1 GHz Opamp-Based Bandpass Filter

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***Abstract* — A biquad bandpass filter, operating at 1 GHz with 9 dB gain, was designed and fabricated in 180nm SiGe BiCMOS technology. It relies on a 11-GHz unity-gain-bandwidth, highly stable opamp implemented using a MOS-HBT cascode stage with cascode p-MOSFET load and common-mode-feedback. The filter has tunable bandwidth and tunable center frequency and marks a radical approach to the design a bandpass filter operating in the gigahertz range.**

***Index Terms* — opamp, bandpass filter, biquad, varactor, SiGe BiCMOS, tunable filter, unity gain bandwidth**

I. INTRODUCTION

Traditionally, on-chip RF bandpass filters have been implemented using RLC, gm-LC, and gm-C circuits. In contrast, this paper presents an opamp-based bandpass filter. The target center frequency of operation is 1 GHz and the target bandwidth is 500 MHz. The opamp design [1] is critical since its unity-gain-bandwidth UGB must extend at least 10 times beyond the center frequency of the filter. An opamp bandpass filter has the advantage of not requiring large area inductors to create the bandpass response [2].

II. CIRCUIT DESIGN

Fig. 1 shows the system block diagram of the chip. It is composed of an input buffer for 50 Ω impedance matching and a biquad filter that accomplishes the bandpass filtering function. The biquad filter consists of a cascade of two opamp integrators with resistive and capacitive feedback, forming a second order bandpass response.

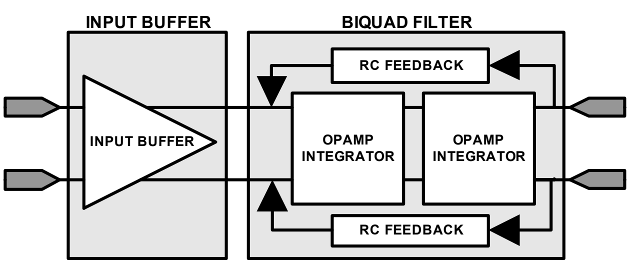

Fig. 1 Biquad bandpass filter system block diagram.

Fig. 2 illustrates a single-ended high Q biquad filter topology [3]. The circuit transfer function (1) shows that the filter has a second order bandpass response with two poles and a zero at the center frequency. The center frequency, bandwidth, and gain are determined by equations (2), (3), and (4) respectively.

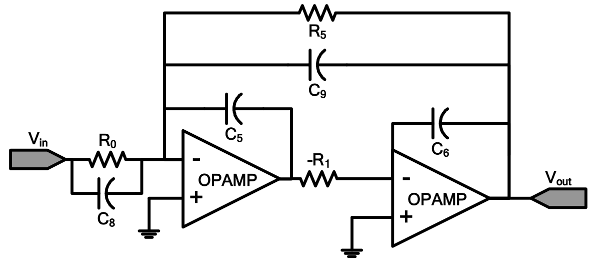

Fig. 2 Single ended high Q biquad filter topology.

$$H(s) = \frac{V_{out}}{V_{in}} = \frac{-R_0 R_5 C_8 s - R_5}{R_0 R_1 R_5 C_5 C_6 s^2 + R_0 R_5 C_9 s + R_0} \quad (1)$$

$$\omega_0 = \frac{1}{\sqrt{R_1 R_5 C_5 C_6}} \quad (2)$$

$$\mathrm{BW}_{3\mathrm{dB}} = \frac{C_9}{R_1 C_5 C_6} \quad (3)$$

$$\text{Center Frequency Gain} = \frac{C_8}{C_9} \quad (4)$$

Operating the bandpass filter with a center frequency of 1 GHz required a custom design of the opamp. Two identical large unity-gain-bandwidth opamps, whose schematic is shown in Fig. 3, were employed. The design was scaled from 130 nm to 180 nm SiGe BiCMOS technology while biasing the MOSFETs and HBTs at peak f_{MAX} current density [1]. The n-MOSFETs are biased at 0.2 mA/µm while the HBTs are biased at 1.1 mA/µm. This bias scheme ensures immunity to threshold voltage, temperature, bias current, and to gate length variation and is critical to realizing robust opamps in deep submicron CMOS/BiCMOS technologies [1]. In order to obtain high bandwidth with excellent phase margin [1][3], a BiCMOS cascode input stage was used. It consists of a 180-nm common-source n-MOSFET and a common base SiGe HBT which have the benefit of eliminating the Miller effect. Despite the relatively slow 180-nm MOSFET with f_T of 60 GHz, because the unity-gain bandwidth is dictated by the SiGe HBT [1], the single-ended opamp has a unity gain bandwidth exceeding 11 GHz, about one decade higher than the center frequency of the filter. A wide-swing p-MOS active load provides the large output resistance needed for high gain and allows for larger output voltage swing than a typical cascode active load [3]. Emitter followers were used in the output stage of the amplifier to provide level shifting between stages, low output impedance, and to form part of a common-mode feedback network.

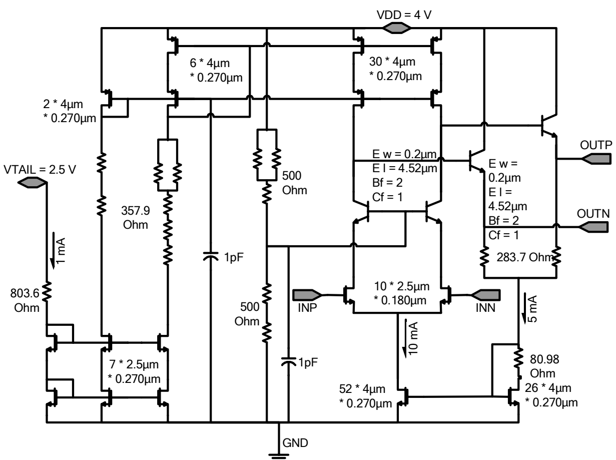

Fig. 3 Differential BiCMOS telescopic cascode opamp cell.

Both fixed frequency and tunable filters were designed and tested. The tunable filter circuit schematic is shown in Fig. 4. It features adjustable center frequency and adjustable bandwidth. PMOS transistors, connected as varactors, were employed for tuning. This was necessary in order to preserve the fixed DC voltage at the input and output of the biquad filter, while still being able to vary the control voltage of the varactor. Fig. 5 illustrates the connection scheme employed to convert a PMOS transistor into a three terminal PMOS varactor. The varactor was created by connecting the gate to one terminal of the opamp and then tying the source and drain together to the other terminal of the opamp. The varactor capacitance is tuned by applying the control voltage to the bulk terminal of the PMOS transistor. This changes its threshold voltage and, in so doing, modifies the capacitance. Due to the non-conventional way of controlling the varactor capacitance, the filter tuning range is limited to about 20%, which is adequate for compensating process and temperature variations. Referring to Fig. 4, adjusting the PMOS bulk terminal labeled VCTR_CTRL changes the center frequency of the filter, while adjusting the terminal VBW_CTRL tunes its bandwidth.

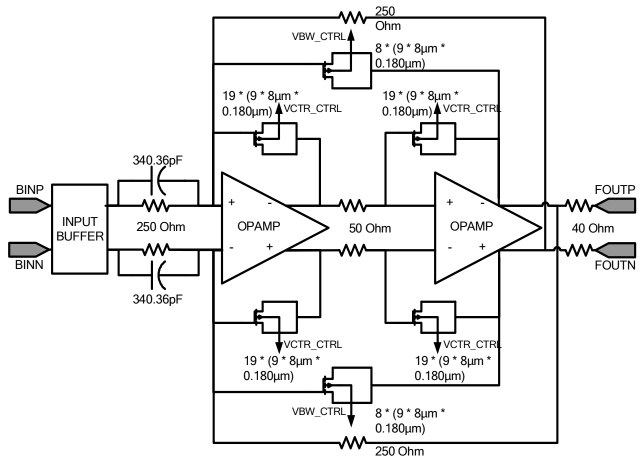

Fig. 4 Tunable bandpass filter schematic with PMOS varactors.

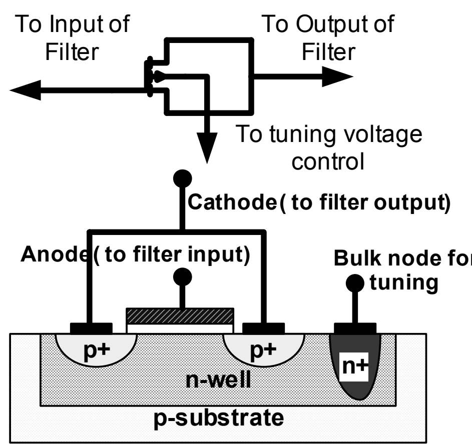

Fig. 5 Schematic connection of the three terminal p-MOS varactor used to accomplish center frequency and bandwidth tuning.

For testing purposes, both filters feature a 50-Ohm input buffer implemented with an n-MOSFET inverter biased at the peak f_T current density of 0.3 mA/µm in order to ensure high linearity and adequate gain. Fig. 6 shows the schematic diagram of the input buffer that was designed for 50 Ω input impedance matching. Due to the limited number of probe pads available, the 50-Ω resistors R0 and R1 could not be directly connected to a voltage rail and thus provide a perfect 50 Ω input impedance match. Instead they had to be connected through two 300 Ω biasing resistors R2 and R3. As a result, the S11 of the filter is somewhat degraded but remains lower than -13 dB.

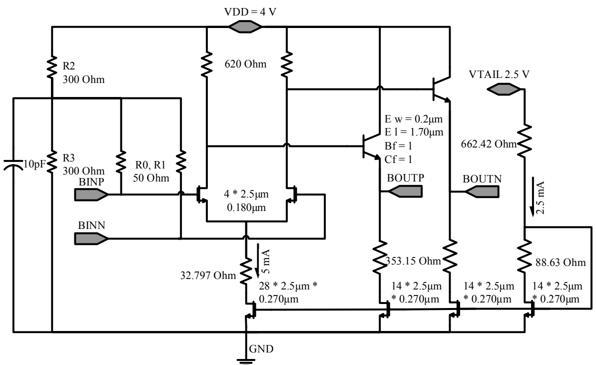

Fig. 6 Input Buffer Schematic.

III. EXPERIMENTAL RESULTS

S-parameter measurements for the two filters and for an opamp were conducted on wafer, single-endedly, where the unused input and output terminals were grounded externally through 50 Ω loads. A microphotograph of the tunable filter can be seen in Fig. 7. As shown in Fig. 8, the measured center frequency of the fixed frequency filter is 1 GHz and the bandwidth is 450 MHz. The center frequency is exactly on target, while the bandwidth is 10% lower than desired.

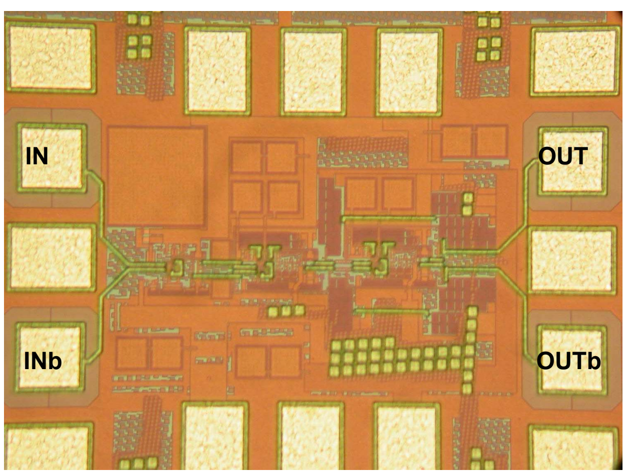

Fig. 7 Fixed frequency filter die photo. For reference, the HF pads are 60 µm x 60 µm.

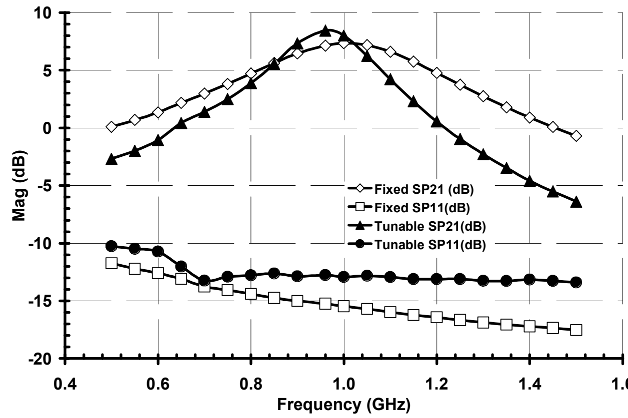

Fig. 8 Measured tunable filter and fixed frequency filter frequency response. S-parameter measurements were taken single-endedly where the second input and output terminals were grounded through 50 Ω loads.

Three sets of measurements were carried out for the tunable filter. The first measurement, shown in Fig. 8, was taken with VCTR_CTRL and VBW_CTRL fixed at 1.5 V. The center frequency is 970 MHz and the bandwidth is 220 MHz. The center frequency is just 0.3% off target. The difference in measured bandwidth versus the design target is likely due to the pessimistic Q of the varactor model. The center frequency is tunable between 890 MHz and 1.1 GHz when sweeping the bulk voltage VCTR_CTRL of the varactor from 1 V to 4 V and fixing VBW_CTRL at 1.5 V. The data from the third measurement are shown in Fig. 9. Sweeping the bulk voltage VBW_CTRL from 1 V to 4 V and fixing VCTR_CTRL at 1.8 V changes the bandwidth from approximately 160 MHz to 260 MHz. Note that adjusting the center frequency has the side effect of changing the bandwidth of the filter and vice versa. This is due to the fact that the varactor capacitance affects both the center frequency (2) and the bandwidth (3).

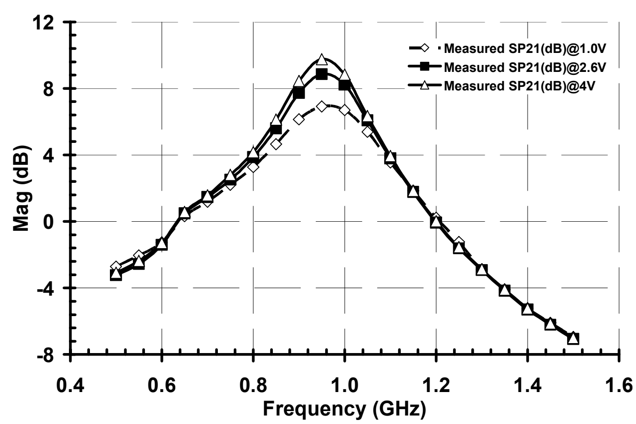

Fig. 9 Measured tunable filter bandwidth tuning.

The performance of the two filters is summarized in Table 1.

Specification	Measured result
Fixed Filter Center Frequency	1 GHz
Fixed Filter Bandwidth	450 MHz
Fixed Filter Power	230 mW
Tunable Filter Center Frequency	970 MHz
Tunable Filter Center Bandwidth	220 MHz
Tunable Filter Power	230 mW

Table 1 Measured filter performance summary.

Fig. 10 shows the single opamp performance measured up to 20 GHz. The measurement was carried out in single-ended mode. The unity gain bandwidth is 11.9 GHz and the circuit is stable with 52° of phase margin. It is important to note that a similar opamp design was implemented in 130-nm SiGe BiCMOS technology as described in [1]. Despite the lower f_T of the 180-nm n-MOSFETs and p-MOSFETs used in this design, with the same tail current density of 0.2mA/μm, the opamp bandwidth remained unchanged because it is determined by the HBT in the BiCMOS cascode. This proves that by using constant current density biasing at peak f_{MAX} current density, designs can be ported from one technology node to another. The measured opamp characteristics are summarized in Table 2.

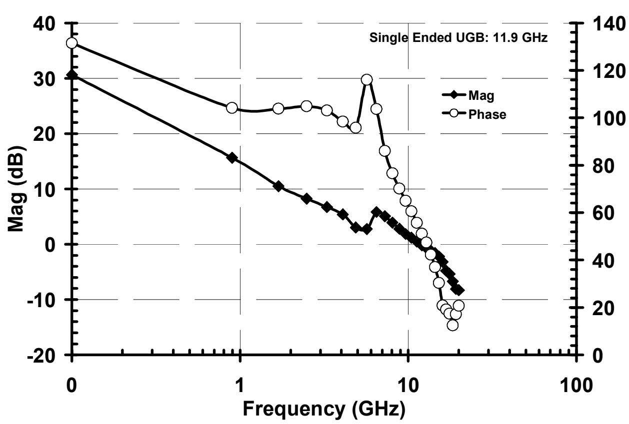

Fig. 10 Measured single-ended opamp gain S_{21} and phase with 10 mA tail current. Opamp schematic is shown in Fig. 3.

Specification	Measured result
Supply Voltage	4.0 V
Single Ended UGB	11.9 GHz
Phase Margin	52.53°
Single Ended DC Gain	30 dB
Diff. DC Gain	36 dB
Power Dissipation	76.5 mW

Table 2 Measured differential opamp performance.

IV. Conclusion

An opamp-based biquad bandpass filter was designed and implemented in 180-nm SiGe BiCMOS technology. Using custom BiCMOS differential amplifiers with 11.9 GHz unity-gain bandwidth, both fixed frequency and tunable filters with center frequency of 1 GHz were designed and fabricated. The measured data show that porting a low frequency filter topology to the gigahertz frequency range is feasible. The main advantage over inductor-based RF filters is at least a factor of 10 reduction in area and die cost.

Acknowledgement

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