

2 GHz Bandstop Active Filter using Operational Transconductance Amplifiers

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Abstract—A three-resonator bandstop active filter is designed in CMOS technology. The filter has a notch at 2 GHz and it exhibits a rejection of -44 dB at that frequency. The general approach taken here was to design the filter using lumped passive elements and then to replace the inductors in the circuit with OTA-based active inductors. Microwave active filters are of interest in order to reduce chip area because at those frequencies the size of a filter is directly related to the wavelength of the signal of interest, which can be on the order of centimeters at the lower end of the microwave spectrum.

I. INTRODUCTION

The integration of entire microwave radio systems on a single chip has been the predominant trend in radio frequency integrated circuit (RFIC) design over the last decade. Among the last remaining components in radio systems that are often found off-chip are the filtering circuits. Thus, the realization of microwave filters in integrated circuit (IC) form is very attractive due to the enormous reduction in the physical size of the filter when compared with microstrip, specially at the lower end of the microwave spectrum where wavelengths are large. An abundant number of microwave IC bandpass filter designs exist using various technologies such as piezoelectric surface and bulk acoustic wave devices [1], CMOS [2], and compound semiconductor transistors [3]. In contrast to bandpass filters, substantially less *bandstop* IC filters have been reported to date.

In this work, we present the design of a 3-resonator microwave active bandstop filter using 0.18 μm CMOS technology. The filter uses lumped element inductor-capacitor (LC) resonators in which passive inductors are replaced by active inductors. The active inductors are implemented using a set of operational transconductance amplifiers (OTA's) arranged in a mutual feedback configuration. Since the active inductor values depend fundamentally on the transconductance, G_m , of the OTA cells, we used a variable current source in order to precisely tune the inductors to their required value. The complete filter was simulated and it exhibits a notch at 2 GHz with a rejection of -44 dB at that frequency.

II. INTEGRATED CIRCUIT DESIGN

The design of the active bandstop filter can be synthesized into a four-step process. The first and second steps are to create

a normalized LC lowpass filter and then to convert it to a highpass prototype using shunt inductors and series capacitors. The third step is to transform the highpass prototype into a bandstop filter by replacing the shunt inductors with series LC resonators and the series capacitors are replaced with parallel LC resonators [4]. The fourth and final step is to replace the passive inductors with active inductors using OTA's. The entire process is summarized graphically in Fig. 1.

A. Lowpass and Highpass Prototype Filters

First, a 3rd order Butterworth normalized lowpass ladder filter ($Z_o = 1 \Omega$ and $\omega_c = 1 \text{ rad/sec}$) was designed. The element values for a normalized Butterworth filter are well known and they can be calculated analytically or found in pre-computed tables [4] [5]. Next, the highpass filter was created by converting capacitors into inductors and vice versa in the lowpass prototype. The new values of the inductances and capacitances in the highpass filter are obtained by taking the reciprocal values of the lowpass filter elements. To complete the highpass filter design, the component values need to be scaled to the desired cutoff frequency, meaning that both inductors and capacitors need to be divided by the factor $2\pi f_h$, where f_h is the -3 dB cutoff frequency of the highpass filter.

B. Highpass to Bandstop Transformation

The bandstop filter is obtained from the highpass prototype by transforming the shunt inductors into series resonators and the series capacitors into parallel resonators. The resonant frequency of all the resonators is the same, $2\pi f_o$, with f_o being the desired filter notch frequency. For the series resonators the relevant equations are,

$$L_{bs} = L_{hp} \quad (1)$$

$$C_{bs} = \frac{1}{L_{bs}(2\pi f_o)^2} \quad (2)$$

and for the parallel resonator case we have,

$$C_{bs}^p = C_{hp} \quad (3)$$

$$L_{bs}^p = \frac{1}{C_{bs}^p(2\pi f_o)^2} \quad (4)$$

where 'bs' stands for bandstop and 'hp' stands for highpass. The procedure is similar to the one used for lowpass to bandpass filter transformations, but there is an important

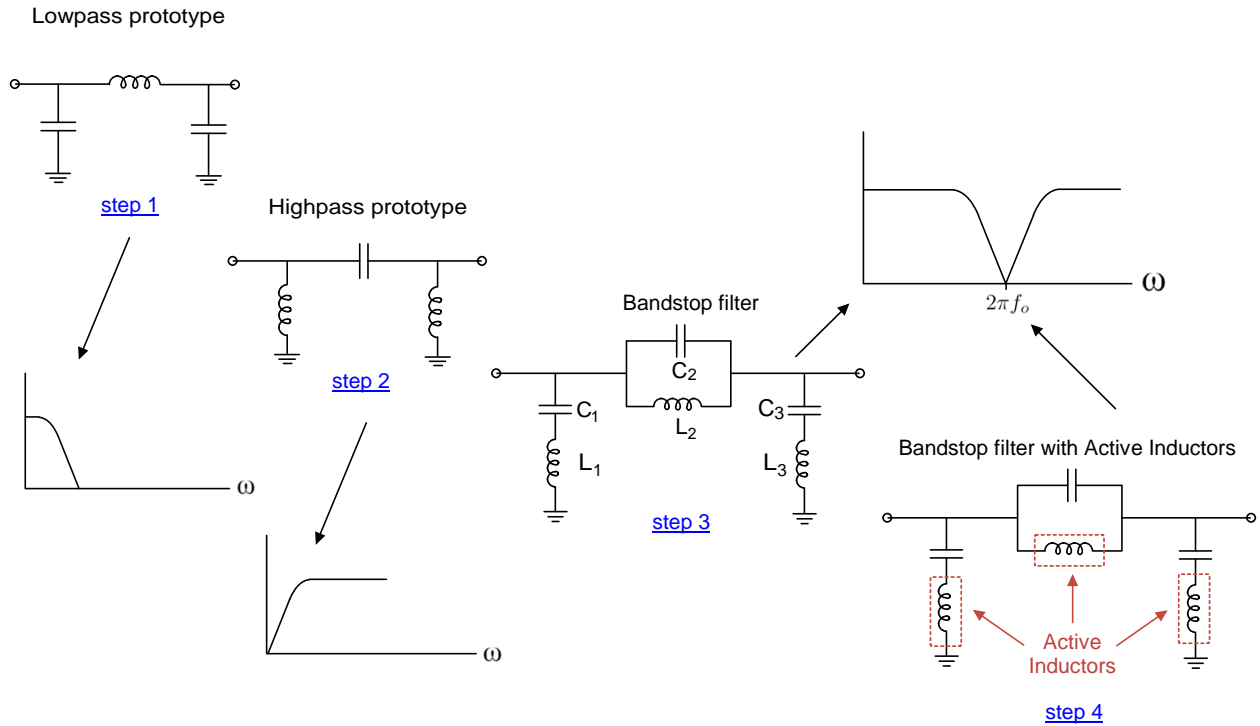


Fig. 1. Active bandstop filter design process

difference: in the bandpass case, Eqn. 1 and Eqn. 3 would need to be divided by a factor of 2. The last step in this phase of the design is to scale the filter characteristic impedance, Z_o , from 1Ω to the desired value.

For a filter with a notch frequency of 2 GHz and a bandstop range of 1.2 GHz, the inductor needed for the parallel resonator is 4.2 nH and the inductor value needed for the two series resonators is 8.6 nH. In principle, the 4.2 nH inductor can be realized on-chip using a spiral structure but the 8.6 nH inductor is very large and is much less attractive for passive on-chip implementation. For this reason we investigated the use of active inductors to replace the passive ones.

C. Active Inductor Design

The active inductors are realized using OTA impedance inverters [6], as shown in Fig. 2. The difference between the two circuits in Fig. 2 is that one is a differential inverter while the other is a single-ended inverter. For both circuits, the input impedance is given by

$$Z_{in} = \frac{1}{G_{m1}G_{m2}Z_L} \quad (5)$$

where G_{m1} and G_{m2} are the transconductances of the individual OTA cells. When the load impedance is capacitive, $Z_L = 1/j\omega C_L$, then the input impedance Z_{in} becomes inductive and this is how the active inductor is made. If the OTA's are identically biased so that $G_{m1} = G_{m2} = G_m$, then the inductance is given by,

$$L_{act} = \frac{C_L}{G_{m1}G_{m2}} = \frac{C_L}{G_m^2}. \quad (6)$$

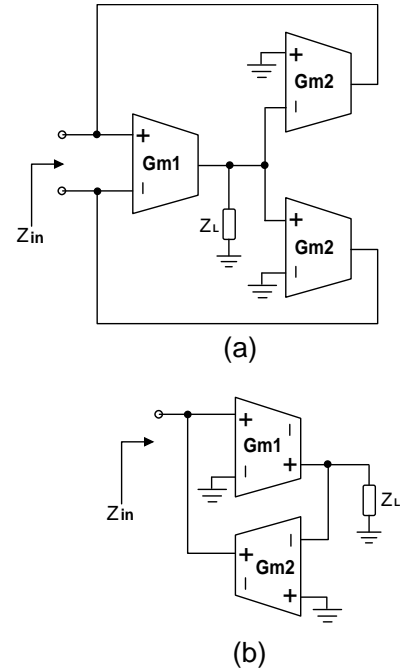


Fig. 2. Impedance inverters (a) differential (b) single-ended. After [6]

The OTA cell used in this work is a differential pair with active load as shown in Fig. 3. It is of significant interest to derive an analytic expression that relates the transconductance, G_m , of this OTA to its bias current control voltage, V_{GS1} .

We begin with the current-voltage relationship for a short-

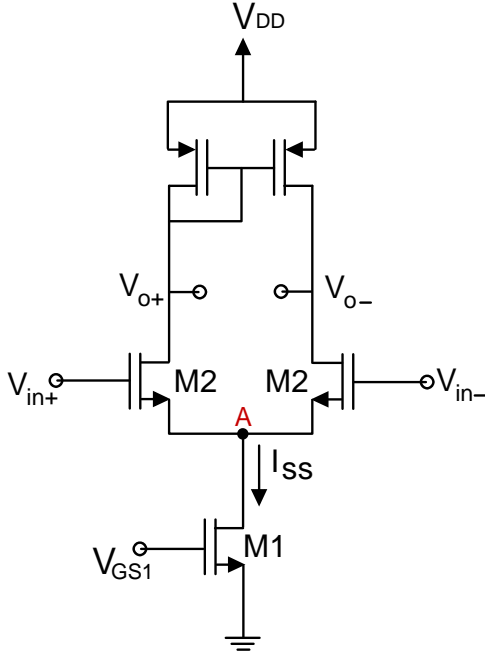


Fig. 3. Operational transconductance amplifier cell (biasing not shown)

channel device in saturation, which is given by [7],

$$i_{DS} = v_{sat} C_{ox} W (V_{GS} - V_{tn} - \frac{1}{2} V_{DSAT}) \quad (7)$$

where v_{sat} is the carrier saturation velocity, C_{ox} is the transistor gate capacitance, W is the gate-width, V_{GS} is the gate-source voltage, V_{tn} is the transistor threshold voltage, and V_{DSAT} is the voltage at which the carrier velocity saturation ensues. An approximate expression for V_{DSAT} that is often used is,

$$V_{DSAT} \approx L v_{sat} / \mu_n \quad (8)$$

where L is the transistor gate-length and μ_n is the electron mobility. For the OTA in Fig. 3, we can use Eqn. 7 to determine its transconductance as follows,

$$G_m = \frac{\partial i_{DS}}{\partial V_{GS2}} = v_{sat} C_{ox2} W_2. \quad (9)$$

Invoking Kirchoff's current law at node A in Fig. 3, one immediately concludes that $I_{ss} = 2i_{DS}$ and thus,

$$I_{ss} = 2v_{sat} C_{ox2} W_2 (V_{GS2} - V_{tn} - \frac{1}{2} V_{DSAT}) \quad (10)$$

Since I_{ss} itself is controlled by transistor M₁, then the following expression also holds,

$$I_{ss} = v_{sat} C_{ox1} W_1 (V_{GS1} - V_{tn} - \frac{1}{2} V_{DSAT}). \quad (11)$$

After doing some algebra with Eqns. 9–11, we determine that the relationship between G_m and V_{GS1} is given by,

$$V_{GS1} = \gamma G_m V_{GS2} + (V_{tn} + \frac{1}{2} V_{DSAT})(1 - \gamma G_m) \quad (12)$$

where $\gamma = 2/v_{sat} C_{ox1} W_1$.

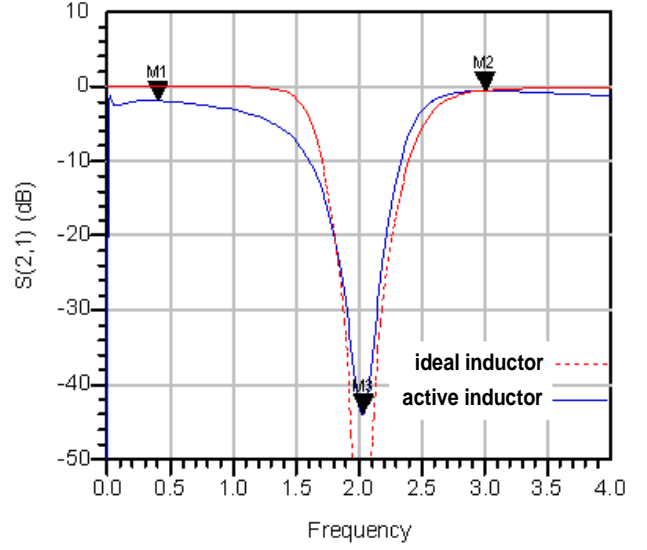


Fig. 4. Simulated bandstop filter response

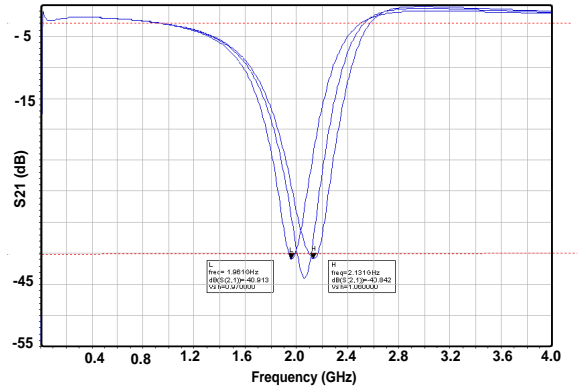


Fig. 5. Tuning capability of the bandstop filter

III. RESULTS

An integrated circuit filter was designed using TSMC's 0.18- μm CMOS technology. All simulations were carried out using the high-frequency simulator Advanced Design System (ADS) from Agilent Technologies. The filter had a notch at 2.0 GHz and a stop-band of 1.2 GHz. The component values are shown in Table I.

To realize the the 8.5 nH inductors the impedance inverter in Fig. 2a was used with a load capacitance of 0.07 pF. Using Eqn. 6, it is found that the required transconductance for the OTA's is $G_m = 2.9$ mS. To obtain this value of G_m , the DC voltage V_{GS1} applied to the gate of transistor M₁ in the OTA cells was 1.0 V. For the 4.2 nH inductance, which is grounded at one end, the inverter in Fig. 2b was chosen because it requires one less OTA. The same load capacitance of 0.07 pF was used and the necessary G_m for the OTA's is 4 mS, requiring a $V_{GS1} = 1.4$ V.

To illustrate the savings in chip-space, the area required by a 0.07 pF capacitor is on the order of $10 \mu\text{m} \times 10 \mu\text{m}$, and this is substantially less than would be required for a spiral

TABLE I
BANDSTOP FILTER COMPONENT VALUES

L ₁	C ₁	L ₂	C ₂	L ₃	C ₃
8.5 nH	0.74 pF	4.2 nH	1.5 pF	8.5 nH	0.74 pF

inductor, which is often greater than $100 \mu\text{m} \times 100 \mu\text{m}$.

Fig. 4 shows the frequency response of the bandstop filter using the active inductors. For comparison we have also plotted the response of the bandstop filter with passive inductors and the agreement is very good. The filter with the active elements has a rejection of -44 dB at 2 GHz.

Since the value of the active inductors can be varied by changing the G_m of the OTA's through the control voltage V_{GS1} , this means that the center frequency of the bandstop filter can be tuned. Fig. 5 shows the tuning capability of the filter. The notch frequency tuning range is 170 MHz and the rejection is better than -40 dB for all frequencies.

The DC current draw of the entire filter was 18 mA from a 1.8 V power supply, resulting in a total power dissipation of 32.4 mW.

IV. CONCLUSIONS

We have designed a bandstop filter with a center frequency of 2 GHz using $0.18 \mu\text{m}$ CMOS technology. The filter employs OTA impedance inverter circuits in order to realize high-valued active inductors. The active inductors are then used to replace passive ones in a 3-resonator bandstop filter.

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