One input voltage and three output voltage universal biquad filters with orthogonal tune of frequency and bandwidth

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ABSTRACT
This research paper contributes the one input three output voltage mode universal biquad filters with linear and electronic control of the natural frequency ($\omega_0$), using two commercially available ICs, LT1228s as active device with two grounded capacitors, five resistors. The presented universal biquad filters can simultaneously provide three voltage-mode filtering functions, low-pass (LP), high-pass (HP) and band-pass (BP) without changing the circuit architecture. Furthermore, the first presented biquad filter provides low impedance at HP, BP voltage output nodes and LP, BP output voltage nodes are low impedance for the second proposed filter which is easy cascade ability with other voltage mode circuits without the employment of buffer circuits. The quality factor ($Q$) of both proposed filters is orthogonally adjusted from the passband voltage gain and $\omega_0$. The proposed filters are simulated and experimented with commercially accessible ICs, LT1228. The simulated and experimental results demonstrate the filtering performances.

Keywords:
Active building block
LT1228
SIMO
Universal filter

INTRODUCTION
The analog filters or some likeness thereof are basic to the operation of the most engineering circuits. It is hence considering a logical concern for anybody handled with electronic circuit design to be able to create filter circuits fit for meeting a given arrangement of specifications. They are emphasized signals in the specified frequency band (wanted signal) while reject other the signals outside that band (unwanted signal). They are broadly utilized for their significant prerequisites in electrical and electronic engineering application and extremely famous in utilizing for a circuit framework of analog signal processing system. Numerous fields utilize filter circuits, for example, communication, instrumentation, measurement, sound system, and control systems [1]. Particularly, several function filters which are called universal filters or multifunction filters have been extensively studied and have become an interesting research topic. The analog biquad filter with single-input multiple-output (SIMO) category is the most renowned filtering circuit which provides numerous output responses in the same circuit topology. Moreover, SIMO filtering scheme does not want other circuits to get the wanted filtering responses for example the input signal selection. With SIMO feature, the single input is employed and provides simultaneously multiple filtering output responses.
Thus, SIMO type filters are more beneficial in comparison with other type filters for the applications which require simultaneously several filtering responses such as in 3-way crossover network [2-5].

The operational transconductance amplifier (OTA) is active building block which has received great attention to employ for synthesis of analog signal processing systems [6]. With electronic adjustability of transconductance gain, the parameters of the OTA based circuits are electronically controllable. Varied commercial OTA ICs are accessible for instance CA3080 [7] and LM13700 [8], and LT1228 [9]. For specific work with real practical test, the use of commercially available OTA is still cheaper and more appropriate compared to the chip fabrication [10]. The LT1228 from Linear Technology Inc is the intriguing one. This active device consists of OTA and current feedback amplifier (CFA). The LT1228’s CFA has high input impedance and so it is a superb buffer for the output of the transconductance amplifier maintains its wide bandwidth over a large vary of voltage gains creating it simple to interface the transconductance amplifier output to different electronic circuitry. It is intended to drive low impedance loads, such as cables, with better linearity at high frequencies. Moreover, the external dc bias current, \( I_b \) can control electronically to its transconductance \( g_m \). Therefore, the LT1228 based circuits are controlled by computer in trendy circuit vogue.

Several SIMO second order voltage mode universal filters using different analog active function blocks have been published in the literature [11-47]. However, these reposed filters have some inconveniences as it has been observed that the active building blocks are not implemented from the commercially available IC [13-31, 36-43]; all capacitors are not grounded [14, 18, 29, 42], the natural frequency is not electronically tuned [14-34, 36], absence of orthogonal tune of the \( \omega_0 \) and \( Q \) [13-16, 18-24, 29, 30, 35, 36, 40, 41, 43, 46, 47], the passband voltage gain is not constant during tuning the natural frequency and quality factor [14, 17, 18, 20, 25, 26, 28, 29, 32-34, 36, 38, 39, 42].

This current paper’s commitment is to propose two universal second order filters with one input voltage and three output voltages. The presented filters are constructed from two LT1228s with five resistors and two grounded capacitors. Three voltage mode filtering functions, HP, LP and BP with constant passband gain during tuning the \( \omega_0 \) and \( Q \) are obtained. The \( \omega_0 \) is electronically adjusted. The \( Q \) or bandwidth is controlled by means of setting the value of the feedback resistors while no disturbing the \( \omega_0 \). The exhibition of the proposed filters is approved through utilizing P spice program and trial.

2. PROPOSED UNIVERSAL FILTERS

2.1. Principle of the used commercially available active element

The active element used in this design is LT1228, commercially available IC from Linear Technology Inc. The LT1228 symbolic diagram with eight terminals is illustrated in Figure 1(a). The \( y \) terminal is output current which has high impedance and the \( \pm \) terminals are inverting and non-inverting input voltages that also have high impedance. The \( x \) and \( w \) terminals are voltage output which have low impedance. Terminal relationships of LT1228 can be shown in hybrid matrix form given in (1).

\[
\begin{pmatrix}
I_{v_{x}} \\
I_{v_{-}} \\
I_f \\
V_x \\
V_y \\
V_n
\end{pmatrix} =
\begin{pmatrix}
0 & 0 & 0 & 0 & 0 & V_x \\
0 & 0 & 0 & 0 & 0 & V_y \\
g_m & g_m & 0 & 0 & 0 & V_y \\
0 & 0 & 1 & 0 & 0 & I_x \\
0 & 0 & 0 & R_f & 0 & I_n
\end{pmatrix}
\]

where \( R_f \) is the internal trans resistance gain and it is infinity in an ideal case. The LT1228’s \( g_m \) is controlled electronically via external DC bias current (\( I_b \)) as (2):

\[
g_m = \frac{I_b}{3.87V_T}
\]

where \( V_T \) is thermal voltage. As defined in (1), the LT1228 characteristics can be represented as equivalent scheme in Figure 1(b). The pin connection of LT1228 is illustrated in Figure 1(c).

2.2. First proposed voltage-mode universal biquad filter using two LT1228s

The proposed biquad filters in Figure 2 are synthesized from two loop integrators. Figure 2(a) illustrates the first proposed universal voltage-mode second order filter comprised of two LT1228s, two grounded capacitors and five resistors. The first lossless integrator is constructed from the capacitor \( C_1 \) and first LT1228 while, the second lossless integrator is realized from capacitor \( C_2 \) and second LT1228. The
resistors \( R_1, R_2, R_3 \) and first LT1228 are constructed as voltage summing circuit. Finally, the resistors \( R_4, R_5 \) and second LT1228 are constructed as voltage amplifier. The input voltage \( (V_{in}) \) is applied at the voltage summing circuit while, the high-pass (HP) filtering voltage node \( (V_{HP1}) \) is at the output of the voltage summing, the low-pass (LP) filtering voltage node \( (V_{LP1}) \) is at the output of first integrator and the band-pass (BP) filtering voltage node \( (V_{BP1}) \) is at the output of voltage amplifier. With this scheme, three second order filtering functions are simultaneously given without changing the filtering construction. Moreover, it is found that the low output impedance at output voltage nodes \( V_{HP1} \) and \( V_{BP1} \) responses is achieved.

![LT1228 Symbolic Diagram](image1)

![LT1228 Pin Connection](image2)

Figure 1. LT1228, (a) Symbolic diagram, (b) Representation, (c) Pin connection

![Proposed Universal Filters](image3)

Figure 2. The proposed universal filters, (a) First filter, (b) Second filter

If \( R_1 = R_2 = R_3 \), the voltage transfer functions (TF) are expressed as (3):

\[
\begin{align*}
V_{\text{HP1}} &= \frac{-s^2}{s^2 + \frac{Kg_{m2}g_{m2}}{C_2}} \cdot V_{\text{in}} \\
V_{\text{LP1}} &= \frac{\frac{g_{m1}g_{m2}}{C_1C_2}}{s^2 + \frac{Kg_{m2}g_{m2}}{C_2}} \cdot V_{\text{in}} \\
V_{\text{BP1}} &= \frac{-sKg_{m2}}{s^2 + \frac{Kg_{m2}g_{m2}}{C_2} + \frac{3g_{m1}g_{m2}}{C_1C_2}} \cdot V_{\text{in}} \\
V_{\text{BP2}} &= \frac{-sKg_{m2}}{s^2 + \frac{Kg_{m2}g_{m2}}{C_2} + \frac{3g_{m1}g_{m2}}{C_1C_2}} \cdot V_{\text{in}}
\end{align*}
\]  

(3)

where \( K = 1 + \left( \frac{R_4}{R_5} \right) \). From (3), the unity passband voltage gain for high-pass and band-pass functions, the one-third passband voltage gain for low-pass function are obtained. Also from them, the high-pass and band-
pass functions are inverting, while the low-pass function is non-inverting. The natural frequency \((\omega_{0l})\) of the first filter as well as the quality factor \((Q_l)\) are obtained as (4):

\[
\omega_{0l} = \sqrt{\frac{3g_{ml}g_{m2}}{C_1C_2}} \quad \text{and} \quad Q_l = \frac{1}{K} \sqrt{\frac{3g_{ml}C_1}{g_{m2}C_2}} \tag{4}
\]

Substituting the transconductances \(g_{ml}\) and \(g_{m2}\) in function of \(I_{B1}\) and \(I_{B2}\) as appeared in (2) into (4), the \(\omega_{0l}\) and \(Q_l\) of the first proposed circuit are as (5):

\[
\omega_{0l} = \frac{1}{3.87V_i} \sqrt{\frac{3I_{B1}I_{B2}}{C_1C_2}} \quad \text{and} \quad Q_l = \frac{R_s}{R_s+R_i} \sqrt{\frac{3I_{B1}C_1}{I_{B2}C_2}} \tag{5}
\]

From (5), it can be remarked that \(\omega_{0l}\) can be electronically controlled by \(I_{B1}\) and \(I_{B2}\). Moreover, the \(Q_l\) can be altered without affecting \(\omega_{0l}\) by changing the value of \(R_s\) (as stated in data sheet of LT1228 [10], \(R_s\) should not be adjusted to keep the constant bandwidth). Other advantages for tuning filter parameter can be achieved by simultaneously adjusting \(I_{B1}=I_{B2}=I_B\) (this feature is easily implemented by using microcontroller or microcomputer) and setting \(C_1=C_2=C\). Then filtering parameters in (5) becomes

\[
\omega_{0l} = \frac{I_B}{3.87V_i} \sqrt{3} \quad \text{and} \quad Q_l = \frac{R_s}{R_s+R_i} \sqrt{3} \tag{6}
\]

It can be remarked from (6) that \(\omega_{0l}\) and \(Q_l\) are independently controlled. Moreover, the \(\omega_{0l}\) can be linearly and electronically adjusted. Also, the control of \(\omega_{0l}\) and \(Q_l\) doesn’t affect the passband voltage gain for all filtering responses.

2.3. Second proposed voltage-mode universal biquad filter using two LT1228s

The second proposed biquadratic filter is also synthesized from two loop integrators as explained above. Figure 2(b) illustrates the scheme of the second proposed filter. It is comprised of two LT1228s, five resistors, two grounded capacitors which is same to first proposed filter. The first lossless integrator is constructed from the capacitor \(C_1\) and first LT1228 while, the second lossless integrator is realized from capacitor \(C_2\) and second LT1228. The resistors \(R_1, R_2, R_3\) and second LT1228 are constructed as voltage summing circuit. Finally, the resistors \(R_4, R_5, R_6\) and first LT1228 are constructed as voltage amplifier. The input voltage \((V_{in})\) is applied at the voltage summing circuit while, the high-pass (HP) filtering voltage node \((V_{HP2})\) is at the output of the voltage summing, the low-pass (LP) filtering voltage node \((V_{LP2})\) is at the output of first integrator and the band-pass (BP) filtering voltage node \((V_{BP2})\) is at the output of voltage amplifier. With this scheme, three other order filtering functions are simultaneously given without changing the filtering scheme. Moreover, it is found that the low output impedance at output voltage nodes \(V_{BP2}\) and \(V_{HP2}\) responses is achieved. If \(R_7 = R_8 = R_t\), the voltage transfer functions (TF) are expressed as (7):

\[
\frac{V_{HP2}}{V_{in}} = \frac{s^2 + \frac{3K_{m1}}{3C_1} + \frac{g_{m1}g_{m2}}{3C_1C_2}}{s^2 + \frac{3K_{m1}}{3C_1} + \frac{g_{m1}g_{m2}}{3C_1C_2}} ; \quad \frac{V_{LP2}}{V_{in}} = \frac{\frac{g_{m1}g_{m2}}{3C_1C_2}}{s^2 + \frac{3K_{m1}}{3C_1} + \frac{g_{m1}g_{m2}}{3C_1C_2}} ; \quad \frac{V_{BP2}}{V_{in}} = \frac{sK_{m1}}{3C_1} \tag{7}
\]

where \(K = 1 + (R_4/R_5)\). From (7), the unity passband voltage gain for low-pass and band-pass functions, the one-third passband voltage gain for high-pass function are obtained. Also from them, the low-pass and band-pass functions are inverting, while the high-pass function is non-inverting. The natural frequency \((\omega_{02})\) of the second filter as well as the quality factor \((Q_2)\) are obtained as (8):

\[
\omega_{02} = \sqrt{\frac{g_{m1}g_{m2}}{3C_1C_2}} \quad \text{and} \quad Q_2 = \frac{1}{K} \sqrt{\frac{3g_{m2}C_1}{g_{m1}C_2}} \tag{8}
\]

Substituting the transconductances \(g_{m1}\) and \(g_{m2}\) in function of the bias currents, \(I_{B1}\) and \(I_{B2}\) as appeared into (2) into (8), the \(\omega_{02}\) and \(Q_2\) of the second proposed circuit are as (9).
\[ \omega_0 = \frac{1}{3.87V_i} \sqrt{\frac{I_{B1}I_{B2}}{3C_1C_2}} \quad \text{and} \quad Q_2 = \frac{R_3}{R_1 + R_2} \sqrt{\frac{3I_{B2}C_1}{I_{B2}C_2}} \]  

From (9), it is found that \( \omega_0 \) can be electronically controlled by \( I_{B1} \) and \( I_{B2} \). Moreover, the \( Q_2 \) can be altered without affecting \( \omega_0 \) by changing the value of \( R_5 \). Other advantage for tuning filter parameter can be achieved by simultaneously adjusting \( I_{B1} = I_{B2} = I_B \) (this feature is easily implemented by using microcontroller or microcomputer) and setting \( C_1 = C_2 = C \). Then filtering parameters in (9) becomes;

\[ \omega_0 = \frac{I_B}{3.87V_i} \sqrt{\frac{I_{B2}}{C_1C_2}} \quad \text{and} \quad Q_2 = \frac{R_3}{R_1 + R_2} \sqrt{3} \]  

It can be noted from (10) that \( \omega_0 \) and \( Q_2 \) are independently tuned. Moreover, the \( \omega_0 \) is linearly and electronically tuned. Also, the control of \( \omega_0 \) and \( Q_2 \) does not affect the passband gain for all filtering responses.

3. PARASITIC EFFECTS

In a genuine application, the impacts of parasitic components in LT1228 are not ignorable that it might influence the performance of the proposed filters. So, the parasitic elements appeared at all terminals of LT1228 will be considered. These parasitic elements are as follows: \( R \) and \( C \) appear in parallel at \( V \) terminal to ground; \( R_n \) and \( C_n \) appear in parallel at \( V \) terminal to ground; \( R_t \) and \( C_t \) appear in parallel at \( y \) terminal to ground; \( R_s \) appears in series at \( x \) terminal; \( R_a \) appears in series at \( w \) terminal and the internal transresistance gain is considered as \( R_{int}/C_t \). As stated in data sheet of LT1228 [10], the feedback resistors \((R_t, R_d \text{ in first filter and } R_t \text{ in second filter})\) from \( w \) to \( x \) terminal in summing and amplifier circuits should be low to reduce the effect of \( C_t \) and \( R_t \) and to get higher operating frequency. Also, if the operational frequency of the proposed filters is expected lower than 10 MHz, the most effect stems from \( R_t, C_t, R_s, C_t, R_l \) and \( C_t \) (the effect of \( R_n, R_o, R_t \) and \( C_t \) is ignore). In consideration of these parasitic elements, the three voltage transfer functions \((V_{HP1}, V_{LP1} \text{ and } V_{BP1})\) of the first presented biquad filter realized in Figure 2(a) are obtained as (11).

\[
\frac{V'_{HP1}}{V_{in}} = -s^2 \left( \frac{G_{11}C_1' + G_{12}C_2'}{C_1'C_2'} \right) s + \frac{G_{21}G_{22}}{C_1'C_2'} ; \quad \frac{V'_{LP1}}{V_{in}} = \frac{G_{m1s}G_{m2}}{C_1'C_2'} - \frac{K_{g1s}G_{s1}}{C_1'C_2'}, \quad \frac{V'_{BP1}}{V_{in}} = \frac{-K_{g2s}G_{s2} - G_{s2}}{C_3'}, \quad \frac{D'(s)}{s} \]

where

\[
D'(s) = s^2 + \left( \frac{G_{11}C_1' + G_{12}C_2'}{C_1'C_2'} \right) s + \frac{G_{21}G_{22} + G_{m1s}G_{m2} + K_{g2s}G_{s1}}{C_1'C_2'} \]

\[C_1' = C_1 + C_{1s}, \quad C_2' = C_2 + C_{2s} + C_{4s}, \quad G_{s1} = \frac{1}{R_{1s}}, \quad \text{and} \quad G_{s2} = \frac{1}{R_{2s}} + \frac{1}{R_3} \]

From (12), the non-ideal natural frequency and quality factor of the first proposed filter become;

\[ \omega_{01} = \sqrt{\frac{G_{21}G_{22} + G_{m1s}G_{m2} + K_{g2s}G_{s1}}{C_1'C_2'}} \quad \text{and} \quad Q_1 = \frac{\sqrt{C_1'C_2'} \left( G_{s1}G_{s2} + 3G_{m1s}G_{m2} + K_{g2s}G_{s1} \right)}{G_{s1}C_1' + G_{s2}C_2' + K_{g2s}C_1'} \]

In consideration of these parasitic elements, the three voltage transfer functions \((V_{HP2}, V_{LP2} \text{ and } V_{BP2})\) of the second presented biquad filter realized in Figure 2(b) are obtained as (14).

\[
\frac{V_{HP2}}{V_{in}} = \frac{\frac{s^2}{3} \left( \frac{G_{11}C_1' + G_{12}C_2'}{C_1'C_2'} \right) s + \frac{G_{21}G_{22}}{3C_1C_2'}}{D''(s)} ; \quad \frac{V_{LP2}}{V_{in}} = \frac{\frac{-G_{m1s}G_{m2}}{3C_1C_2'}}{D''(s)} ; \quad \frac{V_{BP2}}{V_{in}} = \frac{\frac{K_{g1s}G_{s1}}{3C_1C_2'} - \frac{K_{g2s}G_{s2}}{3C_1C_2'}}{D''(s)} \]

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where

\[ D^+(s) = s^2 + \left( \frac{3G_1 C_2 + G_{s1} C_{s1} + K_{g1} C_{s2}^{*}}{3C_1 C_2} \right) + \left( \frac{3G_1 C_2^{*} + G_{s1} C_{s1} + K_{g1} C_{s2}^{*}}{3C_1 C_2} \right) \]

\[ C_1^{*} = C_1 + C_{s1} + C_{s2}, \quad C_2^{*} = C_2 + C_{s1}, \quad G_1^{*} = \frac{1}{R_{s1}} + \frac{1}{R_{s2}} \quad \text{and} \quad G_{s2} = \frac{1}{R_{s2}}. \]

From (15), the non-ideal natural frequency and quality factor of the second proposed filter become;

\[ \omega_{\omega_{2}} = \sqrt{\frac{3G_1 C_2 + K_{g1} G_{s1} + g_{m1} s_{m2}}{3C_1 C_2}} \quad \text{and} \quad Q_2 = \sqrt{\frac{3G_1 C_2^{*} (3G_1 C_2^{*} + g_{m1} s_{m2} + K_{g1} G_{s1})}{3(C_1^{*} C_2^{*} + G_{s1} C_{s1}^{*}) + K_{g1} C_{s2}^{*}}} \]  

(16)

4. SIMULATION RESULTS

To confirm the theoretical ability of the filtering design, the functionality of the presented circuits illustrated in Figure 2 was tested by using PSPICE. The supply voltages were \( V_{CC} = V_{EE} = 5 \, \text{V} \). For the first proposed filter, the bias current was set to \( I_B1 = I_B2 = 123.5 \, \mu\text{A} \). The values of passive component were selected as \( R_1 = R_2 = R_3 = 1 \, \Omega \), \( C_1 = 1 \, \text{nF} \) and \( C_2 = 10 \, \text{nF} \). Using above mentioned device values, the calculated natural frequency and quality factor from (5) is obtained as \( \omega_0 = 107.66 \, \text{kHz} \), \( Q_1 = 2.73 \). The simulation results are obtained as \( f_0 = 105.682 \, \text{kHz} \), \( Q = 2.82 \). The deviation of the theoretical and simulated \( f_0 \) is 1.84%, \( Q \) is 3.29%. There simulated filtering responses of the first presented versatile second order filter are illustrated in Figure 3. The high-pass filtering response with different values of resistor \( R(R_1 = R_2 = R) \) in simming circuit of the first presented biquad filter is illustrated in Figure 4. In this simulation, the values of \( R \) were set to 0.2 \( \Omega \), 0.6 \( \Omega \) and 5 \( \Omega \). It is found that with low value of the feedback resistor \( R \), the bandwidth of the proposed filter is higher than the bandwidth at high value of \( R \) as expected in section of parasitic effect.

![Figure 3. The simulation of gain response of the presented versatile filter in Figure 2(a)](image)

![Figure 4. High-pass response of the filter in Figure 2(a) with different values of \( R(R_1 = R_2 = R) \)](image)

Figure 5 confirms that \( Q_1 \) can be controlled by varying the value of resistance \( R_3 \) without affecting \( f_01 \) as expected in (7), where \( R_2 \) is assigned to 0.2 \( \Omega \), 0.6 \( \Omega \), 5 \( \Omega \). The electronic tune of the natural frequency by simultaneously changing \( I_B1 \) and \( I_B2 \) \((I_B1 = I_B2 = I_B)\) is shown in Figure 6 where value of \( I_B \) was set to 60 \( \mu\text{A} \), 120 \( \mu\text{A} \) and 240 \( \mu\text{A} \). The natural frequency tuned from these bias currents are located at 57.28 kHz, 102.80 kHz and 205.59 kHz. The result in Figure 6 indicates that the natural frequency can be linearly and electronically tuned by the \( I_B \) with constant value of the \( Q_1 \) as expected in (8). The second proposed filter was also simulated with the same voltage supplies and resistance values of the first filter. Other elements were set as follows: \( I_B1 = I_B2 = 350 \, \mu\text{A} \) and \( C_1 = 10 \, \text{nF} \), \( C_2 = 1 \, \text{nF} \). Using these mentioned component values, the theoretical \( f_{02} \) and \( Q_2 \) from (13) are given as \( f_{02} = 101.70 \, \text{kHz} \), \( Q_2 = 2.73 \). The simulation results are obtained as \( f_{02} = 99.77 \, \text{kHz} \), \( Q_2 = 2.88 \). The deviation of the theoretical and simulated \( f_{02} \) is 1.89% and \( Q_2 \) is 5.49%. The simulated three filtering responses are shown in Figure 7. The result in Figure 8 confirms that \( Q_2 \) can be controlled by varying the value of resistance \( R_3 \) without affecting \( f_{02} \), where \( R_3 \) is assigned to 0.2 \( \Omega \), 0.6 \( \Omega \), 5 \( \Omega \). It is observed that \( Q_2 \) can be controlled without affecting the \( f_{02} \). It is the fact that the BJT OTA has linear range when the input amplitude is below 50 mV. To prove the linearity of presented universal biquad filters, the plot of percent of total harmonic distortion (THD) versus input voltage amplitude for band-pass response is illustrated in Figure 9. Both presented filters were designed to achieve the natural frequency of 100 kHz. It can be seen from Figure 9 that the linearity of the second presented...
voltage-mode biquad filter is superior to the first presented biquad filter. It is seen from Figures 2(a) and 2(b) that the band-pass responses of both proposed filters are at the output of the lossless integrators which have the high-pass response as input voltage. The second presented filter’s high-pass voltage gain is one third while the first presented filter’s high-pass voltage gain is unity so this value of $V_{HP1}$ causes the higher distortion than $V_{HP2}$ at above 50 mV of the input voltage signal.

![Figure 5. The BP gain response of the filter in Figure 2(a) with different values of $R_5$](image)

![Figure 6. The BP gain response of the filter in Figure 2(a) with different values of $I_B$ ($I_{B1}=I_{B2}=I_B$)](image)

![Figure 7. The simulation of gain response of the presented versatile filter in Figure 2(b)](image)

![Figure 8. The gain response of the filter in Figure 2(b) with different values of $R_1$](image)

![Figure 9. The percent THD against input magnitude](image)

5. EXPERIMENTAL RESULTS

To assess the exhibitions of the presented universal filters in Figure 2, the experiments were also tested by utilizing two LT1228s. An experimental setup was made by taking $V_{CC}=-V_{EE}=5$ V using GW Instek GPS-3303 power supply, $R_1=R_2=R_3=R_4=R_5=1$ kΩ for both filters, $C_1=1$ nF, $C_2=10$ nF, $I_{B1}=I_{B2}=123.5$ μA for first presented filter and $C_1=10$ nF, $C_2=1$ nF, $I_{B1}=I_{B2}=365$ μA for second filter. The plot of experimental magnitude response (measured from Keysight DSOX1102G oscilloscope) of the first and second presented filters is shown in Figures 10(a) and 10(b), respectively. It is found that the experimental natural frequencies
of the first and second proposed filters are $f_{o1}=114$ kHz and $f_{o2}=109$ kHz, respectively. Figure 11 confirms that $Q_1$ and $Q_2$ can be controlled by varying the value of resistance $R_5$ without affecting $f_{o1}$ as expected in (5) and (9), where $R_5$ is assigned to 0.47 kΩ, 1 kΩ, 3 kΩ. The electronic tune of the natural frequency, $f_{o1}$ by simultaneously changing $I_{B1}$ and $I_{B2}$ ($I_{B1}=I_{B2}=I_B$) for the first presented universal biquad filter is illustrated in Figure 12(a) where value of $I_B$ was set to 67 µA, 123.5 µA and 245 µA. The experimental natural frequency tuned from these bias currents are located at 60 kHz, 114 kHz and 230 kHz. For the second proposed filter, the electronic tune of the natural frequency, $f_{o2}$ by simultaneously changing $I_{B1}$ and $I_{B2}$($I_{B1}=I_{B2}=I_B$) is shown in Figure 12(b) where value of $I_B$ was set to 184 µA, 365 µA and 722 µA. The experimental natural frequency tuned from these bias currents are located at 55 kHz, 114 kHz and 220 kHz. The result in Figure 12 indicates that the natural frequency of both proposed filters can be linearly and electronically tuned by the bias current without affecting the quality factor as expected in (6) and (10). The measurement of the band-pass response for the first presented versatile second order filter is illustrated in Figure 13, where the 50 mV$_{p-p}$ sine wave signals with frequencies, 10 kHz, 115 kHz and 1 MHz was input signal. The comparison of proposed universal biquad filter and other voltage-mode SIMO filtering configurations is shown in Table 1 (see in Appendix). It is found from Table 1 that with two commercially available ICs with grounded capacitors, the $\omega_0$ of proposed filters can be electronically tuned. Also, both filters can provide orthogonal control $\omega_0$ and $Q$. Additionally, the constant passband gain during tuning $\omega_0$ and $Q$ for all responses is achieved.

![Figure 10](image1.png)

Figure 10. The theoretical and experimental frequency response of the proposed filters; (a) Figure 2(a), (b) Figure 2(b)

![Figure 11](image2.png)

Figure 11. The experimental BP gain response of the proposed filters with different values of $R_5$; (a) Figure 2(a), (b) Figure 2(b)
Figure 12. The experimental BP gain response of the proposed filters with different values of $I_B$.
(a) Figure 2(a), (b) Figure 2(b)

Figure 13. The measurement of the band-pass response for the first proposed filter, (a) $f = 10kHz$, (b) $f = 115kHz$, (c) $f = 1MHz$

6. CONCLUSION

In this article, the one input voltage and three output voltages universal biquad filters constructed two LT1228s, five resistors and two grounded capacitors. The proposed circuits are synthesized from two loop integrator topologies. They can simultaneously provide three voltage-mode biquad filtering transfer functions, LP, HP and BP. The availability of orthogonal adjustability of the quality factor and natural frequency is achieved. Also, the natural frequency is electronically and linearly tuned. The effect of non-ideal properties of LT1228 on the performances of the presented filters are considered and studied. The performances of the presented filters have been investigated by simulation and experimental results using commercially available LT1228 with ±5 V supply voltages and the results confirm the theoretical propositions.
## APPENDIX

Table 1. Comparison of proposed universal filter and other voltage-mode SIMO filtering configurations

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<th>Ref.</th>
<th>ABB</th>
<th>No. of ABB</th>
<th>No. of commercial ICs</th>
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<th>Filtering responses</th>
<th>No. of low Zc nodes</th>
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<th>Electronical gain during tuning $\varepsilon_{lo}$ and $Q$ for all responses</th>
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ACKNOWLEDGEMENTS

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One input voltage and three output voltage universal biquad filters with orthogonal ... (May Phu Pwint Wai)
REFERENCES


One input voltage and three output voltage universal biquad filters with orthogonal ... (May Phu Pwint Wai)