss) at the point of mismatch is now given. Clearly, with a lossy cable, a worse mismatch can be tolerated in order to still keep the reflection below the required level at the point of observation.

The curves of Fig. 8-3 indicate that there is a worst case condition for each cable. For example, the curve for 1/2 - inch cable shows that for a length of 500 feet, the minimum allowable return loss is 30 db, corresponding to a VSWR of 1.06! In other words, a tap-off point, such as a bridger amplifier or other start of a distribution line, 500 feet ahead of a point where a mismatch of more than 1.06 exists, will produce visible picture degradation due to reflections with this type of cable. The conditions for 3/4 -inch cable are, of course, more severe. All worst-case conditions are listed in Table 8-1.

The length of cable which requires the best match is called the critical cable length or worst - case condition. Any good system is designed to operate without ghosts under the worst-case conditions, since little benefit results in avoiding the critical cable length because the dip in the curves of Fig. 8 - 3 is rather shallow. For example, it is generally not possible to keep a length from 150 to 1200 feet (for 1/2 - inch cable) free of taps, bridgers, or splices to obtain an increase in permissible mismatch by only 5 db to 1.12, which is also a difficult value to achieve. It takes an excellent cable splice to achieve a VSWR of 1.05 or less. As we shall see, the only workable solution to avoid reflection problems lies in the judicious use of directional couplers.

Fig. 8 - 3 shows that a higher mismatch of approximately 6 db can be tolerated at Channel 13 than at Channel 2. This is due to the increased cable losses at Channel 13. Consequently, electronic equipment must have a superior match at Channel 2 only. The higher permissible mismatch

at Channel 13 might be used to advantage to reduce the noise figure or improve some other more important characteristic at Channel 13. In equipment design, it is always well to keep in mind that all performance parameters are closely tied together; improvement in one results in a degradation of the other. Therefore, it is necessary to keep close track of all specifications so that no one parameter is overspecified. In the past, amplifier specifications, often evolved from cut-and-try methods, invariably lead to specifications far from the ideal, with a consequent degradation in overall amplifier and system performance.

Table 8-1. Critical length of cable and required return loss.

Cable	Channel	Critical Length (Ft.)	Return Loss (db)
RG-59/U	2	170	21
	13	85	14.5
.412"	2	500	29
	13	270	23
1/2"	2	500	30
	13	330	24.5
3/4"	2	600	32.5
	13	420	27.5

8-2. WORST-CASE DESIGN IN THE DISTRIBUTION SYSTEM

Reflections are usually greatest in the distribution system, due to the number of taps and possible echo paths. A typical section of a distribution system (Fig. 8-4) contains two distribution amplifiers and five tapping points which generally serve four houses each. Only one TV set is shown on each tap. All mismatches are labeled with letters from A to L. It is instructive to analyze the various reflections possible in such a system. A ready conversion of return loss to VSWR is possible using Fig. 8-5. Let us consider the signal-to-reflection ratios for particular points of mismatch and for the worst-case frequency, Channel 2.

1. Mismatch at Input of Distribution Amplifier, Single Reflection: The mismatch is at point G. From Table 8-1 the critical cable length for 0.412-inch cable is 500 feet,

and the permissible return loss is 29 db. This value corresponds to a VSWR of 1.075 for the input match of a distribution amplifier at Channel 2. The worst reflection will result at TV sets fed from tap C, which is 500 feet away from the mismatch. An input match of 1.075 is hard to achieve for an amplifier. It is also well to keep in mind that published VSWR readings are often low, compared to actual measurements due to systematic measuring errors.* With 1/2-inch cable, which is occasionally used in distribution systems, the critical length is 500 feet and the return loss is 30 db (VSWR 1.066!). Worst taps are B, C and D.

2. Mismatches at the Taps Themselves: This case is very similar to the first case. For example, a mismatch at F produces a worst-case reflection at B. Therefore, the maximum allowable mismatch for each tap is 1.075 for 0.412-inch cable and 1.066 for 1/2-inch cable.

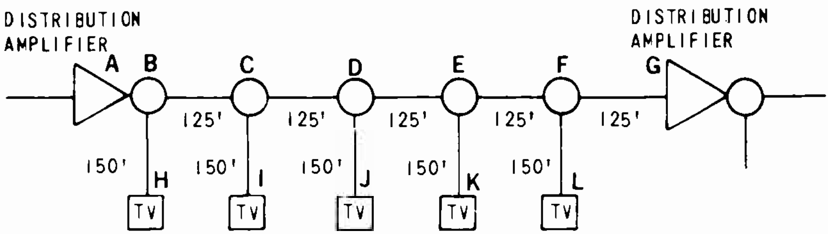


Fig. 8-4. Block diagram of a typical distribution system. Points of mismatch are indicated by letters.

3. Tap to TV Set: The signal travels from the tap to the TV set, is reflected at the TV set, travels back to the tap where it is reflected again to travel back to the TV set to produce the ghost picture. In this case, we have a double reflection. The worst-case condition for RG-59/U is about 170 feet for a single reflection, about 85 feet for a double reflection. The required total return loss is 21 db, which calls for a match at the TV set of 1.2 if the output of the tap is unmatched. A figure higher than this value is often encountered with TV sets.

These three examples show clearly the problems associated with a distribution system. The required VSWR's simply cannot be met with practical components which must sell at a reasonably low cost. It is obvious that only a different concept can lead to a high quality system.

*Refer to Chapter 13.

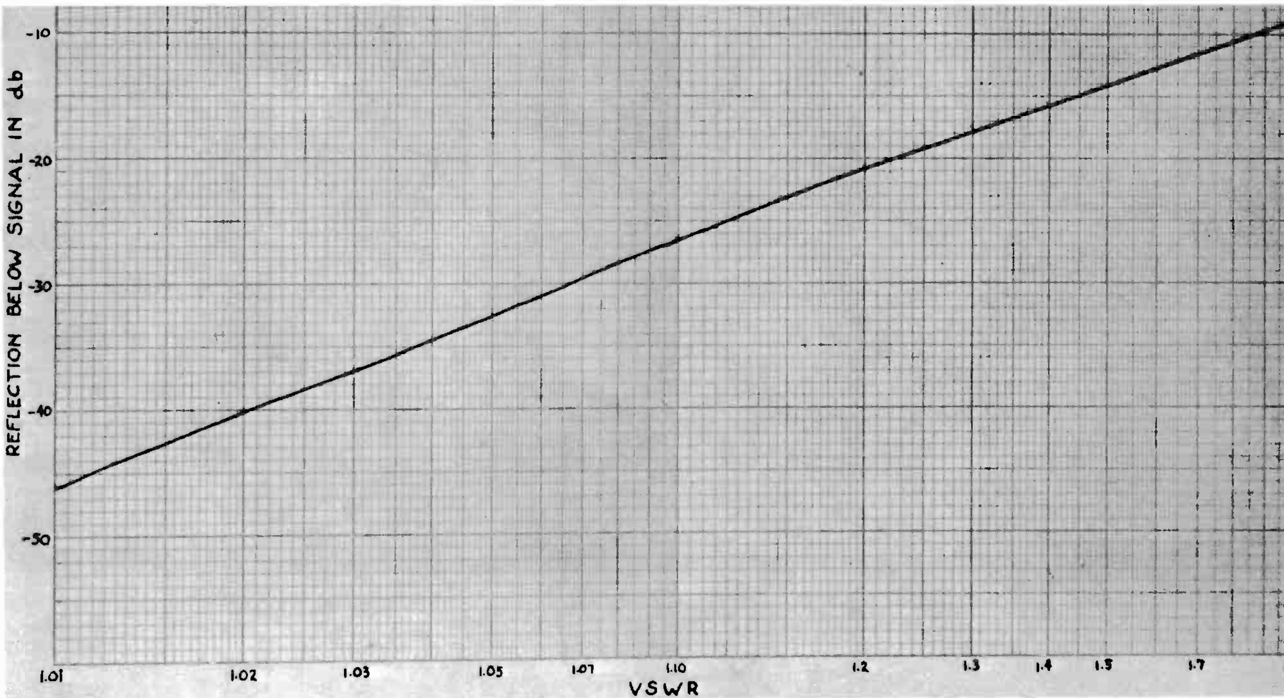


Fig. 8-5. Reflected-to-incident wave vs VSWR curve.

8-3. DIRECTIONAL COUPLERS AND THEIR USE

It would be desirable to have a tapping device which suppresses reflections, and in addition, is matched in every direction. Such a directional coupler can be constructed in many ways. Generally, current and voltage are sensed separately and are added in one direction, subtracted in the other. Theoretically, the difference may be made zero, but practical limitations exist. Such a directional coupler is sketched in Fig. 8-6.

The desirable tap loss is determined by position in the system, and various values can be constructed, ranging from 10 db to more than 30 db for a 4-way directional coupler. The insertion loss, in turn, is determined by the tap loss. There is a theoretical limitation on insertion loss due to the power lost in the tap. The smaller the tap loss, the higher the insertion loss. For a tap loss of 3 db

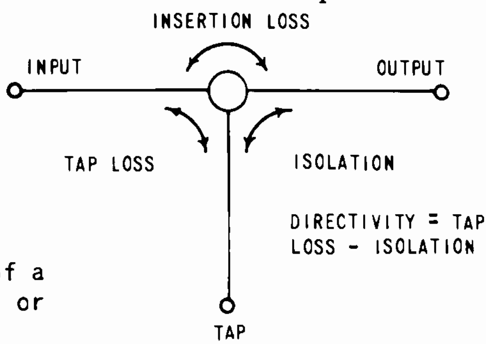


Fig. 8-6. Diagram of a directional coupler or tap.

(1/2 the power), the insertion loss must, of necessity, also be 3 db. Theoretically, a 4-way power splitter has a loss of 6 db (1/4 power). Therefore, a 4-way directional tap with a tap loss of 9 db and an insertion loss of 3 db would be an excellent device.

Actually, insertion losses are proportionately higher (Fig. 8-7), due to unavoidable coil losses. The same relationship between insertion loss and tap loss exists, of course, as well for other taps (capacitive, pressure, transformer, etc.). Fig. 8-7 permits a ready evaluation of the quality of a particular tap or directional coupler.

A directional coupler, in addition, possesses directivity, that is, isolation from the output. While this isolation theoretically might be made infinite, practical limitations due to parasitic capacitances and inductances limit isolation to values from 30 to 40 db. High isolation itself

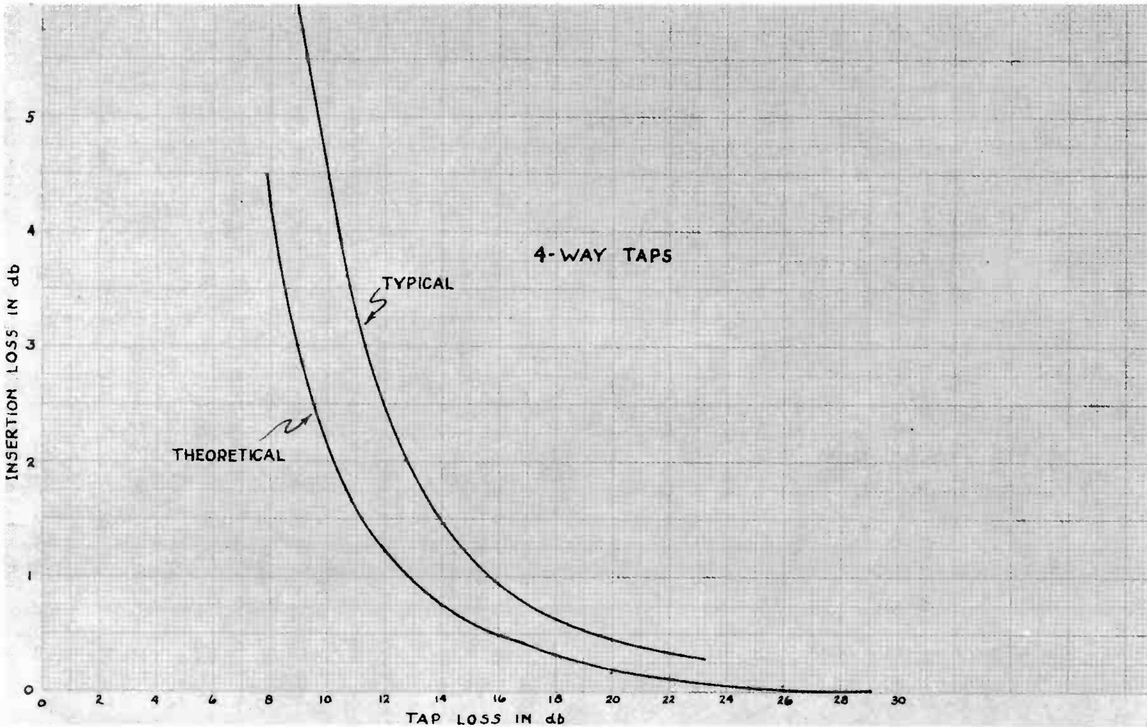


Fig. 8-7. Theoretical and practical insertion loss curves for directional couplers with four outputs.

is of no benefit, rather the difference between isolation and tap loss should be high. This difference is also called directivity, and leads to a substantial increase in the signal-to-echo (S/E) ratio, as is clear from the following example in a typical distribution system (Fig. 8-8).

Only the worst-case condition is considered, 1/2-inch aluminum cable of the critical length of 500 feet at a frequency of 54 MHz. For the first case of the normal tap in Fig. 8-8, assume the signal to be down 15 db at the tap output. The echo then is produced as follows: The signal goes straight through the tap, is attenuated by the insertion loss, is further attenuated in the 500-foot length of cable, is reflected at the point of mismatch, travels back through the

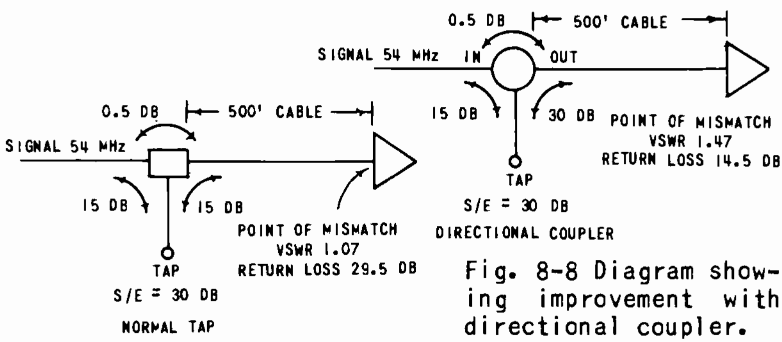


Fig. 8-8 Diagram showing improvement with directional coupler.

cable and goes out through the tap with an additional 15 db loss. From Table 8-1, the required signal-to-echo ratio is 30 db for this length of cable. This value takes into account the loss of the echo when traveling twice through the cable, as well as the time delay of the ghost signal. Since the echo picks up the 0.5 db insertion loss in the tap, the minimum allowable return loss is then 29.5 db, corresponding to a maximum VSWR of 1.07. This value cannot be achieved in a practical system. (For 0.412-inch cable, the maximum VSWR comes out to 1.08.) Practical amplifiers, which can be produced economically, have an input VSWR on the order of 1.2. Even passive devices, such as a well designed standard tap, have a VSWR of about 1.15.*

With the directional coupler in Fig. 8-8 the signal is down 15 db at the tap, exactly as for the normal tap. The echo is produced by signal traveling through the coupler,

*Such taps, measured under very common but incorrect methods, measure a VSWR of 1.1 and are so often advertised.

Table 8-2. Reflections in distribution systems.

Values for Channel 13 in ( ). For letter code see Fig. 8-4.

Case	Mode Of Reflection	Point Of Observation	Mismatch At	Max. VSWR	Remarks
1	Single	I (J)	G	1.43 (1.47)	
2	Single	Worst-Case	Any Input	1.43 (1.47)	
3	Single	B (C)	F	Inf.	Min. Tap Loss 7.5 (7.25) db
4	Single	I	I	Inf.	Min. Isolation 21 (14.5) db
5	Double	I	C	1.35 (1.92)	Mismatch of 4 at I
6	Double	H	G	Input at G:	Output at A:
				1.19 (1.21)	1.92 (3.6)
				1.29 (1.27)	1.43 (1.93)
				1.37 (1.4)	1.23 (1.43)
7	Single	Any	Worst-Case	1.31	Splitter with 15 (10) db isolation
8	Single	Any	Worst-Case	1.31	Bridger directivity 15 (10) db

which is then reflected at the amplifier and goes out at the tap analogous to the standard tap. However, while passing from output to tap of the directional coupler, the signal is attenuated 30 db (an additional 15 db below the signal). Consequently, the return loss at the point of mismatch may be reduced by 15 db, thereby permitting a maximum VSWR of 1.47 (1.55 for 0.412-inch cable). Clearly, this allows a sufficient degree of freedom for arriving at a high-quality system with realistic and economically manufactured components. Directional couplers with a directivity of 15 db at Channel 2 and 10 db at Channel 13 are adequate for all cases and can be economically produced.

The best standard tap is undoubtedly the backmatched transformer tap. While the backmatch cuts down on the reflection in the house drop itself, it has no beneficial effect on reflections in the distribution system. The use of taps other than directional couplers therefore results, of necessity, in a low quality system, which can hardly be considered adequate. The use of such taps is therefore false economy. The cost of directional couplers can be minimized by using the fewest number of couplers with the highest number of taps available(one 4-way coupler should be used instead of two couplers with two taps each). Also, as we have seen, when directional couplers are used, other equipment, such as distribution amplifiers, can be designed more economically. The more realistic input match requirement results, for example, in a far reduced cost for these amplifiers. As is true in so many cases, a really well-engineered high-quality system concept is not necessarily more expensive but many times results in economies by avoidance of overspecifications of irrelevant parameters.

8-4. WORST-CASE CONDITIONS WITH DIRECTIONAL COUPLERS

Even if directional couplers are used throughout the distribution system, attention must be paid to several reflection modes which still can become a problem and must be kept under control. The various cases are summarized in Table 8-2 and will be discussed in detail. In all cases, we assume directional couplers with a directivity of 15 db at Channel 2 and 10 db at Channel 13. For a worst-case condition, all insertion losses are assumed 0 db, distribution cable is 1/2-inch aluminum cable, and RG-59/U is used for the house drops. Refer to Fig. 8-4 for letter code.

Case 1: Mismatch at the input of a distribution amplifier at G. Point of observation at the critical cable length for Channel 2 at 500 feet at TV set I. From Table 8-1, the required return loss is 30 db. With 15 db directivity in the coupler, this allows a maximum VSWR of 1.43 at G. At Channel 13 the most critical point of observation is J and the required total return loss is 24.5 db. With 10 db directivity, a VSWR of 1.47 at G is permissible.

Case 2: Mismatch at the input of any directional tap. Point of observation is critical cable length away. The situation is exactly the same as before, and for this type of reflection an input match of 1.43 at Channel 2, or 1.47 at Channel 13 would be satisfactory.

Case 3: Directional coupler F has an open tap. Point of observation is critical cable length away. A total return loss of 30 db is required. With 15 db directivity this allows a minimum tap loss of 7.5 db in F, because the signal gets attenuated twice by going to the unterminated tap and back up to the distribution line. At Channel 13, with 10 db directivity, the minimum tap loss is 7.25 db.

Case 4: Two TV sets I and I' are fed from a splitter out of directional coupler C. If one set is disconnected, what is the required isolation in the splitter? For RG-59/U, the worst-case condition at Channel 2 requires a return loss of 21 db (see Table 8-1). Therefore, the isolation provided in the splitter must be 21 db. At Channel 13 the isolation must be 14.5 db.

Case 5: Ghosting is also possible by double reflections. For example, reflections are possible in the house drop itself. The TV set normally does not have a match much better than 2, and the VSWR might read 4 if no matching transformer were used. The required return loss is again 21. This calls for a backmatch of 1.35 at each coupler (1.92 at Channel 13).

Case 6: This very important double reflection occurs as follows: The signal is reflected at the input G of the following distribution amplifier, travels back all the way to the preceding line extender, where it is reflected at the output A and then goes in the forward direction out through the directional couplers. In this case, the directivity of the couplers is limited in its effectiveness by the output match of the amplifier. For freedom from ghosts, the total return loss must be 30 db (at Channel 2), and this value must be achieved by the sum of the return losses of both the input and output of a distribution amplifier. Various possibilities are listed in the Table. For all cases, the corresponding readings at Channel 13 are also listed in the Table.

Case 7: A dual-output trunk amplifier feeds two trunk lines. The isolation in the splitter is 15 db at Channel 2 and 10 db at Channel 13, as for directional taps. For a worst-case condition with 3/4 - inch cable, the total return loss must be 32.5 db at Channel 2 (27.5 db at Channel 13), allowing a maximum VSWR of 1.31 at both channels. For an isolation of 10 and 5 db, the maximum mismatch which can be tolerated at the critical cable length is 1.16.

Case 8: Bridger connection to main trunk. This connection must be made via a directional coupler in order to keep reflections in the main trunk out of the distribution system. The maximum allowable mismatch is then as in Case 7. Without a directional coupler, for the worst-case condition, a VSWR 1.045 must be achieved corresponding to a return loss of 32.5. This value is difficult to hold in installed cable, not to mention the problem of connectors and other equipment.

It should be kept in mind that the treatment above is for the worst possible case. Therefore, no additional safety margins must be met. As is amply clear from the preceding detailed analysis, directional couplers are essential for freedom from ghosts. To be economically compatible, it is then necessary to take full advantage of their superior echo suppression characteristics by relaxing other specifications wherever possible; i.e., the match of the couplers does not have to be exceptionally good, and the distribution and other amplifiers can also be designed for improved performance in some other characteristics due to the different match requirement. This approach leads to a CATV system of the highest quality at low cost.

CHAPTER 9

High-Level Distribution

Distribution and system efficiency relates to the number of subscribers which can be supplied from one distribution amplifier, or line extender. For example, in the old, obsolete distribution system with capacitive pressure taps of Fig. 6-2, up to 20 subscribers could be served from one amplifier. Eliminating the capacitive pressure taps and raising the distribution level from 32 to 35 dbmv allows the connection of 24 subscribers (Fig. 6-3) and reduces the variation between Channels 2 and 13 at the subscriber's home from 18.5 to 4 db, certainly a substantial improvement. The correction of the level variations had been achieved mainly through the use of directional taps, while the increased distribution level from 32 to 35 dbmv allowed the connection of 4 additional subscribers.

9-1. DISTRIBUTION EFFICIENCY AND OPERATING LEVEL

As indicated in Fig. 6-3, even in a more favorable distribution system there is an equalized loss (cable) of 13.5 db and a flat loss of 8.2 db. Thus, a significant amount of amplifier gain must be wasted to compensate for flat loss, that is, total insertion loss of all the directional taps. With cable loss unavoidable, it is highly desirable to use all amplifier gain only for cable equalization and to reduce insertion losses to the minimum. As is clear from Fig. 8-7, insertion loss may be reduced by increasing tap loss of directional couplers. However, since the subscriber level must be held constant, an increase in tap loss also necessitates an increase of distribution level. A further analysis of this point leads to the conclusion that nowhere in a CATV system is a high signal operating level desirable except in the distribution system. A high operating level in the main-trunk must lead to increased distortion and reduced cascadeability only. Even in shorter systems where cascadeability and the ultimate in system dynamic range is not required, there is still normally no benefit in raising the operating level of the main trunk. The distribution system, by contrast, benefits directly from a higher signal level, since more subscribers can be connected per amplifier. This is brought about by reducing the wasteful flat loss (insertion loss) and putting it to good use by adding extra subscribers. Consequently, it is the distribution amplifier which should have the highest output capability of any CATV amplifier and it should certainly not be below that of a main-trunk amplifier.

It is rather startling to find that the entire CATV industry developed in the opposite direction for a while, mainly due to a lack of appropriate amplifiers. In the beginning, with vacuum-tube amplifiers, a type of high-level distribution system often was used—however without use of directional taps, or otherwise engineered level diagrams. Generally, the distribution system was treated as the stepchild of the whole CATV system, with little thought or design given to it. Line extenders, as the name implies, were little more than relatively "unimportant," but necessary adjuncts to a CATV system. With typical output capabilities from 35 to 40 dbmv for these line extenders, it was clear that even after a short main-trunk run, a maximum distribution level of only 28 to 32 dbmv was possible, which in turn resulted in a very inefficient and costly distribution system. Such a low-level distribution system is a typical case of false economy and neglected overall system design.

In the meantime, distribution systems have been thoroughly analyzed and amplifiers designed specifically for high-level distribution have become available, often called distribution amplifiers for differentiation from the older, now obsolete, line extenders. It has been shown that such a well designed, high-level distribution system allows more than twice the number of subscribers per amplifier than was previously possible. With amplifier cost for a high-level distribution amplifier only slightly higher than for the old-time line extender, a significant reduction in distribution system cost is achieved, together with a marked im-

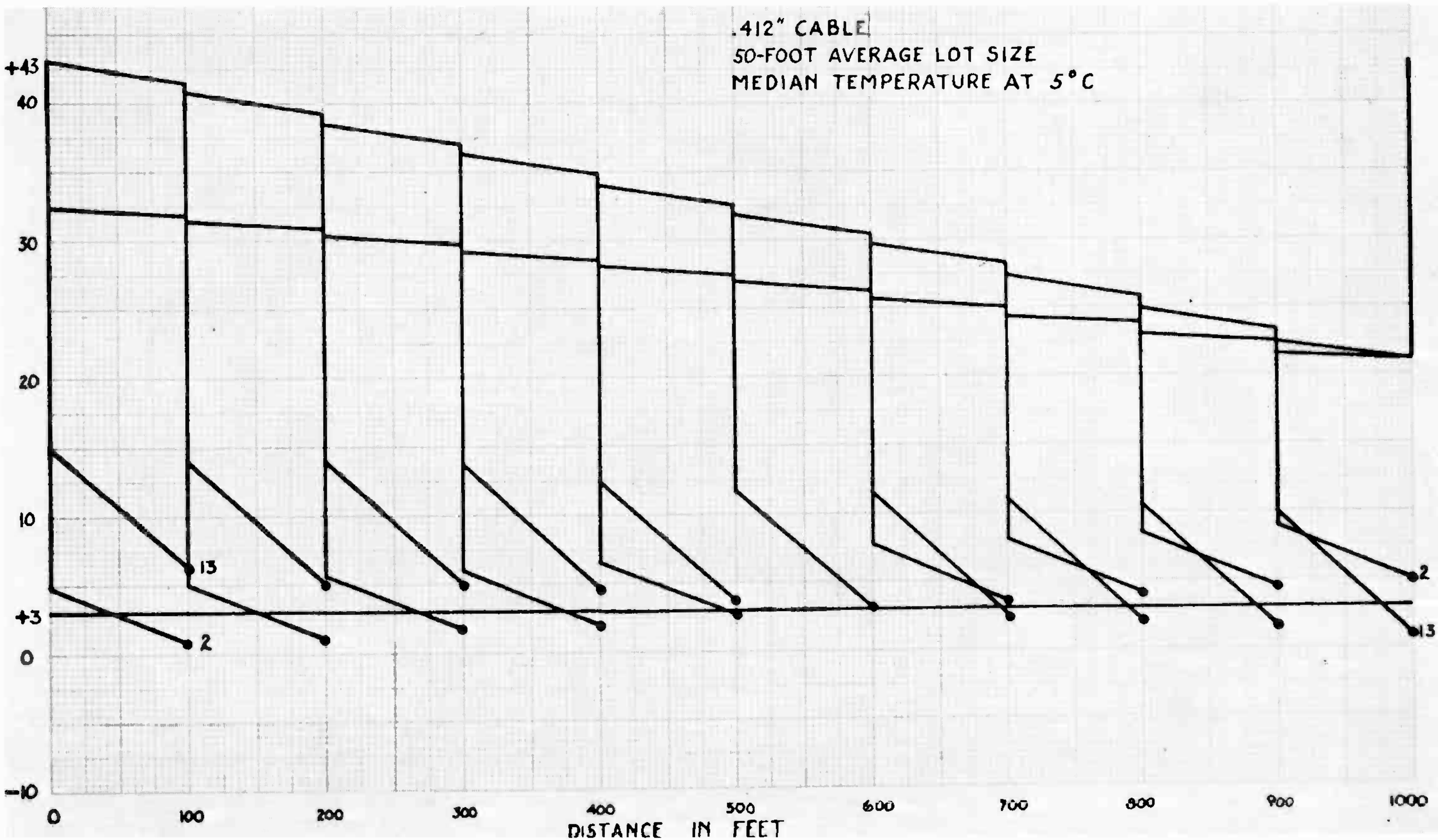


Fig. 9-1. Level diagram for a high-class distribution system.

provement in system performance and reliability due to the reduced number of amplifiers for a given number of subscribers.

9-2. OPTIMUM DISTRIBUTION LEVEL AND LEVEL DIAGRAM

We have determined that a high-level distribution system is desirable to decrease insertion losses. The question now is: What is the optimum signal level to be used in a distribution system?

As distribution level is raised, it is found that from a certain level on, no additional subscribers can be added since all amplifier gain is used to compensate for cable losses, and insertion loss is at this point already reduced to an insignificant amount. To allow the connection of additional subscribers, and to make a further increase in distribution level worthwhile, it is now necessary to increase amplifier gain.

Amplifier gain is, in turn, determined by spacing, tilt variation, and economic considerations. A two-stage amplifier is most desirable for both economic reasons and a high system dynamic range. Optimum spacing for such an amplifier is typically between 20 and 25 db (see Fig. 7-3 for a determination of optimum gain from measurements of a modern two-stage amplifier). This gain range also corresponds closely to the maximum gain available in latest solid-state circuitry, allowing several db for compensation of cable temperature variations and also for unavoidable production spreads in the manufacture of amplifiers.

With this information, it is possible to develop a system level diagram for high-level distribution (Fig. 9-1). In the level diagram shown, a distribution level of 43 dbmv is used. With this value, spacing of distribution amplifiers is at 1,000 feet for .412-inch cable and up to 40 houses may be connected per amplifier, based on an average lot width of 50 feet. Such a distribution system results in excellent system efficiency. Careful analysis of different operating levels shows that even with a distribution level of 50 dbmv no more subscribers can be served, while a level of 41 dbmv, 2 db lower than in Fig. 9-1, results in the loss of 4 houses. Therefore, a 43 dbmv level represents the ultimate for the particular system shown. Quite generally, it can be shown that best efficiency is achieved for a distribution level about 40 db above the nominal subscriber's level. There are minor variations from this value, depending on the gain margin of the distribution amplifier and the quality of the directional taps used in the system.

Fig. 9-1 represents the level diagram of an actual high-class distribution system in use today. Similar diagrams may be drawn for different lot widths and cable; however, the distribution level itself remains unchanged. From such optimized system design it is readily possible to standardize to a large extent with inherent added economies. For example, it is easily verified that 1/2-inch cable used in distribution systems results in about a 20% increase in system cost, since the small savings in amplifier gain per given number of subscribers is more than offset by the increased cable cost. Hence, standardization on .412-inch cable for distribution systems greatly helps in reducing component and system costs.

It should be mentioned, in conclusion, that in the past, bridger amplifiers were often used at very high output levels, often the highest level in the entire system, while line extenders were operated at a low level. This procedure is very questionable when we consider that bridger amplifiers, due to their multiple outputs, have of necessity a reduced output capability as compared to single-output amplifiers. Thus, bridger amplifiers should actually be run at a lower level. A similar case exists with dual-output amplifiers, and it is necessary now to re-examine the choice of operating levels in various portions of a CATV system from an overall viewpoint.

9-3. DUAL-OUTPUT AMPLIFIERS AND CASCADING

Frequently, it becomes necessary to split trunk lines, or distribution cable, in order to serve a particular area more effectively. The haphazard use of splitters leads to an unavoidable system degradation due to addition of extra flat loss, jumper cables, and other discontinuities. Obviously, the best approach has been the use of amplifiers with built-in splitters. Equalization errors can so be avoided. But even then, gain and output capability is reduced for such an amplifier by about 3 db. The reduction in output capability also applies to bridger amplifiers.

There are two basic approaches which may be taken when engineering a system with amplifiers of varying output capability. The first method, often used in the past, ignores the extra distortion caused by working some amplifiers closer to their overload limit on the pretext that if it does not happen too often no problem will arise in the overall system. For example, bridger amplifiers with an output capability at that time of 45 dbmv were used at a 43 or 44 db level. As can be readily be shown, a 1 db margin requires a 6.8 db derating for the rest of the system, corresponding to less than one fourth the system length otherwise possible. Such a procedure is not compatible with high - quality system design. It must also be mentioned that overload margins of 1 or 2 db are inadequate for any practical system, because of the variations with temperature and for other reasons which must be expected.

In a more sophisticated system design, amplifiers are operated at approximately the same safety margins in the distribution system. For example, a dual-output distribution amplifier is operated at a level of 40 dbmv, while a single-output amplifier is operated at a 43 db level (Fig. 9-1). If the difference in output capability is 3 db, equal overload margins result in both cases. Also, derating for cascading is unaffected regardless of how many single- or dual - output amplifiers are used in a particular cascade. Operation at equal margins is easily accomplished. If we remember that a dual output amplifier usually also has a proportionate decrease in gain, we see that operation at equal signal input levels automatically results in equal output overload margins. Due to the reduced gain of a dual output amplifier, it is necessary to shorten the span following such an amplifier. For the example given, a reduction from 1,000 to 800 feet is satisfactory. It is then only necessary to design a new level diagram for this spacing and work out the details. With such a concept, distribution design becomes extremely flexible since no attention must be paid to special derating procedures for different system layouts.

In the main trunk, as we have discussed earlier (Chapter 6-1), a much lower operating level is preferable. Margins from overload for individual amplifiers are much higher. All of the more complicated amplifiers, such as dual-output amplifiers, AGC amplifiers, bridger combinations, etc., usually have a somewhat reduced output capability, generally by less than 3 db. Often it is advantageous to operate the main trunk at identical level and spacing, since gain reductions in amplifiers are usually compensated by circuit design. Precise derating is then made, using the method given in Appendix I.

9-4. MAIN-TRUNK DERATING WITH HIGH-LEVEL DISTRIBUTION

Since the distribution system is operated in high-level distribution at a different safety margin than the main trunk, the question is then: How far must the main trunk be further derated to allow for the higher distortion of the distribution system in order to have an acceptable dynamic range for the whole system?

The procedure for determining overload margins is given easiest by offering a few examples. Assume a modern main-trunk and distribution amplifier with an identical overload level in the vicinity of 60 dbmv and a noise figure of 10 db. The optimum system level from Fig. 6-1 was then 36.5 dbmv, if used throughout the whole system. The maximum cascade without safety margins was 220 amplifiers, and the total safety margin for 100 amplifiers was 7 db.

As we have seen, it is desirable to raise the distribution level for maximum system efficiency. With the concept discussed in the previous section, we can assume single-output amplifiers operating at a 43 dbmv level throughout, since dual-output amplifiers and bridgers are operated at a reduced level proportionate to the decreased output capability. With built-in automatic temperature compensation in distribution amplifiers, it is quite possible to run a considerably longer cascade than with old line extenders. Arbitrarily, we assume a cascade of 10 distribution amplifiers, including a bridger for this example. The combined output capability of 10 amplifiers is 50 dbmv. Operated at 43 dbmv, we have a 7 db safety margin. It will now be necessary to drop the main-trunk level somewhat from the system level found above as 36.5 dbmv. For, as we raise the operating level of part of the system (the distribution system), we must drop the level in the other portion of the system (the main trunk) to arrive at the maximum system

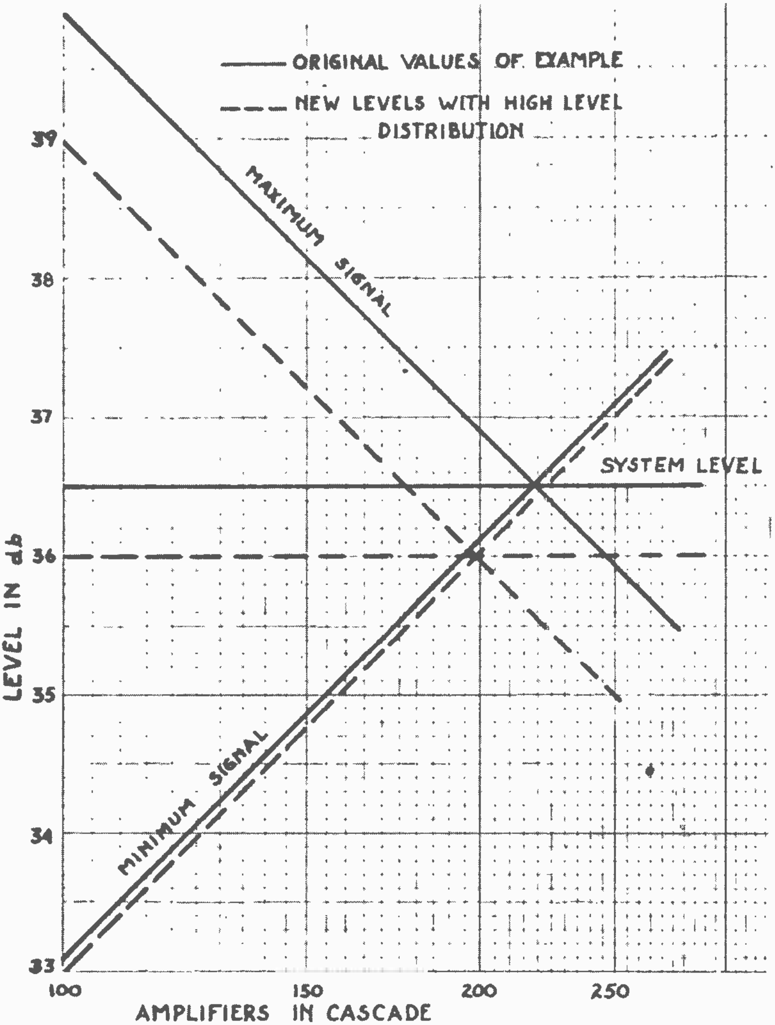


Fig. 9-2. Derating curve for high-level distribution.

dynamic range. The calculation is readily performed using power addition,* and we find for this example that the main trunk must be derated by 0.95 db. This main-trunk derating must be made from the combined overload level of the main trunk only, which is 36.7 dbmv, and is slightly higher than for the cascade including the distribution amplifiers. Thus, system level would have to be reduced to 35.7 dbmv.

*For mathematics, see Appendix IV.

It is now necessary to examine the effect upon noise in order to determine changes in system dynamic range and cascadeability. As is easily shown, the minimum system level decreases only to 36.4 from 36.5 dbmv for this example. There is, therefore, a loss in system dynamic range of 0.7 db, which results in a slight reduction of cascadeability. This derating is shown graphically in Fig. 9-2. The new main-trunk level is then determined at 36.0 dbmv, and the maximum main-trunk cascade without safety margins is reduced to 200 amplifiers; maximum distribution cascade is 10 and distribution level is 43.0 dbmv.

High-level distribution has become a reality with modern, low-distortion amplifiers, and the slight degradation of the already very large dynamic range is insignificant. It is also clear that high-level distribution cannot be achieved with the old line extenders, even in very short systems using ten to twenty main-trunk amplifiers. The example given above for high-quality amplifiers indicates the procedure which generally must be followed for determining system derating for different types of amplifiers, as well as for system levels.

CHAPTER 10

Amplifier Controls

While amplifier controls for gain, tilt, and other functions are commonly provided, the need for these controls must be examined seriously. If an amplifier was correctly aligned for a certain spacing at the factory, what need should there be to make adjustments in the field? Also, can these adjustments be performed satisfactorily in the field?

10-1. THE NEED FOR CONTROLS

On examining these points, we find that the need for adjustments arises basically from two reasons: systematic errors of a faulty system concept and unavoidable tolerance in the installation of a system. For purposes of this discussion, let us assume that the head-end equipment is functioning correctly and its signal output remains constant and at the correct level.

The foremost source of error then lies in the use of jumper cables. For all-band systems, a jumper length of one inch causes problems, and there is no way of avoiding them except by eliminating jumper cables entirely. Other methods, such as aligning an amplifier with the jumper cables attached, are unsatisfactory because the mismatch at the end of both jumpers will be different in the system as compared to the setup on the test bench. Also, on a test bench, additional jumpers are often used for connection to the sweep generator and detector, so that correlation to performance in the system is poor. More on this in Chapter 13.

Obviously, the complete elimination of jumper cables necessitates the use of specially designed amplifiers which have become available since 1964. In many older systems, jumper cables are a major problem. The magnitude of the effect was discussed in Chapter 5 and is determined by the amount of mismatch. The effect on the frequency response depends on the length of the jumper and results in a change of signal level from one channel to the other with, for example, Channel 13 either increased or decreased in level as compared to Channel 7.

By using gain and tilt control, some partial compensation may be possible. However, these controls were never designed for this purpose and, generally, no control can be designed to compensate for response errors due to jumpers because of their unpredictable effect. If jumpers must be used, as is the case in all earlier systems, the use of these controls for frequency response correction results in a partially corrected overall response, although still far inferior to that obtainable by factory alignment of equipment designed for use without jumper cables. Instead of a flatness of ±0.1 db, a typical residual error of ±1 db per amplifier remains in the system after the best adjustment has been made.

Assuming a high quality system not using jumper cables, what other errors must be corrected by adjustments of the amplifier? An analysis shows that all these errors can be treated together as incorrect electrical length of the cable. These errors are due to inaccuracies in spacing, aging of the cable, and temperature changes. Since all of these effects behave the same electrically, they might be compensated for by a single control which affects gain and tilt simultaneously in the proper proportion or, of course, by separate controls. The single control is preferable from the viewpoint of ease as well as accuracy of adjustment. An additional adjustment is often used to compensate for variable changes in system loss due to the use of taps, directional couplers, bridging amplifiers, etc. Before going into a detailed discussion on the function of the various controls, it is desirable to answer our question regarding the ability to make satisfactory field adjustments.

10-2. ACCURACY OF FIELD ADJUSTMENTS

The accuracy of an adjustment in the field depends not so much on the human element, but rather on the inherent accuracies of instrumentation which considerably limit the usefulness of field adjustments. A very common and very faulty procedure consists of going from amplifier to amplifier with a field strength meter and "diddling" all accessible trimmers and controls in a random fashion until some apparently uniform reading is obtained. The basis for this commonly encountered procedure is a poor understanding of system concepts, plus a panic-stricken attempt to make a system operate in the shortest amount of time without true knowledge of the actual source of difficulty. Sometimes such a system is swept (response tested using a sweep generator) afterwards and found to have a horrible response—far from flat. This is then shrugged off as bad cable or connectors and generally ignored as long as mediocre pictures are available at the subscribers' homes. If complaints about picture quality keep coming in, the whole process is repeated over again. In addition, the signal level of the entire system is adjusted as the need arises. If there is too much "snow" (noise), the level is increased; if there is too much "windshield wiping" (distortion), the level is decreased. Needless to say, such methods result in a very marginal system, far inferior to the capability of the individual components of the system.

Obviously, no greater accuracy can be obtained in system adjustments than is possible with the built-in accuracy of instrumentation. The major sources of inaccuracy are the field strength meter, jumper cable, and test points. The most popular field strength meter in use today possesses a published accuracy of only ±3 db; the very best CATV level meters are accurate to ±1.0 db. Test points of older amplifiers are often mismatched (VSWR of about 10:1) and have in addition a tolerance of ±1 db. A jumper cable from the test point to the field strength meter has an additional effect, which is variable and causes some channels to read high and some channels to read low. A typical case is illustrated in Fig. 10-1. The curve shows the error of an unmatched test point when using 5 feet of jumper cable to the meter. A perfect, flat test point is assumed and zero meter error. The total error varies from -1.5 to -7.5 db at various frequencies, depending on the length of the cable. If we keep in mind that a total field adjustment error of 6 db results in half the possible system length, it is clear that all test points must be back-matched. However, even with back-matched test points, there still remain the other sources of error mentioned above. It appears, therefore, very difficult and often futile to obtain any resemblance to a flat system response by field adjustments, particularly if long cascades are involved as in the main trunk.

When we take into account uncompensated temperature changes in the cable, which cause different readings at different times of the day, it is clear that field adjustment should be made only after consideration of all the errors involved, and then only if an adjustment of at least 2 db seems required. A better total accuracy cannot be ex-

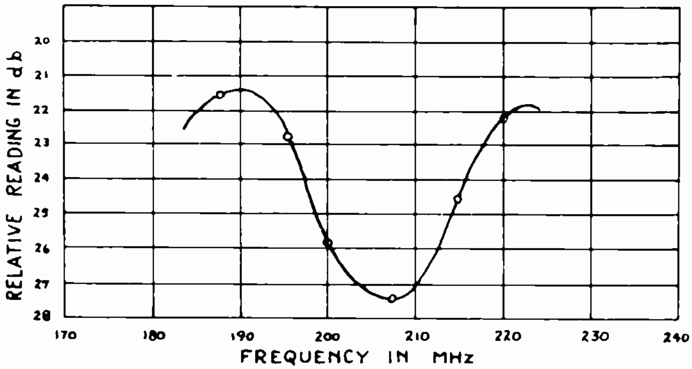


Fig. 10-1. Curve showing error with old-style 20 db test point.

pected with even the very best field equipment available today. To state it in other words, to try and make amplifier adjustments in the field to keep the outputs within, say, 0.1 db of the desired values as read on a field strength meter, is perfectly meaningless. On the contrary, such a procedure insures that the amplifier output itself is set far from flat, while it probably was aligned correctly before. Since these errors are often compounded in CATV systems, the end result can be disastrous.

For many low-quality systems, and all systems using jumper cables must be included in this group, some cut-and-try adjustments as described above may be necessary to make the system operate in some fashion. Due to the unavoidable residual errors per amplifier position, only a relatively small number of amplifiers can then be cascaded, and a severe degradation in system performance is to be expected.

In high-quality systems, alignment within 0.1 db can be achieved at the factory, and also realized in the system if no jumper cables are used. Correction for errors in spacing is then possible by using a single tilt-compensated gain control (described below) at one frequency. Adjustment at a single frequency cuts down on inherent test equipment errors. However, there still remains a rather large unavoidable error per amplifier, far larger than the basic amplifier accuracy of 0.1 db. In order to arrive at the highest degree of perfection, all manual field adjustments must be eliminated in the main trunk and automatic circuitry employed to take care of the necessary correction for spac-

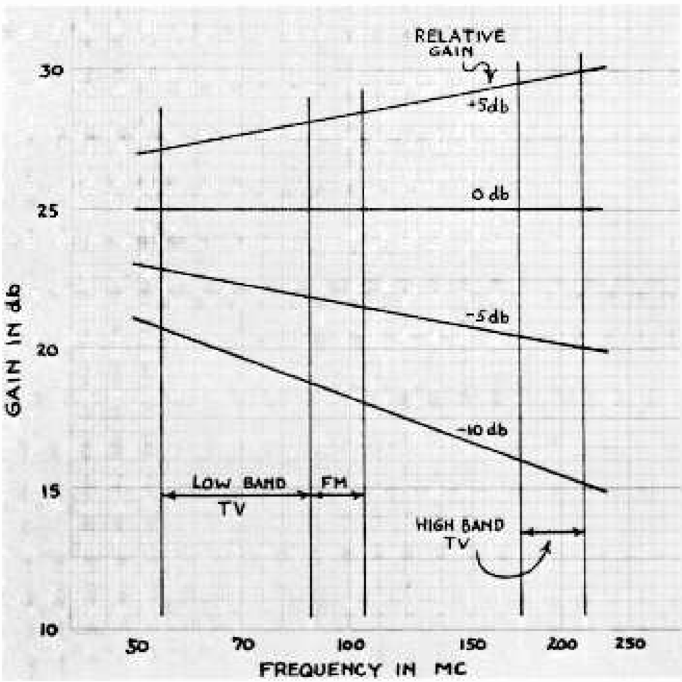


Fig. 10-2. Graph showing action of tilt-compensated gain control.

ing errors and the like. Systems with this concept are described more fully in Chapter 11.

10-3. TYPE AND ACTION OF CONTROLS

Regardless of whether controls are automated or adjusted manually, the basic function is the same. The action of a gain control is to vary the gain of the amplifier at all fre-

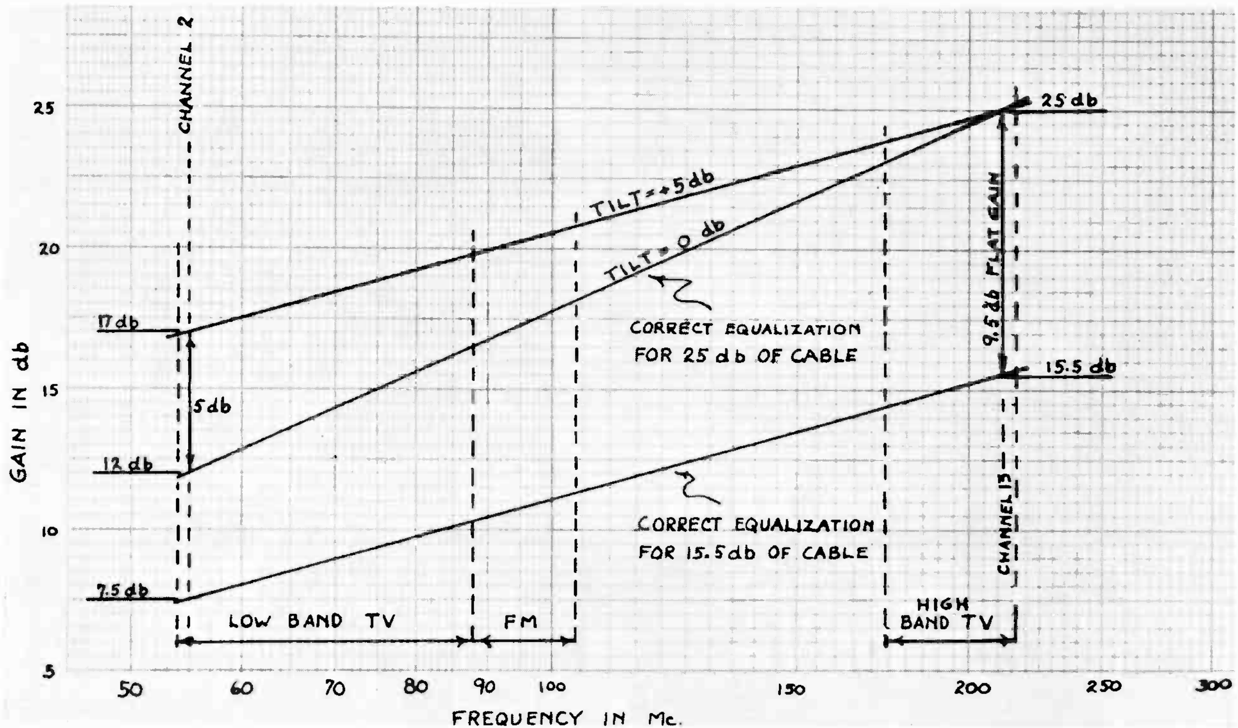


Fig. 10-3. Graph showing function of tilt control.

quencies. In early pieces of CATV equipment, this was a flat control; all frequencies were adjusted simultaneously by the same amount. In many amplifiers, changing the gain also changed the frequency response in random fashion. In both cases, additional adjustments of other controls and trimmers became necessary, even at the factory, to obtain correct equalization for different lengths of cable. Then, in turn, the gain control had to be readjusted and so forth. It is clear that such a procedure cannot be performed in the field.

In the better type of amplifiers, a so-called tilt-compensated gain control has become standard. This control changes the response simultaneously with gain so that different lengths of cable are accurately equalized without the further adjustment of another control (Fig. 10 - 2). The action at Channel 13 is then about twice that at Channel 2. Assuming a flat alignment with 25 db of cable and 25 db of gain, the action of the gain control is therefore as shown in the figure. For example, if the gain is reduced by 10 db at Channel 13, it is reduced only 4.5 db at Channel 2 (-10 db curve). If a cable of 15 db were used instead of 25 db, again a flat response would result. Thus, a single control provides perfect correction for different lengths of cable due to errors in spacing, temperature changes, or aging.

Such a tilt-compensated gain control is analogous to a loudness - compensated volume control familiar from audio circuitry. Naturally, the circuit design for such a tilt-compensated gain control, over such a wide bandwidth as used in CATV, is quite an achievement, although it is just about as difficult to engineer a flat control over this frequency range. A flat control is, of course, undesirable because it necessitates the additional adjustment of a tilt control.

The action of an ideal tilt control is pictured in Fig. 10 - 3. Gain at Channel 13 is held constant, while gain at Channel 2 is changed by 5 db in the figure. A well designed tilt control of this type permits ready adjustment of the proportion of flat gain to equalized gain without affecting the total gain. Such an adjustment is necessary in certain bridger or line extender applications. There is no need for such a control on the main-trunk amplifier unless the gain control is not fully tilt - compensated, or external splitters are used.

The action of the tilt control is clear from Fig. 10-3.

At a normal setting of 25 db spacing, gain at Channel 2 is 12 db and 25 db at Channel 13. With the tilt control set at 5 db, Channel 2 is brought up 5 db to 17 db without changing Channel 13. The slope of equalization is therefore decreased. By calculation, it is easy to show that equalization for only 15.5 db of cable is now provided. Since the total gain was unchanged, we have now 15.5 db equalized gain and 9.5 db flat gain, as compared to 25 db equalized and no flat gain previously. Hence, a tilt control changes the proportion of flat to equalized gain.

Needless to say, many tilt controls in presently used equipment fall short of the desired action, and actually affect gain at Channel 13 simultaneously by giving a hinge point at a wrong frequency. With such a control, successive adjustment between tilt and gain is required. This is cumbersome in the factory, and impossible to adjust accurately in the field.

In lieu of these controls, sometimes pads are used for gain changes in steps. Naturally, for a main trunk, coarse steps are too inaccurate. Also, a pad at the input increases the noise figure; at the output it decreases the overload level. The correct pad position is obviously between stages. in order to keep the effect of changes in gain upon noise figure and overload reasonably small. Pads, of course, cause a flat loss in gain which is undesirable as pointed out above. The logical solution is, therefore,a tilt-compensated gain control between stages in an amplifier. With a well working tilt-compensated gain control, only one such control is needed in a main trunk amplifier.

Any control in addition to a single gain control on a maintrunk amplifier is a silent admission that tilt compensation is not effective or correct, that equalization is not as flat as it should be, and that there are other faults which should have been corrected by better circuitry. Among such controls are separate low-frequency equalization switches or controls, and pads at the input of an amplifier.

Similarly, in distribution systems, input attenuators are undesirable. A tilt control of the type described above is capable of adjusting flat loss continuously without affecting noise figure or overload significantly. This adjustment is needed to compensate for the insertion loss of directional couplers, which might be installed at a later time. As the taps are installed, the gain and tilt controls are reset to obtain the normal distribution level. The procedure is to adjust gain first at some high-band channel, and then tilt at some low-band channel. With properly functioning controls such a two-point adjustment leads to the correct amplifier equalization over the entire frequency range.

In addition to the controls which affect frequency response, there must be an output level adjustment in the AGC amplifier. This adjustment is best accomplished at the factory and depends on total system length. Since voltage regulation is used in solid-state equipment, an accuracy of better than ±1% can be achieved, while a field adjustment would lead to an accuracy of only ±3%. Therefore, this adjustment is normally internal to the amplifier. As we have seen in Chapter 6-1, main-trunk system level does not vary greatly even for different amplifiers and different system lengths. Factory preset system levels have become standard in modern, high-quality amplifiers at levels ranging from 30 to 35 dbmv. There is little advantage to using a different level for short systems unless, of course, the entire system is marginal in performance.

Another control has to do with cable powering. Since only a limited number of amplifiers can be powered from one power supply, it must be possible to interrupt the power on the cable without affecting the TV signals. Normally, a switch, plug, or internal strapping is provided to allow powering an amplifier either from the input or output, or to pass power simultaneously through the amplifier. Actually, two switching positions are sufficient for all practical application—input and through powering. However, for convenience, output powering is often provided also, since it eliminates making a trip to the next amplifier down the line. None of these adjustments are operational controls, and they are set only once during the initial system installation. Response adjustments, on the other hand, have to be made with any change in temperature, unless some other form of compensation is provided, such as a fully automatic system.

CHAPTER 11

Automatic CATV Systems

Automatic CATV, where the human element is completely excluded, is most desirable for precise system operation and freedom from maintenance. As we have seen in the previous chapter, a field adjustment accuracy of ±3 db results in only half the system length otherwise possible, or a degradation in system dynamic range of 6 db, both of which are serious system degradations. Modern amplifiers with their large dynamic range function well even with inexact field adjustments and less accurate system levels. However, it is still desirable to avoid needless system degradation due to human errors, measurement inaccuracy, and other factors. This applies to problems during original system installation as well as to later system maintenance.

11-1. REASONS FOR AGC IN CATV SYSTEMS

An analysis shows that in a complete CATV system every control function can and should be fully automated, with the exception of some adjustments in the feeder (distribution) system. Automation can be perfected to the point where even during initial system installation no field adjustments are required in the main trunk, and construction in building block fashion becomes possible. An automatic CATV system leads to the ultimate in system reliability and freedom from maintenance together with optimum system dynamic range, or system length due to the elimination of field adjustment errors. Also, full automation is a prerequisite for some of the more difficult future system installations which are not practical with the older semi-automatic systems not featuring automatic spacing. Before going into details of automatic CATV, it is worthwhile to briefly examine the variations in a CATV system which are to be corrected.

Typical variations in field strength (fading) of received TV signals are on the order of ±10 db and not exceeding ±15 db.* Control of these variations is readily accomplished by AGC, following the same principle as used in radio and TV circuits. Such AGC circuitry is incorporated in the head end. By the time the TV signals are applied to the main trunk, the maximum signal variation is reduced to less than ±0.5 db for the better type of head-end equipment.

Since the signal levels are now virtually constant, AGC in the normal sense appears to be redundant when used in the cable system. The question is then: What possible sources of error still remain in a CATV system if the output signals available from the head end are held constant?

Several of these errors have already been discussed. They can be distinguished as variations between individual channels, and changes at all TV channels simultaneously. Faulty or wet connectors, cable faults, head-end equipment trouble, drastically wrong amplifier equalization—all of these lead to problems with individual channels. Strip-type AGC (discussed below) is the only form of AGC which will permit some partial automatic compensation for these variations. However, in most cases the fault is serious enough that only correction of the error at the source is sensible. It must always be kept in mind that as system levels deviate from the norm, system dynamic range and quality decrease. A signal decreased somewhere by 6 db picks up a proportionate increase in noise level, even if corrected later on by AGC circuitry or a manual adjustment. Deviations from the optimum system level must be held to a minimum throughout the system. Fortunately, the faults mentioned above are rather pronounced and relatively easy to correct.

Faults with connectors are often due to incorrect installation procedures. This also applies sometimes to cable, particularly with respect to omission of expansion loops, bending radius too small, sharp kinks, and the like. Correction of these faults is accomplished by using the proper installation techniques and by the replacement of defective connectors or cable.

Variations in the head end should, of course, be taken

*See Appendix VII, Reference 22.

care of there. High quality head-end equipment provides a constant output for wide ranges of fading, environmental changes, and other conditions. All other types of head-end gear, particularly vacuum-tube equipment, needs frequent maintenance and readjustment of output levels for satisfactory system performance.

Although normal AGC will decrease variations due to the causes mentioned, it can never fully compensate for these errors. The reason is that AGC, of necessity, must operate either from a pilot carrier or the combined TV signals themselves. In both cases, no correction for variations between channels is possible, but rather the strongest signal or the pilot carrier itself is the one that is held constant by AGC. All other signals are free to change as before. With composite AGC (see below), for example, should the signal of Channel 11 increase by 10 db at the input of an AGC amplifier, the output will reflect only a slight increase (less than 1 db for a good AGC circuit) at Channel 11. However, adjacent channels will then be suppressed by approximately 10 db. This action is unavoidable and indicates that even the best all-band AGC circuit cannot correct directly for arbitrary variations between channels.

Jumper cables (discussed in Chapter 5) cause a more continuous, symmetrical variation over both TV bands. The exact effect is determined by several undefined mismatches and no simple correction is possible. In using jumper cables, the effect on frequency response becomes excessive after only a few amplifiers. One possible method of correction lies in the use of strip amplifiers, which would have to be used in every second or third amplifier position. This rather costly approach, plus the other serious short comings of strip amplifiers, make this concept impractical. It appears that there is no alternate solution other than the complete elimination of jumper cables, external or internal to CATV amplifiers, even if only one inch long.

System changes affecting all TV channels simultaneously include changes of cable attenuation with temperature, changes in amplifier gain with temperature or voltage, and changes in system equalization due to errors in spacing. Of these changes, amplifier variations are readily corrected by better circuitry, and we have to consider in detail only variations related to cable, either due to temperature or incorrect length.

Automatic temperature correction for cable changes consists of automatic gain and tilt control. In addition, a fully automatic system must of necessity provide for automatic correction of spacing errors and automatic control of system level. With the incorporation of the latter two, CATV systems may be constructed in building block fashion with the assurance of far superior system performance than if manual field adjustments were made.

11-2. AGC CONCEPTS FOR CATV

Automatic gain control systems* are a type of servo or feedback system where some error is sensed and corrective action is taken electronically. For example, in the typical AGC system, changes in output level are sensed and

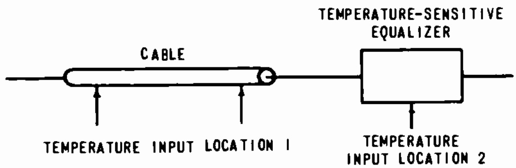


Fig. 11-1. Block diagram of an open-loop AGC system.

fed back inversely proportional to control amplifier gain, so as to keep the output more nearly constant with varying input signal levels. Two basic types of servo systems are distinguished—open- and closed-loop systems. In an open-loop system (Fig. 11-1) the method of correction is not directly related to the error, but to some other parameter loosely related to the error. For example, since cable attenuation changes with temperature it would be possible to develop a circuit or equalizer that also changes with temperature, but in the opposite direction, so as to compensate the cable. An automatic correction concept of this type is an open-loop system, since the change in cable attenuation is not sensed directly, but rather the temperature at the location of the equalizer, which is only loosely, or not at all, related to the effective temperature of the cable. Such an open-loop system is a very imperfect, although automatic, form of correction.

*Appendix VII, References 23, 24.

In a closed-loop system (Fig. 11-2) the error to be corrected is sensed directly and any desired degree of perfection may be achieved. Typical for this concept is any normal AGC system familiar from radio or TV circuits, where output level change, however small, is sensed and the necessary correction is made. A closed-loop system represents the best possible form of servo system, but it usually is considerably more expensive than a simple open-loop system. In CATV systems, cable attenuation changes are to be corrected for high precision; in a closed-loop system, these cable attenuation changes must be sensed directly, rather than some other parameter such as ambient temperature at some amplifier location.

In addition, to the basic open- and closed-loop principle of AGC design, there are several concepts for CATV AGC which relate to bandwidth of AGC circuits as well as choice of control signals.

Early AGC amplifiers had a very crude gain control

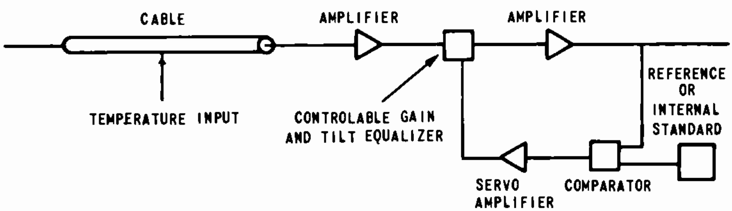


Fig. 11-2. Block diagram of a closed-loop AGC system.

action without tilt-compensation or other means of tilt correction. Often the frequency response changed severely in an arbitrary fashion when gain was changed. Without the circuit design techniques available today, other approaches had to be tried in order to keep variations between TV bands and TV channels under control. This led to the development of split-band AGC amplifiers, as well as AGC-strip amplifiers with separate AGC for each channel.

In a split-band amplifier, filters are used to separate high and low TV bands, which are then passed on to individual AGC amplifier sections. The outputs are then combined by filters. Such a split-band circuit can restore the balance between high and low bands, but it does not restore the correct equalization within each band. Therefore, this principle has been extended with a separate AGC amplifier for each channel, whereby signal separation and addition is again accomplished by filters. This separate strip AGC is only workable where a limited number of TV channels are used—usually, a maximum of seven channels—because practical filters cause too much phase shift at the band edges to be compatible with good quality TV (particularly on color programs). Also, of course, the cost is prohibitive if strip AGC is used for a larger number of channels.

Split-band AGC amplifiers have some merit, especially as an addition to improve the performance of older systems, or in fully automatic systems of extreme length where several hundred amplifiers must be cascaded.* In systems other than fully automatic systems, the use of strip or split-band AGC amplifiers poses new problems in system maintenance, since no simple adjustment of system tilt is then possible in the head end; instead, individual adjustment of each AGC station is necessary. With the typical inaccuracies of field adjustments, the merits of these AGC systems become dubious indeed. In normal modern CATV systems where the common pitfalls such as jumper cables have been avoided, there is little need for split-band or strip-type AGC, and a well designed all-band AGC system represents the best solution. At the same time, such an AGC system can be designed to correct for changes in cable attenuation due to temperature, aging, or incorrect spacing.

In any multi-signal AGC system the choice of the control carrier is very important. Early AGC amplifiers were custom-tuned to one TV channel carried in a particular system. To achieve better consistency, composite AGC came into use for a while where a number of TV channels (usually high band) controlled AGC action. However, with composite AGC, action is a function of the number of channels in use in a particular system. In addition, most systems deriving control voltages from TV signals directly are affected by modulation, which is undesirable because it results in reduced level accuracy.

Another approach is to use a pilot carrier system to overcome the disadvantages of the other concepts. The pilot carrier is usually inserted at the head end and combined with

*In such a system, split-band AGC amplifiers might be used at every tenth AGC location, for example.

the TV signals. With a pilot carrier system, AGC amplifiers are designed to sense deviations in pilot level and to make corrections for the TV signals accordingly. The pilot might be chosen anywhere outside the TV bands in order to avoid interference with TV signals; that is, below Channel 2, between Channels 4 and 5, between low and high band, or above Channel 13. A close scrutiny reveals that the only satisfactory frequency for a pilot carrier is above Channel 13. This can be readily shown by considering the action of correctly designed gain and tilt controls which was discussed in the previous chapter. Automatic CATV calls for automation of a tilt-compensated gain control. Moreover, a correctly functioning tilt control must have a hinge point at Channel 13. With a pilot frequency close to Channel 13, tilt action remains fully operational with AGC. It is easy to see that any other pilot frequency involves a complete system realignment job for the slightest change in system tilt. This includes intricate field adjustments of at least each AGC amplifier which, as we know, cannot be performed satisfactorily in the field. With the pilot carrier operating at the correct frequency (above Channel 13), system tilt is easily reset in the head end or at an individual amplifier without impairing AGC action or AGC range. A pilot-carrier frequency above the highest TV signal is also conducive to better AGC resolution, because attenuation changes are more pronounced at the highest system frequency.

11-3. TEMPERATURE COMPENSATION

Temperature compensation of the CATV amplifiers themselves is readily accomplished by using high-quality electronic components designed to work over the expected temperature range, usually from -50 to +160°F. Any remaining small change is then easily corrected by the judicious use of special temperature-sensitive components. Careful design can stabilize the gain of a typical transistorized amplifier to ± 0.25 db over a temperature range from -50 to +160°F, which is considerably better stability than common for most tube-type equipment. In an open-loop AGC system, the amplifier can be designed to compensate cable changes directly, provided the temperature of both amplifier and cable is the same. Such an open-loop system, because of its low cost, might be used in the feeder system where high precision is not required, or in the main trunk in conjunction with a tightly controlled closed-loop system.

The major system change with temperature is always due to changes in the cable. The change in attenuation with temperature is normally taken as 1% per 5^{o}C or .11% per ^{o}F. Therefore, the electrical length of cable changes +10% to -14% from the normal ambient value for a temperature range from 160^{o} to -50^{o}F, respectively. This is indicated in Fig. 11-3, where the different cable characteristics are indicated by parallel straight lines, which is the exact equivalent of changing the electrical length of the cable, from 1100 feet to 860 feet, for a 1,000-foot length at a temperature of 68^{o}F. At 25 db spacing, the change of +10 to -14% corresponds to a change of +2.5 db to -3.5 db. The question is now how to compensate for this unavoidable change, and at what interval.

Temperature compensation may be achieved by purely passive methods. For example, in the early days of CATV, flat amplifiers were used together with passive equalizers for the cable.* Such an equalizer, often a plug-in type, might be designed with temperature-sensitive electronic parts to compensate for changes in cable temperature. Similarly, the equalizing circuits in an amplifier may be made temperature sensitive and designed to compensate for changes in cable temperature. The fault with temperature compensation methods of this type is that they are never accurate. Obviously, the compensation must be different for different lengths of cable. If there is any error in spacing, the compensation is off and cannot be reset by a simple gain adjustment. More seriously, the compensating circuit is often at a different temperature than the cable. For example, the equalizer or amplifier may be in the shade of a tree, while the cable is exposed to solar radiation. The resulting error is normally such that it cannot be compensated for anywhere else in the system.

It is easy to see that these so-called open-loop systems fail to give the performance required in quality systems. In older, shorter CATV systems, open-loop compensation may make an otherwise impractical system workable.

*This concept is considered obsolete today due to the decreased system dynamic range. (See Chapter 7.)

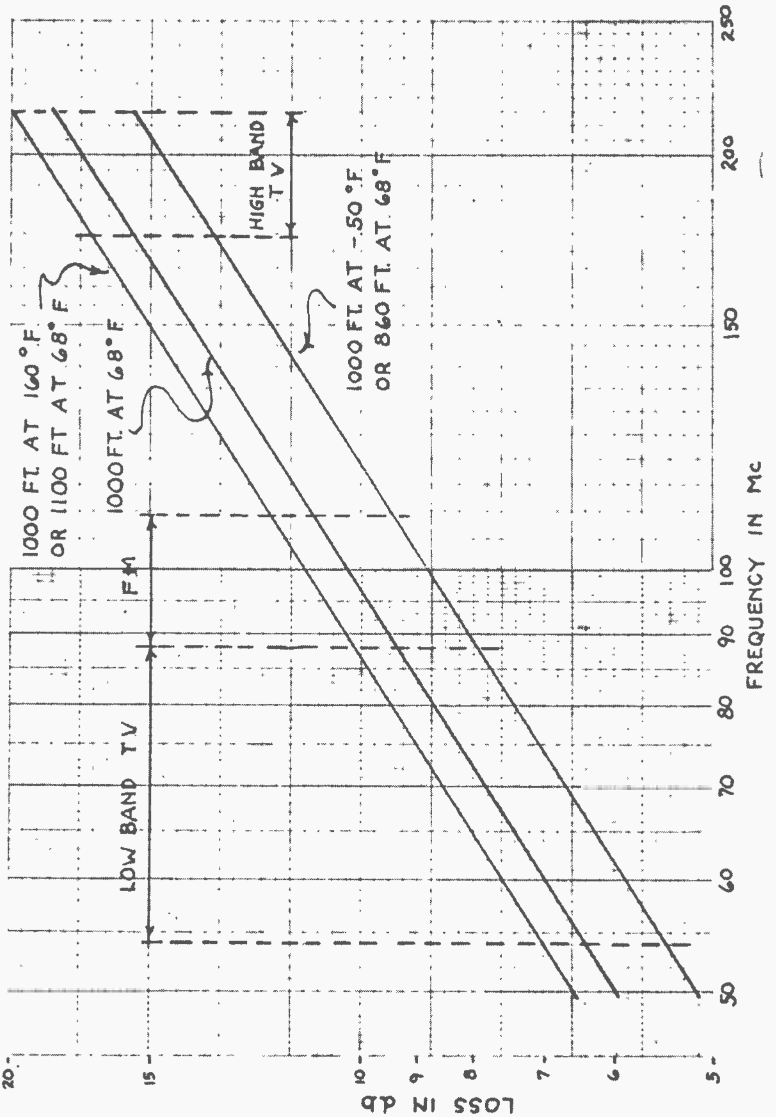


Fig. 11-3. Typical temperature characteristics of coaxial cable.

But these methods fall into the same category as jumper cables and pressure taps, and have no place in a modern system except possibly in conjunction with a well working closed-loop system, or in feeder applications.

In a closed-loop servo system, the error signal obtained relates to the change to be corrected. This error signal, in turn, controls some mechanism which produces the desired correction. In CATV, the change to be corrected is the varying cable attenuation with temperature. First, it is quite clear that it would be impossible to measure the temperature of the cable directly, due to the varying absorption and radiation of heat along the cable.

Fortunately, in a closed-loop system, the temperature of the cable is unimportant, but rather the change in attenuation is of concern. Therefore, if constant signal levels are applied to the cable, the output of the cable or the input to the following amplifier accurately reflects the effective temperature of the cable. For example, at 25 db spacing, an amplifier input level increase of 3.5 db reflects an effective cable temperature of -50°F, provided that the input to the 25 db length of cable was held constant. The change of signal level at the input of an amplifier, therefore, contains all the necessary information for a fully automatic servo system for temperature compensation.

Temperature compensation of the closed-loop type at a single frequency is accomplished by standard AGC, because such a system keeps the output constant for varying input signals. With constant levels from the head-end equipment, all changes are due to temperature, and the sole reason for AGC in CATV systems lies then in the need for temperature compensation. While normal AGC is all that is required for a single frequency, temperature compensation at all frequencies requires the change in gain to be proportioned differently for the different channels. Analysis shows that automation of a tilt-compensated gain control (discussed in the previous chapter) leads to perfect automatic temperature control by automatic gain and tilt control.

11-4. AUTOMATIC SPACING

The main reason that perfect temperature compensation is achieved by an automatic tilt-compensated gain control is that temperature acts on the cable exactly as if the length of the cable were changed, as shown in Fig. 11-3 by the parallel lines. Therefore, the tilt-compensated manual gain control, which was required to correct for errors in spacing or to accommodate differently spaced systems, also provides a perfect manual correction for temperature changes. Converting this control to automatic operation not only results in precise temperature compensation (automatic temperature control), but also corrects for errors in spacing (automatic spacing). If one considers the whole CATV system and makes a thorough analysis of the errors involved, and of system reliability, it becomes quite obvious that automatic spacing is as much a necessity as temperature compensation. It simply is not possible to measure cable length or attenuation to an accuracy of 1 db in the field. There are too many unavoidable sources of error and variables which preclude any semblance of accuracy. This area was already covered to some extent in Chapter 10. Even if an accuracy of 21 db can be achieved, this is still insufficient for highest quality due to the cumulative error in cascading amplifiers.

Finally, the question of how many AGC amplifiers are needed in a system must be treated. We have seen that, due to temperature, a maximum increase of 10% in cable attenuation must be expected. To this value should be added normal spacing errors and some allowance for cable aging. The total maximum increase from all sources might therefore be 15%. This increase in attenuation must be equaled by a corresponding increase in gain of the AGC amplifier. For circuit design reasons, it is impractical to build an amplifier where the maximum gain is more than 10 db higher than the nominal gain. Also, an increase of 10 db has to be offset by a decrease of about 15 db at low temperature. Thus, a total AGC range of 25 db results—quite large considering that highly accurate tilt compensation must be provided at every point.

With 15% corresponding to 10 db change, the maximum spacing between AGC amplifiers is then about 70 db. Thus, with 22 db spacing, every third amplifier at 66 db should be an AGC amplifier. For a system which is to be constructed in an area where temperature compensation must not be provided over the extreme range from -50 to +160°F, AGC amplifier usage may be reduced, for example, to every fifth position. On the other hand, AGC amplifiers having a smaller AGC range may have to be spaced closer for highest system reliability. Other advantages of closer spacing might be that by using a smaller AGC range, a more accurate tilt compensation is possible. However, this increased accuracy of the amplifier, which allows a greater system length, is likely to be obliterated by inaccuracies in the cable unless the system is fully integrated. Also, with more AGC amplifiers in the system, reliability suffers because an AGC amplifier is, of necessity, considerably more complex than a straight manual amplifier.

The concept of AGC in every position is a poor one for the same reason. It must always be kept in mind that a simple amplifier can always be made better in the important characteristics than a complex circuit like an AGC amplifier which must perform additional critical tasks. AGC in every position would, therefore, result in poor reliability, economy, and reduced system dynamic range due to lower output capability and higher noise figure.

11-5. CATV SYSTEM INTEGRATION

Based on the automatic concepts discussed so far, it is possible to standardize and integrate CATV system design and construction. These automatic servo systems include precise system level control, automatic tilt correction, compensation for changes of cable attenuation, and correction for normal errors in spacing. It is then possible in integrated CATV systems to match cable and amplifier fully to each other, so that no cable cutting or splicing is required. Any normally arising errors are then corrected by automatic spacing, which is a feature of the AGC amplifier. No field adjustments are made, and measurements are only taken during the initial systems installation, if at all. The accuracy of the system level is then guaranteed by the AGC amplifier itself, which internally relates the system level to the DC voltage of a reference zener diode. An accuracy of better than 0.1 db (less than 1%) can then be achieved for the system level, which is far superior to attempts of setting up a system the old fashioned way. In its logical conclusion, a fully automatic, integrated CATV system leads to complete automation of the main-trunk and bridger amplifiers.

Spacing for system design is then determined by cable length rather than loss. For example, it is found that attenuation values of 1/2-inch cable are normally within ±3% regardless of the specific cable manufacturer. For system design the cable loss should be taken at a median temperature rather than the extremes. In the past, system layout often followed a different procedure. Cable loss at 68° F was prorated for some high temperature such as 120° F, and then an additional aging factor of 5% was added to come up with a worst-case design for cable loss. Such a procedure was necessary because active equipment available then did not provide for temperature compensation; on the contrary, many early transistor amplifiers lost gain as the temperature went up.

Modern CATV amplifiers are generally made with two types of temperature characteristics—zero T.C. and cable-compensating open-loop units. Zero temperature coefficient is used with pedestal and underground installations, and in-

Table 11-1. Cable length for 23 db spacing.
(Based on average loss at median temperature)

Cable Dia. (Inches)	Length In Feet	Tolerance
.412	1420	± 60
.500	1800	± 80
.750	2560	±120

dicates that the amplifier does not change gain with temperature. Normal open-loop cable compensation means increased amplifier gain with temperature. In either case the above design practice, which was at one time a legitimate worst-case design procedure, now is incorrect. The theory of error analysis indicates that by using the split-error method far better system quality results. This principle is directly related to the correct choice of system median values. For example, if a system were to operate over the temperature range from -50 to ±160° F, a design based on the extreme temperature of 160° F, must already result in an error at the median temperature and end in twice the error at the low temperature. Statistics show that operation at the median temperature occurs much more often than at the extremes of -50 or +160° F. Therefore, by such a design, system operation under the most common normal conditions is less than optimum. With modern amplifiers, system design is performed at the median temperature. For the temperature range given above, the median is 55° F. This median value may change somewhat, depending on system location; however, this change is usually small. For example, a relaxed temperature range from 0° F to 110° F has the same median of 55° F. All design proceeds, then, based on the average cable loss at the median temperature, and the results are shown in Table 11-1. The tolerance refers to a change of ±1 db.

Determination of cable losses in the field is then eliminated. Such a measurement is generally useless anyway, since the effective temperature of the cable is unknown, and because of typical measurement errors. A far greater accuracy and consistency is achieved by working strictly with footage. This concept is particularly attractive with automatic spacing.

It can be shown that with a normal spacing error per span of ±1 db, probabilities for total errors in three spans are as listed in Table 11-2. Thus, the total allowance for automatic spacing of ±3 db is used only in 7.5% of all cases.

Table 11-2. Probability of total error per three spans with individual tolerance of ±1 db.

Total Error	Probability
Less than ±1 db	67%
Less than ±2 db	92.5%
Less than ±3 db	100%

In fully integrated systems, all main-trunk amplifiers are factory prealigned, never to be touched in the field. While these amplifiers may have the normal controls such as gain and tilt, with automatic spacing and temperature compensation, the result is better system performance than is possible by field adjustments. This relates also to system level. As we have seen in Chapter 6, optimum system level is unchanged regardless of the number of amplifiers operated in cascade. Standardization and automation allow also the complete prealignment, with precision factory equipment, of built-in bridgers or bridger combination amplifiers.

While this precision is much less important with modern amplifiers and their high dynamic range, it is still desirable for the ultimate in system reliability and freedom from maintenance, particularly since factory prealignment does not preclude normal field adjustments if so desired.

Continuing this thought of system automation and integration into the feeder system, unfortunately, leads to no further improvement possibilities, because of the great variations in layouts. However, system maintenance in the feeder system is greatly simplified by correctly working gain and tilt controls (see Chapter 8) and some form of open-loop temperature compensation as discussed above.

CHAPTER 12

Principles of Cable Powering

With the older tube systems 110 VAC power was directly applied to every amplifier, often with a separate power meter, depending on requirements of the local electric companies. Due to the high cost of such a setup, amplifiers had to be spaced far apart, much to the detriment of system performance. With the advent of transistors and their much reduced current drain, methods of using the coaxial cable for both signal and power transmission were investigated. Early systems used DC power; however, it was quickly found that corrosion became excessive as higher currents were required. DC current flowing between dissimilar metals, with humidity present, increases galvanic action of such a "battery" by the process of electrolysis.

12-1. METHODS OF POWERING CATV SYSTEMS

In order to reduce the corrosion problem, AC-powering was introduced. Even with AC-powering, electrolysis is possible if direct-coupled half-wave rectifier circuits are used in the amplifiers, since then the current passing through the cable is a half-wave AC current, also called pulsating DC current because it possesses a relatively strong DC component. For best performance, all modern CATV systems now use a form of AC powering which avoids DC components in the cable altogether. Various circuits of both the half- and full-wave rectifier variety are available, which result in complete protection from corrosion.

There is a significant difference between half-wave and full-wave circuits from the viewpoint of powering efficiency. With half-wave circuits, unless amplifiers are manufactured with both polarities and distributed evenly in the system, a much greater voltage drop results because of the doubled current which must flow during the conduction cycle. A powering concept using amplifiers of different polarities which must be balanced in the system for powering reasons is very undesirable, particularly since full-wave circuits are easily incorporated into amplifiers at low cost. With full-wave powering, rectification efficiency in the amplifiers remains constant, and circuit performance is the same with a series loop resistance as with decreased input voltage. From a practical viewpoint, full-wave powering allows a greater distance from the power supply, and the use of fewer power supplies in a system.

Power supplies for CATV systems are preferably of the regulating transformer variety. As discussed below (Section 12-4), these supplies have proved themselves through

Table 12-1. Typical loop resistance for 1000 feet.

Cable	Resistance in Ohms
.412 Thinwall*	3.36
.500 Thinwall	2.27
.750 Thinwall	1.31
.412 Solid Sheath**	2.19
.500 Solid Sheath	1.44
.750 Solid Sheath	.66
5/16 Strand	1.33

*Sealmetic (Anaconda Wire and Cable Co.)

**Canacom (Canada Wire and Cable Co.)

many years of field use as far as reliability and surge protection are concerned. This transformer furnishes the power which is supplied to the cable. In any transmission of electric power, voltage is of utmost importance, the higher the voltage, the greater the efficiency of power transmission. Due to various electrical safety codes, power voltages applied to the coaxial cable had to be limited, although the cable itself is capable of handling much higher voltages. As a good compromise 36 VAC has been adopted as a standard for cable-powered systems. In order to keep the peak voltage low, a quasi-square wave has come into use, since a perfect square wave generates harmonics which are undesirable. Thus, the maximum system powering voltage is given, with the minimum determined by amplifier requirements. Spacing from the power supply is then determined by the cumulative voltage drop in an amplifier cascade.

12-2. LOOP RESISTANCE AND DROP CURVES

In laying out powering systems it is generally best to disregard the shunt resistance of strand wire in order to arrive at a more conservative design. Strand wire reduces the resistance of the outer conductor in normal system construction due to the bonding at amplifier locations via strand

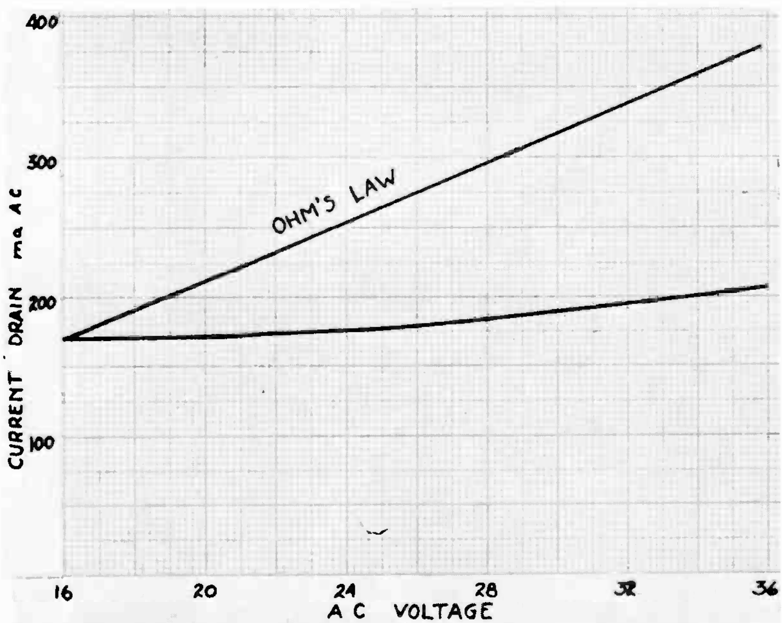


Fig. 12-1. Power characteristics of a typical CATV amplifier.

clamps. This does not apply to underground systems and it is usually best to disregard the action of strand wire altogether. Loop resistances of coaxial cable do not vary greatly, but are different for the two basic cable types, thin-wall and solid-sheath cable. Typical values are listed in Table 12-1.

With loop resistances and the power characteristics of an amplifier known, it is possible to develop voltage drop curves which indicate power reserve in cable powering. The power characteristics of an amplifier relates AC current

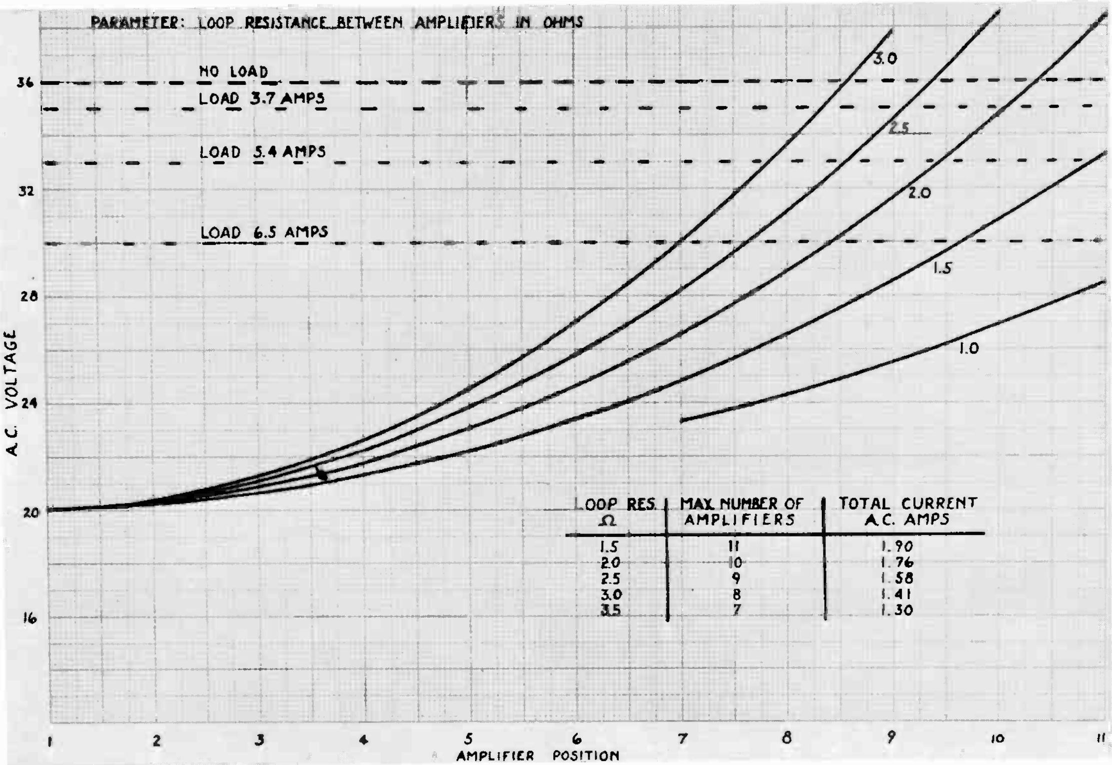


Fig. 12-2. Voltage profiles for the power characteristics of Fig. 12-1.

drain to AC input voltage (Fig. 12-1). CATV amplifiers do not follow Ohm's law due to the servo action to the built-in regulated power supply; that is, current drain is not doubled for twice the voltage. With a high-quality series regulator, current drain is reduced below normal Ohm's law as shown in Fig. 12-1. For a shunt regulator current increases above Ohm's law. This latter type of "brute-force" voltage regulator is not suitable for CATV purposes. Based on the given power characteristics, voltage profiles may be developed for a cascade of CATV amplifiers (Fig. 12-2). This voltage drop indicates the voltage present at each amplifier location for different loop resistances. Loop resistance, in turn, depends on spacing of amplifiers and cable type. For the example given in Fig. 12-2, a minimum useful amplifier input voltage of 20 VAC is assumed. Starting from the last amplifier, the voltage drop is small, because it is caused by the current of only one amplifier. The voltage drop increases with every amplifier as we travel toward the power supply, as does the current in the cable. Also indicated in the graph is the power supply voltage from no load (36 VAC) to a full load condition (30 VAC). The maximum number of amplifiers may then be read where both curves intersect. For example, assume the total load on the power supply is 5.4 amperes, and amplifiers are spaced at a distance of 1,000 feet using .412-inch thin wall cable. Loop resistance is, therefore, 3.36 ohms (from Table 12-1) and extrapolation above the 3.0-ohm curve in Fig. 12-2 shows that seven amplifiers may be cascaded. The total current drain with Fig. 12-1 is 1.3 amperes based on an average current of 185 ma. This leaves 4.1 amperes to power other amplifiers available from the power supply. Other cases may be determined along similar lines from measured power characteristics and a set of voltage profile curves such as in Fig. 12-2. A mathematical treatment is given in Appendix IV.

12-3. LOCATION AND SPACING OF POWER STATIONS

Power supplies should be located strategically at branch points for the best powering efficiency. For example, with high-level distribution (Chapter 9), more distribution amplifiers are cascaded than with the former line extenders, typically 6 to 10. All of these distribution amplifiers are fed from one of several outputs of a bridging amplifier. A bridging amplifier typically has four bridged outputs. Considering all of these factors, it becomes clear that it would be most desirable to locate a power station close to, or adjacent to, a bridger amplifier for best powering efficiency. In our example, one power supply adjacent to a bridging amplifier will furnish power to the bridger itself plus a string of six distribution amplifiers from each of four bridger outputs, or a total of 25 amplifiers. In Section 12-2 the maximum cascade for powering purposes was found to be seven, and this was reduced here to six amplifiers as an added safety measure. Powering 25 amplifiers from one powering location is excellent powering efficiency when we consider (with the data given in Chapter 9) that this corresponds to more than one thousand subscribers. An important rule is, therefore, to locate a power supply adjacent to every bridger location.

The other extreme is an unloaded main-trunk run. In this case, loop resistance is increased, because of the longer main-trunk spacing, and because of increased current drain due to the more complicated amplifiers using extra transistors in such main-trunk amplifiers as AGC and level-control stations, and combination amplifiers which include bridger circuits. On analyzing the voltage drop it is found that with typical CATV amplifiers, power should be applied at intervals of every sixth to tenth amplifier. For extra safety, a second rule of cable powering calls for locating a power supply adjacent to every second AGC location in an unloaded main trunk. All other practical cases fall between these two extremes, and can be treated similarly. In some instances, a smaller booster supply is helpful which allows the addition of a few extra amplifiers for odd layouts.

Power is applied to the cable via a so-called power inserter. This unit consists of a small RF filter to remove interference due to the power line, and circuits which apply the powering voltage in shunt with the RF signal. This power inserter is either a separate unit which is inserted into the cable like a splice, or it might be an integral part of the power supply, or an amplifier. Several modes of powering, which refer to the location of the amplifier, are distinguished by CATV amplifiers which usually have provision to select input, output, or through powering. In "input powering," power is passed from the input connector to the amplifier circuit. No power is applied to the output from the input, although power voltage may be present at the output connector from a different power supply down the line. In "output powering" the situation is reversed, with power for the amplifier derived from a supply feeding the cable somewhere on the output side of the amplifier. Modern power supplies are polarized for "power sharing," which results in added system reliability. In this mode of operation, "through powering" is used where input and output are directly tied together as far as power is concerned.

12-4. LIGHTNING AND SURGE PROTECTION

In systems exposed to the environment as in CATV, special emphasis must be placed on protection against power surges and lightning damage. Excellent performance can be achieved even in bad lightning areas by both system and equipment design. Field tests have been conducted for many years and the knowledge gained has resulted in modern CATV systems which are both lightning and surge proof. The whole concept of surge protection is intimately related to cable powering. It has been shown, for example, that the induced center-conductor voltage is proportional to the low-frequency impedance of the cable system involved. If the cable system is powered from such a low-impedance source as a regulating transformer, lightning-induced voltage surges are reduced considerably. The now obsolete DC powering concept resulted in much higher impedance, and it was much more vulnerable to lightning surges. The first principle is, therefore, to keep surge voltages low by keeping impedance levels low. This applies to induced voltages on the center conductor due to a lightning hit of the outer conductor.

Surges can also be transmitted directly from the power lines. Elimination of these voltages is readily accomplished in the power supply. RF filtering removes high-speed surges and a regulating-type transformer takes out low-speed surges. By proper design of these power supply details, excellent lightning and surge protection is achieved. In addition, various types of lightning arrestors, automatic self-resetting circuit breakers, etc., are often used for extra protection.

Whatever surge voltages remain in the cable system, particularly at locations far away from a power supply, must be eliminated or made harmless by amplifier design. These techniques for surge protection are well known today and are incorporated in the design of modern CATV amplifiers.

CHAPTER 13

Testing CATV Amplifiers

One of the most important functions of a CATV amplifier is to correctly compensate for cable losses, providing both proper gain and frequency response. To achieve the quality level required in modern systems, a flawless test method which avoids all the common sources of error is essential. Chief among the systematic errors are the use of various jumper cables, poor matches, and generally different conditions than would be encountered in the field. All these errors can be avoided only if careful attention is paid to the detailed test instructions which follow.

13-1. EQUALIZATION AND ALIGNMENT

These test methods are particularly designed for the latest high-quality amplifiers for systems not using jumper cables. For these modern systems, a factory alignment to a flatness of ± 0.1 db makes sense, and will pay off in substantially improved performance due to greatly increased cascadeability. For older systems, where jumper cables are used, the same test methods may be used, with or without jumper cables; however, an alignment flatter than ±1.0 db is then a waste of time, since the response would change at least that much when installed in the system. In the future, with improved amplifiers having larger dynamic range, extreme accuracy may no longer be required for normal, short cascades since these amplifiers are very forgiving due to their much lower distortion and noise levels. However, the ultimate in alignment accuracy is still desirable for the highest reliability and freedom from maintenance which is possible with automatic CATV and where long cascades are within the capability of these new amplifiers.

In the following paragraphs, proven, highly accurate test methods are described, using low-cost instruments.

Higher-quality sweep generators, oscilloscopes, etc., may, of course, be used. They do not increase the accuracy of measurement, but they may be more convenient for other reasons such as repair and maintenance, or stability. Special test fixtures mentioned below, such as calibrators, detectors, etc., are required for high alignment accuracy and they may be obtained from several CATV manufacturers. Typical circuits for these are given below.

The first step in the alignment of CATV amplifiers is the calibration of the test setup (Fig. 13-1). The sweep generator, which must cover the range from 50 to 220 MHz in a single band, is connected by a jumper cable to a step attenuator used to obtain calibrated oscilloscope traces. This jumper cable cannot be avoided. Its effect is made small by close impedance matching for a VSWR of 1.1 or better. In addition, the calibrator contains a fixed 10 db

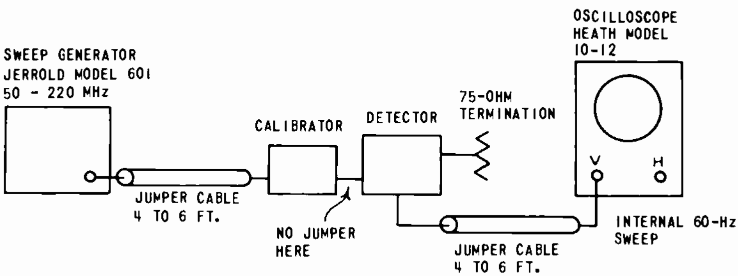


Fig. 13-1. Calibration setup diagram for amplifier alignment.

pad to further isolate reflections, due to the jumper, from the circuit under test.

Any remaining errors are eliminated by indicating the actual calibration trace on the oscilloscope with a grease pencil (Fig. 13-5). While there is no inherent source of error left with this setup, it is still desirable to obtain the flattest possible horizontal trace on the oscilloscope. This is achieved by adjustment of the leveling circuit in the sweep generator, judicious use of the output level control and/or fixed attenuators of the sweep generator, as well as changes in the construction of the detector (different grounding, etc.)

The calibrator, in turn, is connected directly to the detector without coupling pieces or jumper cable, however short. This necessitates, of course, the construction of a special calibrator which has a built-on female cable connector. All wiring inside the chassis box from connector to circuit must be kept to less than 1/2 inch. Where these special pieces of equipment cannot be built or obtained, no attempt should be made to test these high-quality amplifiers; instead, they should be returned to the factory for alignment. The detector has an external 75-ohm termination, and its output feeds the oscilloscope via a jumper cable which is non-critical because it carries only low-frequency signals (better accuracy is possible by using a special, coaxial 75-ohm detector).

All pieces in the test setup, with the exception of the connection from detector to oscilloscope, should have a VSWR of better than 1.1. This is achieved by trimming capacitors marked C_X in Figs. 13-2 and 13-3, and by changes in the layout. The step attenuator of Fig. 13-2 has built-in switchable 0, 1, 2, and a fixed 10 db step. There-

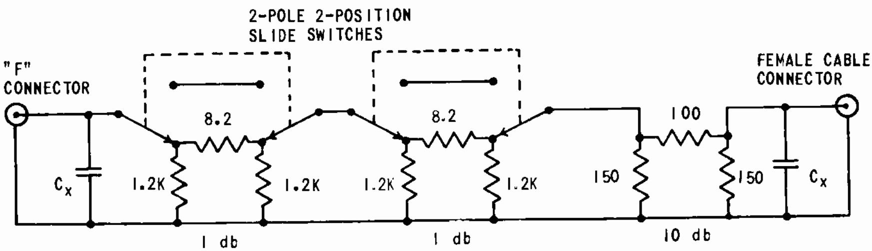


Fig. 13-2. Calibrator schematic.

fore, it is possible to mark three traces on the oscilloscope face at +1, 0 and -1 db. The detector is a standard full-wave voltage doubler provided with a marker input through a 2.2 pf capacitor. The termination is external so that the same setup is suitable for VSWR tests. (With a coaxial detector a special VSWR adapter is used which also serves as a marker inserter.)

No expensive oscilloscope is required, since only low frequencies up to 10 kHz must be displayed. However, provision for 60-Hz line sweep and high sensitivity (10mv/cm) are essential. For precision alignment, a stable reference line of the sweep pattern is essential; consequently, a DC or a DC-restored scope must be used. DC restoration is most easily achieved with a parallel diode at the proper point in the vertical amplifier of the oscilloscope (Fig. 13-4). For proper operation, it is essential

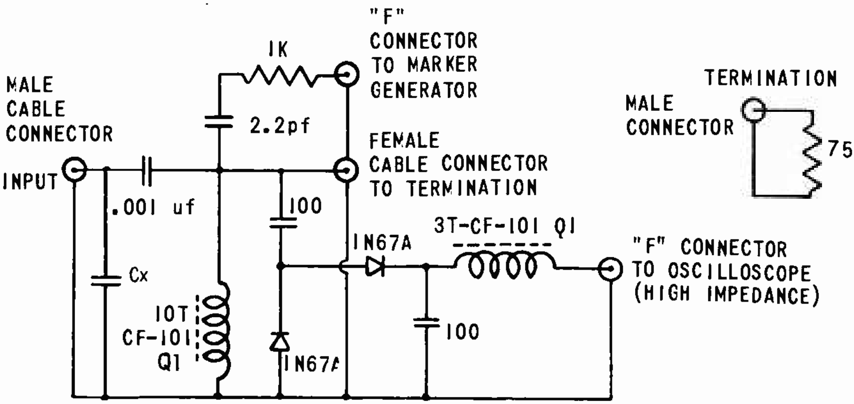


Fig. 13-3. Detector schematic.

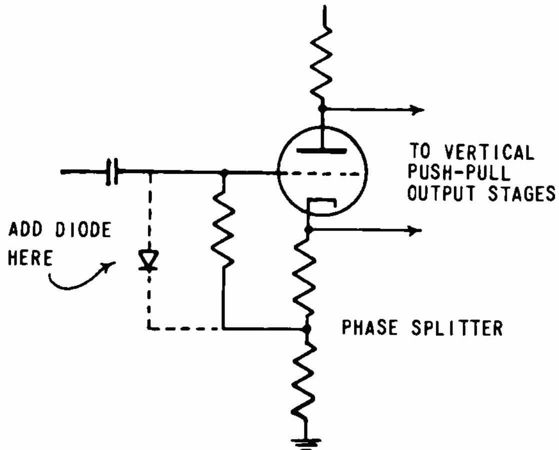


Fig. 13-4. Method of adding DC-restoration to an oscilloscope.

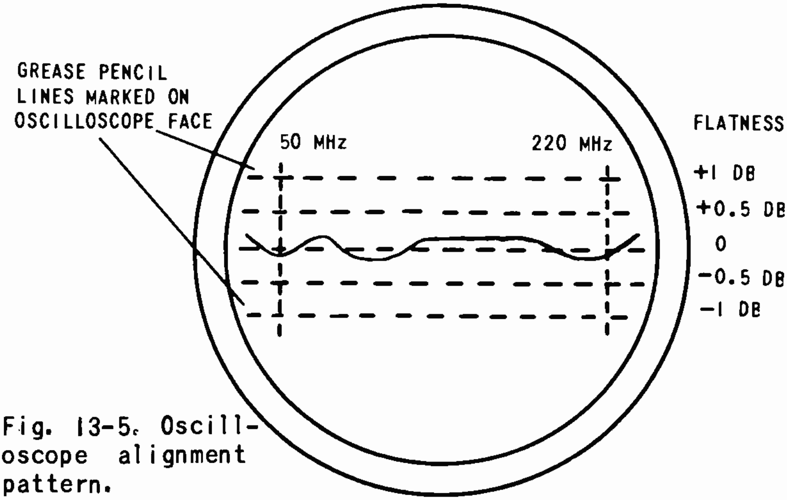


Fig. 13-5. Oscilloscope alignment pattern.

that this diode be connected in the correct polarity, across a point of high impedance and reasonably high signal level, but ahead of the push-pull output section.

The calibration procedure is then as follows: In the setup of Fig. 13-1, adjust the level of the sweep generator and vertical positioning of the oscilloscope to obtain at least a 1/2-inch separation of the three traces obtained with the calibrator. Use maximum vertical gain of the oscilloscope. It does not matter if the DC line (base line) is visible. However, the +1, 0 and -1 db lines should be well separated and located in the center of the screen. Connect a marker generator to the detector and obtain pips at 50 and 220 HMz. Center the sweep generator and adjust oscilloscope horizontal gain to spread out the pattern evenly. Mark the 50 and 220 MHz points with a grease pencil, and also indicate the horizontal traces at +1 and -1 db. Reset the step attenuator to the three positions and check the grease pencil lines. Set the attenuator to the 0 db line.

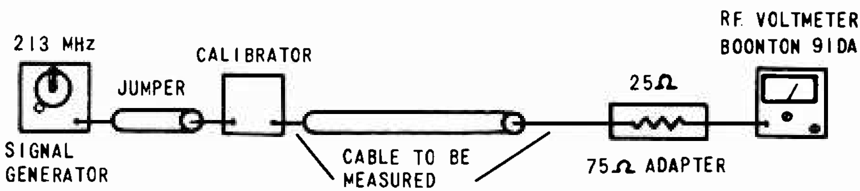


Fig. 13-6. Diagram of setup to measure cable loss.

This completes the calibration, which normally will be accurate for at least a day unless the controls on the sweep generator or oscilloscope are touched inadvertently or if line voltage fluctuations are unduly excessive. Then it will be necessary to readjust the sweep generator and oscilloscope. Rarely will it be necessary to draw new grease pencil lines. A base pattern as shown in Fig. 13-5 should be obtained.

Next it will be necessary to prepare the correct test cables. For example, for 25 db spacing, two lengths of 10 and 15 db each are used. It is therefore necessary to measure these lengths of cable to the required degree of accuracy. The normal field strength meters are unsatisfactory for this purpose, due to their poor range overlap and other built-in inaccuracies. The most accurate test setup at reasonable cost is as shown in Fig. 13-6. This

Table 13-1. Errors and tolerances in alignment setup of Fig. 13-7.

Equipment	Characteristic	Tolerance	Alignment Error	Gain Error
Sweep generator, jumper cable, attenuator and detector	Response not flat[1]	Any	Grease pencil accuracy 0.1 db[2]	None
Step attenuator	Attenuation of 1 & 2 db step	.05 db	None	.05 db
Test cable	Loss	2 db 0.1 db	None[3]	2 db 0.1 db
Detector or Oscilloscope	Nonlinearity	Any	None	None
Amplifier under test	Match	VSWR 2:1	None	None
All equipment	Match	VSWR 2:1	None	None

[1] It is desirable to obtain the flattest possible response before aligning amplifiers, even though this is not reflected in the accuracy. A flat response psychologically is much easier to work with. Therefore, a match with a VSWR of 1.1 should be strived for everywhere.

[2] Please note that this is the basic accuracy of the test setup. For greatest resolution, use maximum oscilloscope gain, so that 1 db corresponds to at least 1/2 inch on the graticule.

[3] See text.

test setup is also convenient for testing attenuators and the like. The accuracy achieved is related directly to the RF voltmeter, a Boonton Electronics Type 91-DA, with a basic accuracy of 2%.

After the cables have been made up, the test and alignment follows the procedure indicated in Fig. 13-7. This is basically the same setup as in Fig. 13-1. The detector and calibrator are simply separated, and the amplifier, with cables, is inserted. Notice that no jumper cables are used anywhere between attenuator and detector. The amplifier adjustments (trimmers, gain, tilt) are then set to obtain the (flattest) response possible, coinciding with the 0 db line on the oscilloscope as shown in Fig. 13-5. The flatness then can be read directly from the graticule, because the logarithmic measure of db is linear to better than 0.1 db over the small range of ±1 db. Therefore, linear interpolation is permissible. For example, if the stretch from 0 to 1 db on the oscilloscope is 1 cm, then 0.1 db corresponds to 1 mm. (In Fig. 13-5, flatness would be ±0.1 db.)

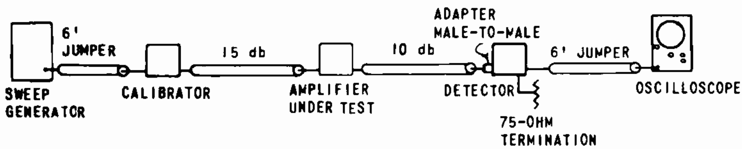


Fig. 13-7. Diagram of alignment and gain test setup.

With this method, no additional gain reading is needed, because obviously the alignment was made so that the total gain of amplifier plus cable is 0 db. Therefore, the amplifier gain is equal to the cable loss which was measured with high precision in the setup of Fig. 13-6.

13-2. GAIN CONTROL AND TILT COMPENSATION

In order to test the range of the gain control of a particular amplifier, it is only necessary to change one of the test cables for one of different length. For example, in Fig. 13-7, the 15 db cable is interchanged for one having 10 db loss at 213 MHz. All adjustments are then repeated as before and the final flatness is read.

It is important to have a clear understanding of gain control range—the range of amplifier gain adjustment through which cable equalization is possible. Gain control range without a flatness specification is meaningless. The specifications for a high-quality amplifier might be:

Flatness	Gain Range
±0.1 db	20 to 25 db
±0.25 db	18 to 26 db
±0.5 db	16 to 27 db
±1.0 db	13 to 28 db
±2.0 db	10 to 29 db

This means that cable equalization to an accuracy of ±0.1 db is achieved for spacings from 20 to 25 db, all the way to a flatness of only ±2 db for a gain range from 10 to 29 db. For high-quality systems, a flatness of ±0.25 db might be the limit. The useful gain range is then from 18 to 26 db.

In some less critical applications, and older systems using jumper cables, a flatness of ±1 db would be sufficient, giving a spacing range from 13 to 28 db. Of course, optimum spacing is determined by signal-to-noise considerations (See Chapters 4 and 5), and the best value of spacing, together with the highest figure of merit, is furnished by the amplifier manufacturer. Gain range then is merely an adjustment for spacing and similar errors, in systems not using automatic spacing. In systems with automatic spacing, no gain control is normally used; rather, only internal adjustments are provided for initial factory alignment. Therefore, no test for gain range is made on these amplifiers.

Another important consideration is the tilt-compensation of the gain control (Chapter 10). This term indicates the use of a single compensated control to adjust the amplifier for different spacing. Obviously, such a control is a necessity in high-quality systems to assure the required accuracy of field adjustments. The older spacing adjustment method, where a multitude of controls (gain, tilt, low-frequency switch, internal trimmers, etc.) had to be touched up for different spacing, was necessitated by poor amplifier design and results in poor system performance, since no satisfactory field adjustment of all these controls is possible. The gain range test in all these cases follows the procedure outlined above,except that after the initial alignment only a single control, the tilt-compensated gain control, is adjusted for different lengths of cable. In this fashion, the tilt compensation of a gain control is checked.

Frequently, a separate adjustment for tilt is needed to compensate for variations in flat loss, particularly in the distribution system. With such dual controls, the procedure is to set gain first at Channel 12 or 13, and then tilt at Channel 2 or 3. A single adjustment of both controls should result in precise alignment if the unit is to be adjusted easily in the field.

The action of a tilt control is tested by inserting additional flat loss. For example, in Fig. 13-7 the input cable of 15 db is changed to 9 db, plus a 6 db coaxial attenuator. This corresponds to a typical cumulative flat loss of directional taps between distribution amplifiers. The gain is still 25 db as before, but the tilt has changed and the tilt control must correct this change. Tilt range is tested similarly by varying the flat loss.

Before going on to other tests, it is worthwhile to consider the inherent sources of error in sweep testing and their elimination in the alignment procedures described. One of the chief causes for faulty alignment lies in reflections and improper cable lengths. Reflections cannot be avoided altogether, and a match with a VSWR of 1.1 must be considered excellent. However, even such a good match causes an error of approximately ±1 db in frequency response for a short length of cable with practically no loss. The effect shows itself as successive peaks and dips, the number increasing with the length of cable.

A short jumper cable of 1 to 2 inches causes mainly a relative shift between high and low band, while longer pieces show variations between channels and thereafter within a channel. Since with a longer cable the loss also increases, reflections are progressively attenuated until finally a point is reached where even a bad mismatch has no visible effect on frequency response. In order to avoid these errors, jumper cables are avoided in the main test section between attenuator and detector. All other jumpers, errors in the sweep generator, detector, etc., are removed from the measurement by using a calibrated sweep. The error in frequency response is therefore determined by the accura-

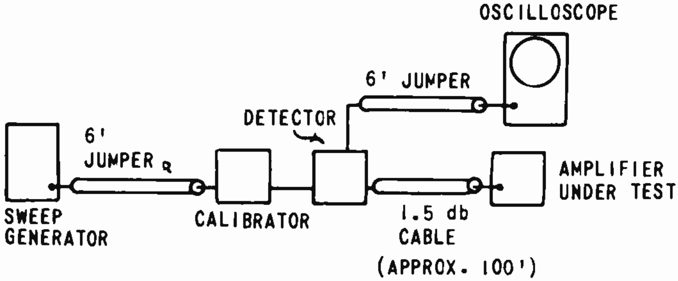


Fig. 13-8. VSWR test setup.

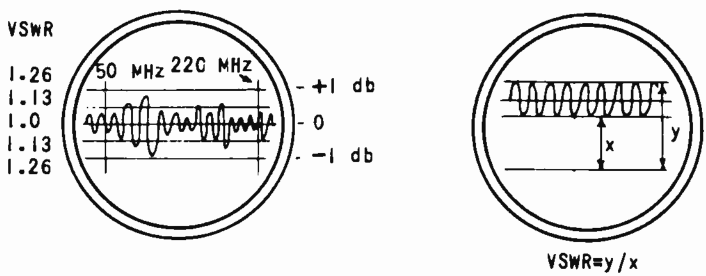


Fig. 13-9. VSWR oscilloscope patterns.

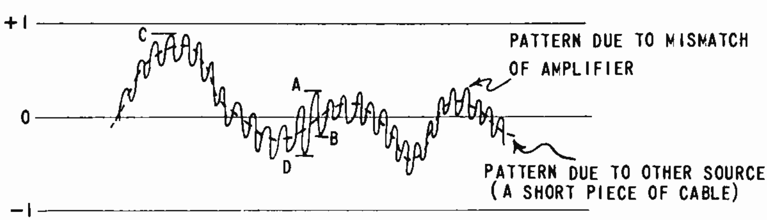


Fig. 13-10. Superimposed VSWR pattern caused by multiple reflections.

cy of the grease pencil lines on the oscilloscope. Generally, this can be held to 0.1 db.

Errors in reading the flatness may be due to an error in the calibrator. This error can be made considerably smaller than 0.1 db by hand trimming resistor values. This is true even if a lower quality db meter is used; even these meters are sufficiently accurate over such a small range as 1 db. The fixed 10 db pad does not contribute any error and a tolerance of ±1 db is satisfactory. Nonlinearities in the detector are eliminated by using a calibrated sweep and DC restoration in the oscilloscope.

The remaining error lies in the length of the test cable. Although an accuracy of 0.1 db is desired for absolute gain measurements, no such great precision is required for frequency response tests. As a matter of fact, since the amplifier performs in the field with inaccurate spacings, it certainly cannot be particular about length of cable on the test bench. This is even more true with modern amplifiers which feature a tilt-compensated gain control. The major cable error lies in the use of other than system cable for alignment purposes, such as RG-59/U. With this type of cable, there are mainly two types of errors: short, sharp dips or spikes which are easily recognized and are ignored in alignment; and a slight change in tilt. This tilt as compared to system cable is typically less than 0.1 db at Channel 2. There is no error in the high-band channels. This small tilt error normally can be ignored in system usage. It could be readily compensated by a small adjustment of the tilt control if so desired. Therefore, for most practical purposes, alignment with RG-59/U cable is satisfactory.

13-3. TESTS OF MATCH AND VSWR

Only the sweep method will be described here since all other methods are less suitable for tests on CATV equipment. The basic errors of commonly used setups are in the nonlinearity of the detector and oscilloscope, inaccuracies in the test cable, and errors due to jumpers. These problems are avoided simply by using a calibrated method; the same setup used for alignment is suitable with no separate calibration necessary. In Fig. 13-8, the termination is removed from the detector, and an accurate 75-

ohm cable is connected with a loss of exactly 1.5 db at 213 MHz. The other end of the cable is terminated by the amplifier under test. Any mismatch at the amplifier end of the cable causes reflections which add or subtract at the detector to form the familiar standing wave pattern on the oscilloscope (Fig. 13-9).

The basic definition of VSWR is given by the ratio of "y" to "x" as indicated in Fig. 13-9. This measurement could be performed directly by reading "x" and "y" on the oscilloscope graticule, were it not for the inherent large errors due to detector and oscilloscope nonlinearities. This inaccuracy is overcome by using a calibrated sweep. It is an easy matter to convert the db steps already indicated on the face plate of the oscilloscope to VSWR. The 1

Table 13-2. Conversion of db to VSWR.

db	VSWR
±2.5	1.78
±2.0	1.58
±1.5	1.41
±1.0	1.26
± .75	1.19
±0.5	1.13
±0.25	1.06

db limits correspond to a VSWR of exactly 1.26; 0.5 db limits (half-way in between) correspond to a VSWR of 1.13, etc. Consequently, VSWR may be read directly to any desired accuracy. For conversion, Table 13-2 may be used.

With this method, the common errors due to nonlinearities are completely eliminated. The pattern obtained must be exactly bisected by the 0 db line as in Fig. 13-9. A pattern as in Fig. 13-10 is obtained if reflections from two sources are superimposed to form one VSWR pattern. Clearly, the VSWR due to the amplifier is indicated by the fine wiggles, and the pattern due to the other source must be disregarded in the measurement. The worst amplifier VSWR in Fig. 13-10 is measured between A and B, not C and D. For greatest accuracy, multiple reflections must be avoided. This is accomplished in the setup of

Fig. 13-8 if no jumper is used between calibrator and detector as is shown.

Another source of error lies in the VSWR cable itself. Obviously, this cable must be of the same impedance as the cable used in the system. This is an important point because nominal 75-ohm cable varies as much as +10%, and this could lead to a worst-case VSWR of 1.23 instead of 1.0, for example. It is therefore advisable to use RG-11/U, or better yet, regular feeder cable for the VSWR test.

The loss in the cable is of extreme importance, since the measured return loss at the detector is better by twice the cable loss as compared to the amplifier mismatch. In order to have a meaningful VSWR measurement, it is

Table 13-3. VSWR with 1.5 db of cable.

Channel 2		Channel 13	
Measured	Actual	Measured	Actual
1.05	1.06	1.05	1.07
1.1	1.12	1.1	1.14
1.15	1.18	1.15	1.22
1.2	1.24	1.2	1.29
1.25	1.31	1.25	1.38

necessary to standardize the loss in the VSWR cable. This loss is made exactly 1.5 db at 213 MHz in order to arrive at a length of approximately 100 feet for the VSWR cable. The exact length is not of importance, but a length of 100 or more feet is desirable in order to obtain good resolution on the oscilloscope. With 1.5 db of cable, the change in return loss is 3 db at 213 MHz, 1.5 db at Channel 2. The actual amplifier VSWR may then be easily computed; however, this is hardly necessary as long as the loss of the VSWR cable is specified. Typical corrections are given in Table 13-3.

Return loss may be read directly as a function of db-variation by using Fig. 20, Appendix VI, which has been computed for cable with 1.5 db loss at 213 MHz. As we have seen in Chapter 8, return loss is the most useful number to indicate match for system design.

With the method described, it is readily possible to read down to a basic accuracy of 1% or a VSWR of 1.01. The VSWR cable should be 1.5, ±0.05 db, to achieve this accuracy.

13-4. MEASUREMENT OF NOISE FIGURE

A typical CATV amplifier aligned for 25 db spacing has a noise bandwidth of 80 MHz. This value can be obtained by graphical methods too involved to be described here. Using this value of 80 MHz and Fig. 3-4, it is possible to obtain the noise figure from the equivalent input noise. The equivalent input noise is obtained from the output noise and the gain of the amplifier. An example will illustrate this method.

Assume an amplifier aligned for 25 db spacing has an output noise level of -32 dbmv at Channel 13 and -48 dbmv at Channel 2, as measured with a meter with a noise bandwidth of 400 kHz. (When measured with a wideband 300-MHz RF voltmeter, the output noise is -12 dbmv.) What are the noise figures?

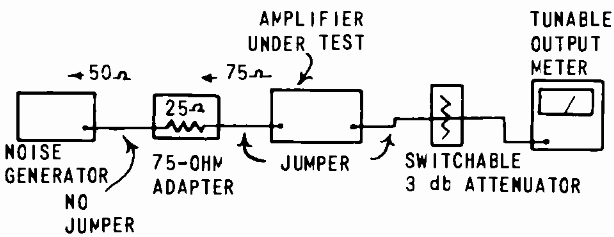


Fig. 13-11. Diagram of noise figure test setup.

With an output noise of -32 dbmv and a gain of 25 db, the input noise is -57 dbmv. Correcting from a 400-kHz to 4-MHz bandwidth, add 10 db, and obtain an equivalent input noise of -47 dbmv. With Fig. 3-4, at a 4-MHz bandwidth, the noise figure at Channel 13 is 12 db. At Channel 2, with 12 db gain, the input noise is -60 dbmv, or -50 dbmv for a 4-MHz bandwidth corresponding to a noise figure of 9 db. The overall noise figure is obtained by using the noise bandwidth of 80 MHz. With a noise output of -12 dbmv, the input is -35 dbmv, using 23 db gain at the median frequency (determined graphically). The overall noise figure is then 11 db.

This method of noise figure determination is reasonably accurate, but quite cumbersome due to the time-consuming task of determining the exact noise bandwidth. Moreover, many meters are of the peak-reading variety and totally unsuitable for noise measurements, leading to errors up to 20 db. Therefore, this method is satisfactory only for the knowledgeable engineer and, in any case, very time consuming. It is used when only a limited amount of equipment is available.

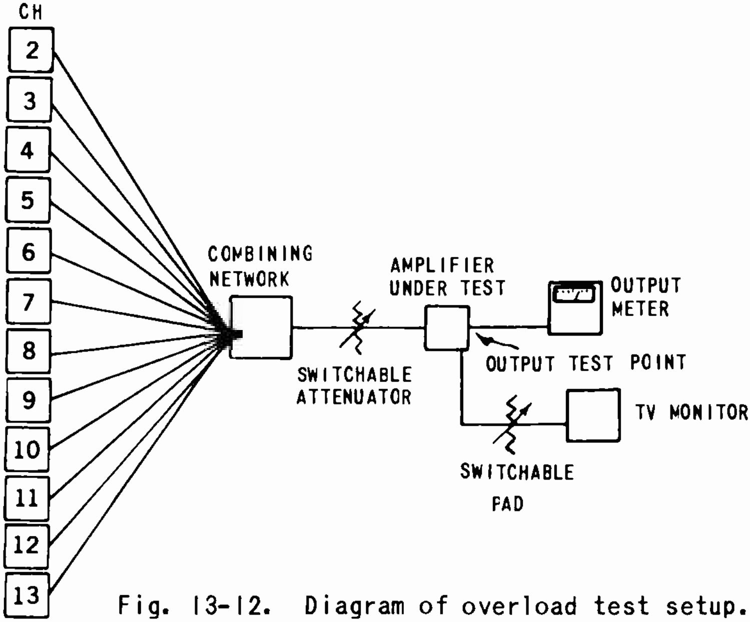


Fig. 13-12. Diagram of overload test setup.

The much preferred direct method uses a noise generator. The principle is to apply a measured amount of wideband noise to the input of the amplifier so that the total amount of noise is doubled. Obviously then, the noise generator produces exactly the same amount of noise as the amplifier itself. This amount of noise is then easily read from the calibrated meter of the noise generator.

Since noise is compared with noise, no determination of noise bandwidth is required. Also, nearly every meter may be used as an output indicator, even peak-reading types, particularly with the refined 3 db method (Fig. 13-11). This 3 db pad is inserted after the amplifier under test in order to obtain the same meter reading with twice the noise power. Consequently, the meter always reads in the same spot and any nonlinearities are excluded from the measurements.

The procedure is then as follows: With the noise generator in "standby" position, obtain a reference noise reading in the desired channel. If manual gain is provided on the output meter, set the pointer to a convenient spot on the dial. Next, with the 3 db pad in, switch the noise generator "on" and adjust noise level to obtain the same reading as before. This setting may be rechecked by flipping rapidly back and forth between the "standby" noise generator and 0 db pad, and the "on" noise generator and 3 db pad. When there is no detectable difference in the output reading, the noise figure is read directly off the noise generator. If a 25-ohm series resistor was used to adapt a 50-ohm noise generator to a 75-ohm system, 1.76 db must be subtracted from the reading to obtain the actual noise figure of the amplifier. The accuracy of the direct noise figure test is generally ±0.5 db.

13-5. TESTING DISTORTION AND OVERLOAD

Distortion in CATV amplifiers is noticed mainly as "windshield wiping," caused by cross-modulation of the desired channel by the sync pulses of an interfering station. The common test method (Fig. 13-12) contains numerous sources of error. Through special frequency converters and amplifiers, TV signals are generated on all 12-channels by using the local signals available directly, and also reconverting them to different TV channels. This combined signal is applied to the amplifier under test, and the output is monitored on a TV set for the onset of visible distortion. The output level at this point is then read and represents the overload level of the amplifier.

While this overload test is very easy to perform, it has severe shortcomings due to the large inherent sources of error which result in a total accuracy not better than ±5 db.

Starting with the head end, the use of live TV signals causes several problems. The background brightness changes from scene to scene, thereby changing the distortion level and also the subjective visibility. Also, there are large variations from station to station and camera to camera, and the signal is often poorly controlled at the station. This applies also to the sync pulse itself which, on some stations, might even disappear at times into the video signal. Obviously, such poorly controlled TV signals are not suitable for measurements.

In order to increase the accuracy, the desired channel may be left unmodulated (blank). This removes the error due to poorly-controlled background level of the desired station. However, it is possible for the TV set to synchronize on the cross-modulated pulses, thereby changing the visibility of distortion. Also, such a condition does not resemble the actual overload mechanism in a system. This can be taken care of by adding a small fudge factor of, say, 4 db to the overload reading to account for the dif- between blank and modulated picture.

There are other errors in the signal source. For example, if Channel 13 is received directly and Channel 12 is obtained by converting Channel 13 to 12, the sync pulses of both channels will be timed identically. Therefore, no windshield wiping between these two channels is possible; the test procedure then completely ignores the normally severe effect of adjacent channel cross-modulation. Since different channels are in use in different cities, this factor alone makes it nearly impossible to directly compare readings taken with different equipment. To avoid the signal source errors, the only possible solution is to eliminate off-the-air TV signals for testing purposes.

But even assuming perfect signal sources, many other large errors are inherent in the test setup of Fig. 13-12. Obviously, any error in the output meter directly affects the overall accuracy. Moreover, the TV set itself causes distortion, which changes drastically with the adjustment of fine tuning and contrast, and of course, with input level. The attenuator ahead of the TV set is used to keep the input level to the TV set constant at all times, generally at some value between 0 and 20 dbmv. As far as contrast and fine tuning is concerned no standardization is possible, and some average setting must be used. Together with the subjective judgment of the operator, these errors are large and cannot be eliminated.

Another often overlooked error is due to system equalization. For simplicity, the inputs to the amplifier under test often are made flat; that is, the level of each TV

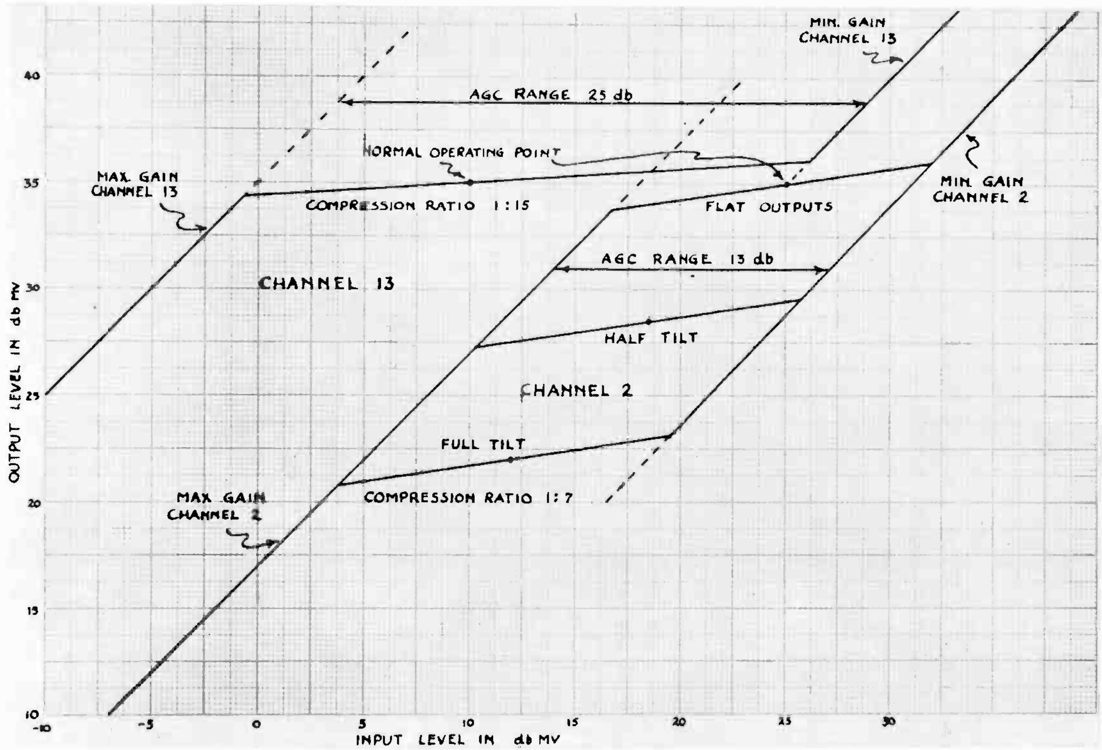


Fig. 13-13. Idealized AGC characteristics for 25 db spacing.

channel is identical at the input of the amplifier. This form of testing corresponds to full tilt discussed in Chapter 6. Since main-trunk amplifiers are often operated at half tilt, the test does not represent actual performance in a system. The same is true for other amplifier types where the test setup differs from actual operation in the field.

In summary, it may be said that errors in this type of overload testing are such as to make this test nearly meaningless, except maybe for comparative purposes. The reason that such a test method might be used at all is that no better distortion instrumentation has been available, since standard cross-modulation tests do not correlate to windshield wiping effect. Accuracy of all other important tests can be made high (±0.1 db for alignment, gain, and VSWR; ±0.5 db for noise), so that only overload and distortion needs a considerably improved technique. Different test methods with far superior accuracy have been developed (see Appendix VII, Reference 25), and distortion analyzers based on these principles are likely to become standard in the near future.

13-6. TESTS OF AGC PERFORMANCE

In AGC amplifiers, in addition to the previously described tests, AGC action must be checked, and in particular, delay points, drop-out level, compression ratio,* range of output level, and AGC tilt. These prerequisites of every AGC circuit can be explained from Fig. 13-13. Consider first the idealized curve for Channel 13. As the input signal to the AGC amplifier (bottom scale) in increased, the output (vertical scale) rises proportionally until the onset of AGC, where the output level remains nearly constant.

The point of AGC start is called delay point because AGC is ineffective (delayed) for signals below this point (at -0.6 dbmv input in Fig. 13-13). The useful AGC range reaches from the delay point to the drop-out point (at an input signal level of 26 dbmv), where AGC action ceases and the amplifier again behaves as a fixed-gain amplifier, although with reduced gain. The useful AGC range thus is limited by two sloping straight lines, one representing maximum gain, the other minimum gain of the amplifier. In the AGC range, the output remains not constant, but rises slightly with input signal.

*Appendix VII, Reference 23.

The ratio of output signal change divided by input signal change is called compression ratio. For example, in Fig. 13-13, the output changes by 1 db for an input change of 15 db; therefore, the compression ratio is 1:15. When measuring compression ratio, it is important that the input signal always remains within the AGC range. Normally, the knee at delay and drop-out points is not as sharp as shown in the idealized curves. In case of doubt, the entire AGC characteristic must be plotted. A well designed AGC system operates normally in the center of the AGC range—that is, at an input level of 10 dbmv in Fig. 13-13—in order to have enough leeway in either direction for correction purposes.

Due to the different efficiencies of some AGC detectors for CW or modulated signals, a different set of AGC curves may result. The test for AGC action should therefore be performed with a modulated TV signal, unless the amplifier is designed to operate with a pilot carrier.

The next important characteristic is the range of output level adjustment. This control (or internal adjustment, as the case may be) determines the system operating level, and some range of adjustment is desirable to permit the use of an optimum level for the particular number of amplifiers operating in cascade. Such a control, if well designed, produces merely a vertical displacement of the curves, with other characteristics unchanged, such as compression ratio, delay, and drop-out point. Then only the output level is changed with no change in AGC action.

A range of adjustment of ±5 db is normally sufficient. However, in integrated systems with high-quality equipment, the output level for the AGC amplifier is factory-set for the optimum system level (see Chapter 6), since no improvement results in system dynamic range by using a higher output level for a shorter cascade. The elimination of this control results in greater reliability by avoiding the possibility of a faulty field adjustment.

The tilt control in an AGC amplifier has a complicated function which is best explained using Fig. 13-13. So far, we have only considered the curve at Channel 13. For a fully-tilted system (flat inputs), the curve for Channel 2 is directly below the Channel 13 curve, displaced by the difference in gain of 13 db (for 25 db spacing). Also shown are the curves for half tilt and flat outputs. The in-between channels have curves located proportionately between the curves of Channels 2 and 13, according to their frequency. The tilt control then has the function of changing the characteristic curves for all channels, except Channel 13, to correspond to a fully, half, or untilted system. It does not matter then if one channel or all 12 are used; the output will always be correct at each channel and is held constant by AGC action.

While this principle of composite AGC works well in theory, it has serious practical disadvantages. First, a composite AGC system, one where all channels together produce AGC action, relies entirely upon the strongest signal. It cannot correct for variations between channels, inevitable when jumper cables are used (and only strip AGC can correct for these variations). Consequently, the strongest signal determines the operation.

Second, the alignment of AGC equalization is extremely critical and can be performed only by a qualified AGC specialist. While special test equipment could be developed to simplify this task and increase the accuracy, this has not been done, mainly because another better AGC principle has already been found.

Third, while such a system works well even if only the low-band channels are used, it is not possible to make a satisfactory adjustment of system tilt in the head end without simultaneously affecting the whole system level because system tilt is then determined by the AGC amplifiers. This is easy to show in an example. Assume that an AGC amplifier is set for full tilt. Now, if for some reason, it is desirable to change the system to half tilt, Channel 2 is then increased by 6.5 db in the head end, and the other channels proportionately, with Channel 13 remaining the same (Fig. 13-13). However, the output of the AGC amplifier at Channel 2 is now held constant by AGC action. An increase in input by 6.5 db results in a decrease in amplifier gain by about 6 db at Channel 2; this corresponds to a decrease in gain of about 12 db at Channel 13, if the gain control is correctly tilt-compensated. This means that the output at Channel 2 was increased by 0.5 db, and at Channel 13, reduced by 12 db (output level 23 dbmv at both Channels 2 and 13).

Instead of achieving a desired tilt correction, the system level itself has been changed with serious consequences on system performance. In order to change system tilt in such a system, each AGC amplifier must be reset separately by its AGC tilt control, a procedure which cannot be performed in the field with the required accuracy.

These shortcomings of composite AGC were offset by field experience; a modification, although not good, led to a system operating in some fashion. This modification consisted of tuning the gain control section of the AGC amplifier to function at a few channels only, usually near the top of the high band. Such a compromise system has also a large number of disadvantages. While the tilt in the headend equipment may now be set correctly for the channels which are not gain controlled, the same fault exists for the channels which have AGC action; this behavior causes a bend in the system frequency response near the frequencies where AGC action becomes effective.

Secondly, as already mentioned, there is no AGC action at all for many channels. Therefore, the AGC amplifiers have to be aligned differently for each system. For example, in a low-band system, the AGC has to be tuned to provide gain control near the top of the low band. If high-band channels are added later, all AGC units must be realigned. Or, a system carrying Channel 12 will have different AGC action from one with Channel 13. Experience with AGC amplifiers of this type shows the typical poor performance and unreliability of this concept. It simply is not possible to manufacture high-quality AGC amplifiers where a faulty concept requires extensive factory and field modifications for each particular application. It is clear from this discussion that composite AGC amplifiers are not suitable for high-quality AGC systems.

As a next logical step, one could develop equipment where AGC action depends on one channel only, usually the highest TV channel carried. While such a concept is a vast improvement, it still has a number of shortcomings. It still would be necessary to have a different AGC amplifier, and a different alignment for each system. Also, in case of later addition of new channels to a system, a major overhaul of all AGC equipment is necessary. Finally, a TV signal itself is not the most suitable signal for AGC due to the varying modulation which, in turn, affects AGC action.

For precise AGC action, therefore, the use of a pilot carrier is essential. The advantages of the pilot carrier system are obvious: identical AGC action regardless of system; precision factory alignment is possible; system tilt is readily adjusted in head-end equipment; independence from modulation changes.

The test for AGC action with a pilot-carrier AGC amplifier is similar to the test already discussed with Channel 13 as shown in Fig. 13-13. The only difference is that a generator, tuned to the pilot-carrier frequency, is applied to the input of the AGC amplifier. Thereafter, the AGC characteristic may be measured with delay and drop-out levels, and compression ratio may be determined.

13-7. TEMPERATURE CORRECTION AND AUTOMATIC SPACING

These tests are closely related. As a matter of fact, although we talk of AGC or level control, really there is no need for such a concept since the signals are sent down the system with a constant level from the head end. The only function of AGC circuits is then actually to correct for changes in cable attenuation with temperature or errors in spacing.

To test for temperature compensation, the inexperienced technician might construct a large temperature test chamber at great expense, install large spools of cable and then proceed to heat or cool this chamber to perform tests on temperature - correcting AGC amplifiers. Such a test would not only be expensive and time consuming, but is loaded with large inherent sources of error, particularly due to the tremendous thermal hysteresis involved in such testing.

Fortunately, much more elegant and precise methods exist for testing temperature compensation. It has long been established that temperature has exactly the same effect on coaxial cable as if its electrical length were changed. Very accurate tests of this effect have been made, and any deviation from this law is certainly less than 1% or 0.1 db if any deviation exists at all. (See also Fig. 11-3). This accuracy is amply sufficient and quite commensurate to other accuracies in tests discussed so far. Consequently, temperature compensation of the cable is tested simply to changing the length of test cable. If temperature compensation is effective, the frequency response and output level must be unaffected by changes in cable.

The maximum probable change in the electrical length of a cable depends on the spacing between AGC amplifiers in a system. For example, with 25 db spacing and an AGC amplifier in every third position, the spacing between AGC amplifiers is 75 db. For a temperature range from -50 to 160° F, the change is then -10.5 to +7.5 db, respectively. If such an AGC amplifier is normally aligned at 25 db, its output must remain unchanged if the test cable is changed from 14.5 to 32.5 db. Again, for high quality equipment designed for use without jumper cables, the test cable is split into two sections.

The test procedure is then as follows: Calibrate the test setup (Fig. 13-1) as described before. Then proceed with the alignment as shown in Fig. 13-7. Next, exchange the input cable (15 db) for one having 4.5 db and 22.5 db loss at Channel 13, respectively. The change in

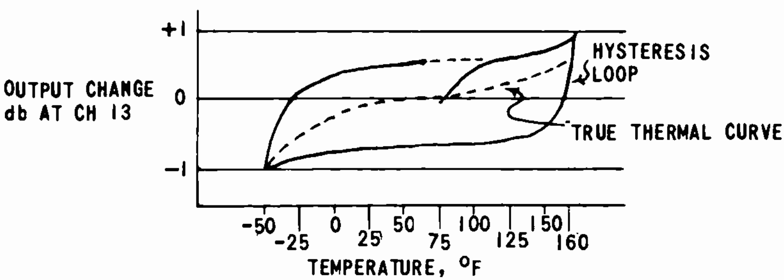


Fig. 13-14. Thermal hysteresis curve.

output and flatness is then read directly from the oscilloscope. With some AGC amplifiers, this test cannot be performed due to the time constant of the particular AGC circuit, which is unable to perform at a sweep rate of 60 Hz. It is then possible to test temperature compensation manually, by changing the AGC bias directly with a potentiometer. For this case, manufacturer's test instructions should be consulted. The same is true for systems using pilot carriers.

The test for temperature compensation, by its nature, is identical to a test for automatic spacing. Normal spacing errors fall into the range of ±3 db; therefore, the much larger range necessary for temperature compensation encompases this range and no separate test is needed. However, since both effects may be additive, the range of temperature compensation should be increased proportionately.

Temperature testing must also include tests on the amplifiers themselves. For this purpose, a small temperature chamber to house an individual amplifier must be used. The test setup in Fig. 13-7 is used to first complete the alignment of the amplifier. Thereafter, the amplifier is placed in a test chamber using jumper cables (about 3 feet each) of heat-proof coaxial cable, such as RG-140/U.

The effect of the jumper cables is taken into account by drawing the new response on the oscilloscope. For this purpose, the standard calibration procedure is followed and the pattern of Fig. 13-5 is obtained, except that there now will be three parallel lines showing peaks and dips spaced 1 db apart. The peaks and dips due to the jumper cables are of no consequence in this test, since only the change of response with temperature is of interest. Note that the amplifier alignment is not touched. Thereafter, the temperature is raised slowly to 160° F, cycled down to -50° F and brought back up to ambient. The output level is read from the oscilloscope at every 10° F on both paths going up and down.

The complete thermal hysteresis loop may now be plotted (Fig. 13-14). The true thermal curve is the average of the hysteresis curve. This curve must not change by more than 0.5 db for high quality equipment, unless the concept calls for an open-loop compensation system as might be used in distribution applications. Then the thermal curve must be the opposite of the cable within 0.5 db. The thermal curve is normally measured at Channels 13 and 2 to give a full picture of changes in slope. These tests are not required routinely, but rather performed by the manufacturer to assure a high quality product.

The test procedures discussed here provide high standards of accuracy, and are designed to be simple in use so that excellent results can be achieved by system technicians.

CHAPTER 14

Increasing Channel Capacity

Although additional program material is generally not available, there has been considerable thought in recent years about methods of adding extra TV Channels to a CATV system for use at some future date or for other services. Twenty, twenty-four, twenty-seven, and even more channels have been promoted by many CATV equipment manufacturers, and as we shall see there is really no upper limit to the possible number of channels in sight; however, with most methods one has to accept degradation in picture quality, along with a large increase in system cost.

There are three basic methods which have been proposed or actually used in system designs. 1. The addition of extra channels to the existing 12 channels in the VHF band, with several variations; that is, the use of the frequency range from 20 t0 54, from 88 (120) to 174, and 216 to 300 MHz. We shall code this method "Extra VHF Channels." 2. Use of UHF Channels throughout. 3. Multiple cable systems. There are additional methods of merit, but since they are not widely discussed at the moment we shall concentrate at this time on the pro's and con's for the basic techniques mentioned above.

14-1. FREQUENCY ALLOCATIONS

When extra channels are added to a 12-channel CATV system, the wisdom of the standard TV channel allocations becomes immediately apparent. With the present frequencies for the 12 VHF channels, interference due to intermodulation products caused by second-order curvature is minimized. Just adding one additional channel causes immediate problems and equipment which previously had acceptable third-order curvature (cross modulation) now, in addition, must also minimize the normally much stronger second-order nonlinearities. Two frequencies, f1 and f2, generate among others the following intermodulation products due to a second-order nonlinearity: f1 plus f2, f1 minus f2, 2f1, 2f2, 2f1 plus f2, 2f1 minus f2, 2f2 plus f1, 2f2 minus f1, 2f1 plus 2f2, 2f1 minus 2f2, with f1 assumed larger than f2 (f1 is greater than f2). Consulting Appendix VI we see that the second harmonic of Channel 2 (2 x 54 MHz) is at the top of the FM band. Therefore, Channel 2 and the other low-band channels have no chance of interferring with each other. Nor can the sum or difference of any low- or high-band channel cause interference (216 minus 54 is 162 MHz; below Channel 7, etc.). An insignificant overlap exists between the second harmonic of Channels 6 and 7 (2 x 88 and 174 MHz). The reader can verify the other products. As it stands, only the normally much weaker third-order nonlinearities might cause interference (and they are, of course, also responsible for cross modulation; for example, 3 x 60 MHz—Channel 3—is 180 MHz—Channel 8—etc.

How can we get extra channels without having to worry about second-order nonlinearities? As can be seen from examining the make-up of the frequency allocations (Channels 2 through 13), frequencies below Channel 2 cannot be used because difference products with Channel 6 would cause interference with other channels. The old Channel 1 is not available because of the IF frequencies used in most TV sets. The range immediately above Channel 13 also cannot be used (Channel 7 together with the TV IF frequency would cause interference). But 450 to 900 MHz is usable (the standard UHF band with 75 available channels). The UHF approach will be examined in 14.4 below.

We can now look into the possibility of dropping some of the old channels and see if we can do better with a totally different arrangement. If we add a subchannel, say, from 27 to 33, we are wiping out standard Channels 2 and 3 by second-order interference, clearly a bad trade—one gained, two lost. On the other end, above Channel 13, we find that we can add extra channels advantageously and still maintain our integrity as far as second-order nonlinearities are concerned. For example, we can use the octave from 120 to 240 MHz for 20 channels, or from 150 to 300 MHz for 25 channels if the upper cut-off frequency of our amplifiers permits. A lower band, in addition, cannot be added because the difference product would fall into the range from 150 to 300 MHz.

Next, we can sacrifice our integrity by avoiding interference due to second-order nonlinearities and go ahead and add additional channels any old place, and either live with poorer picture quality or sharply derate the system (considerably shorter cascade). As an alternate we could ask for a better amplifier. However, even with such an improved amplifier the quality will always be less than with the same amplifier and standard allocations or the simple octave approach. Therefore, it is necessary now to look at the effects of intermodulation in a little more detail.

14-2. INTERMODULATION IN CATV SYSTEMS

Unlike cross modulation, intermodulation is visible without modulation as running diagonal bars (beats). The visibility of the effect depends very much on the difference product. A difference between two carries of 1.0 to 2.5 MHz generally produces minimum interference and normally is not noticeable. A standard mixing chart will clearly reveal the multitude of severe potential interference products when channels are added in addition to the standard 12 channels. Fortunately, beat products add in RMS fashion unlike cross modulation.

For example, assume a cascade of 100 amplifiers. First, overall cross modulation is to be 46 db down. This calls for a cross-modulation level of -86 db for an individual amplifier. Also, intermodulation is to be present in addition. For a total distortion level of -46 db, individual cross-modulation signals must now be -89 db and individual intermodulation components at 69 db. In addition, a further, substantial derating must be made for the numeric increase in number of channels above 12. This derating is dependent on the circuit design of the amplifier, system tilt, and other factors. With this concept, system quality is clearly compromised, but it would be worthwhile to examine what can be done to decrease second-order distortion in an amplifier to the degree needed to get by with this compromise approach.

Second-order distortion, as every circuit designer knows, can be minimized by pushpull design familiar from audio and hi-fi. An experienced CATV wideband amplifier designer can immediately estimate the magnitude of the possible improvement. If an excellent circuit design job were done with multiple balance controls, one could expect to reduce second-order products by about 10 db. This is, of course, quite a bit less than would be expected in an audio amplifier design. The cost of such a CATV pushpull amplifier would be more than doubled, certainly an unsatisfactory solution for such a small improvement.

Another solution would be to limit amplification to octave bands; that is, some form of split band design. For example, 54 to 108 and 120 to 240 MHz for a total of 26 channels. The improvement that can be realized with this approach is substantial. It is a simple matter to build filters in excess of 50 db rejection. Cost would be somewhat less than the pushpull approach, because both amplifier sections (having to handle only one octave band each) could be of a much simpler straightforward design. In addition, cross modulation would be substantially reduced. On the negative side, a small frequency band, say from 108 to 120, would be lost to filter crossover characteristics. Considering everything, it is evident that the split-band amplifier approach is far and away superior to the pushpull concept discussed above.

Let us consider what can be done to a standard amplifier to reduce distortion. Feedback theory tells us that for the right combination of positive and negative feedback, overall distortion reduction is possible. In CATV, because of phase shift, stability, and gain problems, we can expect a distortion reduction of about 7 to 8 db for a gain loss of about 2 db. This technique includes all methods of broadband unilateralization. Another method, which will not be described here, allows a 35 to 40 db reduction with a gain loss on the order of 2 db. Both methods are low cost and should, obviously, be used first, since they increase amplifier cost insignificantly. If distortion reduction is still insufficient, the split-band approach should be used at, roughly, twice the amplifier cost, possibly in combination with the other two methods. There can be absolutely no excuse for a designer to ever seriously consider a pushpull amplifier for CATV. The various approaches are summarized in Table 14-1.

When we now consider again our various approaches to add extra channels to the existing 12, we see that this is now readily possible with the amplifier designs proposed and quite likely at no degradation in quality; in fact, quite to the contrary if some form of split-band amplifier is used. The cost for the electronic equipment in the system is roughly doubled.

Table 14-1. Comparison of 2nd Order Distortion Reduction Methods

METHOD	TYPICAL DISTORTION REDUCTION	GAIN LOSS	COST	COMMENTS
1. Undisclosed	35 to 40 d	0 to -2 d	1 to 2$	-
2. Feedback	7 to 8 d	-2 dB	1 to 2$	Related to neutralization
3. Push-pull	10 d	(30db)	More than double	Poorest
4. Split-band	50 to 60 d	-	About double	Best, also cuts across modulation

14.3 ECONOMIC AND CONVERSION PROBLEMS

Any time we deviate from the standard TV channels, we must provide some means of tuning in those extra signals. Obviously, extra gear must also be provided at the head end. However, in the head end the cost factor and technical questions are handled in a straight-forward manner and are of no major concern. In contrast, at the TV set the problems are numerous. The obvious solution of installing a set converter ran into serious problems. First, costs of manufacturing such a converter were badly underestimated. With more realistic sale prices in the vicinity of $40 and up, the question is who is going to pay for it? Will there be enough subscribers willing to pay an installation fee in the vicinity of $70? Most likely not. And as if this were not enough, most subscribers resent a "black box" sitting on top of their furniture. They insist on styling, clearly marked channels, etc. Other approaches, such as conversion into the UHF range, also proved unacceptable because of the difficulty in channel identification. So, while we now have solutions to build systems with more than 12 channels, we are unable to come up with a set converter which is acceptable from an economic or human engineering logic viewpoint.

Forget about more than 12 channels? Not quite; there are other approaches (see 14.5 below). However, to come back to our "added VHF channel approach," it seems the problem needs to be tackled from a totally different angle. In the author's opinion the time has come when a special tuner for CATV should be optional equipment provided by the TV manufacturers. The market is big enough by now that some U.S. manufacturer should be interested or else most certainly the Japanese will furnish it. Such a CATV tuner should allow ready interchange with an existing VHF-UHF tuner. It should provide for the standard VHF stations plus extra channels in the range from 120 to 240, for example. Also, positive detents must be provided for all channels on the channel selector mechanism. No UHF part would be provided, and input circuitry would be 75-ohm unbalanced. It seems that such a tuner could be manufactured at about the same cost of existing tuners. They could be made available with the trade-in of the old tuner, and in new TV sets the option should be made directly available with the purchase of the TV set. Until such time, it seems that the added VHF channel approach must fail for economic and human reasons at the subscriber's end.

14-4. UHF CHANNELS

In this approach, all channels are converted to UHF. UHF is either transmitted over the whole system, or in another approach only in the feeder system with extra conversion equipment as part of the bridger amplifier. Due to the much higher cable loss, more amplifiers will be needed; however, circuit design need not be unduly complex, and individual amplifier cost might be comparable to present types. The subscriber presumably already has a tuner with UHF on it. From a human engineering viewpoint a multitude of channels in the UHF part would require positive channel selection and indication with a selector switch, and obviously, a well working AFC (automatic frequency control) must be provided in such a set. Unfortunately, at the moment there are only a few European sets that seem to come close to it. And again, we would have to ask for a special option from a TV manufacturer. This UHF system, even with skipping every second channel, could provide some 35 TV channels.

14-5. MULTIPLE CABLE SYSTEMS

Let us now look at the—at first sight—most ridiculous idea of building two CATV systems. Amplifier cost is obviously doubled; however, the overall system cost increase is closer to 50%, because of proportionately reduced construction cost. This would not be too bad and is, in fact, close to a split-band amplifier system cost. What do we gain? Well, first off, standard VHF stations are used—no conversion of any kind is needed; present TV sets are OK. We have now 24-channel capability. There will be two drop cables to each TV set. Crosstalk between the two cables should be 40 db down (see Fig. 8-3), with a minimum of 30 db. A switch or plug must be provided so that the subscriber can select the right cable. From a human engineering standpoint this will be far more acceptable than a converter. In addition, we have compatibility and redundancy. If a subscriber were to choose only the 12-channel service, he could do so. Only one house drop would be installed and his subscription fee could be proportionately reduced. With the other system he would be out of luck, 27 channels or else!

Redundancy does not exist in the true sense; however, should one cable system fail or necessitate extensive maintenance, adjustments, etc., there is still the other cable to provide service. It should be a lot more acceptable to subscribers than if all service were turned off. Now there remains the distinct possibility that sometime in the future two-way transmission will be used. The second cable is then already in. Last, with two cables we are still free to use any of the previously proposed methods, in addition; but now instead of 27 channels, we are talking about 54 channels, etc. In summary, there are at the present time a number of weighty points in favor of the multiple-cable approach.

14-6. ADDITION OF OTHER SERVICES

Basically, a CATV system constitutes a high-class communication link to the individual house, and although we are providing a great number of TV signals there is still considerable room left for just about any conceivable type of transmission, such as thousands of telephone channels, two-way data transmission such as for utility metering, banking, purchasing, and the like. All digital transmission systems are happy with a much reduced signal-to-noise ratio and these signals could be run at a lower level so as not to cause interference in the TV channels. Also, because the bandwidths of these services is considerably less, noise level is also down. For example, from 4 MHz (television) to 3 kHz (telephone) the noise reduction is 31.5 db. Consequently, the signal level can be dropped by the same amount without changing the signal-to-noise ratio. It seems, therefore, that the signal level for each service should be proportional to the square root of the bandwidth of that service, with the signal channel requiring the widest bandwidth used as reference. Some deviation from this rule must be made if the signal-to-noise ratio for each service is to be separately set. No major obstacles exist to any of these schemes, and as soon as the basic CATV amplifier has reached a reasonable state of perfection through further research and development, it will be possible to make necessary modifications for further expanded service.

At the time of this writing (1969), normal 12-channel CATV amplifiers on sale are far from perfect and even though they may be new products, they are often already obsolete in many ways. In the author's opinion, the solid-state knowledge available today has only to a small degree been applied in today's CATV products. It is deplorable that inferior circuits are being multiplied endlessly now in pushpull, split-band and other amplifier circuits, when there is still so much work to be done on the basic amplifier. However, it is possibly only a matter of a few more years until circuits of acceptable perfection, within 90% of the theoretical limit, will become available. Time will tell.

APPENDIX I

Calculation of Cumulative Noise and Overload

For identical amplifiers, an equal amount of noise power N is added in each amplifier location so that at the last, or nth amplifier, the signal-to-noise ratio is:

$$(S/N)_n \text{ in db} = 10 \log(S/nN) = (S/N)_i - 10 \log n \quad (1)$$

where $(S/N)_n$ = signal-to-noise power ratio after nth amplifier

n = number of amplifiers

$(S/N)_i$ = signal-to-noise power ratio of individual amplifier in db

It can be seen that the signal-to-noise ratio of the cascaded system in db is degraded by ten times the logarithm of the number of amplifiers (n) below the signal-to-noise ratio of the individual amplifier in db. For example, after 10 amplifiers, the signal-to-noise ratio is degraded by 10 db, after 100 amplifiers, by 20 db.

This calculation is correct only for identical amplifiers with the same noise level, and as expressed by the same noise figure. For more than 10 amplifiers of different noise figures, a reasonable approximation can be obtained by taking the arithmetical average of the individual signal-to-noise ratios in db, and derating by the value given in (1).

For fewer than 10 unequal amplifiers, this method becomes inaccurate. The cascaded signal-to-noise ratio is then found by:

$$S/(N_1 + N_2 + N_3 + \ldots N_n) = S/(1 + k_2 + k_3 + \ldots + k_n)\, N_1 \quad (2)$$

or in db $(S/N)_n$ in db $= (S/N)_1$ in db $- 10 \log (1+k_2+\ldots+k_n)$ (3)

where, $(S/N)_1$ = Signal-to-noise ratio of first amplifier in db

$k_n = \dfrac{N_n}{N_1}$ = Noise ratio of nth to first amplifier

Since this formula is rather cumbersome to use, it is often faster to use Fig. I-1, which is valid for two amplifiers, and to repeat this process several times. Suppose, for example, a cascade of 20 identical amplifiers, each having a signal-to-noise ratio of 60 db, is preceded by a piece of head-end equipment having a signal-to-noise ratio of 50 db. What is the total signal-to-noise ratio?

From (1), the derating for 20 amplifiers is 13 db; therefore, the signal-to-noise ratio for the cascade is 47 db. This cascade is preceded by the piece of head-end equipment with a signal-to-noise ratio of 50 db. From Fig. I-1, with a 3 db difference, the derating is 1.7 db below the poorer one of both, which is the amplifier cascade with 47 db. Therefore, the combined signal-to-noise ratio is 45.3 db.

Analogous to the increase in noise level with the number of amplifiers connected in cascade, it is also found that the overload level, the maximum output signal level for a given amount of distortion, is similarly reduced for the cascade as compared to an individual amplifier. It has been shown conclusively* that, for CATV systems, the distortion products are not correlated as might be expected, but rather behave in a completely random fashion. Therefore, the calculation for overload degradation proceeds exactly as for random noise, and the previous formulas and Fig. I-1 apply to overload as well as to noise.

For amplifiers with different characteristics in a cascade of less than 10 amplifiers, again a separate derating for both overload and noise must be made. These values are then added to obtain the total overload-to-noise (not signal-to-noise) degradation. For larger cascades of different amplifiers, the average overload-to-noise ratio is determined, and this figure is derated by 20 times the log of the number of amplifiers.

*See Reference 18, Appendix VII.

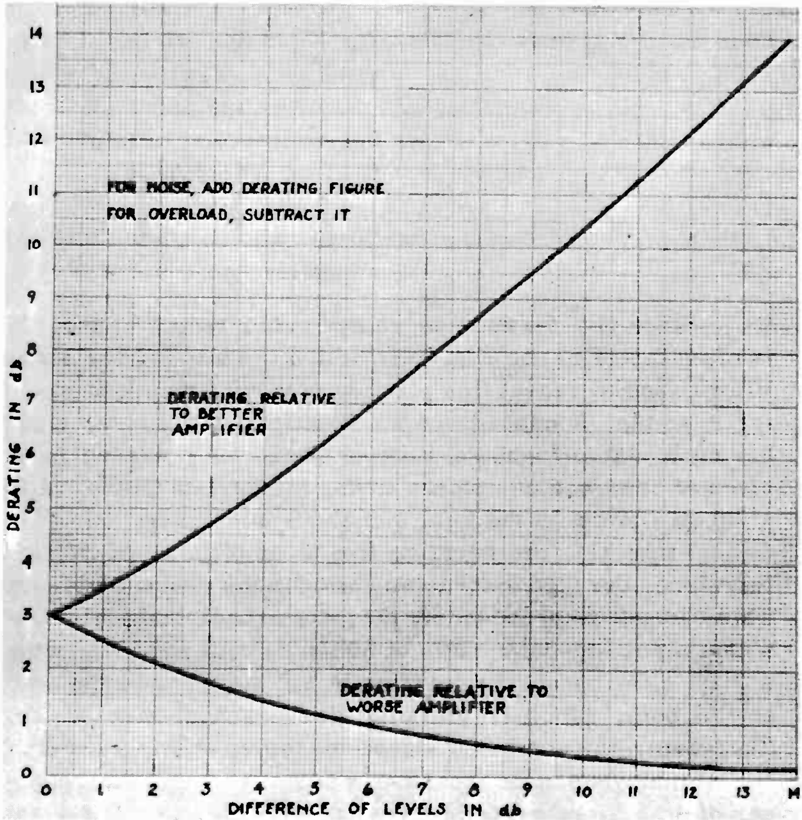


Fig. I-1. Cumulative noise or overload level curves.

The same derating method is used for any cascade of identical amplifiers. A few sample calculations for better familiarization with this method follow:

Example 1: Amplifier A has a noise figure F of 12 db, and an overload level Ø of 42 dbmv. For Amplifier B, F = 15 db and Ø = 41 dbmv. What is the combined overload-to-noise ratio for the cascade for a spacing of 25 db?

The combined noise figure from Fig. I-1 for 3 db difference is 1.7 db higher than the worst noise figure. Therefore, F = 16.7 db. The equivalent input noise level is then -42.3 dbmv. This value is obtained either mentally from the memorized relationship (10 db noise figure is equal to -49 dbmv input noise), or from Fig. 3-4. The output noise with a gain of 25 db is then -17.3 dbmv. The combined overload level with Fig. I-1 is 38.5 dbmv. The overload-to-noise ratio for the cascade is then 55.8 db.

Example 2: 20 amplifiers are operated in cascade. The arithmetic average of the individual dynamic ranges is 71 db. What is the dynamic range for the cascade?

The derating for 20 amplifiers is 13 db each for overload and noise, or a total of 26 db. Therefore, the combined system dynamic range is 71 - 26 = 45 db.

APPENDIX II

Mathematical Derivation of Optimum Spacing

The fact that an optimum spacing exists in repeater applications is easily deduced from simple considerations. Also, it is possible to determine, by trial and error, the optimum value for a single-stage amplifier. For a more thorough understanding, a mathematical analysis is required.

SINGLE-STAGE AMPLIFIERS

The system dynamic range K in db is:

$$K = (\emptyset - 10 \log n) - (N + 10 \log n) - A \qquad (4)$$

where, $\emptyset$ = individual overload output level in dbmv

N = individual input noise level in dbmv

n = number of amplifiers

A = power gain of individual amplifiers

The first term in brackets on the right hand side is the reduction in overload with cascading; the second term is the increase in input noise level. Equation (4) can be rewritten as:

$$A + 20 \log n = \emptyset - N - K = C \qquad (5)$$

where, C = constant

As a second equation, the system gain L may be defined as:

$$L = An \qquad (6)$$

It is now desired to adjust the spacing or gain A for maximum system length L, keeping C constant. From (6), we calculate A and substitute into (5), to obtain:

$$L/n + 20 \log n = C$$

$$L = Cn - 20n \log n \qquad (7)$$

Differentiation of L with respect to n, and equating to zero, leads to:

$$dL/dn = C - 20 \log n - 20n\, d(\log n)\, dn$$

$$= C - 20 \log n - 8.69n\, d(\ln n)/dn$$

$$= C - 20 \log n - 8.69 = 0 \qquad (8)$$

Substituting this result into equation (5), we find:

$$A = 8.69 \text{ db} = 1 \text{ Neper} \qquad (9)$$

The optimum repeater amplifier gain is therefore 8.69 db.

It is truly remarkable that this optimum gain is independent of amplifier and system characteristics. This fact is easily verified from Fig. 4-2. The same result is obtained for amplifiers with high or low overload. Regardless of amplifier type, good or bad, transistor or tube, the best system performance results for the spacing of 8.69 db.

The calculations in equation (5) imply, however, that C is independent of gain A, a condition which is met by single-stage, distributed, and only some multistage amplifiers. In multistage amplifiers, both noise figure and overload level are often a function, of gain, and a separate calculation is made for this case below.

For the simple, single-stage circuit, how critical is the optimum gain for best system performance? In order to calculate other than ideal conditions, the maximum system length, as a fraction of the theoretically optimum system length, can be computed for different amplifier gains and a constant system signal-to-noise ratio. With equations (4) and (5), we have:

$$K = \emptyset - N - A_{opt} - 20 \log L_{opt}/A_{opt}$$

$$= \emptyset - N - A - 20 \log L/A$$

therefore, $$A - A_{opt} = 20 \log (L_{opt}\, A/L\, A_{opt}) \qquad (10)$$

where, $A_{opt} = 8.69$ db

L_{opt} = optimum system length with optimum spacing

A = amplifier gain in db

L = system length in db

A plot of equation (1) was given in Fig. 4-3. This curve gives the degradation in possible system length with other than optimum spacing. From this curve, some rather useful data may be taken. For example, with the common spacing of 25 db, only 44% of the ideal system length is possible. For a single-stage amplifier, this undesirable high gain might occur in some of the older subchannel system designs. A spacing of 15 db allows 83% of the ideal system length, nearly twice as long than with a spacing of 25 db. In all cases, the quality of the output signal is unchanged.

Table 11-1. Degradation of ideal system length.

Spacing in db	% of Possible Length
8.69	100
13	90
15	83
25	44
30	19
40	13

Ninety per cent of the theoretical system length is reached with a spacing of 13 db, and there is little advantage of spacing closer than this figure. The other extreme would be of a gain of 40 db, which permits only one-eighth of the best system length and must be considered poor system design. Typical system length as a function of spacing are given in Table II-1.

Instead of considering degradation of system length for a given dynamic range, we can compute the degradation of system dynamic range, but with a constant system length. It is then easily found that:

$$K_{opt} - K = A - A_{opt} - 20 \log A/A_{opt} \tag{11}$$

Equation (11) was already plotted in Fig. 4-4, and is self-explanatory.

TWO-STAGE AMPLIFIERS

For amplifiers where overload level and noise figure depend on amplifier gain, and this is the case for most multistage amplifiers, a separate analysis must be made. For example, in a 2-stage circuit, as gain of the first stage is decreased, eventually the noise figure rises. Similarly, as the second stage gain is increased, overload in the first stage becomes less significant and the overload level of the whole amplifier is increased.

These considerations are of little importance if the individual stage gain is high. Unfortunately, in wide-band circuits used for CATV purposes, stage gains are usually low, and these factors must be considered. It will be sufficient to calculate the 2 - stage circuit to get a feeling of how the optimum spacing must be changed.

In a 2 - stage circuit, the maximum signal power output is given by:

$$P_{so} = \frac{\emptyset_1 \emptyset_2 A_2}{\emptyset_1 A_2 + \emptyset_2} \tag{12}$$

where, $\emptyset_1$, $\emptyset_2$ = overload levels, first and second stage
A_1, A_2 = power gains, first and second stage

This equation can be derived from:

$$1/e_{\emptyset}^2 = (1/e_1^2) + (1/e_2^2)$$

which has been derived for cascaded overload.* This re-

*See Reference 18, Appendix VII.

lationship leads in turn to $P_\emptyset = P_2/(P_1+P_2)$, where $P_\emptyset$ is the combined maximum output, and P_1 and P_2 are individual outputs for stage 1 and stage 2. With $P_1 = \emptyset_1 A_2$ and $P_2 = \emptyset_2$, equation (12) results. For the noise of the amplifier, we find:

$$P_{no} = N_1 A_1 A_2 + N_2 A_2, \tag{13}$$

where, P_{no} = total noise power output

N_1, N_2 = input noise power, stage 1 and 2

The overload-to-noise ratio K for a cascade of n amplifiers using equations (12) and (13), is then:

$$K = \frac{\emptyset_1 \emptyset_2}{(\emptyset_1 A_2 + \emptyset_2)\ (N_1 A_1 + N_2) n^2} \tag{14}$$

The total system gain is found to be:

$$L = (A_1 A_2)^n \tag{15}$$

Equations (14) and (15) are numeric equations and correspond to logarithmic equations (5) and (6) for the single-stage case. Using these two equations, we can find the optimum amplifier gain or spacing. Mathematically, it is easiest to eliminate n by taking the natural logarithm of (15). This leads to:

$$n = \ln L / \ln A_1 A_2 \tag{16}$$

Substituting (16) into (14), we have:

$$K = \frac{\emptyset_1 \emptyset_2 \ln^2 A_1 A_2}{(N_1 A_1 + N_2)\ (\emptyset_1 A_2 + \emptyset_2) \ln^2 L} \tag{17}$$

It is now necessary to maximize K as a function of both A_1 and A_2; that is, the partial derivatives dK/dA_1 and dK/dA_2 must both be zero. This process leads to the following equations:

$$\ln A_1 A_2 = 2\ (1 + N_2 / N_1 A_1) \tag{18}$$

and $\ln A_1 A_2 = 2\ (1 + \emptyset_2 / \emptyset_1 A_1)$ (19)

We normalize $N_2 / N_1 = N$ and $\emptyset_2/\emptyset_1 = \emptyset$, and after a short calculation, we obtain the final result:

$$\ln A_2 = 1 + \emptyset / A_2 - 0.5 \ln (N / \emptyset) \tag{20}$$

$$A_1 = NA_2 / \emptyset \tag{21}$$

Since these expressions are considerably more complicated than for the single-stage case, it is best to examine the result on a few examples:

Example 1: First, it would be best to set the limits to equal the single-stage case and verify the previous result. We set $N_2=0$, $\emptyset$=infinity. With this substitution, we have from (20), $\ln A_2=1=\ln A_1$. $A=A1\ A_2=e^2=8.69$ db as before.

Example 2: Calculation of a 2-stage circuit with identical performance data for each stage; therefore $\emptyset_1 = \emptyset_2$; $N_1= N_2$; $A_1 = A_2$. From (20), $\ln A_2 = 1 + 1/A_2$. This is a transcendental equation which has a solution for $A_2 = 3.59$. $A_1 A_2 = A_2^2 = 12.9$ or 11.0 db gain for the whole amplifier (5.5 db gain per stage).

Example 3: We assume a 6 db increased noise and overload level for the second stage, which is a reasonable assumption if a higher power tube or transistor is used for the output stage. Therefore $N = \emptyset = 4$; hence $\ln A_2 = 1 + 4/A_2$. The solution is $A_1 = A_2 = 5.57$, or 7.5 db gain per stage, 15 db for the whole amplifier.

The results of these examples are tabulated in Table 4-3, and in Table II-2.

Various other cases can be calculated, but the more important practical amplifiers are likely to fall somewhere in between, and the examples given will permit a reasonable estimate. It is also well to keep in mind that the peak of the curve in Fig. 4-3 is reasonably shallow; that is, the exact value itself is not too critical. Also, in CATV work, due to the required cable equalization, the gain would be different at different channels, so that a reasonable compromise must be made. Generally, close to optimum conditions are met near Channel 2 in practical amplifiers available today. For every particular amplifier, optimum spacing may be determined directly from the measured amplifier characteristics as discussed in Chapter 4. The values given in Table II-2 should be considered only rough guidelines, since it is virtually impossible to measure noise figure, gain, and overload, for each individual stage in a multistage amplifier, to the accuracy required for meaningful results using formulas (20) and (21). Therefore, examples were picked which likely include the range of practical circuits.

Table II-2. Optimum gain (spacing) for various amplifiers.

Amplifier Type	Optimum Gain	Gain for 90% Maximum System Length
One stage	8.69 db	13 db
Two identical stages	11.0 db	15 db*
2nd stage 6 db higher overload and noise	15.0 db	19 db*
Three identical stages	13.0 db*	16 db*

*Estimate

From a practical viewpoint, the theoretical analysis always sets the objective goal for the circuit and system designer. The design of a new CATV amplifier is then completed according to the latest state of the art. For this particular amplifier, optimum spacing is then determined from measurements of overload and noise level. A check is then possible to see how closely the design approaches the theoretical possible. This process may seem circuitous; however, there is no short cut. In any event, the optimum spacing for each amplifier is readily determined from the cascaded amplifier figure of merit. Optimum spacing, as well as amplifier dynamic range, should be part of the specification of every CATV amplifier.

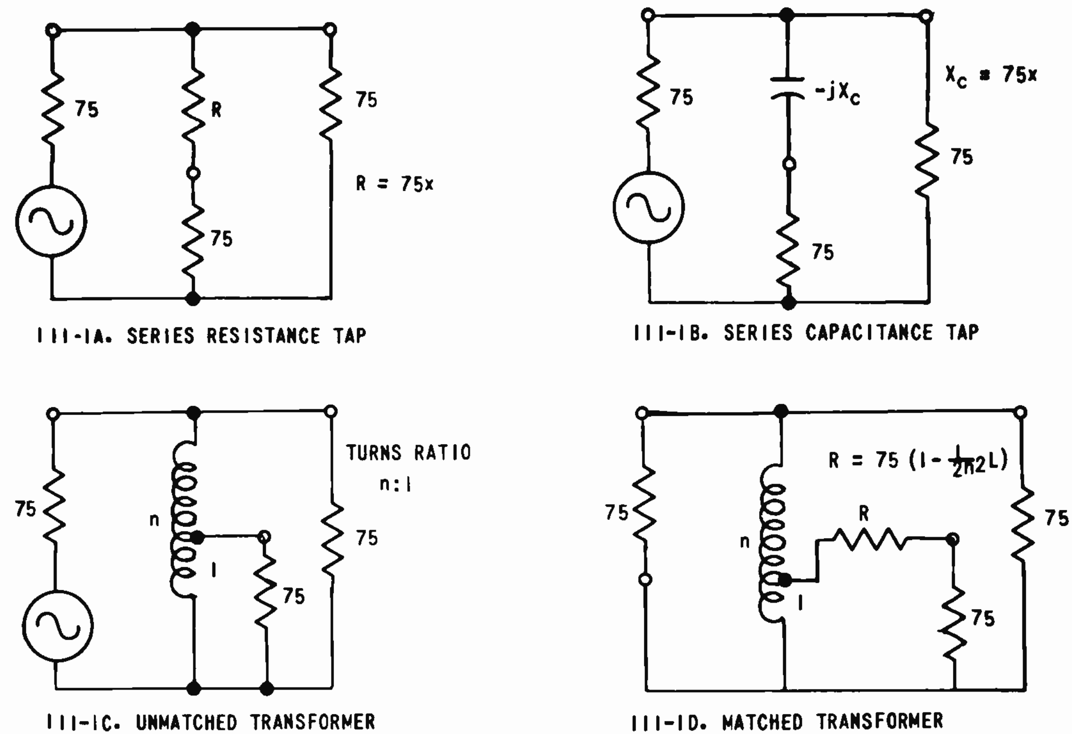
APPENDIX III

Taps in 75-Ohm Systems

In tapping matched systems a power loss and mismatch is introduced into the system, both of which are related to the tap loss. Only when hybrid coils or directional couplers are used can the mismatch in the 75-ohm system be designed out. A power (insertion) loss still exists related to tap loss. Directional taps have already been discussed in Chapter 8 and here only the simpler types of taps are discussed as given in Fig. III-1, based on series impedance and transformer taps. The formulae for the four types of taps are given in Table III-1.

All taps are plotted in the graphs of Figs. III-2 to III-4. The last graph gives a comparison of the various taps. For the series impedance type of tap, the return loss is always 6 db higher than the tap loss. The capacitive tap has the smaller insertion loss, because no power is dissipated in the capacitor. The frequency response of the resistive tap is better. The two transformer taps are only distinguished by the addition of the bridging-out resistor shown in Fig. III-1d. Both taps have a flat frequency response with the insertion loss higher for the backmatched type due to power lost in the resistor.

Comparison of the various types is made for a given tap loss in Fig. III-4. The lowest insertion loss and highest return loss is achieved in the unmatched transformer tap. Best all-around characteristics are achieved in the backmatched transformer tap.



III-1A. SERIES RESISTANCE TAP

III-1B. SERIES CAPACITANCE TAP

III-1C. UNMATCHED TRANSFORMER

III-1D. MATCHED TRANSFORMER

Fig. III-1. Forms of simple taps.

Table III-1. Electrical characteristics of simple taps.

Type of Tap	Series Resistance	Series Capacitance	Unmatched Transf.	Matched Transf.
Tap Loss	$2/(2x+3)$	$2/\sqrt{9+4x^2}$	$2n/(1+2n^2)$	$1/2n$
Insertion Loss	$(2x+2)\ (2x+3)$	$2\sqrt{1+x^2}/\sqrt{9+4x^2}$	$2n^2/(1+2n^2)$	$1-1/4n^2$
Return Loss	$1/(2x+3)$	$1/\sqrt{9+4x^2}$	$1/(1+2n^2)$	$1/4n^2$
X	$R/75$	$X_c/75$	-	-
n	-	-	Turns Ratio	Turns Ratio

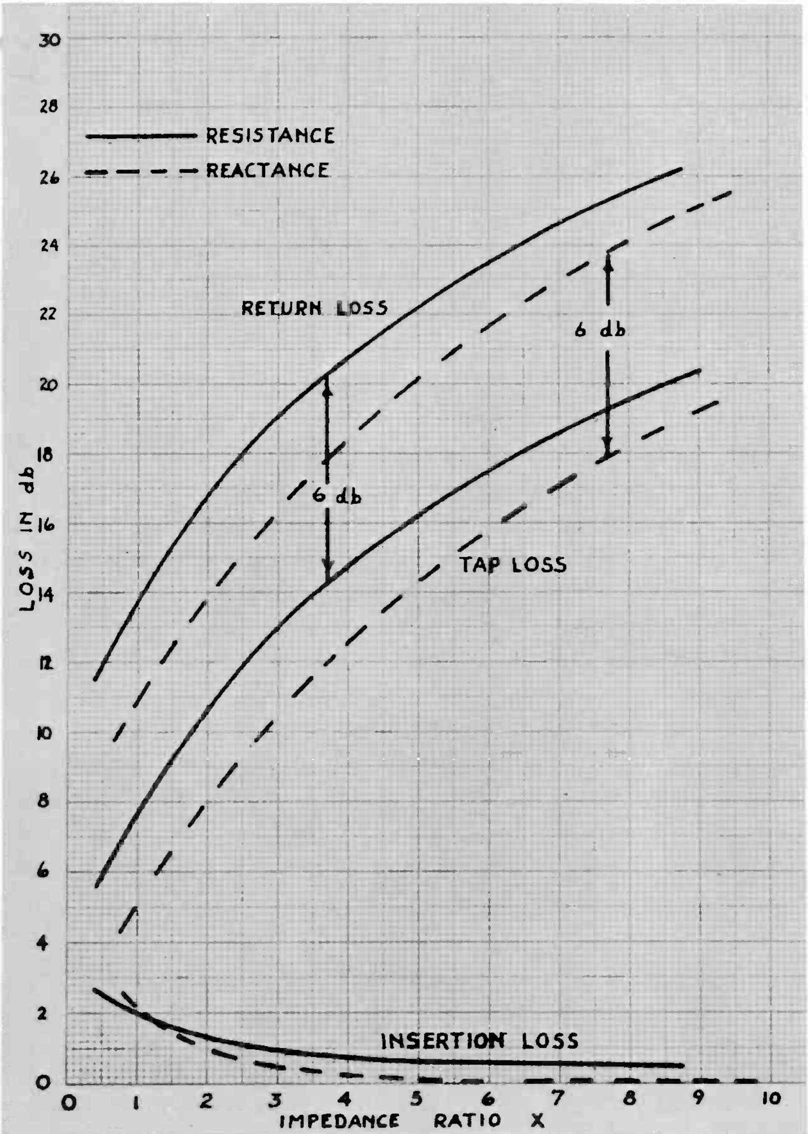


Fig. III-2. Series-impedance tap design graph.

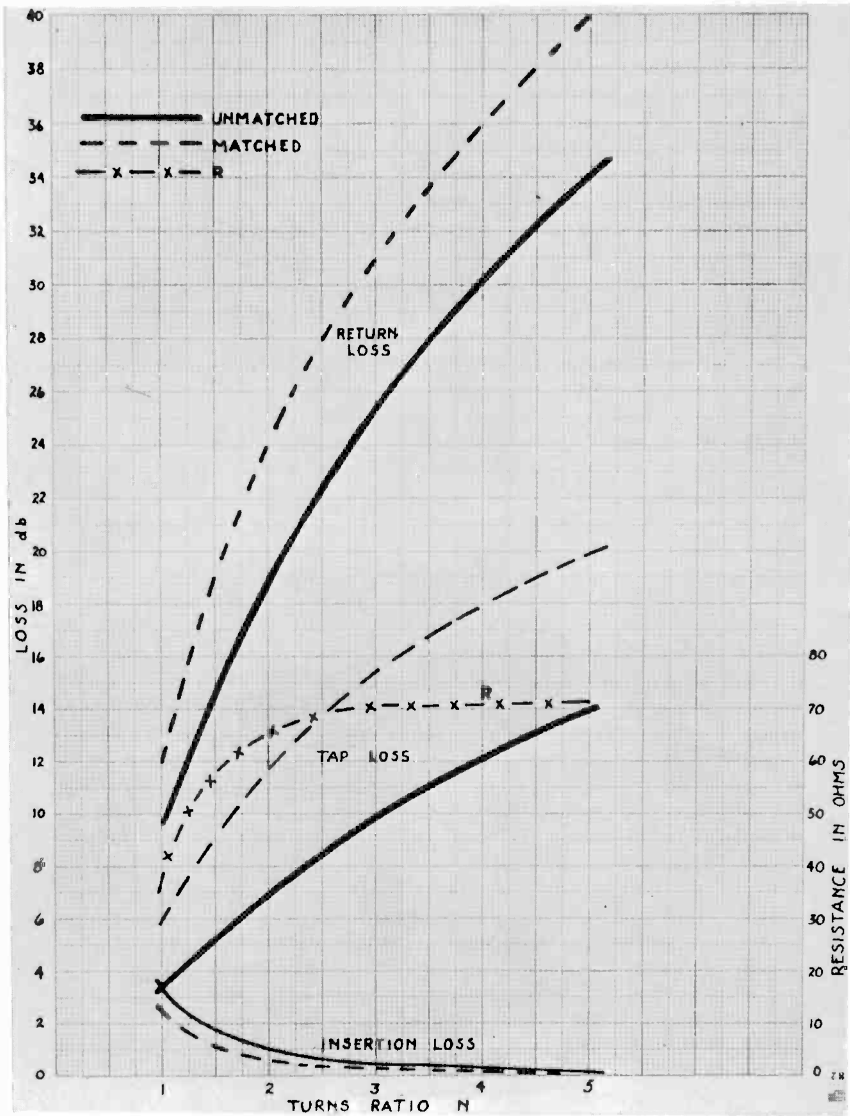


Fig. III-3. Transformer tap design graph.

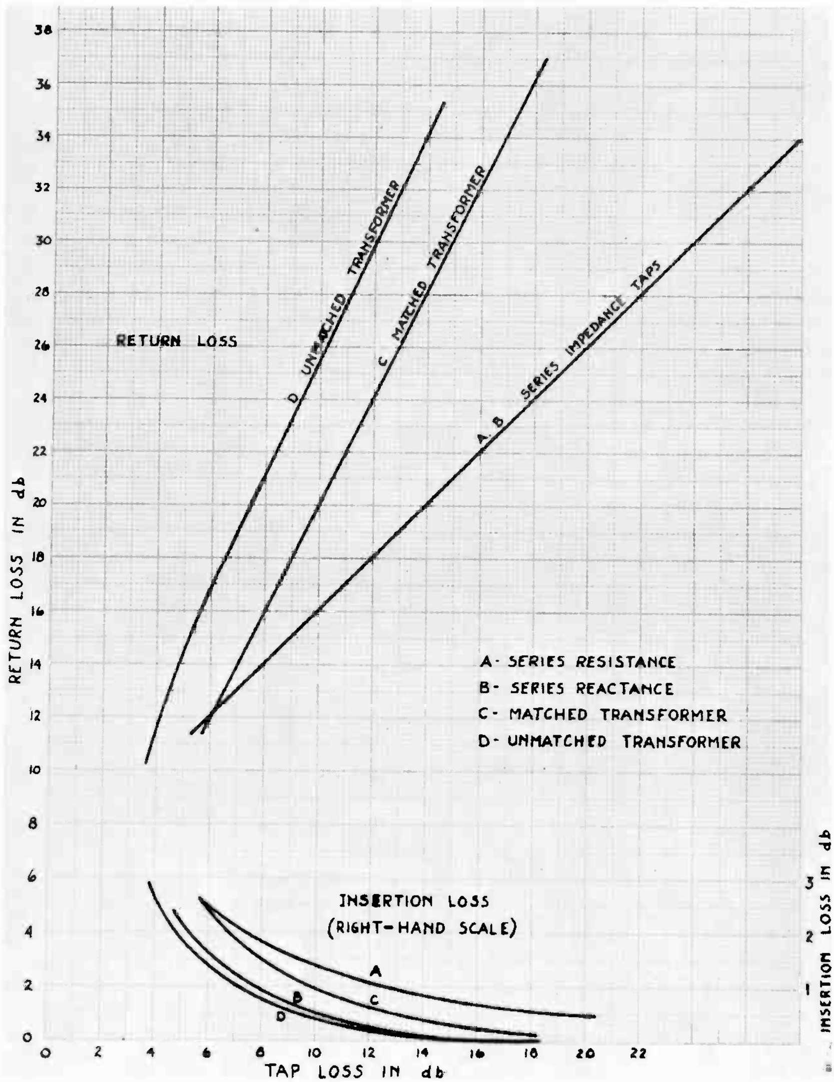


Fig. III-4. Graphical comparison of tap types.

A design example will illustrate the use of the curves. Design a tap with 15 db loss at 213 MHz. From Fig. III-2 we find $x = 4.1$ for the resistive and 5.4 for the capacitive tap. The resistance R in Fig. III-1 is then 308 ohms. Capacitive reactance is 405 ohms, corresponding to a capacitance of 1.7 pf at Channel 13 (Fig. III-2). For the transformer taps we find from Fig. III-3 a turns ratio of 1 to 1.5 for the unmatched type, and a ratio 1 to 2.8 for matched type together with a series resistor of 70 ohms.

APPENDIX IV

CATV Mathematics

I. <u>Systems and Amplifiers</u>

System level (dbmv) $L = (\emptyset + F + G)/2 - 9.5$ (1)

Max. cascade n $= \text{num}\left[(\emptyset - F - G + 19)/20\right]$ (2)

$= \text{num}\left[(D_a - 40)/20\right]$

Max. system length (db): $D = nG$ (3)

where $\emptyset$ = amplifier overload level in dbmv

F = amplifier noise figure in db

G = amplifier gain in db

n = number of amplifiers in cascade

D_a = amplifier dynamic range

with safety margins:

S_o = safety margin from overload in db

S_n = safety margin from noise in db

System level (dbmv): $L = (\emptyset + F + G + S_n - S_o - 19)/2$ (4)

Max. cascade n $= \text{num}\left[(\emptyset - F - G + 19 - S_n - S_o)/20\right]$

$= \text{num}\left[(D_a - 40 - S_n - S_o)/20\right]$ (5)

Max. system length (db): $D = nG$ (3)

Amplifier dynamic range (db): $D_a = \emptyset - F - G + 59$ (6)

System dynamic range (db): $D_s = D_a - 20 \lg n$ (7)

System overload level (dbmv): $\emptyset_s = \emptyset - 10 \lg n$ (8)

System input noise level (dbmv): $N_{si} = F + 10 \lg n - 59$ (9)

System output noise level (dbmv):

$$N_{so} = F + 10 \lg n + G - 59 \qquad (10)$$

System signal-to-noise ratio (db):

$$(S/N)_s = (\emptyset - F - G + S_n - S_o)/2 - 10 \lg n + 49.5 \qquad (11)$$

$$= (D_a + S_n - S_o)/2 - 10 \lg n + 20$$

II. <u>Derating and Combined Overload and Noise Levels</u>

$$b = 10 \lg \left[1/ \left[1 - 1/\text{num } (a/10) \right] \right] \qquad (12)$$

where a = operating margin below overload of amplifier A in db (or above noise level of amplifier A in db)

b = required operating level margin below amplifier B in db (or above noise level of amplifier B in db)

Combined overload level of amplifiers A and B, with loss between amplifiers equaling gain:

$$\emptyset_{a+b} = \emptyset_a - 10 \lg \left[1 + \text{num } \left[(\emptyset_a - \emptyset_b)/10 \right] \right] \quad O_a > O_b \qquad (13)$$

$$= \emptyset_b - 10 \lg \left[1 + 1/\text{num } \left[(\emptyset_a - \emptyset_b)/10 \right] \right]$$

Combined noise level:

$$N_{a+b} = N_a + 10 \lg \left[1 + \text{num}\left[(N_b - N_a)/10\right]\right] \quad N_b > N_a \quad (14)$$

$$= N_b + 10 \lg \left[1 + 1/\text{num}\left[(N_b - N_a)/10\right]\right]$$

where $\emptyset_a$ = overload level of amplifier A

N_a = noise level of amplifier A

For more than ten amplifiers the following formulas may be used:

Average overload level $\overline{\emptyset} = (O_1 + \emptyset_2 + \emptyset_3 + \ldots\ldots \emptyset_n)/n$ (15)

Average noise level $\overline{N} = (N_1 + N_2 + N_3 + \ldots\ldots N_n)/n$ (16)

where the indices refer to the individual amplifiers.

III. Mismatch, VSWR and Return Loss

Reflection coefficient:

$$r = (Z_L/Z_o - 1)/(Z_L/Z_o + 1), \text{ If } Z_L > Z_o \quad (17)$$

$$= (Z_o/Z_L - 1)/(Z_o/Z_L + 1), \text{ If } Z_o > Z_L$$

$$= E_r/E_i \text{ at point of mismatch}$$

where Z_o = characteristic impedance of cable

Z_L = load impedance

E_r = reflected voltage

E_i = incident voltage

Voltage Standing Wave Ratio (VSWR): $S = \dfrac{E_{max}}{E_{min}}$ (18)

at point of mismatch $S = (1+r)/1-r)$ (19)

$$= Z_L/Z_o \quad Z_L > Z_o$$

$$= Z_o/Z_L \quad Z_o > Z_L$$

$$E_r/E_i = (S-1)/S+1) \tag{20}$$

$= r$ at receiving end of line only

Return Loss (db): $R = 20 \lg (1/r)$ (21)

Return loss (db) after attenuator or cable with x db loss:

$$R = 20 \lg r - 2x \tag{22}$$

Mismatch Loss (db) $= 10 \lg (1-r^2)$ (23)

IV. Cable Attenuation and Tilt

Gain at Channel 2 (db): $G_2 = 0.48\ G + 0.52\ x$ (24)

Tilt (db): $T = 0.52\ G - 0.52\ x$ (25)

Level vs. system tilt: (26)

	Output Level, Channel 2	Input Level, Channel 2
Full Tilt	$L - 0.52\,G + 0.52\ x$	$L - G$
Half Tilt	$L - 0.26\,G + 0.26\ x$	$L - 0.74\,G - 0.26\ x$
Flat Outputs	L	$L - 0.48\,G - 0.52\ x$

Overload vs. System Tilt: (27)

	Overload Level, Channel 2	Overload Level, Channel 13
Full Tilt	$\emptyset_2$	$\emptyset_{13}$
Half Tilt	$\emptyset_2 + 0.13\ G$	$\emptyset_{12} - 0.13\ G$
Flat Outputs	$\emptyset_2 + 0.26\ G$	$\emptyset_{13} - 0.26\ G$

where G = amplifier gain at Channel 13 (spacing in db)

G_2 = amplifier gain at Channel 2 in db

L = System level (output level, Channel 13) in dbmv

$\emptyset_{13}$ = overload level, Channel 13 in dbmv

$\emptyset_2$ = overload level, Channel 2 in dbmv

x = flat gain in db

V. Miscellaneous Formulae

Length (in feet) traveled in 1 nsec (10^{-9} sec) $= 0.986\ \eta$ (28)

where η = propagation constant in free air

1 nsec $\sim$ 1 light - foot

Wavelength (in feet: $L = 986\ \eta / f$ (29)

where f = frequency in MHz

Volt-drop for cable powering

$$V_D = iR\,n\,(n+1)/2 \tag{30}$$

where i = average current drain
R = loop resistance
n = number of amplifiers

Temperature Conversion: $^{o}C = 5\,(^{o}F - 32)/9$ (31)

Mean square thermal noise:

$$E_n^{\,2} = 1.63 \times 10^{-20}\ B\,R \text{ at } 75^{o}\text{ F} \tag{32}$$

where B = bandwidth in Hz
R = resistance in ohms

Equivalent noise sideband input (ENSI)

$$= m\,E_{si}\sqrt{P_{no}/P_{so}} \tag{33}$$

where m = modulation index
E_{si} = signal input voltage
P_{no} = noise output power
P_{so} = signal output power

Cross-modulation index:

$$m_k = 3\,a_3\,m_2\,E_2^{\,2}/a_1 \tag{34}$$

where a_1, a_3 = coefficients of power series
E_2 = peak voltage of interfering signal
m_2 = modulation index of interfering signal

Cross-modulation factor:

$$k = 3\,a_3\,m_2\,E_2^{\,2}/a_1\,m_1 \tag{35}$$

where m_1 = modulation index of desired signal

Modulation index:

$$m = \frac{E_{max} - E_{min}}{E_{max} + E_{min}} \tag{36}$$

$$= (E - E_{min}) / E \text{ for down modulation,}$$

where $E = (E_{max} + E_{min}) / 2$
$= \text{unmodulated signal level}$

$$Z_o = 138 \log (D/d) / \sqrt{E_e} \tag{37}$$

$$yx = \frac{100}{\sqrt{Ee}} \% \tag{38}$$

where
Z_o = characteristic impedance in ohms
Ee = dielectric constant of insulation
D = inside diameter of outside conductor, inches or cm
d = diameter of center conductor, inches or cm
n = velocity of propagation in percent

VI. Examples and Comments

Equations (1) through (11) are contained in Fig. VI-14 which is easy to use and allows sufficient accuracy. Similarly, equations (12) through (14) are contained in Fig. I-1, equations (17) through (23) in Figs. VI-18, VI-19, VI-20.

Equations (24) through (27) are based on average cable data for 1/2" cable. There is a very slight deviation for .412 or .750" cable. A few examples will illustrate the use of the equations.

Example 1

Calculate optimum system level for an amplifier with an output capability of 55 dbmv, a noise figure of 12 db at 22 db gain. A total safety margin of 6 db is to be used with 4 db margin from overload and 2 db margin from noise. What is the maximum cascade and system length?

From equation (4), we find

$$\begin{aligned} L &= (55+12+22+2-4-19)/2 \\ &= 68/2 \\ &= 34 \text{ dbmv} \end{aligned}$$

From (5), $$\begin{aligned} n &= \text{num}\left[(55-12-22+19-2-4)/20\right] \\ &= \text{num } 1.7 \\ &= 50 \text{ amplifiers} \end{aligned}$$

From (3), $$D = 50 \times 22 = 1100 \text{ db}$$

Example 2

Calculate amplifier and system dynamic range for the system of Example 1. What is the system signal-to-noise ratio?

From (6), we have amplifier dynamic range

$$\begin{aligned} D_a &= 55-12-22+59 \\ &= 80 \text{ db} \end{aligned}$$

From (7), system dynamic range

$$\begin{aligned} D_s &= 80-20 \text{ lgn} \\ &= 80-34 \\ &= 46 \text{ db} \end{aligned}$$

From (11), system signal-to-noise ratio

$$\begin{aligned} (S/N)_s &= (55-12-22+2-4)/2 - 10 \lg 50 + 49.5 \\ &= 19/2 \quad -17+49.5 \\ &= 42 \text{ db} \end{aligned}$$

Example 3

Two amplifiers are to be cascaded. Amplifier A has an overload level of 55 dbmv, Amplifier B overloads at 40 dbmv. Amplifier A is to be run at 53 dbmv. How far must Amplifier B be derated?

From equation (12) with $a = 2$ db
we find

$$\begin{aligned} b &= 10 \lg \left[1/\left[1-1/\text{num } 0.2\right]\right] \\ &= 10 \lg \left[1/\left[1-0.631\right]\right] \\ &= 10 \lg 2.71 = 4.33 \text{ db} \end{aligned}$$

Amplifier B must be run at 35.67 dbmv

Example 4

A 75 - ohm system uses a 50-ohm uhf connector. What is the reflection coefficient, return loss, and VSWR? What is the VSWR 5 db of cable away? What is the mismatch loss?

From (19), we have VSWR $= 1.5$

From (17), reflection coefficient r $= 0.5/2.5$
$= 0.2$

From (21), return loss R $= 20 \lg 5$
$= 14$ db

5 db of cable away, the incident wave is 5 db higher and the reflected wave 5 db lower. Return loss is, therefore, 24 db. Reflection coefficient from (21) is found as

$$1/r = \text{num}\frac{24}{20} = \text{num } 1.2 = 15.85, \text{ or}$$

$$r = 1/15.6 = .063$$

From (19), VSWR $s = 1.063/.937 = 1.13$

From (23), mismatch loss $= 10 \lg (1 - 0.04) = 10 \lg .96$
$= -0.18$ db

Example 5

A system is run full-tilt with 3 db flat loss and a total gain of 20 db. What is the output level at Channel 2 if the level at Channel 13 is 43 dbmv?

From (26), we find level $= 43 - 0.52 \times 22 + 0.52 \times 3$
$= 43 - 11.44 + 1.56$
$= 34.0$ dbmv

Example 6

What length of RG 59/U cable corresponds to a quarter wavelength at 220 MHz?

From (29), with $\eta = 0.66$ we find

wavelength $L = 986 \times 0.66/220$
$= 2.95$ feet

quarter wavelength $= .738$ feet $= 8.85$ inches

Example 7

The voltage drop in the last span of a cascade of amplifiers is 0.5 volts. What is the total drop for a cascade of 6 amplifiers? Assume that the minimum amplifier input voltage is 20 VAC. What is the required minimum supply voltage?

From (30), $V_D = 0.5 \times 6 \times 7/2 = 10.5$
The minimum required supply voltage is 30.5 VAC

NOTE: Constant current for all amplifiers was assumed here. For other power characteristics, the average current should be chosen (current at median supply voltage).

Example 8

A coaxial cable with foamed dielectric ($\varepsilon = 1.5$) has an inside diameter of the outer conductor of 447 mils and a diameter of the center conductor of 98 mils. What is the characteristic impedance and velocity of propagation?

From (37), we find $Z_o = 138 \lg (447/98)/\sqrt{1.5}$
$= 138 \times 0.659/\sqrt{1.5}$
$= 74.3$ ohms

From (38), $\eta = 1/\sqrt{1.5}$
$= 81.6\%$

APPENDIX V

Typical Equipment Specifications

The following specifications are intended as a guide for good quality, transistorized systems. They do not necessarily represent the ideal "Integrated System" concept mentioned in this book.

Main Trunk Amplifier

Bandwidth	50 to 220 MHz
Equalization for	75-ohm coax
Flatness of equalization	± 0.25 db
Gain at Channel 13	22 db
Noise figure	10 db
Multichannel overload level	55 dmbv
Tilt control	± 3 db
Gain control	± 5 db
Temperature compensation to	± 0.5 db
Input match	1.25
Output match	2.0
Test points (input and output)	-20 db
AC power voltage, 60 Hz	20 to 40 VAC
Current drain	125 to 250 ma AC

Level Control (AGC) Amplifier

Specifications as above, except	
Output change,	
for input change of ± 10 db	± 1 db
AGC control	
(to set output level)	± 5 db
Current drain	250 to 500 ma DC

Bridging Amplifier

Bandwidth	50 to 220 MHz
Equalization for	75-ohm coax
Flatness of equalization	± 0.5 db
Gain at Channel 13	15 db
Noise figure	15 db
Multichannel overload level	50 dbmv
Isolation from main trunk via directional coupler	10 db
Number of outputs	4
Tilt control	± 3 db
Gain control	± 5 db
Input match	1.25
Output match	1.5
AC power voltage	20 to 40 VAC
Current drain	200 to 400 ma AC

Bridger Combination Amplifier

Main-trunk or AGC section similar to main-trunk or AGC amplifier bridger section similar to bridging amplifier coupling to main trunk via directional coupler.

Distribution Amplifier (High Level)

Bandwidth	50 to 220 MHz
Equalization for	75-ohm coax
Flatness of equalization	± 0.25 db
Gain at Channel 13	22 db
Gain adjustment for	15 to 25 db of cable
Tilt adjustment for	0 to 9 db of flat loss
Noise figure	15 db
Multichannel overload level	55 dmbv
Input match	1.25
Output match	1.5
Built-in 4-way directional tap	
AC power voltage	20 to 40 VAC
Current drain	100 to 200 ma

Line Extender (Low Level)

Similar to distribution amplifier, except	
Gain at Channel 13	15 to 20 db
Multichannel overload level	40 dbmv
Current drain	50 to 100 ma

Directional Couplers

Match, all terminals	1.15
Directivity, Channel 13	10 db
Directivity, Channel 2	15 db
Insertion loss	See Fig. 8-7

Head-End Equipment

Tuner Section

Noise figure*	6 db
Input VSWR	1.25
Bandwidth	6 MHz
Flatness	± 0.25 db
Adjacent Ch. rejection	60 db
Adjacent Ch. cross modulation:	
Less than 1% at equal inputs of	50 dbmv
Rejection of spurious signals	60 dbmv
Frequency stability	0.01%
AGC compression ratio	1 : 40
AGC range	35 db
Video output signal	1v p-p
Video output impedance	75 ohms
4.5-MHz FM output	0.1v
Output impedance	75 ohms
Sound output	0 dbmv
Impedance	600 ohms

Controls: Video Level, Sound Level, Input Balance, TV Channel, AGC Range.

*May be achieved by separate antenna preamplifiers.

AC power voltage	20 to 40 VAC (100 to 130 VAC)
Current drain	250 to 500 ma AC (50 to 65 ma AC)

Modulator Section

Picture carrier output	50 dbmv
Range adjustment	± 10 db
Output impedance	75 ohms
Spurious frequency rejection	60 db
Frequency tolerance (crystal control)	± 0.005%
AM Modulation capability	100%
Maximum envelope distortion	3%
Video input for full modulation	0.5v p-p
Video s/n ratio, more than	40 db
Video bandwidth	5 MHz
Differential gain, less than	± 0.5 db
Differential phase, less than	$\pm 2.5^{\circ}$
Vestigial sideband characteristic	± 0.5 db
Isolation from sound section	40 db
Sound carrier output	40 dbmv
Range of adjustment	± 10 db
AM on sound carrier, less than	1%
Spurious frequency rejection	60 db

FM Modulator

Audio drive for ± 25 kHz deviation	-6 dbmv
Input impedance	600 ohms
Harmonic distortion, ± 25 kHz deviation	1% at 1 kHz or less
Pre-emphasis time constant	75 microsec
Audio s/n ratio	60 db

AC power requirements for complete modulator:

Voltage	20 to 40 VAC (100 to 130 VAC)
Current drain	250 to 500 ma AC (50 to 65 ma AC)

System Specifications

(20 to 80 Amplifiers in cascade)

Main Trunk

Spacing, Channel 13	22 db*
Input levels, all channels	12 dbmv
Output levels, Channel 13	34 dbmv
System mode	Full Tilt
AGC position	Every third amplifier

Distribution System (high level)

Bridger outputs	40 dbmv
Spacing of distribution amplifiers	1000 ft. of .412" cable
Flat loss	0 to 9 db
Input levels, all channels	22 dbmv
Output levels, Channel 13	43 dbmv
Output levels of directional taps	12 dbmv nominal
Subscriber level	3 dbmv nominal
Number of subscribers per amplifier	40

Distribution System (low level)

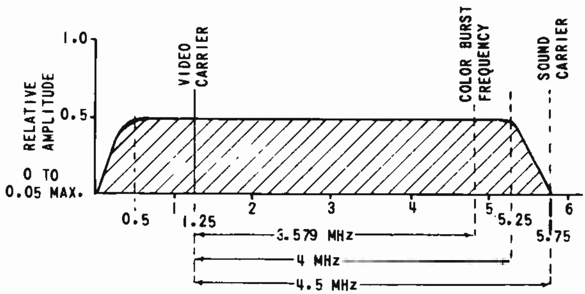
Similar to high level, except

Spacing of line extenders	600 ft. of .412" cable
Input levels, all channels	15 dbmv
Output level, Channel 13	32 dbmv
Number of subscribers per amplifier	24

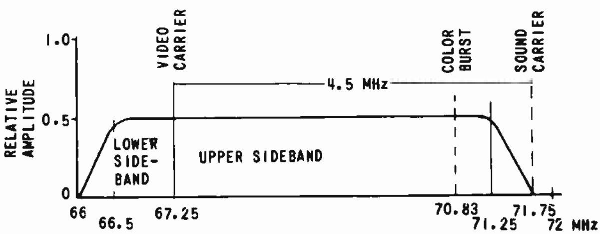
*In integrated systems, spacing is expressed in footage rather than db loss, e.g., 1750' for 1/2" cable.

APPENDIX VI

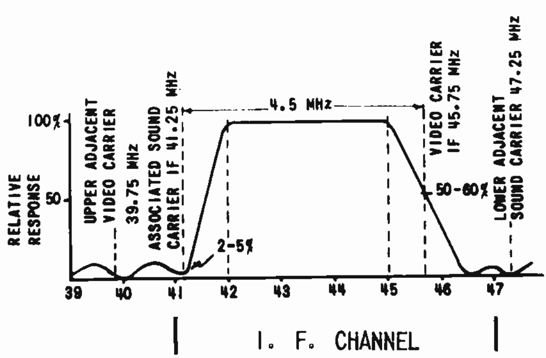
Miscellaneous CATV Data & Charts



1. Standard TV channel showing sound-video carrier relationship.



2. Channel 4 sound-video carrier relationship.

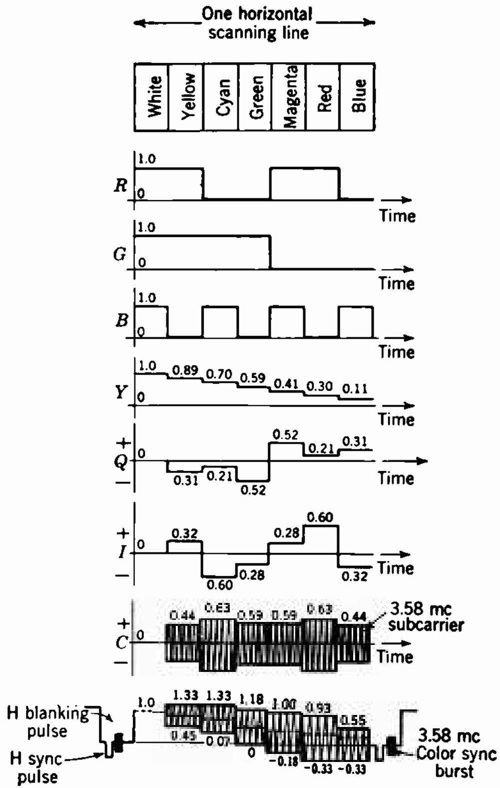


3. Standard receiver IF response curve.

4. Television broadcast channel frequencies and wavelengths.

Channel	Band	Picture Carrier	Sound Carrier	Half-Wave Lengths in Inches			
	MHz	MHz	MHz	Free Air	Dipole	Coax Vel. Prop.0.84	Coax Vel. Prop.0.67
2	54-60	55.25	59.75	103.5	97.5	87.	69.4
3	60-66	61.25	65.75	93.7	88.2	78.7	62.8
4	66-72	67.25	71.75	85.5	80.4	71.9	57.3
5	76-82	77.25	81.75	74.7	70.2	62.7	50.1
6	82-88	83.25	87.75	69.5	65.4	58.4	46.6
7	174-180	175.25	179.75	33.4	31.4	28.1	22.4
8	180-186	181.25	185.75	32.2	30.3	27.1	21.6
9	186-192	187.25	191.75	31.2	29.3	26.2	20.9
10	192-198	193.25	197.75	30.3	28.5	25.4	20.3
11	198-204	199.25	203.75	29.4	27.7	24.7	19.7
12	204-210	205.25	209.75	28.5	26.8	23.9	19.1
13	210-216	211.25	215.75	27.7	26.1	21.9	18.6
14	470-476	471.25	475.75	12.5	11.75	10.5	8.4
83	884-890	885.25	889.75	6.7	6.3	5.6	4.5
*	470+A to 476+A	471.25 +A	475.75 +A	591/B	556/B	496/B	396/B

*For UHF channels between Channels 14 and 83, obtain A = 6 (x-14), where x = channelnumber, and B = 473+A. Read values from last line of table.



5. Composite color video signal. (Drawings from Basic Television, B. Grob, permission McGraw Hill Book Co.)

Red Video Signal

Green Video Signal

Blue Video Signal

Luminance Signal

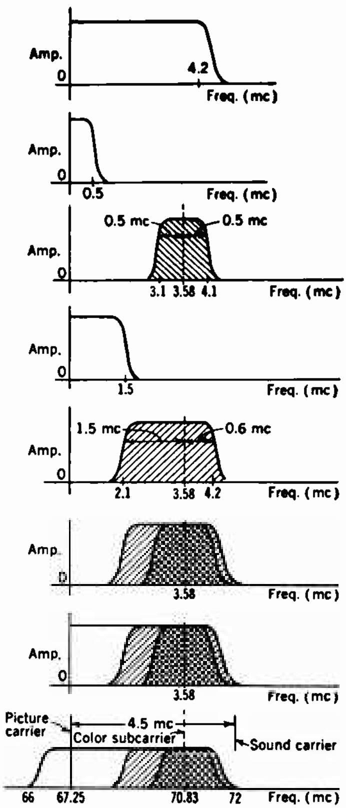
Q - Signal

I - Signal

Chrominance Signal, Modulated by I- and Q-Signals In Quadrature

Total Color-Plexed Video Signal With Added Y-Modulation and Sync Pulses

6. Color bandwidth requirements. (Drawings from Basic Television, B. Grob, permission McGraw Hill Book Co.)



Y - Signal

Q - Signal

Sidebands of 3.58 mc Chrominance Subcarrier Due to Q-Signal

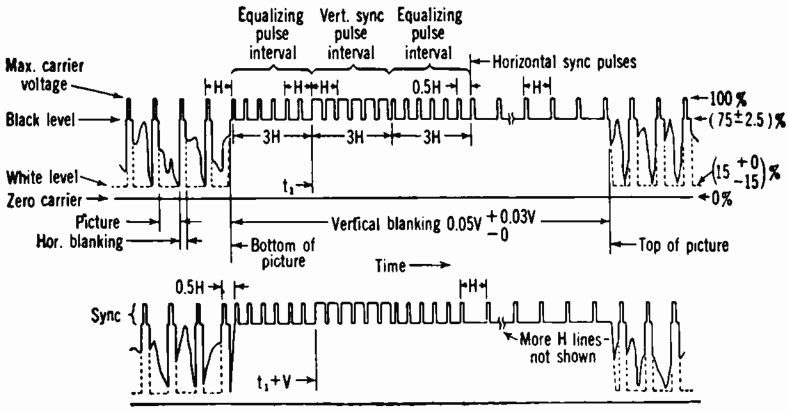
I - Signal

I - Modulated Chrominance Signal

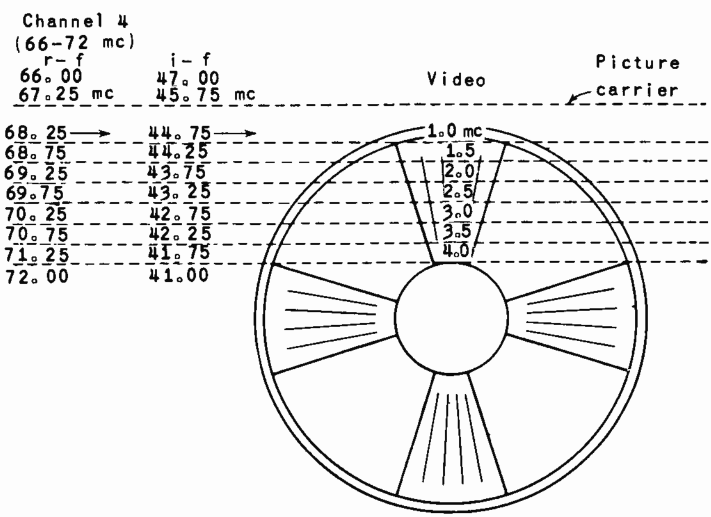
Total Chrominance Signal

Complete Color-plexed Video Signal

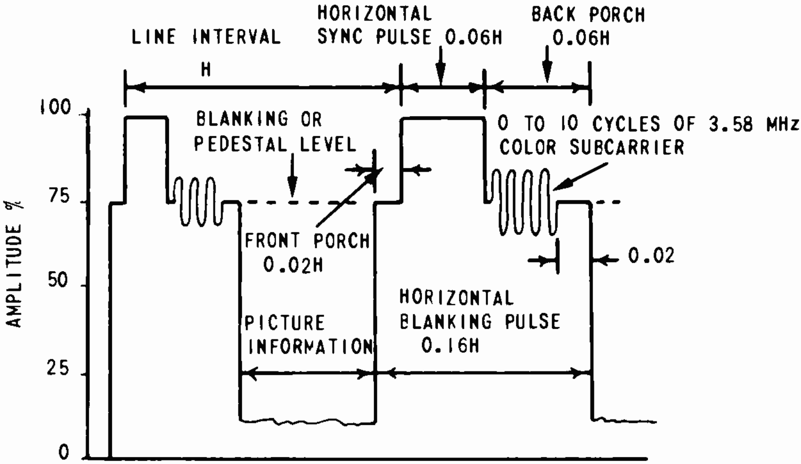
Sideband Frequencies For Channel 4



7. Composite TV waveform showing sync and blanking pulses. V = 1/60 sec. H = 1/15,750 sec.



8. Resolution chart illustrating continuity of RF, IF, and video signal frequencies for Channel 4. (above drawings from Basic Television, by B. Grob permission of McGraw Hill Book Co.)



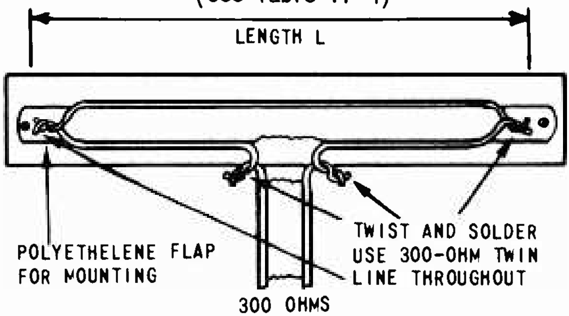
9. Detailed sketch of horizontal blanking and sync pulses.

34. PROPOSED CATV CHANNEL ALLOCATIONS

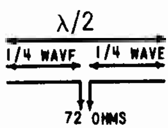
CHANNEL	FREQUENCY RANGE (MHz)	CHANNEL	FREQUENCY RANGE (MHz)
2	54-60	7	174-180
3	60-66	8	180-186
4	66-72	9	186-192
5	76-82	10	192-198
6	82-88	11	198-204
A	120-126	12	204-210
B	126-132	13	210-216
C	132-138	J	216-222
D	138-144	K	222-228
E	144-150	L	228-234
F	150-156	M	234-240
G	156-162	N	240-246
H	162-168	O	246-252
I	168-174	P	252-258

10. Construction of simple 300-ohm folded dipole.

(See Table VI-4)

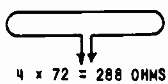


11. Antenna types.

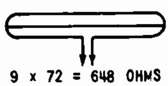


Half-wave Dipole

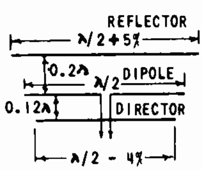
$L = \frac{555}{f}$ (inches MHz)



Folded Half-wave Dipole

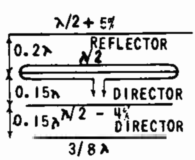


High Impedance
Folded Dipole



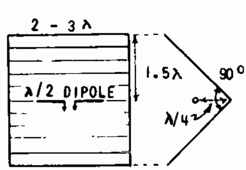
Half-wave Dipole With
Director and Reflector

= Wavelength
Gain 7 db
Z = 1/8 of Dipole



Yagi Antenna With 2 Directors

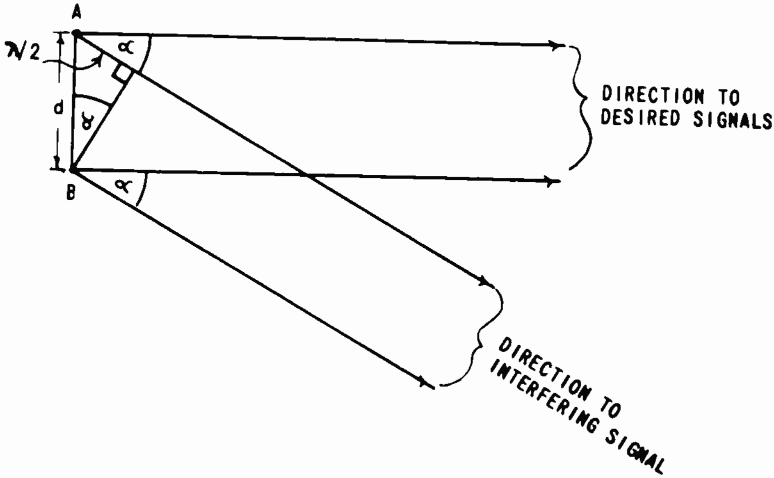
Gain 10 db
Front-to-back Ratio 15 db



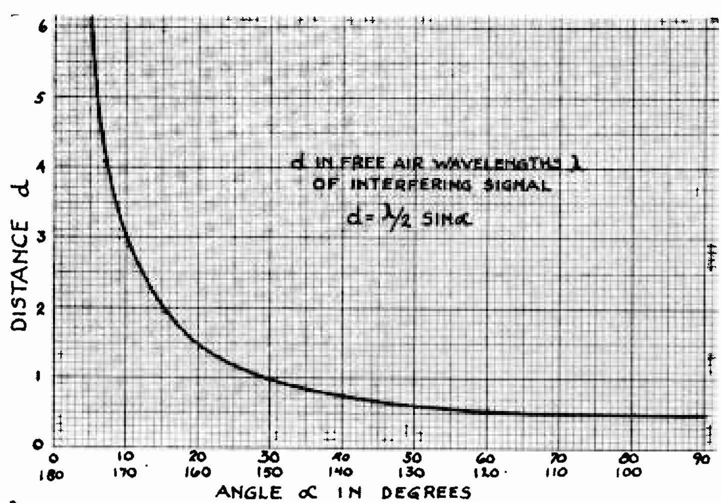
Dipole With Corner Reflector

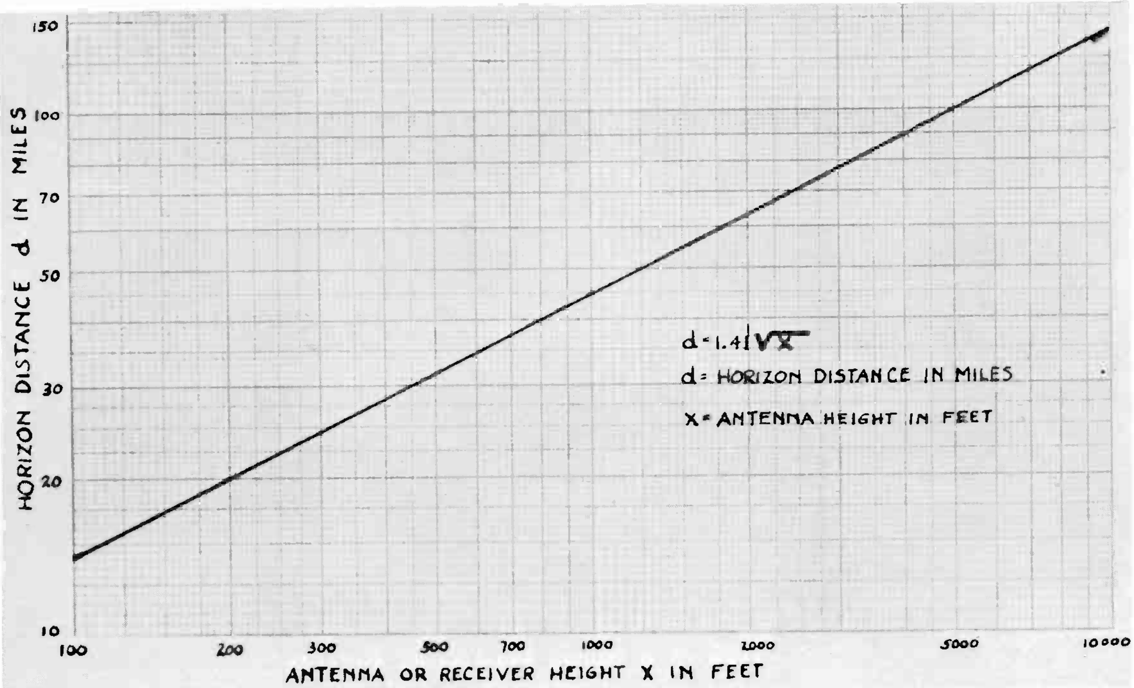
Gain 10 db
Front-to-back Ratio 20 db

12. Spacing of two antennas for the suppression of one interfering signal arriving at an angle to the desired signal.



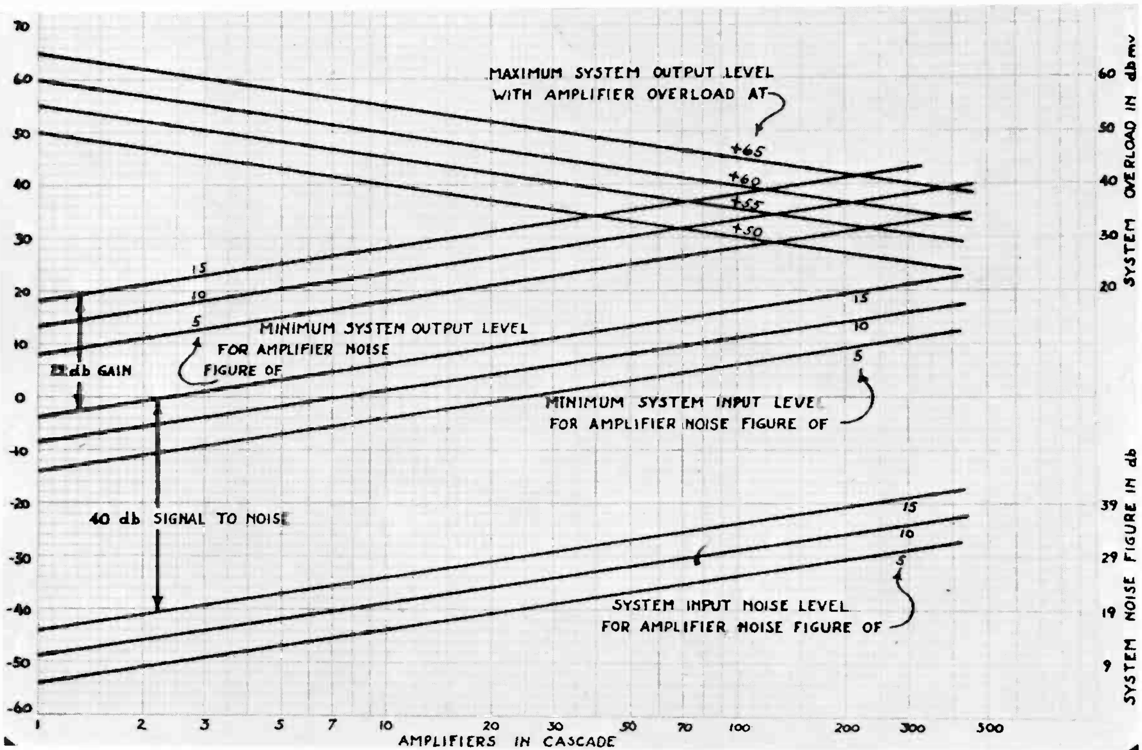
Antenna locations at A and B, spaced distance d apart. Spacing must be such, that signal at A is delayed by one half wavelength, hence $d = \lambda/2 \sin \alpha$. Read d in wavelengths of the interfering station for any angle from graph below.





13. Horizon distance.

UHF Line-of-Sight propagation depends on horizon distance, which is determined by antenna height x. The radio horizon distance is about 15% longer than the optical distance because the radio waves follow slightly the earth's curvature. This is taken into account in the graph left (in profile charts by increasing the earth radius by a factor of 4/3). If both transmitter and receiver antenna are elevated, radio distance is the sum of the individual horizon distances. For other than smooth earth, the horizon distance is best determined graphically using profile charts.

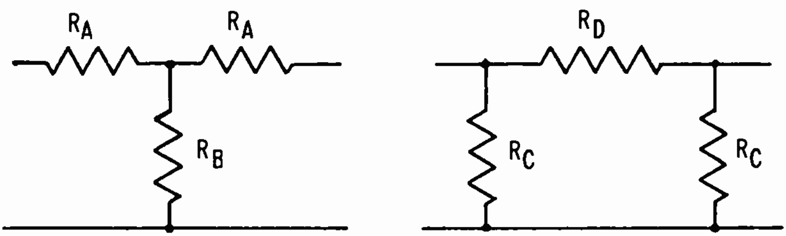


14. System derating chart.

15. 75-ohm symmetrical attenuators.

Loss in db	T - attenuator		Π - attenuator	
	R_A	R_B	R_C	R_D
0.5	2.16	1300	2610	4.32
1.0	4.32	650	1305	8.68
1.5	6.46	431	872	13
2.0	8.60	323	652	17.4
3.0	12.8	212	438	26.4
4.0	16.8	157	331	35.8
5.0	21.0	124	267	45.5
6.0	24.9	100	226	55.9
7.0	28.6	83.2	196	67.1
8.0	32.3	71	174	79
9.0	35.8	60.8	157	92.2
10.0	38.9	52.6	144.2	106.5
12.0	44.9	40.2	125	140
14.0	50	31	112	181
16.0	54.3	24.3	103	230
18.0	58.2	19	96.5	292
20.0	61.2	15.2	91.6	371

DIAGRAM



$R_A = 75(k-1)/(k+1)$

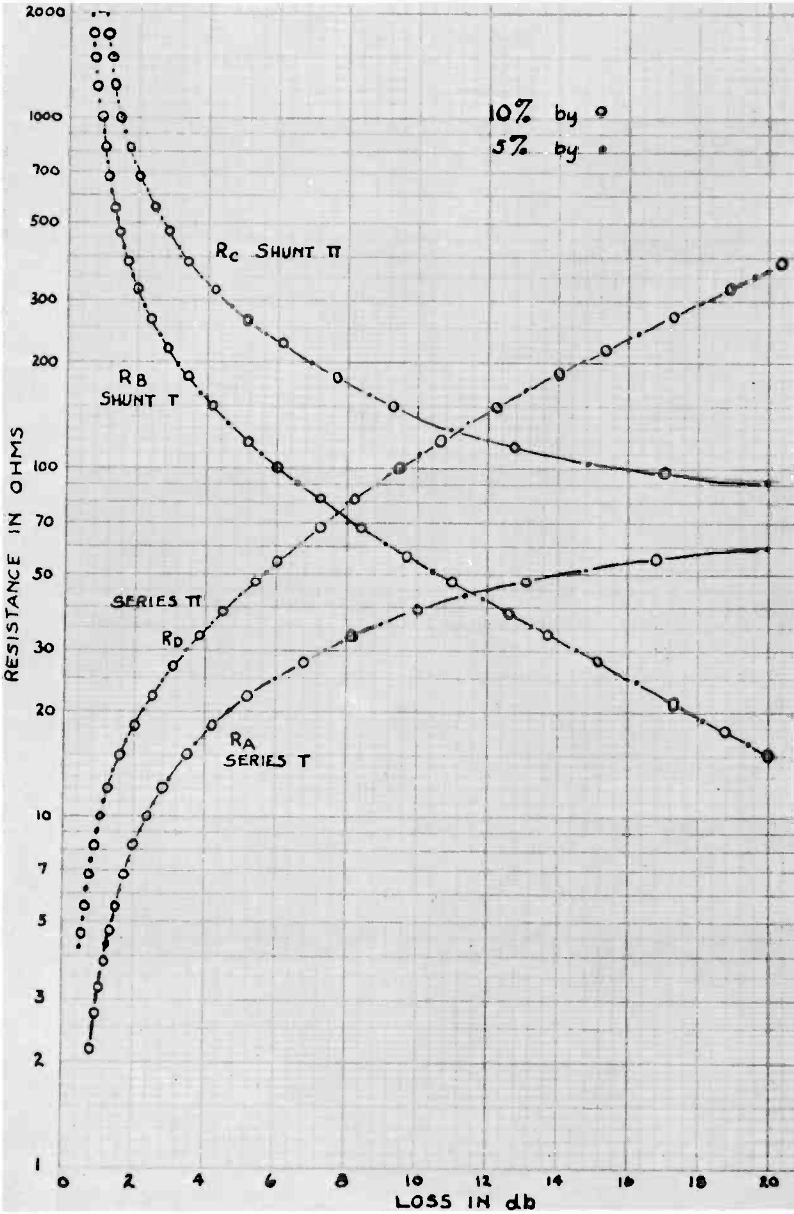
$R_B = 150k/(k^2-1)$

$k = e_i/e_o$

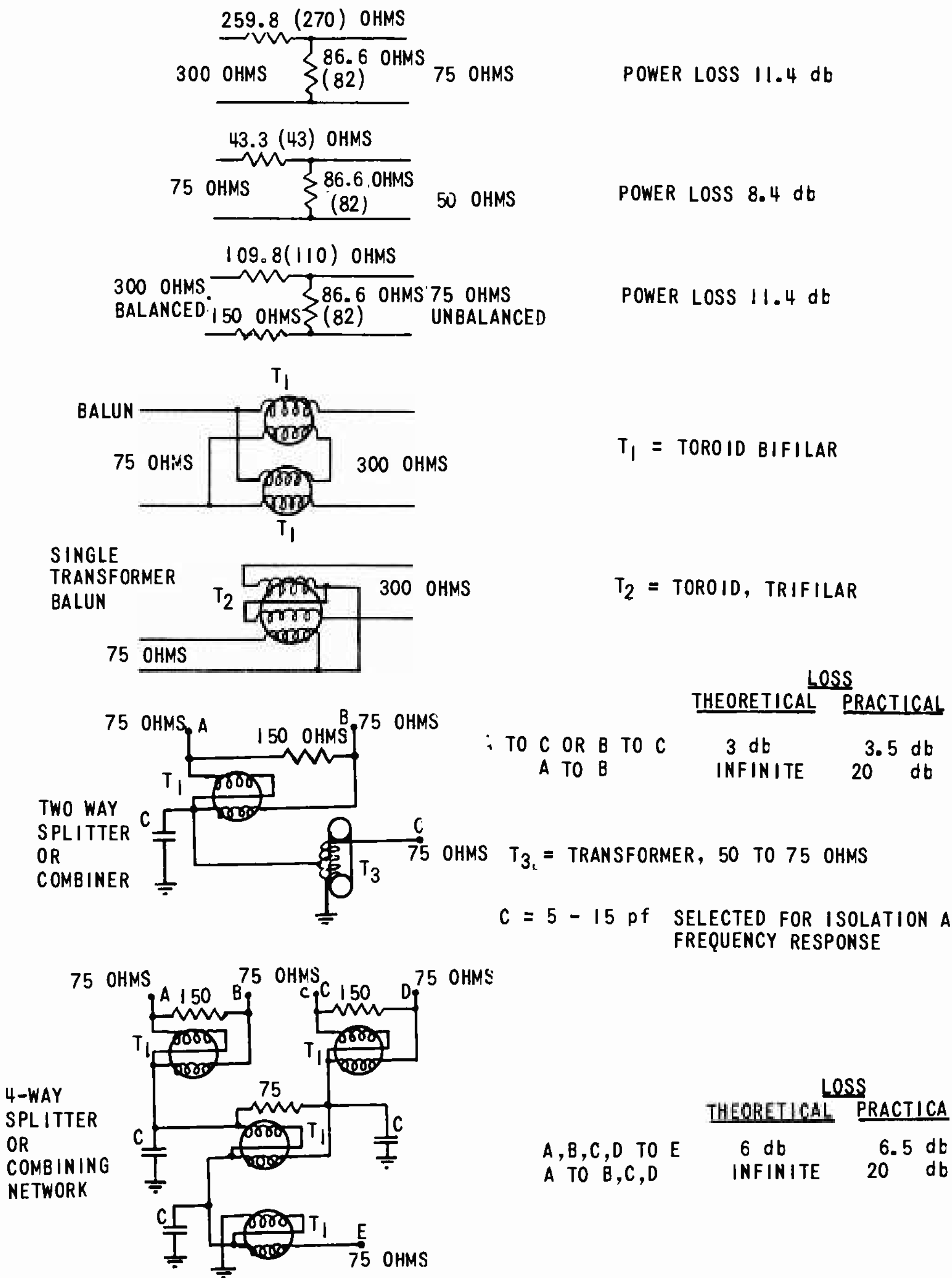
$R_C = 75(k+1)/(k-1)$

$R_D = 75(k^2-1)/2k$

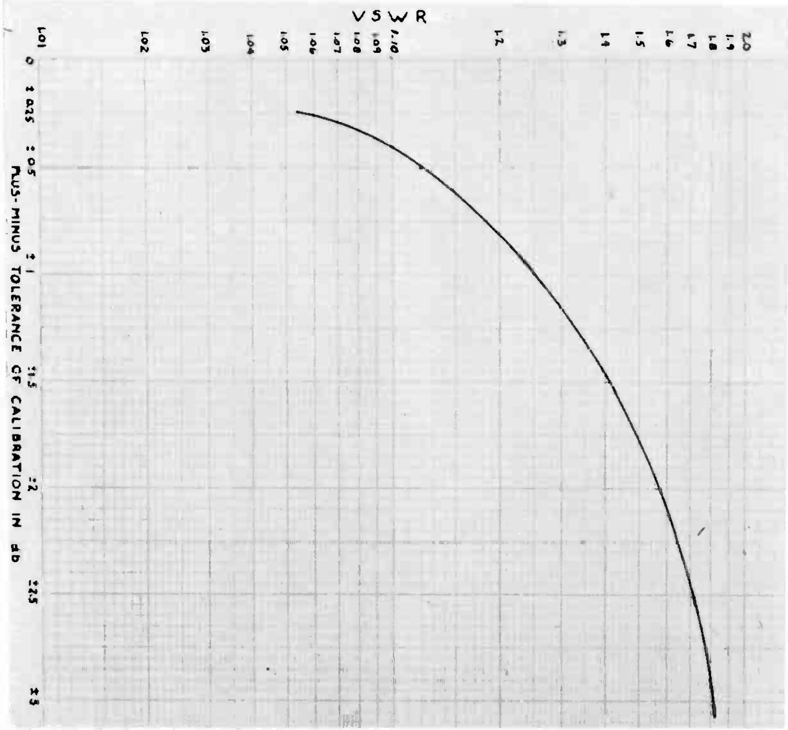
db Loss = 20 lg k



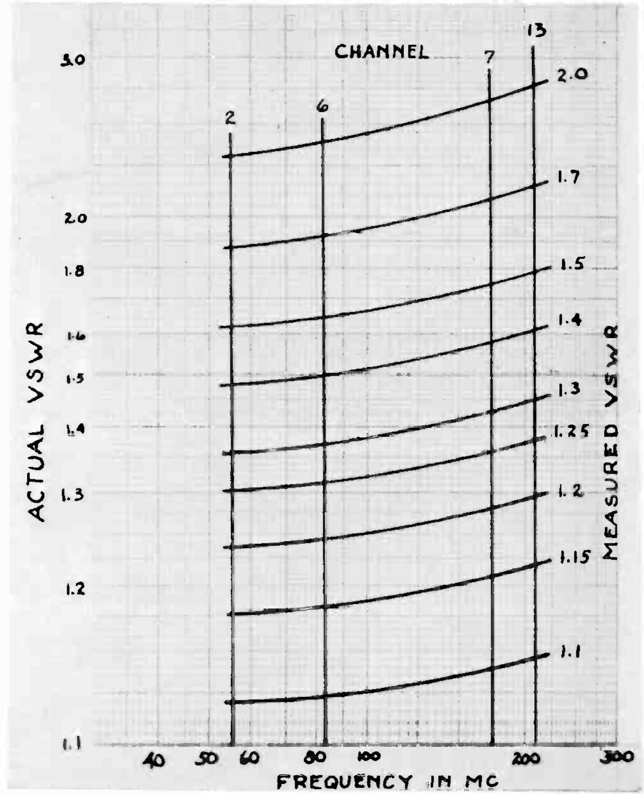
16. Accuracy of symmetrical attenuators.



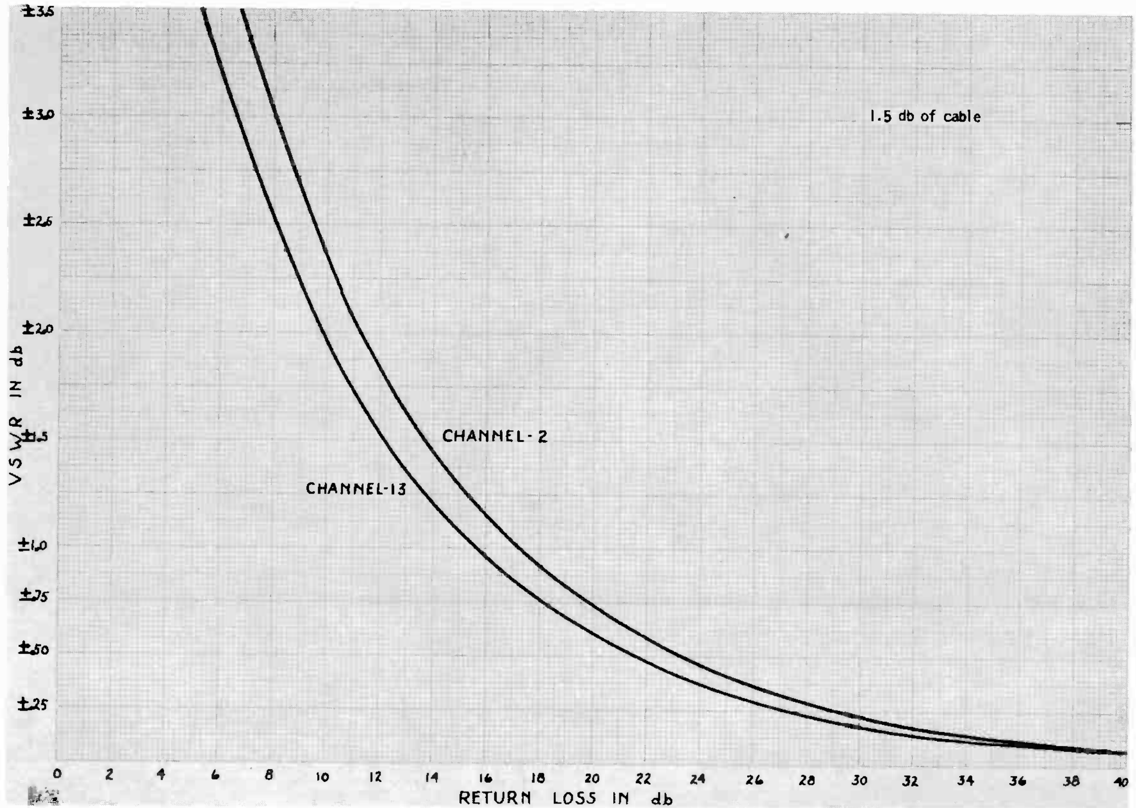
17. Minimum-loss pads, baluns and hybrid coils.



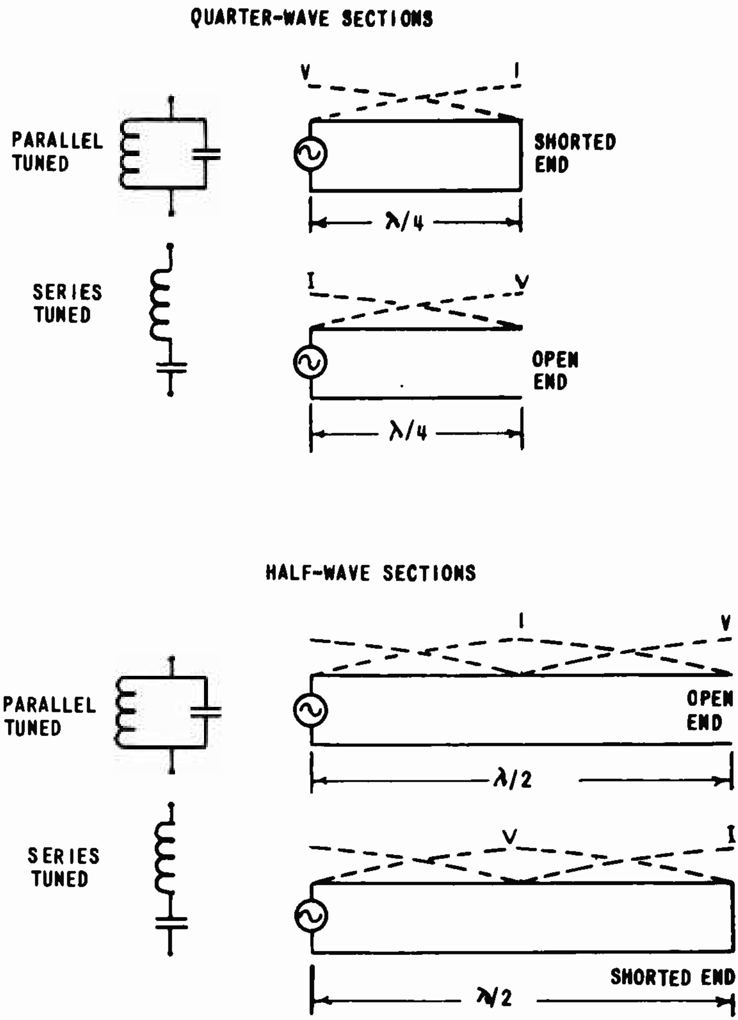
18. VSWR vs db variation.



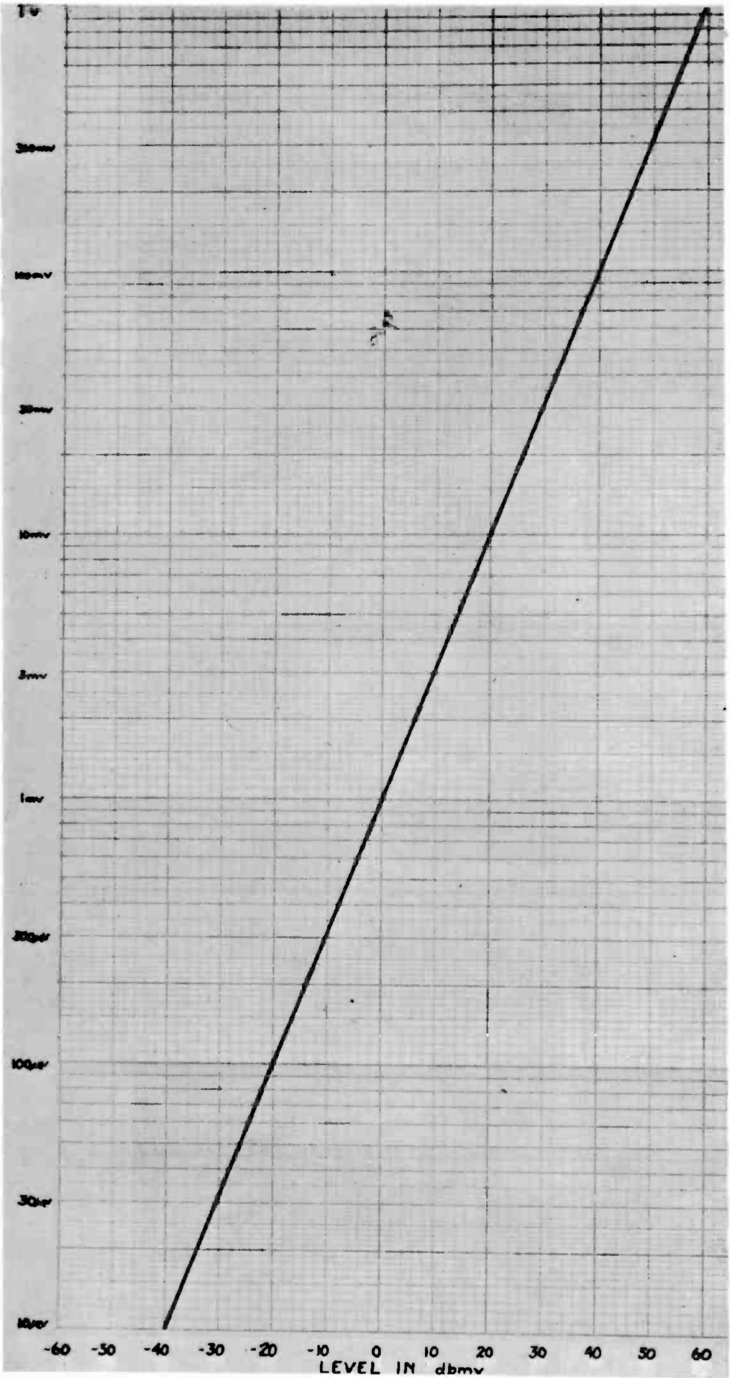
19. VSWR correction for 1.5 db of cable.



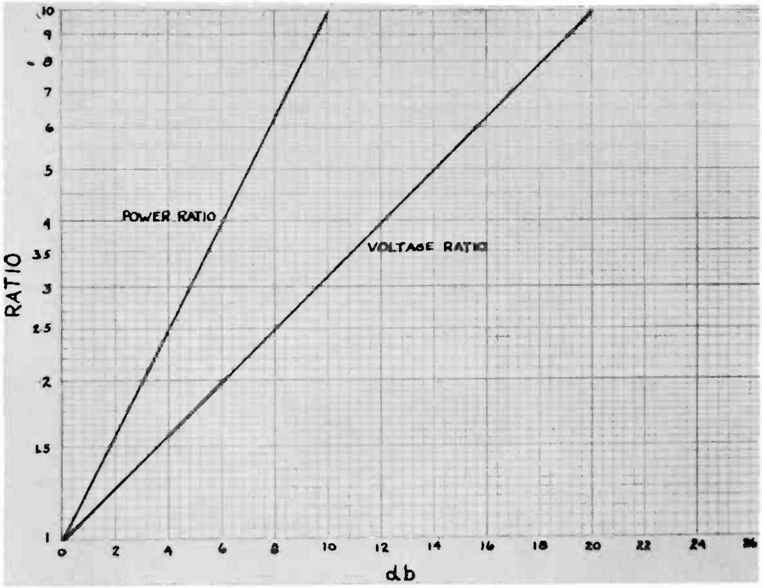
20. Return loss vs db variation.



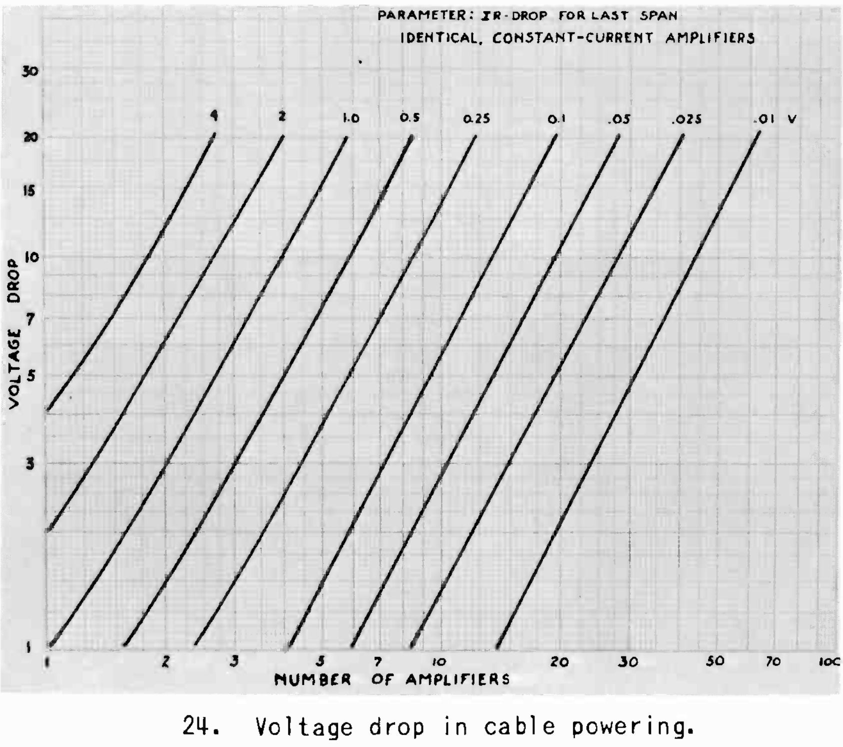
21. Tuned transmission lines.



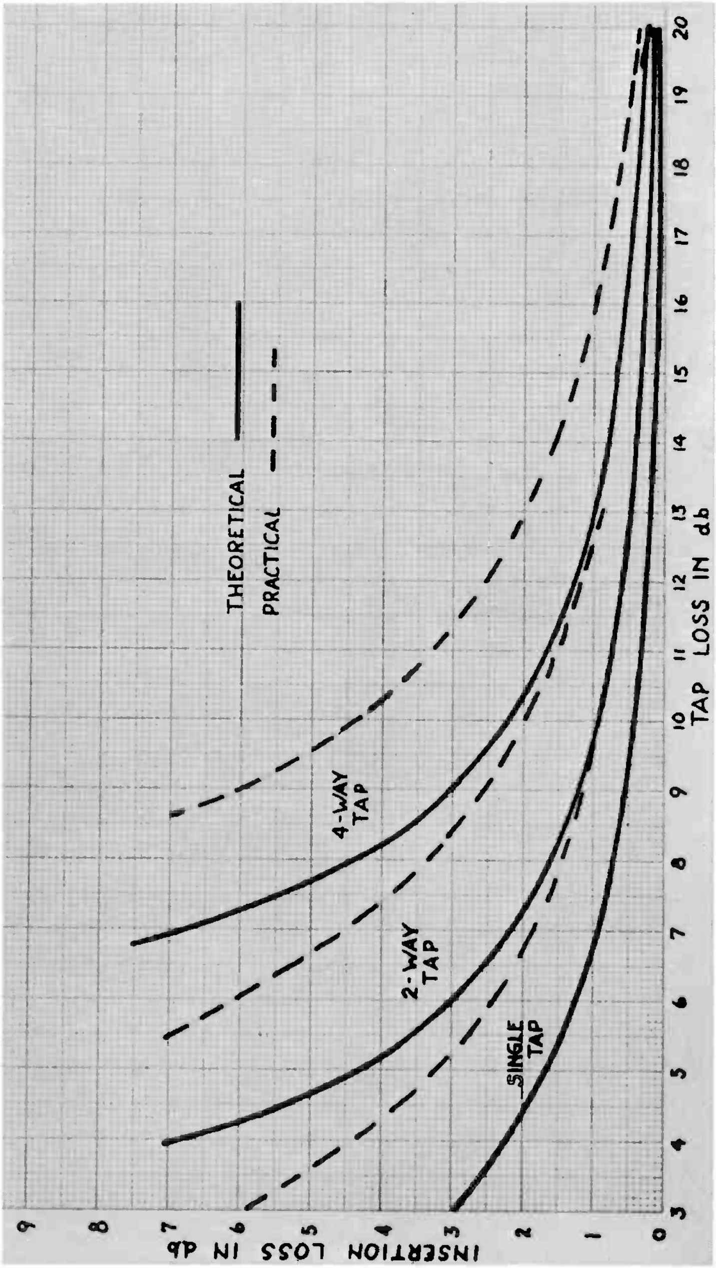
22. dbmv to voltage.



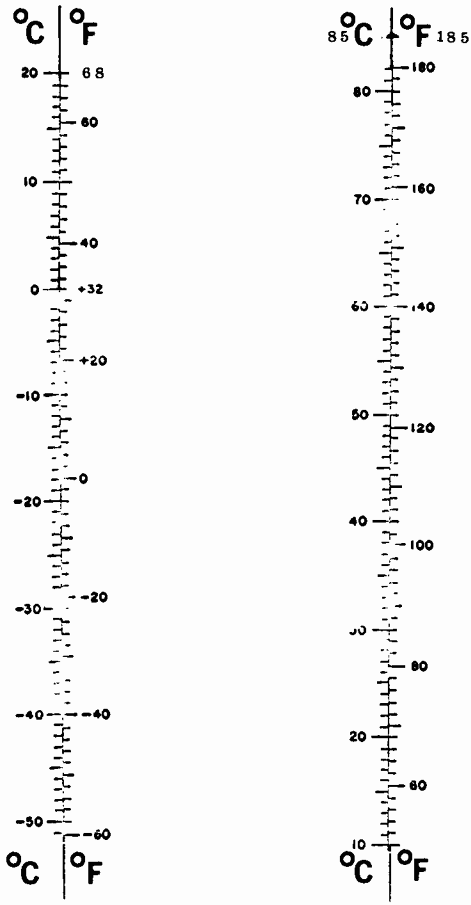
23. db to voltage or power.



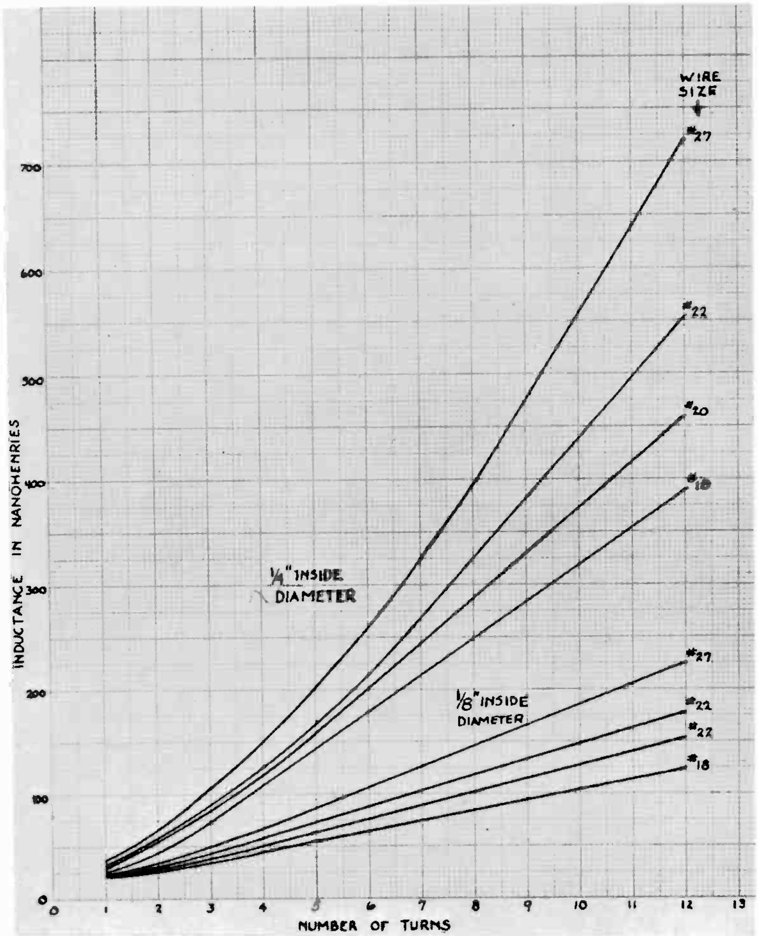
24. Voltage drop in cable powering.



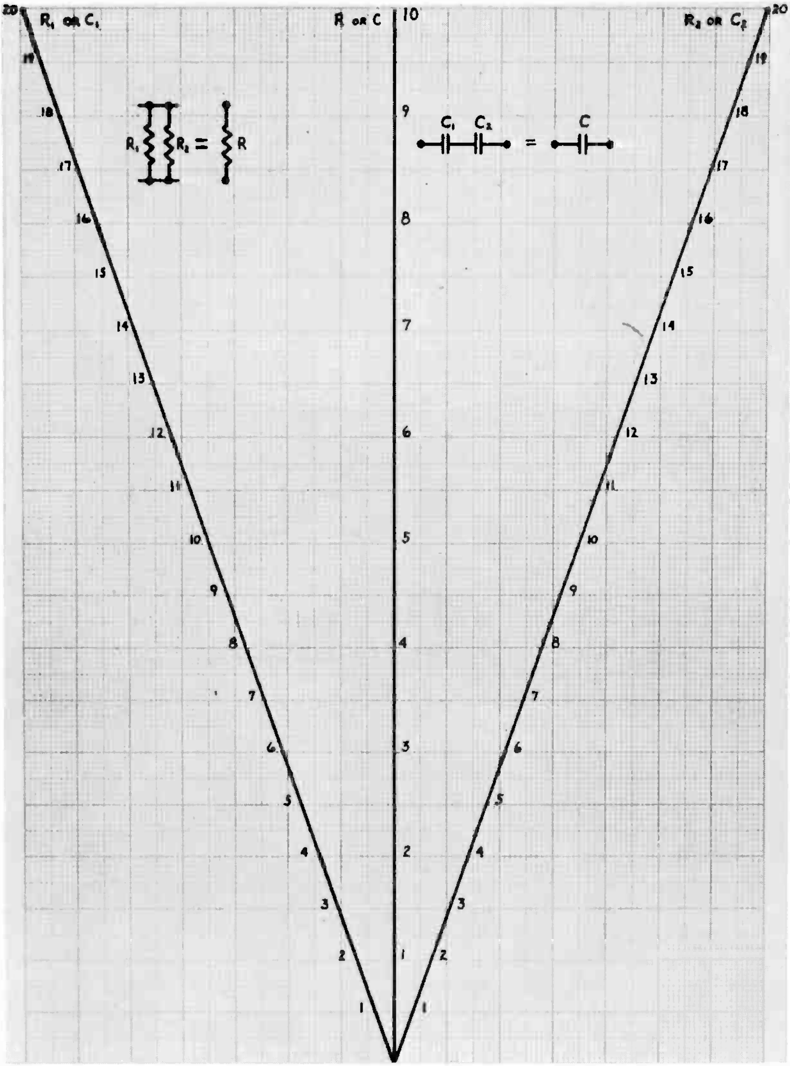
25. Insertion loss of taps.



26. Conversion from centigrade to Fahrenheit.



27. Solenoids.



28. Parallel resistors or series capacitors.

29. Operating levels in dbmv for different system tilts at 25 db spacing.

Channel	Picture Frequency	FULL-TILT Input	FULL-TILT Output	HALF-TILT Input	HALF-TILT Output	FLAT OUTPUTS Input	FLAT OUTPUTS Output
2	55.25	7.5	19.3	14.1	25.9	20.7	32.5
3	61.25	7.5	20.0	13.75	26.25	20.0	32.5
4	67.25	7.5	20.7	13.4	26.6	19.3	32.5
5	77.25	7.5	21.7	13.0	27.2	18.3	32.5
6	83.25	7.5	22.3	12.7	27.5	17.7	32.5
7	175.25	7.5	30.0	9.0	31.5	10.0	32.5
8	181.25	7.5	30.5	8.7	31.7	9.5	32.5
9	187.25	7.5	31.0	8.4	31.9	9.0	32.5
10	193.25	7.5	31.4	8.2	32.1	8.6	32.5
11	199.25	7.5	31.8	8.0	32.3	8.2	32.5
12	205.25	7.5	32.2	7.7	32.4	7.8	32.5
13	211.25	7.5	32.5	7.5	32.5	7.5	32.5

Note: These values give optimum system signal-to-noise ratio for a cascade of up to 50 amplifiers with an individual overload level of 50 dbmv, a noise figure of 10 db, and at a spacing of 25 db.

30. System levels in dbmv for integrated systems.

Channel	Maintrunk Input	Maintrunk Output	Distribution Input	Distribution Output
2	12	22.5	22	32.5
3	12	23.4	22	33.2
4	12	24.2	22	34.0
5	12	25.3	22	35.0
6	12	26.1	22	35.6
7	12	32.1	22	41.2
8	12	32.4	22	41.4
9	12	32.7	22	41.7
10	12	33.0	22	41.9
11	12	33.2	22	42.2
12	12	33.5	22	42.4
13	12	33.7	22	42.7
Pilot (220 MHz)	12	34.0	22	43.0
Spacing	1750 of 1/2"		1000' of .412"	
Tilt	Full Tilt		Full Tilt	

NOTE: This operation is used with automatic or manual maintrunks spaced at 22 db at 41°F and with high level distribution systems for average spacing of 1000 feet of .412 inch cable, with 3 db flat and 3 db equalized insertion loss. It applies to amplifiers with an overload level of 58 dbmv and a noise figure of 12 db in cascades up to 80 amplifiers with 6 db safety margin.

33. NORMALIZED LOSSES OF COAXIAL CATV CABLES

	FREQUENCY (MHz)	LOSS FACTOR (db)		FREQUENCY (MHz)	LOSS FACTOR (db)
	10	.20	Ch 7	175.25	.91
	20	.29	Ch 8	181.25	.925
	50	.47	Ch 9	187.25	.94
Ch 2	55.25	.49	Ch 10	193.25	.955
Ch 3	61.25	.52	Ch 11	199.25	.97
Ch 4	67.25	.545	Ch 12	205.25	.985
Ch 5	77.25	.59	Ch 13	211.25	1.00
Ch 6	83.25	.615		250	1.10
				300	1.21

To obtain precise cable loss, multiply actual loss at 211.25 MHz with loss factor. For other frequencies between 10 and 300 MHz normal interpolation may be used.

31. Transmission line data 75-ohm A/N standard RG/U cables.

Solid Polyethelene (Vel. Prop. 66.5%)

Type	Jacket O.D.	Dielectric O.D.	Center Conductor	Shield	Capacitance	Loss in db/100 ft		Weight
RG-No.	Inches	Inches	A.W.G.		pf/ft	100	200 MHz	lbs/100 ft
6 A/U	.332	.185	21	DBL	20	2.9	4.3	8.2
11 A/U	.405	.285	7/26	SGL	20.5	2.15	3.2	9.6
13 A/U	.425	.285	7/26	DBL	20.5	2.15	3.2	12.6
59 A/U	.242	.146	.023	SGL	20.5	4	5.7	3.2

Solid Telflon (Vel. Prop. 69.5%)

140 A/U	.233	.146	.025	SGL	21	4	5.7	4.5
144 A/U	.410	.285	7/.018	SGL	20.5	2.15	3.2	12.0
187 A/U	.110	.060	7/.004	SGL	20	5.9	9.7	1.8

Foamed Polyethelene (Vel. Prop. 78%)

(11 A/U)	.405	.285	14	SGL	17.3	1.5	2.3	9.0
(59 A/U)	.242	.107	20	SGL	17.3	2.7	4.0	3.0

300-ohm Twin Transmission Line

Type	Conductors	Vel. Prop.	Capacitance	Loss at 100 MHz db/100 ft	
	A.W.G.	%	pf/ft	dry	wet
Flat	7 x 25	80	4.6	1.5	7.6
Tubular	7 x 28	80	4.4	1.1	2.8
Open Wire	18	98	2.6	0.40	1.2

32. Typical characteristics of CATV cables.

Nominal Diameter Inches	Cable Type	Center Conductor Dia. in Mils	Dielectric	Dielectric Outside Dia. in Mils	Minimum Bending Radius	Outside Diameter Jacketed	DC-Loop Resistance Per 1000'	Loss at 213 MHz db Per 100'
.412	Solid Sheath Aluminum	76	FP	355	4	470	2.19	1.6
.500	"	98	FP	450	5	575	1.44	1.3
.750	"	143	FP	663	7.5	825	.66	.9
.412	Thin-Wall Aluminum	76	FP	350	2	465	3.36	1.6
.500	"	99	FP	450	2.5	565	2.28	1.3
.750	"	146	FP	675	3	810	1.31	.9
.412	Thin-Wall Copper	76	FP	350	2	465	3.23	1.5
.500	"	99	FP	450	2.5	565	2.17	1.2
.750	"	146	FP	675	3	810	1.24	.85
RG-59A/U	Copper Braid	32	FP	146	.25	242	6.15	4.2
RG-59/U	" "	25	SP	107	.25	242	6.45	5.7

CODE: FP - Foamed Polyethelene, Vel. of Prop = 81%

SP - Sclid Polyethelene, Vel. of Prop. = 66%

APPENDIX VII

Literature and References

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2.) L. E. Middlekamp, Reduction of Co-Channel Television Interference by Very Precise Offset Carrier Frequency, IRE Transactions on Broadcast Transmission Systems, December 1958.

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Transmission line data 237, 246