A Simple Reconfigurable BiCMOS Active Inductor and Its Implementation in A Phase-Shifter Unit Cell

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***Abstract*—An active inductor based phase-shifter unit cell is proposed. The active inductor is designed with BiCMOS process technology and is implemented with only one heterojunction bipolar transistor and one field-effect transistor without any requirement of complicated transconductance amplifier designs. A phase-shifter unit cell is implemented with a high-pass T-Section with two varactors in series and the active inductor as a shunt. Relative phase variation is achieved by tuning the active inductor or by varying the effective junction capacitance of the varactors. Maximum relative phase variations of $23.7°$ and $38.6°$ are achieved at 4 GHz by exclusively tuning the gate voltage and varactor capacitance, respectively. The relative phase variations at 18 GHz are $6.0°$ and $8.5°$, respectively, for the same exclusive conditions.**

I. Introduction

Active Inductors are of interest today as an alternative to spiral inductors for various monolithic microwave integrated circuit applications. The reason for this is the excessive space requirement for passive inductors, which inflates the cost of the final product. Applications including oscillators, filters, phase shifters, etc. prefer the use of active inductors in-spite of various forms of loss, due to the form factor and an ongoing research focuses on the possibility of lowering them. Few well known CMOS based active inductors are discussed in [1].

Phase shifters are one of the necessary components for phased-array antennas and radars. Active inductor based phase shifters is a niche area and few developments are discussed in [2]–[6]. It is interesting to note that most of the designs use field-effect transistors (FET) and publications on the use of modern bipolar transistors are rare in the area of phase shifters. Few works which otherwise use active inductors using bipolar transistors are discussed in [7]–[9].

With the advancement of Silicon Germanium (SiGe) process technology it has the potential to deliver performance comparable to the Gallium Arsenide (GaAs) process, but SiGe process technology has a lower manufacturing cost similar to Silicon process technologies. The design discussed in this paper uses IHP SG13S BiCMOS process technology which is useful for designs with both bipolar and field-effect transistors in it. The tunable active inductor proposed in this paper uses a single heterojunction bipolar transistor (HBT) and a single FET for negative impedance transformation.

II. Active Inductor Circuit Design

The tunable active inductor proposed in this paper uses one HBT ($Q1$) and a FET ($M1$) in the design. A capacitor C_{ind} is connected from the base of the HBT to its emitter. It is shown later in this section that the capacitance C_{ind} helps to transform the input susceptance to a value similar to a lossy inductor. The FET acts as a feed-back control for both the DC bias and the AC input signal to the active inductor block. The circuit design is shown in Fig. 1. The bias voltage V_G is a DC value which can be changed to emulate a variable inductor. A DC current source, which includes a RF choke, denoted by I_C supplies the base and collector current to the HBT. The supply voltage is denoted by V_{DD}.

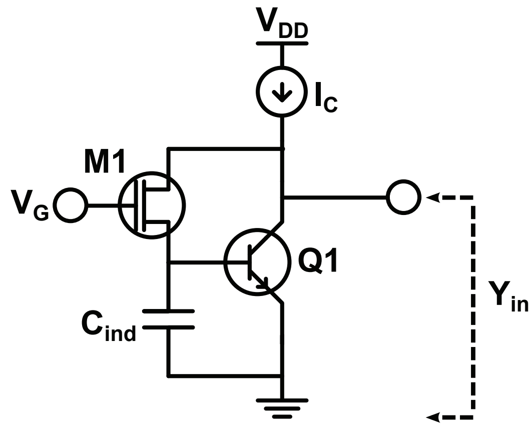

Fig. 1: Circuit design of the proposed tunable active inductor.

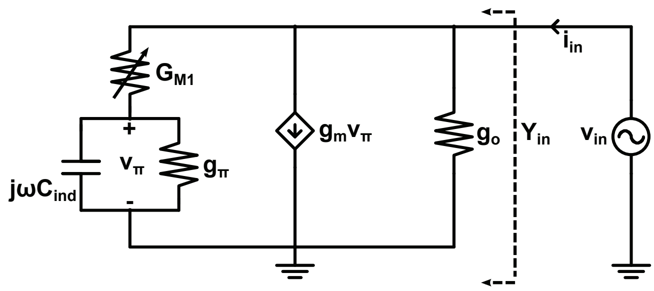

Fig. 2: Small signal model of the active inductor.

The small signal equivalent is shown in Fig. 2. For simplicity, the drain to source conductance is only mentioned in the model and is denoted by G_{M1}. The input admittance is calculated and the susceptance of it is given by:

$$Im\{Y_{inp}\} = -\frac{\omega C_{ind} G_{M1}\left(g_m - G_{M1}\right)}{\left(g_m + G_{M1}\right)^2 + \left(\omega C_{ind}\right)^2} \tag{1}$$

where g_m is the forward transconductance of the HBT (Q1) and ω is the angular velocity. The lossy part of the admittance is given by:

$$Re\left\{Y_{inp}\right\} = \frac{G_{M1}\left[\left(g_m+g_\pi\right)\left(g_\pi+G_{M1}\right)+\left(\omega C_{ind}\right)^2\right]}{\left(g_m+G_{M1}\right)^2+\left(\omega C_{ind}\right)^2} \tag{2}$$

where g_π is the input conductance of $Q1$. This small signal model is valid when the operating frequency $f \leq f_T/10$. In the SG13S process technology the transit frequency of the HBT is over 250 GHz, And hence the model will hold good for any operating frequencies below 25 GHz.

Active tuning is achieved by changing the gate voltage of $M1$ which varies both the DC base current I_B and the AC small signal current i_b. This is due to the biasing design which is based on voltage feedback loop. The transconductance is given by the ratio $g_m = \beta I_B/V_T$ where β is the current gain and V_T is the thermal voltage. With the change of the gate voltage of $M1$, the base current I_B changes and hence the g_m changes. In the small signal analysis, the transconductance is given by the ratio $g_m = \beta i_b/v_{be}$ and changing the gate voltage of $M1$ will also control the small signal current i_b and thus control g_m.

It is interesting to note here that the active inductance block can achieve a self resonance when the value of g_m (which itself is a function of G_{M1}) becomes equal to G_{M1}. At very low values of g_m ($g_m < G_{M1}$), the inductive property of the block is lost.

Fig. 3 shows a zoomed in portion of the layout of the active inductor.

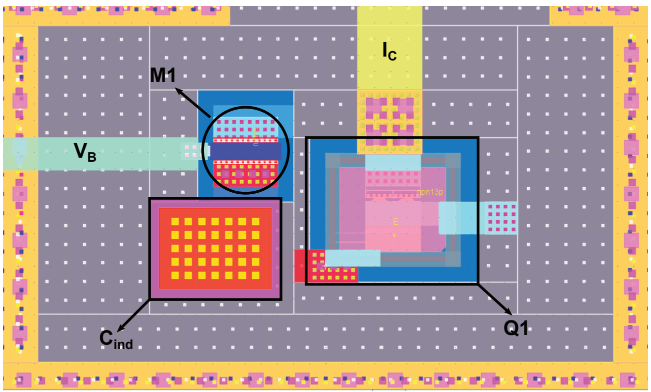

Fig. 3: Zoomed in view of the layout of the active inductor.

III. T Section Phase Shifter Unit Cell

A high pass T-section phase-shifter unit cell is considered in this paper. The topology of the unit cell is shown in Fig. 4. Two series capacitors (C_{se}) connect ports $P1$ and $P2$ with the active inductor L_{sh} shunting the node between two capacitors to the ground. For more control on the phase shift, varactor diodes are used instead of fixed capacitors.

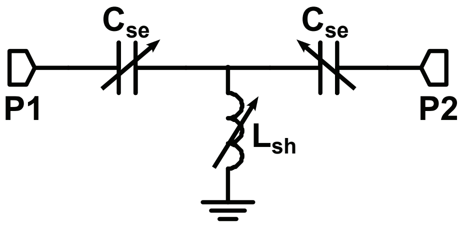

Fig. 4: The unit cell T-Section phase-shifter unit cell.

The transmission phase at port $P2$ of the unit cell shown in Fig. 4 can be approximated by:

$$\angle S21 \approx 90^\circ - tan^{-1}\left[\frac{2\omega L_{sh}Z_0 - \frac{2Z_0}{\omega C_{se}}}{\frac{2L_{sh}}{C_{se}} - \frac{1}{\omega^2 {C_{se}}^2} + {Z_0}^2}\right] \tag{3}$$

ignoring the lossy component of the active inductor. Z_0 is the characteristic impedance of ports $P1$ and $P2$. From Eq. (1), the value of imaginary component of the active can be approximated into L_{sh} and it is given by:

$$L_{sh} \approx \frac{C_{ind}}{G_{M1}\left(g_m - G_{M1}\right)} \tag{4}$$

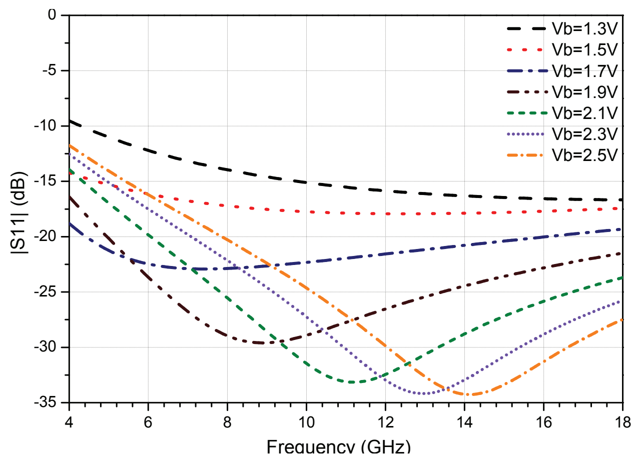

Fig. 5: Return loss of the phase-shifter unit cell with respect to gate voltage sweep of M1.

IV. Results

The active inductor circuit and the phase-shifter unit cell was simulated using AWR Microwave Office with a supply voltage of $V_{DD} = 1V$ and the gate voltage V_G of $M1$ was tuned between 1.3V and 2.5V with the capacitance fixed at 2pF. The average power consumption of the active inductor was 2.25mW. The return loss between 4 GHz and 18 GHz is shown in Fig. 5. The insertion loss for the same bandwidth is shown in Fig. 6. The change of transmission phase with respect to gate voltage is shown in Fig. 7. The relative phase variation due to voltage sweep at 4 GHz is 23.7° and at 18 GHz it is 6.0°. At a centre frequency of 11 GHz the relative phase variation is 11.0°.

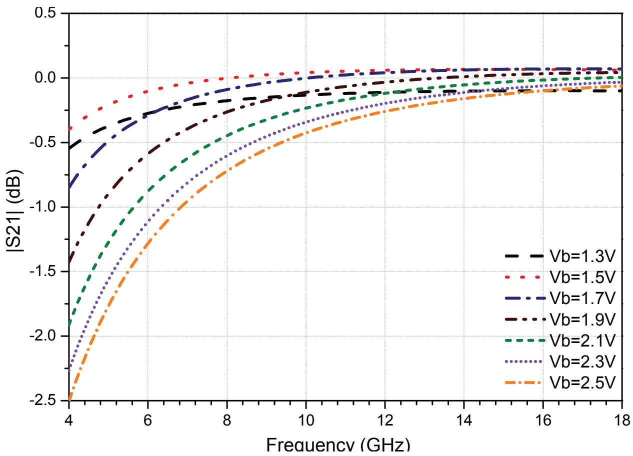

Fig. 6: Insertion loss of the phase-shifter unit cell with respect to gate voltage sweep of M1.

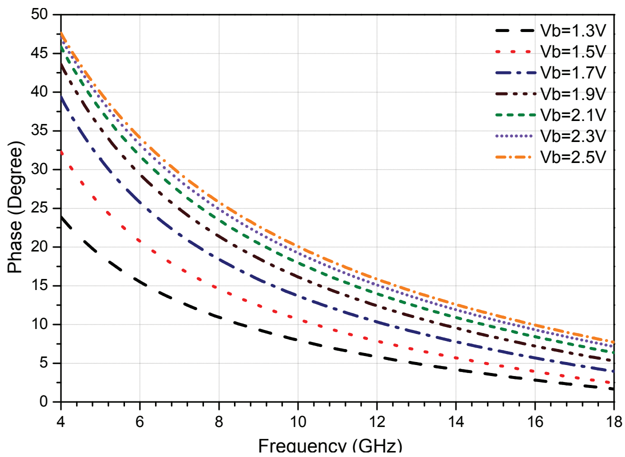

Fig. 7: Relative phase shift with respect to gate voltage sweep of M1, C_{se} fixed at 2pF.

The susceptance of C_{se} can be changed with additional bias-tees at ports $P1$ and $P2$ and further change of transmission phase is possible. One such simulation was done with the gate voltage V_G fixed at 2V and the capacitance of the varactor diodes changed between 0.8pF to 2.4pF. The resultant plot is shown in Fig. 8. At 4 GHz the relative phase variation is $38.6°$ and at 18 GHz it is $8.5°$. At a centre frequency of 11 GHz the relative phase variation is $13.4°$.

With the increase of gate voltage V_G, the return loss is affected at lower frequencies and the insertion loss of the unit cell reaches 2.5dB at 4 GHz. The insertion loss is less than 0.5dB at frequencies above 10 GHz for all values of gate voltages.

V. Conclusion

A phase-shifter unit cell using an active inductor is presented in this paper. The active inductor uses a heterojunction bipolar transistor for negative impedance conversion. It has the advantage of lower noise than field-effect transistors. A field-effect transistor is used here only to control the base current of the transistor. A second advantage of using bipolar transistor based active inductor is the stability of the pseudo-inductance

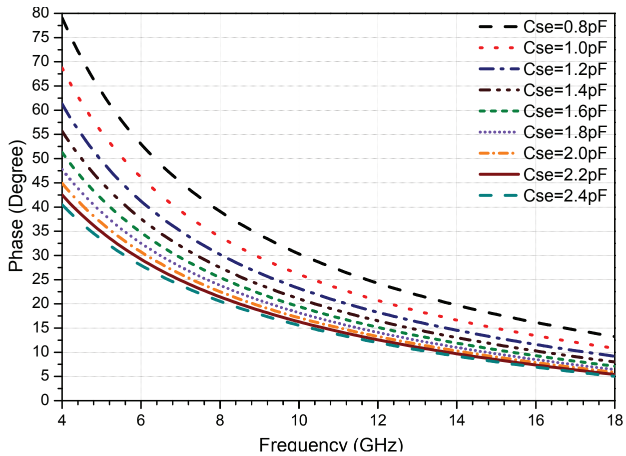

Fig. 8: Relative phase shift with respect to the sweep in varactor capacitance, V_B fixed at 2V.

which depends on the transconductance of the transistor used. The transconductance of a bipolar transistor only depends on the thermal voltage. The transconductance for a field-effect transistor depends on the gate to source (overdrive) voltage and it can shift even with a slight change in the gate voltage. The proposed unit cell can be laid-out in a small area and multiple cascades can help achieve a wider relative phase shift.

Acknowledgment

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