An integrable electronic-controlled quadrature sinusoidal oscillator using CMOS operational transconductance amplifier

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This paper describes two approaches to implementing third-order oscillators. The first approach proposes a third-order oscillator using transconductors and capacitors. They are cascaded as two lossy and one lossless integrator circuit. This approach is a feedback from transconductance gain $g = k$. This first circuit is based on a basic transconductor with a simple configuration including 16 transistors, four current sources and three capacitors. The second approach proposes a third-order oscillator using transconductors, capacitors and a transresistor circuit. These are cascaded as lossy integrator circuits and are feedback with voltage gain $v = 8$. This voltage gain can be designed using a transconductor circuit and a transresistor circuit. This second circuit consists of 18 transistors, four current sources and three capacitors. Since both circuits use no resistors, they are suitable for further fabrication. These circuits use a $\pm 3$ V power supply.

1. Introduction

Linear transconductors or voltage-to-current converter circuits are fundamental building blocks of analogue circuits and systems. They are used in analogue filters, voltage controlled-resistance circuits, A/D or D/A converters etc. Recently, sinusoidal oscillators have been created in various approaches (Senani 1985, 1993, Boutin 1986, Abuelma and Almaskati 1987, Vazquez et al. 1990, Chen et al. 1991, Bhaskar and Senani 1993). For analogue signal processing, the operational transconductance amplifier (OTA) is interesting. The OTA can be used for sinusoidal oscillators and analogue filters (Malvar 1982, Abuelma and Almaskati 1987, Sinencio et al. 1988). In this paper, we propose a new approach to implementing the two different sinusoidal oscillators. Both proposed circuits are based on a third-order network because a high-order network has high accuracy and high quality factor ($Q$). It gives us good frequency response with low distortion. The frequency output of the third-order oscillator is a high-accuracy, high-purity sine wave and multiple phase of output. Both approaches to oscillators are easily configured, frequency controllable and further integrable.

2. The operational transconductance amplifier (OTA)

The OTA has input as voltage, output as current. The simple CMOS OTA uses only four transistors and a current source, as shown in figure 1. From figure 1, the transconductance is given by
3. Principle of the oscillator

Sinusoidal oscillators must normally have their loop gain ($L_G$) set to 1. The principle of operation is shown as a block diagram in figure 2. This diagram consists of an amplifier ($A$), network ($H(s)$) and summing junction. Positive feedback has been assumed so that the transfer function is equal to

$$\frac{v_o}{v_{in}} = \frac{H(s)}{1-kH(s)} = \frac{H(s)}{1-LG} \tag{2a}$$

The operating principle of the oscillator can be obtained from (2a). When there is no input ($v_{in} = 0$) for finite output, theoretically $v_o$ must be equal to infinity. This case can occur if and only if the loop gain ($L_G$) = 1. The oscillator can be realized by loop gain ($kH(s)$) = 1 or by the denominator ($1 - L_G$) = 0 accordingly.

Generally, for the third-order phase-shift oscillator, such as OPAMP, the feedback gain is very high ($\approx |29|$). For this reason, the size of transistors is very large. In addition, this is not suitable for the integrated form although the output is quite highly accurate. These are restrictions of the third-order sinusoidal oscillators.

The transfer function of the third-order oscillator can be written as

$$1 - L_G = \frac{N(s)}{D(s)} = 0 \tag{2b}$$

![Figure 1. Simple operational transconductance amplifier (OTA).](image1)

![Figure 2. Block diagram of feedback system](image2)
where the numerator $N(s)$ is a third-order polynomial. With its coefficients defined as $a_i$, $N(s)$ can be written as

$$N(s) = a_0 s^3 + a_1 s^2 + a_2 s + a_3$$  (3)

where $s = j\omega$. Thus

$$0 = N(j\omega) = -j\omega^3 a_0 - \omega^2 a_1 + j\omega a_2 + a_3$$  (4)

From (4), the coefficients are found to be

$$a_3 - a_1 \omega^2 = 0 \quad \text{and} \quad a_2 - \omega^2 a_0 = 0$$  (5)

From (5), the conditions for oscillation are given by

$$a_0 a_3 - a_1 a_2 = 0$$  (6)

and then the oscillation frequency is

$$\frac{a_3}{a_1} = \frac{a_2}{a_0} = \omega^2$$  (7)

4. The first OTA sinusoidal oscillator

The principle of lossy and lossless integrators is used to implement a third-order filter as shown in figure 3. The transfer function of the system in figure 3 can be written as

$$\frac{v_o}{v_{in}} = \frac{\alpha_1 \alpha_2 / \alpha_3}{s^3 + s^2 (\alpha_1 + \alpha_2) + s \alpha_1 \alpha_2}$$  (8)

or

$$\frac{v_o}{v_{in}} = \frac{\alpha_1 \alpha_2 / \alpha_3}{s (s^2 + s (\alpha_1 + \alpha_2) + \alpha_1 \alpha_2)}$$  (9)

From figure 3, the lossy integrator can be implemented as in figure 4 and the transfer function can be written as

$$H(s) = \frac{v_o}{v_{in}} = \frac{(g_m/C)}{s + (g_m/C)}$$  (10)

The lossless integrator can be connected in a cascade of a second-order low-pass filter as shown in figure 5. Its transfer function can be written as

$$\frac{v_o}{v_{in}} = \frac{g_{m1} g_{m2} / C_1 C_2}{s^2 + (g_{m1}/C_1 + g_{m2}/C_2)s + (g_{m1} g_{m2}/C_1 C_2)}$$  (11)

Figure 3. Third-order filter uses for first oscillator.
From (11), the second-order low-pass filter equation can be implemented as a third-order filter by cascade of a lossless integrator, as shown in figure 6. From the integrator circuit of figure 6, the transfer function can be written as

\[ \frac{v_o}{v_{in}} = - \frac{g_m}{sC} \]  

(12)

When we use figure 5 to cascade with figure 6, as shown in figure 7(a), the transfer function can be written as

\[ \frac{v_o}{v_{in}} = \frac{-g_{m1}g_{m2}g_{m3}/C_1C_2C_3}{s^3 + (g_{m1}/C_1 + g_{m2}/C_2)s^2 + (g_{m1}g_{m2}/C_1C_2)s} \]  

(13)

\( v_o \) is feedback connected to \( v_{in} \). The third-order low-pass filter shown in figure 7(a) becomes the sinusoidal oscillator in figure 7(b). In this case, the loop gain \( L_G \) should be equal to 1. From (2)–(7), the coefficients of the polynomial are \( a_0 = 1, \)
$a_1 = g_{m1}/C_1 + g_{m2}/C_2, \ a_2 = g_{m1}g_{m2}/C_1 C_2$ and $a_3 = g_{m1}g_{m2}g_{m3}/C_1 C_2 C_3$, and the oscillating conditions can be written as

$$\frac{g_{m3}}{C_3} = \left(\frac{g_{m1}}{C_1} + \frac{g_{m2}}{C_2}\right)$$

(14)

and the oscillation frequency is

$$\omega_n = \sqrt{\frac{g_{m1}g_{m2}}{C_1 C_2}}$$

(15)

Suppose the transconductances are $g_{m1} = g_{m2} = g_m$ and $C_1 = C_2 = C_3 = C$. From (14), $g_{m3} = 2g_m$, the oscillation frequency becomes $\omega_n = g_m/C$ and its electronic-controllable frequency is controlled by $g_m$.

5. The second OTA sinusoidal oscillator

The previous method for the OTA sinusoidal oscillator uses the approach of lossy and lossless integrators which are cascaded. This method uses only lossy integrators, which are cascaded as a third-order filter as shown in figure 8. The transfer function can be written as

$$\frac{v_o}{v_{in}} = \frac{\alpha_1 \alpha_2 \alpha_3}{s^3 + s^2(\alpha_1 + \alpha_2 + \alpha_3) + s(\alpha_1 \alpha_2 + \alpha_2 \alpha_3 + \alpha_1 \alpha_3) + \alpha_1 \alpha_2 \alpha_3}$$

(16)

Figure 8. Third-order filter uses for second oscillator.
From Figure 8, we connect an amplifier with gain \( k \); then the feedback to input is as shown in Figure 9. The loop gain is

\[
LG = \frac{k \alpha_1 \alpha_2 \alpha_3}{s^3 + s^2(\alpha_1 + \alpha_2 + \alpha_3) + s(\alpha_1 \alpha_2 + \alpha_2 \alpha_3 + \alpha_1 \alpha_3) + \alpha_1 \alpha_2 \alpha_3}
\] (17)

From (2)–(7), these oscillating conditions are given by

\[
k = -\left(\frac{\alpha_1}{\alpha_3} + \frac{\alpha_1}{\alpha_2} + \frac{\alpha_2}{\alpha_3} + \frac{\alpha_2}{\alpha_1} + \frac{\alpha_3}{\alpha_1} + \frac{\alpha_3}{\alpha_2} + 2\right)
\] (18)

and the oscillation frequency becomes

\[
\omega_n^2 = \alpha_1 \alpha_2 + \alpha_2 \alpha_3 + \alpha_1 \alpha_3
\] (19)

or

\[
\omega_n^2 = \frac{(1-k) \alpha_1 \alpha_2 \alpha_3}{\alpha_1 + \alpha_2 + \alpha_3}
\] (20)

Suppose \( \alpha_1 = \alpha_2 = \alpha_3 = \alpha_a \); then equations (18)–(20) become

\[
k = -8
\] (21)

\[
\omega_n^2 = 3 \alpha_a^2 \quad \text{or} \quad \omega_n = \sqrt{3} \alpha_a
\] (22)

6. The OTA voltage amplifier

This amplifier consists of an OTA and a resistor as shown in Figure 10. The gain is

\[
\frac{v_o}{v_{in}} = -g_{m4} R_{\text{eq}}
\] (23)

The circuit in Figure 10(b) is an electronic resistor (Wang 1990). The transistors are operated in the saturation region. In Figure 10(b), matched transistors M1 and M2 are diode connected. The input \( I_{\text{in}} \) is applied to the central node of the circuit, developing a voltage \( V_o \) at the node. Using the square law characteristic, the drain currents in M1 and M2 can be expressed as

\[
I_{D1} = \frac{\mu C_{OX}}{2} \left( \frac{W}{L} \right) (V_{\text{DD}} - V_o - V_T)^2
\] (24)

\[
I_{D2} = \frac{\mu C_{OX}}{2} \left( \frac{W}{L} \right) (V_o - V_{SS} - V_T)^2
\] (25)
where $V_{DD} = -V_{SS}$, $\mu$ is mobility of carriers, $C_{OX}$ is the gate capacitance per unit area, $V_T$ is the threshold voltage and $W$ and $L$ are the channel width and length, respectively.

Considering (24) and (25) as two square terms in the left-hand side of (24) and using Kirchhoff’s current law (KCL) and the current constraint at the node, a simple algebraic manipulation gives the transresistance result

$$R_{eq} = \frac{V_0}{I_{in}} = \frac{L}{2\mu C_{OX} W (V_{DD} - V_T)} \quad (26)$$

This second sinusoidal oscillator must be used with the gain $k = -8$. So we set the resistance value in figure 10(b) to $R_{eq} = 8/g_{m4}$. The third-order filter can be constructed as in figure 11. The various parameters have been set to be $\alpha_1 = \alpha_2 = \alpha_3 = \alpha$, $g_{m1} = g_{m2} = g_{m3} = g_{m4} = g_m$ and $C_1 = C_2 = C_3 = C$, so that the loop gain equals

$$LG = \frac{-8g_m^3/C^3}{s^3 + s^2(3g_m/C) + s(3g_m^2/C^2) + g_m^3/C^3} \quad (27)$$

The oscillating condition in equations (2)–(7), $N(s)$, will be given by

Figure 10. (a) OTA voltage amplifier. (b) Electronic resistor circuit.

Figure 11. Second OTA sinusoidal oscillator.
\[ N(s) = 0 = s^3 + s^2 \left(3g_m/C\right) + s \left(3g_m^2/C^2\right) + 9g_m^3/C^3 \]  

(28)

The oscillation frequency is

\[ \omega^2 = \frac{3g_m^2}{C^2} \quad \text{or} \quad \omega = \frac{\sqrt{3g_m}}{C} \]  

(29)

The electronically controllable frequency is controlled by \( g_m \).

7. Simulation and experimental results

The principles of the first and second designs can be realized by a simple OTA as shown in figures 12 and 13 respectively.

Figure 12. Complete first sinusoidal oscillator.

Figure 13. Complete second sinusoidal oscillator.
Figure 12 shows the complete first sinusoidal oscillator using a simple OTA. The aspect ratio of the transistors is $W/L = 30 \, \mu m/5 \, \mu m$ for M1, M2, M3, M5, M6, M7, M12, M16, $W/L = 25 \, \mu m/5 \, \mu m$ for M4, M8, M11, M15, and $W/L = 50 \, \mu m/5 \, \mu m$ for M9, M10, M13, M14. The current sources vary in the range $I = 10 \, \mu A$–$200 \, \mu A$ and capacitors vary from $C = 15 \, pF$ to $C = 100 \, nF$. The oscillation frequency is obtained in figure 14 and the different points of the sinusoidal signals are shown in figure 15. Their phases are about 90° different.

Figure 13 shows a complete second sinusoidal oscillator using a simple OTA. The aspect ratio of the transistors is $W/L = 30 \, \mu m/5 \, \mu m$ for M2, M3, M5, M6, M7, M9, M10, M11, $W/L = 25 \, \mu m/5 \, \mu m$ for M4, M8, M12, $W/L = 600 \, \mu m/5 \, \mu m$ for M13, M14, and $W/L = 5 \, \mu m/5 \, \mu m$ for M17, M18. The current sources vary from $I = 10 \, \mu A$ to $I = 200 \, \mu A$, $I_A = 600 \, \mu A$, and capacitors vary from $C = 15 \, pF$ to

Figure 14. First oscillation frequencies with current $I$ varied.

Figure 15. Different points sinusoidal signal of first oscillator with $C = 10 \, nF$, $I = 200 \, \mu A$, $f = 3.9 \, kHz$. 
$C = 100 \text{nF}$. The oscillation frequency is shown in figure 16 and the different points of the sinusoidal signals are shown in figure 17; their phases are about 90° different.

The MOS model is shown in figure 18. The simulation frequency output of both oscillators is shown in figures 19 and 20. The experimental frequency spectrum output of both oscillators is shown in figures 21 and 22. The fundamental frequencies of the first and second oscillator are 20 kHz and 30 kHz, respectively, with a total harmonic distortion (THD) of $-35 \text{ dB}$ lower than their fundamental frequencies. The THDs is shown in Table 1, along with the noise levels of both oscillators. The THDs are quite low depending on the current $I$ controlled. A comparison of this paper and previous work is given in Table 2. The output voltage against varied current $I$ of both oscillators is shown in figure 23. The bandwidth of both circuits can also be confirmed when $C = 15 \text{ pF}$.

Implementation of the first and second oscillators can be obtained by CMOS complementary pairs MC14007 on a bread-board discrete circuit, as shown in figure 24(a) and (b) respectively. Capacitors of 0.01 μF are used and the waveforms at different points are shown in figure 25(a) and (b), using 20 μs/div and 10 μs/div.

![Figure 16. Second oscillation frequencies with current $I$ varied.](image)

![Figure 17. Different points sinusoidal signal of second oscillator with $C = 10 \text{nF}$, $I = 200 \text{μA}$, $f = 4.7 \text{kHz}$.](image)
Third-order oscillators using CMOS OTA

Figure 18. PSpice MOS model for the proposed oscillators.

Figure 19. Simulation frequency spectrum of first oscillator with $f = 1.89 \text{ MHz}$, $C = 15 \text{ pF}$, $I = 200 \mu\text{A}$.

Figure 20. Simulation frequency spectrum of second oscillator with $f = 197 \text{ MHz}$, $C = 20 \text{ pF}$, $I = 200 \mu\text{A}$.
respectively. The frequency output of both first and second oscillators is quite high, being a pure sine wave of about 20 kHz and 30 kHz respectively.

8. Conclusions

This paper presents new approaches to the sinusoidal oscillator based on the third-order principle. The authors do not see a need for any resistors suitable for further fabrication on chip. These oscillators have been used at ±3 V with simple configurations. The output frequency can be adjusted by the current $I$. The experimental results are confirmed by PSpice with the level 2 European Silicon Structure model. The highest oscillating frequency is about 2 MHz. The hardware circuit can be confirmed by CMOS complementary pairs, MC14007. Its oscillating frequency outputs are quite good.
### Table 1. Total harmonic distortion of (a) first oscillator, (b) second oscillator.

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<th>THD</th>
<th>$I = 10 \mu A$</th>
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<th>$I = 150 \mu A$</th>
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<td>$C = 15 \text{ pF}$</td>
<td>$-25.67 \text{ dB}$</td>
<td>$-29.7 \text{ dB}$</td>
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Table 2. Comparison of proposed oscillator with those of previous papers.

Figure 23. Output voltage of proposed oscillators against varying $I$ when $C = 15 \text{ pF}$.

Figure 24. Implementation of (a) first oscillator, (b) second oscillator.
References


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*Figure 25. Nodes 4 and 7 output waveform of (a) first oscillator, (b) second oscillator.*