2.5 GHZ SECOND- AND FOURTH-ORDER INDUCTORLESS RF BANDPASS FILTERS

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ABSTRACT

A new design approach for realising low-power low-voltage high-Q high-order RF bandpass filter is proposed. Based on the gyrator-C inductor topology, a 2^{nd}-order biquadratic bandpass filter can be realised by adding a series capacitor to the input port of the gyrator. High-Q 2^{nd}-order and 4^{th}-order fully differential RF bandpass filters operating in the 2.4-GHz ISM (Industrial, scientific and medical) frequency band under a 2-V single power supply voltage with low power dissipation are reported.

1. INTRODUCTION

Modern wireless communication systems, such as cellular, cordless, GPS and Bluetooth, have increased the demand for low-cost and high-performance RF receivers operating with low supply voltage and power consumption. The most promising key to these demands is to realise a fully integrated RF receiver in CMOS technology. One of the main problems of implementing a CMOS RF front-end is the integration of high quality inductors, typically required for high frequency band-selection and image rejection filters. Currently integrated passive inductor still suffers from losses due to the CMOS substrate, rendering low quality factor. To implement high-Q band-selective filters, Q-enhancement circuitry is typically used to boost up the inductor quality factor, but at the cost of noise, linearity and power consumption. Furthermore, passive inductor requires large chip area and thus increasing the cost of RF circuits. By contrast, active inductor can offer greater tuneability and high quality factor under small chip area.

Recently, there has been an increasing interest in the design and implementation of CMOS inductorless RF bandpass filters [1]-[5] However, formerly reported filters are usually of second-order and dissipating more than 10 mW. With the emphasis on low voltage, low power and tuneability, this paper proposes a new design approach that is suitable for realising high-Q RF high-order bandpass filters. Based on the gyrator-C inductor topology, section 2 introduces the proposed method for realizing a 2^{nd}-order biquadratic bandpass filter. Section 3 describes the implementation of the 2^{nd}-order biquad filter for RF applications, while section 4 presents the design of a 4^{th}-order RF bandpass filter. Finally, conclusion is given in section 5.

2. BIQUADRATIC BANDPASS FILTER BASED ON THE GYRATOR-C INDUCTOR TOPOLOGY

Based on the classical gyrator-C active inductor topology, a 2^{nd}-order biquadratic bandpass transfer function can be realized by adding a series capacitor to the input port of the active inductor, as shown in Figure 1. The bandpass transfer function obtained at the other port of the gyrator can be expressed as given in eqn. (1),

$$\frac{V_{out}}{V_{in}}(s) = \frac{s\left(\frac{G_{m1}}{C_1}\right)}{s^2 + s\left(\frac{G_{o1}}{C_1} + \frac{G_{o2}}{C_2 + C_{in}}\right) + \frac{G_{m1}G_{m2}}{C_1(C_2 + C_{in})}} \tag{1}$$

where G_{mi} and G_{oi} are respectively the transconductance and output conductance of the transconductors. The centre frequency ω_o, quality factor Q_o and passband gain K_o can be expressed as given by eqns. (2)-(4), where it can be deduced that large transconductance and small output conductance values are necessary for realising high-frequency and high-Q bandpass filter. Note that a 2^{nd}-order highpass function is obtained at node V_1 in Figure 1.

$$\omega_o = \sqrt{\frac{G_{m1}G_{m2}}{C_1(C_2 + C_{in})}} \tag{2}$$

$$Q_o = \frac{\omega_o}{\frac{G_{o1}}{C_1} + \frac{G_{o2}}{C_2 + C_{in}}} \tag{3}$$

$$K_o = \frac{G_{m1}/C_1}{\frac{G_{o1}}{C_1} + \frac{G_{o2}}{C_2 + C_{in}}} \tag{4}$$

Without C_{in}, the same bandpass transfer function can also be realised by considering the currents of the gyrator as signals. However the current-mode implementation as demonstrated in [6] is somewhat awkward to be used for realising a high-order filter, as the DC biasing of each biquad must coordinate properly. Contrastingly, the proposed voltage-mode biquad can simply be cascaded to realize a higher-order bandpass filter, thanks to the input capacitive coupling. The DC biasing of each biquad can be adjusted independently, thus greatly simplifying the design of a higher-order filter.

3. SECOND-ORDER RF BANDPASS FILTER

3.1. Single-ended implementation

For RF applications, the 2nd-order bandpass filter shown in Figure 1 can be realised by using a VHF active inductor. Figure 2 shows the realisation of two single-ended 2nd-order bandpass filter by using the VHF CMOS active inductors reported in [1] and [2]. Both implementations achieve the same transfer function with the following parameters (all symbols have their usual meanings).

$$\omega_o \approx \sqrt{\frac{g_{m1}g_{m2}}{c_{gs2}\left(c_{gs1}+C_{in}\right)}} \tag{5}$$

$$Q_o \approx \frac{\omega_o}{\frac{g_{ds1}}{c_{gs2}}+\frac{g_{m1}}{c_{gs1}+C_{in}}} \tag{6}$$

$$K_o = \frac{\frac{g_{m1}}{c_{gs2}}\left(\frac{C_{in}}{c_{gs1}+C_{in}}\right)}{\frac{g_{ds1}}{c_{gs2}}+\frac{g_{m1}}{c_{gs1}+C_{in}}} \tag{7}$$

It can be deduced that Q_o and K_o will be less than unity, thus Q-enhancement is necessary. The Q-factor and passband gain can be improved by applying cascode and/or negative resistance cancellation techniques [1], [2]. By contrast the circuit in Figure 2(a) renders lower supply voltage and power dissipation, whereas the circuit in Figure 2(b) allows a greater flexibility since g_{m1} and g_{m2} can be tuned independently.

Note that the input impedance matching can be obtained by adding, prior to C_{in}, an L-matching network, which can be implemented by using a series bonding inductor and an off-chip shunt capacitor [2].

3.2. Q-enhanced fully differential implementation

In this paper, the circuit shown in Figure 2(b) was chosen as a design example. A Q-enhanced fully differential version of the circuit is shown in Figure 3. Cascode transistor, M_{p3}, and cross-coupled transistor, M_{p4}, reduce g_{ds2} and g_{ds1} respectively. Note that this effectively increases the Q-factor of the active inductor. The DC bias voltages V_a and/or V_b can be used to tune the Q of the inductor [2], and hence the Q of the filter. The DC bias currents I_1 and/or I_2 can be used to tune the inductance, and hence the centre frequency of the filter.

Using a 0.35-μm CMOS process, the proposed 2nd-order fully differential bandpass filter was designed to operate under a 2-V single power supply voltage. All transistors have the minimum channel length of 0.35-μm, while the channel widths are as follows: W_1 = 10 μm, W_2 = 20 μm, W_3 = 120 μm, and W_4 = 0.35 μm. At 2.5-GHz center frequency, C_{in} is set to 10 fF and the cascode current sources I_1 and I_2 draw 180 and 500 μA respectively, thus the filter only dissipates 2 mW. Unless stated otherwise, all HSPICE simulation of the biquad were performed under the above biasing condition.

Figures 4 and 5 show the simulated AC magnitude responses of the filter. The Q-tuning capability of the filter is demonstrated in Figure 4, where adjusting V_b from 0 to 0.25 V increases both the Q-factor from 50 to over 3000 and the peak voltage gain from 4 to 40 dB. Note that the center frequency only shifts by 10 MHz or 0.4 %. The center frequency tuning capability with constant gain and bandwidth adjustment of the filter is shown in Figure 5, where adjusting I_2 from 460 to 800 μA increases the center frequency from 2.45 to 2.85 GHz. The 400-MHz or 15 % frequency tuning range is limited by the fact that the biasing of the circuit has changed to the condition at which the transistor M_1 is no longer in saturation region and thus high Q-factor cannot be maintained.

As an evaluation of linearity and dynamic range, an intermodulation test was performed by applying two-tone sinusoidal signals at 2.42 and 2.45 GHz to the biquad filter working at 2.45-GHz center frequency, 80-MHz bandwidth (i.e. RF signal band for Bluetooth receivers), and 21-dB peak voltage gain. The third-order input-referred intercept point (IIP_3) was found at 0 dBV (0 dBV = 1 V_{pk}) Total in-band input-referred noise voltage is equal to 150 μV_{rms} yielding 41-dB spurious free dynamic range (SFDR), which is too low for most applications. One way of improving noise and SFDR is to increase C_{in}, but at the cost of higher power dissipation.

4. FOURTH-ORDER RF BANDPASS FILTER

In a typical heterodyne RF receiver, a higher-order band-selective filter is required to provide good image rejection. As mentioned in section 2, one of the advantages of the proposed biquad filter topology is that the realisation of a higher-order filter is straightforward, simply by cascade connection of biquad stages. This is demonstrated by the realisation of a 4^{th} –order bandpass filter operating at 2.51-GHz centre frequency and 70-MHz bandwidth, as shown in Figure 6. The input capacitors of the 1^{st} and 2^{nd} biquad stages are 10 and 1 fF respectively. The 1^{st} biquad has 2.46-GHz centre frequency, 30.4-dB peak gain and 30-MHz bandwidth, while the 2^{nd} biquad works at 2.5-GHz centre frequency, 34.5-dB gain and 2-MHz bandwidth. The 4^{th}-order filter dissipates only 4.4 mW under a 2-V single power supply voltage.

Figures 7 and 8 show the simulated AC frequency responses of the 4^{th}-order filter. The filter exhibits 20-dB passband gain with 1-dB ripple and 10-dB attenuation at 60-MHz frequency offset from the centre frequency. Assume that the filter is to be used in an RF receiver with 100-MHz intermediate frequency, an image rejection of 35 dB can be obtained at 200 MHz away from the centre frequency. The Q-tuning capability of the 4^{th}-order bandpass filter is demonstrated in Figure 8, where only V_b of the 2^{nd} biquad is adjusted from 0.1 to 0.3 V. This changes the Q-factor changes from 30 to 40.

Linearity and dynamic range were evaluated in the same manner as the 2^{nd}-order biquad. Two-tone sinusoidal signals at 2.46 and 2.5 GHz were applied to the 4^{th}-order filter. The IIP_3 was found at -5 dBV as shown in Figure 9. The total in-band input-referred noise voltage is equal to 350 μV_{rms}, yielding a SFDR of 35-dB. The simulated performance of the 4^{th}-order bandpass filter is summarized in Table 1.

5. CONCLUSION

A new design approach suitable for realizing higher-order biquadratic RF bandpass filters with low supply voltage and power dissipation has been proposed. The proposed method renders very simple implementation of RF bandpass filter, allowing operation at high frequencies with low power supply voltage and reduced chip area. The proposed method looks very promising for realizing RF band-selected filters for short/medium wireless data applications such as Bluetooth, where the dynamic range requirements are not too stringent.

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6. REFERENCES

[1] Y. Wu, X. Ding, M. Ismail, and H. Olsson, "Inductorless CMOS RF bandpass filter," *Electronics Letters*, Vol. 37, No. 16, pp. 1027-1028, Aug. 2001.
[2] A. Thanachayanont, "A 1.5-V CMOS fully-differential inductorless RF bandpass amplifier," *Proc. IEEE ISCAS'2001*, Vol. 1, pp. 49-52, 2001.
[3] D. Leung and H. Luong, "A fourth-order CMOS bandpass amplifier with high linearity and high image rejection for GSM receivers," *Proc. IEEE ISCAS'1999*, Vol. 2, pp. 589-592, 1999.
[4] C. H. Leong and G. W. Roberts, "A sixth-order UHF bandpass filter using silicon bipolar active inductors," *Proc. IEEE ISCAS'1998*, Vol. 1, pp. 127-130, 1998.
[5] R. Kaunisto, K. Stadius, V. Porra, "A.3 GHz silicon-BJT active resonator and filter," *Proc. IEEE ICECS'1998*, Vol. 3, pp. 197-200, 1998.
[6] U. Yodprasit and K. Sirivathanant, "A compact low-power vertical filter for very-high-frequency applications," *Proc. IEEE ISCAS'2001*, Vol. 1, pp. 164-167, 2001.

Table 1. Performance of the 4^{th}-order bandpass filter

Parameters		Simulation results
Technology		0.35-μm CMOS
Power supply voltage		2 V
Total supply current		2.2 mA
Centre frequency		2.4 –2.6 GHz
-3dB Bandwidth		20 – 200 MHz
Maximum passband gain		0 – 40 dB
f_o = 2.5 GHz	In-band noise	350 μV
Q_o = 36	IIP3	-5 dBV
K_o = 20 dB	SFDR	35 dB

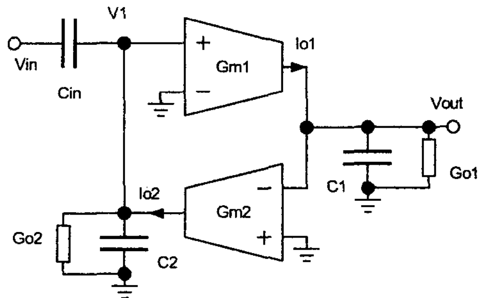

Figure 1. Second-order biquadratic bandpass filter based on the gyrator-C architecture.

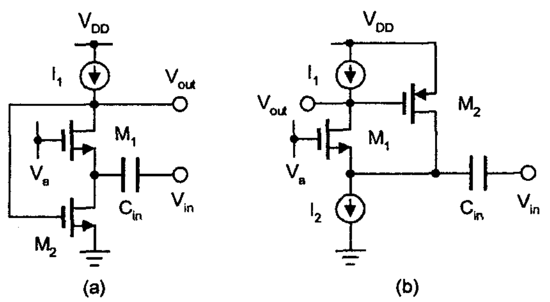

Figure 2. Realisations of single-ended 2^{nd}-order RF bandpass filter.

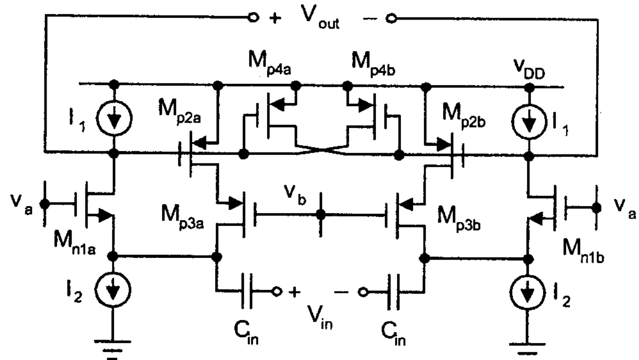

Figure 3. Fully differential 2nd-order RF bandpass filter with Q-enhancement.

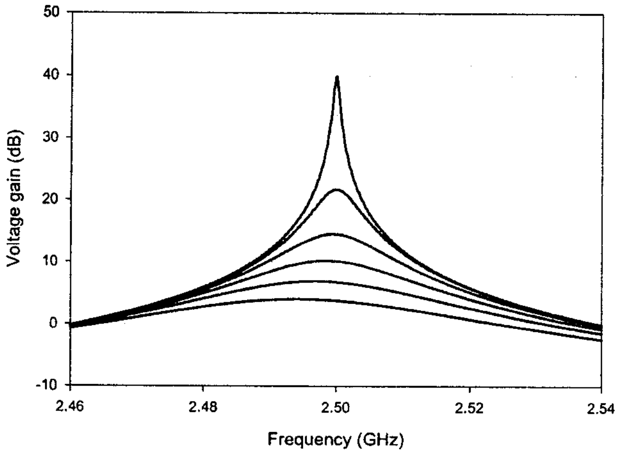

Figure 4. Q-tuning of the 2nd-order filter.

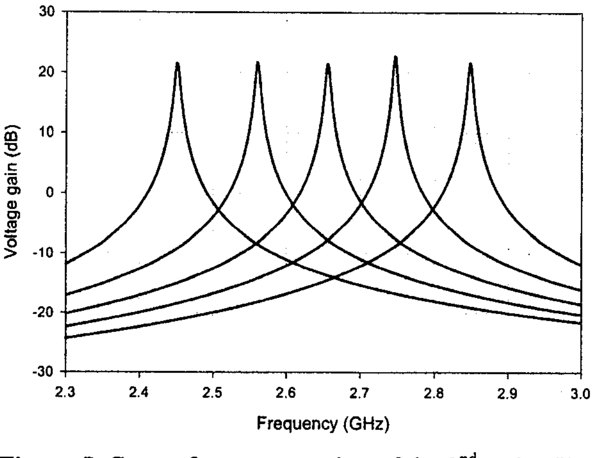

Figure 5. Centre frequency tuning of the 2nd-order filter with adjustment of gain and bandwidth.

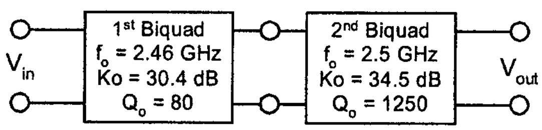

Figure 6. Block diagram of a 4th-order filter operating at 2.51-GHz centre frequency and 70-MHz bandwidth.

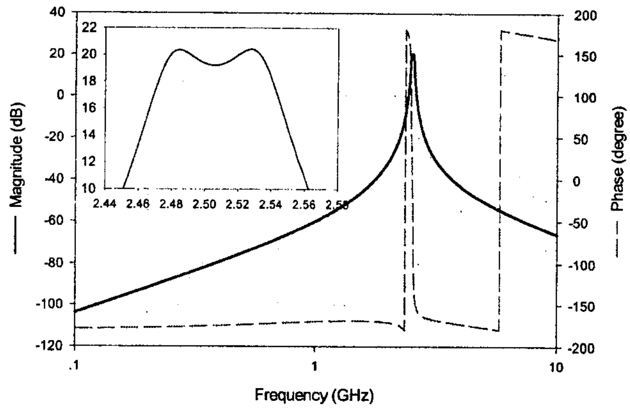

Figure 7. Magnitude and phase responses of the 4th-order bandpass filter.

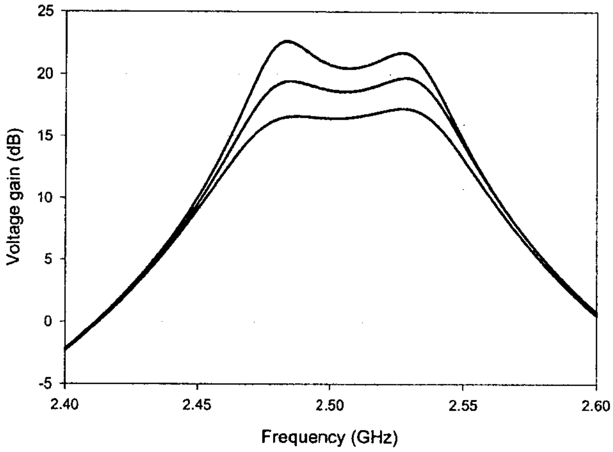

Figure 8. Q tuning of the 4th-order filter.

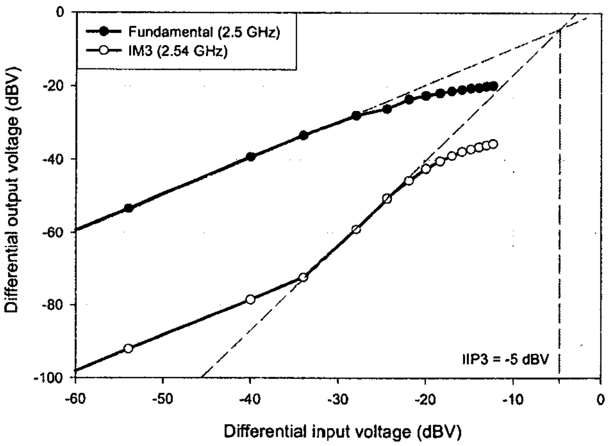

Figure 9. Third-order intermodulation of the 4th-order filter.