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REALIZATION OF COCHLEAR FILTERS BY VLT SWITCHED CAPACITOR BIQUADS

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ABSTRACT

We describe here the realization of cochlear filters using switched capacitor biquads. This approach is made possible by a new design technique, called *charge-differencing* (CD), which easily reduces 25% (or more) the silicon area required to implement very large time-constant (VLT) biquads. In this technique, filter time constants are controlled by both the product of two capacitor ratios and the *differences* of capacitor values, making the capacitor spread ratio very small. The new switched capacitor biquads are also stablized against op-amp nonideal characteristics, such as input offsets and finite gains, using a two-phase gain- and offset-compensation method.

1. INTRODUCTION

The auditory system can detect, recognize, and localize complex sounds accurately and rapidly, even in the presence of high levels of ambient noise. Several factors are responsible for this superior performance, most important among them are the unique properties of the cochlea. Functionally, the cochlea can be thought of as a bank of asymmetrically shaped filters, with bandwidths and center frequencies that change systematically along its length. Examples of the amptitude frequency, amptitude spatial, and spatial-temporal responses of the simulated 4th-order cochlear filters are shown in Fig. 1, 2 and 3 respectively. In order to mimic the performance of the auditory system, cochlear models such as this have been experimented with over the last few years as front-ends in speech analysis and recognition systems. A principle obstacle to their wider acceptance, however, is their heavy computational cost. Consequently, hardware implementations of these models have been an attractive option to achieve real-time performance. So far, successful attempts have employed analog designs, specifically, using subthreshold and operating transconductance amplifiers to build a cascade of second-order filter stages[1, 2]. A second approach is to use a bank of switched-capacitor filters (SCF's) [3]. SCF's in general have a wide dynamic range, and extremely precise and reliable response characteristics that would obviate the need for any post-fabrication tuning.

2. CHARGE-DIFFERENCING TECHNIQUE IN DESIGNING COCHLEAR FILTERS

Several difficulties arise in designing SCF's for cochlear processing. The most serious is the need for a frequency range which is broad (0.03-20 kHz) and stretches to relatively low frequencies. Conventional biquad designs[4, 5] require a capacitor spread ratio of approximately $1/(\Omega_0 T)$, where Ω_0 is the pole frequency of the biquad and T is the sampling period. For low frequency channels, this ratio becomes very large, and VLT circuits have to be used[6, 7]. Yet even these circuits may not be adequate for the case with extremely low pole frequency. Furthermore, existing VLT circuits are achieved by signal attenuation, rendering them vulnerable to op-amp nonideal characteristics such as input offset-voltage v_{os} and finite gain A.

A charge-differencing technique is introduced to reduce the capacitor spread ratio. Moreover, with the implementation of gain- and offset-compensation (GOC) technique, the effect due to v_{os} and A will be much reduced[8].

2.1 GOC Charge-differencing Integrator

A charge-differencing integrator is shown in Fig. 4. Gain- and offset-compensation is achieved by c_o, an offset-storing capacitor. It functions as follows: when switches 1 are closed, the charge $v_{in}(n)c_1$ is accumulated in the capacitors c_A and c_{a2}. The output voltage is then sampled by c_{a1}, i.e., the charge $(v_{in}(n)c_{a1}c_1)/(c_A + c_{a2})$ is transfered into c_{a1}. When the switches 2 are closed, a charge $((v_{in}(n)+v_{os})c_1)$ is effectively pulled back to ground from capacitors c_A and c_{a1}. Since c_A received the charge $(v_{in}(n)c_A c_1)/(c_A+c_{a2})$ in the previous phase, c_{a1} has to compensate the net difference charge $((v_{in}(n)c_{a2}c_1)/(c_A + c_{a2}) + v_{os}c_1)$ before it redistributes the charge received in the previous phase to c_A. With $A = \infty$, time domain analysis gives

$$
\begin{aligned}
v_{out}(n+\frac{1}{2}) = & - \frac{c_1(c_{a1}-c_{a2})}{(c_A+c_{a1})(c_A+c_{a2})}v_{in}(n) \\
& + v_{out}(n-1/2) \\
& + (\frac{c_A+c_{a1}+c_1}{c_A+c_{a1}} - \frac{c_A+c_1}{c_A+c_{a2}})v_{os}
\end{aligned}
$$

where the index n indicates the time interval $[nT,(n+1/2)T)$. Without c_o, the dependence on v_{os} will be much larger. Now, the gain of the integrator, and hence the time-constant of integration, is controlled by both the product of two capacitor ratios and the difference of capacitor values, making the capacitor spread ratio very small.

The charge-differencing technique is based on the difference of two quantities. Let us consider the quantity $c = a \pm b$. The sensitivity of c with respect to a is S^c_a=(dc/c)/(da/a)=$a/(a \pm b)$. If $a \approx b > 0$, it is obvious that the case of differencing give much higher S^c_a. The same principle applies to the charge-differencing biquads, which means that $c_{a1} - c_{a2}$ cannot be arbitrarily small. A reasonable choice for the low frequency cochlear filters is $c_{a1} = 1.5$ and $c_{a2} = 1$ unit, and the sensitivity with respect to c_{a1} or c_{a2} would be increased by only a few times.

2.2 GOC Charge-differencing Biquads

With the above GOC charge-differencing integrator, two types of biquads (referred to as the E and F circuits in [4], and Type I and Type II biquads in this paper) can be built (Fig. 5 and 6). Type I biquad gives simpler design equations, while those of Type II biquad are slightly more complicated. Their transfer functions ($v_{os} = 0$) are given by:

$$
\begin{aligned}
\mathsf{H}_{bq}(z) &= \frac{v_{out}(z)}{v_{in}(z)} \\
&= -\frac{A-(A+B-C)z^{-1}+(B-D)z^{-2}}{E-(E+F-G)z^{-1}+(F-H)z^{-2}}
\end{aligned}
$$

where

$$
\begin{aligned}
A &= \frac{c_5}{c_2}(\frac{c_B+c_{b1}}{c_{b1}-c_{b2}}) \\
B &= \frac{c_5}{c_2}(\frac{c_B+c_{b1}}{c_{b1}-c_{b2}}) + \frac{c_6}{c_A+c_{a1}} \\
C &= \frac{c_1+c_6}{c_A+c_{a2}} \\
D &= \frac{c_6}{c_A+c_{a1}} + \frac{c_1}{c_A+c_{a2}}.
\end{aligned}
$$

For the Type I biquad,

$$
\begin{aligned}
E &= \frac{(c_B+c_{b1})(c_B+c_{b2})}{c_2(c_{b1}-c_{b2})} \\
F &= \frac{c_3}{c_A+c_{a1}} + \frac{(c_B+c_{b1})(c_B+c_{b2})}{c_2(c_{b1}-c_{b2})} \\
G &= \frac{c_3+c_4}{c_A+c_{a2}} \\
H &= \frac{c_3}{c_A+c_{a1}} + \frac{c_4}{c_A+c_{a2}}.
\end{aligned}
$$

For the Type II biquad,

$$
\begin{aligned}
E &= \frac{(c_B+c_{b1})(c_B+c_{b2}+c_4)}{c_2(c_{b1}-c_{b2})} \\
F &= \frac{c_3}{c_A+c_{a1}} + \frac{(c_B+c_{b1}+c_4)(c_B+c_{b2})}{c_2(c_{b1}-c_{b2})} \\
G &= \frac{c_3}{c_A+c_{a2}} \\
H &= \frac{c_3}{c_A+c_{a1}}.
\end{aligned}
$$

3. DESIGN EXAMPLE: RESPONSE CHARACTERISTICS OF A COCHLEAR BANDPASS FILTER

To demonstrate the advantages of the new design, we simulated the circuit of the lowest frequency cochlear filter as shown in Fig. 7. The transfer function of this channel can be approximated by the following transfer function:

$$
\begin{aligned}
\mathsf{H}(s) &= \mathsf{K}\frac{s^2/(2\pi f_1)^2}{1+s/[Q_1(2\pi f_1)]+s^2/(2\pi f_1)^2} \\
&\cdot \frac{1}{1+s/[Q_2(2\pi f_2)]+s^2/(2\pi f_2)^2}
\end{aligned}
$$

where, K is a gain factor, $Q_1 = 0.9$, $f_1 = 50Hz$, $Q_2 = 2.6$, $f_2 = 100Hz$, and sampling frequency $f_s = 500KHz$. Note that the bandpass filter is realized by cascading one highpass and one LP01 lowpass biquad. To obtain an optimum dynamic range, the capacitor values (after scaling) of the cochlear filter are $c_{A.II} = 26.7$, $c_{B.II} = 31.4$, $c_{2.II} = 1$, $c_{3.II} = 1.18$, $c_{4.II} = 1.34$, $c_{5.II} = 31.4$, $c_{a1.II} = 1.5$, $c_{a2.II} = 1$, $c_{b1.II} = 1.77$, $c_{b2.II} = 1.18$, $c_{A.I} = 48.6$, $c_{B.I} = 18.7$, $c_{2.I} = 1$, $c_{3.I} = 2.6$, $c_{4.I} = 19.7$, $c_{6.I} = 1$, $c_{a1.I} = 3.9$, $c_{a2.I} = 2.6$, $c_{b1.I} = 1.5$, $c_{b2.I} = 1$, and $c_{o1.II} = c_{o2.II} = c_{o1.I} = c_{o2.I} = 1$.

This saves approximately 25% of the silicon area required to implement the same transfer function using other VLT biquads[6] which are claimed to be the most area-efficient so far. The SWITCAP simulation result is also shown in Fig. 8.

4. ACKNOWLEDGEMENTS

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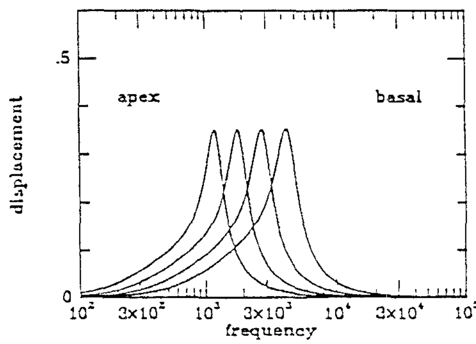

Figure 1: Amptitude frequency response of several cochlear filters.

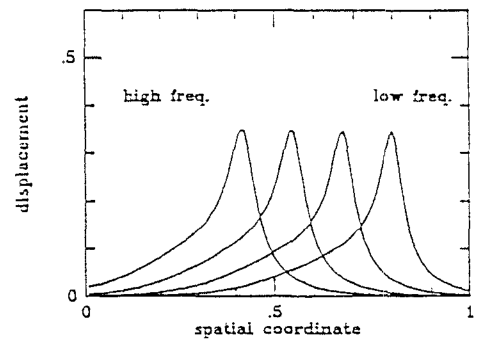

Figure 2: Amptitude spatial response of several cochlear filters.

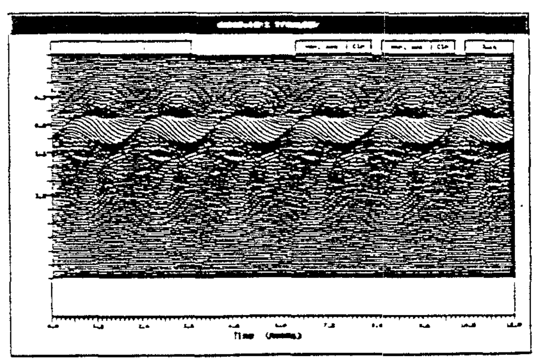

Figure 3: Spatial-temporal response of several cochlear filters.

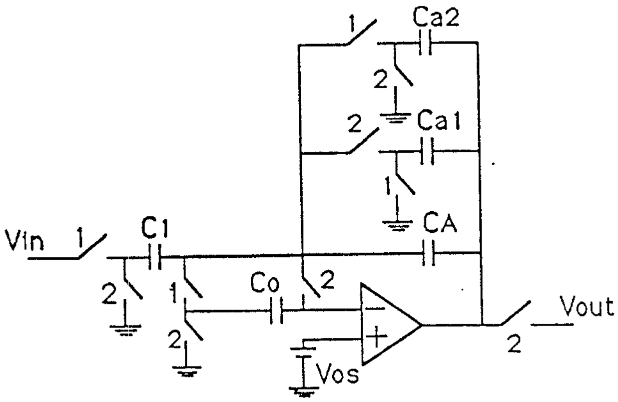

Figure 4: GOC charge-differencing integrator.

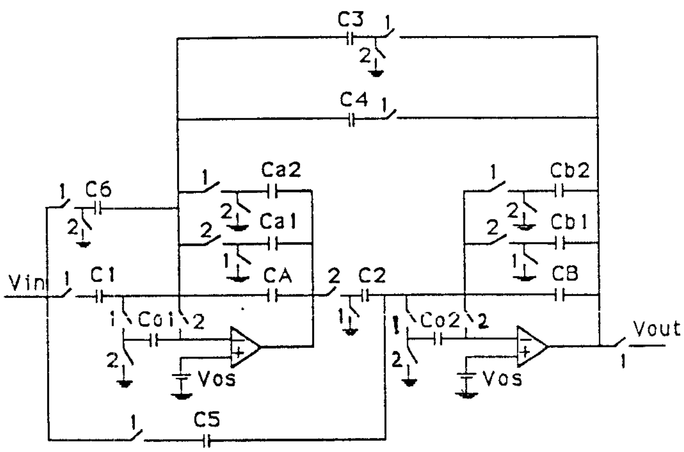

Figure 5: Type I biquad.

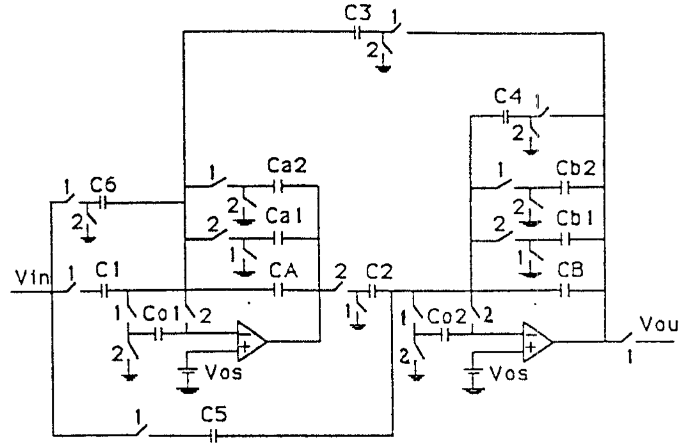

Figure 6: Type II biquad.

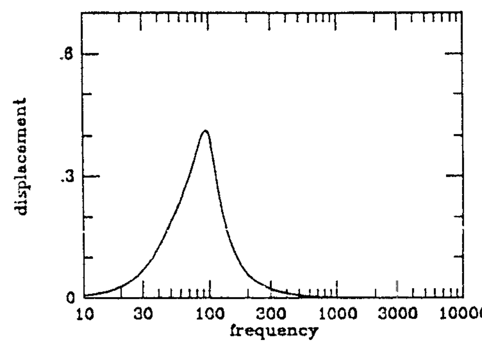

Figure 8: SC cochlear filter response.

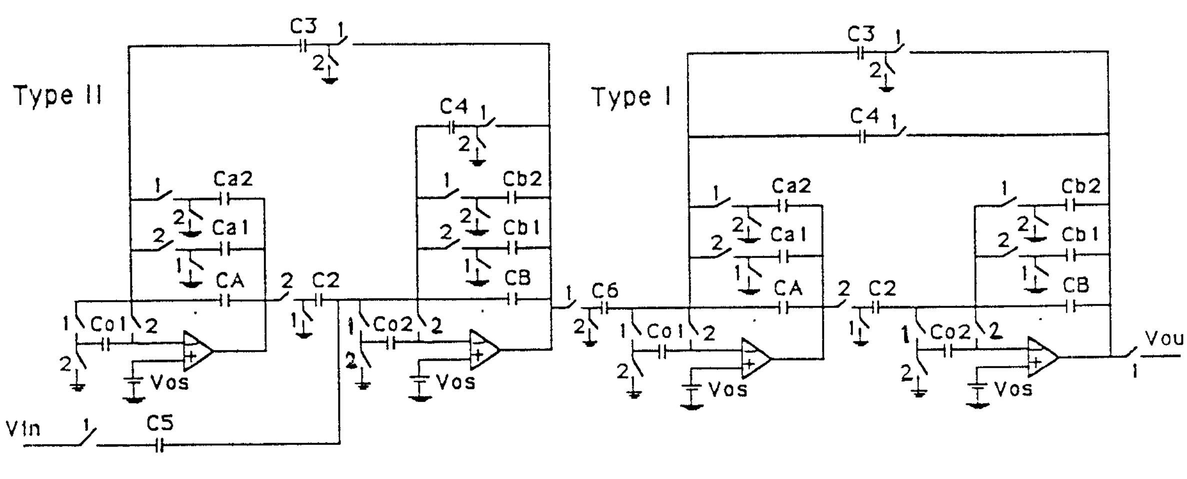

Figure 7: SC cochlear filter circuit.