DOKUZ EYLÜL UNIVERSITY
GRADUATE SCHOOL OF NATURAL AND APPLIED SCIENCES

NEW HIGH PERFORMANCE PHASE EQUALIZER DESIGN

by

Çağlar HENDEN

August, 2006

İZMİR

NEW HIGH PERFORMANCE PHASE EQUALIZER DESIGN

A Thesis Submitted to the
Graduate School of Natural and Applied Sciences of Dokuz Eylül University
In Partial Fulfillment of the Requirements for the Degree of Master of Science
in Electrical and Electronics Engineering

by
Çağlar HENDEN

August, 2006
İZMİR

M.Sc THESIS EXAMINATION RESULT FORM

We have read the thesis entitled **"NEW HIGH PERFORMANCE PHASE EQUALIZER DESIGN"** completed by **ÇAĞLAR HENDEN** under supervision of **ASSOC. PROF. DR. UĞUR ÇAM** and that in our opinion it is fully adequate, in scope and in quality, as a thesis for the degree of Master of Science.

Assoc. Prof. Dr. Uğur ÇAM

(Supervisor)

(Jury Member)

(Jury Member)

Prof. Dr. Cahit HELVACI

Director

Graduate School of Natural and Applied Sciences

ACKNOWLEDGMENTS

I would like to thank my supervisor Assoc. Prof. Dr. Uğur ÇAM for his valuable guidance and support during the course of this thesis. I also want to thank to Gulsen Karaman for her help during the thesis. Finally, I want to thank my parents for their support throughout my life.

Çağlar HENDEN

NEW HIGH PERFORMANCE PHASE EQUALIZER DESIGN

ABSTRACT

In this thesis, theoretical background is given for video signals and video signal processing. Active filter blocks for video systems are introduced and simulated using Pspice. The delay errors of the video filters are also shown. New active blocks are introduced such as operational transconductance amplifier (OTA), current conveyor (CCII) and operational transresistance amplifier (OTRA). Wide band OPAMP, OTA, CCII and OTRA are used as all-pass filters and their spice simulations are given. The active blocks are then used as phase equalizer for an anti-aliasing filter to regulate delay error. The introduced active filters are evaluated and their performances are compared.

Keywords: Phase Equalizers, Allpass Filters, Video Frequency Filters, Operational Transconductance Amplifier, Current Conveyor, Operational Transresistance amplifier.

YENİ YÜKSEK PERFORMANSLI FAZ DENGELEYİCİ TASARIMI

ÖZ

Bu tezde video sinyalleri ve video sinyali işleme hakkında bilgiler verilmiştir. Video sistemlerinde kullanılan aktif filtre blokları tanıtılmış ve bu blokların Spice simülasyonları yapılmıştır. Aktif filtrelerin frekans bağımlı gecikmeleri gösterilmiştir. Video uygulamalarında kullanılmak üzere OTA, CCII ve OTRA gibi yeni aktif elemanlar tanıtılmıştır. Yüksek performanslı OPAMP, OTA, CCII ve OTRA tüm geçiren filtre olarak kullanılmıştır ve bu blokların Spice simülasyonları yapılmıştır. İlgili bloklar antialiasing filtrenin frekans bağımlı gecikmesini düzenlemek için faz dengeleyici olarak kullanılmıştır. Tanıtılan aktif filtreler değerlendirilmiş ve performansları karşılaştırılmıştır.

Anahtar Kelimeler: Faz Dengeleyiciler, Tüm geçiren filtreler, Video Frekansı Filtreleri, OTA, CCII, OTRA.

CONTENTS

CHAPTER ONE
INTRODUCTION

A filter is a circuit designed to pass a certain band of frequencies while attenuating all frequencies outside of this band. Filters can also be used to regulate the delay of a system. In this thesis, we will investigate different filter types that can be used to regulate delay in video processing systems.

Video frequency filters can be mainly grouped into two headlines. These are Antialiasing Filters, Reconstruction Filters. We will explain these filters in chapter III. In chapter IV, we will give some examples of the video filters and their main drawback, which is delay. To overcome the delay problem, we will investigate several all-pass filter blocks with spice simulations.

The first block that has been investigated is the conventional operational amplifier. OPAMP is used as the active device in the vast majority of the active filter literature and we will use it in classical video filters applications in chapter four and chapter seven. But it has some limitations and drawbacks, such as GBW limitation, limited maximum Quality Factor, Slew Rate problem. So we want to use new active blocks in video frequency signal applications, such as OTAs, CCIIs and OTRAs.

OTAs' structures offer improvements in design simplicity and programmability when compared to op amp based structures as well as reduced component count. Many of the basic OTA based structures use only OTAs and capacitors and, hence, are attractive for integration. Component count of these structures is often very low. From a practical viewpoint, the high-frequency performance of discrete bipolar OTAs is quite good.

Third active block we have used in this thesis is current conveyor. Instead of classical video filters with passive elements and op-amps, we have simulated CCII based all-pass filter. Detailed explanations and an all-pass filter application of CCIIs in video frequencies will explain in chapter six. Pspice simulation results as a delay equalizer in video filters will be simulated later in chapter seven.

The fourth and the last active block we have investigated is OTRA. The best candidate for the active element to implement the transimpedance all-pass filter is the OTRA, which is a high gain current-input, voltage output device. Unfortunately, the commercial realizations do not provide internal ground at the input port and they allow the input current to flow in one direction only. The former disadvantage limited the functionality of the OTRA.

Finally a conclusion will be given.

CHAPTER TWO
ANALOG FILTER FUNDAMENTALS

An analog filter uses analog electronic circuits made up from components such as resistors, capacitors and op-amps to produce the required filtering effect. Such filter circuits are widely used in such applications as noise reduction, video signal enhancement, graphic equalizers in hi-fi systems, and many other areas. There are well-established standard techniques for designing an analog filter circuit for a given requirement. At all stages, the signal being filtered is an electrical voltage or current, which is the direct analogue of the physical quantity (e.g. a sound or video signal or transducer output) involved. (Schaumann, R. & Valkenburg, 2001)

It is well known that there are two types of analog filters, which one is a passive filter and another one is an active filter.

2.1 Passive Filters

The oldest forms of electronic filters are passive analog linear filters. Passive filters consist of passive elements, which are resistors, inductors and capacitors. Passive filters transfer functions magnitude does not exceed one. Both cut-off frequency and the passband magnitude of passive filters were altered with the addition of a resistive load at the output of the filter.

2.2 Active Filters

Active filters consist of passive elements with also active elements such as operational amplifier, operational transconductance amplifier and first and second-generation current conveyor. Active filters provide a control over amplification. Active circuits are used to implement filter designs when gain, load variation, and physical size are important parameters in the design specifications.

2.3 Comparison of Passive and Active Filters

There are some advantages of passive filters as well as various disadvantages.

The advantages of passive filters are:

a. Easier to design and implement
b. Lower cost
c. Does not need power supplies

The drawbacks of passive filters are:

a. Very sensitive to the component value tolerances
b. For low frequencies, the values of R and C can be quite large, leading to physically large components
c. A first or second-order filter may not give adequate roll-off
d. If gain is required in the circuit, it cannot be added to the filter itself. Passive filters transfer functions magnitude does not exceed unity while active filters provide a control over amplification.
e. The filter can potentially have high output impedance. Since the resistor value is typically large, to keep the capacitors a reasonable value, the next stage device can see significant source impedance.
f. Cut-off frequency and the passband magnitude of passive filters were altered with the addition of a resistive load at the output of the filter.

Thus, we use active circuits to implement filter designs when gain, load variation, and physical size are important parameters in the design specifications. For example, active circuits can produce bandpass and bandreject filters without using inductors. This is desirable because inductors are usually large, heavy, and costly, and they may introduce electromagnetic field effects which compromise the desires frequency response characteristics.

CHAPTER THREE
VIDEO SIGNAL FUNDAMENTALS AND VIDEO FREQUENCY FILTERS

3.1 Color Spaces

A color space is a mathematical representation of a set of colors. The three most popular color models are RGB (used in computer graphics); YIQ, YUV, or YCbCr (used in video systems); and CMYK (used in color printing). All of the color spaces can be derived from the RGB information supplied by devices such as cameras and scanners.

The red, green, and blue (RGB) color space is widely used throughout computer graphics. Red, green, and blue are three primary additive colors (individual components are added together to form a desired color) and are represented by a three-dimensional, Cartesian coordinate system in figure 3.1. (Keith Jack, 2001)

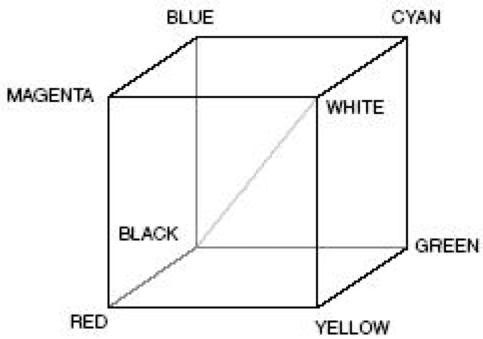


Figure 3.1 RGB color cube

However, RGB is not very efficient when dealing with “real-world” images. All three RGB components need to be of equal bandwidth to generate any color within the RGB color cube. The result of this is a frame buffer that has the same pixel depth and display resolution for each RGB component. Also, processing an image in the RGB color space is usually not the most efficient method. For example, to modify the intensity or color of a given pixel, the three RGB values must be read from the frame buffer, the intensity or color calculated, the desired modifications performed, and the new RGB values calculated and written back to the frame buffer. If the system had access to an image stored directly in the intensity and color format, some processing steps would be faster. For these and other reasons, many video standards use luma and two color difference signals. The most common are the YUV, YIQ, and YCbCr color spaces. Although all are related, there are some differences. (Keith Jack, 2001)

3.1.1 YUV

The YUV color space is used by the PAL (Phase Alternation Line), NTSC (National Television System Committee), and SECAM (Sequentiel Couleur Avec Mémoire or Sequential Color with Memory) composite color video standards. The basic equations to convert between R'G'B' gamma-corrected RGB and YUV are:

$$\begin{aligned} Y &= 0.299R' + 0.587G' + 0.114B' \\ U &= -0.147R' - 0.289G' + 0.436B' = 0.492(B' - Y) \\ V &= 0.615R' - 0.515G' - 0.100B' = 0.877(R' - Y) \end{aligned} \qquad \text{Eq.(3.1)}$$

3.1.2 YIQ Color Space

The YIQ color space is derived from the YUV color space and is optionally used by the NTSC composite color video standard. (The "I" stands for "in-phase" and the "Q" for "quadrature," which is the modulation method used to transmit the color information.)

3.1.3 YCbCr Color Space

The YCbCr color space was developed as part of ITU-R BT.601 during the development of a worldwide digital component video standard. YCbCr is a scaled and offset version of the YUV color space. There are several YCbCr sampling formats, such as 4:4:4, 4:2:2, 4:1:1 and 4:2:0 that are also described.

3.1.3.1 4:4:4 YCbCr Format

Figure 3.2 illustrates the positioning of YCbCr samples for the 4:4:4 format. Each sample has a Y, a Cb, and a Cr value. Each sample is typically 8 bits (consumer applications) or 10 bits (pro-video applications) per component.

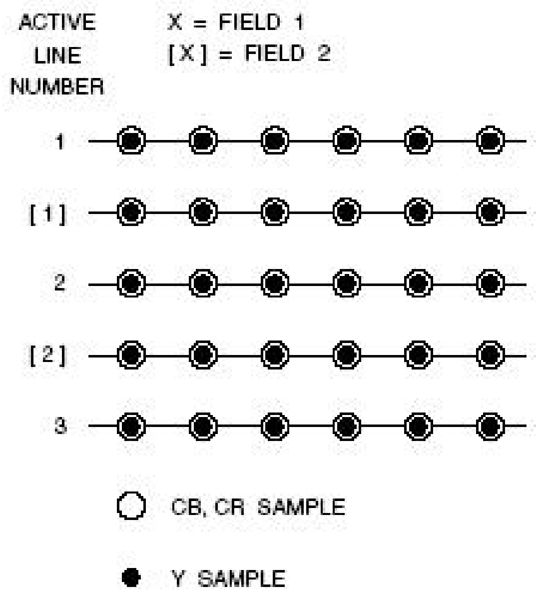


Figure 3.2 4:4:4 Co-Sited Sampling. The sampling positions on the active scan lines of an interlaced picture.

3.1.3.2 4:2:2 YCbCr Format

Figure 3.3 illustrates the positioning of YCbCr samples for the 4:2:2 format. For every two horizontal Y samples, there is one Cb and Cr sample. Each sample is typically 8 bits (consumer applications) or 10 bits (pro-video applications) per component.

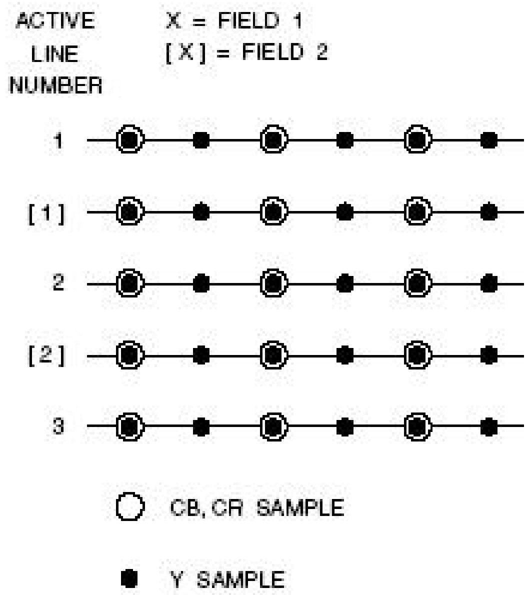


Figure 3.3 4:2:2 Co-Sited Sampling. The sampling positions on the active scan lines of an interlaced picture.

3.1.3.3 4:1:1 YCbCr Format

Figure 3.4 illustrates the positioning of YCbCr samples for the 4:1:1 format (also known as YUV12), used in some consumer video and DV video compression applications. For every four horizontal Y samples, there is one Cb and Cr value. Each component is typically 8 bits. Each sample therefore requires 12 bits.

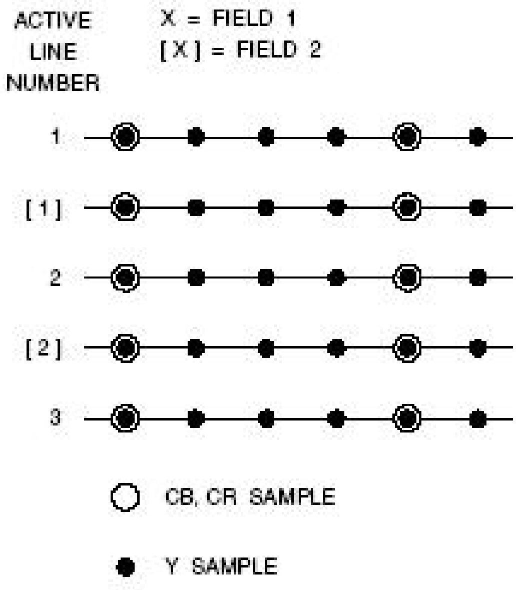


Figure 3.4 4:1:1 Co-Sited Sampling. The sampling positions on the active scan lines of an interlaced picture.

SAMPLE 0	SAMPLE 1	SAMPLE 2	SAMPLE 3	SAMPLE 4	SAMPLE 5
Y7 - 0	Y7 - 1	Y7 - 2	Y7 - 3	Y7 - 4	Y7 - 5
Y6 - 0	Y6 - 1	Y6 - 2	Y6 - 3	Y6 - 4	Y6 - 5
Y5 - 0	Y5 - 1	Y5 - 2	Y5 - 3	Y5 - 4	Y5 - 5
Y4 - 0	Y4 - 1	Y4 - 2	Y4 - 3	Y4 - 4	Y4 - 5
Y3 - 0	Y3 - 1	Y3 - 2	Y3 - 3	Y3 - 4	Y3 - 5
Y2 - 0	Y2 - 1	Y2 - 2	Y2 - 3	Y2 - 4	Y2 - 5
Y1 - 0	Y1 - 1	Y1 - 2	Y1 - 3	Y1 - 4	Y1 - 5
Y0 - 0	Y0 - 1	Y0 - 2	Y0 - 3	Y0 - 4	Y0 - 5
CB7 - 0	CB5 - 0	CB3 - 0	CB1 - 0	CB7 - 4	CB5 - 4
CB6 - 0	CB4 - 0	CB2 - 0	CB0 - 0	CB6 - 4	CB4 - 4
CR7 - 0	CR5 - 0	CR3 - 0	CR1 - 0	CR7 - 4	CR5 - 4
CR6 - 0	CR4 - 0	CR2 - 0	CR0 - 0	CR6 - 4	CR4 - 4

12 BITS PER SAMPLE

-0 = SAMPLE 0 DATA
-1 = SAMPLE 1 DATA
-2 = SAMPLE 2 DATA
-3 = SAMPLE 3 DATA
-4 = SAMPLE 4 DATA

Figure 3.5 4:1:1 Frame Buffering Format

3.1.3.4 4:2:0 YCbCr Format

Rather than the horizontal-only 2:1 reduction of Cb and Cr used by 4:2:2, 4:2:0 YcbCr implements a 2:1 reduction of Cb and Cr in both the vertical and horizontal directions. It is commonly used for video compression. (Keith Jack, 2001)

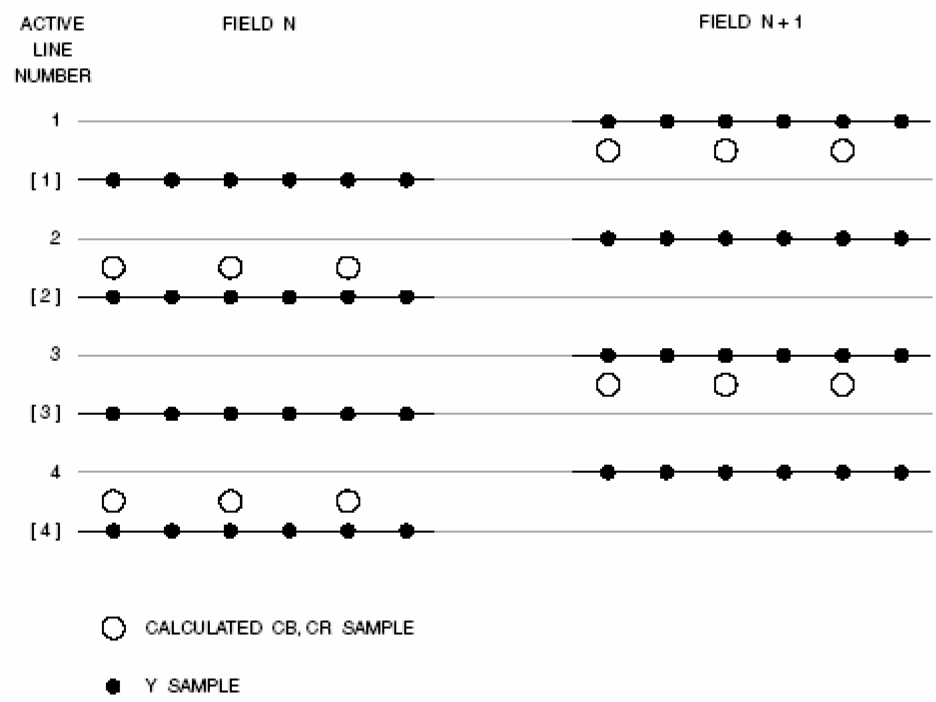


Figure 3.6 4:2:0 Sampling for MPEG2. The sampling positions on the active scan lines of an interlaced picture

3.2 Digital Component Video Background

In digital component video, the video signals are in digital form (YCbCr or R´G´B´), being encoded to composite NTSC, PAL, or SECAM only when it is necessary for broadcasting or recording purposes. The European Broadcasting Union (EBU) became interested in a standard for digital component video due to the difficulties of exchanging video material between the 625-line PAL and SECAM systems. A series of demonstrations was carried out to determine the quality and suitability for signal processing of various methods. From these investigations, the main parameters of the digital component coding, filtering, and timing were chosen and incorporated into ITU-R BT.601. BT.601 has since served as the starting point for other digital component video standards. (Keith Jack, 2001)

3.2.1 Coding Ranges

The selection of the coding ranges balanced the requirements of adequate capacity for signals beyond the normal range and minimizing quantizing distortion. 8 or 10 bits per sample are used for each of the YCbCr or R´G´B´ components. Although 8 bit coding introduces some quantizing distortion, it was originally felt that most video sources contained sufficient noise to mask most of the quantizing distortion. (Keith Jack, 2001)

3.2.2 BT.601 Sampling Rate Selection

Sampling of analog R´G´B´ or YUV video signals is done in “line lock” system. This technique produces a static orthogonal sampling grid in which samples on the current scan line fall directly beneath those on previous scan lines and fields, as shown figures 3.2 through 3.6. Another important feature is that the sampling is locked in phase so that one sample is coincident with the 50% amplitude point of the falling edge of analog horizontal sync. This ensures that different sources produce samples at nominally the same positions in the picture. Making this feature common simplifies conversion from one standard to another.

For 525-line and 625-line video systems, several Y sampling frequencies were initially examined. Because the lowest sample rate possible (while still supporting quality video) was a goal, a 12MHz sample rate was preferred for a long time, but eventually was considered to be too close to the Nyquist limit, complicating the filtering requirements. When the frequencies between 12 MHz and 14.3MHz were examined, it became evident that a 13.5MHz sample rate for Y provided some commonality between 525 and 625 line systems. Cb and Cr, being color difference signals, do not require the same bandwidth as the Y, so may be sampled at one-half the Y sample rate, or 6.75 MHz. The accepted notation for a digital component system with sampling frequencies of 13.5, 6.75, and 6.75 MHz for the luma and color difference signals, respectively, is 4:2:2 (Y:Cb:Cr). (Keith Jack, 2001)

One frame consists of 2 fields. For 50Hz systems frames has a frequency of 25Hz. So each frame has 40ms period. For 625 line systems each line has a period of 64µs (40ms/625). From this information we can find that with 13.5MHz sampling, each scan line contains 858 samples (525-line systems) or 864 samples (625-line systems) and consists of a digital blanking interval followed by an active line period. Both the 525 and 625 line systems use 720 samples during the active line period (53.3µ for 625 line system). Having a common number of samples for the active line period simplifies the design of multi-standard equipment and standards conversion. With a sample rate of 6.75 MHz for Cb and Cr (4:2:2 sampling), each active line period contains 360 Cr samples and 360 Cb samples.

Initially, BT.601 supported only 525- and 625-line interlaced systems with a 4:3 aspect ratio (720x480 and 720x576 active resolutions). Support for a 16:9 aspect ratio was then added (960x480 and 960x576 active resolutions) (720x16/12) using an 18MHz (13.5x16/12) sample rate. (Keith Jack, 2001)

3.3 Video Resolution

It is common to see video resolutions of 720 x̣480 or 1920 x̣1080. However, those are just the number of horizontal samples and vertical scan lines, and do not necessarily convey the amount of unique information. For example, an analog video signal can be sampled at 13.5 MHz to generate 720 samples (≅53µs-active region) per line. Sampling the same signal at 27 MHz would generate 1440 samples per line. However, only the number of samples per line has changed, not the resolution of the content. Therefore, video is usually measured using "*lines of resolution*". In essence, how many distinct black and white vertical lines can be seen across the display? This number is then normalized to a 1:1 display aspect ratio (dividing the number by 3/4 for a 4:3 display, or by 9/16 for a 16:9 display). Of course, this results in a lower value for widescreen (16:9) displays, which goes against intuition.

Standard definition video usually has an active resolution of 720 x̣480 or 720 x̣576 interlaced. This translates into a maximum of about 540 lines of resolution, or a 6.75 MHz bandwidth (720 / 53.3 µs =13.5MHz sampling-Nyquist frequency). Standard NTSC, PAL, and SECAM systems fit into this category. For broadcast NTSC, with a maximum bandwidth of about 4.2 MHz, this results in about 330 lines of resolution. The format of the signals varies from country to country. In the United

States and Japan, the NTSC format is used. NTSC stands for National Television Systems Committee, which is the name of the organization that developed the standard. In Europe, the PAL format is common. PAL (phase alternating line), developed after NTSC, is an improvement over NTSC. SECAM is used in France and stands for sequential coleur avec memoire (with memory). It should be noted that there is a total of about 15 different sub-formats contained within these three general formats. Each of the formats is generally not compatible with the others. Although they all utilize the same basic scanning system and represent color with a type of phase modulation, they differ in specific scanning frequencies, number of scan lines, and color modulation techniques, among others. (Keith Jack, 2001)

3.4 Computer Video Timing

The various computer formats (such as VGA, XGA) differ substantially, with the primary difference in the scan frequencies. These differences do not cause as much concern, because most computer equipment is now designed to handle variable scan rates. This compatibility is a major advantage for computer formats in that media, and content can be interchanged on a global basis. The Video Electronics Standards Association (VESA) defines the timing for progressive analog R´G´B´ signals that drive computer monitors. Some consumer products are capable of accepting these progressive analog R´G´B´ signals and displaying them. Common active resolutions and their names are:

640 × 400 VGA

640 × 480 VGA

854 × 480 SVGA

800 × 600 SVGA

1024 × 768 XGA

1280 × 768 XGA

1280 × 1024 SXGA

1600 × 1200 UXGA

Common refresh rates are 60, 72, 75 and 85 Hz, although rates of 50–200 Hz may be supported. Graphics controllers are usually very flexible in programmability, allowing trading off resolution versus bits per pixel versus refresh rate. As a result, a large number of display combinations are possible. (Keith Jack, 2001)

3.5 Video Frequency Filters

Originally, video filters were passive L-C circuits surrounded by amplifiers, but the increased gain-bandwidth product (GBW) of modern op-amps makes it possible to combine them with R-C circuits to achieve smaller, more accurate designs. Active filters developed a bad reputation because of problems getting repeatable results until sensitivity analysis methods provided solutions for these problems in the 1960s. (WEB_1, 2003)

Today, op-amps and specialized design software for the computers make it possible to design high-bandwidth active filters. Unfortunately, this doesn't take into consideration the problems associated with video. For that, we'll need to understand the applications and formats used, as well as the nuances of active filter design. In an effort to do that; we'll look at today's major active-filter applications in broadcast and graphic (PC) video, which are:

Anti-Aliasing Filters: These are placed before an analog-to-digital converter (ADC), to attenuate signals above the sample rate (Nyquist Frequency) of the ADC.
Reconstruction Filters: These are placed after a digital-to-analog-converter (DAC), to filter the video and to amplitude-correct for the DAC response. They are also called sin (X)/(X), or zero-order-hold corrector.

Normally, these filters are designed for the steepest possible response to reject everything above the cutoff frequency. To avoid distorting the complex video waveform, the filter must be phase linear, where the linearity is specified as group delay variation; and the amount allowed will depend on the video format. The principal video formats in use today are listed as following.

Primaries or Native Video: These are the gamma-corrected primaries (R', G', B'), or linear R, G, B, depending on whether the video is broadcast or graphics. All these signals have the same bandwidth, and the group-delay variation must be sub-pixel to suppress visual artifacts.

Component Video: There are two component-video formats; color difference (Y, Pb, and Pr) found in broadcast and luma/chroma (Y-C), or S-VHS, found in VCRs. Both have a luma (Y) bandwidth that's greater than the color (Pb/Pr or C) bandwidth making it difficult to keep the signals time-coincident.

Composite Video: Composite video (CVBS) is only found in SDTV, not HDTV. It's a single channel format, and is the most forgiving of group-delay variation; in fact, group delay is seldom specified except for the RF Modulator path. (WEB_1, 2003)
Video Filters are typically low pass filters that are used to clean up artifacts caused by the data conversion process. There are two types of video filters that we will be interested in: Anti-Aliasing and Reconstruction.

3.5.1 Anti-aliasing Filters

Aliasing is a signal distortion caused by sampling, which can have an adverse effect on picture quality. This occurs when the analog signal is sampled by an analog to digital converter (ADC). An “Alias” or “fold back” is a replication of the original signal created in the frequency domain as a harmonic of the original signal and the sampling clock. If the original signal contains frequency components greater than half of the sampling (Nyquist) rate, then the original signal and the alias will sum to create an error, which is a type of distortion. In any case a bandwidth limit is needed on the digitized signal to restrict the frequency components to the original signal’s analog bandwidth. (Tole, J.W., 2000)

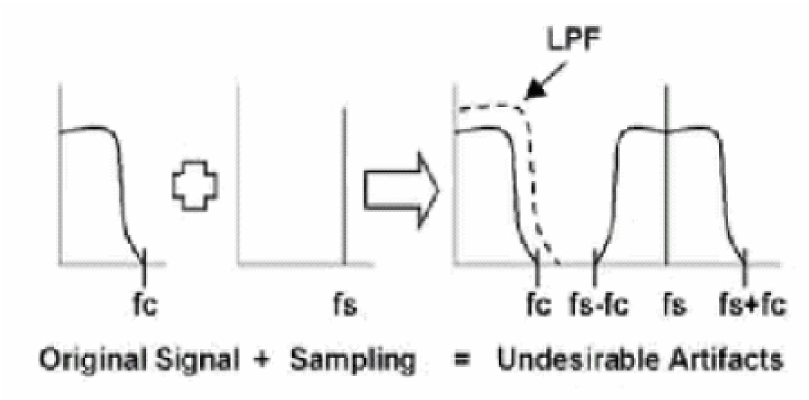
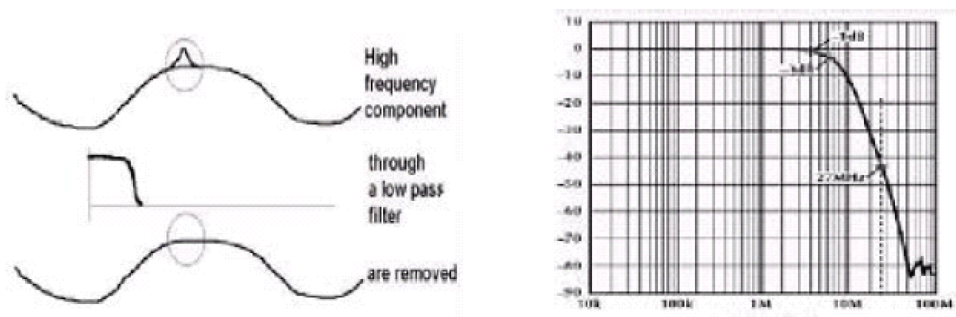


Figure 3.7 Basics of Anti-Alising

Aliasing causes several undesirable distortions to a video picture. Since the folded spectrum adds to the original spectrum it will sometimes be in phase and sometimes out of phase, causing ripples that depend on the position of the picture element relative to the clock. The net effect is that picture elements, edges, highlights, and details will "wink" in amplitude as they move across a picture if they have high frequency content above the Nyquist frequency of the sampler. (Tole, J.W., 2000)

Anti-aliasing reduces the bandwidth of the signal to a value appropriate for the sample processing system. Some of the detail information is lost, but only the information that cannot be displayed is removed. Assuming that the passband contains the picture information that is to be viewed, the only distortion that occurs is due to amplitude and phase variations of the anti-aliasing filter in the passband.



Removing Artifacts Filter Characteristics

Figure 3.8 Basic Antialiasing Filter Application and its frequency response

3.5.1.1 The Aliasing Phenomenon

In data acquisition systems, aliasing is a sampling phenomenon that can cause gross errors in results and reduce the accuracy of the data collected by an A/D card, which converts the analog output of a sensor into a digital number that can be read by the acquisition system's computer. It occurs whenever an input signal has frequency components at or higher than half the sampling rate. If the signal is not correctly band limited to eliminate these frequencies, they will show up as aliases or spurious lower frequency errors that cannot be distinguished from valid sampled data. The alias signals are actually at a higher frequency, but are converted by the sampling process to a false frequency below half the sampling rate. For example, with a sampling rate of 1,000 Hz, a signal at 800 Hz will be aliased to 200 Hz (the false lower frequency). Thus, aliasing is a phenomenon that occurs when a high frequency component effectively takes on the identity of a lower frequency. Slight fluctuations in the measured environment or the measured signal can cause alias signals to move, leaving errors in different locations throughout our data each time we use an A/D converter. One solution to the aliasing problem is to sample the signal at a very high rate and then filter out the high frequencies with digital techniques. But, such over sampling of data increases system costs by requiring faster A/D conversion for digital processing, more memory, and higher bandwidth buses. It also leads to higher analysis costs by creating more data to process and interpret.

3.5.1.2 How to Avoid Aliasing

A more practical alternative is to limit the bandwidth of the signal below one-half the sample rate with a low-pass or anti-alias filter, which can be implemented on each input channel in front of the A/D converter. Low-pass filtering must be done before the signal is sampled or multiplexed, since there is no way to retrieve the original signal once it has been digitized and aliased signals has been created.

To avoid aliasing with a low-pass filter, two processes actually must occur:

a. As dictated by the Nyquist theory, the input signal must be sampled at a rate of at least twice the highest frequency component of interest within the input signal.

b. Any frequency components above half the sampling rate (also called the Nyquist frequency) must be eliminated by an anti-alias filter before sampling. (Tole, J.W., 2000)

Under ideal conditions, a low-pass filter would exactly pass unchanged all slower signal components with frequencies from DC to the filter cutoff frequency. Faster components above that point would be totally eliminated, reducing the signal disturbance. But, real filters do not cut off sharply at an exact point. Instead, they gradually eliminate frequency components and exhibit a falloff or rolloff slope. These attenuation slopes typically range from 45 dB/octave to 120 dB/octave and "bottom out" at some finite value of stop band rejection, typically 75 to 100 dB. A simple illustration of these processes can be seen in the case of the 800Hz frequency aliasing to 200 Hz. Suppose that the 800 Hz is an unwanted interfering signal caused by an unwanted mechanical vibration. To prevent its alias from causing significant data errors at 200Hz, the 800Hz frequency must be removed by a low pass filter. If the cutoff point is set near 450 Hz, a filter with a steep roll off slope will eliminate the 800 Hz frequency, making the false 200 Hz frequency disappears. The input frequencies of interest below the filter cutoff (450 Hz) will still pass through the system unchanged.

High-frequency components can result from the inherent noise of the system itself and from noise or interference including 50Hz or 60Hz pickup, broadcasting stations, and mechanical vibrations. High-frequency components also are inherent in any sharp transitions of the measured signal. Lowpass filters generally can eliminate alias errors produced by these sources as long as the filters precede the A/D converter.

3.5.1.3 Benefits of Using Low-Pass Filters

The aliasing phenomenon becomes a problem in A/D conversion systems when an input signal contains frequency components above half the A/D sampling rate. These higher frequencies can "fold over" into the lower frequency spectrum and appear as erroneous signals that cannot be distinguished from valid sampled data. The best approach to eliminating false lower frequencies is to use a low-pass filter, which inhibits aliasing by limiting the input signal bandwidth to below half the sampling rate. A low-pass filter, which is applied to each input channel in front of the A/D card, also eliminates unwanted high-frequency noise and interference introduced prior to sampling. It reduces system cost, acquisition storage requirements, and analysis time by allowing for a lower sampling rate. Finally, a low-pass filter serves as an important element of any data acquisition system in which the accuracy of the acquired data is essential. (Tole, J.W., 2000)

3.5.2 Reconstruction Filters

Following digital processing, signals are typically converted back into the analog domain by a digital to analog converter (DAC), which is called reconstruction. High band spectral artifacts are introduced during the process that can distort picture quality. Reconstruction filters remove these artifacts. The filter’s reconstruction performance is based on how well the high band spectral artifacts are removed without distorting the valid signal spectral contents within the passband. Video signals are affected by these artifacts through a variation of the amplitude of small detail elements in the picture (such as highlights or fine pattern details), as the elements move relative to the sampling clock. The result is similar to the problem of aliasing and causes a “winking” of details as they move in the picture. (Tole, J.W., 2000)

When an ADC samples a signal, it creates multiple, recurring images, centered on the sample-clock's harmonics. The job of the reconstruction filter is to remove all but the baseband component. If the anti-aliasing filter did its job, the output of the DAC looks like image A in figure 3.9, and the images to the right are what we want to remove. Simple enough, except for one thing, the DAC samples are impulses and exist for only for an instant in time. To improve this, the DAC "holds" the impulse for a clock period, creating the familiar staircase waveform seen at a DAC output. This "hold" is a digital filter with a sin (x)/(x) characteristic, shown in figure 3.10, from whence it got its name, Sinc corrector. Notice at $0.5F_s$, the Nyquist frequency, the response is down 4db. Overlaying this on Figure 3.5 shows the insertion loss as a crosshatched area that includes the signal. The second job of the reconstruction filter is to restore this loss. The good news is the "hold" has a pole of attenuation centered on F_s, so we don't have to filter the sample clock, but applications still use the attenuation at F_s as a reference. (WEB_1, 2003)

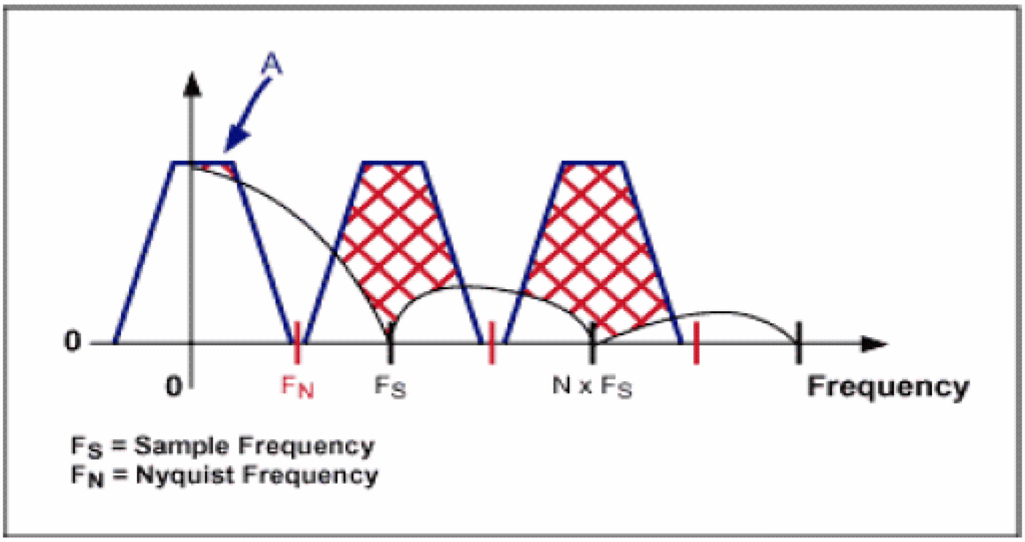


Figure 3.9 DAC output spectrum in terms of sampling frequency FS and Nyquist frequency FN.

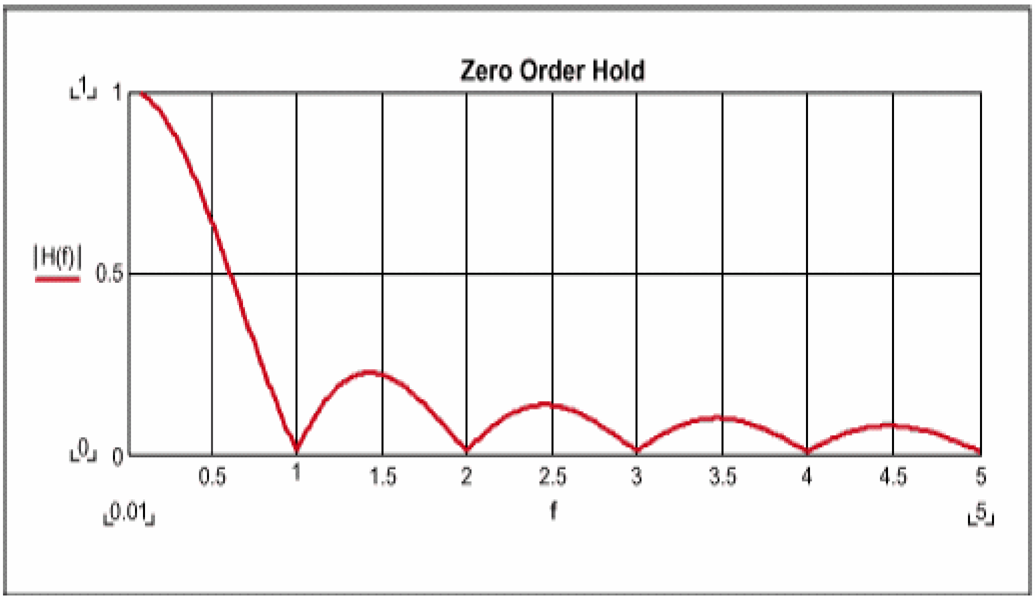


Figure 3.10 The sin(x)/(x) response of the DAC "hold" versus the sample frequency.

CHAPTER FOUR

VIDEO FILTER CIRCUITS THAT ARE USED IN VIDEO PROCESSING

Whether you use a video filter for antialiasing or reconstruction, the filter must have a lowpass characteristic to pass the video-frame rate. The industry categorizes lowpass filters by their amplitude characteristic or by the name of the polynomial, such as Bessel, Butterworth, Chebyshev, or Cauer, which describes it. Figure 4.1 shows these characteristics normalized to a 1-rad bandwidth. (Stutz, B., & Bekgran, M., 2003)

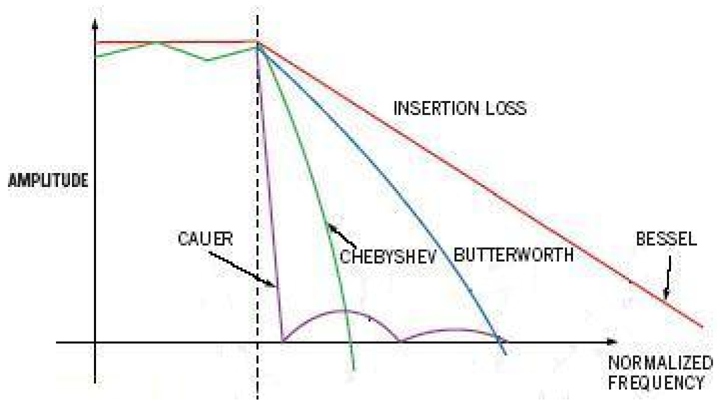


Figure 4.1 Different Filter characteristics

Increased gain-bandwidth product (GBW) of modern op-amps makes it possible to combine them with R-C circuits to achieve smaller, more accurate designs. In this section, we analyze the classical video frequency filters with passive elements (resistor, inductor, capacitor) and Op-Amp.

4.1 Antialiasing Filters

As we have seen in chapter 3 these devices are placed before an analog-to-digital converter (ADC) to attenuate signals above the Nyquist frequency, which is one half the sample rate of the ADC. These filters are usually designed with the steepest possible response to reject everything above the cutoff frequency. For applications, such performance is achieved using analog filters combined with digital filters and an over sampling ADC. (Stutz, B., & Bekgran, M., 2003)

4.1.1 Antialiasing Filter for XGA Graphics

For applications such as PC Graphics, very little filtering is required. For PC Video, the XGA resolution (1024*768 at 85Hz) has a sampling rate of 94.5 MHz and a Nyquist frequency of 47.25MHz. A Rauch realization of a 20MHz, 4-pole Butterworth filter can be used. (Stutz, B., & Bekgran, M., 2003)

A four-pole antialiasing filter for XGA graphics is simulated in figure 4.2. The circuit is based on LT1363 high speed op-amp.

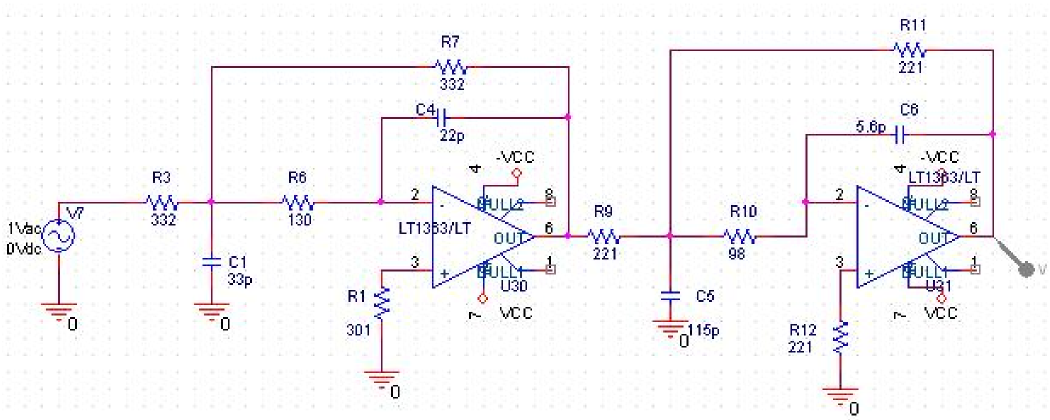


Figure 4.2 4-Pole 20MHz Butterworth Filter for XGA Graphics Antialiasing uses a Rauch Circuit

Gain response of the filter is shown in figure 4.3 with a cut of frequency at 20MHz.

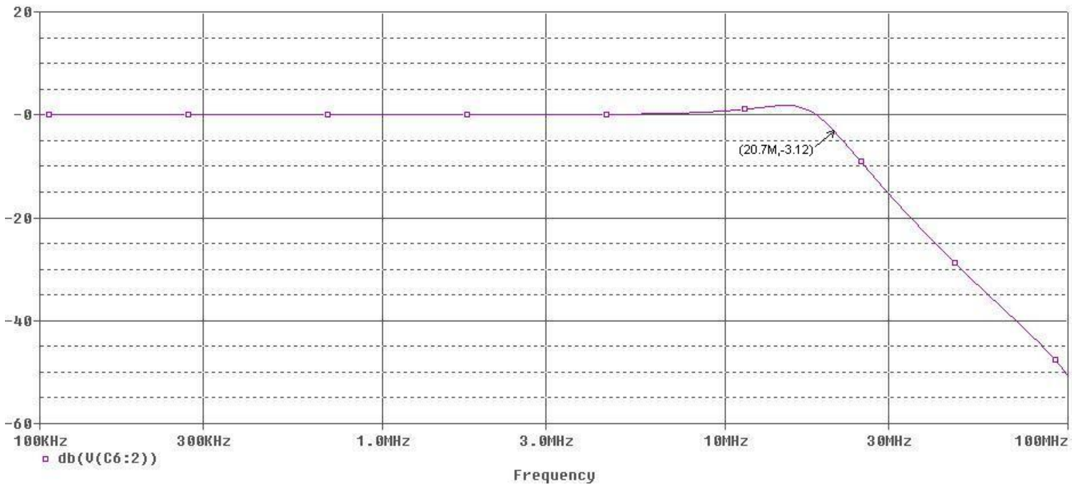


Figure 4.3 4-Pole 20MHz Butterworth Filter for XGA Graphics Antialiasing Gain Response

4.1.2 Antialiasing Filter for ITU-601

For Antialiasing filters, a template determines selectivity for ITU-601. Proposes and publishes video standards for Broadcast in the EU and ITU-R BT.601 is Universal Sampling spec for SDTV and HDTV Broadcast Video which is shown in figure 4.4. The specified bandwidth is 5.75 MHz, with an insertion loss of 12dB at 6.75 MHz and 40dB at 8 MHz, and with a group delay variation of 3ns over 0.1dB bandwidth, which will be discussed in chapter 7. Such performance is too difficult for an analog filter alone, but 4x over sampling modifies the requirements to 12dB at 27MHz and 40dB at 32MHz. Using software and normalizing curves a five-pole Butterworth filter with a -3dB bandwidth of 8.45MHz satisfies the requirement for selectivity. (Stutz, B., & Bekgran, M., 2003)

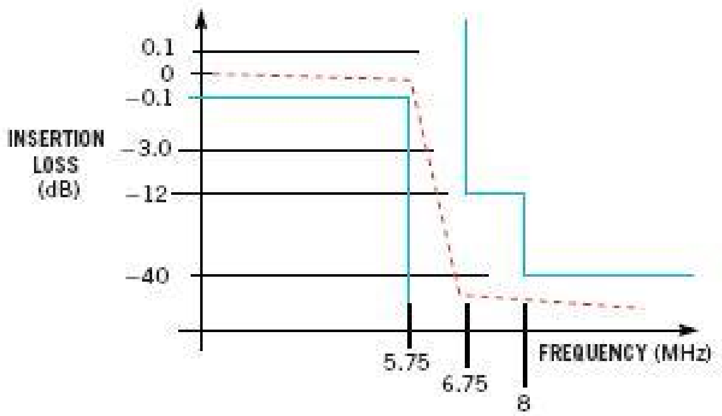


Figure 4.4 Antialiasing Requirements in accordance with the ITU-R BT.601-5 standard.

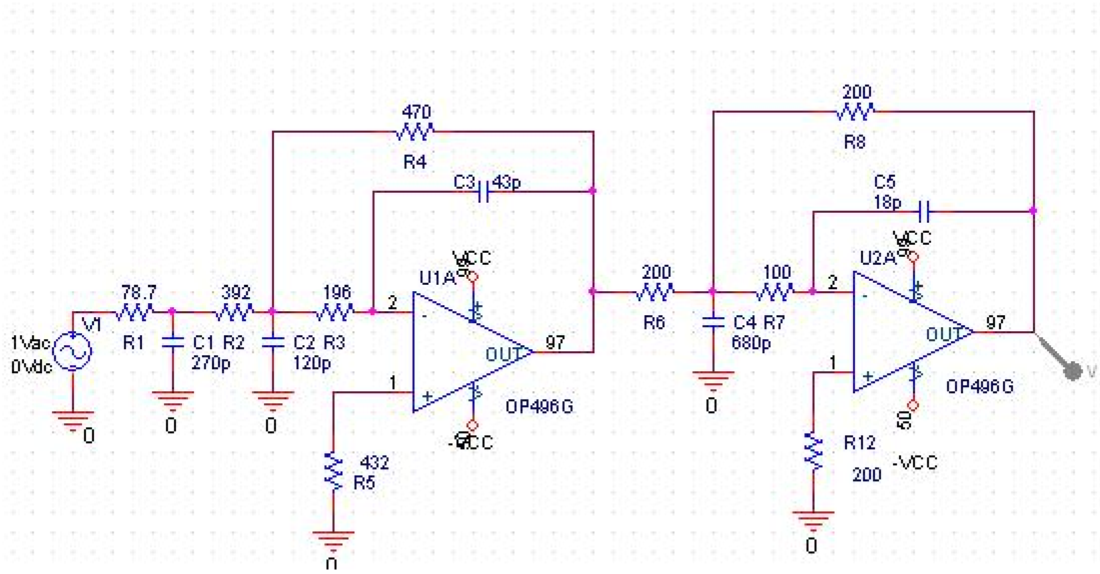


Figure 4.5 5-Pole 8.45 MHz Butterworth Filter for ITU-601 Antialiasing, using a Rauch circuit

Output response of the filter is shown in figure 4.6.

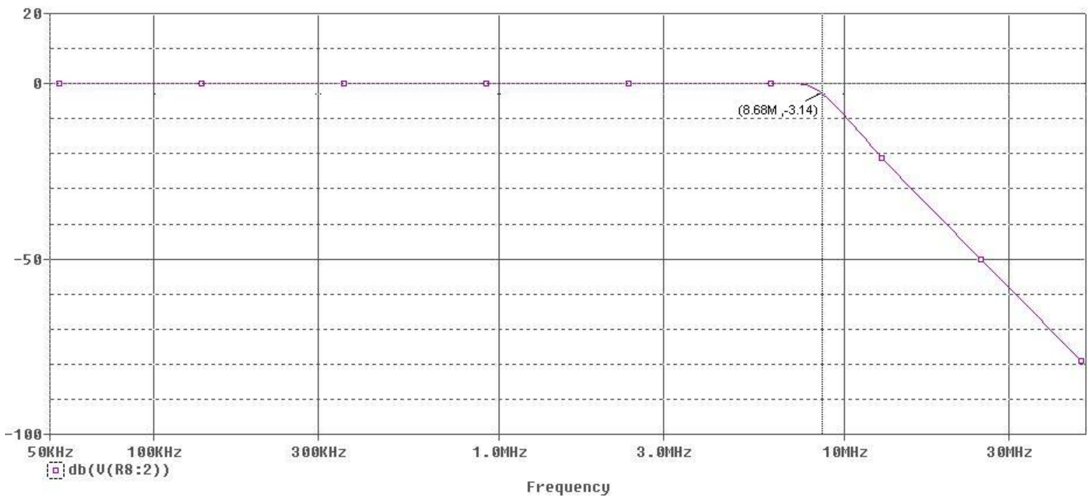


Figure 4.6 5-Pole 8.45 MHz Butterworth Filter for ITU-601 Antialiasing Output Response

4.2 Reconstruction Filters

As we have seen in chapter 3 reconstruction filters also called (sinx)/x or zero-order hold correctors, these filters are placed after a digital-to-analog converter (DAC) to remove multiple images created by sampling, though not to remove the DAC clock. Reconstruction filters are seldom as selective as antialiasing filters, because the DAC's hold function also acts as a filter an action that lowers the required selectivity, but introduces loss in response. When a signal is sampled, the samples are composed of multiple recurring signal images centered on harmonics of the sample clock. A reconstruction filter removes all but the baseband sample doesn't remove. If the antialiasing filter has served its purpose, the DAC outputs look like image "A" in figure 4.7, and then all samples to the right of it should be removed. Thus, reconstruction is similar to antialiasing except that, because each sample exists only for an instant, the DAC holds each for one clock period, thereby creating the familiar staircase approximation to a sloping line. (WEB_1, 2003)

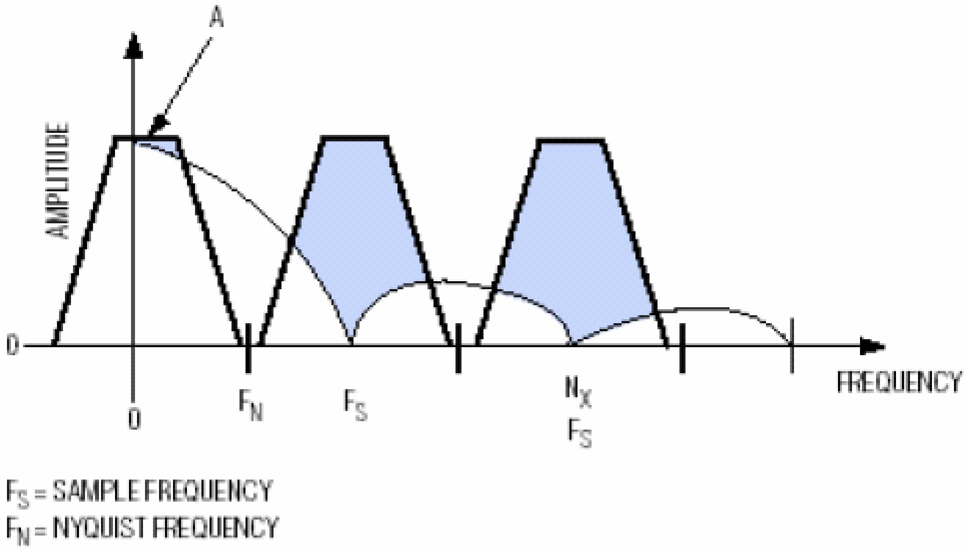


Figure 4.7 Typical DAC output spectrum is shown in terms of the sampling (F_S) and Nyquist (F_N) frequencies.

The hold function corresponds to a digital filter whose characteristic is similar to that of a Butterworth or Bessel filter in figure 4.8 (a). The response is decreased by 4dB at half the sample frequency. The second objective of a reconstruction filter is to restore that loss, which requires an amplitude equalizer like the circuit shown in figure 4.9. The equalizer is based on a delay stage and has a response like a Bessel filter. It can be designed from the DAC sample rate; f_s. Figure 4.8 (b) shows the DAC's frequency response with and without an amplitude equalizer. Like the delay stage, it can be included in any reconstruction filter. (WEB_1, 2003)

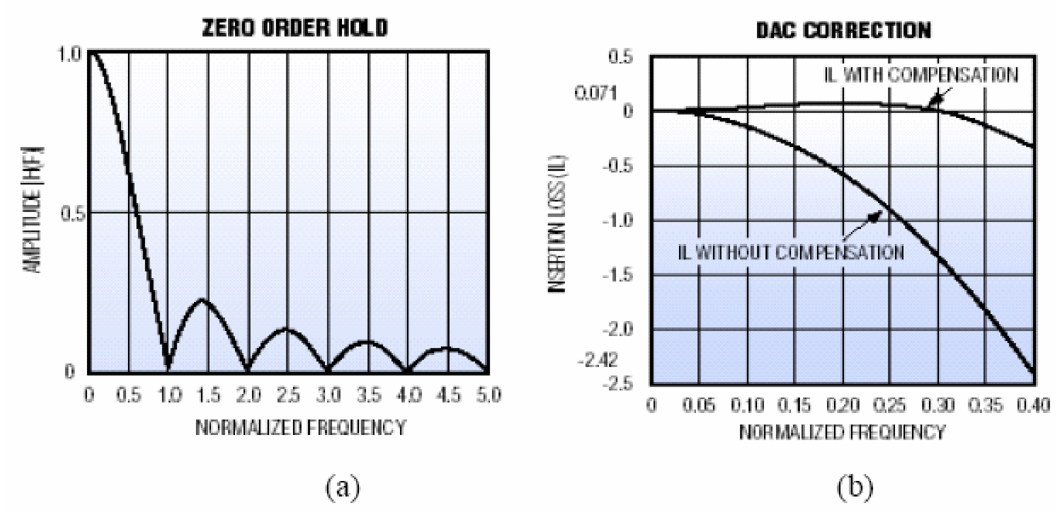


Figure 4.8 (a) The “hold” function of a DAC produces a (sinx)/x response with nulls at multiples of the sampling frequency. (b) A DAC output (b) is shown with and without the (sinx)/x correction.

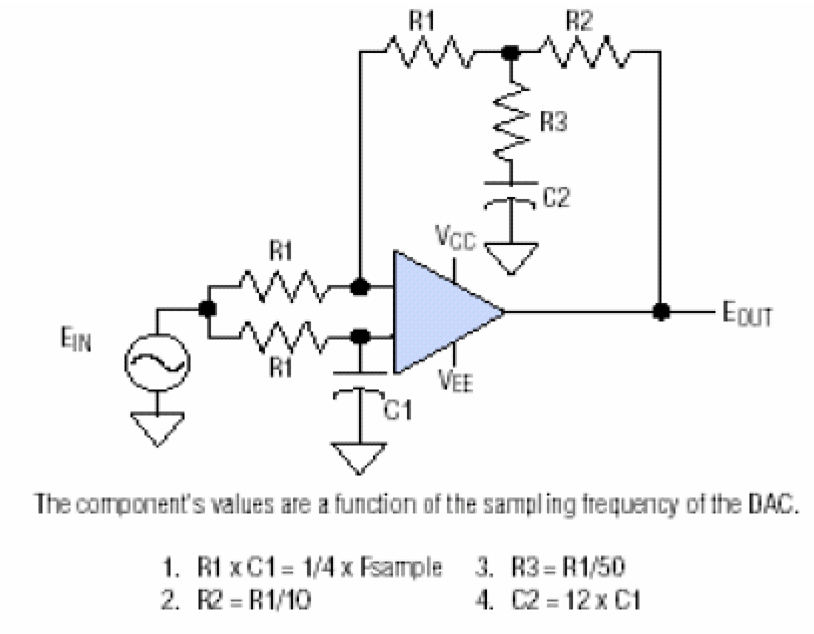


Figure 4.9 Amplitude Equalizer Circuit

The hold response also has a pole centered on the sample clock, which completely removes the clock. Nevertheless, most reconstruction applications refer to clock attenuation.

4.2.1 3 Pole PAL reconstruction filter

The circuit in Fig. 4.10 can be used for video reconstruction filtering for composite (CVBS) or S-Video signals in standard definition digital TV (SDTV) applications. The most common requirement for NTSC / PAL reconstruction is an attenuation of 20dB at 13.5 MHz and 40dB at 27 MHz, where w_c depends on the applicable video standard. A 3-pole Butterworth with Sallen-Key configuration is chosen for two reasons. First, its gain (+2) drives a back-terminated cable. Second, the group delay variation can be adjusted to optimize performance without a delay equalizer, which will be discussed in chapter 5. The circuit is designed to drive a 75-ohm termination, common in video applications, with an overall gain of 1. Like the anti-aliasing filter before a DAC, this filter is used to remove the higher frequency replicas of a signal following a DAC. (Stutz, B., & Bekgran, M., 2003)

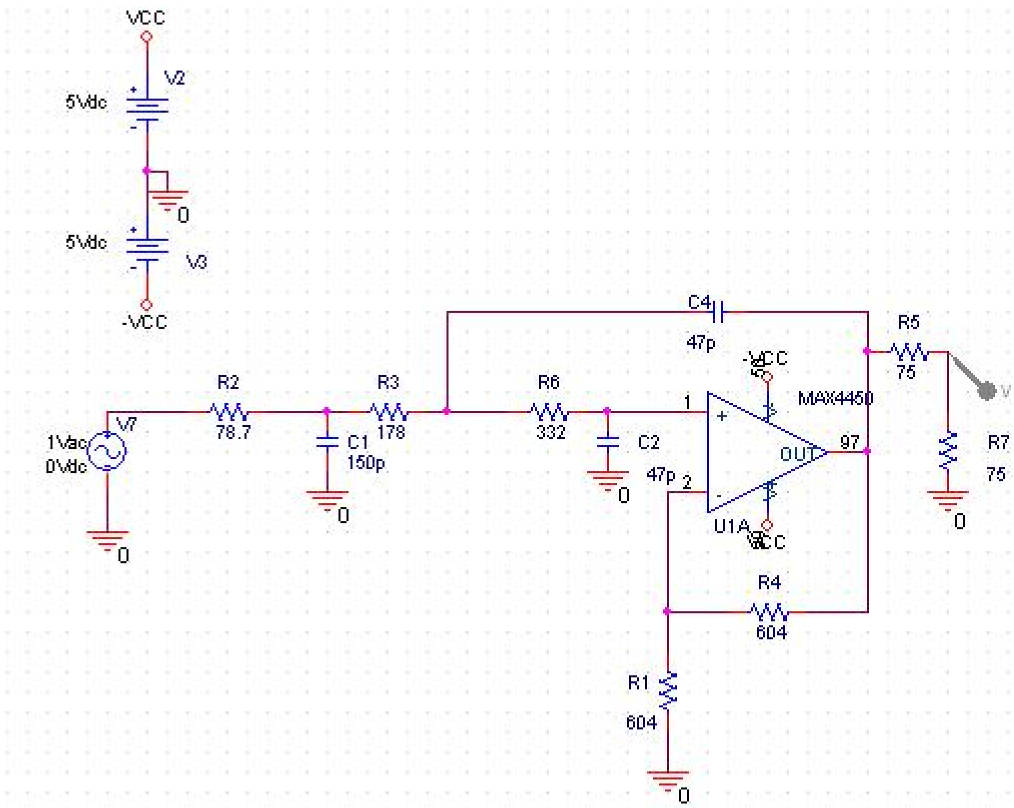


Figure 4.10 3-Pole Butterworth Filter with Sallen Key Configuration for PAL Reconstruction Output Response

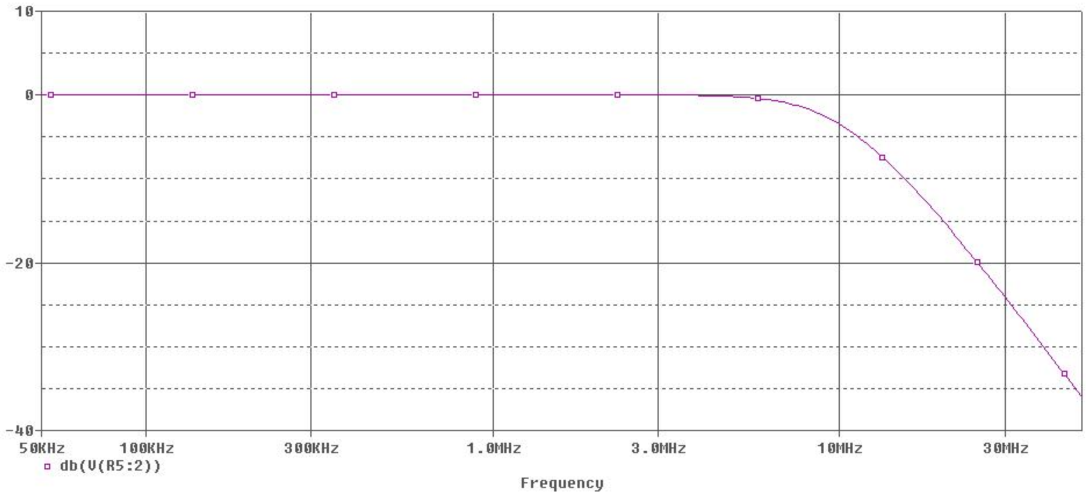


Figure 4.11 3-Pole Butterworth Filter with Sallen Key Configuration for PAL Reconstruction Output Response using MAX4450

To preserve quality in the video waveform, one should minimize group-delay variations in the filter and also any group-delay differential between filters, which will further be discussed in chapter 5. That capability requires a means for adjusting the filters' group delay without affecting its bandwidth. In figure 4.10, the addition of a resistor in series with C_2 and R can create a lag lead network. Keeping the sum of R and R_8 constant and equal to the original R-value preserves bandwidth by preserving the dominant-pole frequency. Increasing the R-value, on the other hand, introduces a 'lead' term that lowers group delay by reducing the rate of change in phase. These applications usually include digital amplitude correction for the DAC, which can be easily be added, if necessary. (Stutz, B., & Bekgran, M., 2003)

4.2.2 3 Pole XGA Reconstruction Filter

Figure 4.12 illustrates a circuit for XGA, a 20MHz three-pole Butterworth filter in the Sallen-Key configuration. This filter has a gain of 2 to drive a back terminated, 75Ω coaxial cable. (Stutz, B., & Bekgran, M., 2003)

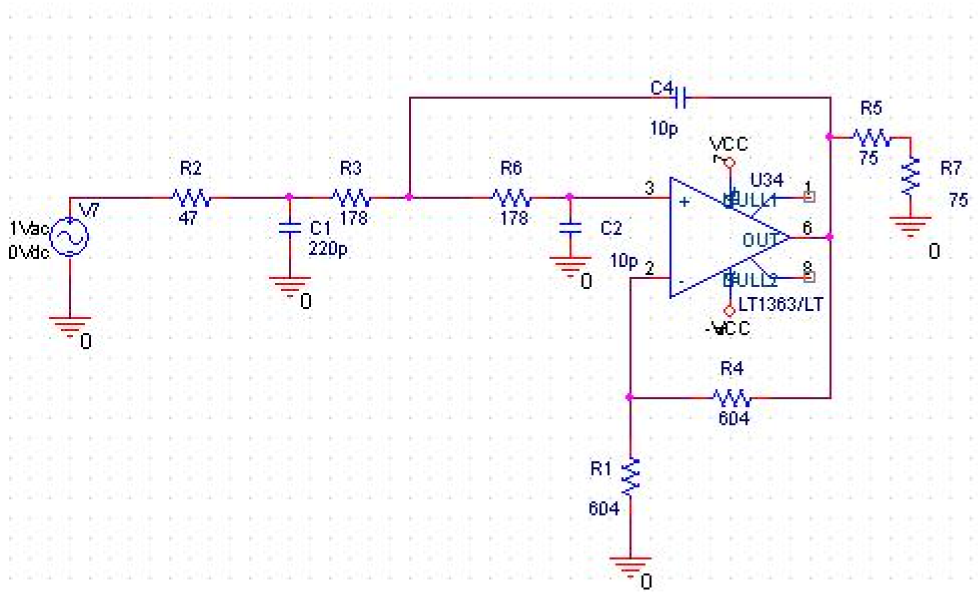


Figure 4.12 3-Pole 20MHz Butterworth Filter for XGA Reconstruction without DAC correction

Output response of this filter is shown in figure 4.13.

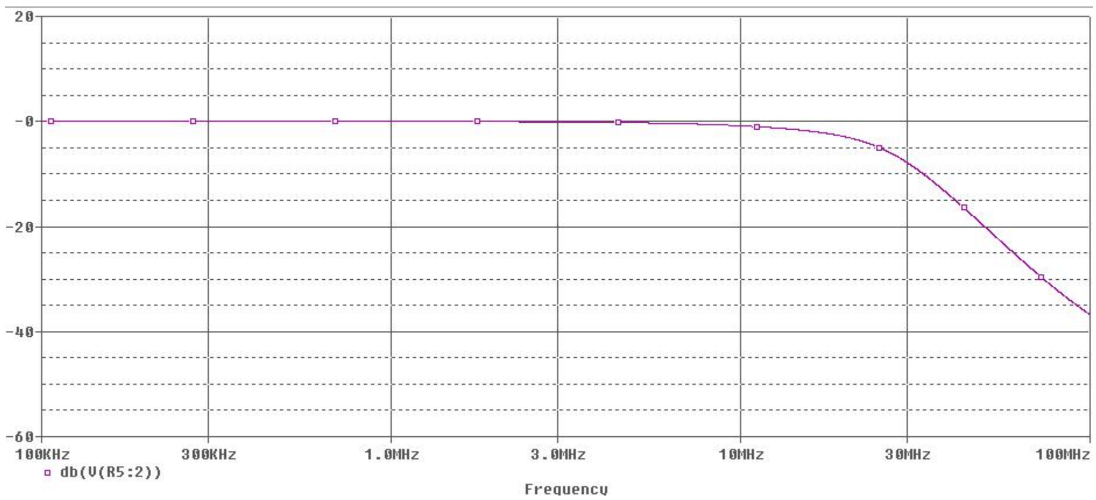


Figure 4.13 3-Pole 20MHz Butterworth Filter for XGA Reconstruction Output Response

CHAPTER FIVE
DELAY FILTERS

5.1 What is a Delay

Delay is one of the basic and natural attributes of signal processing. A signal cannot appear at the output of a system earlier than it is applied at the input. Delay occurs naturally in the transmission of signals through a system. (Schaumann, R. & Valkenburg, 2001)

We are interested in delay filters that approximate the operation of the filter to a pure delay without changing the magnitude or shape of the input signal.

5.2 Time Delay

Delay is a time domain quantity but filters are designed from frequency domain specifications such as magnitude and phase. Let $V_1(t)$ applied to a circuit that provides a delay of D seconds.

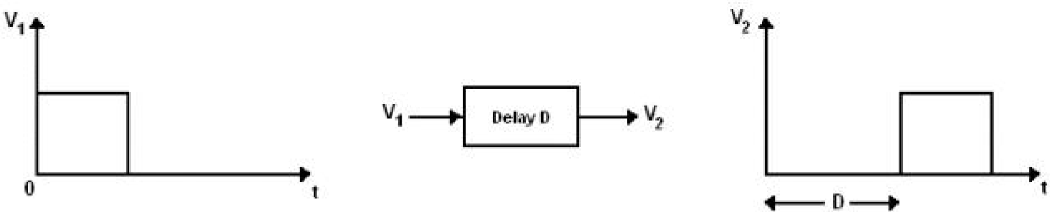


Figure 5.1 Time Delay

$$V_2(t)=V_1(t-D) \qquad \text{Eq.(5.1)}$$

$$V_2(t)=A\sin[w(t-D)+ø] \qquad \text{Eq.(5.2)}$$

$$V_2(t)=A\sin[wt-wD+ø] \qquad \text{Eq.(5.3)}$$

The input and output signals differ only by a phase angle:

$$\Theta = -wD \qquad \text{Eq.(5.4)}$$

If all Fourier components of the input signal are delayed by the same amount D and not changed in amplitude, then the output will indeed be a delayed replica of the input. Let:

$$V1 = A\angle\, ø \text{ and } V2 = A\angle(\, ø - wD) \qquad \text{Eq.(5.5)}$$

Then the transfer function is:

$$V_2/V_1 = 1\angle(-wD) \qquad \text{Eq.(5.6)}$$

or in exponential form:

$$T(jw) = \frac{V_2(jw)}{V_1(jw)} = 1e^{-jwD} \qquad \text{Eq.(5.7)}$$

Delay in time can be defined as the negative derivative of the phase:

$$D(w) = -\frac{d\theta(w)}{dw} \qquad \text{Eq.(5.8)}$$

This is also known as group delay, signal delay or envelope delay. (Schaumann, R. & Valkenburg, 2001)

In an ideal delay filter, the phase is linear with respect to the frequency and has a negative slope. The magnitude and delay is constant. Under such circumstances a signal will be delayed without distortion.

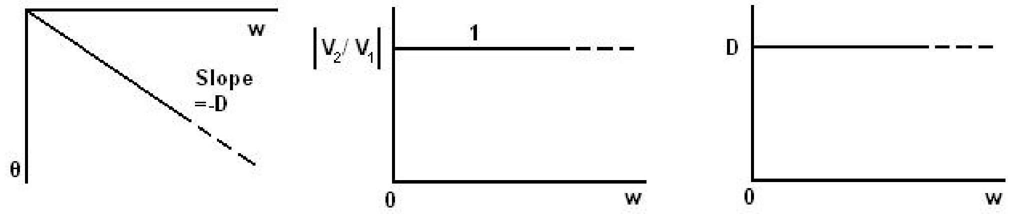


Figure 5.2 Definition of the operation of an ideal delay filter

5.3 Comparison of Filter Responses

Bessel-Thomson response results from imposing the requirements that

$$\left.\frac{dDn(w)}{dw}\right|_{w=0} = 0 \text{ and } D_n(0) = 1 \qquad \text{Eq.(5.9)}$$

and that the delay be maximally flat at w=0. In designing Bessel-Thomson filters, we merely accept the magnitude response that resulted and found that $|T_n(jw)|=1$ only at low frequencies. In contrast, the Butterworth, Chebyshev and Cauer responses were derived without reference to delay. $D_n(w)$ can be found for any transfer function by simply determining the phase function θ(w) and then differentiating with respect to w.

The Butterworth response is derived from the requirement that the magnitude function be maximally flat for small w. If we determine the delay characteristics, the result is as shown in Figure 5.3 for n=2 to n=10. We see that the delay has a peak value, especially for larger values of n, at the normalized frequency w=1. This may be contrasted to the Bessel-Thomson response shown in Fig.5.4, which achieves the maximally flat response form. The reason for the peaking delay of the Butterworth response is found in the higher pole quality factors resulting in sharper cut off and less linear phase.

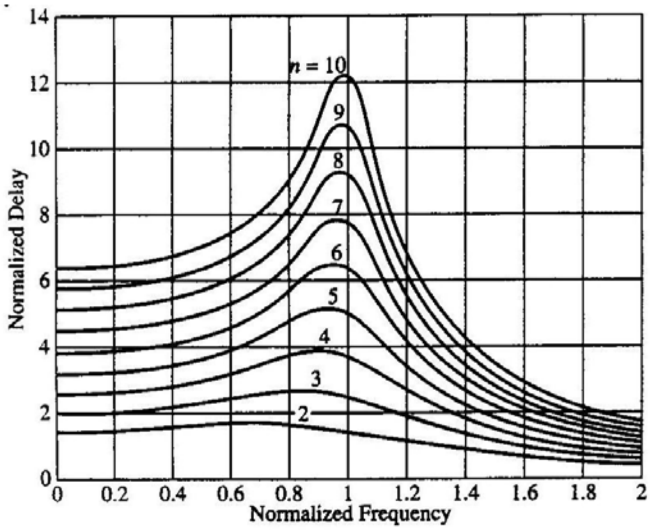


Figure 5.3 Delay Characteristic of Butterworth Filter of orders 2 through 10

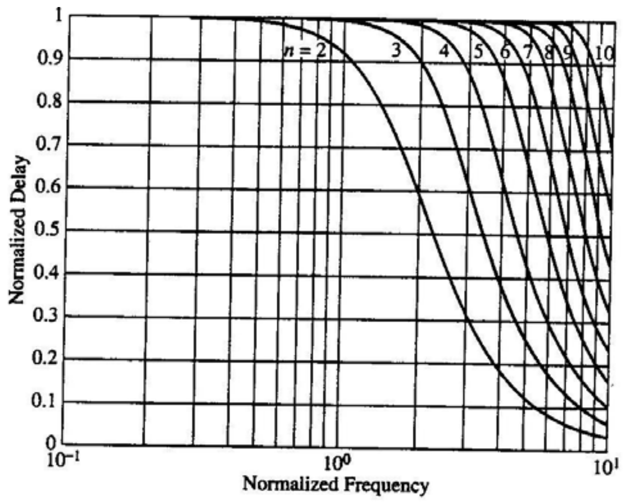


Figure 5.4 Delay of Bessel-Thomson filters of orders 2 through 10

The delay response for Chebyshev filters is more complicated. They were derived from the requirement of equal-ripple attenuation in the passband, and this resulted in a phase response that was much less linear than the phase of a filter with maximally flat magnitude. In the Chebyshev case, the transfer functions are determined by the ripple width α_{max} as well as by their order n, and so are the delays. Figure 5.5 shows the delay characteristic of the Chebyshev responses for n=3, 5, 7, and 9. In figure 5.5 (a) and (b) Chebyshev delay for 0.5dB and 2dB ripple can be seen respectively, and notice that the delay gets worse, i.e., less constant, not only with increasing degree but also that larger ripple results in worse delay performance.

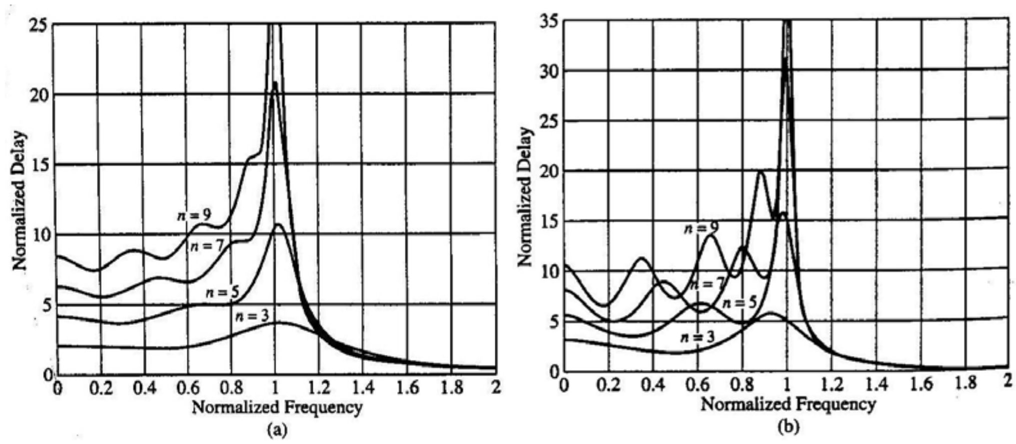


Figure 5.5 Delay of Chebyshev filters of orders 5, 7 and 9 (a) 0.5-dB ripple; (b) 2dB ripple

The reason is the larger quality factor: Q increases with increasing ripple. Chebyshev delay responses will likely cause considerable delay distortion. But in making this statement we are again cautioned that the degree of the total filter must be kept in mind. (Schaumann, R. & Valkenburg, 2001)

If we simulate the phase and delay response of the 3-pole Butterworth PAL reconstruction filter in Figure 4.10, the phase and delay response of the circuit will be as in figures 5.6 and 5.7 respectively.

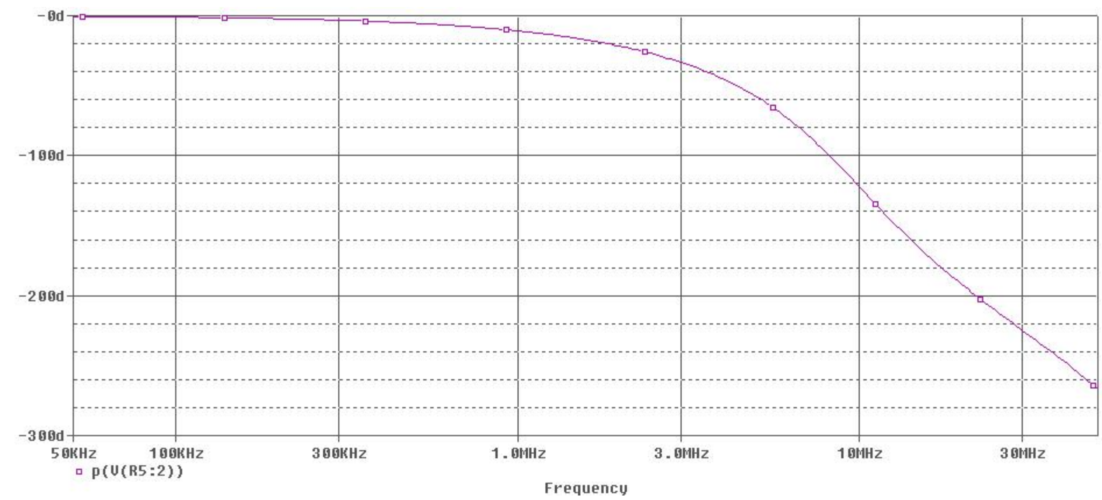


Figure 5.6 Phase Response of Butterworth Filter of Order 3 based on MAX4450

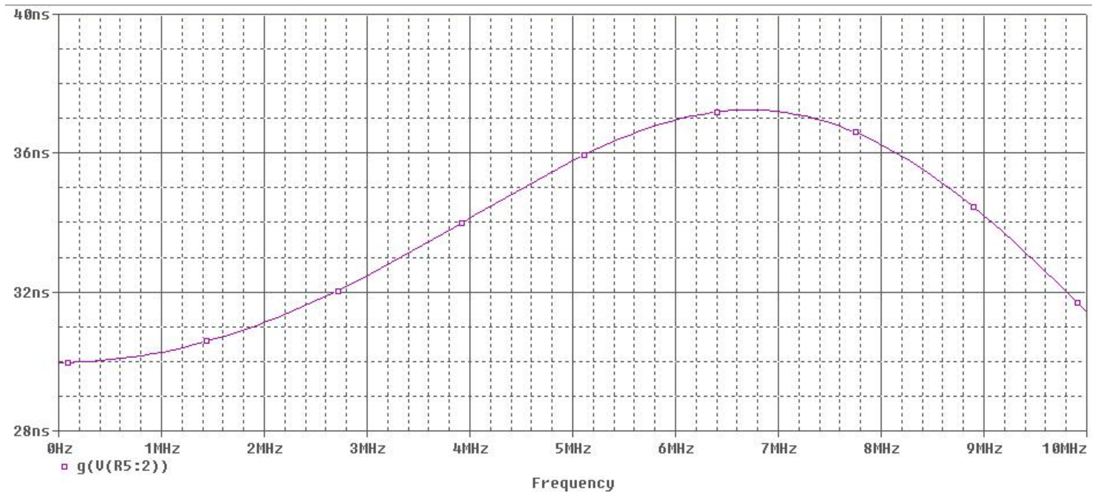


Figure 5.7 Delay Response of Butterworth Filter of Order 3 based on MAX4450

If we simulate the 4-pole 20MHz Butterworth XGA antialiasing filter in figure 4.2, the phase and delay response of the circuit will be as in figures 5.8 and 5.9 respectively.

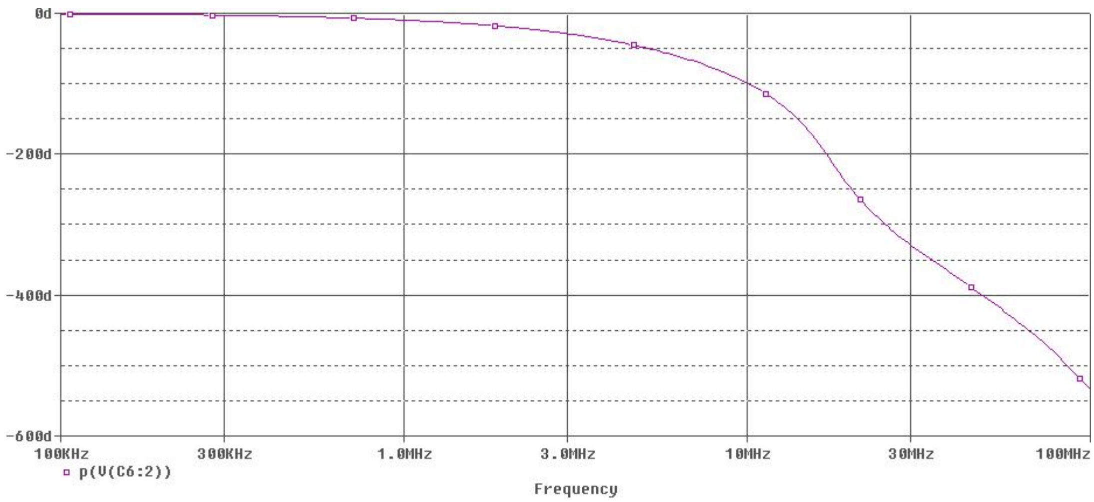


Figure 5.8 Phase Response of 4 Pole 20MHz Butterworth Filter XGA Antialiasing Filter

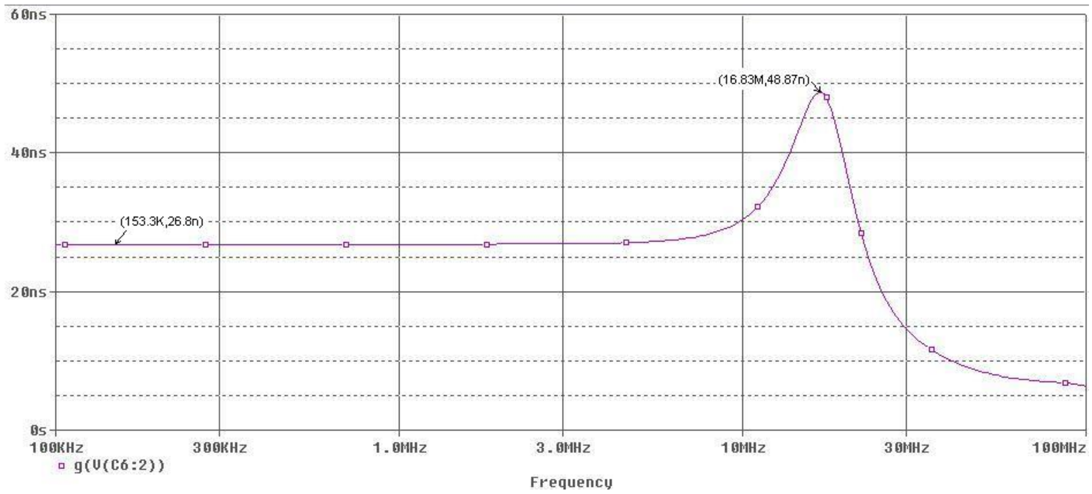


Figure 5.9 Delay Response of 4 Pole 20MHz Butterworth Filter XGA Antialiasing Filter

If we simulate the 5-pole 8.45MHz Butterworth antialiasing filter in figure 4.5, the phase and delay response of the circuit will be as in figures 5.10 and 5.11 respectively.

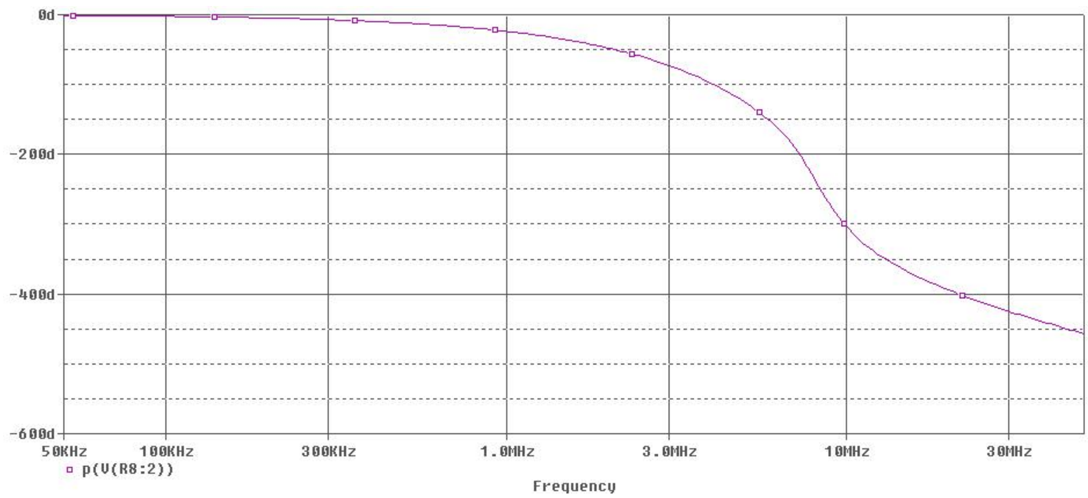


Figure 5.10 Phase Response of 5 Pole Butterworth Filter based on MAX4451

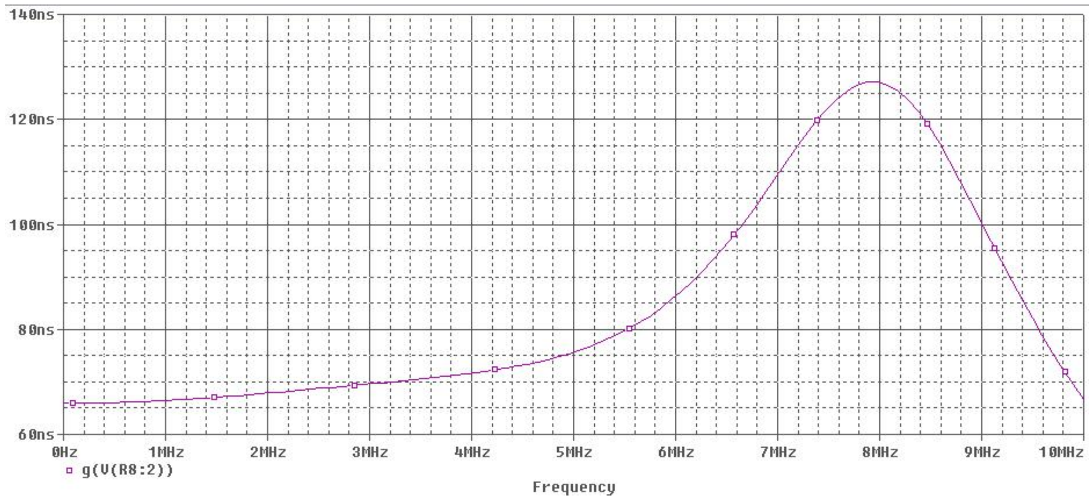


Figure 5.11 Delay Response of 5 Pole Butterworth Filter based on MAX4451

For filtering applications, if substantial stopband attenuation were required coupled with good delay performance, a delay filter would be necessary.

5.4 Approximating an Ideal Delay Function

With the goal of obtaining a filter that realizes the ideal magnitude $|T(jw)|=1$ and the ideal delay D=1, lets start the approximation from a low pass filter in Eq.(5.10).

$$T(s) = \frac{K}{D(s)} \qquad \text{Eq.(5.10)}$$

where K is a constant and the polynomial D(s) is designed to let T(s) approximate a constant delay in either a maximally flat or an equiripple format. The frequency range over which the delay is constant within some specified error depends on the degree n of the approximation. Since the function T(s) in Eq.(5.10) is a lowpass function, it is clear that errors in magnitude from the ideal $|T(jw)|$ will grow as frequencies increase. In many applications this is acceptable or even desirable because the lowpass behavior of the delay function helps to eliminate unwanted high frequency noise. However, since high frequency components of the signal are attenuated as well, we must expect that sharp corners of signals, v(t) are eliminated or at least reduced. If the signal shape must be preserved, it may be desirable to be able to delay the signal without limiting its frequency band. Let us discuss how a circuit can be designed that provides a pure delay. Let us first start from the general expression T(s)=N(s)/D(s).

$$T(jw) = \frac{N(jw)}{D(jw)} = \frac{|N(jw)|\,e^{j\theta_N(w)}}{|D(jw)|\,e^{j\theta_D(W)}} = \frac{|N(jw)|}{|D(jw)|}e^{j[\theta_N(w)-\theta_D(w)]} \qquad \text{Eq.(5.11)}$$

Using Eq.(5.11), the delay is then

$$D(w) = -\frac{d\left[\theta_N(w) - \theta_D(w)\right]}{dw} \qquad \text{Eq.(5.12)}$$

and the magnitude,

$$\frac{|N(jw)|}{|D(jw)|} = \left|\frac{N(jw)}{D(jw)}\right| \qquad \text{Eq.(5.13)}$$

which should be equal to a frequency independent constant K, preferably K=1. To make the ratio of the magnitudes of the two complex numbers N(jw) and D(jw) equal to unity for all frequencies, N(jw) and D(jw) must be conjugate complex, that is N(jw)=D(-jw) or N(s)=D(-s). The transfer function is then

$$T(s) = \frac{D(-s)}{D(s)} \qquad \text{Eq.(5.14)}$$

They are labeled as allpass functions because with $|T(jw)|=1$ they pass all frequency components of a signal equally well. The phase angles of conjugate complex numbers are equal in magnitude but opposite in sign. Consequently, from Eqs.(5.11) and (5.12) the phase angle and delay of an all-pass filter are

$$\theta_{AP}(w) = \theta_N(w) - \theta_D(w) = -2\theta_D(w) \qquad \text{Eq.(5.15)}$$

and

$$D_{AP}(w) = \frac{-2d\theta D(w)}{dw} \qquad \text{Eq.(5.16)}$$

In other words, phase and delay have twice the value that is contributed by the denominator of the allpass function alone. We should note that the allpass filter has twice the delay of an all pole lowpass filter of equal degree. (Schaumann, R. & Valkenburg, 2001)

5.5 Delay Equalization

In designing all pass filters we define an error in percent by which the filter's delay deviated from the ideal D=1. Since D is a function of frequency, we need to specify at which frequency w_1 the maximally acceptable delay error occurs. (Schaumann, R. & Valkenburg, 2001)

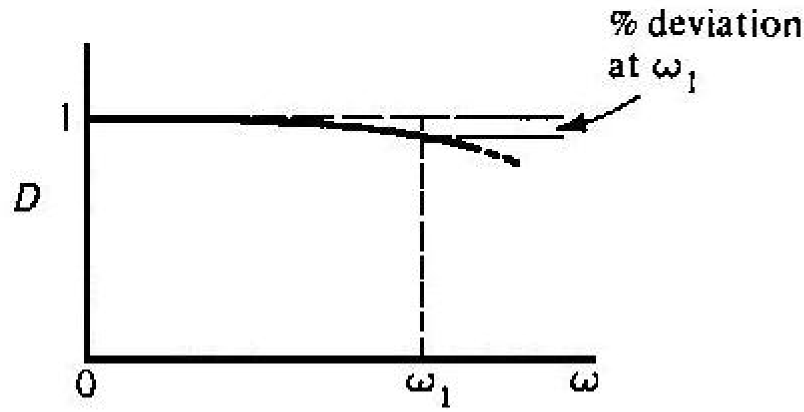


Figure 5.12 Definition of Delay Error

There are applications in which not only a required magnitude response must be realized but a constant delay is necessary as well. One example is a signal that has sharp transitions in the time domain, such as a train of pulses. As we know from Fourier analysis, such a signal contains many components of different magnitudes and phases over a band of frequencies. To reconstitute a pulse with little or no distortion at the filter output the delay must be constant over the frequency band, otherwise severe distortion in the shape of the pulse will result. We can design a filter that has a specified magnitude, such as a maximally flat, and that at the same time provides a constant delay over the filter's passband. The process is referred to as delay equalization. (Schaumann, R. & Valkenburg, 2001)

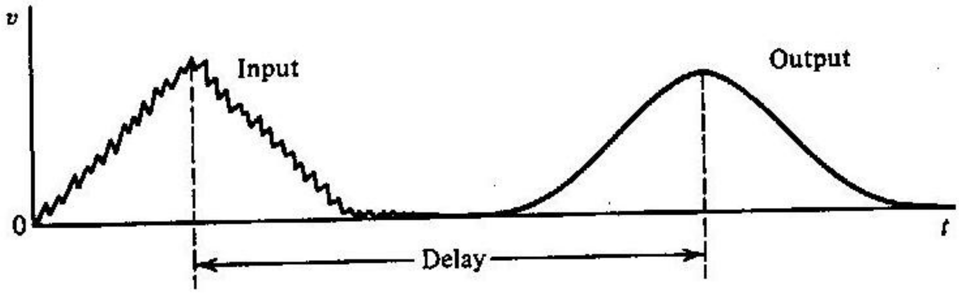


Figure 5.13 Desired filtering process of a signal in the time domain

As an example of an application is the signal-filtering requirement in Fig.(5.13) where a triangular signal is corrupted by some high frequency "noise" that must be removed. To eliminate the noise, a lowpass filter is required, for example with a fifth order Butterworth response that passes all important frequency components that make up the triangular signal in the time domain.

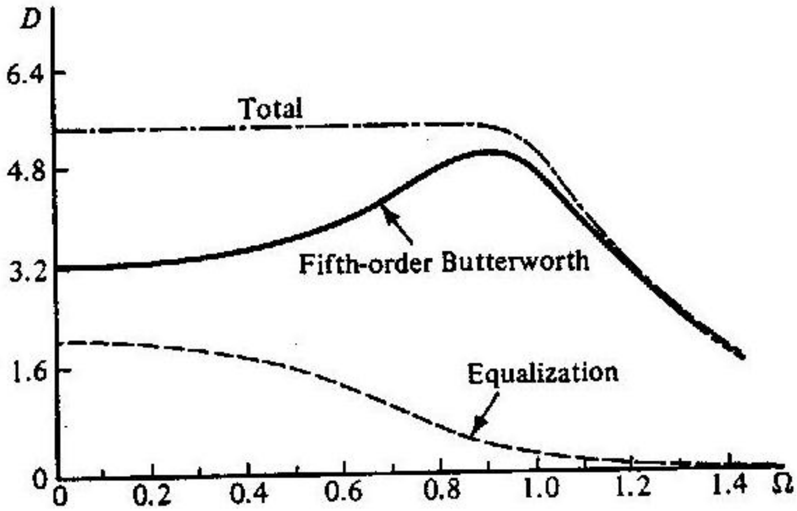


Figure 5.14 Delay of a fifty order Butterworth filter, desired total delay, and necessary delay to be added to that of the Butterworth filter for equalization

But the filter must not distort the components relative phases so that the signal shape is not changed. Thus, as is shown in Fig.(5.13) at the output we clearly recognize the shape of the input signal; it is essentially undisturbed, although sharp corners, such as the peak, are rounded because the lowpass filter removed the highest frequencies. Figure (5.14) shows the delay requirement and the correction needed. (Schaumann, R. & Valkenburg, 2001)

5.5.1 Equalization Procedures

Designing a filter for a certain magnitude response gave us a circuit in which we had no control over the phase, because after choosing the type of the magnitude response, such as Chebyshev or Butterworth, all poles and zeros are determined, and there remains no degree of freedom to meet any additional filter specifications. Our only option is to select an approximation that gave the "least objectionable" phase or delay. We already saw that choosing a filter with low Q values reduces delay variations.

If the obtained delay varies too much for the signal-processing requirement at hand, we can only modify the transfer function. But we must do this in such a way that the filter's magnitude response is not destroyed. A solution to the problem is to connect the filter T_M with desired magnitude response, $|T_M(jw)|$, in cascade with a circuit T_{AP} whose magnitude response is equal to unity, $|T_{AP}(jw)|=1$, but whose phase varies with frequency. (Schaumann, R. & Valkenburg, 2001)

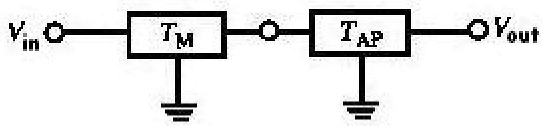


Figure 5.15 Cascade connection of a filter TM(s) with the desired magnitude response and an allpass filter TAP(s) to equalize the dalay.

The transfer function of the cascade configuration in Fig.(5.15) is T(s)=TM(s)TAP(s), or on the jw axis:

$$T(jw) = T_M(jw) \cdot T_{AP}(jw)$$
$$= |T_M(jw)||T_{AP}(jw)|e^{-j\theta_M(w)}e^{-j\theta_{AP}(w)} = |T_M(jw)|e^{-j[\theta_M(w) - j\theta_{AP}(w)]} \qquad \text{Eq.(5.17)}$$

From the Eq.(5.17) we see that the magnitudes multiply with no contribution from the allpass module because $|T_{AP}(jw)|=1$, and that the phases add. Since the delay is obtained from the negative derivative of the phase, the delays D_M and D_{AP} add as well:

$$D(w) = -\frac{d[\theta_M(w) + \theta_{AP}(w)]}{dw} = -\frac{\theta_M(w)}{dw} - \frac{\theta_{AP}(w)}{dw} = D_M(w) + D_{AP}(w) \qquad \text{Eq.(5.18)}$$

Note that the total delay D in this process will increase because D is larger than either D_M or D_{AP}. If $D(w)=D_0$ is to be constant over some specified frequency range, and $D_M(w)$ is known from the transfer function that provides the desired magnitude response, we must find an allpass function to provide:

$$D_{AP}(w) = D_0 - D_M(w) \qquad \text{Eq.(5.19)}$$

This equation can, in general, not be satisfied exactly for all frequencies, but can only be approximated for frequencies in a finite interval. The wider the frequency range of interest and the smaller the permitted delay error is, the more expensive the solution becomes because it requires allpass filters of increasingly high order. (Schaumann, R. & Valkenburg, 2001)

$$T_{AP}(s) = \frac{N(s)}{D(s)} = \frac{D(-s)}{D(s)} \qquad \text{Eq.(5.20)}$$

To provide a magnitude that is equal to unity for all frequencies, as we saw in Eq.(5.21), we need a transfer function in which the numerator is obtained from the denominator by replacing s by –s. The pole zero patterns of such allpass functions for n=1 to 4 are shown in Fig. 5.16. The pattern in Fig. 5.16a belongs to a first order allpass with one pole and one zero at equal distances from the origin. That shown in Fig. 5.16b for a second order allpass has two poles and two zeros in quadrantal symmetry. Fig. 5.16c for a third order allpass can be thought of as the superposition of the previous two pattern; the fourth order allpass pattern in Fig. 5.16d follows from two second order ones. (Schaumann, R. & Valkenburg, 2001)

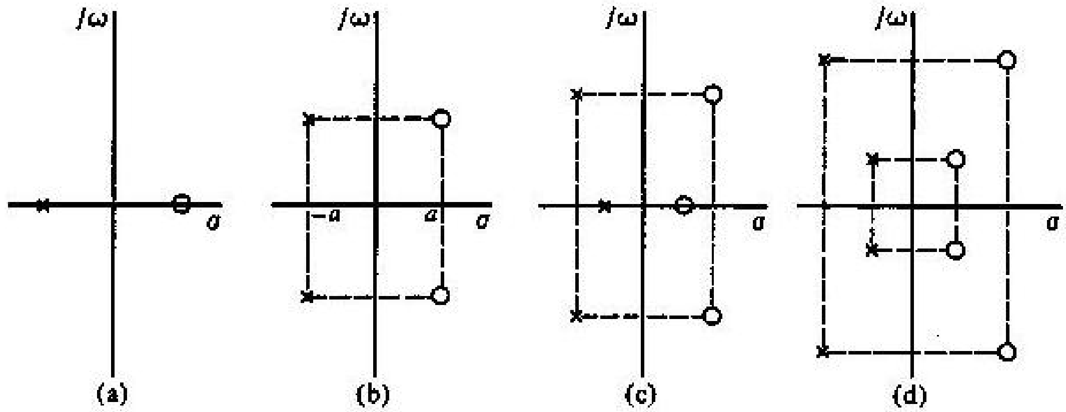


Figure 5.16 Typical pole-zero patterns of allpass filters of orders 1, 2, 3 and 4

The design of a filter with constant delay (over a limited frequency range) is accomplished in two steps:

1. Design the filter with the required magnitude response $|T_M(jw)|$.
2. Calculate the delay $D_M(w)$ that the filter provides and supplement it by the delays of cascaded first or second order allpass modules such that the total delay, the sum of $D_M(w)$ and $D_{AP}(w)$, is approximately constant over the prescribed frequency range.

5.5.2 Equalization with First Order Modules

Consider the first order allpass function

$$T_1(s) = \frac{K(s - \sigma_1)}{s + \sigma_1} \qquad \text{Eq.(5.21)}$$

corresponding to the poles zero pattern in Fig. 5.16a Its magnitude and phase are

$$|T_1(jw)| = K \quad \text{and} \quad \theta_1 = \theta_{numeratior} - \theta_{deno\min ator} = -2\tan^{-1}(\frac{w}{\sigma_1}) \qquad \text{Eq.(5.22)}$$

Observe that in an allpass numerator and denominator have the same phase so that the allpass phase is simply twice the denominator phase. The delay is computed as

$$D_1(w) = -\frac{d\theta}{dw} = \frac{\frac{2}{\sigma_1}}{1+\left(\frac{w}{\sigma_1}\right)^2} \qquad \text{Eq.(5.23)}$$

This function is plotted in Fig. 5.17; we note that the delay decreases with increasing frequency. Its maximum is at w=0 and has the value

$$D_{1,\max} = D_1(0) = \frac{2}{\sigma_1} \qquad \text{Eq.(5.24)}$$

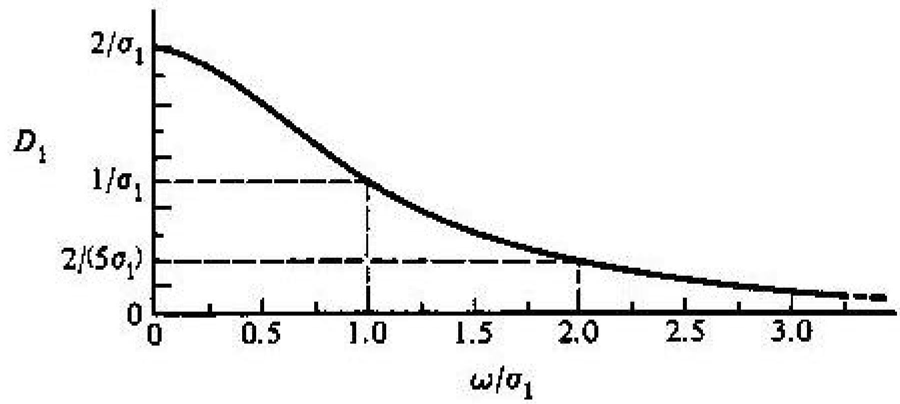


Figure 5.17 Delay of a first-order allpass filter

CHAPTER SIX
ALL PASS FILTERS WITH DIFFERENT ACTIVE BLOCKS

All pass filters are one of the most important building blocks of many analog signal-processing applications. They are generally used for introducing a frequency dependent delay while keeping the amplitude of the input signal constant over the desired frequency range. In majority of the active filter designs the conventional operational amplifiers are used. For design purposes generally the op amp is assumed to be ideal and op amp parameters are taken as $A_V = \infty$, $R_{in} = \infty$, $R_O = 0$. Also large amount of feedback are used to make the filter gain essentially independent of the gain of the op amp. In practice, op amps have some limitations and drawbacks such as GBW limitation, limited maximum Quality Factor, Slew Rate problem. So new active blocks were introduced.

The active devices that have been used in literature for the realizations of first order all pass filters include:

•Operational Amplifier (OP-AMP)
•Current Conveyor (CCII)
•Operational Transconductance Amplifier (OTA)
•Operational Transresistance Amplifier (OTRA)
•Four-terminal floating nullor (FTFN)

6.1 All-pass Filter Using Single Operational Amplifier (OP-AMP)

For design purposes, the assumption that the op amp is ideal ($A_v=\infty$, $R_{in}=\infty$, $R_o=0$) is generally made, and large amounts of feedback are used to make the filter gain essentially independent of the gain of the op amp. Unfortunately operational amplifier limitations preclude the use of these filters at high frequencies. However, evolution of integrated transresistance, transconductance and current amplifiers for better filter characteristics has not kept pace with that of the operational amplifiers, although a few devices in these alternate categories are commercially available.

CA3080 and LM13600 as transconductance amplifiers and LM3900 as transresistance amplifier can be given as example. (Geiger, R.L. & Sinencio, S., 1985)

For a first order op amp based all pass filter the circuit in figure 6.1 can be given as an example.

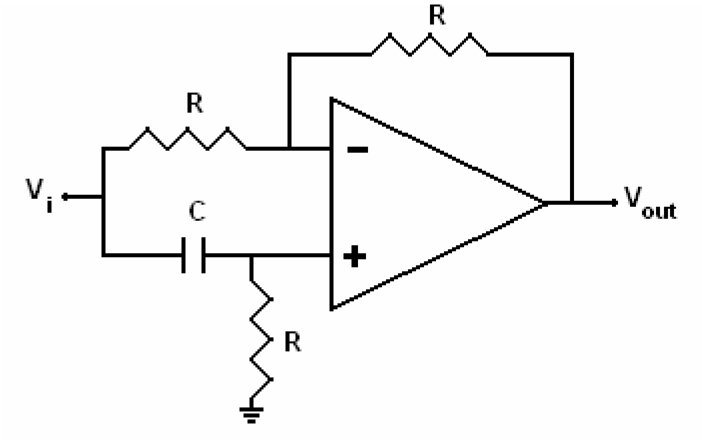


Figure 6.1 First Order OP AMP All Pass Filter

The transfer function of the filter can be given as:

$$\frac{V_0}{V_i} = \frac{\left[s - \frac{1}{RC}\right]}{\left[s + \frac{1}{RC}\right]} \qquad \text{Eq.(6.1)}$$

From the Eq.(6.1), we can find the phase equality as:

$$\theta = 180^{O} - 2\arctan(wRC) \qquad \text{Eq.(6.2)}$$

We know that the group delay is the negative derivative of the phase. From the Eq.(6.2) the group delay can be calculated as:

$$\tau = -\frac{d\theta}{dw} = \frac{2RC}{\left[1+(wRC)^2\right]} \qquad \text{Eq.(6.3)}$$

For an ideal op-amp as the active device, the gain vs frequency response will be as in figure 6.2.

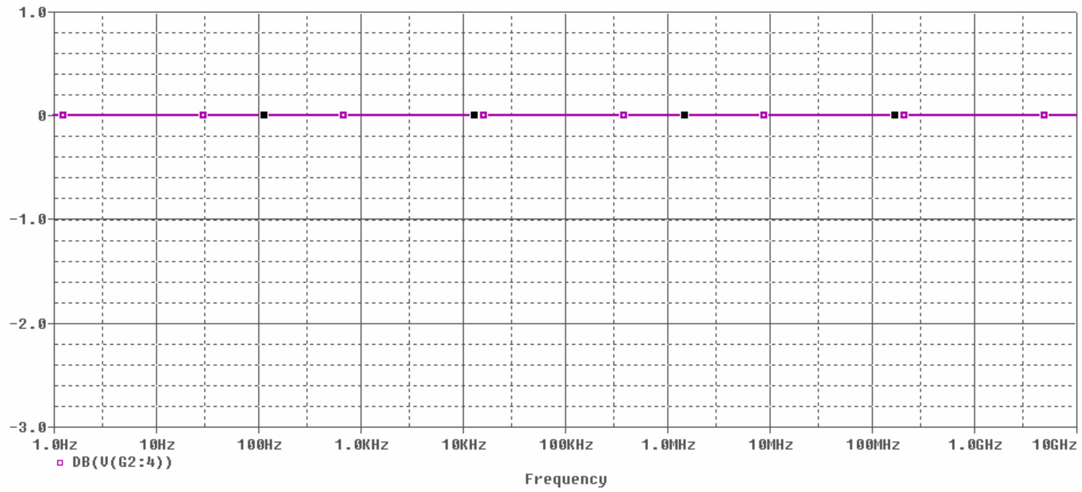


Figure 6.2 Gain Response of an All-pass Filter based on Ideal OP-AMP

If we use a commercially available LM324 as the active device, the frequency response of the filter will be as in figure 6.3. As can be seen from the result, the low GBW parameter of LM324 has effected the performance of the overall filter in high frequencies so conventional LM324 op-amp can not be used for high frequency applications.

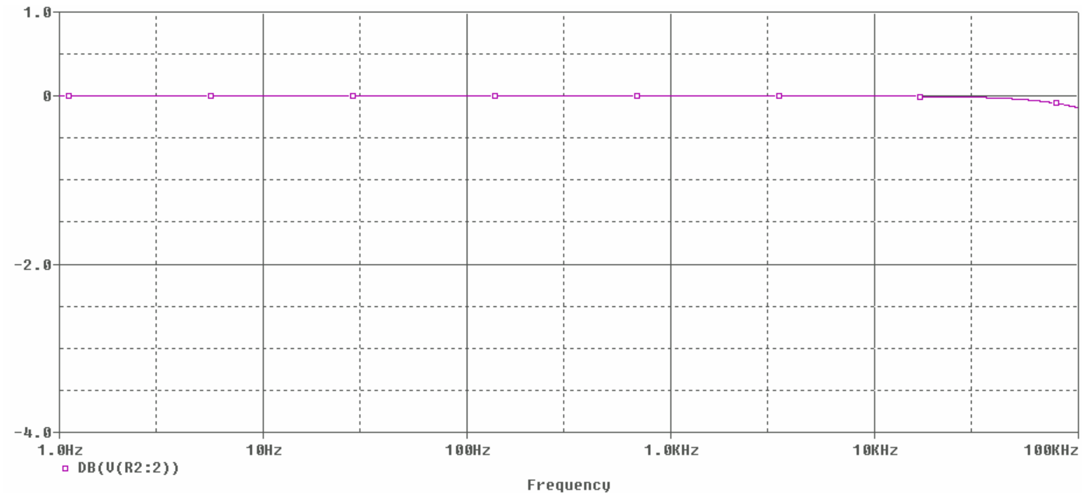


Figure 6.3 Gain Response of an All-pass Filter based on LM324

6.2 All-pass Filters Using Single Operational Transconductance Amplifiers (OTA)

From a practical viewpoint, the high-frequency performance of discrete bipolar OTAs is quite good. The transconductance gain, g_m, can be varied over several decades by adjusting an external dc bias current, I_{ABC}. The major limitation of existing OTAs is the restricted differential input voltage swing required to maintain linearity. Although feedback structures in which the sensitivity of the filter parameters are reduced, major emphasis will be placed upon those structures in which the standard filter parameters of interest are directly proportional to g_m of the OTA. Thus, the g_m will be a design parameter much as are resistors and capacitors. Since the transconductance gain of the OTA is assumed proportional to an external dc bias current, external control of the filter parameters via the bias current can be obtained. (Geiger, R.L. & Sinencio, S., 1985)

6.2.1 OTA Model

OTA is a transconductance type device, which means that the input voltage controls an output current by means of the device transconductance g_m. This makes the OTA a voltage-controlled current source (VCCS). Transconductance gain, g_m, is programmed by an external amplifier bias current. Transconductance is assumed to be proportional to I_{ABC}.

$$g_m = hI_{ABC} \qquad \text{Eq.(6.4)}$$

The proportionality constant h is dependent upon temperature, device geometry, and the process.

The symbol and ideal small signal equivalent circuit for OTA is shown in figure 6.4. Unlike op amp since the OTA is a current source, the output impedance of the device is high. In the OTA model, input and output impedances are assumed to be infinity.

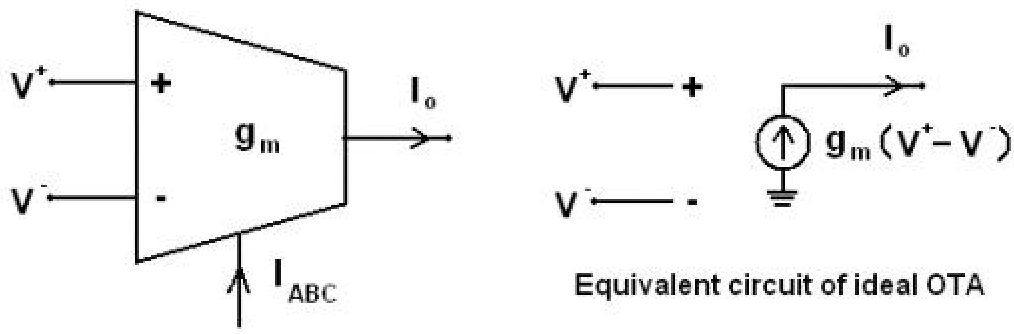


Figure 6.4 Symbol of OTA & Equivalent circuit of OTA

The output current is a function of the applied voltage difference between positive and negative terminals and the transconductance. The output current can be given as:

$$I_0 = g_m(V^+ - V^-) \qquad \text{Eq.(6.5)}$$

An output voltage can be derived from this current by simply driving a resistive load. It is possible to design circuits using the OTA that do not employ negative feedback. Instead of employing feedback to reduce the sensitivity of a circuit's performance to device parameters, the transconductance is treated as a design parameter, much as resistors and capacitors are treated in op-amp based circuits.

6.2.2 OTA All-pass Filter

Considering the all pass filters that has been introduced in Geiger R. L., Sanchez-Sinencio, E. (1985), the circuit in figure 6.5 behaves as a phase equalizer.

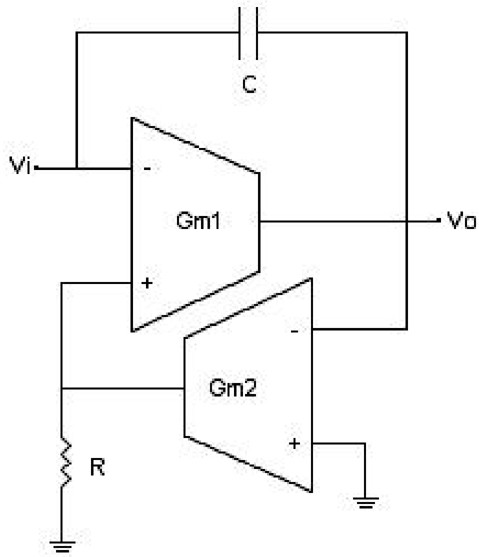


Figure 6.5 First Order All Pass Filter based on OTA

The transfer function of the all pass filter is:

$$H(s) = \frac{sC - g_{m1}}{sC + g_{m1} \times g_{m2} \times R} \qquad \text{Eq.(6.6)}$$

As can be seen from the transfer function, the all pass filter requirement for the OTA filter is $R = 1/g_{m2}$ and g_{m1} can be used to adjust the phase shift.

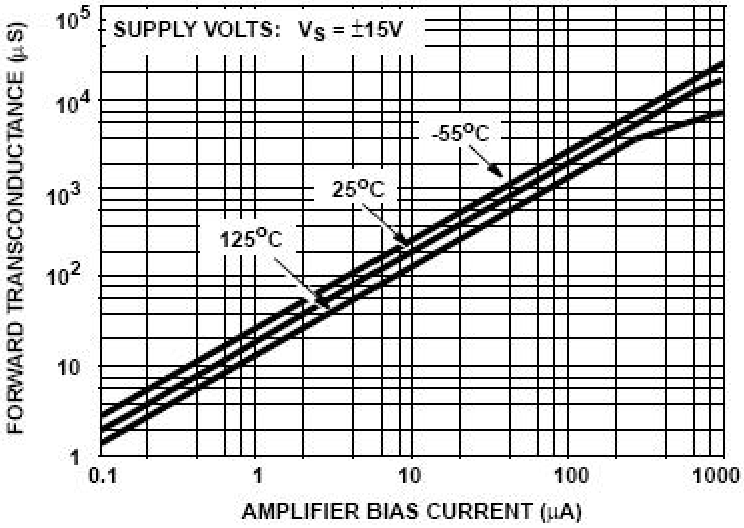


Figure 6.6 Transconductance vs Amplifier Bias Current of CA3080

Figure 6.6 shows the transconductance vs amplifier bias current of CA3080. By adjusting the bias current, the all pass filter requirement, $g_{m2} = 1/R$, can be obtained.

For monolithic applications, the resistor R can be replaced with a third OTA in figure 6.7. (Geiger, R.L. & Sinencio, S., 1985)

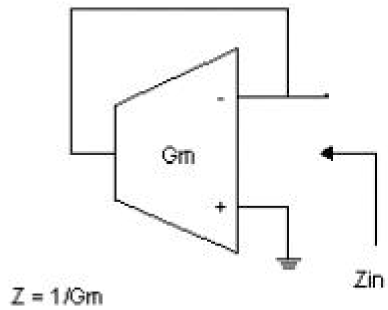


Figure 6.7 Load Application with OTA

The phase characteristic for the circuit can be figured as in figure 6.8.

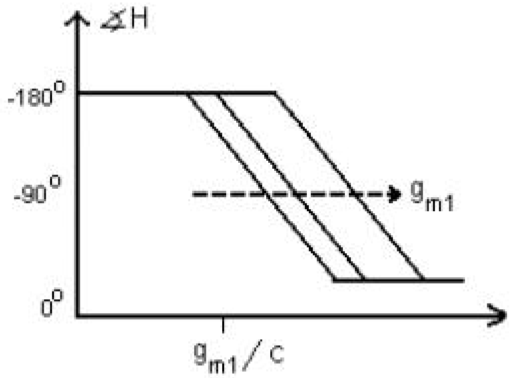


Figure 6.8 Transfer characteristic for the circuit

Figure 6.9 shows the schematics of the first order all pass filter based on ideal OTA. The gain parameter of g_{m2} is 0.01 while g_{m1} and g_{m3} are set to 0.02. The output of the ideal OTAs are terminated with 1MΩ resistors since the input impedances are infinity and output is a current source. Gain, phase and delay response of the OTA filter are shown in figures 6.10, 6.11 and 6.12 respectively.

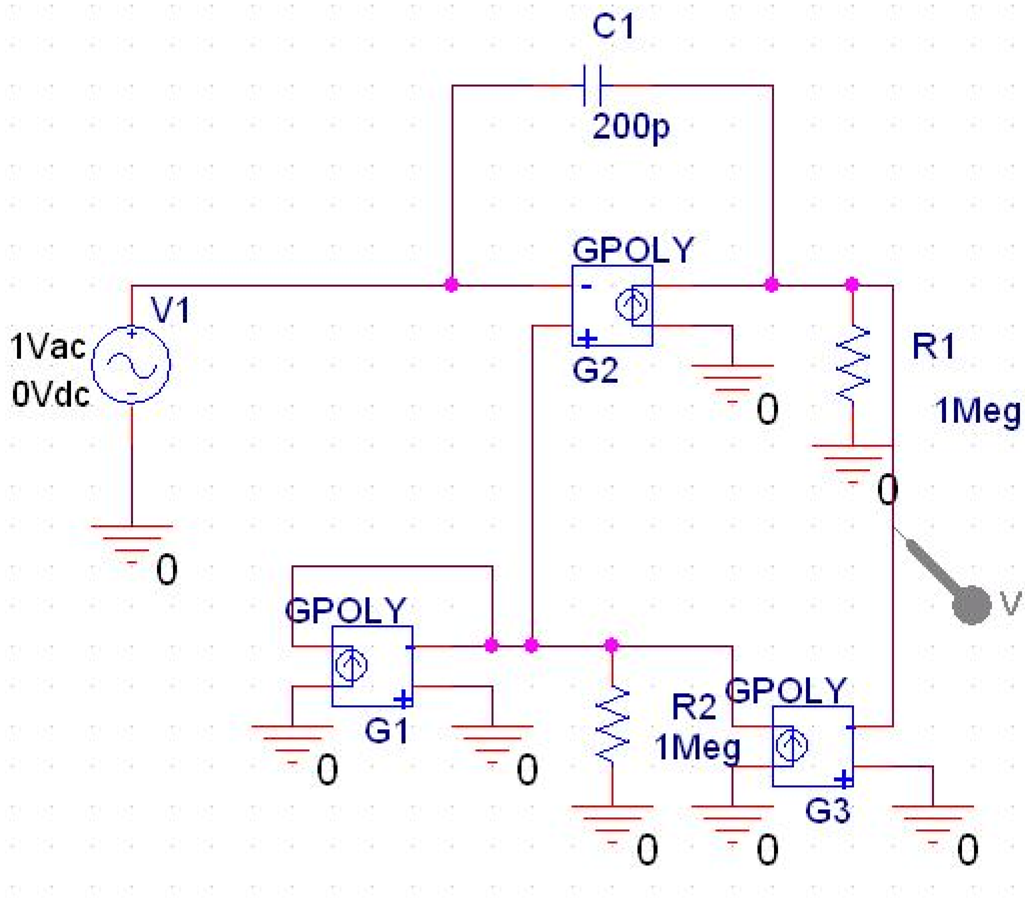


Figure 6.9 Schematics of first order ideal OTA filter

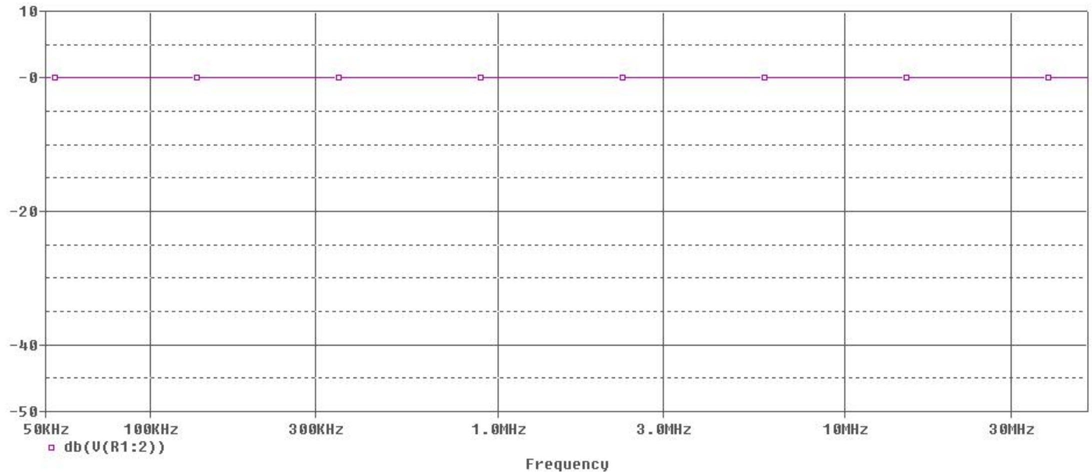


Figure 6.10 Frequency response of first order ideal OTA filter

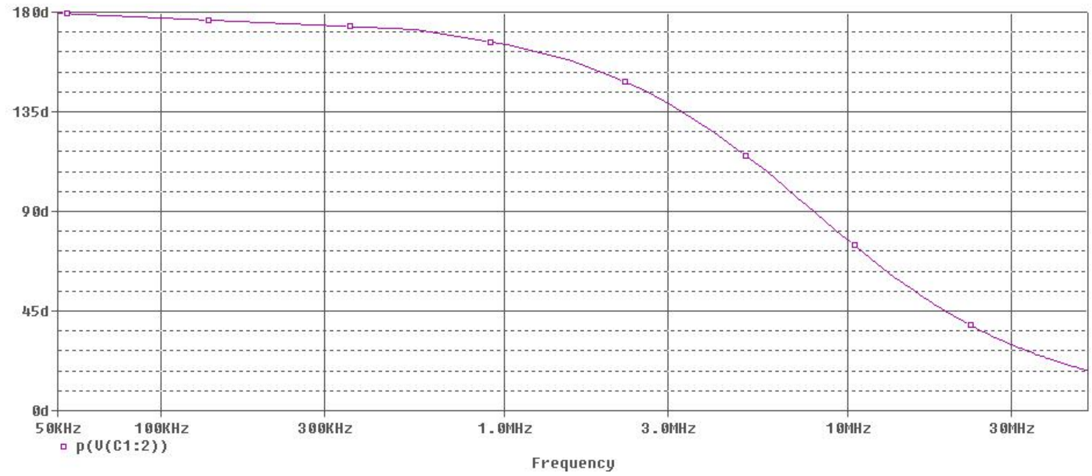


Figure 6.11 Phase characteristics of first order ideal OTA filter

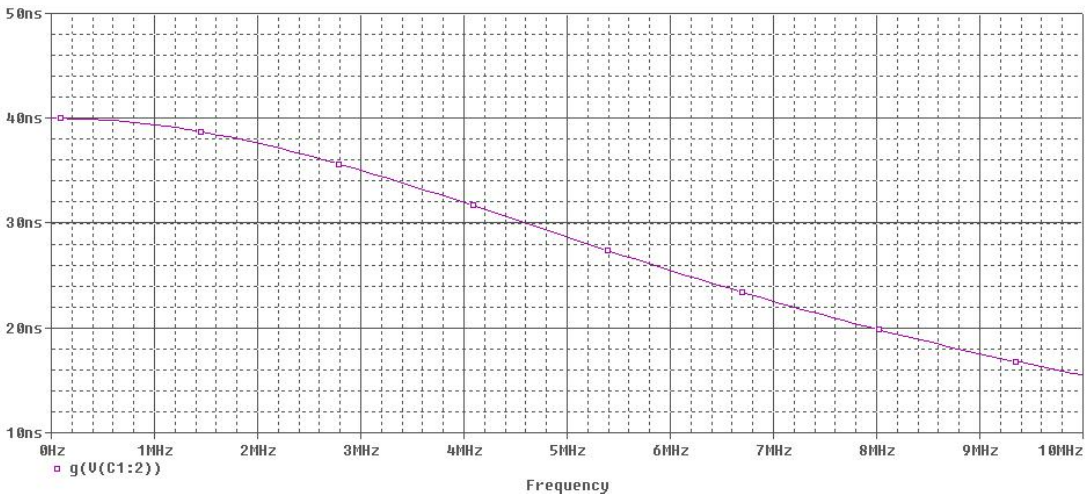


Figure 6.12 Delay characteristics of first order ideal OTA filter

Commercially available MAX435 is a high-speed wideband transconductance amplifier. The output of MAX435 is a current that is proportional to the applied differential input voltage. The circuit gain is set by the ratio of two impedances (the user selected transconductance element Z_t and the output load impendence Z_L) and an internally set current gain factor (K). K is typically equal to 4 for MAX435. Figure 6.13 shows the first order all pass filter based on MAX435.

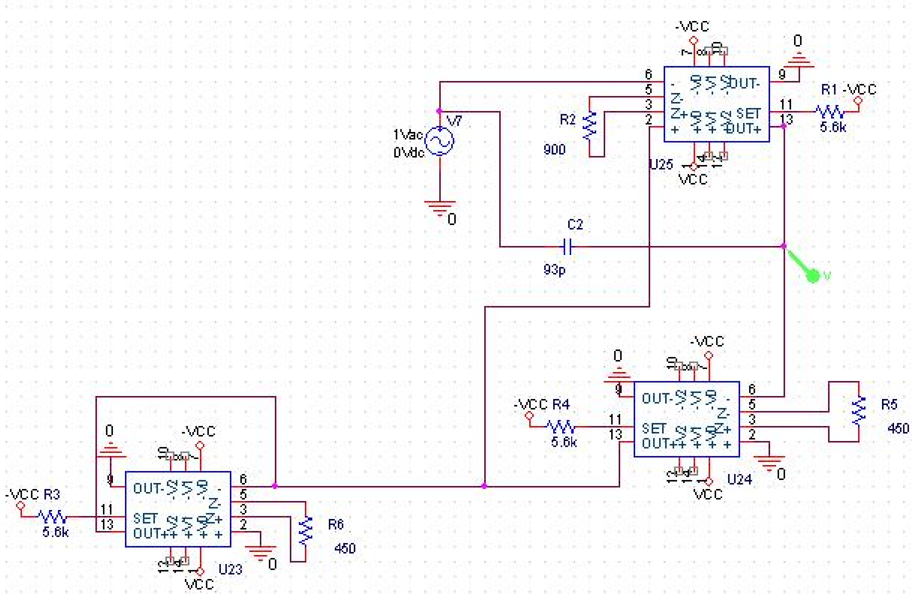


Figure 6.13 Schematics of first order OTA filter based on max435

The spice simulation of gain, phase and delay for the OTA filter are shown in figures 6.14, 6.15 and 6.16 respectively.

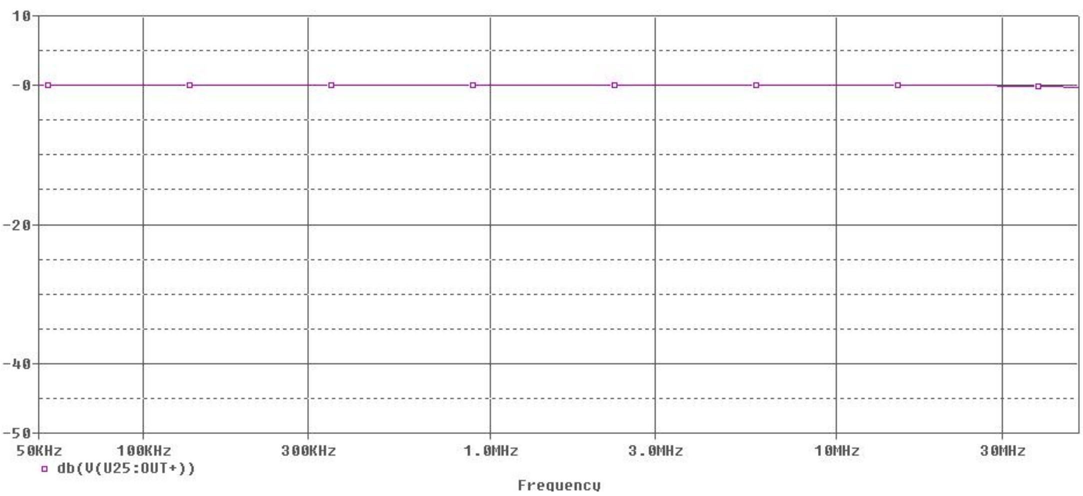


Figure 6.14 Frequency response of first order OTA filter based on max435

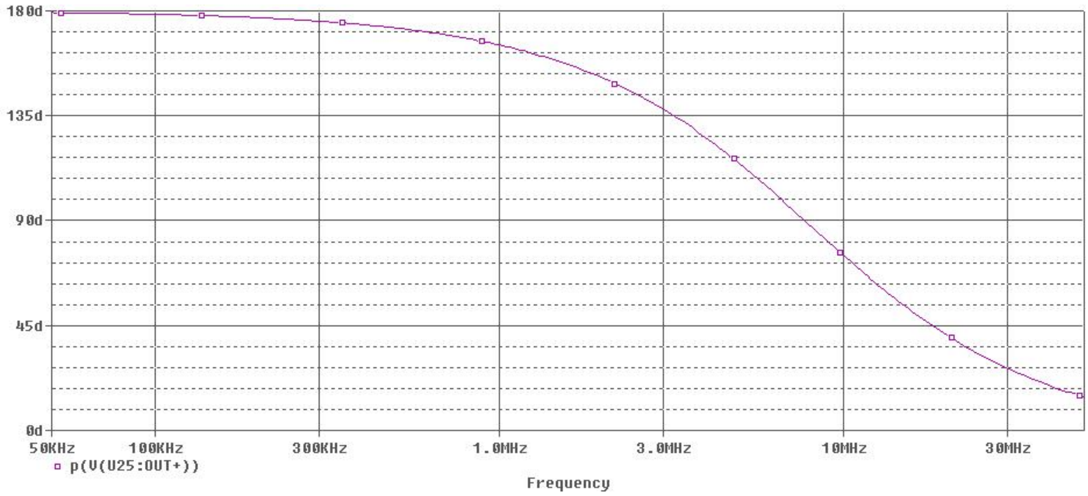


Figure 6.15 Phase response of first order all pass filter based on max435

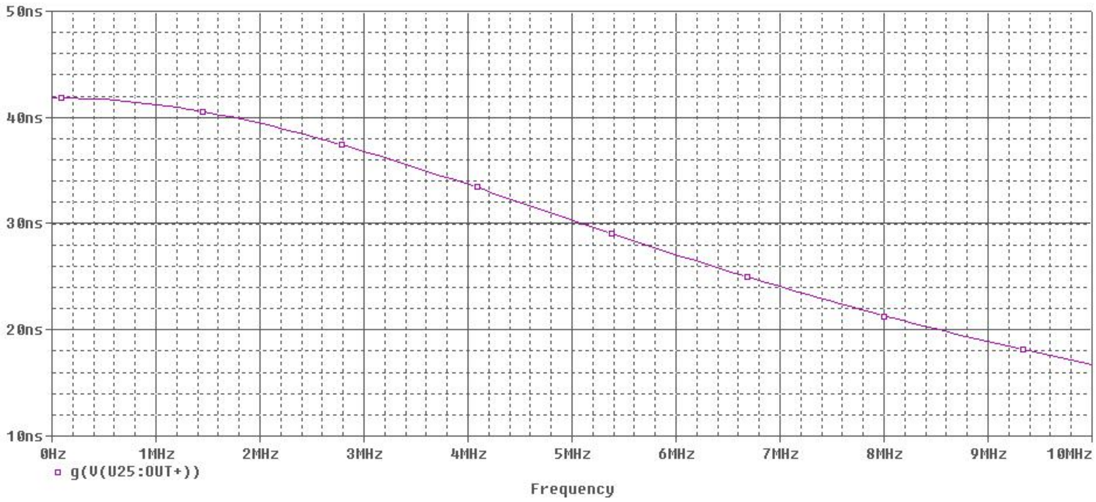


Figure 6.16 Delay of first order all pass filter based on MAX435

There are various advantages of using OTA in all pass filters:

•Reduced component count
•Design simplicity and programmability
•High frequency performance of discrete bipolar OTAs
•Transconductance gain, g_m can be varied by adjusting an external dc bias current, I_{ABC}.

The major disadvantage of existing OTAs is the restricted differential input voltage swing required to maintain linearity. Recent research results with significant improvements in the input characteristics of OTAs can be attained. For the CA3080 the differential input voltage swing is limited to $30mV_{p\text{-}p}$.

6.3 All-pass Filters Using Current Conveyors

Current conveyors are unity gain active elements exhibiting high linearity, wide dynamic range and better high frequency performance compared with their voltage mode counterparts. Since they do not suffer from the deliberately introduced low frequency dominant pole of an operational amplifier, their usable frequency range is much higher.

A current conveyor is a device which when arranged with other electronic elements in specific circuit configuration can perform many useful analog signal-processing functions. Current conveyors find application covering a broad area ranging from filter, oscillator and immittance simulator design to integrators and differentiators. (Çiçekoğlu, O. & Kuntman, H. & Berk, S., 1999)

In many ways the current conveyor simplifies circuit design in much the same manner as the conventional op-amp. This stems largely from the fact that the current conveyor offers an alternative way of abstracting complex circuit functions, thus aiding in the creation of new and useful implementations. This together with the fact that the actual terminal behavior of the current conveyor, like the opamp, approaches its ideal behavior quite closely. This, implies that one can design current conveyor circuits that work at levels that are quite close to their predicted theoretical performance. The current conveyor offers several advantages over the conventional op-amp; specifically a current conveyor circuit can provide a higher voltage gain over a larger signal bandwidth under small or large signal conditions than a corresponding op-amp circuit in effect a higher gain-bandwidth product.
(Sedra, A.S. & Roberts, G.W. (1990).

6.3.1 First Generation Current Conveyor (CCI)

Current-conveyors are three-port networks with terminals X, Y and Z as represented in figure 6.17. The network of the first generation current-conveyor CCI has been formulated in a matrix form as follows, Eq.(6.7)

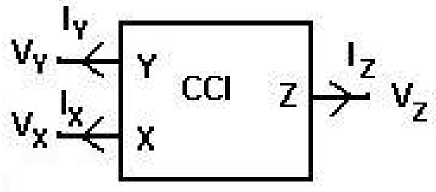


Figure 6.17 First generation current conveyor symbol and its signal definitions

$$\begin{bmatrix} i_y \\ v_x \\ i_z \end{bmatrix} = \begin{bmatrix} 0 & 1 & 0 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix} \begin{bmatrix} v_y \\ i_x \\ v_z \end{bmatrix} \qquad \text{Eq.(6.7)}$$

In other words, the first generation current conveyor CCI forces both the currents and the voltages in ports X and Y to be equal and a replica of the currents is mirrored (or conveyed) to the output port Z (Sedra, A.S. & Roberts, G.W. (1990))

6.3.2 Second Generation Current Conveyor (CCII)

In many applications, only one of the virtual grounds in terminals X and Y of the first generation current-conveyor is used and the unused terminal must be grounded or otherwise connected to a suitable potential. This grounding must be done carefully since a poorly grounded input terminal may cause unwanted negative impedance at the other input terminal. Moreover, for many applications a high impedance input terminal is preferable. For these reasons, the second generation current-conveyor was developed. It has one high and one low impedance input rather than the two low impedance inputs of the CCI. (Sedra, A.S. & Roberts, G.W. (1990))

Second generation current-conveyor differs from the first generation conveyor in that the terminal Y is a high impedance port, i.e. there is no current flowing into Y. While the Y-terminal of the second-generation current-conveyor is a voltage input and the Z terminal is a current output, the X-terminal can be used either as a voltage output or as a current input. Therefore, this conveyor can easily be used to process both current and voltage signals unlike the first generation current-conveyor or the operational amplifier. A further enhancement to the second generation current-conveyor is that there are two types of conveyors: in the positive current-conveyor CCII+, the currents i_x and i_z have the same direction as in a current-mirror and in the negative current-conveyor CCII- the currents i_x and i_z have opposite direction as in a current buffer. (Sedra, A.S. & Roberts, G.W. (1990))

The second-generation current-conveyor is in principle a voltage-follower with a voltage input, Y, and a voltage output, X, and a current-follower (or current-inverter) with a current input X and a current output Z connected together. CCII+ has only recently become commercially available in integrated circuit form, AD844 from Analog Devices and PA630 from Phototronics. A second-generation current conveyor symbol is shown in figure 6.18.

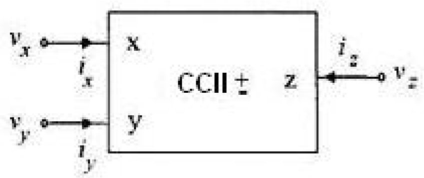


Figure 6.18 Symbol for CCII

Considering the non-idealities arising from the physical implementation of the CCII+ illustrated in figure 6.18, its terminal relationship can be given as:

$$V_X = \beta V_Y , \qquad I_Y = 0 , \qquad I_Z = \alpha I_X \qquad \text{Eq.(6.8)}$$

where α and β are current and voltage gains respectively which can be expressed as α=1- ε_i and β=1- ε_v. (Toker, Özcan, Kuntman, & Çiçekoğlu, 2001)

The matrix representation of CCII+ can be shown as in equation 6.9.

$$\begin{bmatrix} i_y \\ v_x \\ i_z \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ 1+\varepsilon_v & 0 & 0 \\ 0 & 1+\varepsilon_i & 0 \end{bmatrix} \begin{bmatrix} v_y \\ i_x \\ v_z \end{bmatrix} \qquad \text{Eq.(6.9)}$$

where | ε_i| <<1 and | ε_v|<<1 represent the current and voltage tracking errors of the current conveyor, respectively.

6.3.3 Third Generation Current Conveyor (CCIII)

The CCIII symbol is shown in figure 6.19 and the port relation of CCIII is given by Eq.(6.10)

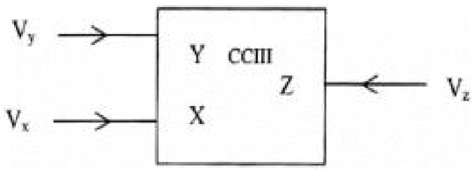


Figure 6.19 Symbol of CCIII

$$I_y = -I_x,\ V_x = V_y,\ I_z = pI_x \qquad \text{Eq.(6.10)}$$

where p=1 for CCIII+ and p=-1 for CCIII-. (Maheshwari, S. & Khan, I. A., 2003)

6.3.4 CCII All-pass Filter Examples

Two basic types of first-order all-pass filters can be defined as follows:

(i) First type all-pass filter:

$$T_1(s) = \frac{V_{out}}{V_{in}} = \frac{G - sC}{G + sC} = \frac{1 - sCR}{1 + sCR} \qquad \text{Eq.(6.11)}$$

(ii) Second type all-pass filter:

$$T_2(s) = \frac{V_{out}}{V_{in}} = -\frac{G - sC}{G + sC} = -\frac{1 - sCR}{1 + sCR} \qquad \text{Eq.(6.12)}$$

These filters have the following phase responses respectively for an ideal case:

$$\theta 1(w) = -2\arctan(wCR) \qquad \text{Eq.(6.13)}$$

$$\theta_2(w) = 180^o - 2\arctan(wCR) \qquad \text{Eq.(6.14)}$$

(Toker, Özcan, Kuntman, & Çiçekoğlu, 2001)

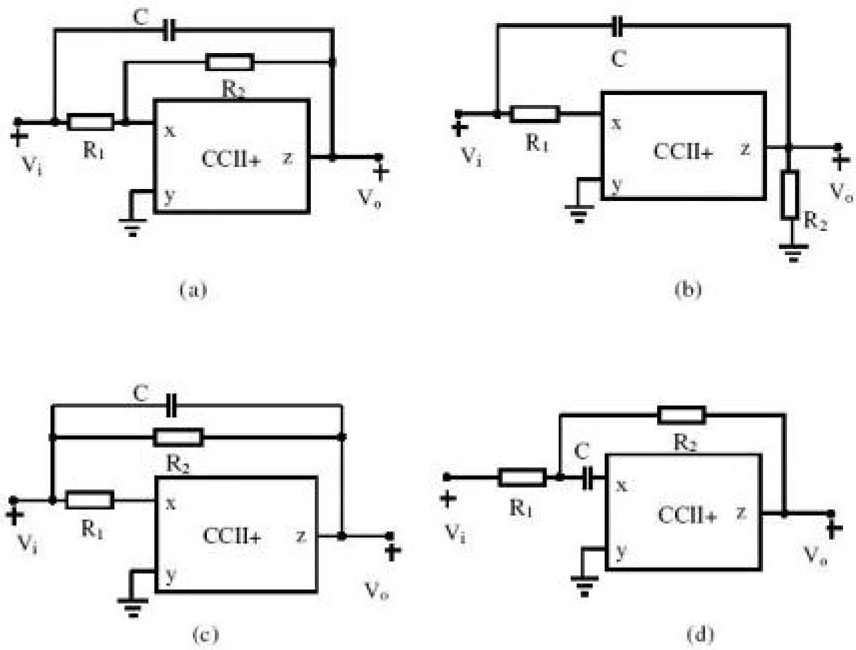


Figure 6.20 All pass filters based on CCII

The configurations in figure 6.20 are operating as current followers and therefore the voltage tracking errors do not appear in the transfer function equations. For the errors due to α, easy compensation is possible by modification of passive element values. (Toker, Özcan, Kuntman, & Çiçekoğlu, 2001)

Table 6.1 shows the transfer functions and element matching conditions of the all pass filters.

Table 6.1 Transfer functions and element matching conditions of the all-pass-filters

Configuration	Type	Transfer Function	Condition	For α=1
Topology I	**Second-type**	$\frac{Vo}{Vi}=-\frac{\alpha}{1+\alpha}\frac{R_2}{R_1}\frac{1-s(CR_1/\alpha)}{1+s[C(R_2/1+\alpha)]}$	$\frac{R_1}{\alpha}=\frac{R_2}{1+\alpha}$	$R_1=R_2/2$
Topology II	**Second-type**	$\frac{Vo}{Vi}=-\alpha\frac{R_2}{R_1}\frac{1-s(CR1/\alpha)}{1+sCR_2}$	$\frac{R_1}{\alpha}=R_2$	$R_1=R_2$
Topology III	**Second-type**	$\frac{Vo}{Vi}=-\frac{\alpha R_2-R_1}{R_1}\frac{1-s[CR1R2/(\alpha R2-R1)]}{1+sCR_2}$	$2R_1=\alpha R_2$	$R_1=R_2/2$
Topology IV	**First-type**	$\frac{Vo}{Vi}=-\frac{\alpha}{1+\alpha}\frac{R_2}{R_1}\frac{1+s\alpha R_2}{1+(1+\alpha)sCR_1}$	$(1+\alpha)R_1=\alpha R_2$	$R_1=R_2/2$

Load impedances affect all of the four filters, and to avoid this, a voltage buffer can be placed at the output of the circuits. The voltage gain of the voltage buffer appears in the transfer function equations as a multiplier. Looking at table 6.1 the effect of the voltage tracking error of the voltage buffer can be easily compensated, except for topology IV. (Toker, Özcan, Kuntman, & Çiçekoğlu, 2001)

The parasitics of the CCIIs affects the circuit performance. The most important parasitic of CCII in the circuits presented in this work is the series input resistance r_X of the *x* terminal. The resistors connected in series to the *x* terminal of the conveyors, are suitable for high frequency operation because, by reducing the value of this resistor by the same amount of r_X, its effect is fully compensated. Circuits employing capacitors in series to the *x* terminal of the conveyor are not suitable for high frequency operation. For this case a transfer function zero occurs, which degrades the frequency response. (Toker, Özcan, Kuntman, & Çiçekoğlu, 2001)

In Çiçekoğlu, O., Kuntman, H., Berk, S. (1999) based on single current conveyor, twenty-two different first order all pass filter topologies are presented. Although all circuits below ideally exhibit the same voltage transfer function, they differ in the number of passive components and represent different physical circuits.

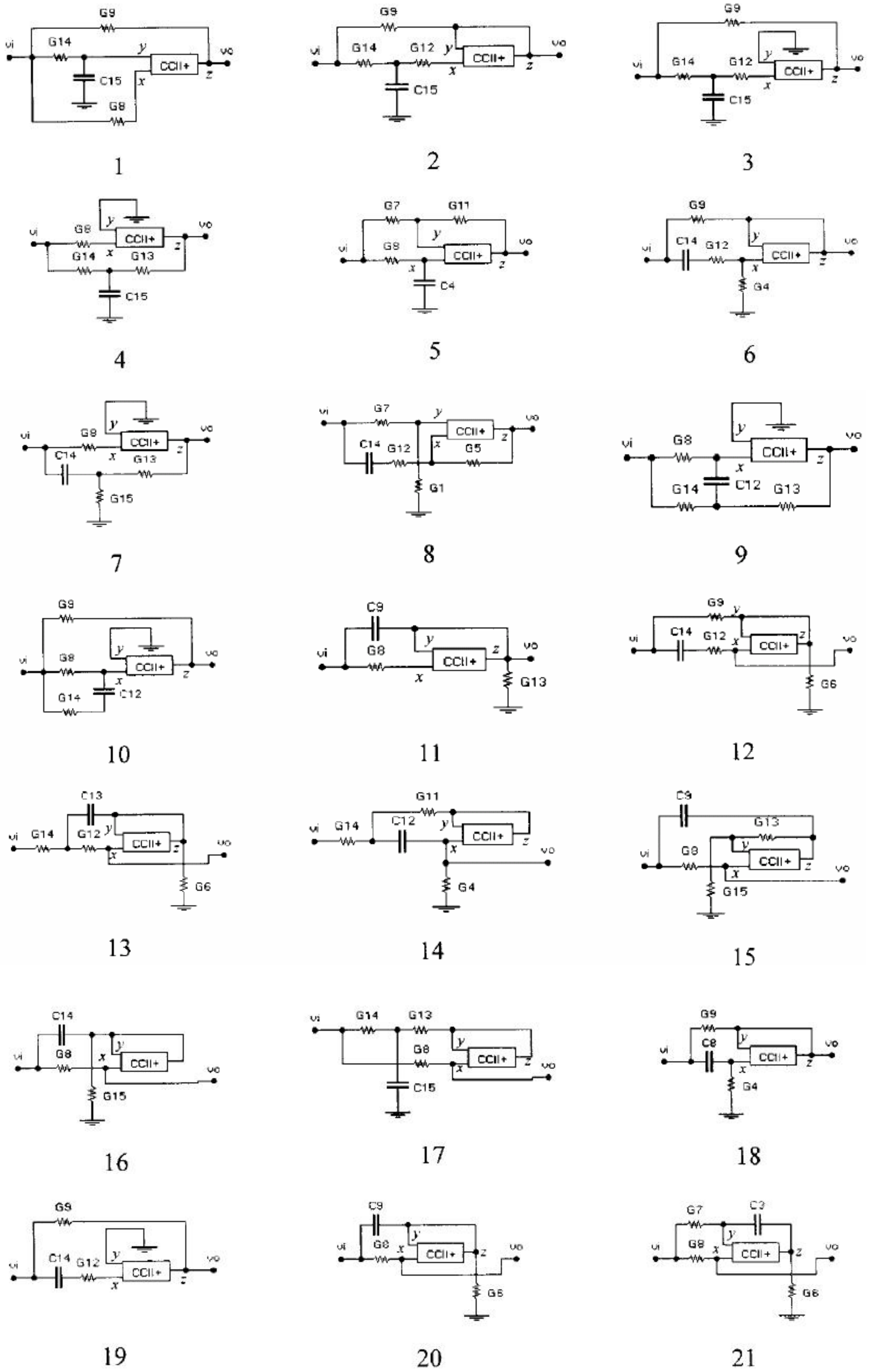


Figure 6.21 All-pass filters using CCII+

Non-ideal transfer functions of the circuits taking the effect of active errors are different. Furthermore, the non-ideal transfer functions taking the finite resistance at the z terminal and the non-zero output resistance at the x terminal of the current conveyor are, in general, different. (Çiçekoğlu, O. & Kuntman, H. & Berk, S., 1999)

The circuits in figure 6.21 may be compared considering the following properties.

- number of passive components
- employment of grounded passive components, resistors and capacitors
- number of component matching conditions
- value of the gain *K*, constant, unity or variable
- value of τ, independent adjustability of τ and *K*
- sensitivities of *K* and τ to passive components

Further properties may be associated with the circuits. For example, circuit 13 and circuit 14 have a resistance ($1/G_{14}$) at the input, which is in series with the output resistance of the input voltage source. Therefore, easy compensation for the non-zero output resistance of the source is possible by adjusting G_{14}. (Çiçekoğlu, O. & Kuntman, H. & Berk, S., 1999)

Circuit 5 in figure 6.21 is simulated using PSPICE with AD844 supported by Analog devices. AD844 is a current feedback op-amp, which may be operated as CCII+ with the additional advantage of having low impedance output which buffers the voltage at the z output. This buffer is designed to drive low impedance loads such as terminated cables, and can deliver ±50mA into a 50Ω load while maintaining low distortion, even when operating at supply voltages of only ±6V. The AD844 exhibits excellent differential gain and differential phase characteristics, making it suitable for a variety of applications with bandwidths up to 60 MHz.

In the filter in figure 6.22 the component values are chosen as:

R_7=100KΩ, R_{11}=100KΩ, R_8=100Ω, C_4=205pF

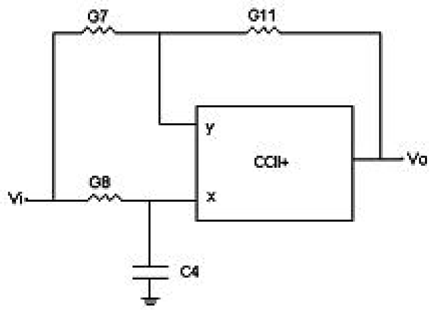


Figure 6.22 First order All-pass Filter, Circuit 5

The transfer function of the circuit is:

$$T(s) = \frac{\frac{sC_4G_7}{G_{11}} + G_7 - G_8}{sC_4 - G_7 + G_8} \qquad \text{Eq.(6.15)}$$

To make the filter all pass filter there is a restriction:

$G_7 = G_{11}$ component matching Eq.(6.16)

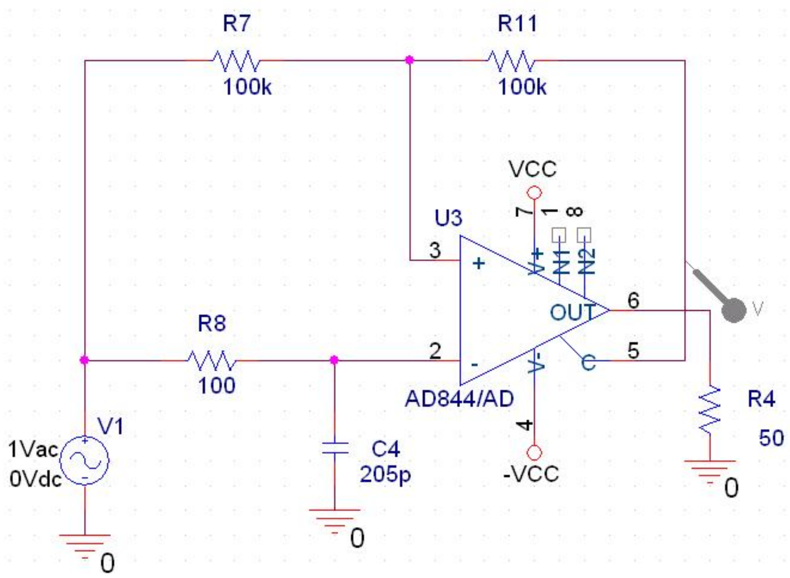


Figure 6.23 First Order All-pass Filter Based on AD844

The gain, phase and delay response of the AD844 filter are given in figures 6.24, 6.25 and 6.26 respectively.

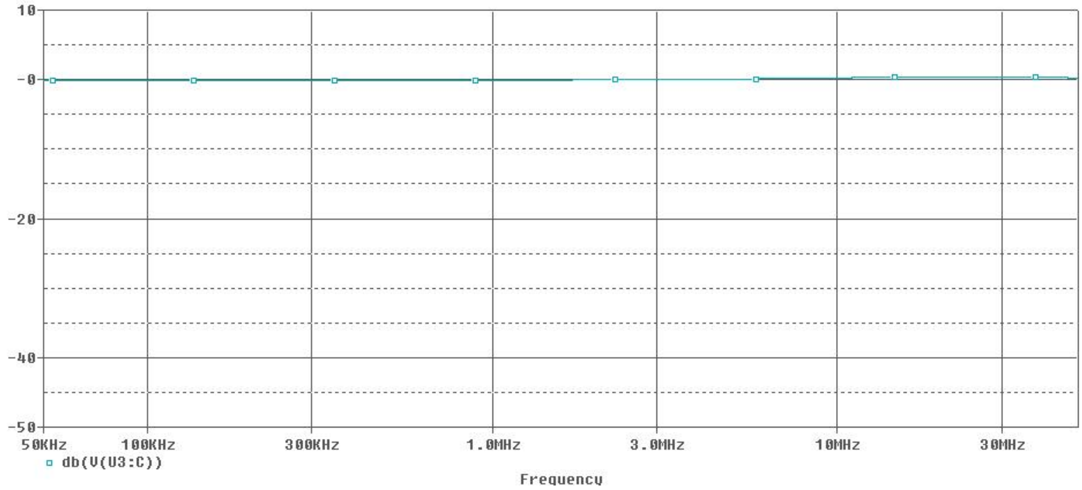


Figure 6.24 Frequency Response of First Order All-pass Filter Based on AD844

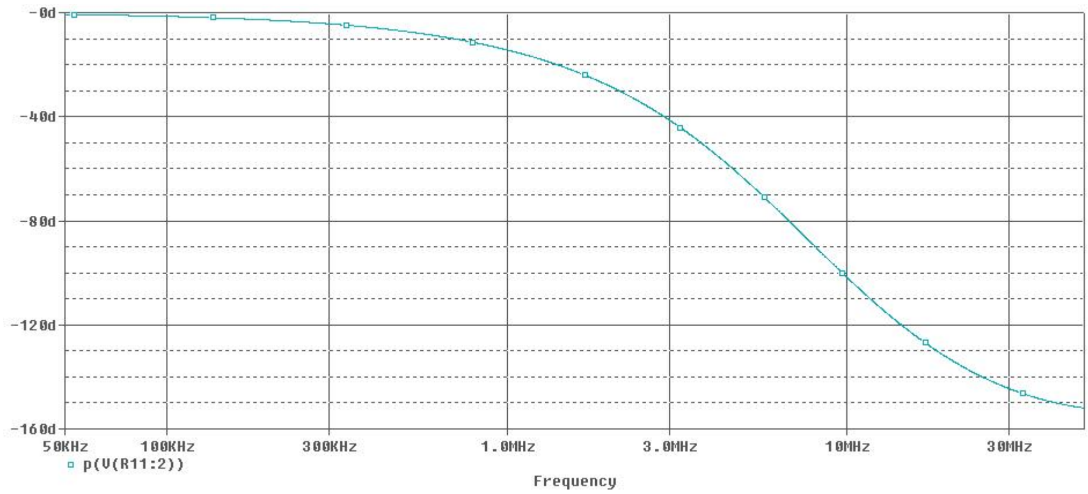


Figure 6.25 Phase Response of First Order All-pass Filter Based on AD844

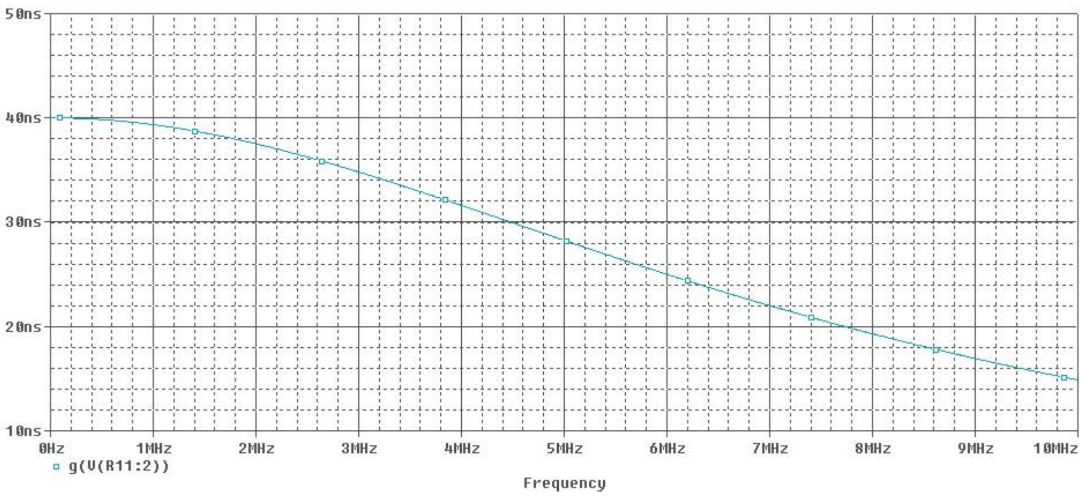


Figure 6.26 Delay Response of First Order All-pass Filter Based on AD844

6.4 All-pass filter Using Single OTRA

The best candidate for the active element to implement the transimpedance allpass filter is the OTRA, which is a high gain current-input, voltage-output device. The use of this device in a negative feedback loop should make it possible to obtain a very accurate transfer function. The input terminals are internally grounded leading to circuits that are insensitive to the stray capacitances. (Çam, Çiçekoğlu, Gülsoy & Kuntman, 2000)

Although OTRAs have been commercially available from several manufacturers under the name of current differencing amplifier or Norton amplifier, the commercial realizations do not provide internal ground at the input port and they allow the input current to flow in one direction only. The former disadvantage limited the functionality of the OTRA, whereas the latter forced the use of external DC bias current, leading to complex and unattractive designs. (Çam, Çiçekoğlu, Gülsoy & Kuntman, 2000)

For ideal operation, the transresistance gain R_m approaches infinity forcing the input currents to be equal. Thus the OTRA must be used in a feedback configuration in a way that is similar to the op-amp. (Çam, Çiçekoğlu, Gülsoy & Kuntman, 2000)

The symbol of OTRA is shown in figure 6.27 and the matrix representation of OTRA can be given as in Eq.(6.17).

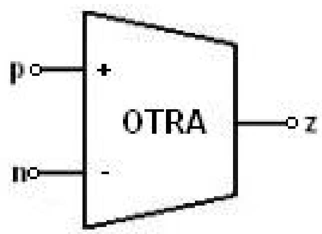


Figure 6.27 Symbol of OTRA

$$\begin{bmatrix} V_p \\ V_n \\ V_z \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ 0 & 0 & 0 \\ R_m & -R_m & 0 \end{bmatrix} \begin{bmatrix} I_p \\ I_n \\ I_z \end{bmatrix} \qquad \text{Eq.(6.17)}$$

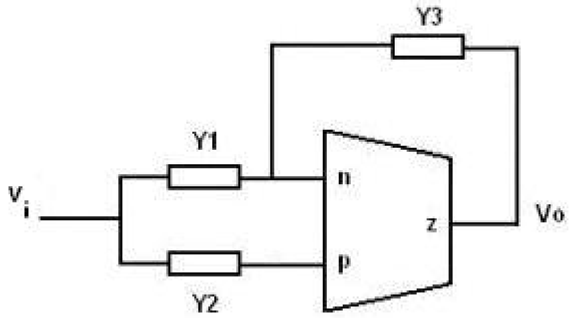


Figure 6.28 All-pass filter with single OTRA

The circuit proposed in Kılınç, S. & Çam, U. (2005) is an all-pass filter based on single OTRA.

The transfer function of the circuit is:

$$T(s) = \frac{V_{out}}{V_{in}} = \frac{Y_2 - Y_1}{Y_3} \qquad \text{Eq.(6.18)}$$

This transfer function allows to designer four different types of realizations for first-order allpass filter by choosing the admittances appropriately.

(i,ii) For $Y_1 = \frac{G}{2}$, $Y_2 = \left(\frac{1}{G} + \frac{1}{sC}\right)^{-1}$, $Y_3 = \frac{G}{2}$ and for $Y_1 = G$, $Y_2 = sC$, $Y_3 = G + sC$ the transfer functions, respectively, become

$$T_{1,2}(s) = \frac{sC - G}{sC + G}, \qquad \text{Eq.(6.19)}$$

(iii,iv) For $Y_1 = \left(\frac{1}{G} + \frac{1}{sC}\right)^{-1}$, $Y_2 = \frac{G}{2}$, $Y_3 = \frac{G}{2}$ and for $Y_1 = sC$, $Y_2 = G$, $Y_3 = G + sC$ the transfer function, respectively, become

$$T_{3,4}(s) = -\frac{sC - G}{sC + G} \qquad \text{Eq.(6.20)}$$

The allpass filters have the following phase responses:

$$\varphi_{1,2}(\omega) = 180^o - 2\arctan(\frac{\omega C}{G}) \qquad \text{Eq.(6.21)}$$

$$\varphi_{3,4}(\omega) = -2\arctan(\frac{\omega C}{G}) \qquad \text{Eq.(6.22)}$$

Figure 6.29 illustrates an all pass filter for case IV based on LM359.

LM359 is a dual high speed Norton amplifier. The current mirror circuitry, which provides the non-inverting input for the amplifier, also facilitates DC biasing the output. The basic Operation of this current mirror is that the current (both DC and AC) flowing into the non-inverting input will force an equal amount of current to flow into the inverting input. The mirror gain (A_I) specification is the measure of how closely these two currents match. DC biasing of the output is accomplished by establishing a reference DC current into the (+) input, $I_{IN}(+)$, and requiring the output to provide the (-) input current.

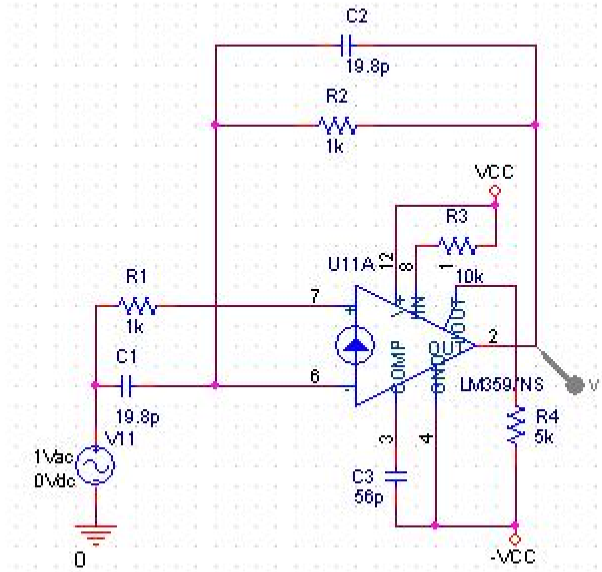


Figure 6.29 First Order All pass filter based on LM359

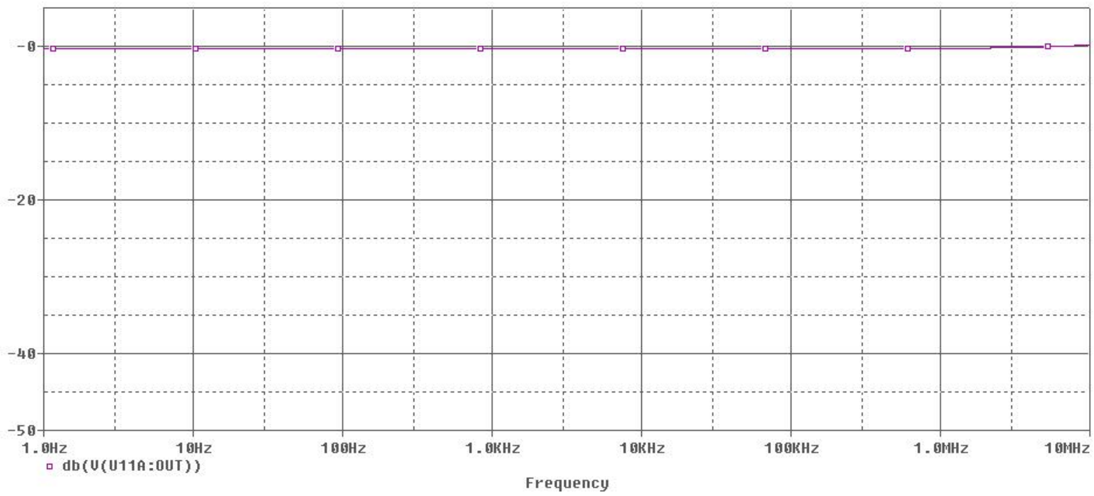


Figure 6.30 Frequency Response of First Order All-pass Filter Based on LM359

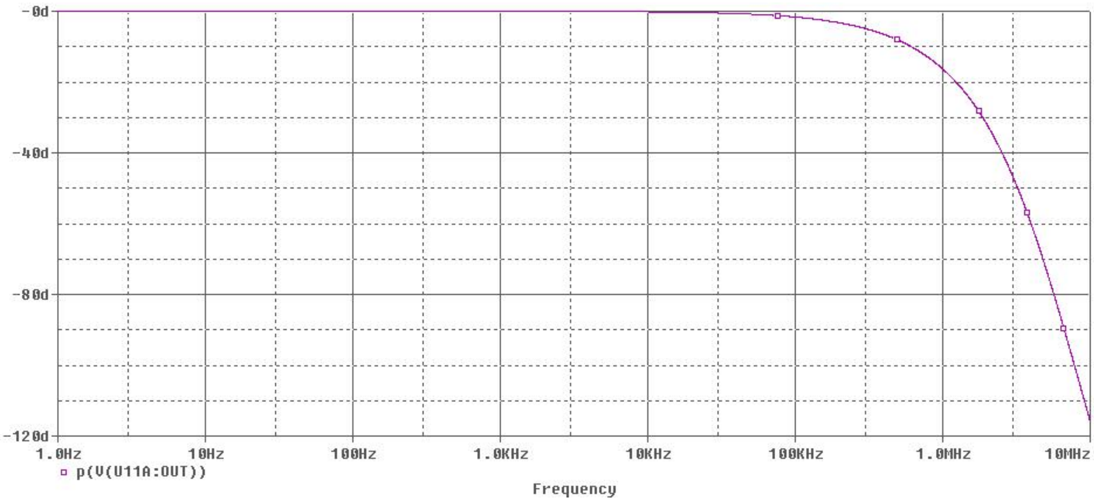


Figure 6.31 Phase Response of First Order All-pass Filter Based on LM359

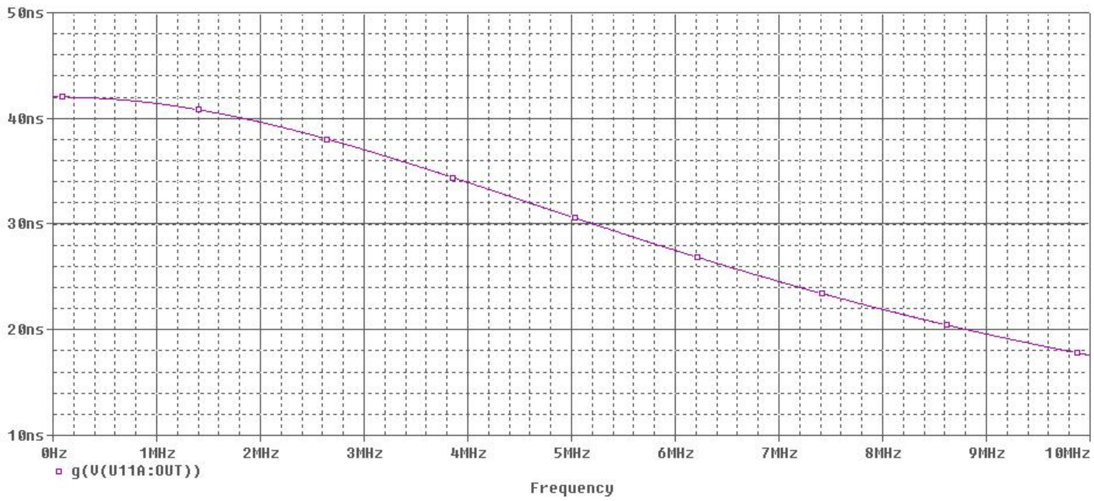


Figure 6.32 Delay Response of First Order All-pass Filter Based on LM359

Advantages of Using OTRA can be listed as follows:

•The negative feedback loop makes it possible to obtain a very accurate transfer function.

•The input terminals are internally grounded leading to circuits that are insensitive to the stray capacitances.

•Both input and output terminals are characterized by low impedance, therefore eliminates the response limitations due to capacitive time constants. (Çam, Çiçekoğlu, Gülsoy & Kuntman, 2000)

CHAPTER SEVEN
ALL PASS FILTERS IN VIDEO SIGNAL PROCESSING

All applications and formats require a video filter to be “phase “linear”, a condition that group delay (delay versus frequency) specifies. The degree of required phase linearity depends on the application and the video format. For example, antialiasing filters and component formats have tighter phase linearity than reconstruction applications and composite video, NTSC, PAL/DVB, ITU, SMPTE, and VESA specify the requirements for the various applications and formats. (Stutz, B., & Bekgran, M., 2003)

In chapter 4 video filters for antialiasing and reconstruction are introduced. Normally, you would choose a filter with the best selectivity and the fewest poles to minimize cost, but the additional need for phase linearity limits the available choices.

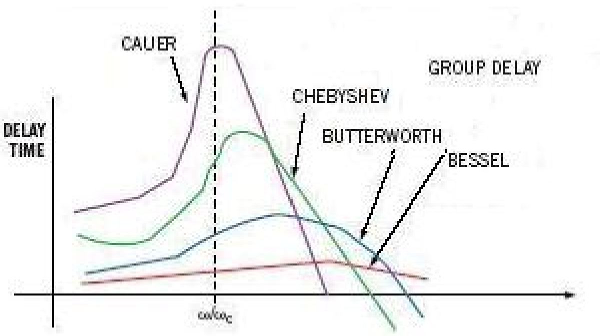


Figure7.1 Group Delay Characteristics of Different Filter Types

7.1 Phase Linearity and Group Delay

As we have seen in chapter 5, a filter’s phase linearity is specified as the envelope delay or group delay versus frequency. A flat group delay indicates that all frequencies are delayed by the same amount, which preserves the shape of the waveform in the time domain. Thus, absolute group delay is less important than the variation in group delay.

Group delay is undesirable but not unacceptable for video, so you may ask how much is acceptable and why it is acceptable. The answer depends on the application and the video format. For example, ITU-470 loosely specifies group delay for composite video, but ITU-601 specifies it tightly to ensure “generational stability,” both for MPEG-2 compression and to control phase jitter before serialization. So you may need to know what filter characteristics to look for to ensure phase linearity.

As we have already seen in chapter 5 the group-delay curves show a peak near the cutoff frequency ($\omega/\omega_C=1$). The steep phase change near the cutoff frequency causes this problematic peak. To get an idea of scale, a threepole, 6-MHz Butterworth filter has a group-delay variation of 20 to 25 nsec over its bandwidth. Adding poles or increasing the filter’s selectivity increases that variation. Other, more exotic filters that minimize group-delay variation include Bessel, Thompson-Butterworth, and LeGendre. Nevertheless, the Butterworth characteristic is most common for video. (Stutz, B., & Bekgran, M., 2003)

7.2 Group Delay Problems

All formats and applications are sensitive to group delay variation. The degree of sensitivity depends on the number of signals and their bandwidths. Composite NTSC/PAL has only one signal, and ITU-470 specifies group delay. Those requirements are easy to meet. RGB and component video both have multiple signals. The RGB signals have equal bandwidths, but component video signals do not, making group delay matching easy with RGB but difficult with component video.

Because the Pb and Pr signals have half the bandwidth of the luma (Y) signal, their group delay is double that of the Y signal. One approach is to slow the Y signal by adding delay stages. Another is to equalize bandwidths by doubling the sample rates of Pb and Pr. That approach raises the 4:2:2 sampling rate to 4:4:4, allowing you to treat the signal as RGB. (The 4:2:2 sampling originally indicated the number of times the color subcarrier was oversampled. ITU-601 replaced the subcarrier frequency with 3.375 MHz. The 4:2:2 sampling occurs at 13.5MHz and 6.75 MHz.) Reconstruction applications discard or average the additional Pb and Pr samples during antialiasing. (Stutz, B., & Bekgran, M., 2003)

The other component video format (SVHS) is confusing to some. The Y channel is the same as in YPbPr, but the chroma signal, C, looks like it needs bandpass filtering rather than lowpass filtering. As for YPbPr signals, bandpass filtering causes group delay and timing problems, so don't do it. Unless you're encoding analog signals, you can lowpass filter Y and C with the same filter. S-VHS is more forgiving of bandwidth limitations than of problems caused by trying to equalize the delay. You typically see S-VHS in reconstruction applications, for which the main concern is correct timing between Y and C. (Stutz, B., & Bekgran, M., 2003)

7.3 Antialiasing Filter for ITU-601

For ITU601 antialiasing filters the specified bandwidth is 5.75MHz±0.1dB with an insertion loss of 12 dB at 6.75MHz and 40dB at 8MHz. For antialiasing filters, a template such as the one in figure 7.2 for ITU-601 determines the sensitivity for delay. The specified group-delay variation is ±3nsec over the 0.1dB bandwidth. These specifications can be met by four times oversampling, which modifies the requirements to 12dB at 27MHz and 40dB at 32MHz. Then we found that a five-pole Butterworth filter with a -3dB bandwidth of 8.45MHz satisfies the requirement for selectivity (Stutz, B., & Bekgran, M., 2003)

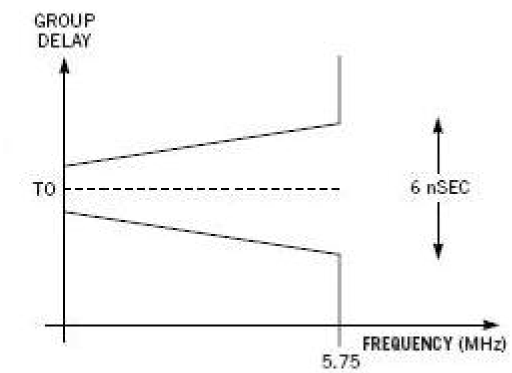


Figure 7.2 ITU-R BT601-5 antialiasing requirement

If we examine the group delay of the circuit in figure 4.5, we find that it doesn't satisfy the requirements of group delay for "ITU-R BT601-5 antianliasing filters". A delay variation of 16.7ns over the 0.1dB bandwidth can be observed as shown in figure 7.4.

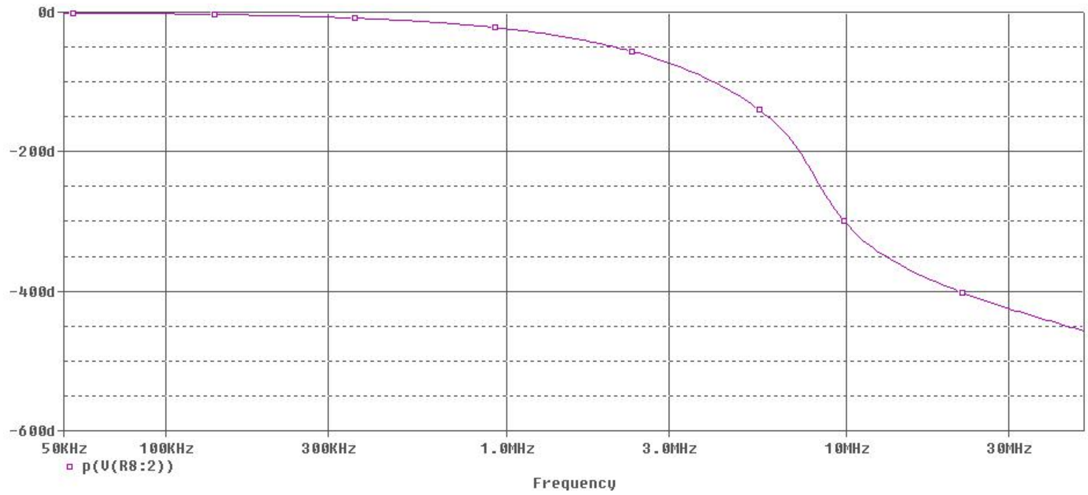


Figure7.3 Phase response of 5 Pole Butterworth Antialiasing Filter

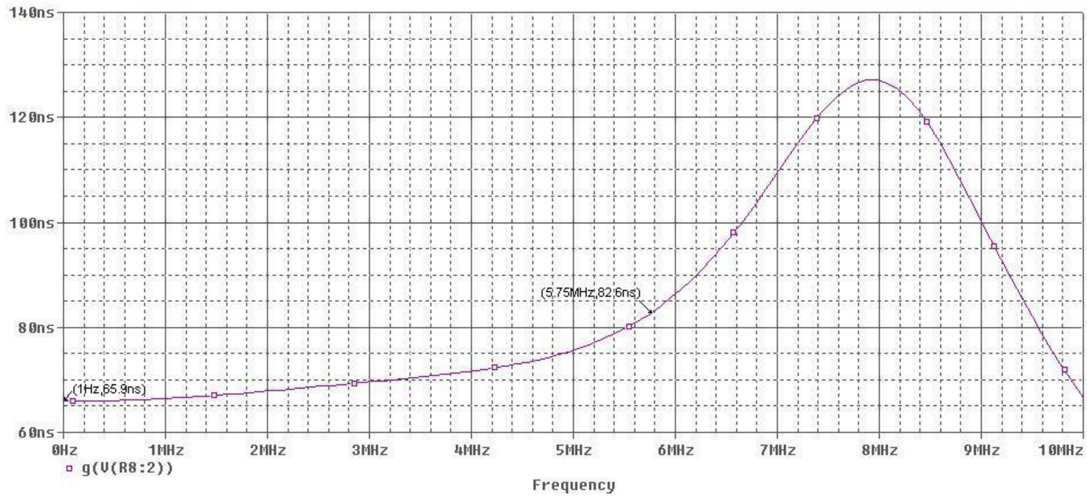


Figure 7.4 Delay response of 5 Pole Butterworth Antialiasing Filter

A delay stage is needed to meet delay requirements of antialiasing filter for ITU-601.

7.3.1 OP-AMP Delay Stage

In choosing an op-amp for the delay stage, the important op-amp parameters 0.1dB, 2Vp-p bandwidth must be taken into attention. MAX4451 is a high-speed opamp that can support the requirements. The gain, phase and delay responses of ITU601 antialiasing filter are shown in figure 7.6, 7.7 and 7.8 respectively.

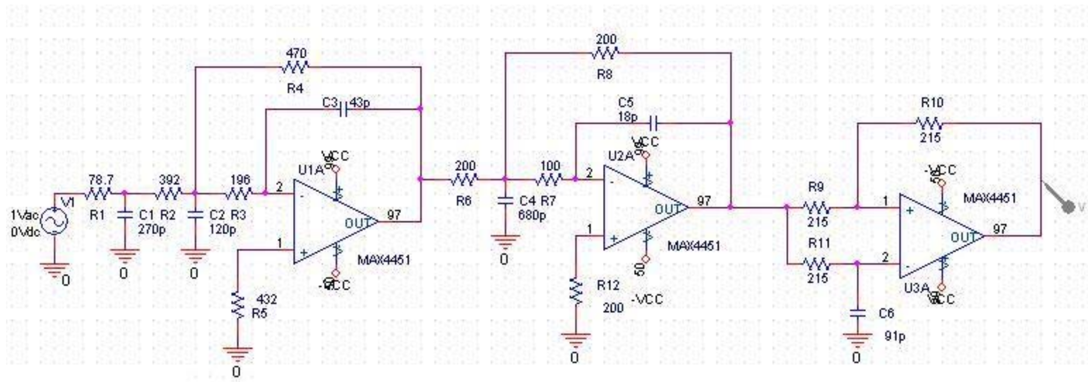


Figure 7.5 A five-pole, 5.75MHz Butterworth filter for ITU-601 antialiasing uses a Rauch circuit with delay equalizer based on MAX4451

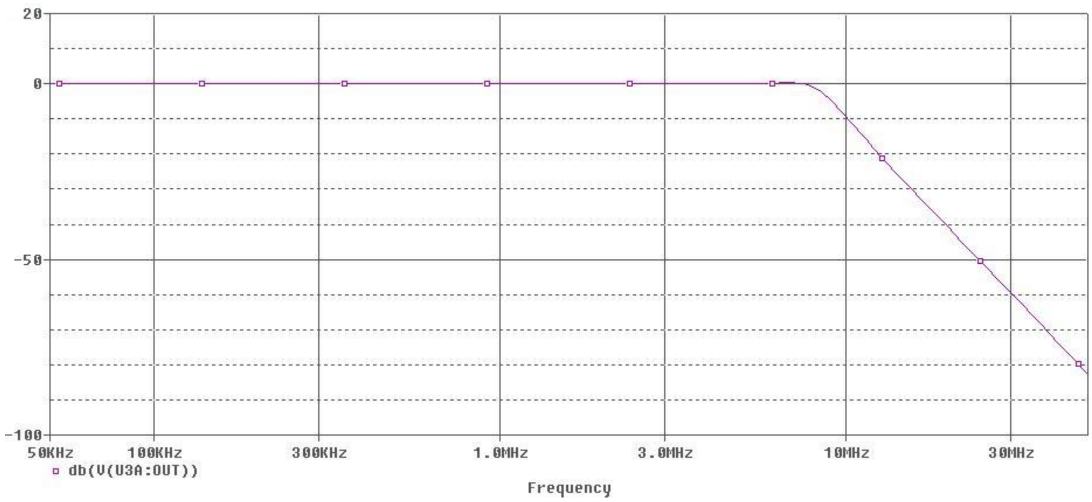


Figure 7.6 Five-pole, 5.75MHz Butterworth filter for ITU-601 antialiasing uses a Rauch circuit with delay equalizer based on MAX4451-Gain response

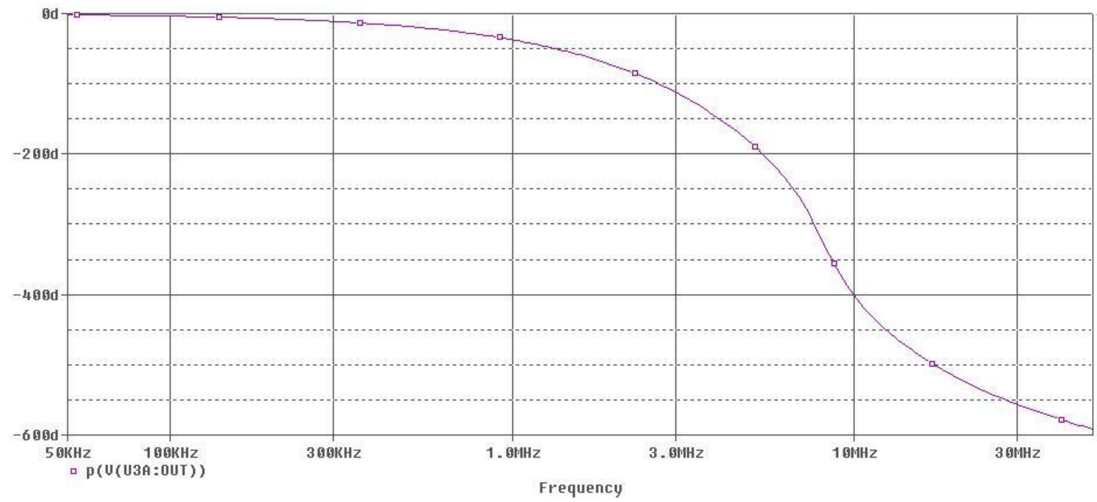


Figure7.7 Five-pole, 5.75MHz Butterworth filter for ITU-601 antialiasing uses a Rauch circuit with delay equalizer based on MAX4451-Phase response

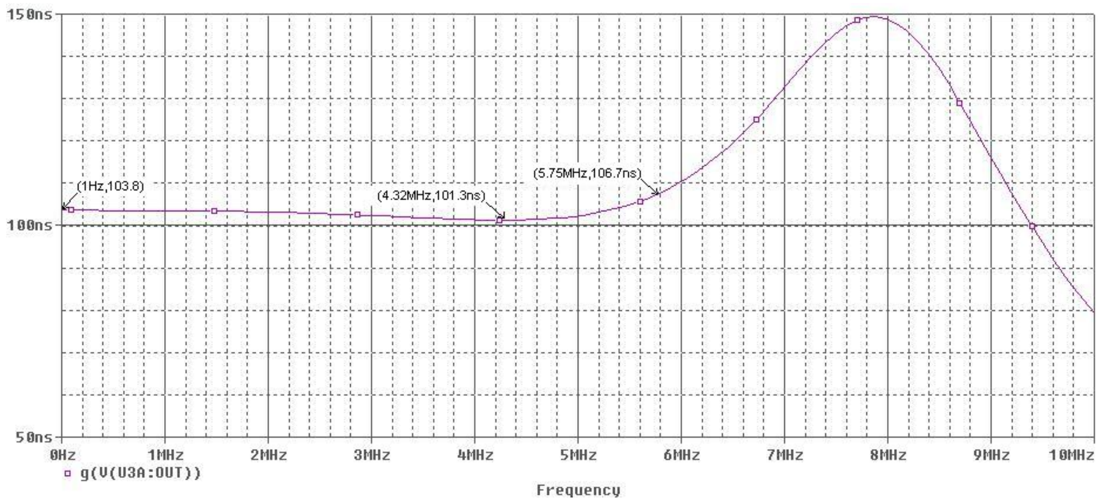


Figure7.8 Five-pole, 5.75MHz Butterworth filter for ITU-601 antialiasing uses a Rauch circuit with delay equalizer based on MAX4451-Delay response

7.3.2 Current Conveyor Delay Stage

Since current conveyors exhibit high linearity, wide dynamic range and have a good high frequency performance, they can be used as a delay stage for video filters. AD844 based all pass filter in figure 6.22 is used as a delay equalizer for the 5-pole Butterworth filter to meet the ITU601 antialiasing requirements.

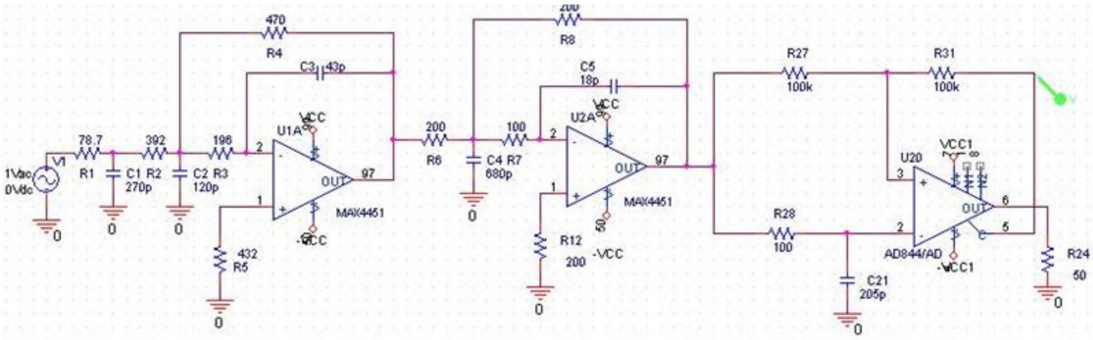


Figure 7.9 5-Pole 5.75 MHz Butterworth Filter for ITU-601 Antialiasing, using a Rauch circuit with delay equalizer based on AD844

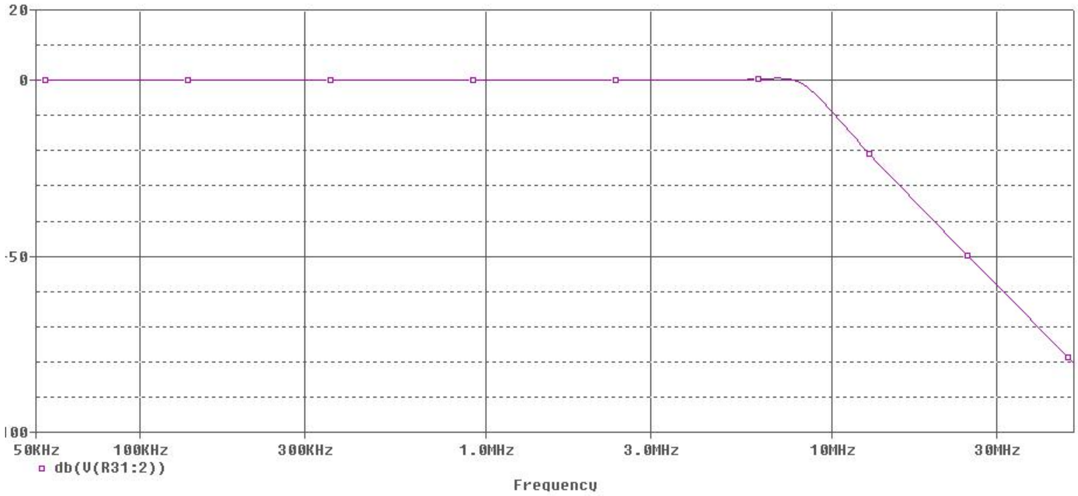


Figure 7.10 5-Pole 5.75 MHz Butterworth Filter for ITU-601 Antialiasing, using a Rauch circuit with delay equalizer based on AD844-Gain Response

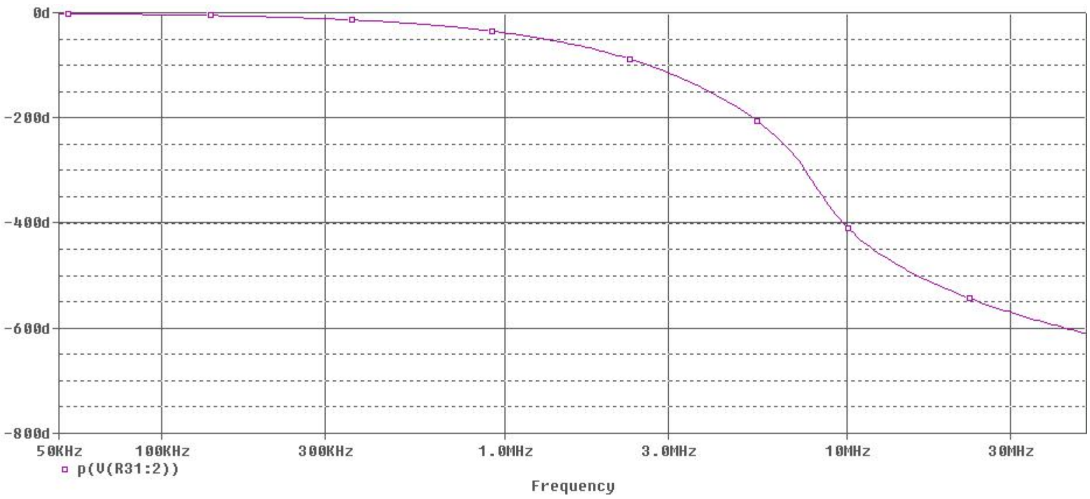


Figure 7.11 5-Pole 5.75 MHz Butterworth Filter for ITU-601 Antialiasing, using a Rauch circuit with delay equalizer based on AD844-Phase Response

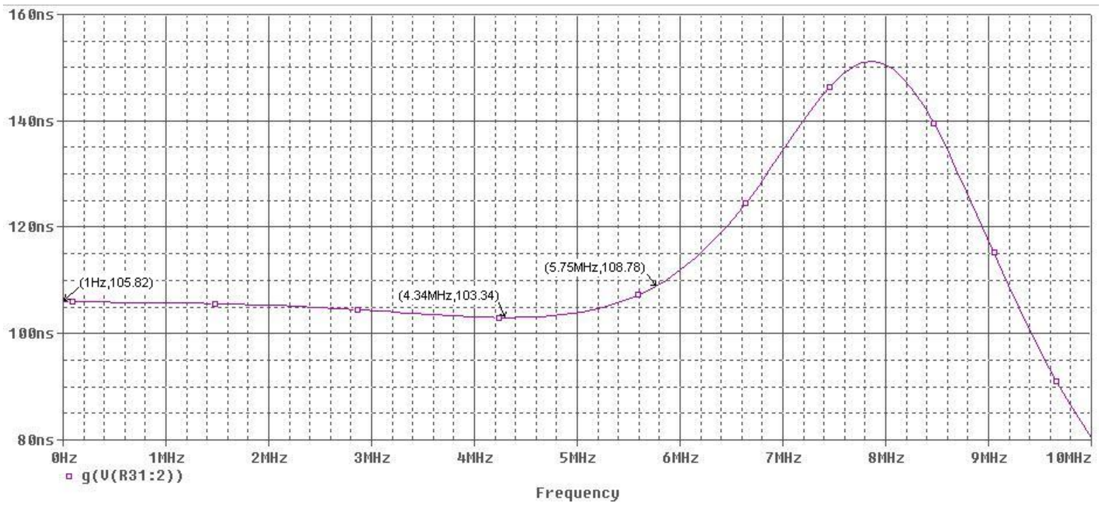


Figure 7.12 5-Pole 5.75 MHz Butterworth Filter for ITU-601 Antialiasing, using a Rauch circuit with delay equalizer based on AD844-Delay Response

7.3.3 OTA Delay Stage

Since high frequency performance of OTAs are good, the all pass filter in figure 6.13 can be used as a delay stage. The delay stage is based on high speed wideband transconductance amplifier MAX435.

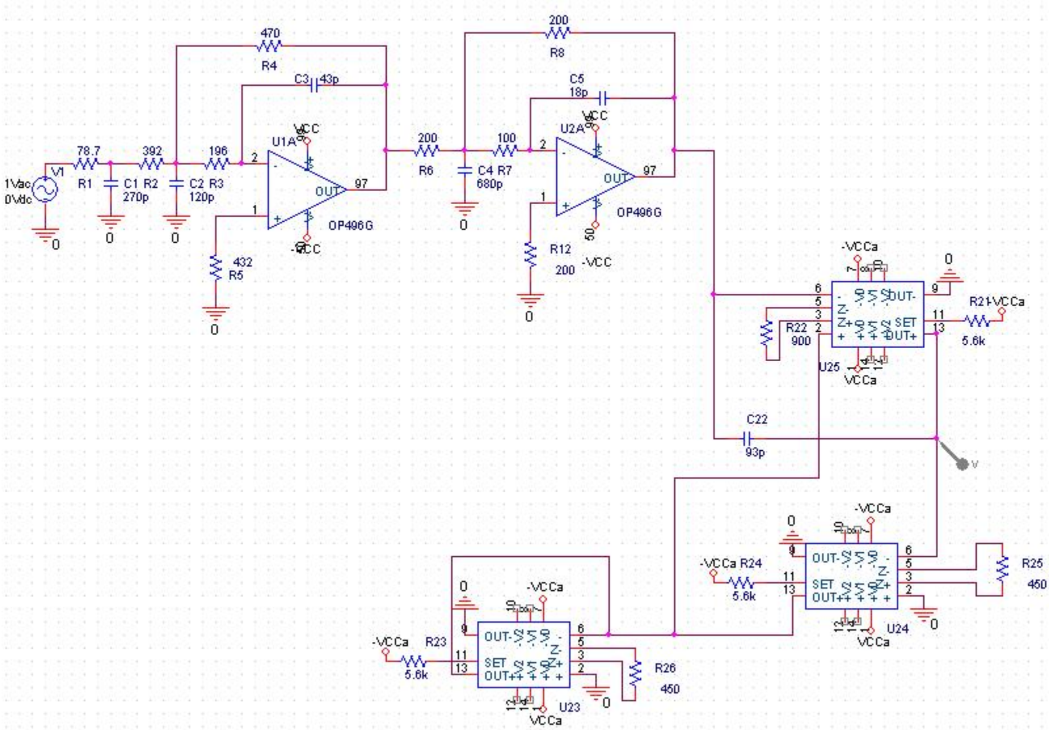


Figure 7.13 5-Pole 5.75 MHz Butterworth Filter for ITU-601 Antialiasing, using a Rauch circuit with delay equalizer based on MAX435

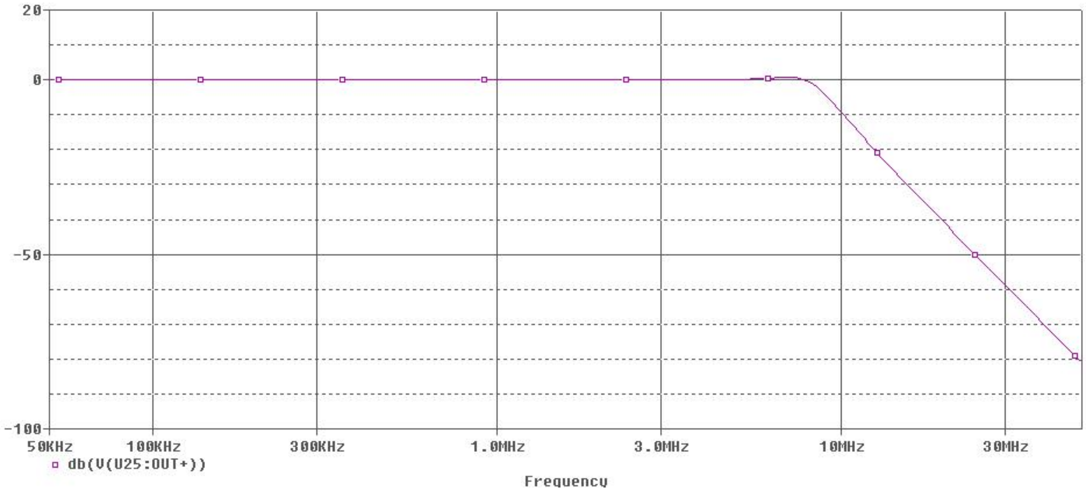


Figure 7.14 5-Pole 5.75 MHz Butterworth Filter for ITU-601 Antialiasing, using a Rauch circuit with delay equalizer based on MAX435-Frequency Response

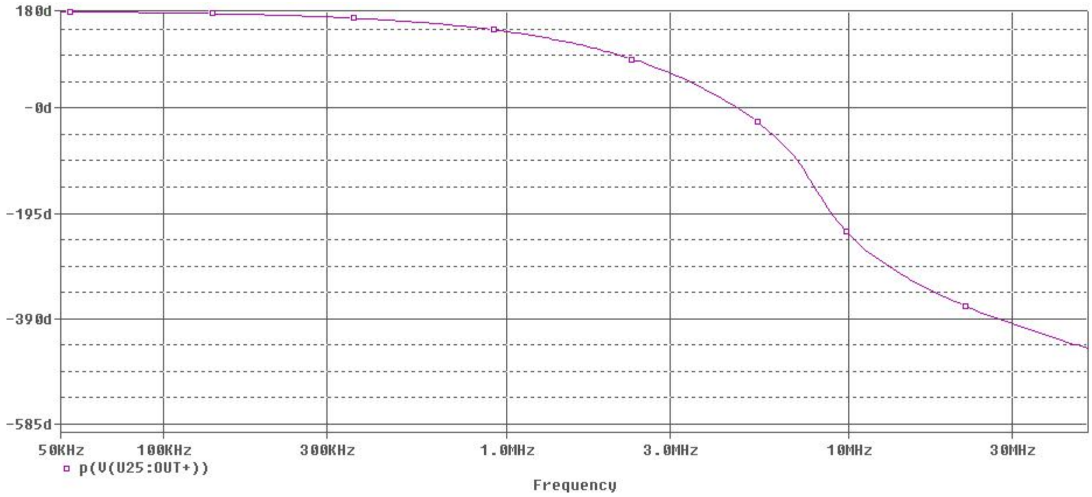


Figure 7.15 5-Pole 5.75 MHz Butterworth Filter for ITU-601 Antialiasing, using a Rauch circuit with delay equalizer based on MAX435-Phase Response

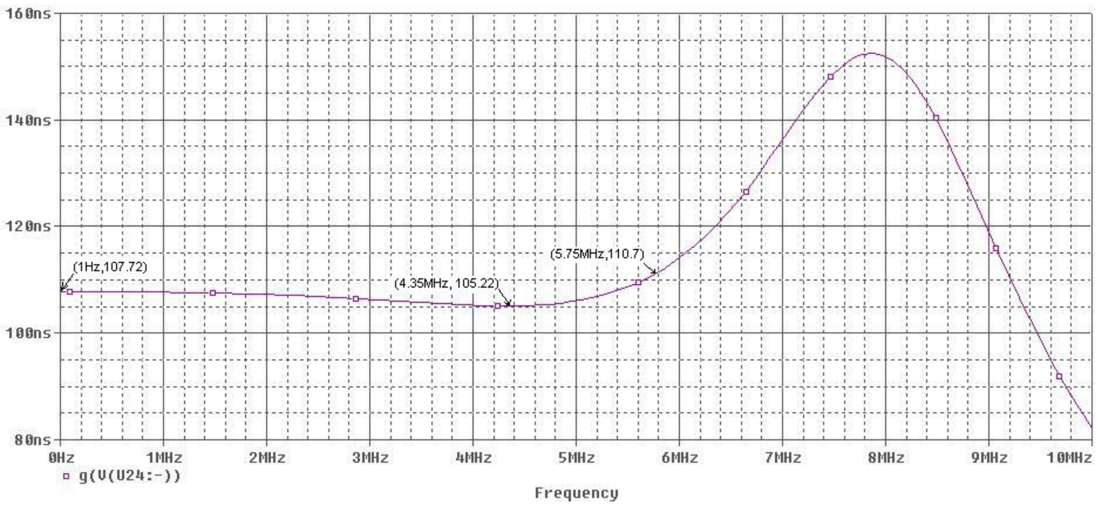


Figure 7.16 5-Pole 5.75 MHz Butterworth Filter for ITU-601 Antialiasing, using a Rauch circuit with delay equalizer based on MAX435-Delay Response

7.3.4 OTRA Delay Stage

Very accurate transfer function can be obtained in negative feedback loop applications since OTRA is a high gain current-input, voltage-output device. The high frequency performance of LM359 makes it possible to use it in video applications. Therefore, the allpass filter in Fig. 6.29 can be used as a delay equalizer in video applications.

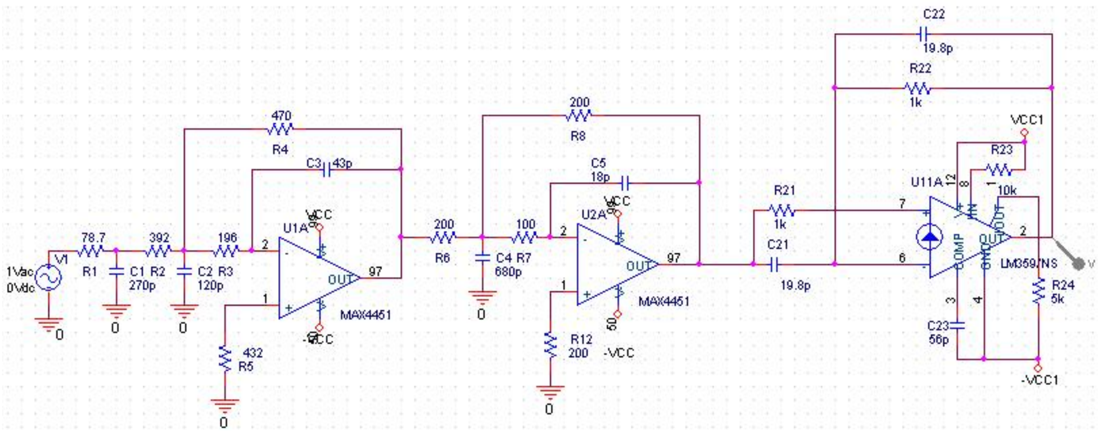


Figure 7.17 5-Pole 5.75 MHz Butterworth Filter for ITU-601 Antialiasing, using a Rauch circuit based on LM359

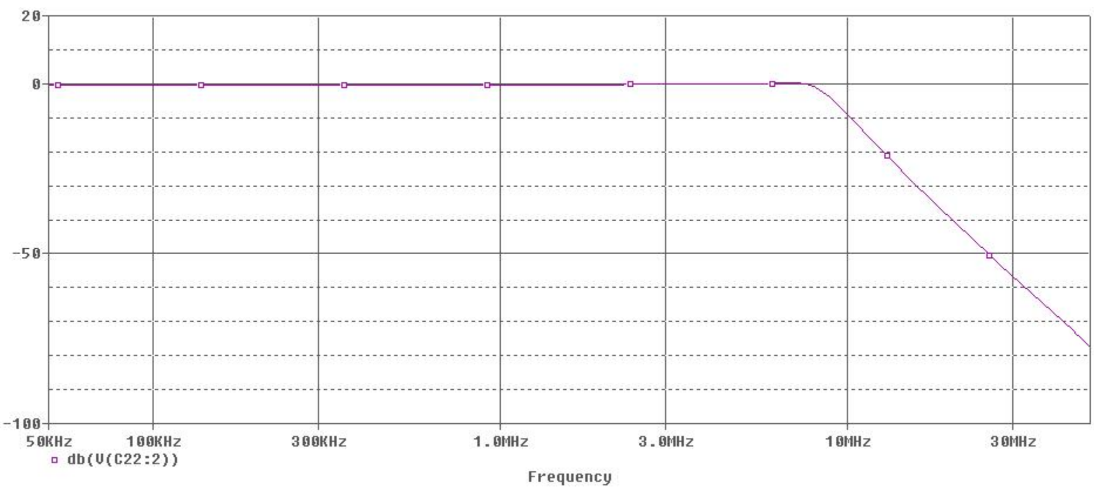


Figure 7.18 5-Pole 5.75 MHz Butterworth Filter for ITU-601 Antialiasing, using a Rauch circuit with delay equalizer based on LM359-Frequency Response

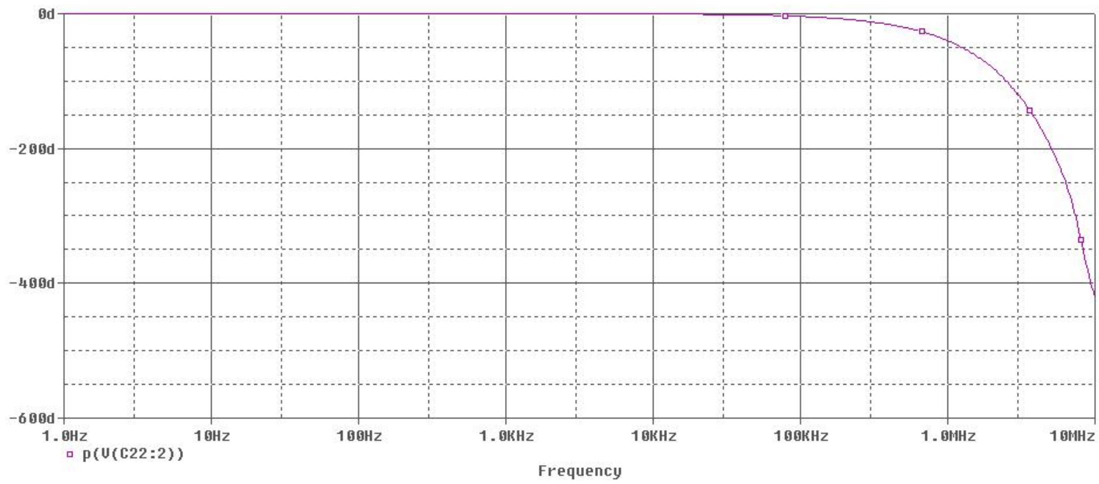


Figure 7.19 5-Pole 5.75 MHz Butterworth Filter for ITU-601 Antialiasing, using a Rauch circuit with delay equalizer based on LM359-Phase Response

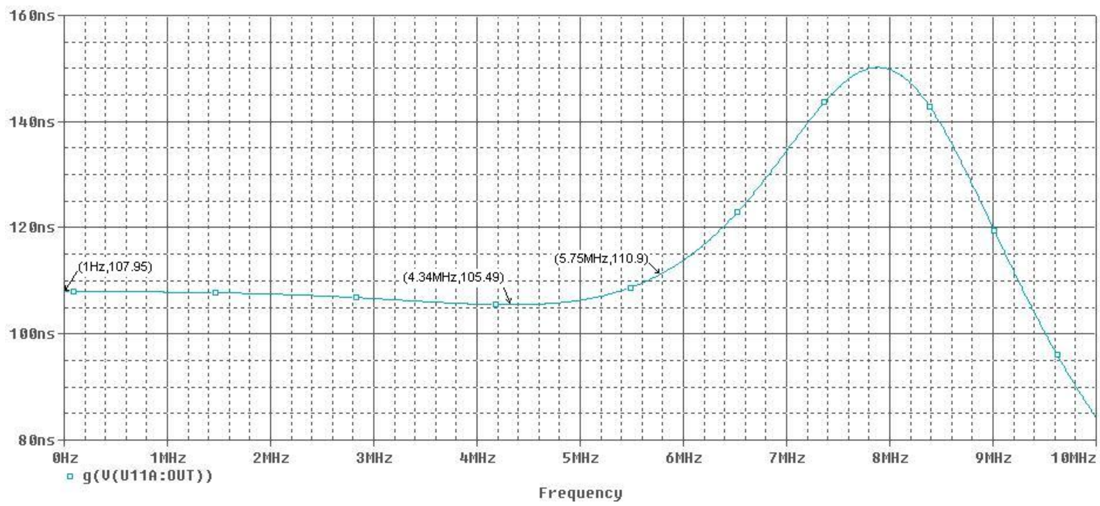


Figure 7.16 5-Pole 5.75 MHz Butterworth Filter for ITU-601 Antialiasing, using a Rauch circuit with delay equalizer based on LM359-Delay Response

CONCLUSION

Classical Op-amps have some drawbacks and limitations such as GBW limitation, limited maximum Quality Factor and Slew Rate problem. In this thesis, new active components have been introduced to be used as phase equalizers in video filters. Other then wide band OP-AMP, these active blocks include operational transconductance amplifiers, current conveyors and operational transresistance amplifiers. As OPAMP, OTRA, OTA and CCII; MAX4451, LM359, MAX435 and AD844 have been chosen respectively which are commercially available.

New active blocks that are used have several advantages over classical video filters, which are based on OPAMP.

MAX435 has been used for new active video filter based on OTA. It is a wideband transconductance amplifier with true differential, high impedance inputs. Its unique architecture provides accurate gain without negative feedback, eliminating closed-loop phase shift. The output of the MAX435 is a current that is proportional to the applied differential input voltage. Circuit gain is set by the ratio of two impedances and an internally set current gain factor (K). This feature allows the OTA to be programmed externally. The all-pass filter based on OTA is also configurable. Regarding the effects such as temperature, adjusting the current gain settings of MAX435 without changing the circuitry can modify the group delay. Its input and output impedances are very high. It can operate in open loop.

AD844 has been used for new active video filters with CCIIs. Its current feedback architecture results in much better ac performance, high linearity and an exceptionally clean pulse response. CCII provides a closed-loop bandwidth, which is determined primarily by the feedback resistor and is almost independent of the closed-loop gain. The AD844 is free from the slew rate limitations inherent in traditional op amps and other current-feedback op amps. Also rise and fall times are essentially independent of output level.

As OTRA, LM359 has been used. LM359 has a high frequency performance and provides user programmable amplifier operating characteristics. Its amplifier is broad banded to provide a high gain bandwidth product and fast slew rate. Since the input resistance of OTRA is low, it has low parasitics. Although OTRA has several advantages the need for using external DC bias current leads to complex and unattractive designs. In biasing LM359, the following properties must be considered:

- DC biasing of the output is accomplished by establishing a reference DC current into the (+) input, and requiring the output to provide the (b) input current.
- The DC input voltage at each input is a transistor VBE (j 0.6 VDC) and must be considered for DC biasing.
- The current mirror at the input terminals provides the current (both DC and AC) flowing into the non-inverting input to force an equal amount of current to flow into the inverting input.

As phase equalizers in video processing, we can use these active devices because of above reasons. If we want to increase the accuracy, we can use low tolerance passive components. The circuits have been simulated at typical condition which is at room temperature. Two factors that disturb the phase equalization most is the temperature and supply voltage ripples. One must consider these parameters in designing.

REFERENCES

Çam, U. & Çiçekoğlu, O. & Gülsoy, M. & Kuntman, H. (2000). New Voltage and Current Mode First-order All-pass Filters Using Single FTFN. *Frequenz, 54*, 7-8

Çiçekoğlu, O., Kuntman, H., Berk, S. (1999). All-pass filters using a single current conveyor. *INT. J. Electronics, 86*, 947-955.

Geiger R. L., Sanchez-Sinencio, E. (1985). Active Filter Design Using Operational Transconductance Amplifiers: A Tutorial. *IEEE Circuits and Devices Magazine,1*

Jack, K. (2001). *A Handbook for the Digital Engineer (3rd ed.)*. VA: LLH Technology Publishing.

Khan, I. A. & Maheshwari, S. (2004) Novel first-Order Current-Mode All-Pass Sections Using CCIII. *Active and Passive Elec. Comp., 27*, 111-117

Kılınç, S. & Çam, U. (2005). Cascadable allpass and notch filters employing single operational transresistance amplifier. *Computers and Electrical Engineering, 31*, 391-401.

Schaumann, R. & Van Valkenburg, M.E. (2001). *Design of Analog Filters*. New York, Oxford University Press.

Sedra, A.S. & Roberts, G.W. (1990). Current Conveyor Theory and Practice. London, Peter Peregrinus Limited.

Stutz, B. & Bekgran, M. (2003). Active filters for video meet antialiasing and reconstruction requirements. *EDN, 48*, 75

Toker, A. & Özcan, S. & Kuntman, H. & Çiçekoğllu, O. (2001). Supplementary allpass sections with reduced number of passive elements using single current conveyor. *INT. J. Electronics, 88*, 969-976

Tole, J.W. (2000), International IC - China Conference Proceeding, Fairchild Semiconductor.

WEB_1 (2003). MAXIM's website. http://dbserv. maxim-ic.com / appnotes. cfm/ appnote_number/2159, 19/04/2004.