LATERAL MICROWAVE TRANSFORMERS AND INDUCTORS IMPLEMENTED IN A Si/SiGe HBT PROCESS

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Abstract - **Experimental results are presented on a set of microwave inductors and transformers fabricated in a lateral spiral design utilizing two metal layers rather than a single metal layer as used in conventional planar magnetic devices. The fabrication process utilizes a production Si/SiGe HBT technology with standard metallization and a thick polyimide dielectric. Inductors with peak Q's between 2.6 - 5 and inductance values between 1 – 3 nH are presented. Transformers with a loss of less than 5 dB when corrected for impedance mismatch and a measured coupling coefficient (k) of 0.6 at 5.5 GHz and 0.4 up to 12.5 GHz are also discussed.**

I. INTRODUCTION

The development of low-loss microwave transformers and inductors is important to the implementation of high-frequency monolithic integrated circuits in silicon technology. They can be used to implement a variety of important functions including impedance transformers, couplers, and baluns. Historically, these magnetic elements in silicon technology have had marginal performance due to the relatively high conductivity of typical silicon substrates. In addition, the ohmic loss of these structures is often high because of the relatively thin Al-based metallization employed in most VLSI processes.

In this paper, we present the performance of bilayer planar transformers and inductors produced with standard silicon VLSI Al-Cu metallization. In this process, the top metal layer is the standard back-end metallization, and the last metal layer is separated from the top metal layer by 12 µm of thick polyimide. We will first discuss the design of microwave transformers and inductors, followed by a description of the fabrication process, and finally a discussion of the experimental results.

II. SILICON-BASED TRANSFORMER AND INDUCTOR DESIGN

The design of microstrip structures and inductors are well-known in planar microwave circuit technology [1-2],[4-5],[7],[9-11]. Transformers are also important passive structures and have been reported by several authors including [3].

The loss due to the transformer or inductor is an important performance criterion. The sources of loss fall into three categories: ohmic loss, dielectric loss and spurious radiation. In our previous work [2], we showed that below 40 GHz, the loss in microstrip lines in this process is dominated by ohmic loss in the conductor due to the skin effect, with the dielectric loss being low. Little spurious radiation is seen due to the relatively thin spacing between the signal lines [4]. Expressions for the conductor loss and the dielectric loss can be found in [5].

The transformers and inductors presented here have been implemented using two metal layers in a lateral spiral design as shown in Figure 1(b-c). One possible disadvantage is that the lateral windings contain many vias, increasing the series resistance. Between the two metal layers, we have chosen a low-loss polyimide dielectric.

A circuit model of the transformers can be seen in Figure 2. In this model, M, is the mutual inductance. L_1 and L_2 are the leakage inductances. Neglecting capacitors C_1, C_2, and C_3, the mutual inductance, M, is proportional to the imaginary part of Z_{21} divided by ω, the angular frequency. The coupling coefficient is then defined as:

$$k = \frac{M}{\sqrt{L_{11}L_{22}}} \quad (1)$$

where,

$$L_{ii} = L_i + M \quad (2)$$

L_{11} and L_{22} are the imaginary parts of Z_{11} and Z_{22}, respectively, divided by ω.

An ideal transformer has no leakage inductance and thus from Equation 1 has a coupling coefficient, k, equal to 1. In the transformer model, the leakage inductances L_1 and L_2 cause k to be less than 1. Physically, the leakage inductances are due to lines of magnetic flux that link one winding but not both. Generally, the coupling coefficient can be increased by any of the following: increasing the number of windings, winding tighter coils, interwinding the coils, or providing magnetic material between the coils. Increasing the number of windings increases the number of flux linkages and thus M. Winding the coils tighter or interwinding the coils reduces the leakage inductance. Providing a magnetic material will increase the flux linkage between the coils by confining the magnetic field and reducing the leakage inductance [13].

III. FABRICATION TECHNOLOGY

The layers are shown in Figure 1(a). Schematic diagrams of the transformers and inductors appear in Figures 1(b-c). The last metal (LM) and top metal (TM) layers consist of 2.7 μm of sputtered Al-Cu. The top metal layer is separated from the substrate by approximately 3 μm of SiO_2. The dielectric is DuPont type 5811 [6] polyimide, spun-on, and then cured to a final thickness of approximately 12 μm. Via holes are created and the resulting sidewall angle allows for excellent step coverage. This structure is similar to previously reported MCM fabrication results on transmission line structures [7], but here the lines are fabricated directly on a silicon substrate. The conductivity of the metal is approximately 2.89×10^7 S/m. The process has been implemented on the IBM 200 mm silicon VLSI fabrication line as part of the Si/SiGe HBT BiCMOS technology [8].

IV. EXPERIMENTAL RESULTS

LATERAL TRANSFORMERS

Transformers with 10, 15 and 18 turns were fabricated, each with signal conductor widths of 45 μm and winding segment lengths of 700 μm. The performance metrics of interest include the coupling coefficient (k), loss, and resonant frequency. The evaluation was carried out from 0.5 GHz to 40 GHz using standard microwave on-wafer circuit testing techniques [7]. Cascade 150 μm GSG probes were used with an HP 8510 network analyzer. SOLT calibration was used.

Feed line and pad parasitics have been removed with the use of an on-wafer microstrip calibration structure. Since the calibration structure has a short electrical length, it is approximated at each frequency by a symmetric lumped pi-network [12]. A MATLAB script calculates the pi-network values from the measured network parameters. Utilizing this method, the calibration structure line length and associated pad parasitics have been removed from the transformer and inductor data presented here. In our previous work in this fabrication process [2], we demonstrated that below 40 GHz the conductor loss dominates the total loss. Also, the polyimide dielectric has an effective relative permittivity of 2.8 and a loss tangent of approximately 0.01.

Figure 4 is a plot of the measured transformer S_{21} in dB as a function of frequency. The unmatched loss in the 15-turn transformer is about 7 dB from 5 GHz to 20 GHz and exhibits the flatest wideband response of the set. The loss here is greater than that presented in [3] but has a wider bandwidth. Figure 5 shows the maximum available gain. It is calculated as:

$$G_{ma} = \left|\frac{S_{21}}{S_{12}}\right|\left(k_s - \sqrt{k_s^{\,2} - 1}\right) \qquad (3)$$

where, k_s, the stability factor is:

$$k_s = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |S_{11}S_{22} - S_{12}S_{21}|^2}{2|S_{12}||S_{21}|} \qquad (4)$$

The calculated parameter, G_{ma}, reflects the gain of the system when the source and load reflection coefficients are conjugately matched to S_{11} and S_{22}, respectively. The best gain achieved is -2.5 dB at 3 GHz with the 15-turn transformer. Also, the maximum available gain of the 10- and 15-turn transformers is about -4 dB up to 22 GHz. Figure 6 is a plot of the measured coupling coefficient, k. The 18-turn transformer has the highest coupling coefficient (0.6) but only has a useable bandwidth between 1-6 GHz. The 15-turn transformer has a lower coupling coefficient (0.55) but has a larger bandwidth, between 1-9 GHz. The 10-turn transformer has the lowest coupling coefficient (0.45) but the largest bandwidth – up to 13 GHz. Figure 7 is a plot of the mutual inductance, M and Figure 8 is a plot of the inductance L_{11}.

LATERAL INDUCTORS

Polygonal spiral inductors on silicon substrates implemented on a single metal layer with only the pass-thru on a second metal layer have been developed by several authors including [2] and [8]. In our work, the inductors were constructed in a similar manner to the transformers. Three inductors with segment lengths 250 μm, 400 μm, and 700 μm were fabricated each with 10 turns. The inductor layout is shown in Figure 1b. Data was taken between 0.5 GHz - 40 GHz.

High Q values are desirable but are limited by the metallization and dielectric losses. After dembedding the pad parasitics in a manner similar to the transformers, the inductance and Q can be calculated from the two-port admittance parameters as:

$$Q = \frac{\mathrm{Im}\left\{\frac{1}{Y_{11}}\right\}}{\mathrm{Re}\left\{\frac{1}{Y_{11}}\right\}} = \frac{-\mathrm{Im}\{Y_{11}\}}{\mathrm{Re}\{Y_{11}\}} \quad (5)$$

$$L = \frac{1}{\omega}\mathrm{Im}\left\{\frac{1}{Y_{11}}\right\} \quad (6)$$

A simplified equivalent circuit is shown in Figure 3. Using these definitions, the calculation of inductance and Q includes the capacitances C_1 and C_2 since Y_{11} includes both of these terms. Figure 9 shows the extracted inductance, and Figure 10, the Q. The short 250 µm segment length inductor has a peak Q of 5 at 4 GHz and an inductance of 1 nH. The inductors with longer segments have a higher inductance but also higher capacitances C_1 and C_2, which lower the self-resonant frequencies. The Q's for the inductors with longer segment lengths are lower since the series resistance increases more for a longer segment than the inductance due to C_1 and C_2. The Q's of these inductors are modest compared to single layer inductors in the same process [2]. In our previous work we demonstrated square spiral inductors with peak Q's as high as 22 at 10 GHz.

Since both the transformers and inductors have numerous vias (twice the number of turns), the DC resistance of the via holes is important. Using DC needle probes, this value was measured to be 0.05Ω per via.

V. CONCLUSIONS

Transformers and inductors have been demonstrated in a production silicon VLSI HBT technology. Inductors with a peak Q of 5 at 4 GHz and inductance of 1 nH have been demonstrated. Although the single layer planar inductors in [2] may be better in many applications, the lateral inductors presented have acceptable performance. Wide bandwidth 10-turn and 15-turn transformers built in a lateral orientation have been shown with an unmatched loss of 7 dB. These devices offer good performance between 5 - 20 GHz for silicon MMIC applications.

VI. ACKNOWLEDGMENTS

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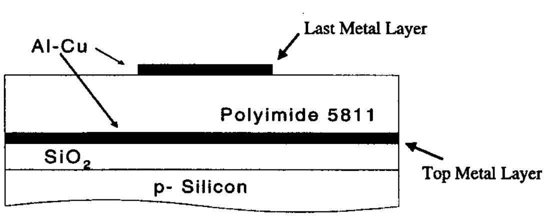
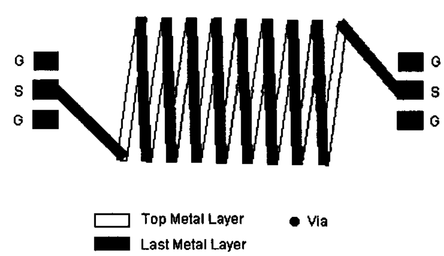


(a)



(b)

Figure 1. a) Substrate layers. **b)** Inductor layout.

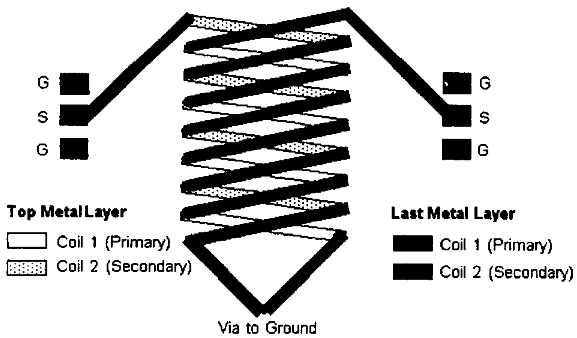


Figure 1c. Transformer layout.

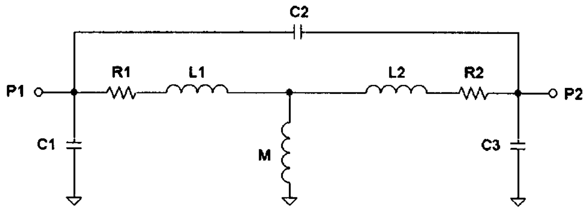


Figure 2. Transformer equivalent circuit model.

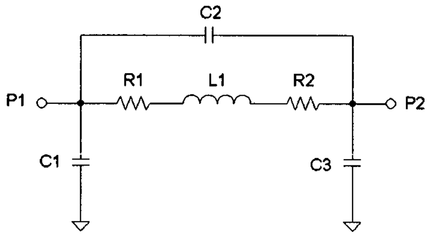


Figure 3. Inductor equivalent circuit model.

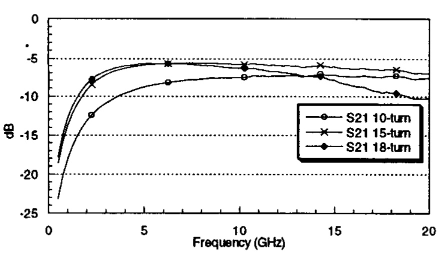


Figure 4. Measured transformer S_{21}.

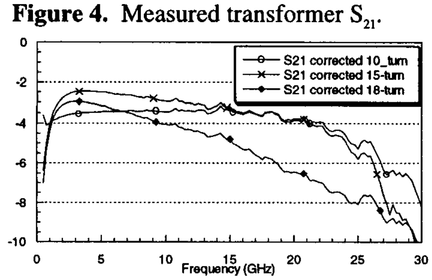


Figure 5. Transformer maximum available gain.

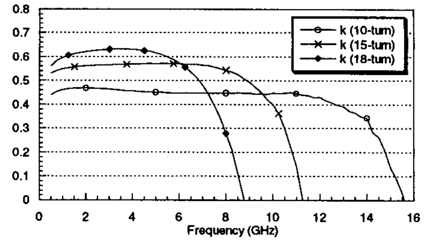


Figure 6. Transformer coupling coefficient (k).

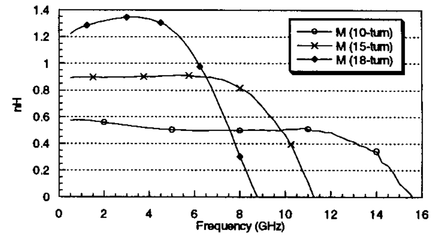


Figure 7. Transformer mutual inductance (M).

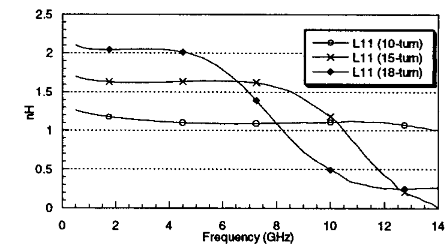


Figure 8. Transformer inductance (L_{11}).

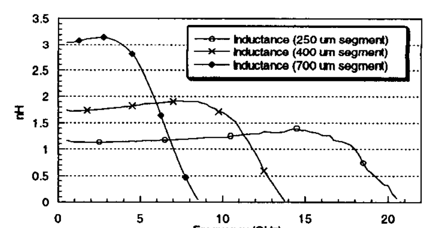


Figure 9. Inductance of lateral inductors.

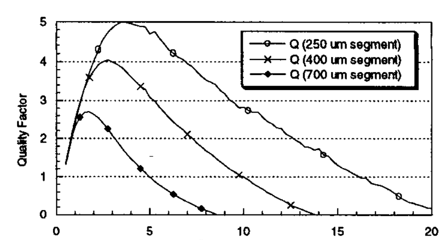


Figure 10. Inductor quality factor.