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Active-only variable-gain low-pass filter for dual-mode multiphase sinusoidal oscillator application

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Abstract: This research presents an active-only low-pass (LP) filter whose pole frequency and gain are independently tunable by means of bias current manipulation. The realization of the proposed LP filter required one operational amplifier, two operational transconductance amplifiers, and two MOS transistors. The multiphase sinusoidal oscillator (MSO) was subsequently realized by cascading n LP filters ($n = 3$) and loop-backing the output of the last LP filter to the input of the first LP filter. Simulations were carried out and the results revealed that the MSO could simultaneously achieve low impedance voltage and high impedance current outputs without the circuit topology alteration. The frequency of oscillation and the condition of oscillation could also be electronically tuned without disturbing each other. Furthermore, the sinusoidal signal of the MSO exhibited relatively low total harmonic distortion of about 0.8% at 2.75 MHz. To verify, a prototype of the active-only LP-based MSO ($n = 3$) was constructed and experiments were performed. The experimental findings were agreeable with the simulation results.

Key words: Low-pass filter, sinusoidal oscillator, tunable, low-voltage, active-only

1. Introduction

Essential to the communications and special-function integrated circuits (ICs) is the analog signal processing capability, which typically requires the amplifier, filter, and oscillator. Furthermore, modern ICs have incorporated programmable features and downsized the die area. Nonetheless, certain passive elements, e.g., the resistors and capacitors, are still present in the majority of modern ICs, contributing to the die area minimization problem.

In [1,2], a continuous time filter was proposed using both the active and passive elements; however, it suffered from the lack of tunability feature and poor high-frequency performance. In [3], the VLSI current-mode circuit was adopted for the high-performance filter and it has since gained wide acceptance due to the low voltage, small die area, and tunability features.

In [4], the author discussed different current-mode filters of diverse active elements. In [5], the current-mode universal filter was realized using the second-generation current conveyor (CCII) with the resistors and capacitors. In addition, the capacitors as the passive element were integrated into OTA- [6] and CC-DDCC-based [7] filters to generate the frequency response. Meanwhile, the universal filters with only-active elements (i.e. without the passive elements) were proposed in [8,9]. It is an interesting method for realizing the filter

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by making it unnecessary to use the passive elements. The active-only filters were generally designed based on lossless integrators. Specifically, the design of the filters was based on the operational amplifier (OA) and operational transconductance amplifier (OTA) and could achieve the multifunction filtering and tunability with low voltages. However, the complex structure of the lossy integrator using two OAs and three OTAs was reported [10]. Due to the high number of active elements used it required a large die area and power consumption.

In [11,12], the active-R circuit was adopted to implement the sinusoidal oscillator, in which the second-order function was realized by the OA internal pole and some resistors. However, the frequency of oscillation (FO) and the condition of oscillation (CO) were not independently adjustable. In [13], an active-only oscillator circuit using OA and OTA was realized to achieve the FO-CO independency. In addition, a multiphase oscillator was deployed in the phase modulator [14] and vector signal generator [15]. The cascading of all-pass filters with an integrator [16] and two lossless integrators [17] in the absence of the passive elements was implemented for the quadrature sinusoidal oscillators (QSOs).

Multiphase sinusoidal oscillators (MSOs) could achieve multiple (>2) outputs and are deployed in a variety of applications. The common MSO realization method is the cascading of lossy integrators, including the CCII [18,19], the OA [20], the current differencing transconductance amplifier (CDTA) [21], and the current amplifier (CA) [22]. Nevertheless, some of these circuits required capacitors and resistors, contributing to the circuit design challenge.

Thus, the resistorless-based MSOs using the OTA [23] and the current-controlled current conveyor transconductance amplifier (CCCCTA) [24] were proposed; however, capacitors were still required. In [25–27], an all-pass filter based on various active devices and a passive element were introduced. The cascading of all-pass filters was utilized to realize the MSO [26,27], but the capacitors were necessary for the pole frequency. More recently, the active-only QSO based on the cascading of all-pass filters was proposed [28]. Despite the active-only all-pass filter being introduced, the gain of filter was unable to adjust. The variable gain is an important feature for oscillator realization and due to the nonadjustability of the gain the condition of oscillation (CO) was unachievable. Importantly, no research exists on active-only MSOs with the independent tunability of the FO and CO.

This current research thus proposes an active-only low-pass (LP) filter whose pole frequency and gain are independently adjustable, and the MSO circuit was subsequently realized by cascading the LP filters. The active-only LP-based MSO could achieve low impedance voltage and high impedance current outputs, in addition to the independent adjustment of the CO and FO. Table 1 tabulates the previous research on three-phase MSOs and this current research.

2. Theoretical discussion

2.1. OA as the lossless integrator

The OA is an active element that produces the output with a very high-voltage gain [8]. It has been utilized in various applications, e.g., in amplifiers and filters. The open-loop gain ($A_v(s)$) of the OA can be expressed as a function of the LP filter, as shown in Eq. (1). Considering Eq. (1), at a very low pole frequency (ω_i), ω_i can be neglected and the function approximated and rewritten as a lossless integrator function.

$$A_v(s) = \frac{V_O(s)}{V_{in}(s)} = \frac{B}{s+\omega_i} \approx \frac{B}{s} \tag{1}$$

Here, B is the gain bandwidth and ω_i is the pole at very low frequency of the OA.

Table 1. Existing research on the three-phase MSO compared to this current research.

Ref.	Active element	Type of cascaded filter	Number of active elements	Number of R + C	CO + FO electronic tunability	High impedance current output	Low impedance voltage output
[17]	CCII	LPF	3	6 + 3	No + No	No	No
[18]	CCCII	LPF	3	0 + 6	No + Yes	Yes	No
[19]	OA	LPF	3	6 + 3	No + No	No	Yes
[20]	CDTA	LPF	5	0 + 3	Yes + Yes	Yes	No
[21]	CA	LPF	3	0 + 3	No + Yes	Yes	No
[22]	OTA	LPF	4 + 2 MOS	0 + 3	Yes + Yes	No	No
[23]	CCCCTA	LPF	3	0 + 3	Yes + Yes	Yes	No
[24]	CCCDTA	APF	3	3 + 3	Yes + Yes	Yes	No
[25]*	OA+OTA	APF	3OA + 6OTA	0 + 0	No + Yes	Yes	No
Proposed	OA+OTA	LPF	3OA + 6OTA + 6MOS	0 + 0	Yes + Yes	Yes	Yes

*The research in [25] concerns the QSO topology, from which the MSO topology of this current research is adopted.

2.2. Active-only low-pass filter

The LP filter, or the lossy integrator, can be realized using the negative loop-back of the lossless integrator, as shown in Figure 1a. In addition, to realize the gain-adjustable LP filter, an additional proportional block (k) has to be integrated, as shown in Figure 1b.

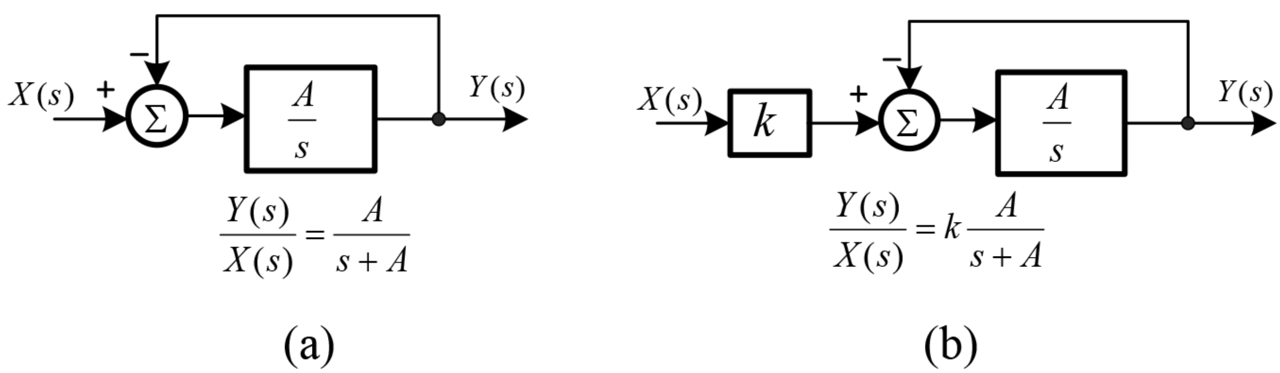

Figure 1. The low-pass filter realized by the lossless integrator (a, b).

Figure 2a illustrates the active-only unity gain LP filter, which is generally realized by one OA and two multioutput OTAs. For the variable-gain LP filter, it required one more electronic resistor as shown in Figure 2b. In Figure 2b, two current replicas ($+i_{in}$, $-i_{in}$) of the input current (i_{in}) could be obtained at the remaining output terminals of OTA$_2$. The current replicas are required for the current-mode MSO. Meanwhile, the voltage and current relationships of the active-only LP filter can be expressed as:

$$V_1 = i_{in}/g_{m2} \tag{2}$$

$$V_3 g_{m1} R = V_2 \tag{3}$$

$$V_3 = (V_1 - V_2)\left(\frac{B}{s}\right) = \frac{-i_{out}}{g_{m1}} \tag{4}$$

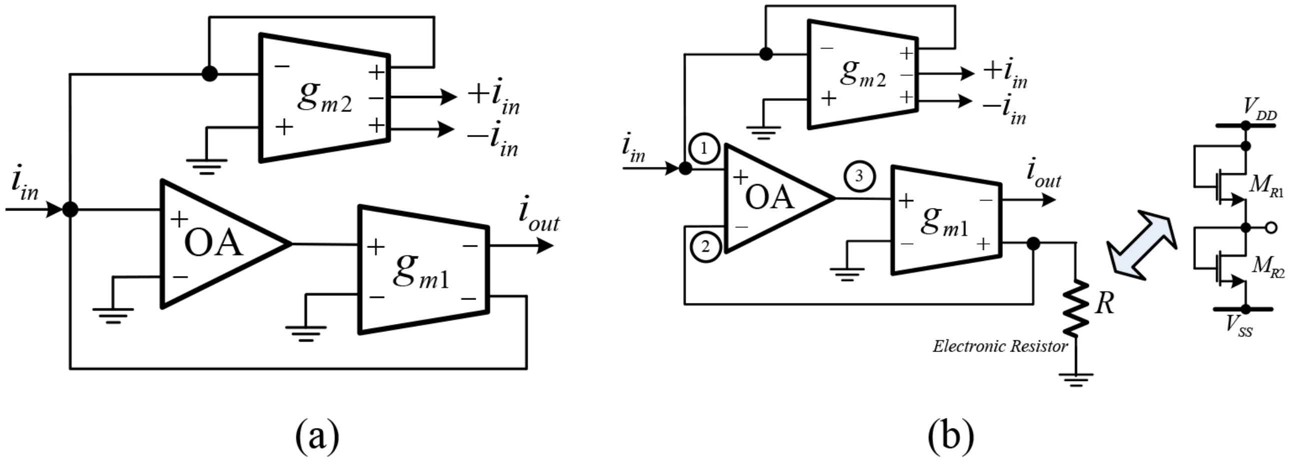

Figure 2. The active-only low-pass filter: (a) variable-gain, (b) unity-gain.

Thus, the current transfer function can be expressed as:

$$\frac{i_{out}}{i_{in}} = \frac{-1}{Rg_{m2}}\left(\frac{1}{s/g_{m1}RB+1}\right) \tag{5}$$

Here, $k = -1/Rg_{m2}$ is the gain and $1/\tau = \omega = g_{m1}RB$ is the pole frequency (ω). Note that the gain and the pole frequency of the LP filter could be independently tuned by varying the transconductance, g_{m2} and g_{m1}, respectively. In fact, this experimental research utilized two MOS transistors in place of the electronic resistor [29] and obtained the resistance (R) as:

$$R = \frac{L}{2\mu C_{OX}W(V_{DD}-V_T)} \tag{6}$$

Here, μ, C_{ox}, W, L, V_T, and V_{DD} represent the surface mobility, channel oxide capacitance, channel width and channel length, threshold voltage of the MOS transistor, and voltage supply, respectively.

2.3. The n-cascaded LP-based oscillator

In this experimental research, the MSO was realized by cascading of the LP filters [22], as shown in Figure 3.

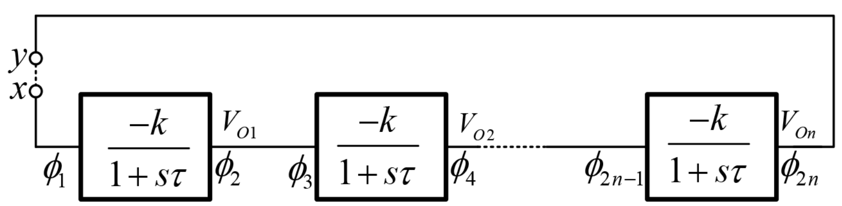

Figure 3. The MSO with the n-cascaded negative LP filters (n = odd).

Specifically, the LP network was realized by cascading n negative-LP filters, and the MSO was realized by looping the output of the final LP filter back to the first LP filter. It is imperative that n be an odd integer to obtain the negative closed-loop (MSO) function, as shown in Eq. (7).

$$H(j\omega_0) = \left(\frac{-k}{1+j\omega_0\tau}\right)^n = 1 \tag{7}$$

In Figure 3, given the different phases (ϕ), each LP filter can produce 2ϕ and the phase function of the low-pass network (LPN) can be expressed as:

$$\angle H(j\omega_0) = 2n\phi = 2n(0 - \tan^{-1}(\omega_0\tau)) = 2\pi \tag{8}$$

Then the frequency of oscillation (FO) can be rewritten as:

$$\omega_0 = \frac{1}{\tau}\tan\left(\frac{\pi}{n}\right) \tag{9}$$

Furthermore, the condition of oscillation (CO) can be derived from Eq. (9) and expressed as:

$$k = \sqrt{1 + (\omega_0\tau)^2} \tag{10}$$

3. Circuit description

In this research, the proposed active-only LP filter was realized using one OA, two multioutput OTAs, and two MOS transistors. Due to the die area minimization, the CMOS OA [30] (Figure 4) and the simple three-output CMOS OTA (Figure 5) were used for the LP filter design.

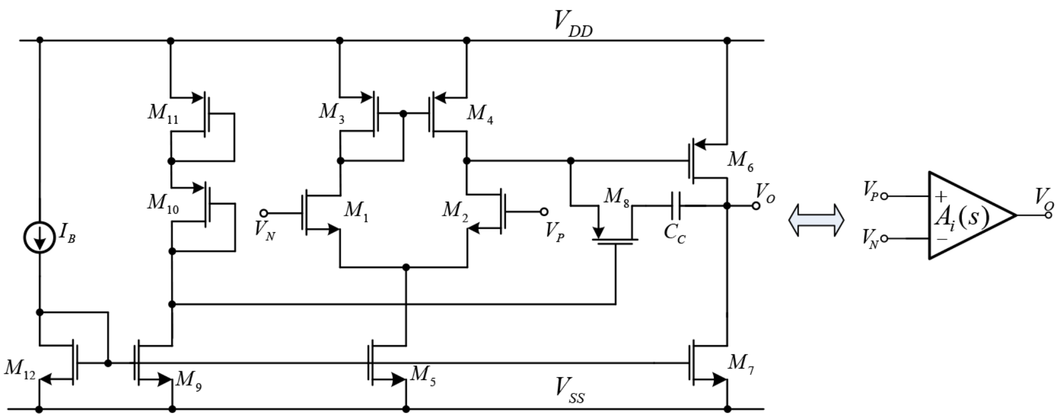

Figure 4. Schematic of the CMOS OA.

Figure 6 illustrates the schematic of the three negative LP sections (A, B, C), given that n is an odd integer for the negative closed-loop MSO (i.e. $n = 3$). The three voltage (V_o) and six current outputs (i_o) could thus be achieved.

From Eqs. (9) and (10), the FO and CO for $n = 3$ can be expressed as in Eqs. (11) and (12), respectively.

$$\omega_0 = \sqrt{3}g_{m1}RB \tag{11}$$

$$k = 1/Rg_{m2} = 2 \tag{12}$$

4. Nonideality study

The performance of the active-only LP filter is subject to the OTA output current gain (α), the parasitic resistance (R_{Pi}), and capacitance (C_{Pi}). Figure 7 illustrates the schematic of the proposed LP filter with

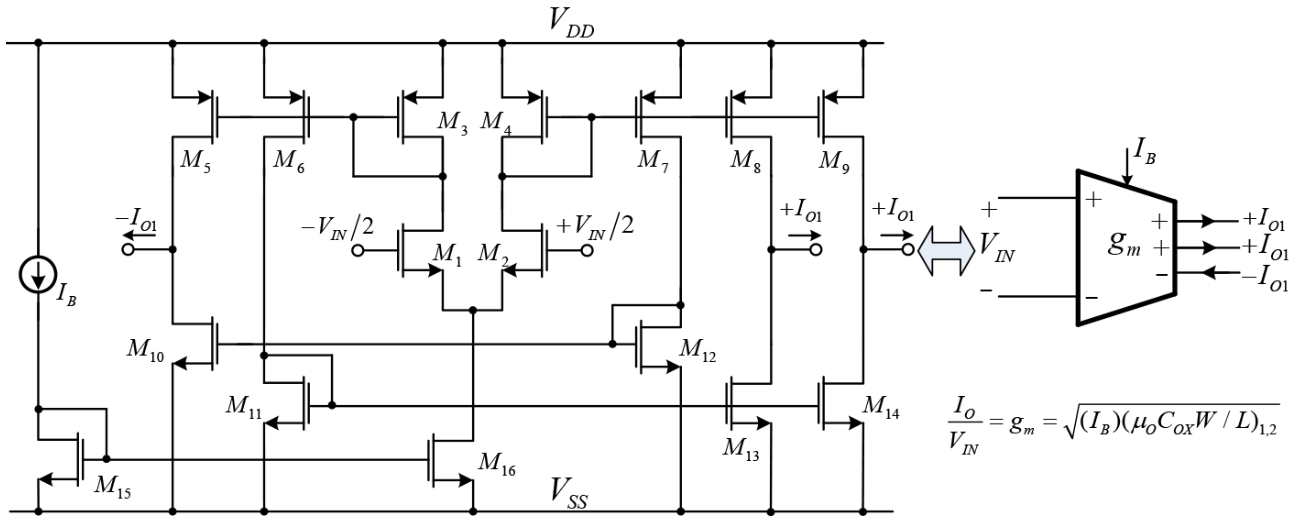

Figure 5. Schematic of the simple three-output CMOS OTA.

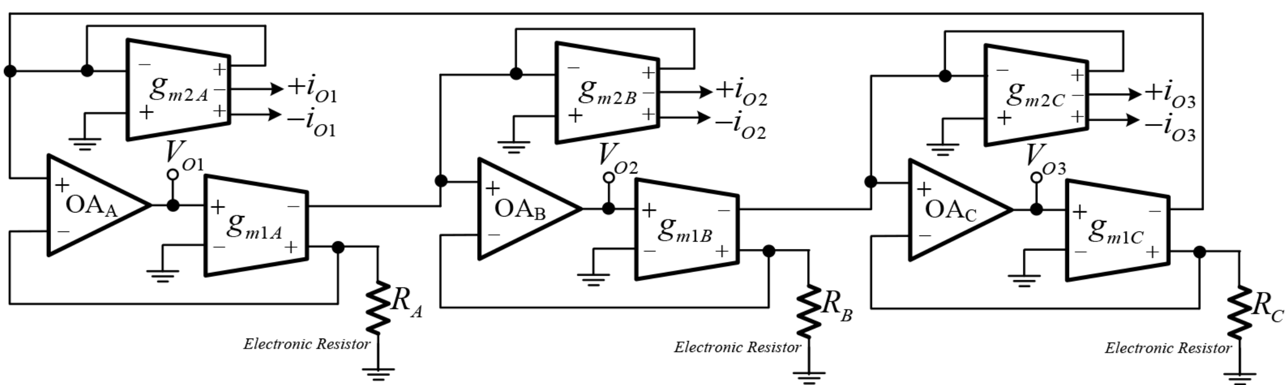

Figure 6. The active-only LP-based MSO (n = 3).

the OTA current gain (α) and the parasitic elements (R_{Pi} and C_{Pi}). In the figure, $R_{P1} = R_i + R_{O1}$, $R_{P2} = R_i + R_{O2}$, $C_{P1} = C_i + C_{O1}$, and $C_{P2} = C_i + C_{O2}$, given that R_{Pi} and C_{Pi} are the parasitic elements, where R_i and C_i are the resistance and capacitance of the OA input and R_{Oi} and C_{Oi} the resistance and capacitance of the OTA_i output.

4.1. Effect of the nonideal OTA current gain (α)

Neglecting the parasitic resistance (R_{Pi}) and capacitance (C_{Pi}), the transfer function of the LP filter, given the nonideal current gain (α), can be rewritten as:

$$H_{n1}(s) = \left(\frac{-\alpha_1}{\alpha_2 R g_{m2}}\right) \frac{R g_{m1} B}{s + \alpha_1 R g_{m1} B} \tag{13}$$

From Eq. (13), the pole frequency (ω) of the LP filter is minimally influenced by the OTA_1 current gain (α). Similarly, the gain (k) is slightly affected by the current gains of OTA_1 and OTA_2.

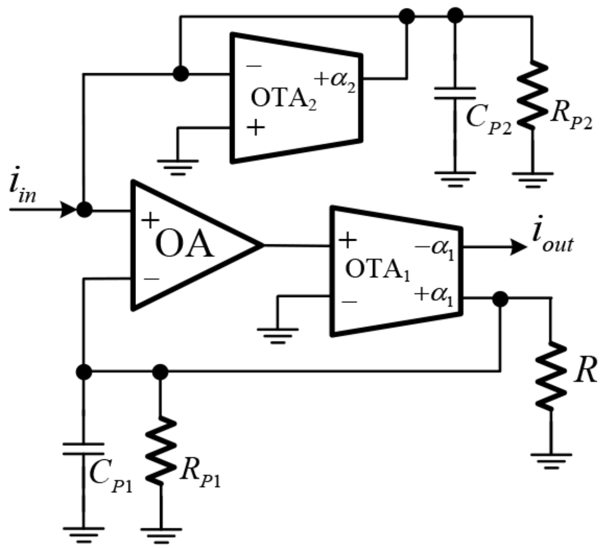

Figure 7. Schematic of the low-pass filter with nonideal parasitic elements.

4.2. Effect of parasitic resistance (R_P)

Neglecting the nonideal current gain (α) and parasitic capacitance (C_{Pi}), the transfer function of the LP filter, given the nonideal parasitic resistance (R_{Pi}), can be rewritten as:

$$H_{n2}(s) = \frac{-(g_{m1}BR_{P1}R_{P2} \;\; + Rg_{m1}BR_{P1})}{Rg_{m1}BR_{P2} + Rg_{m2}g_{m1}BR_{P1}R_{P2} + s\,(g_{m2}R_{P1}R_{P2} + Rg_{m2}R_{P1} + R_{P2} + R\;\;)} \tag{14}$$

Given that the parasitic resistances (R_{Pi}) are larger than R (i.e. $R << R_{P1}, R_{P2}$), the transfer function in Eq. (14) can be rewritten as:

$$H_{n2}(s) \approx \frac{-g_{m1}BR_{P1}}{(Rg_{m1}B)(1 + g_{m2}R_{P1}) + s\,(g_{m2}R_{P1} + 1)}$$

From Eq. (14), it is obvious that the transconductance range of OTA$_2$ is $g_{m2} >> 1/R_{P1}$.

4.3. Effect of parasitic capacitance (C_P)

Neglecting the nonideal current gain (α) and parasitic resistance (R_{Pi}), the transfer function of the LP filter, given the nonideal parasitic capacitance (C_{Pi}), can be rewritten as:

$$H_{n3}(s) \approx \frac{-(g_{m1}B + sRg_{m1}BC_{P2})}{Rg_{m2}g_{m1}B + s(g_{m2} + Rg_{m1}BC_{P1})} \tag{15}$$

From Eq. (15), the suitable electric resistor (R) is governed by the parasitic capacitance (C_{Pi}), which in turn is dictated by the OA and OTA$_1$, given that $R << (1/g_{m1}BC_{P1}), (1/g_{m1}BC_{P2})$.

5. Simulation and experimental results

In this research, the active elements were of the TSMC 0.25 μm CMOS technology [7] with ± 1.5 V power supply. The 3 pF compensated capacitor (C_C) and 30 μA bias current (I_B) were used to realize the OA with a

3.6 MHz gain-bandwidth (GBW). In addition, the OA, the simple three-output OTA, and the electronic resistor were realized using the MOS transistors (Table 2).

Table 2. Dimensions of the MOS transistors for the OA, OTA, and electronic resistor.

OA		OTA	
Transistors	W/L (μm/μm)	Transistors	W/L (μm/μm)
M1, M2, M10	1/1	M1, M2	5/2
M3, M4, M11	15/1	M3-M9	15/1
M5, M12	4.5/1	M10-16	10/1
M6	94/1	Electronic resistor	
M7	14/1	MR1, MR2	10/10
M8	6/1		
M9	2/1		

The LP filter could achieve the independent tuning of the pole frequency (ω) and the gain (k), given that $k > 1$. Figure 8 illustrates the magnitude and phase response of the negative LP filter, given varying I_{B1} and $I_{B2} = 80$ μA. The results showed that, given I_{B2} of 80 μA, the pole frequency could be varied corresponding to its phase response at 135 degrees (570 kHz to 2 MHz) by adjusting I_{B1} (6.25–100 μA) without disturbing the gain ($k = 5$ dB or $k = 1.8$).

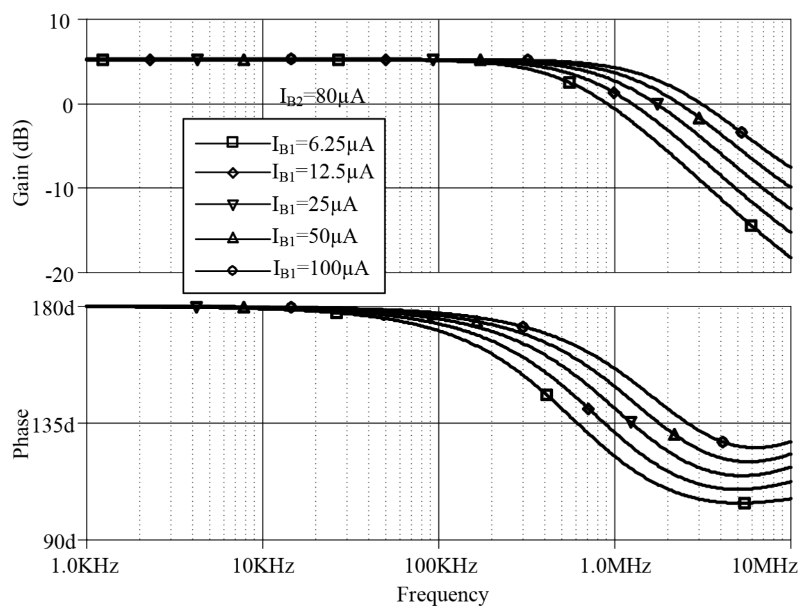

Figure 8. Magnitude and phase response of the negative LP filter given $I_{B2} = 80$ μA and varying I_{B1}.

Figure 9 depicts the magnitude and phase response of the negative LP filter, given $I_{B1} = 10$ μA and varying I_{B2}. The results revealed that, given I_{B1} of 10 μA, the gain could be varied by adjusting I_{B2} (20–320 μA) without disturbing the pole frequency and its phase response at 135 degrees is also constant at 900 kHz.

It is obvious from Figures 8 and 9 that the gain and pole frequency could be independently tuned by manipulating the particular bias current.

Figure 10 shows the noise output of the negative LP filter, given $I_{B2} = 80$ μA and varying I_{B1}. The results show that low output noises are observed lower than 130 pV/$\sqrt{Hz}$ along the frequency. Figure 11

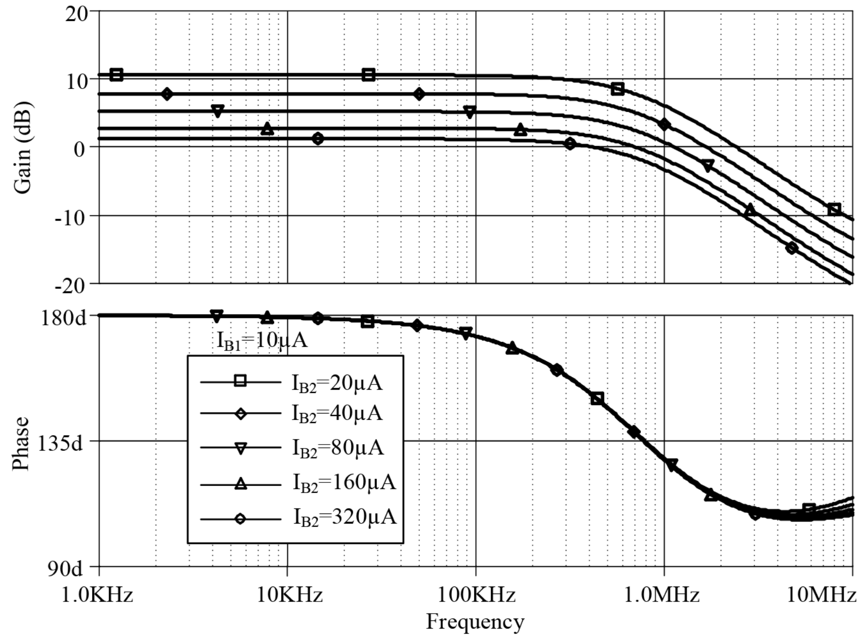

Figure 9. Magnitude and phase response of the negative LP filter given $I_{B1} = 10\ \mu$A and varying I_{B2}.

illustrates the THD measurement based on $I_{B1} = 30\ \mu$A and $I_{B2} = 40\ \mu$A ($f_0 \approx 1$ MHz and $k \approx 2$). Sinusoidal current inputs of in-band frequency (10 kHz and 100 kHz) are applied while varying the amplitude. The THD of the proposed active-only variable gain LP filter is satisfied at less than 1% for 10 μA of input current.

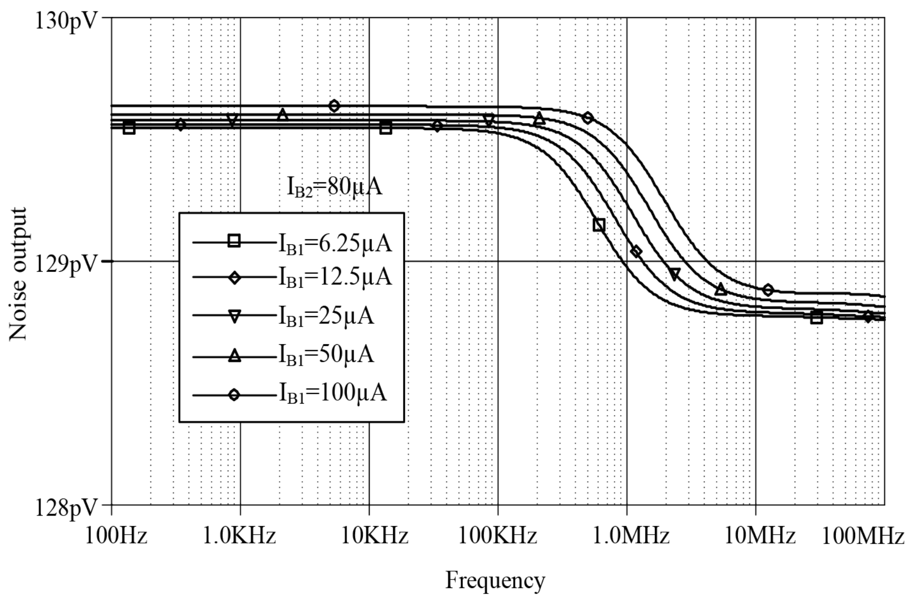

Figure 10. Noise output of the negative LP filter given $I_{B2} = 80\ \mu$A and varying I_{B1}.

In the simulation, given the $n = 3$ MSO configuration (Figure 6) with $I_{B1} = 30\ \mu$A, $I_{B2} = 39.7\ \mu$A for $k = 2$, and the three voltage and six current output terminals respectively connected to the 200 kΩ and 1 Ω

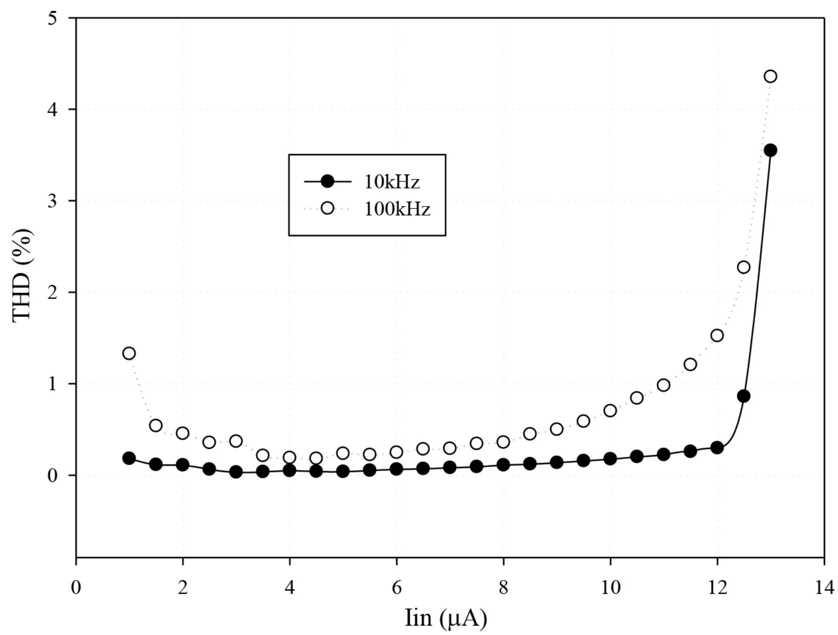

Figure 11. THD measurement of the negative LP filter given $I_{B1} = 30$ μA and $I_{B2} = 40$ μA.

load resistors, the proposed MSO could achieve the frequency of oscillation (FO) of approximately 2.75 MHz. In Figure 12, the voltage outputs (V_{o1} to V_{o3}) were 120° equally spaced in phase and 150mV in amplitude, while the current outputs ($\pm I_{o1}$ to $\pm I_{o3}$) were 60° equally spaced in phase and 18 μA in amplitude.

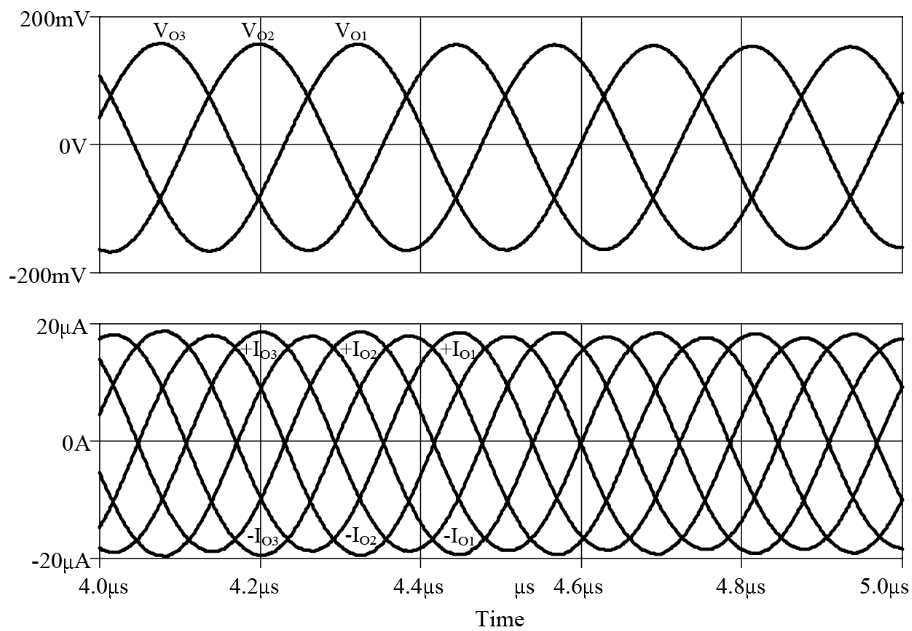

Figure 12. The voltage and current sinusoidal outputs of the proposed active-only LP-based MSO ($n = 3$).

Figure 13 depicts the corresponding frequency spectra of the sinusoidal signal outputs. In the figure, at the voltage terminal, the 3rd harmonic component was 1.25 mV, while the fundamental component was 152.5 mV. At the current terminal, the 3rd harmonic component was 325.87 nA, while the fundamental component was 18.6 μA. The total harmonic distortions (THDs) of the voltage and current terminals at 2.75 MHz were around 0.8% and 1.75%, respectively.

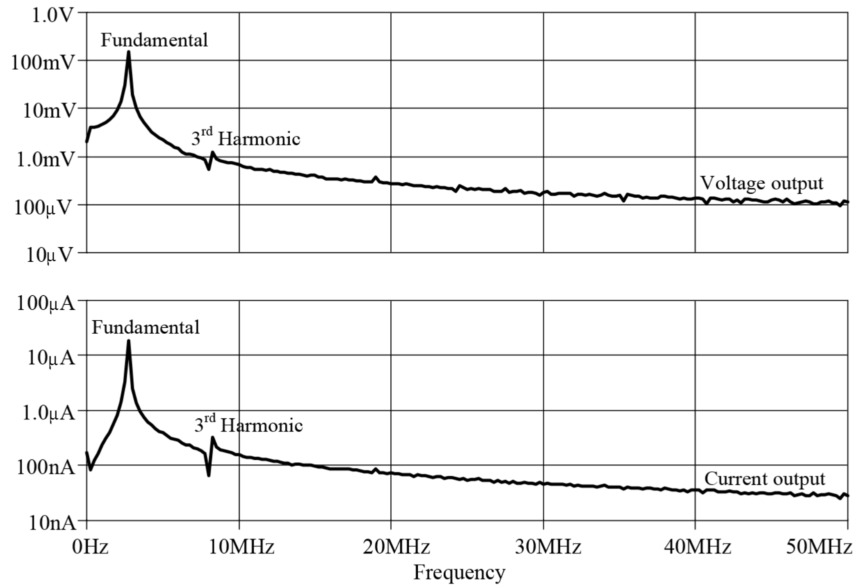

Figure 13. The voltage and current spectra of the proposed active-only LP-based MSO ($n = 3$).

To validate the simulation results, an active-only LP-based MSO ($n = 3$) prototype was fashioned using the commercially available OTA (LM13600) and OA (LM324). Figures 14 and 15 respectively illustrate the schematic of the experimental active-only LP-based MSO and the corresponding prototype.

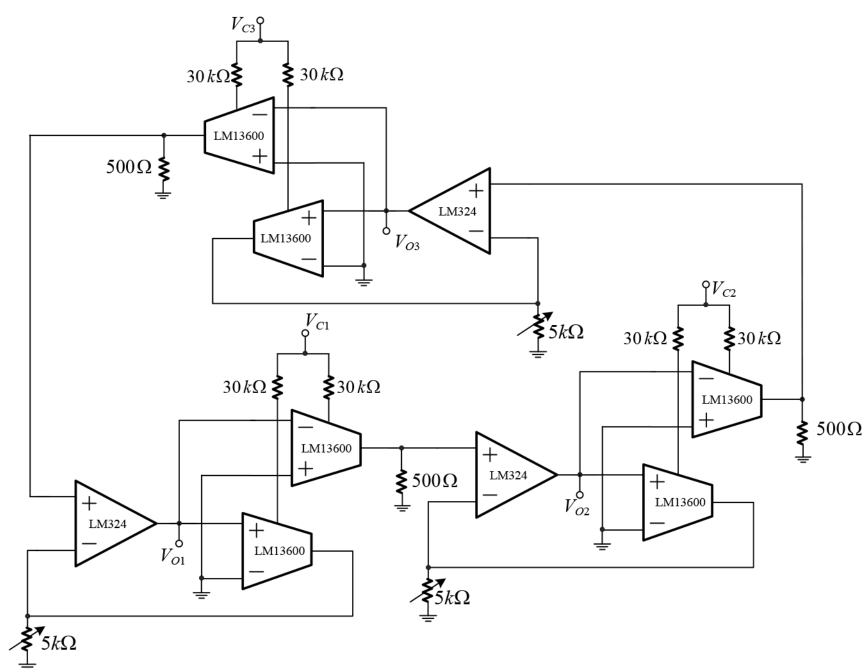

Figure 14. The schematic of the experimental LP-based MSO ($n = 3$).

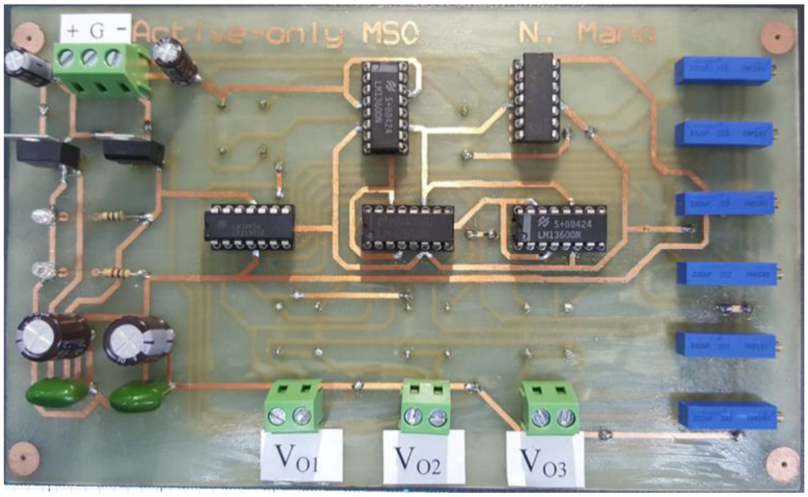

Figure 15. Prototype of the active-only LP-based MSO (n = 3).

In Figure 14, the OTA bias currents of 50 μA were fed through V_{Ci} and the 30 kΩ resistors, while the OA gain-bandwidth was around 1 MHz. For the sake of simplicity, the electronic resistor (R) and OTA$_2$ (a grounded resistor for the gain manipulation) were respectively replaced with a passive resistor (R = 500 Ω) and a 5 kΩ variable resistor. The three voltage outputs were obtained at around 700 kHz with 120° roughly spaced in phase and 120 mVp-p in amplitude, as shown in Figure 16, which is agreeable with the simulation results. Figure 17 illustrates the spectrum of one voltage output with the 3rd harmonic distortion of around –21 dB.

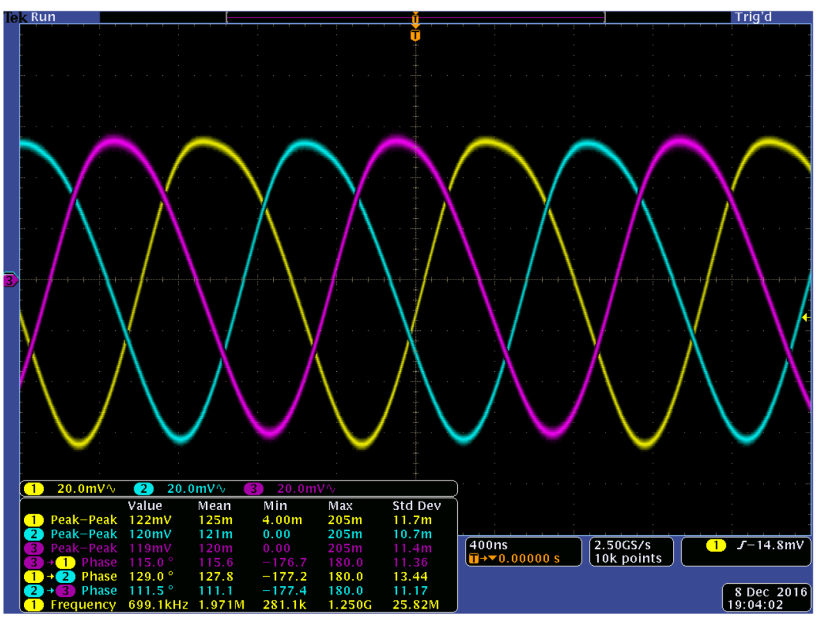

Figure 16. The experimental result of the three voltage outputs.

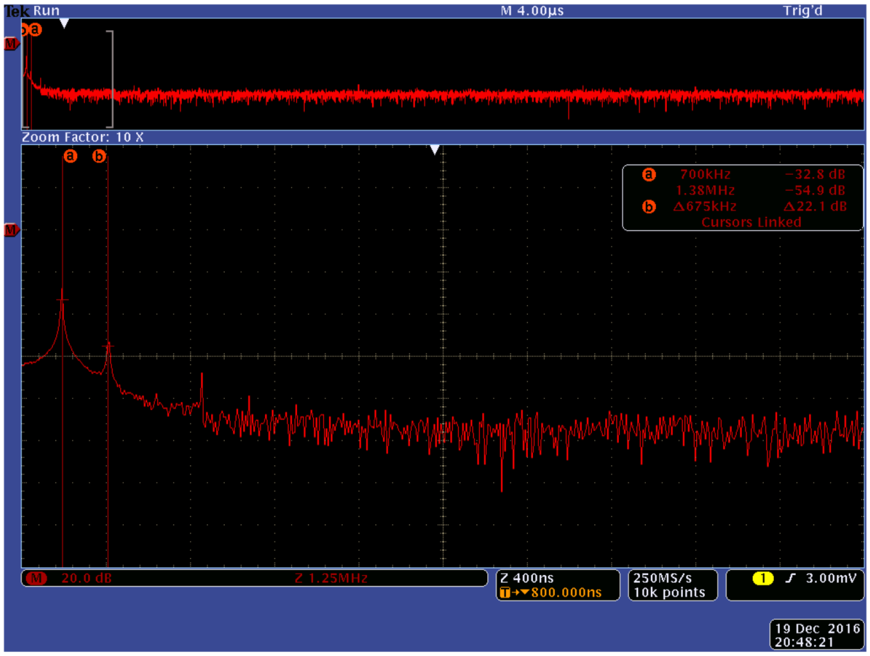

Figure 17. The experimental result of the spectrum in one voltage output.

6. Conclusion

This experimental research has proposed an active-only LP filter (i.e. without the external passive element) whose pole frequency and gain can be independently tuned by means of the bias current manipulation. The proposed LP filter consisted of one OA, two MO-OTAs, and one MOS-based resistor. The MSO was subsequently realized by cascading the LP filters ($n = 3$). Simulations were carried out and the results showed that the MSO could concurrently achieve low impedance voltage and high impedance current outputs without circuit topology alterations. The FO and the CO could also be electronically tuned without disturbing each other. In addition, the sinusoidal signal of the MSO exhibited a relatively low THD of around 0.8% at 2.75 MHz. To validate, the active-only LP-based MSO ($n = 3$) prototype was constructed and experiments were performed. The simulation results are agreeable with the experimental findings.

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