DESIGN OF A VIDEO MEASUREMENT SYSTEM

BY

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DEDICATION

I wish to thank my parents, Mr. and Mrs. D. R. Bailey for raising me and making it possible for me to attend engineering school.

I also wish to thank Professor Pratt and Professor Ricci for their encouragement and support during my studies leading to my MSEE degree.

Also, I am grateful for the off campus EE program which made it possible for me to pursue a graduate degree while working at a full time job.

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1.0 INTRODUCTION

This thesis describes the design of a video measurement system using digital signal processing methods. The basic system is intended for use with the NATIONAL TELEVISION SYSTEM COMMITTEE (NTSC) type of video system transmitted in North and Central America as well as Japan. However the system can be modified rather simply to test the PHASE ALTERATION LINE system (PAL).

From the time a video system is installed until the end of it's useful life it requires periodic testing to verify proper operation. The term system can refer to a single device such as a television camera or tape machine or a complete satellite distribution system where the signal source can be on one continent and the receiver on another. In any case a TV signal is generated somewhere and received somewhere.

Over the past several decades a number of industry accepted procedures have evolved to test various parameters of a TV signal. These involve the use of standard test signals with standard measurement techniques. The world's three main TV systems are all similar in that they are composed of lines transmitted as analog signals with synchronizing pulses. The line and field rates are the same for PAL and the SEQUENTIAL A' MEMOIRE system (SECAM) while the line and field rates for NTSC are different. All three systems use different methods of transmitting color information.

In the United States the standard test signals and the methods of interpreting these test signals are contained in a document named "NTC 7". NTC 7 was jointly written by AT&T plus the three commercial networks and PBS. The document was published by PBS. NTC 7 was written during the 1970's when all or almost all long distance network television distribution was done using terrestrial systems, hence the term "Long Lines" which usually referred to multi hop microwave radios. About the only use of satellites at that time was transcontintental or intercontinental services such as a "back haul" from the location where a program was produced to a network origination point.

With the advent of satellites for network distribution of TV programming in the USA, Canada and other countries, the methods and signals described in NTC 7 have remained valid. However certain measurement tolerances have changed to reflect the higher quality of the signals delivered by satellite.

The techniques described in NTC 7 involve the use of a special oscilloscope (a television waveform monitor) to make various measurements on the test signals. This thesis describes algorithms developed using digital signal processing techniques to make certain of the measurements described in NTC 7.

NTC 7 consists mainly of tests that can be performed on a video system while it is carrying television programming. These are the in service tests. It is these in service tests which are addressed in this thesis.

2.0 THESIS DESCRIPTION

In past years, standard test signals have been developed for testing communications systems which carry television signals. These test signals are compatible with the television signals themselves.

The test signals addressed in this paper are inserted in the Vertical Blanking Interval (VBI) of the TV signal. The VBI is the part of the signal which corresponds to the time between the bottom of the screen and the top of the screen after the picture is scanned from top to bottom. Using the VBI to carry test signals has the advantage that the system carrying the TV signals does not need to be taken out of service to be tested. Traditionally, these measurements are made by viewing the test signals on TV waveform monitors. With the recent development of Digital Signal Processing (DSP) technology, and corresponding development of computers and processors, it is now feasible to use these techniques to perform these measurements. It is to this end that this thesis is directed.

This thesis describes a system which will perform these measurements using DSP techniques.

For the ease of the reader these measurements are described briefly below. More detail on these measurements may be found in NTC7.

There are two test signals used in NTC7. They are copied from NTC7 and shown here in figures 1 and 2.

Insertion Gain is the amount that the amplitude of a television signal is increased or decreased by a transmission system. The level going into the system is set at 1 volt, or 140 IRE from line bar to negative tip of sync pulse. The line bar is 100 IRE above the base line and the sync pulse is 40 IRE below the base line. At the receiving end, the level of the line bar is measured above the base line.

Measurement of Line-Time Waveform Distortion is accomplished by again using the Line Bar Test Signal. The measurement is made by normalizing the center of the bar to 100 IRE and measuring the leading and trailing edges not including the first and last microsecond. This is sometimes referred to as "Bar Tilt", and is an indica-

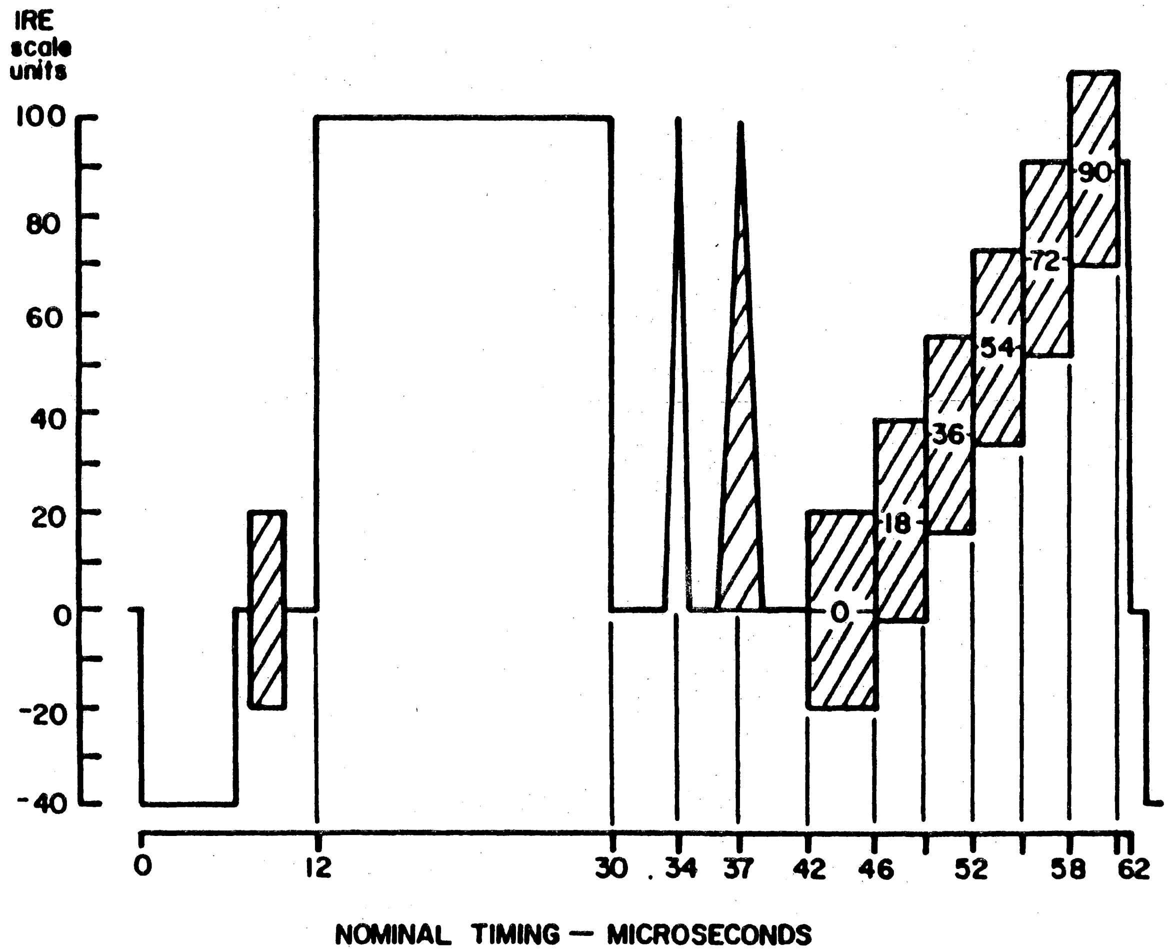


FIGURE 1 THE COMPOSITE TEST SIGNAL
(COPIED FROM NTC 7)

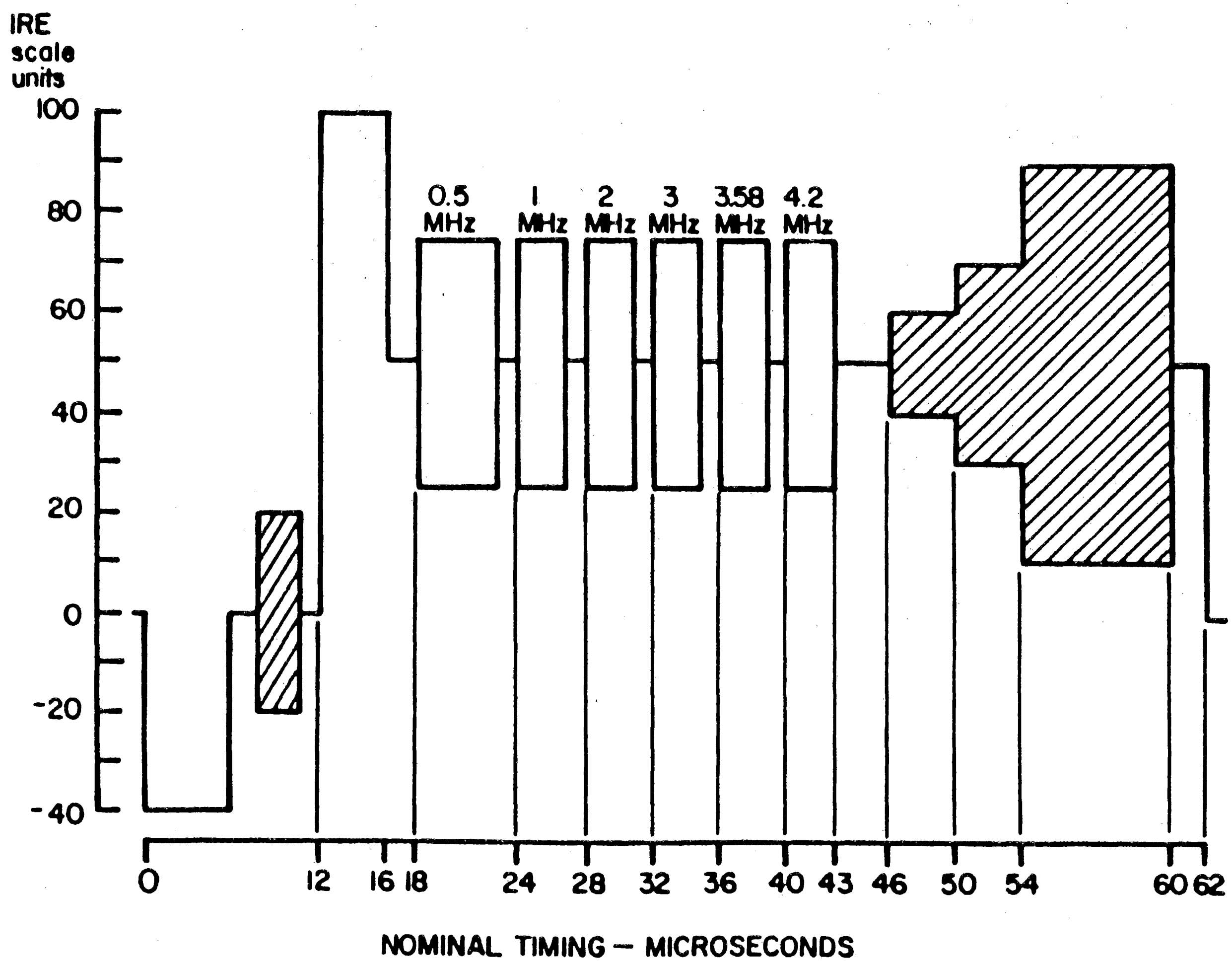


FIGURE 2 THE COMBINATION TEST SIGNAL
(COPIED FROM NTC 7)

tion of low frequency anomalies.

<u>Short-Time Waveform Distortion</u> is measured using the Line Bar Test Signal plus the 2T Pulse. Two measurements are made, both with the center of the Line Bar Test signal normalized to 100 IRE. First the height of the 2T Pulse is measured. Next the ringing or overshoot is measured at the leading edge of the bar and after the trailing edge of the bar. These peak-to-peak measurements are made within the first 1 microsecond of the bar and within the first 1 microsecond after the signal transitions from the bar to the base line.

<u>Chrominance-Luminance Gain Inequality</u> is measured using the 12.5 T Pulse. This pulse has two components, a chrominance portion and a luminance portion with a half amplitude time of 1562.6 nanoseconds. This parameter provides a way of stating the relative amplitudes of the 12.5 T Pulse and Line Bar Signal after they have passed through a television transmission system. The Line Bar Test Signal is normalized to 100 IRE at the receiving end of the transmission system prior to making the measurement. Chrominance-Luminance Gain Inequality is the

top of the 12.5 T Pulse when the Line Bar Test Signal is normalized to 100 IRE.

Chrominance-Luminance Delay Inequality is also measured using the 12.5 T Pulse. In this case however, the amplitude of the 12.5 T Pulse is normalized to 100 IRE at the receiving end of the transmission system before commencement of the measurement. Chrominance-Luminance Delay Inequality is a way of specifying the relative time shift between the luminance and chrominance portions of the 12.5 T Pulse after passage through a transmission system.

Measurement of Gain/Frequency Distortion is done using the frequency bursts at 0.5 MHz, 1.0 MHz, 2.0 MHz, 3.0 MHz, 3.579545 MHz and 4.2 MHz. Before commencement of the measurement, the reference pulse to the left of the frequency bursts is normalized to 100 IRE. The peak-to-peak amplitude of each of the bursts is measured. This measurement amounts to a frequencty sweep at discrete frequencies.

Measurement of <u>Luminance-Non Linear Distortion</u> is made using the Modulated 5-Riser Staircase Test Signal. It is a measure of the relative heights of the steps.

The <u>Chrominance Non-Linear Gain Distortion</u> measurement is made using the 3-Level Chrominance Test Signal. It is defined as the ratios of the smallest and largest peak-to-peak amplitudes to the middle burst peak-to-peak amplitude, with the middle burst peak-to-peak amplitude normalized to 40 IRE.

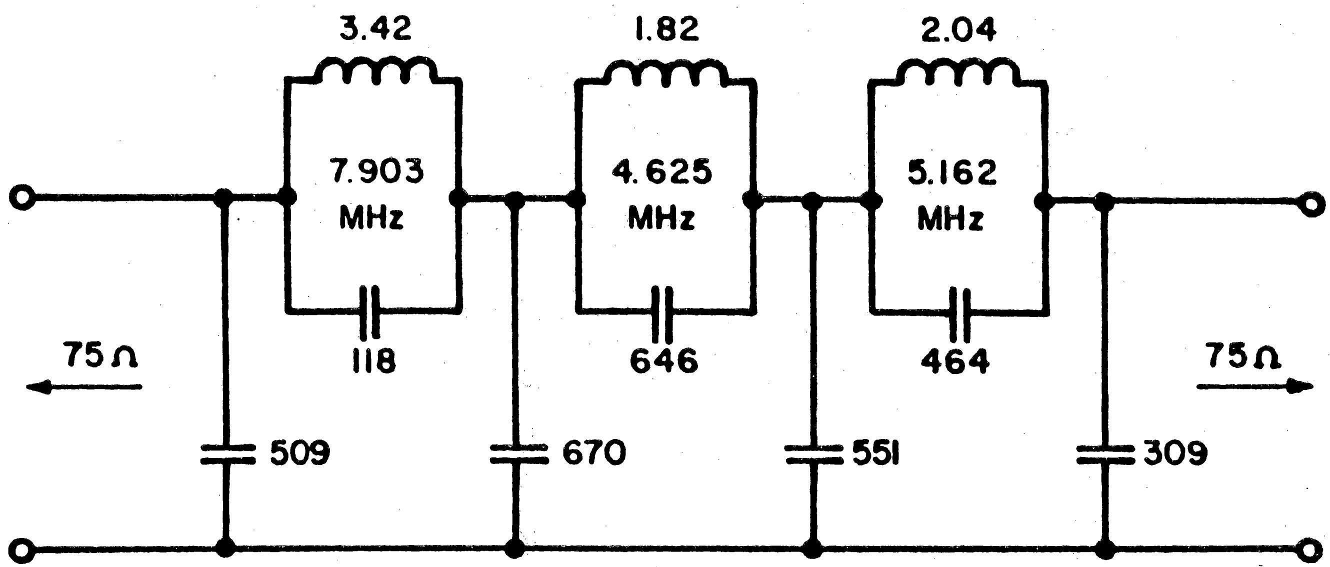
<u>Chrominance Non-Linear Phase Distortion</u> also makes use of the 3-Level Chrominance Test Signal. The phase of each of the three bursts is measured and the peak-to-peak phase variation of the three burst levels is calculated.

Measurement of <u>Differential Gain</u> uses the Modulated 5-Riser Staircase Test Signal. It is the difference between the peak-to-peak amplitude of the largest chrominance pulse (normalized to 100 IRE) and the peak-to-peak amplitude of the smallest chrominance pulse, as the luminance signal (the 5-Riser Staircase) is varied in amplitude.

The Modulated 5-Riser Staircase is used in the measurement of Differential Phase. Diferential Phase is the peak-to-peak phase change of the chrominance signal as the amplitude of the luminance signal (the 5-Riser Staircase) is changed in amplitude.

Measurement of Chrominance To Luminance Intermodulation is performed using the 3-Level Chrominance Signal. Chrominance-To-Luminance Intermodulation is defined as the variation of the luminance signal as a result of the chrominance signal, in this case The 3-Level Chrominance Signal. Traditionally the measurement is made by low pass filtering the chrominance portion of the signal and observing the resultant deviation of the luminance portion of the signal, which in this case is the DC level of the 3-Level Chrominance Signal.

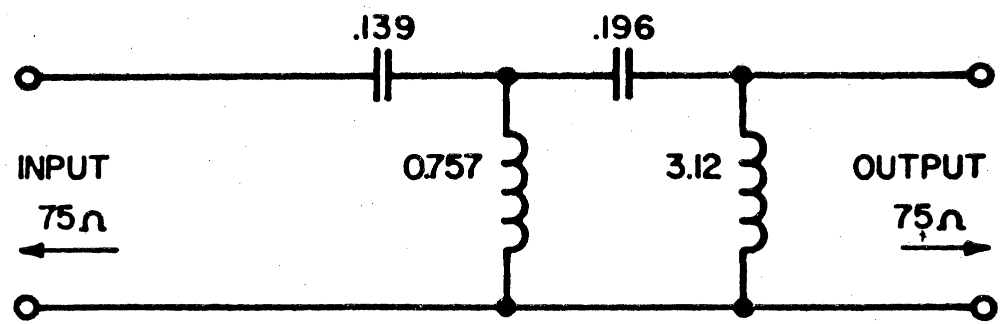
In the case of television, signal-to-noise is measured in terms of Peak-To-Peak Video (100 IRE = 0.714 volts) to RMS Weighted Noise Ratio in a defined channel. There are three filters used for this purpose. They are from page 52 of NTC7 and are shown here as figures 3, 4 and 5. The low pass filter shown in figure 3, is used to limit the spectral content of noise above 4.2 MHz. The high pass filter shown in figure 4 is used to limit noise below 10 KHz. This minimizes contributions from sources such as hum and other low frequency noise generators. Figure 5 shows the schematic for the weighting filter. This is a special low pass filter which simulates the response of the human visual system to high frequency noise. Thus rms noise measurements are made through all three of these filters in series.



Inductances are given in uH, and capacitances in pF. Q measured at 5 MHz is between 80 and 125 for all inductors.

Low-pass filter for use in noise measurements (f_c = 4.2 MHz)

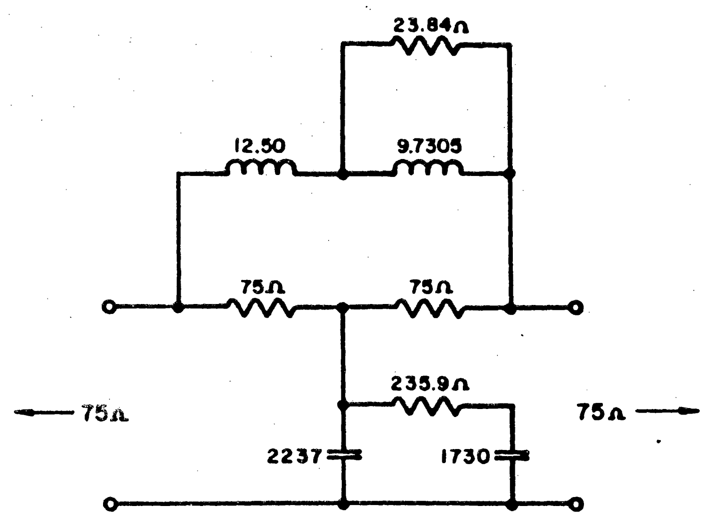
FIGURE 3 VIDEO LOW PASS CHANNEL SHAPING FILTER
(COPIED FROM NTC 7)



Inductances are given in MH, capacitances in pF. Q measured at 10 kHz should be 100 or more.

High-Pass Filter (f_c = 10 kHz)

FIGURE 4 10 KHz HIGH PASS FILTER
(REMOVES LOW FREQUENCY HUM AND NOISE)
(COPIED FROM NTC 7)
PLEASE NOTE:(ACTUALLY CAPACITANCES ARE IN MICRO FARADS)



Capacitances are given in pF, inductances in uH.
Capacitor and resistor tolerance ±1%.

$$\text{Insertion loss (dB)} = 10 \log_{10} \frac{[1+(f/f_1)^2][1+(f/f_2)^2]}{[1+(f/f_3)^2]},$$

where $f_1 = 0.270$ MHz, $f_2 = 1.37$ MHz, and $f_3 = 0.390$ MHz.

Random noise weighting network

FIGURE 5 VIDEO WEIGHTING FILTER
(SIMULATES RESPONSE OF VISUAL SYSTEM TO NOISE
(COPIED FROM NTC 7)

3.0 HARDWARE DESCRIPTION

This thesis describes the design of a television measurement system. The system consists of both hardware and software. It is composed of three primary parts, a Front End Processor (FEP) which converts several lines of a television vertical blanking interval into digital form, a computer to which the digital data representing the lines is downloaded, and the software residing in the computer which analyses the data to make the measurements.

The computer used in this application is an HP 71B with communication to and from the front end processor via the HPIL which is a Hewlett Packard interface standard for low power/battery equipment. This allows the development of a portable test system for testing channels carrying video signals.

The schematic for the front end processor (FEP) is shown in figure 6. The front end processor also controls the analog to digital converter, which in this case is a TDC1048E1C circuit board from TRW. The TDC1048E1C is shown on the front end processor schematic, but it is in fact a separate board.

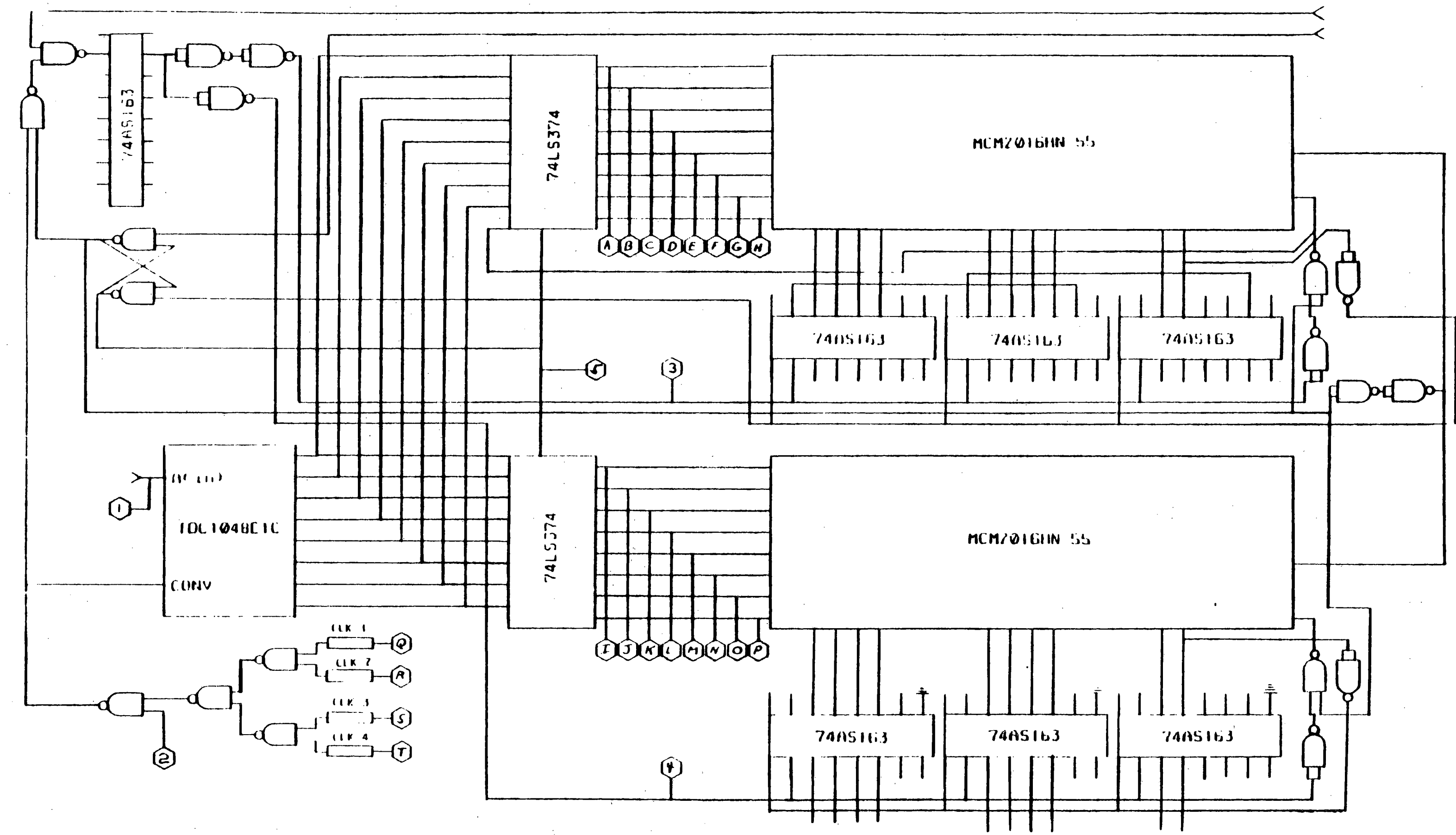


FIGURE 6(A) FRONT END PROCESSOR (FEP)
(DIGITIZES ONE OR MORE TV LINES, AND DOWNLOADS THIS DATA TO A COMPUTER)

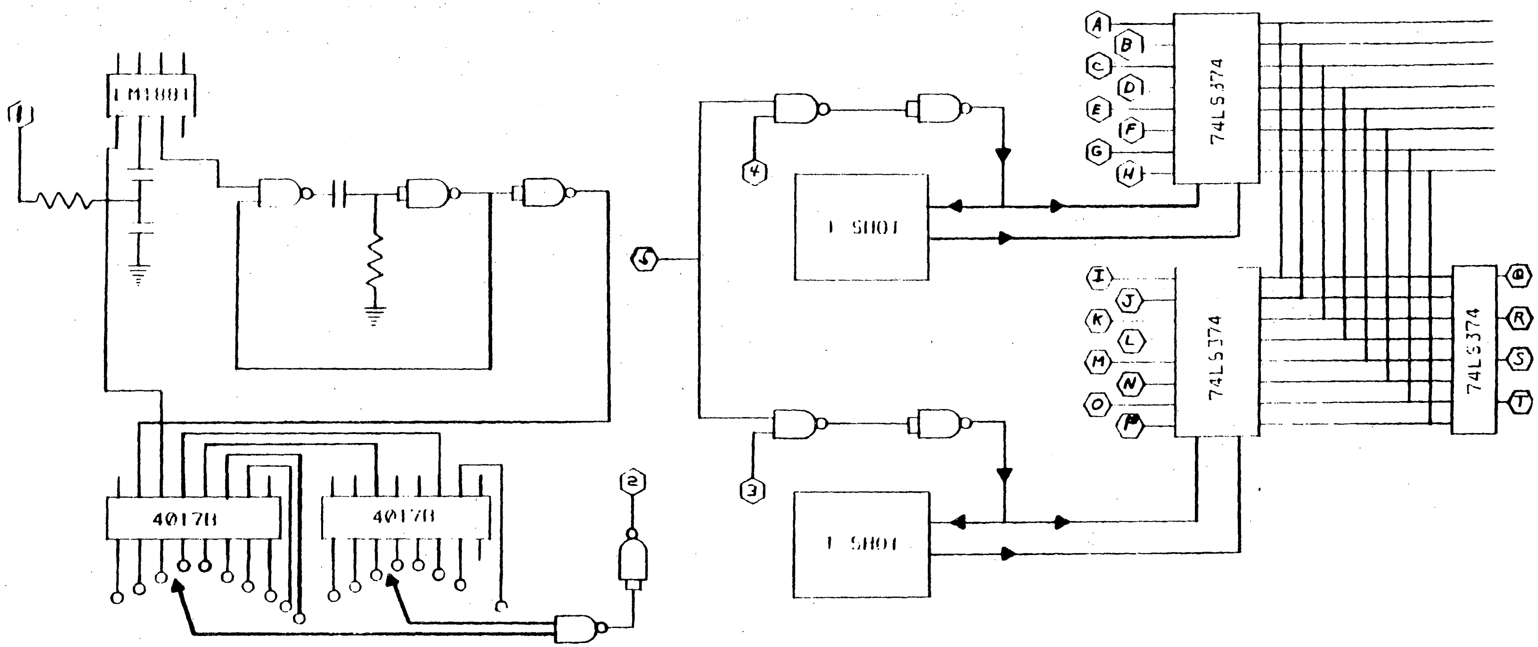


FIGURE 6(B)   FRONT END PROCESSOR (FEP)
(DIGITIZES ONE OR MORE TV LINES, AND DOWNLOADS THIS DATA TO A COMPUTER)

4.0 VIDEO SIGNAL TO NOISE RATIO

This section describes the measurement of video signal to noise ratio. Digital filters are developed in this section which match classical analog filters through which noise power is traditionally measured in a television channel.

By convention, Video Signal To Noise Ratio is defined as peak to peak video (0.714 volts) to rms weighted noise ratio, with the noise power measured through a weighting filter plus channel shaping filters.

Since these filters are used only for the measurement of noise power, only the magnitude frequency responses are important, not the phase responses.

An unused line or lines in the vertical blanking interval will be selected by the FEP, sampled at 13.5 MHz and stored in temporary memory. Samples will be taken until the buffer of the FEP is filled. This occurs after 2044 samples. One television line is 63.56 microseconds. At the 13.5 MHz sample rate, 2044 samples correspond to 151 microseconds or 2.38 lines.

After the FEP buffer is filled, this data will be downloaded to a computer. This data will then be used as input to the digital filter described in this section. The data corresponding to the synchronizing pulses will be ignored. The data samples for the last 1.38 lines will be used if the lines are inactive, otherwise they will be ignored. The data will be input to the digital filter in the same order that it was taken by the analog to digital converter.

The output samples of the digital filter will be squared and summed. The resulting sum will be averaged and the square root of the average will be taken. Thus the rms value will be calculated directly and expressed as a dB ratio referenced to 100 IRE = 0.714 volts.

After the digital filter has processed this data, the FEP will be enabled by the computer. Then the FEP will obtain another 2044 samples.

The 10 KHz high pass filter will be implemented as an analog filter. For noise measurements it will be in series with the system anti-aliasing filter.

The analog filter only has four components. It is rather difficult to implement it as an equivalent digital filter because the corner frequency is 10 KHz and the sampling frequency is 13.5 MHz. An attempt was made to implement this filter as an IIR digital filter, but when implemented on a computer it had rather high levels of internally generated noise. The rms noise level was approximately 40 dB below the peak signal level. This digital filter adds an undesirable level of degradation to the noise level of the system, which limits the use of the system for low noise test signals such as studio sources. See Appendix 1 for the design of this filter.

4.1 IMPLEMENTATION OF DIGITAL FILTERS EQUIVALENT TO ANALOG FILTERS

A digital filter equivalent to the analog weighting filter and analog low pass channel shaping filter will be implemented. The Frequency Sampling Technique (FST) as explained in [6] will be used. The resulting filter has 35 taps.

In order to proceed with design of the digital combination (weighting-low pass) filter, the magnitude frequency responses of the analog weighting filter and analog low pass channel shaping filter, from NTC 7 will

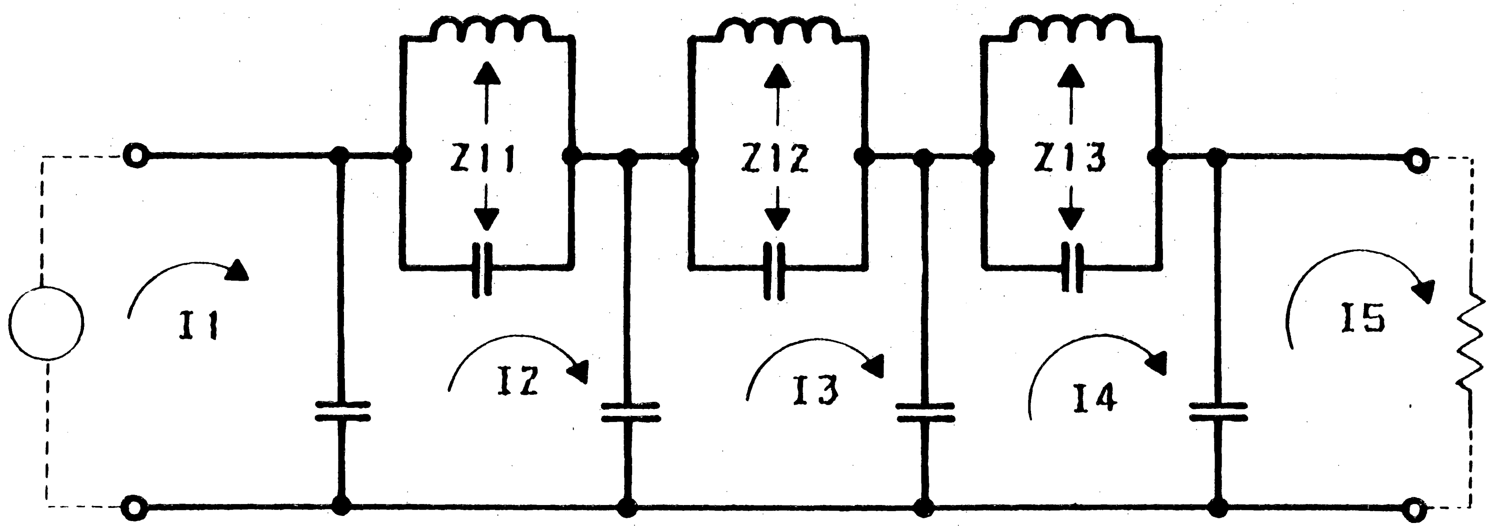


FIGURE 7 VIDEO LOW PASS CHANNEL SHAPING FILTER FROM NTC7 SHOWING COMBINED IMPEDENCES AND ARROWS FOR LOOP EQUATIONS (SEE FIGURE 3)

be determined.

The schematic for the analog 4.2 MHz low pass channel shaping filter specified in NTC 7 is shown in figure 7. The arrows indicate directions for writing loop equations. The impedences Z11, Z12 and Z13 are ideal imajinary impedences resulting from the combinations of the indicated ideal capacitors and inductors.

The loop equations can be written by inspection and are shown below.

For J1: 0 = Z1(J1-J4) + Z11(J1) + Z3(J1-J2)

0 = (Z1+Z3+Z11)(J1) - Z3(J2) - Z1(J4)

For J2: 0 = Z3(J2-J1) + Z12(J2) + Z5(J2-J3)

0 = -Z3(J1) + (Z3+Z12+Z5)J2 - Z5(J3)

For J3: 0 = Z5(J3-J2) + Z13(J3) + Z7(J3-J5)

0 = -Z5(J2) + (Z5+Z7+Z13)J3 - Z7(J5)

For J4: 1V = 75 ohm(J4) + Z1(J4-J1)

1V = -Z1(J1) + (Z1+75 ohm)J1

For J5: 0 = Z7(J5-J3) + 75 ohm(J5)

$$0 = -Z7(J3) + (Z7+75\ \text{ohm})J5$$

Using these loop equations, the matrix equation shown below is generated.

$$\begin{bmatrix} (Z1+Z11+Z3) & (-Z3) & 0 & (-Z1) & 0 \\ (-Z3) & (Z3+Z12+Z5) & (-Z5) & 0 & 0 \\ 0 & (-Z5) & (Z5+Z7+Z13) & 0 & (-Z7) \\ (-Z1) & 0 & 0 & (Z1+75) & 0 \\ 0 & 0 & (-Z7) & 0 & (Z1+75) \end{bmatrix} \begin{bmatrix} J1 \\ J2 \\ J3 \\ J4 \\ J5 \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \\ 0 \\ 1 \\ 0 \end{bmatrix}$$

The magnitude frequency response for the analog low pass channel shaping filter is determined by solving the matrix equation for J5 at a given frequency, and multiplying J5 by 75 ohms. This magnitude frequency response (Ha(s)) is shown in figure 8. As shown, this filter has some amount of ripple in the pass band. When designing the digital filter, this pass band ripple will be removed from the frequency samples.

The magnitude frequency response of the analog weighting filter is given in equation form in NTC 7 (shown here in figure 5). A plot of this equation (Hb(s)) is shown in figure 9.

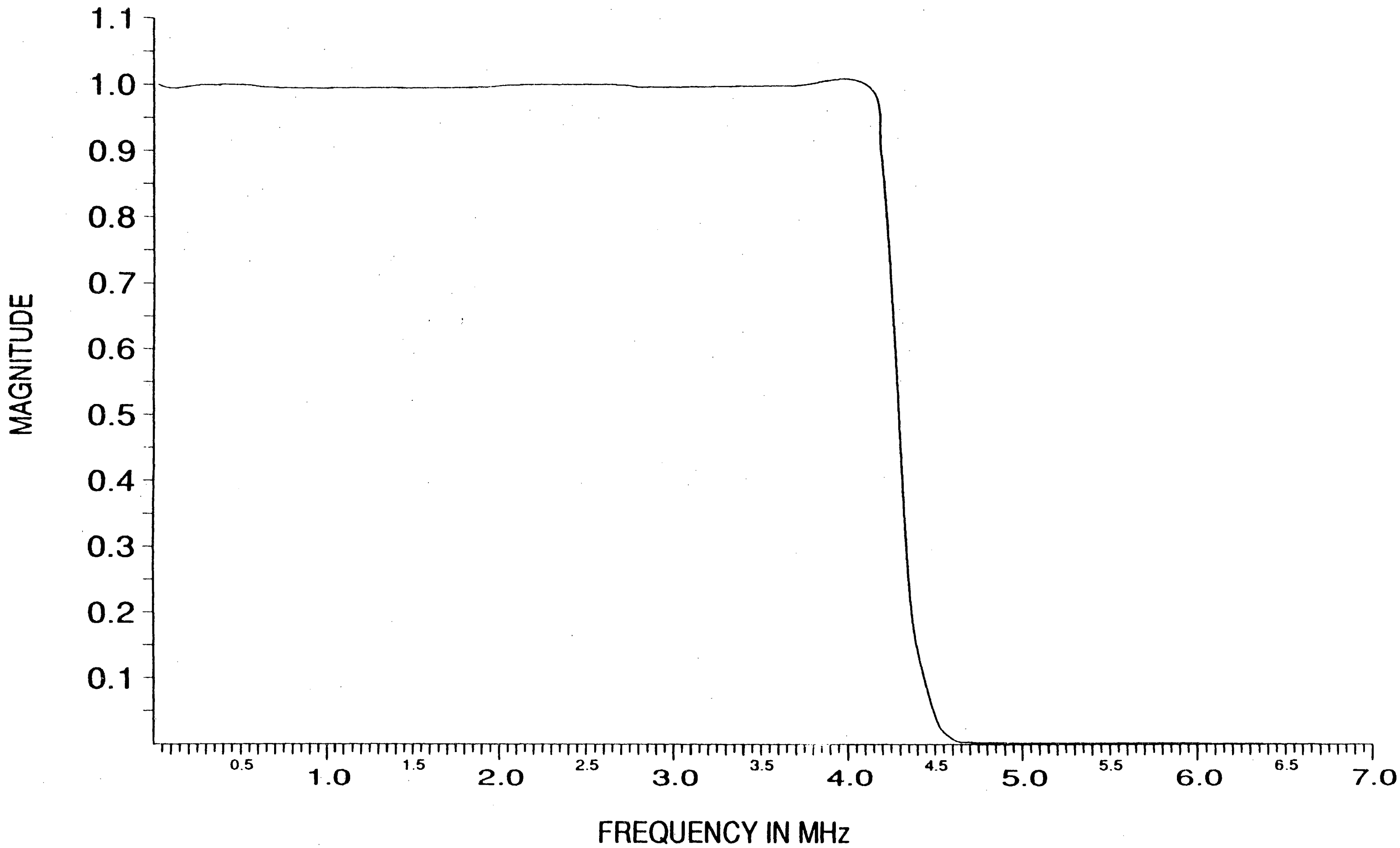


FIGURE 8 MAGNITUDE FREQUENCY RESPONSE (Ha(S))
OF ANALOG VIDEO LOW PASS CHANNEL SHAPING FILTER

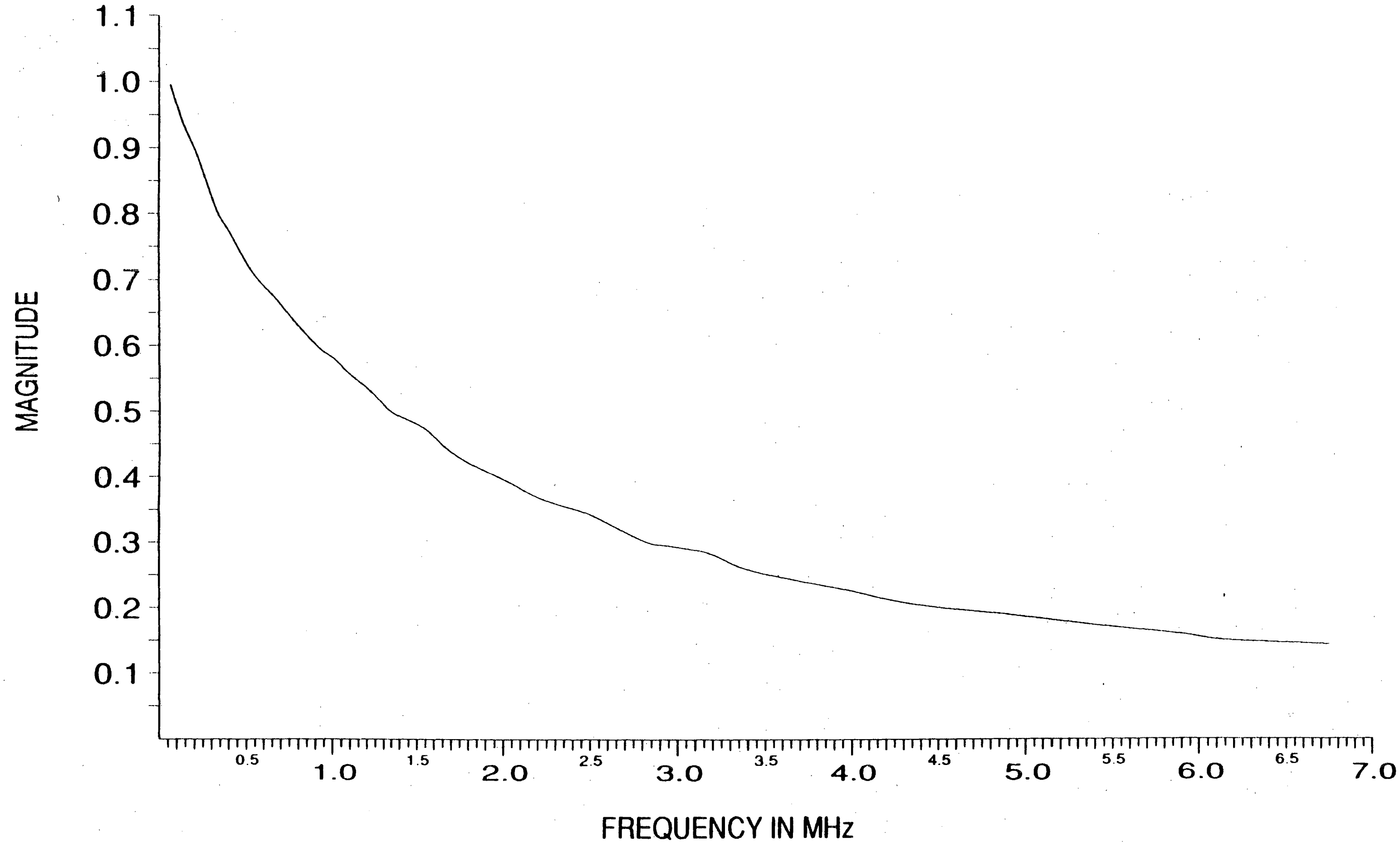


FIGURE 9 MAGNITUDE FREQUENCY RESPONSE (Hb(S))
OF ANALOG VIDEO WEIGHTING FILTER

The frequency sampling method is used to design the digital FIR filter. Frequency samples are taken at (6.75MHz/17)N for N, from 0 to 17. Each frequency sample value is determined when the magnitude frequency response of the weighting filter at the sample frequency is multiplied by the magnitude frequency response of the 4.2 MHz low pass channel shaping filter at the same frequency.

The analog magnitude frequency response (so sampled) is reflected about F(s)/2 (F(s) = the sampling frequency = 13.5 MHz). Since F(s) = 2(PI), and F(s)/2 = PI, the resulting response is symmetric about PI with period of 2(PI). Thus the sample points are symmetric about PI (point 17 is at PI) and there are 35 frequency samples.

The Inverse Discrete Fourier Transform of these frequency samples is performed to calculate the impulse response of the digital filter. In this case a 35 length impulse response is produced. The resulting filter is linear phase [6]. Also the resulting 35 impulse coefficients are symmetric about the center point, coefficient 17. This can cut the number of multiplications almost in half during implementation.

The following frequency samples were used to generate coefficients for the filter. The frequency samples A(1) - A(35) are symmetric about A(17). Thus only the first 17 samples are listed.

A(0) = 1	A(9) = 0.24858801443
A(1) = 0.77073778507	A(10) = 0.22637223263
A(2) = 0.63177462482	A(11) = 0.14714195122
A(3) = 0.53613232060	A(12) = 0
A(4) = 0.45904019717	A(13) = 0
A(5) = 0.39706758328	A(14) = 0
A(6) = 0.34749256830	A(15) = 0
A(7) = 0.30763271112	A(16) = 0
A(8) = 0.27524231802	A(17) = 0

The coefficients thus generated are listed below. Only the first 17 are listed since the coefficients are symmetric about H(17).

H(0) = 0.00309665366	H(9) = 0.00586240767
H(1) = 0.00650984924	H(10) = 0.01809817911
H(2) = 0.00251774223	H(11) = 0.00673905953
H(3) = 0.00390115726	H(12) = 0.00908855067
H(4) = 0.00804158075	H(13) = 0.03677332410
H(5) = 0.00241929534	H(14) = 0.02488299145
H(6) = 0.00478891808	H(15) = 0.03804685548
H(7) = 0.01142818342	H(16) = 0.17603778929
H(8) = 0.00327539680	H(17) = 0.27698413181

A magnitude frequency response plot for this filter

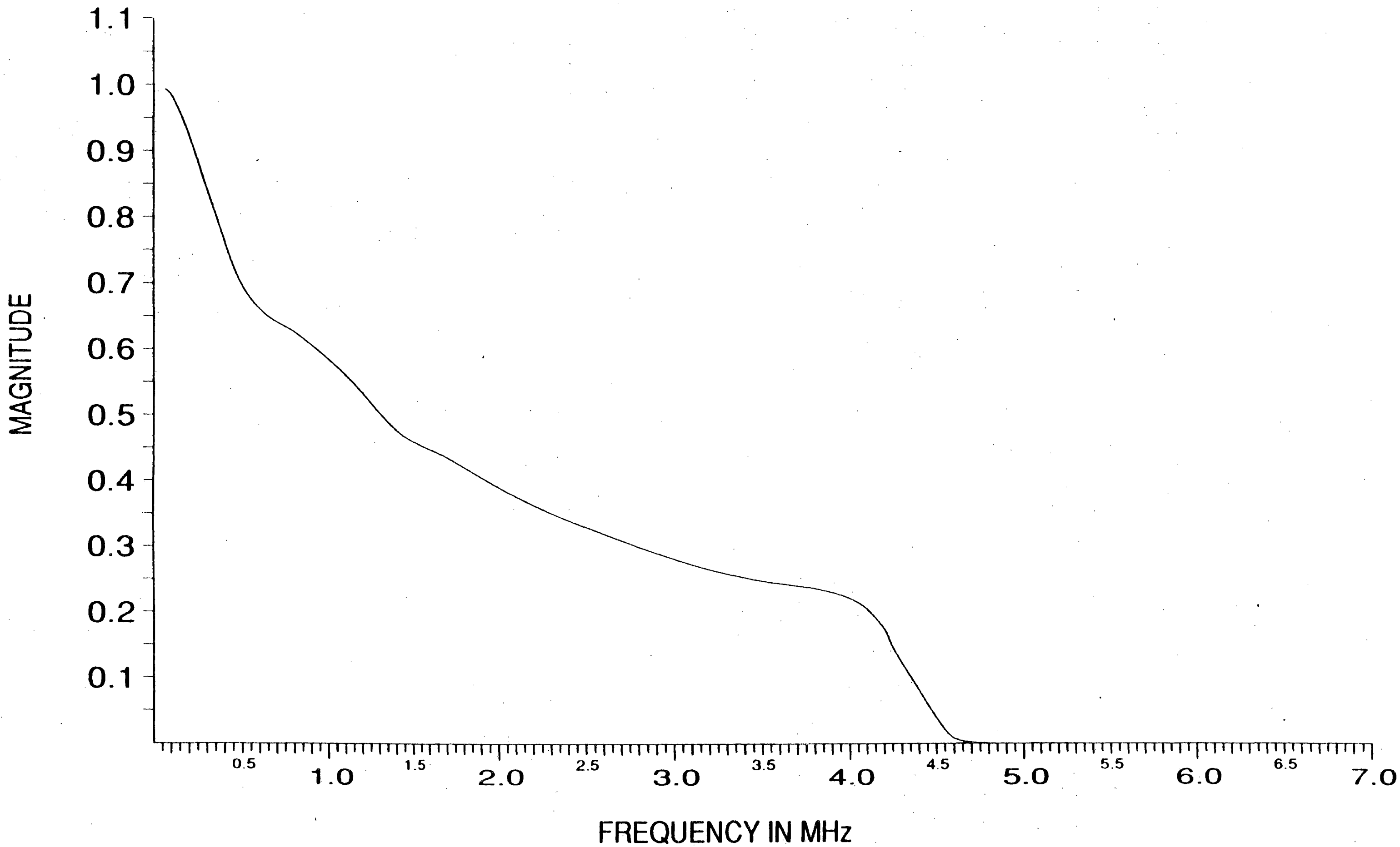


FIGURE 10 MAGNITUDE FREQUENCY RESPONSE OF DIGITAL FIR FILTER, WHICH IS DIGITAL VERSION OF ANALOG VIDEO LOW PASS FILTER AND ANALOG VIDEO WEIGHTING FILTER

is shown in figure 10. It can be seen that this digital filter closely matches the analog weighting filter (figure 9) in the pass band of the channel shaping filter (figure 8). The digital filter frequency response is slightly higher in the stop band than the analog channel shaping filter. This stop band error will be reduced by the system anti-aliasing filter.

4.2 CALCULATION OF NOISE GENERATED IN THE DIGITAL FILTER

The digital filter described above will be implemented in direct form using finite precision coefficients. These coefficients are rounded approximations to the ideal infinite precision coefficients. As a result, the implemented coefficients add noise when multiplied by an input sample. This noise can be modeled as an independent random variable with uniform distribution [6]. Thus this noise can be represented as a white noise source with variance σ^2 = (2^-2B)/3 where B is the number of bits used to represent the coefficient [6].

In [7] it is shown that the mean squared value of voltage times conductance equals power. Also if the DC level equals zero, variance and the mean squared value are equal. In this case the DC level is zero and σ is the rms noise power.

Thus σ represents the rms noise power generated at each tap from coefficient rounding. If B = 32, the rms noise power at each tap is σ = 1.34e-10 or 10log(1.34e-10) = -98.72 dB (referenced to 1, the maximum peak signal level)

The rms noise power added by all 35 taps of the filter is 10log(35X1.34e-10) = -83.27 dB.

5.0 VIDEO TEST SIGNALS

As previously described in this thesis, there are two standard test signals defined in NTC7. There is the Composite Test Signal, (figure 11) and the Combination Test Signal (figure 12). Both figures show sample numbers, with sample number 1 = t(1) at the start of the leading edge of the horizontal synchronizing pulse. Several sampling rates will be used and the sample numbers corresponding to the different sample rates are indicated. Unless otherwise noted, sample numbers will refer to a sample rate of 14.31818 MHz.

These signals shown in figures 11 and 12 may be placed on the same line of sequential fields of the vertical blanking interval or on sequential lines of the same field of the vertical blanking interval. Each test signal is used to perform certain tests on the video transmission system.

The measurements and the methods of making the measurements are described below:

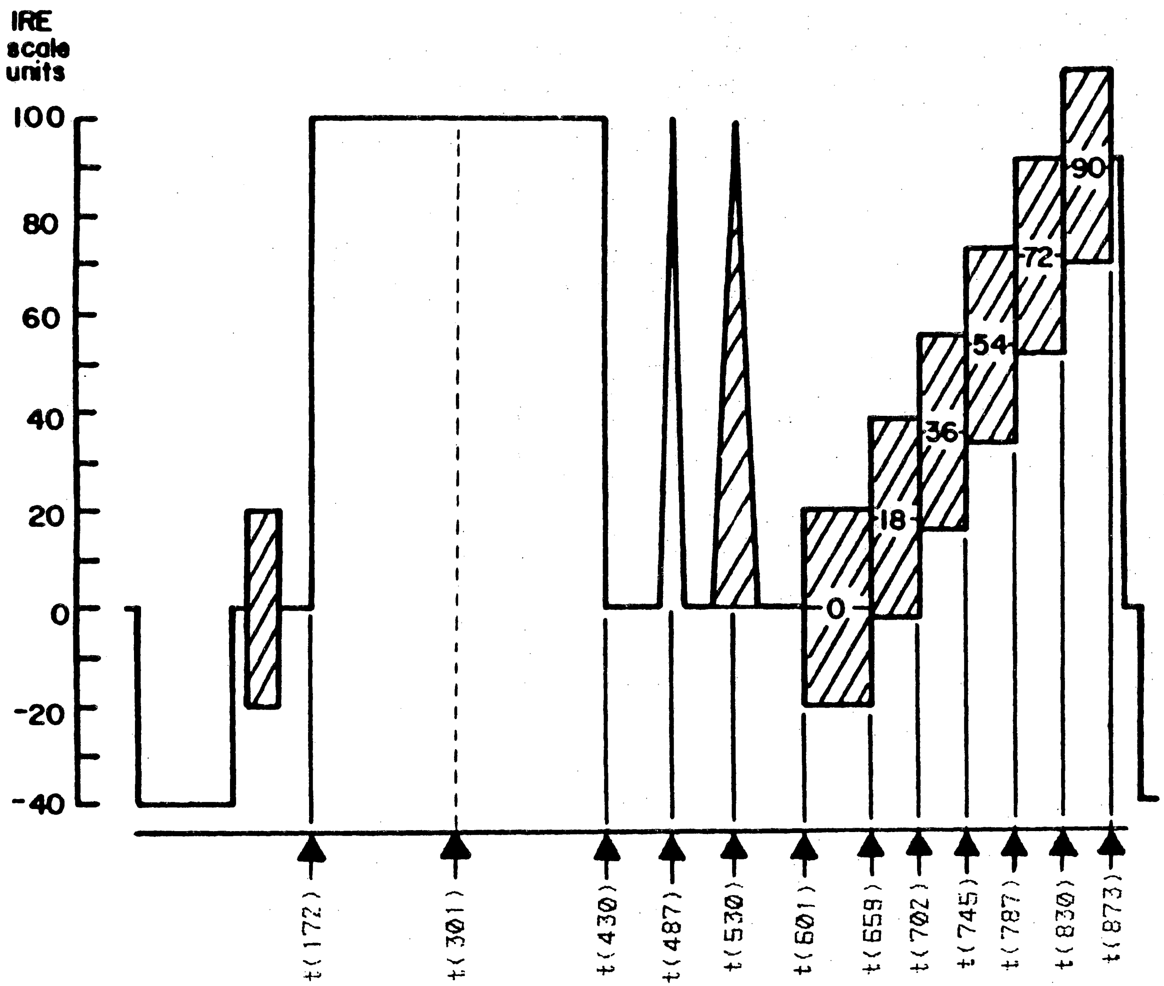


FIGURE 11 THE COMPOSITE TEST SIGNAL WITH SAMPLE POINTS (SEE FIGURE 1)

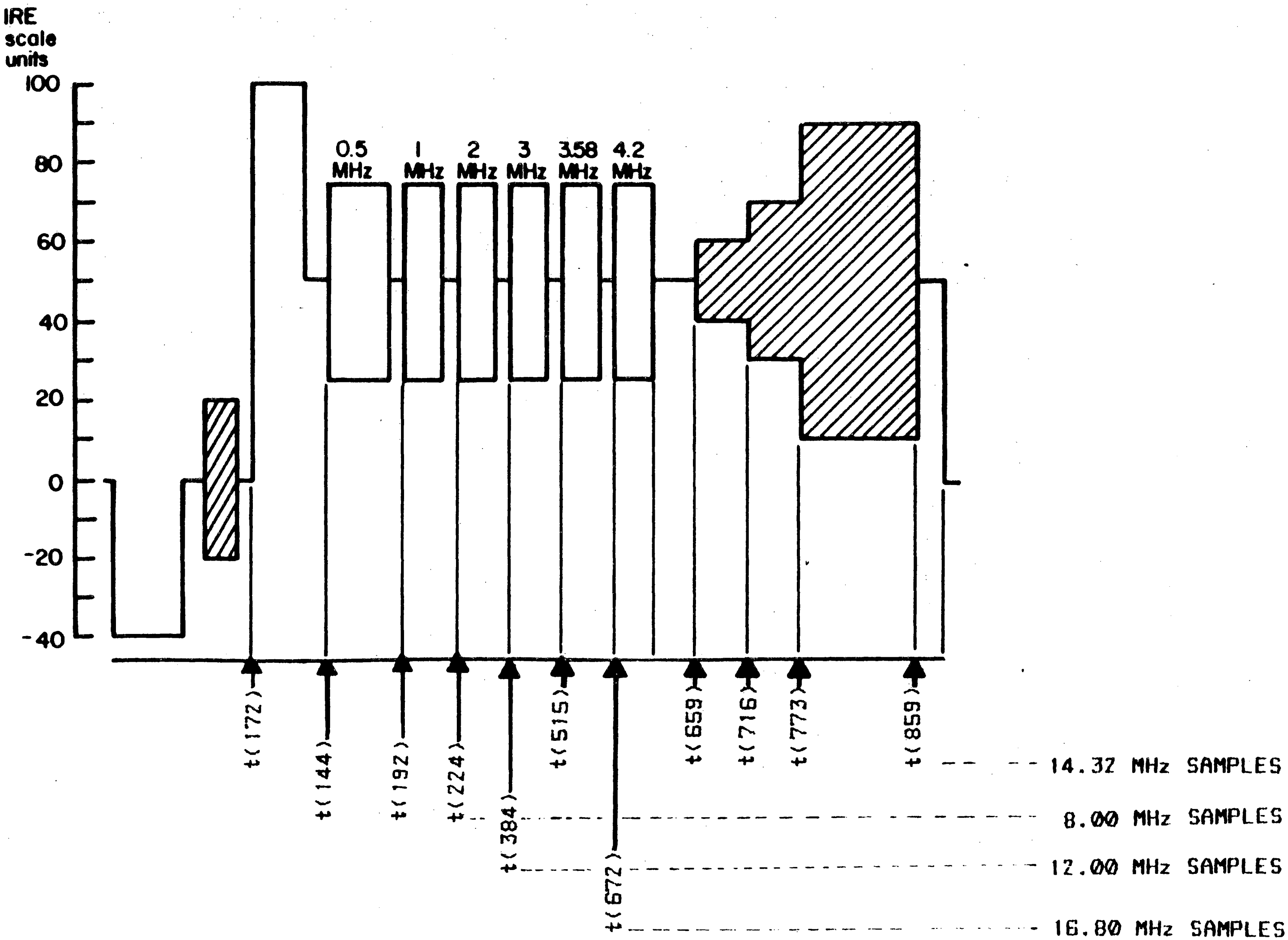


FIGURE 12 THE COMBINATION TEST SIGNAL WITH SAMPLE POINTS
(SEE FIGURE 2)

5.1 INSERTION GAIN

Insertion Gain is the amplitude of a television signal as indicated by the amplitude of the Line Bar (t(172) to t(429)) above the base line in the Composite Test Signal.

Although the sampling frequency is 14.31818 MHz, it is not critical for this measurement, but this frequency will be used for other measurements where the frequency is critical. The center of the bar corresponds to t(301). The baseline is represented by sample t(165) which is 11.5 microseconds from the leading edge of the synchronizing pulse. The value for the bar center will be determined by averaging near sample t(301), from t(251) to t(351). Thus insertion gain is:

$$t(301) = \frac{\sum_{i=252}^{351} t(i) - t(165)}{100}$$

Note that t(301) has been assigned the average value calculated above, as it will be used again.

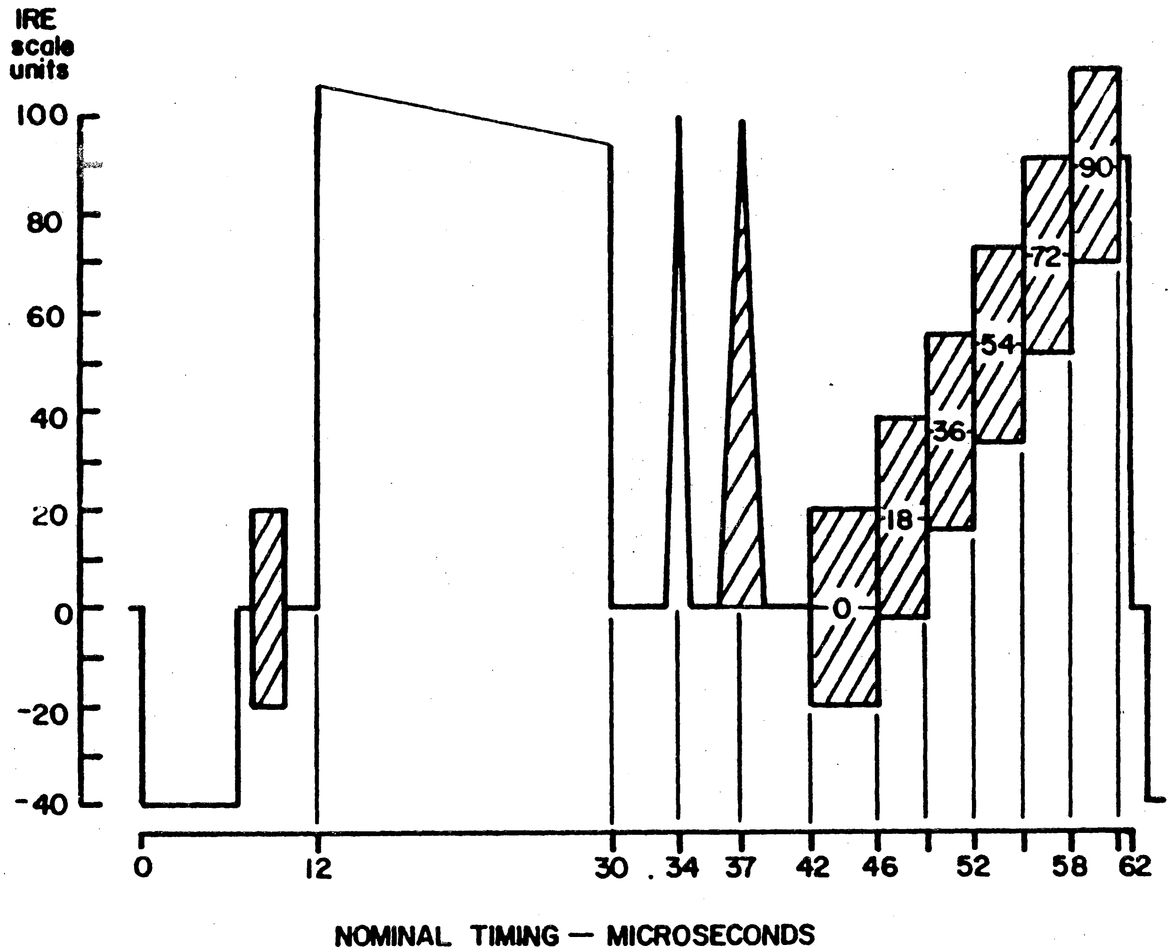


FIGURE 13 THE COMPOSITE TEST SIGNAL SHOWING LINE-TIME WAVEFORM DISTORTION (SEE NTC7)

5.2 LINE-TIME WAVEFORM DISTORTION

The anomaly of Line-Time Waveform Distortion (sometimes referred to as "Bar Tilt") is shown in figure 13. This measurement is performed by again using the Line Bar Test Signal.

The sampling rate is 14.31818 MHz. Again this frequency is not critical.

Since the first and last microseconds of the bar are ignored, the sampling points are 13 microseconds and 29 microseconds from the leading edge of sync. This corresponds respectively to t(186) and t(415). The equation to calculate Line Time Waveform Distortion is:

$$\frac{t(186) - t(415)}{t(301)} \quad X \quad 100$$

5.3 SHORT-TIME WAVEFORM DISTORTION

Short-Time Waveform Distortion is measured using the Line Bar Test Signal plus the 2T Pulse, t(483) to t(490), shown in figure 11. The height of the 2T pulse is measured referenced to the amplitude of the Line Bar center. Thus the 2T pulse height is expressed as a percentage of the Line Bar center. This is shown below in mathematical form:

$$\text{For } 483 \leq i \leq 490; \text{ THE MAXIMUM VALUE OF: } \frac{t(i)}{t(301)}$$

There is a second part of Short-Time Waveform Distortion. This is any ringing or overshoot which exists at the leading edge of the bar and after the trailing edge of the bar. These leading and trailing edge measurements are added to produce a peak-to-peak measurement. These measurements are made within the first 1 microsecond of the bar, t(172) to t(186), and within the first 1 microsecond after the bar, t(430) to t(444).

Short-Time Waveform Distortion is the maximum difference between t(172) to t(186) and the level of the bar center, plus the maximum difference between the baseline, t(165), and t(430) to t(444). These two values are shown below in mathematical form:

$$\text{For } 172 \leq i \leq 186; \text{ The Maximum Value Of } t(i) - t(301)$$

For $430 \leq i \leq 444$; The Maximum Value Of t(165) - t(i)

5.4 CHROMINANCE-LUMINANCE GAIN INEQUALITY

Chrominance-Luminance Gain Inequality (CLGI) is measured using the 12.5 T Pulse, t(519) to t(542), figure 11. This pulse is composed of two components, a chrominance portion and a lower frequency luminance portion which amplitude modulates the chrominance portion. [2]

CLGI provides a relative measure of high frequency (Chrominance) and low frequency (Line Bar) magnitude responses. High frequency magnitude response is obtained by measuring the maximum amplitude of the chrominance portion of the 12.5 T Pulse relative to the amplitude of the Line Bar, t(301). The Line Bar amplitude indicates low frequency magnitude response. (The Line Bar represents lower frequency than the luminance portion of the 12.5 T Pulse.) Thus CLGI is the maximum magnitude of the 12.5 T Pulse above the baseline referenced to the magnitude of the Line Bar center t(301) above the baseline.

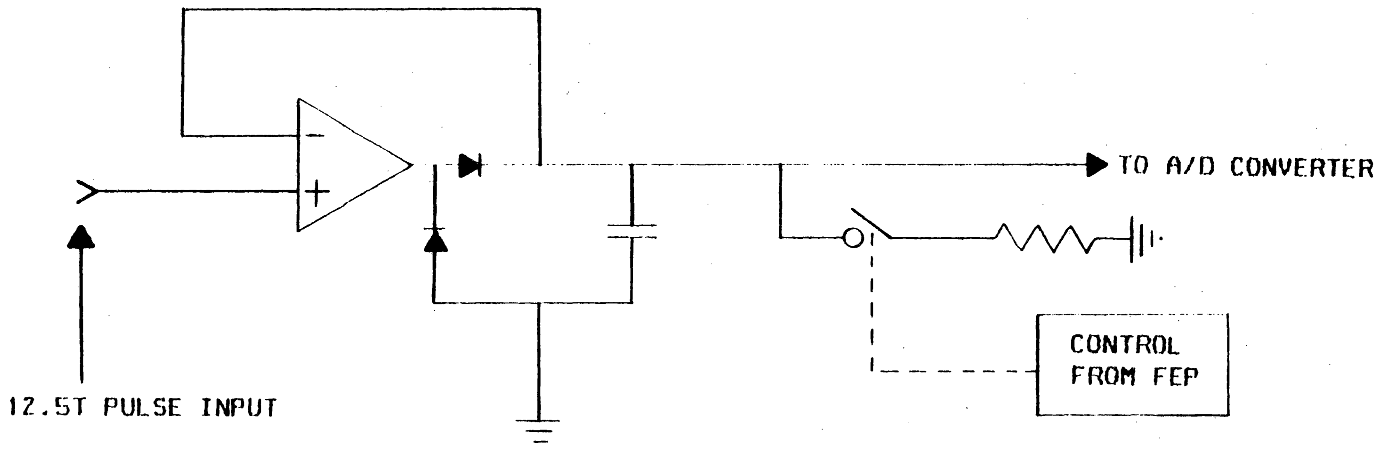


FIGURE 14 POSITIVE PEAK DETECTOR
(MEASURES PEAK AMPLITUDE OF 12.5T PULSE)

This measurement will be made by sampling the output of a positive peak detecter from t(519) to t(542). The schematic to perform this function is shown in figure 14.

The input signal e(in) feeds the peak detector [8] which stores the amplitude of each increasing chrominance pulse. After the peak of the 12.5 T Pulse occurs the voltage stored on the capacitor does not increase.

Thus the voltage stored on the capacitor (less any leakage voltage) will represent the maximum level of the 12.5 T Pulse. In this way, the peak which is the maximum level of the 12.5 T Pulse is stored.

The computer will take the maximum level of t(519) to t(542) as the amplitude of the 12.5 T Pulse.

Thus CLGI is the maximum of t(i) for $519 \leq i \leq 542$

With t(301) normalized to 100 IRE. (The capacitor will be chosen so that it's discharge rate will not affect the measurement significantly.)

5.5 CHROMINANCE-LUMINANCE DELAY INEQUALITY

Chrominance-Luminance Delay Inequality (CLDI) is also measured using the 12.5 T Pulse.

CLDI is a measure of the difference in transit times of the two components of the 12.5 T Pulse as they pass through a television system.

The basis for this measurement is the algorithm described in NTC7 for calculation of CLDI. Differences in propogation delay of the two components of the 12.5 T Pulse cause the 12.5T Pulse to be displaced above and below the baseline as indicated by the Y1 and Y2 values shown in figure 15. This figure is copied from NTC7.

The values for Y1 and Y2 are determined by the negative peak detector and associated circuitry shown in figure 16.

The clock associated with this figure is 14.31818 MHz and is synchronous with and four times the chrominance pulse .

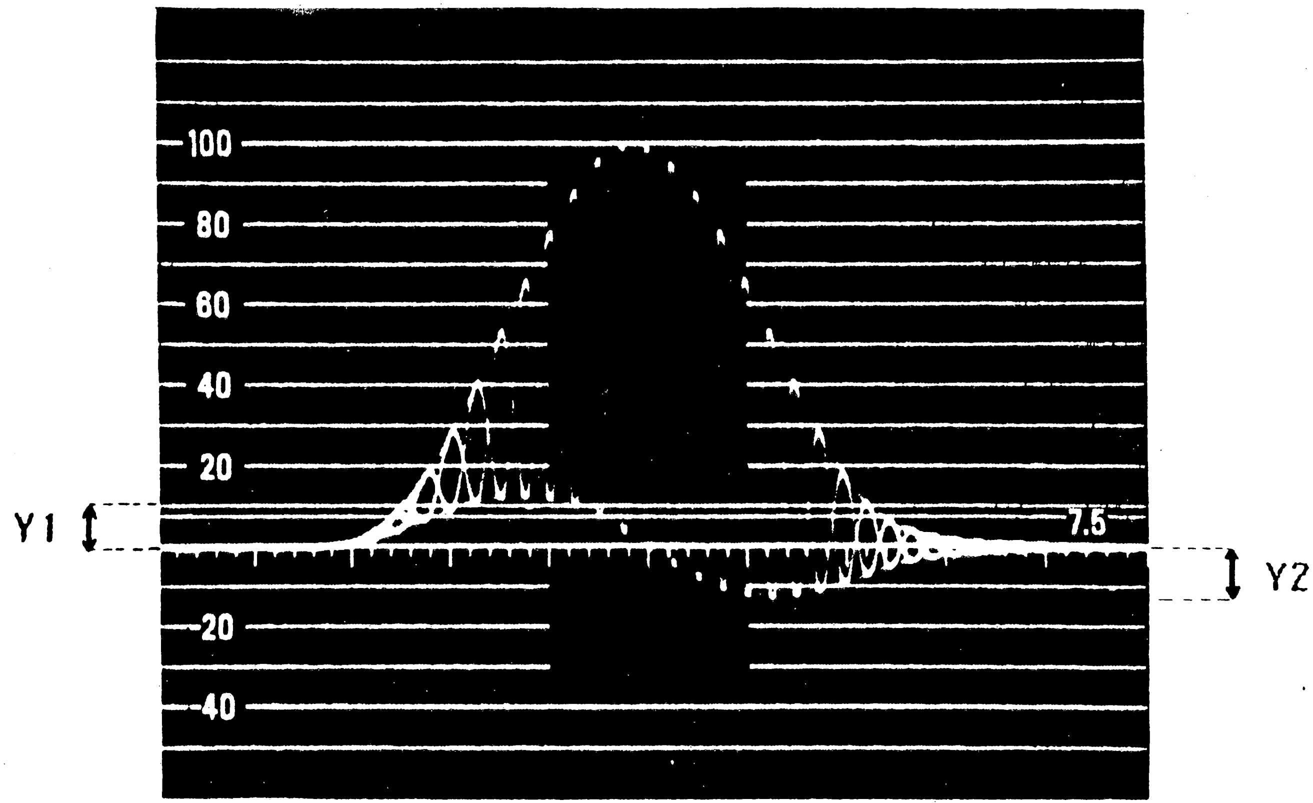


FIGURE 15 EXPANDED VIEW OF 12.5T PULSE SHOWING Y1 AND Y2 VALUES

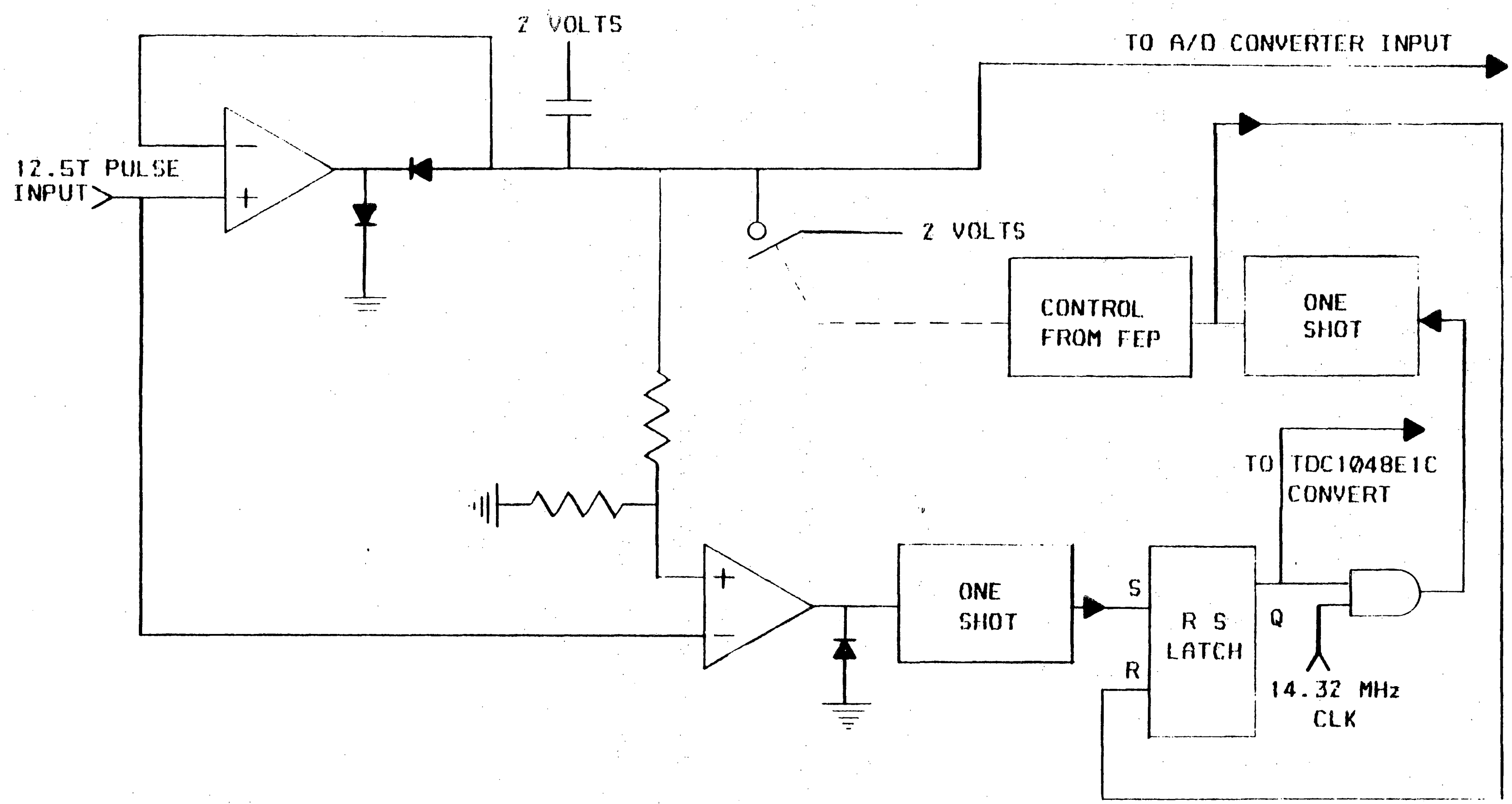


FIGURE 16 NEGATIVE PEAK DETECTOR
(MEASURES NEGATIVE PEAKS OF 12.5T PULSE)

From t(519) to t(542), the negative peak detector charges the capacitor to the negative value of each chrominance pulse. The analog to digital converter samples this voltage on the following clock pulse rising edge, and the capacitor is discharged following the propogation delay through the and gate, the one shot and the control logic for the analog switch. In this way the negative peak values are obtained for each chrominance pulse.

The value for Y1 is the maximum value above the baseline for a negative peak, and the value for Y2 is the maximum value below the baseline for a negative peak. In mathematical form it is:

Y1 = The Maximum Negative Peak Above The Baseline

Y2 = The Maximum Negative Peak Below The Baseline

CLDI is determined from the following relationship from NTC7:

$$CLDI = 20 \quad (Y1)(Y2)$$

where the result is in nanoseconds.

5.6 GAIN/FREQUENCY DISTORTION

Measurement of Gain/Frequency Distortion is done using the frequency bursts at 0.5 MHz, 1.0 MHz, 2.0 MHz, 3.0 MHz, 3.579545 MHz and 4.2 MHz shown in figure 12. The amplitude of each of the bursts is measured with the reference at the start of the active line (t(172 to t(229)) normalized to 100 IRE. This measurement amounts to a frequency sweep at discrete frequencies.

Because of the number of different frequencies involved, a number of different sampling frequencies will be used. The FEP described in chapter 3 has on board oscillators at a number of different frequencies. These oscillators control the sampling rate under computer control. In operation, the computer will obtain a line of samples (from the FEP) at each sampling rate and ignore all samples other than those which pertain to the burst which is to be measured.

The sampling frequencies for the different bursts are:

Burst Frequency	Sampling Frequency
4.2 MHz	16.8 MHz

3.579545 MHz	14.31818 MHz
3.0 MHz	12.0 MHz
2.0 MHz	8.0 MHz
1.0 MHz	8.0 MHz
0.5 MHz	8.0 MHz

The same sampling frequency is used for the bursts at 0.5 Mhz, 1.0 MHz and 2.0 MHz. In the case of the 1.0 MHz burst every second sample will be used and in the 0.5 MHz case every fourth sample will be used.

This measurement will be made using the Discrete Fourier Transform (DFT) as described in Appendix 2. There a method is described to calculate the amplitude and phase of a signal at a known single frequency. The method consists of taking synchronous samples at four times the frequency, and implementing the algorithm as shown. The algorithm is explained in detail in Appendix 2.

The algorithm makes use of four components, two real and two imaginary. The imaginary components are indicated by j. After taking four synchronous samples per period the algorithm is:

P = t(1) - jt(2) - t(3) + jt(4);

where P is in rectangular coordinates.

P is expressed as a magnitude and phase when it is converted to polar form. In this case only the resultant amplitudes are used, the resultant phases are disregarded.

Seven successive samples will be taken of the 0.5 MHz through the 4.2 MHz bursts. The algorithm will be implemented on these seven samples (four successive samples at a time) and the results averaged.

The reason for using seven samples is explained. It is decided to take the same number of samples from each burst. The 2.0 MHz burst consists of only two cycles. Thus a maximum of nine samples could be taken if the first and last samples coincide with the start and stop of the burst. Choosing seven samples allows some sampling phase error.

Implementation of the algorithm on samples of the 0.5 MHz burst results in the following equations:

$$P1 = \left| t(145) - jt(146) - t(147) + jt(148) \right|$$

$$P2 = \left| t(146) - jt(147) - t(148) + jt(149) \right|$$

$$P3 = \left| t(147) - jt(148) - t(149) + jt(150) \right|$$

$$P4 = \left| t(148) - jt(149 \quad - t(150) + jt(151) \right|$$

The resultant magnitude of the 0.5 MHz burst is:

$$\frac{P1 + P2 + P3 + P4}{4}$$

In a similar manner the algorithm is implemented for the other bursts, where four synchronous samples are taken during each period.

To investigate what could happen with the algorithm if the sampling rate were not synchronous, the following analysis is performed: As a starting point sample a cosine at 90 degree increments. If the first sample is at 90 degrees and the last sample is at 360 degrees, the algorithm produces a magnitude of 2 at 90 degrees. Assume 1 percent jitter in a positive direction between samples. (This is probably much worse than would be expected in a well designed system.) Thus the first sample would occur at 90 degrees, the second sample would occur at 180 + 3.6 = 183.6 degrees, the third at 277.2 degrees and the fourth at 370.8 degrees. This produces a magnitude of 1.98 at 93.62 degrees ($1.98\angle\underline{93.62}$), thus producing a magnitude error of 1 percent.

Alternately, run the same test with a sine wave. The sample points will be the same. The algorithm produces $2\ \angle\underline{0}$ under conditions of no sampling rate error. With the same sample rate errors used for the cosine the algorithm produces $2.01\ \angle\underline{7.16}$. This is equivalent to a magnitude error of 0.5 percent.

5.7 LUMINANCE NON-LINEAR DISTORTION

Measurement of Luminance-Non Linear Distortion is made by using the modulated 5-Riser Staircase Test Signal t(601) to t(873) of figure 11. This test signal is composed of a stairstep signal on which is superimposed a chrominance signal. Luminance Non-Linear Distortion is a measure of the interaction between the two signals and is obtained by measuring the relative heights of the steps.

The rise time of the steps is nominally 250 nanoseconds which corresponds to a period of 500 nanoseconds or a frequency of 2.0 MHz.

The height of each step will be measured by noting the fact that the D C level of each step may be calculated by synchronously sampling the chrominance signal four times per period for several periods, summing the sample values and dividing by the number of samples. Four samples per period are taken over four periods. The resulting sixteen samples will be added and the sum will be divided by sixteen. The resultant average will be the D C level of the step.

The step size is the D C level of a step minus the D C level of the next lower step. The first step size is the D C level of the first step minus the baseline value, t(165).

Luminance Non-Linear Distortion is the ratio of the largest step size to the smallest step size.

$$\text{Luminance Non-Linear Distortion} = \frac{\text{Largest Step X 100}}{\text{Smallest Step}}$$

Using the leading edge of the synchronizing pulse as the reference point, the samples and the algorithms to calculate the D C level of each step are shown below:

$$\text{D C Level Of Step 1} = \frac{\sum_{x=622}^{637} t(x)}{16}$$

$$\text{D C Level Of Step 2} = \frac{\sum_{x=673}^{688} t(x)}{16}$$

$$\text{D C Level Of Step 3} = \frac{\sum_{x=715}^{730} t(x)}{16}$$

$$\text{D C Level Of Step 4} = \frac{\sum_{x=758}^{733} t(x)}{16}$$

$$\text{D C Level Of Step 5} = \frac{\sum_{x=801}^{816} t(x)}{16}$$

$$\text{D C Level Of Step 6} = \frac{\sum_{x=844}^{859} t(x)}{16}$$

5.8 CHROMINANCE NON-LINEAR GAIN DISTORTION

The Chrominance Non-Linear Gain Distortion measurement (CNGD) is made using the 3-Level Chrominance Test Signal, t(659) to t(859)of figure 12. This signal has chrominance at three different peak-to-peak levels.

CNGD is defined as the ratios of the amplitudes of the lowest level, t(659) to t(716), and highest level, t(773) to t(859) to the amplitude of the middle level t(716) to t(773). The amplitude of the middle level is normalized to 40 IRE.

The samples will be synchronous at 14.31818 MHz.

The measurement is computed by using fifteen synchronous samples from each burst level. The four point DFT algorithm described in Appendix 2 is used on these samples (four sequential points at a time, in increments of one sample). Thus the algorithm will be run twelve times for each burst level and calculate twelve values for the amplitude and phase. Only the magnitude, or amplitude values will be used. The twelve amplitude values will be averaged.

The peak-to-peak amplitude of the low level signal is A1.

The peak-to-peak amplitude of the middle level signal is A2.

The peak-to-peak amplitude of the high level signal is A3.

Thus Chrominance Non-Linear Gain Distortion is :

$$\frac{A1}{A2} \quad \text{and} \quad \frac{A3}{A2}$$

The following samples are used to calculate the peak-to-peak levels:

For the low level burst: t(682) To t(696)

For the middle burst: t(739) To t(753)

For the high level burst: t(811) To t(825)

For the low level signal:

$$\frac{\sum_{x=682}^{693} t(x) - jt(x+1) - t(x+2) + jt(x+3)}{12}$$

For the middle level signal:

$$\frac{\sum_{x=739}^{750} t(x) - jt(x+1) - t(x+2) + jt(x+3)}{12}$$

For the high level signal:

$$\frac{\sum_{x=811}^{822} t(x) - jt(x+1) - t(x+2) + jt(x+3)}{12}$$

5.9 CHROMINANCE NON-LINEAR PHASE DISTORTION

Chrominance Non-Linear Phase Distortion also makes use of the 3-Level Chrominance Test Signal. See figure 12. The phase of each of the three signal levels is measured and the peak-to-peak phase variation of the three signal levels is calculated.

The measurement will be computed by using sixteen samples from each of the three signal levels to calculate the phase of each burst using the four point DFT algorithm described in Appendix 2.

The sixteen samples will be grouped into four sequential groups of four sequential samples. The four point DFT algorithm will be applied to each group of four samples to compute the magnitude and phase. Only the phase will be used. The four calculated phases from each signal level will be averaged to calculate the phase for each signal level. The samples at 14.31818 MHz will be synchronous at four times the chrominance frequency.

Chrominance Non-Linear Phase Distortion is the maximum peak-to-peak phase variation of the three signal levels.

Algorithms to calculate the phase of each signal level are shown below.

For the low level signal, A1:

$$\angle 1 = \angle(t(679) - jt(680) - t(681) + jT(682))$$

$$\angle 2 = \angle(t(683) - jt(684) - t(685) + jt(686))$$

$$\angle 3 = \angle(t(687) - jt(688) - t(689) + jt(690))$$

$$\angle 4 = \angle(t(691) - jt(692) - t(693) + jt(694))$$

$$\angle A1 = \frac{\angle 1 + \angle 2 + \angle 3 + \angle 4}{4}$$

For the middle level signal A2,:

$$\angle 5 = \angle(t(736) - jt(737) - t(738) + jt(739))$$

$$\angle 6 = \angle(t(740) - jt(741) - t(742) + jt(743))$$

$$\angle 7 = \angle(t(744) - jt(745) - t(746) + jt(747))$$

$$\angle 8 = \angle(t(748) - jt(749) - t(750) + jt(751))$$

$$\angle A2 = \frac{\angle 5 + \angle 6 + \angle 7 + \angle 8}{4}$$

For the high level signal, A3:

$$\angle 9 = \angle(t(808) - jt(809) - t(810) + jt(811))$$

$$\angle 10 = \angle(t(812) - jt(813) - t(814) + jt(815))$$

$$\angle 11 = \angle(t(816) - jt(817) - t(818) + jt(819))$$

$$\angle 12 = \angle(t(820) - jt(821) - t(822) + jt(823))$$

$$\angle A3 = \frac{\angle 9 + \angle 10 + \angle 11 + \angle 12}{4}$$

5.10 DIFFERENTIAL GAIN

Measurement of Differential Gain also uses the Modulated 5-Riser Staircase Test Signal. Differential Gain is the difference between the largest peak-to-peak chrominance amplitude and the smallest peak-to-peak chrominance amplitude as the D C level or step is changed.

The DFT algorithm described earlier will be used with sixteen samples during each stairstep level to compute the amplitude of the chrominance signal at each staircase level. The largest and smallest peak-to-peak amplitudes will thus be determined.

Differential Gain is thus:

Max Peak-To-Peak Chrom - Min Peak-To-Peak Chrom

The algorithms are shown below for calculation of the peak-to-peak amplitude of the chrominance at each staircase level:

Peak-To-Peak Chrominance At Staircase Level 1 =

$$\frac{\sum_{x=624}^{635} t(x) - jt(x+1) - t(x+2) + jt(x+3)}{12}$$

Peak-To-Peak Chrominance At Staircase Level 2 =

$$\frac{\sum_{x=675}^{686} t(x) - jt(x+1) - t(x+2) + jt(x+3)}{12}$$

Peak-To-Peak Chrominance At Staircase Level 3 =

$$\frac{\sum_{x=717}^{728} t(x) - jt(x+1) - t(x+2) + jt(x+3)}{12}$$

Peak-To-Peak Chrominance At Staircase Level 4 =

$$\frac{\sum_{x=760}^{771} t(x) - jt(x+1) - t(x+2) + jt(x+3)}{12}$$

Peak-To-Peak Chrominance At Staircase Level 5 =

$$\frac{\sum_{x=803}^{814} t(x) - jt(x+1) - t(x+2) + jt(x+3)}{12}$$

Peak-To-Peak Chrominance At Staircase Level 6 =

$$\frac{\sum_{x=846}^{857} t(x) - jt(x+1) - t(x+2) + jt(x+3)}{12}$$

5.11 DIFFERENTIAL PHASE

Measurement of Differential Phase uses the

Modulated 5-Riser Staircase Test Signal. Differential phase is the peak-to-peak phase variation of the chrominance signal as the staircase level changes.

The DFT algorithm described earlier will be used with sixteen samples during each stairstep level to compute the phase of the chrominance signal at that staircase level.

Differential Phase is thus:

Maximum Positive Phase + Maximum Negative Phase

The algorithms are shown below for calculation of chrominance phase at each staircase level:

Chrominance Phase At Staircase Level 1 =

$$\angle 1 = \angle(t(622) - jt(623) - t(624) + jt(625))$$

$$\angle 2 = \angle(t(626) - jt(627) - t(628) + jt(629))$$

$$\angle 3 = \angle(t(630) - jt(631) - t(632) + jt(633))$$

$$\angle 4 = \angle(t(634) - jt(635) - t(636) + jt(637))$$

$$\angle S1 = \frac{\angle 1 + \angle 2 + \angle 3 + \angle 4}{4}$$

Chrominance Phase At Staircase Level 2 =

$\angle 5 = \angle(t(673) - jt(674) - t(675) + jt(676))$

$\angle 6 = \angle(t(677) - jt(678) - t(679) + jt(680))$

$\angle 7 = \angle(t(681) - jt(682) - t(683) + jt(684))$

$\angle 8 = \angle(t(685) - jt(686) - t(687) + jt(688))$

$$\angle S2 = \frac{\angle 5 + \angle 6 + \angle 7 + \angle 8}{4}$$

Chrominance Phase At Staircase Level 3

$\angle 9 = \angle(t(715) - jt(716) - t(717) + jt(718))$

$\angle 10 = \angle(t(719) - jt(720) - t(721) + jt(722))$

$\angle 11 = \angle(t(723) - jt(724) - t(725) + jt(726))$

$\angle 12 = \angle(t(727) - jt(728) - t(729) + jt(730))$

$$\angle S3 = \frac{\angle 9 + \angle 10 + \angle 11 + \angle 12}{4}$$

Chrominance Phase At Staircase Level 4

$\angle 13 = \angle(t(758) - jt(759) - t(760) + jt(761))$

$\angle 14 = \angle(t(762) - jt(763) - t(764) + jt(765))$

$\angle 15 = \angle(t(766) - jt(767) - t(768) + jt(769))$

$\angle 16 = \angle(t(770) - jt(771) - t(772) + jt(773))$

$$\angle S4 = \frac{\angle 13 + \angle 14 + \angle 15 + \angle 16}{4}$$

Chrominance Phase At Staircase Level 5

$$\angle 17 = \angle(t(801) - jt(802) - t(803) + jt(804))$$

$$\angle 18 = \angle(t(805) - jt(806) - t(807) + jt(808))$$

$$\angle 19 = \angle(t(809) - jt(810) - t(811) + jt(812))$$

$$\angle 20 = \angle(t(813) - jt(814) - t(815) + jt(816))$$

$$\angle S5 = \frac{\angle 17 + \angle 18 + \angle 19 + \angle 20}{4}$$

Chrominance Phase At Staircase Level 6 =

$$\angle 21 = \angle(t(844) - jt(845) - t(846) + jt(847))$$

$$\angle 22 = \angle(t(848) - jt(849) - t(850) + jt(851))$$

$$\angle 23 = \angle(t(852) - jt(853) - t(854) + jt(855))$$

$$\angle 24 = \angle(t(856) - jt(857) - t(858) + jt(859))$$

$$\angle S6 = \frac{\angle 21 + \angle 22 + \angle 23 + \angle 24}{4}$$

5.12 CHROMINANCE-TO-LUMINANCE INTERMODULATION

Measurement of Chrominance-To-Luminance Intermodulation (CLI) is performed using the 3-Level Chrominance Signal, t(659) to t(859), figure 12. The 3-Level Chrominance Signal sits on a lower frequency luminance signal of 50 IRE nominal amplitude. CLI is defined as the variation (intermodulation) of the luminance signal by the chrominance signal, in this case the 3-Level Chrominance Signal. It appears as a deviation from flatness of the luminance signal in the presence of the chrominance signal.

The measurement result is the deviation from the reference level of this low frequency signal expressed in IRE.

The reference level (where there is no chrominance signal) extends from t(616) to t(659). The reference level is determined from the relationship:

$$\frac{\sum_{x=634}^{643} t(x)}{10}$$

The 3-Level Chrominance Signal extends from t(659) to t(859). Since CLI exists along the luminance signal where the 3-Level Chrominance Signal also exists, it is best to calculate the D C level of the chrominance signal for each period and not average over several periods

The average or D C level of the chrominance portion of the 3-Level Chrominance Signal is determined for each chrominance period by synchronously sampling the chrominance signal four times per period and summing the four samples as shown below:

$$P(x) = \sum_{x=616}^{656} t(x) + t(x+1) + t(x+2) + t(x+3)$$

Chrominance-To-Luminance Intermodulation is:

For $616 \leq x \leq 656$

The maximum value of $\left|\text{Reference Level} - P(x)\right|$

CONCLUSION

The idea of this thesis is to design a system which can test a video channel using digital techniques. The test system is based on the NTC7 standard test procedures to measure industry accepted parameters. The measurements defined in NTC7 are analog in nature. The measurements implemented in the test system described in this thesis are performed in the sampled domain using certain techniques from Digital Signal Processing, such as the FFT and Digital Filtering.

The use of DSP techniques has certain advantages over analog methods. The principal advantage is the minimization of linear hardware. A classical analog implementation of this system would likely require several circuit boards with analog circuitry requiring close tolerences on resistors and capacitors. To manufacture such a device would require stocking and possibly testing of many different components.

DSP techniques require linear electronics only on the input up to and including the analog-to-digital converter. After this converter, processing is done with digital circuits. Digital circuits have the advantage of only having two stable states. This greatly simplifies circuit design and board layout when compared with analog circuits.

Initially the intended implementation was to be with the TMS32010 series of microprocessors by Texas Instruments. However as the design proceeded it became apparent that certain advantages could be obtained by implementing the system as a Front End Processor (figures 6(a) and 6(b)) which will digitize one or more lines of a tv signal and download the data to a computer.

There are several reasons for this decision, both technical and marketing. It is simpler to implement the software on a standard computer than the TMS32010. Also there are savings to be realized in the development of displays and chassis (although the Front End Processor does involve hardware development). Also the use of a computer allows the use of off the shelf graphics software for display purposes which would be difficult and expensive to develop for only one product.

The next step is to finish building the Front End Processor and implement the algorithms described in this thesis on a computer.

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APPENDIX 1

INTRODUCTION

This appendix describes the design of a digital IIR filter comparable to the analog 10 KHz high pass filter shown in figure 4. The Bilinear Transform (BLT) technique will be used. The BLT technique was chosen because it produces a filter with a small number of coefficients.

CHARACTERIZATION OF ANALOG HIGH PASS FILTER FREQUENCY RESPONSE

The development starts with the calculation of the magnitude frequency response of the analog high pass filter. For simplicity, the source driving the high pass filter is assumed to have zero ohms impedance.

The functions representing the magnitude frequency response of the analog high pass filter in the s domain are shown in equations A1.1,A1.2 and A1.3.

$$H(s) = H1(s) \times H2(s) \qquad (A1.1)$$

$$H1(s) = \frac{S^2 R_2 L_2}{S^2 R_2 L_2 + \frac{SL_2}{C_2} + \frac{R_2}{C_2}} \qquad (A1.2)$$

$$H2(s) = \frac{S^2 R_2 L_2 L_1 + \frac{L_1}{C_2}\left[R_2 SL_2\right]}{\frac{R_2 + SL_2}{S^2 C_1 C_2} + \frac{R_2 L_2}{C_1} + \left[R_2 + SL_2\right]\left[\frac{L_1}{C_1} + \frac{L_1}{C_2}\right] + S^2 R_2 L_2 L_1} \qquad (A1.3)$$

As shown above, the transfer function for the high pass filter has been factored into the product of two separate functions, H1(s) and H2(s).

SCALING OF ANALOG FILTER PRIOR TO BLT APPLICATION

The Bilinear Transform maps the left half of the analog (s) plane to the inside of the unit circle on the digital (z) plane. The right half of the analog plane is mapped outside the unit circle on the digital plane. Also, the imajinary axis is mapped to the unit circle, with the positive part of the imajinary axis mapped from o to PI on the unit circle, and the negative part of the imajinary axis mapped from 0 to -PI on the unit circle

[3].

All digital frequencies are below F(S)/2 = PI, where F(S) is the sampling frequency.

A one to one correspondence exists between analog and digital frequencies as shown by equation A1.4 [3]. This equation also shows tangent warping between the analog and digital planes. Warping is not a problem because the corner frequency of the low pass filter is low in the linear range of equation A1.4, and the pass band is flat.

The analog filter must be scaled before application of the BLT, so the -3dB point on the analog filter (at 10 KHz) will produce the desired -3dB point at 10 kHz on the digital filter. The sampling frequency is F(s) = 1/T, where F(s) = 13.5E6 Hz.

$$w(A) = \tan\left[\frac{w(D)T}{2}\right] = \tan\left[\frac{(2)(PI)(10000)}{2 \text{ X } 13.5E6}\right] \qquad (A1.4)$$

Where:

w(A) = Radian Frequency in the Analog Domain

w(D) = Radian Frequency in the Digital Domain

(2)(PI)(10000) is the analog corner frequency, (-3dB point) in radians/sec

To scale, multiply C and L of the original analog filter

by:

$$U = \frac{(10,000)(2)(PI)}{w(A)} \tag{A1.5}$$

This produces an analog filter of the same shape as the original filter but with a much lower corner frequency.

This is the function to which the BLT will be applied. It will translate analog frequencies between 0 frequency and F(s)/2 to the digital domain between 0 and PI radians per second.

IMPLEMENTATION OF BILINEAR TRANSFORM ON H1(S) AND H2(S)

Implementation of the BLT is accomplished by making the substitution s = ((z-1)/(z+1)) (this is a version of the BLT) in the equations for H1(s) and H2(s). The result of this substitution is shown below in equation A1.6 for H1(s) and equation A1.7 for H2(S). These substitutions produce functions of H(z), from functions of H(s).

$$H1(z) \;=\; \frac{[z-1]^2\, R_2 L_2}{[z-1]^2\, R_2 L_2 + [z-1][z+1]\dfrac{L_2}{C_2} + [z-1]^2\dfrac{R_2}{C_2}} \qquad \text{(A1.6)}$$

$$H2(z) \;= \qquad \text{(A1.7)}$$

$$\frac{\dfrac{[z-1]^4}{[z+1]^4} R_2 L_2 L_1 + \dfrac{[z-1]^3}{[z+1]^3}\dfrac{L_2 L_1}{C_2} + \dfrac{[z-1]^2}{[z+1]^2}\dfrac{R_2 L_1}{C_2}}{\dfrac{[z-1]^4}{[z+1]^4} R_2 L_2 L_1 + \dfrac{[z-1]^3}{[z+1]^3}\dfrac{L_2 L_1}{C_1} + \dfrac{L_2 L_1}{C_2} + \dfrac{[z-1]^2}{[z+1]^2}\left[\dfrac{R_2 L_1}{C_1} + \dfrac{R_2 L_1}{C_2} + \dfrac{R_2 L_2}{C_1}\right] + \dfrac{[z-1] L_2}{[z+1] C_1 C_2} + \dfrac{R_2}{C_1 C_2}}$$

These equations can be written in the form shown below in equations A1.8 and A1.9. They have been made causal by multiplying by z. The coefficients are shown in table A1.1 where each L and C was multiplied by U defined in equation A1.5.

$$H1(z) \;=\; \frac{a(0) - a(1)z^{-1} + a(2)z^{-2}}{b(0) + b(1)z^{-1} + b(2)z^{-2}} \qquad \text{(A1.8)}$$

$$H2(z) \;=\; \frac{a(0) + a(1)z^{-1} + a(2)z^{-2} + a(3)z^{-3} + a(4)z^{-4}}{b(0) + b(1)z^{-1} + b(2)z^{-2} + b(3)z^{-3} + b(4)z^{-4}} \qquad \text{(A1.9)}$$

The function for H2(z) has been factored into the product of two second order functions $H2^{\prime}(z)$ and $H2^{\prime\prime}(z)$. Thus the overall filter function is the product of three second order functions shown below in equation A1.10. The coefficients for this filter are listed in table A1.2.

$$H(z) = H1(z) \times H2^{\prime}(z) \times H2^{\prime\prime}(z) \tag{A1.10}$$

$$H2^{\prime}(z) = \frac{a(0) + a(1)z^{-1} + a(2)z^{-2}}{b(0) + b(1)z^{-1} + b(2)z^{-2}} \tag{A1.11}$$

$$H2^{\prime}(z) = \frac{a(3) + a(4)z^{-1} + a(5)z^{-2}}{b(3) + b(4)z^{-1} + b(5)z^{-2}} \tag{A1.12}$$

DETERMINATION OF STABILITY FOR H1(z), H2 (z) and H2 (z)

The complex poles and zeros were calculated for

TABLE 1

COEFFICIENTS FOR H1(z) AND H2(z)

FOR H1(z):

a(0) = 0.997484569	b(0) = 1
a(1) = -1.004969139	b(1) = -1.994964664
a(2) = 0.997484569	b(2) = 0.994973614

FOR H2(z):

a(0) = 0.996452654	b(0) = 1
a(1) = 3.980793140	b(1) = -3.987855499
a(2) = 5.963672417	b(2) = 5.963640214
a(3) = -3.970776028	b(3) = -3.963713669
a(4) = 0.991444098	b(4) = 0.987928954

TABLE 2

COEFFICIENTS FOR H2 (z) AND H2 (z)

FOR H2 (z):

a(0) = 1	b(0) = 1
a(1) = -1.99930466618	b(1) = -1.99303989802
a(2) = 0.99930353110	b(2) = 0.99304032880

FOR H2 (z):

a(3) = 1	b(3) = 1
a(4) = -1.99565999735	b(4) = -1.99481560125
a(5) 0.99566706480	b(5) = 0.99485380282

H1(z), H2 (z) and H2 (z).

Below are shown the complex zeros for H1(z):

H(1) = (1.000000000, 0.000002832)

H(2) = (1.000000000, -0.000002832)

The complex poles for H1(z) are:

J(1) = (0.997482332, 0.001616008)

J(2) = (0.997482332, -0.001616008)

These pole/zero values are in rectangular coordinates. The magnitude of each pole is less than 1 and they are therefore inside the unit circle which implies that the transfer function H1(z) is stable for this causal filter. (Care must be taken during implementation of a digital filter, or coefficient approximation can cause the poles to migrate outside the unit circle causing the transfer function to become unstable.)

The complex zeros for H2 (z) are:

H(1) = (0.99929809570, 0)

H(2(= (1, 0)

The complex poles for H2 (z) are:

J(1) = (0.99303817749,0)

J(2) = (1,0)

The complex zeros for H2 (z) are:

H(3) = (0.99782943726, 0.00259475382)

H(4) = (0.99782943726, -0.00259475382)

The complex poles for H2 (z) are:

J(3) = (0.99740982056, 0.00487940062)

J(4) = (0.99740982056, -0.00487940062)

These poles are all expressed in rectangular coordinates and their magnitudes are less than 1 which means that $H2'(z)$ and $H2''(z)$ are also stable causal filters.

THE RESULT OF IMPLEMENTING THE FILTER WITH FINITE COEFFICIENTS AND ROUNDING THE RESULTANT PRODUCTS

The implementation of a IIR digital filter is represented in equation A1.13, using infinite precision arithmetic.

$$y(n) = \sum_{K=0}^{M} x(n-k)a(k) - \sum_{k=1}^{L} y(n-k)b(k) \qquad \text{(A1.13)}$$

When a practical digital filter algorithm is implemented, past and present input and output samples are multiplied by finite length coefficients. When the samples are represented by A bits and the coefficients are represented by B bits, the resulting product has A +

B bits. In the case of FIR filters the resultant full precision output will have A + B bits. In the case of a full infinite precision IIR filter, the precision or number of bits of resolution would increase with each pass of an output sample through the algorithm. However with real machines, the resultant products are rounded or truncated to A or B bits of resolution. In either case an error term is generated which causes the output to be the sum of the error plus what the output would have been if infinite precision arithmetic were performed [4].

The noise from each of the two pole filter sections is added as shown in figure A1.1 [4].

When this filter was implemented on a computer, the output rms round off error was approximately 40 dB below the output peak-to-peak level for a sine wave input signal. This round off error is deemed to be too high and is much higher than the round off error generated in the other filters.

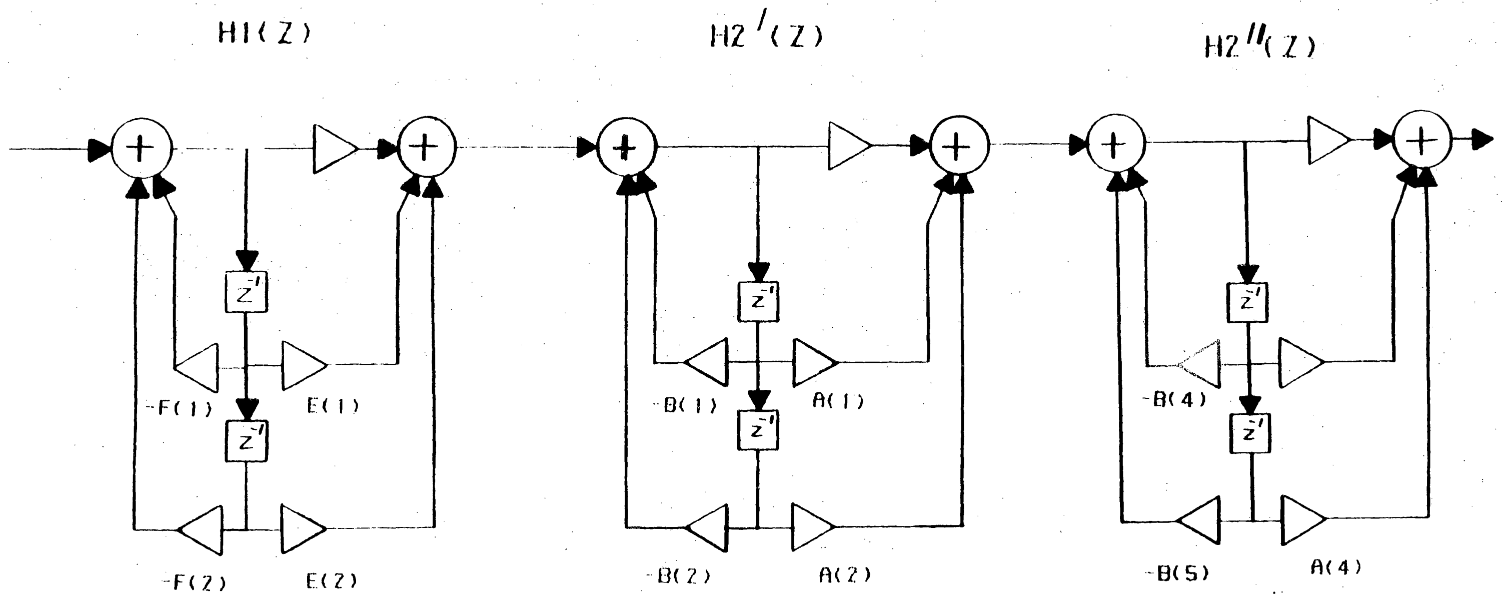

FIGURE A1.1 DIRECT FORM DIGITAL FIR FILTER STRUCTURE
(IMPLEMENTS 3 CASCADED 2 POLE SECTIONS TO FORM HIGHPASS FILTER)

APPENDIX 2

This appendix describes how amplitude and phase can be calculated from four synchronous samples of a single frequency signal, such as a sinewave, when four samples are taken during one period.

The basis for this algorithm is the Discrete Fourier Transform (DFT). The DFT is shown below:

$$A(p) = \sum_{n=0}^{N-1} x(n)W(N)^{pn}$$

Where: $W(N) = e^{\frac{-j(2)(PI)}{N}}$

Assume that four synchronous samples, x(0), x(1), x(2), and x(3) have been taken during one period of a sinewave. The DFT will be applied to these four samples. Applying the DFT in this case with p = 1 results in:

$$A(1) = x(0) + x(1)\, e^{\frac{-j(PI)}{2}} + x(2)(-1) + x(3)\, e^{\frac{-j(3)(PI)}{2}}$$

$A(1) = x(0) - jx(1) - x(2) + jx(3)$; where A(1) is in rectangular form.

Below the algorithm is applied to a cosine with the sample points indicated:

x(0) = cos(90 degrees); x(1) = cos(180 degrees);
x(2) = cos(270 degrees); x(3) = cos(360 degrees).

These sample points produce a magnitude of 2.00 at an angle of 90 degrees.

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