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A Bio-Realistic Analog CMOS Cochlea Filter With High Tunability and Ultra-Steep Roll-Off

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Abstract—This paper presents the design and experimental results of a cochlea filter in analog very large scale integration (VLSI) which highly resembles physiologically measured response of the mammalian cochlea. The filter consists of three specialized sub-filter stages which respectively provide passive response in low frequencies, actively tunable response in mid-band frequencies and ultra-steep roll-off at transition frequencies from pass-band to stop-band. The sub-filters are implemented in balanced ladder topology using floating active inductors. Measured results from the fabricated chip show that wide range of mid-band tuning including gain tuning of over 20dB, Q factor tuning from 2 to 19 as well as the bio-realistic center frequency shift are achieved by adjusting only one circuit parameter. Besides, the filter has an ultra-steep roll-off reaching over 300 dB/dec. By changing biasing currents, the filter can be configured to operate with center frequencies from 31 Hz to 8 kHz. The filter is order, consumes power and occupies chip area. A parallel bank of the proposed filter can be used as the front-end in hearing prosthesis devices, speech processors as well as other bio-inspired auditory systems owing to its bio-realistic behavior, low power consumption and small size.

Index Terms—Analog VLSI, auditory filter, bio-inspired circuits, CMOS cochlea, floating active inductor.

I. INTRODUCTION

The cochlea in the inner ear of mammals has remarkable filter functions. It converts sound pressure into multi-channels of band-passed outputs, where the sensitivity of each channel is dynamically tuned according to the input intensity and the out-of-band frequency components are greatly suppressed with an ultra-steep roll-off at the stop-band [1]. These filtering features of cochlea make it capable of adapting to wide dynamic range of sound input and performing high-resolution frequency decomposition. In recent years, many bio-inspired systems employing filters that emulate the cochlea filter functions have been implemented, which are used in a variety of applications including hearing prosthetic devices [2]–[4], speech/sound recognition systems [5]–[10] as well as RF spectrum analyzers [11] and channel multiplexers [12]. In this work, we aim to build a cochlea filter in analog VLSI which closely resembles the frequency response of biological cochlea. The use of analog circuitry for front-end signal processing improves power-efficiency [13] (see position of the cochlea filters in Fig. 1). Besides, the progressive results from physiological experiments on the biological cochlea [16]–[25], [27], [28] have provided us with deeper understanding of the cochlea behavior which brings more inspiration to build filters that behave faithfully to biology.

In fact, analog VLSI models of the cochlea have been studied for over two decades and a number of systems have been implemented [29]–[42]. These systems generally consist of filter banks based on second-order sections (SOS) in different configurations including cascade [29]–[35], parallel [36]–[38] and 2-dimensional (2-D) topology [39]–[42]. The filter cascade structure models the wave propagation in the basilar membrane of cochlea using from 32 to 120 stages of SOS connected in series [29], [30], [32]–[34]. Gain and filter roll-off steepness are accumulated with the long cascade, which provides good similarity with biological frequency response. Nevertheless, the cascade structure suffers from the accumulation of both noise and delays, and also, failure of one stage in the cascade will
affect all of its following stages. On the contrary, the parallel structure avoids these drawbacks by employing independent filter channels, and the number of SOS stages in each channel is limited to 1 or 2 [36]–[38]. However, the filter complexity is significantly reduced in each channel, its frequency response is no longer comparable with that observed in the biological cochlea. The 2-D topology solves the problems in both noise and delay accumulation and bio-fidelity. It agrees with the parallel structure in the aspect that SOS stages are configured in a parallel manner, and only one SOS stage is used for each channel [39]–[42]. However, the channels are not independent of each other but are coupled through a resistive network which models the effects of cochlea fluid, and the coupling between the channels results in fairly faithful response compared with biology [40]–[42]. Nevertheless, while the usability of the 2-D topology designs has not been convincingly proven, the cascade and parallel structures have already been used in a variety of audio processing tasks [43]–[45] and particularly, the feasibility of the parallel structure in cochlea implant products has been sufficiently demonstrated [36], [37], [46].

In this work, we aim to improve the performance of the parallel structure in terms of bio-faithful frequency tuning and roll-off steepness. A filter channel which closely emulates the frequency response measured from biological cochlea is developed to replace the simple band-pass filters used in the existing parallel designs [36]–[38]. The filter is designed by directly following the implications from recent physiological experiments. Inspired by the fact that the biological cochlea has separate response features from low to high frequencies [1], the proposed design consists of three cascaded sub-filter stages, which respectively resemble the passive response in low frequencies, active response in mid-band frequencies and steep roll-off in transition frequencies. The use of specialized sub-filters increases overall efficiency, reducing filter tuning complexity and filter order required to achieve bio-faithful response. The sub-filters are built in balanced ladder topology using floating active inductors, which reduces the design complexity. The proposed cochlea filter is superior to the gammatone-based design [47] in terms of ease of analog VLSI implementation and bio-realism of frequency response. Besides, the filter can directly interface with microphone output, unlike the design reported in [48] which has to be operated with a floating current source as input.

The paper is organized as following. The system-level filter structure is introduced in Section II. The details of circuit implementation are discussed in Section III. The circuit non-idealities are analyzed in Section IV. The measured chip results are discussed in Section V, followed by a conclusion of the paper in Section VI.

II. SYSTEM DESIGN

Physiological experiments [1], [16]–[18], [21] indicate that magnitude frequency response of the basilar membrane in biological cochlea has an asymmetric shape and highly active behavior. It has a gentle slope in the low-frequency band (in the region of 20 dB/dec [1], [17], [18]) and a highly steep roll-off at the stop-band (330 dB/dec or even higher [1], [16]–[18]).

In the mid-band, the gain, selectivity and center frequency increases with decreasing input strength: the increase of gain can reach in the range of 20 – 40 dB [16]–[18], [21], the maximum Q factor can reach as high as 10 [16]–[18], [21], and the center frequency increases by over 40% [1], [16]–[18], [21]. Generally speaking, the response is gentle and passive in low frequencies, selective and active in mid-band frequencies and steep in the transition from pass-band to stop-band, which indicates the frequency response of cochlea can be divided into three stages from low to high frequencies as first suggested in [1]. Based on this observation, a filter architecture composed of three cascaded sub-filters, each of which represents one stage of the cochlea response, is proposed as shown in Fig. 2. A biquad band-pass filter (BPF) whose center frequency determines the passive center frequency of the entire cochlea filter presents the gentle and passive response, a biquad low-pass filter (LPF) with tunable gain, Q factor and center frequency presents the active and selective response, and an elliptic filter which has sharpest transition among all filter types presents the steep roll-off. To match the steepness of roll-off slope, the elliptic filter is designed as 5th-order and the entire cochlea filter is a 9th-order system.

III. CIRCUIT IMPLEMENTATION

A. Basic Cell: Floating Active Inductor (FAI)

Among the three filters proposed to build a cochlea filter, it is the high order elliptic filter that brings most design difficulty and challenge. For audio frequencies, implementation of passive LC-ladder topology in VLSI is unrealistic due to unfeasibly large size of passive inductors. Active RC [49] and switch-capacitor implementations [50] require the same number of op-amps with the filter order, and are thus constrained by power consumption. Reported log-domain implementations mostly use bipolar transistors targeting at high frequency applications [51], [52], while the CMOS implementations are either low-order [53], [54] or not proven with chip results [55]–[57].

Therefore, we developed a floating active inductor (FAI) as the basic cell to build active LC ladders, which is shown in Fig. 3. The FAI is inspired by several existing active inductor designs in [58], and has been modified so that it operates in floating mode as required by LC ladders. Also, the circuits are designed to operate in weak inversion so as to achieve the long
The FAI is based on gyrator-C topology. The transistor pair MN and MP respectively provide the forward and reverse transconductance to form a gyrator, and $C_{gy}$ is the capacitor loaded at the gyrating ports. The transistor pair MX functions as a compensation transconductor which reduces the resistive loading that the sources of the MP transistors add on the gyrating ports. The quality factor of the FAI is tunable by adjusting the transconductance difference between the MP pair and MX pair.

Deriving from [59], the equivalent impedance of the FAI is given by

$$Z(s) = \frac{4C_{gy} \cdot s}{g_{MP} g_{MN}} + \frac{2(g_{MP} - g_{MX})}{g_{MP} g_{MN}}$$  \hspace{0.5cm} (1)

where $g_{MP}$, $g_{MN}$ and $g_{MX}$ are the transconductance of the transistor pairs. Substrates of MN and MP and MX pairs are all connected to power supplies, and as sources of MX0-MX1 and MN0-MN1 pairs are respectively tied together, the body effect does not have much influence on their differential transconductance. However, MP0 and MP1 have separate sources and thus the body effect transconductance $g_{mb}$ should be considered. The transistors operate in weak inversion and thus $g_{mb}$ approximately equals $n - 1/n \cdot I_D/U_T$ [60] where $n$ is slope factor, $I_D$ is biasing drain current and $U_T$ is thermal voltage. Besides, the currents in MN and MX are made equal as shown in Fig. 3. Thus, the equivalent inductance and resistance can be written as

$$L = \frac{4C_{gy} n U_T^2}{I_{MP} I_{MX}}$$  \hspace{0.5cm} (2)

$$R = \frac{2 U_T (n I_{MP} - I_{MX})}{I_{MP} I_{MX}}$$  \hspace{0.5cm} (3)

Equations (2) and (3) show that the FAI inductance and resistance can be tuned by adjusting $I_{MP}$ and $I_{MX}$.

The drain-source conductance ($g_{ds}$) due to channel length modulation is not considered. Because in weak inversion $g_{ds} = I_D/\lambda L$ will be at least one hundred times smaller than the transistor transconductance $g_{mn} = I_D/n U_T$ according to the process parameters ($\lambda$ is channel length modulation parameter and $L$ is transistor channel length).

B. Triple-Stage Cochlea Filter Design Based on FAI

Based on the FAI cell, 2nd order BPF, 2nd order LPF, 5th order elliptic filters and their cascaded cochlea filter channel are built as shown in Fig. 4. The BPF is built by loading a fully differential OTA (FDOTA: Fig. 5) with two FAIs and a capacitor, and the LPF is built upon FAI-C voltage divider with a fully differential difference amplifier (FDDA: Fig. 6) as input buffer, while the elliptic filter is built according to filter design handbook [61] with single-end OTAs (OTA: Fig. 7) providing equivalent source resistance. There are a total of eight FAIs in each channel, and their $I_{MP}$ currents are made equal through current mirrors, while $I_{MX}$ currents are set separately. We denote the $I_{MP}$ currents as $I_{SAT}$, and $I_{MX}$ as $I_{BPF}$, $I_{LPF}$ and $I_{ELL}$ respectively. The FDOTA and the single-end OTA also operate in weak inversion and are biased with $2I_{SAT}$ currents. The FAIs in BPF and LPF are all loaded with equal capacitance of $C_D$, while $C_{BPF}$ is set as 1.2 $C_D$ and $C_{LPF}$ is set as $C_D$.

Thereby the transfer functions of BPF and LPF are derived as

$$H_{BPF}(s) = \frac{5 I_{SAT} C_0}{6 n U_T C_0} \cdot \frac{s + \omega_{0, BPF}}{s^2 + \frac{\omega_{0, BPF}^2}{Q_{BPF}} s + \omega_{0, BPF}^2}$$  \hspace{0.5cm} (4)

$$\omega_{0, BPF} = \frac{n I_{SAT} I_{BPF}}{2 I_{BPF} C_0 \sqrt{\frac{5}{6}} I_{SAT} I_{BPF}}$$  \hspace{0.5cm} (5)

$$H_{LPF}(s) = \frac{\omega_{0, LPF}^2}{s^2 + \frac{\omega_{0, LPF}^2}{Q_{LPF}} s + \omega_{0, LPF}^2}$$  \hspace{0.5cm} (6)

The transfer function of the elliptic filter can be derived by obtaining poles and zeros from filter design tables [61]. In our case, the transfer function of the elliptic filter is given by

$$H_{ELL}(s) = \frac{0.069581 \omega_{3 dB} \cdot (s^2 + 7.3381 \omega^2_{3 dB}) (s^2 + 3.1407 \omega^2_{3 dB})}{(s + 0.96232 \omega_{3 dB}) (s^2 + 1.2360 \omega_{3 dB} s + 1.1724 \omega^2_{3 dB}) (s^2 + 0.34350 \omega_{3 dB} s + 1.4269 \omega^2_{3 dB})}$$  \hspace{0.5cm} (7)

The transfer function of the elliptic filter can be derived by obtaining poles and zeros from filter design tables [61]. In our
design, the elliptic filter is 5th order, with the reflection coefficient $\rho = 5\%$, the modular angle $\theta = 36^\circ$ and power loss factor $K^2 = \infty$, which in theory can achieve pass-band ripple of 0.01 dB, steepness factor of 1.7013 and minimum stop-band attenuation of 40.81 dB (equivalent to cut-off slope of 176.8 dB/dec). The transfer function of the elliptic filter is given in (8) at the bottom of the preceding page, where $\omega_{-3 \text{ dB}}$ is the $-3$ dB corner frequency.

We can also obtain the parameters of filter elements by referring to the filter design tables in [61]. By setting $\omega_{-3 \text{ dB}}$ as $\sqrt{2}$ times of $\omega_{0,BPF}$ (center frequency of BPF), the values of required inductance and capacitance can be derived as following:

\[
\begin{align*}
L_{FAI2} &= 0.8551 \left( \frac{nU_T}{I_{STAT}} \right)^2 \sqrt{\frac{I_{STAT}}{n_{BPF}}} C_0 \\
L_{FAI4} &= 1.113 \left( \frac{nU_T}{I_{STAT}} \right)^2 \sqrt{\frac{I_{STAT}}{n_{BPF}}} C_0
\end{align*}
\]  

Combining (2) and (9), the values of $C_{gy}$ in FAI2 and FAI4 can be derived

\[
\begin{align*}
C_1 &= 0.5252 \sqrt{\frac{I_{ELL1}}{n_{BPF}}} C_0 \\
C_2 &= 0.7651 \sqrt{\frac{I_{STAT}}{n_{BPF}}} C_0 \\
C_3 &= 2.296 \sqrt{\frac{I_{STAT}}{n_{BPF}}} C_0 \\
C_4 &= 1.373 \sqrt{\frac{I_{STAT}}{n_{BPF}}} C_0 \\
C_5 &= 2.293 \sqrt{\frac{I_{STAT}}{n_{BPF}}} C_0
\end{align*}
\]  

In the cochlea filter channel, only $I_{LPF}$ is actively tuned, while $I_{BPF}$ and $I_{ELL1}$ kept constant. Based on exhaustive simulations with extracted parameters in software, we choose to make $I_{BPF}$ equals $I_{STAT}$ so that the BPF has a relatively low Q factor and gentle response, and $I_{ELL1}$ equals $1.2 I_{STAT}$ so that the FAIs in the elliptic filter are inductive enough to maintain the steep roll-off while enough margin is left to avoid negative-damping and instability. In addition, we use $x = I_{LPF}/I_{STAT}$ to denote the tuning factor of the LPF and also the entire cochlea filter, which can be dynamically controlled to mimic the active response of the biological cochlea. The transfer functions can be greatly simplified if we use $\omega_{pc}$ (the passive center frequency of the cochlea filter) to replace $\omega_{0,BPF} = 1/4nU_T C_0 \sqrt{5/3nI_{STAT}I_{BPF}}$. Combining
(4), (6) and (8) with the above mentioned design settings \( \omega_{3dB} = \sqrt{2} \omega_{BPF}, I_{BPF} = I_{STAT}, x = I_{LFF}/I_{STAT}, I_{ELL1} = 1.2 I_{STAT} \), the transfer function of the cascaded cochlea filter channel can be derived as (12) at the bottom of the page. As the slope factor \( n \) in AMS 0.35 \( \mu \)m process is approximately 1.25, (12) can be further simplified into (13) at the bottom of the page. The proposed filter has nine poles and five zeros and its normalized zero-pole plot is shown in Fig. 8. As there are no zeros in the right half of the s-plane, the cochlea filter is a minimum-phase filter, as is the biological cochlea [63].

IV. ANALYSIS OF NON-IDEALITIES

A. Mismatch

Naturally, the operation of the FAI is subject to transistor mismatches which cause circuit offsets. The mismatches can be classified into two categories: horizontal and vertical. The horizontal mismatch is the mismatch between transistors in differential pairs MP0-MP1, MN0-MN1 and MX0-MX1 which results in a DC current in the FAI running from port TA to TB. Taking the differential pair MN as an example, shown in Fig. 9(a), the mismatch between MN0 and MN1 will cause deviation in their transconductance, which results in a DC current running from port TA to port TB even if TA and TB are tied to the same DC voltage \( V_d \). The value and variance of the offset DC current can be derived as

\[
I_{oa} = \Delta gm(V_g - V_a) \approx \Delta gmU_r I_D I_a
\]

where \( I_0 \) is the transistor off current when \( V_g = V_a = 0 \) and is expressed as

\[
H_{tot}(s) = (2.31\omega_{pc} - \frac{s + 1.55n/\sqrt{n}}{s^2 + 1.55\frac{n}{\sqrt{n}}\omega_{pc}s + \omega_{pc}^2})
\]

\[
\cdot 1.2\omega_{pc}^2
\]

\[
\cdot \left[ \frac{0.0988 \omega_{pc}}{s^2 + 1.36\omega_{pc}} (s^2 + 14.7\omega_{pc}^2) (s^2 + 6.28\omega_{pc}^2) \right] (s^2 + 1.75\omega_{pc}s + 2.35\omega_{pc}^2) (s^2 + 0.486\omega_{pc}s + 2.85\omega_{pc}^2)
\]

\[
H_{tot}(s)
= \frac{0.274\omega_{pc}}{(s^2 + 0.347\omega_{pc}s + \omega_{pc}^2)(s^2 + 13.26\omega_{pc}s + 23.5\omega_{pc}^2)(s^2 + 0.486\omega_{pc}s + 2.85\omega_{pc}^2)}
\]

\[
\cdot \frac{1}{(s^2 + 0.347\omega_{pc}s + \omega_{pc}^2)(s^2 + 13.26\omega_{pc}s + 23.5\omega_{pc}^2)(s^2 + 0.486\omega_{pc}s + 2.85\omega_{pc}^2)}
\]
Fig. 9. Mismatch analysis of the FAI. (a) Offset current flowing between \( T_A \) and \( T_B \) results from horizontal mismatch between transistors in the differential pair. [Horizontal mismatch.] (b) Offsets currents separately flowing at \( T_A \) and \( T_B \) result from vertical mismatch between the current sources. [Vertical mismatch.]

\( I_m = I_{\text{spec}} \cdot e^{-\frac{V_{th}}{nU_T}}. \quad (15) \)

\( V_{th} \) is the transistor threshold voltage and \( I_{\text{spec}} \) is the specific current defined as \( 2\beta nU_T^2 \) where \( \beta \) is the transconductance parameter. The ratio between \( I_D \) and \( I_{\text{spec}} \) is the inversion coefficient which is far less than unity in weak inversion [62].

The same analysis can be performed on transistor pair MP and MX, and the total offset currents resulting from horizontal mismatch can be expressed in the following as summation of three transistor pair mismatches:

\[
I_{\text{osH}} = \Delta g_{MN} U_T \ln \frac{I_{MX}}{I_{MN}} + \Delta g_{MN} U_T \ln \frac{I_{MX}}{I_{PM}} \quad (16)
\]

The vertical mismatch on the other hand refers to the mismatch between the current sources in the upper side and current sinks in the lower side. As shown in Fig. 9(b), the difference between the currents in the upper and lower side will flow out of the FAI through \( TA \) and \( TB \). These two offset currents can be written as

\[
I_{\text{osV}(A)} = 2\Delta I_{MFA(A)} + 2\Delta I_{MX}. \quad (17)
\]

B. Noise

The noise model of the FAI is illustrated in Fig. 10, from which noise current density at the port \( TA \) and \( TB \) can be derived

\[
\tilde{I}_n^2(\omega) = \tilde{I}_{nM}^2 \left[ \frac{C_{gs}^2}{C_{gs}} + \frac{g_{MN}^2 \varepsilon_{MN}^2 + g_{MN}^2 \varepsilon_{MN}^2}{(g_{MX} - g_{MP})^2 + 4\omega^2 C_{gs}^2} \right]. \quad (18)
\]

Note that the noise model in Fig. 10 does not include the source degeneration that \( C_{gs} \) and \( MX \) have on the MP transistor. The actual noise transconductance for \( \varepsilon_{MN}^2 \) is smaller than \( g_{MP} \) and thus (18) is in fact the worst-case noise estimation. According to the noise parameters given by the foundry, the transistor noise corner frequency is derived as \( \omega_{\text{corner}} = 3\pi K_F A_F^{-1}/4qC_{gs}WL \) where \( K_F \) and \( A_F \) are flicker noise parameters, \( q \) is electron charge and \( C_{gs} \) is gate oxide capacitance. For the cochlea filter, the biasing current \( I_D \) is mapped with the passive center frequency \( \omega_{pc} \), and thus we find the point where \( \omega_{pc} \) equals \( \omega_{\text{corner}} \) as follows:

\[
\omega_{pc0} = \frac{3\pi K_F (3.46U_T C_{gs}) A_F^{-1}}{4qC_{gs}WL} \quad (19)
\]

Equation (19) is suitable for PMOS transistor, while for an NMOS transistor \( WL \) should be replaced with \( L^2 \) according to the noise model provided by the foundry. Besides, parameter \( A_F \) is between 1 and 2, and thus the \( \omega_{pc} \) will be higher than \( \omega_{\text{corner}} \) if it exceeds \( \omega_{pc0} \). Therefore a set of values for \( WL \) and \( L^2 \) which makes \( \omega_{pc0} \) lower than 20 Hz can be derived so that the passive center frequency of the cochlea filter is always higher than the noise corner frequency and circuit noise is dominated by thermal noise. This setting not only simplifies the following noise calculation but also contributes to better noise performance as the cochlea filter has peak gain at a frequency equal or higher than \( \omega_{pc} \). Calculation with the foundry parameters indicates the transistor dimension should meet the requirement that \( WL > 16 \sim 17 \mu m^2 \) for PMOS transistors and \( L^2 > 9 \sim 10 \mu m^2 \) for NMOS transistors.

Considering only thermal noise, (18) is rewritten and simplified with the cochlear filter parameters as follows:

\[
\tilde{I}_n^2(\omega) = \frac{8kT I_{STT} A_F^{-1} 2x}{3U_T} \left[ 1 + \frac{2x}{(x - 1.25)^2 + 0.521(\frac{\omega}{\omega_{pc}})^2} \right]. \quad (20)
\]

C. Comprehensive Analysis

Based on the analysis above, a more complete FAI model with mismatch and noise considerations is derived and illustrated in Fig. 11. The DC offset currents affect the DC operation point of circuits and thus as shown in Fig. 4, at least one port of each FAI is connected to low impedance source to release the offset currents. In the BPF and LPF, all FAIs have one
port shorted to ground (BPF) or FDDA buffer output (LPF), releasing $I_{osh}$ and one branch of $I_{osV}$. The remaining branch of $I_{osV}$ flows through the FAI and causes a DC shift of $I_{osV} \cdot R_{FAI}$ which is in the worst case less than 1 mV if the variances of parameters do not exceed 1%. FAIs in the elliptic filter should be analyzed separately. The FAI2s have one port connected to the OTA followers while the FAI4s have one port connected to the other port of FAI2s. Thus $I_{osh}$ and $I_{osV}$ of both FAI2 and FAI4 flow through the OTA follower and cause DC shift of $(I_{osh} + 2I_{osV})nUT/ISTAT$ which is in the worst case less than 40 mV if the variances of parameters do not exceed 1%. Besides, the offset currents in FAI4 flow through FAI2 in addition to one branch of $I_{osV}$ of FAI2 itself and cause DC shift of $I_{osV} \cdot R_{FAI}$ which is in the worst case less than 3 mV if the variances of parameters do not exceed 1%. In the end, one branch of $I_{osV}$ flows through FAI4 and causes DC shift of $I_{osV} \cdot R_{FAI}$ which is less than 0.1 mV if the variances of parameters do not exceed 1%. Simulations prove the above-mentioned level of DC shift has insignificant effects on the circuit operation. Thus by referring to the parameter matching equations provided by the foundry, the width and length of the transistors in the FAIs are optimized so that the probability of the parameter variance being greater than 1% is limited to 1%.

Taking into consideration the parameter matching, the noise corner frequency setting explained in the previous section and also the weak inversion requirement, the dimensions of the transistor pairs in FAI are set to $W = 180 \mu m$ and $L = 3.5 \mu m$ for PMOS pairs MP and MX while $W = 100 \mu m$ and $L = 3.5 \mu m$ for NMOS pair MN.

For the noise analysis, the input referred noise of the BPF and LPF in the cochlea filter can be derived as following:

\[
V_{n_{i,BPF}}^2(\omega) = 0.0651 \cdot \frac{\omega^2_{PC} C_0}{\omega_{PC}^2 C_0^2} \tag{21}
\]
\[
V_{n_{i,LPF}}^2(\omega) = \frac{2^2 \omega^2_{LPF}(\omega)}{x^2 \omega_{PC}^2 C_0^2} \left( \frac{0.17 \omega^2}{\omega_{PC}^2} + 0.333(1.25 - x)^2 \right). \tag{22}
\]

Noise calculation for the elliptic filter is far more complicated, but as the BPF and LPF provide all the gain for the cochlea filter, the noise from the elliptic filter is less significant compared with BPF and LPF when referred to the input. Therefore, neglecting the noise from elliptic filter, input referred noise density of the cochlea filter channel is the sum of BPF input referred noise and LPF input referred noise divided by gain of BPF and can be written as (23), at the bottom of the page, combining (4), (20), (21) and (22) and using $\xi$ to represent $\omega/\omega_{PC}$. As explained above, cochlea filter bandwidth has been set as $\sqrt{2}\omega_{PC}$, and thus integrated input referred noise of the cochlea filter is calculated as follows:

\[
V_{n_{i,noise}}^2 = \int_0^{\sqrt{2}} \frac{V_{n_{i,LPF}}^2(\xi)}{\xi^2 C_0^2} d\xi = \frac{440kT}{\omega_{PC} C_0} \left( \frac{1}{x} - \frac{1}{6} \right)^2 + 0.65]. \tag{24}
\]

Equations (24) indicates that the filter input-referred noise decreases with increasing center frequency and tuning factor. A cochlea filter with passive center frequency of 100 Hz has $47 \mu V_{rms}$ input noise in the low Q mode ($x = \frac{0.75}{6}$) and $27 \mu V_{rms}$ in the high Q mode ($x = 1.24$).

V. RESULTS

Based on the design explained above, a cochlea filter channel has been fabricated using AMS $0.35 \mu m$ 2-poly 4-metal process, as shown in Fig. 12. A prototype PCB and an NI PXI platform are built to characterize the filter, as illustrated in Fig. 13.

The static current of the cochlea filter $I_{STAT}$ is set with different values from 37.14 pA to 9.915 nA so as to make the filter operate in nine frequency regions corresponding to the octave audio bands from 31 Hz to 8 kHz. The BPF current $I_{BPF}$ is set where $\omega_{PC}$ is approximately $10 \sim 20\%$ smaller than the corresponding octave frequency value. The LPF current $I_{LPF}$ is tuned in the region from where the LPF center frequency overlies with $\omega_{PC}$ (LQ mode), to where the LPF peak gain is maximized (HQ mode). The elliptic filter current $I_{ELLI}$ is set as $1.2I_{STAT}$ as explained in Section III.

\[
V_{n_{i,LPF}}^2(\xi) = \frac{1.15kT}{\omega_{PC} C_0} \left\{ \frac{(1.25 - x)^2 + 2x + 0.521\xi^2[(1.86\xi^2 + 1) - 1]}{x^2(\xi^2 + 0.14)} + 6.521 + \frac{2}{0.12 + \xi} \right\}. \tag{23}
\]
A. Frequency Response

1) Magnitude Response: Frequency responses of the nine filter bands in magnitude are measured and plotted in Fig. 14. Apart from the LQ and HQ modes mentioned above, responses of the cochlea filters in medium Q (MQ) mode are also measured, where $I_{TF}$ is adjusted so that center frequency (CF) of the entire filter is located approximately at the corresponding octave frequency value. As shown in Fig. 14, the passive and gentle low-frequency band, active and selective mid-band and steep roll-off are achieved in all of the filters. Although a 40dB of peak gain variation range has been measured from chinchilla cochlea [16], [18], more physiological measurements in recent years report approximately 20 ~30 dB of gain variation [17], [19]–[21], [23]. The high-frequency amplitude plateau in biological cochlea [28] is also found in the 31 Hz, 63 Hz, 125 Hz, 500 Hz, 1 kHz and 2 kHz bands. Besides, it is observed that CFs of LQ mode locate approximately 10 ~20% leftwards from the CFs of MQ, while CFs of HQ mode locate 10 ~20% rightwards. In other words, the CFs become higher together with increasing peak gain and selectivity, which agrees with biological cochlea behavior [16]–[21]. The detailed results are listed in Table I. Noticing that as parasitic resistance of wires exists in real VLSI implementation, the range of tuning factor $x$ in Table I is wider than the value used in previous section (0.75 ~1.24). To improve the precision of current measurement, current mirrors with 1:100 ratio are used on chip. Therefore, the actually measured off-chip currents are 100 times as much as those values in Table I.

2) Phase Response: As both biological cochlea and the proposed filter have minimum-phase property, their phase responses should also be similar when the magnitude responses are matched. The phase responses of the 31 Hz and 8 kHz bands are illustrated in Fig. 15. As also observed in the physiological results, the filter phase lag increases with frequency in the pass-band [16]–[19], [21], [23]. The LQ phase lag at the passive centre frequency is slightly over half cycle [17], [18], [20], [21], while the HQ phase lag at the active centre frequency is approximately one cycle [17], [21]–[23]. The LQ response has more phase lag at frequencies lower than $\omega_{ac}$ (active center frequency) while the HQ response has more phase lag at frequencies higher than $\omega_{ac}$ [17]–[21].

3) Group Delay: Fig. 15 indicates that the group delay reaches maximum at $\omega_{ac}$ where the phase response curve has steepest slope. Fig. 16 shows the maximum group delay in unit of periods across different center frequencies. As expected, the group delay increases from LQ to HQ mode. The maximum group delays in LQ and MQ modes are approximately 5 periods while the maximum group delay in HQ mode is in the region of 10 periods. Fig. 17 shows the physiologically measured maximum group delay in human cochlea [64]. The measurement in [64] is based on the stimulus-frequency emission method which stimulates the cochlea with low-intensity input. Therefore the results in [64] correspond to the HQ response in this work. The comparison in Fig. 17 shows the cochlea filter has similar order of group delay with human cochlea.

B. Time Domain Response

Impulse responses of the cochlea filter in 31 Hz and 8 kHz bands are given in Fig. 18, combined with response from BPF, LPF and elliptic filter separately. As the cochlea filter is a composite of three filters in cascade, its overall impulse response is the convolution of three individual responses. From Fig. 18, it is observed that for HQ mode, envelopes of response are not smooth and there is a trough near the third ringing crests, while for LQ mode, however, the envelopes are fairly smooth. The reason for this phenomenon is that, as shown in the decomposed response plots, BPF and elliptic filters settle much quicker than LPF in HQ mode, and thus although their responses are significant enough to affect the convolved overall response in the early stage, after 100 ms in Fig. 18(a) and 0.4 ms in Fig. 18(b), the overall responses are fully dominated by LPF. Therefore the impulse responses appear to have two stages of behavior, the convoluted response and the LPF-dominant response, separated by the settling of BPF and elliptic filter. As for the LQ mode, the LPF settles even faster than the BPF and elliptic filter, and thus the overall responses are smooth over time.

The post-dominance of LPF in HQ mode results in a shift of ringing frequency. As shown in Fig. 18, initial ringing periods in convolved responses ($t_{dr}$) are wider than ringing periods in the LPF-dominant responses ($t_{dr}$). This effect agrees with the frequency gliding phenomenon observed in physiological measurements, where instantaneous frequency of biological cochlea response to clicks is not constant but increases over time until settled at steady state [24]–[27]. The gliding phenomenon is a standard for cochlea model evaluation suggested by the physiologists [24], [26]. Its origin has been proved to be independent of the nonlinear active process [24], [26]. Similarly, the frequency shift in this cochlea filter is not based any active control and thus provide a basis for future research on how the gliding effects influence signal processing in the cochlea.

C. Noise Measurement

Output noise spectrum from the 31 Hz and 8 kHz bands, with comparison between LQ, MQ and HQ modes are illustrated in Fig. 19. The 50 Hz harmonics shown are due to the ripples of
TABLE I
FREQUENCY RESPONSE SPECIFICATIONS OF COCHLEA FILTER IN DIFFERENT CONFIGURATIONS

| CF    | CF Variation | GD  | ERB  | $I_{STAT}$ | $x$     | PG  | PG Variation | Q factor | Roll-off Slope |
|-------|--------------|-----|------|------------|---------|-----|--------------|----------|----------------|
| Filter 1 |             |     |      |            |         |     |              |          |                |
| LQ    | 25 Hz       | -19.3% ~+12.9% | 12.2 ms | 4.3 Hz     | 37.14 mA | 0.68 | 7.73 dB      | 0.08     | 2.08           |
| MQ    | 31 Hz       |              | 12.6 ms | 4.8 Hz     | 3.07 dB  | 1.07 | 12.48 dB     | 1.39     | 2.63 dB        |
| HQ    | 35 Hz       |              | 3.9 ms  | 2.97 Hz    | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| Filter 2 |             |     |      |            |         |     |              |          |                |
| LQ    | 53 Hz       | -15.9% ~+17.5% | 53.9 ms | 6.5 Hz     | 78.41 mA | 0.95 | 14.26 dB     | 1.45     | 29.14 dB       |
| MQ    | 63 Hz       |              | 63.5 ms | 7.5 Hz     | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| HQ    | 74 Hz       |              | 168 ms  | 7.8 Hz     | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| Filter 3 |             |     |      |            |         |     |              |          |                |
| LQ    | 105 Hz      | -16% ~+12% | 32.8 ms | 4.3 Hz     | 143.7 mA | 0.82 | 9.97 dB      | 1.45     | 29.14 dB       |
| MQ    | 125 Hz      |              | 35.1 ms | 4.8 Hz     | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| HQ    | 140 Hz      |              | 73.6 ms | 2.97 Hz    | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| Filter 4 |             |     |      |            |         |     |              |          |                |
| LQ    | 220 Hz      | -12% ~+10% | 16.8 ms | 6.5 Hz     | 279.1 mA | 0.92 | 11.52 dB     | 1.23     | 17.72 dB       |
| MQ    | 25 Hz       |              | 16.8 ms | 6.5 Hz     | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| HQ    | 275 Hz      |              | 33.1 ms | 7.5 Hz     | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| Filter 5 |             |     |      |            |         |     |              |          |                |
| LQ    | 430 Hz      | -12.2% ~+11.2% | 10.6 ms | 4.3 Hz     | 554.0 mA | 0.88 | 13.14 dB     | 1.26     | 21.32 dB       |
| MQ    | 490 Hz      |              | 11.4 ms | 4.8 Hz     | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| HQ    | 545 Hz      |              | 17.4 ms | 2.97 Hz    | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| Filter 6 |             |     |      |            |         |     |              |          |                |
| LQ    | 0.9 Hz      | -14.3% ~+14.3% | 4.03 ms | 386 Hz     | 1.189 mA | 0.89 | 10.79 dB     | 1.24     | 17.74 dB       |
| MQ    | 1.05 Hz     |              | 4.82 ms | 386 Hz     | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| HQ    | 1.2 Hz      |              | 7.65 ms | 101 Hz     | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| Filter 7 |             |     |      |            |         |     |              |          |                |
| LQ    | 1.7 Hz      | -15% ~+10.8% | 1.56 ms | 6.86 Hz    | 2.156 mA | 1.31 | 21.38 dB     | 1.54     | 32.22 dB       |
| MQ    | 2.0 Hz      |              | 1.74 ms | 6.02 Hz    | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| HQ    | 2.8 Hz      |              | 4.04 ms | 0.238 Hz   | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| Filter 8 |             |     |      |            |         |     |              |          |                |
| LQ    | 3.5 Hz      | -10.3% ~+12.8% | 70.6 µs | 1.69 Hz    | 4.291 mA | 0.91 | 11.28 dB     | 1.29     | 18.95 dB       |
| MQ    | 3.9 Hz      |              | 950 µs  | 1.33 kHz   | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| HQ    | 4.4 Hz      |              | 2.04 ms | 0.375 kHz  | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| Filter 9 |             |     |      |            |         |     |              |          |                |
| LQ    | 7.1 Hz      | -14.5% ~+10.8% | 223 µs  | 4.01 kHz   | 9.915 mA | 0.68 | 8.152 dB     | 1.20     | 19.03 dB       |
| MQ    | 8.3 Hz      |              | 369 µs  | 2.94 kHz   | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |
| HQ    | 9.2 Hz      |              | 1.24 ms | 1.28 kHz   | 3.07 dB  | 1.39 | 2.63 dB      | 1.39     | 2.63 dB        |

*CF = Center Frequency; GD = Group Delay; ERB = Equivalent Rectangular Bandwidth; PG = Peak Gain.

D. Distortion Measurement

1) Total Harmonic Distortion (THD) and Signal-to-Noise-and Distortion Ratio (SINAD): Fig. 21 shows the plots of THD and SINAD against input level based on the measured results from the 31 Hz and 8 kHz bands. The SINAD of both filters generally maintain above the 12 dB SINAD threshold for intelligent hearing before THD reaches the edge of 5% limit. The HQ mode has most significant harmonic distortion due to high LPF gain and thus high signal amplitude at the elliptic filter input. As predicted in the noise spectrum [Fig. 19(b)], the 8 kHz band has more harmonic distortion than the 31 Hz band. Based on the 5% THD limit, the maximum input level is plotted against filter center frequencies, so is the maximum SINAD. Fig. 22 shows the filter linearity tends to degrade with higher frequencies.

2) Two-Tone Inter-Modulation Distortion: Inter-modulation distortion test is performed and the results are shown in Fig. 23. The third-order inter-modulation product $2f_1 - f_2$ appears to be the most prominent distortion component because it is designed to coincide with the filter center frequency. The inter-modulation distortion is also found in the biological cochlea, which proves that the biological hearing system can tolerate worst-case spurious free dynamic range (SFDR) [28]. Fig. 24 plots the filter maximum input range measured using the 17 dB SFDR limit. It shows in most frequency bands the maximum input range is further reduced compared with the results based on the 5% THD limit. Nevertheless, the input range of the 8 kHz band appears even higher than the results in Fig. 22 and there is no significant degradation compared with the other power supply. Besides, the noise spectrum of the HQ mode of 8 kHz band has prominent peaks at CF harmonics. This indicates that the 8 kHz band has more harmonic distortion than the 31 Hz band especially in the HQ mode. As shown in Table I, the DC current in FAI scales with CF. Therefore, the operation of transistors in the FAI moves from weak inversion towards moderate inversion when the center frequency increases. However, the DC operating points of the filter are designed based on the weak inversion assumption, and the circuit linearity will be affected by the DC variation. The CF harmonics should be theoretically well attenuated by the elliptic filter. However, the transistor in moderate inversion has lower $gm/|V_{DS}|$ compared with weak inversion which makes the FAI inductance deviate from designed values and consequently degrade the elliptic filter performance. The harmonic distortion issue will be discussed further in Section V-D.

The input-referred noise density at the center frequency of the 9 filter bands in LQ, MQ and HQ modes are illustrated in Fig. 20. Compared with noise predication function of (24), the measured results in MQ and HQ modes agree in terms of the fact that input referred noise decreases for higher center frequencies. However, LQ mode shows noise does not vary much from low to high frequencies, because the filter selectivity is not high enough to overcome the added power supply harmonics from the increase of filter bandwidth. Besides, Fig. 20 also proves the increase of tuning factor $x$ results in lower input referred noise except the 31 Hz band where MQ has even wider equivalent rectangular bandwidth than LQ mode as shown in Table I.
Fig. 14. Measured frequency response of 9 cochlea filters covering octave audio bands from 31 Hz to 8 kHz. The LQ, MQ and HQ modes respectively correspond to the biological response with high, normal and low intensity sound stimulus. Tuning of the filter from LQ to HQ is achieved by adjusting only one circuit parameter ($I_{LR}$).

Fig. 15. Measured phase response of the 31 Hz band and 8 kHz band.

Fig. 16. Measured maximum group delay across different center frequencies.

Fig. 17. The HQ maximum group delay curve in comparison with physiologically measured results from human cochlea. The figure is adapted from the Fig. 5 in [64].

The distortion products of interest in the inter-modulation measurement are in-band signals while the CF harmonics measured in the THD test are out-of-band signals. Therefore, the high harmonic distortion measured in the 8 kHz band is probably due to the degraded stop-band attenuation.

E. Critical Bandwidth

An equivalent rectangular band-pass filter model is illustrated in Fig. 25, which helps us to understand the critical band and frequency discrimination feature of the designed cochlea filters. The calculated equivalent rectangular bandwidth (ERB) and measured $-3$ dB bandwidth of the cochlea filters are plotted versus corresponding CF in Fig. 26, together with the approximated bandwidth of human auditory filters derived from the formula given by Glasberg and Moore [65] for comparison. It shows that $-3$ dB bandwidth is generally narrower than the ERB, but their discrepancy is not significant. The exact bandwidth values are listed in Table I. We observe from Fig. 26 that the ERB curve given by Glasberg and Moore from psychoacoustical research lies in-between the ERB curves of the MQ and HQ modes. However as the cochlea filter can be continuously tuned, it is possible to find a condition between the MQ and HQ modes where the ERB versus CF curve corresponds with better agreement to psycho-acoustical results [65].

F. Testing With Acoustic Signals

A segment of acoustic signal (mixed sounds from musical instruments of the horn and bass drum) is applied to the cochlea filter. As the frequencies around 63 Hz, 1 kHz and 2 kHz have highest intensity, the cochlea filter is tested in these three frequency bands accordingly and the results are shown in Fig. 27. It is notable that, the noise around center frequency is selectively amplified rather than the signal in the HQ output of the 1 kHz band, as the cochlea filter currently does not have the capability to distinguish between signal and noise. This problem can be potentially solved with the addition of an SNR estimation mechanism [66]. Also, the results show that the octave distribution bands.
Fig. 18. Measured impulse response from (a) 31 Hz and (b) 8 kHz band cochlea filters. The decrease of ringing period \( t_{eq} \) with time agrees with the gliding phenomenon in biological cochlea.

Fig. 19. Measured output noise spectrum from (a) 31 Hz and (b) 8 kHz band filters. Like the frequency response shown in Fig. 14, the spectrum shape has stable low-frequency band (does not vary with tuning), tunable mid-frequency band and steep roll-off at stop-band.

Fig. 20. Plot of measured integrated input-referred noise versus center frequency.

The decrease of ringing period \( t_{eq} \) with time agrees with the gliding phenomenon in biological cochlea.

G. Summary

Table II summarizes the measured specifications of the cochlea filter chip. Note that the power dissipation of the FDDA and the output buffers does not scale with filter center frequency. Consequently, the power dissipation of the 31 Hz filter is only one-third less than that of the 8 kHz filter. Table III gives scores in terms of auditory filter model following the criteria given by Lyon [67]. Noticing that the current cochlea filter has not been integrated with automatic-Q-control (AQC) mechanism, but as the measured results indicate the filter can be actively tuned according to sound level, a potential ‘+’ credit is given in the ‘dynamic’ criterion.

VI. CONCLUSION

Design and experimental results of a bio-realistic analog cochlea filter have been presented, the highlights of which can be summarized as following:
- The filter is highly faithful with measured response from physiological experiment on mammalian cochlea, with
Fig. 22. Measured maximum input range and SINAD across different center frequencies.

Fig. 23. Two-tone inter-modulation distortion measured from (a) 31 Hz and (b) 8 kHz band filters. Two signals in equal amplitude (10 mV) with primary frequencies and such that are applied to the cochlea filter.

Fig. 24. Maximum input range measured across different center frequencies using the 17dB SFDR limit.

Fig. 25. Equivalent rectangular band-pass filter model of the cochlea filters. The rectangular filters pass the equal amount of energy with the corresponding cochlea filters in Fig. 14.

Fig. 26. ERB and –3 dB bandwidth of the cochlea filters in comparison with approximated ERB of human auditory filters [65].

passive and gentle response in low-frequency band, active and selective response in mid-band and a sharp transition from pass-band into stop-band. Besides, similarity of the filter in phase response and impulse response has also been demonstrated.

- The filter can operate at center frequencies from as low as 31 Hz to 8 kHz. Experimental results show that the operation in deep low frequency is even more robust than in high frequencies.
- Filter efficiency has been improved by the specialized triple-stage design, to the extent that:
  - The active behavior observed in biology is emulated by tuning only one circuit parameter (tuning factor $x$), and positions of only one pair of poles (poles of LPF) are shifted in tuning. The reduced tuning complexity will prospectively increase the robustness and dynamic performance of the proposed cochlea system in Fig. 1.
  - In previous second-order section based filters, the biorealistic 330 dB/dec roll-off requires the filter order to be at least 16 [68]. The same steepness is achieved in our 9th order design by using a sharp cut-off elliptic filter. The lower filter order leads to less power consumption and smaller chip area for each channel, and thus a larger number of channels can be implemented in a parallel filter bank, which will prospectively increase the functionality of the cochlea system shown in Fig. 1.

On the other hand, additional improvement and future work are still necessary including:

- The filter dynamic range is constrained by the limited linear range of the tanh transconductance in FAIs. Linearisation techniques such as multi-tanh [69] may be investigated in the future.
- Filter array with center frequencies distributed in more applicable manner such as one-third octave or bark scale should be implemented so as to fulfill practical auditory processing tasks.
- Other mechanisms including local control (AQC) and signal-noise distinguishing will be integrated with the filter and ultimately a system illustrated in Fig. 1 will be implemented.

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Fig. 27. Measured time-frequency spectrogram of outputs from cochlea filters in response to mixed signals of the horn and bass drum. The signal is applied together with quantization noise from the 13-bit DAC in NI-6732 analog output. Outputs from the three bands and their combinations prove the frequency selectivity of the cochlea filters. Also, the quantization noise has been attenuated, especially by the filter in HQ mode. (a) Input signals. (b) Output from 63Hz filter. (c) Output from 1 kHz filter. (d) Output from 2 kHz filter. (e) Combination of the three filter outputs. (f) Combination of the three filter outputs.

| TABLE II | SUMMARY OF CHIP SPECIFICATIONS |
|----------|--------------------------------|
| Fabrication process | AMS 0.35 μm 3.3V 2P4M |
| Channel area     | 0.9mm² (0.99mm X 0.91mm)     |
| Center frequency | 31Hz~8kHz                     |
| Power dissipation | 59.3μW @31Hz; 90.0μW @8kHz    |
| Peak gain variation | 18.8dB @31Hz; 21.1dB @8kHz |
| Cut-off slope     | 125.5dB/dec(LO) ~ 336.2dB/dec(HQ) @31Hz; 123.6dB/dec(HQ) @8kHz |
| Phase delay @CF   | 210.5degree(LO) ~ 373.1degree(HQ) @31Hz; 204.8degree(LO) ~ 347.4degree(HQ) @8kHz |
| Min. input noise  | 93.34mVrms @31Hz; 34.32mVrms @8kHz |
| Max. input swing (THD< 5%) | 82.0 mVp-p @31Hz; 31.2 mVp-p @8kHz |
| Max. input swing (SFDR>17dB)) | 60 mVp-p @31Hz; 61 mVp-p @8kHz |
| SINAD           | 26.38dB (HQ) ~ 31.67dB (LO) @31Hz; 17.13dB (HQ) ~ 23.56dB (LO) @8kHz |
| Expected dynamic range with AQC | 49.8dB @31Hz; 50.2dB @8kHz |

The lower bound is determined by the HQ input-referred noise and the upper bound is determined by the LO linear range.

| TABLE III | SCORES AS AUDITORY FILTER MODEL [67] |
|------------|-------------------------------------|
| 1. Simple   | Laplace domain 6. Stable tail +    |
| 2. BW control | + 7. Runnable +                   |
| 3. Peakskirts | + 8. Waves -                      |
| 4. Symmetry | + 9. Impulse resp. -               |
| 5. Gain Variation | + 10. Dynamic + (potentially)     |

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