Oscillator topology: Fig. 1 shows the oscillator circuit. The oscillator was fabricated in a 0.4µm digital CMOS technology. Fig. 2 shows a chip photograph. The average capacitance in the resonance circuit is controlled by the tuning voltage V_{cap} . In our design, the capacitance is composed of a pair of PMOS transistors. In addition, a second control input V_{ind} controls the average inductance in the resonance circuit. The outer inductances in Fig. 1 are realised as a centre-tap coil to increase the inductance and the quality factor.

Fig. 2 Chip photograph of VCO

Fig. 3 Measured oscillation frequency against fine tuning voltage V_{cap} for different values of V_{ind}

Fig. 4 Single-sideband phase noise spectrum

The oscillation frequency depends on both the fixed reactances C_{fix} , L_{fix} , and the variable reactances C_{var} , L_{var} . Assuming an inverse square root dependence of the oscillation frequency on C and L , the ratio f_{max}/f_{min} of the boundaries of the tuning range equals $[(C_{var} + C_{fix})/C_{fix}]^{1/2}$ for capacitive tuning and $[(L_{var} + L_{fix})/L_{fix}]^{1/2}$ for inductive tuning. Since for on-chip oscillators the relative contribution of L_{fix} can be very small compared to C_{fix} , the

achievable tuning range is much higher for inductive tuning than for capacitive tuning.

Measurement results: The measured tuning characteristics are shown in Fig. 3. The capacitive tuning range is between 100 and 150 MHz. The inductive tuning range amounts to 650 MHz. Fig. 4 shows the phase noise spectrum for an oscillation frequency of 1.9 GHz. The oscillator core draws 13 mA current. Power consumption can be reduced at the expense of higher phase noise by increasing the value of the resistor in Fig. 1.

Conclusions: We have presented a fully integrated wideband VCO fabricated in a digital 0.4µm CMOS technology. The VCO has a control input with moderate gain for channel selection, and a second control input to compensate for temperature and lot-to-lot variations in the device parameters. The latter control input may be embedded in a very slow PLL, while the former can be used for channel selection with a fast PLL.

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Electronics Letters Online No: 20010245

DOI: 10.1049/el:20010245

10 January 2001

F. Herzel, H. Erzgräber and P. Weger (IHP, Im Technologiepark 25, D-15236 Frankfurt (Oder), Germany)

E-mail: herzel@ihp-ffo.de

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Integrator and differentiator with time constant multiplication using current feedback amplifier

Jiin-Long Lee and Shen-Iuan Liu

A set comprising an active RC integrator and differentiator with time constant multiplication is presented. The proposed integrator and the differentiator can be used for low-frequency signal processing applications. Moreover, no extremely large-valued passive components are needed. In the integrator, the unlimited multiplication of time constants is allowed. Both have been implemented using commercially available current feedback amplifiers. Experimental results demonstrate the theoretical predictions.

Introduction: Active RC integrators and differentiators are basic building blocks of analogue signal processing, such as instrumentation systems, waveform generation applications and filtering circuits [1, 2]. Large time constant integrators and differentiators are useful for the design of proportional plus integral plus derivative controllers [3, 4]. Many active RC network integrators and differentiators with increased time constants based on conventional operational amplifiers have been reported in the literature [5, 6]. Current feedback amplifiers (CFAs) have been proved to overcome the shortcomings of conventional operational amplifiers, such as slow rate limitation and non-constant closed loop band-

width for different gains. In this Letter, a set of integrator and differentiator with time constant multiplication using CFAs is presented.

Fig. 1 Proposed integrator using non-inverting CFA

Fig. 2 Proposed differentiator using inverting CFA

Circuit descriptions: In general, a CFA is a four-terminal amplifier and its characteristics can be modelled as the following equations:

$$V_x = V_y \quad i_y = 0 \quad i_z = \pm i_x \quad \text{and} \quad V_o = V_z \quad (1)$$

where the plus and minus signs denote a non-inverting CFA and an inverting CFA, respectively. The proposed integrator using a CFA is shown in Fig. 1. If one takes the odd number resistors ($R_1, R_3, R_5, \dots$) equal to nR , even number resistors ($R_2, R_4, \dots$) equal to R , and the m th node as the input, a general form of the transfer function can be expressed as

$$H_m(s) = \frac{V_o}{V_m} = \frac{1}{snRC \frac{1}{2m\sqrt{n^2+4n}} [(n+2+\sqrt{n^2+4n})^m - (n+2-\sqrt{n^2+4n})^m] + 1} \quad (2)$$

where n is the ratio between odd number and even number resistors and m is an arbitrary integer larger than 2. Given any resistor ratio and m in Fig. 1, a multiplication factor can be obtained by using eqn. 2. For instance, if $m = 3$ (V_3 is an input) the transfer function can be given as

$$\frac{V_o}{V_3} = \frac{1}{snRC[(n+2)(n+1)+n+1]+1} = \frac{1}{s\tau+1} \quad (3)$$

where $\tau = nRd[(n+2)(n+1)+n+1]$. According to eqn. 3, the time constant is multiplied by a factor of $(n+2)^2 - 1$. If $n = 10$, a multiplication factor of 143 can be realised.

The proposed differentiator using a CFA is shown in Fig. 2. The routine analysis gives

$$V_o = sC \frac{R_5 \left(1 + \frac{R_2}{R_1}\right) + R_2 \left(1 + \frac{R_4}{R_3}\right) \left(\frac{R_4}{R_1} + \frac{R_2}{R_3} + 1\right)}{1 - \frac{R_2 R_5}{R_1 R_4}} V_{in} \quad (4)$$

If $R_2 = R$, $R_4 = 2R$, $R_1 = R_3 = R_5 = 2nR$, eqn. 4 can be expressed as

$$V_o = s4RC \left[n \left(1 + \frac{1}{2n}\right) + (n+1)(n+1) \right] \quad (5)$$

For instance, if $R_2 = 1\text{ k}\Omega$, $R_4 = 2\text{ k}\Omega$, $R_1 = R_3 = R_5 = 10\text{ k}\Omega$, the time constant is multiplied by a factor of 166, and the unity-gain frequency is lowered by the same factor.

Fig. 3 Input and output waveforms of proposed integrator

Fig. 4 Input and output waveforms of proposed differentiator

Experimental results: To verify the theoretical analysis, the proposed integrator and differentiator have been implemented using commercially available current feedback amplifiers (AD844). Fig. 3 shows a typical test result of the integrator when $m = 3$. The rectangular waveform V_1 (channel 1) ($= 4.0\text{V}$ peak-to-peak, 300Hz) was applied to the circuit in Fig. 1 with $C_1 = 900\text{pF}$, $R_1 = R_3 = R_5 = 10\text{ k}\Omega$, and $R_2 = R_4 = 1\text{ k}\Omega$. The output waveform of the proposed integrator is also shown in Fig. 3 (channel 2). Its operating frequency range is from 200Hz to 1MHz , with phase error $< 12^\circ$. Fig. 4 shows a typical test result of the differentiator. The triangular waveform V_1 (channel 1) (1.8V peak-to-peak, 300Hz) was applied to the circuit of Fig. 2 with $C = 17\text{nF}$, $R_2 = 500\Omega$, $R_4 = 1\text{ k}\Omega$ and $R_1 = R_3 = R_5 = 3.27\text{ k}\Omega$. The output waveform of the proposed differentiator is also shown in Fig. 4 (channel 2). Its operating frequency range is from 100Hz to 10kHz with a phase error $< 6^\circ$.

Conclusion: A set comprising an active RC integrator and differentiator with time constant multiplication using CFAs has been developed. Since the equivalent resistors have at least a double multiplication effect, it is also suitable to realise low-frequency applications. Experimental results have been given to prove the feasibility of the proposed circuits.

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2 January 2001

Electronics Letters Online No: 20010252

DOI: 10.1049/el:20010252

Jiin-Long Lee and Shen-Iuan Liu (Department of Electrical Engineering and Graduate Institute of Electronic Engineering, National Taiwan University, Taipei, Taiwan 10617, Republic of China)

E-mail: lsi@cc.ee.ntu.edu.tw

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Regulated cascode-based vertical filter

U. Yodprasit and K. Sirivathanant

An approach to realise a low-power, high-order, very-high-frequency current-mode filter is proposed. The filter utilises vertical connection of regulated-cascode (RGC) biquads to reduce current consumption. Simulation results of a fourth-order Chebyshev lowpass filter are demonstrated.

RGC biquad: Fig. 1 shows the basic RGC and its equivalent circuit indicating a direct implementation of a current-mode two-integrator-loop biquad. The characteristic equations of the bandpass (bp) and lowpass (lp) functions can be expressed as

$$\omega_{o,bp} \approx \omega_{o,lp} = \sqrt{\frac{g_{m1}g_{m2}}{C_{gs1}C_{gs2}}} \quad (1)$$

$$\omega_{z,lp} = \frac{g_{m1}g_{m2}r_{ds2}}{C_{gs2}} \quad (2)$$

$$Q_{bp} \approx Q_{lp} = \frac{\omega_{o,bp,lp}}{\frac{1}{C_{gs2}r_{ds1}} + \frac{1}{C_{gs1}(\frac{1}{g_{m1}})}} \approx \frac{\omega_{o,bp,lp}}{\omega_{T1}} \quad (3)$$

where $\omega_{o,bp}$, $\omega_{o,lp}$ and $\omega_{z,lp}$ are, respectively, the centre frequency of the bandpass function, the cutoff frequency and the zero frequency of the lowpass function, and ω_{T1} denotes the transition frequency of M1. The biquad differs from an ideal biquad only that the bandpass current, $i_{o,bp}$, is negatively fed back to the input terminal through C_{gs2} , hence decreasing of the bandpass and lowpass Q factors (Q_{bp} and Q_{lp}) which, from eqn. 3, cannot exceed unity in normal centre/cutoff frequencies ($\omega_{o,bp,lp} \ll \omega_{T1}$). To compensate for the loss, a negative resistor (Q -tuning network 1) is placed between the inputs of the bottom biquad shown in Fig. 2. A major role in loss compensation of the negative resistor is to provide positive feedback of the bandpass current by scaling $i_{o,bp}$ of M1 through the translinear loop formed by M1 and M7 of both sides. Thus, the compensating current can be controlled by adjusting the transconductance of M7, whereas the transconductance of M1 is used for frequency adjusting. The compensated Q factors can be derived as

$$Q_{bp,comp} \approx Q_{lp,comp} = \frac{\omega_{o,bp,lp}}{\frac{(g_{m1}-g_{m7})}{C_{gs1}} + \frac{1}{C_{gs1}r_{ds2}} + \frac{1}{C_{gs2}r_{ds1}}} \approx \frac{\omega_{o,bp,lp}}{\frac{(g_{m1}-g_{m7})}{C_{gs1}}} \quad (4)$$

Fig. 1 Conventional RGC filter

RGC-based vertical filter: A fourth-order, 0.25dB ripple, Chebyshev lowpass filter, based on the proposed technique, has been realised as shown in Fig. 2. The lowpass filter utilises two bal-

anced biquads having different Q factors and cutoff frequencies to realise Chebyshev response. For an example of 1GHz cutoff frequency of the filter, the cutoff frequency and the Q factor of the bottom biquad are set to 1.08GHz and 2.5 whereas they are set to 675MHz and 0.66 for the top biquad. Note that for Q -factor adjusting, the Q factor of the bottom biquad is greater than unity so that Q -tuning network 1 is needed. By contrast, the Q factor of the top biquad has to be < 0.7 and since it might be greater than this value in practical design loss should be further introduced via negative feedback provided by the Q -tuning network 2. For cutoff-frequency adjusting, owing to vertical connection, M2s share the same biasing current, I_{O2} , hence different cutoff frequencies can be obtained by adjusting I_{O11} and I_{O12} . This offers an advantage over conventional vertical filters [1] since linear capacitors are not necessary and hence they are electronically tunable and digital process compatible. Moreover, common-mode feedback (CMFB) network is not required for each biquad since the DC input voltage can be well defined by V_{GS1} . This greatly simplifies circuit topology because a CMFB network must be carefully designed to achieve low noise and larger bandwidth than the filter.

Fig. 2 Fourth-order Chebyshev lowpass filter

Fig. 3 Cutoff frequency tuning and magnified passband

a Cutoff frequency tuning
b Magnified passband
--- 900 MHz
— 1 GHz
... 1.1 GHz

Table 1: Summarised performances of lowpass filter

f_o	SFDR	DR	P_{diss}
Hz	dB	dB	mW
900M	39	35	1.5
1G	39	37	1.8
1.1G	42	41	2.1

Simulation results: The filter was simulated by SPECTRE using 0.35 μ m complementary metal-oxide-semiconductor (CMOS) process and 2.4V supply voltage. Figs. 3a and b show frequency tun-