CMOS First-Order All-Pass Filter With 2-Hz Pole Frequency

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Abstract—A CMOS fully integrated all-pass filter with an extremely low pole frequency of 2 Hz is introduced in this paper. It has 0.08-dB passband ripple and 0.029-mm²Si area. It has 0.38-mW power consumption in strong inversion with ± 0.6 -V power supplies. In subthreshold, it has 0.64- μ W quiescent power and operates with ± 200 -mV dc supplies. Miller multiplication is used to obtain a large equivalent capacitor without excessive Si area. By varying the gain of the Miller amplifier, the pole frequency can be varied from 2 to 48 Hz. Experimental and simulation results of a test chip prototype in 130-nm CMOS technology validate the proposed circuit.

Index Terms—All-pass filter (APF), amplifier, Miller multiplier, operational transconductance amplifier (OTA), voltage follower.

I. INTRODUCTION

ALL-PASS FILTERS (APFs) have wide usage in signal processing [1]. They can be used in group delay (GD)/phase equalizers, frequency-selective filters with a high-quality factor, etc. [2]. It is also an important block of quadrature and multiphase oscillators [3]. As it can provide time delay [4], it can be used in differential modulation and beamforming schemes [2] as well. A first-order APF introduces a -90° phase shift at a frequency f_0 (called pole frequency), whose value is inversely proportional to an RC time constant. Most reported integrated APFs have a pole frequency f_0 in the kilohertz–gigahertz range [2], [5]–[10]. The goal of this paper is to implement a fully integrated APF with very low f_0 in the range of Hz. This low f_0 frequency is necessary for some applications like in rotating disk viscometer [11], sensor-based medical devices that perform parallel processing [12] of the sensed biological signal which has frequencies below 100 Hz [13]. Viscometer can detect blood viscosity [14], which helps to determine many diseases [15] such as diabetes

mellitus, malaria hypertension [16], and ischemic stroke [17]. The rotating disk viscometer needs uniformly rotating magnetic field which can be generated by two stable quadrature sinusoidal waveforms in the range of 0.5–50 Hz. APFs can generate quadrature sinusoidal waveforms [3]. APFs can also be used to build linear phase equalizers which act as a delay line. This is necessary to provide synchronization between different parallel processing blocks of real-time signal processing medical devices such as electroencephalography [12] systems.

The frequency f_0 in the proposed APF is given by an expression of the form $f_0 = 1/(2\pi RC)$. Hence, to obtain a low-frequency f_0 in the Hz range, very large capacitor C and resistor R values are required. However, implementation of large capacitor and resistor values in standard CMOS technology with available conventional resistors and capacitors would take extremely large silicon area. In this paper, large valued resistors are implemented using the leakage resistance of the reverse biased source-bulk and drain-bulk p-n junction within a pMOS transistor operating in the cutoff region. A large valued resistor can be emulated using switched-capacitor topology as well, where a two nonoverlapping clock needs to be generated [18]. A linear differential Miller amplifier is used to emulate large capacitors (~ 100 pF) using Miller effect from a moderately valued physical capacitor C ($\sim$ pF). These large valued resistances and capacitors can provide a very low cutoff frequency. By varying the gain of the Miller amplifier, variable Miller capacitors can be realized, which in turn provide adjustable pole frequencies which can be used to tune f_0 to compensate for manufacturing and temperature variations.

This paper is organized as follows. Section II describes the proposed APF circuit. A Miller multiplier used in the circuit is described in Section III. A comprehensive noise analysis of the proposed circuit is presented in Section IV. Simulation results for operation in strong inversion and subthreshold region are given in Section V. Section VI provides all experimental results. Finally, in Section VII, the conclusion of this paper is drawn.

II. DESCRIPTION OF THE PROPOSED ALL-PASS FILTER

Due to its versatility, the linearized operational transconductance amplifier (OTA) is the basic building block of many linear and nonlinear analog circuits, such as continuous-time oscillator and filters [19], [20]. The proposed circuit is based

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Fig. 1. Transistor level implementation of the proposed APF.

on a linear OTA architecture. The transistor level implementation of the APF is shown in Fig. 1. The differential input stage uses resistive degeneration to transform the differential input voltage $V_d = V_{in}^+ - V_{in}^-$ linearly into a signal current with value $i = V_d/R_{OTA}$. The complementary signal currents, in M_1 , M_{1P} (i and $-i$) are mirrored to the output branches of the OTA through two paths that add their currents in cascoded transistors M_{3C} , M_{3CP} . In one of the paths (M_3 , M_{3P}), the current i is replicated with unity gain. In the second path (through M_{3DL} , M_{3DLP}), the current is mirrored with a current gain -2 and filtered by first-order RC low-pass filters connected between nodes V_x^+ , V_x^- , and nodes V_B and V_A at the gates of M_{3DL} and M_{3DLP} . Cross-connections are used to achieve inversion of the low-pass filtered currents (i_{LPF}) in M_{3DL} and M_{3DLP} . The LPFs have a 3-dB frequency $\omega_0 = 1/(R_{large}C_M)$ where C_M is a Miller capacitance and R_{large} are very high valued resistors. As the low-pass filtered current in M_{3DL} (M_{3DLP}) is twice the current of M_3 (M_{3P}) and inverted as well, the resultant current i_s is given by

$$\begin{aligned} i_s(\omega) &= 2i_{LPF}(\omega) - i(\omega) \\ &= (V_d(\omega)/R_{OTA}) \left(\frac{2}{(1 + j\omega/\omega_0)} - 1 \right) \\ &= (V_d(\omega)/R_{OTA}) \frac{(1 - j\omega/\omega_0)}{(1 + j\omega/\omega_0)}. \end{aligned} \quad (1)$$

The output voltage of the filter is given by

$$\begin{aligned} V_{Out}(\omega) &= (Z_L || R_{Out}) i_{out}(\omega) \\ &= 2(Z_L || R_{Out}) i_s(\omega) \end{aligned} \quad (2)$$

where R_{Out} is the output impedance of the OTA, buiding block of the APF. The load impedance of the filter is given by

$$Z_L = (R_L || X_C) = R_L / (1 + j\omega C_L R_L). \quad (3)$$

Hence, using (1)–(3) the transfer function is given by

$$H(j\omega) = \frac{2}{R_{OTA}} \frac{(R_L || R_{Out})}{\left(1 + \frac{j\omega}{\omega_{Pout}}\right)} \left(\frac{(1 - j\omega/\omega_0)}{(1 + j\omega/\omega_0)} \right). \quad (4)$$

Fig. 2. Implementation of large resistor.

R_{Out} can be expressed by $R_{Out} = g_m r_o^2/2$ where g_m is transconductance gain and r_o is output resistance of the output transistors. Assuming $R_{Out} \gg R_L$, the output pole of the filter is given by $\omega_{pOut} \approx 1/R_L C_L$. Now, if ω_0 is extremely small compared to the output pole ω_{pOut} ($\omega_0 \ll \omega_{pOut}$) for frequencies $\omega \ll \omega_{pOut}$ (4) can be approximated by

$$H(j\omega) = 2(R_L/R_{OTA}) \frac{(1 - j\omega/\omega_0)}{(1 + j\omega/\omega_0)}. \quad (5)$$

Hence, (5) corresponds to the transfer function of a first-order APF with gain $K_{fil} \approx 2R_L/R_{OTA}$ and pole frequency $f_0 = 1/(2\pi R_{large}C_M)$. For $R_{OTA} = 2R_L$, the APF has unity gain.

The two important characteristics of the first-order filters are its phase and GD. The phase response of the filter obtained from (5) is given by

$$\phi = -2 \tan^{-1}(\omega/\omega_0). \quad (6)$$

The GD of the APF is given by (7). GD variations are a measure of the phase nonlinearity. If there is a linear phase variation with frequency, GD must be constant. A first-order APF does not provide linear phase shift. However, by increasing the order of filter linear phase shift can be approximated

$$GD = -\frac{d\phi}{d\omega} = \frac{2}{1 + (\omega/\omega_0)^2} \left(\frac{1}{\omega_0} \right). \quad (7)$$

The bandwidth (BW_{APF}) of the APF is determined by the high-frequency dominant pole f_{pOut} located at the output node V_{Out} . The f_{pOut} can be expressed by (8) when $R_L \ll R_{Out}$ and C_L is the load capacitance

$$f_{pOut} = \frac{1}{2\pi R_L C_L}. \quad (8)$$

The pole frequency f_0 with value in the order of Hz can be achieved by utilization of the leakage resistance of two diode-connected pMOS transistors operating in cutoff mode. Fig. 2 shows the implementation of these large valued resistances. From Fig. 2, it can be observed that two p-n junctions are formed. One is between the source and the n-well. The second one is between n-well and drains region. As the n-well is connected to V_{DD} , the p-n junctions are reverse biased. The leakage resistance of these reversed biased p-n junctions provides a very large resistance when the diode-connected pMOS is in cutoff region. These two large valued resistors are connected to the gate of the M_{3DL} and M_{3DLP} in Fig. 1 and

Fig. 3. (a) Block diagram of differential Miller amplifier. (b) Circuit implementation of linear Miller amplifier.

provide well-controlled dc gate bias voltage. These two transistors are called quasi-floating gate transistors [21], [22]. This technique has been used to implement extremely large resistors ($R_{\text{large}} \sim 10\text{--}100\text{ G}\Omega$) for many applications [23]. Miller multiplication effect is also used in this paper to emulate large capacitors from relatively low valued physical capacitors depending on the gain of the amplifier (in the example presented here $C_M \approx 310\text{ pF}$). The Miller multiplier circuit is described in Section III.

III. MILLER MULTIPLIER

A large capacitance without excessive silicon area requirements is implemented using Miller multiplication [24], as shown in Fig. 3(a). Here, capacitors (C) are connected between the input and output of an inverting amplifier with gain $-A_M$. This leads to an input Miller capacitance with value $C_M = C(1 + A_M)$. The linear amplifier used for this purpose is shown in Fig. 3(b). It has a similar structure as reported in [25] and [26]. It is comprised a linear voltage to current converter using two cascoded flipped voltage

Fig. 4. Equivalent circuit of the input stage of Miller amplifier.

followers (CFVFs) [27] in a differential configuration at the input stage. A pMOS cascode transistor $M_{CP1,CP2}$ connected between the drain of $M_{N1,N1P}$ and the gate of $M_{B1,B1P}$ in Fig. 3(b) provides local negative feedback. The drain currents in M_{N1} and M_{N1P} are constant and independent of the input voltage. Hence, neglecting second-order effects V_{GSMN1} and V_{GSMN1P} are constant. The CFVFs act as high-performance buffers with very low output resistance $r_{\text{outCFVF}} = 1/(g_m g_m r_o)$ at the sources of M_{N1} and M_{N1P} . In the example shown later, the value of r_{outCFVF} is $72\ \Omega$ and $R_M = 1.5\text{ k}\Omega$. Hence, $r_{\text{outCFVF}} \ll R_M$. This validates that the input stage of the capacitance multiplier is linear. The equivalent circuit of the input stage is shown in Fig. 4. The differential input voltage $V_{\text{indiff}} = (V_B - V_A)$ at the gates of M_{N1P} and M_{N1} is transferred to R_M . This leads to a signal current given by

$$i_M = (V_B - V_A) / R_M = V_{\text{indiff}} / R_M. \quad (9)$$

Current i_M flows in R_M and in M_{B1} and M_{B1P} . Transistors M_{N2} , M_{N2P} are scaled up by a factor of two with respect to transistors M_{B1} , M_{B1P} . Hence, the current i is replicated in the output branches (M_{N3} and M_{N3P}) with a value twice of i_M and flows in the load resistors R_{LMA} . Hence, the single-ended signal output voltages are given by

$$V_{\text{outMSE}} = V_D = -V_E = 2i_M R_{LMA}. \quad (10)$$

The voltage gain of the Miller amplifier with respect to differential input signal has a value $A_{M\text{diff}} = V_{\text{outMSE}} / V_{\text{indiff}} = 2R_{LMA} / R_M$. Hence, the voltage gain of the capacitance multiplier of one output with respect to a single-ended input is given by

$$A_M = -V_{\text{outMSE}} / (V_{\text{indiff}}/2) = -4(R_{LMA} / R_M). \quad (11)$$

In this paper, values of $R_{LMA} = 12\text{ k}\Omega$ and $R_M = 1.5\text{ k}\Omega$ are used for the capacitance multiplier. The bias current has a value of $I_{BM} = 40\ \mu\text{A}$, and the voltage gain of the capacitance multiplier is $A_M \approx -32$. The resultant Miller multiplication factor is $K_M = (1 - (-A_M)) = 33$. The capacitor C has a value of 10 pF . This results in a Miller capacitance $C_M = 330\text{ pF}$. The silicon area occupied by this amplifier with two 10-pF capacitors (C) is 0.012 mm^2 . If two 330-pF capacitors had been used instead of the Miller multiplier, it would occupy 0.245 mm^2 . Hence, 21 times area was saved at the expense of $192\ \mu\text{W}$ consumed by the Miller multiplier.

If required, the pole frequency f_0 of the APF can be programed by adjusting R_{LMA} , which controls the Miller

multiplication factor $K_M = 1 - A_M = 1 + 4R_{\text{LMA}}/R_M$. This is useful if f_0 needs to be tuned to compensate for manufacturing and/or temperature variations. Temperature dependence and manufacturing tolerances are common problems in all continuous time integrated filters [28], [29]. The programmability can be done by implementing variable R_{LMA} using switches as shown in Fig. 5. They connect one terminal of capacitor C to a tap in R_{LMA} . This causes the effective R_{LMAeff} to take values from 0 to $7R_1$, which corresponds to a capacitance multiplication factor K_M varying from 1 to 33. The unit resistor value is $R_{\text{LMAunit}} = R_1 = 1.71 \text{ k}\Omega$ and $R_M = 1.5 \text{ k}\Omega$. One way to tune the APF is to use an external accurate reference clock and scale down its frequency using digital frequency dividers to derive a reference delay that can be tuned to the delay of the proposed APF (this is dependent on the pole frequency). The APF would be tuned with the gain of the Miller Amplifier until the delay derived from the APF and the reference delay derived from the external clock match.

IV. NOISE ANALYSIS OF APF

Total noise currents at nodes V_x^+ , V_x^- generated in the input stage are mirrored to the output stage by transistors $M_{3,3p}$ with

a 1:1 gain factor and by transistors $M_{3DL,3DLP}$ with a 1:2 gain factor.

$$\overline{i_{\text{inAPff}}^2} = 2[\overline{i_{\text{nBPf}}^2} + \overline{i_{1f}^2} + \overline{i_{2f}^2}]. \quad (12)$$
$$\frac{1}{i_{nBPf}} = \frac{K}{C_{OX}(WL)_{BP}} \frac{g_{mBP}^2}{f} \left(\frac{R_{OTA}/2}{1/g_{m1} + R_{OTA}/2} \right)^2 \quad (13)$$
$$\begin{aligned} \overline{i_{1f}}^2 &= \frac{K}{C_{\text{OX}}(WL)_1} \frac{1}{f} \left(\frac{g_{m1}}{1 + g_{m1}R_{\text{OTA}}/2} \right)^2 \\ &\approx \frac{K}{C_{\text{OX}}(WL)_1} \frac{1}{f} \left(\frac{1}{R_{\text{OTA}}/2} \right)^2 \end{aligned} \quad (14)$$

$$\overline{i_{2f}^2} = \frac{K}{C_{OX}(WL)_2} \frac{g_{m2}^2}{f} \quad (15)$$

$$\overline{i_{3f}}^2 = \frac{K}{C_{OX}(WL)_3} \frac{g_{m3}^2}{f} \quad (16)$$

$$\frac{1}{i_{3\text{DL}f}^2} = \frac{K}{C_{\text{OX}}(WL)_{3\text{DL}}} \frac{g_{m3\text{DL}}^2}{f} \quad (17)$$

$$\overline{i_{4f}}^2 = \frac{K}{C_{OX}(WL)_4} \frac{g_{m4}^2}{f}. \quad (18)$$

The total output flicker noise current per unit BW resulting from the output stage transistors and the currents mirrored

from the input stage are

$$\overline{i_{\text{OutAPFF}}^2} = 2 \left(\overline{i_{n\text{BPF}}^2} + \overline{i_{1f}^2} + \overline{i_{2f}^2} \right) + \overline{i_{3f}^2} + \overline{i_{4f}^2} + \overline{i_{3\text{DLf}}^2}. \quad (19)$$

The output flicker noise voltage per unit BW is

$$\overline{V_{\text{outf}}^2} = \overline{i_{\text{OutAPFF}}^2} R_L^2.$$

From the transfer function of the filter given in (5), the gain of the filter is 1 from dc till 100 kHz. Hence, the input referred flicker noise voltage at ω_0 is

$$\overline{V_{\text{inf}}^2} = \overline{i_{\text{OutAPFF}}^2} R_L^2. \quad (20)$$

The source of another noise is thermal noise of MOSFET and resistors R_{OTA} and R_L . The total thermal input noise current spectral density originating from M_1 , M_{BP} , M_2 , and R_{OTA} reaching nodes V_x^+ , V_x^- are given by

$$\overline{i_{\text{inAPFT}}^2} = 2 \left[\overline{i_{n\text{BPT}}^2} + \overline{i_{1T}^2} + \overline{i_{2T}^2} + \overline{i_{\text{ROTA}}^2} \right]. \quad (21)$$

Thermal noise currents contributions per unit BW at nodes V_x^+ , V_x^- from the tail current source M_{BP} and input differential pair are given in the following equations:

$$\begin{aligned} \overline{i_{1T}^2} &= \frac{4kT\gamma}{g_{m1}} \left(\frac{g_{m1}}{1 + g_{m1}R_{\text{OTA}}/2} \right)^2 \\ &\approx \frac{4kT\gamma}{g_{m1}} \left(\frac{1}{R_{\text{OTA}}/2} \right)^2 \end{aligned} \quad (22)$$

$$\overline{i_{n\text{BPT}}^2} = 4kT\gamma g_{m\text{BP}} \left(\frac{R_{\text{OTA}}/2}{1/g_{m1} + R_{\text{OTA}}/2} \right)^2. \quad (23)$$

The thermal noise current contribution at nodes V_x^+ , V_x^- from R_{OTA} can be expressed by (24). k is Boltzmann's constant

$$\overline{i_{\text{ROTA}}^2} = \frac{4kT}{R_{\text{OTA}}/2} \left(\frac{R_{\text{OTA}}/2}{R_{\text{OTA}}/2 + 1/g_{m1}} \right)^2. \quad (24)$$

The thermal noise current contributions per unit BW from the elements in the output stage (M_3 , M_4 , $M_{3\text{DL}}$, and R_L) are given by

$$\begin{aligned} \overline{i_{nT3,4,3\text{DL},R_L}^2} &= 2 \left[4kT\gamma g_{m3} + 4kT\gamma g_{m4} + 4kT\gamma g_{m3\text{DL}} + \frac{4kT}{R_L} \right]. \end{aligned} \quad (25)$$

The output thermal noise voltage per unit BW resulting from the noise currents in the input stage mirrored to the output stage and the currents generated in the output stage is given by

$$\overline{V_{\text{outT}}^2} = \overline{i_{\text{OutAPFT}}^2} R_L^2 \quad (26)$$

where

$$\overline{i_{\text{OutAPFT}}^2} = 2 \left(\overline{i_{n\text{BPT}}^2} + \overline{i_{n1T}^2} + \overline{i_{n2T}^2} + \overline{i_{n\text{ROTA}}^2} + \overline{i_{nT3,4,3\text{DL},R_L}^2} \right). \quad (27)$$

Like flicker noise, it can be asserted that as the filter has magnitude gain of 1 the input referred thermal noise voltage per unit BW is $\overline{V_{\text{inT}}^2} = \overline{i_{\text{OutAPFT}}^2} R_L^2$. Hence, total input referred noise voltage of the APF per unit BW is $\overline{V_{\text{intotal}}^2} = \overline{V_{\text{inf}}^2} + \overline{V_{\text{inT}}^2}$.

Fig. 7. Single-ended magnitude response of Miller amplifier.

Fig. 8. Magnitude response of impedance C_M at the input of Miller amplifier.

For the design reported here with unit transistor size $(W/L)_N = 6.65/0.27$, $(W/L)_P = 33.75/0.27$, $g_{m1} = 864 \mu\text{A/V}$, and $f = 2 \text{ Hz}$, process-dependent constant $K = 1 \times 10^{-25}$, $C_{\text{ox}} = 1 \times 10^{-15} \text{ f}/\mu\text{m}^2$, $R_{\text{OTA}} = 10 \text{ k}\Omega$, and $R_L = 5 \text{ k}\Omega$ flicker noise spectral density is $75.9 \mu\text{V}/\sqrt{\text{Hz}}$ at f_0 . K and C_{ox} are obtained from the technology design kit. For thermal noise calculation Boltzmann's constant k is $= 1.38 \times 10^{-23}$, and T is 300°K . Thermal noise is $80 \text{ nV}/\sqrt{\text{Hz}}$ at f_0 . As expected, $1/f$ noise is dominant over thermal noise in the range of f_0 .

V. SIMULATION RESULTS

A. Simulation in Strong Inversion

The circuits of Figs. 1 and 3(b) were designed and simulated in a 130-nm CMOS technology with values $R_{\text{OTA}} = 10 \text{ k}\Omega$, $R_M = 1.5 \text{ k}\Omega$, capacitor $C = 10 \text{ pF}$, unit transistor sizes (in micrometers) of $33.76/0.27$ (pMOS) and $6.67/0.27$ (nMOS), $\pm 0.6\text{-V}$ dual supply voltages (V_{DD} and V_{SS}), a load $R_L = 5 \text{ k}\Omega$, $C_L = 25 \text{ pF}$, a bias current $I_B = 20 \mu\text{A}$, and $R_{\text{LMA}} = 12 \text{ k}\Omega$. The magnitude response A_M of the Miller amplifier is shown in Fig. 7. From Fig. 7, it can be observed that the single-ended gain for the inverting amplifier is $30 \text{ dB} \approx 31.62 \text{ V/V}$ instead of 32 V/V . The moderate reduction happens due to body effect from gate to source of transistors M_{N1} and M_{N1P} at the input of the Miller multiplier Fig. 3(b). This result complies with the theoretical value of the gain of the inverting amplifier which is given by 32 V/V . The magnitude and phase of the impedance of C_M at the input of the Miller multiplier are shown in Figs. 8 and 9, respectively. It can be seen that as the frequency increases

Fig. 9. Phase response of impedance C_M at the input of Miller amplifier.

TABLE I

POLE FREQUENCY AND THD (300-mV AMPLITUDE SINUSOIDAL SIGNAL) AT FOUR CORNERS

	tt	fs	ss	ssf
Pole frequency (Hz)	2.03	2.02	2.08	2.04
THD (dB)	-48	-44	-48	-40

Fig. 10. Transient response at four different corners.

impedance decreases at a constant rate of 20 dB/decade and that the phase is close to -90° up until 20 kHz. This behavior complies with the actual behavior of the reactance of a capacitor. At 2 Hz, the impedance is approximately 256 M Ω . As $X_C = 1/(j\omega C)$, the value of C is 310 pF at 2 Hz. Therefore, Miller multiplication makes it possible to emulate a 310-pF capacitor from a 10-pF base capacitor. Total harmonic distortion and pole frequency of the APF at four different corners (tt, ss, fs, and ssf) are given in Table I. The transient response of the APF is shown in Fig. 10 at four different corners. The simulated value of the pole frequency is given in Fig. 11 at four different corners. The standard deviation of the pole frequencies at four different corners is 0.03 Hz. A consistent pole frequency of the filter at four different corners has been obtained. THD has been calculated for each corner for a 300-mV amplitude sinusoidal input signal and given in Table I. The standard deviation of THD is 3.8 dB. Fig. 12 shows the equivalent input noise of the APF. From Fig. 12, it can be observed that as expected, at low-frequencies flicker noise is predominant. The measured value of the flicker noise at 2 Hz (84 $\mu\text{V}/\sqrt{\text{Hz}}$) is very close to its theoretical

Fig. 11. Phase response of APF at four different corners.

Fig. 12. Equivalent input noise of APF.

value (75.9 $\mu\text{V}/\sqrt{\text{Hz}}$) obtained from (20). Total rms noise can be found from [31]

$$V_{n,\text{rms}}(F_L, F_H) = V_{\text{nw}} \sqrt{F_C \ln(F_H/F_L) + (F_H - F_L)} \quad (28)$$

where V_{nw} is the white noise spectral density determined from simulations, F_L is the lowest frequency of interest in $1/f$ region, and F_C is corner frequency where power density of flicker noise and thermal noise are equal. F_C is 800 kHz as shown in Fig. 12. F_H is the equivalent noise BW that corresponds to $\text{QBW}_{\text{noise}} = (\pi/2)f_{3\text{dB}}$. Using $F_H = 1.57\text{ MHz}$, $F_L = 0.1\text{ Hz}$, $F_C = 800\text{ kHz}$, and $V_{\text{nw}} = 228\text{ nV}/\sqrt{\text{Hz}}$ in (28) a value $V_{n,\text{rms}} = 0.8\text{ mV}_{\text{rms}}$ is calculated.

The pole frequency can be varied by changing the gain of the Miller amplifier. The simulated magnitude and phase responses corresponding to the eight possible effective load resistances $R_{\text{LMA}} = 0, R_1, 2R_1, 3R_1, 4R_1, 5R_1, 6R_1, 7R_1$ are shown in Figs. 13 and 14. The pole frequencies f_0 vary from 2.03 to 48.62 Hz by varying the gain A_M for different values of R_{LMA} of the amplifier.

B. Simulation in the Subthreshold Region

The APF can work in the subthreshold region as well with an essentially lower bias current $I_B = 100\text{ nA}$. For this current, the output pole frequency has a value of $f_{\text{pOut}} = 2\text{ kHz}$. This is still at a much higher frequency than frequency f_0 which is close to 2.8 Hz. In this case, the supply voltage

Fig. 13. Magnitude response for different values of gain for varying R_{LMA} .

Fig. 14. Phase response for different values of gain for varying R_{LMA} .

Fig. 15. Simulated magnitude response of the APF in subthreshold.

can take significantly lower values of ± 200 mV. The circuit was simulated with I_B of 100 nA, the same transistor sizes as in Section V-A, and with resistors, R_{OTA} , R_L , R_M , and R_{LMA} each scaled up by a factor of 100. Total quiescent power dissipation in this case is $0.64 \mu\text{W}$. Figs. 15 and 16 show the frequency response of the APF operating in these conditions. It has 0.14-dB passband ripple. The f_0 frequency is at 2.76 Hz. At 2.76-Hz equivalent input noise is $80 \mu\text{V}/\sqrt{\text{Hz}}$. The equivalent input noise response is shown in Fig. 17. Fig. 18 shows the transient response of the filter for 50-mV amplitude and 2.76-Hz sinusoidal input signal.

VI. EXPERIMENTAL RESULTS

A test chip prototype was fabricated (through and thanks to Metal Oxide Semiconductor Implementation Service) in a $0.13\text{-}\mu\text{m}$ CMOS n-well process using the same parameters described in Section V for operation in strong inversion. A

Fig. 16. Phase response of APF at subthreshold.

Fig. 17. Equivalent input noise of APF at subthreshold.

Fig. 18. Transient response of APF at 2.7-Hz 50-mV amplitude sinusoidal signal at subthreshold.

Fig. 19. Experimental magnitude response of APF.

bias current $I_B = 20 \mu\text{A}$ and ± 600 mV were used to test the chip. The load capacitance $C_L = 25$ pF corresponds to the capacitance of the scope probes and the test board. Fig. 19 shows the measured magnitude response of the APF.

Fig. 20. Experimental phase response of APF.

Fig. 21. Simulated GD response of APF.

Fig. 22. Micrograph of the fabricated APF.

The experimental magnitude response is constant. At 100 kHz, a magnitude deviation of 0.08 dB is caused by the output pole. Magnitude decreases by 3-dB close to the output pole at $f_{pOut} = 1$ MHz. The value of the f_{pOut} , BW of the APF obtained from experiment conforms the theoretical value (1.2 MHz) obtained from (8). The measured f_0 is 2 Hz as shown in Fig. 20. Fig. 21 shows the simulated GD response. The theoretical value of GD calculated from (7) is 0.08 s obeys the simulation result. The micrograph of the fabricated circuit is given in Fig. 22. The proposed circuit occupies 0.029 mm^2 of silicon area.

Figs. 23–25 show the experimental transient responses of the APF for 0.2, 2, and 25 Hz for the 400-mVpp amplitude

Fig. 23. Experimental transient response at 0.2 Hz.

Fig. 24. Experimental transient response at 2 Hz.

Fig. 25. Experimental transient response at 25 Hz.

sinusoidal signal. It can be observed that at a low-frequency $f = 0.2$ Hz input and output signals are in phase. At 2 Hz, there is a 90° phase shift between input and output signals which have the same amplitude. Finally, the input and output signals are 180° out of phase at 25 Hz. These results validate the proposed circuit. The quiescent power consumed by circuit of Fig. 1 for a bias current $I_B = 20 \mu\text{A}$ is $192 \mu\text{W}$. The Miller multiplier consumed $192\text{-}\mu\text{W}$ quiescent power. Hence, total $384\text{-}\mu\text{W}$ quiescent power is consumed to obtain a 2-Hz pole frequency. Table II gives comparison results of previously published APFs and the proposed APF. From Table II, it can be ascertained that the proposed APF has the lowest f_0 frequency.

TABLE II
COMPARISON OF SOME PUBLISHED APF WITH THIS PAPER

	[5]	[2]	[6]	[7]	[8]	[9]	[10]	This Work	
Process Technology	130nm CMOS	130nm CMOS	0.7 μ m CMOS	0.5 μ m CMOS	0.5 μ m CMOS	4007 MOS transistor arrays	AMS 0.35 μ m CMOS	130nm CMOS (Exp)	130nm CMOS (Sim)
Supply Voltage(V)	1.5	1.5	-	± 2.5	± 2	± 2.5	1.5	± 0.6	± 0.2
Bias Current (μ A)	-	-	-	-	-	75-350	0.20	20	0.10
f_0 (Hz)	10G	≈ 6 G	9.68k	159k	2k-2M	≈ 50 -70k	22k	2	2.76
BW(Hz)								1M	2k
Power Dissipation (mW)	24	18.5	-	-	-	-	0.0003	0.38	0.00064
Si area	-	0.062 mm ²	-	-	-	-	-	0.029 mm ²	-
THD (dB)	-	-	-	-	-	-	-	-48*	-40**
Noise	36 pa/ $\sqrt{\text{Hz}}$ @10GHz	-	-	-	-	-	0.03mVrms	84 μ V/ $\sqrt{\text{Hz}}$ @ 2Hz	80 μ V/ $\sqrt{\text{Hz}}$ @ 2.76Hz
Ripple in pass-band	NA	NA	0.11dB	-	-	-	-	.08dB	0.14dB
Experiment-al validation	No	Yes	Yes	No	No	Yes	No	Yes	No
*THD has been calculated for the 300mV amplitude sinusoidal signal @2Hz ** THD has been calculated for the 50mV amplitude sinusoidal signal@2.76Hz									

VII. CONCLUSION

A first-order APF with extremely low f_0 frequency (Hz) was introduced and verified experimentally with a test chip prototype in 130-nm CMOS technology. To the authors' knowledge, no other APF with such low f_0 frequency has been reported in the literature. The proposed circuit is based on a linear OTA and provides very low passband ripple. It achieves extremely large time constants ($\sim$ seconds) with low silicon area requirements due to the utilization of quasi-floating gate transistors and Miller capacitors. By varying the gain of the Miller amplifier, variable f_0 can be achieved. Experimental results substantiate that the proposed APF provides phase quadrature at 2 Hz and a constant gain from dc to 100 kHz. It was verified with simulations that the proposed circuit can also work in the subthreshold region with essentially reduced supply voltages and power dissipation.

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