TUNABLE ANALOG LOUDSPEAKER CROSSOVER NETWORK

Eduardo Rapoport, Fernando A. P. Baruqui and Antonio Petraglia

Program of Electrical Eng., COPPE, EE - Federal University of Rio de Janeiro
CP 68504, CEP 21945-970, Rio de Janeiro, RJ - Brazil
{rapoport,baruqui,petra}@pads.ufrj.br

ABSTRACT

This paper proposes a new implementation of tunable analog loudspeaker crossover networks suitable to monolithic integrated circuit (IC) realizations. The crossover frequency is linearly set by tuning voltages to adequate the audio reproduction to different environments and loudspeaker characteristics over the entire audio spectrum. Based on structurally allpass networks, the proposed approach presents very low sensitivity to IC device mismatching. This is verified by simulation and comparison with an alternative design. Details of the tuning procedure are provided. Experimental verification of the structure low sensitivity and tuning capabilities is also shown.

1. INTRODUCTION

Because of the practical difficulty of a single loudspeaker to meet the frequency response requirements in the whole audio frequency range (20 Hz $\sim$ 20 kHz), crossover networks (e.g., [1]-[4]), are widely used to split the incoming audio signal into two [1]-[3] (or more [4]) frequency bands, as illustrated in Fig. 1. Consequently, each loudspeaker driver (see Fig. 2) is fed within the appropriate frequency band.

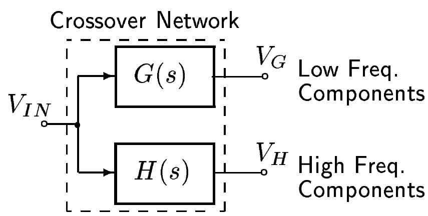

Figure 1: Two-way crossover network.

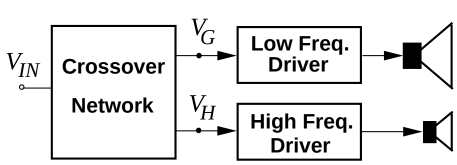

Figure 2: Loudspeaker crossover system.

While both analog [1]-[3] and digital [4] crossover systems have been reported, the one described in [2] includes a tuning scheme by which the crossover frequency can be adjusted to compensate for loudspeaker and environment differences. Using RC-active circuits, the crossover network has been implemented in [2] as sum and difference of allpass outputs, as sketched in Fig. 3. In practice, however, variations in the circuit elements may severely destroy the crossover frequency response symmetry with respect to the crossover frequency (see Sec. 4), contributing distortion in the audio reproduction.

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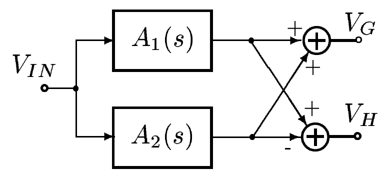


Figure 3: Implementation of the crossover network using allpass building blocks.

This paper proposes the implementation of the class of crossover networks depicted in Fig. 3, but based on structurally allpass filters to maintain symmetry in frequency response, even in the occurrence of element variations. Moreover, the structure advanced here employs OTA-C techniques [5], so that filter transfer functions depend on transconductances (g_m) and capacitances only. The crossover frequency is adjusted by tuning voltages that modify transconductances, exploiting the linear dependence of g_m on V_{DS} [6] to ensure precise and easy control of the crossover frequency. By contrast, the scheme described in [2] relies on simultaneous adjustments of three distinct resistors, thus requiring a sophisticated tuning procedure.

2. CROSSOVER NETWORK DESIGN

An important requirement for the crossover network design is that the filter pair $[G(s), H(s)]$ be allpass complementary, that is

$$G(s) + H(s) = A(s) \tag{1}$$

where $A(s)$ is an allpass transfer function. As a result

$$|G(j\omega) + H(j\omega)| = 1, \quad \text{for all } \omega \tag{2}$$

indicating that the phasor sum magnitude of the crossover filters equals unity over all the audio frequency range. In absence of loudspeaker driver errors, and if group delay distortion can be tolerated by the listener, this property ensures that the sound pressure level at the listener is proportional to the system input signal, independently of frequency [7].

Additionally, the frequency responses of $G(s)$ and $H(s)$ must be sufficiently selective to avoid leakage of large amplitude frequency components outside the band each loudspeaker was designed to reproduce. Such selectivity requirements may be achieved by constraining the transfer function pair so that it satisfies the power complementary property:

$$|G(j\omega)|^2 + |H(j\omega)|^2 = 1, \quad \text{for all } \omega \tag{3}$$

Transfer functions that simultaneously satisfy Eqs. (2) and (3) are called doubly complementary pairs, and can be designed by odd-order classical filter approximations (e.g., Butterworth, Chebyshev and elliptic). The necessary and sufficient conditions for the existence of realizable doubly complementary filter pairs have been outlined in [2]. Here we summarize main results and focus on efficient realizations of tunable crossovers networks based on allpass decomposition for OTA-C realizations.

2.1. Doubly Complementary Crossovers

The decomposition of $G(s)$ and $H(s)$ in terms of the allpass transfer functions in Fig. 3, that is,

$$G(s) = \frac{1}{2}(A_1(s) + A_2(s)) \tag{4}$$

$$H(s) = \frac{1}{2}(A_1(s) - A_2(s)) \tag{5}$$

is achieved by first expressing the loss function of $G(s)$, for instance, as

$$T(s) = \frac{1}{G(s)G(-s)} = 1 - C^2(s) \tag{6}$$

where $C(s)$ is the characteristic function. Then the left half plane roots of $C(s) = 1$ and $C(s) = -1$ are assigned as poles of $A_1(s)$ and $A_2(s)$, respectively. Finally, these allpass transfer functions are substituted in Eqs. (4) and (5) to form the transfer functions $G(s)$ and $H(s)$.

2.2. Structurally Allpass Realization

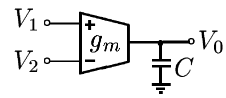
It has been shown that an n-th-order allpass transfer function $A(s)$ can be decomposed as [5]

$$A(s) = 1 - \frac{2}{1 + Y(s)} \tag{7}$$

where $Y(s)$ is an admittance function, which can be expanded as

$$Y(s) = a_1 s + \cfrac{1}{a_2 s + \cfrac{1}{a_3 s + \cdots \cfrac{1}{a_n s}}} \tag{8}$$

A ladder-based structure can thus be derived observing that the terms $1/a_i s$ are implemented by the association of a transconductor and a capacitor in integrator configuration, as shown in Fig. 4. Observe from Eq. (8) that an n-th-order allpass filter employs n (grounded) capacitors yielding a canonic crossover network. Note also that $A_1(s)$ and $A_2(s)$ are added in Eq. (1) to implement $G(s)$, and at almost no additional cost are subtracted to implement $H(s)$. Therefore, the implementation of Fig. 1 would require approximately twice the number of capacitors, assuming $G(s)$ and $H(s)$ are also realized by canonic structures.



$$\frac{V_0}{V_1 - V_2} = \frac{1}{\frac{C}{g_m}s} = \frac{1}{a_i s}$$

Figure 4: OTA-C integrator implementing $1/a_i s$.

3. DESIGN EXAMPLE

To verify the sensitivity of the crossover network advanced here and compare it with the one proposed in [2], a *3rd-order* elliptic pair $[G(s),\, H(s)]$ has been designed and implemented as in Fig. 3, having a maximum of 0.1 dB ripple in the passband and at least 16.4 dB attenuation in the stopband. The lowpass transfer function is:

$$G(s) = \frac{0.621s^2 + 1.97}{s^3 + 1.88s^2 + 2.36s + 1.97} \tag{9}$$

whose characteristic function is

$$C(s) = \frac{s^3 + 0.975s}{0.621s^2 + 1.97} \tag{10}$$

yielding the allpass transfer functions

$$A_1(s) = \frac{s^2 - 0.628s + 1.58}{s^2 + 0.628s + 1.58} \tag{11}$$

$$A_2(s) = \frac{-s + 1.23}{s + 1.23} \tag{12}$$

Applying the decomposition in Eq. (8) one obtains

$$Y_1(s) = 1.59s + \frac{1}{0.398s} \tag{13}$$

$$Y_2(s) = 0.801s \tag{14}$$

The complete crossover circuit schematic diagram is shown in Fig. 5. Notice that the allpass (normalized) transconductances are equal to either 1 or 2, allowing the application of layout techniques to improve both matching among transconductors and tuning accuracy, as discussed in the following paragraphs.

4. SIMULATION AND EXPERIMENTAL RESULTS

A prototype crossover network has been built with RCA CA3080 bipolar OTAs and capacitors of 32.8 nF, 8.2 nF and 16.5 nF. Details of the tuning process are given next.

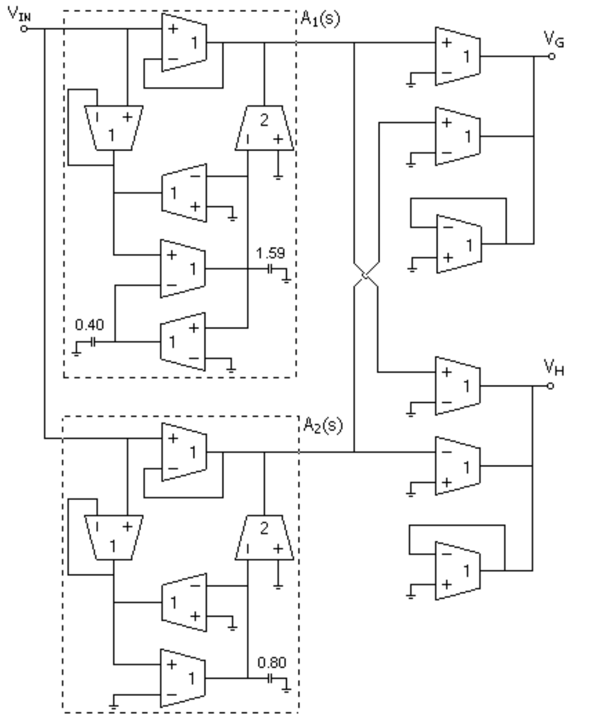


Figure 5: Complete crossover schematic diagram. Note inside the transconductors that either $g_m = 1$ or $g_m = 2$.

4.1. Tuning Procedure

Since the basic circuit element is the integrator of Fig. 4, where $V_0/(V_1 - V_2) = g_m/sC$, the crossover frequency can be easily modified by scaling all g_m values by a common factor, say α. Thus we have $V_0/(V_1 - V_2) = \alpha g_m/sC$ for each integrator, and $[G(j\omega/\alpha),\ H(j\omega/\alpha)]$ for the crossover pair in Fig. 1.

The CA3080 transconductance relates to its biasing current according to $g_m = 20 I_{bias}$, and can therefore be scaled by externally providing I_{bias}. This can be established by a resistance and a biasing voltage, R and V_{bias} in Fig. 6, so that $I_{bias} = (V_{bias} - 0.7)/R$ and consequently $g_m = 20(V_{bias} - 0.7)/R$. As the transconductances in Fig. 5 are $g_m = 1$ and $g_m = 2$, they can all be simultaneously scaled using a single controlling element, V_{bias}, as illustrated in Fig. 6.

4.2. Tuning in Integrated Circuit Implementation

In a monolithic realization, each transconductance can also be controlled by a biasing current, since $g_m \propto \sqrt{I_{bias}}$. All transconductances are then adjusted by a single element through current mirrors precisely implemented, as illustrated in Fig. 7. The high accuracy of this procedure follows from the fact that the currents to be copied are either equal or twice the reference current.

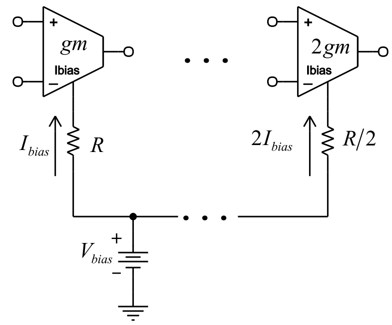


Figure 6: Tuning scheme for the discrete prototype crossover.

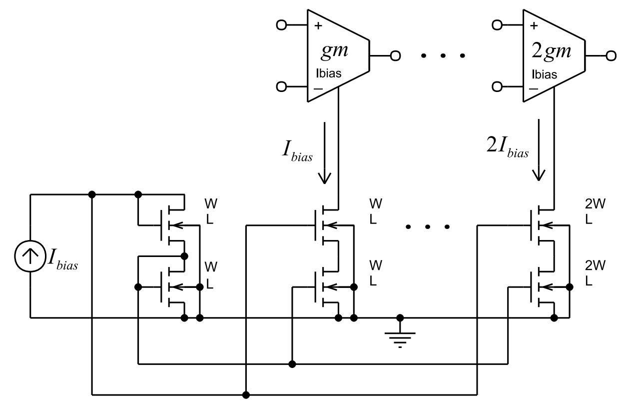


Figure 7: Tuning scheme in integrated circuit implementation.

4.3. Experimental Verification

A crossover frequency of 550 Hz was initially established with $V_{bias} = 5$ V, providing transconductances of 6.13 mS and 12.3 mS. Then V_{bias} was changed to 0 V and -5 V, moving the crossover frequency to 355 Hz and 175 Hz, respectively. The experimental frequency responses are shown in Fig. 8.

4.4. Sensitivity Results and Comparisons

Sensitivity simulations have been carried out for both the proposed crossover network (Fig. 5) and the one in [2], considering random errors with 1% standard deviation in resistances, conductances and capacitances. To show benefits of the allpass structuralness property of the allpass filters in Fig. 5, ideal adders and subtractors have been assumed in both Fig. 5 and [2]. It should be pointed out that while the adder and subtractor in Fig. 5 employ transconductors with equal g_m's, the ones in [2] use resistors of different values. As a consequence, the addition and subtraction operations are more accurately implemented in Fig. 5 than in [2]. Figs. 9(a) and 9(b), respectively, present the $\pm\sigma$ boundaries enclosing about 68.3 % of all passband and stopband frequency responses. Observe that near the (normalized) pole frequencies of around 0.9 rad/sec for the lowpass filter and 1.6 rad/sec for the highpass filter, the proposed structure

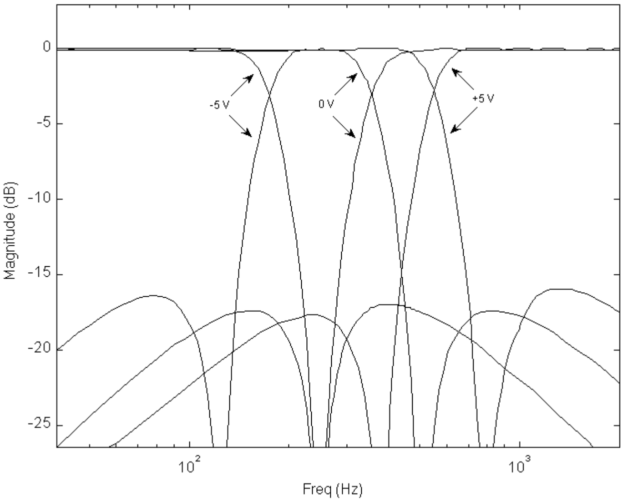


Figure 8: Experimental results showing tuning capabilities of the proposed design for $V_{bias} = -5$ V, 0 V, and +5 V.

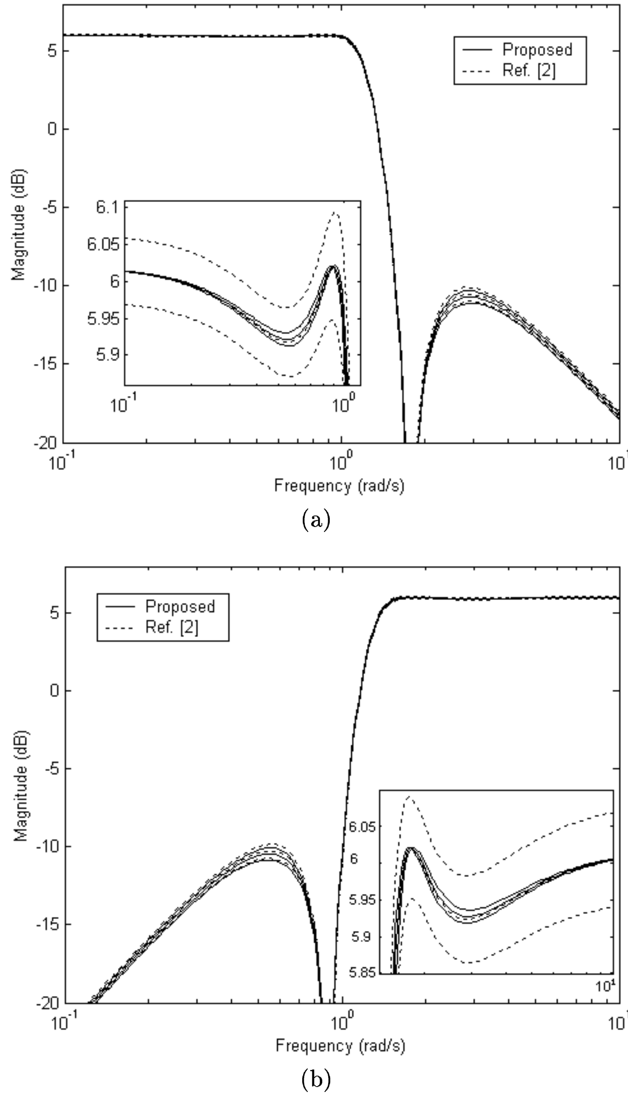


Figure 9: Sensitivity performance comparison in the passband (a) and stopband (b): proposed (solid) and Ref. [2] (dashed).

has zero deviation, a consequence of the allpass structuralness property. Also, notice in the passband detail that in addition to its larger passband sensitivity, the RC-active structure in [2] presents a translation in amplitude. Since the filters $G(s)$ and $H(s)$ in [2] are also generated by sum and subtraction of $A_1(s)$ and $A_2(s)$, a gain increase in $H(s)$ implies in a gain reduction in $G(s)$, and vice-versa, causing random amplitude asymmetry between the audio channels.

5. CONCLUSIONS

A tunable OTA-C crossover network for audio loudspeaker systems was advanced. Comparisons with a previously reported approach indicated that the proposed structure had smaller sensitivity with respect to random component errors, and simpler and more effective tuning procedure of the crossover frequency. Details of the tuning circuitry suitable to IC realization were shown. Experimental results obtained with a discrete prototype were also presented, verifying theoretical predictions.

6. REFERENCES

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