SECTION VI
PASSIVE AND ACTIVE ANALOG FILTERING

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SECTION VI

PASSIVE AND ACTIVE ANALOG FILTERING

HANK ZUMBAHLEN

INTRODUCTION TO FILTER DESIGN AND IMPLEMENTATION

Filtering is an important part of analog signal processing. Filtering can be used to reduce unwanted signals, limit bandwidth, help recover wanted signals, minimize aliasing in sampled data systems, and smooth the output of DACs. There are five classes of filters. *Lowpass* filters pass all frequencies below the cutoff frequency and block all frequencies above the cutoff frequency. *Highpass* filters are the inverse of the lowpass filters. They block the low frequencies and pass those above the cutoff frequency. *Bandpass* filters pass those frequencies between the lower cutoff and upper cutoff frequencies and reject all others. *Bandstop* filters are the inverse of bandpass filters. They reject frequencies between the cutoff frequencies and pass all others. *Allpass* filters pass all frequencies equally but introduce a predictable phase delay to the signal.

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CLASSES OF PASSIVE AND ACTIVE FILTERS

- Lowpass
- Highpass
- Bandpass
- Bandstop
- Allpass

Figure 6.1

Traditional filters were passive, that is designed with no active elements. Active components were too costly and had very poor performance. Inductors, capacitors, and resisters were used to synthesize the filter. This approach has several difficulties because inductors become physically large for low frequency filters and have poor characteristics at high frequencies. There is a great deal of interaction between the different sections of the filter. Impedance levels must be precisely controlled. Close component tolerances are difficult to manufacture and maintain. Despite these limitations passive filters are still dominant at high frequencies, primarily due to dynamic performance limitations of op amps.

Active filters answer some of the limitations of the passive filter by offering isolation between stages and eliminating the need for inductors. Their use at high frequencies is limited by the dynamic performance of the active elements.

A filter can be specified in terms of five parameters as shown in Figure 6.4. The *cutoff frequency* F_c is the frequency at which the filter response leaves the error band (or the -3dB point for a Butterworth filter). The *stopband frequency* F_s is the frequency at which the minimum attenuation in the stopband is reached. The *passband ripple* A_{max} is the variation (error band) in the passband response. The *minimum passband attenuation* A_{min} defines the signal attenuation within the passband. The *order* M of the filter is the number of poles in the transfer function.

PASSIVE FILTERS

- Designed with Inductors, Capacitors, Resistors
- Large Inductors Required for Low Frequency Filters
- Interaction Between Filter Stages
- Component Tolerances Difficult to Manufacture and Maintain
- Still the Only Solution at High Frequencies Due to Active Component Limitations

Figure 6.2

ACTIVE FILTERS

- Eliminate Need for Inductors
- Good Interstage Isolation
- High Frequency Use Limited by Op Amp Dynamic Performance

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Figure 6.3

KEY FILTER DESIGN PARAMETERS

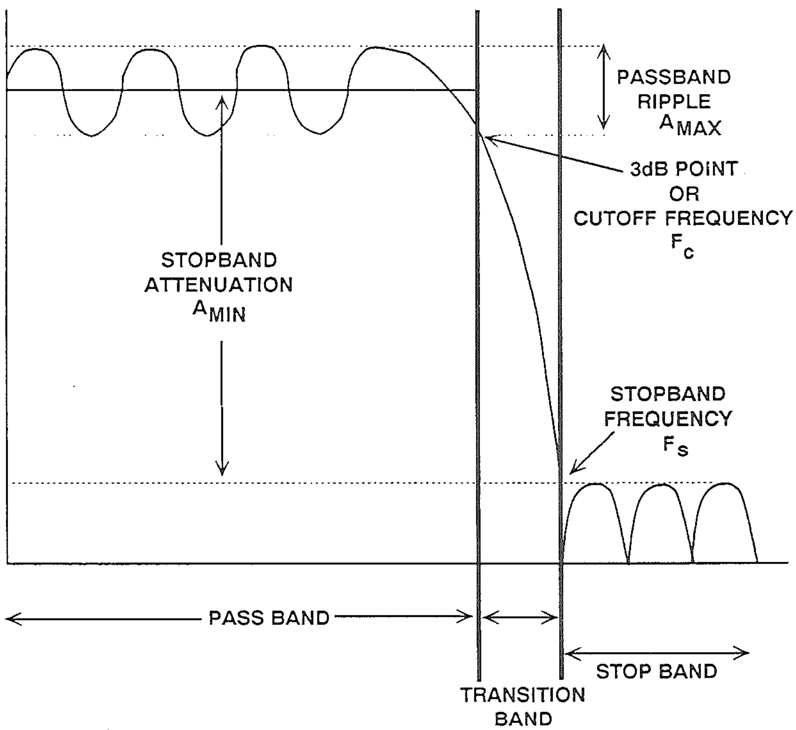


Figure 6.4

FILTER SPECIFICATIONS

- Cutoff Frequency, F_c
- Stopband Frequency, F_s
- Passband Ripple, A_{max}
- Stopband Attenuation, A_{min}
- Filter Order, M

Figure 6.5

Typically, one or more of the above parameters will be variable. For instance, if you were to design an antialiasing filter for an ADC you will know the cutoff frequency, the stopband frequency, and the minimum attenuation. You can then go to a chart or computer program to determine the other parameters.

There are many transfer functions that may satisfy the requirements of a particular filter. The *Butterworth* filter is the best compromise between attenuation and phase response. It has no ripples in the passband or the stopband and is called the *maximally flat filter* because of this. The Butterworth filter achieves its flatness at the expense of a relatively wide transition region from passband to stopband.

The *Chebyshev* filter has a smaller transition region than the same-order Butterworth filter, but it has ripples in either its passband or stopband. This filter gets its name because the Chebyshev filter minimizes the height of the maximum ripple—this is the Chebyshev criterion.

The Butterworth filter and the Chebyshev filter are all-pole designs. By this we mean that the zeros of the transfer function are at one of the two extremes of the frequency range (0 or ∞). For a lowpass filter the zeros are at $f = \infty$. We can add finite frequency transfer function zeros as well as poles to get an *Elliptical Filter*. This filter has a shorter transition region than the Chebyshev filter because it allows ripple in both the stopband and passband. The Elliptical filter also has degraded phase (time domain) response.

These are by no means all possible transfer functions, but they do represent the most common.

POPULAR FILTER DESIGNS

- Butterworth: All Pole, No Ripples in Passband or Stopband, Maximally Flat Response
- Chebyshev: All Pole, Ripple in Passband, Shorter Transition Region than Butterworth for Given Number of Poles
- Elliptical: Ripple in Both Passband and Stopband, Shorter Transition Region than Chebyshev, Degraded Phase Response, Poles and Zeros

Figure 6.6

Once the order of the filter and the specifications of filter have been determined, the design charts (see Reference 1) or computer programs are consulted, and the linear and quadratic factors of poles for the transfer function are determined. All filters, regardless of order, are made up of one- or two-pole sections. The single pole section is defined by its cutoff frequency, which is the -3dB point. The pole pair in a two-pole filter section is defined by its resonant frequency (F_0) and Q, which indicates the peaking of the section. Sometimes alpha (α) is used instead of Q ($Q=1/\alpha$).

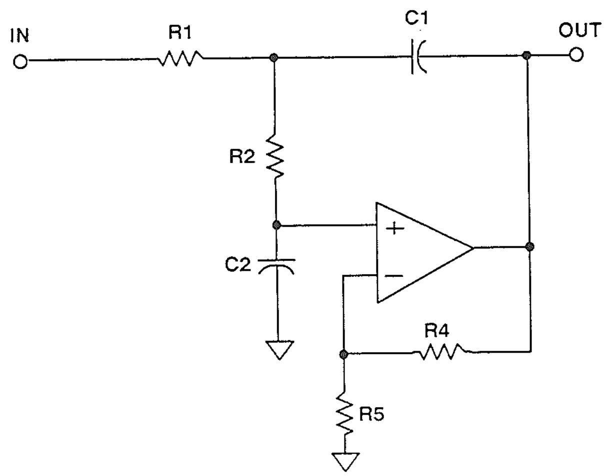
When the values of F_0 and Q are defined, the configuration for the realization of the filter is then chosen: Butterworth, Chebyshev, or Elliptical.

For passive filters, these values, along with the filter characteristic impedance determine the inductor, capacitor, and resistor values.

For active filters, it is necessary to decide which of the realizations to use. The three most common are the *Sallen-Key* (voltage controlled voltage source), *multiple feedback*, and *state variable.* Each realization has its own advantages and disadvantages.

The Sallen-Key configuration shown in Figure 6.7 is the least dependent on the performance of the op amp, and the signal phase is maintained. For this filter the ratio of the largest resistor value to the smallest resistor value and the ratio of the largest capacitor value to the smallest capacitor value is low. The frequency term and Q terms are somewhat independent, but they are very sensitive to the gain parameter. The Sallen-Key is very Q-sensitive to element values for high Q sections. The design equations are given in Figure 6.7.

VOLTAGE CONTROLLED VOLTAGE SOURCE (SALLEN-KEY) REALIZATION



H = Circuit Gain Below Cutoff

α = Damping Ratio = 1/Q

F_O = Cutoff Frequency

Choose C1

$K = 2\pi F_C C1$

$M = \frac{\alpha^2}{4} + H - 1$

$C2 = M\ C1$

$R1 = \frac{2}{K\alpha}$

$R2 = \frac{\alpha}{2MK}$

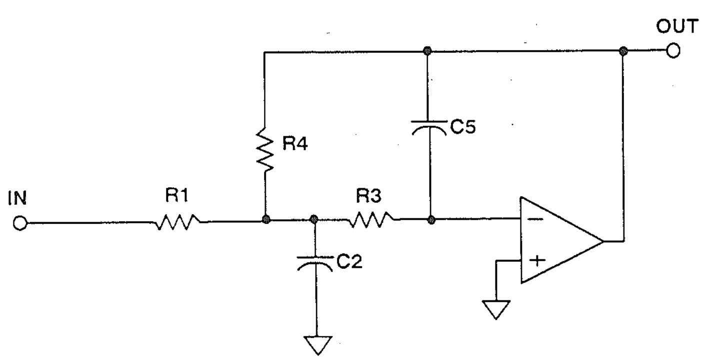
Choose R5

$R4 = R5(H - 1)$

For H = 1, R4 = 0, R5 = Open

Figure 6.7

MULTIPLE FEEDBACK REALIZATION



F_O = Cutoff Frequency

α = Damping Ratio = 1/Q

H = Absolute Value of Circuit Gain

Choose C5

$K = 2\pi F_O C1$

$C2 = \frac{4C5}{\alpha^2}(H + 1)$

$R1 = \frac{\alpha}{2HK}$

$R3 = \frac{\alpha}{2K(H + 1)}$

$R4 = HR1$

Figure 6.8

The multiple feedback filter shown in Figure 6.8 uses an op amp in the inverting configuration. The dependence on the op amp parameters is greater than in the Sallen-Key realization. It is hard to generate high Q sections due to the limitations of the open loop gain of the op amp. The maximum to minimum component value ratios are higher than in the Sallen-Key realization. The design equations are also given in Figure 6.8.

The state-variable realization shown in Figure 6.9 offers the most precise implementation, at the expense of many more circuit elements. All parameters can be adjusted independently, and lowpass, highpass, and bandpass outputs are available simultaneously. The gain of the filter is also independently variable. Since all parameters of the state variable filter can be adjusted independently, component spread is minimized. Also variations due to temperature and component tolerances are minimized. The design equations for the state variable filter are given in Figure 6.9.

Another active filter technique that has recently become more popular is the *Frequency Dependent Negative Resistor* (FDNR), which is a subset of the *General Impedance Converter* (GIC). In the FDNR the passive realization goes through a transformation by 1/s. Therefore inductors, whose impedance is sL, transform into a resistor of value L. Similarly, a resistor of value R becomes a capacitor of value R/s. A capacitor of impedance 1/sC transforms into a frequency dependent variable resistor, which is given the designation D. Its impedance is $1/s^2C$. The transformations to the FDNR configuration and the GIC implementation of the D element are given in Figure 6.10.

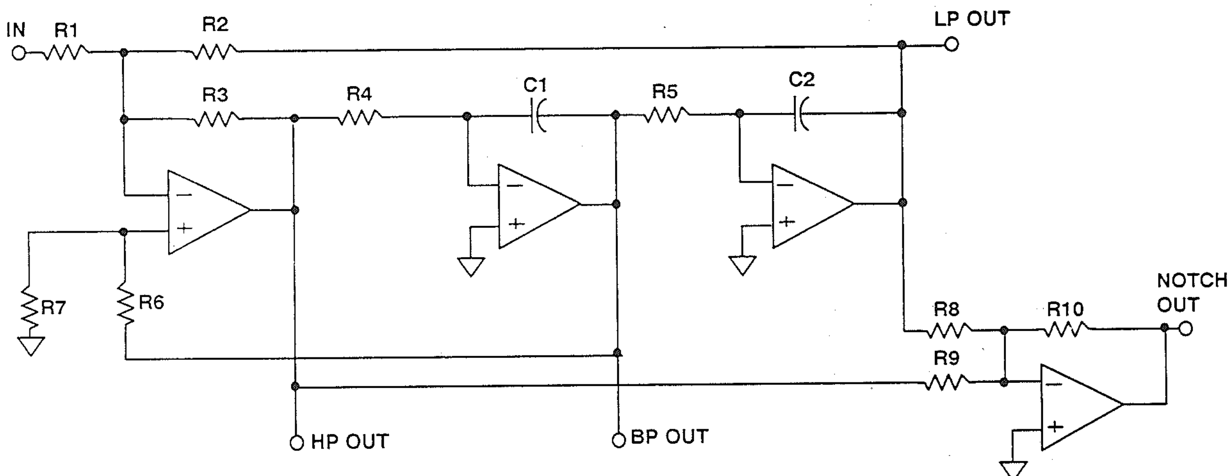
The advantage of the FDNR filter is that there are no op amps in the signal path which can add noise. It is also relatively insensitive to component variation. The advantages of the FDNR come at the expense of an increase in the number of components required.

For all of the realizations discussed above, the tabulated filter values are in terms of the lowpass function normalized to a frequency of 1 radian/second with an impedance level of 1Ω. To realize the final design, the filter values are scaled by the appropriate frequency and impedance.

Similarly, the lowpass prototype is converted to a highpass filter by scaling by 1/s in the transfer function. In practice this amounts to capacitors becoming inductors with a value 1/C and inductors becoming capacitors with a value of 1/L for passive designs. For active designs resistors become capacitors with a value of 1/R, and capacitors become resistors with a value of 1/C.

Transformation to the bandpass response is a little more complicated. If the corner frequencies of the bandpass are widely separated (by more than 2 octaves) the filter is made up of separate lowpass and highpass sections. In the case of a narrowband bandpass filter the design is much more complicated and is usually done using a computer program or design tables.

STATE VARIABLE REALIZATION



LOWPASS GAIN = -R2/R1

HIGHPASS GAIN = -R3/R1

$$\text{BANDPASS GAIN} = \frac{R6 + R7}{R1R7\left(\frac{1}{R1} + \frac{1}{R2} + \frac{1}{R3}\right)}$$

$$F_O = \frac{1}{2\pi}\sqrt{\frac{R3}{R2 \cdot R4 \cdot R5 \cdot C1 \cdot C2}}$$

$$Q = \frac{1}{\alpha} = \frac{R6 + R7}{R7}\left(\frac{1}{\frac{1}{R1} + \frac{1}{R2} + \frac{1}{R3}}\right)\sqrt{\frac{R4 \cdot C1}{R2 \cdot R3 \cdot R5 \cdot C2}}$$

FOR NOTCH FREQUENCY = F_Z

FOR $F_O = F_Z$, $\frac{R2 \cdot R9}{R3 \cdot R8} = 1$

FOR $F_O > F_Z$, $\frac{R2 \cdot R9}{R3 \cdot R8} < 1$

FOR $F_O < F_Z$, $\frac{R2 \cdot R9}{R3 \cdot R8} > 1$

$$\frac{F_Z^2}{F_O^2} = \frac{R2 \cdot R9}{R3 \cdot R8}$$

Figure 6.9

FREQUENCY DEPENDENT NEGATIVE RESISTOR 1/S IMPEDANCE TRANSFORMATION

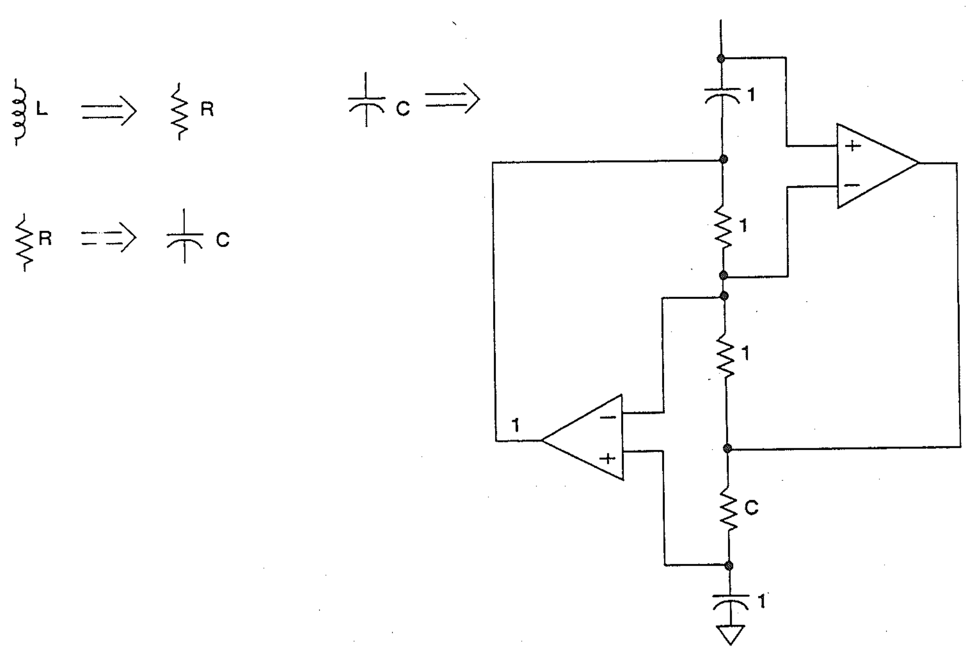


Figure 6.10

SOME ACTIVE FILTER REALIZATIONS

- Sallen-Key: Good Phase Response, Least Dependent on Op Amp Performance, Sensitive to Element Values for High Q Sections
- Multiple Feedback: Less Sensitive to Element Values, High Q Sections Difficult due to Op Amp Open Loop Gain Limitations
- State-Variable: Most Precise, More Components, All Parameters Independently Adjustable
- Frequency Dependent Negative Resistance (FDNR): Op Amps not in Signal Path, More Components, Relatively Insensitive to Component Variations

Figure 6.11

ANTIALIASING FILTER DESIGN EXAMPLE

We will now design passive and active antialiasing filters based upon the same specifications. The active filter will be designed in four ways: Sallen-Key, multiple feedback, state variable, and Frequency Dependent Negative Resistance (FDNR). We choose the Butterworth filter in order to give the best compromise between attenuation and phase response.

The specifications for the filter are as follows:

ANTIALIASING FILTER SPECIFICATIONS

- Cutoff Frequency F_C = 8kHz
- Stopband Attenuation F_S at 50kHz = 70dB
- Best Balance Between Attenuation and Phase Response Choose Butterworth Design
- From Design Charts, for f = 6.25 (50kHz/8kHz), M = 5

Figure 6.12

Consulting the design charts (Reference 1, p. 82), we see that for 70dB of attenuation at a frequency of 6.25 (50kHz/8kHz) a fifth order filter is required.

We now consult the tuning tables (Reference 1, p. 341) and find:

ALPHA AND F0 VALUES FROM TUNING TABLES

Stage	Alpha	F0
1	---	1.000
2	1.618	1.000
3	0.618	1.000

Figure 6.13

The first stage is a real pole, thus the lack of an alpha value. It should be noted that this is not necessarily the order of implementation in hardware. In general you would typically put the real pole last and put the second order sections in order of decreasing alpha (increasing Q).

For the passive design we will choose the zero input impedance configuration. From the design table (Reference 1, p. 313) we find the following normalized values for the filter:

NORMALIZED PASSIVE FILTER VALUES FROM TABLES

L1 = 1.5451	C2 = 1.6944
L3 = 1.3820	C4 = 0.8944
L5 = 0.3090	

Figure 6.14

These values are for a 1 rad/second filter with a 1 Ω termination. To scale the filter we divide all reactive elements by the desired cutoff frequency, 8kHz (50265 rad/sec). We also need to scale the impedance. For this example, we choose a value of 1000 Ω. To scale the impedance we multiply all resistor and inductor values and divide all capacitor values by the impedance scaling factor. After scaling, the circuit looks like Figure 6.15.

For the Sallen-Key active filter, we use the design table shown in Figure 6.7. The values for C1 in each section are chosen to give reasonable resistor values. The implementation is shown in Figure 6.16. For the Sallen-Key realization to work correctly, it is assumed to have a zero-impedance driver and a return path for dc. Both of these criteria are approximately met when you use an op amp to drive the filter.

Figure 6.17 shows a multiple feedback realization of our filter. It was designed using the equations in Figure 6.8.

The state variable filter is shown in Figure 6.18, and the Frequency Dependent Negative Resistance (FDNR) realization is shown in Figure 6.19. In the conversion process from passive to FDNR, the D element is normalized for a capacitance of 1F. We then scale the filter to a more reasonable value (0.01μF in this case).

In all of the filters above the values shown are the exact calculated values. These exact values are rarely obtainable. We must therefore either substitute the nearest standard value or use series/parallel combinations. Any variation from the ideal values will cause a shift in the filter response characteristic, but often the effects are minimal. The computer can be used to evaluate these variations on the overall performance and determine if they are acceptable.

In active filter applications using op amps, the dc accuracy of the amplifier is often critical to optimal filter

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EXAMPLE FILTER PASSIVE IMPLEMENTATION

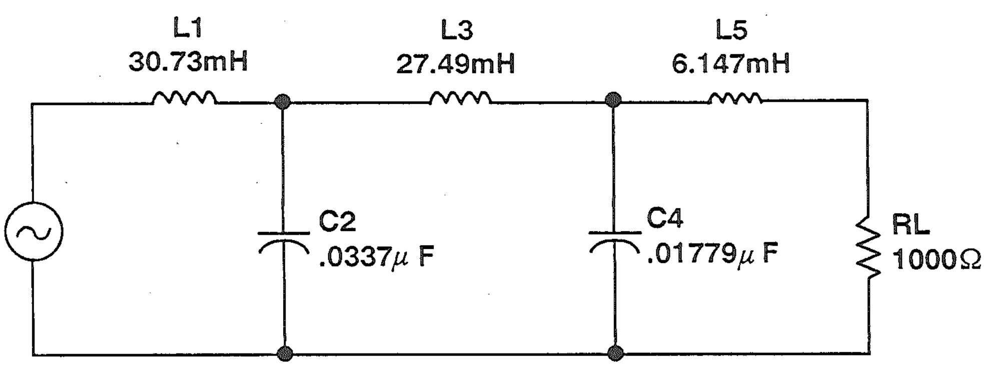

Figure 6.15

EXAMPLE FILTER
SALLEN-KEY IMPLEMENTATION

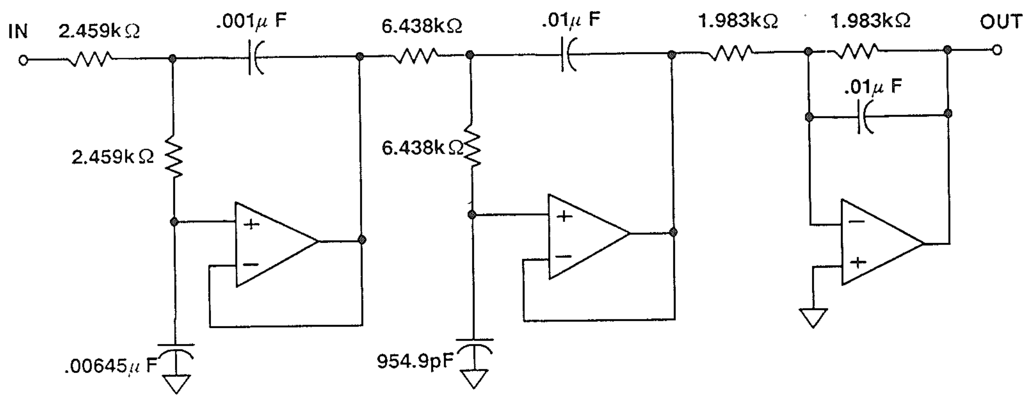

Figure 6.16

EXAMPLE FILTER
MULTIPLE FEEDBACK IMPLEMENTATION

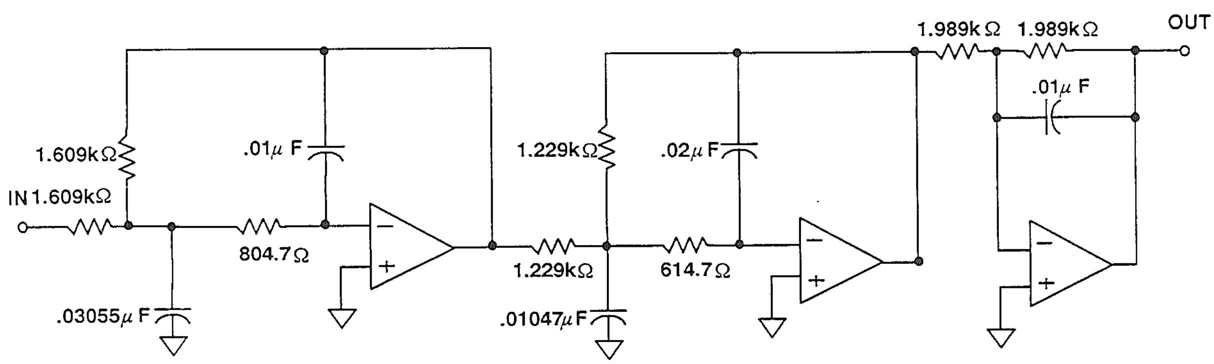

Figure 6.17

EXAMPLE FILTER
STATE VARIABLE IMPLEMENTATION

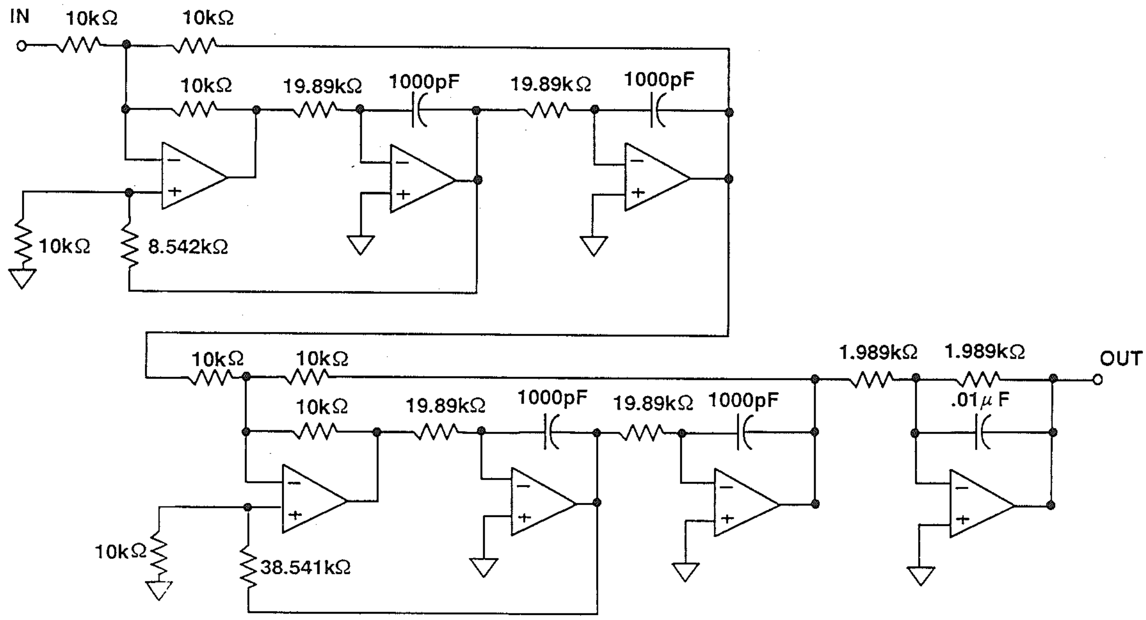

Figure 6.18

EXAMPLE FILTER
FDNR IMPLEMENTATION

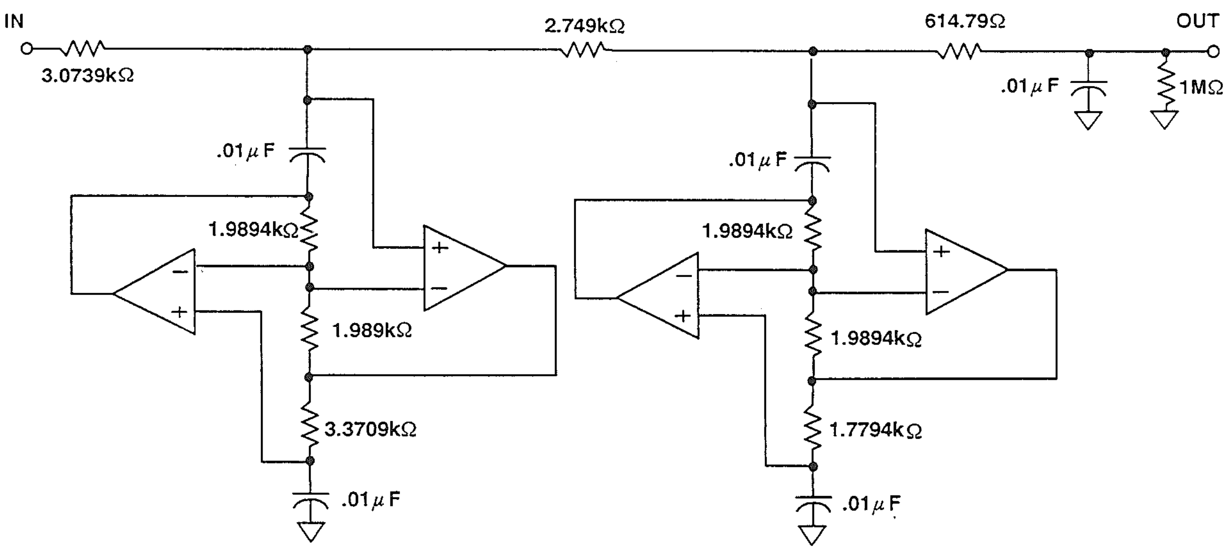

Figure 6.19

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performance. The amplifier's offset voltage will be passed by the filter and may be amplified to produce excessive output offset. For low frequency applications requiring large value resistors, bias currents flowing through these resistors will also generate an output offset voltage.

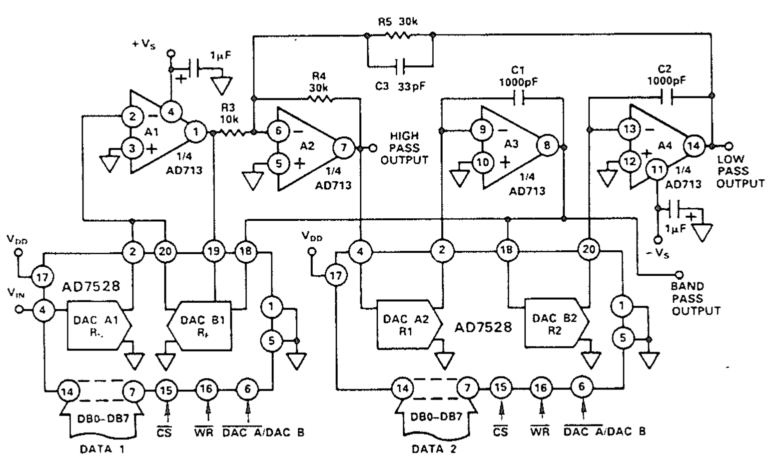
In addition, at higher frequencies, an op amp's dynamics must be carefully considered. Here, slewrate, bandwidth, and open loop gain play a major role in op amp selection. The slewrate must be fast as well as symmetrical to minimize distortion.

A Programmable State Variable Filter

A programmable state variable filter using DACs is shown in Figure 6.20. DACs A1 and B1 control the gain and Q of the filter characteristic, while DACs A2 and B2 must accurately track for the simple expression for f_c to be true. This is readily accomplished using two AD7528 DACs and one AD713 quad op amp. Capacitor C3 compensates for the effects of op amp and gain-bandwidth limitations.

This filter provides lowpass, highpass, and bandpass outputs and is ideally suited for applications where digital control of filter parameters is required. The programmable range for component values shown is f_c = 0 to 15kHz, and Q = 0.3 to 4.5.

A PROGRAMMABLE STATE VARIABLE FILTER CIRCUIT



CIRCUIT EQUATIONS

$C_1 = C_2, R_1 = R_2, R_4 = R_5$

$$f_C = \frac{1}{2\pi R_1 C_1}$$

$$Q = \frac{R_3}{R_4} \cdot \frac{R_F}{R_{FBB1}}$$

$$A_0 = -\frac{R_F}{R_S}$$

NOTE:
DAC equivalent resistance equals

$$\frac{256 \times \text{(DAC Ladder resistance)}}{\text{DAC Digital Code}}$$

Figure 6.20

SEVEN-POLE FDNR 20kHz ANTIALIASING FILTER

Figure 6.21 shows a 7-pole antialiasing filter for a 2x oversampling (88.2kSPS) digital audio application. This filter has less than 0.05dB passband ripple and 19.8 ± 0.3μs delay, dc-20kHz. The filter will handle a 5V rms signal (V_S = ±15V) with no overload at any internal nodes. The frequency response of the filter is shown in Figure 6.22.

20kHz FDNR AUDIO ANTIALIASING FILTER

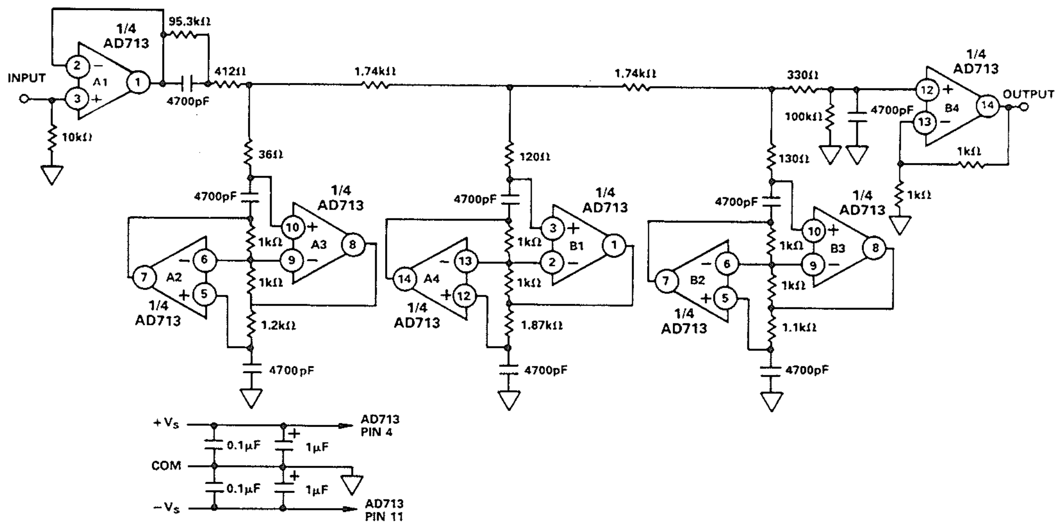


Figure 6.21

AUDIO ANTIALIASING FILTER RESPONSE

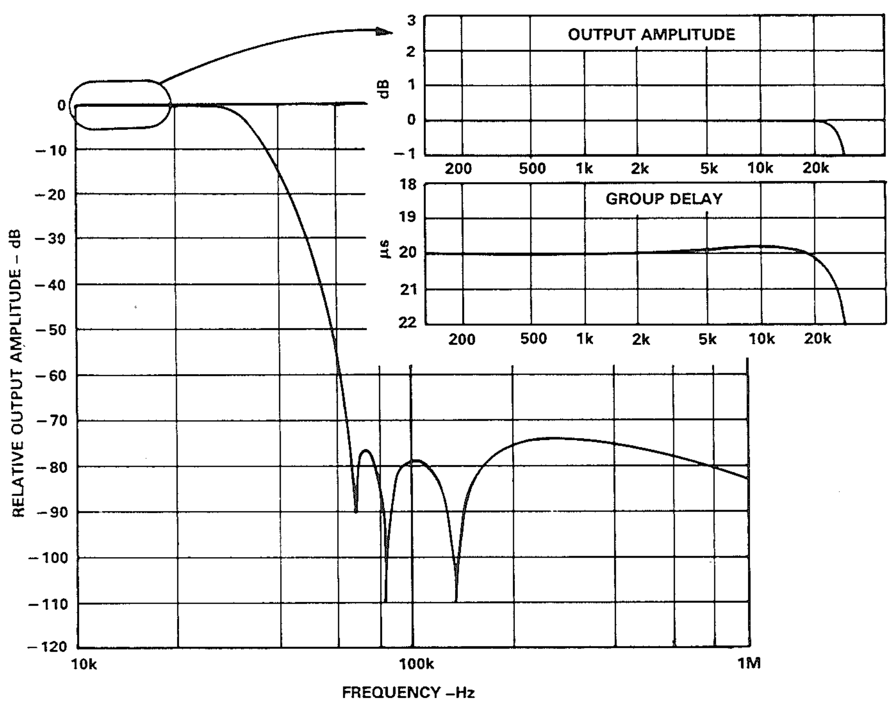


Figure 6.22

2 MHz Biquad Bandpass Filter Using a 30MHz Quad Amplifier

Figure 6.23. shows a circuit for a biquad bandpass filter with a 2MHz center frequency. This type of filter is often used in ultrasound receivers to detect a 2MHz signal while rejecting all others. The OP-467 is ideal for such an application because of its wide bandwidth and quad package. With four amplifiers, the OP-467 allows this circuit to be built using only one IC.

The 30MHz bandwidth is sufficient to accurately produce the 2MHz center frequency, as the measured response shows in Figure 6.24. Notice that the center frequency is exactly 2MHz and the gain is unity. A lower speed amplifier would cause the center frequency to shift significantly. For example, using an op amp with a 10MHz gain-bandwidth product results in 20% shift in the center frequency to 1.6MHz, even though the same component values are used. When the bandwidth is too close to the filter's center frequency, the amplifiers' internal phase shift causes excess phase shift at 2MHz, which alters the filter's response. In fact if the chosen op amp has a bandwidth close to 2MHz, the combined phase shift of the three op amps will cause the loop to oscillate. The OP-467 has a high enough bandwidth such that it contributes only a small amount of phase shift at 2MHz.

A 2MHz BIQUAD BANDPASS FILTER USING A 30MHz QUAD AMPLIFIER

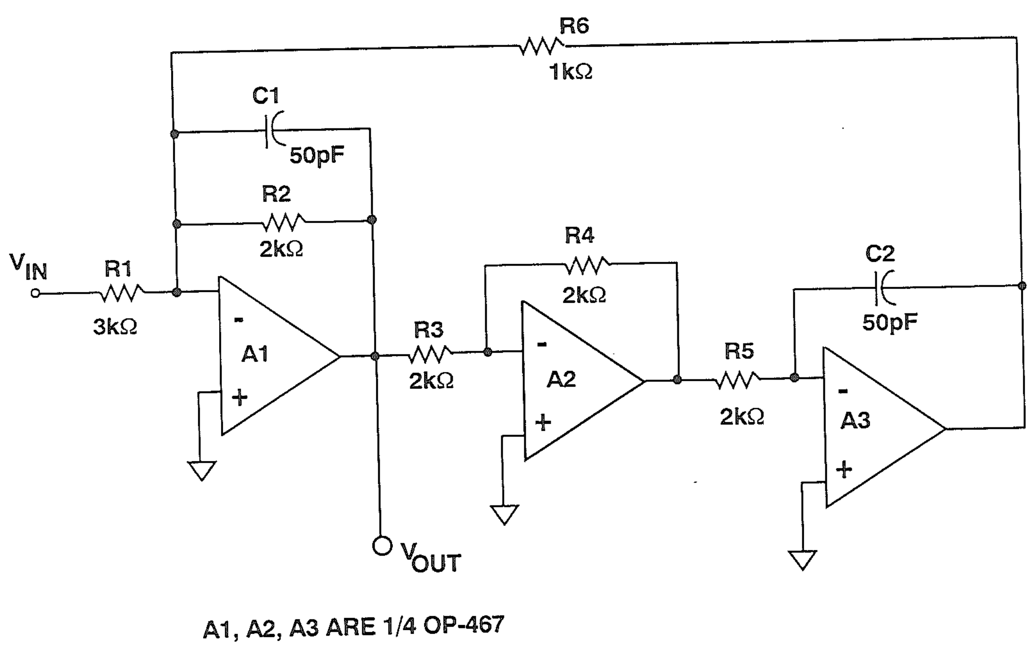


Figure 6.23

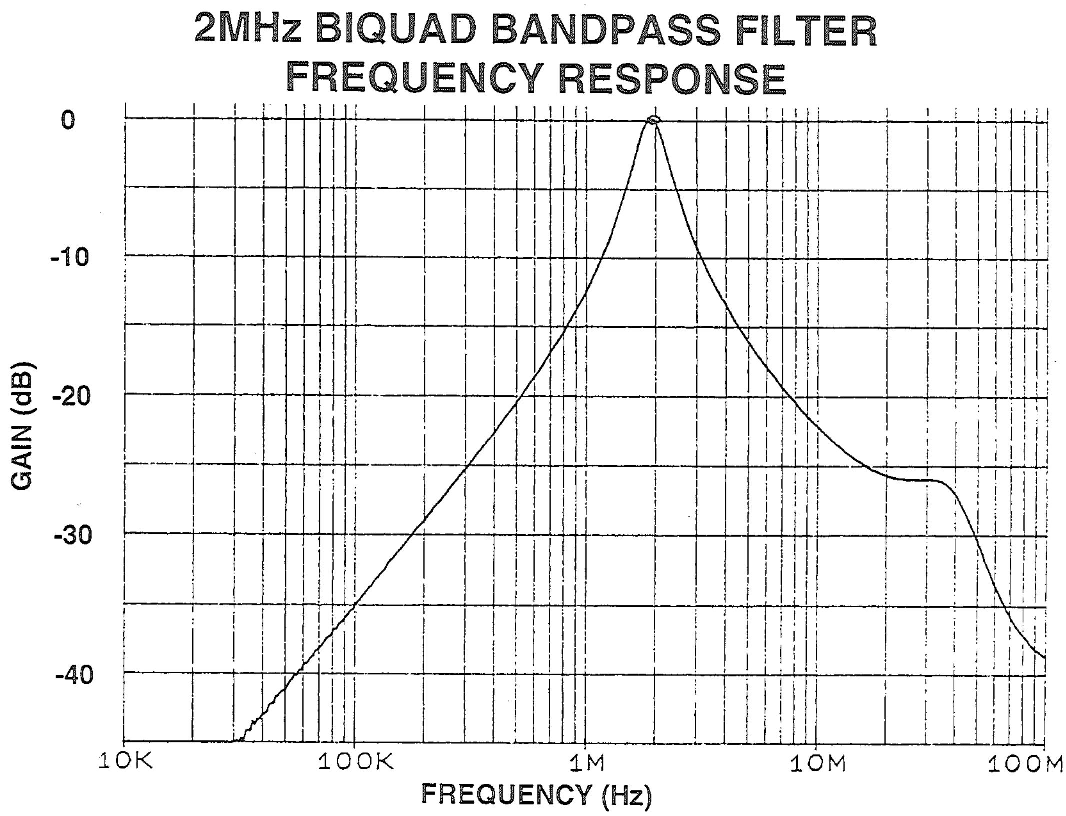


Figure 6.24

Careful consideration must be given to the layout of this circuit as with any high speed circuit. As mentioned above, if the phase shift is too large at 2MHz, the filter's response will be altered, or worse, it will oscillate. Any parasitic capacitance will cause additional phase shift. Thus care must be taken to minimize this capacitance, especially on the inverting inputs to the amplifiers. The traces connected to this node should be kept short and the ground plane removed from this area.

The OP-467 allows the entire circuit to be built with only one package, which was not possible before. Additionally, the one additional op amp in the package can be used elsewhere in the circuit to provide gain or buffering. If the op amp is not used, then it should be configure as a voltage follower with the positive input grounded and no load on the output. This ensures that it will not oscillate.

PRACTICAL PROBLEMS IN FILTER IMPLEMENTATION

In the introductory section we dealt with filters as mathematical functions. The filter designs were assumed to be implemented with "perfect" components. It's only when the filter is built with real-world components that design tradeoffs which must be made.

In building a filter with an order greater the two, multiple second and/or first order sections are used. The frequencies and Qs of these sections must align precisely or the overall response of the filter will be affected. For example, the antialiasing filter design example previously discussed is a 5th order Butterworth filter, made up of a second order section with a frequency (Fo) = 1 and a Q = 1.618, a second order section with a frequency (Fo) = 1 and a Q = .618 and a first order section with a frequency (Fo) = 1 (for a filter normalized to 1 rad/sec). If the Q or frequency response of any of the sections is off slightly, the overall response will deviate from the desired response. It may be close, but it won't be exact.

PRACTICAL CONSIDERATIONS IN FILTER IMPLEMENTATION

- Higher Order Filters Require Accurate First and Second Order Sections: Q and Frequency Response
- Passive Component Inaccuracies, Parasitics, and Drift
- Active Component Frequency Response, Input and Output Impedance, and Distortion

Figure 6.25

RESISTOR CONSIDERATIONS

- Accuracy (Using Standard Values) and Temperature Coefficient
- Select Values Between 100Ω and 1MΩ
- Use Metal Film if Possible

Figure 6.26

PASSIVE COMPONENTS (RESISTORS, CAPACITORS, INDUCTORS)

Passive components are the first problem. When designing filters, values of components are required that are not available commercially. Resistors, capacitors, and inductors come in standard values. While custom values can be ordered, the tolerance will still be +/- 1% at best. An alternative is to build the required value out of a series and/or parallel combination of standard values. This increases the cost and size of the filter. Not only is the cost of components increased, but so are the manufacturing costs, both for loading and for tuning the filter. Furthermore, its success will still be limited by the number of parts that are used, their tolerance, and their tracking. A more practical way is to use a circuit analysis program to determine the response using standard values. The program can also evaluate the effects of component drift over temperature. The values of the sensitive components are adjusted using parallel combinations where needed, until the response is within the desired limits.

In addition to the initial tolerance of the components, you must also evaluate the effects of temperature drift. The temperature coefficients of the various components may be different in both magnitude and sign. Capacitors, especially, are difficult in that not only do they drift, but the temperature coefficient is also a function of temperature as shown in Figure 6.27.

Capacitors also have temperature coefficients that vary with the value of capacitance, being higher on larger values. Some of the plastic film capacitors (notably polycarbonate) also change their nominal value permanently when heated. Again, some capacitors, mainly the plastic film types, have a limited temperature range.

CAPACITANCE CHANGE VERSUS TEMPERATURE

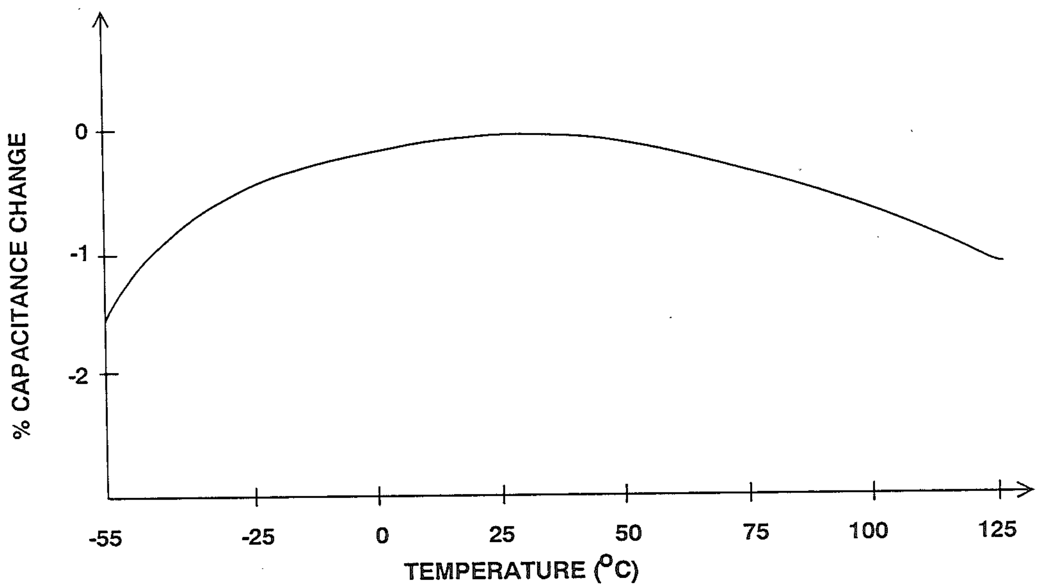


Figure 6.27

CAPACITOR CONSIDERATIONS

- Accuracy (Using Standard Values) and Temperature Coefficient
- Parasitics: Inductance, Leakage, Dielectric Absorbtion
- Select Values Between 1pF and 10µF
- Avoid Electrolytics
- Polystyrene (plastic film) is Preferable

Figure 6.28

The frequency and Q of a filter are determined by the component values. Obviously, if the component value is drifting, the frequency and the Q of the filter will drift which, in turn, will cause the frequency response to vary. This is especially true in higher order filters. Higher order means that you will have higher Q sections. Higher Q sections means that component values are more critical, since the Q is typically set by the ratio of two components, usually capacitors.

While there is infinite choice of the values of the passive components for building filters, in practice there are physical limits. Capacitor values below 1 pF and above 10 μF are not practical. Electrolytic capacitors should be avoided in circuits requiring any sort of accuracy. Electrolytic capacitors are also very leaky. If they are operated without a polarizing voltage, they become non-linear when the ac voltage reverse biases them. Even with a dc polarizing voltage, the ac signal can reduce the instantaneous voltage to 0 or below. Large values of film capacitors are physically very big.

Resistor values of less than 100 Ω should be avoided, as should values over 1 MΩ. Very low resistance values (under 100 Ω) require a great deal of drive current and dissipate a great deal of power. Both of these should be avoided. Very large values tend to be more prone to parasitics. Noise also increases with the square root of the resistor value.

Parasitic capacitances due to circuit layout and other sources affect the performance of the circuit. They exist between two traces on a PC board (on the same side or opposite side of the board), between leads of adjacent components and just about everything else you can (and in most cases can't) think of. These capacitances are usually small, so their effect is greater at high impedance nodes. Thus they can be controlled most of the time by keeping the impedance of the circuits down. Remember that the effects of stray

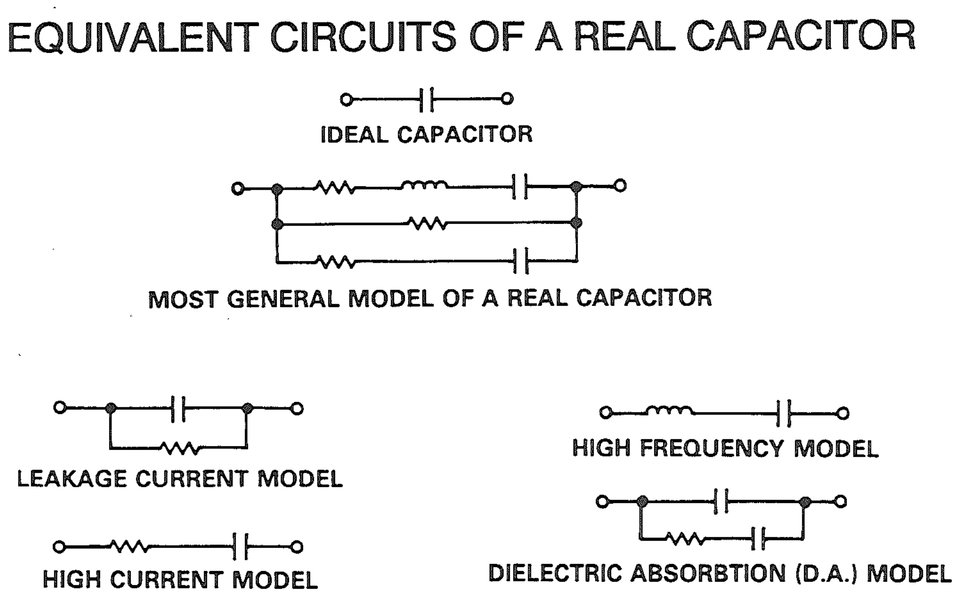


Figure 6.29

FEATURES OF COMMON CAPACITORS

TYPE	TYPICAL DIELECTRIC ABSORPTION	ADVANTAGES	DISADVANTAGES
NPO Ceramic	0.1%	Small Case Size Inexpensive Good Stability Wide Range of Values Many Vendors Low Inductance	DA too High for More than 8-Bit Applications
Polystyrene	0.001% to 0.02%	Inexpensive Low DA Available Wide Range of Values Good Stability	Destroyed by Temperature > +85°C Large Case Size High Inductance
Polypropylene	0.001% to 0.02%	Inexpensive Low DA Available Wide Range of Values	Destroyed by Temperature > +105°C Large Case Size High Inductance
Teflon	0.003% to 0.02%	Low DA Available Good Stability Operational Above +125°C Wide Range of Values	Relatively Expensive Large High Inductance
MOS	0.01%	Good DA Small Operational Above +125°C Low Inductance	Limited Availability Available only in Small Capacitance Values
Polycarbonate	0.1%	Good Stability Low Cost Wide Temperature Range	Large DA Limits to 8-Bit Applications High Inductance
Polysulfone	0.1%	Good Stability Low Cost Wide Temperature Range	Large DA Limits to 8-Bit Applications High Inductance
Monolithic Ceramic	>0.2%	Low Inductance Wide Range of Values	Poor Stability Poor DA
Mica	>0.003%	Low Loss at HF Low Inductance Very Stable Available in 1% Values or Better	Quite Large Low Values (<10nF) Expensive
Aluminium Electrolytic	High	Large Values High Currents High Voltages Small Size	High Leakage Usually Polarized Poor Stability Poor Accuracy Inductive
Tantalum Electrolytic	High	Small Size Large Values Medium Inductance Reliable	Quite High Leakage Usually Polarized Expensive Poor Stability Poor Accuracy

Figure 6.30

RESISTOR COMPARISON CHART

TYPE		ADVANTAGES	DISADVANTAGES
DISCRETE	Carbon Composition	Lowest Cost High Power/Small Case Size	Poor Tolerance (5%) Poor Temperature Coefficient (1500ppm/°C)
	Wire-Wound	Excellent Tolerance (0.01%) Excellent TC (1ppm/°C) High Power	Reactance May be a Problem Large Case Size Most Expensive
	Metal Film	Good Tolerance (0.1%) Good TC (<1 to 100ppm/°C) Moderate Cost	Must be Stabilized with Burn-In Low Power
	Bulk Metal or Metal Foil	Excellent Tolerance (to 0.005%) Excellent TC (to <1ppm/°C) Low Reactance	Low Power Very Expensive
	High Megohm	Very High Values (10^8 - 10^{14} Ω) Only Choice for Some Circuits	High Voltage Coefficient (200ppm/V) Fragile Glass Case Expensive
NETWORKS	Thick Film	Low Cost High Power Laser-Trimmable Readily Available	Fair Matching (0.1%) Poor TC (>100ppm/°C) Poor Tracking TC (10ppm/°C)
	Thin Film on Glass	Good Matching (<0.01%) Good TC (<100ppm/°C) Good Tracking TC (2ppm/°C) Moderate Cost Laser-Trimmable Low Capacitance	Delicate Often Large Geometry Low Power
	Thin Film on Ceramic	Good Matching (<0.01%) Good TC (<100ppm/°C) Good Tracking TC (2ppm/°C) Moderate Cost Laser-Trimmable Low Capacitance Suitable for Hybrid IC Substrate	Often Large Geometry
NETWORKS	Thin Film on Silicon	Good Matching (<0.01%) Good TC (<100ppm/°C) Good Tracking TC (2ppm/°C) Moderate Cost Laser-Trimmable Suitable for Monolithic IC Construction	Some Capacitance to Substrate Low Power
	Thin Film on Sapphire	Good Matching (<0.01%) Good TC (<100ppm/°C) Good Tracking TC (2ppm/°C) Laser-Trimmable Low Capacitance	Higher Cost Low Power

Figure 6.31

capacitance are frequency dependant, being worse at high frequencies because the impedance drops with increasing frequency.

Parasitics are not just associated with outside sources. They are also present in the components themselves.

A capacitor is more than just a capacitor in most instances. A real capacitor has inductance (from the leads and other sources) and resistance as shown in Figure 6.29. This resistance shows up in the specifications as leakage and poor power factor. Obviously, we would like capacitors with very low leakage and good power factor (see Figure 6.30). Resistors also have inductances.

In general it is best to use plastic film (preferably polystyrene) or mica capacitors and metal film resistors, both of moderate to low values in our filters.

One way to reduce component parasitics is to use surface mounted devices. Not having leads means that the lead inductance is reduced. Also, being physically smaller allows more optimal placement. A disadvantage is that not all types of capacitors are available in surface mount.

LIMITATIONS OF ACTIVE ELEMENTS (OP AMPS) IN FILTERS

The active element of the filter will also have a pronounced effect on the response.

In developing the various topologies (Multiple Feedback, Sallen-Key, State Variable, etc.), the active element was always modeled as a "perfect" operational amplifier. That is to say it has:

1) infinite gain
2) infinite input impedance
3) zero output impedance

none of which varies with frequency. While amplifiers have improved a great deal over the years, this model has not yet been realized.

The most important limitation of the amplifier has to due with its gain variation with frequency. All amplifiers are band limited. This is due mainly to the physical limitations of the devices with which the amplifier is constructed. Negative feedback theory tells us that the response of an amplifier must be first order (-6 dB per octave) when the gain falls to unity in order to be stable. To accomplish this, a real pole is usually introduced in the amplifier so the gain rolls off to < 1 by the time the phase shift reaches 180 degrees (plus some phase margin, hopefully). This roll off is equivalent to that of a single pole filter. So in simplistic terms, the transfer function of the amplifier is added to the transfer function of the filter to give a composite function. How much the frequency dependant nature of the op amp affects the filter is dependant on which topology is used.

The Sallen-Key configuration, for instance, is the least dependant on the frequency response of the amplifier. All that is required is for the amplifier response to be flat to just past the frequency where the attenuation of the filter is below the minimum attenuation required. This is because the amplifier is used as a gain block. Beyond cutoff, the attenuation of the filter is reduced by the rolloff of the gain of the op amp. This is because the output of the amplifier is phase shifted, which results in incomplete nulling when fed back to the input.

The state variable configuration uses the op amps in two modes, as amplifiers and as integrators. As amplifiers, the constraint on frequency response is the same as for the Sallen-Key, that is flat out to the minimum attenuation frequency. As an integrator, however, more is required. A good rule of thumb is that the open loop gain of the amplifier must be greater than 10 times the closed loop

CONSIDERATIONS FOR OP AMPS

- Finite Gain-Bandwidth
- Input Impedance
- Output Impedance
- Distortion

6

Figure 6.32

SALLEN KEY SENSITIVITY TO OP AMP

- Least Dependent on Op Amp Frequency Response.
  Op Amp Used As Gain Block
- Op Amp Flat to Slightly Beyond Filter Stopband Frequency
- Can Use Current Feedback Op Amps in Sallen Key Configuration

Figure 6.33

gain. This should be taken as the absolute minimum requirement. What this means is that there must be 20 dB loop gain, minimum. Therefore, an op amp with 10 MHz unity gain bandwidth can be used to make a 1 MHz integrator. What happens is that the effective Q of the circuit increases as loop gain decreases. This phenomenon is called Q enhancement. The mechanism for Q enhancement is similar to that of slew rate limitation. Without sufficient loop gain, the op amp virtual ground is no longer at ground. In other words, the op amp is no longer behaving as a op amp. Because of this, the integrator no longer behaves like an integrator.

The multiple feedback configuration also places heavy constraints on the active element. Q enhancement is a problem in this topology as well. As the loop gain falls, the Q of the circuit increases, and the parameters of the filter change. The same rule of thumb as used for the integrator apply to the multiple feedback topology (loop gain should be at least 20 dB).

In the FDNR realization, the requirements for the op amps are not as clear. To make the circuit work, we assume that the op amps will be able to force the input terminals to be the same voltage. This implies that the lcop gain be a minimum of 20 dB at the resonant frequency. Also it is generally considered to be advantageous to have the two op amps in each leg matched. This is easily accomplished using dual op amps. It is also a good idea to have low bias current devices for the op amps, so FET input op amps should be used, all other things being equal.

STATE VARIABLE SENSITIVITY TO OP AMP

- Op Amps Used as Amplifiers and Integrators
- Amplifiers: Flat to Beyond Stopband Frequency
- Integrators: 20dB Minimum Loop Gain (Open Loop Gain 10 Times Closed Loop Gain), or Q Enhancement Occurs
- 10MHz Unity-Gain Bandwidth Op Amp Can Make a 1MHz Integrator

Figure 6.34

MULTIPLE FEEDBACK SENSITIVITY TO OP AMP

- Same Rule of Thumb as for Integrators: 20dB Minimum Loop Gain to Prevent Q Enhancement

6

Figure 6.35

In addition to the frequency dependant limitations of the op amp, several other of its parameters may be important to the filter designer.

One is input impedance. As we said, the filter topologies assume "perfect" amplifiers. This implies that the input impedance is infinite. This is required so that the input of the op amp does not load the network around it. This means that we probably want to use FET amplifiers with high impedance circuits. There is also a small frequency dependant term to the input impedance, since the effective impedance is the real input impedance multiplied by the loop gain. This usually is not a major source of error, since the network impedance of a high frequency filter should be low.

Similarly, the op amp output impedance affects the response of the filter. The output impedance of the amplifier is divided by the loop gain, therefore the output impedance will rise with increasing frequency. This may have an effect with high frequency filters if the output impedance of the stage driving the filter becomes a significant portion of the network impedance.

The fall of loop gain with frequency can also effect the distortion of the op amp, since there is less loop gain available for correction. In the multiple feed back configuration the feedback loop is also frequency dependant, which may further reduce the feedback correction, resulting in increased distortion. This effect is counteracted somewhat by the reduction of distortion components in the filter network (assuming a lowpass or bandpass filter).

All of the discussion so far is based on using classical voltage feedback op amps. Current feedback, or transimpedance, op

FDNR SENSITIVITY TO OP AMP

- Loop Gain Should be 20dB Minimum at Resonant Frequency
- Match Op Amps in Each Leg (Use Duals)
- Use FET Input Op Amps for Low Bias Currents

Figure 6.36

OTHER OP AMP CONSIDERATIONS IN FILTERS

- Input and Output Impedance
- Distortion at Low Loop Gains
- Noise
- Dynamic Range (Overload), Q times Gain < Loop Gain
- Current Feedback Op Amps May Only be Used in Sallen Key Configuration

Figure 6.37

amps offer improved high frequency response but are unusable in any topology except the Sallen-Key. The problem is that capacitance in the feedback loop of a current feedback amplifier usually causes it to become unstable. Also, most current feedback amplifiers will only drive small capacitive loads. Therefore, it is difficult to build classical integrators using current feedback amplifiers. Some current feedback op amps have an external pin which may be used to configure them as a very good integrator, but this configuration does not lend itself to classical active filter designs. Current feedback integrators are non-inverting, which is not acceptable in the state variable configuration. Also, the bandwidth of a current feedback amplifier is set by its feedback resistor which would make the Multiple Feedback topology difficult to implement. Another limitation of the current feedback amplifier in the Multiple Feedback configuration is the low input impedance of the inverting terminal. This would result in loading of the filter network. Sallen-Key filters are possible with current feedback amplifiers, since the amplifier is used as a non-inverting gain block. New topologies which capitalize on the current feedback amplifiers superior high frequency performance and compensate for its limitations will have to be developed.

The last thing that you need to be aware of is exceeding the dynamic range of the amplifier. Qs over 0.707 will cause peaking in the response of the filter (see Figure 6.38). For high Qs, this could cause overload of the input or output stages of the amplifier with a large input. Note that relatively small values of Q can cause significant peaking. The Q times the gain of the circuit must stay under the loop gain (plus some margin, again 20 dB is a good starting point). This holds for multiple amplifier topologies as well. Be aware of internal node levels, as well as input and output levels. As an amplifier overloads, its effective Q decreases, so the transfer function will appear to change even if the output appears undistorted. This shows up as the transfer function changing with increasing input level.

6

EFFECT OF FILTER Q ON GAIN PEAKING

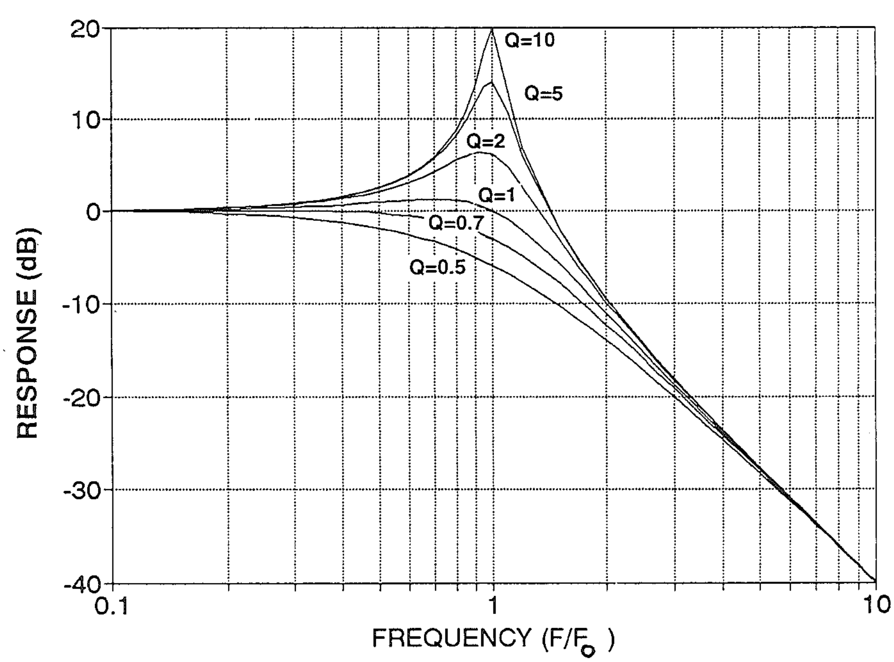


Figure 6.38

We have been dealing mostly with lowpass filters in our discussions, but the same principles are valid for highpass, bandpass, and band reject as well. In general, things like Q enhancement and limited gain/bandwidth will not effect highpass filters, since the resonant frequency will probably be low in relation to the cutoff frequency of the op amp. Remember, though, that the highpass filter will have a low pass section, by default, at the cutoff frequency of the amplifier. Bandpass and bandreject (notch) filters will be affected, especially since both tend to have high values of Q. The general effect of the op amp's frequency response on the filter Q is shown in Figure 6.39.

As an example of the Q enhancement phenomenon, consider the Spice simulation of a 10 kHz bandpass Multiple Feedback filter with Q = 10 and gain = 1, using a good high frequency amplifier (the AD847) as the active device. The circuit diagram is shown in Figure 6.40. The open loop gain of the AD847 is greater than 70 dB at 10 kHz as shown in Figure 6.41. This is well over the 20 dB minimum, so the filter works as designed as shown in Figure 6.42. We now replace the AD847 with an OP-90. The OP-90 is a dc precision amplifier and so has a limited bandwidth. In fact, its open loop gain is less than 10 dB at 10 kHz (see Figure 6.41). This is not to imply that the AD847 is in all cases better than the OP-90. It is a case of misapplying the OP-90. From the output for the OP-90, also shown in Figure 6.42, we see that the magnitude of the output has been reduced, and the center frequency has shifted downward.

Q ENHANCEMENT

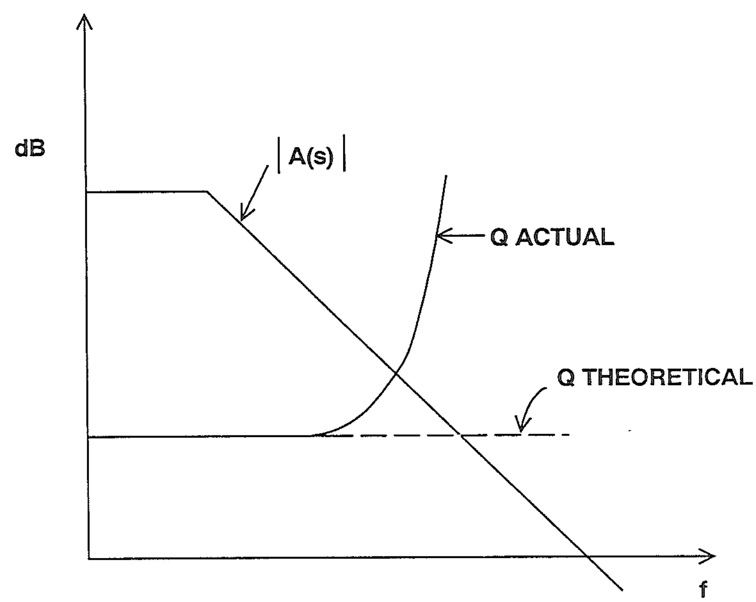
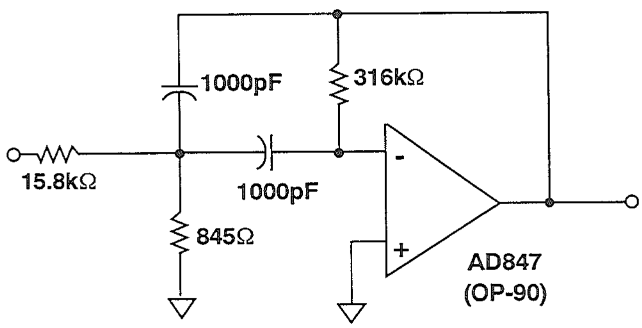


Figure 6.39

10KHz MULTIPLE FEEDBACK BANDPASS FILTER



Q = 10
H = 1

Figure 6.40

OPEN LOOP GAIN OF AD847 AND OP-90 OP AMPS

AD847

OP-90

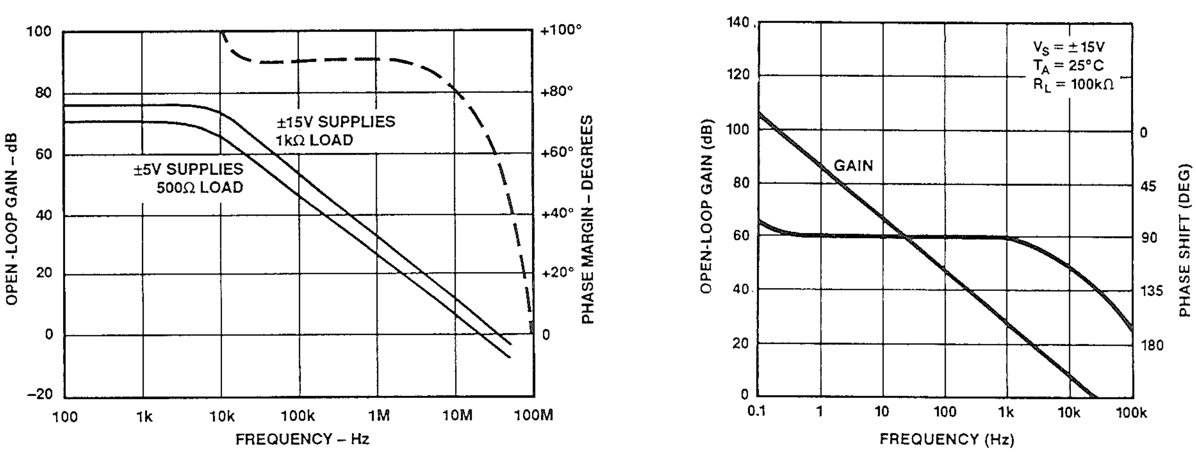


Figure 6.41

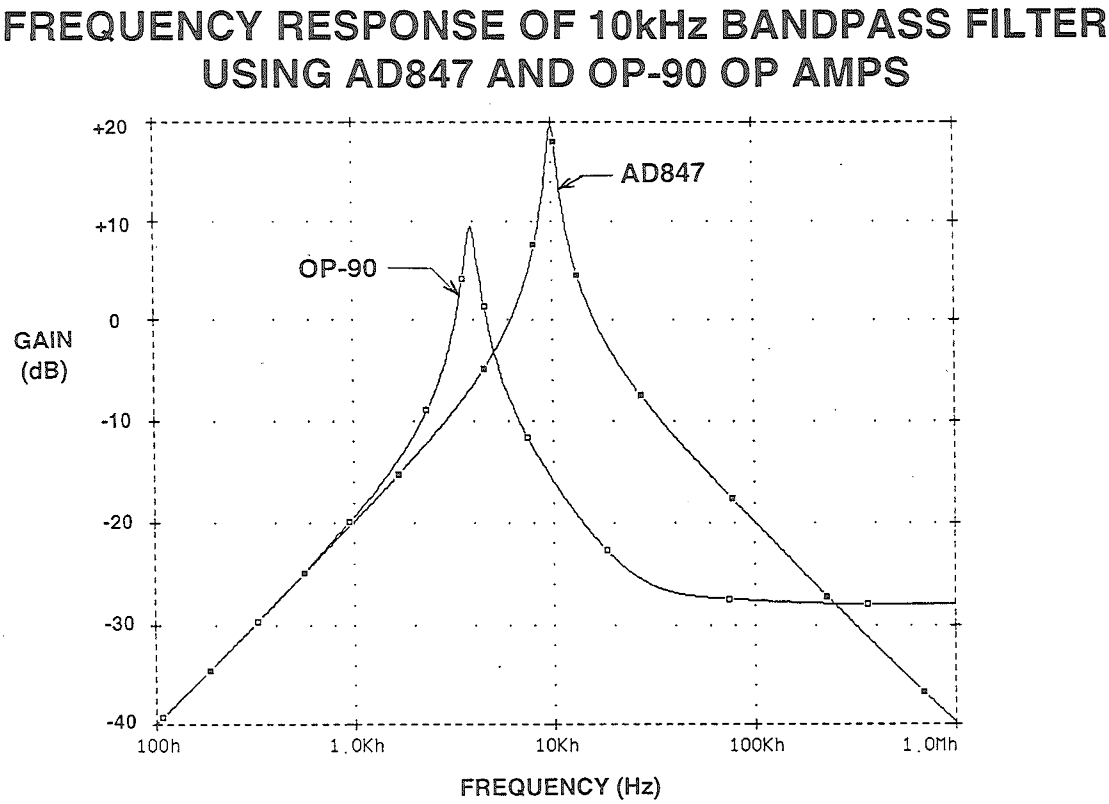


Figure 6.42

A 12MHz Sallen-Key Filter Using a Current Feedback Amplifier

As a demonstration of a high frequency filter we will design a 12 MHz Sallen-Key filter using a transimpedance amplifier (the AD9617) as the active device (see Figure 6.43). Here we see that the corner frequency of the filter is slightly less that the design value of 12MHz (see Figure 6.44). Also note that the attenuation is only good to about 100 MHz. This is where the loop gain of the amplifier starts to decrease. Now, since the output of the amplifier is phase shifted relative to its input, there is incomplete nulling of the signal. Also, as the impedance of the filter network decreases, it reaches the point where it starts to bypass the amplifier.

12 MHz SALLEN-KEY LOWPASS FILTER WITH BUTTERWORTH RESPONSE USES CURRENT FEEDBACK AMPLIFIER

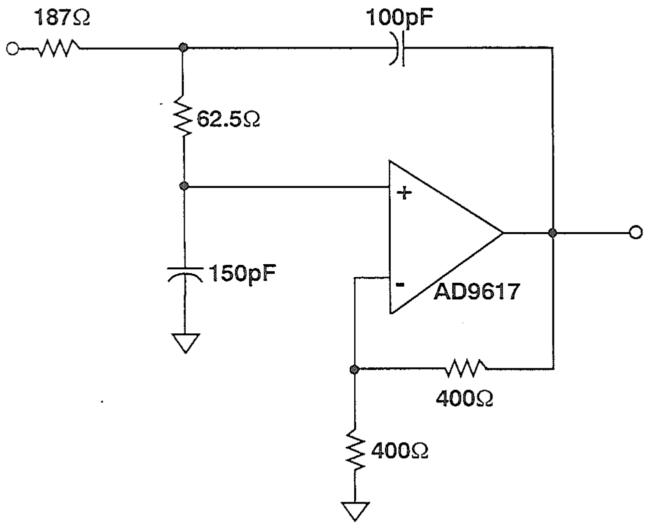


Figure 6.43

FREQUENCY RESPONSE OF 12MHz SALLEN-KEY LOWPASS FILTER

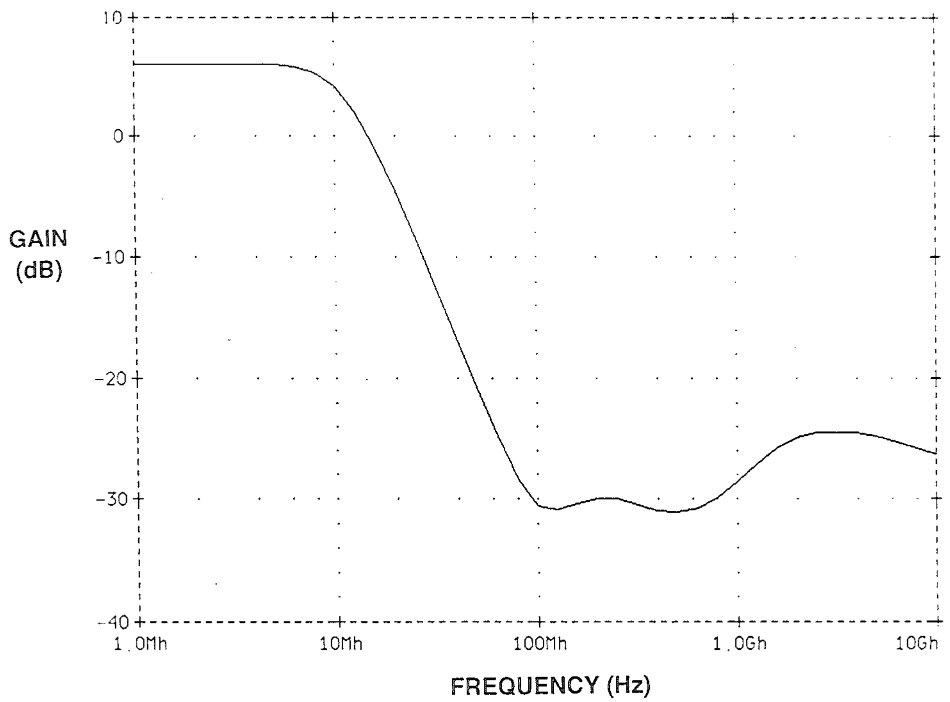


Figure 6.44

6

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