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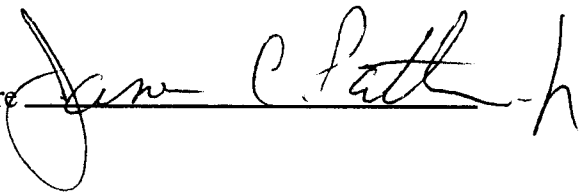
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DEVELOPMENT OF A PRACTICAL METHOD OF MEASURING THE MODULATION DEPTH OF THE NTSC TELEVISION VISUAL CARRIER

A Thesis

Presented for the

Master of Science

Degree

The University of Tennessee, Knoxville

James C. Patterson Jr

May 1990

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Thanks to all.

ABSTRACT

Cable television systems in the United States use tens-of-thousands of NTSC modulators to modulate television programming for use by their customers. Proper operation of these modulators requires careful adjustment of the controllable output parameters of these modulators. The output parameter discussed herein is the modulation depth of the visual carrier.

Until now, little effort has been spent on developing an easy method of determining the modulation depth. Research of modulation measurement methods reinforces this statement. Background material is covered and used to develop a measurement strategy. A circuit realization follows this development. A prototype of this circuit is constructed and tested. The test results are very encouraging. Although there are accuracy limitations in this particular circuit, the validity of the measurement method is reinforced.

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INTRODUCTION

Today in the United States, over 8,390 Cable Television (CATV) systems[1] are in operation, serving more than 50% of households having a television.[2] Almost all of these CATV systems use NTSC modulators to properly generate television signals from audio and video sources. These modulators are responsible for: assigning the signals to the proper frequency spectrum location, properly modulating the input signals, and providing proper output power levels. The latter two are easily adjusted on commercially available modulators, however, the proper adjustment of these parameters can be difficult. Power output level measurement is well-defined and easy to adjust. Modulation levels, however, are difficult to measure and adjust.

This thesis is concerned with the accurate measurement of the modulation depth of the visual carrier provided by these modulators. Present modulation depth measurement methods are difficult and expensive, so most adjustment is achieved through subjective trial-and-error. The goal of this thesis is to investigate the nature of this modulated signal to arrive at a solution to this measurement problem.

This is an opportune moment to explain the method this thesis will use to convey information. The audience of a thesis is generally given to be the author's peers. In the case of an engineering author, this audience would usually be other engineers. The peerage, or audience, for this work will range from college-degreed engineers to experience-trained technicians. This extreme diversity is a peculiarity of the CATV industry. With this in mind, every effort is made to convey clear, concise information to every reader at every level. It is hoped that any reader will realize some useful information for their efforts.

Several methods will be used to achieve this. Worded and graphical explanations of such things as Fourier analysis and modulation theory are used for the benefit of technicians that may not have a good theoretical background of these subjects. Likewise, considerable

effort is spent explaining the basics of television and Cable television for the benefit of engineers that are not directly involved with the CATV industry.

This presentation begins with Chapter 1 - Television. This chapter explains how television works. Chapter 2 is an explanation of CATV systems and should be helpful to anyone operating outside the CATV industry. Chapter 3 concerns itself with Amplitude Modulation theory. Chapter 4 discusses modulation measurement methods currently in use. Chapter 5 contains the development of a measurement method. Chapter 6 is concerned with a practical circuit realization and the performance results of this circuit.

CHAPTER 1

TELEVISION

The invention and success of television was the stimulus that brought about the need for Cable television systems. Chronologically speaking, since television was developed first, television should be discussed first. A historic perspective is necessary to fully explain the technical details of today's modern television systems.

The word "television" means: "to see far".[3] Television is a medium of broadcast consisting not only of sound, but also of "moving pictures" accompanying that sound. This was probably a very early definition of television used by the general public. In the early twentieth-century, cinemas or movies had grown in popularity and stature, having matured from a curiosity to almost a necessity, even though they were silent. Marconi's wireless radio invention was rapidly developing into the radio broadcast industry, which was becoming the premiere source of information and entertainment. Much of the newly developed technology of the radio industry was used to enable movies to have sound accompanying the pictures. The public was enthralled with this new development and seemed to be developing a desire for "technology" in the home.

This succession of developments was probably the catalyst for developing "radio with pictures" or television. As early as the late 1800's, electro-mechanical television was being demonstrated.[4] In the 1920's, Philo T. Farnsworth[5] was developing a design for all-electronic television which was quite a departure from the electro-mechanical design work done by others of the time. At almost the same instant, V.K. Zworykin[6] was working on basically the same design. Zworykin was more concerned in image capture, and Farnsworth was more interested in the presentation of the images. This parallel development of television caused a great stir ending in a court battle over the patent rights to television. To this day there

is little agreement as to which man is truly the "father of television".

Despite the legal problems, television began to develop as an experimental broadcast medium. Public demonstrations of television were given at the 1939 World's Fair[7], and development continued until the outbreak of World War II. Postwar development of television was accelerated by war-time technical achievements and an eager public. By the early 1940's, several experimental broadcast stations were in operation. The success of these experimental stations led the Federal Communications Commission (FCC) to attempt to standardize the transmitted television signal. This led to the formation of the National Television System Committee (NTSC).[8] Until this time, a variety of transmitting frequencies and modes had been in use. The NTSC was to create a standard for television broadcast that would make all television broadcasting compatible with all television receivers.

The following explanation of the NTSC standard is approached from a tutorial viewpoint. The NTSC was an ad-hoc committee with membership consisting of representatives of the major television broadcasters and television receiver manufacturers. As expected, each representative felt that their particular experimental methods were best. This meant that many compromises had to be reached. Even the FCC rejected the first set of standards, insisting on increased resolution. Eventually, the NTSC standards were adopted. A historically accurate account of the standardizing process would be awkward and cumbersome, and as such, not a good approach to explaining television. A more logical tutorial can be constructed by the fact that the "answers to the questions" are already known!

The first major standard set was that television imaging would be all-electronic and not an electro-mechanical hybrid. This choice being made, attention was turned to imaging in an all-electronic environment. A television picture tube with a phosphor coated screen has an electrically-steered beam of electrons that strike the phosphor. This process creates "light", wherever the beam is steered, by excitation of energy from the colliding electrons.[9] A "raster-scan" process was adopted by the NTSC for imaging on the picture tube. Figure

1-1 is a diagram of a simple picture tube. Note that the beam scans from the left to the right across the screen. When finished, it returns to the left, drops down a prescribed distance, and

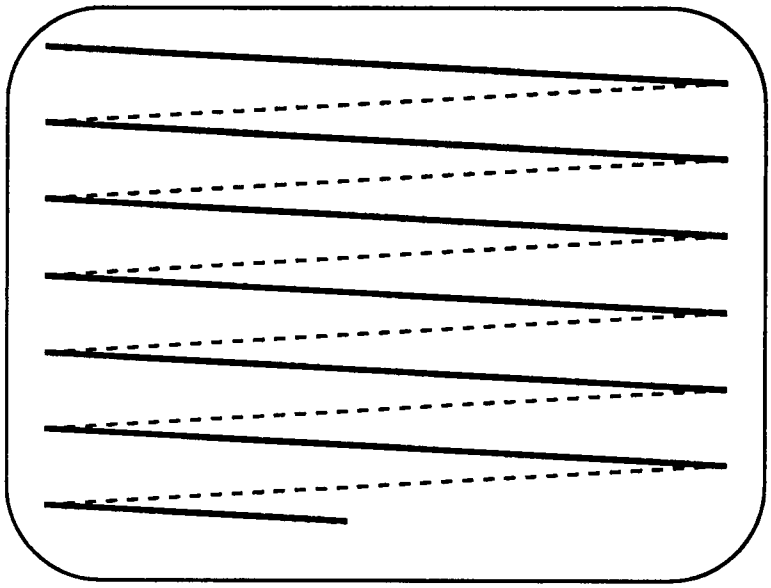


FIGURE 1-1

RASTER-SCANNING OF A PICTURE TUBE

then scans again. This action creates the "raster," a series of lines drawn across the screen at regular intervals. The scan actually slopes downward during each scan. Figure 1-1 has this slope greatly exaggerated for clarity. Varying the strength of the electron beam creates varying levels of light output from the screen.This phenonema is used to create black, white, and gray areas on the screen. By keeping the lines close together, an image can be formed by scanning the screen from top to bottom. The selection of raster-scan led to several technical specifications. Note that the raster-scan is of a periodic nature. The scan begins in the upper-left-hand corner and scans lines until the lower-right-hand corner is reached. The

scan then begins again in the upper-left-hand corner and starts the process over. This will occur at regular intervals whether or not the image changes and even if an image is not present.

A single, full screen scan will create a temporary image on the screen. This image is much like a single frame of a movie. By successively replacing the images, a sense of motion can be created. A single, full screen scan is also referred to as a "frame." If these frames are replaced at a rate of about sixty per second or faster, the human eye will not perceive the "flicker" caused by the frame replacement and the image will appear flicker-free , insuring that motion will be natural. Unfortunately, 1/60 of a second for a full-screen scan with enough lines for good resolution was a technically unfeasible proposition to the NTSC due to the large bandwidth requirements. Motion pictures change frames at 24 frames-per-second (fps) and use a shutter to simulate 48 fps to create a flicker-free image by taking advantage of the persistence of human vision.[10] The NTSC decided to slow the frame rate to 1/30 of a second. To avoid the flicker induced when a complete frame is replaced at this slower rate, the frame is subdivided into two fields with the new field scan lines falling between the lines of the previous field. This creates a gradual change of image from one frame to another. This means that the electron beam goes through the top-to-bottom scanning process twice to create a single frame. The first scan begins at the first line, line 1, and then proceeds to scan every other line (3,5,7,...), which gives the name the odd field. Upon completion of the odd field, the scan begins again with line 2 and proceeds every-other line (2,4,6,...) until completed. By using even numbers, the even field name is obvious. (It should be noted that some texts will refer to the fields in a slightly different manner. Lines are numbered consecutively from 1 to 525, with lines 1 to 262½ in field 1 and lines 262½ to 525 in field 2.) Figure 1-2 graphically represents this "interlaced" method of image production. If necessary, the 30 Hz field rate could also be derived from the 60 Hz AC electrical mains, but primarily a 30 Hz rate reduces the distracting portion of the interference induced by the 60 Hz AC main.

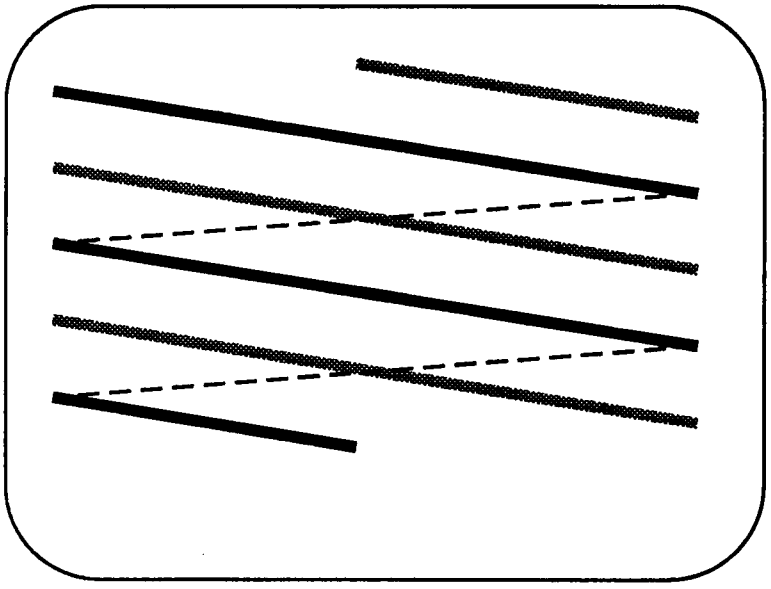


FIGURE 1-2

INTERLACED RASTER-SCANNING OF A TELEVISION PICTURE TUBE

Now with a frame rate of 30 Hz, how many lines should be in a frame? The NTSC decided that a frame should have 525 horizontal lines. This would give ample resolution and conserve bandwidth in the transmission of the television signal. The 525-line frame, when changing at a rate of 30 times a second, gives the line rate of 15,750 Hz. (or 1 line every 63.5 micro-seconds).

We now have frames, fields, and lines. The next step is to integrate them into a signal that a receiver can decipher. A method had to be found to synchronize the incoming information and the raster-scanning process. Figure 1-3 shows the visual information of a single line. This signal would directly control the intensity of the electron beam in the picture tube to control the black, white, and gray content of the line. A method had to be decided on to inform the receiver of the beginning of this line. The method chosen was the horizontal

synchronizing (sync) pulse. Figure 1-4 shows several lines of a field, each with it's own horizontal sync pulse.

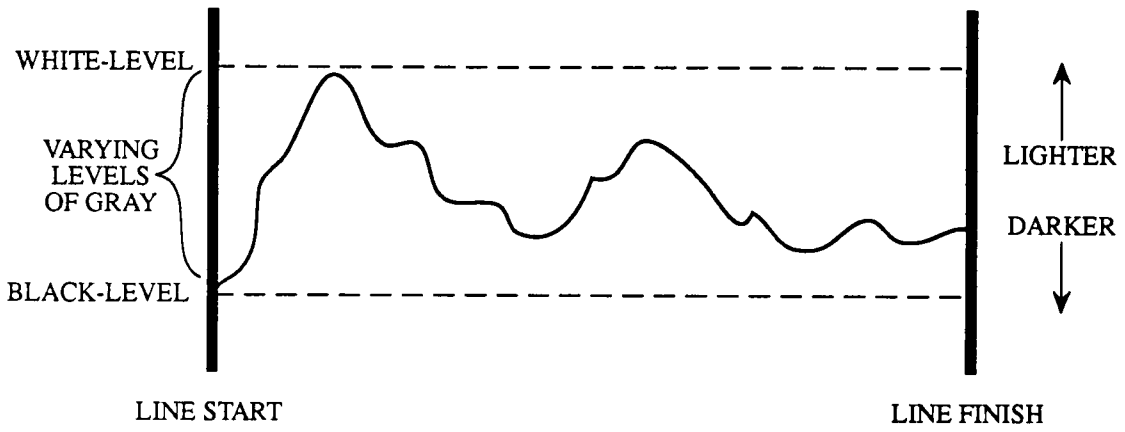


FIGURE 1-3

A SINGLE LINE OF VIDEO INFORMATION

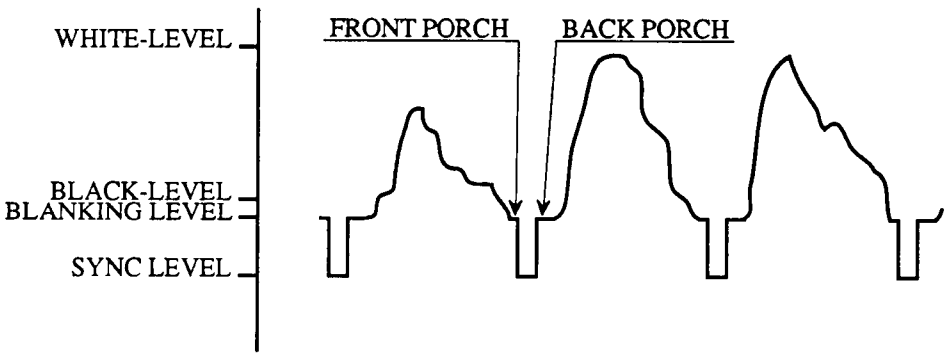


FIGURE 1-4

THE RELATIONSHIPS BETWEEN VIDEO AND HORIZONTAL SYNC PULSES

There are several important items shown by Figure 1-4. Note that the sync pulse is below the black-level. Not only does the line trace from left-to-right, but it must also return to the left to begin a new line. If the electron gun is not turned off, this return (retrace) will over-write the line just scanned. By keeping all sync pulses below the black-level, they will at no time be visible in the image. The area below the black-level is usually referred to as the blanking level. Blanking level is at the black level, or slightly below black level, which insures blanking. The constant amplitudes just preceding and following the sync pulse are referred to as the front porch and back porch, respectively. These are included as references of black-level for the receiver and to insure the sync pulse is blanked when it occurs. With a system for informing the receiver of the beginning of each horizontal line now standardized, a method of informing the receiver of exactly which line is being received must be selected.

Vertical sync pulses were standardized to "place mark" the beginning of each field. Vertical sync pulses are broad pulses when compared to horizontal sync pulses. This allows the two pulses to be distinguished by the receiver. Figure 1-5 is a simplified diagram of a vertical sync pulse for odd and even fields.

The receiver continuously integrates the incoming signal. When a vertical sync pulse is encountered, the value of the integrand is substantially larger than the integrand of a horizontal sync pulse because of the differences in the areas. Likewise, the receiver uses a differentiator to determine the presence of horizontal sync pulses. The sharp leading edge of the sync pulse causes a large impulse output from the differentiator. This is then used to synchronize the receiver's horizontal oscillator circuitry. Note that the leading edge of a vertical sync pulse is always a distance "H" from the last horizontal sync pulse. The distance "H" is the line period, about 63.5 micro-seconds. This helps keep the horizontal oscillator synchronized and running, as most are free-running oscillators. This free-running oscillator design is inexpensive and accurate, but it must be constantly resynchronized for proper operation. To provide continued oscillator synchronizing pulses, serrations are "cut" into the

vertical sync pulse every H distance from the vertical sync pulse leading edge. This creates the leading edge necessary for the differentiator to generate horizontal synchronizing information , but the small serrations do cause integrator errors.

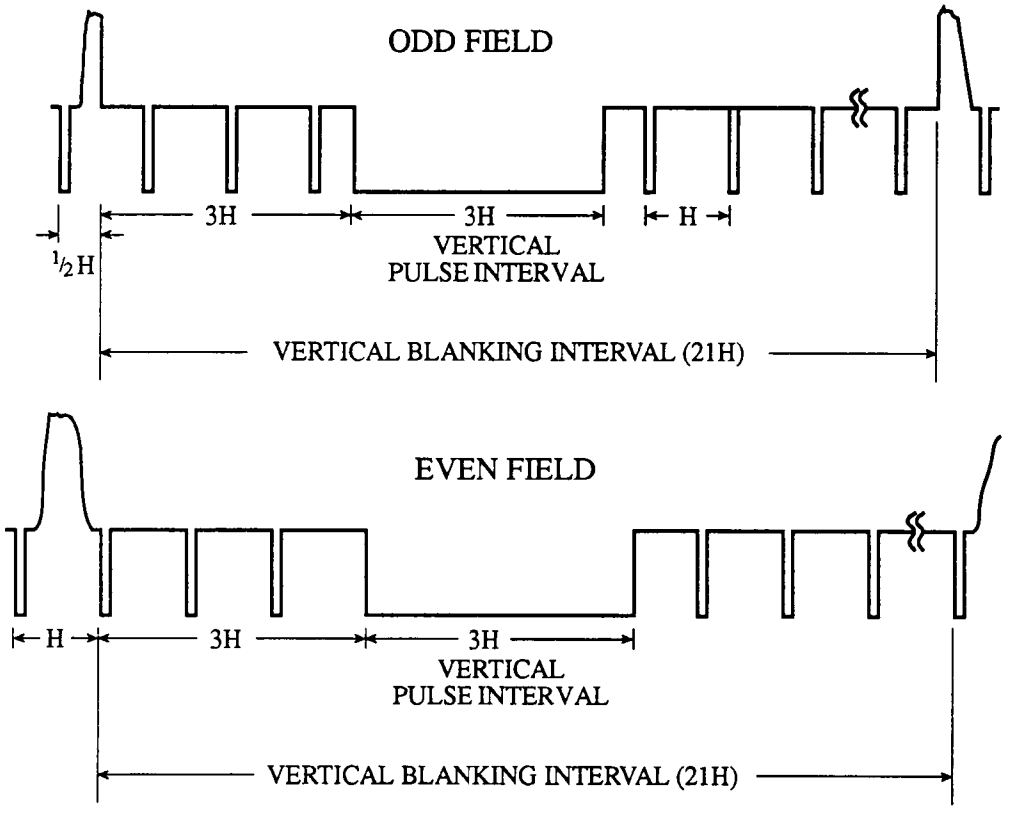


FIGURE 1-5

THE VERTICAL INTERVALS OF THE EVEN AND ODD FIELDS

A question frequently arising from this explanation of vertical sync pulses is: How does the receiver recognize which is the odd field and which is the even field? The answer is it can not tell the difference! All vertical sync pulses are treated in the same manner.

Examining Figure 1-5, from left to right, the receiver encounters the first horizontal sync pulse in the vertical blanking interval, then another, and another. Finally, the vertical sync pulse is encountered, the integrator charges, and upon reaching a predetermined charge level the receiver directs the electron beam towards the upper-left-hand corner from it's present location at the bottom of the picture tube. This journey cannot be instantaneous, and the serrations are keeping the horizontal oscillator synchronized and scanning in a normal fashion. Because of this, the beam starts upward, encounters a horizontal sync pulse, scans a blank line, and continues on up at the end of this line until another horizontal sync pulse is encountered. This continues 20 times until the beam reaches it's upper-left-hand position. The scanning continues from this point in a normal but blanked fashion until reaching the first line with image information. Horizontal scanning continues as normal until the receiver encounters the odd field vertical sync pulse. As with the even field, blanking occurs for 3 horizontal lines and then the vertical sync pulse is encountered. Note that this vertical sync pulse occurs half-way through the horizontal scan. Since this leading edge is between a horizontal line, it does not interfere with horizontal oscillator synchronization, but a serration in the vertical sync pulse must follow at the proper timing interval to maintain horizontal synchronization. Again the electron beam is moved upwards, but this time from the middle of the last line. This will cause the beam to start at the top of the screen, but in the middle not the left-hand corner. Horizontal scanning continues, but the first line encountered with active information starts mid-way through the scan. In both fields the electron beam begins at the same vertical position, with the odd field beginning in the center of the screen. This generates the interlacing action in an accurate and simple fashion. An undesirable effect of the vertical sync pulse of the two fields not being exactly the same is that this introduces error into the integration process. Since the odd vertical sync pulse follows a horizontal sync pulse by only 0.5H, it will integrate slightly faster than the even vertical sync pulse. This problem is overcome by adding equalizing pulses before (pre-equalizing) and after (post-equalizing)

the vertical sync pulse. The equalizing pulses and the serrations of the vertical sync pulse significantly complicate the vertical interval and this is illustrated in Figure 1-6

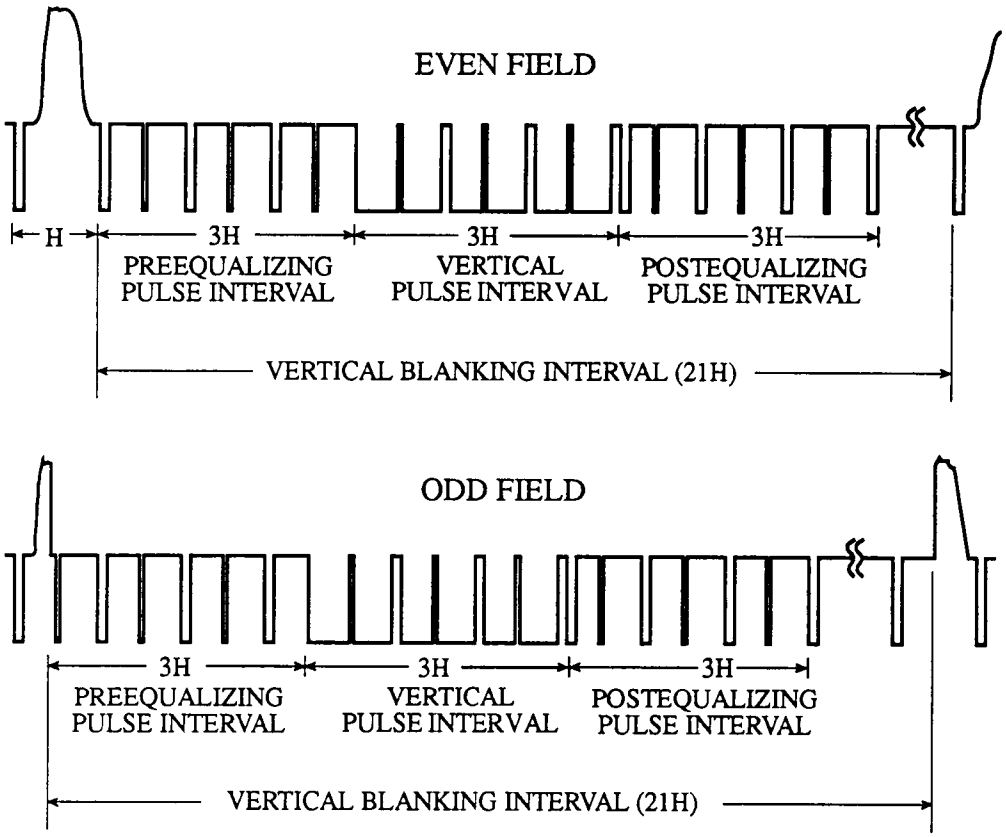


FIGURE 1-6

VERTICAL INTERVALS COMPLETE WITH EQUALIZING PULSES

Video, or more accurately composite video, is the signal that contains the image and timing information. In NTSC standard, video occupies a frequency space from just above DC to approximately 4.2 MHz.[11] This low frequency, high bandwidth signal cannot efficiently be broadcasted, so another method of transmission was needed. Amplitude

modulation (AM) was being used in standard AM radio broadcasting. AM uses a carrier frequency that is much higher than the signal frequency. The signal is encoded onto the carrier so that it appears to "ride" the carrier. This carrier could be, for example, 50 MHz. At this frequency, broadcast is more reliable and practical and has a more predictable range. Standard double-sideband AM video would use about 10 MHz of frequency spectrum, allowing broadcasters to be "channelized," or separated in frequency every 10 MHz.[12] A bandwidth saving technique, vestigial sideband, would allow the channel to occupy less frequency spectrum space, approximately 5 MHz. This would allow more channels to exist in less frequency space.

Other modes of transmission included frequency modulation (FM). FM was a modulation method that was receiving a great deal of attention at the time of the NTSC. FM has the distinction of having greater noise immunity than does AM. However, FM requires significant bandwidth for transmission. FM television would require about 30 MHz of frequency spectrum,[13] and given the transmitting technology of the NTSC era, this would have left room for only one broadcast channel. This led to the adoption of vestigial side-band AM for the transmitted video. FM was chosen as the method of transmitting the accompanying audio programming.

With transmission modes set, the FCC allotted blocks of the VHF spectrum for television use exclusively. The blocks were chosen to be 6 MHz wide to allow sufficient space between adjacent channels for efficient receiver discrimination. These blocks were numbered consecutively, starting with 1, for the convenience of the consumer. Yes, at the beginning there was a channel 1, and early receivers have the channel 1 tuner position. Channel 1 was lost in 1948 to land-mobile use, and the number 1 was omitted from subsequent receivers. In the years since this first allocation, more and more frequency spectrum has been devoted to television. Figure 1-7 is a diagram of the present-day television

frequency spectrum allocation.[14] An important note is that this allocation is for the exclusive use of television, and is not shared with any other service.

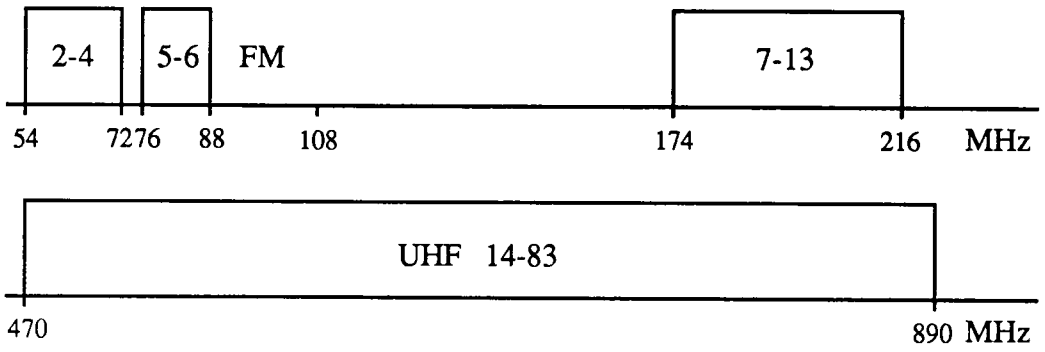


FIGURE 1-7

BROADCAST TELEVISION CHANNEL ALLOCATIONS

At this point, a complete, all-electronic, well-defined monochromatic television system has been described and standardized. The NTSC proceedings filled hundreds of pages of text, so this explanation has taken literary license to abridge this information.

Two major changes have occurred to the standard since it's initial adoption by the FCC. The first was the advent of color television. The only requirement the FCC required to be met was that the color television standard be compatible with all existing monochromatic receivers. The chosen standard met the compatibility criterion with some minute changes to the original monochromatic standard. To reduce the potential for interference, the horizontal scanning frequency was reduced from 15750 Hz to 15734.26 Hz; this 0.1% change was still acceptable to monochromatic receivers. Similarly, the phase modulated color (chroma) information was placed on a subcarrier at 3.579545 MHz. All color and monochromatic receivers were now compatible with all television transmissions. The

second major change was the inclusion of test signals in the vertical interval. The vertical interval test signals (VITS) gave broadcasters the ability of continuously monitoring and testing their transmissions without interrupting programming.

These developments set the stage for explosive growth in the television marketplace. Televisions were becoming more and more commonplace, but now other problems were beginning to surface. Television stations were few in number and reception of their signals could be a problem. This problem would lead to the creation of a billion dollar industry: cable television.

CHAPTER 2

CABLE TELEVISION BASICS

Modulation measurement is the previously stated problem, but exactly how and where is it used? This section and subsequent sections will be used to answer this question. This particular application is in the area of Cable Television (CATV), so some basic information and definitions of CATV are necessary. If the reader is a CATV professional, this section may be skipped.

What is a Cable System? The answers to this question are as diverse as the personalities of the people attempting to answer this question. Community Antenna Television (hence the CATV abbreviation) was the most common name given to this particular technology in its early days. This soon proved too long and cumbersome and was renamed. These names range from the generic "Cable System" to the all too cute "Cablevision". Regardless of the name that the reader is most familiar with, it is all basically the same from a technological standpoint. Community Antenna Television is probably the most descriptive of all the names. At its inception, CATV was used to provide a community with difficult to receive off-the-air television programming. The era was post World-War Two and war surplus provided the first coaxial cable used in CATV. These vintage CATV systems consisted of an off-the-air television reception site, or headend, where the television signals were intercepted. From the headend, these channels were amplified, filtered, mixed with other channels, and sent down a transmission line. This transmission line was usually the previously mentioned coaxial cable, but occasionally was of the open-line (ladder-line) type. As the television signals propagated down the transmission line, losses in the line decreased (attenuated) the signals. If the transmission distances were large, amplifiers were inserted at periodic intervals to keep the signal at reasonable levels. While all this was

occurring down the line, the line was also being "tapped" into to provide customers with these signals. These taps were usually crudely constructed affairs, but their function was to minimize the interference between the tap and the main signal line. From this tap, a signal was fed down a transmission line to the customer's home. In the home, a matching device was used to connect the transmission line to the television. These early CATV systems were "low-band" systems only capable of carrying channels 2 through 6. Some early systems carried only one channel, as television broadcast was also in it's early days of growth and few television stations existed.

CATV has been in existence for nearly forty years now, but just recently could it be considered a maturing technology. This maturing now allows a CATV system to be discussed in terms of standards that are adhered-to by most of the CATV industry. In its infancy, CATV was invented and implemented by people best described as "Thomas Edison-like" individuals. Most CATV systems were small, and the same person would serve as engineer, technician, lineman, contractor, secretary, manager, politician, and any other job that might need to be performed. There was a great interest in electronics in this post-war era. Large numbers of war veterans had their first exposure to electronics in the military, and this led to the post-war flourishing of amateur radio and electronics-based magazines. This was the profile of the early CATV engineer and is still representative of the group.

Modern CATV systems have sprung from these humble beginnings. CATV standards have been formed from technological changes, customer preferences, and outside forces such as the price of copper. The following is an attempt to acquaint the reader with the modern CATV system from the generic standpoint. Most of this information will be relevant to any modern cable system, with only small differences from system to system. Figure 2-1 shows a portion of a CATV system layout using industry standard symbols.[15] The major areas of interest are the Headend, Trunk line, Feeder line, and Home terminal. Each of these areas will be addressed individually.

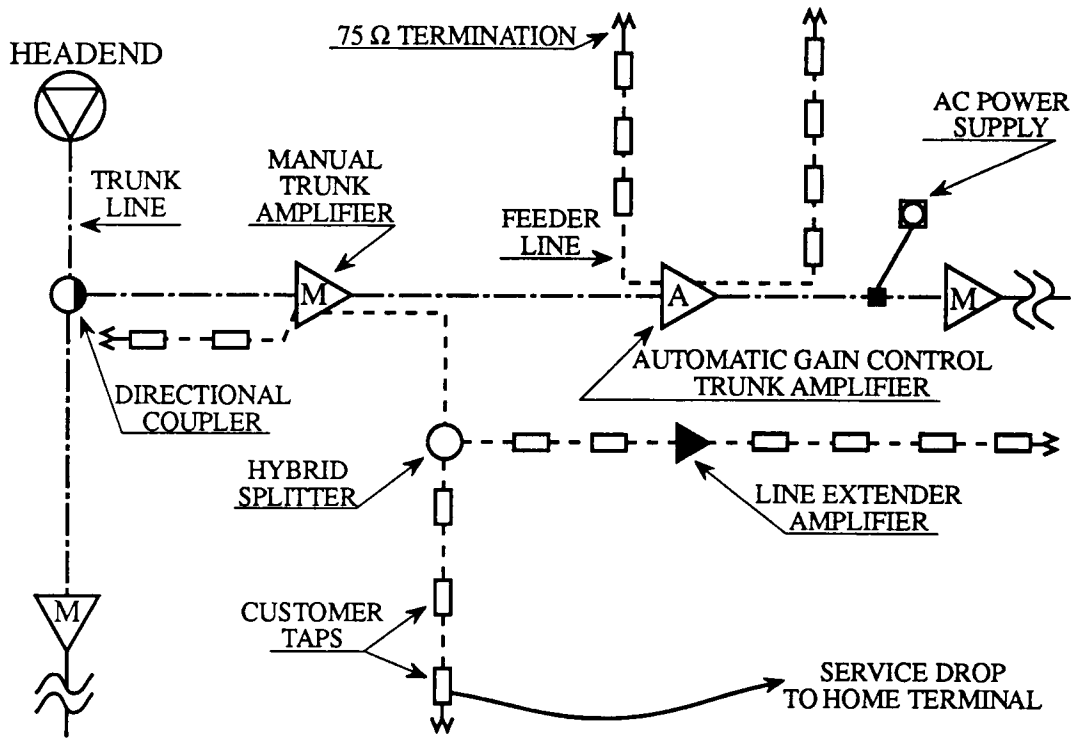


FIGURE 2-1

A GENERIC CATV SYSTEM LAYOUT DIAGRAM

The modern headend has the same mission as did its early predecessor. The headend is the source of all signals that are present on the CATV system. These signals are available from a variety of sources: off-air, microwave, satellite, and locally originated. Off-air signals are the conventionally received off-the-air television broadcast stations. These signals are usually received by the use of conventional television reception antennas, although some CATV antennas can be unwieldy clusters of up to sixteen antennas. Weak signals will be amplified at the antenna (pre-amplified). The antenna signal is then fed into the headend where processing occurs. Signal processing takes many forms. It can be as simple as filtering and amplification, or as complicated as demodulation-processing-modulation. Most modern CATV systems use a processing method that falls between these extremes.

Figure 2-2 is a diagram of the modern CATV channel processor. The antenna signal is fed into the processor input section. This input section usually has a low-noise "front-end" section, and light filtering to reduce unwanted signals. This signal is now fed into the down-

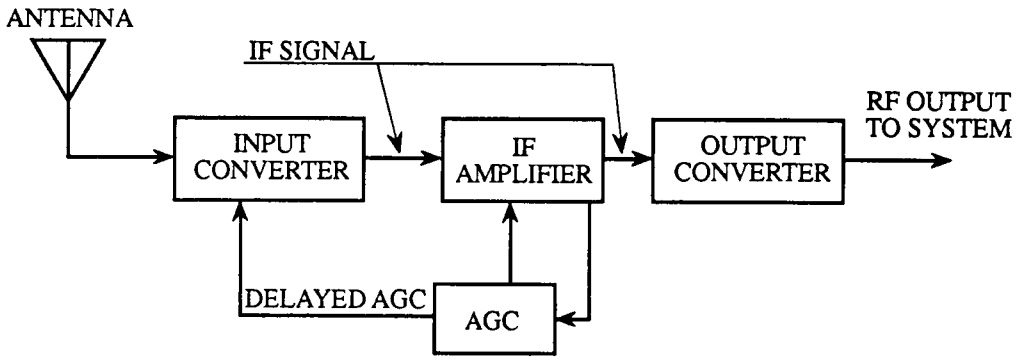


FIGURE 2-2

A BLOCK DIAGRAM OF A CATV CHANNEL PROCESSOR

converter section where it is heterodyne-converted to an intermediate frequency (IF). Most often this frequency is the "standard" IF used by the television receiver industry; 45.75 MHz for the visual carrier and 41.25 MHz for the aural carrier. This particular frequency selection allows the use of componentry used by the television receiver industry. The IF is where most of the filtering and amplification is done. The most unwanted signal of all is that of an adjacent television station. At the high frequencies of television broadcast, sharp, high-Q, stable filtering is very difficult to achieve and each channel of interest uses a custom filter adjusted to that particular channel. Filtering at the lower IF frequencies is much easier to achieve, and has the added benefit of economy. IF filters will work with any channel that is down-converted, so they are not individually tuned to the particular broadcast channel, but

are instead tuned to the IF. A sample of the IF is used by the automatic gain control (AGC) section to control the output of the processor. The AGC keeps the output level constant even though the input level may vary. The delayed AGC control to the input section prevents "front-end" overload, but is inactive with low input signals. This helps insure low noise operation over a broad range of input signal levels. The amplified and filtered IF signal is now fed to the up-converter section. The up-converter heterodyne-converts the IF up in frequency to a channel assignment. Note that this channel assignment need not be the same as the input channel assignment and by proper up-converter selection virtually any channel assignment is possible. This is very important when processing UHF signals that are normally outside the pass-band of the CATV system or when the off-air signal is so strong that it will interfere with itself in the CATV system. Following the up-converter is the output section. This section amplifies the output of the up-converter and usually does some light RF filtering. The output of this processor is then fed to a mixer, which is usually a passive device, and then on to a post-amplifier if needed.

Occasionally an off-air signal is desired, but due to distance, natural obstruction, or some other problem, reception is impossible. Microwave transmission can be the answer to this problem. Figure 2-3 is of a typical microwave distribution system. The off-air signal is collected much in the same way as it is in the headend site and is processed to the IF stage. From IF, the signal is demodulated, filtered and amplified. This demodulated signal, video and audio, is used to excite the microwave transmitter. Microwave transmission is usually in the range of 1000 MHz (1 GHz) to 12,000 MHz (12 GHz). Frequency-modulation is used where long-distance and/or high quality are necessary, and less costly amplitude-modulation is used otherwise. The microwave signal is transmitted from a highly-directive antenna to a reception site at the headend where the microwave signal is down-converted, demodulated, and modulated to a channel assignment frequency. If the distance from the hub-site to the head-end is too great, intermediate reception-amplification-retransmission sites may be

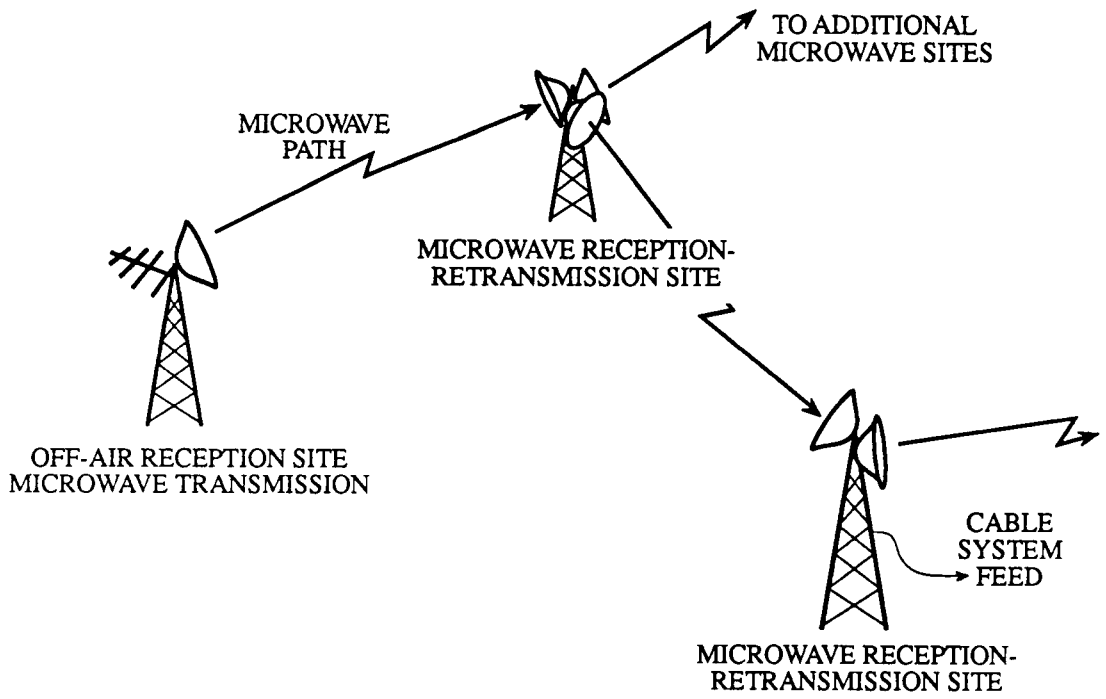


FIGURE 2-3

A DIAGRAM OF MICROWAVE LINKS

required to amplify the signal. When this situation arises, CATV systems will sometimes band together to provide each other with service. This allows the closest system to receive the microwave signal, pick-off some for it's own use, and retransmit the signal on to the next system. Several regional links of this nature have been used and interestingly Home Box Office, CATV's first premium movie service, started as a regionally microwaved service. Microwave systems are very expensive, and this is the reason they are not in wide use.

Home Box Office was previously mentioned as the first CATV regional microwave service. In 1975, the technology was in place for this service to begin satellite delivery of its programming. Although once very expensive, low-cost satellite reception equipment is now commonplace. Satellite technology has been responsible for a decline in the use of

microwave systems and distant off-air reception. Satellite reception is very much the same as microwave reception in frequency and modulation. An audio and video signal are FM modulated onto a microwave carrier and directed toward a satellite in geosynchronous orbit about the earth. The satellite receives this signal, converts the frequency, and then retransmits this new frequency back toward earth. The earth terminal is responsible for receiving this signal. A highly-directive antenna (usually referred to as a dish) is used to collect the very weak signals of the satellite. A special, very low-noise pre-amplifier amplifies this signal which in turn is fed into the receiver. The receiver demodulates this microwave signal and outputs both video and audio. The video and audio must now be modulated to the desired channel frequency assignment. Satellite television reception has proved to be a high-quality means of receiving new and innovative programming for use in the CATV system.

The term "local origination" is almost self-explanatory; it is programming that is originated by the CATV system. This programming can be videotaped events, live action, or computer generated displays of such things as the weather or stock market reports. All local origination programming is ultimately provided to the headend in the form of audio and video. The headend modulates this signal to the desired channel frequency assignment for use by the system.

Microwave reception, satellite reception, and local origination have a common element that is the focal point of this thesis: modulators. The audio and video signals output from these signal sources must be modulated to a channel frequency assignment. This requires more than a heterodyne frequency conversion. North American television is generated and transmitted to follow the NTSC standards. This allows all televisions to receive all broadcasts. The modulator must follow this standard. Full explanations of this standard is in Chapter 1, but one point that is important is modulation adjustment. The output of the headend is television signals. These signals have levels associated with them and there exists instrumentation with which to measure and set these levels. Modulation adjustment,

however, has no such instrumentation and is usually adjusted by "eye." This points out the need for this type of instrumentation.

Figure 2-4 is a simplified block diagram of a CATV modulator. The video signal from any of the previously discussed sources is applied to a video amplifier section. This buffered video is then applied to the video modulator. A diode-ring modulator is most commonly employed in this section, with the video input modulating the injected 45.75 Mhz carrier. Controlling the video input level determines the extent, or depth, of this modulation. The video modulated IF signal is now applied to the vestigial sideband filter (VSB), which is today most commonly a surface acoustic wave (SAW) filter. From here, the VSB visual IF signal is sent to the IF amplifier.

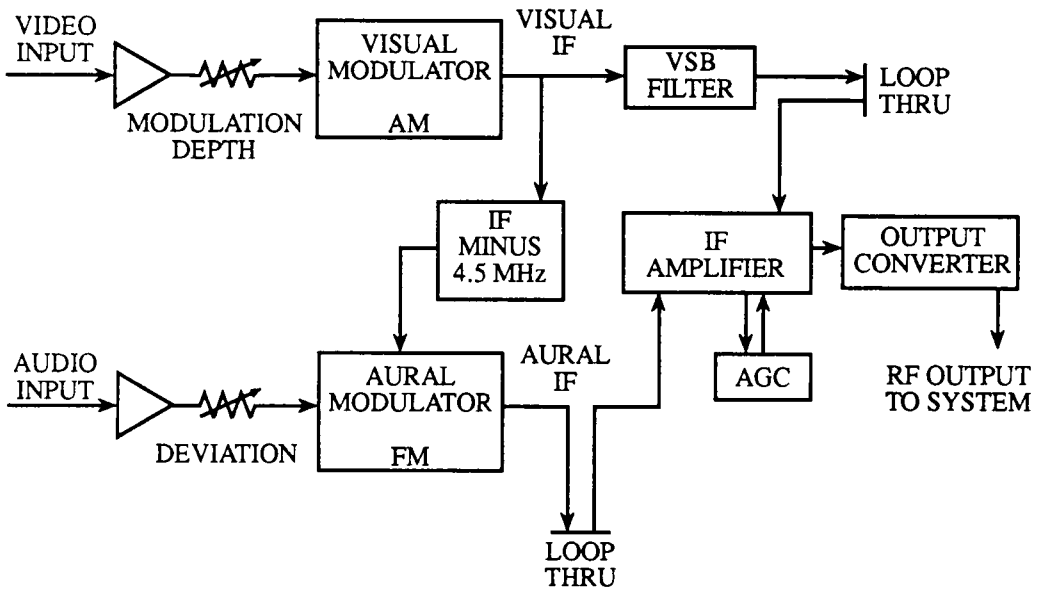


FIGURE 2-4

A BLOCK DIAGRAM OF A CATV MODULATOR

In a parallel fashion, the audio input is applied to an audio buffer before being injected into the FM modulator. The FM modulator's carrier is phase-locked 4.5 MHz below the visual IF carrier. The extent, or deviation, of this frequency modulation is usually controlled by the level of the audio signal being injected into the aural modulator. After bandpass filtering, the aural IF is applied to the IF amplifier, where it is combined with the visual IF as a subcarrier.

At this point, the signal is identical to the IF of the off-air processor. The IF signal is amplified, with the AGC ensuring the output level is constant. From IF amplifying, the signal is applied to the output converter, where it is up-converted to the proper channel assignment, RF amplified, and lightly filtered. The output of the modulator is then fed to a mixer which is usually a passive device, and then on to a post-amplifier if needed. These last steps are identical to the last steps in a processor. Most manufacturers will use the same output converter and portions of the IF amplifiers in the manufacture of both processors and modulators.

The variety of signals that have been mixed in the headend are now ready for transport by the transmission line called the trunk line. The trunk line is a 75-ohm, coaxial cable that usually has a solid outer sheath. The trunk line can be considered to be the "backbone" of the cable system. The trunk carries the signal from the headend into the areas to be served. The trunk line is never tapped to serve a customer. The signal levels are kept at very distinct, critical levels. It is this care that preserves the quality of the signals being transported by the trunk.

As with all things in nature, a compromise must be made in the design of CATV systems. The limiting factors in CATV system design is the trade-off between noise and cross-modulation distortions.[16] At the headend, each processor and modulator puts out a television channel, usually with a very good signal-to-noise ratio and a relatively high output power level - although this level is rarely above one volt peak-to-peak across 75-ohms. From

the individual processors and modulators, the individual channels are mixed and sent on to the trunk lines. As these signals travel through the trunk line they are attenuated. When the signal level drops too low, the signal-to-noise ratio is adversely affected, so some amplification is needed. It is this first amplifier, and all subsequent amplifiers, which create the need to compromise. The signal input to the amplifier has a very low amplitude (about 3200 microvolts), but this amplitude represents as many as 60 channels occupying more than 400 MHz of frequency spectrum. Each channel represents a very complex collection of AM, FM, and PM signals, with very little frequency spectrum separating them. This makes the amplifier very prone to something collectively referred to as cross-modulation distortion.

Cross-modulation distortion is the CATV "catch-all" phrase for a variety of distortions, with the worst probably being inter-modulation distortion. In this, the channels will "beat" together to form additional signals which are algebraic sums and differences of the original channels. This is an unfortunate effect of all amplifying methods, and is more of a problem in broadband or high-gain situations. At low gains, cross-modulation distortion is usually very low and tolerable. As gain increases, so does cross-modulation distortion. The cross-modulation distortion increases at a rate which is greater than the gain increase, which creates a practical limit to the gain and output level of this, or any, amplifier. Remembering that there is a signal-to-noise problem associated with low signals, there exists practical high and low signal levels for good performance.

The trade-off is as follows: high signal levels have good signal-to-noise performance, but also have high levels of cross-modulation distortion, whereas low signal levels have good cross-modulation distortion performance, but high levels of noise degradation. To accommodate this trade-off situation, trunk lines are driven at very low levels, just high enough for reasonable signal-to-noise performance. This allows a greater number of amplifiers to be used one after the other (or cascaded). This maximizes the practical length of a trunk line. Conversely, feeder lines are operated at higher levels to provide an adequate signal level to

serve all the taps in the line. As a result, the feeder line has a very limited cascade of amplifiers before the cross-modulation distortion becomes intolerable.

At a trunk amplifier, a small portion of the trunk signal is applied to a high-gain amplifier referred to as a bridger amplifier. This amplifier feeds the feeder lines attached to this particular trunk station. The feeder lines have taps, where needed, to serve the customers. These taps are specifically designed to present a constant 75-ohm impedance to the feeder line and to each customer drop line. The constant impedance also serves to isolate the customer drop lines from each other. This reduces the potential for interference between drops. As the signal in the feeder line is attenuated, amplifiers referred to as "line extenders" are inserted to provide additional signal gain. Since the feeder is high-level, and the line extenders are also high-level devices, there is a limit to the number of line extenders that can be cascaded before cross-modulation distortion becomes a problem.

This is an opportune moment to further explore the role of the coaxial transmission line. As mentioned before, the standard CATV transmission line is the 75-ohm coaxial line. All CATV trunk, feeder, and drop cables have this 75-ohm characteristic impedance. From basic transmission line theory, the maximum power transfer will occur when this 75-ohm line is matched with 75-ohm devices, and minimum reflections will occur if these lines are resistively terminated with a 75-ohm resistor. These are two very important considerations. A poorly matched line can be very lossy and the reflections caused by the mismatch will degrade the final picture quality received by the customer. A frequent point of confusion often arises from the CATV industry use of the 75-ohm characteristic impedance when discussing "standard" coaxial cables (ie. RG-59 or RG-6). Most texts will quote a 73-ohm characteristic impedance for these cables. Early in the development of CATV, some 73-ohm cable was used, but the CATV industry gradually settled on the 75-ohm characteristic impedance that is now universally accepted. Most cable manufacturers have accepted the 75-ohm characteristic impedance, so it is important to note this fact when using these coaxial

cables for purposes outside the CATV industry. Most manufacturers will add a letter designation, such as RG-59A or RG-6A, to the 75-ohm version of that particular cable.

With the discussion of coaxial cable, some discussion of transmission lines is in order. Figure 2-5 is a representation of a coaxial cable. The coaxial cable is constructed of two

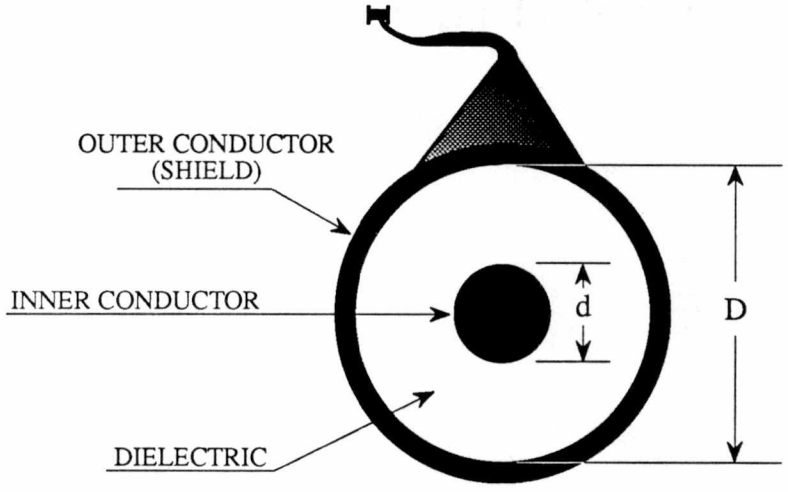


FIGURE 2-5

A SECTION-VIEW OF A COAXIAL CABLE

conductors separated by an insulating dielectric with one of the conductors being a hollow cylinder completely enveloping the other conductor in the axial dimension - hence the term co-axial. The symmetry of this situation leads to equation 2-1, which is the expression for the characteristic impedance of any coaxial cable.[17] The choice of a dielectric and diameters are all that are required to construct a coaxial cable with a particular characteristic impedance. The D and d variables are as given by Figure 2-5, and ε_r is the dielectric constant of the insulation. The material composition of the inner and outer conductors has no bearing on the characteristic impedance. Physical size and material composition affects some coaxial cable

$$Z_0 = \frac{138}{\sqrt{\varepsilon_r}} \log \frac{D}{d} \qquad (2\text{-}1)$$

parameters. High frequency attenuation is a function of physical size, and dielectric and conductor material. A large cable has lower attenuation than a smaller cable. At high frequencies, the loss mechanisms in coaxial cable are: conductor heating, dielectric heating and radiation losses. Larger cables tend to lessen the conductor and dielectric heating losses. Most recent advances in coaxial cable loss have been improved dielectric compounds. The choice of dielectric material is very important to the performance of the coaxial cable; it must present a low-loss path between the two conductors and still mechanically support the conductors. In installing this cable, it is very important not to kink or dimple the outer conductor (shield) and it is important not to displace the inner conductor from the coaxially-symmetric position. Any of these flaws will cause the cable to exhibit an impedance other than the characteristic impedance in the flawed areas, which can cause mis-match problems. At DC and low frequencies, the conductivity of the inner and outer conductors is the dominant source of attenuation. At these lower frequencies, losses are primarily due to resistive losses. For CATV purposes, these resistive losses determine the limitations of the cable for carrying the 60 Hz AC used to power the CATV system's active equipment.

A CATV system is constructed entirely of 75-ohm coaxial cable. There are many reasons for maintaining a constant impedance: ease of design, construction, and maintenance, and the benefit of mass produced materials, just to mention a few. All CATV equipment (ie amplifiers, splitters, directional couplers, etc...) present a 75-ohm impedance at all ports of the device which can be connected to a CATV system. Internally, each of these devices may have impedances which are far removed from the 75-ohm system impedance, and indeed different designs from different manufacturers are significantly different internally, yet all present a 75-ohm impedance at all ports. This allows like devices from different

manufacturers to be interchanged without an impedance matching problem, although performance or quality can vary from manufacturer to manufacturer.

CATV system testing is simplified by the choice of a constant system impedance. All test equipment is characterized by having a 75-ohm input impedance. This ability to work in a constant impedance environment brought about some terminology and techniques that greatly simplify design, construction, and maintenance of CATV systems. Probably the most important and the most confusing is the decibel (dB). Phonetically, the dB is usually referred to as the "dee-bee".

To paraphrase: "A dB is a dB is a dB, except...". The exception is that most uses of dB are not true to the definition of dB, so care must be taken in using dB nomenclature. As a strict definition, a dB is a dimensionless number used to express the ratio of two values. Equation 2-2 is the expression for dB for power ratios, and equation 2-3 is the expression for dB for voltage ratios.

$$dB = 10 \log \frac{P_1}{P_2} \tag{2-2}$$

$$dB = 20 \log \frac{V_1}{V_2} \tag{2-3}$$

The log tends to highly "compress" the values of the ratio. This makes large ratio ranges compress into a more manageable scale, but with a corresponding loss of resolution. The dB that is most familiar is not this "true" defined dB, but rather a dB that does indeed have dimensions associated with it. By letting either P_2 or V_2, from equations 2-2 and 2-3 respectively, equal a measurement reference value of power or voltage, the resulting dB value

then takes on dimensions and has an absolute, defined value. The dBmV is the unit most often used in CATV applications. The dBmV is defined by equation 2-4, note that the reference is 1000 microvolts measured across 75-ohms. This latter requirement is continuously met because CATV systems are consistently 75-ohm.

$$\text{dBmV} = 20 \log \frac{V}{1000\ \mu V} \tag{2-4}$$

The choice of the 1000 microvolt reference level is not as arbitrary as it may seem. An FCC-commissioned study by the Television Allocations Study Organization (TASO)[18] recommended that a 1000 microvolt signal provided to a television receiver would give very acceptable picture quality. CATV systems took this as indicating the minimum acceptable signal level for any television receiver. Due to the nature of the log scale, in equation 2-4, when V is equal to 1000 microvolts, the resulting value of dBmV is zero. If V is less than 1000 microvolts, the value of dBmV is negative. If V is greater than 1000 microvolts, the resulting value of dBmV is positive. Hence, a television receiver should be provided with 0 dBmV or more.

A very good question at this point is why go to this trouble of dBmV conversion if the signal can already be measured in microvolts? The utility of the dBmV is not clearly evident until it is used in practical situations. Perhaps a good place to start is scale-compression. Table 2-1 is a short table of some values of dBmV and their corresponding values in microvolts. Table 2-1 shows that a very large range of signal levels can be compressed into a very small scale. There is a loss of resolution, but most instruments have an accuracy no better than ±1/4 dB, so resolution is not a real problem. It is much easier to

remember that a trunk amplifier's output is 32 dBmV, than to remember it is 39,810 microvolts!

TABLE 2-1

A TABULAR COMPARISON BETWEEN dBmV AND MICROVOLTS

dBmV	Microvolts
-10	316.2
-5	562.3
0	1000
5	1,778
10	3,162
15	5,623
20	10,000
25	17,780
30	31,620
35	56,230
40	100,000

Another situation in which dBmV is useful is in CATV system design. For instance, a 2-way splitter "looses" 3 db of signal in the split. The signal is not really lost, 3 db is simply halving the signal between the outputs. As an example, the splitter in Figure 2-6 has an input level of 26 dBmV. If 3 db is lost in splitting this signal, each output port has 26 - 3 = 23 dBmV of output signal level.

Working the same example with microvolts, converting 26 dBmV to 19,950 microvolts, leads to a slightly more complicated situation. Given that the power splits equally, how does this affect the voltage? Voltage may be substituted into the dB defining equation 2-2 by squaring the voltage. This is how equation 2-3 is derived for voltages instead of using powers. With this in mind, equation 2-5 for dB can now be solved because dB = 3, which is the ratio of power in (V_{IN}^2) to power out (V_o^2).

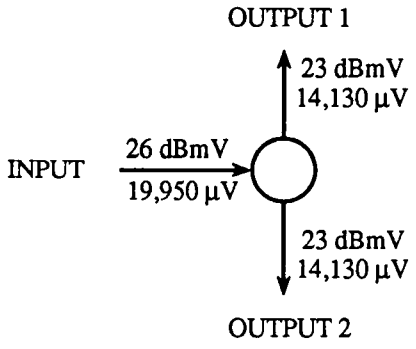


FIGURE 2-6

A CATV SIGNAL SPLITTER DESIGN EXAMPLE

$$3\ \text{dB} = 10 \log \frac{(V_{IN})^2}{(V_O)^2} \tag{2-5}$$

$$\text{Solving for } \frac{(V_{IN})^2}{(V_O)^2} \text{ yields} \tag{2-6}$$

$$\log^{-1}\left(\frac{3}{10}\right) = \frac{(V_{IN})^2}{(V_O)^2} \tag{2-7}$$

$$2 = \frac{(V_{IN})^2}{(V_O)^2} \tag{2-8}$$

$$V_O = \frac{V_{IN}}{\sqrt{2}} \approx 0.707\ V_{IN} \tag{2-9}$$

With V_{IN} = 19,950 microvolts, V_O = 14,130 microvolts. (2-10)

Similar calculations would be necessary for all devices used in the system, whether lossy or active. This is a very cumbersome method with which to design because of the size of the numbers and the unusual, sometimes fractional, relationship of the input and outputs. Fortunately, the log function makes gain and loss relationships simply an addition or subtraction exercise. As an example of an active device, Figure 2-7 is a trunk amplifier with actual input and output levels in dBmV and microvolts.

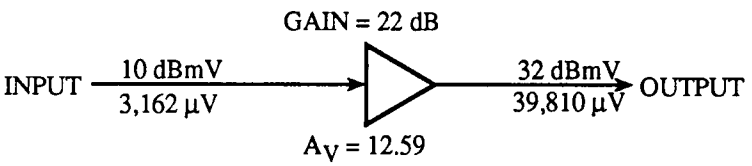


FIGURE 2-7

A CATV AMPLIFIER DESIGN EXAMPLE

Another way in which dBmV simplifies design is in determining cable loss. Manufacturers provide attenuation information in terms of dB per foot. It is a simple matter to multiply this figure by the footage to arrive at the attenuation. Perhaps note has been taken of the use of both dB and dBmV since their introduction, in a fashion that might indicate they are interchangeable. This is an incorrect assumption. A dBmV is a unit of measure and as such has dimensions. The dB is a dimensionless ratio. In active and passive devices, the gains and losses are expressed in dB. This is because it is a measure of the ratio of input to output

for the device, and as such has no dimensions. When discussing a particular input or output level, this is measured in dBmV.

The CATV system headend is generating television signals and the trunk and feeder lines are transporting these signals. All that is left is delivery of the CATV system signals to the customer. The taps in the feeder lines are used to deliver the signal, via a drop line, to the customer. The drop line is a miniature version of the trunk and feeder lines, but with a greater degree of flexibility.

The signals have now been delivered to the customer, but there may be one more step to insure proper reception by the customer's television receiver. Chapter 1 on television has a figure that explains the frequency spectrum that modern NTSC television uses in the United States. This figure is reproduced in Figure 2-8, with the frequency spectrum utilized by modern CATV systems, the "cable channels", superimposed onto the figure.[19]

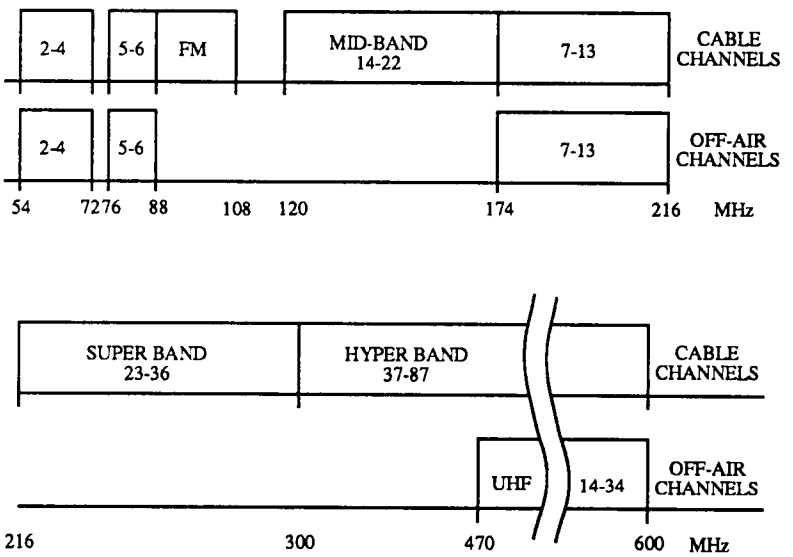


FIGURE 2-8

FREQUENCY SPECTRUM ALLOCATIONS FOR OFF-AIR BROADCASTING AND CATV SYSTEMS

Conventional television channels were added as demand and technology dictated. Since these channel additions occurred at various times, the television spectrum appears to be "broken-up" and not continuous. All the frequency spectrum between the three major channel divisions (VHF Low, VHF High, and UHF) are occupied by other services. Theoretically, a cable system does not have to adhere to this spectrum allocation as coaxial cable is non-radiative. With careful construction and maintenance, a CATV system can indeed approach near non-radiative performance. This allows the CATV system to "fill-in" the gaps that exist in the off-air broadcast spectrum allocations. These additions are shown in Figure 2-8 and the designations are listed for the various frequency groups. The important thing about this allocation is that a method must exist for a conventional television receiver to "tune-in" these additional channels. The answer is the converter.

The "cable box," or converter, does the job of converting all the channels of a CATV system to a single, conventional television channel. The converter has some means of selecting a cable channel, which it will convert to a conventional channel (usually channel 2,3, or 4). The customer's television is tuned to this channel and all tuning is then performed by the converter. A special converter may also be used to decode or unscramble a channel or group of channels. The most popular coding method consists of suppressing the sync pulses at the headend and having this special converter re-insert the sync pulses. A more sophisticated converter, called an addressable converter, can be controlled remotely, without the need for physical access to the converter.

Figure 2-9 is a block diagram of a modern CATV channel converter. The input is applied to the up-converter section where light filtering and RF amplification occurs first. The tuning control adjusts the voltage controlled oscillator (VCO), the output of which is mixed with the input signals. The output of this process is an IF which is an up-converted version of the input, with all the mixer products above the highest input frequency. This broadband IF is then applied to the IF filtering and amplifying section. Here the 1000-1006

MHz IF signal is selected. This means to obtain the channel of interest, it must first be mixed with the proper VCO output frequency so that the mixing process causes this channel to mix to the 1000-1006 MHz IF frequency. The subtractive mixing product is used here so that the other mixing products fall outside the input frequency passband. The filtered and amplified IF is then fed to the down converter. The down converter mixes a fixed oscillator signal with the IF to generate the output. The selection of the fixed oscillator frequency determines the output channel, and the subtractive portion of this mixing process is used to eliminate spectrum inversion and unwanted mixing products. The signal from the down converter mixer is fed to a filter fixed to the appropriate output channel passband. In most modern converters this filter is usually a SAW filter.

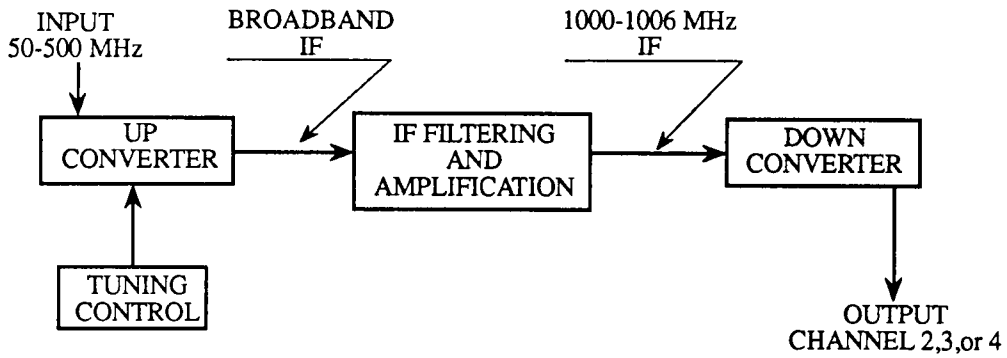


FIGURE 2-9

A BLOCK DIAGRAM OF A TYPICAL CATV CONVERTER

As with all CATV devices, the input and output impedance of the converter is 75-ohms. A short length of drop cable connects the converter output to the television receiver. If the television receiver has a 75-ohm input, the drop cable can be connected directly to the receiver. Otherwise, a 300-ohm balanced-line input to the receiver is achieved through the

use of a "matching transformer." This small device converts the 75-ohm unbalanced input to a 300-ohm balanced-line output that connects directly to the receiver. This device is passive and achieves the impedance transformation by transformer action.

One chapter does not allow for a complete, exhaustive discussion of cable television; however the basics have been covered. The primary area of interest is the headend channel modulator. A reasonable knowledge of dBmV, cable channel allocations, noise and cross-modulation distortions, and cable system operation will assist in reaching the main goal of modulation measurement.

CHAPTER 3

AMPLITUDE MODULATION

Amplitude modulation was the modulation system chosen by the NTSC for the visual portion of the television transmission. A good, basic understanding of amplitude modulation (AM) is necessary for a complete understanding of the problem set forth by this thesis. Most texts teach AM theory with a mathematical approach, and such will be the case here, but first it is important to develop an intuitative, physical "feel" about AM.

Scale is an excellent starting point. As previously mentioned in the television section, broadcast transmission at video frequencies is difficult and inefficient, but by modulating this video onto a much higher-frequency carrier these problems are overcome. For scaling purposes, broadcast video occupies a bandwidth from approximately DC (0 Hz) to 4.2 MHz (4,200,000 Hz) and the carrier frequencies this video will modulate are 54 MHz (54,000,000 Hz) to over 500 MHz (500,000,000 Hz). A practical demonstration of the scale differences between the signal and the carrier is illustrated in Figure 3-1. All four figures represent the screen of an oscilloscope. An oscilloscope can "slow-down" a signal and display a small portion of this signal on a screen. In Figure 3-1(a), the top trace is the unmodulated carrier and its frequency is 55.25 MHz, the visual frequency of channel 2, the lowest frequency used in television broadcast. Note that it is easy to distinguish each cycle of the carrier even though there are 55,250,000 of these cycles occurring each second. The bottom trace in Figure 3-1(a) is the sinusoidal signal which is used to represent the video. The frequency of 4.2 MHz is the highest used in broadcast video, but is much lower than the carrier, so the changes in time of this signal are not yet apparent. Figure 3-1(b) is the same carrier and signal, but the oscilloscope time base is "sped-up." Notice how the carrier cycles are more closely packed and it is becoming difficult to distinguish the individual cycles. Likewise, note that the signal

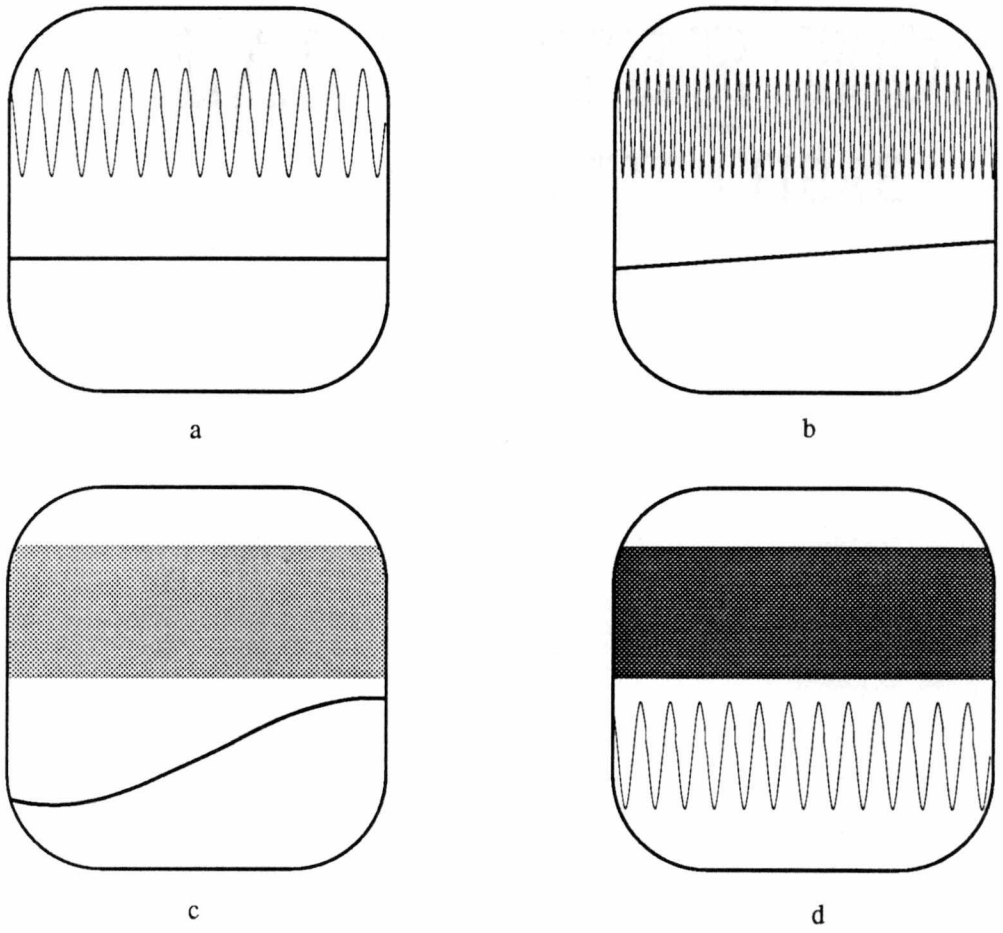


FIGURE 3-1

A ILLUSTRATION OF THE FREQUENCY-SCALING DIFFERENCES BETWEEN THE CARRIER AND THE MODULATING SIGNAL

now shows a perceptible change, but not yet a full cycle. In Figure 3-1(c), the oscilloscope is sped-up even more - notice that this causes the carrier to be a "blur" of closely spaced cycles, but the cyclic nature of the signal is only now becoming apparent. In Figure 3-1(d), the cyclic nature of the carrier is obscured by the resolution of the oscilloscope and the human eye. At this point the signal can now be viewed and its cyclic nature is very apparent.

This has been a practical exercise to illustrate the differences in scale of the frequency

of the carrier and the frequency of the signal. Indeed, Figure 3-1 was drawn from actual oscilloscope traces. From a relative standpoint, the carrier "views" the signal as an extremely slowly-varying voltage, and the signal "views" the carrier as a non-changing voltage, due to the carrier's rapid transition speed. In the period of time that one cycle of signal exists, the carrier will go through over 13 cycles, and would be even more at the higher frequencies of the other television channels. This is an important concept to grasp for a thorough understanding of AM.

Amplitude modulation is defined as a system of modulation in which the amplitude of the carrier is made proportional to the instantaneous amplitude of the modulating voltage. Figure 3-2 is a mechanical representation of the AM process.

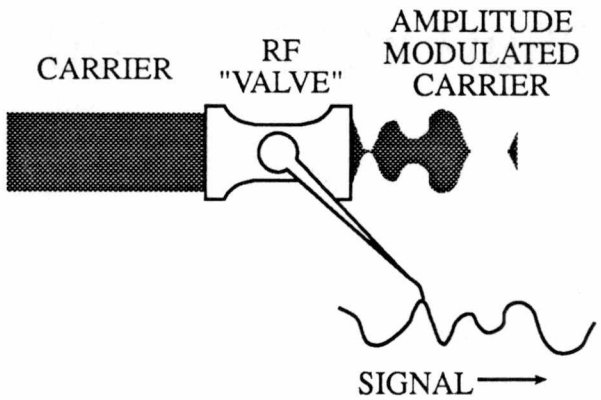


FIGURE 3-2

A MECHANICAL ANALOGY OF AMPLITUDE MODULATION

Figure 3-2 may appear whimsical, but it does convey a physical sense to the process of amplitude modulation. Note that the "RF valve" control handle "rides" the peak of the signal; positive amplitude peaks allow more carrier to pass through, whereas negative

amplitude peaks attenuate the carrier significantly. If negative peaks reach a certain level, the carrier will be completely cut-off. At this point the carrier is over-modulated, a condition that is usually undesirable. Likewise, a positive peak of sufficient amplitude can exceed the carrier "head-room" and also create an over-modulated situation. Later, this simple model will be shown to have some small inconsistencies with the mathematical AM model, but is an accurate representation of the specialized AM used in television broadcast.

Moving from mechanical analogy to mathematical modeling and using a method suggested by Kennedy [20], let the instantaneous carrier voltage be represented by e_c and the instantaneous modulating signal be represented by e_m. For the sake of simplicity, assume both carrier and modulating signal are sinusoidal, giving:

$$e_c = E_c \sin \omega_c t \tag{3-1}$$

$$e_m = E_m \sin \omega_m t. \tag{3-2}$$

Where E_c and E_m are the peak values of the carrier voltage and modulating signal voltage, respectively. The ω_c and ω_m are the frequencies of the carrier and modulation signal, respectively; and noting that $\omega_{c >> } \omega_m$ in most cases. In this case ω_c and ω_m are angular frequencies measured in radians, but can be easily converted to the more familiar Hertz by use of the relation: $\omega = 2\pi f$. Notice that the phase angle has been omitted since it is unchanged by the amplitude modulation process.

From the basic definition of AM, the peak value of the carrier, E_c, must be made proportional to the modulating voltage e_m. This proportionality must be one so that the

absence of modulating signal leaves the amplitude of the unmodulated carrier, e_c. Figure 3-3 illustrates this simple modulation situation.

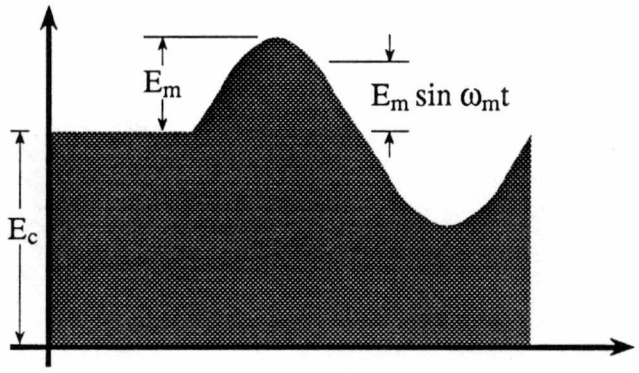


FIGURE 3-3

THE REPRESENTATION OF AMPLITUDE MODULATION USED IN THE THEORETICAL DEVELOPMENT

As in the mechanical analogy, if $E_m > E_c$, over-modulation occurs. The ratio of E_m to E_c is usually referred to as the modulation index m:

$$m = \frac{E_m}{E_c}. \tag{3-3}$$

For distortionless operation $E_m > E_c$, therefore m is a number between 0 and 1, and can be represented as a percentage of modulation.

From equation 3-3 and Figure 3-3, it is possible to write an equation for the amplitude of the modulated carrier at the arbitrary point A. Note that the value of the carrier is constant

and equal to E_c. The amplitude of the modulation signal is varying as given in equation 3-2. From this it follows that

$$A = E_c + E_m \sin \omega_m t \ . \tag{3-4}$$

Rearranging equation 3-3 gives

$$E_m = m\, E_c \ . \tag{3-5}$$

Substituting equation 3-5 into equation 3-4 yields

$$A = E_c + mE_c \sin \omega_m t \ , \tag{3-6}$$

or,

$$A = E_c \, (1 + m \sin \omega_m t) \ . \tag{3-7}$$

This argument assumed the carrier was constant at value E_c, whereas the carrier e_c is varying as in equation 3-1. A is now the peak value of the carrier which will vary as in equation 3-8, substituting into equation 3-8 yields the instantaneous voltage of the carrier e_c:

$$e_c = A \sin \omega_c t \tag{3-8}$$

Substituting equation 3-7 into equation 3-8 yields

$$e_c = E_c (1 + m \sin \omega_m t) \sin \omega_c t . \tag{3-9}$$

Using the trigonometric expression,

$$\sin x \sin y = \frac{1}{2} [\cos (x - y) - \cos (x + y)] , \tag{3-10}$$

to expand equation (9) yields

$$e_c = E_c \sin \omega_c t + \frac{mE_c}{2} \cos (\omega_c - \omega_m) t - \frac{mE_c}{2} \cos (\omega_c + \omega_m) t . \tag{3-11}$$

Equation 3-11 is the expression for the instantaneous value of the carrier at any time, t, and the simple situation of a single-sinusoid modulating signal. Equation 3-11 has several interesting terms. The first term, $E_c \sin \omega_c t$, is our original expression for the carrier, with frequency ω_c and peak value E_c. The second term $mE_c/2 \cos (\omega_c - \omega_m)t$does not have the same frequency, but has a lower frequency $\omega_c - \omega_m$. Likewise, the last term, $mE_c/2 \cos (\omega_c + \omega_m)t$ is at a third-higher frequency $\omega_c + \omega_m$. These second and third terms are called "sidebands," and they contain all of the modulated information. This is because the first term is simply the unmodulated carrier, and although the instantaneous value changes with time, no information

is contained in this variation. Figure 3-4 is a frequency spectrum representation of equation 3-11. Note that with m at the maximum value of one, the sidebands are at their maximum value of $E_c/2$, half the carrier voltage. Also, since all the modulated information is contained

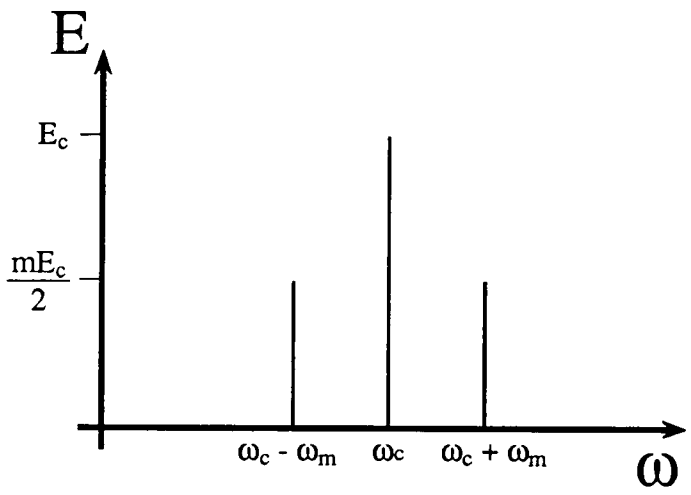


FIGURE 3-4

FREQUENCY SPECTRUM OF A SINGLE SINEWAVE AMPLITUDE MODULATED CARRIER

within the sidebands, the carrier is superfluous and can be deleted from broadcast. Similarly, since the information in both sidebands is identical, one of them may be deleted. These deletions can save significant amounts of bandwidth, but can lead to transmitter and receiver complexities. Equation 3-11 and Figure 3-4 are representative of the AM used in standard AM radio broadcast service in the United States.

Equations 3-11 and Figures 3-3 and 3-4 are only applicable when the modulating signal is a single-frequency sinusoid. At first glance this would appear to be a very limiting case, as most useful modulation is usually a collection of random frequency and random

amplitude signals, such as music or speech. Fourier analysis can assist in expanding this AM theory for use with more "real-world" modulating signals. Fourier analysis is a mathematical method of constructing periodic waveforms. It works by algebraically summing sine waves to construct the waveform of interest.[21] Fig 3-5 shows a squarewave being constructed by sinewaves.

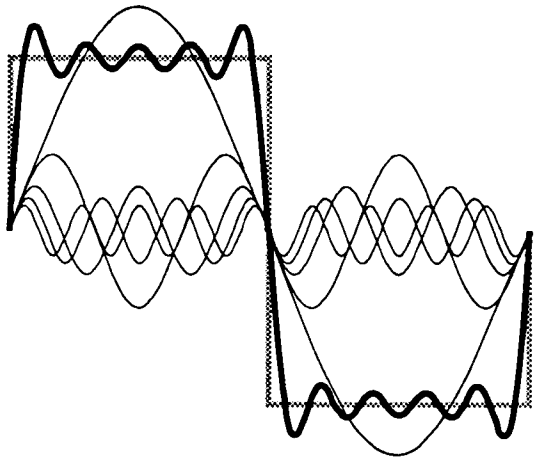


FIGURE 3-5

CONSTRUCTION OF A SQUAREWAVE BY THE SUMMING OF SINEWAVES

There are several points to remember about the use of Fourier analysis:

1. The fundamental frequency of the waveform is also the lowest frequency sine wave used.

2. The fundamental frequency sine wave in (1.) always has the largest amplitude of all the sine waves used to create the waveform, with each successive sine wave decreasing in magnitude from the previous.

3. All sine waves summed will have a frequency relationship to the fundamental frequency (f) of: f + 2f + 3f+ .·· +nf, with some waveforms requiring only the even or odd numbered frequencies.

4. Except for sinusoid-based waveforms, all waveforms require an infinite sum of sinusoids to create a perfect copy of the waveform; however since the magnitude of each successive sinusoid decreases, it is usually practical to create a waveform with a finite number of sine waves.

Infinite summations appear to be more difficult to deal with than they truly are. Consider again the squarewave in Figure 3-5. Equation 3-12 is the Fourier expression of this waveform, and a small series expansion is shown. As can be seen, equation 3-12 is merely a sum of sinusoidal waves and sinusoids fit very nicely into equation 3-11 and the general AM theory just covered. Each harmonic sinusoid is applied to equation 3-11 and the harmonic amplitude and frequency determine the sideband characteristics for each harmonic. Placing all these sideband terms together gives a complete picture of the frequency spectrum of an AM squarewave.

The choice of squarewave modulation is both instructional and intentional. The squarewave has a high level of harmonic content and is one of the most difficult waveforms to generate without using excessive amounts of bandwidth. The similarity between the squarewave modulation signal and video sync pulses should begin to be apparent. Fourier analysis can be very complicated from the mathematical standpoint, however it can be very important in determining bandwidth requirements for a particular AM situation.

This general AM theory can now be applied to the specifics of television. Figure 3-3 illustrated theoretical AM by a sinusoid signal and might be considered symmetrical

modulation (symmetrical about the carrier level). Television broadcast AM differs from this by using something usually referred to as negative modulation. Television modulation places all the amplitude variation between full carrier level and zero carrier level, corresponding to zero modulation and 100% modulation, respectively.[22] Negative modulation is the inverse of the modulating signal; increasing signal amplitude decreases the carrier level and decreasing signal amplitude increases the carrier level. Figure 3-6 is representative of negative modulation. Note that 0% modulation corresponds to a DC level and is used to maintain the integrity of the DC content of video.

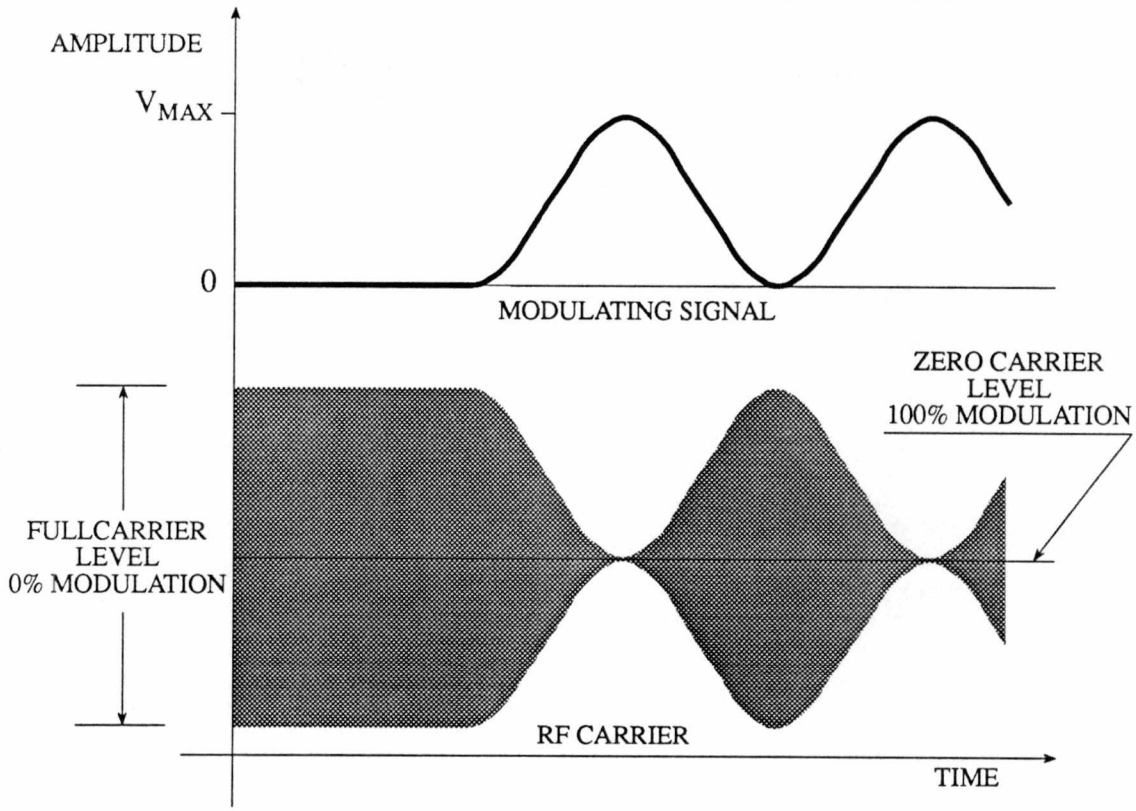


FIGURE 3-6

AN EXAMPLE OF NEGATIVE AMPLITUDE MODULATION

Television AM more accurately agrees with the mechanical analogy of Figure 3-7. By reversing the "valve" action to create negative modulation and using video as the modulating signal, Figure 3-7 is representative of television RF AM.

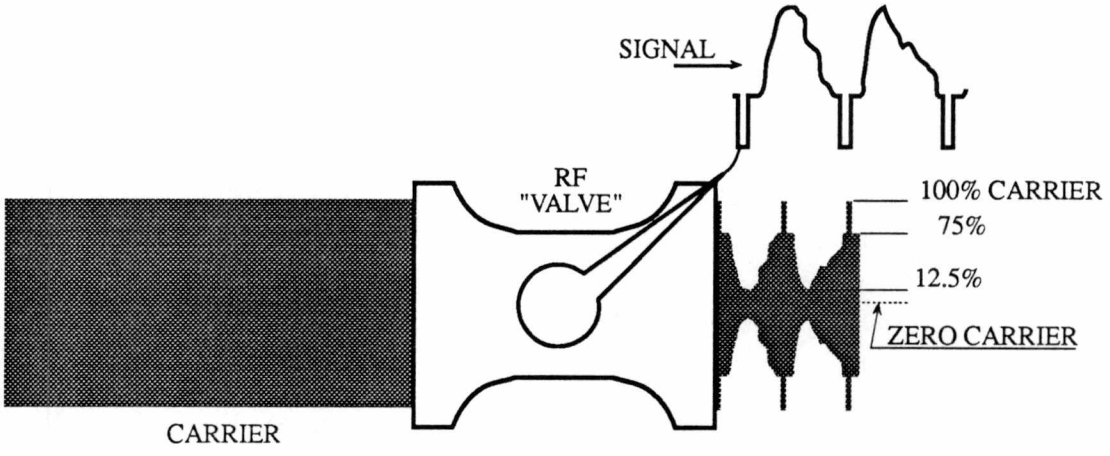


FIGURE 3-7

A MECHANICAL ANALOGY OF TELEVISION NEGATIVE AMPLITUDE MODULATION

Although Figure 3-7 is a mechanical analogy, RF "valves" exist as electronic attenuators in real-life situations. At high frequencies, PIN diode attenuators are used to modulate the amplitude of carriers much as the "RF valve" of Figure 3-7 does.[23] Most all active and passive electronic devices lend themselves to the use of "valve" analogies.

AM is a transmission mode widely used in broadcasting. Fourier analysis allows AM to be expressed in mathematical terms. This is a very complicated process for a video modulating signal. In the case of video modulation, a physical "feel" for the AM process is important for understanding television RF AM.

CHAPTER 4

MODULATION MEASUREMENT METHODS

An overview of modulation measurement techniques in use today is in order. This is important in formulating an approach to this measurement problem that either improves on present methods or simply uses present methods as a research resource toward a totally new method. This is also important to insure that duplication of previous work is not unnecessarily performed, and also guards against possible patent infringement.

Several methods exist to adjust the modulation level of a CATV modulator. Each method has differing amounts of accuracy and complexity.

Probably the most widely used method is the previously mentioned "eyeball" method. In this method, a television is monitored and used to adjust the modulation level by "eye." This is best accomplished by comparing an off-air signal to the signal under adjustment. This method is the lowest in cost, but is also the least consistently accurate method. Interestingly enough however, an experienced technician can usually set modulation levels within 5% of optimum. This is further aided by the white-clip circuitry incorporated into most modern CATV modulators. Although not a substitute for proper modulation level adjustment, it does guard against extreme over-modulation situations.

Another method of modulation monitoring is direct RF envelope measurements.[24] In this method a high-frequency oscilloscope is used to directly observe the RF output of the modulator. Maximum and minimum envelope excursions can then be used to calculate the modulation depth. This measurement method can be quite complex and expensive in practice. Direct RF viewing at the channel frequency requires a very sharp bandpass filtering network to avoid interference from the channels that surround the channel under test. To an extent this problem can be overcome by testing the modulator signal before it is mixed with

the other channels in the headend, but even this requires that the aural carrier be deleted so as not to affect the accuracy of this test. Direct RF measurement requires a very sharp filtering network for each channel tested; an unreasonable situation in a modern, multi-channel CATV system. This problem can be overcome by the use of IF monitoring. In this, the IF signal is monitored in the same way as the RF signal. The advantages are that the IF is of a lower frequency than any channel and some modulators split the IF into visual and aural components negating the need for aural filtering. An important note is that some early CATV modulators do not have IF loop-thru connections that allow easy interception of the IF signal. Figure 4-1 is a graphic representation of the oscilloscope display waveform created by direct view of a properly filtered IF or RF signal. It should be noted that Figure 4-1 is a

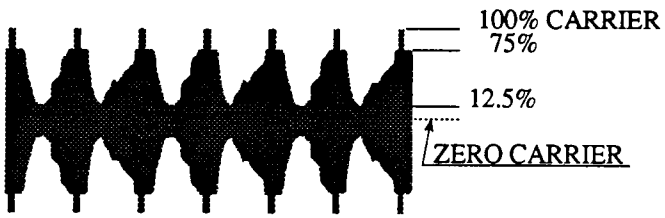


FIGURE 4-1

THE AMPLITUDE MODULATED RF ENVELOPE FOR MONOCHROMATIC TELEVISION

monochromatic signal with no color information and a fully attenuated aural carrier. There are several problems associated with the use of the RF or IF envelope method. Both RF and IF signals are AM vestigial sideband signals, which have an attenuated sideband and a well-defined high frequency roll-off. A television receiver compensates for this with a high frequency peaking circuit. This situation leads to a difficulty in obtaining peak level displays of high frequency visual and color signals. This problem can be overcome somewhat by

monitoring the IF signal before it enters the vestigial sideband filter, but this requires that an incursion be made into the CATV modulator at the "board," or circuit level. IF harmonics are probably present at this point in the circuit and can make the oscilloscope display unintelligible. If all of these problems are overcome, the complexities of a video signal, especially a color signal, can make interpretation of the oscilloscope display very difficult.

A measurement problem that has not yet been discussed has to do with the marketplace in which CATV operates, and not with technical aspects. RF and IF direct view testing is almost always intrusive. Intrusive from the standpoint that it can require that the CATV modulator be out of service during the testing and adjustment process. In an industry that is primarily service oriented, intrusive testing is always avoided where possible.

Direct RF and IF envelope measurements are complex and require very expensive test equipment. The test is usually intrusive and the results can be very difficult to interpret. Subjectively comparing this method of measurement with the "eyeball" method, in terms of complexity and expense, would probably lead to choosing the "eyeball" method over the direct view method. Other measurement options do exist, though.

A widely accepted method of measuring and adjusting modulation levels of CATV modulators is the use of modulation meters.[25] These meters are physically a part of the modulator assembly and are marked with modulation percentage scales. The accuracy of these meters varies greatly from manufacturer to manufacturer and even from model to model of the same manufacture. Most meter circuits do not monitor the actual modulated RF or IF signal. Instead, the meter is connected to the video input that is used to modulate the IF carrier. The metering circuit monitors the peak levels of the incoming video to determine the degree of modulation. In simple terms, an increase in video input signal will increase the degree of modulation of the IF carrier. A decrease in the video input signal likewise corresponds to a decrease in the degree of IF carrier modulation. This method falls victim to any non-linearity in the modulating system. The metering circuitry monitors the video

input and not the actual modulated signal, so modulating non-linearities are not necessarily evident in the meter indication. Some CATV modulator manufacturers have overcome this problem to a degree by designing metering circuits that exhibit non-linearities that match the non-linearities of the modulating system. Modulation meters are built into almost all CATV modulators and are of some use in monitoring modulation levels, but the accuracy of the meter should be determined by another modulation measuring method prior to using the metering system as a testing method. Most modulation meters are not accurate and stable enough to serve as accurate test instruments.

A method recently gaining in popularity is the use of the spectrum analyzer.[26] From the Chapter 3 AM discussion, it was shown that the sideband power levels could be used to determine modulation levels. Although a true statement, in reality the complexities of video modulation coupled with the randomness of the visual signal make this method presently unrealizable. Before explaining the spectrum analyzer method of modulation measurement, some explanation of the function of a spectrum analyzer is in order.

The concepts of spectrum analysis and frequency spectrum are sometimes confusing to people not actively involved in the use of either concept. Figure 4-2 is an isometric diagram of what is actually occuring in the translation from the time domain to the frequency domain.[27] Most people are familiar with the operation of an oscilloscope and it's use in viewing time-varying signal amplitudes. This is view(a) of Figure 4-2, a simple sinusoid signal with a time-varying amplitude. The oscilloscope does not relay any information about the frequency of the signal, although it can be computed by the relationship between frequency and period. View(b) is a slightly different way of viewing the same signal. The reader should imagine that they are standing at the time axis looking at the signal approaching them. At this point the time variance of the signal amplitude is not as apparent as in view(a). Here the approaching signal appears to be a straight line varying from zero amplitude to a maximum value. If the signal frequency is high, the variation in amplitude will blur and

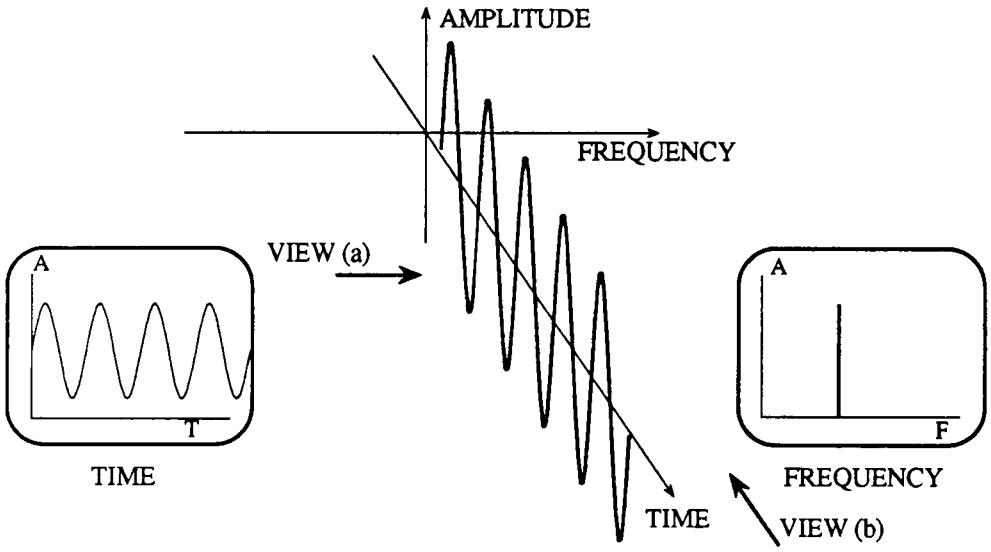


FIGURE 4-2

THE RELATIONSHIP BETWEEN TIME AND FREQUENCY

appear to be a straight line from zero amplitude to the maximum amplitude. The spectrum analyzer will "fold" the signal about the zero point creating the view(b) display on the screen with the bottom of the screen representing zero amplitude. This simple figure does not illustrate the utility of the spectrum "view," however Figure 4-4 uses a slightly more complicated signal to further demonstrate the use of the spectrum view.

Figure 4-4 uses the squarewave previously constructed in Figure 3-5 of Chapter 3. This exercise was concerned with showing that non-sinusoidal waveforms are constructed by an infinite sum of sinusoidal signals. Figure 4-3 is a reproduction of this exercise.

Figure 4-3 illustrates a squarewave being generated by 5 discrete sine waves, and although not a perfect squarewave, it is still recognizable. Figure 4-4 is an isometric representation of this signal in both frequency and time domains. As previously stated, view(a) is an oscilloscope view of the signals. Notice that unlike Figure 4-3, the individual sine wave components are not visible, only the resultant of the algebraic summing of these

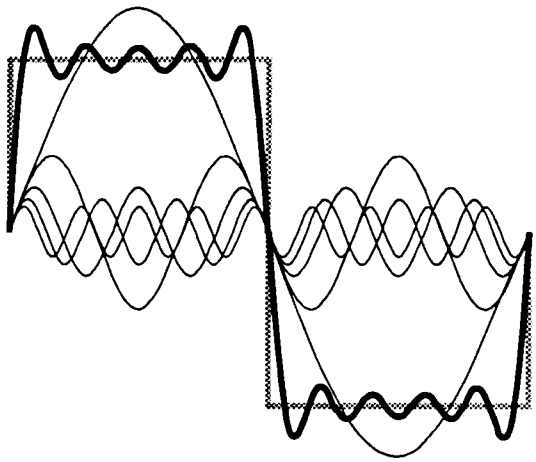


FIGURE 4-3

THE CONSTRUCTION OF A SQUAREWAVE BY THE SUMMING OF SINEWAVES

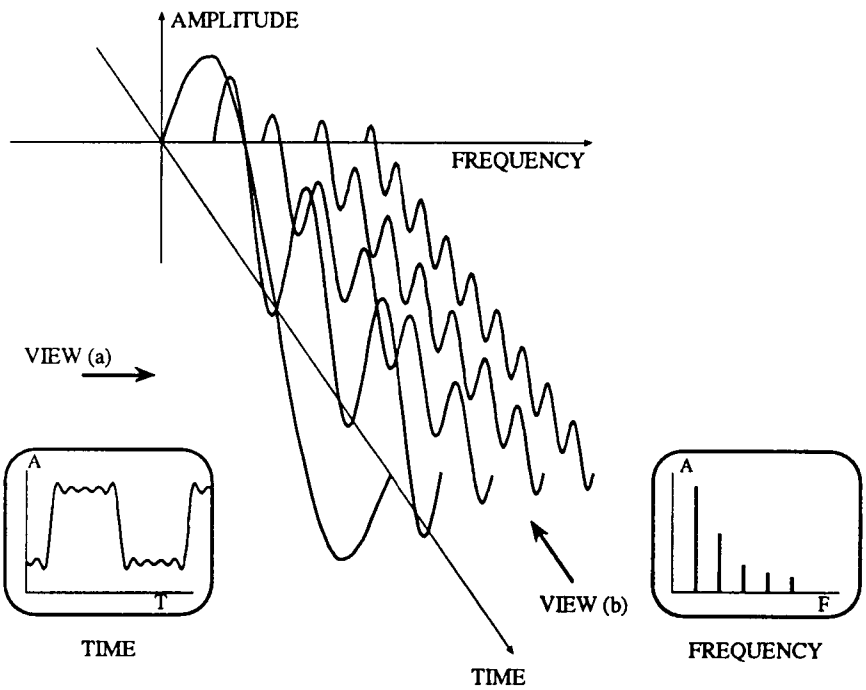


FIGURE 4-4

THE TIME AND FREQUENCY RELATIONSHIPS FOR A SQUAREWAVE

discrete components is visible. From the view(b) vantage point the oncoming signals look to be 5 discrete straight vertical lines extending from zero amplitude to the maximum amplitude. From this vantage point, it would be difficult to determine the time domain appearance of such a collection of sine components. The equivalent is true for the view(a) situation also. The spectrum analyzer allows the user to take up a view(b) vantage point and furnishes frequency spectrum and amplitude information.

A spectrum analyzer is a very powerful test instrument that allows the user to view the frequency spectrum and magnitude of the input signal. RF spectrum analyzers do this by "sweeping" the frequencies of interest and making level measurements continously through the sweep. Figure 4-5 is a simple diagram of a spectrum analyzer. The portion marked "R" can be thought of as a window that sweeps from the low frequency boundary to the high frequency boundary. As the window sweeps, the spectrum analyzer measurement circuits "see" only what is in the window and generate a voltage proportional to the average or peak signal level within the window. This plots as points on the viewing screen. The sweep and the beam on the screen are synchronized so that the display is representative of the spectrum tested. Instead of declaring a low and high frequency limit, spectrum analyzers use a center frequency adjustment and the low and high frequency limits are established by use of a "span" control. This allows any frequency to be centered on the display and examined very closely by limiting the span to a very narrow excursion about the center frequency. Likewise, the span may be increased to allow a great deal of frequency spectrum to be examined.

In using the spectrum analyzer to measure modulation, a broad span is not used. Instead, the spectrum analyzer is set up in a mode of operation referred to as "zero-span." This converts the spectrum analyzer to a tuned receiver, with the tuned frequency being the spectrum analyzer center frequency. In practice the television carrier frequency is center-tuned on the spectrum analyzer , and then the dispersion is zero-spanned. At this point, the image appearing on the spectrum analyzer screen is an amplitude versus time display. The

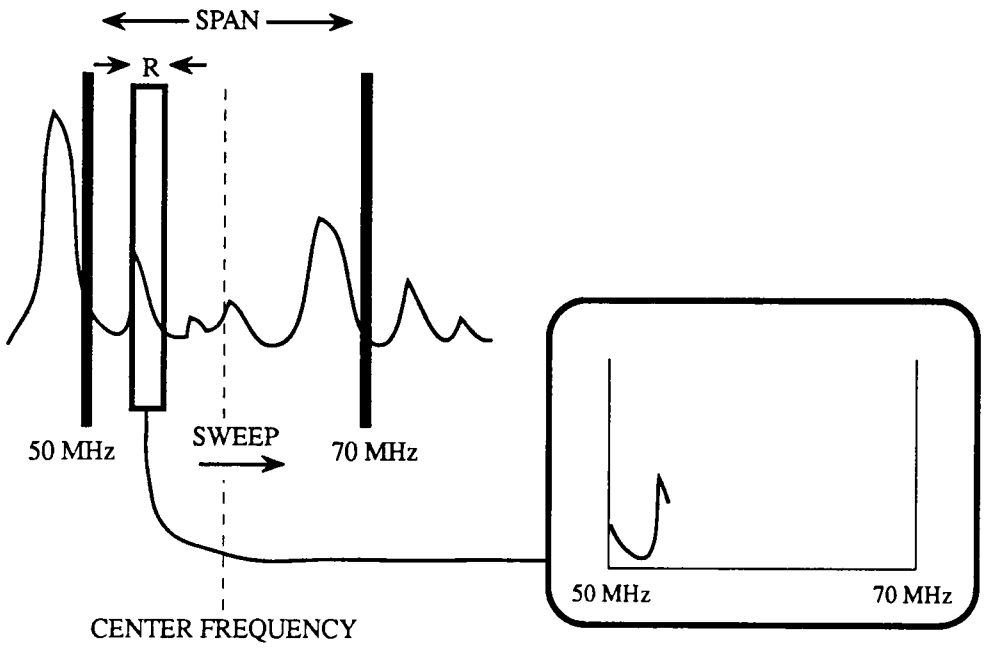


FIGURE 4-5

A SIMPLE SPECTRUM ANALYZER

width of the window "R" is usually referred to as the IF bandwidth or the resolution. In this test the resolution is set to about 3 MHz to insure adequate bandwidth to accurately reproduce the video waveform. The long persistence screen will display a maximum trace line that corresponds to the sync peaks. Below this, another line corresponding to the front and back porches will be created. If the input video signal is not a test signal, but is a normal video program, the complexity of the display is increased. Switching the spectrum analyzer amplitude scale to the linear mode allows the measurements to be made. Figure 4-6 is representative of the spectrum analyzer test using normal video programming.

Modulation adjustment can be made by either insuring that the random white peaks are one-eighth the magnitude of the sync tip level, or by insuring that the blanking level is three-quarters the magnitude of the sync tip level. Modulation adjustment using a spectrum

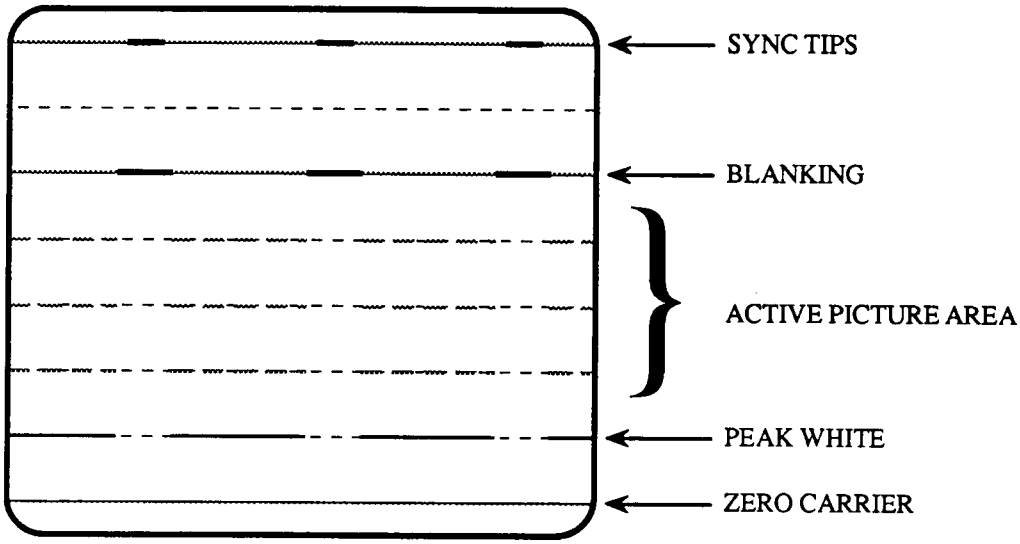


FIGURE 4-6

THE DISPLAY OF A SPECTRUM ANALYZER WHEN USED TO MEASURE MODULATION LEVELS

analyzer is relatively simple, but a spectrum analyzer capable of performing this test will cost more than a new luxury automobile. This expense makes this method impractical for most small CATV operations.

A final method of measuring modulation is by the use of a "zero-chop" demodulator.[28] In this test, an oscilloscope or waveform monitor is used to display the output of the demodulator that is tuned to the channel under test. During the vertical interval, circuitry in the zero-chop demodulator switch the input signal off for a short time. At this time the demodulator output is essentially a DC level corresponding to zero carrier. The appearance of this reference on the waveform monitor is then used as a reference, much in the same way as in the spectrum analyzer test explained previously. Figure 4-7 represents the waveform

monitor display of this test method. Although not as expensive as the spectrum analyzer, this is the second most expensive test method and requires a high level of training to perform accurately.

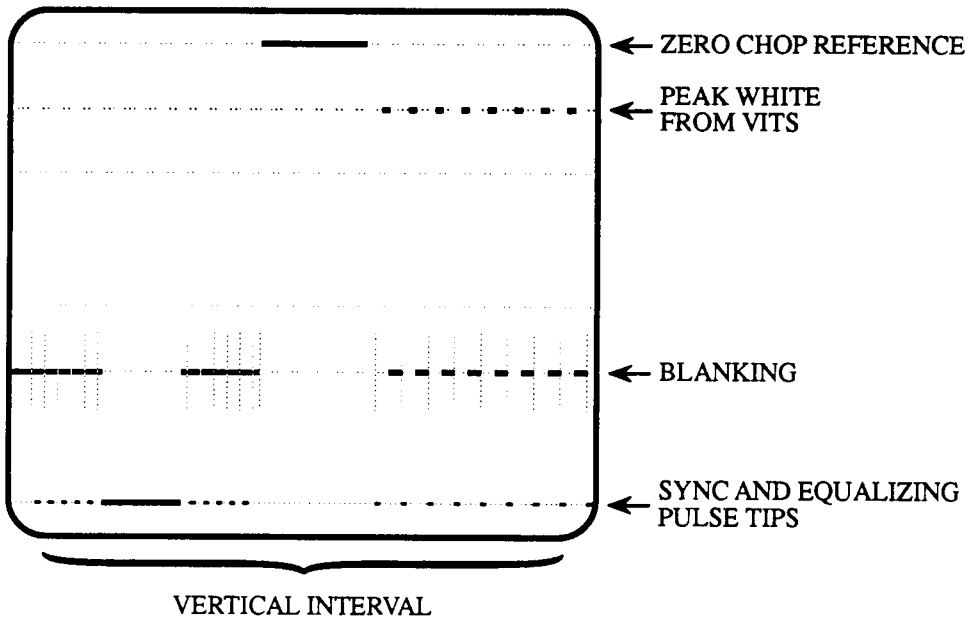


FIGURE 4-7

THE DISPLAY OF A WAVEFORM MONITOR USED WITH A ZERO-CHOP SWITCH TO MEASURE MODULATION LEVELS

The methods of modulation measurement in use at the present reinforce the statement that little effort has been expended to find an accurate, inexpensive method of measurement. Each method has strong points and each suffers from deficiencies. The information presented on these methods will be of assistance in formulating a measurement method.

CHAPTER 5

DEVELOPMENT OF A MODULATION MEASUREMENT METHOD

The groundwork having been laid, the task at hand is the development of a modulation measurement method. The development process is where the philosophy and biases of the designer, or designers will surface. It is very important to realize that the approach taken here is certainly not the only one that will accomplish the task at hand. Indeed, there are countless other ways to perform this task; each with it's own advantages and disadvantages. This chapter will be presented in a logical, tutorial fashion much like the development of the NTSC standards of Chapter 1. Actual development did not progress in this fashion. The "false-starts," "dead-ends," and poor assumptions of actual development can be instructive, but they tend to be lengthy and confusing. Several important developmental "mistakes" will be discussed for their tutorial merit.

"How good is good enough?" This sentence should be the guideline for anyone undertaking the development of anything. This slogan finds utility is all facets of life, not just in the engineering disciplines. This seemingly subjective question is precipitated by the fact that achieving perfection is impossible. This question should be applied using perfection as a baseline of comparison. "How much error can be tolerated?" "How much current is excessive?" "Which devices are best?" These are some of the questions that will need to be answered, both before and during the development process.

The developmental process may begin almost anywhere, but perhaps it is best to begin with the "big picture." In the "big picture," an all-encompassing macroscopic view is taken of the task at hand. The most important is the goal: development of a practical method of modulation measurement of the NTSC visual carrier. Simple as it may seem, it is elemental to solving the problem. The following is a list of primary concerns that are derived from this goal:

1 Cost?

2 Accuracy?

3 Method of application?

4 Marketability?

The list could easily be expanded, but first an examination of these four.

Cost is always a consideration. Chapter 4 lists several accurate measurement methods that were very expensive. This would tend to violate the "practical" portion of the main goal. Likewise, the inaccuracy of the inexpensive methods are also not practical. Middle-ground must be found. This particular development will not be concerned with cost from the stand-point of mass production. This development is primarily a research vehicle and as such will not be more expensive than a production item. This will give some needed latitude in this development, but care must always be exercised in all development decisions regarding cost.

Accuracy is a very important consideration. Is 1% good enough? Perhaps 2% ? This is always a difficult question. Usually a balance must be struck between accuracy and cost. In this case, however, accuracy will be dictated by the accuracy of the signals used to produce the test. If the reference signal is accurate to within 1%, for instance, it would be impossible to measure anything against this and achieve accuracy greater than 1%.

The method of application is a broad catch-all term that includes such things as the operating environment and the actual physical make-up of the device. This device will be used primarily in a CATV headend to adjust CATV modulators. Most CATV headends have very good environmental control, so a wide operating range is not necessary. CATV is viewed as a service, and as such the test method should not be intrusive. The test method should be simple and repeatable. In the "real world," a test method that is very accurate but very complicated is usually quickly displaced by a less desirable but easier method. This could be referred to as the "human nature" aspect of the design development process.

Marketability is a test that really applies very little in this particular case. As previously stated, this development is basically a research tool. However, one should not discount the value of this development in leading to a potentially marketable device.

Having established the major concerns of the development, it is now time to proceed with the particulars. First, what method will be used? Figure 5-1 is a block diagram of a CATV modulator. By examining the function of this device, a measurement method may be uncovered. Is the input video signal a potential source? Chapter 4 explained that most modulation meters use this input signal to measure modulation. By deriving the proper

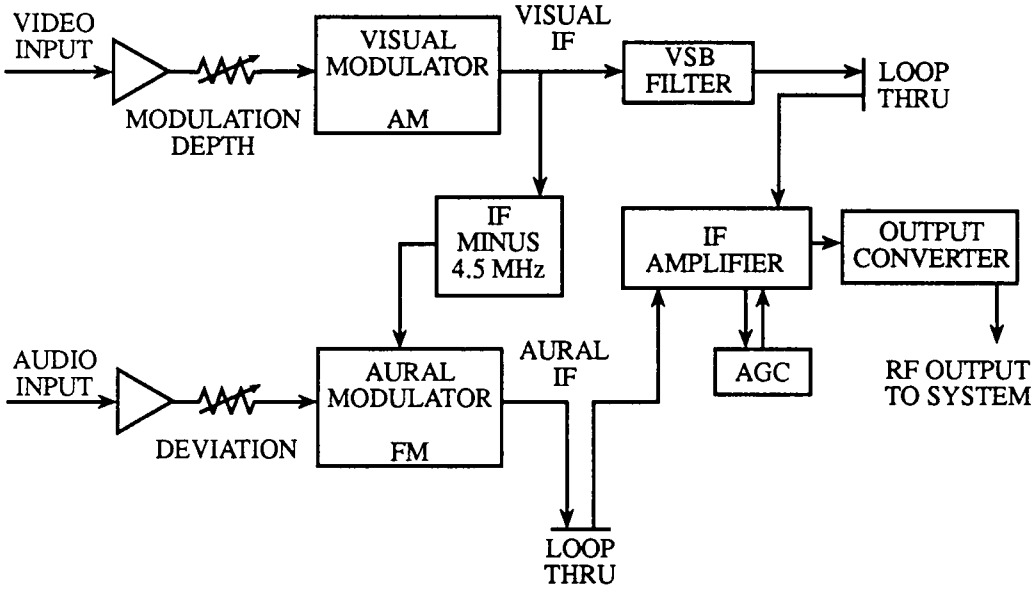


FIGURE 5-1

A BLOCK DIAGRAM OF A CATV MODULATOR

transfer function, it would be possible to relate the level of modulation output from the modulator to the level of the input signal. The largest obstacle to this approach is that this

transfer function would be entirely different for each modulator tested. This would necessitate that the measuring device be initially calibrated with each modulator. This assumes that a highly accurate measuring method already exists, which would indicate that there is no need for this device.

The video input appears to be a "dead-end," so on through the modulator to the IF. Chapter 4 also treated direct IF examination. Direct "board-level" IF probing does not meet the practicality requirement. Using the IF loop-thru connections is possible, but this must be done without intrusion. This would necessitate the use of directional couplers in the IF loop-thrus of all the modulators. This is still within the realm of practicality, but may incur the "human nature" reaction due to the increased test complexity. This gives the first potential test signal; the IF loop-thru. It is important to continue to see if an even better test signal exists.

The next readily available signal is the actual channel output of the modulator. Most CATV modulators have 75-ohm test points in addition to the output ports. This is a built-in, non-intrusive test point that is ideal for the purposes of this development. The output signal has the disadvantage of being at a higher frequency than the IF signal. The output can be one of over 60 discrete channels. This makes a tuning device mandatory. The output also contains a sound carrier and color information, which further complicates measurement methods.

At this point the choices seem very slim; either IF or channelized RF for the signal under test. At least these approachs avoid the problems of nonlinear modulation. However the IF test signal is not subject to the sync compression that can occur in the output stages. Sync compression is a situation that occurs when the output is driven to its maximum. In the case of television, sync tips represent the maximum output signal level with all other information being below the sync tip levels. When the output gain is pushed to the maximum, the output devices cannot amplify the sync tips any more, but there is still dynamic headroom available to amplify the other information that exists below the sync tips. This has the effect of holding the sync tips at the maximum output level and allows the remaining information

to "gain" toward the sync tips. This appears to compress the sync (when in reality the sync is stationary and the rest of the information is expanding toward the sync tips). With this in mind and the complexity of physical IF hook-up, perhaps RF output monitoring is the best test signal available.

With the choice of a test signal made, a method of testing this signal must be found. The most logical place to begin is with the output signal straight from the modulator. At this point it will be assumed that the sound carrier is suppressed; that the modulated signal is monochromatic; and that no harmonics exist in the output signal. These conditions make the output signal as simple as possible. Figure 5-2 is reproduced from Chapter 4 and represents the output signal, subject to the prior restrictions. Perhaps this visualization will aid in determining a measurement method. A variety of percentages are labeled on Figure 5-2 corresponding to sync tip, blanking, and peak white levels. All of these measurements are referred to the zero carrier level. A first thought might be to attempt to measure the envelope, much like the direct RF measurement method presented in Chapter 4. Perhaps this warrants some investigation.

Direct RF measurement methods are few in number. The techniques that do exist have one thing in common: indirect measurement.[29] It seems contradictory to speak of a

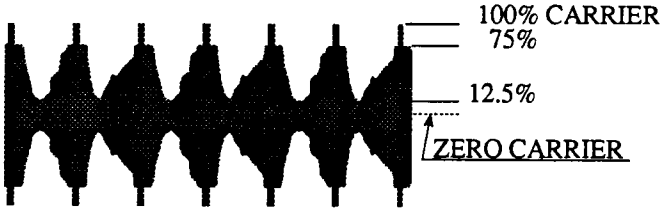


FIGURE 5-2

A GRAPHIC REPRESENTING THE TELEVISION RF ENVELOPE

direct measurement being made indirectly, so an explanation is in order. At this point in time there exists no technique or device that can directly make RF measurements. Most RF measuring techniques convert the RF signal into a more readily measured quantity. The simplest RF measurement method is the thermocouple or bolometer method. In this method a thermocouple is affixed to an RF terminating resistor. As the RF is incident on the termination it tends to heat as it dissipates the RF energy. The heating of the termination is usually very small, but instrumentation does exist that can measure the low signal levels we are dealing with. The indirectness here is that the RF is converted to heat and the heat is the measured quantity. Although devices do exist to measure these low signal levels, the thermocouple's response to changes in the heating level is very slow when compared with the rate of change of the video signal. This effect leads to a response that tends to be an average of the incident RF signal level. This is not a measured quantity in which we are particularly interested. Perhaps the answer lies in another method.

Probably the most popular method of RF measurement is by the use of "square-law" devices. A square law device has an output voltage or current that is proportional to the input RF level.[30] The simplest of such devices is the diode detector. Figure 5-3 is a diagram of a simple diode detector.

In a matched transmission line situation such as this, the output can easily be made proportional to the input power or voltage level. A bias network assures that the diode is biased in it's square-law region. As the RF incident on the diode increases, so does the diode current and vice-versa. If the capacitor C in Figure 5-4(a) is chosen to be large, it will tend to average the current changes and present an average value at the output. If C is removed, the output will be a rectified version of the input signal, again with a frequency too high for direct measurement. If a value of C is chosen somewhere between these two extremes, most of the input level changes will appear in the output. Figure 5-4 is illustrative of these

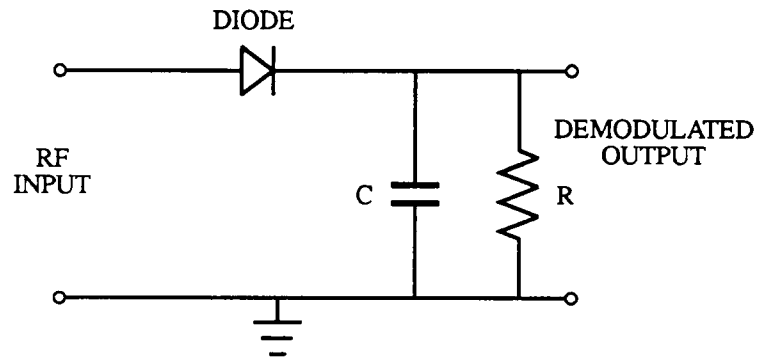


FIGURE 5-3

A SIMPLE DIODE DETECTOR

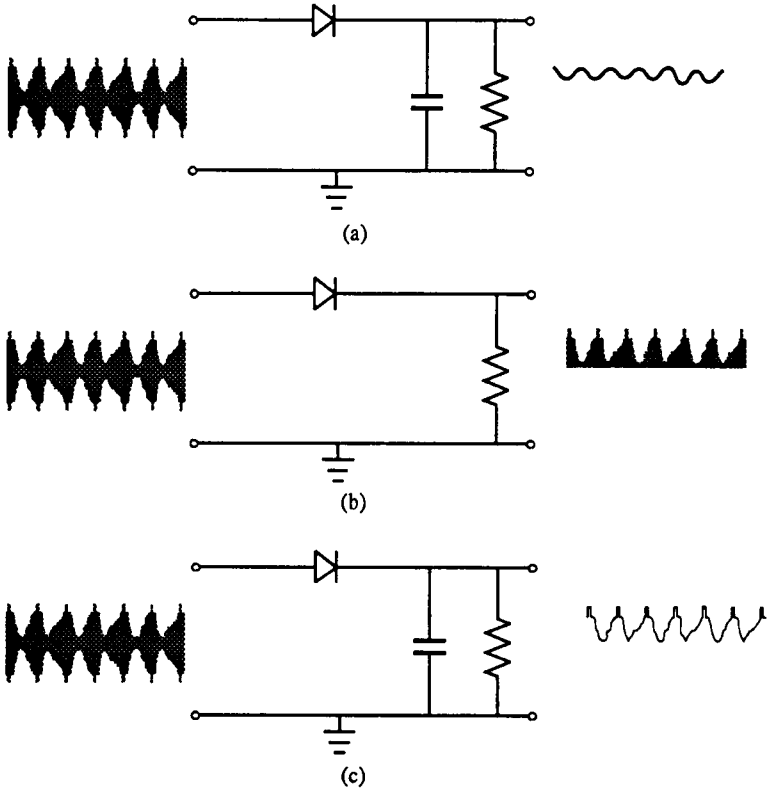


FIGURE 5-4

THE EFFECTS OF CAPACITANCE ON THE OUTPUT OF DIODE DETECTORS

situations. From Figure 5-4 it is obvious that Figure 5-4(c) will allow the measurement of the RF envelope levels.

The conversion method described above is usually referred to as detection or demodulation. Detection is used to separate the modulating information from the carrier in AM systems. It appears that some form of demodulation is called for in this case. It will also be necessary for the output to follow the input very closely to get accurate measurements.

A demodulator is needed, so some thought must be given to it's design and function. Perhaps the first consideration should be noise. In a coaxial system, proper operation requires that the transmission lines be terminated resistively. All resistors have thermal noise associated with them. This thermal noise will form a "noise floor," or a minimum noise level which is inescapable. Resistive thermal noise is given by equation 5-1.[31]

$$N_r = 4KTR\,\Delta f \qquad (5\text{-}1)$$

In equation 5-1, K is Boltzman's constant, T is the temperature in Kelvin, R is the resistance is Ohms, and Δf is the bandwidth of interest. Care should be taken in applying equation 5-1 to noise calculations due to the Δf term. Equation 5-1 is really an expression of the spectral density of the noise and as such Δf is not as straight forward as might be perceived. In CATV design the Δf bandwidth is assumed to be a "brickwall" bandwith, and although not physically realizable, it does provide a relative figure of merit measurement. Historically, the FCC and TASO set this bandwidth standard at 4 MHz, and indeed most television receivers have video passbands of about 4 MHz so the Figure has "real world" applications. The CATV industry has adopted this bandwidth for all CATV uses, and all test instrumentation is either designed with this in mind or correction factors are given to accommodate this standard.[32] Equation 5-1 is solved in equation 5-3 yielding:

$$N_r = 4\,(1.38 \times 10^{-23} \frac{J}{°K})\,(300\ °K)\,(75\Omega)\ \Delta f \qquad (5\text{-}2)$$

$$N_r = 1.242 \times 10^{-18} \frac{nV^2}{Hz}\,(4\ MHz) = 4.968\ pV^2\ . \qquad (5\text{-}3)$$

Using the answer found in equation 5-3 and converting it to the CATV standard of dBmV occurs in equation 5-4. The noise voltage is divided due to the 0.5 voltage gain of a matched transmission line.

$$dBmV = 20 \log \frac{V}{1\ mV} = 20 \log \frac{\left(\frac{\sqrt{0.4968\ nV^2}}{2}\right)}{1\ mV} = -59.06\ dBmV \qquad (5\text{-}4)$$

This establishes the noise floor at about -59 dBmV. Will this be a problem in the demodulation process? The operating environment of the modulator will provide the answer.

In analyzing the effects of noise, some investigation into the modulator is necessary. Most CATV modulators are located in a CATV headend and as such are high output devices. Typical output levels range from about 40 dBmV to in excess of 60 dBmV. Output distortions are minimized by the fact that these are single channel devices, thus allowing these high output levels. Most CATV modulators have easily accessible test points. These test points are usually internal directional couplers placed directly in the output path with the tap being 30 dB below the output signal. This gives an output test point level between 10 and 30 dBmV and assuming a perfectly noiseless modulator will yield the maximum signal-to-noise ratio of 10 - (-59) = 69 dB. The noise floor input to the modulator is also -59 dBmV because standard video impedance is also 75-ohms. From this perfect situation, expansion can be

made to account for devices with noise. Most modulators will be fed a video signal from a satellite receiver. A well-designed satellite reception system is capable of very high signal-to-noise ratios. This will not be the noisiest input signal. Local origination signals and off-air intercepted microwave links are probably the noisiest input signals. For computational purposes, assume a local origination system with a very poor signal-to-noise ratio of 45 dB. Modulator noise figure is almost a contradiction of terms since the input signal is video and the output is modulated RF. However, for the sake of computation, assume the modulator contributes noise with a 6 dB noise figure. The output carrier-to-noise ratio is now 39 dB. This carrier-to-noise ratio is still large enough to insure reasonable accuracy, but this figure is very low by CATV standards.

Perhaps a different approach to the noise problem is needed. Beginning with the CATV customer, a carrier-to-noise ratio of 43 dB will have visible noise in the picture. This should be an unacceptable situation in the CATV system. Most CATV designs target customer signal-to-noise ratios at a minimum of 45 dB. If the customer signal-to-noise ratio must be 45 dB, the carrier-to-noise ratio at the headend must be greater. The reason it must be greater is that as the signal is distributed from the headend, carrier-to-noise ratio degradation occurs at every amplifying station. In practical system design, a headend carrier-to-noise ratio greater than 55 dB is required to maximize cascade length and still insure a reasonably high quality level of service to the customer.

Carrier-to-noise ratios greater than 50 dB can be expected at the headend test site. This means that test point output may be attenuated to 0 dBmV before the noise floor has any effect on the signal-to-noise ratio of the test signal. By maintaining this signal level situation, noise should not be a factor in the demodulator design.

The noise considerations have also provided some insight into the possible range of input signals. Test point outputs range from 10 to 30 dBmV, and at no time should be allowed

to be lower than 0 dBmV. This is a good time to take a development inventory. Figure 5-5 is a block diagram of the progress up to this point.

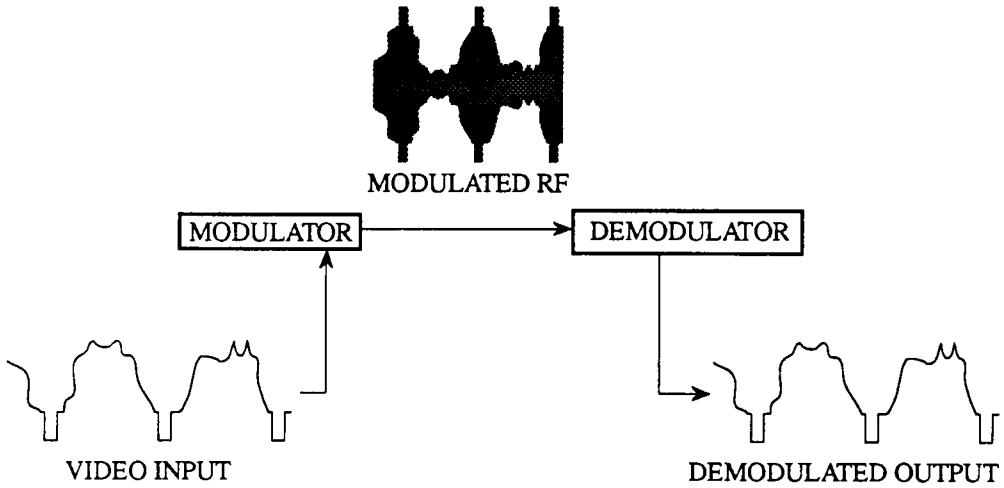


FIGURE 5-5

A BLOCK DIAGRAM OF THE OUTPUT REQUIRED BY THE MODULATION MEASUREMENT SYSTEM

As indicated in Figure 5-5, the demodulator output is proportional to the RF envelope, or more accurately half of the RF envelope. Now, how can this proportional output be used to measure modulation depth? The most obvious answer is to use the zero carrier reference to measure the ratio between the sync peaks and the peak white. This appears to be a simple task. This is where a good look at the video signal is in order. For this test method to work accurately, the demodulated output should resemble the input of the modulator as closely as possible. Figure 5-6 illustrates the video output of such a demodulator.

Notice that the sync tips of Figure 5-6 are well-defined and periodic in nature. A sync tip appears every 63.5 microseconds, and every 31.75 microseconds during the vertical

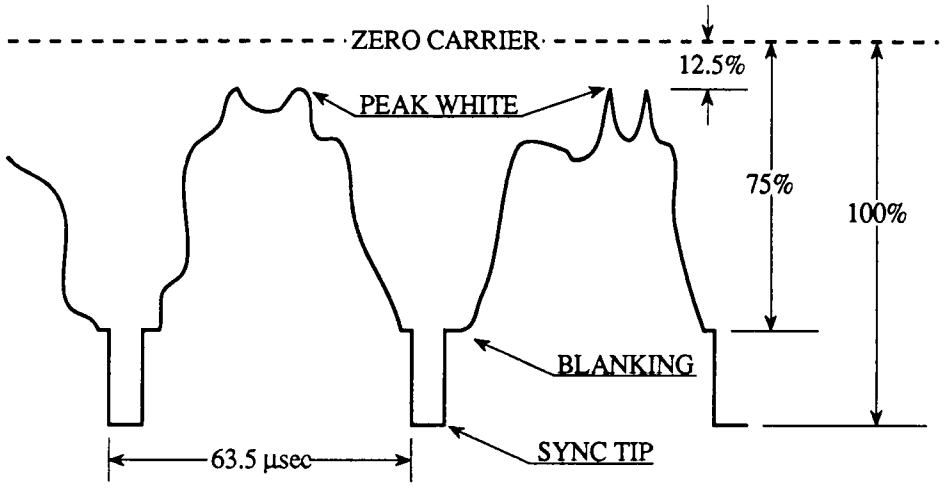


FIGURE 5-6

NTSC VIDEO TIMING AND AMPLITUDE STANDARDS

interval. This should be a reasonably easy quantity to measure. Peak white levels are not as straight forward. The video content that lies between the sync pulses is random in nature. Several lines can pass without a peak white level existing. The entire vertical interval has no such peak white occurrence (with the exception of the vertical interval test signals). White peaks may be a very elusive level to capture and measure. This will probably call for a very sophisticated peak-hold circuit. Perhaps an alternative exists.

The sync tips were ideal for measurement because of their periodic nature. Another periodic signal is the blanking level adjacent to the sync pulses. The front and back porches comprise this level which is collectively referred to as blanking. By FCC standards, blanking level should be 75% of sync tip level, referred to the zero carrier level. This is illustrated in Figure 5-6. The sync tip, blanking, and peak white levels are usually very well controlled by the programming originators. Properly modulated video will preserve this relationship when demodulated and measured with respect to the zero carrier level. With this in mind, if the video program input to the modulator is well controlled by the originator, the sync tip

blanking ratio is an indication of modulation depth. This situation seems to place the burden of modulation depth control on the programming source. This is true of any modulation measuring system. If the source of video allows peak white levels to exceed the 12.5% limit, for instance, and the modulator is set with this falsely high white peak level, the demodulated signal of a television receiver will reflect this with a video signal that is out of proportion. In this case the sync pulse will occupy less than the normal 25% of the signal level. The television receiver expects accurately proportioned video and may respond to this video malady by either loosing sync or brightening dark scenes. Without a source of consistent properly proportioned video, modulation adjustment is moot.

It can now be assumed that sufficient information exists in the output of a demodulator to measure the depth of modulation. Further investigation of AM demodulation is warranted. Several methods of AM demodulation are available to the designer. The simple diode demodulator in Figure 5-3 is very misleading in that a real diode demodulator is much more sophisticated. The diode demodulator is the simplest of all demodulation methods, and as such suffers with the poorest performance. Diode input signal levels need to be about -20 to -10 dBm for linear performance. This is about 28 to 38 dBmV, and would indicate a need for gain prior to the diode. Diode demodulation suffers from an extremely non-linear response which is worst at low signal levels. A diode demodulator also suffers from a large conversion loss. These problems are overcome by significantly increasing the input signal levels. Input levels of 3 to 5 volts peak-to-peak are not uncommon in these designs. These large input levels necessitate pre-demodulator gains from about 44 dB to 64 dB, referred to the modulator test point output. Large RF gains of this magnitude are not trivial endeavors. The input level must also be very stable for linear performance. This necessitates the use of automatic gain control (AGC). Using the previous gains as a guide, the control range would have to be at least 64 - 44 = 20 dB. Additional range would be necessary to accommodate actual testing situations. For maximum AGC performance, it would need to be gated to the

sync tips. The sync tips represent the peak RF level and gating provides a means of using only the sync tips as a level reference. This makes the AGC immune to changes in the average picture level (APL), but also significantly increases the AGC circuit complexity.

The active transistor demodulator is an improvement on the diode demodulator in terms of performance. The transistor demodulator, however, is also more complex than the diode demodulator. Performance is even better in the quasi-synchronous demodulator and best yet is the synchronous demodulator.[33] The synchronous demodulator is by far the best performing demodulator available today. The price of this performance is extreme circuit complexity over and above the gain and AGC requirements of all other demodulator methods.

Taking all of the information on demodulation into consideration leads to the conclusion that demodulator design is beyond the scope of this development. Indeed, a well thought-out demodulator design could easily fill volumes and consume thousands of man-hours. Just a few years ago this would have been an insurmountable obstacle. The proliferation of home video in recent years has furnished a simple answer to the demodulator problem. All home video cassette recorders (VCR) are equipped with video outputs, as are a large number of modern television receivers. Stand-alone demodulator units are also available at a fraction of the cost of previously available demodulators. All of these devices have 75-ohm RF input impedances and standard 75-ohm video output impedances. They all also contain the tuners that are necessary for checking the large number of cable channels available. Purchasing one of these units does not violate the practicality test because most CATV systems probably own or have access to such devices. A development inventory of our progress is illustrated in Figure 5-7.

There is a price to be paid for this "off the rack" answer to the demodulator problem. As designers, we have lost control of the zero carrier level reference. In designing a demodulator, it probably would have been possible to preset a zero carrier reference. For

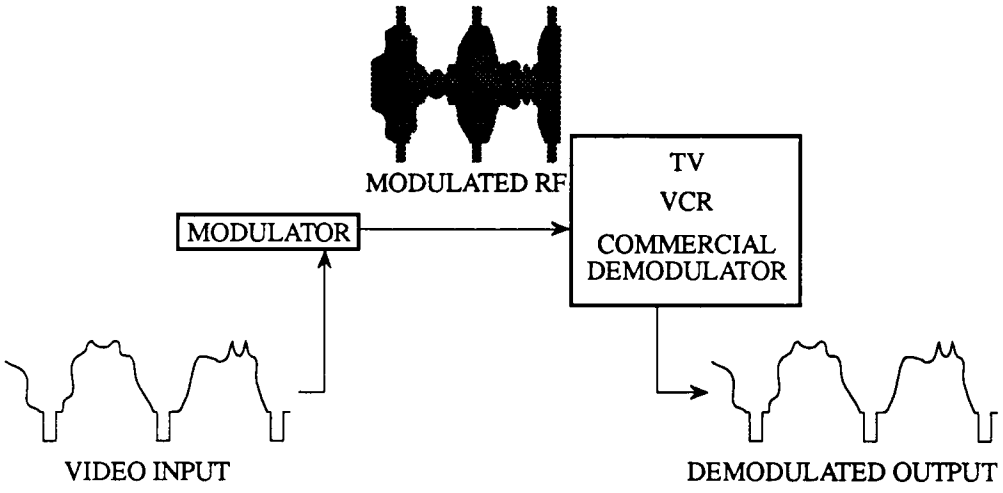


FIGURE 5-7

A BLOCK DIAGRAM OF A PROPOSED MODULATION MEASURING SYSTEM

AGC purposes, most commercial demodulators clamp the sync tips at a constant level. This allows the zero carrier level to "wander" between discrete maximums and minimums. Even if a zero carrier reference was output by the demodulator, it would probably be lost in the AC video coupling. All video outputs are AC coupled to prevent any interaction between the driver and the driven load. The zero carrier level would probably be represented by a DC offset voltage which is not available at the output.

As already mentioned, the video output of these demodulators is AC coupled. This means that the positions of the sync tips, blanking, and peak white, relative to ground, will vary with the average picture level (APL). Figure 5-8 is illustrative of the effect of APL on an AC coupled video output.

The bottom line is that the zero carrier level is now lost. Either an alternative reference or an alternative method of recreating the zero carrier reference must be found. A 1946 technical article by T.J. Buzalski [34] provides some insight into this problem. Buzalski writes of using a device similar to the zero-chop switch mentioned in Chapter 4 to preserve DC levels

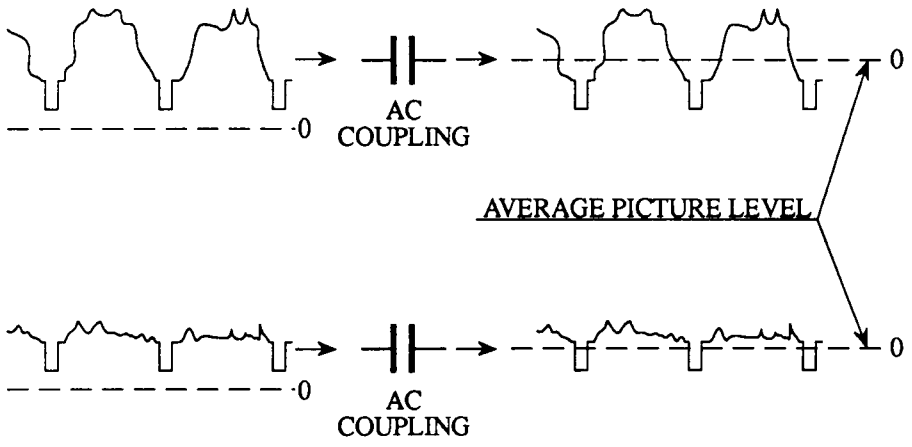


FIGURE 5-8

THE EFFECTS OF AC COUPLING ON AVERAGE PICTURE LEVEL

in AC oscilloscope amplifiers. By switching the input signal off and on an output level equal to the zero carrier will appear. By carefully controlling the switching process, the zero carrier level may be captured for reference purposes. Figure 5-9 shows the effect of such a switching operation on the input RF signal. If the switching-off time is short compared to the low frequency time constant of the output AC coupling, the reference will pass through the AC output coupling. If the switching operation occurs between sync pulses, the fundanental frequency of the pulse created by this operation is equal to the sync pulse frequency, about 15,750 Hz. This is a high frequency compared to the low frequency response of most AC coupled video outputs. Most AC coupled video systems will pass frequencies as low as 50 Hz.

The effect of the zero-chop switching on the output of the demodulator should resemble Figure 5-10. Figure 5-10 is a very optimistic representation of what will actually occur when a television or VCR is subjected to such a switching operation. Careful thought

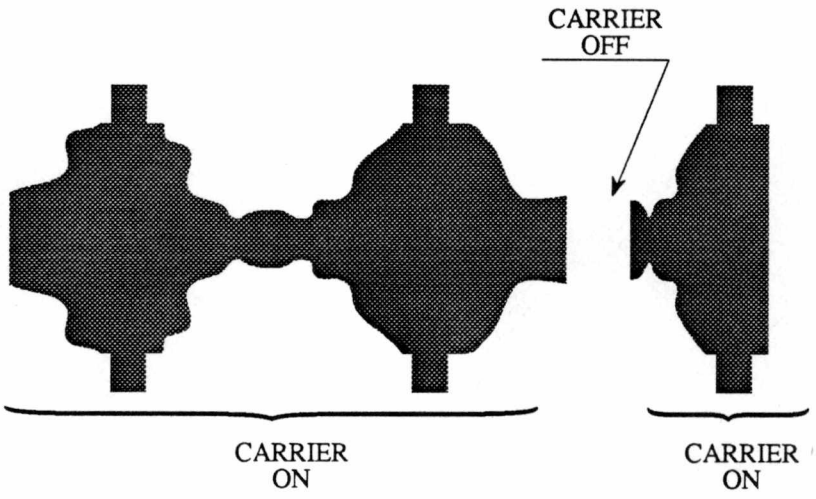


FIGURE 5-9

THE EFFECT OF ZERO-CHOP SWITCHING ON THE RF ENVELOPE

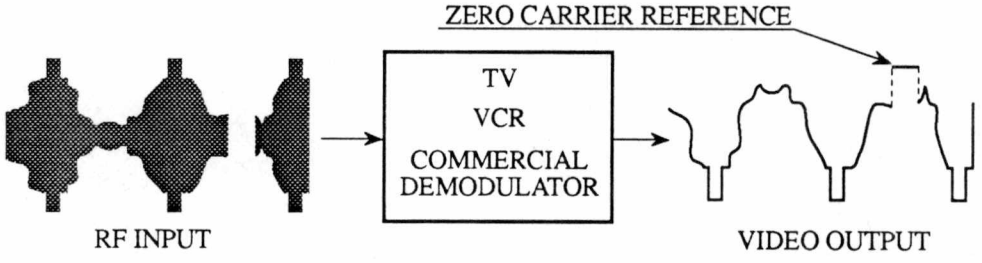


FIGURE 5-10

THE EFFECTS OF ZERO-CHOP SWITCHING ON DEMODULATED VIDEO

will need to be given to the design of the switching apparatus to avoid interfering with the operation of the demodulator.

All modern VCRs, televisions, and commercial demodulators use sync-gated AGC. The merits of such a system have been previously discussed. If the switch were to open during one of the AGC gating periods, the AGC would lock on the zero carrier level and react by turning the system gain up as high as possible. This would probably saturate the demodulator

until the next AGC gating period could correct this over-gain situation. Some demodulators will also gate on the equalizing pulses found in the vertical interval. These too must be avoided. A commercial zero-chop demodulator avoids this problem by preventing AGC gating during the vertical interval. The switch is opened during the vertical interval. This open period is long enough that electro-mechanical switches are usually used to provide the switching. Typical "off" switch states are 400 microseconds long. When the switch is returned to the "on" state, the demodulator again allows AGC gating to resume. Controlling the gating of the AGC of the demodulator is beyond our ability since we are not designing the demodulator. An alternative answer to this problem must be found.

Figures 5-9 and 5-10 may be somewhat prophetic in the way that they show the theoretical switching operation. In both instances the switching is completed in the interval between sync pulses. If switching can occur between the sync pulses, the demodulator might be able to recover before the next gated-sync pulse. This would avoid interaction with the AGC system, as any such interaction could invalidate the data produced. Three distinct problems surface almost immediately: switch design, timing control, and demodulator reaction to zero-chop.

The video sync pulses shown in Figures 5-9 and 5-10 are 63.5 microseconds apart and switching must take place within this time frame. Subtracting out the sync pulse width and the blanking width leaves about 52.5 microseconds. A 10 microsecond switch off-time would probably be a good first choice as a balance between the time needed to capture the zero carrier reference level and the time needed for demodulator recovery. A 10 microsecond switching time is too short for any electro-mechanical switching devices. The switch must then be electronic. Electronic switching methods include: field-effect transistors (FET) switches, diode switches, bipolar junction transistor switches, and metal-oxide semiconductor (MOS) switches. The choice of switching method can be made by analyzing the situation in which it will be applied.

Figure 5-11 is a block diagram of the zero-chop switch application. From this diagram it can be seen that by the placement of the switch it must be capable of switching the RF signal, which can range from 50 to 500 MHz. This places quite a demand on the switching device. The switch is also in a 75-ohm transmission line environment and as such it must match the 75-ohm input and output impedances. Another concern is that the switch preserve the impedance levels throughout the switching operation. This is a very stringent requirement, but it is necessary to insure that no interaction can exist between the switch and either the input driver or the output demodulator. It is impossible to predict the reaction of either device to a sudden change in impedance levels, but any reaction could be such that it invalidates the accuracy of the measurement. This is but one of the problems arising from using an "outside" demodulator. The only option is to avoid any possibility of interaction by preserving the constant impedance levels at both the output and input of the switch.

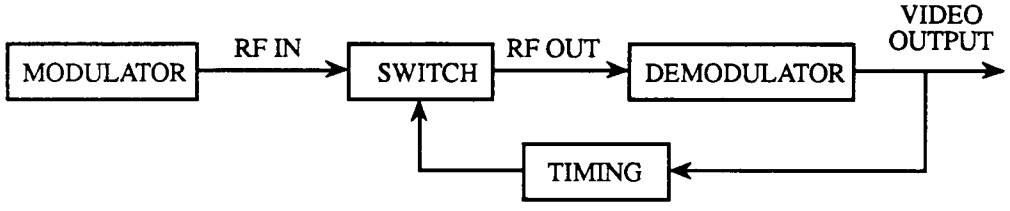


FIGURE 5-11

A BLOCK DIAGRAM OF A PROPOSED MODULATION MEASURING SYSTEM

Research indicates that these requirements reduce the number of control devices to two: FET switches and PIN diode switches. An FET constant impedance switch can be constructed by using special RF FETs with very low "on" resistances. These "on" resistances can be as low as 25-ohms. There are two major problems with the FET switch. The first

problem is that all FETs exhibit a tendency of generating a large transient "spike" upon switching.[35] This spike could easily interact with either the input or output devices. The other problem with FET switches is that the devices are rather exotic and tend to be difficult to procure. The major advantages of FET switches are simplicity of design and very low switching current requirements.

A PIN diode is a very specialized device. It is manufactured to appear as a diode at low frequencies and appear as a resistance at high frequencies. The construction of the PIN diode is shown in Figure 5-12

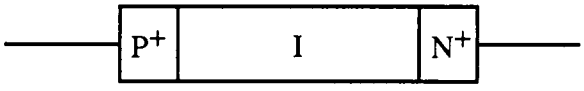


FIGURE 5-12

THE CONSTRUCTION OF THE PIN DIODE

A PIN diode is constructed much the same as a PN junction diode except that a large section of intrinsic material separates the P and N materials. This is the origin of the PIN nomenclature. The P and N layers are heavily doped and usually referred to as P+ and N+ layers. At high frequencies, in a no-bias or reverse-bias situation, the intrinsic layer is highly resistive due to it's physical size and intrinsic composition. As forward bias is applied, the P+ and N+ regions inject minority carriers into the intrinsic region. This causes a drop in the resistivity of the intrinsic region. RF resistivity changes of 10,000-ohms to 1-ohm are possible with diode current changes of 1 microampere to 100 milliamperes, respectively. The PIN diode's operation at low frequencies is primarily attributed to the lifetime of the carriers.[36]

The PIN diode's application as a switch or attenuator should be obvious by the large RF resistance changes possible. The simplest applications of PIN diodes are as either series or shunting control elements. This use is very popular in microwave design, but presents a severe impedance mis-match. A double-pole double-throw arrangement may be used , but a severe mis-match occurs during the switching interval. The only design that continously matches the input and output impedance is an attenuator design. The first thought may be that an attenuator is a poor quality switch. However the converse is also true; a switch is just a very good attenuator. Switch quality is usually specified as a specific number of dBs of isolation, which is the same number of dBs of attenuation. An attenuator capable of a high level of attenuation should meet our needs.

Attenuator designs that continously match transmission line situations are usually limited to two: "π" and "T" networks. Figure 5-13 illustrates both networks. In both networks, values of R_p and R_s can be selected that achieve input and output matching. In Figure 5-13(a), as R_s approaches 75-ohms and Rp approaches a very large resistance, most of the input current I_1 is shunted through R_s. The small current through R_p is indicative of the

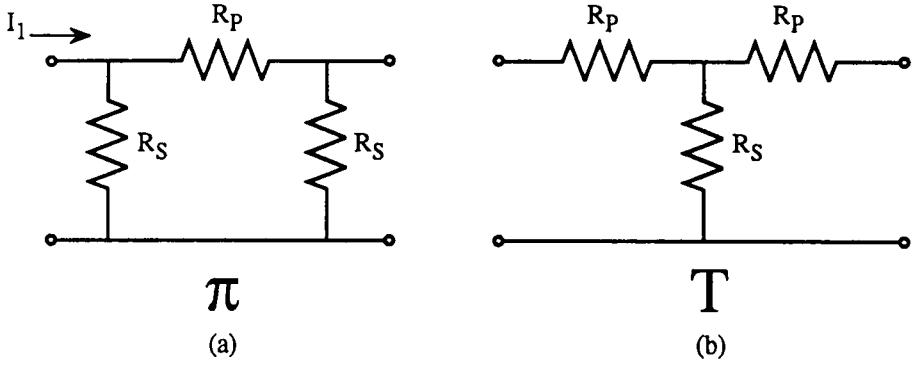


FIGURE 5-13

"π" AND "T" CONSTANT IMPEDANCE ATTENUATOR CIRCUITS

attenuation this case provides. At the other extreme, if R_p approaches zero-ohms and R_s becomes very large, the loss through the attenuator becomes very small as most of I_1 will pass through R_p to the output. The T-network behaves in a similar manner. Table 5-1 lists values of R_s and R_p for several attenuation values with an input and output impedance match of 75-ohms.

TABLE 5-1

A COMPARISON OF THE RESISTANCE VALUES REQUIRED FOR A PARTICULAR LEVEL OF ATTENUATION IN A π OR T ATTENUATOR

	π		T	
Attenuation (dB)	R_s	R_p	R_s	Rp (ohms)
0	∞	0	∞	0
0.2	6465.500	1.725	3255.000	0.900
3	438.600	26.400	212.850	12.825
6	225.750	56.030	100.350	24.900
10	144.375	106.725	52.725	38.925
20	91.650	371.250	15.150	61.350
30	79.875	1184.850	4.725	70.425
40	76.500	3750.000	1.500	73.500
50	75.450	11859.600	0.450	74.500

Applying PIN diodes to the circuits of Figure 5-13 yields Figure 5-14. Notice that each resistor is now replaced with a PIN diode which is characterized as a variable resistance at high frequencies. If the values of the resistances of the PIN diodes are varied in the same

manner illustrated in Table 5-1, a variable attenuator is constructed with constant input and output impedance. To control the change in resistivity, the current flowing in each PIN diode element must be controlled.

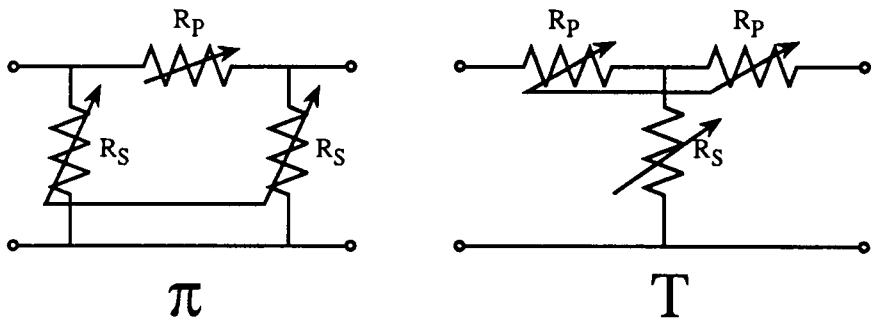


FIGURE 5-14

CIRCUIT DIAGRAMS OF CONSTANT-IMPEDANCE VARIABLE ATTENUATORS

Examining Table 5-1 carefully shows that a low value of resistance in some of the PIN diode elements is required for a low attenuation level. This is true for both π and T-networks. For a very large attenuation the T-network also requires a low value of resistance of some of the PIN diodes. The π-network relaxes that requirement at high attenuation levels. Designing a π or T PIN diode attenuator requires tailoring the PIN diode currents to give the correct resistivity, with respect to each other, to meet both attenuation and impedance requirements. The matching requirements of the π-network are more stringent at low attenuation levels than at high attenuation levels. The matching requirements of the T-network are stringent at both attenuation extremes and have smaller resistivity changes than the π-network. These facts make the π-network somewhat easier to design. A bridged T-network can overcome some

of the difficulties in matching the required resistance values, but the π-network is still somewhat easier in application.

The choice of an attenuator as a switch is not without complications. The question, "How much attenuation is necessary to simulate a good switch?" must be answered. Once again analysis begins with the quantity of measure in which we are most interested: video. As previously illustrated, video has four distinct levels associated with it: sync tip, blanking, peak white and zero carrier. Figure 5-15 is a repeat of previous illustrations providing information on the video expected from the demodulator. The sync tips represent the largest signal excursion of both the video and the modulated RF carrier. This level has been denoted as the 100% level. This gives a blanking level of 75%, peak white level of 12.5% and a perfect zero carrier level of 0%. Perfect zero carrier level is the result of perfect switching, which is an attenuation level of ∞ dB. The question at hand is how much practical attenuation is necessary to preserve the accuracy of this testing method?

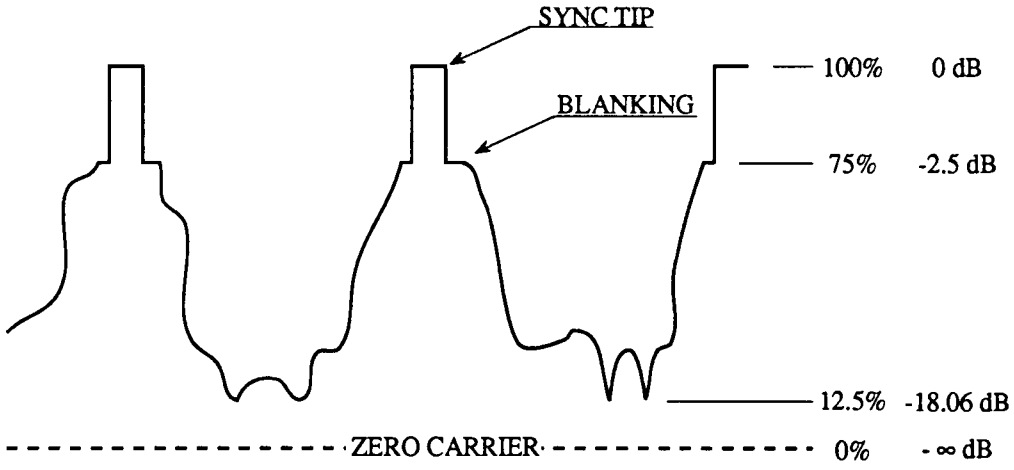


FIGURE 5-15

NTSC VIDEO AMPLITUDE SPECIFICATIONS EXPRESSED IN dB

$$dB = 20 \log \frac{V_1}{V_2} \qquad (5\text{-}5)$$

In using Figure 5-15, assume that the percentage levels represent voltages. If such is the case, the dB is defined by equation 5-5.

By substituting in the relative percentages for V_1, and setting V_2 equal to 100%, an expression is derived to relate the difference in dB between the levels shown in Figure 5-15. Equation

$$dB = 20 \log \frac{V_1\%}{100\%} \qquad (5\text{-}6)$$

5-6 includes this change.Equation 5-6 is used to compute the dB levels shown in Figure 5-15.

In a real attenuator, the zero carrier cannot be ∞ dB below the sync tips. The task is to determine the error introduced by a zero carrier level that is less than ∞ dB below the sync tips. Assume, for instance, that instead of 0%, the zero carrier is incorrectly placed at 1%, relative to the real zero carrier level. Our proposal has been to use the relationship of the sync tip level and the blanking level as an indication of the modulation depth. In a properly modulated situation of Figure 5-14, the ratio of sync tip to blanking is 1:0.75, referred to the 0% zero carrier level. If the zero carrier is incorrectly established 1% above the actual location of the zero carrier level, the sync tips now measure 99% and the blanking is measured as 74%. This gives a 1:0.747 ratio. This is a 0.34% error compared to the 1:0.75 ratio. Can this error be tolerated?

Once again the operating environment needs to be analyzed to get a reference of comparison. The FCC requires that television broadcasters meet the following requirements:

Peak white is 12.5 (± 2.5) % of sync tip level and blanking is 75 (± 2.5) % of sync tip level. This wide tolerance may seem to be "sloppy" by most engineering standards, but most broadcasters hold the tolerance to ± 1%.[37] This higher level of accuracy is primarily achieved by an increase in the level accuracy of the modulating video signal the broadcaster uses. Most CATV programming sources use equipment that is equivalent to this to control their video, so similar performance can be expected. This means that a ± 1% error in the blanking and peak white levels can be expected in practice. This limits the accuracy of any measurement method. Although the accuracy of the video measurement could be greater than ± 1%, the accuracy of the overall measurement is dependent on the accuracy of the video levels created by the programming source. In this particular case, any inaccuracy introduced by the measuring method will add to the ± 1% error expected in the input signal. Remember that the FCC requirements are relative to the sync tip level. Equation 5-7 is a computation of this error in terms of the 1:0.75 sync tip to blanking ratio that will be used to compute the depth

$$\text{ratio error} = \left[\frac{\left(\frac{0.75 \pm .01}{1}\right) - 0.75}{0.75}\right] \text{X } 100 = \pm 1.333\% \qquad (5\text{-}7)$$

of modulation. Equation 5-7 illustrates that a ± 1% error in the input video signal will create a ± 1.333% error in the measuring method we have chosen. Now a comparison of this error needs to be made to the error that can be introduced by an imperfect switch. An improperly positioned zero carrier level will induce error in both the sync tip and blanking levels.

$$\text{ratio error} = \left[\frac{\left(\frac{0.75 \pm .01 - Z_e}{1 - Z_e}\right) - 0.75}{0.75}\right] \text{X } 100\ \% \qquad (5\text{-}8)$$

Equation 5-8 takes this into account with the added zero carrier level error labeled as Z_e. In equation 5-8 the zero carrier level error, Z_e, is subtractive only. This is because of the unusual situation of the additive error being the impossible situation of attenuation greater than ∞ dB. This causes the error expression to no longer be symmetric about zero. For instance, the extreme tolerances compute a ratio error of + 1.01, - 1.68 % for a zero carrier level error of 1%. Although this choice of Z_e increased the negative error, it also decreased the positive error. The overall error range has also increased very little. The extreme tolerances for Z_e = 2% are + 0.68, - 2.04 % ratio error. This is a maximum measurement error of 2% taking all sources of error into account. This is a reasonable minimum performance level to aspire to. Using equation 5-6 to convert 2% zero carrier error into dB yields a zero carrier level that is 33.98 dB below the sync tips. For a 1% zero carrier level error, it is 40 dB below sync tip level. In a somewhat round about way, the attenuation capabilities of the PIN diode switch have just been specified. The minimum acceptable level of attenuation is 33.98 dB, with accuracy improving with any increase in attenuation. The utility of using the dB scaling is that the attenuation value is not dependent on the input or output levels, but is relative to them. Assuming the AGC of the demodulator does not interfere, a 33.98 dB attenuated input level should correspond to a 33.98 dB attenuation in the output level and create a zero carrier level

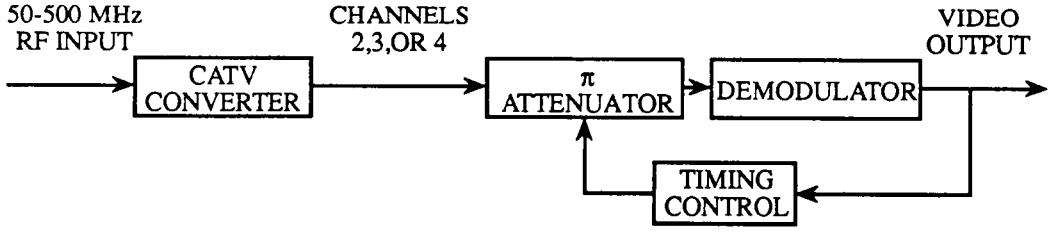


FIGURE 5-16

A BLOCK DIAGRAM OF A PROPOSED MODULATION MEASURING SYSTEM

that is within 2% of the actual level.

This now specifies the zero-chop switch. Figure 5-16 details the progress to this point.

The CATV converter is added to the input of Figure 5-16 for several reasons. The CATV converter downconverts the incoming signals to either channels 2, 3, or 4. Designing a PIN diode attenuator that performs well at this relatively low frequency range is considerably easier than designing an attenuator for a 50 to 500 MHz frequency range. The upper operating frequency of CATV systems is continually increasing and an input converter will insure that this design will perform regardless of any future increases in upper operating frequencies. An input CATV converter also relaxes the requirements of the tuner used in the VCR, television, or commercial demodulator. Some CATV systems use signal scrambling methods that require use of a special CATV converter to recover the programming. This allows the decoded signal to be tested to insure proper operation of the decoding process. Lastly, CATV converters are inexpensive, common-place items in most CATV systems, so using one does not violate the practicality requirement of this development.

Figure 5-16 also indicates the next direction of development: timing control. It has previously been mentioned that the zero-chop switch must not operate during any portion of either the horizontal sync pulse or an equalizing pulse. If this were to occur, the AGC of the demodulator would be seriously upset. The timing control will have the duty of insuring that this situation cannot exist. A method must be found to allow the timing control to detect these pulses. To answer this, once more we must journey into the operating environment.

Chapter 1 was a discussion of television and it's operation. Mention was made that detection of horizontal sync pulses was easily accomplished by differentiating the video signal. This method also detects equalizing pulses. Vertical pulses may be detected by integrating the video signal. The back porch may be detected by the falling-edge of the horizontal sync pulse. In television, this process is collectively referred to as sync separation.

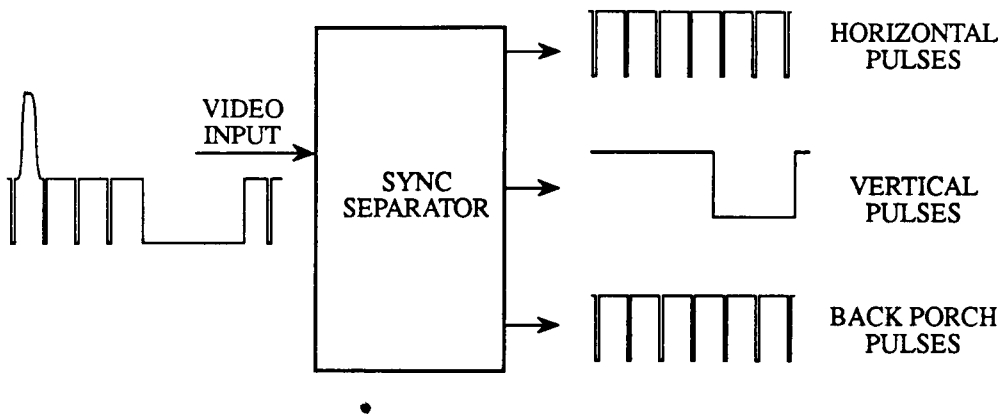


FIGURE 5-17

THE OPERATION OF A SYNC SEPARATOR

Numerous circuits exist to accomplish these tasks, and some integrated circuits will perform all of these tasks in a single package. Figure 5-17 is a block diagram of the output of a sync separator.

The outputs of the sync separator of Figure 5-17 are used as triggers for the timing control. First, a description of the timing control functions. The timing control must turn the PIN diode switch "off" for about 10 microseconds. It would be convenient if some easy method of adjusting this "off" time could be included in the design. The timing control must be able to "place" this "off" period between sync pulses, somewhere in the active picture area. The placement of this "off" period should also be adjustable. The method used for timing control will depend on personal biases. A digitally oriented answer is to use the vertical pulse to start a counter. This counter would count horizontal sync pulses. The counter could be programmed to overflow just prior to the next set of equalizing pulses, and wait for the vertical sync pulse to again initiate counting again. This would avoid any collisions with the equalizing pulses. Control of the placement of the "off" time within the active picture area may then be done in at least two ways. A digital counting method would be to phase-lock

an oscillator to the horizontal pulse. The oscillator would need to operate about 64 times higher than the horizontal sync rate. This would insure a "placement" resolution of about 1 microsecond. A counter is again employed to count the pulses of this oscillator and programmed to overflow at the desired "placement" area within the active picture area. When this counter indicates the proper "placement" of the off state, a "one-shot" can be triggered that will provide control of the off time. Alternatively, a counter could be employed to hold the off state of the switch a predetermined number of oscillator pulses before returning to the on state. This digital answer is very accurate, but is also very complicated.

Another equally accurate, but significantly simpler method is to use two "one-shots." This is a good time to explain the function of a one-shot. A one shot is more accurately referred to as a monostable multivibrator. It is a device that will remain in its preferred, or stable, state if allowed to. By triggering the one-shot, it reverts to its unstable state for a period of time that can be predetermined. The unstable state usually produces an output in the form of a pulse. The horizontal oscillator of most television receivers use the one-shot principle. Upon receiving a horizontal sync pulse, the horizontal oscillator will go through a single sweep of it's unstable state, after which it will return to the stable state and await another

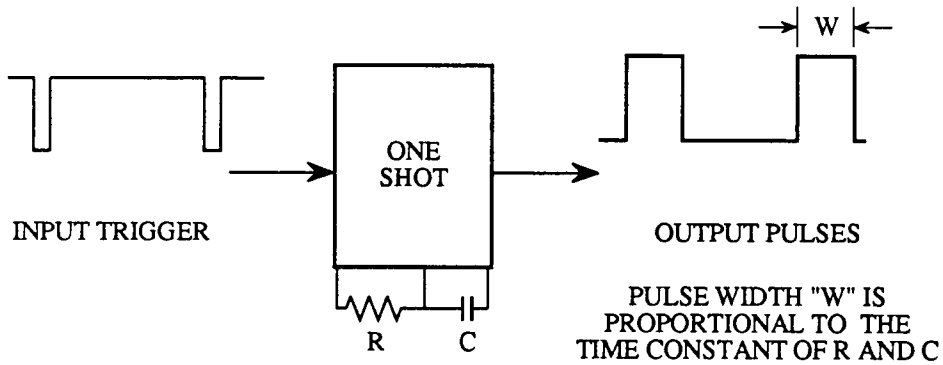


FIGURE 5-18

THE INPUT-OUTPUT RELATIONSHIP OF A ONE-SHOT

horizontal sync pulse. A simple block diagram in Figure 5-18 illustrates the input and output relationship of a one-shot. Most one-shots use a resistor and capacitor in combination to determine the frequency, or length, of the unstable state.

In this development, a one-shot can be triggered by the horizontal pulse output of the sync separator. The one-shot then remains in the unstable state until the active picture area occurs. The resistor and capacitor of the one-shot can be adjusted to insure that this occurs. At this time, the change of state of this one-shot triggers another one-shot. This one-shot then

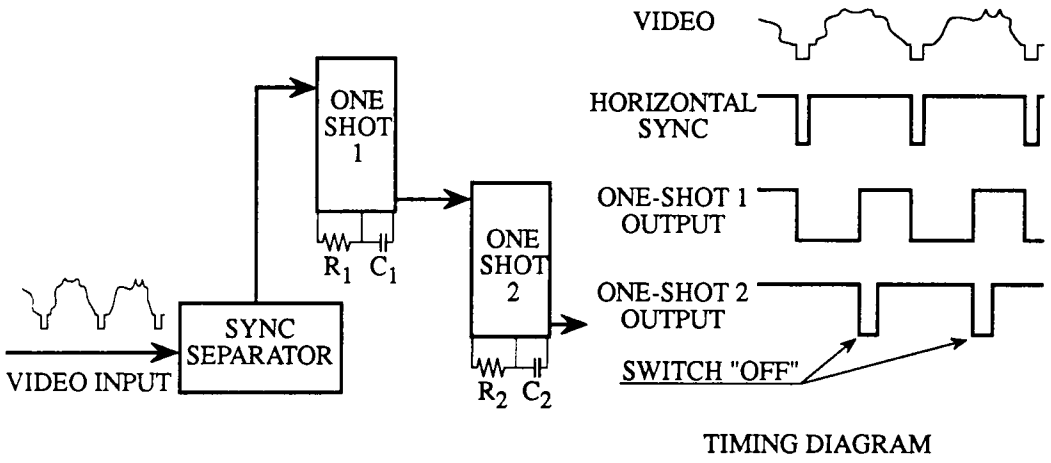


FIGURE 5-19

A BLOCK DIAGRAM OF THE ZERO-CHOP TIMING CONTROL

delivers the timed control signal needed by the PIN diode attenuator to switch into the "off" state. After about 10 microseconds, this one-shot returns to its stable state and the switch is returned to its "on" state. Figure 5-19 is a simple diagram of this process.

In Figure 5-19, the choices of R_1C_1 and R_2C_2 will determine the "placement" of the switch pulse in the active picture area and the length of time that the switch is off, respectively.

This is a quick, easy, and simple answer, but it does not address the problem of switching during an equalizing pulse. The answer to this requires no additional circuitry, just the correct application of timing. The explanation of this process is aided by the timing diagram of Figure 5-20 and the fact that a one-shot can be retriggerable. Some one-shots do not respond

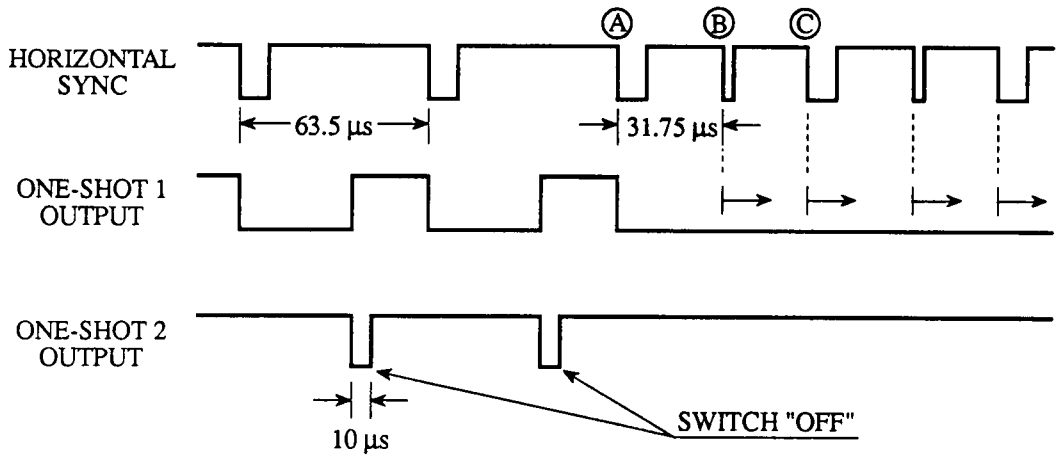


FIGURE 5-20

THE USE OF RETRIGGERABLE ONE-SHOTS TO AVOID EQUALIZING PULSES

to trigger pulses unless they are in their stable states. A retriggerable one-shot will respond to additional trigger pulses while in their unstable state. In most one-shots, this causes the unstable state to be lengthened by another timing period beginning at the point in time of the retrigger. Figure 5-20 shows the horizontal sync pulses at the beginning of the equalizing period. Assume that one-shot 1 is retriggerable. The R and C of one-shot 1 have been selected for a pulse width greater than 31.75 microseconds - in this example 33 microseconds. As previously, the horizontal sync pulse triggers one-shot 1 which then outputs a pulse 33 microseconds long. The ending of this 33 microsecond pulse triggers one-shot 2 which outputs a 10 microsecond pulse to the PIN diode attenuator. This occurs exactly as it did in

the case of Figure 5-19. The difference is when one-shot 1 encounters the horizontal sync pulse labeled as point A. As usual, this horizontal sync pulse triggers one-shot 1, which outputs a pulse that is 33 microseconds long. Before one-shot 1 can return to its stable state, it encounters another trigger pulse in the form of the first equalizer pulse. This retriggers one-shot 1, which means it will now continue the pulse begun at point A for an additional 33 microseconds beginning at point B. At point C, a horizontal sync pulse is encountered which also retriggers one-shot 1. The combination is horizontal sync pulses and equalizing pulses keep retriggering one-shot 1; not allowing it to return to the stable state. Until one-shot 1 returns to its stable state, one-shot 2 is not triggered. This prevents an "off" state from occuring during the equalizing interval. What finally causes one-shot 1 to return to its stable state? Nothing, or more accurately the lack of something. As the vertical interval progresses, one-shot 1 encounters the last equalizing pulse and again retriggers for 33 microseconds. Before it can return to its stable state, it encounters a horizontal sync pulse which retriggers one-shot 1 for 33 microseconds. Since the equalizing interval is over, one-shot 1 will not encounter a retrigger 31.75 microseconds after the last retrigger and will "time-out" and return to it's stable state. This will then trigger one-shot 2 and the process of Figure 5-19 continues.

The methods of Figure 5-20 are very simple to implement, but are not without a price. In delaying the "off" state more than 31.75 microseconds from the beginning of the horizontal sync pulse, the first half of the active picture area is unusable for the task of zero-chop switching. This places tighter constraints on the zero-chop switching speed and recovery time. This could also potentially interfere in the recovery of the demodulator. If the demodulator cannot stabilize, after the zero-chop and before the occurrence of the horizontal sync pulse, the result would be a potential interference between the zero-chop switching and the demodulator AGC system. This must be carefully developed to avoid any such

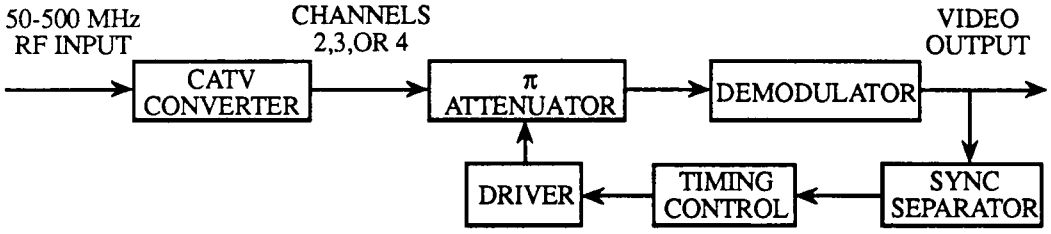


FIGURE 5-21

A BLOCK DIAGRAM OF A PROPOSED MODULATION MEASURING SYSTEM

interference.

The first development assessment shown in Figure 5-5 has changed considerably. At this point in the development process, Figure 5-21 illustrates the progress so far.

Once again an item has been "slipped in" on the reader. The box marked "DRIVER" has been added to the development between the timing control and the PIN diode attenuator. The reason for this addition is that most one-shots cannot provide the necessary output drive to supply the PIN diodes with the current necessary to control them. The driver is responsible for providing this current and should not be very complicated. For this reason, design of the driver will be left until the PIN diode attenuator is designed and tested.

Figure 5-21 labels the output as simply being "video." This is the output that has been desired from the beginning, but it is not the end of the development. A system of measuring the video signal for the benefit of the testing method is the problem under consideration. It should be remembered that the testing method relies on a comparison between the sync tip and blanking levels.

The video signal that is provided by the demodulator is wideband, high quality, color-encoded video. EIA video standards require that this output signal be 1 volt peak-to-peak and it has already been established that it is AC coupled. The standard video output also has a

75-ohm characteristic impedance. If this signal is viewed on an oscilloscope it would not appear the same as the diagrams of video used in this presentation. Indeed, it would appear to be an unintelligible combination of signals. This is primarily due to two factors: audio and color. The audio is contained in an FM subcarrier located 4.5 MHz from the visual carrier. This subcarrier appears to "ride" the video waveform. The color encoding information is a PM subcarrier located at 3.58 MHz, relative to the visual carrier. The color subcarrier has a significantly wider bandwidth than the sound subcarrier. The color subcarrier is also closer to the video information. These are the reasons the color signals interfere so greatly with the video information. Television receivers use a variety of methods to separate the video, audio, and color. Most of these methods use filtering. Most texts agree that most of the video information needed in this testing method exists in the first MHz of video bandwidth. A slightly more analytical approach will be used here to insure that the filtered video is representative of the video and that the audio and color subcarriers do not interfere with the video measurements.

The horizontal sync pulses and blanking are almost periodic in nature. These are the two quantities of interest of the testing method. In making the assumption that these are indeed periodic signals, the Fourier transform may be used to mathematically represent the frequency spectrum occupied by both signals. The Fourier series was casually explained in Chapter 3 and is used as a mathematical method of representing the frequency spectrum occupied by a periodic waveform. The assumption that the horizontal sync pulses and blanking are periodic will lead to a degree of error, but it should be an acceptable level of error. In a straight forward method, the horizontal sync of Figure 5-22(a) may be broken up into two distinct periodic pulses. The pulses appear to be squarewaves, but television standards require these to be slightly trapezoidal to conserve bandwidth. Figure 5-22(b) is the horizontal sync information with an active picture area that is monochromatic and flat gray.

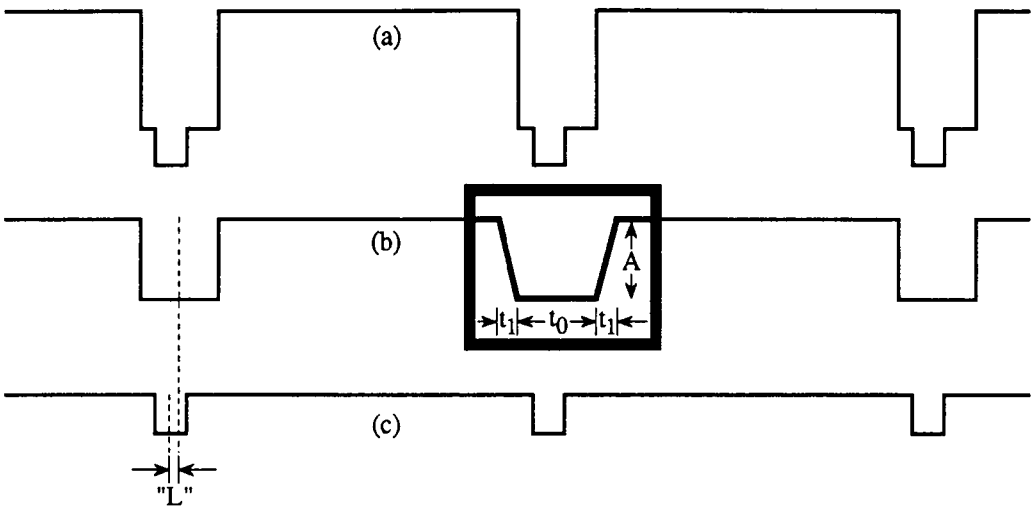


FIGURE 5-22

THE DIAGRAM OF THE WAVEFORMS USED TO SIMULATE A FLAT-FIELD VIDEO SIGNAL

Figure 5-22(b) shows the blanking and horizontal sync pulses separated into two distinct waveforms. At this point, the spectrum of the vertical interval is ignored. The spectrum of any real video information in the active picture area is also not accounted for. This should not invalidate this spectrum approximation since it does include the elements of the video information in which we are interested.

Equation 5-9 is the Fourier transform of a periodic trapezoid, with the pertinent variables labeled in Figure 5-22(b). By using this formula, the spectrum of the sync and blanking may be computed and plotted. Figure 5-23 is a plot of the calculated frequency

$$C_n = \frac{2A(t_0 + t_1)}{T} \frac{\sin\left(\frac{\pi n t_1}{T}\right)}{\frac{\pi n t_1}{T}} \frac{\sin\left[\frac{\pi n(t_0 + t_1)}{T}\right]}{\frac{\pi n(t_0 + t_1)}{T}} \qquad (5\text{-}9)$$

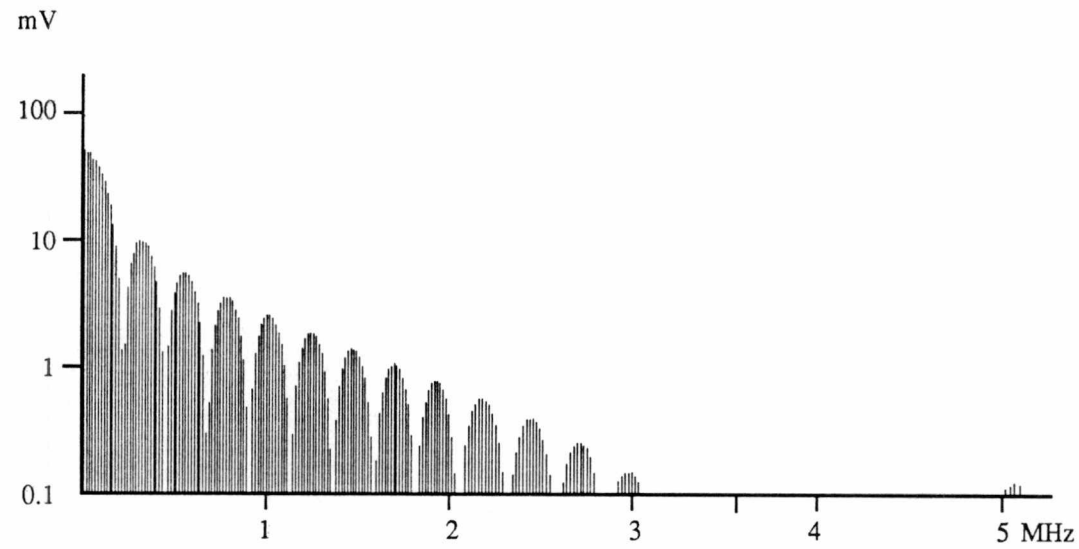


FIGURE 5-23

THE COMPUTED SPECTRUM OF THE HORIZONTAL SYNC PULSES

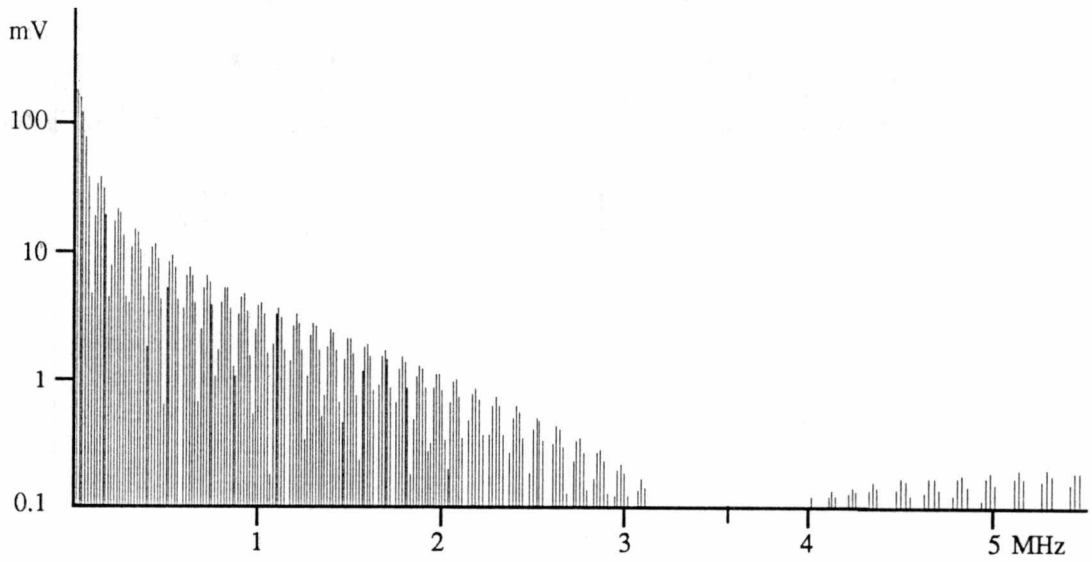


FIGURE 5-24

THE COMPUTED SPECTRUM OF THE BLANKING AND LUMINANCE

spectrum of the sync. Figure 5-24 is a plot of the calculated frequency spectrum of the blanking. Notice that the amplitude is plotted logarithmically and as such the drop in amplitude with increasing frequency is more dramatic than it may first appear.

The spectrum of Figures 5-23 and 5-24 may be added to create the spectrum of the complete video waveform of Figure 5-22(a). The spectrums cannot be simply added in this case because the waveforms are not symmetric. This means that the sync and blanking are out-of-phase. The quantity labeled "L" in Figure 5-22(a) indicates this phase difference. Either the sync or the blanking can be mathematically modified to account for this phase difference. This modified spectrum may then be added to the remaining spectrum to create the spectrum of the video waveform illustrated in Figure 5-22(a).

The phase change is computed in the following manner. Figure 5-22(a) assumes the FCC standard video signal, with a 1.3 and 4.6 microsecond front and back porch pulses,

$$\text{phase angle difference} = 2\pi n\left(\frac{1.65}{63.5}\right) \text{radians} \tag{5-10}$$

respectively. This means that the sync pulses are off-center by 1.65 microseconds. This corresponds to a phase angle difference given in equation 5-10, with n representing the harmonic number.

By applying equations 5-9 and 5-10, the complete spectrum of the video waveform of Figure 5-22(a) may be computed. Such a spectrum is shown in Figure 5-25.

Figure 5-26 is a spectrum of the waveform of Figure 5-22(a) with the addition of color spectrum information. This particular color information is a flat yellow color field. This spectrum information is from reference 38. Notice that the color spectrum is centered about the 3.58 MHz subcarrier frequency and the amplitude rolls off in both directions. In this

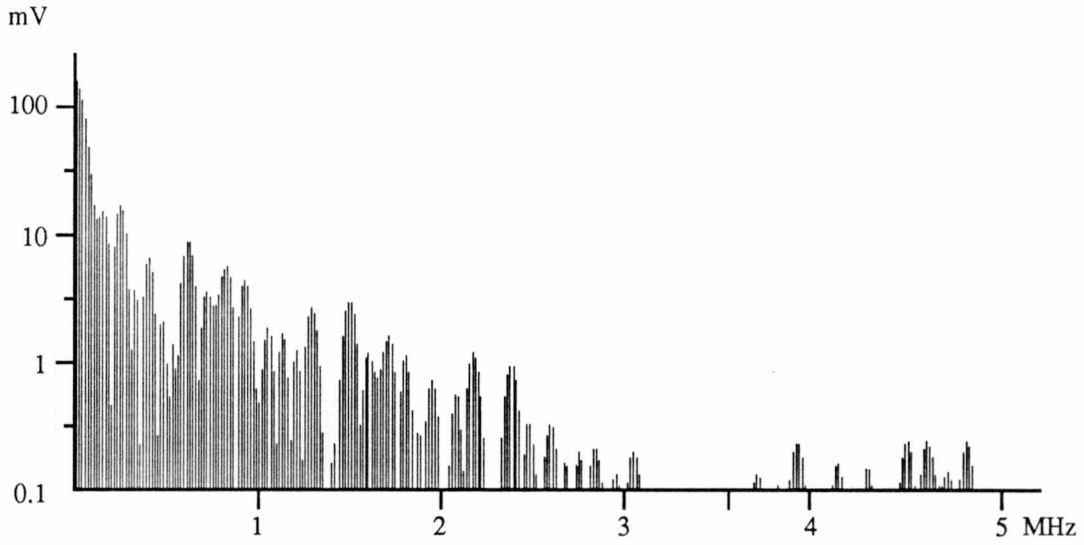


FIGURE 5-25

THE COMPUTED SPECTRUM OF HORIZONTAL SYNC, BLANKING, AND LUMINANCE

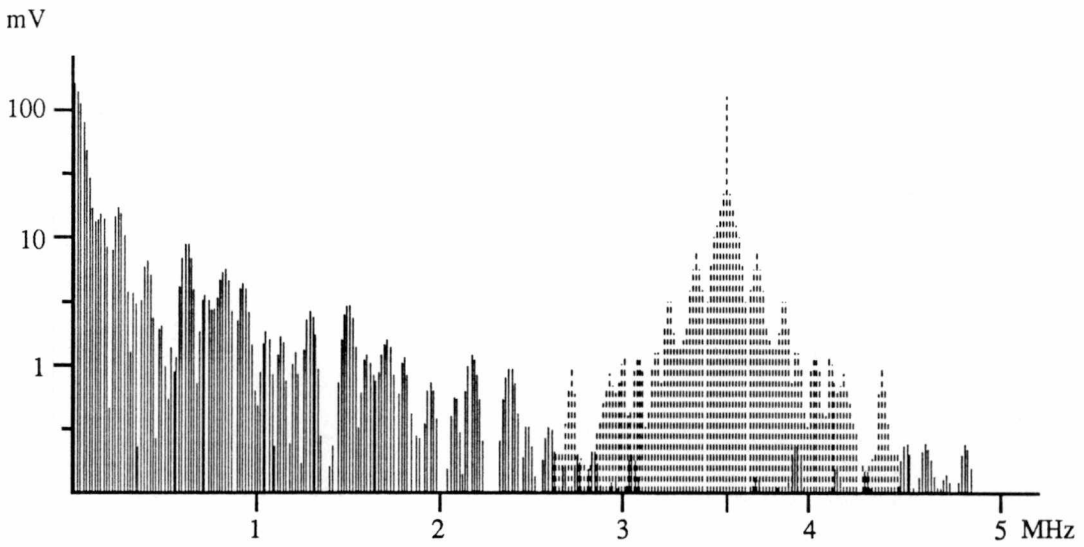


FIGURE 5-26

THE COMPUTED SPECTRUM WITH THE ADDITION OF A FLAT-YELLOW FIELD

Figure the interleaving of the spectrums is evident and was alluded to in Chapter 1. This lessens the interference between the color and video information. Figure 5-26 may be used to make some assumptions about the spectrum needed by the video and color signals.

From the information of Figure 5-26, a low pass filter can be used to pass most of the spectrum containing the video signal and at the same time attenuate the interfering color and audio subcarriers. Several implementations of the low pass filter characteristics are plotted in Figure 5-27. The Chebyshev and Elliptic low pass filters can have a much "steeper" skirt which causes the filter to roll-off much quicker than the Butterworth implementation. The price for this increased skirt performance is the ripple shown in the pass band. Using a filter with ripple in the pass band may alter the characteristics of the video signal. A Butterworth low pass filter with no ripple in the pass band should be the first choice of a filter if it can meet

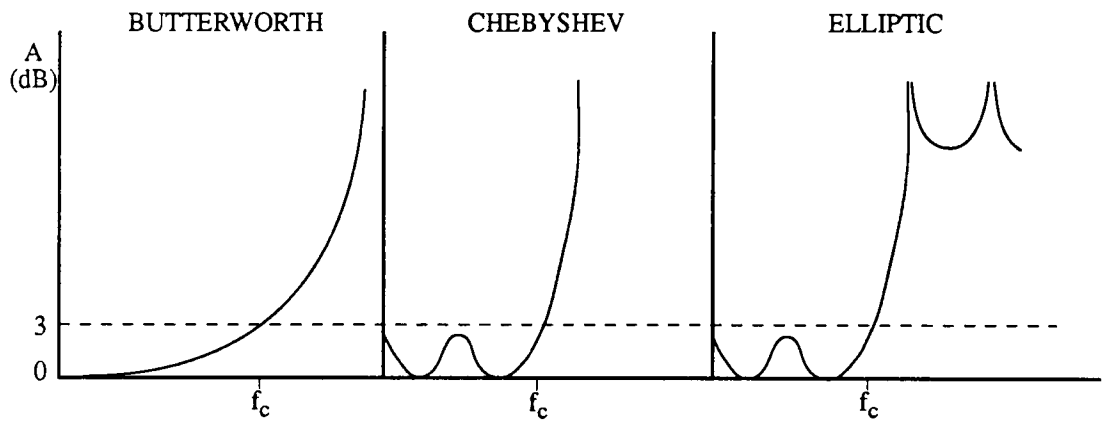


FIGURE 5-27

THE OUTPUT CHARACTERISTICS OF THREE LOW PASS FILTERS

the filtering requirements.

Equation 5-11 gives the frequency response of the Butterworth low pass filter.[39] In

$$A = 10 \log \left[1 + \left(\frac{f}{f_c}\right)^{2k}\right] \qquad (5\text{-}11)$$

equation 5-11, "A" is the attenuation in dB, f_c is the 3 db frequency, and k is the number of circuit elements. Butterworth low pass filters are usually implemented by using tables containing normalized circuit element values. This information is then converted to the

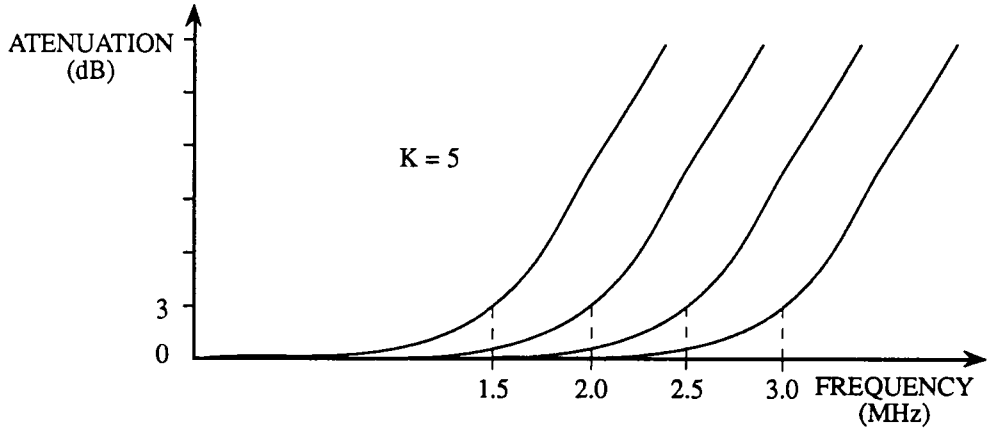


FIGURE 5-28

BUTTERWORTH LOW PASS FILTER CHARACTERISTICS FOR SEVERAL 3 dB FREQUENCIES

correct values by using the frequency and impedance of the filter as scaling factors. Using equation 5-11 and assuming a practical number of circuit elements (5), generates the plot of the low pass filter response for a variety of 3 dB frequencies. This is plotted in Figure 5-28.

The information in Figure 5-28 can be used to estimate the 3 db frequency that will properly attenuate the color spectrum. Referring to the spectrum of Figure 5-26, the color

spectrum level at the color subcarrier frequency is roughly equal to the amplitude of the fundamental of the video spectrum. This is a worst-case situation that will not be encountered in the CATV environment because of the frequency response of the CATV modulator. From this, an estimate of attenuation may be arrived at. Assuming that a 0.1% contribution by the 3.58 MHz subcarrier may be tolerated, this is a 30 dB difference between the fundamental and the 3.58 MHz subcarrier. Working equation 5-11 backward indicates a 3 dB point of 1.8 Mhz will meet the required 30 dB of attenuation of the 3.58 MHZ subcarrier. Setting the number of filter elements to 5 and the 3 dB point to 1.8 MHz specifies the Butterworth low pass filter. The input impedance of the low pass filter is 75-ohms and the filter is scaled to this specification. This will require a 75-ohm output termination for proper filter response.

The output of the low pass filter is used as the input to the sync separator. This avoids the problem of false triggering caused by the color and audio subcarriers. The output of this filter should resemble the video waveforms used in the Figures in this chapter. This output must now be measured for use in the measurement technique we are using. The sync tip, blanking, and zero carrier levels must be "captured" and retained for comparison with each other. This capturing can be accomplished by circuits referred to as "sample and hold" devices. Before this is done, some gain should be introduced into the system to offset the losses of the low pass filter. This will also give a larger range of levels to capture. This amplification should also introduce some DC offset to the video signal. This will insure that all three levels are always between ground and the positive supply voltage. This insures that the three levels are always positive. This gain will probably be between 4 and 8 and could easily be accomplished by a single stage common emitter amplifier. This will result in an inverted video output, but this is not a problem. It may be an advantage as the zero carrier level reference will be the lowest voltage of the three. Figure 5-29 illustrates all of this in an updated block diagram.

The matter at hand is to capture the level of the three signals of interest in Figure 5-

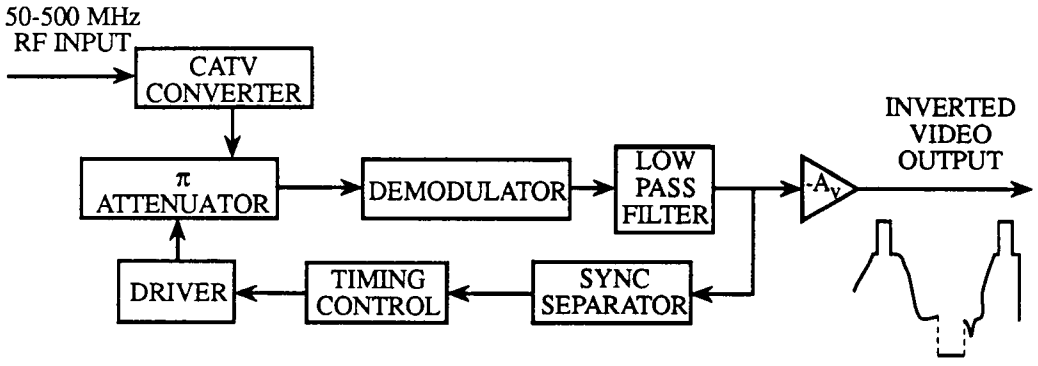


FIGURE 5-29

A BLOCK DIAGRAM OF A PROPOSED MODULATION MEASURING SYSTEM

29. As was previously mentioned, sample-and-hold techniques can be used to acquire these signal levels. A brief explanation of sample-and-hold techniques and their application to this problem is in order.

A sample-and-hold circuit will follow an input signal level and output the value of the input signal when the "hold" command is issued. Figure 5-30 is a simple diagram of the

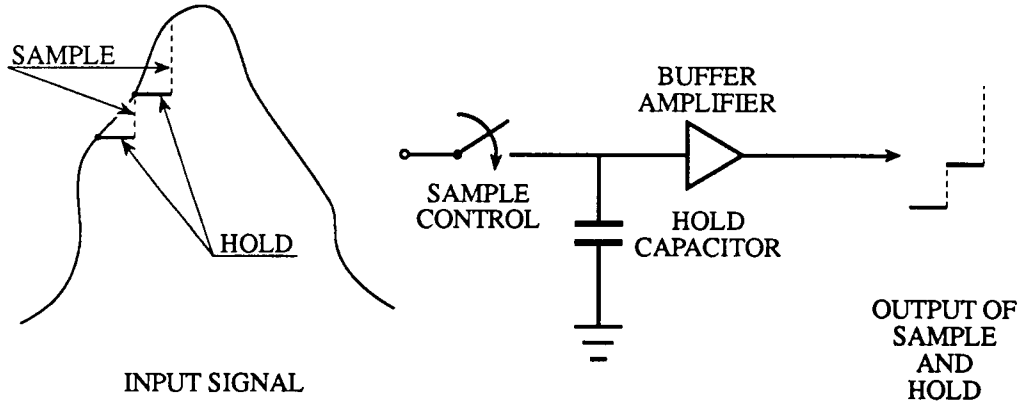


FIGURE 5-30

THE INPUT-OUTPUT RELATIONSHIP OF SAMPLE-AND-HOLD DEVICES

sample-and-hold process and a simple sample-and-hold circuit representation.

Although simple, Figure 5-30 is indicative of the sample-and-hold process and physical implementation. Several caveats do exist in the application of sample-and-hold. In a real sample-and-hold implementation, the sample switch requires a finite amount of time to close and settle. This time is collectively referred to as the aperture time and can be as long as 500 nanoseconds in a good quality device. This means an accurate sample will occur only after the aperture time. The amount of time needed to charge the hold capacitor to the sampled value of the input depends on the change in input since the last sample and the rate of change of the input during the sample mode. During the hold periods the voltage on a real hold capacitor will "leak" off. This is referred to as "sag." A hold capacitor must be chosen small enough to quickly respond to the sampled signal, but it must also be chosen large enough so that sag does not adversely affect the accuracy of the sampled data. Sample-and-hold devices have design graphs relating the capacitor size, sample time, and input rate of change. The last important design information about a sample-and-hold device is the dynamic sampling error. This occurs when a rapid change in the input signal occurs during application of the hold mode. All sample-and-hold devices require a finite amount of time to fully execute the hold command, and any input signal change during this time period can have an unpredictable effect on the held output.

In applying sample-and-hold devices for our purposes, some method of triggering the hold mode of these devices must be found. The sync tip, blanking, and zero carrier levels will each have a sample-and-hold device dedicated to the measurement of that particular level. The timing information available from the sync separator is the most obvious source of timing control signals. Triggering the sample-and-holds associated with measuring the sync tip and blanking levels are the simplest. The sync tip sample-and-hold can be triggered directly by the horizontal sync pulse of the sync separator. The leading-edge of the sync pulse may have some "ringing" associated with it, but will settle by the end of the sample-and-hold aperture

time. If the sample is made for most of the sync tip period, the error introduced by any ringing should disappear. The sync separator could trigger a one-shot. This one-shot would then stay in the unstable state for the sample period. Upon return to the stable state, the sample-and-hold would enter the hold mode. The sync tip is typically 4.7 microseconds long, so setting the one-shot for a 4 microsecond sample period should adequately sample-and-hold the sync tip level. Notice that the dynamic sampling error has been held to a minimum by attempting to sample in an area that has very little change. Acquisition of the blanking level is done in a similar manner. The 4.6 microsecond back porch also indicates the use of a 4 microsecond sample time.

Care must be taken in sampling during the vertical interval. The equalizing pulses create the only problem. A system much like the one used to avoid zero-chop switching collisions with these equalizing pulses must be used. The blanking sample-and-hold can avoid this problem by using the zero-chop method of equalizing pulse avoidance. The sync tip presents a more difficult obstacle. The sample-and-hold process will occur on the equalizing pulse before it can be stopped unless control is based on the previous occurrence of the equalizing pulses. In this design, the vertical pulse of the sync separator triggers one-shot 1 in Figure 5-31. This one-shot remains unstable until well past the equalizing pulse interval. The output of one-shot 1 then triggers one-shot 2. One-shot 2 is connected to the clear input of the sampling control one-shot. When one-shot 2 is unstable, the sampling control one-shot is enabled and performs as usual. Just prior to the beginning of the next equalizing pulse interval, one-shot 2 "times out" reverting to the stable state. This inhibits the sampling control one-shot and prevents any sampling from occuring during the equalizing pulse interval. The vertical pulse triggers one-shot 1 and the process begins again.

Sampling the zero carrier reference level is a little more complicated than the sampling of either the sync tip or blanking levels. Because of the settling time the demodulator may require to recover from the switching "off" process, sampling should not

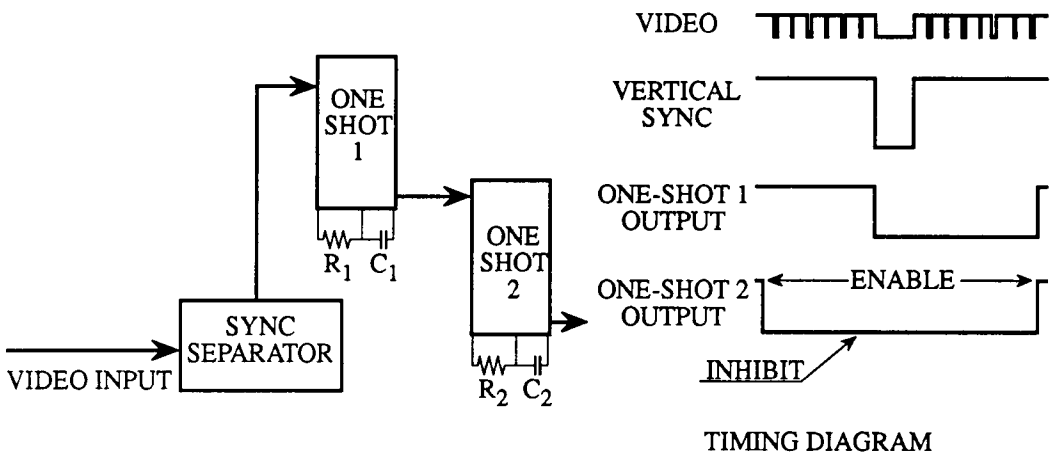


FIGURE 5-31

A METHOD OF AVOIDING COLLISIONS WITH EQUALIZING PULSES

occur until this recovery occurs. If the zero-chop lasts a total of 10 microseconds, and 4 microseconds is a adequate sample time, this leaves 6 microseconds in which to allow the demodulator to settle. A delay in the sampling time can be generated by using two one-shots as before. The timing control output to the switch is used to trigger the first one-shot, which 3 microseconds later triggers the sample time one-shot. This then samples for 4 microseconds and leaves 3 microseconds before the zero-chop switch changes states. This should provide ample time to avoid either switching "on" or "off" transients.

With this in mind, the sampled signals can now be compared. A simple op-amp circuit may be used to acquire the difference between the sync tip and blanking levels, and the reference. At this point we have the measurement data needed to perform the test. The last problem is to devise a method of interpreting this data.

Two ways exist to interpret that data. Since the depth of modulation is proportional to the ratio of the sync tip and blanking levels, a divider can be used to directly acquire the ratio of the data. A divider uses the mathematical fact of equation 5-12. This implies that

division may be performed by subtraction. Converting the data to logarithmic values is the most difficult portion of this process. Linear logarithmic operations that are not temperature dependent can be difficult to design. This would however be the best method of comparing the measurement data. An alternative is the use of a nulling circuit. When the television carrier is properly modulated, the sync tip to blanking ratio is 1:0.75 . By designing an instrumentation amplifier that attenuates the sync tip level by 25% and subtracts it from the blanking level, the output is a zero level when the sync tip to blanking ratio is exactly 1:0.75. Although much simpler than a divider, the null circuit has two distinct problems. Scaling can be difficult for values of modulation depth that are not perfect. A null circuit can also put out a zero voltage level with no input, which can lead to erroneous conclusions by the user. The

$$\frac{A}{B} = \log^{-1}\left[\log(A) - \log(B)\right] \tag{5-12}$$

output of either method is then used to drive some form of output device. The simplest output device is an analog meter. Design of this output device is left until the divider or null circuit is designed.

Figure 5-32 is a final representation of the development work done in this chapter. The development has produced results that can be used in a circuit implementation. As actual circuit design progresses, some aspects of this development will be subject to change or modification. Chapter 6 will be concerned with turning this development information into an actual design.

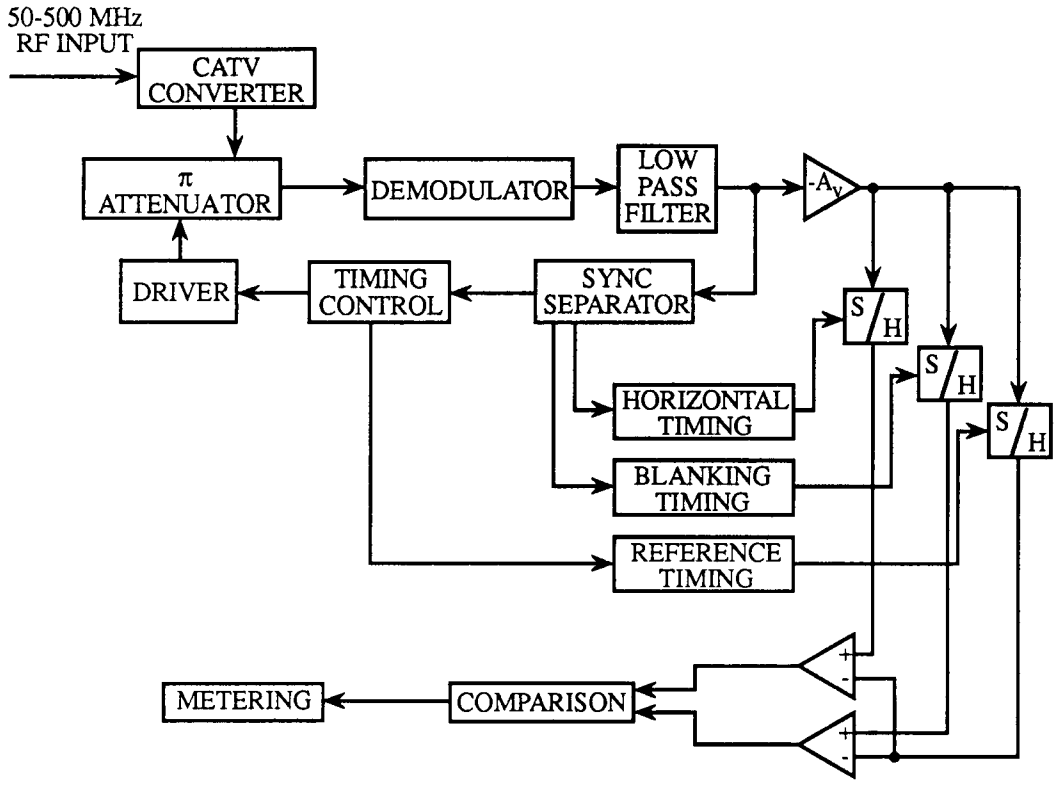


FIGURE 5-32

A BLOCK DIAGRAM OF THE PROPOSED MODULATION MEASURING SYSTEM

CHAPTER 6

CIRCUIT REALIZATION AND EXPERIMENTAL RESULTS

This chapter will be concerned with using the development information of Chapter 5 to design and construct an actual circuit realization of the modulation depth testing method put forth by this presentation. The development order of the actual circuits will closely follow the input to output order in which Figure 5-32 is presented in Chapter 5. This figure is reproduced here as Figure 6-1.

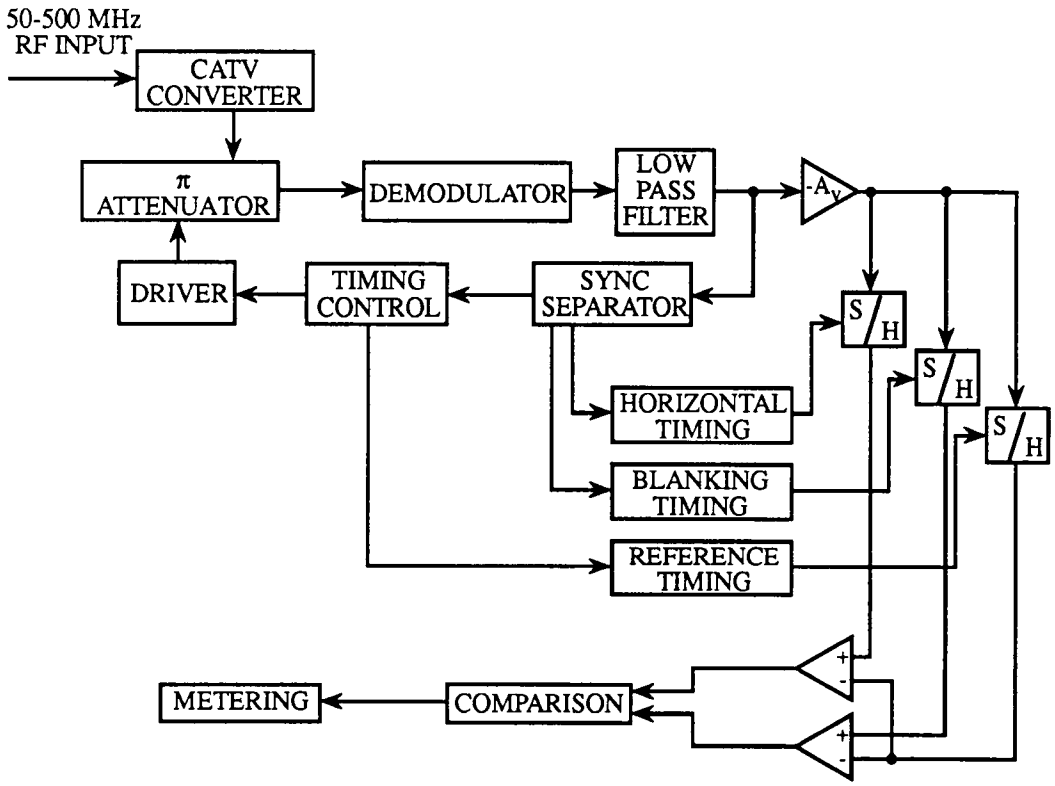


FIGURE 6-1

A BLOCK DIAGRAM OF THE PROPOSED MODULATION MEASURING SYSTEM

The first device encountered in Figure 6-1 is the CATV converter. As previously mentioned, these are very common devices in most CATV systems. As some CATV systems use nonstandard methods of programming transmission, this converter may be necessary in these systems to convert the input signal to the conventional standard television transmission. Most CATV converters have a through gain of from 0 to 8 dB. This will need to be taken into account to avoid any possibility of overloading any stages following the converter.

Orderly development having already been promised, it is unfortunate to have to "jump" out of order this soon, but it is necessary to avoid any "back-tracking" later. Chapter 5 correctly states a variety of reasons for not attempting to develop a dedicated demodulator for this testing system. It would be a more ideal situation to be able to include a dedicated demodulator in the testing system because of the greater degree of control this offers the designer. A demodulator has in fact been located that will adequately perform the needed tasks. This is a single integrated circuit implementation of an IF-detector combination circuit. It is literally a RF in-video out device. It is also designed to operate at the normal television IF frequencies (45.75 MHz for the visual carrier), but operation is possible at frequencies up to 70 MHz. This means that a tuner is not necessary. A CATV converter is already required in this design, so the demodulator will be tuned to operate at the output frequency of the converter. The channel chosen in this design will be channel 3. This information is necessary so that the converter may be properly exploited in its new role as the system tuner.

The operating requirements of the converter have significantly increased. The converter will have the responsibility of tuning the entire testing system. In a VCR, television, or commercial demodulator, the IF from the tuner is filtered to prevent interference from the other television channels. Most interference is contributed by the adjacent channels. Using channel 3 as an IF will require filtering to prevent the same type of interference. The demodulator will have to be additionally shielded to prevent interference from any off-air channel 3 broadcasts.

The CATV converter up- or down-converts the desired channel to channel 3. Most converters have light band pass filtering at the output, but the adjacent channels are not attenuated enough for use as an IF. The output of the converter will have to be additionally filtered before it is suitable as an IF source. Figure 6-2 is the output characteristics of a modern high performance CATV converter. This particular model benefits from the use of a SAW

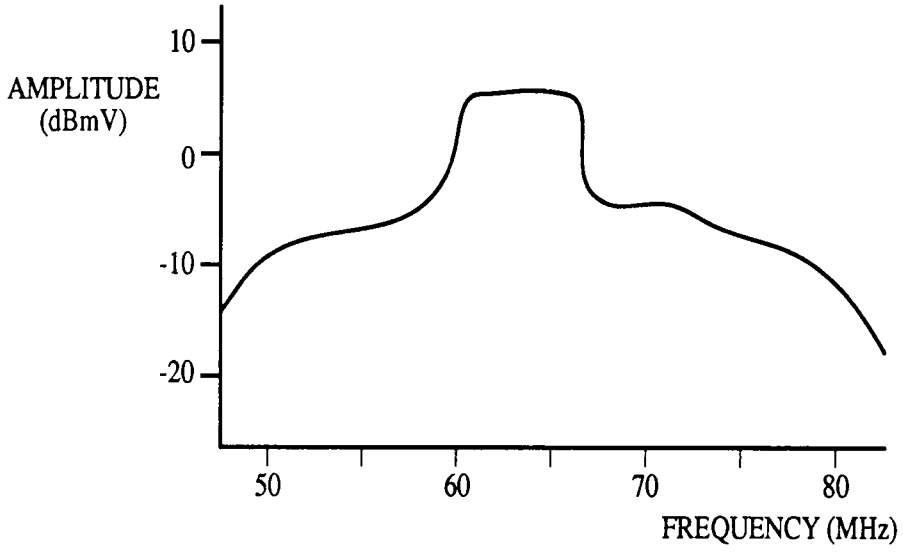


FIGURE 6-2

THE FREQUENCY RESPONSE OF A CATV CONVERTER

filter in the output. If this were a mass produced design, a high quality channel 3 SAW filter would be the best selection for an IF filter. SAW filters of this nature are difficult to procure, so a substitute will be used. CATV systems use a wide variety of filters to provide customers with a high-quality product. Because of this, RF filters tend to be commonplace around CATV systems. The most inexpensive filter used is a notch filter. This is sometimes referred to as a "trap." A notch filter is a filter that will reject a band of frequencies while leaving the

rest of the spectrum unaffected. Figure 6-2 illustrates that the upper and lower adjacent channels are the only ones that need to be removed. This can most economically be done by a series of traps, each tuned to a particular interfering carrier. Because the traps have an input and output impedance of 75-ohms, matching is not a problem. Figure 6-3 is the output spectrum of the converter and shows the frequencies of the unwanted carriers.

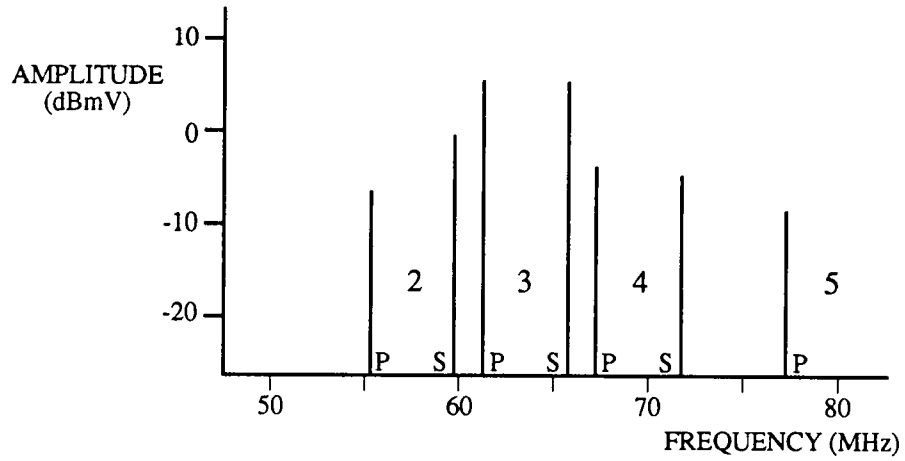


FIGURE 6-3

THE FREQUENCY SPECTRUM OF THE CARRIERS OUTPUT FROM A CATV CONVERTER

Referring once more to Figure 6-1, the next item that needs attention is the zero-chop switch. The development in Chapter 5 presented the reasoning behind the choice of PIN diodes, so now the task is to implement this attenuator. The basic π PIN diode attenuator is shown in Figure 6-4. The bias networks attached to each diode are used to control the current in each diode, which controls the attenuation. A simpler implementation of a pin attenuator is available without the need for independent, active bias controls. Figure 6-5 is a diagram of such a circuit.

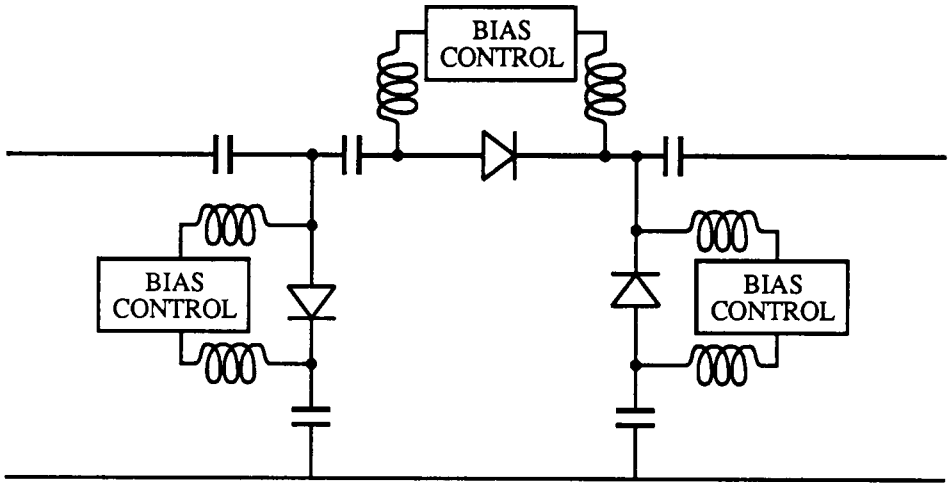


FIGURE 6-4

A BLOCK DIAGRAM OF A π ATTENUATOR

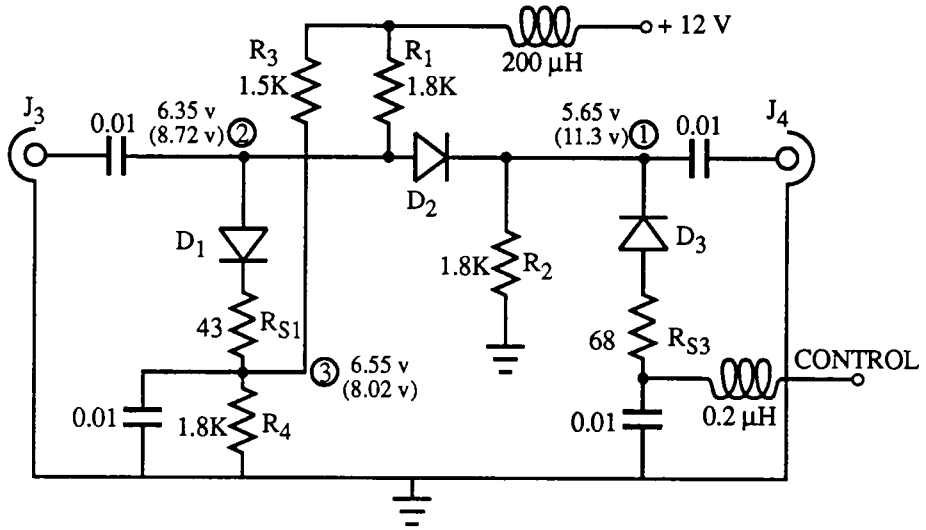


FIGURE 6-5

THE π ATTENUATOR CIRCUIT

The PIN diode attenuator illustrated in Figure 6-5 is probably the most used AGC control circuit in use today.[40] The current through each diode is provided by only one control voltage. This circuit warrants a closer examination. Analysis of this circuit is relatively simple at the attenuation extremes where the diodes may be assumed to be "turned-on" or "turned-off" completely. The region between these two extremes is very difficult to analyze. The PIN diode follows the familiar diode equation listed as equation 6-1. This equation relates diode current to diode voltage. This equation does not have an ordinary algebraic solution. This equation is known as a transcendental equation and can only be solved by

$$I = I_S \left[e^{\left(\frac{qV}{kT}\right)} - 1 \right] \qquad (6\text{-}1)$$

either fixing one variable or by iterative numeric methods. This complication leads to most manufacturers publishing actual circuit implementations, and some provide analytic data necessary for numeric simulation. The diode current must then be related to the RF resistance and the best choices made for the device to present a constant impedance.

A simple analysis of Figure 6-5 is as follows. First assume that the control voltage is zero. This corresponds to the zero attenuation situation. From this it is known that D_1 and D_3 must present a high level of RF resistance, so they are turned-off. Diode D_2 must present a low RF resistance, so it must be turned on. Design curves show the forward-biased voltage drop of the PIN diode to be about 0.7 volts. This establishes the current through D_2 by equation 6-1. From the manufacturer's design tables, this current corresponds to a RF

$$I_{D2} = \frac{12 - 0.7}{1.8K + 1.8K} = 3.14\ mA \qquad (6\text{-}2)$$

resistance of about 12-ohms. This resistance can then be used to estimate the insertion loss by assuming the resistance of D_1 and D_3 to be much greater that 75-ohms. Equation 6-3 is a dB estimate of the insertion loss. The current through D_2 also establishes a voltage at node

$$\text{Insertion loss} = 20 \log \frac{V_{out}}{V_{in}} \approx 20 \log \frac{75}{75 + 12} = 1.29 \text{ dB} \qquad (6\text{-}3)$$

1 of about 5.65 volts and a voltage of about 6.35 volts at node 2. The voltage at node 3 is established by the voltage division of R_3 and R_4, which is a voltage of 6.55 volts. In examining the three node voltages, D_1 and D_3 are indeed reverse-biased and turned-off. The circuit is essentially dormant until the control voltage approaches the voltage of node 1. As this occurs, diode D_3 begins to turn-on and allows some current to flow through D_3 as related by equation 6-1. Turn-on does not occur quite as quickly as might be assumed, as D_3 begins to turn-on the D_3 current begins displacing the D_2 current through R_2. Node voltage, however, also rises due to the rise in control voltage. This prevents D_3 from turning-on completely when the control voltage reaches 6.35 volts. As the node 1 voltage rises and D_3 contributes more current for the drop across R_2, the current flowing through D_2 begins to decrease. The situation is such that the RF resistance of D_3 is decreasing and the RF resistance of D_2 is increasing, which is what is desired. The sum of R_1 and R_2 determined the initial current of D_2, which in turn established the insertion loss. The ratio of R_1 to R_2 is the important issue at this point. This ratio must be chosen so as to vary the currents in D_2 and D_3, and in turn the RF resistances, to maintain a constant impedance level. As the control voltage increases, the current through D_2 is decreasing. This current decrease causes the node 2 voltage to rise toward the supply voltage. As node voltage 2 is rising, diode D_1 begins to turn-on. It is very important that diode D_1 turn-on at the same rate as D_3 to insure a constant impedance is

maintained. This matching is provided by the ratio of R_3 and R_4. The control voltage eventually reaches the point in which D_1 and D_3 are turned-on and D_2 is turned-off.

The turn-on of D_1 and D_3 and turn-off of D_2 is the other extreme of the attenuation scale, maximum attenuation. Again analytical analysis of the diode currents and node voltages is possible. Solving the correct mesh equations and assuming a 0.7 volt drop across D_1 indicates a 1.82 milliampere current flowing through D_1. This current level corresponds to an RF resistance of about 35-ohms. This resistance needs to be very close to 75-ohms for proper impedance match. The small series resistor R_{S1} will provide the additional 43-ohms necessary to insure a proper match. The current through D_3 is much greater than the current though D_1. To calculate the D_3 current, assume the control voltage can go to the +12 supply voltage and forward diode drop is 0.7 volts. This yields a D_3 current of 6.3 milliamperes. This is an RF resistance of about 8-ohms and indicates a series resistor of 67-ohms for proper impedance matching. In practice, these series resistors will be chosen by using return loss measurements for an optimum match.

Figure 6-6 is a plot of the loss versus control voltage for this attenuator design. Notice that the largest rate of change of attenuation occurs between 8 and 9.3 volts. These are the control voltages that cause the most rapid turn-on and turn-off of the PIN diodes. Figure 6-6 is not very linear, but our concern is for minimum insertion loss and maximum attenuation while maintaining a reasonably constant impedance. This design appears to meet the development requirements with a 0.75 dB insertion loss and an ultimate attenuation of 44 dB. The lower than predicted insertion loss illustrates the error contributed by analytical analysis and tabulated device information. An advantage of this circuit is that a single control voltage is all that is needed to completely control the attenuation of this design. The supply voltage of +12 volts is chosen because of the requirements of the demodulator. Due to the high frequency of operation, this circuit is constructed with a ground plane, and RF connections are made by F-81 type RF connectors. The 200 microhenry inductor placed in the supply line

is to reduce the insertion loss of the attenuator through the supply. The 0.2 microhenry air-core inductor accomplishes the same task of preventing RF losses through the control driver. The 0.01 microfarad coupling capacitors present a very low reactance to the RF signals due to the high frequency involved, and effectively block the DC bias and control signal.

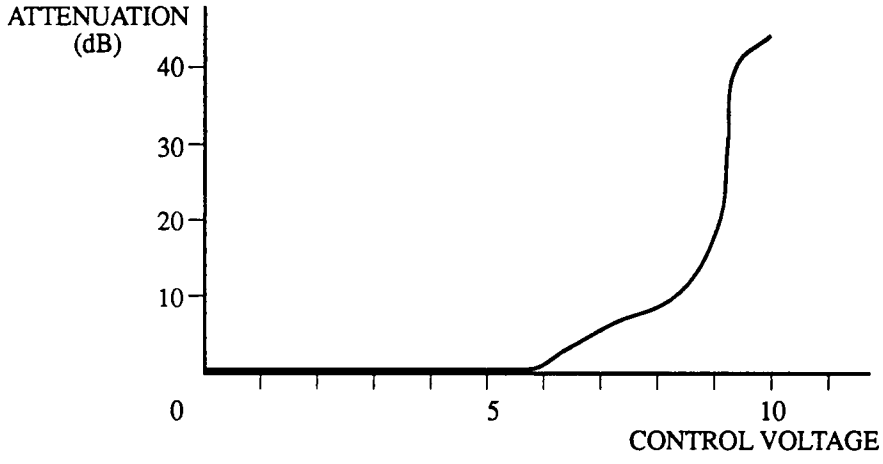


FIGURE 6-6

THE ATTENUATION PERFORMANCE OF THE π ATTENUATOR

The signal from the zero-chop switch is now introduced to the demodulator. The demodulator used in this experiment is not a self-contained demodulator like the one suggested in the development chapter. Instead, it is a single integrated circuit that will be adapted to the task at hand. An important note is that this demodulator will be designed for 75-ohm input and output impedances, and the input and output will follow EIA guidelines. This ensures that the remaining circuitry would also function with a self-contained demodulator.

The National Semiconductor® LM1823 is an IF amplifier and synchronous demodulator contained in a single package. This demodulator is ideal for use in this application. It is compact, simple to implement, has a self-contained AGC system, and will operate above the standard television IF frequency. The fact that the detector is a synchronous detector is perfect for our purposes because it can best handle the extremes of the zero-chop switching with minimal distortions. The following is a discussion that is primarily focused on the modification of this device to meet our needs. The principles by which this device operates will be discussed where necessary for an understanding of the adaptation process.

The National Semiconductor® Linear Applications Manual[41] supplies most of the application particulars of this device and includes some of the internal circuit details. All design examples are for the standard television IF of 45.75 Mhz, so changes in some of the external circuit elements will need to be designed. The LM1823 uses a balanced RF input which is fed into the IF amplifying section. The IF section has more than 74 dB of gain available, with a small degree of control available to the designer. The IF output and detector input are external to the device and may be used by the designer. The detector has a conversion gain of 34 dB and uses a phase-locked-loop (PLL) source for a true synchronous detector design. The detector section also produces the gated AGC control voltage for use by the IF section, which nominally has 55 dB of control range. Several aspects of the AGC system are available to the designer. This is a very sophisticated device that is beyond the scope of this presentation to discuss in detail, but complete external circuit details and the reasoning behind these circuits will be detailed.

The RF input to the LM1823 is a balanced common-base configuration to make the use of SAW filter input devices simpler to interface. This application does not use a SAW filter input, but a balanced input is mandatory for proper operation. The 75-ohm unbalanced input can be converted to a balanced input in a variety of ways. An unbalanced-to-balanced transformer is not difficult to fabricate from a ferrite bobbin and some small gauge enameled

wire. This approach was tested and worked very well. A very simple alternative is to use a CATV matching transformer. Most consumer televisions still use 300-ohm balanced inputs and these matching transformers are necessary to match the 75-ohm cable service to the 300-ohm television input. Because of the scale of matching transformer manufacture, they are very inexpensive. The impedance match is by a 4:1 impedance transformation, so the LM1823 input is reflected by this ratio to the RF input. The use of this device did not appear to alter the performance of the device. The ease of use, price, and high degree of shielding makes this an ideal choice.

With the RF now in the IF amplifier, the gain of the IF amplifier must be determined. This is a very unusual situation to explain. The AGC system is sync keyed, and in the absence of a video signal will turn the system gain to the maximum. With a total system gain from RF to video of 109 dB, the output in this situation is a large amount of noise. With the 75-ohm noise floor of -59 dBmV and a noiseless conversion, output noise is 50 dBmV, or about 0.32 volts. The fact that this is a noisy conversion increases this output noise level significantly. The point seems moot since the appearance of a video signal at the output will cause proper AGC action to occur. This is a true statement, however the input RF level must be large enough to exceed the system noise before the AGC can recognize the video sync tips. When the AGC keying system recognizes these sync tips, the gain of the IF is controlled and the amount of noise contributed by the LM1823 becomes negligible compared to the noise associated with the RF input. This unusual situation creates a minimum input level before proper demodulation can occur. The IF gain can be reduced, which in turn reduces the total system noise, and will increase the system sensitivity to low input signals. This gain reduction will allow signals as low as 0 dBmV to be demodulated. This gain reduction is effected by leaving pins 3 and 4 of the LM1823 open and reducing R_7 to 160-ohms. All pin number references are made with respect to Figure 6-7.

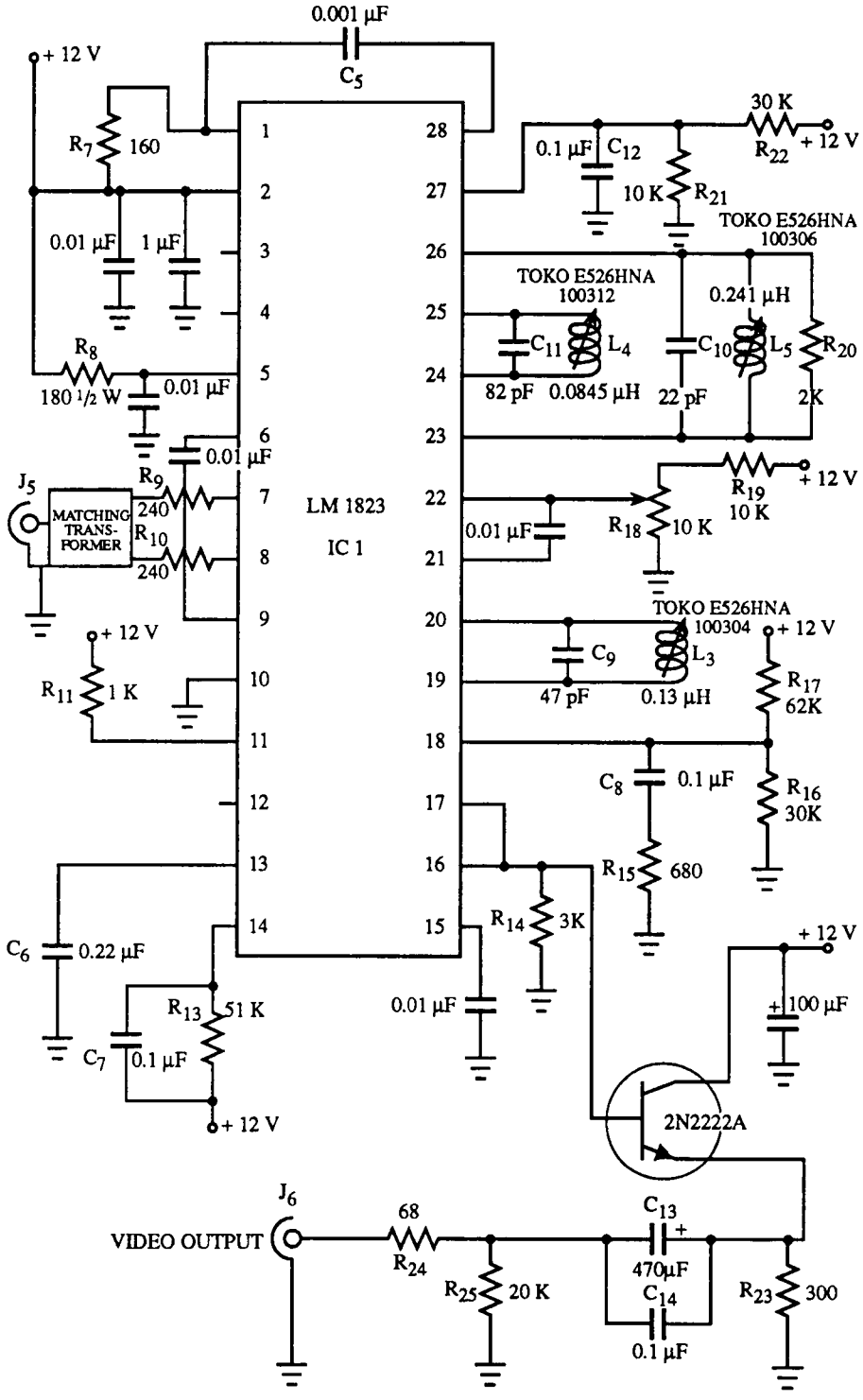


FIGURE 6-7

THE DEMODULATOR CIRCUIT

The supply voltage of the LM1823 is +12 volts, and pin 2 is tied to this supply. Resistor R_8 controls the current into an onboard reference. Pins 7 and 8 are the balanced RF inputs and pin 10 is ground. Pin 11 is an AGC output pin for use with a tuner. Pin 12 adjusts the pin 11 characteristics, and since a tuner is not used, pin 12 is tied to the positive supply. Pin 13 has a capacitor tied to it for AGC filtering and pin 14 uses an RC network to establish the AGC gate generator time constant. Pin 1 is the IF amplifier output and is coupled to the detector input pin 28 by capacitor C_5. Pin 27 is the AFC output and although not used in this application, the resistive network attached to this pin is included to properly load the LM1823. Pin 22 uses a resistive divider to adjust small phase errors in the detector and the network attached to pin18 acts as a PLL filter. All of these pins have little to do with the choice of RF input frequencies, so their application note values are sufficient.

The remainder of pins on the LM 1823 are frequency dependent. It is important to understand some of the particulars of synchronous detection. The IF signal is split, with some going to a limiter and the remainder going to the phase detector. The IF signal that goes to the limiter is stripped of video modulation and is used by a PLL to lock a voltage controlled (VCO) oscillator to the carrier frequency of the IF. The output of the VCO is applied to the phase detector along with the originally split IF signal. This phase detector uses synchronous, or in-phase, inputs. Since both inputs are at the same frequency and the VCO signal has a constant amplitude, the output of the phase detector follows the amplitude differences of the inputs and tracks the amplitude modulation. It is the limiting and VCO control functions which are frequency dependent.

The tuned circuit between pins 24 and 25 is used by the limiter to limit the frequency of operation of the limiter to the desired IF. The resonant frequency of a parallel tuned LC circuit is given in equation 6-4. The value of f_r is the visual carrier frequency of channel 3, 61.25 MHz. Solving equation 6-4 will yield the inductive and capacitive components

$$f_r = \frac{1}{2\pi\sqrt{LC}} \tag{6-4}$$

necessary to resonate at the channel 3 frequency. The "Q" of this LC circuit will determine the selectivity of the limiter. Most texts will define the Q of a parallel LC circuit as in equation 6-5.The application notes give very little information about the Q needed for the limiter to

$$Q = R\sqrt{\frac{C}{L}} \tag{6-5}$$

function at peak performance. Equation 6-5 can be used to estimate the values of C and L for the Channel 3 frequency by comparison with the C and L values recommended for the 45.75 MHz IF. If the change in inductance is small, R can be assumed to be a constant in both cases and does not contribute to the comparison. This leaves the relation expressed in equation 6-6 which relates the 45.75 MHz components, C_L and L_L, to the 61.25 MHz channel 3 components, C_3 and L_3.Equation 6-6 should not be taken as more than an estimating aid for choice of the necessary L and C components. If L_3 is made much larger or smaller than L_L, the resistivity change in the inductor will contribute significantly. Equation 6-6 allows selection of components in a fashion that is better than the alternative: a guess. With this in mind, equations 6-4 and 6-6 are used to determine values for C_3 and L_3 , which are labeled C_{11} and L_4 in the circuit diagram of Figure 6-7.

$$\sqrt{\frac{C_3}{L_3}} = \sqrt{\frac{C_L}{L_L}} \tag{6-6}$$

Pins 23 and 26 also use a parallel LC tank circuit to tune the AFC control section. In this case the AFC is not used, so it might be possible to delete this tank circuit entirely. The LM1823 design information does not provide much information about the effects of such a deletion. This tank circuit may be necessary to properly load a portion of the internal circuit. The absence of this tank could then lead to an unbalance that could affect the portions of the circuit that are being used. Without experimental information, the conservative design approach is to include this tank circuit. Values of L and C are chosen in a manner similar to the manner described for the limiter tank components. These L and C components are labeled L_5 and C_{10}, respectively, in the circuit diagram of Figure 6-7.

The last frequency dependent circuit is an LC tank circuit across pins 19 and 20. This tank circuit is responsible for tuning the PLL section. From the previous explanation of synchronous demodulation, a VCO is phase-locked to the limited input carrier. The PLL is responsible for retuning the VCO to match this incoming frequency. The VCO must be oscillating near this frequency for the PLL to be able to "pull" the VCO into the proper frequency. If this natural frequency of oscillation is too far from the incoming reference, the PLL will not be able to change the VCO frequency enough to match the incoming carrier. The LC tank across pins 19 and 20 is responsible for setting this natural frequency, which is of course the same as the input frequency of 61.25 MHz. The component selection for this tank circuit is aided by design information from the manufacturer. The Q of the tank will control the "pull-in" range of the PLL. A small range insures high stability of the PLL, while a broad range insures that the PLL will lock on signals that are slightly off frequency. A Q was chosen to give approximately $\pm$ 0.75 MHz of pull-in range. This range is necessary because of the expected diversity of the input frequencies. This diversity is from several sources, including output frequency errors in the CATV converter, intentional CATV frequency offsets, and temperature dependency. The L and C component values, L_3 and C_9, are shown in the circuit diagram of Figure 6-7.

Pin 16 of the LM1823 is the video output, and connecting it to pin 17 provides an AGC referred to this output video. The video output of this device is about 3 volts peak-to-peak and is DC offset. The output driver is a low impedance Darlington pair, but the maximum output current level set by the manufacturer prohibits directly driving a 75-ohm load. The Q_1 transistor stage is an emitter-follower buffer used to provide the current necessary to drive a 75-ohm load. The DC offset of the video output is used to bias the base of Q_1. This DC coupling simplifies the design and increases the performance of this buffer. R23 is the emitter loading resistor and is a small value due to the low base DC offset voltage. The collector is connected directly to the +12 volt supply and the 100 microfarad capacitor is only one of many capacitors used to prevent circuit cross-talk through the supply. The output impedance of Q_1 is essentially the emitter dynamic resistance, r_e, which is given by equation 6-7.[42] In this case, r_e is approximately 2.4-ohms. Capacitor C_{13} is used to block the DC component of Q_1. The value chosen has a low frequency response, when terminated properly, that is well below what is necessary for good video reproduction. The small capacitor, C_{14}, compensates for the large resistive losses of C_{13} at high frequencies. The large resistor, R_{25}, prevents any voltage from appearing at terminal J6 when an unterminated condition exists. The R_{24} resistor is a standard value that is used to match the output of Q_1 to the 75-ohm operating environment. The output of J_6 is conventional video with about a 1.25 volt peak-to peak measurement.

$$r_e \approx \frac{26\,Mv}{I_E} \qquad (6\text{-}7)$$

Due to the high gains and high frequencies involved, the LM1823 circuit shown in Figure 6-7 was constructed on a ground plane printed circuit board. This situation also

required the use of several 0.01 microfarad capacitors for bypassing. These capacitors are shown in Figure 6-7, but they are not assigned circuit numbers. A reduced printed circuit mask is shown as Figure 6-8. The inductor values are nominal and a manufacturer part number is shown. Any other non-generic devices also have manufacturer part numbers associated with them.

Figure 6-1 indicates the need for a low pass filter. From the development of Chapter 5, most of the design information is known for this filter. A table of normalized Butterworth filter component values is consulted for the values illustrated in Figure 6-9.[43]

The component values of Figure 6-9 must now be converted to the frequency and impedance levels needed in this application. The impedance is of course 75-ohms and the 3 dB frequency from Chapter 5 is 1.8 MHz. This allows the use of scaling equations 6-8 and 6-9 to determine the actual component values. In equations 6-8 and 6-9, R is 75-ohms and f_c is 1.8 MHz. This provides the component values listed in Figure 6-10. The values shown

$$L = \frac{R\, L_{NORMALIZED}}{2\pi f_c} \tag{6-8}$$

$$L = \frac{C_{NORMALIZED}}{2\pi f_c R} \tag{6-9}$$

in Figure 6-10 are not the exact mathematical values from the equations, but are the component values of the standard value components used in this implementation. It should also be understood that the 75-ohm series input resistor is assumed to be provided by the device driving this filter and is not a part of the filter. The 75-ohm output termination is provided to insure proper filter response and is necessary because the devices driven by this

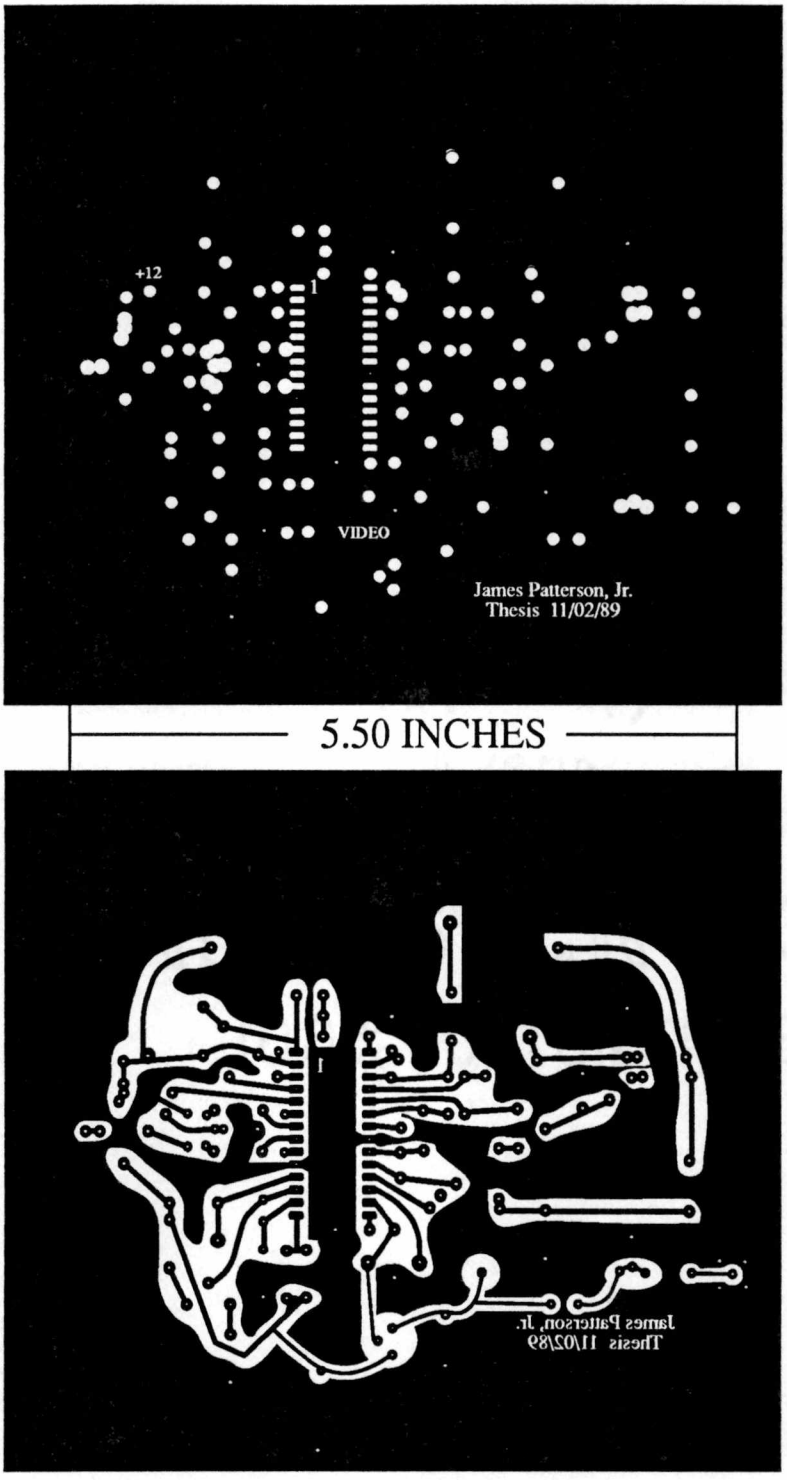


FIGURE 6-8

PRINTED CIRCUIT MASKS FOR THE DEMODULATOR

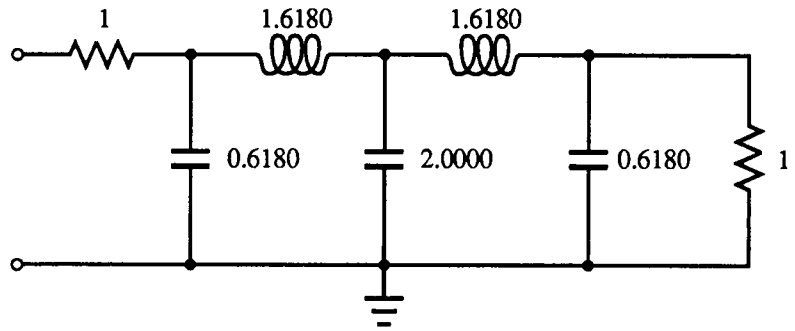


FIGURE 6-9

NORMALIZED LOW PASS FILTER CIRCUIT

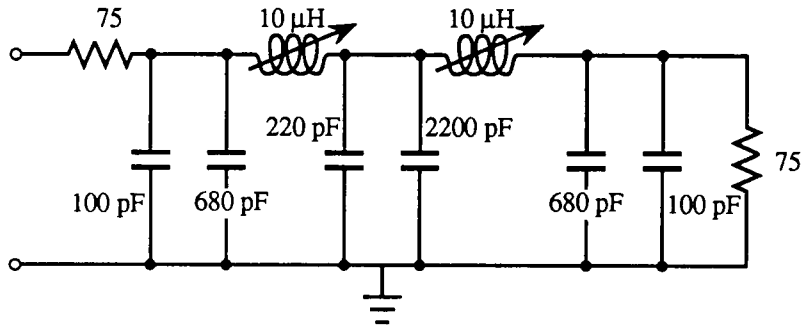


FIGURE 6-10

THE LOW PASS FILTER CIRCUIT

filter have high impedances compared to 75-ohms. For this reason the output of the filter does not use an RF or video connector, a simpler single-wire connection to the following circuits is sufficient.

Now that a well-filtered source of video is available, the sync separator demands attention. This is because most of the remaining circuitry depends on the sync separator for

timing information. The design of a sync separator is very easy thanks to the availability of sophisticated integrated circuits. The LM1881 is one of these integrated circuits. This device requires only a single +12 volt supply and a video input between 0.5 and 2 volts peak-to-peak. The LM1881 has an input impedance of 10K-ohms and will not load the filter output. The LM1881 is shown in Figure 6-11 along with the output timing diagrams. The output pulses of the LM1881 range from a low state of nearly ground to a high state of almost the supply voltage, much like the output states associated with CMOS logic devices. The LM1881 has an absolute maximum supply voltage of 13.2 volts, so care must be taken to prevent the +12 volt supply from exceeding this maximum value. The RC tank circuit connected to pin 6 provides timing information to the LM1881 about the vertical sync pulse and the value shown is for all NTSC video.

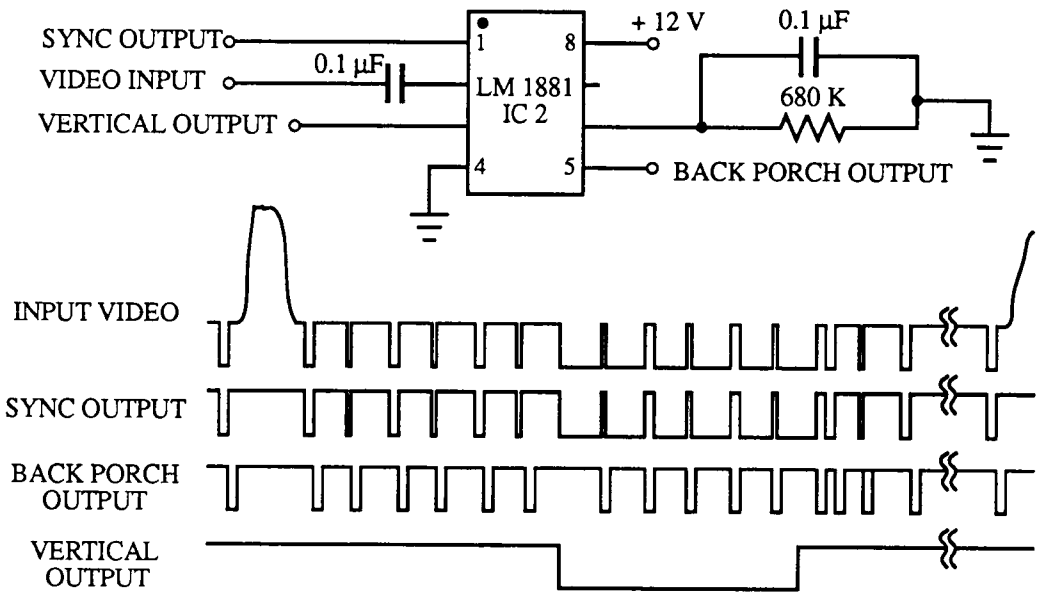


FIGURE 6-11

THE SYNC SEPARATOR CIRCUIT

The design of the timing control components is dictated by the LM1881. The manufacturer provides little information about interfacing the LM1881 to other devices. Without this information, it cannot be assumed that this device is capable of driving standard TTL logic devices. With the ground-to-supply logic swings of the LM1881 outputs, 4000-series CMOS logic devices seem to be a compatible choice. The low input current requirements of this logic family will not load the LM1881 excessively, and the logic swings of the LM1881 are directly compatible with this this logic family. This means that no interfacing drivers are needed and the logic devices may be directly connected to the LM1881 outputs. The 4000-series CMOS devices will also function with a single-ended +12 volt supply, so a logic voltage power supply is unnecessary. When a timer is needed, most designers will use the 555-type timer. This timer will not be easy to use with this particular application due to the fact that the input trigger pulse width exceeds the output pulse width. A 555-type timer will not respond in a predictable manner in this situation. A 4528 CMOS timer will function in this capacity and has a number of triggering and output options. The 4528 does lack the current driving capability of the 555-type timer and will require an interfacing driver be used in this situation.

The design of the zero-chop switch timing control is the next task. The Chapter 5 development of the timing control for the zero-chop switch did most of the difficult design work. All that is left is to implement the timing information provided in the Chapter 5 development. The control of the zero-chop switch is by one-shots. The CD4528 is a dual retriggerable monostable multivibrator that is ideal for this application. Being CMOS, this device will function with the +12 volts supply, will not load the LM1881, and two timers are in each package. A primary advantage of this timer is that the trigger is edge-triggered. The timer triggers by a low-to-high or a high-to-low transition, and the output is independent of the trigger pulse width. The timing of this device is set by an RC combination. A formula for the pulse width does not exist, but tables are provided by the manufacturer to select the

resistance and capacitance. Even with this table, the final pulse width must be set by use of a potentiometer which is replaced by the correct value of fixed resistance.

Figure 6-12 is a diagram of the timing control for the zero-chop switch. The horizontal sync output of the LM1881 is connected to the negative edge trigger of the first one-shot, which is one-half of IC3. The values of R_{30} and C_{17} are chosen for a pulse width of 33 microseconds. The pin 6 output of one-shot 1 is then fed into the negative-edge trigger of one-shot 2. One-shot 2 is the other half of IC3. This is the switch "off" timing pulse and is set for an initial value of 12 microseconds. R_{31} and C_{18} are chosen to attain this pulse width. The output of one-shot 2 is pin 10 and is used to control the driver of the zero-chop switch. Good logic construction techniques need to be followed with each timer having good grounding and supply decoupling.

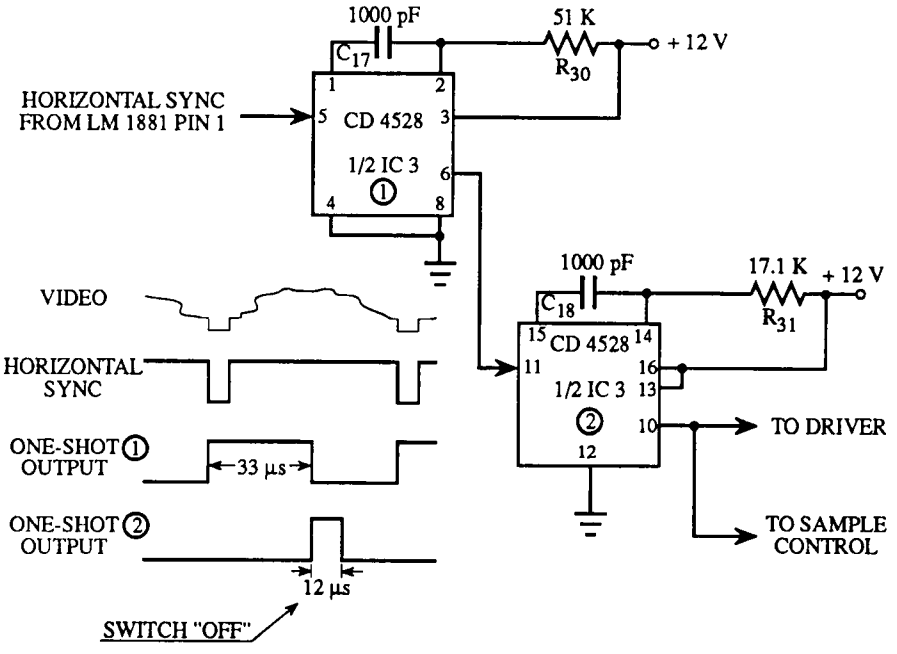


FIGURE 6-12

THE π ATTENUATOR CONTROL CIRCUIT

The driver is the circuit used to interface the timing control to the zero-chop switch. The 4528 does not have the output current necessary to drive the zero-chop switch directly. Several circuits could fulfill the design needs, but this design will use a two transistor logic switch. A complimentary output exists on the 4528 so a single transistor switch could be used, but the two transistor switch is a more reliable design. A linear driver, such as an emitter follower, could also be used as a driver. Figure 6-13 is a diagram of the driver circuit. Analytical means can be used to estimate the correct resistance values, but actual circuit testing is necessary to insure reliable performance. The reason for this is that most devices have a range that the parameters can exist in and this is usually not accurate enough for anything but a starting point. Q_2 is an NPN transistor. A reasonable current is assumed for the collector at saturation and the value of R_{33} is computed from this. The small signal current gain divides the collector current to compute the base current necessary to put the transistor into saturation. The base voltage is assumed to be 0.7 volts above ground and the high level input level is used to compute the base resistor R_{32}. This resistor must then be adjusted so that transistor Q_2 saturates on a high level input, but does not go into heavy saturation. The base current must also be low for cut-off, but the input levels are low enough to insure complete cut-off. The Q_2 logic switch is inverting and the Q_3 logic switch will also invert this for an overall non-inverted switching operation. The Q_3 transistor is a PNP device and is biased in a similar manner as Q_3. The base resistors of Q_3 are carefully selected for correct base current, and some degree of adjustment is available with the inclusion of the R_{35} potentiometer. The Q_3 collector resistors are also adjustable to insure full current and voltage swings necessary to operate the zero-chop switch. The collector of Q_3 is directly connected to the control of the PIN diode switch by means of an air core inductor. This arrangement is sufficient to drive the switch satisfactorily.

The demodulator and zero-chop switch are now operational and the filtered output is available for the next stage of the development: inverting gain. The intent of this amplifying

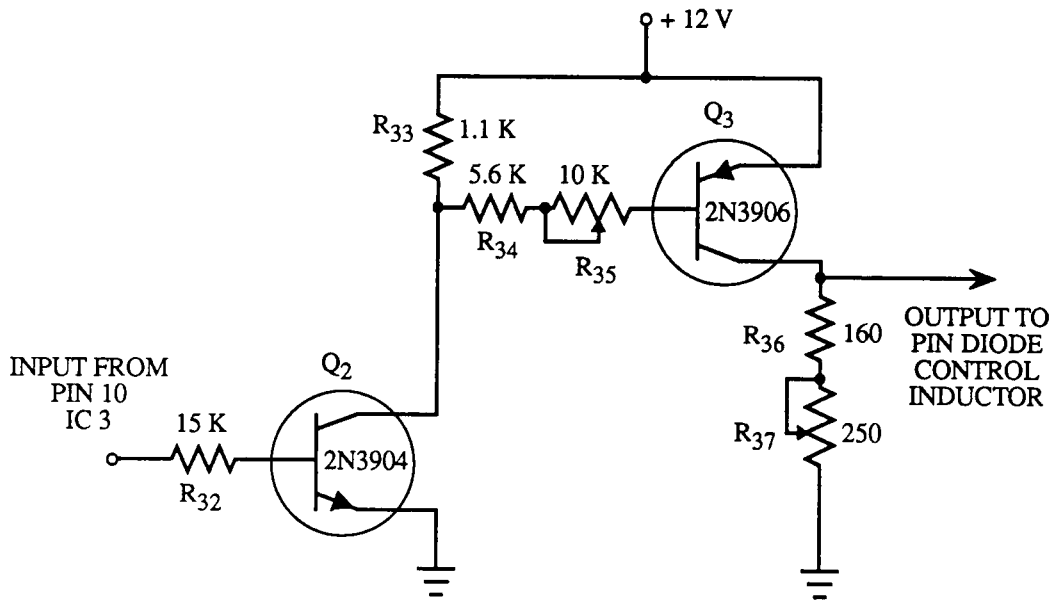


FIGURE 6-13

THE π ATTENUATOR DRIVER CIRCUIT

section is to provide some gain and DC offset to the filtered signal. This is easily accomplished by a single transistor common-emitter amplifier. The input will have to be AC coupled because of the low pass filter, but the output will be DC coupled to provide the required DC offset. The gain of this stage is controlled by allowing a small amount of emitter resistance to be unbypassed. This effectively controls the gain of the circuit and, unlike input dividers, has little effect on the signal-to-noise ratio of the signal. The gain of a configuration like this is given by equation 6-10, with the resistors numbered as in Figure 6-14 which is the diagram of this amplifier. The gain of this amplifier must be set to accept input levels from about 0.5 to 2 volts peak-to-peak. The voltage gain in this circuit is about -6.5. The input bias resistors R_{40} and R_{41} are chosen so that the base is very near zero volts. This helps to insure that the output has sufficient headroom and that the output is always positive. This will help later because a divider functions much better when all the signals are positive. Q_4 has a high

$$-A_v \approx \frac{R_{42}}{r_{eQ4} + R_{43}} \approx \frac{R_{42}}{R_{43}} \qquad (6\text{-}10)$$

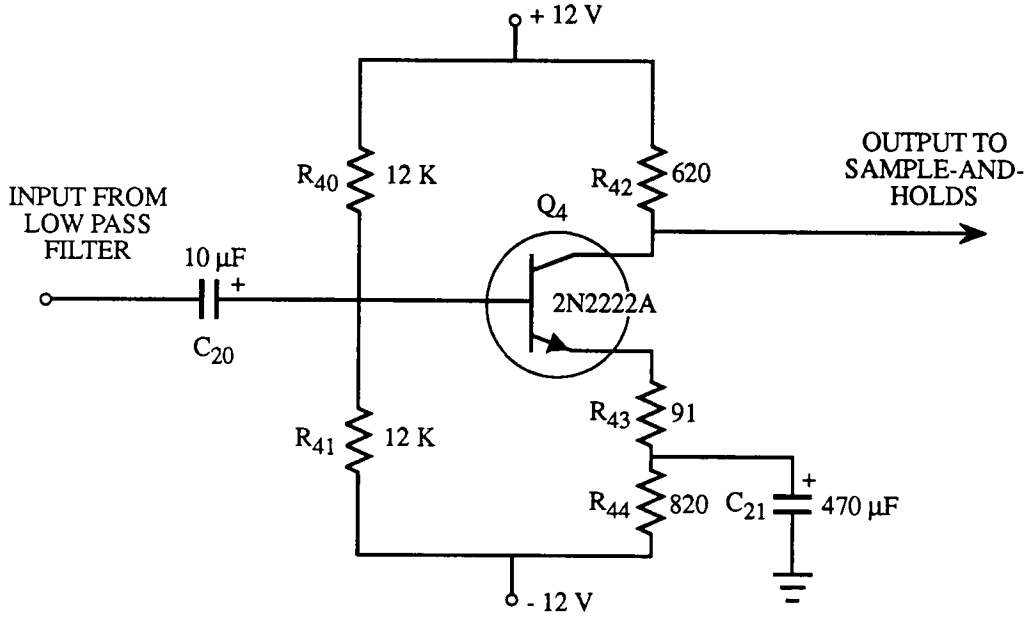


FIGURE 6-14

THE INVERTING VIDEO AMPLIFIER CIRCUIT

impedance input, so there is no loading on the low pass filter. The amplifier drives sample-and-hold circuits with high impedances, so there is very little external load on Q_4.

Video output is now available to the sample-and-hold circuits for measurement. It is necessary to design timing controls for the sampling. The sampling control for the sync sample-and-hold circuit is simple to design. Chapter 5 already determined that a one-shot with a 4 microsecond pulse width was sufficient to control the sampling. A 4528 is again used as a one-shot and is illustrated in Figure 6-15. The triggering input is the horizontal sync output of the LM1881 sync separator. The negative-edge trigger of one-shot 1 of IC4 is pin

5. R_{46} and C_{23} set the sampling pulse width at 4 microseconds. This pulse is output on pin 6 and directly drives the sampling logic control of the sync sample-and-hold circuit. Pin 3 is the "clear" control and will be used to prevent sampling during the equalizing pulses. When pin 3 is high the one-shot functions as normal, when pin 3 is low the output pulse is disabled. The sampling control for the blanking is identical except the trigger signal from the LM1881 sync separator is the back porch pulse. Figure 6-15 also contains this circuit. Resistor R_{47} and C_{24} are selected for a 4 microsecond pulse output. Notice that the resistances of R_{47} and R_{46} are different for the same pulse width. This is primarily due to the capacitor tolerance and is the reason for careful selection of the resistor. The control signal is output on pin 10 and controls the sampling of the blanking sample-and-hold directly. Pin 13 is the clear input and is connected to pin 3 to prevent sampling during equalizing pulses.

The equalizing pulse sampling problem has just been mentioned and this is a good place to design this avoidance circuit. The equalizing pulses must be avoided to insure

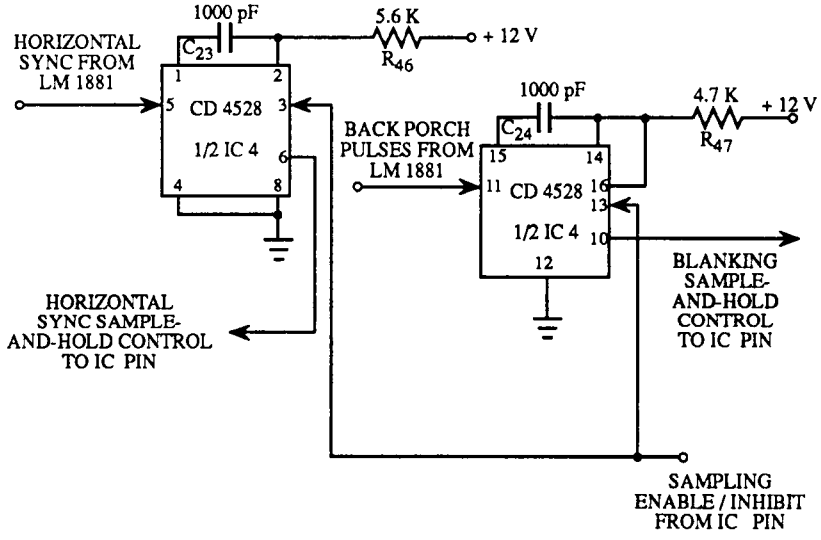


FIGURE 6-15

THE SYNC AND BLANKING SAMPLING CONTROL CIRCUIT

accurate sampling. Chapter 5 developed a method of avoidance using two one-shots. Figure 6-16 is a diagram of the circuit realization of this development. The vertical pulse from the LM1881 sync separator is the negative-edge that triggers one-shot 1 of IC6, which is also a 4528. R_{49} and C_{26} are chosen to output a pulse width of 500 microseconds, which is well beyond the last equalizing pulse. The pin 6 negative-edge triggers input pin 11 of one-shot 2, which is the other half of IC6. The pulse width of this one-shot is set by R_{50} and C_{27}. This pulse width lasts for most of the field, so it is quite long: about 15.7 milliseconds. This pulse width is so great that accurately determining exactly when the pulse will end is difficult. This tolerance is accounted for by assuming a "time-out" that is $\pm$ 4H. This long pulse is applied directly to the clear pins, 3 and 13, of IC4. When this pulse is occurring, IC4 functions normally. When the pulse ends, IC4 is inhibited and does not function as a one-shot until the clear line goes to the high state.

The last portion of sampling control has to do with sampling the zero reference. This reference is generated by the action of the zero-chop switch and because of the abruptness of the switching may create transients during a portion of the reference. These transients will exist at the beginning and end of the switching operation. To avoid these transients, two one-shots are used to delay sampling until after the start-up transients and to stop the sampling process prior to the shut-down transients. The circuit that performs this operation is shown in Figure 6-17. This circuit is of the same nature as the zero-chop switching timing controller. The onset of zero-chop switching is relayed to one-shot 1 of IC5 from pin 10 of IC3. This positive-edge triggers one-shot 1 which has R_{52} and C_{29} selected for a pulse width of about 3 microseconds. The negative-edge of this pulse triggers one-shot 2 of IC5 which generates the 4 microsecond pulse that controls the sampling of the reference level. This pulse times-out about 3 microseconds before the switch "off" period ends. The sampling control pulse appears on pin 10 of IC5 and directly controls the logic input pin 8 of the sample-and-hold circuit.

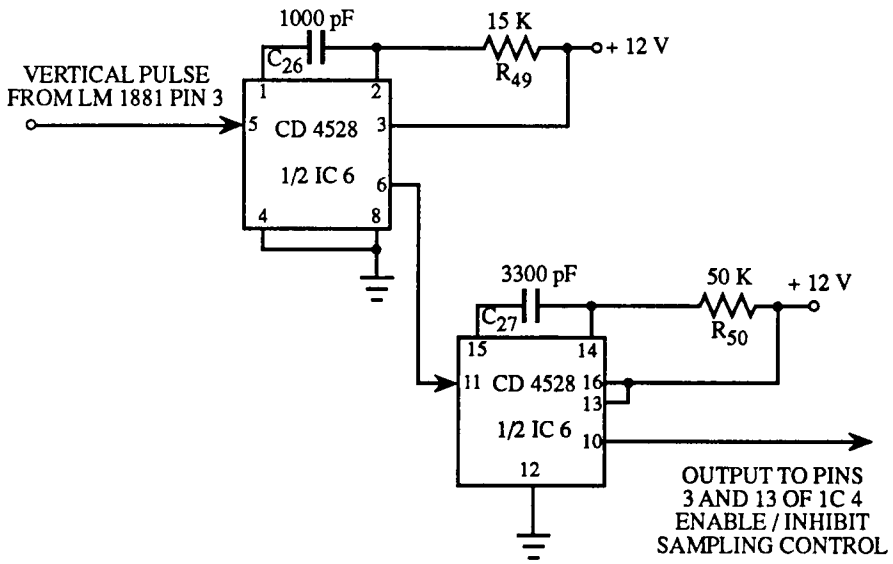


FIGURE 6-16

THE CIRCUIT THAT PREVENTS EQUALIZING PULSE COLLISIONS

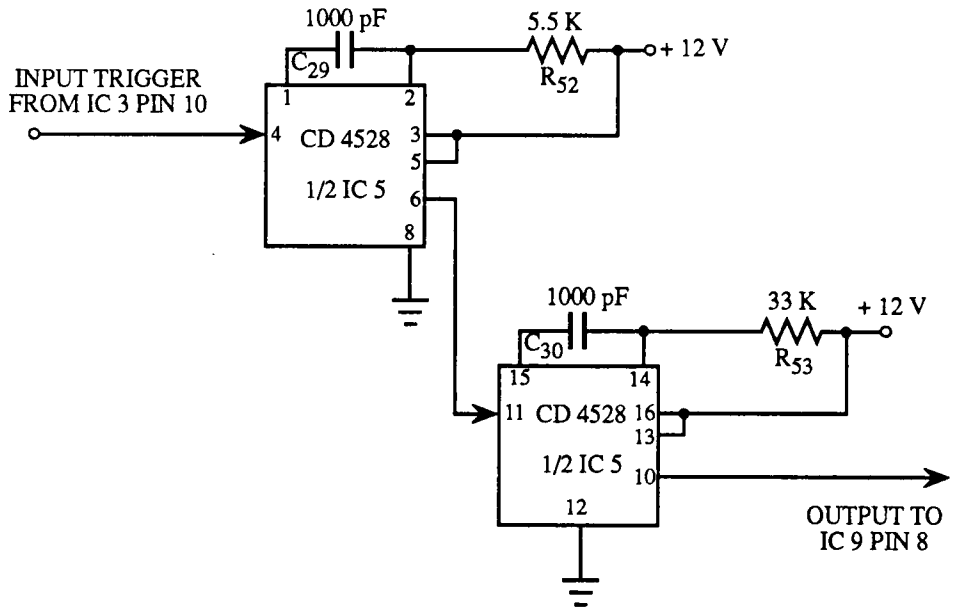


FIGURE 6-17

THE REFERENCE SAMPLING CONTROL CIRCUIT

The sampling control signals are now available to the sample-and-hold circuits. The LF398 is the sample-and-hold device used in this application. This device is very easy to apply and has the speed necessary to capture these short-duration signals.The sample-and-hold circuit is shown in Figure 6-18. The LF398 has a sampling logic control which is adjustable to a variety of triggering signals. The logic control must be set-up to place the control signal +7 to -40 volts with respect to the logic reference. Application information on CMOS logic references from the manufacturer appears to be in error by allowing the +7 volt limit to be exceeded by any supply voltage greater than about 11 volts. In a conservative manner, this reference was altered to prevent the logic high from exceeding +6 volts above the logic reference. The resistive divider composed of resistors R_{55} and R_{56} is used to set the logic reference. The capacitor attached to pin 6 is the hold capacitor and is a polystyrene design to reduce leakage. The signal being measured is the inverted video from the amplifier and is fed into pin 3 of all three LF398 sample-and-holds. The various sampling controls will determine the portions of this input signal that are sampled. It was first thought that the output of the reference sample-and-hold could be applied to the normally grounded side of the hold capacitor of the sync and blanking sample-and-hold circuits and preclude the need for subtractive circuitry. This did not work, probably because the LF398 generates a virtual ground reference from the ± 12 volt supplies, although this is not known with any certainty. Pin 5 is the output pin on the LF398 and contains the sampled information. This output is then applied to the subtractive circuits.

The subtractive circuits are responsible for determining the levels between the sync and the reference and the level between the blanking and the reference. The reference is the sampled level during the zero carrier reference zero-chop. The sync and blanking levels are sampled during their respective intervals. The subtraction is accomplished by the use of a simple op-amp circuit shown in Figure 6-19. The reference is applied to the inverting input of the op-amp which has an inverting gain of two, but also has an input resistive divider with

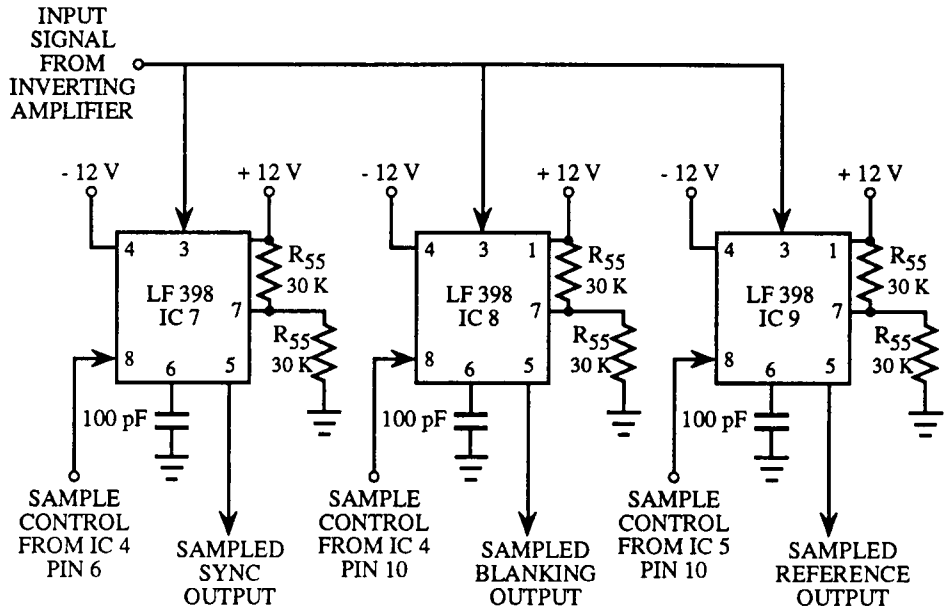


FIGURE 6-18

THE SAMPLE-AND-HOLD CIRCUITS

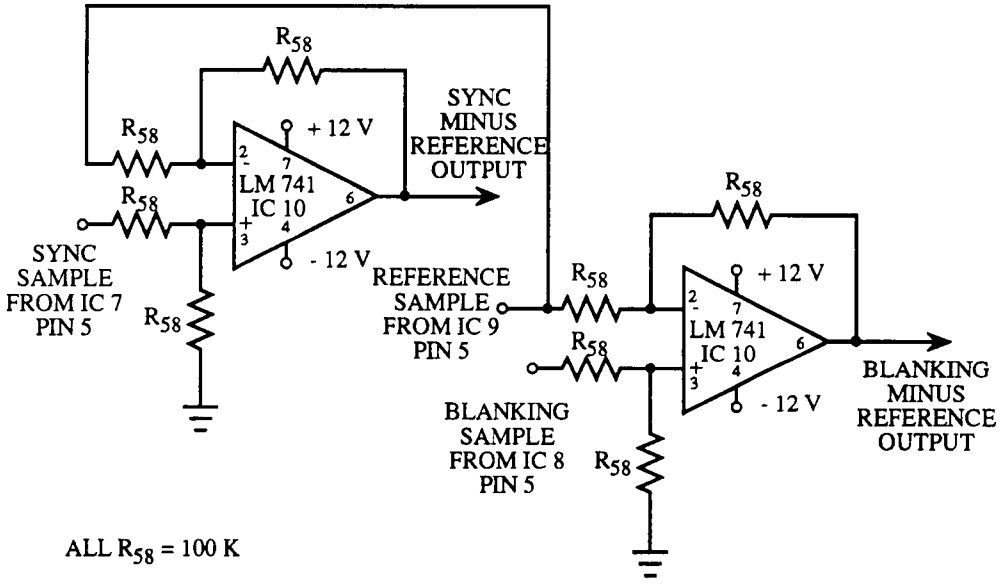


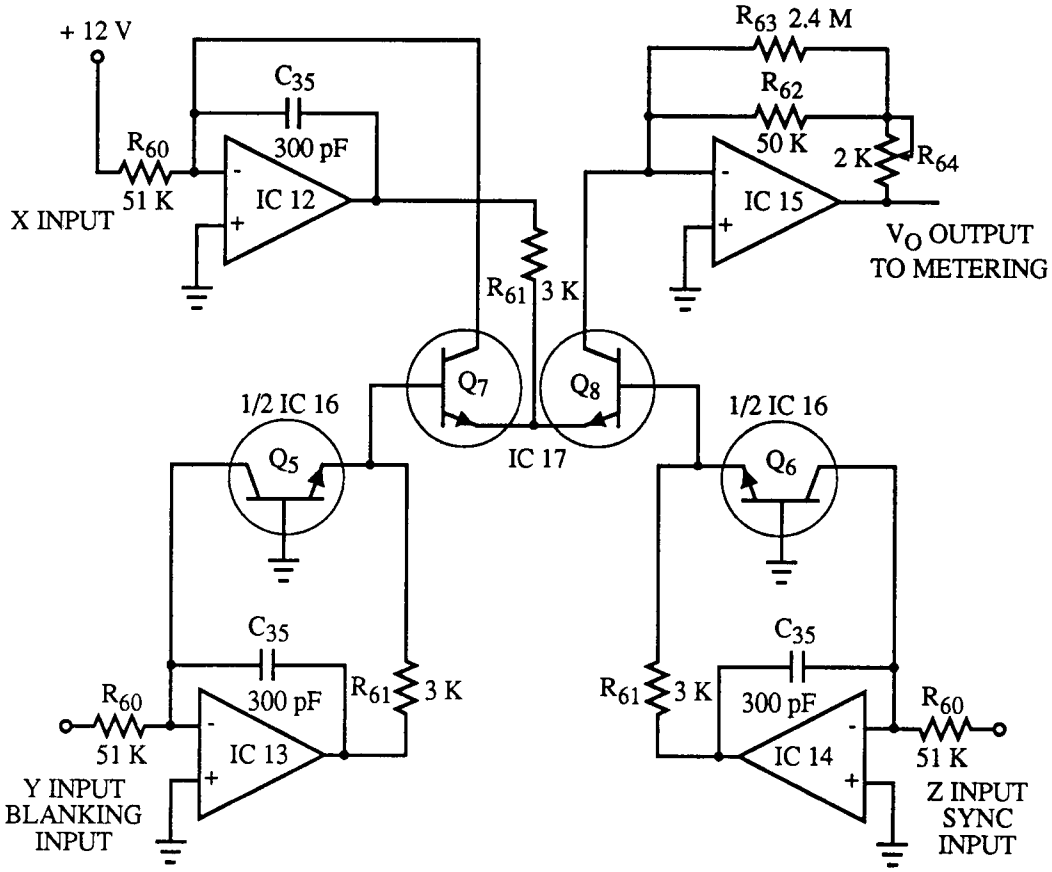
FIGURE 6-19

THE SUBTRACTIVE CIRCUIT

a gain of 0.5. This gives an inverting gain of 1. The sync or blanking is applied to the non-inverting input, which is set for a gain of 1. This will tend to subtract the input signals, giving a output that is either the sync minus the reference or the blanking minus the reference. The resistors, all labeled R_{58}, are identical and are so because of maintaining a consistent input impedance on both inputs. The op-amps in the design are LM741, but OP-27 op-amps are used in the actual implementation. The LM741 should also work correctly in this design, but the offset voltage should be checked closely.

The measured levels are now ready for comparison. This will be accomplished with an analog divider. This is a National Semiconductor application[44] using op-amps and matched transistors. Most divider circuits include thermistors to offset the temperature dependence of the transistor junction used as the source of the logarithm. This circuit is one of the few that relies on matched transistors for temperature compensation. The circuit diagram of this implementation is shown in Figure 6-20, and is not quite as complicated as it first appears.

In analyzing the circuit in Figure 6-20, some of the circuit components are not directly affecting the process occurring. The capacitors marked C_{35} are used for compensation of the op-amp and although necessary for high performance, these capacitors may be ignored to preform a basic analysis.[45] Resistors R_{61} are also a portion of this compensation. Beginning with op-amp IC13, this op-amp is configured as a grounded-base transistor log amplifier. The input voltage of IC13 creates an input current by virtue of the virtual ground at the inverting input. This situation forces the op-amp to maintain a collector current through the transistor Q_5 that is equal to the input current. This is due to the feedback. The emitter-base voltage varies logarithmically with variations in the collector current.[46] Equation 6-11 is a simplified equation for the input-output relationship for this log amplifier. The I_s term in equation 6-



IC 12 = 1C 13 = IC 14 = 1C 15 = LM 308
IC 16 = 1C 17 = LM 394

FIGURE 6-20

THE ANALOG DIVIDER CIRCUIT

11 is the reverse saturation current, and the kT/q term is indicative of the temperature dependence. The circuit in Figure 6-20 is premised on making these two terms constant in all the log amplifiers by using very well matched transistors. If this is assumed, the expression then becomes that shown in equation 6-12.

Equation 6-12 simplifies the analysis of the amplifiers. Equation 6-12 describes the

$$V_o \approx -\frac{kT}{q} \log\left(\frac{V_{in}}{R_{60}\ I_s}\right) \tag{6-11}$$

$$V_o \approx -\log\left(\frac{V_{in}}{R_{60}}\right) \tag{6-12}$$

output of amplifier 13 and is the "Y" input term. This equation also describes the output of amplifier 14, which is the "Z" input term. The amplifier associated with IC12 does not have a grounded-base, but rather the base is attached to the Q_5 emitter. This references the log action of Q_7 on the log of the "Y" input. This causes the log of the "X" input to be added to the log of the "Y" input. This summed log output is then applied to the emitter of Q_8. The emitter-base voltage of Q_8 is the difference between the "X" and "Y" summation and the "Z" input. This leads to the the generation of a collector current from Q_8 that is proportional to the inverse log of the emitter-base voltage. This leads directly to equation 6-13, which also assumes that the log action contains the same constants as in equation 6-12. This causes these constants to "drop-out" in the inverse log process.The current I_{in} is then converted to voltage in IC15. This leads to the input-output relationship of equation 6-14. The R_{60} and R_{62} values are chosen in such a way as to reduce equation 6-14 to the final form in equation 6-15. The circuit used ties the "X" input to the positive supply for scaling, and the sync is input to the "Z" terminal, and the blanking is input to the "Y" terminal. This will output a voltage of 9 volts when the proper modulation depth is obtained.

$$I_{in} = \log^{-1}\left[\left(\log\left(\frac{Y}{R_{60}}\right) + \log\left(\frac{X}{R_{60}}\right)\right) - \log\left(\frac{Z}{R_{60}}\right)\right] \tag{6-13}$$

$$V_o = \left[\frac{\left(\frac{X}{R_{60}}\right)\left(\frac{Y}{R_{60}}\right)}{\left(\frac{Z}{R_{60}}\right)} \right] (R_{62}) \qquad (6\text{-}14)$$

$$V_o = \frac{(X)(Y)}{(Z)} \qquad (6\text{-}15)$$

The divider output voltage is used to drive the 50 microampere meter in Figure 6-21. A resistive divider consisting of R66 and R67 is used to place a reference voltage at 10 volts. This is done to reverse the action of the meter and cause an increase in modulation depth to be displayed as a clock-wise movement of the meter pointer. Resistor R68 is chosen to set the current levels through the meter. This is used to calibrate the meter. Mathematically computing the divider output voltage for various levels of modulation is not very complicated and Table 6-1 is a listing of some of these values.

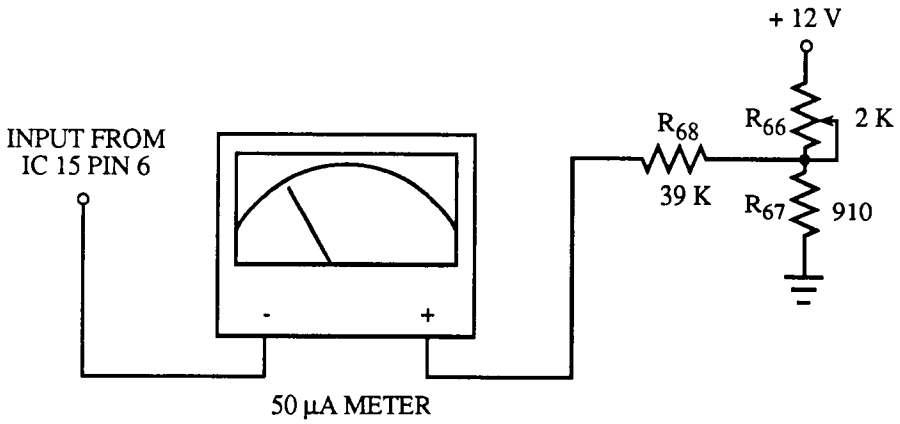


FIGURE 6-21

THE OUTPUT METERING CIRCUIT

TABLE 6-1

THE COMPUTED DIVIDER OUTPUT VOLTAGES FOR VARIOUS VALUES OF MODULATION DEPTH

PERCENTAGE OF MODULATION DEPTH (Peak White)	DIVIDER OUTPUT (Volts)
0	8.57
5	8.74
10	8.92
12.5	9.00
15	9.09
20	9.26
25	9.43
30	9.60
35	9.77

Measuring the performance of the circuit is very important as it experimentally verifies the measurement technique. The following performance tests were designed to check the accuracy and linearity of the various sections of the circuit, both individually and as a whole. This will give some insight into possible performance enhancements.

The first test will be concerned with all the sections following the demodulator. This is an input-output test using video as the input signal and comparing it with the divider output. This is an indication of the accuracy and linearity of this section of the circuit. The graph in

Figure 6-22 plots the error measured in this test. Although the error illustrated in this figure is not insignificant, it does fall within the previously discussed error requirements.

Figure 6-22 indicates an error in measurement that is reasonable. Further investiga-

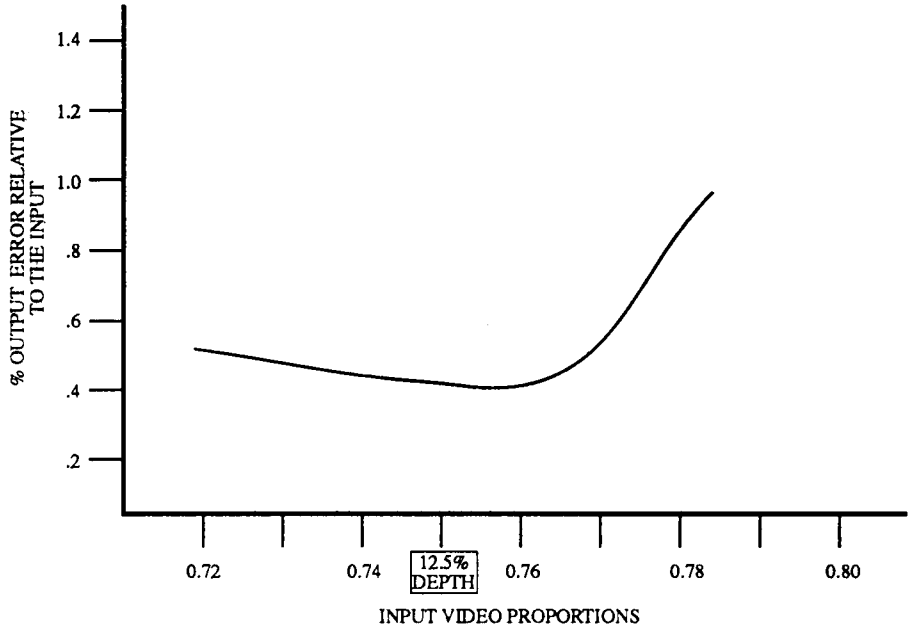


FIGURE 6-22

A PLOT OF THE ERROR OF THE VIDEO MEASURING SECTION

tion indicates a primary source of error being the sample-and-hold process. Hold errors of up to 80 millivolts were measured and contribute about 60% of the total error of Figure 6-22. This error is not a "hold step," but is an error between successive samples. There are several potential causes for this error. The video sample could be affected by poor power supply filtering, and in essence "riding" the power supply ripple. Another cause of error could be the quality of the video test signal generated by the video generator. Instrumentation was not available to measure for this situation. The last highly probable source of error is interference

from subcarriers that the low pass filter should delete. Figure 6-23 is a graph of the frequency response of the low pass filter.

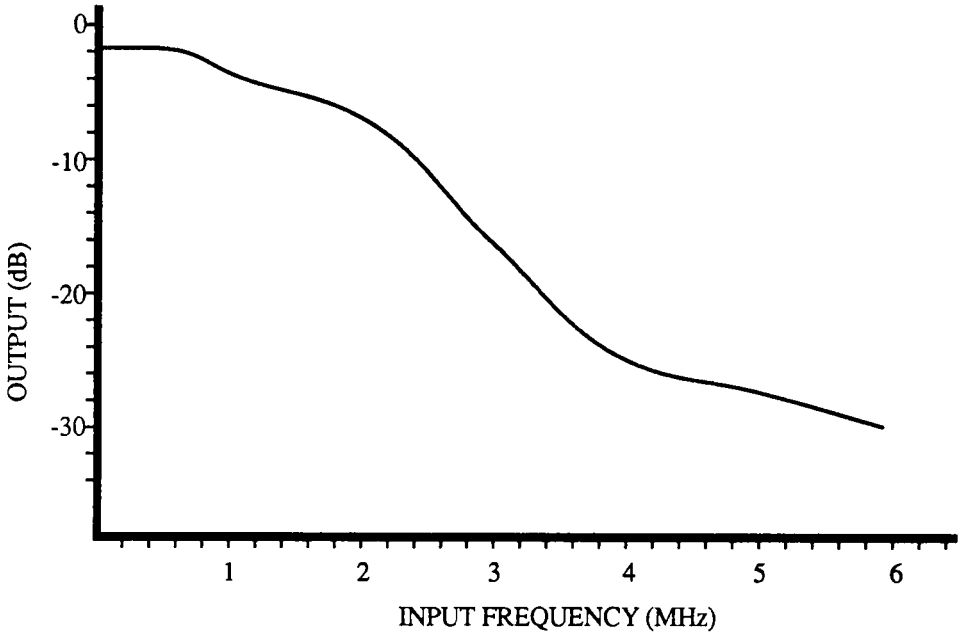


FIGURE 6-23

THE MEASURED FREQUENCY RESPONSE OF THE LOW PASS FILTER

Figure 6-23 indicates that the performance of the low pass filter is less that the design specified. This is probably due to incorrect component values. Suppressing the chroma subcarrier at the test generator yielded about a 25% error reduction, which tends to support the assumption of error from interfering subcarriers. Further reductions in error could probably be accomplished by closely adjusting the characteristics of the subtractive circuit. The validity of this statement could not be tested, due primarily to the large dynamic ranges of the signals in these circuits.

The final test is the RF-input to meter-output, overall system performance test. It is important to note the method used to perform this test. A highly accurate modulation measuring system, such as a spectrum analyzer, was not available for this test. Instead, broadcast television stations were used as test signals. Four high quality VHF off-air television channels were available at the test site. Two of these stations had signal levels sufficient to check the antenna preamplifier for sync compression. This test indicated no sync compression problem associated with the use of this preamplifier. These amplified signals were fed to a variable RF attenuator and then to the input of the converter. The RF attenuator was varied to check the accuracy of the system at varying levels of input signal. Figure 6-24 plots this error against the input levels. This error plot is an average of the output of the

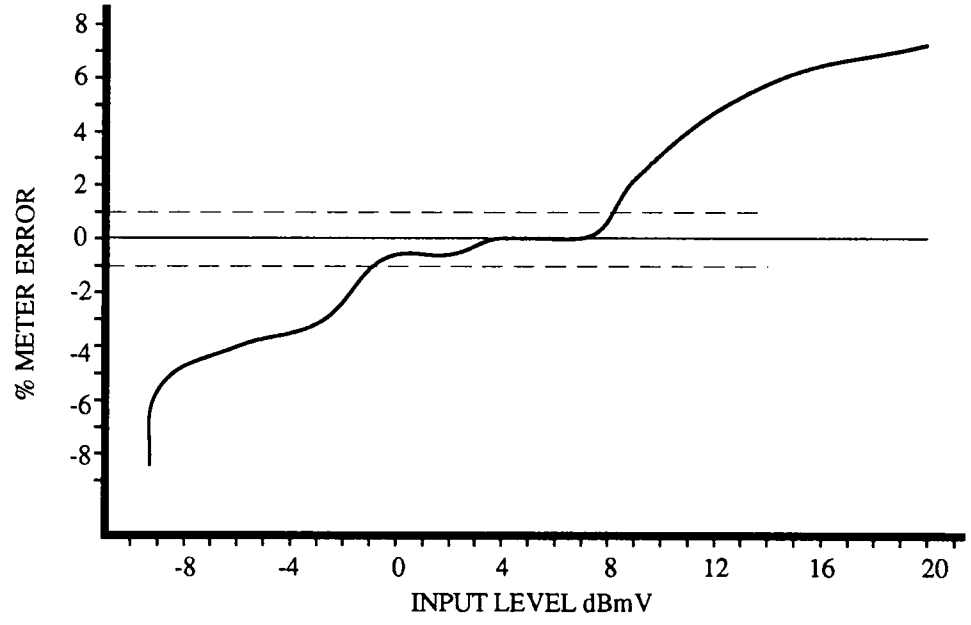


FIGURE 6-24

TOTAL SYSTEM ERROR VERSUS INPUT SIGNAL LEVEL

circuit in response to the four input signals. Figure 6-25 is an enlarged view of a portion of Figure 6-24 and shows the four signal responses independently. The signal responses were very close, indicating a high probability that the input test signals form a valid testing method. The disadvantage of this testing method is that it cannot check a range of input modulation levels. This is not quite as important as it seems since the proper modulation depth is exactly 12.5% peak white. A relatively high degree of error in measurement away from this value can be tolerated as long as the 12.5% indication is accurate. The output was measured with a digital volt meter for higher accuracy, but since the output is not well-filtered DC a slight measurement error could have been introduced. This was not checked.

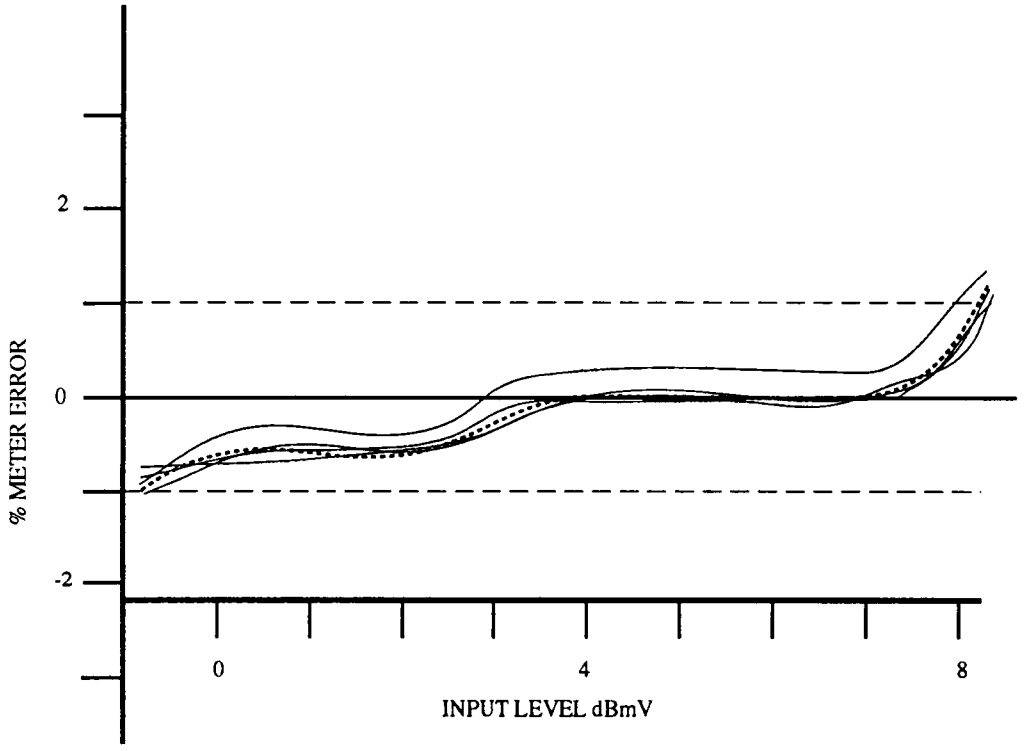


FIGURE 6-25

AN EXPANDED VIEW OF FIGURE 6-24 WITH ALL FOUR TEST SIGNALS OUTPUTS PLOTTED

Figure 6-24 indicates a high error situation exists for large variations in signal input level. This error was traced to the LM 1823 demodulator. An oscilloscope displayed a significant signal compression as the input level changed. A correction for this problem within the LM 1823 could not be found. This indicates the need to closely monitor the level of the input signal to insure accurate results.

The preceding information allows some conclusions to be drawn. The first is that this development was truly a success. Experimental results confirmed that the measurement method is a valid one. The accuracy indicated by the limited testing methods is very encouraging. Hindsight reveals several improvements that could be made to the circuit to improve its performance. The use of a counter to avoid equalizing pulse collisions would be a more accurate method than the one used. An input RF AGC system could be employed to assure the input RF levels fall within the most accurate portion of the testing range.

This thesis has covered a great deal of material to reach this point. Several chapters were devoted solely for tutorial use, before development actually began. The development process was also a lengthy process. A measurement method was indeed found, and the validity of the video proportion measuring method proved experimentally. The circuit realization did not perform perfectly, but perhaps more importantly, it was a success.

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