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An Active-RC Complex Filter with Tuning Circuits for Wireless Sensor Networks

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Abstract

A fourth order Butterworth active-RC complex filter with tuning circuits for wireless sensor networks (WSN) is presented in this paper. A method to synthesize an active-RC complex filter is proposed. A tuning system for the complex filter is designed, which is based on tuning the bit number of capacitors bank to compensate RC variations. The operational amplifier (OPA) is designed, which the gain bandwidth (GBW) is about 380MHz, the gain of the DC is 58.1dB, the phase margin is 58° *and the power dissipation of the operational amplifier is about 0.6mW. The complex filter is implemented in BCD 0.18 um CMOS process. The complex filter has a bandwidth of 2 MHz, and centre frequency is 2MHz. The results of simulation show that the complex filter image rejection ratio is -48.2dB, pass-band gain is 0.379dB, the power consumption is about 4.98mW.*

Keywords*: active-RC complex filter, wireless sensor networks, tuning circuits, operational amplifier*

1. Introduction

Wireless sensor networks have become more popular in some areas such as data collection, data sampling and so on. Because wireless sensor networks are used to monitor a given field of interest for changes in the environment, so they are very useful for military, environmental and scientific applications [1-3]. Wireless sensor networks consist of lager amounts of wireless sensor nodes, so there are a lot of noises in the wireless sensor networks, therefore, strong adaptability, comprehensive sensing coverage and high fault tolerance are the important performance indicators. In order to make wireless sensor networks have the performance indicators, the integrated filters are necessary for the WSN [4-6].

An active-RC complex filter not only can suppress the image noise, but also has excellent performances, such as linearity, dynamic range, and low power consumption under low voltage supply. Therefore, the active-RC complex filter is widely used in the modern wireless sensor networks [7].

This paper introduces a method of how an active-RC low pass filter synthesized to an active-RC complex filter. In order to ensure the center frequency of complex filter is 2MHz, the tuning system is designed, which is composed of analog circuits and digital circuits. Operational amplifier is an importance active device in complex filter, and OPA directly impacts on the performances of the complex filter, Therefore, a low power consumption, wide gain bandwidth of operational amplifier is designed for complex filter. The complex filter is implemented in BCD 0.18um CMOS process, the results of simulation show that the complex filter has a good performance.

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2. Architecture of Wireless Sensor Network Receiver

Today, the low intermediate frequency and zero intermediate frequency architectures are widely applied in wireless sensor networks receivers [8]. Zero intermediate frequency conversion receivers suffer from flicker noise and DC offset problems which cause significant degradation in signal to noise ratio [9-10]. As the shown in Figure 1, in low intermediate frequency architecture, not only the low intermediate frequency conversion receiver has a good trade-off in flicker noise and DC offset, but also when the down-conversion to low intermediate frequency, complex filter has the ability to suppress image noise.

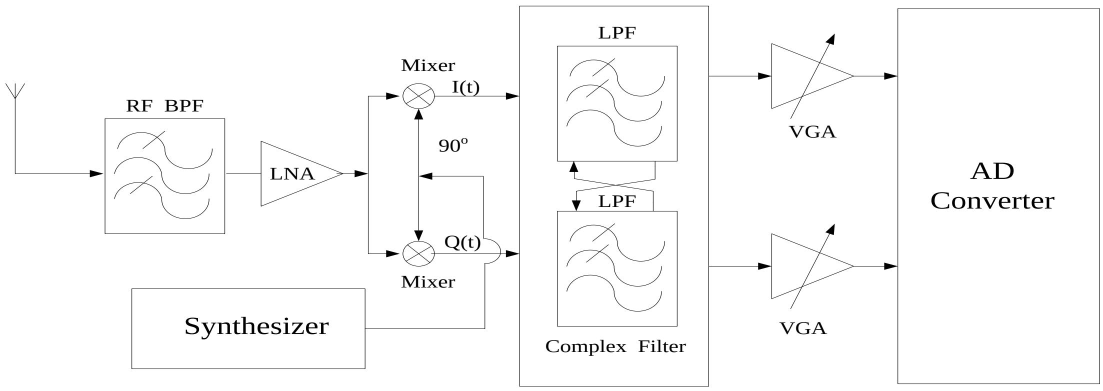

Figure 1. The Architecture of WSN Receiver

As shown in Figure 2, when the RF signals pass through the RF BPF and low noise amplifier, the positive and negative sidebands of the useful signals and image noise are yielded. Through the quadrature down-conversion, the signals become positive sideband signals. In order to reject the image noise, the complex filter is necessary.

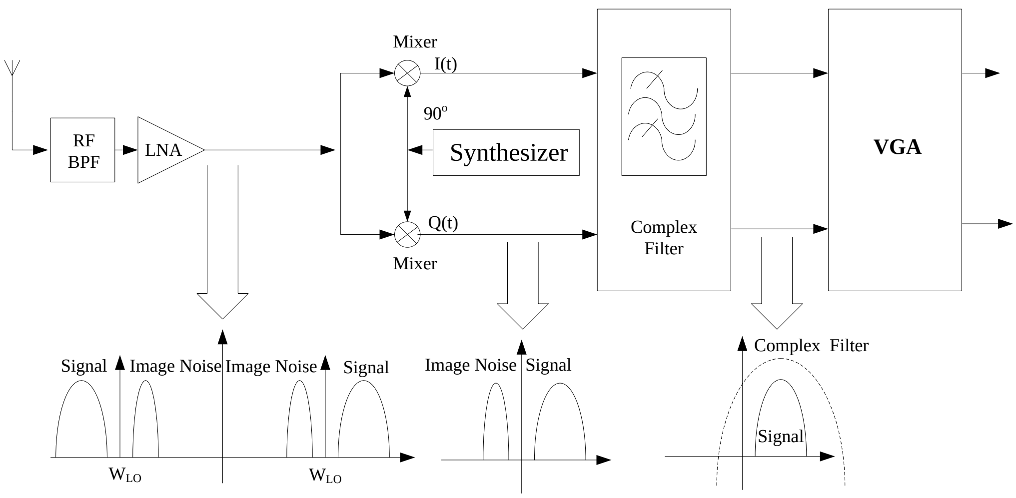

Figure 2. Diagram of Signal Processing

3. Complex Filter Design

3.1. The First Order Complex Filter Design

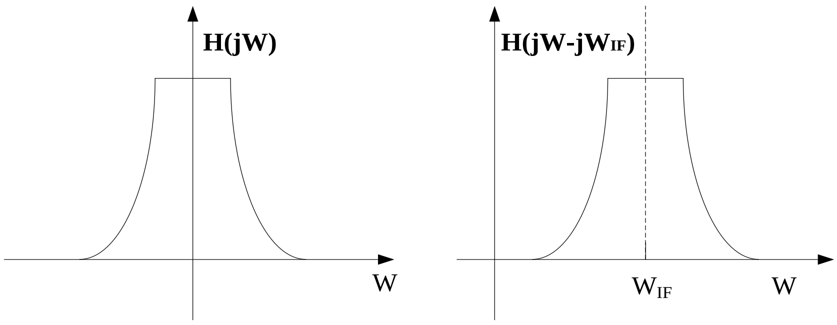

Figure 3. Low Pass Filter to Complex Filter

As shown in Figure 3, a frequency-shifting technology shifts the midpoint frequency of prototype low pass filter to the center frequency W_{IF} along the direction of the frequency axis. As a result, the frequency response is symmetrical with respect to the center frequency W_{IF} . The transfer function of the prototype low pass filter can be expressed as the following equation:

$$H_{LPF}(s) = \frac{W_A}{s + W_B} \qquad (1)$$

Through the frequency-shifting technology, the transfer function of complex filter can be obtained as:

$$H_{CF}(s) = \frac{W_A}{(s - jW_{IF}) + W_B} \qquad (2)$$

Figure 4 shows that the complex filter is designed by returning the *Vout* of low pass filter with a coefficient $j(W_{IF}/W_A)$ to the input of the low pass filter. As the shown in Figure 4, the transfer function is expressed as:

$$Vout = (\frac{W_A}{s + W_B}) \cdot (Vin + j\frac{W_{IF}}{W_A}Vout) \qquad (3)$$

$$Vout = \frac{W_A}{(s - jW_{IF}) + W_B}Vin \qquad (4)$$

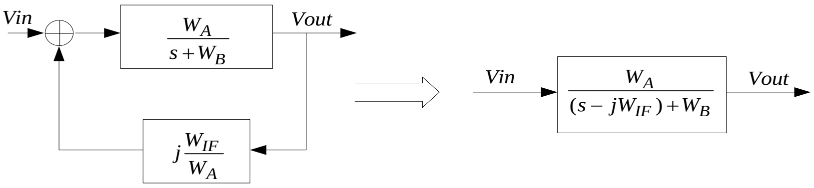

Figure 4. Block Diagram of the Complex Filter

As the shown in Figure 5，the transfer function of the first order complex filter can be expressed as:

$$VI_{out} = (VI_{in} - \frac{W_{IF}}{W_A} VQ_{out}) \cdot (\frac{W_A}{s + W_B}) \quad (5)$$

$$VQ_{out} = (VQ_{in} + \frac{W_{IF}}{W_A} VI_{out}) \cdot (\frac{W_A}{s + W_B}) \quad (6)$$

From equations (5) and (6), the transfer function of the first order complex filter $H_{CF}(jw)$ can be calculated as:

$$H_{CF}(s) = \frac{VI_{out} + jVQ_{out}}{VI_{in} + jVQ_{in}} = \frac{W_A}{s - jW_{IF} + W_B} \quad (7)$$

Where, $$W_A = \frac{1}{R_A C_P}, \quad W_B = \frac{1}{R_B C_P}, \quad W_{IF} = \frac{1}{R_{IF} C_P}. \quad (8)$$

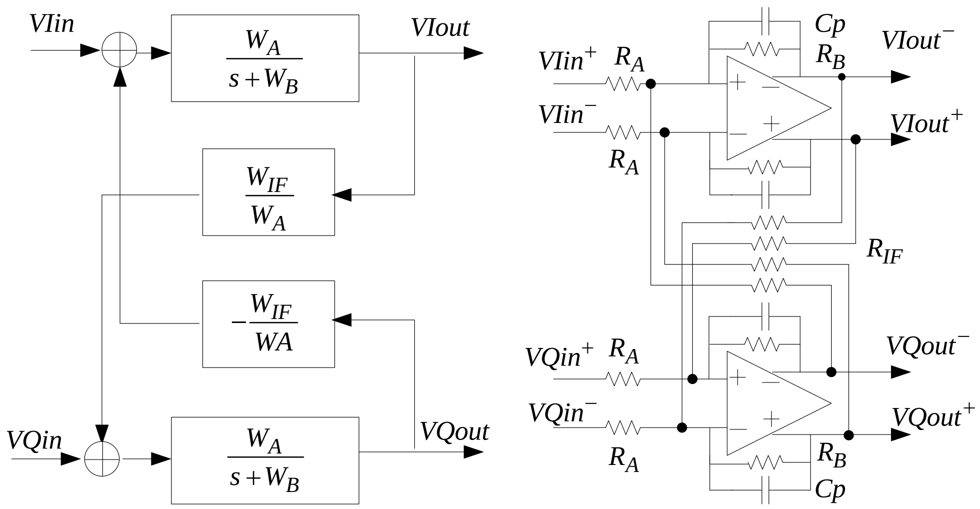

Figure 5. Implement Circuits of the First Order Complex Filter

Comparing equations (4) and (7), the first order complex filter can be designed by frequency-shifting technology and can be implemented by active devices.

3.2. The Fourth Order Complex Filter Design

The fourth order complex filter is shown in Figure 6, which is cascaded by two second-order complex filters. Therefore, the transfer function of the fourth order complex filter can be expressed as the following equation:

$$H_{CF4}(s) = H_{CF21}(s) \cdot H_{CF22}(s) \quad (9)$$

Where, $H_{CF21}(s)$ and $H_{CF22}(s)$ are the transfer functions of the second order complex filters. The $H_{CF21}(s)$ is also cascaded by two first-order complex filters, and the $H_{CF21}(s)$ is expressed as:

$$H_{CF21}(s) = \frac{W_{A21}}{(s - j\lambda_1 W_{IF}) + W_{B21}} \cdot \frac{W_{A21}}{(s - j\lambda_2 W_{IF}) + W_{B21}} \tag{10}$$

Where, $\lambda_1 W_{IF}$ and $\lambda_2 W_{IF}$ are the shifting-frequencies.

The $H_{CF22}(s)$ can be expressed as:

$$H_{CF22}(s) = \frac{W_{A22}}{(s - j\beta_1 W_{IF}) + W_{B22}} \cdot \frac{W_{A22}}{(s - j\beta_2 W_{IF}) + W_{B22}} \tag{11}$$

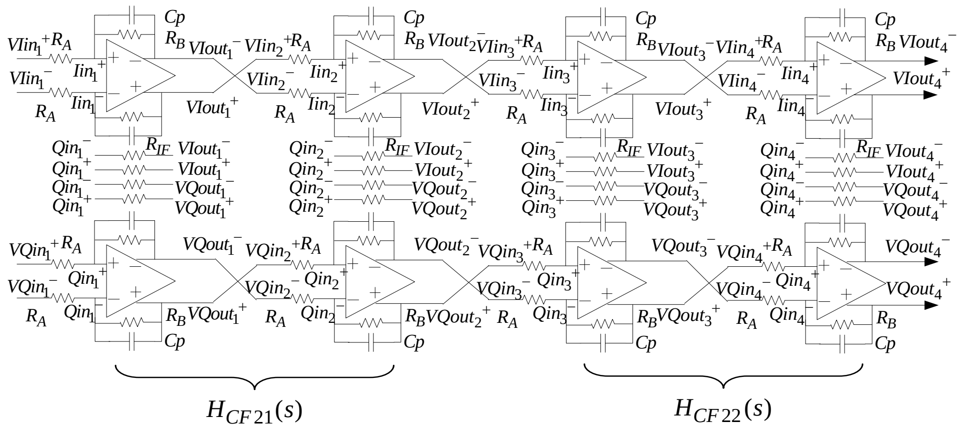

Figure 6. The Fourth Order Complex Filter

In equation (10), assuming the $\lambda_1 W_{IF} + \lambda_2 W_{IF} = 2W_{IF}$, the $H_{CF21}(s)$ can be expressed as:

$$\begin{aligned} H_{CF21}(s) &= \frac{W_{A21}}{(s - j\lambda_1 W_{IF}) + W_{B21}} \cdot \frac{W_{A21}}{(s - j\lambda_2 W_{IF}) + W_{B21}} \\ &= \frac{W_{A21}^2}{(s - jW_{IF})^2 + 2W_{B21}(s - jW_{IF}) + (1 - \lambda_1\lambda_2)W_{IF}^2 + W_{B21}^2} \end{aligned} \tag{12}$$

The prototype fourth order Butterworth low pass filter is expressed as:

$$\begin{aligned} H_{LPF4}(s) &= H_{LPF21}(s) \cdot H_{LPF22}(s) \\ &= \frac{a_{10}}{b_{12}s^2 + b_{11}s + b_{10}} \cdot \frac{a_{20}}{b_{22}s^2 + b_{21}s + b_{20}} \end{aligned} \tag{13}$$

Where, $H_{LPF21}(s)$ and $H_{LPF21}(s)$ are prototype second order Butterworth low pass filters. Thus, the $H_{LPF21}(s)$ can be expressed as:

$$H_{LPF21}(s) = \frac{a_{10}}{b_{12}s^2 + b_{11}s + b_{10}} \tag{14}$$

Then, after frequency-shifting W_{IF} , the he $H_{LPF21}(s)$ can be expressed as:

$$H_{(LPF-CF)21}(s) = \frac{\frac{a_{10}}{b_{12}}}{(s - jW_{IF})^2 + \frac{b_{11}}{b_{12}}(s - jW_{IF}) + \frac{b_{10}}{b_{12}}} \tag{15}$$

Comparing the transfer functions $H_{CF21}(s)$ and $H_{(LPF-CF)21}(s)$, assuming that equation (12) is equivalent to (15), these can be obtained:

$$W_{A21} = \sqrt{\frac{a_{10}}{b_{12}}}, \quad W_{B21} = \frac{1}{2}\frac{b_{11}}{b_{12}}$$

$$\lambda_1 \cdot \lambda_2 = (W_{B21}^2 - \frac{b_{11}}{b_{12}})/W_{IF}^2 + 1$$

$$\lambda_1 = 1 - \sqrt{1 - \lambda_1\lambda_2}$$

$$\lambda_2 = 1 + \sqrt{1 - \lambda_1\lambda_2} \tag{16}$$

According to equation (8):

$$R_{A21} = \frac{1}{\sqrt{\frac{a_{10}}{b_{12}}}C_P}, \quad R_{B21} = \frac{1}{\frac{1}{2}\frac{b_{11}}{b_{12}}C_P}$$

$$R_{IF1} = \frac{1}{\lambda_1 W_{IF} C_P}, \quad R_{IF2} = \frac{1}{\lambda_2 W_{IF} C_P} \tag{17}$$

As the same method, the $H_{CF22}(s)$ can be obtained as:

$$R_{A22} = \frac{1}{\sqrt{\frac{a_{20}}{b_{22}}}C_P}, R_{B22} = \frac{1}{\frac{1}{2}\frac{b_{21}}{b_{22}}C_P}$$

$$R_{IF3} = \frac{1}{\beta_1 W_{IF} C_P}, \quad R_{IF4} = \frac{1}{\beta_2 W_{IF} C_P} \tag{18}$$

4. Operational Amplifier Design

Because the centre frequency of the complex filer is 2MHz and the bandwidth is 2MHz, therefore, the cut-off frequency W_C of the complex filter is 3MHz. In order to attain the digital demodulation circuit requirements, the ripple in pass-band is less than0.5dB, the GWB of operational amplifier should be greater than $100W_C$. Thus, the operational amplifier is two-stage fully differential OPA with miller compensation capacitance and common mode feedback as shown in Figure 7. In such structure, the first stage provides high gain, the second stage offers a large voltage swing.

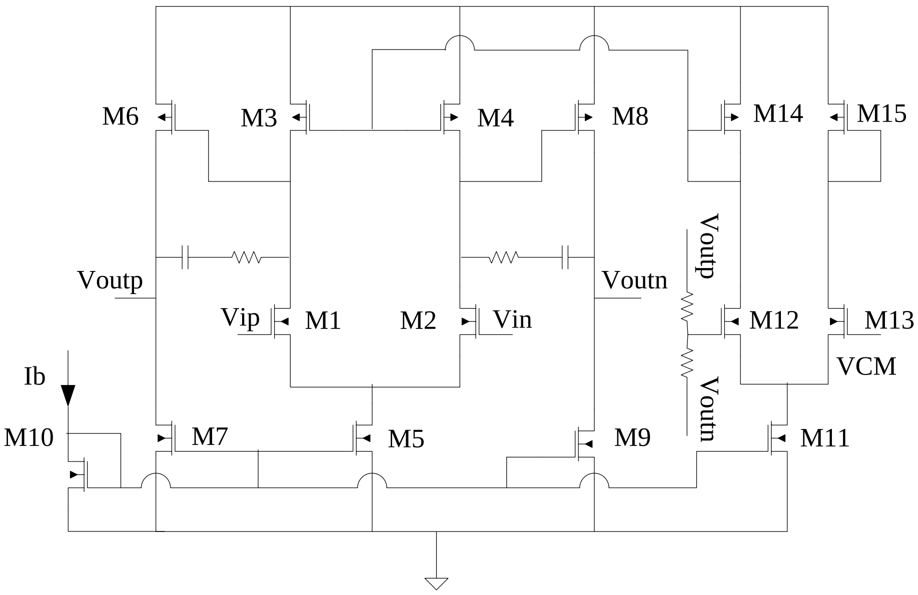

Figure 7. Schematic of Operational Amplifier

The simulation results of different process corners are shown in Table 1. As the shown in Table 1, the minimum GWB of operational amplifier is 335MHz, which is greater than $100W_C$, the maximum power consumption is 0.63mW, the minimum gain is 56.2dB, and the minimum phase margin is $57°$. These parameters make sure the operational amplifier to satisfy the requirements of complex filter.

Table1. Simulation Results of the Complex Filter

	tt	ss	ff
GBW	384MHz	335MHz	456MHz
gain	58.1dB	58dB	56.2dB
phase	58°	57°	59°
Power	0.58mW	0.55mW	0.63mW

5. Tuning System Design

As the variations of manufacturing process and environment temperature, the values of capacitors and resistances have significant changes, which may reach 20%, and the RC time constant of the complex filter has a change. Thus, the center frequency and bandwidth of the complex filter are drifted. In order to conquer the problem, a tuning system is designed.

As the shown in Figure 8, the tuning system is composed of power block, analog block and digital block.

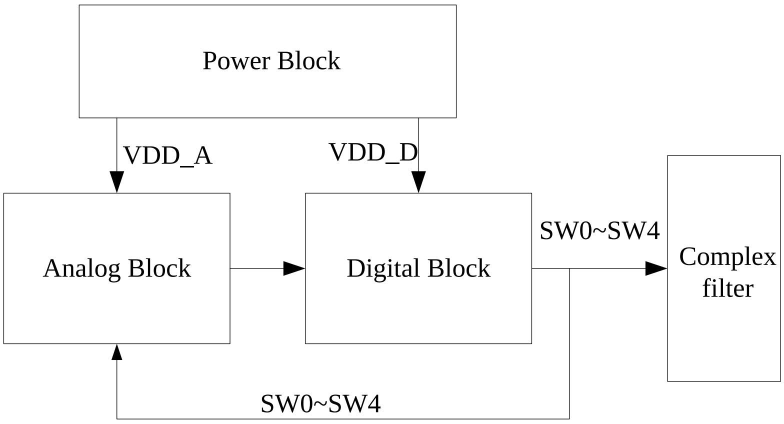

Figure 8. Block Diagram of the Tuning System

The analog block is shown in Figure 9. In order to reduce mismatches, the reference resistance R_{ref} should be chosen as the same kind of the resistance of the complex filter, and the capacitors of tuning system should also be the same as the capacitors of complex filter.

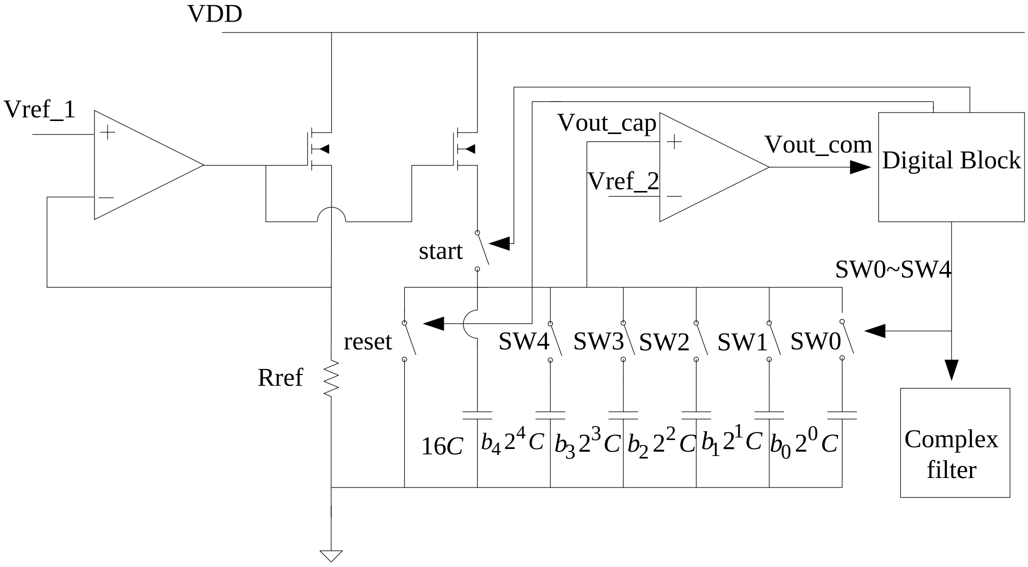

Figure 9. Schematic of Tuning System Analog Block

The capacitor bank of the tuning system can be calculated as:

$$C = C_{fix} + C_0 \sum_{i=0}^{i=4} b_i \cdot 2^i$$

$$C_{\max} = C_{fix} + 31C_0$$

$$C_{\min} = C_{fix} \qquad (19)$$

In order to cover 20% variation of the capacitors and resistances, the C_{fix} should be obtain as

$$C_{fix} = 16C_0 \quad (20)$$

The output voltage of the capacitor bank can be calculated as:

$$V_{out\_cap} = \frac{V_{ref1}}{R_{ref}} \cdot \frac{1}{C_{fix} + b_4 2^4 C_0 + b_3 2^3 C_0 + b_2 2^2 C_0 + b_1 2^1 C_0 + b_0 2^0 C_0} \cdot T \quad (21)$$

AS the shown in Figure 10, when the start signal is beginning, the $SW_4 \sim SW_0$ is 10000, if $V_{out\_cap} > V_{ref2}$, it proves that the output voltage of capacitor bank is large and the value of the capacitor bank is small, thus, in the next clock, the $SW_4 \sim SW_0$ is 11000, and if $V_{out\_cap} < V_{ref2}$, it proves that the voltage of capacitor bank is small and the value of the capacitor bank is large, in the next clock, the $SW_4 \sim SW_0$ is 01000. The process will repeat until the completion of the tuning. The digital logic control block is shown in Figure 10.

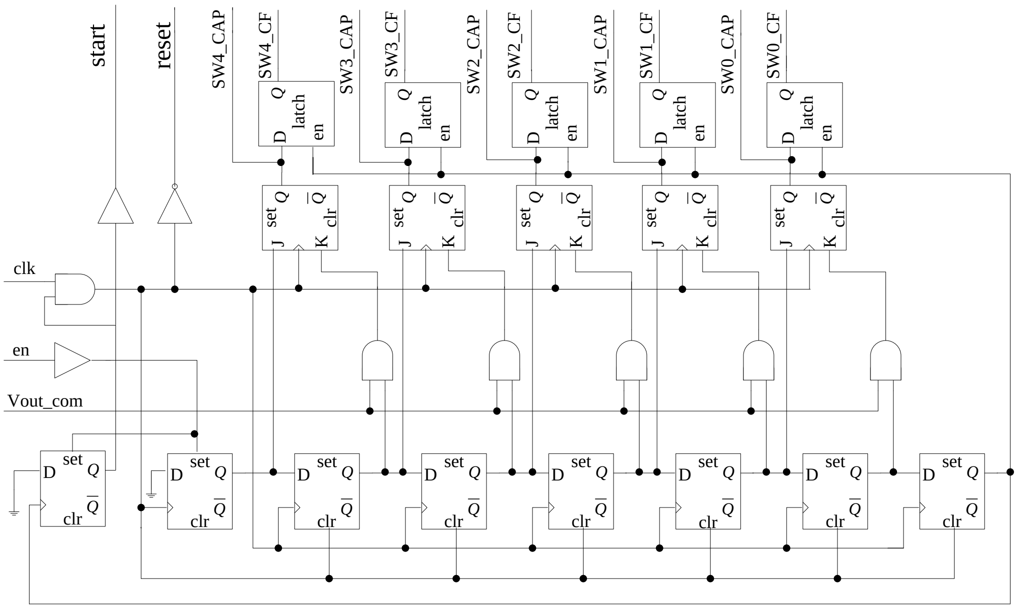

Figure 10. Schematic of Digital Logic Control Block

6. Implementation and Simulation

The complex filter and the tuning circuits have been implemented in BCD 0.18 um CMOS process. In order to simulate the frequency characteristic, the quadrature sinusoidal signals are sent into the complex filter. The quadrature sinusoidal signals are in four different phases, the phase of signals are $0°$, $180°$, $90°$ and $270°$. The simulation is shown in Figure 11.

Figure 11. Simulation of the Complex Filter

The Figure 11 shows that the center frequency is 2MHz, the pass-band gain is about 0.379dB, the pass-band is between 1MHz and 3MHz, the gain at the 1MHz is about -2.95dB, and the gain at the 3MHz is about -3.8dB. The power consumption of the complex filter is about 4.98mW and the image rejection ratio of the complex filter is more than -48.2 dB.

After tuning and before tuning waveforms are shown in Figure 12. From the simulation waveforms can be obtained that center frequency of complex filter has been adjusted to 2MHz.

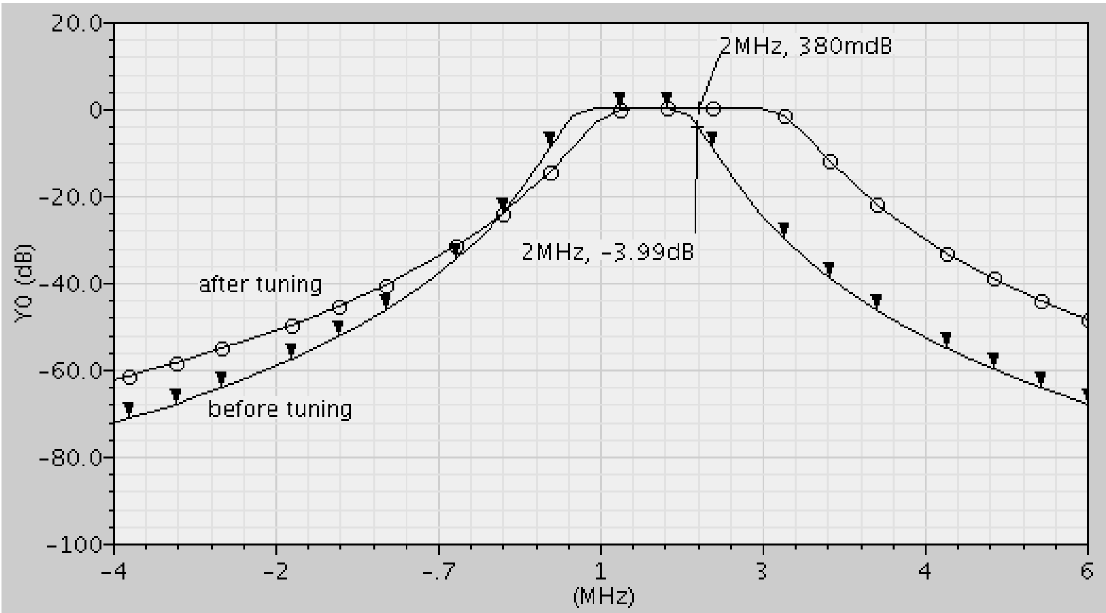

Figure 12. Simulation of After Tuning and Before Tuning

The performance of summary of the complex filter is shown in Table 2

Table 2. Performance Summary of the Complex Filter

Technology	BCD 0.18 um
Supply Voltage	1.8V
Center frequency	2 MHz
Pass-band Gain	0.379dB
Bandwidth	2 MHz
Gain @ 1MHz	-2.95dB
Gain @ 3MHz	-3.8dB
Image Rejection Ration	-48.2dB
Power consumption	4.98mW

7. Conclusion

This paper proposes an active-RC complex filter for wireless sensor networks, which is implemented in BCD 0.18um CMOS process. A method to synthesize an active-RC complex filter is introduced. In order to adjust the RC time constant, the tuning system is designed for complex filter. A two-stage fully differential operational amplifier is designed, the GBW of the operational amplifier is more than 380MHz and the power consumption is about 0.6mW. The results of simulation show that the center frequency of the complex filter is 2MHz, the bandwidth is 2MHz, the image rejection ration is -48.2dB, and the power consumption is about 4.98mW.

Acknowledgements

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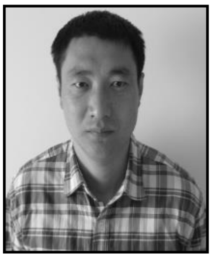

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