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Isomorphic Circuits of Independent Amplitude Tunable Voltage-Mode Bandpass Filters and Quadrature Sinusoidal Oscillators

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Abstract: This paper presents isomorphic circuits of voltage-mode (VM) non-inverting bandpass filters (NBPFs) and VM quadrature sinusoidal oscillators (QSOs) with independent amplitude control functionality. The proposed VM NBPFs and VM QSOs exhibit low-output impedance and independent amplitude control, which are important for easily cascading the VM operation and independent control of the amplitude gain. The proposed isomorphic circuits employ three LT1228 commercial integrated circuits (ICs), two grounded capacitors, two grounded resistors and one floating resistor. The use of grounded capacitors is beneficial for the implementation of the IC. Both NBPFs have a high-input impedance and have a wide range of independent amplitude tunable passband gain without affecting the quality factor (Q) and center frequency (f_o). The Q and f_o parameters of the proposed NBPFs are orthogonal tunability. By feeding back each input signal to the output response of the NBPF, two VM fully uncoupled QSOs are also proposed. The proposed VM fully uncoupled QSOs have two quadrature sinusoidal waveforms with two low-output impedances and one independent amplitude tunable sinusoidal waveform. The frequency of oscillation (FO) and the condition of oscillation (CO) are fully uncoupled and controlled electronically. The performances of the proposed isomorphic circuits have been tested with a ± 5 volt power supply and are demonstrated by experimental measurements which confirm the theoretical assumptions.

Keywords: analog circuit design; non-inverting bandpass filter; fully uncoupled quadrature oscillator; LT1228

Citation: Wang, S.-F.; Chen, H.-P.; Ku, Y.; Chen, W.-Y. Isomorphic Circuits of Independent Amplitude Tunable Voltage-Mode Bandpass Filters and Quadrature Sinusoidal Oscillators. *Appl. Sci.* **2021**, *11*, 7431. <https://doi.org/10.3390/app11167431>

Academic Editors:
Wen-Hsiang Hsieh, Jia-Shing Sheu
and Minvydas Ragulskis

Received: 26 July 2021

Accepted: 10 August 2021

Published: 12 August 2021

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1. Introduction

The LT1228 is a commercial integrated circuit (IC) using bipolar or complementary metal oxide semiconductor (CMOS) technology. It combines a transconductance amplifier (OTA) and a current feedback amplifier (CFA). LT1228 is one of the attractive components used for realizing voltage-mode (VM) analog circuits. It has wider electronic tunability, higher accuracy, higher frequency applicability and simplicity of implementation, so many topologies based on LT1228 application circuits have been published in the literature [1–8]. In addition, many active filters and oscillators based on different active components have also been presented in the literature, and they have been widely used in signal generation applications [9–20].

An active VM non-inverting bandpass filter (NBPF) is one of the important filters for analog signal processing, and its transfer function can be expressed as

$$\frac{V_{\text{NBPF}}}{V_{\text{in}}} = \frac{ds}{as^2 + bs + c} \quad (1)$$

where the coefficients a , b , c and d are real numbers, V_{in} is the NBPF input signal and V_{NBPF} is the output response. According to the feedback theory, the VM quadrature sinusoidal oscillator (QSO) can be synthesized by connecting the input voltage of the NBPF to the output response. Therefore, Equation (1) can be rewritten as Equation (2):

$$[as^2 + (b - d)s + c]V_{NBPF} = 0 \quad (2)$$

Based on Equations (1) and (2), the VM QSO can be realized by zeroing the input voltage of the NBPF, and the characteristic equation (CE) of the VM QSO can be expressed as Equation (3):

$$CE : as^2 + (b - d)s + c = 0 \quad (3)$$

The two poles of Equation (3) are obtained as follows:

$$s_{1,2} = \frac{-(b - d) \pm \sqrt{(b - d)^2 - 4ac}}{2a} = \frac{(d - b)}{2a} \pm j\sqrt{\frac{c}{a} - \left(\frac{d - b}{2a}\right)^2} \quad (4)$$

In Equation (3), the oscillator will oscillate if the condition of oscillation (CO) and the frequency of oscillation (FO) are as given by Equations (5) and (6), respectively:

$$CO : b = d \quad (5)$$

$$FO : \omega_O^2 = \frac{c}{a} \quad (6)$$

For a positive feedback system, if the two poles of Equation (4) are located in the right half of the complex plane, the circuit can be built, and the sinusoidal oscillation can be maintained. Therefore, the CO of Equation (5) is modified to become Equation (7):

$$CO : b \leq d \quad (7)$$

Recently, two interesting VM circuits were proposed in the literature [21,22]. The circuit proposed in [21] employs five OTAs combined with two grounded capacitors. The circuit proposed in [22] also employs five OTAs combined with two grounded capacitors. However, the passband gain of these two circuits cannot be adjusted independently without affecting the quality factor (Q) and resonance angular frequency (ω_o), and it cannot be transformed into an electronically controllable QSO with an independent amplitude control function [21,22]. An interesting VM QSO with gain controllability of the quadrature voltage output was recently proposed in [23]. The circuit in Figure 5 of [23] employs three LT1228s, two grounded capacitors, two grounded resistors and three floating resistors, but the resistances of the circuit in Figure 5 of [23] are not the smallest. According to the feedback theory, this study proposes isomorphic circuits of the VM NBPFs and VM QSOs with independent amplitude control functionality. Each of the isomorphic circuits contains three LT1228 commercial ICs, two grounded capacitors, two grounded resistors and one floating resistor. Table 1 lists the available circuits and compares them according to the relevant standards. The additional performance comparisons with previous circuits are summarized in Tables 2 and 3. As shown in Tables 1–3, the proposed VM NBPFs and VM QSOs exhibit high-input and low-output impedance and independent amplitude control, which are important for easily cascading the VM operation and independent control of the amplitude gain. The range of the measured QSO amplitude can be adjusted from 1.51 to 24.05, and the measured oscillation frequency can be varied from 82.15 kHz to 1629 kHz. This means that the proposed circuits are sufficient for the typical impedance sensor applications [23]. Because the proposed isomorphic circuits have independent amplitude control functions and a wide tunable frequency range, the applications will be beneficial in the literature [23,24].

Table 1. Comparison between recent electronic tunable filter/oscillator circuits.

Parameter	Figures 1 and 2 in [21]	Figures 1 and 4 in [22]	Figure 5 in [23]	First Proposed in This Work	Second Proposed in This Work
Number of active devices	5 (OTA)	5 (OTA)	3 (LT1228)	3 (LT1228)	3 (LT1228)
Number of passive elements	2 (2C)	2 (2C)	7 (2C, 5R)	5 (2C, 3R)	5 (2C, 3R)
Orthogonal tunability of ω_0 and Q	yes	yes	NA	yes	yes
High-input and low-output impedance of the filter	no	no	NA	yes	yes
Independent tunability of the filter amplitude	no	no	NA	yes	yes
Electronic and linear tune of oscillation frequency	no	yes	yes	yes	yes
Oscillator CO and FO with fully uncoupled electronic tuning law	no	yes	yes	yes	yes
Oscillator's amplitude tunability	no	no	yes	yes	yes

Note: Q: quality factor; ω_0 : resonance angular frequency; CO: condition of oscillation; FO: frequency of oscillation; and NA: not available.

Table 2. The additional performance comparisons with previous VM filters.

Parameter	Figure 1 in [21]	Figure 1 in [22]	First Proposed in This Work	Second Proposed in This Work
Supply voltage	± 2 V	± 15 V	± 5 V	± 5 V
Measured power dissipation	0.14 W	1.2 W	0.22 W	0.22 W
Designed center frequency	159.16 kHz	217 kHz	165.8 kHz	165.8 kHz
Measured maximum operating voltage gain	0 dB	0 dB	27.86 dB	27.58 dB
Measured output P1dB (dBm)	−14.6	NT	−11.4	−9.77
Measured IMD3 (dBc)	−42.86	−31.16	−40.1	−39.66
Measured third-order intercept point (dBm)	−5.47	NT	−7.95	−5.753
Measured phase noise (dBc/Hz)	NT	NT	−83.86	−73.87
Figure of merit	NT	NT	3.07×10^6	3.08×10^6

Note: P1dB: 1-dB compression point; IMD3: third-order intermodulation distortion; and NT: not tested.

Table 3. The additional performance comparisons with previous VM oscillators.

Parameter	Figure 2 in [21]	Figure 4 in [22]	Figure 5 in [23]	First Proposed in This Work	Second Proposed in This Work
Supply voltage	± 15 V	± 15 V	± 5 V	± 5 V	± 5 V
Measured power dissipation	NT	1.2 W	1.98 W	0.22 W	0.22 W
Measured total harmonic distortion (%)	NT	NT	<1.26	0.81	0.33
Measured frequency tunable range (kHz)	NT	150.1~2653	8.21~1117.51	82.15~1629	80.83~1626
Measured amplitude tunability range	NA	NA	1.97~15.92	1.51~24.05	1.51~24.03
Measured phase noise (dBc/Hz)	NT	−73.23 (@ 1000 Hz)	NT	−40.18 (@ 30 Hz)	−34.7 (@ 30 Hz)

Note: NA: not available; and NT: not tested.

2. The Proposed Isomorphic Circuits

Each of the isomorphic circuits' symbols of an LT1228 commercial IC from Linear Technology [25] is shown in Figure 1a, and its packed IC is shown Figure 1b. The LT1228 IC contains an OTA with a specially devised CFA. Figure 1c shows the equivalent circuit of

the LT1228 IC, and the ideal port relations of LT1228 can be characterized by the following hybrid matrix [23]:

$$\begin{bmatrix} I_{V+} \\ I_{V-} \\ I_y \\ V_x \\ V_w \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \\ g_m & -g_m & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & R_T & 0 \end{bmatrix} \begin{bmatrix} V_+ \\ V_- \\ V_y \\ I_x \\ I_w \end{bmatrix} \quad (8)$$

where g_m is the transconductance gain of the LT1228 IC and R_T is the transresistance gain of the CFA. Ideally, R_T approaches infinity, and the g_m is controlled by the direct current (DC) bias current (I_B) of LT1228. The g_m value is equal to 10 times the value of I_B and can be expressed as [23,25]

$$g_m = 10I_B = \frac{V_{DD} - V_{EE} - 2V_{BE}}{R_B} \quad (9)$$

where V_{DD} is the positive voltage supply and V_{EE} is the negative voltage supply. V_{BE} is a bias voltage of the bipolar junction transistor and is approximately equal to 0.65 volts. R_B is a bias control resistor used to obtain the DC bias current I_B . The proposed two isomorphic circuits are shown in Figures 2 and 3. Each of the circuits contains three LT1228 commercial ICs, two grounded capacitors, two grounded resistors and one floating resistor. The use of grounded capacitors is beneficial for the implementation of the IC. Each input voltage of the VM NBPf (Figures 2a and 3a) is applied to the positive voltage terminal of the LT1228, so it has a high-input impedance and can be easily cascaded without any voltage buffers. The output impedances of terminals V_{BP2} and V_{BP5} (Figures 2a and 3a, respectively) are very small, so they can also be directly connected to the next stage without any voltage buffers. Moreover, the VM NBPf responses of V_{BP3} and V_{BP6} shown in Figures 2a and 3a, respectively, can achieve a wide range of independent amplitude adjustable passband gains without affecting the parameters of f_o and Q .

Figure 1. LT1228. (a) Schematic symbol. (b) Pin. (c) Equivalent circuit.

Figure 2. The first proposed isomorphic circuit of the VM NBPf and VM QSO. (a) VM NBPf circuit. (b) VM QSO circuit.

Figure 3. The second proposed isomorphic circuit of the VM NBPf and VM QSO. (a) VM NBPf circuit. (b) VM QSO circuit.

Routine analysis of the VM NBPf circuit (Figure 2a) yielded the following three NBPf responses:

$$\frac{V_{BP1}}{V_{in}} = \frac{sC_1g_{m2}R_1}{s^2C_1C_2R_1 + sC_1 + g_{m1}g_{m3}R_1} \quad (10)$$

$$\frac{V_{BP2}}{V_{in}} = \frac{sC_1g_{m2}R_1}{s^2C_1C_2R_1 + sC_1 + g_{m1}g_{m3}R_1} \quad (11)$$

$$\frac{V_{BP3}}{V_{in}} = \left(\frac{sC_1g_{m2}R_1}{s^2C_1C_2R_1 + sC_1 + g_{m1}g_{m3}R_1} \right) \left(1 + \frac{R_2}{R_3} \right) \quad (12)$$

Based on the three NBPf responses from Equations (10)–(12), the NBPf parameters of f_o and Q are given by

$$f_o = \frac{1}{2\pi} \sqrt{\frac{g_{m1}g_{m3}}{C_1C_2}} \quad (13)$$

$$Q = R_1 \sqrt{\frac{C_2g_{m1}g_{m3}}{C_1}} \quad (14)$$

From Equations (13) and (14), the NBPf parameters of f_o and Q have orthogonal tunability by tuning the g_{m1} for f_o first and then the grounded resistor R_1 for Q without affecting parameter f_o .

The passband gains of the three NBPF responses are given by

$$H_{BP1} = H_{BP2} = g_{m2}R_1 \quad (15)$$

$$H_{BP3} = g_{m2}R_1 \left(1 + \frac{R_2}{R_3}\right) \quad (16)$$

From Equations (15) and (16), it is important to note that the passband gains of the NBPF responses are independently adjustable by g_{m2} without affecting the NBPF parameters of f_o and Q . Moreover, the resistors R_2 or R_3 in Equation (16) can achieve a wide range for the independent amplitude adjustable passband gain of the NBPF without affecting parameters f_o and Q . It is interesting to note that, as shown in Figure 2a, by feeding back the input signal to the output response of the NBPF, a VM fully uncoupled QSO can be realized as shown in Figure 2b. This means that the input signal V_{in} is connected to V_{BP1} (Figure 2a), and Equation (10) becomes

$$\left. \frac{V_{BP1}}{V_{in}} \right|_{V_{BP1}=V_{in}} = 1 = \frac{sC_1g_{m2}R_1}{s^2C_1C_2R_1 + sC_1 + g_{m1}g_{m3}R_1} \quad (17)$$

Therefore, the CE of Figure 2b is expressed as

$$s^2C_1C_2R_1 + sC_1(1 - g_{m2}R_1) + g_{m1}g_{m3}R_1 = 0 \quad (18)$$

Based on Equation (18), the CO and FO of the proposed VM QSO (Figure 2b) are given by Equations (19) and (20), respectively:

$$\text{CO: } g_{m2}R_1 \geq 1 \quad (19)$$

$$\text{FO: } f_o = \frac{1}{2\pi} \sqrt{\frac{g_{m1}g_{m3}}{C_1C_2}} \quad (20)$$

According to Equations (19) and (20), the CO can be fully controlled independently without affecting the FO by adjusting g_{m2} or R_1 , and the FO can be also fully controlled independently without affecting the CO by adjusting g_{m1} or g_{m3} . Hence, the VM QSO (Figure 2b) provides a fully uncoupled tuning law for the CO and FO. Routine analysis of the QSO structure, shown in Figure 2b, yielded the following two quadrature voltage outputs and a sinusoidal output waveform with independent amplitude control:

$$\left. \frac{V_{o1}}{V_{o2}} \right|_{s=j\omega_o} = -\frac{g_{m1}}{sC_1} \bigg|_{s=j\omega_o} = \frac{g_{m1}}{\omega_o C_1} e^{j90^\circ} = \sqrt{\frac{C_2g_{m1}}{C_1g_{m3}}} e^{j90^\circ} \quad (21)$$

$$V_{o3} = \left(1 + \frac{R_2}{R_3}\right) V_{o2} \quad (22)$$

According to Equation (21), the magnitude ratio and phasor of output voltages V_{o1} and V_{o2} are given as

$$\left| \frac{V_{o1}}{V_{o2}} \right| = \sqrt{\frac{C_2g_{m1}}{C_1g_{m3}}} \bigg|_{C_1=C_2, g_{m1}=g_{m3}} = 1 \quad (23)$$

$$\angle \frac{V_{o1}}{V_{o2}} = 90^\circ \quad (24)$$

Equations (23) and (24) show that the voltage magnitude ratio of V_{o1} and V_{o2} is in unity, and the voltage phase shift of V_{o1} and V_{o2} is 90° . Thus, the proposed first VM QSO has a unity voltage amplitude ratio and a quadrature voltage phase shift.

Similarly, routine analysis of the VM NBPf circuit (Figure 3a) yielded the following three NBPf responses:

$$\frac{V_{BP4}}{V_{in}} = \frac{sC_3g_{m2}R_4}{s^2C_3C_4R_4 + sC_3 + g_{m1}g_{m3}R_4} \quad (25)$$

$$\frac{V_{BP5}}{V_{in}} = \frac{sC_3g_{m2}R_4}{s^2C_3C_4R_4 + sC_3 + g_{m1}g_{m3}R_4} \quad (26)$$

$$\frac{V_{BP6}}{V_{in}} = \left(\frac{sC_3g_{m2}R_4}{s^2C_3C_4R_4 + sC_3 + g_{m1}g_{m3}R_4} \right) \left(1 + \frac{R_5}{R_6} \right) \quad (27)$$

Based on three NBPf responses of Equations (25)–(27), the f_o , Q , H_{BP4} , H_{BP5} and H_{BP6} are given by

$$f_o = \frac{1}{2\pi} \sqrt{\frac{g_{m1}g_{m3}}{C_3C_4}} \quad (28)$$

$$Q = R_4 \sqrt{\frac{C_4g_{m1}g_{m3}}{C_3}} \quad (29)$$

$$H_{BP4} = H_{BP5} = g_{m2}R_4 \quad (30)$$

$$H_{BP6} = g_{m2}R_4 \left(1 + \frac{R_5}{R_6} \right) \quad (31)$$

From Equations (28)–(31), the NBPf parameters of f_o and Q have orthogonal tunability, and the passband gains are independently adjustable by g_{m2} without affecting the NBPf parameters f_o and Q . Note that the resistors R_5 and R_6 in Equation (31) can independently achieve a wide range of NBPf adjustable passband gain without affecting the NBPf parameters f_o and Q .

By feeding back the input signal to the output response of the VM NBPf as in Figure 3a, a VM fully uncoupled QSO can be realized as shown in Figure 3b. This means that the input signal V_{in} is connected to V_{BP4} (Figure 3a), and Equation (25) becomes

$$\left. \frac{V_{BP4}}{V_{in}} \right|_{V_{BP4}=V_{in}} = 1 = \frac{sC_3g_{m2}R_4}{s^2C_3C_4R_4 + sC_3 + g_{m1}g_{m3}R_4} \quad (32)$$

Therefore, the CE from Figure 3b is expressed as

$$s^2C_3C_4R_4 + sC_3(1 - g_{m2}R_4) + g_{m1}g_{m3}R_4 = 0 \quad (33)$$

Based on Equation (33), the CO and FO of the proposed VM QSO (Figure 3b) are given by Equations (34) and (35), respectively:

$$\text{CO: } g_{m2}R_4 \geq 1 \quad (34)$$

$$\text{FO: } f_o = \frac{1}{2\pi} \sqrt{\frac{g_{m1}g_{m3}}{C_3C_4}} \quad (35)$$

According to Equations (34) and (35), the CO can be fully controlled independently without affecting the FO by adjusting g_{m2} or R_4 , and the FO can also be fully controlled independently without affecting the CO by adjusting g_{m1} or g_{m3} . Hence, the VM QSO shown in Figure 3b provides a fully uncoupled tuning law for the CO and FO. Routine analysis of the VM QSO structure (Figure 3b) yielded the following two quadrature voltage outputs and a sinusoidal output waveform with independent amplitude control:

$$\frac{V_{o4}}{V_{o5}} = \left. \frac{g_{m1}}{sC_3} \right|_{s=j\omega_o} = \frac{g_{m1}}{\omega_o C_3} e^{-j90^\circ} = \sqrt{\frac{C_4g_{m1}}{C_3g_{m3}}} e^{-j90^\circ} \quad (36)$$

$$V_{o6} = (1 + \frac{R_5}{R_6})V_{o5} \quad (37)$$

According to Equation (36), the magnitude ratio and phasor of output voltages V_{o4} and V_{o5} are given as

$$\left| \frac{V_{o4}}{V_{o5}} \right| = \sqrt{\frac{C_4 g_{m1}}{C_3 g_{m3}}} \bigg|_{C_3=C_4, g_{m1}=g_{m3}} = 1 \quad (38)$$

$$\angle \frac{V_{o4}}{V_{o5}} = -90^\circ \quad (39)$$

Equations (38) and (39) show that the voltage magnitude ratio of V_{o4} and V_{o5} is in unity, and the voltage phase shift of V_{o4} and V_{o5} is -90° . Thus, the proposed second VM QSO has a unity voltage amplitude ratio and a quadrature voltage phase shift.

3. Simulation and Experimental Measurements

The proposed isomorphic circuits of the VM NBPFS and VM QSOs were simulated with Cadence OrCAD PSpice simulation software version 16.6 to confirm the theory and implemented by three off-the-shelf LT1228 ICs, two grounded capacitors and three resistors to verify the performance of the isomorphic circuits. The isomorphic circuits adopted a DC bias of ± 5 V for the voltage supplies, and the voltage output was measured in the time domain with a Tektronix DPO 2048B oscilloscope, the voltage output was measured in the frequency domain with a Keysight E5061B-3L5 network analyzer, and the voltage output in the frequency spectrum was measured with a Keysight-Agilent N9000A CXA signal analyzer. The measurement photos of the isomorphic circuits are shown in Figures 4 and 5.

Figure 4. The (a) top view and (b) bottom view of the measured prototype from Figure 2.

Figure 5. The (a) top view and (b) bottom view of the measured prototype from Figure 3.

3.1. First Proposed Isomorphic Circuit Simulation and Experimental Results

The theoretical, simulated and measured NBPFS responses of V_{BP2} and V_{BP3} (Figure 2a) are shown in Figures 6 and 7, respectively, as depicted in Equations (11) and (12). In Figure 6, the NBPFS voltage gain at the voltage output of V_{BP2} could be adjusted independently by g_{m2} without affecting parameters f_0 and Q , as depicted in Equations (13)–(15). In Figure 7, the wide range of independent amplitude adjustable NBPFS voltage gain at the voltage output of V_{BP3} could also be independently adjusted by R_2 without affecting the NBPFS parameters f_0 and Q , as depicted in Equations (13), (14) and (16). The component values used in Figure 6 were $C_1 = C_2 = 0.96$ nF, $g_{m1} = g_{m3} = 1$ mS and $R_1 = R_2 = R_3 = 1$ k Ω .

for the theoretical value of $f_o = 165.8$ kHz, and only g_{m2} changed, with different values of 1, 1.5, 2 and 3 mS. This resulted in an NBPf at a voltage output of V_{BP2} with voltage gains of 0, 3.52, 6.02 and 9.54 dB. Figure 8 shows the measurement results of the V_{BP2} gain response. The component values used in Figure 7 were $C_1 = C_2 = 0.96$ nF, $g_{m1} = g_{m2} = g_{m3} = 1$ mS and $R_1 = R_3 = 1$ k Ω for a theoretical value of $f_o = 165.8$ kHz, and only R_2 changed, with different values of 1, 4, 11 and 19 k Ω . This resulted in an NBPf at a voltage output of V_{BP3} with voltage gains of 6.02, 13.98, 21.58 and 26 dB. Figure 9 shows the measurement results of the V_{BP3} gain response. To display the independent electronically adjusted passband gains from Figure 2a, the value of g_{m2} was changed from 1 mS to 4 mS, and the NBPf voltage gain at the voltage output of V_{BP2} could be changed from 0 dB to 12.04 dB. As shown in Figure 10, the proposed NBPf in Figure 2a exhibited independent electronic adjustment of passband gains at the voltage output of V_{BP2} . To display the maximum operating passband gains of Figure 2a, the value of R_2 was changed from 1 k Ω to 25 k Ω , and the NBPf voltage gain at the voltage output of V_{BP3} could be changed from 6.02 dB to 28.3 dB. As is shown in Figure 11, the proposed NBPf from Figure 2a exhibited a wide range of independent amplitude adjustable passbands at a voltage output of V_{BP3} . Figure 12 shows the theoretical, simulated and measured NBPf responses on the V_{BP2} output terminal with different values of R_1 when $C_1 = C_2 = 0.96$ nF, $R_2 = R_3 = 1$ k Ω and $g_{m1} = g_{m2} = g_{m3} = 1$ mS. Figure 13 shows the measurement results of the V_{BP2} gain response with different Q values. In this case, the four values of R_1 were changed to 1, 1.5, 2 and 3 k Ω , and the measured values of Q were 1.17, 1.53, 2.08 and 3. As shown in Figures 12 and 13, the Q parameter of the NBPf could be independently controlled by adjusting the different values of R_1 without affecting the parameters of f_o , as depicted in Equations (13) and (14). For an experimental test, Figure 14 shows the input and output voltage waveforms of the NBPf response at the V_{BP2} output terminal, which could be extended to an amplitude of 120 mVpp without significant distortion. As shown in Figure 14, the input voltage waveform was a 165.8-kHz sinusoidal waveform, and the output voltage waveform was measured to be a 161.7-kHz sinusoidal waveform with a phase difference of 0.23° .

To evaluate the performance of the proposed isomorphic circuit, the total harmonic distortion (THD) and the intermodulation distortion (IMD) of the isomorphic circuit were measured. The THD is defined as the ratio of the sum of the powers of all harmonic components (except the fundamental frequency) to the power of fundamental frequency. The IMD is defined as the two-tone test signals of different frequencies mixed together to form an additional signal at the excitation frequency. The THD and IMD can be expressed as follows [26,27]:

$$THD = \sqrt{\frac{\sum_{n=2}^N V_n^2}{V_1^2}} \times 100\% \quad (40)$$

where V_1 is the fundamental frequency voltage content and V_n ($n = 2, 3, \dots, N$) is the n th harmonic voltage content:

$$IMD(f_1, f_2) = \frac{\sum_{k=1}^2 p(f_2 - kf_1) + p(f_2 + kf_1)}{p(f_2)} \times 100\% \quad (41)$$

where $p(f_2 \pm (n - 1)f_1)$ is the root mean square value of the n th-order intermodulation component at the sum and difference of the two tones and $p(f_2)$ is the root mean square value of the fundamental component at the excitation frequency f_2 . Figure 15 shows the frequency spectrum of the NBPf response at the V_{BP2} output terminal. The NBPf response center frequency measured at the output of V_{BP2} was about 165.8 kHz. The THD, including the fundamental harmonic through the sixth harmonic components of Figure 15, was about 1.08%. Figure 16 shows the THD result of the NBPf response at the V_{BP2} output terminal with different input voltages by keeping a constant center frequency of 165.8 kHz.

Figure 17 shows the spectrum of the NBPF at the output of V_{BP2} , which was obtained by applying two-tone signals f_1 and f_2 around the theoretical value of $f_0 = 165.8$ kHz for IMD characterization. In Figure 17, the low-frequency tones of $f_1 = 164.8$ kHz and the high-frequency tones of $f_2 = 166.8$ kHz were used with an input amplitude of 63 mVpp. The measured value of the third-order IMD (IMD3) was about -40.1 dBc, and the third-order intercept (TOI) was about -7.95 dBm. Figure 18 shows the 1-dB compression point (P1dB) of the NBPF measured at the output of V_{BP2} between the output power and the input power when the center frequency was 165.8 kHz. The measured output power P1dB was about -11.4 dBm. Figure 19 shows the phase noise performance of the NBPF measured at the output of V_{BP2} . The measured value of the phase noise of V_{BP2} was less than -83.86 dBc/Hz at a 30-Hz offset.

Figure 6. The theoretical, simulated and measured NBPF responses at the voltage output of V_{BP2} when g_{m2} was varied.

Figure 7. The theoretical, simulated and measured NBPF responses at the voltage output of V_{BP3} when R_2 was varied.

Figure 8. The measured magnitude responses of the V_{BP2} when g_{m2} was varied.

Figure 9. The measured magnitude responses of the V_{BP3} when R_2 was varied.

Figure 10. The theoretical, simulated and measured results of the electronic amplitude adjustable NBP voltage gains at the voltage output of V_{BP2} .

Figure 11. The theoretical, simulated and measured results of the independent amplitude adjustable NBP voltage gains at the voltage output of V_{BP3} .

Figure 12. The Q parameters of the NBP at the output of V_{BP2} by varying R_1 while maintaining f_0 .

Figure 13. The measured magnitude responses of the V_{BP2} with different Q values by varying R_1 while maintaining f_0 .

Figure 14. The measured time domain input and output voltage waveforms of the NBPf at the output of V_{BP2} .

Figure 15. The frequency spectrum results of the NBPf at the output of V_{BP2} .

Figure 16. THD analysis results of the NBPf at the output of V_{BP2} with different input voltages.

Figure 17. The NBPF at the V_{BP2} output spectrum for a two-tone intermodulation distortion test.

Figure 18. The measured P1dB of the NBPF at V_{BP2} with the input power when f_0 was 165.8 kHz.

Figure 19. The measured phase noise performance of the NBPF at V_{BP2} .

The figure of merit (FoM) of the filter can be defined as [28]

$$\text{FoM} = \frac{\text{Dynamic range} \times f_0}{\text{Power dissipation} \times \text{Supply voltage}} \quad (42)$$

According the NBPF frequency spectrum measured in Figure 15, the dynamic range between the fundamental tone and the largest spurious of the spurious-free dynamic range was approximately 40.79 dB. Thus, the FoM of the proposed VM NBPF at the voltage output of V_{BP2} was approximately computed to be 3.07×10^6 .

To investigate the VM fully uncoupled QSO in Figure 2b, Figure 20 shows the measured waveforms of the quadrature voltage outputs V_{o1} and V_{o2} in the time domain. The component values used in Figure 20 were $C_1 = C_2 = 0.96$ nF, $g_{m1} = g_{m2} = g_{m3} = 1$ mS and $R_1 = R_2 = R_3 = 1$ k Ω for the theoretical value of $f_0 = 165.8$ kHz. The measured FO in Figure 20 was 162.3 kHz, which was close to the theoretical value of 165.8 kHz. Figure 21 shows the spectrum of the VM fully uncoupled QSO at the V_{o2} output terminal. The measured FO was 160.95 kHz, which was close to the theoretical value of 165.8 kHz. The difference between the amplitudes of the fundamental and second harmonics was 41.87 dB, and the calculated THD was about 0.81%. As shown in Figure 21, the third harmonic and subsequent harmonics were not visible in the spectrum because these harmonics were lower than the noise floor. This means that the THD of the first proposed VM fully uncoupled QSO was small. Figure 22 shows the phase noise performance of the QSO measured at the output of V_{o2} . The measured value of the phase noise of V_{o2} was less than -40.18 dBc/Hz at a 30-Hz offset. Figure 23 shows the FO tuning range of the measured output voltage V_{o2} , where $C_1 = C_2 = 0.96$ nF, $g_{m2} = 1$ mS and $R_1 = R_2 = R_3 = 1$ k Ω , and only $g_{m1} = g_{m3}$ were varied from the values of 0.5–10 mS. As is shown in Figure 23, the measured oscillation frequency was electronically and linearly varied from 82.15 kHz to 1629 kHz. In Figures 20–23, the g_{m2} value should have been slightly adjusted to satisfy the CO as depicted in Equation (19). Figure 24 shows the relationship between the measured amplitude ratio of the quadrature output voltages V_{o1} and V_{o2} and the FO as depicted in Equation (23). Figure 25 shows the relationship between the measured phase difference of the quadrature output voltages V_{o1} and V_{o2} and the FO as depicted in Equation (24). Figure 26 shows the experimental results of the voltage gain of V_{o3} when only R_2 was changed while maintaining $C_1 = C_2 = 0.96$ nF, $g_{m1} = g_{m2} = g_{m3} = 1$ mS, $g_{m2} = 1.05$ mS and $R_1 = R_3 = 1$ k Ω . As is shown in Figure 26, the proposed VM fully uncoupled QSO (Figure 2b) exhibited a wide range of independent amplitude adjustable V_{o3} voltage gains. The measured operating amplitude control range could be adjusted from 1.51 to 24.05. Table 4 summarizes the performance of the first proposed isomorphic circuit.

Figure 20. The measured waveforms of the quadrature voltage outputs V_{o1} and V_{o2} in the time domain.

Figure 21. The frequency spectrum results of the VM fully uncoupled QSO at the V_{o2} output terminal.

Figure 22. The measured phase noise performance of the VM QSO at the V_{o2} output terminal.

Figure 23. The experimental results of V_{o2} for $g_{m1} = g_{m3}$.

Figure 24. The measured amplitude ratio of the quadrature output voltages V_{o1} and V_{o2} versus the tuning FO.

Figure 25. The measured phase difference of quadrature output voltages V_{o1} and V_{o2} versus the tuning FO.

Figure 26. The experimental results of the V_{o3} voltage gain when R_2 was varied.

Table 4. Performance parameters of the performance of the first proposed isomorphic circuit.

Non-Inverting Bandpass Filter Factor	
Supply voltage (V)	± 5
Power dissipation (W, measurement)	0.22
Central frequency (kHz, design/measurement)	165.8/161.7
Maximum operating voltage gain (dB, design/measurement)	28.3/27.86
Measured total harmonic distortion at $V_{in} = 120 \text{ mV}_{PP}$ (%)	1.08
Measured output 1-dB compression point (dBm)	−11.4
Measured third-order intermodulation distortion at 63 mV_{PP} (dBc)	−40.1
Measured third-order intercept point (dBm)	−7.95
Measured phase noise at 30 Hz (dBc/Hz)	−83.86
Measured spurious-free dynamic range (dB)	40.79
Figure of merit	3.07×10^6
Fully Uncoupled Quadrature Oscillator Factor	
Supply voltage (V)	± 5
Power dissipation (W, measurement)	0.22
Number of sinusoidal voltage output used	3
Oscillation frequency (kHz, design/measurement)	165.8/162.3
Measured total harmonic distortion (%)	0.81
Measured phase noise at 30 Hz (dBc/Hz)	−40.18
Measured operating oscillation frequency range (kHz)	82.15~1629
Measured operating amplitude control range	1.51~24.05

3.2. Second Proposed Isomorphic Circuit Simulation and Experimental Results

The theoretical, simulated and measured NBPF responses of the V_{BP5} and V_{BP6} (Figure 3a) are shown in Figures 27 and 28, respectively, as depicted in Equations (26) and (27). In Figure 27, the NBPF voltage gain at the voltage output of V_{BP5} could be adjusted independently by g_{m2} without affecting the parameters f_o and Q as depicted in Equations (28)–(30). In Figure 28, the wide range of the independent amplitude adjustable NBPF voltage gain at the voltage output of V_{BP6} could also be independently adjusted by R_5 without affecting the NBPF parameters f_o and Q as depicted in Equations (28), (29) and (31). The component values used in Figure 27 were $C_3 = C_4 = 0.96$ nF, $g_{m1} = g_{m3} = 1$ mS and $R_4 = R_5 = R_6 = 1$ k Ω for the theoretical value of $f_o = 165.8$ kHz, and only g_{m2} changed, with different values of 1, 1.5, 2 and 3 mS. This resulted in an NBPF at a voltage output of V_{BP5} with voltage gains of 0, 3.52, 6.02 and 9.54 dB. Figure 29 shows the measurement results of the V_{BP5} gain response. The component values used in Figure 28 were $C_3 = C_4 = 0.96$ nF, $g_{m1} = g_{m2} = g_{m3} = 1$ mS and $R_4 = R_6 = 1$ k Ω for a theoretical value of $f_o = 165.8$ kHz, and only R_5 changed, with different values of 1, 4, 11 and 19 k Ω . This resulted in an NBPF at a voltage output of V_{BP6} with voltage gains of 6.12, 13.9, 21.3 and 25.52 dB. Figure 30 shows the measurement results of the V_{BP6} gain response. To display the independent electronically adjusted passband gains from Figure 3a, the value of g_{m2} was changed from 1 mS to 4 mS, and the NBPF voltage gain at the voltage output of V_{BP5} could be changed from 0 dB to 12.04 dB. As is shown in Figure 31, the proposed NBPF from Figure 3a exhibited independent electronic adjustment of the passband gains at a voltage output of V_{BP5} . To display the maximum operating passband gains from Figure 3a, the value of R_5 was changed from 1 k Ω to 25 k Ω , and the NBPF voltage gain at the voltage output of V_{BP6} could be changed from 6.02 dB to 28.3 dB. As is shown in Figure 32, the proposed NBPF from Figure 3a exhibited a wide range of independent amplitude adjustable passbands at the voltage output of V_{BP6} . Figure 33 shows the theoretical, simulated and measured NBPF responses on the V_{BP5} output terminal with different values of R_4 when $C_3 = C_4 = 0.96$ nF, $R_5 = R_6 = 1$ k Ω and $g_{m1} = g_{m2} = g_{m3} = 1$ mS. Figure 34 shows the measurement results of the V_{BP5} gain response with different Q values. In this case, the four values of R_4 were changed to 1, 1.5, 2 and 3 k Ω , and the measured values of Q were 1.14, 1.59, 2.06 and 2.98, respectively. As is shown in Figures 33 and 34, the Q parameter of the NBPF could be independently controlled by adjusting the different values of R_4 without affecting the parameters of f_o , as depicted in Equations (28) and (29). For the experimental test, Figure 35 shows the input and output voltage waveforms of the NBPF response at the V_{BP5} output terminal, which could be extended to an amplitude of 120 mVpp without significant distortion. As shown in Figure 35, the input voltage waveform was a 165.8-kHz sinusoidal waveform, and the output voltage waveform was measured as a 165.8-kHz sinusoidal waveform with a phase difference of 0.57° . Figure 36 shows the frequency spectrum of the NBPF response at the V_{BP5} output terminal. The NBPF response center frequency measured at the output of V_{BP5} was 165.8 kHz. The THD, including the fundamental harmonic through the sixth harmonic components of Figure 36, was about 1.12%. Figure 37 shows the THD result of the NBPF response at the V_{BP5} output terminal with different input voltages while keeping a constant center frequency of 165.8 kHz. Figure 38 shows the spectrum of the NBPF at the output of V_{BP5} , which was obtained by applying two-tone signals f_1 and f_2 around the theoretical value of $f_o = 165.8$ kHz for IMD characterization. In Figure 38, the low-frequency tones of $f_1 = 164.8$ kHz and the high-frequency tones of $f_2 = 166.8$ kHz were used with input amplitudes of 63 mVpp. The measured value of the IMD3 was about -39.66 dBc, and the TOI was about -5.753 dBm. Figure 39 shows the P1dB of the NBPF measured at the output of V_{BP5} between the output power and the input power when the center frequency was 165.8 kHz. The measured output power P1dB was about -9.77 dBm. Figure 40 shows the phase noise performance of the NBPF measured at the output of V_{BP5} . The measured value of the phase noise of V_{BP5} was less than -73.87 dBc/Hz at a 30-Hz offset. According to the NBPF frequency spectrum measured in Figure 36, the dynamic range between the fundamental tone and the largest spurious of the spurious-free dynamic

range was approximately 40.86 dB. Thus, the FoM of the proposed VM NBPF at the voltage output of V_{BP5} was approximately computed to be 3.08×10^6 .

Figure 27. The theoretical, simulated and measured NBPF responses at the voltage output of V_{BP5} when g_{m2} was varied.

Figure 28. The theoretical, simulated and measured NBPF responses at the voltage output of V_{BP6} when R_5 was varied.

Figure 29. The measured magnitude responses of the V_{BP5} when g_{m2} was varied.

Figure 30. The measured magnitude responses of the V_{BP6} when R_5 was varied.

Figure 31. The theoretical, simulated and measured results of the electronic amplitude adjustable NBP voltage gains at a voltage output of V_{BP5} .

Figure 32. The theoretical, simulated and measured results of the independent amplitude adjustable NBP voltage gains at a voltage output of V_{BP6} .

Figure 33. The Q parameters of NBPf at the output of V_{BP5} by varying R_4 while keeping f_0 .

Figure 34. The measured magnitude responses of the V_{BP5} with different Q values by varying R_4 while keeping f_0 .

Figure 35. The measured time domain input and output voltage waveforms of the NBPf at the output of V_{BP5} .

Figure 36. The frequency spectrum results of the NBPf at the output of V_{BP5} .

Figure 37. THD analysis results of the NBPf at the output of V_{BP5} with different input voltages.

Figure 38. The NBPf at the V_{BP5} output spectrum for a two-tone intermodulation distortion test.

Figure 39. The measured P1dB of the NBPf at V_{BP5} with the input power when f_o was 165.8 kHz.

Figure 40. The measured phase noise performance of the NBPf at V_{BP5} .

To investigate the VM fully uncoupled QSO in Figure 3b, Figure 41 shows the measured waveforms of the quadrature voltage outputs V_{o4} and V_{o5} in the time domain. The component values used in Figure 41 were $C_3 = C_4 = 0.96$ nF, $g_{m1} = g_{m2} = g_{m3} = 1$ mS and $R_4 = R_5 = R_6 = 1$ k Ω for the theoretical value of $f_o = 165.8$ kHz. The measured FO in Figure 41 was 161.3 kHz, which was close to the theoretical value of 165.8 kHz. Figure 42 shows the spectrum of the VM fully uncoupled QSO at the V_{o5} output terminal. The measured FO was 164 kHz, which was close to the theoretical value of 165.8 kHz. The difference between the amplitudes of the fundamental and second harmonics was 49.67 dB, and the calculated THD was about 0.33%. As is shown in Figure 42, the third harmonic and subsequent harmonics were not visible in the spectrum, because these harmonics were lower than the noise floor. This means that the THD of the second proposed VM fully uncoupled QSO was small. Figure 43 shows the phase noise performance of the QSO measured at the output of V_{o5} . The measured value of the phase noise of V_{o5} was less than -34.7 dBc/Hz at a 30-Hz offset. Figure 44 shows the FO tuning range of the measured output voltage V_{o5} , where $C_3 = C_4 = 0.96$ nF, $g_{m2} = 1$ mS and $R_4 = R_5 = R_6 = 1$ k Ω , and only $g_{m1} = g_{m3}$ were varied at values from 0.5 mS to 10 mS. As is shown in Figure 44, the measured oscillation frequency was electronically and linearly varied from 80.83 kHz to 1626 kHz. In Figures 41–44, the

g_{m2} value should have been slightly adjusted to satisfy the CO, as depicted in Equation (34). Figure 45 shows the relationship between the measured amplitude ratio of the quadrature output voltages V_{o4} and V_{o5} and the FO as depicted in Equation (38). Figure 46 shows the relationship between the measured phase difference of the quadrature output voltages V_{o4} and V_{o5} and the FO as depicted in Equation (39). Figure 47 shows the experimental results of the voltage gain of V_{o6} when only R_5 was changed while maintaining $C_3 = C_4 = 0.96$ nF, $g_{m1} = g_{m3} = 1$ mS, $g_{m2} = 1.05$ mS and $R_4 = R_6 = 1$ k Ω . As is shown in Figure 47, the proposed VM fully uncoupled QSO (Figure 3b) exhibited a wide range of independent amplitude adjustable V_{o6} voltage gains. The measured operating amplitude control range could be adjusted from 1.51 to 24.03. Table 5 summarizes the performance of the second proposed isomorphic circuit.

Figure 41. The measured waveforms of the quadrature voltage outputs V_{o4} and V_{o5} in the time domain.

Figure 42. The frequency spectrum results of the VM fully uncoupled QSO at the V_{o5} output terminal.

Figure 43. The measured phase noise performance of the VM QSO at the V_{o5} output terminal.

Figure 44. The experimental results of V_{o5} for $g_{m1} = g_{m3}$.

Figure 45. The measured amplitude ratio of quadrature output voltages V_{o4} and V_{o5} versus the tuning FO.

Figure 46. The measured phase difference of quadrature output voltages V_{O4} and V_{O5} versus the tuning FO.

Figure 47. The experimental results of the V_{O6} voltage gain when R_5 was varied.

Table 5. Performance parameters of the performance of the second proposed isomorphous circuit.

Non-Inverting Bandpass Filter Factor		
Supply voltage (V)		± 5
Power dissipation (W, measurement)		0.22
Central frequency (kHz, design/measurement)		165.8/165.8
Maximum operating voltage gain (dB, design/measurement)		28.3/27.58
Measured total harmonic distortion@ $V_{in} = 120 \text{ mV}_{PP}$ (%)		1.12
Measured output 1-dB compression point (dBm)		−9.77
Measured third-order intermodulation distortion at 63 mV _{PP} (dBc)		−39.66
Measured third-order intercept point (dBm)		−5.753
Measured phase noise at 30 Hz (dBc/Hz)		−73.87
Measured spurious-free dynamic range (dB)		40.86
Figure of merit		3.08×10^6
Fully Uncoupled Quadrature Oscillator Factor		
Supply voltage (V)		± 5
Power dissipation (W, measurement)		0.22
Number of sinusoidal voltage output used		3
Oscillation frequency (kHz, design/measurement)		165.8/161.3
Measured total harmonic distortion (%)		0.33
Measured phase noise at 30 Hz (dBc/Hz)		−34.7
Measured operating oscillation frequency range (kHz)		80.83~1626
Measured operating amplitude control range		1.51~24.03

4. Concluding Remarks

This paper proposes isomorphic circuits with independent amplitude control function VM NBPBs and VM fully uncoupled QSOs. The proposed VM NBPBs and VM fully uncoupled QSOs employ three LT1228 commercial ICs, two grounded capacitors and three resistors. The proposed isomorphic circuits exhibit low-output impedance and independent amplitude control, which are important for easily cascading the VM operation and independent control of the amplitude gain. The Q and f_o parameters of all the proposed NBPB responses are orthogonally adjustable. The passband gain of all the proposed NBPBs can be adjusted independently without affecting Q and f_o . The measured passband gains of the first proposed NBPB response can be independently tuned to 27.86 dB, and the measured phase noise is less than -83.86 dBc/Hz at a 30-Hz offset. The measured spurious-free dynamic range of the first proposed NBPB response was approximately 40.79 dB, and the calculated FoM was 3.07×10^6 . The measured passband gains of the second proposed NBPB response can be independently tuned to 27.58 dB, and the measured phase noise is less than -73.87 dBc/Hz at a 30-Hz offset. The measured spurious-free dynamic range of the second proposed NBPB response was approximately 40.86 dB, and the calculated FoM was 3.08×10^6 . By feeding back each input signal to the NBPB response, the VM fully uncoupled QSO could be obtained. During FO tuning, the voltage amplitude ratio of the quadrature output was equal. The FO could be electronically and linearly controlled by the bias current of the LT1228 IC, and one sinusoidal waveform could be independently controlled without any additional active devices. The experimental measurements confirmed the workability of the isomorphic circuits. Since the proposed isomorphic circuits have independent amplitude control functions for the NBPBs and QSOs, these isomorphic circuits are expected to have a wide variety of applications in the future, such as in phase-sensitive detection, signal processing, instrumentation, telecommunication and power systems.

Author Contributions: S.-F.W. and H.-P.C. conceived and designed the theoretical verifications; Y.K. revised the manuscript to improve the quality of English; H.-P.C. analyzed the results and wrote the paper; W.-Y.C. performed the simulations and experiments. All authors have read and agreed to the published version of the manuscript.

Funding: This research received no external funding.

Institutional Review Board Statement: Not applicable.

Informed Consent Statement: Not applicable.

Data Availability Statement: Not applicable.

Conflicts of Interest: The authors declare no conflict of interest.

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