A widely tunable active-RC complex filter for multi-mode wireless receivers with automatic frequency tuning

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Abstract: A 5th-order Chebyshev active-RC filter has been implemented in TSMC 0.13 μ CMOS technology. With the feature of multi-mode, it has 8 different bandwidth modes including 6 Real Low-pass modes and 2 Complex band-pass modes for wireless sensor network applications. Combined with a special kind of Miller compensation method, the OTA's stability is observably improved. Also, to eliminate the effect of temperature and process variation, an automatic frequency tuning scheme is applied. The filter measures a 0 dB gain and 0.9 dB ripple in passband. It also measures an input-referred noise of 28.8 nV/ $\sqrt{\text{Hz}}$ and in-band-IIP3 of 25 dBm while consuming 15 mA current from 1.3 V supply. The tuning circuit has a precision of 2% and covers a range of 30%.

Keywords: complex filter, multi-mode, auto tuning, Miller compensation

Classification: Integrated circuits

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1 Introduction

Wireless sensor network (WSN) has been widely recognized as a promising technology with a variety of applications both in martial and civilian fields. A WSN system consists of large numbers of spatially distributed nodes which can collaborate with one another to enable the reliable monitoring of the external environment. The position of sensor nodes need not be engineered or predetermined. This allows random deployment in inaccessible terrains or disaster relief operations [1]. On the other hand, WSN’s requirements of low cost, low power, high flexibility and high reliability make strict demands for RF transceiver design.

A typical low-IF receiver architecture used in the WSN application is shown in Fig. 1. To adapt WSN’s requirement of flexibility, Low-IF and zero-IF architectures are both utilized in this design. Received signal from antenna is amplified by low noise amplifier (LNA) and then split into quadrature paths followed by down-converting and filtering [2]. As the major component of filtering, Complex filter is an essential component in receiver to select signal and attenuate image signal and adjacent interference in low intermediate frequency mode.

Concerning about the design of an integrated filter, two major issues need to be preferentially considered, the selection of LC prototypes and the implementation models. Regarding the selection of LC prototype, there are concerns about pass-band ripple, stop-band rejection and group delay. Since Orthogonal Frequency Division Multiplexing (OFDM) scheme is adopted in the system, which can tolerate a non-uniform group-delay response, so Chebyshev filters are more suitable for the rectangular spectra and it requires less orders than Butterworth filters [3]. For implementation models, two major options are Gm-C and Active-RC filter. Even though Active-RC filter’s power dissipation is larger in most cases, but to meet the requirement of high linearity, an Active-RC filter would be a better option [4].

This paper is organized as follows. We shall first briefly discuss the filter architecture in section II. This will be followed by the detail of circuit implementation including OTA design and tuning scheme in section III. Simulation and measurement results are presented in section IV respectively. Finally, conclusions are drawn in section V.

Fig. 1. Low-IF receiver architecture

2 Filter architecture

Regardless of the actual realization methods, e.g., LC, or active-RC, the ladder topology is almost always more widely deployed than the cascaded biquads due to its low sensitivity properties to component variations [5]. In this design, a 5th order Chebyshev active-RC filter is synthesized by using leap-frog method. The transfer function of typical bandwidth is derived from system specifications (equation 1).

$$H(s) = \frac{2261}{(s^2 + 1.714s + 40.39)(s^2 + 4.488s + 20.18)(s + 2.774)} \quad (1)$$

Since the receiver will potentially be used as a low-IF receiver, rejection of image signal and adjacent interferers should be deliberately considered during filter design process. As illustrated in Fig. 2, output signal from RF is mixed with quadrature local oscillator (ω_{LO}) signal and the resulting signal is frequency-shifted so that the desired and image signals fall into same frequency with conjugate representations [6]. In order to extract information contained in desired spectrum and suppress the image signal, a complex band-pass filter (CBPF) is required.

Fig. 2. Signals in Low-IF receiver

Complex filter, also known as poly-phase filter, can be implemented by transforming s of LPF to $s - j\omega_c$ in the s -plane. As the fundamental block of fully differential Active-RC filter, integrator has a transfer function of

$$H_{lp} = \frac{1}{jRC\omega} \quad (2)$$

Fig. 3. b1) Complexification of the integrator by frequency shifting
b2) Fully-differential active-RC complex integrator.

Generally, poles of a real filter are symmetric with the real axis, and the responses of the filter to positive frequency and negative frequency are the same [7], so shift the frequency response of integrator by ω_0 , we can get the band-pass transfer function

$$H_{bp} = \frac{1}{jRC(\omega - \omega_0)} \quad (3)$$

and it can be realized by introducing a complex feedback loop.

As depicted in Fig. 3(a), x and y , respectively, are the input and output of two orthogonal channels with $x_Q = -jx_I$ (channel y in the same way with it), we assume that the feedback factor from the output of channel I to the input of channel Q is K , so Fig. 3(a) can be expressed as:

$$(x_i - y_q K)H(\omega)_{lp} = y_i \quad (4)$$

$$H(\omega)_{bp} = \frac{1}{jRC\left(\omega - \frac{K}{RC}\right)} = \frac{1}{jRC(\omega - \omega_0)} \quad (5)$$

and $-K$ is acquired from the formula above.

$$K = \omega_0 RC \quad (6)$$

For implementation, we use cross-coupled resistors for feedback. The cross-branches produce feedback currents, which are proportional to the output voltages, to be summed with input currents. Apply Kirchhoff's current law at point A, the resistance value is calculated as $\frac{1}{\omega_0 C}$.

On the basis of the above derivation, we choose the appropriate parameter of RC. This configuration provides a frequency-shift that moves the center frequency

to 350 kHz, with variable bandwidths of 60 kHz and 180 kHz. Fig. 4 demonstrates the complete fully differential active-RC complex filter.

Fig. 4. Fully-differential active-RC complex filter with compensation technique

3 Circuit implementation

3.1 Compensation method in OTA design

Fig. 5 shows a two stage CMOS fully-differential operational amplifier. The two-stage scheme enables the respectively design of output swing and differential gain. But the stability issue needs comprehensive deliberation. Especially in this multi-mode application with down-converted signal band up to 14 MHz, achieving a high open-loop gain of OTA and maintaining reliable stability may be difficult. Unfortunately, the conventional Miller compensation method in common use is unable to achieve relatively wide frequency utilization. To solve this problem, a special kind of Miller Compensation has been introduced.

Firstly, the conventional compensation method is presented in Fig. 6(a), R_z , which is used to cancel the zero on the right-plane introduced by the forward, and C_c are connected in series, with the equivalent capacitance of C_c at node O equals to $(A_2 + 1) C_c$. This splits the dominant pole and the non-dominant pole off to gain high phase margin.

Fig. 5. Proposed fully-differential OTA circuit

The proposed compensation network, as shown in Fig. 6(b), can reach higher open-loop gain when compared to the conventional one at the frequency of 14 MHz, or in another way, expand the bandwidth of OTA [8]. C_1 and C_2 are two capacitors with the same capacitance equal to $2C_c$. By dividing the output voltage with C_1 and R_a , there's less AC voltage across C_2 and less current needed for charging. Since the voltage division varies with frequency, and the equivalent impedance of C_2 is much higher than R_a at low frequency, the open-loop gain decreases 4.5 dB per decade until a gain of 15.7 dB is reached at 133.5 MHz, which is called the crossed frequency. Two frequency response curves intersect when the impedance of R_a equals to C_2 . The crossing frequency can be calculated as

Fig. 6. Comparison between two compensation method
(a) Conventional compensation (b) proposed compensation

$$f_{cross} = \frac{1}{4\pi R_a C_c} \quad (7)$$

At high frequency, the influence of R_a is negligible for its higher impedance compare to C_2 , so the frequency response can be similar to conventional Miller compensation's result. R_b is used for keeping the CMFB loop stable. Comparison result of open-loop frequency response of OTA with the conventional and the improved Miller compensation is given in Fig. 7.

This proposed OTA has a 63.3 dB DC gain and a 48.1 dB AC gain at 14 MHz. In terms of stability, 57.7° differential PM and 68.4° common mode PM are attained.

Fig. 7. Simulated amplitude and phase frequency response of the OTA with the conventional (curve 1) and the proposed (curve 2) Miller compensation schemes

Table I. Simulation results of OTA

	Condition	Min	Typical	Max	Units
I_{vdd}		1.05	1.07	1.09	mA
I_{off}		0.19	37	1150	nA
Voltage gain	@DC	58	63	73	dB
	@4 MHz		60		
	@14 MHz		48		
UGF	DM Loop		869		MHz
	CM Loop		56		
Phase Margin	DM Loop	52	58	59	degree
	CM Loop	66	68.4	69	
CMRR	@DC		97		dB
	@20 MHz		62		
Output Voltage Swing (V_{p-diff})			0.5		V

3.2 Tuning scheme

In active-RC filters, the frequency characteristic varies due to some factors like fabrication process, operating temperature and supply voltage. An automatic frequency tuning scheme should be employed on the chip to ensure characteristic's stability. Under normal circumstances, there are two major options for RC constant tuning, active and passive component tuning. We choose passive capacitor tuning since it builds high linearity for two reasons. Firstly, the change in the capacitance does not affect the linearity of the system, because it does not change the operating points of the transistors. Secondly, the capacitance of the MIM capacitor is linear since it is independent of the voltage across the capacitor [9]. To ensure the good relative precision with the filter, the tuning module has identical resistors and capacitors implemented on the same chip. By detecting and automatically adjusting its own RC constant, the tuning scheme outputs digital control words to change filter's RC constant. In order to meet the high precision of 2% tuning accuracy and $\pm 30\%$ range, we use the Binary Search Algorithm (BSA) with 5 control bits output.

The concept of capacitor tuning is illustrated in Fig. 8(a). The complete structure consists of an error amplifier, a comparator, a variable resistor, capacitor-array and digital tuning circuit.

The reference voltage is obtained from bandgap circuit and connected to the error amplifier and comparator. In order to derive a constant charging current, we use a OTA with high enough gain to guarantee the voltage on the variable resistor equals to V_{ref} , and I_2 is a duplication of I_1 with its value equals to V_{ref}/R_{ref} .

By charging and discharging the C-array through different phases and paths, the BSA module refers the compared signal of V_{cap} and V_{ref} to adjust the value of C-array. When the two noted voltages equal, the calibration process is ultimately fulfilled. To stretch the process a litter further, details of the workflow are elaborated. In the first phase, switch S_1 is closed with S_2 , S_3 open, C-array discharges through S_1 . Then, C-array charges through S_2 by I_2 . After that, S_3

Fig. 8. (a) Architecture of tuning circuit (b) Schematic of capacitors bank

closes and the BSA module change the value of C-array according to the output of comparator. If V_{cap} and V_{ref} are not compared to be equal, state machine go back to phase one and the cycle will be executed again. When $V_{cap} = V_{ref}$, condition $RC = nT_{clk} = T_{ch}$ will be met according to equation 8. Since T_{ch} is a constant, the RC will be tuned to a fixed constant eventually.

$$V_{cap} = \frac{Q}{C} = \frac{I_2 \Delta t}{C} = \frac{\frac{V_{ref}}{R} n T_{clk}}{C} = V_{ref} \quad (8)$$

3.3 Capacitors array design

Schematic of capacitors array is shown in Fig. 8(b), 5 bits of capacitor control words $CAP[0] \sim CAP[4]$ and 1 bit of complex mode enable signal decide the capacitance value. Equation 9 expresses the effective capacitance of C-array in LPF mode.

$$C_{array} = C_{fixed} + CAP[0] \times C_u + 2CAP[1] \times C_u + \dots + 2^4 CAP[4] \times C_u \quad (9)$$

To cover a range of $-20 \sim +20\%$ variation of resistors and capacitors in different corners, we choose

$$\frac{C_{max}}{C_{min}} = \frac{C_{fixed} + 31C_u}{C_{fixed}} = 2 \quad (10)$$

approximately. When the complex mode is disabled, tuning step of the capacitors array is 64.16 fF, nominal capacitance is 3.33 pF, and the tuning range is 2.374~4.363 pF, with tuning accuracy equals to 1.92%. When complex mode is enabled, all parameters listed above should be multiplied by 3, with total capacitance of the array equals to 13.089 pF.

The scheme uses binary search algorithm to speed up the convergence of tuning process. It starts at an initial value of $CAP = 01111$. According to comparator's

output, the algorithm narrows the searching range by dividing the interval into two parts with equal numbers at the end of each cycle until $V_{\text{ref}} = V_{\text{CMP}}$. Once the tuning process is finished, the module sends a Cal.done signal and control word CAP to filter and the output register remains the same until the recalibration signal arrives. It only takes about 10 μs to complete the tuning process under a typical circumstance. A transient response in this case is shown in Fig. 9.

Fig. 9. Transient response of tuning circuit

4 Measurement results

The transceiver with multi-mode complex filter has been fabricated in TSMC 0.13 μm CMOS technology, the die photo is shown in Fig. 10. Metal-insulator-metal (MIM) capacitors and poly resistors are used for fabricating capacitors and resistors respectively. The filter occupies an area of $0.98 \times 1.09 \text{ mm}^2$ and the tuning module's size is $0.14 \times 0.27 \text{ mm}^2$.

Fig. 10. Microphotography of the transceiver

When working under a 1.3 V supply voltage, the filter consumes 15 mA current. The passband gain is about 0 dB, the stop band rejection is 25 dB at $1.5f_{\text{1dB}}$ and 34 dB at $2f_{\text{1dB}}$.

The frequency response in LPF mode after tuning is demonstrated in Fig. 11(a). We can see that the result is very close to the specification. For the reason that the network analyzer we are using has a start frequency of 300 kHz, which is not be able to measure the filter frequency response in CBPF mode, a spectrum analyzer is used instead. Result shown in Fig. 11(b) indicates that it is still satisfying in CBPF modes.

Fig. 11. (a) measured frequency response of the filter in LPF mode
(b) measured frequency response of the filter in CBPF mode

Table II. Performance summary of this work

Supply voltage	1.3 V
Power consumption	19.5 mW
Inband-IIP3	25 dBm
Noise Figure	28.8 nv/ $\sqrt{\text{Hz}}$ (LPF Mode @10 MHz) 127 nv/ $\sqrt{\text{Hz}}$ (CBPF Mode, @350 kHz)
Passband gain	0 dB
IRR	40 dB
Tuning Accuracy	$\pm 2\%$
Tuning Range	$\pm 30\%$
Typical Tuning Time	13 μs

In terms of linearity, when measured with two input tones of 6 MHz and 8 MHz, the filter achieves in-band 25 dBm IIP3 with 50 Ω source impedance. The input referred noise density is 127 nv/ $\sqrt{\text{Hz}}$ (@350 kHz) and 28.8 nv/ $\sqrt{\text{Hz}}$ (@10 MHz) when in CBPF and LPF mode respectively. The measurement result also shows that the IRR exceeds the expected performance of 40 dB. Lastly, Table II summarizes the measurement results of the filter.

Table III. Comparison with previous reported works

Reference	JSSC2007 [10]	JSSC2009 [11]	JSSC2015 [5]	This work
Technology	0.13 μ CMOS	0.18 μ CMOS	40 nm CMOS	0.13 μ CMOS
Type	LPF	LPF	BPF	LPF/CBPF
Architecture	Active-RC	Gm-C	Active-RC	Active-RC
Supply voltage	1.2 V	1.2 V	1.5 V	1.3 V
cutoff frequency	0.35~23.5 MHz	0.5~20 MHz	7~56 MHz (centered at 140 MHz)	60, 180 KHz 2.8~14 MHz
Number of modes	-	-	4	8
IIP3	20 dBm	22.3 dBm	25.9 dBm	25 dBm
Power	21.6 mW	11.1 mW	50.5 mW	19.5 mW

5 Conclusion

This paper presents a 5-th Chebyshev complex filter with auto tuning scheme. This filter has 8 different modes including complex-BPF and real-LPF, and the bandwidth varies from 60 k to 14 M. It is tested that the multi-mode property makes the filter very suitable for WSN application. It draws 15 mA current under a 1.3 V supply voltage. The filter achieves high linearity of 25 dBm in-band IIP3 with 50 source impedance. Furthermore, an accurate frequency tuning scheme is applied. The tuning accuracy is 2% and it covers a range of 30%.

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