Active Crossover Networks for Noncoincident Drivers

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The spatial separation between drivers in a loudspeaker system affects the radiation pattern over the frequency range where more than one driver contributes to the total acoustic output. An analysis of conventional crossover networks shows that the main lobe of the radiation pattern shifts in direction and increases in amplitude. A new network transfer function, which can easily be realized with operational amplifiers, eliminates this problem. Additional active delay networks are used to compensate for offsets in the acoustical planes from which the individual drivers radiate. The audibility of phase distortion is investigated with the conclusion that it is undetectable for the proposed types of networks.

INTRODUCTION: The design of a frequency-dividing network to feed the individual drivers in a loudspeaker system is a difficult task if optimum results are desired. In the case of a passive dividing network between a single amplifier and several drivers, problems arise because the impedance of these drivers is frequency dependent. Most likely this impedance is not purely real over the frequency range where the transition in acoustic output from one driver to the other has to be made. Thus simple filter theory cannot be used. Furthermore the drivers are likely to have different efficiencies so that attenuation has to be incorporated into the dividing network. This increases the source impedance which the driver sees and reduces the damping of its mechanical resonance.

Because maximum damping is needed for the woofer to control its low-frequency behavior, no attenuation can be allowed in the low-pass filter which drives the woofer. This then means that the tweeter has to have efficiency equal to or higher than the woofer and restricts the choice of units that could be combined with a particular woofer or necessitates the use of matching transformers.

Finally there is the problem of designing the dividing network so that it presents a tolerable impedance to the power amplifier driving it. The active crossover network gets around all of these problems and offers some additional benefits. Each driver is driven directly from its own power amplifier whose gain may be adjusted to equalize differences in efficiencies between the drivers. This gives complete freedom in selecting drivers.

The output impedances of the power amplifiers directly control the damping of the drivers without intervening crossover networks, and each can be optimized for its particular driver if necessary.

The load presented to each amplifier is less frequency dependent. System distortion is reduced because the woofer channel may be driven into clipping without also affecting the midrange or tweeter channel.

Because the signal is divided into the separate frequency ranges at the input to the power amplifier where one deals with low-level signals and a well-defined interface between the electronics, it now becomes possible to choose from a wide variety of dividing networks which can be conveniently realized with operational amplifiers.

The difficulty now becomes to decide which network is optimum. This paper is an attempt to classify the different networks which have been used, to point out some of their properties which have not been considered before, and to propose a unique class of filters which appears to be the best engineering compromise. This compromise is based upon a consideration of the radiation patterns produced by different networks operating with noncoincident drivers as well as the transient responses of those networks.

THE PROBLEM

The design problem is illustrated in Fig. 1. Two sound sources H and L contribute to the sound pressure at point P_1 in space. This point is "on axis" of the cabinet C–C, but at unequal distances from the drivers because the effective plane of radiation for driver L is offset by the distance d_2 from that of driver H.

The objective is to apply such an input signal to H and L that the sound pressure at P_1 is independent of frequency. To simplify the analysis, the sound pressure at point P_0 will be considered first. This point has equal distance from H and L and is assumed to be located on the axis of both drivers.

The signal arriving at this point has the transfer function

$$F_0 = F_H + F_L. \quad (1)$$

F_H is the transfer function of the driver H and its associated high-pass filter, while F_L is the corresponding low-pass function. The driver axis can be made to coincide with the cabinet axis either by physically mounting driver H by a distance d_2 behind the plane C–C or by electrically delaying the signal which is applied to H by the time which it takes a signal to propagate the distance d_2 in air. A simple circuit for producing such electrical delay will be shown later.

The problem is thus reduced to that of Fig. 2. Of interest is the sound pressure not only on axis but also at angles α off axis. This is modified by the separation of the drivers d_1 which causes different path lengths to point P and thus changes the phase with which the sound pressures from H and L combine at P. It is also influenced by the relative phase of the electrical signals driving H and L.

First the sound pressure for the on-axis point P_0 will be considered. The combined acoustic output should be as frequency independent as possible. It will be described by the transfer function F_0. Later on the effect of the transfer function F_0 upon the radiation pattern will be investigated.

THREE FILTER TYPES

The transfer function

$$F_0 = F_H + F_L \quad (1)$$

will meet the requirement of frequency independence to varying degrees, depending on the crossover network design. F_0 can be classified to belong to one of the following three types of functions:

1) F_0 is frequency independent in amplitude and phase;

2) F_0 is frequency independent in amplitude only and exhibits a frequency-dependent phase shift;

3) F_0 is frequency dependent both in amplitude and phase.

The first type is the transfer function for the "constant-voltage" crossover network:

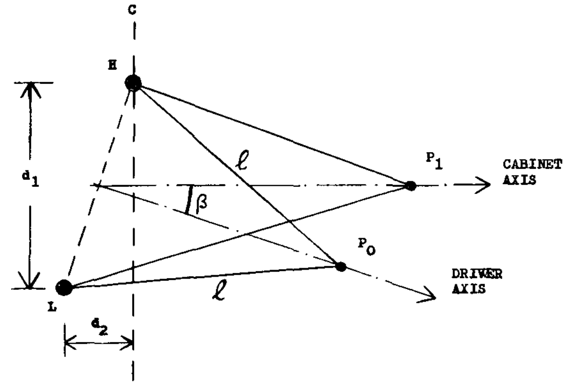


Fig. 1. Radiation from two sources H and L which are separated by a distance d_1 and offset by d_2 in their planes of radiation.

$$F_0 = F_H(s) + F_L(s) = 1 \quad (2)$$

where s denotes the complex frequency,

$$s = \sigma + j\omega \quad (3)$$

where

$$\omega = 2\pi f \quad (4)$$

the radian frequency. Eq. (2) would seem to give the ideal crossover network because it introduces no frequency response or transient distortion.

The second type of transfer function describes the "all-pass" crossover network:

$$F_0 = F_H(j\omega) + F_L(j\omega) = 1\, e^{j\varphi(\omega)}. \quad (5)$$

Its steady-state frequency response is unity, but it has a frequency-dependent phase shift $\varphi(\omega)$ which produces delay or transient distortion.

The third type of network is a "compromise" between the constant-voltage and the all-pass networks:

$$F_0 = F_H(j\omega) + F_L(j\omega) = [1 + r(\omega)]\, e^{j\varphi(\omega)}. \quad (6)$$

Some frequency response ripple $r(\omega)$ is traded for reduced delay distortion due to $\varphi(\omega)$.

EXAMPLES FOR THE THREE FILTER TYPES

The constant-voltage crossover network has been described in detail elsewhere [1]–[4]. The combination of a first-order Butterworth low-pass filter with a first-order Butterworth high-pass filter gives the simplest constant-voltage crossover:

$$F_L = \frac{1}{1 + s_n} \quad (7)$$

$$F_H = \frac{s_n}{1 + s_n} \quad (8)$$

$$F_H + F_L = 1. \quad (2)$$

Here s_n is the complex frequency normalized to the nominal crossover frequency f_c:

$$s_n = \frac{s}{2\pi f_c}. \quad (9)$$

This is not a very practical filter because of its slow cutoff behavior of 6 dB per octave. A more useful filter with 12-dB per octave slopes is synthesized from the following transfer functions [1]:

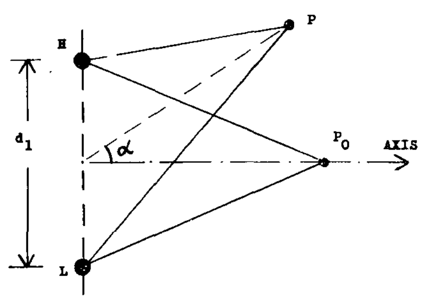


Fig. 2. Radiation from coplanar sources H and L.

$$F_L = \frac{1 + 3.7\, s_n}{1 + 3.7\, s_n + 3.7\, s_n^2 + s_n^3} \qquad (10)$$

$$F_H = \frac{3.7\, s_n^2 + s_n^3}{1 + 3.7\, s_n + 3.7\, s_n^2 + s_n^3}\,. \qquad (11)$$

It will be shown that the constant-voltage networks produce undesirable radiation patterns when the drivers are noncoincident.

The criterion for an all-pass crossover network is

$$F_0 = F_L(j\omega) + F_H(j\omega) = 1\, e^{i\varphi(\omega)}. \qquad (5)$$

The third-order Butterworth low-pass and high-pass filters form such networks. The low-pass transfer function is

$$F_L = \frac{1}{1 + 2\, s_n + 2\, s_n^2 + s_n^3} \qquad (12)$$

and the high-pass is

$$F_H = s_n^3\, F_L. \qquad (13)$$

The combined transfer function is

$$F_0 = F_L \pm F_H = \frac{1 \pm s_n^3}{1 + 2\, s_n + 2\, s_n^2 + s_n^3}\cdot \qquad (14)$$

Note that either sign can be chosen simply by reversing the phase of one of the drivers.

After substituting $s_n = j\omega_n$, where $\omega_n = \omega/2\,\pi f_c$, the combined frequency response is then

$$|F_0(j\omega_n)| = \left| \frac{1 \mp j\omega_n^3}{1 + 2j\omega_n - 2\,\omega_n^2 - j\omega_n^3} \right|$$

$$= \sqrt{\frac{1 + \omega_n^6}{(1 - 2\omega_n^2)^2 + (2\omega_n - \omega_n^3)^2}} = 1. \qquad (15)$$

The phasing of the drivers makes no difference to the steady-state amplitude response. This is so because the phase difference between F_L and F_H is 270° for the plus sign and 90° for the minus sign. The two outputs are always in phase quadrature.

This can be seen readily from the complex frequency plane of Fig. 3. The poles of F_L are the same as those of F_H. The high-pass filter has three additional zeros at the origin of the complex frequency plane. These zeros cause an additional phase shift of 3 × 90° relative to the low-pass filter for any frequency ω.

It can be shown that the group delay

$$t_g = -\frac{d\varphi}{d\omega} \qquad (16)$$

is different for the two connections:

$$t_{g+} = 2\,\frac{1 + \omega_n^2}{1 - \omega_n^2 + \omega_n^4} \qquad (17)$$

$$t_{g-} = 2\,\frac{1}{1 + \omega_n^2}\cdot \qquad (18)$$

The in-phase connection of the drivers results in four times as much delay at the crossover frequency $\omega_n = 1$ as the out-of-phase connection, which therefore should be chosen to minimize transient distortion.

The third-order Butterworth filter is particularly useful for passive crossover networks because of its steep 18-dB per octave cutoff slopes and low delay distortion, but it gives a radiation pattern which is not symmetrical with respect to the driver axis because the outputs from H and L are in phase quadrature.

The "compromise" network is characterized by

$$F_0 = F_L(j\omega) + F_H(j\omega) = [1 + r(\omega)]\, e^{j\varphi(\omega)} \qquad (6)$$

and

$$t_g = -\frac{d\varphi}{d\omega} \qquad (16)$$

optimized. The transfer functions F_L and F_H might be Gaussian or Butterworth/Thomson filters which would cause some ripple r in the frequency response of F_0 but reduce the frequency dependency of the group delay. The second-order Butterworth function which is often used for crossovers will be considered as part of this class. For the low-pass network,

$$F_L = \frac{1}{1 + \sqrt{2}\, s_n + s_n^2}\,\cdot \qquad (19)$$

The high-pass network is

$$F_H = s_n^2 F_L \qquad (20)$$

and the combined output

$$F_0 = F_L \pm F_H = \frac{1 \pm s_n^2}{1 + \sqrt{2}\, s_n + s_n^2}\,\cdot \qquad (21)$$

The frequency response becomes

$$|F_0(j\omega_n)| = \left| \frac{1 \mp \omega_n^2}{1 + j\omega_n\sqrt{2} - \omega_n^2} \right| = \frac{1 \mp \omega_n^2}{\sqrt{1 + \omega_n^4}}\cdot \qquad (22)$$

If the drivers are connected in phase, the combined output will cancel at the crossover frequency $\omega_n = 1$. For the out-of-phase connection a 3-dB rise in frequency response results at the crossover frequency.

The group delay is the same for either connection:

$$t_g = \sqrt{2}\,\frac{1 + \omega_n^2}{1 + \omega_n^4}\cdot \qquad (23)$$

This group delay is 40% larger at the crossover frequency $\omega_n = 1$ than that of the third-order Butterworth derived network, Eq. (18). The 3-dB peak in frequency response and slower cutoff rate of 12 dB per octave would seem to make this network a poor choice except for cost.

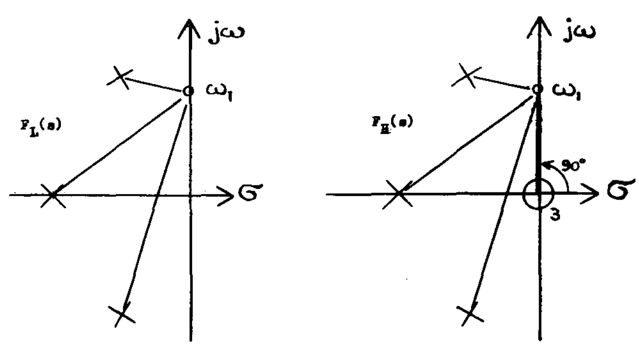


Fig. 3. Complex frequency plane representation of the third-order Butterworth low-pass and high-pass filters.

RADIATION PATTERN

It should be emphasized again that the discussion of crossover networks so far only considered the sound pressure at points P_0 which are equidistant from both drivers (Figs. 1 and 2). For this case the constant-voltage crossover network is clearly optimum because it introduces no phase shift or amplitude variation to the combined outputs from H and L. Now the frequency response at some arbitrary point P will be analyzed. For simplification this point is assumed to be far enough away from the drivers so that lines drawn from H and L to P are essentially parallel (Fig. 4). The difference in path length between P and the two drivers is

$$l = d_1 \sin \alpha. \tag{24}$$

This causes a phase shift difference between the output signals coming from H and L of

$$\varphi_l = 360° \frac{l}{\lambda} = 360° \frac{d_1}{\lambda} \sin \alpha. \tag{25}$$

Here λ is the wave length of the frequency radiated from H and L. The electrical networks driving H and L have their own phase shift φ_L and φ_H which depends on the type of filter used. The total phase difference $\Delta\varphi$ between the signals from H and L at point P then becomes

$$\Delta\varphi = \varphi_H - \varphi_L + 360° \frac{d_1}{\lambda} \sin \alpha. \tag{26}$$

Signals will add in phase at point P in space whenever

$$\Delta\varphi = \pm n\, 360° \tag{27}$$

and subtract when

$$\Delta\varphi = \pm (2n + 1)\, 180° \tag{28}$$

where $n = 0,1,2,\cdots$. The strongest interaction between the drivers will occur at the crossover frequency where both contribute equal-amplitude signals. For frequencies which are much higher or lower than the crossover point the radiation pattern of the system is determined only by the radiation pattern of the driver which is active in that range.

RADIATION PATTERN FOR DIFFERENT FILTER TYPES

The previously discussed crossover functions will be analyzed for their radiation patterns at the crossover frequency. The constant-voltage network is represented by:

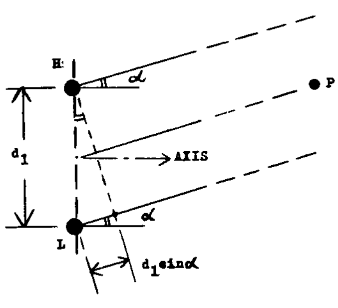


Fig. 4. Definitions for calculation of far-field radiation pattern.

$$F_L = \frac{1 + 3.7\, s_n}{1 + 3.7\, s_n + 3.7\, s_n^2 + s_n^3} \tag{10}$$

$$F_H = \frac{3.7\, s_n^2 + s_n^3}{1 + 3.7\, s_n + 3.7\, s_n^2 + s_n^3}. \tag{11}$$

The all-pass network consists of a third-order Butterworth low-pass and high-pass network connected out of phase:

$$F_L = \frac{1}{1 + 2\, s_n + 2\, s_n^2 + s_n^3} \tag{12}$$

$$F_H = -\, s_n^3 F_L. \tag{29}$$

The second-order Butterworth function is used for a compromise filter with 3-dB ripple:

$$F_L = \frac{1}{1 + \sqrt{2}\, s_n + s_n^2} \tag{19}$$

$$F_H = -\, s_n^2 F_L. \tag{30}$$

It will be assumed that the spacing d_1 between the drivers equals one wavelength at the crossover frequency, i.e.,

$$\frac{d_1}{\lambda} = 1. \tag{31}$$

This could be the case of a 10-inch (254-mm) frame for the woofer and a 4-inch (101-mm) frame for the tweeter mounted as close together as possible so that d_1 = 7-inches (177-mm) and considering a crossover frequency of 1.7 kHz.

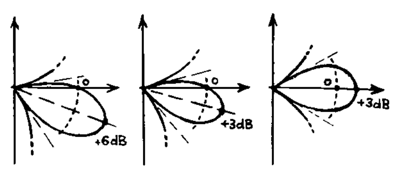
For this case the phase difference at P becomes, from Eqs. (26) and (31),

$$\Delta\varphi = (\varphi_H - \varphi_L) + 360° \sin \alpha. \tag{22}$$

Table I summarizes the characteristics of the networks.

The constant-voltage design exhibits a radiation pattern with a 6-dB peak 20° below the axis and signal cancellation

Table I. Characteristics at the crossover frequency for a relative driver spacing of $d_1/\lambda = 1$ and $d_2 = 0$.

	Constant Voltage	Allpass	Compromise
Magnitude of F_H or F_L	0 dB	−3 dB	−3 dB
Relative phase $\varphi_H - \varphi_L$	120°	90°	0°
Cutoff slope	12 dB/oct	18 dB/oct	12 dB/oct
Angle α for maximum amplitude	−20°	−15°	0°
Maximum amplitude	+6 dB	+3 dB	+3 dB
Angle α for signal cancellation	+10° −56°	+15° −49°	±30°
Schematic radiation pattern			

at 10° above the axis. The network has its desired properties only for points on axis which are on the skirt of the radiation pattern between signal cancellation and the 6-dB peak.

This is clearly not an ideal filter because the radiation pattern shifts with frequency. At low and high frequencies the maximum of radiation occurs along the driver axis. For frequencies in between the pattern tilts downward and increases 6 dB in amplitude.

The all-pass network as exemplified by the third-order Butterworth shows a slightly better behavior because it peaks by only 3 dB at 15° below the axis. Furthermore the frequency range over which it shows this directional shift in the pattern is narrower than for the constant-voltage filter because it has steeper cutoff slopes. This reduces the overlap and interaction between drivers.

The second-order Butterworth design as the "compromise" filter has a symmetrical radiation pattern. It is the only filter of the three which would result in a loudspeaker system with an acoustic axis that does not move as frequency changes. Unfortunately though it still has a 3-dB peak at the crossover frequency.

The narrowing in the beam width of the radiation pattern in these examples was caused by the distance d_1 between the drivers. Increasing d_1/λ will increase the number of peaks and cancellations and result in a multibeam radiation pattern. This may mask other faults in a crossover design. For best performance though the drivers must be as close together as possible, i.e., $d_1/\lambda < 1$.

The tilting in the pattern is controlled by the phase difference $\varphi_H - \varphi_L$ between the drive signals. In order to avoid the frequency-dependent tilt this phase difference has to be zero as in the example of the second-order Butterworth network, or the drivers have to be coaxially mounted.

OPTIMUM CROSSOVER FUNCTION

From the foregoing observations of the three different networks it is now possible to set down the requirements for the filter function which would give the optimum performance when the drivers are separated by a distance d_1, as is the case for most practical loudspeaker systems.

1) The phase difference $\varphi_H - \varphi_L$ between the drive signals at the crossover frequency has to be zero in order to avoid tilting in the radiation pattern.

2) The output amplitude from the high-pass and low-pass section has to be 6 dB down at the crossover frequency so that the sum of the two is unity and no peaking occurs.

3) The phase difference $\varphi_H - \varphi_L$ has to be the same for all frequencies so that the symmetry of the radiation pattern is preserved above and below the crossover frequency.

This last requirement is the same as saying that high-pass and low-pass filters must change their phase at the same rate with frequency, that is, they must have identical group delay. High-pass and low-pass filters with identical poles in the complex frequency plane and zeros at $s = 0$ for the high-pass filter will satisfy this criterion. For $\varphi_H - \varphi_L \equiv 0$ there has to be an even number of zeros, and for a -6-dB crossover amplitude the poles have to be double poles.

It has been pointed out by R. Riley that the cascade of two identical Butterworth filters will meet all the above requirements. Thus all the available design information for active Butterworth filters can be directly used for this type of crossover network.

The resulting network is an all-pass filter. For example, cascading two first-order Butterworth filters to obtain a second-order, 12-dB per octave slope crossover network:

$$F_L = \frac{1}{(1 + s_n)^2} \tag{33}$$

$$F_H = \left(\frac{s_n}{1 + s_n}\right)^2 \tag{34}$$

$$F_0 = F_L \pm F_H = \frac{1 \pm s_n^2}{(1 + s_n)^2} \tag{35}$$

$$|F_0(j\omega_n)| = \left| \frac{1 \mp \omega_n^2}{1 - \omega_n^2 + j^2\omega_n} \right| = \frac{1 \mp \omega_n^2}{1 + \omega_n^2}. \tag{36}$$

Again the drivers have to be connected out of phase, otherwise cancellation would occur.

Phase reversal of one of the drive signals can serve as an easy performance test. The amount of on-axis cancellation which can thus be obtained is a direct measure of how well the outputs from H and L are matched in amplitude and phase.

NETWORK REALIZATION

Of practical interest are the two- and four-pole crossover networks with 12- and 24-dB per octave cutoff slopes. Fig. 5 shows possible designs which give the amplitude and group delay responses shown in Fig. 6. It should be noted that the 24-dB per octave filter function is difficult to realize as a passive network because of the double complex poles.

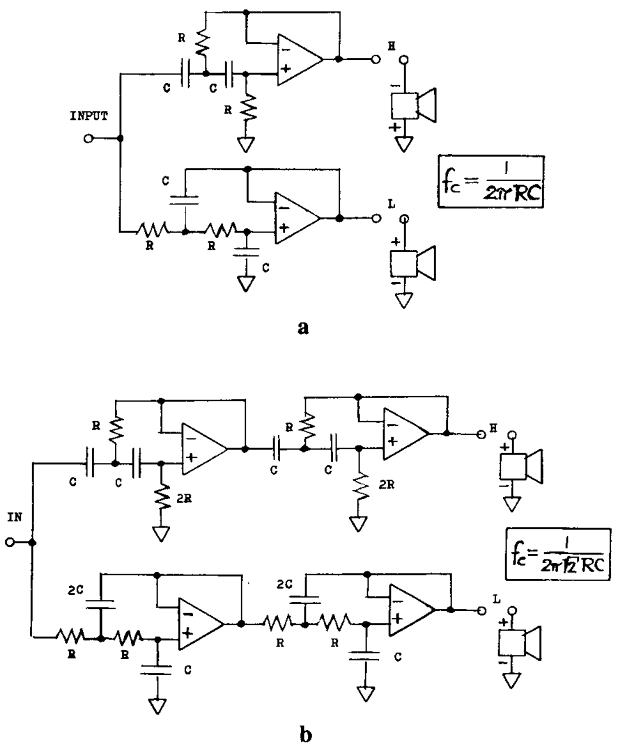


Fig. 5. Realization of optimum crossover function. **a.** 12-dB per octave filter. **b.** 24-dB per octave filter.

CORRECTING DRIVER OFFSET

In the analysis so far it has been assumed that both drivers radiate from the same acoustical plane and that the distance d_2 in Fig. 1 is zero. Going back to the earlier example of a 10-inch (254-mm) woofer and a 4-inch (101-mm) tweeter frame, the distance d_2 might be estimated to be the offset between voice coils when both drivers are mounted in the same cabinet panel C–C. Assuming d_2 = 3 inches (75 mm) and d_1 = 7 inches (177 mm), the radiation pattern will be tilted at the crossover frequency by the angle

$$\beta = \arctan \frac{d_2}{d_1} = 23^\circ . \tag{37}$$

In order to preserve the benefits of the new crossover design this frequency-dependent rotation of the radiation pattern has to be eliminated. Probably the easiest way to accomplish this is to delay the electrical signal to driver H by the time

$$t_g = \frac{d_2}{v} = \frac{3 \text{ inches} \cdot \text{seconds}}{12\ 624 \text{ inches}} = 240\ \mu\text{s}. \tag{38}$$

First-order all-pass networks of the configuration in Fig. 7 can be cascaded to give sufficient delay. The transfer function for this circuit is

$$F(s) = \frac{1 - sCR}{1 + sCR} \tag{39}$$

and the delay is

$$t_g = -\frac{d\varphi}{d\omega} = \frac{2RC}{1 + (\omega RC)^2} \tag{40}$$

$$t_g = \frac{1}{\pi f_0} \frac{1}{1 + (f/f_0)^2} \tag{41}$$

with

$$f_0 = \frac{1}{2\pi RC}. \tag{42}$$

Because this delay is frequency dependent except when $f << f_0$, care has to be taken that the frequency f_0 is sufficiently higher than the crossover frequency f_c. Requiring that $f_0 \geq 3 f_c$, the delay t_g' which can be obtained per stage becomes

$$t_g' \leq \frac{1}{10 f_c} \tag{43}$$

For a 1.7-kHz crossover frequency $t_g' = 60\ \mu$s, and thus four stages have to be cascaded to obtain the 240-μs delay needed in the example.

It is obvious that a large amount of circuitry is required when the acoustical planes of the two drivers are very different. But without this correction the main lobe of the radiation pattern would shift away from the cabinet axis whenever both drivers contribute to the total acoustic output, namely, in the crossover region of the frequency spectrum.

For optimum crossover network design the drivers have to be made to radiate from the same acoustical plane which should be parallel to the cabinet front panel. Then the cascaded Butterworth sections of Fig. 5 will give a frequency-stable main axis of the radiation pattern and constant amplitude for the summed signals from H and L.

AUDIBILITY OF DELAY DISTORTION

The optimum network has all-pass characteristics. The amplitude response is unity, but the phase of the summed signal changes with frequency. The network has a frequency-dependent group delay

$$t_g = \frac{2}{1 + \omega_n^2} \tag{44}$$

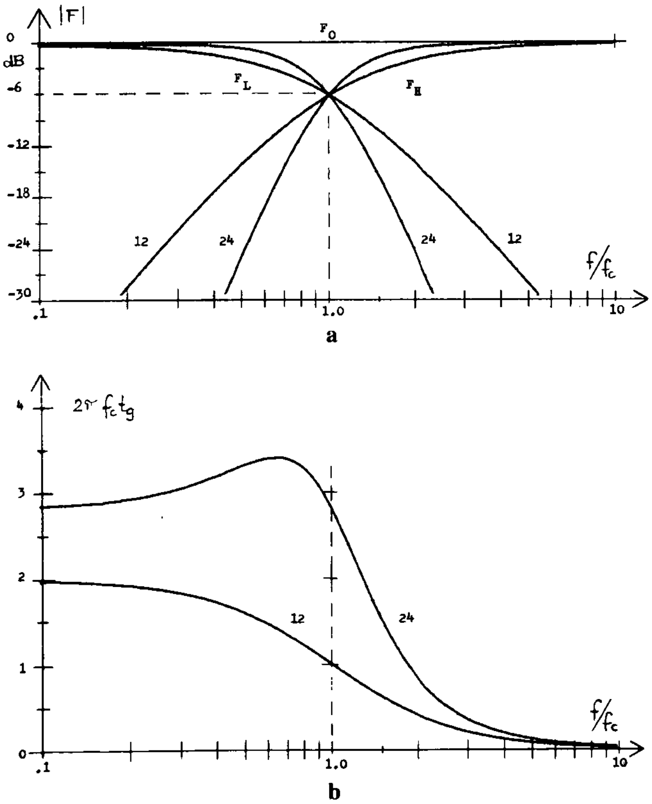


Fig. 6. Optimum crossover function response. **a.** Amplitude response. **b.** Group delay response.

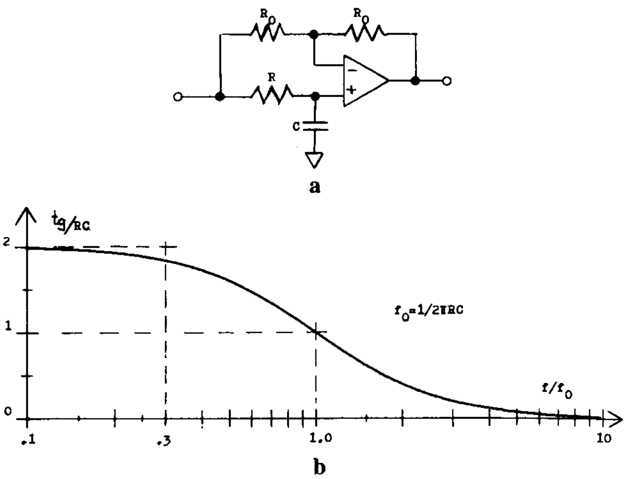


Fig. 7. **a.** First-order all-pass delay network. **b.** Its group delay response.

for the 12-dB per octave crossover and

$$t_g = 2\sqrt{2}\,\frac{1 + \omega_n^2}{1 + \omega_n^4} \qquad (45)$$

for the 24-dB per octave networks.

Whenever a signal that is made up of more than one spectral component is subjected to a frequency-dependent delay, the relative phase between the spectral components is shifted and a phase-distorted time function of the signal results. The question arises, when does the delay distortion become audible? The circuit of Fig. 7 can be used to introduce varying amounts of delay distortion into a signal path. Fig. 8 gives a configuration for higher order all-pass networks where $F_1(s)$ might be a band-pass filter. Using headphones it has been found that delay distortion is generally not audible even when the waveforms observed on an oscilloscope look greatly distorted [5].

There are a few exceptions which have limited practical significance.

1) Some change can be noticed even with a first-order all-pass filter when listening to clicks of very low repetition rate (< 5Hz).

2) A second-order all-pass network can introduce a ringing sound on clicks and square waves if its Q is high enough.

3) A change in sound character can be detected with high sound level square waves and first-order all-pass networks. This may involve nonlinearities in the ear.

Because the optimum filters described have a gradually changing group delay characteristic, no ringing can be noticed and they introduce no audible delay distortion on program material.

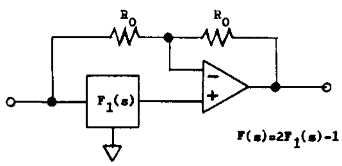


Fig. 8. General circuit for realizing all-pass transfer functions.

CONCLUSION

Active crossover design in the past has not considered the effects of mounting distances between drivers. The crossover networks based upon cascaded Butterworth sections have optimum radiation characteristics for systems where the distance between the drivers is not small compared to a wave length at the crossover frequency. This is the case for the majority of loudspeaker systems.

To utilize these networks it is necessary that both drivers radiate from the same acoustical plane. This can be achieved either mechanically by appropriate mounting or electrically by delay networks. Furthermore the drivers should be mounted one above the other and as close to each other as possible to obtain wide dispersion in the horizontal plane and a minimum number of radiation lobes in the vertical plane.

While the active-network approach may be costly and only applicable to top performance loudspeaker systems, it can result in a system with improved dispersion characteristics and thus a very firm stereo image.

On appropriate program material it will not only give a smooth lateral spread of the image but also a noticeable enhancement of depth perspective.

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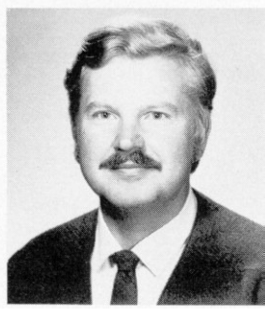

Siegfried H. Linkwitz was born in Bad Oeynhausen, Germany, in 1935. He received the Diplom-Ingenieur degree in electrical engineering from Darmstadt University in 1961. Upon graduation he joined Hewlett Packard Company in California as a research engineer for the design of microwave test equipment. He was involved with the development of signal generators, phase measuring RF voltmeters, and spectrum analyzers. At present Mr. Linkwitz is program manager in microwave spectrum analysis.

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