A NEW TRANSMISSION LINE APPROACH FOR DESIGNING SPIRAL MICROSTRIP INDUCTORS FOR MICROWAVE INTEGRATED CIRCUITS*

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Abstract

Spiral microstrip inductors with up to two full turns have been modeled by using parallel coupled and single transmission lines. With this method, the electrical characteristics of these inductors, which may be fabricated as elements of MIC and MMIC circuits, can be designed for reactances up to 200 Ω with a maximum error of 15 percent for frequencies up to the K_u band.

Introduction

The design of microstrip inductors for microwave integrated circuits has been traditionally based on two fundamental methods: the lumped-element approach and the distributed-line approach. In the lumped-element approach, formulas for free space inductance are used, with corrections for ground plane effects [1]. These frequency independent formulas are useful only when the inductor total length is a small fraction of a wavelength and when the capacitance between turns can be neglected. Equivalent lumped-element models for spiral inductors have been deduced from 2-port S parameter measurements [2]. These models do not lend themselves readily to design applications because their elements are difficult to evaluate in advance. In the conventional distributed design, an inductor is realized and analyzed as a single microstrip line. This approach may require a large area on the circuit. For example, to realize a 5-nH inductor at 4 GHz on a 0.318-mm GaAs substrate, a 3.56-mm-long, 100-Ω line is needed. To save space, a meander line may be used, but sufficient spacing between the turns is required to avoid negative inductive coupling.

In this paper, a new distributed-line technique is presented for modeling spiral microstrip inductors. It enables the fabrication of compact elements consisting of as many as two full turns and the prediction of their distributed-line characteristics up to the K_u band, with a maximum error of 15 percent. The theoretical analysis is based on the dispersive parameters of single and parallel coupled microstrip lines [3]. Measured values of reactance for spiral inductors on GaAs substrate are compared with the theoretical predictions. The role of bends in very narrow microstrip lines has not been thoroughly explored, but a better understanding of their effect would enhance the accuracy of the proposed transmission line model. The quality factors, Q, of the inductors are also calculated from the transmission line model and are compared with the values obtained by resonant techniques.

Analysis

A 2-turn microstrip inductor may be viewed as a section of two parallel coupled microstrip lines that are interconnected at opposite ends, as shown in Figure 1. The single line sections between nodes 1 and 2 and between nodes 4 and 5 represent the feedlines of the inductor. The parallel coupled sections bounded by nodes 2, 3, and 4 constitute the intrinsic inductor. In Figure 2, a 1-3/4-turn element is defined by the lines bounded by nodes 2, 3, 4, and 5. Here, the parallel coupled sections are interconnected by a single line, between nodes 3 and 4. When using this modeling technique, it is stipulated that the inner and outer turns of the physical element are parallel and approximately equal in length. The smaller the width of the lines and the separation between them, the better the expected correspondence between the physical inductor and the equivalent circuit.

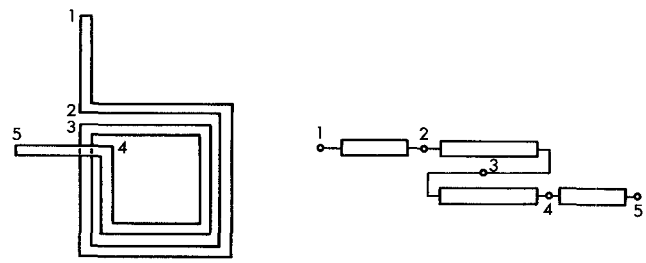

Figure 1. 2-Turn Inductor and Its Distributed-Line Model

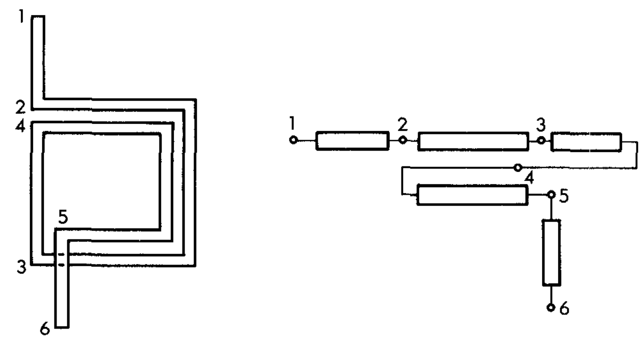

Figure 2. 1-3/4-Turn Inductor and Its Distributed-Line Model

The electrical characteristics of the intrinsic 2-turn inductor can be derived from the general 4-port network of coupled transmission lines. In Figure 3, the current and voltage relationships of the pair of lines can be described by the admittance matrix equation

$$\begin{bmatrix} I_1 \\ I_2 \\ I_3 \\ I_4 \end{bmatrix} = \begin{bmatrix} Y_{11} & Y_{12} & Y_{13} & Y_{14} \\ Y_{21} & Y_{22} & Y_{23} & Y_{24} \\ Y_{31} & Y_{32} & Y_{33} & Y_{34} \\ Y_{41} & Y_{42} & Y_{43} & Y_{44} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{bmatrix} \quad (1)$$

The description of the 2-turn inductor is found by imposing the boundary conditions

$$I_2 = -I_3$$

$$V_2 = V_3$$

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which, in Figure 3, are expressed by the insertion of the dashed line between nodes 2 and 3. As a result, the following reduced matrix equation is obtained:

$$\begin{bmatrix} I_1 \\ I_4 \end{bmatrix} = \begin{bmatrix} Y'_{11} & Y'_{14} \\ Y'_{41} & Y'_{44} \end{bmatrix} \begin{bmatrix} V_1 \\ V_4 \end{bmatrix} \tag{2}$$

where

$$Y'_{11} = Y_{11} - \frac{(Y_{12} + Y_{13})(Y_{31} + Y_{21})}{(Y_{22} + Y_{23} + Y_{32} + Y_{33})} \quad \text{etc.}$$

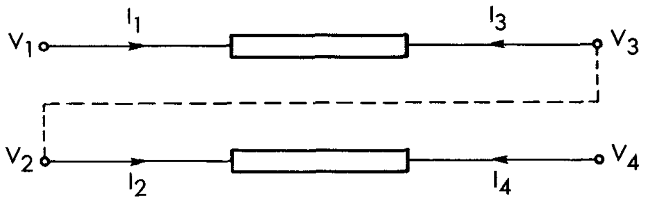

Figure 3. Pair of Parallel Coupled Transmission Lines

With port 4 grounded, Y'_{11} is equal to the input admittance of the 2-port circuit. The admittance Y'_{11} can be found from the preceding equation and from the definition of the Y parameters as functions of the even and odd mode characteristic admittances, Y_{oe} and Y_{oo}, and the even and odd mode electrical lengths, θ_e and θ_o, of the coupled lines [3].

The foregoing analysis with a different set of boundary conditions for equation (1) can be used to analyze the case of a 1-3/4-turn inductor.

Experimental Results

Rectangular spiral inductors were fabricated on GaAs substrates, 0.318 mm thick. The lines were made with 100 Å Cr/ 1,000 Å Au, sputtered and 3 µm Au plated. The design goal was to minimize the size of the elements and the difference between the lengths of the inner and the outer turns of the coupled sections. Figure 4 shows the layout of a 1-3/4-turn inductor designed for 1.5 nH in the linear range of its reactance. The line width is 11.4 µm (w/h = 0.04), and the spacing between the turns is 8.9 µm (s/h = 0.03). The element without feedlines measures 0.2 mm x 0.25 mm. The 90° bends of the inner turn are mitered to 50 percent, and the 45° bends of the outer turn are unmitered. From theoretical and experimental studies [4],[5], the bends in the circuit were expected to cause a reduction in the electrical length and in the inductance. By extrapolating the theoretical curves in Figure 6 of Reference 5, the effective loss of length in a 90° bend was estimated to be 0.05 mm. For coupled lines, a computer-aided analysis resulted in a loss of 0.05 mm per bend. The loss of two successive unmitered 45° bends was approximated by that of one mitered 90° bend.

An initial equivalent circuit was set for the inductor of Figure 4 according to the physical dimensions of the inductor (see Figure 5a) and the estimated effect of the bends. The circuit was optimized to conform with the measured reactance, resulting in the equivalent circuit of Figure 5b. The single line sections and the coupled sections were effectively shorter than those of the physical model by 0.05 mm and 0.25 mm, respectively. The equivalent circuit for this inductor set the basis for the modeling of other inductors with the same layout but with different side lengths. Therefore, their equivalent circuits were subjected to the same absolute reductions in line lengths as those described in Figure 5. Figure 6 shows the calculated and the measured reactance values of several inductors. For reactances about 200 Ω or less, the maximum difference between the calculated and the measured values is 15 percent. At higher reactances, the electrical length of the inductors approaches one-fourth of a wavelength. Consequently, small discrepancies between the model and the real inductor may lead to large errors in the calculation of the reactance. For example, at 10 GHz, an error of 0.03 nH in the evaluation of the inductance of the bond wire bridge is equivalent to a phase error of 2.5° in a 50-Ω system. This error is amplified to an uncertainty of 30 percent in the calculation of an inductor with L ≈ 4 nH at 10 GHz.

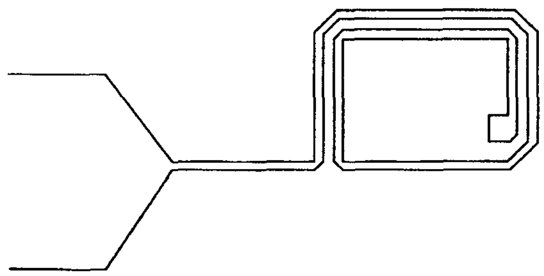

Figure 4. 1.5-nH Inductor on 0.318-mm GaAs Substrate (w = 11.4 µm, s = 8.9 µm)

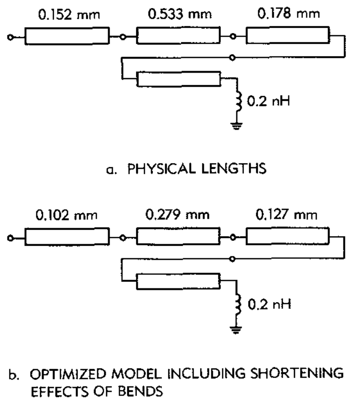

Figure 5. Transmission Line Model of the 1.5-nH Inductor Shown in Figure 4 (w = 11.4 µm, s = 8.9 µm)

In an attempt to reduce the effect of sharp bends, new inductors were made with round corners. Figure 7 shows a 2.9-nH inductor with w/h = 0.06 and s/h = 0.05. A transmission line model fitted to the measured reactance of two round-bend inductors indicated that the total loss of electrical length, 0.25 mm, was lower by 0.10 mm from the loss in the sharp-bend inductors.

The Q of the inductors were measured at several frequencies by embedding the elements in linear 50-Ω line resonators, such as the one in Figure 8. The Q values were calculated from the measured values of Q_{total} and a priori data on the losses of the 50-Ω lines. The experimental values are shown in Figure 9, together with Q values that were calculated from the transmission line model.

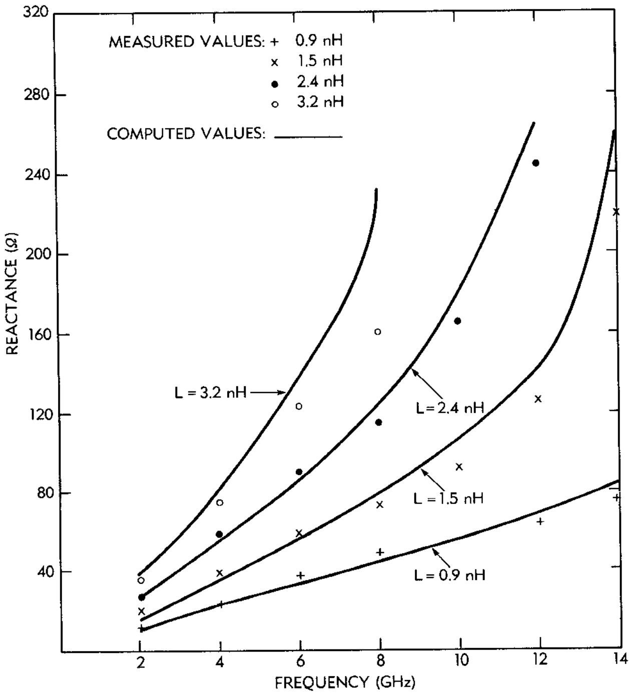

Figure 6. Calculated and Measured Reactance of Spiral Microstrip Inductors on GaAs ($h = 0.318$ mm, $w = 11.4$ μm, $s = 8.9$ μm)

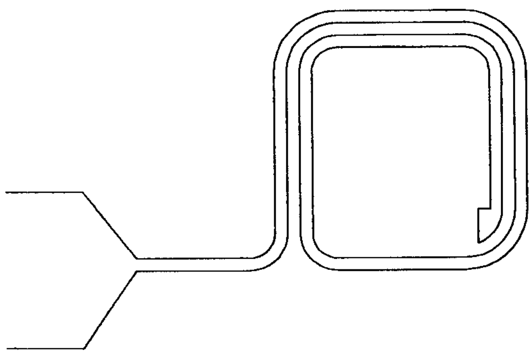

Figure 7. 2.9-nH Inductor on GaAs ($h = 0.318$ mm, $w = 19.1$ μm, $s = 8.9$ μm)

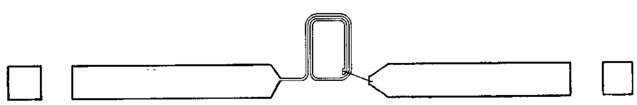

Figure 8. Inductor Imbedded in 50-Ω Linear Resonator

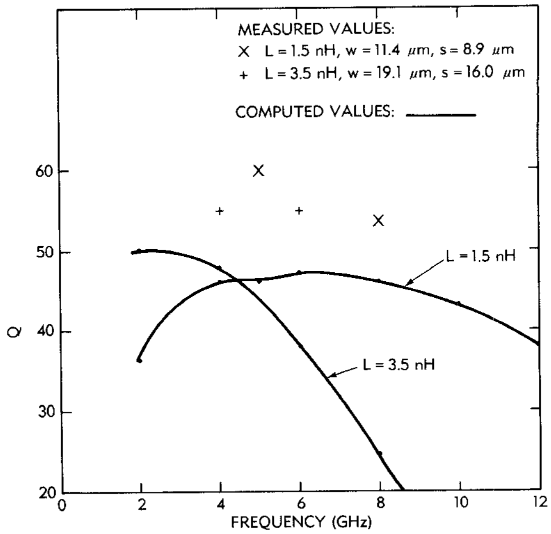

Figure 9. Measured and Calculated Q of Spiral Inductors on GaAs ($h = 0.318$ mm, $\tan \delta = 0.0004$)

Conclusion

A new distributed-line approach to the modeling of microstrip spiral inductors has enabled their design and the prediction of their reactance within 15 percent up to the K_u band for values up to 200 Ω. The ability to fabricate spiral inductors and to predict their distributed-line performance bears a significant space savings, especially at frequencies at which a straight or a meander line realization becomes too long. Furthermore, the distributed-line model facilitates the use of these elements in wideband applications where their inductance does not remain constant. With a better understanding of bends in narrow single and coupled lines, the models for the spiral inductors are expected to improve.

References

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