Chapter 12

Audio Signal Processing and Anti-Aliasing Filters

12.1 Introduction

Two important areas of analog filter applications are included in this chapter: audio signal processing (ASP) and study of anti-aliasing/reconstruction (AA/ReF) filters.

In ASP, the frequency range of operation is from a few Hz to around 20 kHz. All the different types of filters, that is, LP (low pass), BP (band pass), HP (high pass), BE (band elimination) and AP (all pass) filters, can be used. Hence, in most applications, general purpose OAs (operational amplifiers) may be used while taking care of noise signals. Generally, order of the filters may not be very high; whereas, in AA/ReF, only LPFs are used for which the frequency range of operation depends on specific applications. Hence, the selection of OA is to be considered carefully. Order of the filter can also be high.

One of the major objectives in ASP is to improve audio signals employing different approaches. Section 12.2 discusses this aspect in terms of crossover networks and filters eliminating infrasonic and ultrasonic undesired signals. Section 12.3 discusses a specific case which addresses the unequal power output requirements of the satellite speaker and the subwoofer speaker system. Equalization using mid treble boost/cut and mid bass boost/cut, which are part of programmable equalizers are studied in Section 12.4. Section 12.5 takes up certain aspects of the Record Industry Association of America (RIAA) in brief. An application of phase approximated filters in ASP is explained in Section 12.6.

The essential nature of AA/ReF filters is described in Section 12.7 citing the example of a filter design for a 12-bit ADC. A detailed case is presented in Sections 12.8 and 12.9 showing the application of high frequency AA and ReF filters used in video signal processing.

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12.2 Improvement in Audio Performance

Problems in the frequency response of audio systems and loudspeakers are resolved using active analog audio filters. Of course, the ways these filters are used depend on the required end result. For example, in some cases, filters are required to produce a response which is nearly the inverse of the response of the (say) loudspeaker or any other audio speaker. Hence, when the two responses are combined, it results in a nearly flat response. In some applications, undesirable signal peaks coming from music systems, cell phones, and so on, are to be reduced. Figure 12.1 illustrates such a compensation in a PDA (public digital assistance) speaker response [12.1]. Obviously, such compensations are used to improve the audio response; the compensations make the signals more pleasing and intelligible than the original.

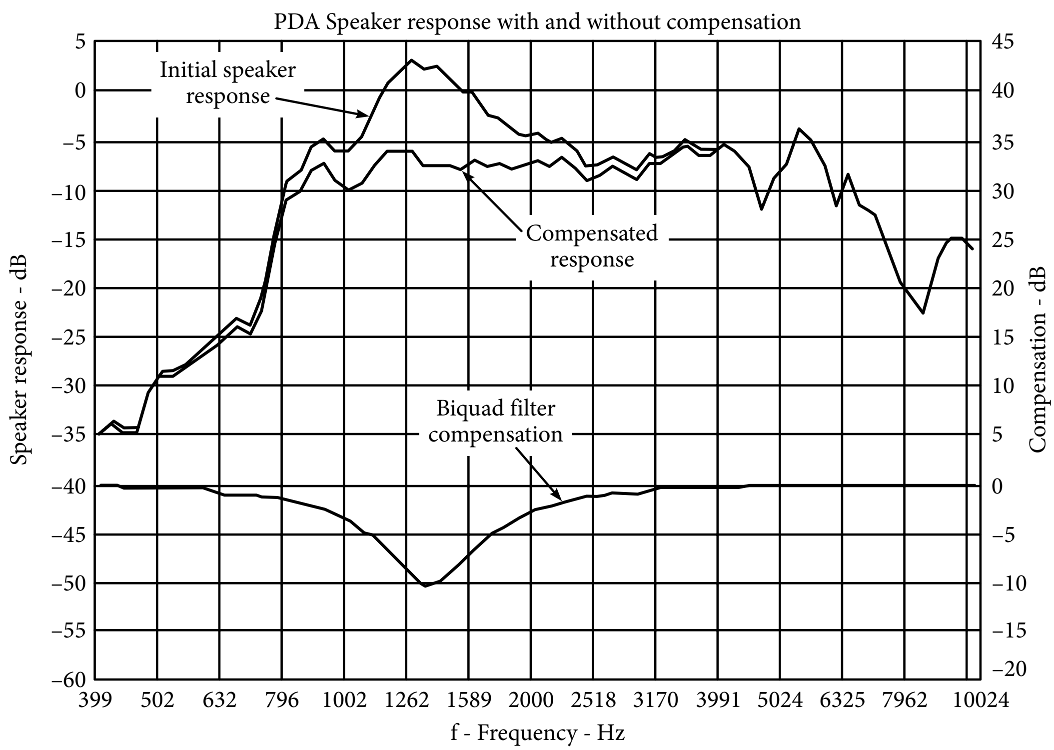

Figure 12.1 PDA response with and without compensation {Courtesy: Texas Instruments}.

Audio filters may be first-, second- or higher-order with a number of choices being available as shown in the earlier chapters, and vast literature is available elsewhere. Before considering some practical examples of audio filters, a brief introduction about sound measuring instruments is desirable [12.2].

In sound level meters (SLMs), the electrical signal from a transducer (say microphone) is fed to a pre-amplifier, and then to a weighted filter over a specified range of frequencies,

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if needed. After that, the filtered signal is further amplified and the output signal is either directly measured or it acts as input to some other device. It is to be noted that noise (sound) measuring instruments are widely used in the practice of occupational hygiene [12.2].

Frequency analyzers are used to determine the distribution of the overall signal level over a range of frequencies. For occupational hygiene, the most usual analysis for noise studies is done in the octave band. A number of frequency analyzers are available for use with SLMs. Quality analyzers can perform frequency analysis in all desired frequency bands simultaneously. Characteristics of the filters play an important role in quality analyzers. In most cases, filter attenuation is at the rate of 24 dB per doubling of frequency after the 3 dB level.

Improvement in audio performance is a subject of continuous study. The application report [12.3] is one such example. Along with a few other applications of LM 833 OA, the report contains some useful continuous-time filters.

12.2.1 Active crossover networks for loudspeakers

Loudspeaker systems generally consist of two or more transducers for different frequency bands of the audio frequency spectrum. Initially, passive filters were used as crossover filters. However, these filters required inductors and capacitors of large values. Such passive filters created noise and distortion; the system efficiency was also considerably poor.

A straightforward approach was to use low-level filters to divide the frequency spectrum. Separate power amplifiers then follow each driver or group of drivers. Though this arrangement appears to be obvious, it is very difficult to find a commercially feasible active dividing network. Fortunately, the objective could be achieved by employing a *constant voltage* crossover circuit as shown in Figure 12.2(a) [12.3]. The term *constant voltage crossover* means that once outputs of the LP and HP sections are added, the exact replica of the input signal will be produced. For the circuit in Figure 12.2(a), analysis gives the following relations:

$$V_{\mathrm{HP}} = -(V_{\mathrm{in}} + V_{\mathrm{LP}}),\ V_1 = -(V_{\mathrm{in}} + V_{\mathrm{LP}})/sCR,\ V_2 = (V_1 + 4V_{\mathrm{HP}})/sCR \tag{12.1}$$

The resulting transfer functions for the LP and HP functions are as follows:

$$\frac{V_{\mathrm{LP}}(s)}{V_{\mathrm{in}}(s)} = \frac{a_1 s + 1}{a_3 s^3 + a_2 s^2 + a_1 s + 1} \tag{12.2a}$$

$$\frac{V_{\mathrm{HP}}(s)}{V_{\mathrm{in}}(s)} = \frac{a_3 s^3 + a_2 s}{a_3 s^3 + a_2 s^2 + a_1 s + 1} \tag{12.2b}$$

With the following selected values of coefficients, $a_3 = 1$, $a_2 = 4$ and $a_1 = 4$, denominator for both the mixed LP and HP filters become third-order as shown in equation (12.3).

$$D(s) = (s + 1)(s^2 + 3s + 1) \tag{12.3}$$

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For a second-order filter, expression for the critical frequency

$$f_c = (1/2\pi RC) \tag{12.4}$$

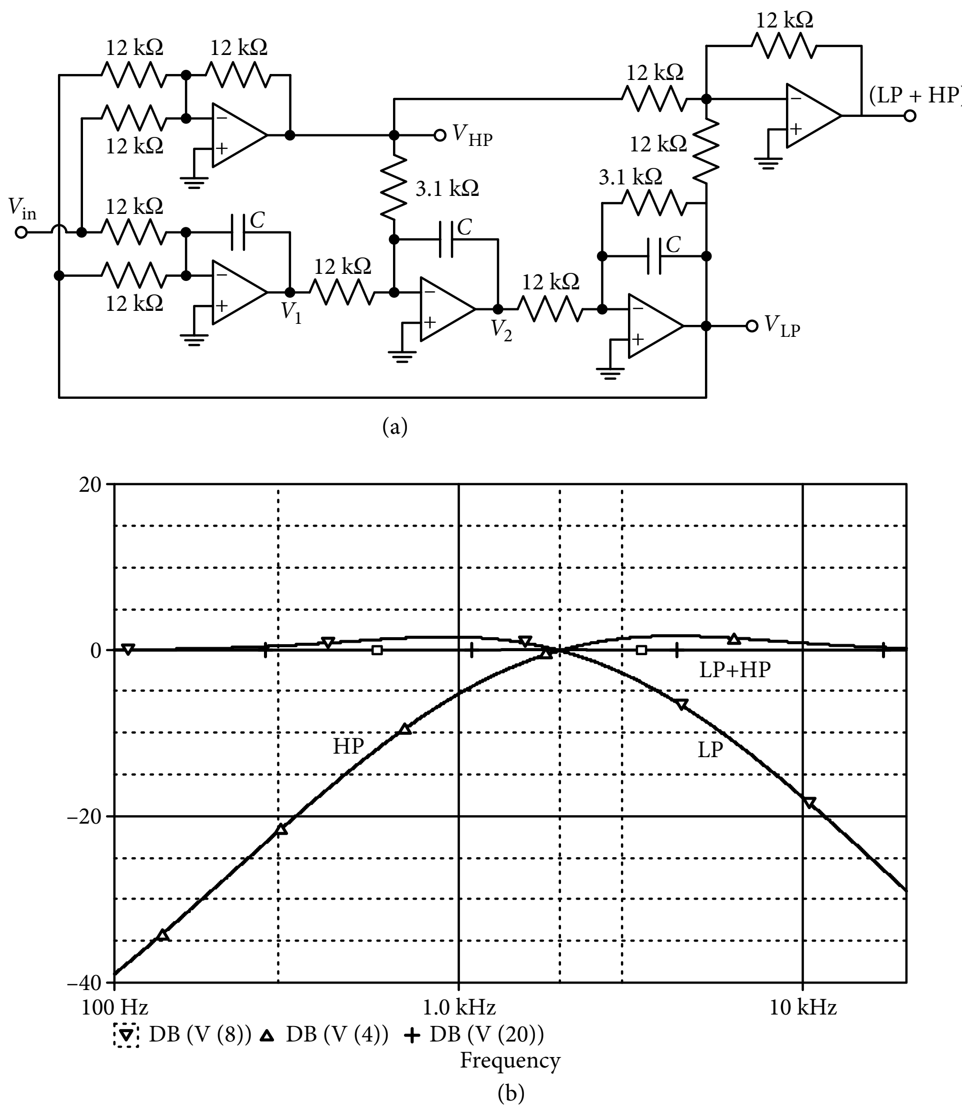

Figure 12.2 (a) Active crossover filter for loudspeakers {Courtesy: Texas Instruments}. (b) Simulated response of the crossover filter shown in Figure 12.2(a).

Hence, for a selected crossover frequency of 2 kHz, with $R = 12$ kΩ we get $C = 6.628$ nF. To get the value of the coefficients a_1, a_2 and a_3, as mentioned earlier, two resistors will have their value as 0.25 $R = 3$ kΩ. If crossover frequency is to be changed, either the three capacitors are to be changed or all the resistors are to be changed simultaneously.

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Figure 12.2(b) shows the simulated response of the LP and HP constant voltage crossover outputs and the summed output of the two responses. The final output is flat with a 0 dB gain as desired. The simulated crossover frequency is 1.9979 kHz.

12.2.2 Infrasonic and ultrasonic filters

There are a number of significant sources of noise signals above and below the limits of audibility. For example, the most significant source of noise in the low-frequency range is due to a phonograph arm/cartridge/disc combination. A low frequency, large amplitude signal at 0.556 Hz is also generated when the disc warps on $33\frac{1}{3}$ rpm records.

Under certain undesirable operating conditions, there are considerable noise signals beyond the audible frequency range as well. These ultrasonic noise signals affect the circuit elements in the audio band as well, and needs to be eliminated. Circuits shown in Figures 12.3 and 12.4 attenuate infrasonic and ultrasonic noise signals. The circuit in Figure 12.3 is a third-order Butterworth HPF with a cut-off frequency of 15 Hz. It is expected that the filter provides attenuation over 28 dBs at 5 Hz, while at 30 Hz, information is to be attenuated by only 0.1 dB. For the HPF shown in Figure 12.3, the following relations are obtained:

$$V_1(sC_1 + G_1 + sC_2) - V_2sC_2 - V_{\text{in}}sC_1 = 0 \tag{12.5a}$$

$$V_2(sC_2 + G_2 + sC_3) - V_1sC_2 - V_{\text{out}}sC_3 = 0 \tag{12.5b}$$

$$V_{\text{out}}(G_3 + sC_3) - V_2sC_3 = 0 \tag{12.5c}$$

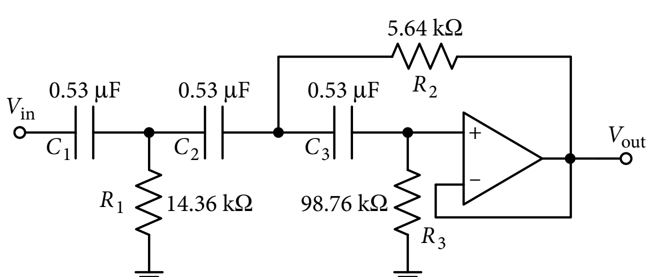

Figure 12.3 Third-order Butterworth HP filter for eliminating infrasonic noise {Courtesy: Texas Instruments}.

From equation (12.5), the transfer function of the third-order Butterworth filter is obtained as:

$$\frac{V_{\text{out}}}{V_{\text{in}}} = \frac{s^3}{s^3 + d_1s^2 + d_2s + d_3} \tag{12.6a}$$

With $d_1 = \left(\frac{1}{C_1R_1} + \frac{1}{C_1R_3} + \frac{1}{C_2R_3} + \frac{1}{C_3R_3}\right)$ (12.6b)

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$$d_2 = \left(\frac{1}{C_1 R_1 C_2 R_3} + \frac{1}{C_1 R_1 C_3 R_3} + \frac{1}{C_1 R_2 C_3 R_3} + \frac{1}{C_3 R_3 C_2 R_3} \right) \tag{12.6c}$$

$$d_3 = \frac{1}{C_1 R_1 C_2 R_2 C_3 R_3} \tag{12.6d}$$

$$\frac{V_{\text{out}}}{V_{\text{in}}} = \frac{s^3}{s^3 + \left(\frac{1}{R_1} + \frac{3}{R_3} \right) s^2 + \left(\frac{1}{R_1} + \frac{1}{R_2} \right) s \frac{2}{CR_3} + \frac{1}{R_1 R_2 R_3 C^3}} \text{for } C_1 = C_2 = C_3 = C \tag{12.6e}$$

Normalized transfer function of the third-order HP Butterworth filter is:

$$H(s) = \frac{s^3}{s^3 + 2s^2 + 2s + 1} \tag{12.7}$$

Using the coefficient matching approach with equations (12.6) and (12.7), the normalized element values are obtained as:

$$R_1 = 0.718,\ R_2 = 0.282,\ R_3 = 4.938 \text{ and } C = 1 \tag{12.8}$$

Frequency de-normalization is done by a factor of $15 \times 2\pi$ rad/s and by an impedance scaling factor of 20 kΩ. It resulted in the element values for the filter as:

$$R_1 = 14.36\ \text{k}\Omega,\ R_2 = 5.46\ \text{k}\Omega,\ R_3 = 98.76\ \text{k}\Omega \text{ and } C = 0.5303\ \mu\text{F} \tag{12.9}$$

The circuit in Figure12.4 is a fourth-order Bessel filter having negligible attenuation of the phase with an expected magnitude attenuation of 0.65 dB at 20 kHz because its design cut-off frequency is 40 kHz.

The transfer function of a fourth-order LP Bessel filter is obtained from Table 4.2 as:

$$H_B(s) = \frac{105}{(s^2 + 5.79242\text{s} + 9.14013)(s^2 + 4.20758 + 11.4878)} \tag{12.10}$$

The Bessel filter is realized as a cascade of two second-order filters. The normalized parameters of the two sections are evaluated from the equation (12.10) as:

$$w_{o1} = 3.023 \text{ rad/s},\ Q_1 = 0.522,\ w_{o2} = 3.389 \text{ rad/s},\ Q_2 = 0.806 \tag{12.11}$$

Else, parameters of the two sections are also available from table 4.3. The low frequency gain for both the sections will be unity. For the realization of the two sections, the circuit shown in Figure 7.22 is used with unity gain. Using equations (7.64) and (7.65), the normalized elements are calculated as:

$$G_{11} = 2,\ G_{12} = 7.204,\ C_{11} = 0.7518,\ C_{12} = 1 \tag{12.12a}$$

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$$G_{12} = 5,\ G_{22} = 5.98,\ C_{12} = 1.465,\ C_{22} = 1 \tag{12.12b}$$

Frequency de-normalization by a factor of $2\pi \times 40$ krad/s and impedance scaling factor of 3.977 kΩ was used. Calculated values are shown in Figure 12.4.

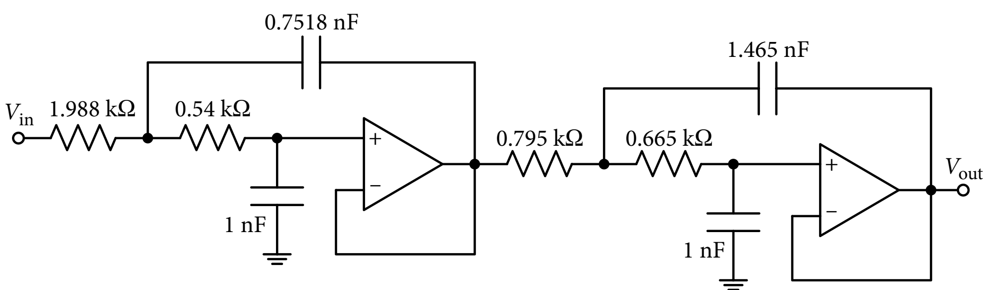

Figure 12.4 Fourth-order Bessel filter for eliminating ultrasonic noise {Courtesy: Texas Instruments}.

The filters in Figures 12.3 and 12.4 have been simulated using economical 741 OAs. However, to realize these filters with extremely low third-harmonic distortion, LM 833 is preferred [12.3]. The simulated amplitude response of the two filters in cascade, where LPF precedes HPF is shown in Figure 12.5. The simulated cut-off frequency of the HPF is 14.9 Hz, and at 5 Hz attenuation is 28.6 dBs; whereas at 30 Hz, attenuation is only 0.063 dB. For the LPF case, cut-off frequency is 46.62 kHz and attenuation is only 0.539 dB at 20 kHz.

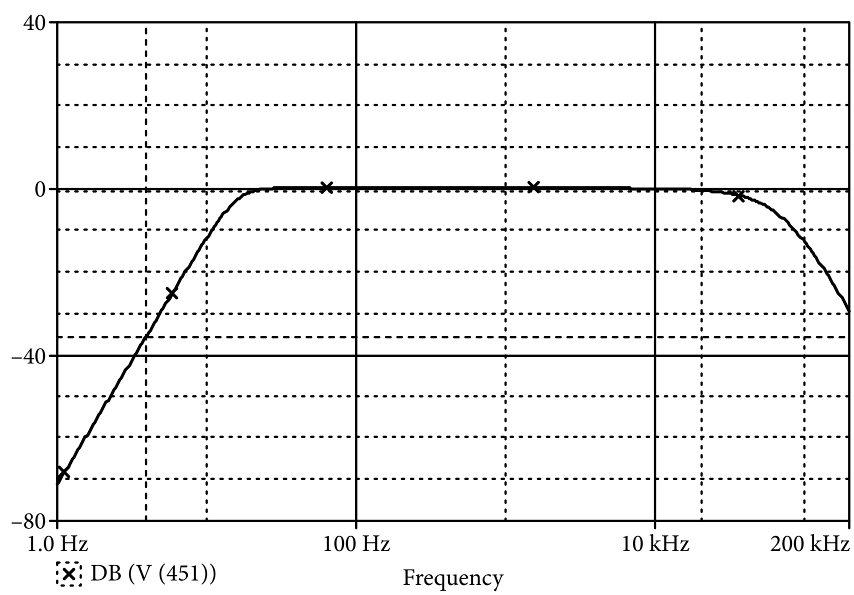

Figure 12.5 The combined response of the infrasonic and ultrasonic filters in Figures 12.3 and 12.4.

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12.3 Satellite/Subwoofer Speaker Systems

We now address the unequal output power requirements of satellite speakers and the subwoofer. Limitations of the conventional solutions have been claimed to be overcome in an application note [12.4]. Solution of the unequal power requirement in audio systems is shown in Figure 12.6. Here, a MAXIM 9791 Window Vista®-complaint 2 × 2 W stereo amplifier with stereo head driver and the MAXIM 9737 is used, which is a 1 × 7 W mono class D amplifier.

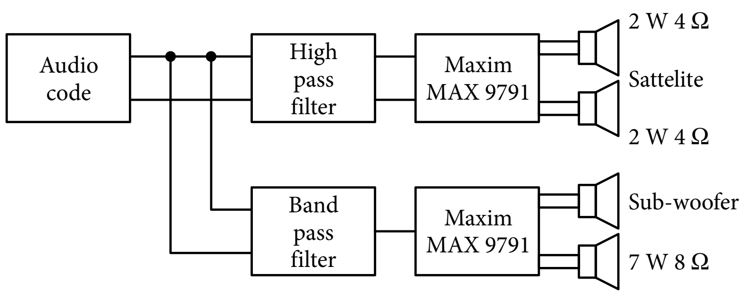

Figure 12.6 Utilization of analog filters in an audio system with unequal output power {Courtesy: Maximum Integrated}.

Presently, we are interested in the utilization of HP and BP filtering shown in the arrangement in Figure 12.6.

HPF with a 3-dB frequency of 500 Hz with a unity gain Butterworth response is used. It employs the filter structure of Figure 7.8(a). Utilization of equation (7.33) gives normalized component values:

$$C_1 = C_3 = C_4 = 1,\ R_2 = 0.4714 \text{ and } R_5 = 2.121 \tag{12.13}$$

For a cut-off frequency of 500 Hz, a frequency scaling factor of 1000π rad/s is applied, and an impedance scaling factor of 50 kΩ is selected. The resultant element values are given in equation (12.14a).

$$C_1 = C_3 = C_4 = 6.363 \text{ nF},\ R_2 = 23.56 \text{ k}\Omega \text{ and } R_5 = 106.15 \text{ k}\Omega \tag{12.14a}$$

A resistance of 470 Ω is connected in series with the capacitance C_1, which is required to minimize the capacitive loading caused by the filter stage. Figure 12.7(a) shows the circuit structure and the simulated response in Figure 12.7(b) confirming the design. Two more responses with cut-off frequencies of 400 Hz and 600 Hz are also shown in the figure. Corresponding values of the elements for these responses are:

$$f_c = 400 \text{ Hz},\ C_1 = C_3 = C_4 = 9.943 \text{ nF},\ R_2 = 18.84 \text{ k}\Omega,\ R_5 = 84.92 \text{ k}\Omega \tag{12.14b}$$

$$f_c = 600\text{Hz},\ C_1 = C_3 = C_4 = 4.419 \text{ nF},\ R_2 = 28.27 \text{ k}\Omega,\ R_5 = 127.38 \text{ k}\Omega \tag{12.14c}$$

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Similarly, the BF filter which is used for the subwoofer is constructed by combining a second-order LPF and a second-order HPF. Cut-off frequencies of the HPF and the LPF range between 80 Hz and 100 Hz, and between 100 and 500 Hz, respectively. For the LPF, the second-order structure of Figure 7.6 is used. From equation (7.29), relations for the parameters are obtained as:

$$\omega_o^2 = \frac{G_3 G_4}{C_2 C_5}, \frac{\omega_o}{Q} = \frac{(G_1 + G_3 + G_4)}{C_2} \text{ and dc gain} = \frac{G_1}{G_4} \tag{12.15}$$

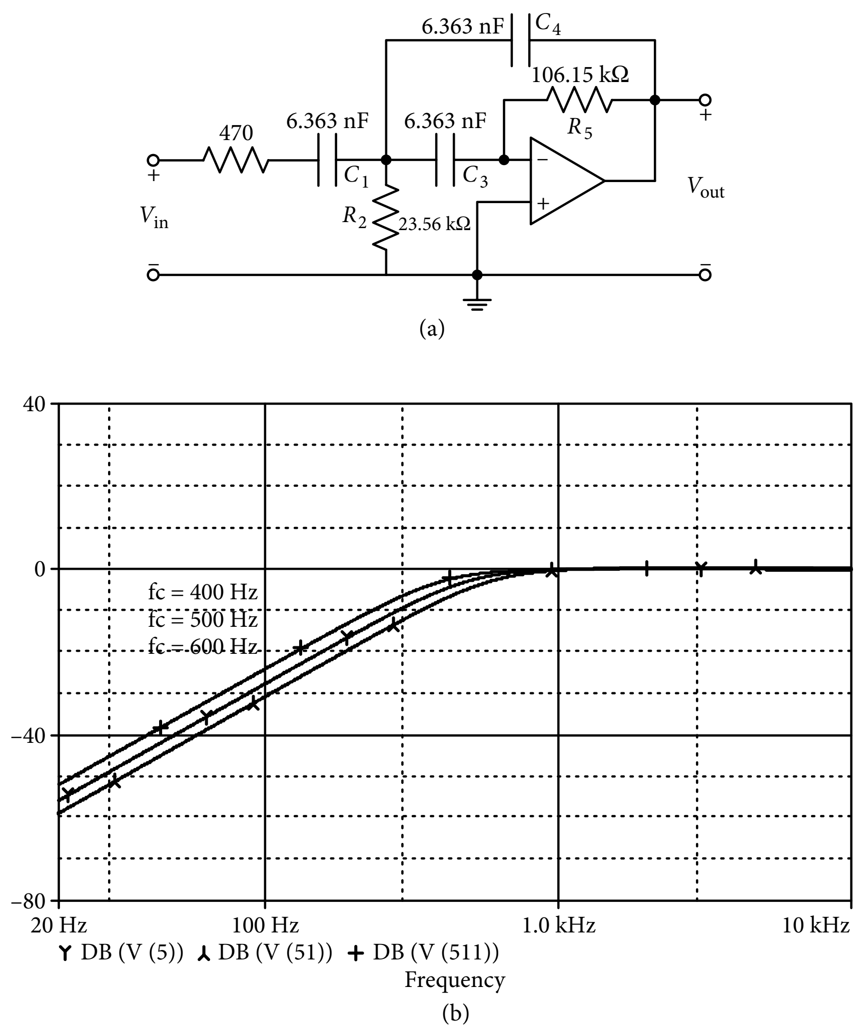

Figure 12.7 (a) High pass filter circuit structure of Figure 7.8(a), employed in Figure 12.6. (b) Responses of the filter shown in Figure 12.7(a).

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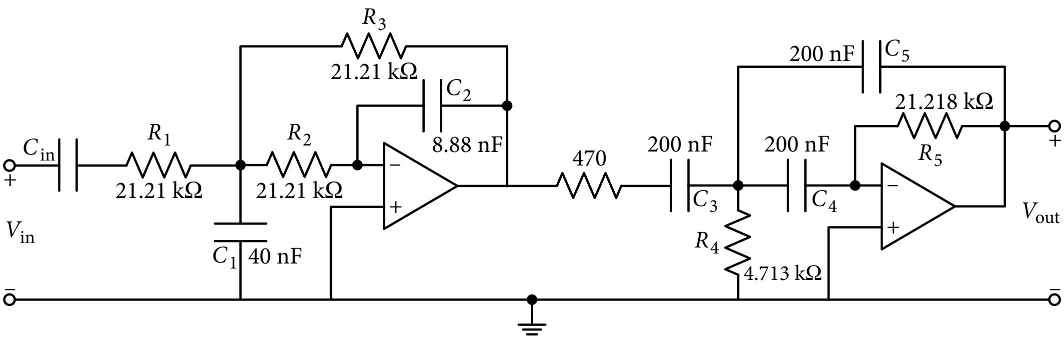
For unity gain LP Butterworth response, with a cut-off frequency of 2500 rad/s (397.7 Hz), application of equation (12.15) yields the component values as $R_1 = R_2 = R_3 = 21.218$ kΩ, $C_1 = 40$ nF, $C_2 = 8.888$ nF, and an input capacitance in series with R_1. For the HP section, the filter structure shown in Figure 7.8(a) is used. Application of equation (7.33), for the cut-off frequency of 500 rad/s (79.5 Hz) yields the component values as $C_3 = C_4 = C_5 = 200$ nF, $R_4 = 4.713$ kΩ, $R_5 = 21.218$ kΩ, and a resistance $R_{in} = 470$ Ω is connected in series with capacitance C_1. The combined circuit structure is shown in Figure 12.8(a). The simulated response of the combined LP and HP filters as a BPF is shown in Figure 12.8(b). Two more BP responses are shown in Figure 12.8(b), with the modified values of capacitors $C_3 = C_4 = C_5 = 250$ nF and 150 nF.



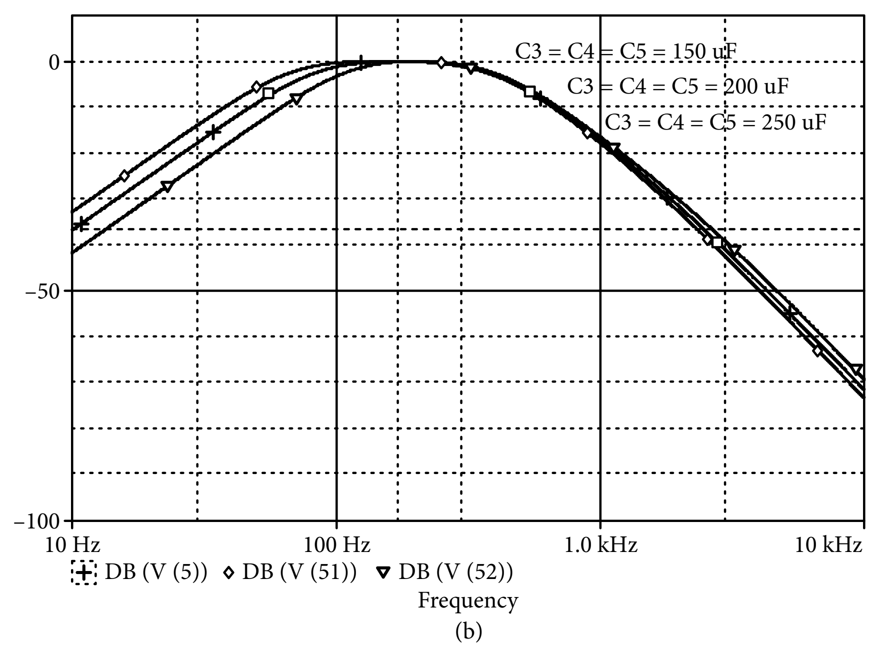


Figure 12.8 (a) Band pass filter circuit which is used with the sub-woofer in Figure 12.6. (b) Simulated magnitude responses of the band pass filter.

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12.4 Four Band Programmable Equalizer

LE8 is a logic-controlled equalizer from Lectrosonics. The equalizer claims to provide an economical solution to the microphone equalization and notch filtering in an automatic modular audio processor system [12.5]. The heart of the LE8 equalizer is a microprocessor which controls all functions of the digitally programmed equalizer (EQ) and notch filters.

The EQ contains bass boost/cut, treble boost/cut filters as well as mid bass boost/ cut and mid-band treble boost/ cut band pass and inverted band pass types of filters. The bass/treble cut/boost filters can be realized with simple bilinear circuits and their description is included in Section 2.3.2 of Chapter 2.

The suggested center frequency of the mid bass is 300 Hz and pole-Q is 0.33; the corresponding values for the mid treble is 3 kHz and 0.33. Mid bass boost and mid treble boost being band pass type can be realized using second-order sections but for respective cut filters, it is preferable to cascade appropriate bass/treble boost/cut. Hence, for the sake of modularity, all the mid section filters are explained here with the same approach.

For mid bass boost realization, having a center frequency of 300 Hz, the treble cut circuit shown in Figure 12.9(a) is followed (cascaded) with a treble boost circuit shown in Figure 12.9(b). For the realization of the rest of the functions, namely, mid treble cut, mid treble boost and mid bass cut, the arrangement is given in Table 12.1, along with component values and cut-off frequency f_c of each section. The simulated response of the four types of filters is shown in Figure 12.10(a) and 12.10(b). As all the modules in Table 12.1 were designed for 10 dBs of boost or cut, their cascading results in a little less amount of boost or cut for the resulting mid frequency filter. Moreover, the center frequency of the filters will depend on the critical frequencies of the modules used; these have to be on either side of the 300 Hz or 3 kHz frequency band; for example, for the mid treble boost, calculated $f_c = 30$ kHz for the treble cut and $f_c = 1$ kHz for the treble boost. Now, all that remains is to find a rigorous mathematical relation between the cut-off frequencies of the two modules with the pole-Q and cut-off frequencies of the circuits forming the mid treble or mid bass cut/boost

Table 12.1 Component values and arrangement for the four mid frequency EQs

Components	Mid bass boost $f_o = 300$ Hz	Mid treble cut $f_o = 3$ kHz	Mid treble boost $f_o = 3$ kHz	Mid bass cut $f_o = 300$ Hz
	Treble cut $f_c = 900$ Hz	**Bass boost** $f_c = 300$ Hz	**Treble cut** $f_c = 9$ kHz	**Bass boost** $f_c = 100$ Hz
R_1 kΩ	200	63.2	200	63.2
R_2 kΩ	200	200	200	200
R_3 kΩ	92.3	92.7	63.2	92.7
C_1 nF	0.6047	1.813	0.067	16.32

Contd.

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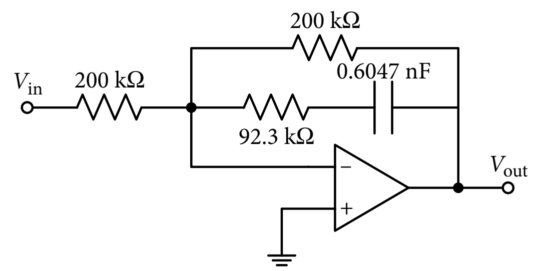
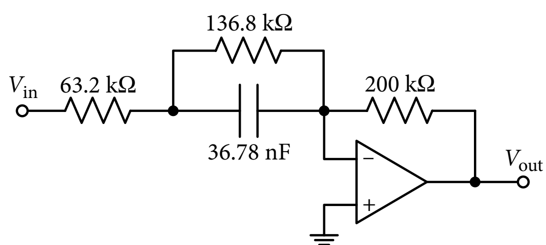
	Treble boost	Bass cut	Treble boost	Bass cut
	$f_o = 100$ Hz	$f_c = 30$ kHz	$f_o = 900$ Hz	$f_o = 900$ Hz
R_4 kΩ	63.2	63.2	100	100
R_5 kΩ	136.8	136.8	216	216
R_6 kΩ	200	63.2	316	100
C_2 nF	36.78	0.1255	2.327	0.7758



(a)



(b)

Figure 12.9 (a) A treble cut realization with cut-off frequency of 900 Hz. (b) A treble boost realization with cut-off frequency of 100 Hz.

Notch Filters: In general, specifications given for the design of notch filters include notch frequency and pole-Q. It is expected that attenuation at the notch will be as large as practically possible. However, in case of the equalizer LE8, it is to be noted that the required attenuation at the notch frequency is to be only 12 dBs so that it compensates for the effect of ringing, which has a magnitude of 12 dBs. The required three notch filters are to have programmable notch frequency, with possible variations in notch frequency from 50 Hz to 500 Hz (frequency resolution of 1.76 Hz) and from 500 Hz to 5 kHz (frequency resolution of 19.4 Hz); the notches are 1/7 of an octave wide.

Programming of the notch frequency is done by varying the components used; these components are controlled by the microprocessor and the circuit structure of the filter. The approach used in Section 8.6 (that of obtaining a multi output biquad using a summing amplifier) is used here to obtain a notch filter.

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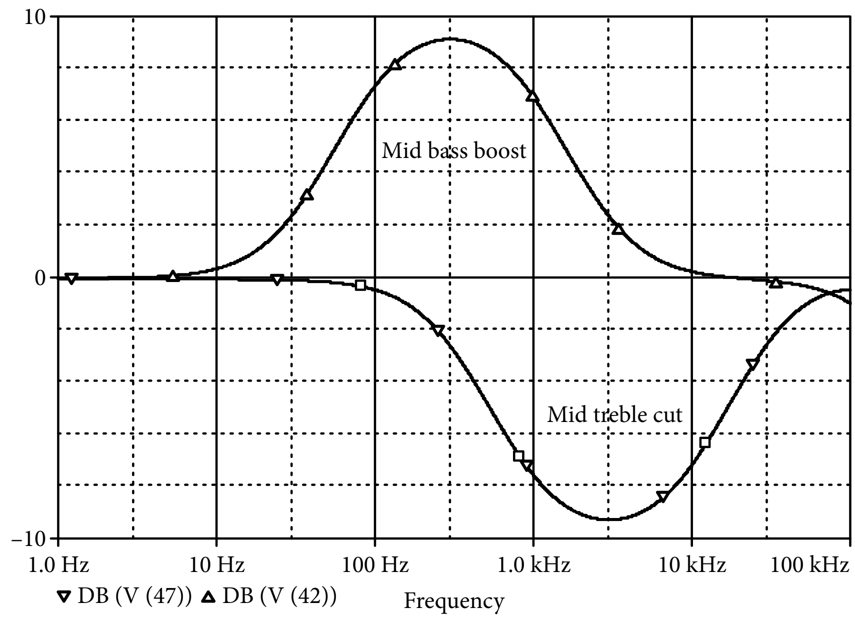
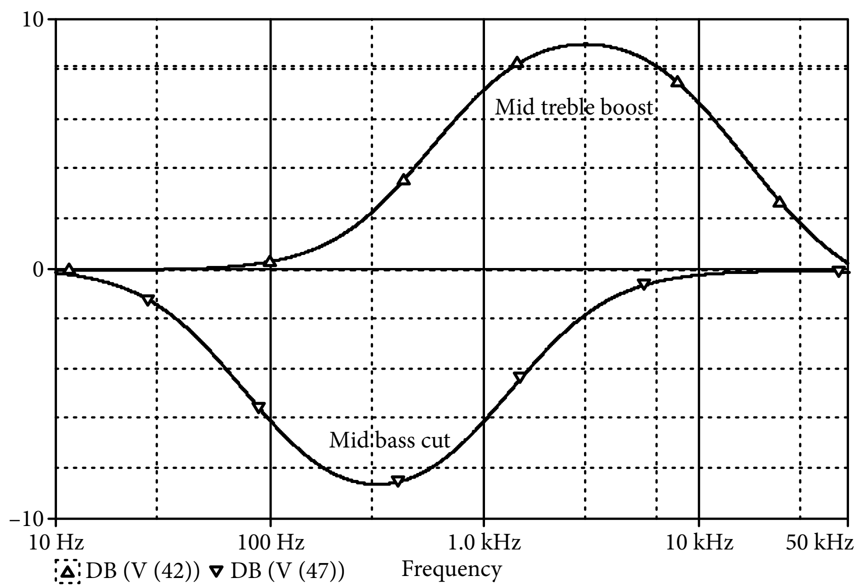


(a)



(b)

Figure 12.10 (a) Mid bass boost and mid treble cut responses using the cascade scheme. (b) Mid bass cut and mid treble boost responses using the cascade scheme.

In Example 8.3, the notch filter is realized with $Q = 5$ at a notch frequency of 3.18 kHz and 0.318 kHz. The filter obtains an attenuation of 40.17 dBs at a notch frequency of 3.18 kHz.

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The notch becomes effective by making the coefficient of s in the numerator as zero in a near ideal case. Attenuation at the notch may be decreased while creating a non-ideal situation by making the coefficient of s in the numerator non-zero.

Using the structure shown in Figure 8.14(a) for the notch frequency of 3.18 kHz, the same values of components were used. Only those component values required change which changed the width of the notch to 1/7 of an octave. Hence, Q is taken as 3.5, which results in the resistor $QR = 17.5$ kΩ. To make the non-ideal attenuation condition, coefficient β was made 0.250 (instead of 1/3.5; equation (8.34)), which resulted in the resistor $R_\beta = 14$ kΩ (or 23.3 kΩ by trial). The simulated response of the notch is shown in Figure 12.11, with the required attenuation of 12 dBs and notch width of 849 Hz or $Q = 3.72$.

Variation in the notch frequency from 3.18 kHz to 0.318 kHz was easily done by changing capacitors C_1 and C_2 from 0.01 μF to 0.1 μF. Its simulated response is also shown in Figure 12.11. Attenuation remains unchanged at 12 dBs and $Q = 3.73$.

In all audio circuits, it is desirable to use low noise OAs.

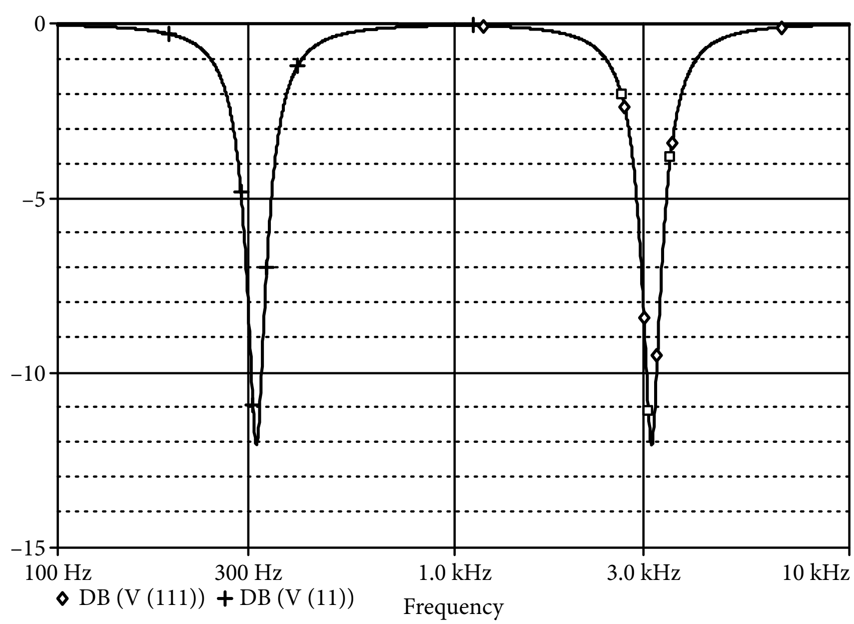

Figure 12.11 Notch filters with 12 dBs attenuation at different frequencies using the circuit structure of Figure 8.14(a).

12.5 Record Industry Association of America (RIAA) Equalization

For good quality hi-fi audio equipment, the phonographic record pickup has to be of magnetic type. However, when we play a disc, it generates a non-linear frequency response. In general,

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disc recording equipment does not give an exactly linear frequency response. Hence, to improve the performance, signals below 50 Hz and in the range of 500 to 2120 Hz, are recorded in a non-linear fashion. The nature and level of non-linearity is set by the Record Industry Association of America (RIAA). The RIAA curve, depicting the mentioned non-linearity is shown in Figure 12.12. The curve is described by 3 poles in time (frequency) terms as [12.6]:

$$T_1 = 3180\ \mu s\ (50\ Hz),\ T_2 = 318\ \mu s\ (500\ Hz),\ T_3 = 75\ \mu s\ (2120\ Hz)$$

These poles can be described as HP with T_1, LP with T_2 and again HP with T_3.

In 1992, the International Electrotechnical Commission (IEC) proposed a new superior version of the RIAA curve by adding another pole, a HP at 7950 µs (20 Hz). An additional pole was included to reduce the subsonic output of the amplifiers caused by the rumble of the turn table [12.6].

From the aforementioned discussion, it can be inferred that when a disc is played through a magnetic pickup system, its output must go to the power amplifier via a pre-amplifier; the frequency equalization curve of the pre-amplifier should be the exact inverse of the curve shown in Figure 12.12.

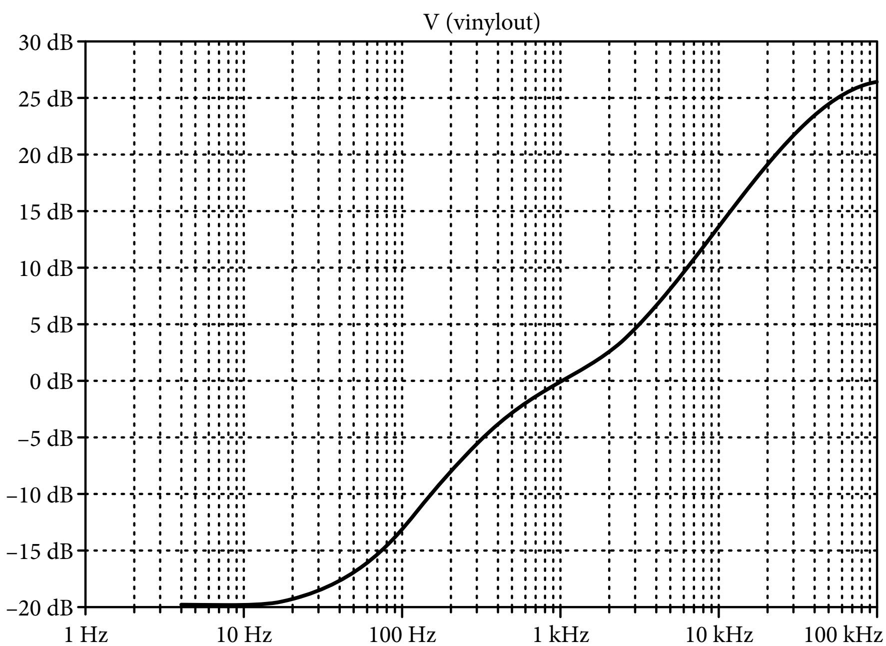

Figure 12.12 RIAA curve depicting non-linearity while recording [12.6].

12.5.1 RIAA phono pre-amplifier

Figure 12.13 shows an equalization schematic proposed by ON Semiconductors which consists of a pre-amplifier, an input buffer, a 5-band equalizer and a mixer using 5532 low

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noise OA [12.7]. The first stage is a pre-amplifier as the signal magnitude level in a magnetic pickup is very low. In the mid-band range, signal magnitude is only a few mV. Hence, a low noise pre-amplifier IC (rather than a simple OA) is necessary. Figure 12.14(a) shows one such configuration that uses a low noise OA, NE 5532A (any other low noise OA can also be used).

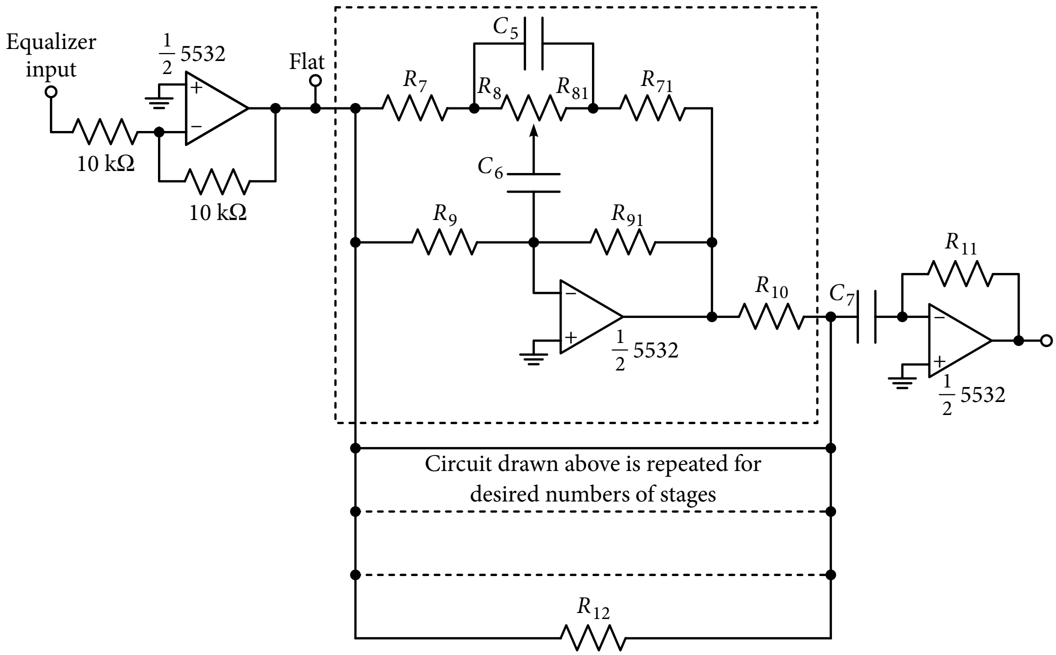
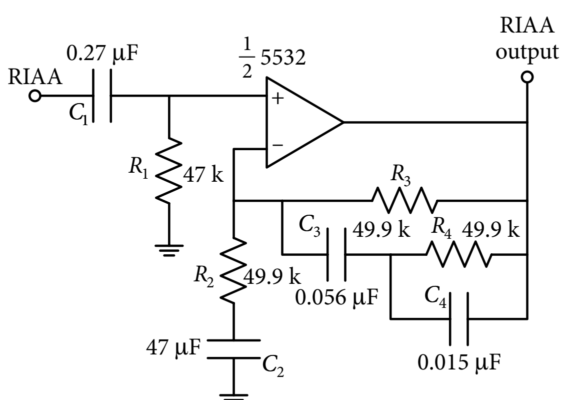


Figure 12.13 RIAA: Equalizer schematic [12.7]. {Used with permission from SCILLC dba ON Semiconductor}.



(a)

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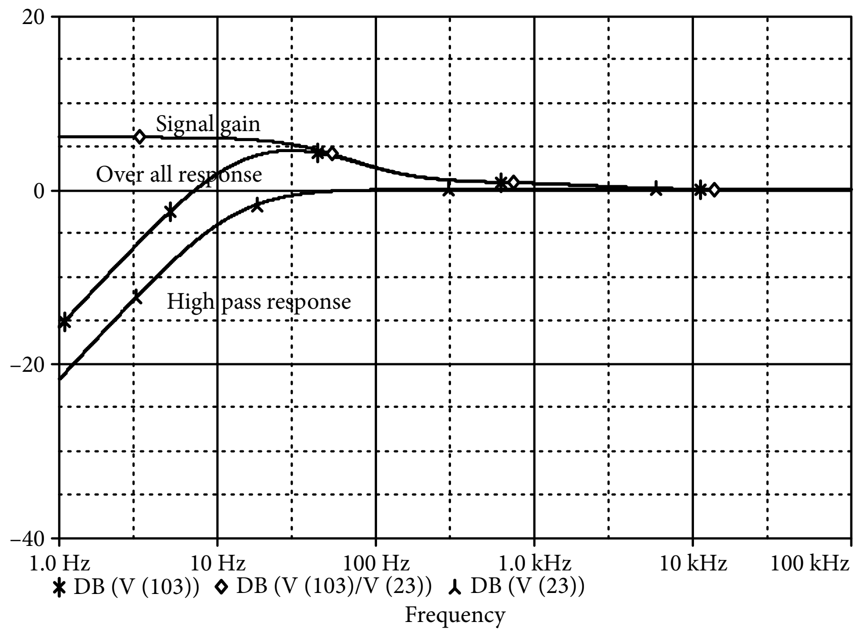

(b)

Figure 12.14 (a) RIAA phono pre-amplifier circuit [12.7]. (b) Magnitude response of the phono pre-amplifier and associated high pass filter. {Used with permission from SCILLC dba ON Semiconductor}.

The phono pre-amplifier is in effect a non-inverting amplifier with negative feedback applied through the potential divider formed by resistors R_2 and R_3. A combination of R_3, C_3, C_4, R_4 and R_2, C_2 network determines the signal gain. At high frequencies, capacitors, C_2, C_3 and C_4 have low impedance, so the gain is $\{1 + (R_3||R_4)/R_2\}$, leading to nearly unity. At low frequencies, impedance of C_3 starts increasing and causes the ac gain to increase. At very low frequencies, the gain is limited to $(1 + R_3/R_2)$. The simulated response of the circuit is shown in Figure 12.14(b). Its low frequency gain is 6 dBs and the high frequency gain is unity, confirming the analysis. However, a combination of R_1 and C_1 acting as HPF with $f_c = (1/2\pi * 0.27\ \mu * 47\ \text{k}) = 12.5$ Hz changes the overall characteristics, reducing the gain at very low frequencies as shown in the simulated response; it also shows the HPF response.

12.5.2 Simultaneous cut and boost circuit

The main constituent of the equalizer in Figure 12.13 is an active band pass/notch filter which is shown within dotted lines. It can cut or boost by positioning the potentiometer to the right or left. Values of the resistors and capacitors for such circuits are available in literature. A table depicting the component values for a wide frequency range is also available in reference [12.7]. Frequency of the filter is changed by changing capacitors C_5 and C_6, where C_5 equals 10 times C_6 and the values of the resistors R_8 and R_{10} are related to R_9 by a factor of 10. The circuit was simulated for two center frequencies of 54 Hz and 541 Hz. Values of the elements which are common at both frequencies are as shown in equation (12.16a) [12.7]:

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$$R_8 = 50\ \text{k}\Omega,\ R_7 = R_{71} = 5.1\ \text{k}\Omega,\ R_9 = R_{91} = 510\ \text{k}\Omega \tag{12.16a}$$

At 54 Hz, with $C_5 = 0.22\ \mu$F and $C_6 = 0.022\ \mu$F, the respective values of the selected components for boost and cut case are:

$$R_8 = 2.5\ \text{k}\Omega,\ R_{81} = 47.5\ \text{k}\Omega,\ \text{and}\ R_8 = 47.5\ \text{k}\Omega,\ R_8 = 2.5\ \text{k}\Omega \tag{12.16b}$$

At 541 Hz, with $C_5 = 0.022\ \mu$F and $C_6 = 0.0022\ \mu$F, the respective values of the selected components for boost and cut case are:

$$R_8 = 15\ \text{k}\Omega,\ R_{81} = 35\ \text{k}\Omega,\ \text{and}\ R_8 = 35\ \text{k}\Omega,\ R_{81} = 15\ \text{k}\Omega \tag{12.16c}$$

The simulated response of the cut and boost circuit is shown in Figure 12.15, reflecting the amount of cut and boost on the relative values of R_8 and R_{81}.

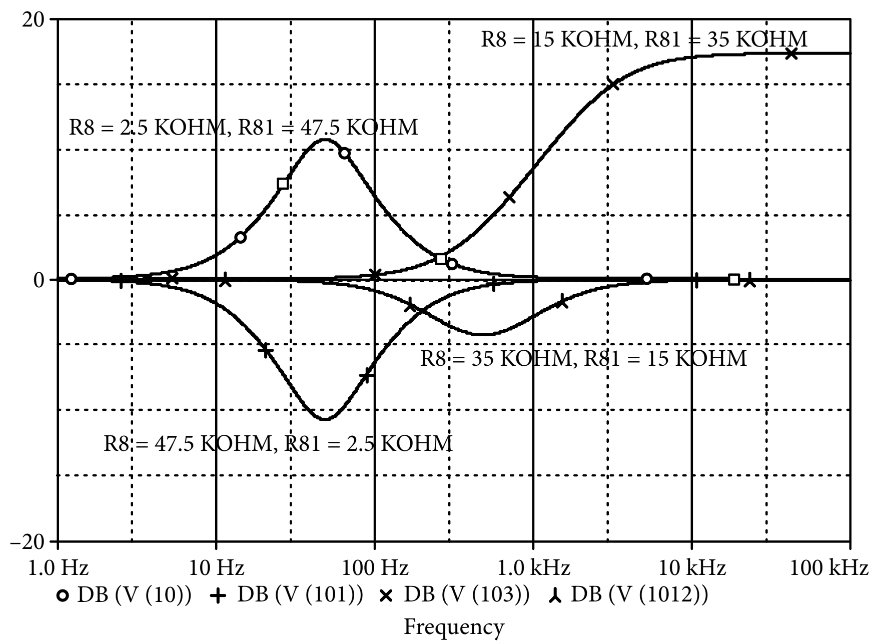

Figure 12.15 Simulated response of the simultaneous cut and boost circuits.

Tone Control Circuit: Tone control circuits are one of the most widely used types of variable frequency filter circuits. With the help of tone control circuits, one can change the frequency response of an audio system according to the requirement. A circuit similar in nature to the circuit of the previous section is also available; it boosts or cut signals at low or high frequencies.

One of the popular approaches to get an active tone control circuit employs a passive tone control network in the negative feedback loop of an OA. Such a system gives an overall signal gain rather than attenuation, when its controls are in a flat position. One such realization is shown in Figure 12.16(a); it provides 20 dBs of bass or treble boost or cut, set by the variable

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resistances (R_2 + R_3) and (R_7 + R_8). With capacitors C_2 to C_5 as 0.00625 μF its turn over frequency is 1 kHz as shown in the simulated response in Figure 12.16(b).

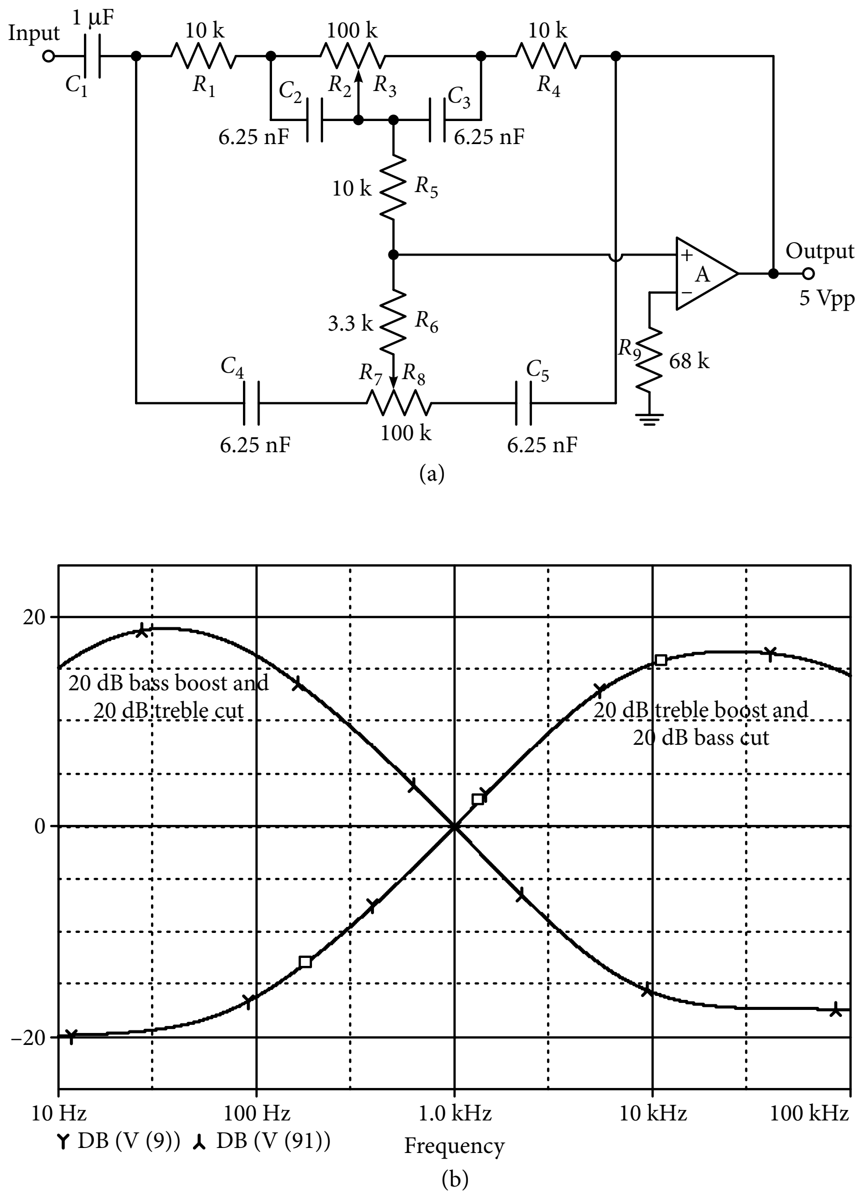

Figure 12.16 (a) Tone control circuit [12.7]. (b) Response of the tone control circuit shown in Figure 12.16(a). {Used with permission from SCILLC dba ON Semiconductor}.

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12.6 Phase Delay Enhancement in 3D Audio

For the observance of sound in stereophonic form, speakers are usually placed at considerable distances. However, such an arrangement is not possible in cell phones or hand-held devices. In such cases, stereophonic sound effect is simulated by what is known as *trans aural cross talk* cancellation. In this scheme we employ wave interference in such a way that the left channel sound is canceled in the right ear and the right channel sound is canceled in the left ear.

Even a small difference in the arrival of sound waves at the left and right ears helps in determining the direction of the sound. This difference of time or time delay is known *as inter-aural time delay* (ITD). The ITD operates on the spectral properties associated with the construction of internal and external parts of the ears and determines the *head-related transfer function* (HRTF). The HRTF accounts for the audio source location in terms of the distance to listener's head, the separation distance between the two ears and frequencies of the sound [12.8].

The basic idea of simulating the 3D audio effect is by combining each source with the pair of HRTF corresponding to the direction of the source [12.9]. However, most multimedia products do not use the directional information for fully 3-D effects. They simply employ phase-delay circuits creating HRTF that produces an effect of receiving the sound as if it comes from a different direction. Therefore, the speakers which are placed closely are perceived as being placed at different locations.

Let us, in a typical hand-held device where the distance between speakers is less than 7 cm, assume that the listener's head width is 20 cm. It is shown mathematically that for a distance of 50 cm between the listener's ear and the device, a phase difference (or phase delay) in sound signals should be between 78.5° and 101.5°A [12.8]. Such phase differences are obtained using a simple AP filter section; this finds considerable application in audio electronics simulation. One of the most commonly used circuits for the purpose is the active first-order circuit shown in Figure 4.11(a). Currently, it is required to construct a phase shift filter with f_o = 10 kHz for the linear signal (or for proper channel) and f_o = 1 kHz for the quadrature signal. It is desired to have a phase shift of 90° between the quadrature signal over the frequency range of 1 kHz to 10 kHz.

Figure 12.17(a) shows a cascaded, first-order all pass filter that will achieve a phase shift of nearly 90° between L (left) and Q (right) outputs for the frequency range of 1 kHz to 10 kHz. Figure 12.17(b) shows the simulated phase response of the circuit with a reasonable approximation to 90° of phase shift.

To improve the 3D effect, more stages can be added and aligned appropriately to improve the frequency range and increase the closeness to the 90° phase shift between L and Q outputs. Audio ICs such as MAX 9775 incorporate such a phase delay circuit and audio amplifier on a single chip.

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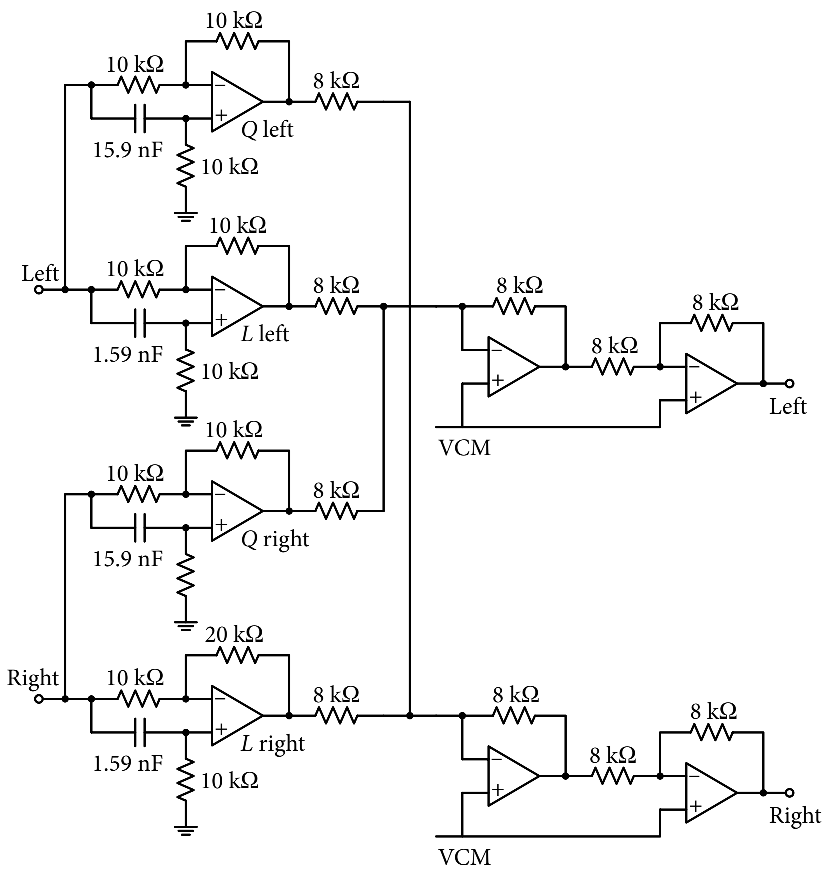
12.7 Anti-aliasing Filters

Precision analog to digital convertors (ADCs) constitute an essential part in a variety of applications, such as instrumentation and measurement, process control and motor control, where an analog signal is to be digitally processed. Currently used SAR (successive approximation register) ADCs go up to 18-bit or even higher resolution, while $\Sigma - \Delta$ ADCs can be 24- or 32-bit resolution at 100s of kc/s. Consequently, there are considerable design challenges and considerations associated with implementing analog and digital filters into the ADC signal processing chain to achieve optimum performance [12.10]. Figure 12.18 shows a typical data acquisition signal chain which employs analog and digital filtering technique or a combination of both. Presently, the interest is not to study the design challenges for a practical filter in detail, which can be seen in reference [12.10], but to show simple examples of analog filter used in data acquisition.



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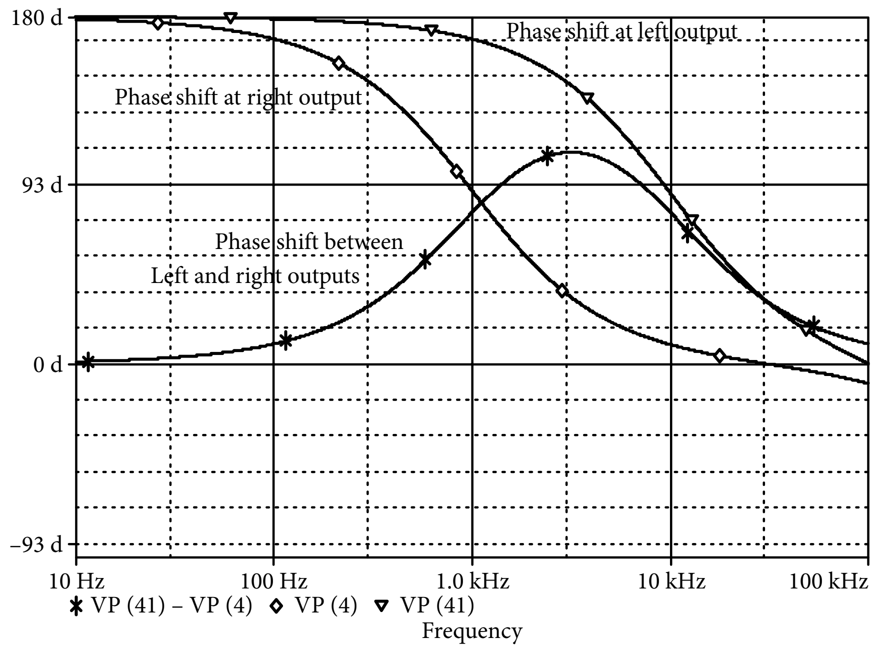

(b)

Figure 12.17 (a) Cascaded first-order active all pass circuits to achieve a phase shift of 90° between left and right signals {Courtesy: Maximum Integrated}. (b) Variation in phase shift for right and left signals and their phase difference.

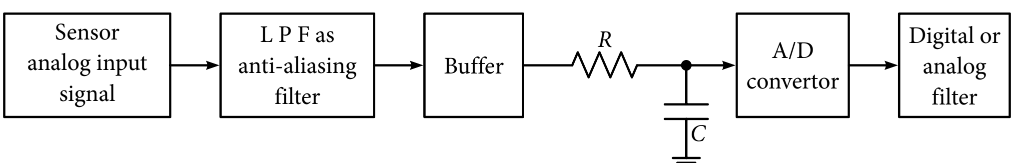

Figure 12.18 A typical data acquisition system [12.10].

The analog filter is placed before the ADC in order to remove noise that is superimposed on the analog signal. Sometimes, there are noise peaks as well and it becomes necessary to eliminate these peaks; otherwise, noise peaks have the potential to saturate the analog modulator of the ADC. In addition, analog filtering is more suitable than digital filtering prior to ADC for higher speed systems (> 5 kHz). In these types of systems, the analog filter can reduce noise in the out-of-band frequency region, which reduces fold-back signals.

ADCs are usually operated with a constant sampling frequency. For sampling frequency f_c, typically called the Nyquist frequency, all input signals below $f_c/2$ folds back into the bandwidth of interest with the amplitude preserved. The frequency folding phenomena can be eliminated or attenuated by moving the corner frequency of the filter lower than $f_c/2$ or

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by increasing the order of the filter. In both the cases, the minimum gain of the filter at $f_c/2$ should be less than the signal-to-noise ratio (SNR) of the sampling system. For example, for a 12-bit ADC, the ideal SNR is 74 dB (> 6 dB per bit) less than the pass band gain. There are a number of papers available describing the utility and design of anti-aliasing filters. For example, Kyu et al.'s paper [12.11] discusses the design and implementation of a fourth-order Butterworth LPF using MATLAB and Circuit Maker for the data acquisition system. The following is also a typical example while having a comparison when LPF is implemented with different approximation approaches.

Anti-aliasing Filter Design: A Case Study[12.12]

In this example, the bandwidth of interest of the analog signal is from dc to 1 kHz; the output of the LPF will be fed to 12-bit SAR ADC. The sampling rate of the ADC will be 20 kHz. Hence, the design parameters for the LPF will be as follows:

i. Cut-off frequency must be 1 kHz or higher.

ii. Filter attenuates the signal by 74 dBs at 10 kHz

iii. Signal magnitude remains near constant in the pass band without any gain or attenuation.

(a) Implementation with Butterworth Approximation.

Using the design equations of the Butterworth approximation with f_c = 1kHz and a stop band frequency of approximately 5 kHz, we get the order of the filter as four. In the references note, a cascade of two Sallen and Key structures shown in Figure 7.22 was used (any other structure or design approach can also be used). Element values of the two second-order sections are:

$$\text{First stage: } R_1 = 2.94 \text{ k}\Omega,\ R_2 = 26.1 \text{ k}\Omega,\ C_1 = 33 \text{ nF and } C_2 = 10 \text{ nF} \quad (12.16)$$

$$\text{Second stage: } R_1 = 2.37 \text{ k}\Omega,\ R_2 = 15.4 \text{ k}\Omega,\ C_1 = 100 \text{ nF and } C_2 = 6.8 \text{ nF} \quad (12.17)$$

OA MCP 601 was used in both the stages. Simulated magnitude and phase responses of the filter are shown in Figure 12.19. The filter's gain at 10 kHz is 79.92 dBs against the theoretical value of 80 dBs; the gain remains flat in the pass band at unity.

(b) Implementation using Chebyshev Approximation

For the same specifications, but with 4 dBs of attenuation allowed in the pass band, the design equations give the required order of the filter as three. Hence, a cascade of a passive first-order and an active second-order (Sallen and Key) circuit was used; it is shown in Figure 12.20(a). The order of the filter is one less than in the Butterworth case; hence, it is economical. The simulated magnitude and phase responses are shown in Figure 12.20(b) which shows an attenuation of 74.12 dBs at 10 kHz against the theoretical value of 70 dBs. Along with the ripple present in the pass band, its phase response is worse than the Butterworth case.

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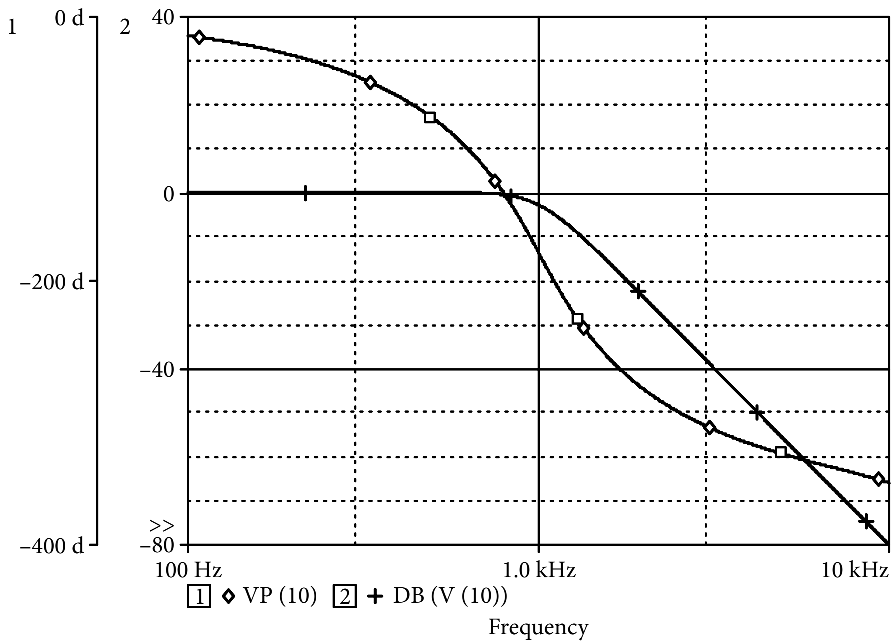
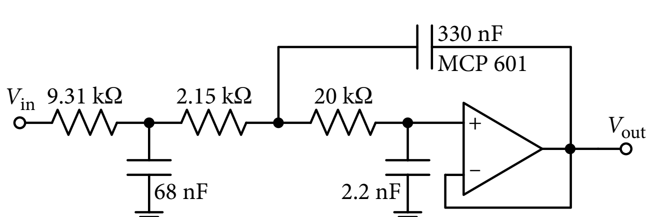


Figure 12.19 Response of fourth-order anti-aliasing filter using Butterworth approximation.

12.7.1 Digitally tuned anti-aliasing filter

LTC 1564 is a digitally controlled anti-aliasing/re-constructional filter. It is a high resolution eighth-order LPF that gives approximately 100 dBs of attenuation at 2.5 times of the corner frequency $f_{c.}$ (f_c which is digitally controlled and ranges 10 kHz to 150 kHz in steps of 10 kHz). It also includes a digitally programmable gain amplifier (1V/V to 16 V/V in 1V/V steps). In addition to the mentioned attenuation, it has an SNR of 100 dBs. The filter has been shown to drive a 16-bit 500 ksps ADC for a complete 16-bit ADC interface [12.13].

The LTC 1564 is in a small 16-pin SSOP (Shrink Small Outline Package) and operates at a single or split supply range of 2.7 V to 10.5 V. Its pin connection diagram is shown in Figure 12.21.



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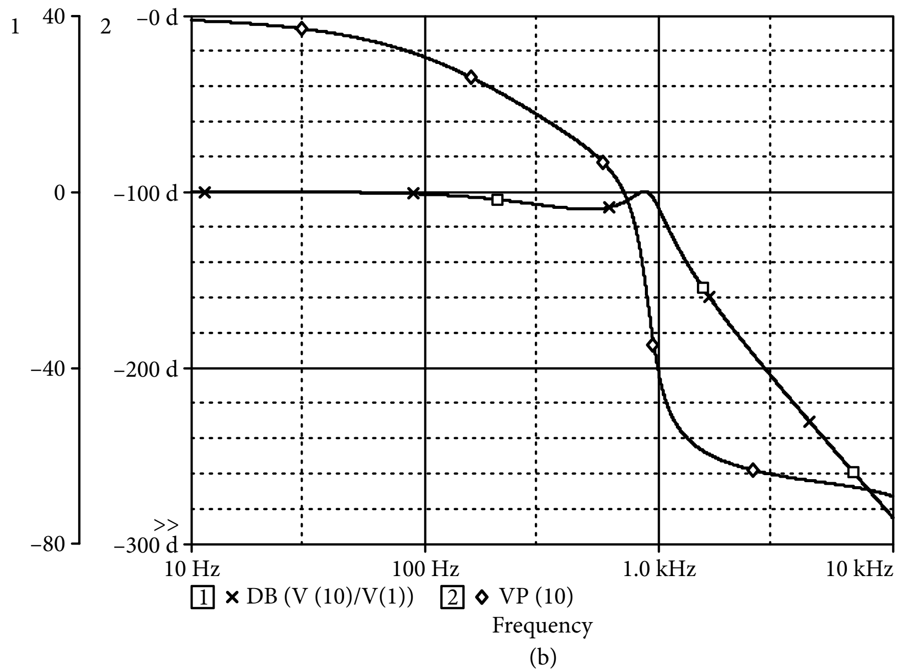

Figure 12.20 (a) Anti-aliasing filter using Chebyshev approximation. (b) Magnitude and phase response of the anti-aliasing filter shown in Figure 12.20(a).

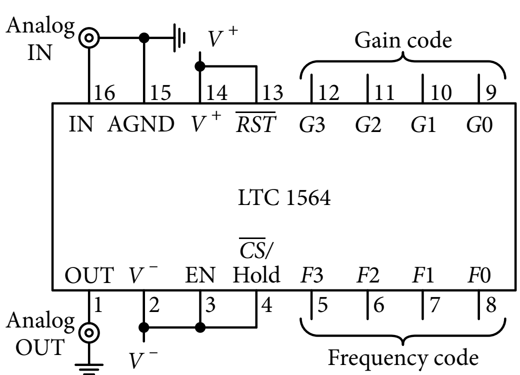

Figure 12.21 Pin connection diagram of an LTC 1564: a digital tuned anti-aliasing filter.

12.8 Active Filters for Video Signals: A Case study [12.14]

The following filters have been fabricated by Maxim Integrated using MAX 4450 OAs.

1. An ITU-601 anti-aliasing filter (AAF).
2. A 20 MHz AAF and reconstruction filter (ReF).
3. An HDTV ReF.

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It is now well understood that AAF is used before an ADC to attenuate signals above the Nyquist frequency or half the rate of the sampling frequency of the ADC. In general, AAFs are designed with maximum roll-off in the stop band that is economically permissible. For ITU-601 and other similar applications, the objective is achieved by using analog filters combined with digital filters and an over-sampling ADC. In the case of ReFs (also called sin *x/x* or zero-order-hold correctors), they are placed after DACs to remove multiple images created by sampling. These filters do not require a steep roll-off in the stop band.

Both AAFs and ReFs have an LP characteristic to pass the video frame rate. Obviously, LPFs with best selectivity and lowest order are chosen. However, phase linearity requirement is also a serious consideration in video filter designs. It is for this reason that Butterworth approximation is most often used, though it may not be the most economical compared to Chebyshev and other approximations.

Group Delay Problems with Component Videos: In different formats and applications, the degree of sensitivity to group delay variations depends on the number of signals and their bandwidth. For example, a composite NTSC/PAL has only one signal with a group delay defined in ITU-470. Hence, the requirement is not difficult to meet; whereas for other component videos, each have multiple signals with equal or different bandwidths, requiring better group delay management.

Selection of OA to be Used: Since a video application operates in a wide bandwidth, and its typical magnitude is 2 $V_{\text{p-p}}$, gain bandwidth for an OA is an important consideration. In this case study, either Sallen and Key (Section 7.7), or multiple feedback or Rauch filter (Section 7.3) is used. Hence, the gain bandwidth of the OA is decided by the following relation of phase argument of a Rauch filter (effectively, it is the same for the Sallen and Key type filters).

$$\text{Arg}[H(j\omega)] = \pi - (\omega_c/GBW_{\text{rad}})(1 + R_f/R_i) \tag{12.18}$$

where R_f and R_i are the gain setting resistances and GBW_{rad} is the OA gain bandwidth product and ω_c is the cut-off frequency of the filter.

For AAFs, filter selectivity is determined by a template for ITU-601, as shown in Figure 12.22. It gives a specified bandwidth of 5.75 MHz ± 0.1 dB with an insertion loss of 12 dB at 6.75 MHz and 40 dB at 8 MHz. Variation in group delay is limited to ± 3 ns over a 0.1 dB bandwidth. This specification is difficult to meet easily using an analog filter alone; however, oversampling (×4) modifies the requirement to 12 dB at 27 MHz and 40 dB at 32 MHz; permitting the use of analog filters.

Using the procedure discussed in Chapter 3, it is easily observable that a 5-pole Butterworth filter with 3 dB cut-off frequency of 8.45 MHz will satisfy the aforementioned specifications. However, for satisfying the group delay specification, an important parameter of OA is the 0.1 dB, $2V_{p\text{-}p}$ bandwidth which is much lower [12.15]. Instead of designing the circuit, the circuits given in the application note 'Reference Schematic 31721' [12.14] of the Maxim Integrated is used. Figure 12.23(a) shows the schematic of the 5-pole, 5.75 MHz Butterworth filter for ITU-601 AAF, using a Rauch circuit with a delay equalizer employing MAX 4451

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OAs. PSpice simulation of the magnitude and group delay is shown in Figure 12.23(b). In the simulated response, attenuation at 27 MHz and 40 MHz is 49.2 dBs and 67 dBs, respectively, and the overall group delay variation is of 0.1 dB up to 6.52 MHz.

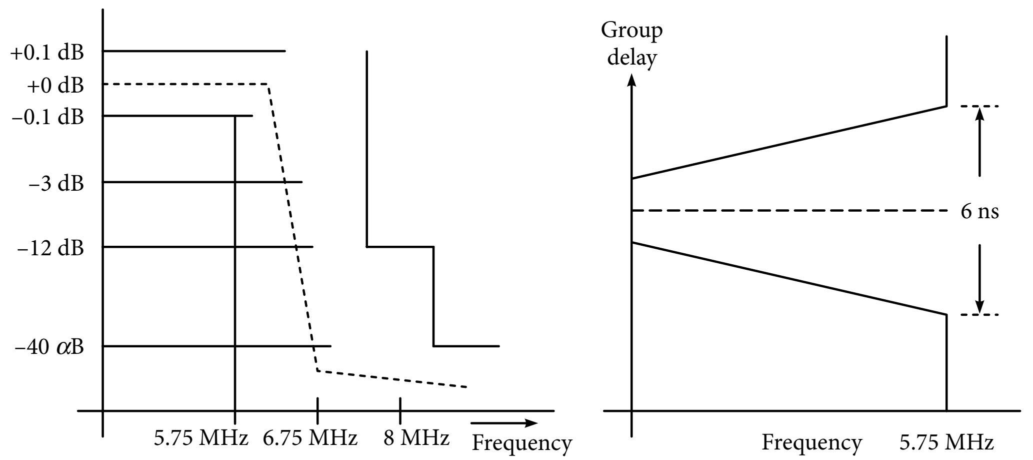

Figure 12.22 A template for determining the selectivity of an anti-aliasing filter {Courtesy Maxim Integrated}.

The next video filter example is that of a PC video, VESA. An XGA resolution (1024 × 758 at 85 Hz) has a sampling rate of 94.5 MHz. To have more than 35 dB attenuation at the Nyquist frequency of 47.75 MHz, a Rauch realization of a 20 MHz 4-pole Butterworth filter is used. Its circuit realization with element values employing MAX 4450 OA is shown in Figure 12.24(a). Figure 12.24(b) shows its simulated magnitude and group delay responses. The simulated response shows only 22 dBs of attenuation at 47.25 MHz.

12.9 Reconstruction Filter

An ReF removes all but the base band sample from the DAC output. Specification for NTSC/PAL ReF states that attenuation of more than 20 dB is required at 13.5 MHz and more than 40 dBs at 27 MHz; cut-off frequency depends on the applicable video standard. A 3-pole Butterworth with Sallen and Key configuration (gain value of 2) is used. Figure 12.25(a) and 12.25(c) shows filters circuits for NTSC and PAL designs, which are expected to satisfy gain and group delay specifications. Figures 12.25(b) and 12.25(d) show respective magnitude and group delay responses. In the simulated response of Figure 12.25(b), attenuation is 17.6 dBs at 13.5 MHz and 29.4 dBs at 27 MHz; somewhat less than the specifications. However, for the PAL case, the simulated response in the figure shows attenuation of 20.1 dBs at 13 MHz and 36.2 dBs at 27 MHz.

Another circuit for XGA is a 3-pole 20 MHz Butterworth filter. It uses the Sallen and Key configuration with gain 2, and it is shown in Figure 12.26(a). Complimenting the AAF of Figure 12.24, its simulated response is shown in Figure 12.26(b)

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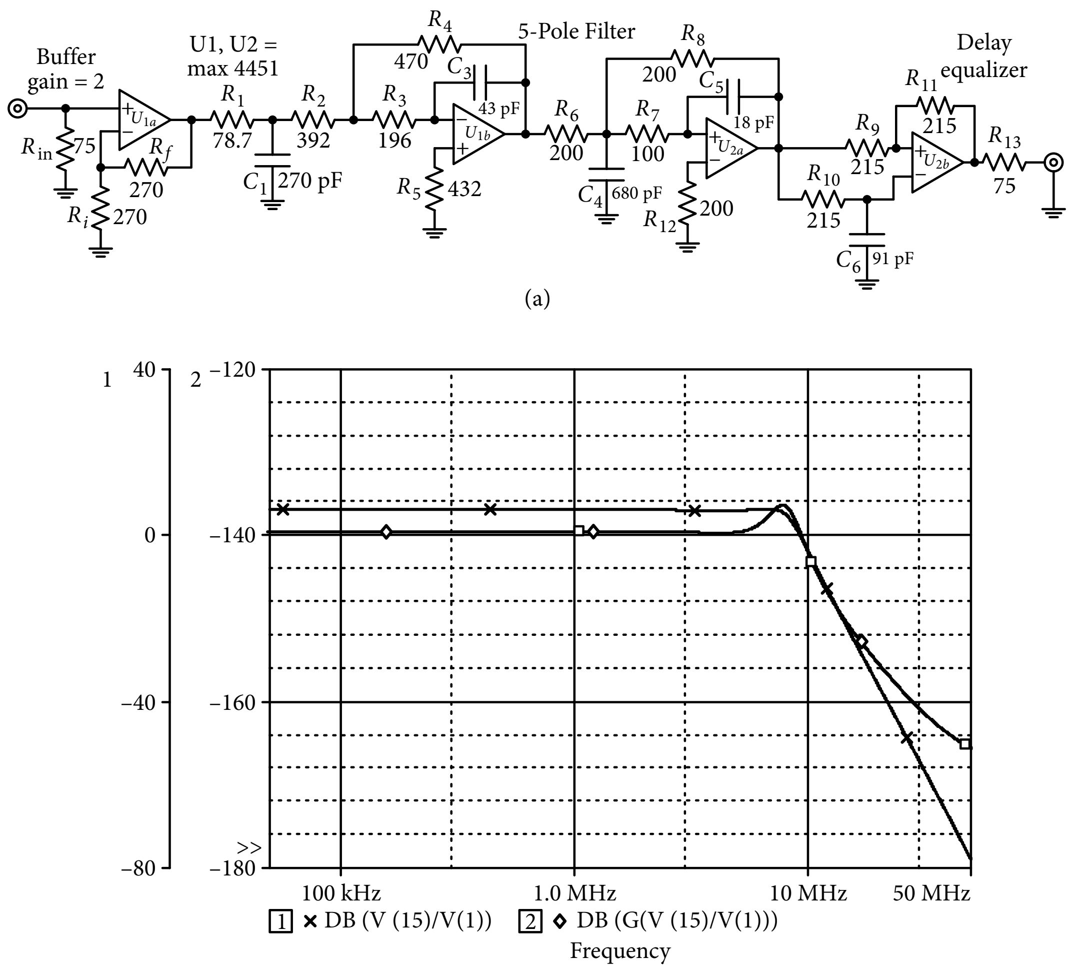

(b)

Figure 12.23 (a) A 5-pole 5.75 MHz Butterworth anti-aliasing filter {Courtesy: Maxim Integrated}. (b) Magnitude and phase response of the filter shown in Figure 12.23(a).

The last application in this case study is a reconstruction filter for HDTV. It is based on the template in the SMPTE-274 and 296M and its corner frequency is 29.7 MHz. A 5-pole Sallen and Key LP filter with a cut-off frequency of 30 MHz will have attenuation of more than 40 dBs at 74.25 MHz. Moreover, with the inclusion of group delay compensation, the filter will have a gain of +2 to drive a back-terminated 75 Ω co-axial cable. The complete circuit with element values is shown in Figure 12.27(a). Figure 12.27(b) shows the simulated magnitude and group delay responses. The filter's simulated low frequency gain is 2, cut-off frequency is 27.2 MHz; the attenuation of 41.6 dBs is achieved at 74.25 MHz.

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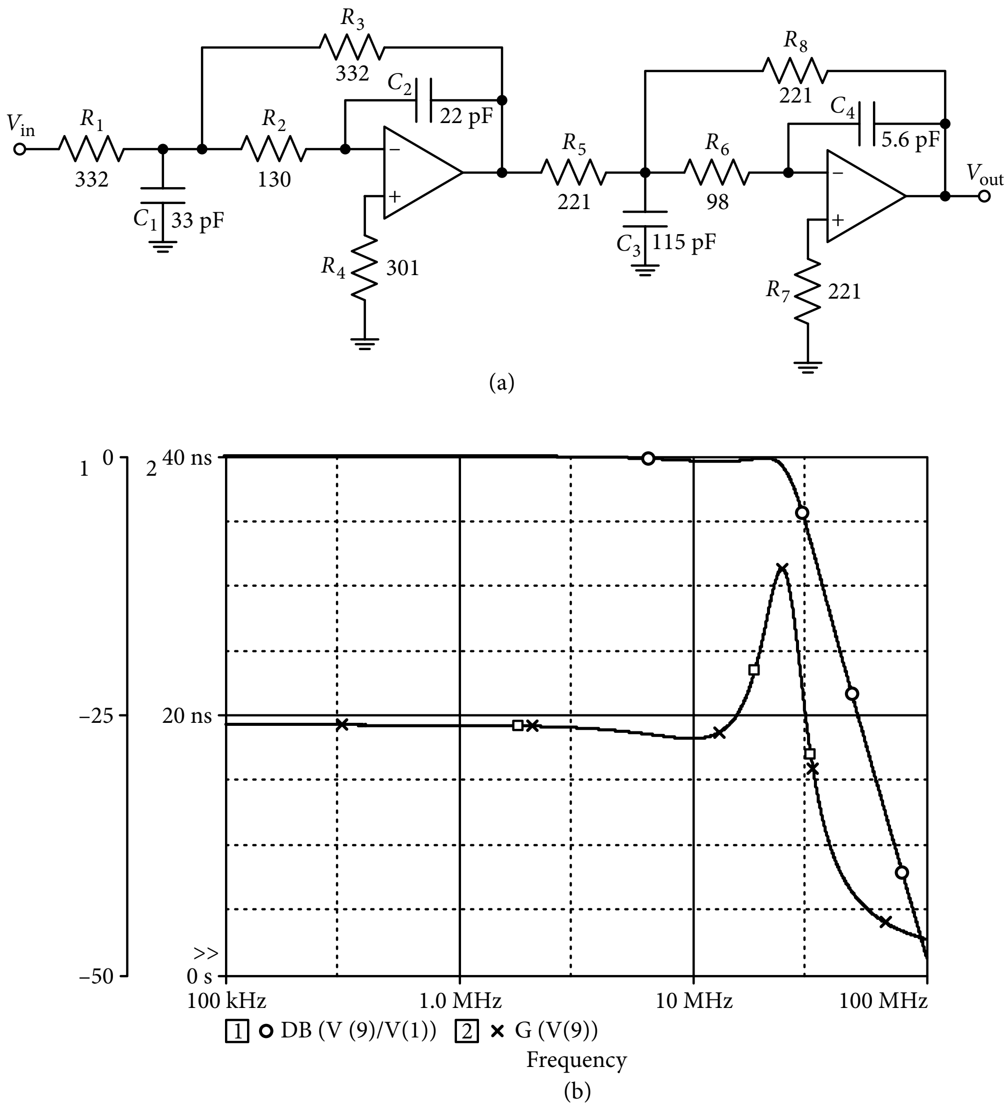

Figure 12.24 (a) A 20 MHz, 4-pole Butterworth filter for a PC video {Courtesy: Maxim Integrated}. (b) Magnitude and phase response for the filter shown in Figure 12.24(a).

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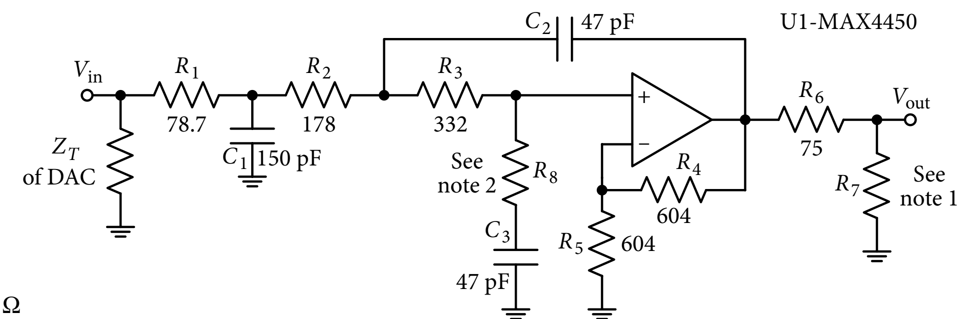
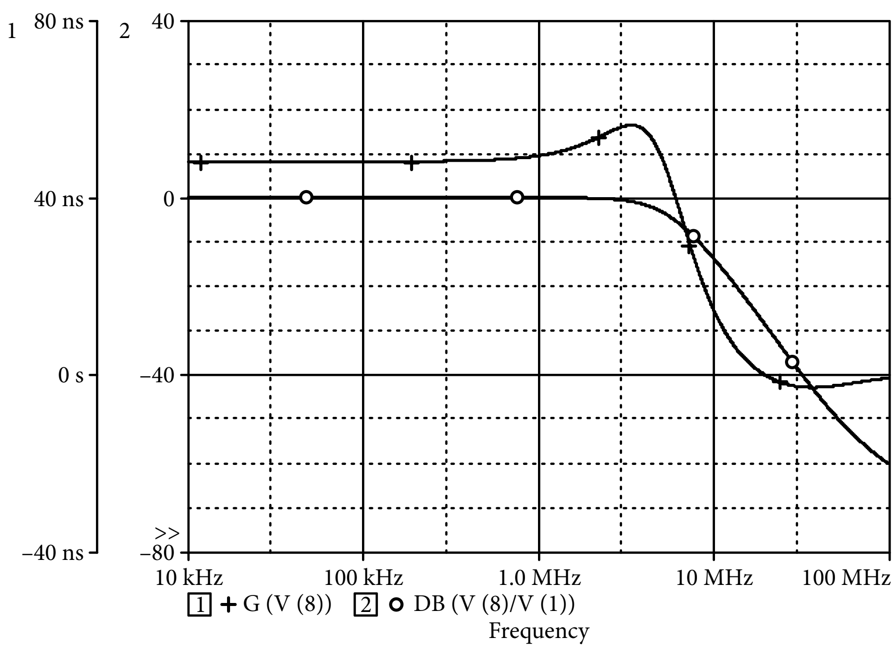
12.9.1 Video line driver for consumer video applications

A 5.25 MHz, third-order Butterworth filter having a gain of 2 V/V is realized using a MAX4390 OA. The proposed filter is capable of driving a 75 Ω back-terminated coaxial cable that is useful for video anti-aliasing and reconstruction filtering for composite (CVBS) or S-Video signals in standard definition digital TV applications. The third-order filter is to have an insertion loss of more than 20 dBs at 13.5 MHz and more than 40 dBs at 27 MHz [12.16].



Note1: 75 Ω
Note 2: Adjust group delay

(a)



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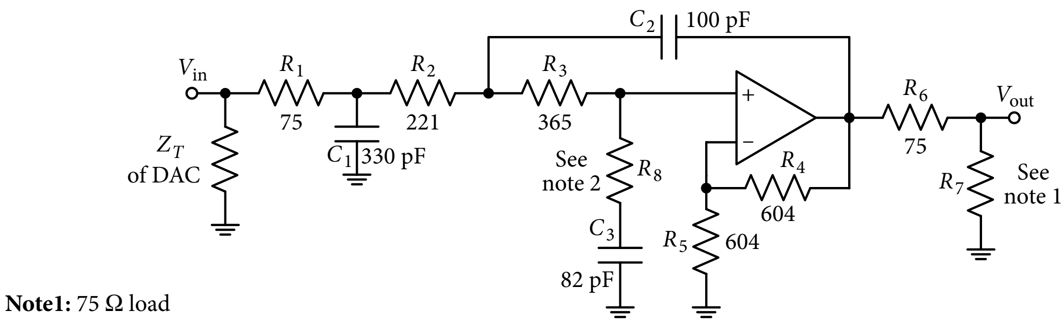
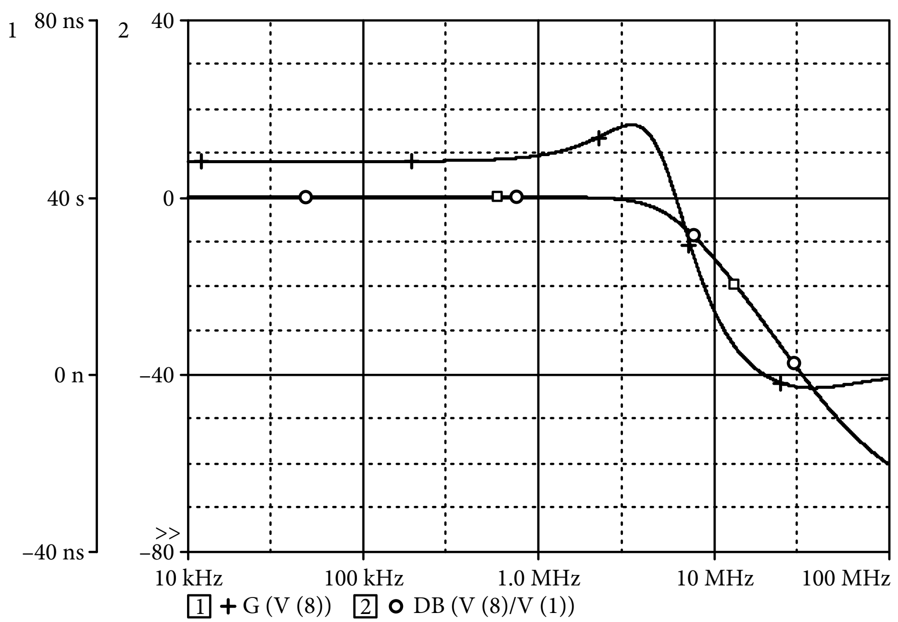


(c)



(d)

Figure 12.25 (a) A reconstruction filter for PAL version {Courtesy: Maxim Integrated}. (b) Simulated response of the reconstruction filter shown in Figure 12.25(a). (c) A reconstruction filter for NTSC version {Courtesy: Maxim Integrated}. (d) Simulated response of the reconstruction filter shown in Figure 12.25(c).

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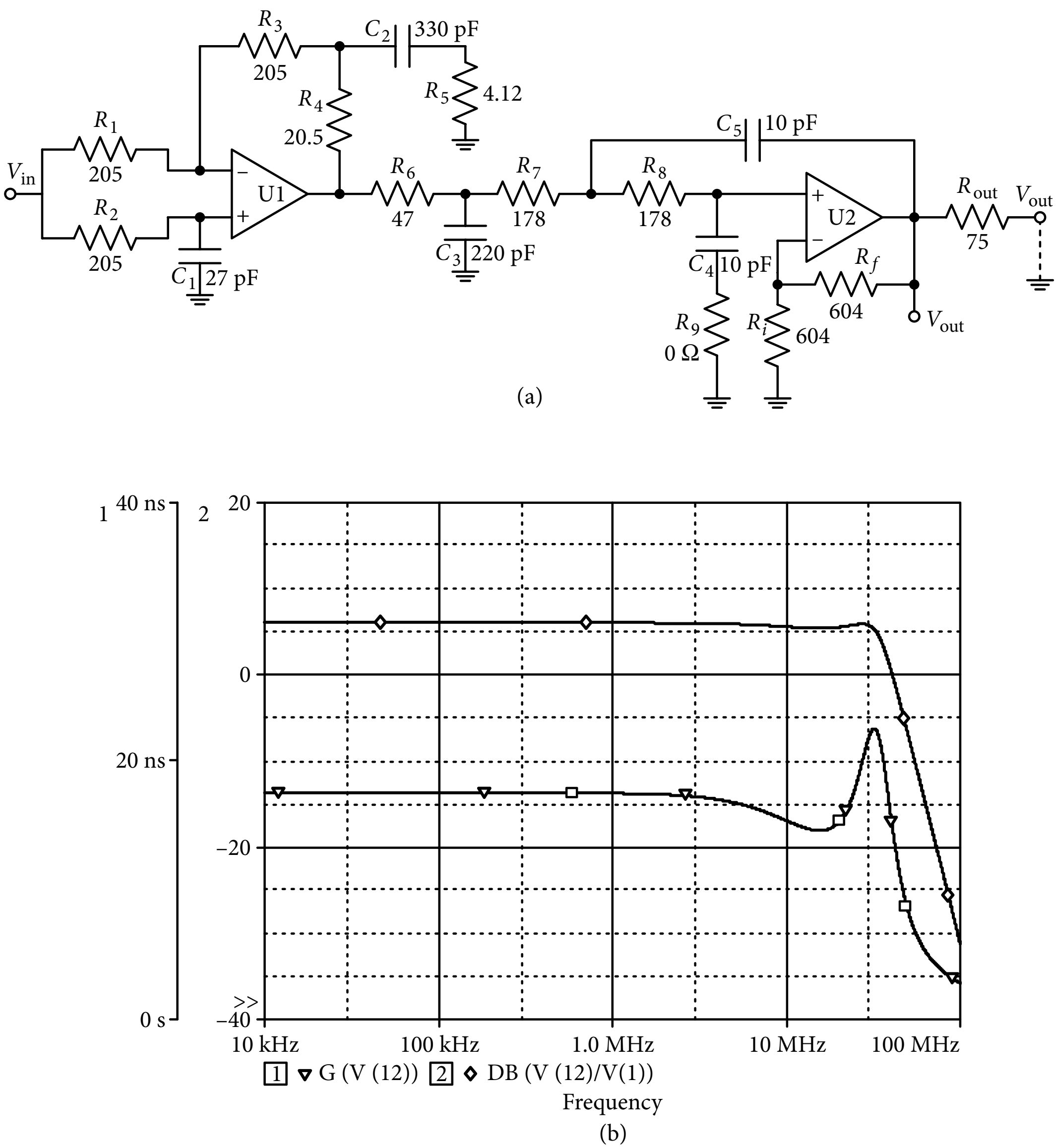

Figure 12.26 (a) A 20 MHz, 3-pole Butterworth reconstruction filter for XGA {Courtesy: Maxim Integrated}. (b) Simulated magnitude and phase responses of the filter shown in Figure 12.26(a).

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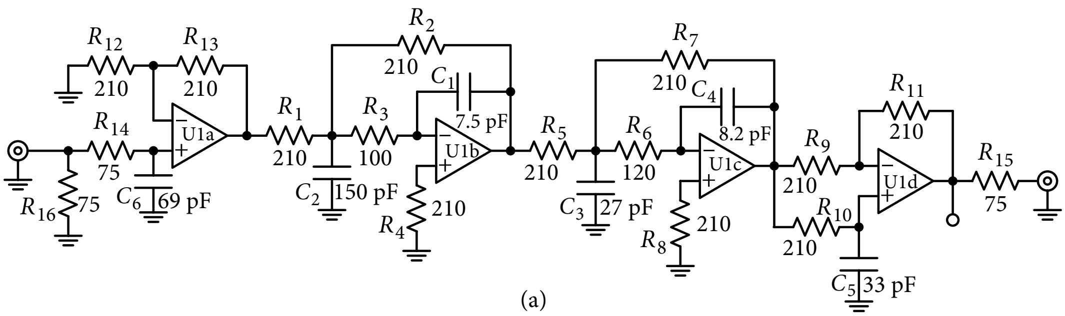
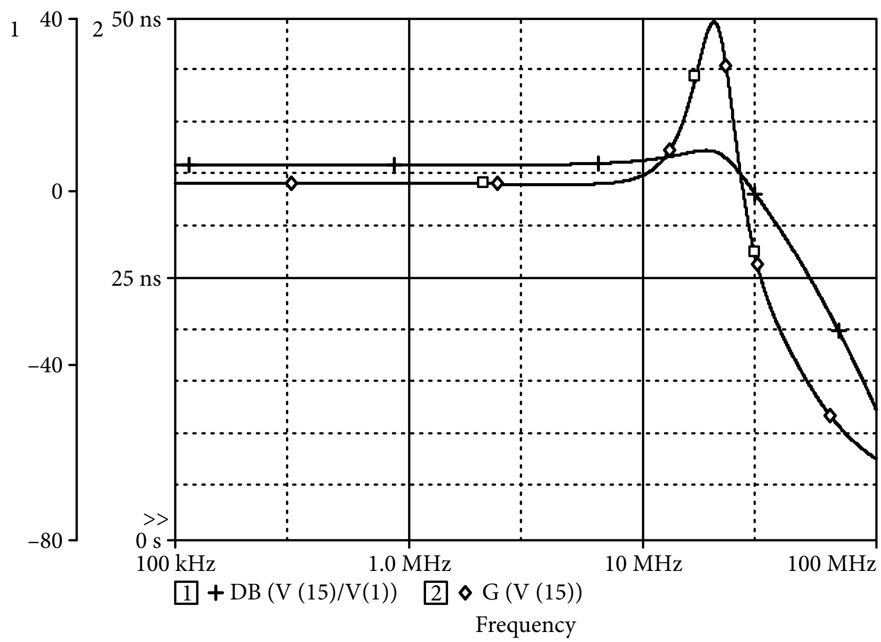


(a)



(b)

Figure 12.27 (a) A reconstruction filter for HDTV {Courtesy: Maxim Integrated}. (b) Simulated response of the filter shown in Figure 12.27(a).

For a standard quality video wave form reconstruction, group delay of the filter needs to be minimized and any group delay differential between it also needs to be minimized. The Sallen and Key filter, with a slight modification as shown in Figure 12.28(a) is used where addition of resistance R_8 in series with C_3 and R_3 creates a lag-lead network. The sum of R_3 and R_8 is kept constant equal to the original value that keeps the bandwidth of the dominant pole constant. When the value of R_3 is increased, it introduces a *lead* term, which lowers the group delay value. Figure 12.28(b) shows the simulated magnitude response with $R_3 + R_8 = 332\ \Omega$ and group delay variation over the pass band for three combination of resistances as shown in the simulation results.

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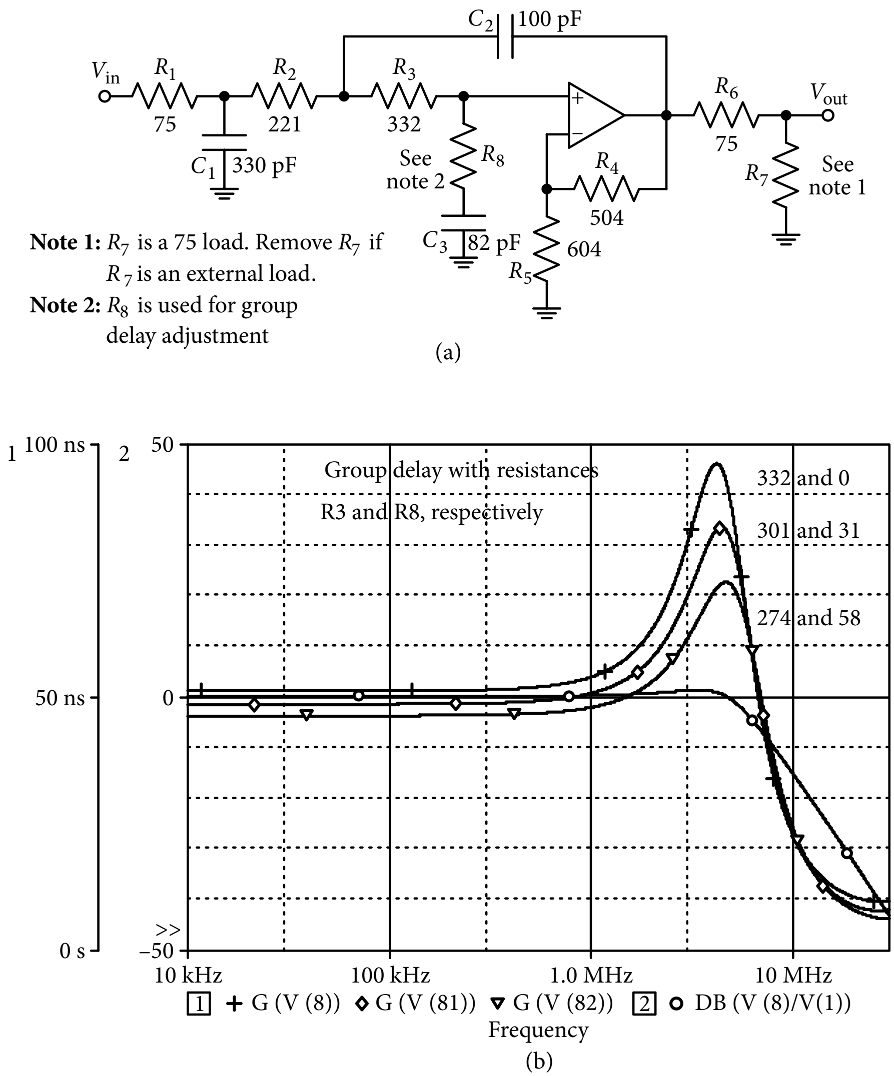

Figure 12.28 (a) Standard quality video wave form reconstruction filter {Courtesy: Maxim Integrated}. (b) Magnitude and group delay response of the 5 MHz video filter using modified Sallen and Key second-order filter.

The Ackerberg–Mossberg (AM) circuit shown in Figure 8.8 is used to provide a similar response using MAX 4390 OAs. A first-order LP section with a dc gain of 2 is cascaded to the second-order AM section as shown in Figure 12.29(a). Pole-location of the third-order Butterworth filter is as shown here:

$$s_1 = -0.5,\ s_{2,3} = -1 \pm j\,0.866 \tag{12.19}$$

To realize the real pole s_1 for the first-order filter section in Figure 12.29(a), the design value of the elements is obtained using equations (1.25) and (1.26). The obtained element values

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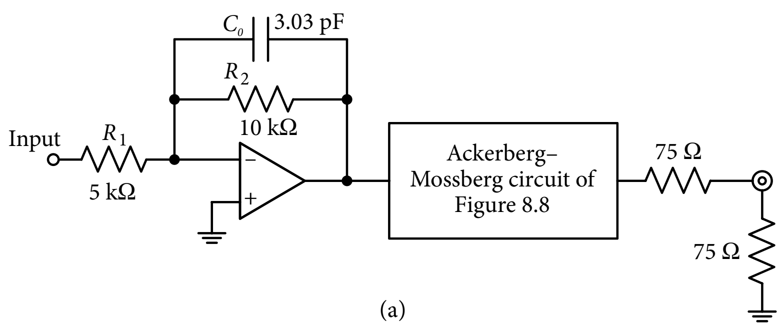
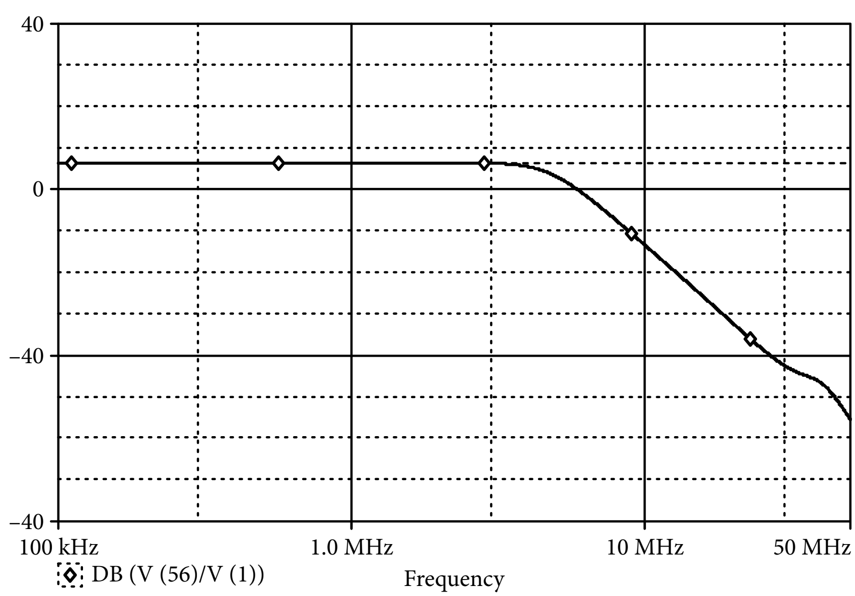
for a dc gain of 2 are shown in Figure 12.29(a). For the AM second-order filter, relations in equation (8.23) and (8.25) are used; the obtained element values are as follows:

$$R = 10\ \text{k}\Omega,\ C = 3.0303\ \text{pF},\ QR = 10\ \text{k}\Omega,\ (R/k) = 10\ \text{k}\Omega,\ R^* = 10\ \text{k}\Omega.$$

Figure 12.29(b) shows the simulated response of the third-order filter with a 3 dB frequency of 5.045 MHz. The design value of pole-Q is unity in this case. Its attenuation is 21.7 dBs at 13.5 MHz and 40.5 dBs at 27 MHz; the realized attenuation is only a fraction better than the design. Figure 12.29(c) also shows the simulated variation in group delay response for $Q = 1.2$ and 1.4. For realizing $Q = 1.2$ and 1.4, all the component values remain the same except QR, which changes to 12 kΩ and 14 kΩ, respectively.



(a)



(b)

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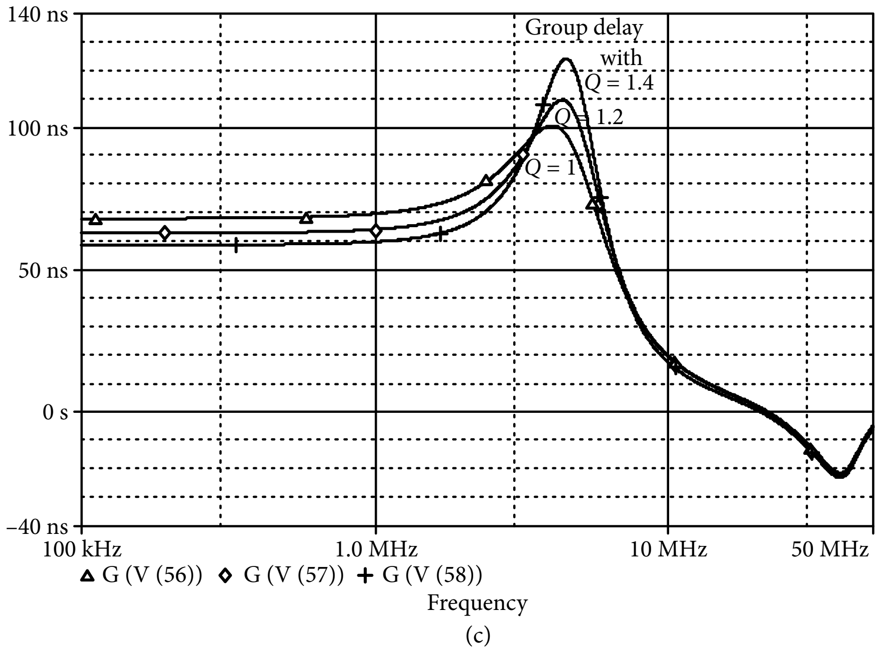

Figure 12.29 (a) Another scheme for reconstructing the standard quality video wave form using the Ackerberg–Mossberg filter configuration. (b) Simulated magnitude response of the 5 MHz video filter based on the AK biquad employed in Figure 12.29(a) and (c) variation in group delay.

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Practice Problems

12-1 Design and test active crossover network of Figure 12.2(a) for the crossover frequency of 10 krad/s.

12-2 (a) Design a third-order Butterworth filter for eliminating infrasonic noise signal having cut-off frequency of 200 rad/s; employ the structure shown in Figure 12.3.

(b) Design a fourth-order Bessel filter for eliminating ultrasonic noise signals having cut-off frequency of 250 krad/s.

(c) Obtain magnitude attenuation of the Bessel filter at 125 krad/s.

(d) Simulate combined magnitude response of the infrasonic and the ultrasonic filters. What are the frequencies at which gain becomes half of its maximum value?

12-3 To address the unequal output power requirement of the satellite speaker and the subwoofer, following filters are required. An HPF for a cut-off frequency of 3 krad/s and a BPF by combining a HPF and a LPF. Range of the respective cut-off frequencies of the HPF and LPF are from 500 rad/s to 600 rad/s and from 600 rad/s to 3 krad/s.

12-4 Repeat the problem 12-3 for the HPF to have the cut-off frequency of 4 krad/s and the composite BPF, for which range of frequencies for HPF is from 550 rad/s to 650 rad/s, and for the LPF frequency range is from 650 rad/s to 3.5 krad/s.

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12-5 Obtain a mid bass boost filter circuit by combining a 12-dB treble cut having cut-off frequency of 800 Hz and a 12-dB treble boost having cut-off frequency of 200 Hz. What is the resultant peak mid bass boost and the frequency at which it occurs?

12-6 Obtain a mid treble cut filter circuit by combining a 12-dB bass boost having cut-off frequency of 200 Hz and a 12-dB bass cut having cut-off frequency of 20 kHz. What is the resultant peak mid treble cut and the frequency at which it occurs?

12-7 Obtain a mid treble boost filter circuit by combining a 12-dB treble cut having cut-off frequency of 8 kHz and a 12-dB treble boost having cut-off frequency of 800 Hz. What is the resultant peak mid treble boost and the frequency at which it occurs?

12-8 Obtain a mid bass cut filter circuit by combining a 12-dB bass boost having cut-off frequency of 200 Hz and a 12-dB bass cut having cut-off frequency of 800 Hz. What is the resultant peak mid bass cut and the frequency at which it occurs?

12-9 Design a notch filter, which has notch at 3 krad/s and width of the notch is 1/7 of an octave. Attenuation at the notch is to 6 dBs. Assume a suitable value of the quality factor.

12-10 Design a programmable notch filter which has the width of the notch as 1/7 of an octave and attenuation at the notch is to be 12 dBs to compensate the effect of ringing. Notch frequency of the filter are 2.5 kHz and 3.4 kHz. Assume suitable value of quality factor.

12-11 First-order APFs are cascaded to provide a phase shift of 90° between the two signals reaching left and right ears of a listener. Signal frequencies are in the range of 1500 rad/s and 15 krad/s. Design a suitable circuit and obtain the phase shift to the signal coming to the left and right ear, and the frequency range for which difference in the phase shift between the two outputs is 78° or more.

12-12 Analog signal with a maximum frequency of 1 krad/s is fed to a 10-bit SAR ADC. Sampling frequency rate of ADC will be 16 kHz. Design a maximally flat analog filter which would have sufficient attenuation (6 dB/bit) at the Nyquist frequency.

12-13 Repeat problem 12-12 if the analog signal frequency range extends up to 2 krad/s.

12-14 Repeat problem 12-12 employing Chebyshev approximation with maximum ripple width of 1 dBs in the pass band.

12-15 Repeat problem 12-13 employing Chebyshev approximation with maximum ripple width of 1 dB in the pass band.

12-16 Derive the transfer function for the circuit in Figure 12.25(a).

12-17 A third-order reconstruction filter is to be designed using Ackerberg–Mossberg configuration with following specifications. Insertion loss of more than 25 dBs at 15 MHz and more than 45 dBs at 30 MHz.

Test the circuit and find out simulated intrinsic loss at 15 MHz and 30 MHz.

12-18 What will be the value of group delay for the filter used in problem 12-17 with $Q = 1$, 1.2 and 1.5 at 1 MHz, 7.5 MHz and 15 MHz.

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