Monolithic Spiral Transformers: A Design Methodology

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Abstract—A method to design spiral transformers is outlined. It utilizes simple inductance estimates in order to initially design the dimensions for the desired transformer, after which an EM simulation is used to extract s-parameters of the structure; a compact model is then extracted from the s-parameter results. A wideband compact model is proposed that is accurate up to the self-resonance frequency (SRF) of the structure. A 3-to-1 transformer is designed and its measured performance verified in a 0.18 $\mu$m RF CMOS process.

I. INTRODUCTION

Spiral transformers are increasingly used in integrated RF applications; however, there are no commercial tools currently available that can be used to easily design unique spiral transformer structures. With quick calculations, the physical dimensions of a transformer can be obtained, thereby saving time in the design cycle due to reduction in iterations of an EM simulation. A method which uses Grover calculations to estimate these physical parameters is outlined in Section II. Next, in Section III, a wideband compact model of the inductor is described, which achieves accurate performance up to the SRF of the transformer. In Section IV, a 3-to-1 transformer is measured and compared to the extracted wideband compact model. Finally, conclusions are provided in Section V.

II. DESIGN METHODOLOGY

One of the most difficult challenges associated with designing a monolithic spiral transformer is that there are few rapid ways in which to design and characterize it without using iterations in a complex electro-magnetic (EM) simulator. In the next section, a design strategy is laid out that leads to first-pass success in spiral transformer designs. The methodology reduces the need for time- and computation-intensive EM simulations by using traditional Grover partial inductance calculations [1, 2]. The partial inductance calculations use a concentric winding approximation in order to simplify the calculations [3]. Next, a full EM characterization of the structure is performed, to verify that the design provides acceptable performance with substrate parasitic effects included.

The Grover calculations are convenient because they easily relate the physical parameters of a structure to inductance values. In order to perform the calculations, the structure’s physical dimensions are used to break down a section of wire into a single filament. The partial self-inductance of each filament can then be calculated along with the partial mutual inductance of each filament to all other filaments. In order to make the application of the calculations easier to perform, the structure will first be broken up into concentric windings. The structure shown in Figure 1(a) is a series-parallel interleaved transformer, which is useful for achieving turns ratios ≠ 1, while maintaining a relatively small winding inductance. Note that the primary is wound such that its constituent rings are effectively in series with one another, while the secondary is wound such that its constituent rings are effectively in parallel [4]. Figure 1(b) demonstrates how the structure can be broken down into several concentric windings for easy calculation. After the structure is broken down, each winding is further reduced into sub-segments, as shown Figure 2.

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The following calculations are a reduced set of those available in [1] for the special case of an octagonal spiral. The partial self-inductance of a straight conductor is given by:

\[
L = 10^{12} \frac{\mu_0 l}{2\pi} \left[ \ln \left( \frac{2l}{GMD} \right) - 1.25 + \frac{AMD}{l} + \left( \frac{\mu}{4} \right) \gamma \right]
\]  

(1)

where \( L \) is the inductance in pH, \( \mu_0 \) and \( \mu \) represent the magnetic permeability of free-space and the conductor respectively, \( l \) is the length of the segment in \( \mu m \), \( AMD \) and \( GMD \) are the geometric mean distance and arithmetic mean distance of the conductor cross-section in \( \mu m \), respectively, and \( \gamma \) is a frequency correction factor for skin depth. For a rectangular cross-section, \( AMD \) and \( GMD \) of the inductors are:

\[
AMD = \frac{w + t}{3}
\]  

(2)

\[
GMD = 0.2235(w + t)
\]  

(3)

where \( w \) and \( t \) are the width and thickness dimensions of the conductor in \( \mu m \). After the partial self-inductances are calculated, the mutual inductances need to be calculated.

The partial mutual-inductance of two parallel elements, Figure 3, is given by:

\[
M_{par,even} = 200000 \ln \left( \frac{l}{GMD} \right) + \sqrt{1 + \left( \frac{l}{GMD} \right)}
\]  

(4)

where \( M \) is in pH, and the other dimensions are the same as in (1). The next type of mutual inductance which is necessary for calculation is unequal parallel elements as seen in Figure 4.

The partial mutual-inductance of two such segments (assuming \( m = p \)) is obtained by solving (4) for two different lengths:

\[
M_{par,uneven} = M_{par,even} \left( l_n + l_m \right) - M_{par,even} \left( l_p \right)
\]  

(5)

Lastly, the mutual inductance of two elements inclined at an angle, Figure 5, needs to be calculated.
\[ M_{\text{ind}} = 20000 \cos \alpha \left[ n \tanh \left( \frac{m}{n+y} \right) \right] \ldots \] (6)

\[ + m \tanh \left( \frac{n}{m+y} \right) \] (6)

The partial mutual-inductance of the conductors in Figure 5(b) is given by:

\[ M_{\text{mut}} = 2\cos \alpha \left[ M_{\text{self}} (x+m, y+n) + M_{\text{mut}} (x, y) \right] \ldots \]

\[ - \left[ M_{\text{self}} (x+m, y) + M_{\text{mut}} (y+n, x) \right] \] (7)

With the Grover calculations complete, the elements can be summed together to represent the total inductance of the primary and secondary, \( L_P \) and \( L_S \), as well as the mutual inductance, \( M \):

\[ L_P = \sum_{i=1}^{8} L_i + \sum_{i=9}^{16} L_i + 2 \sum_{i=1}^{8} \sum_{j=2}^{8} M_{i,j} + 2 \sum_{i=9}^{16} \sum_{j=10}^{16} M_{i,j} \] (8)

\[ L_S \approx \sum_{i=17}^{24} L_i + \sum_{i=17}^{24} \sum_{j=18}^{24} M_{i,j} \] (9)

\[ M = \sum_{i=1}^{16} \sum_{j=17}^{32} M_{i,j} \] (10)

This shows that the primary inductance, \( L_P \), which is wound in series is the sum of all of the elements of the partial self inductance for each loop of the primary, summed with all of the elements of partial mutual-inductance from one element of the primary to another element of the primary. The secondary inductance, \( L_S \), is connected in parallel and, thus, the inductance is dominated by the partial self- and mutual-inductance of the inner turn of the secondary winding.

With the ability to specify an inductance from physical parameters, a layout can now be produced and verified using an EM simulation.

**III. WIDEBAND COMPACT MODEL**

A wideband compact model is often desirable for simulations, especially when the simulation will be performed in the time domain. Typically EM simulators output a frequency domain s-parameter file, which is acceptable for small-signal frequency simulations; however, it is not optimal for large-signal and time-domain simulations. This result is due to the non-linearities in the simulation, which tend to expand the necessary bandwidth of the s-parameter dataset, in order to achieve accurate results. In cases where the time-domain simulator does not have an extrapolation feature, this would lead to EM simulations with higher maximum frequencies, and hence a significant increase in the time required for running the EM simulation.

Many compact models have been developed for inductors [5-7], and the features of these inductors are easy to port over to transformer models, as shown in [4]. The proposed compact model utilizes the characteristics of the prior transformer models and adds the ability to model skin effect [8] and bulk eddy currents [7].

The proposed model is shown in Figure 6. The inductances in the model are estimated using the Grover calculations, whereas the parasitics are initially calculated as in previous models [6], with the exception of the parameters \( x \) and the coupling factor to the bulk inductor (not shown), which are empirically fit using an optimization engine. Once all of the components are initially calculated, an optimization of all of the model parameters, with the exception of the inductance, is performed in order to insure a good fit between the model and the s-parameter data from the EM simulation.

![Proposed transformer wideband compact model.](image1)

**IV. MEASUREMENT RESULTS**

A 3-to-1 transformer is designed in a 7 layer Al-Cu Metal, 0.18 \( \mu \)m CMOS RF process is shown in Figure 7.

![3-to-1 series/parallel interleaved transformer chip microphotograph.](image2)
The transformer is measured using a Cascade probe station and an Agilent performance network analyzer. The transformer’s s-parameters are characterized to 20 GHz, and the winding inductance and mutual inductance are extracted:

\[ L_p = \text{Im} \left( \frac{Y_{22}}{\omega \det Y} \right) \]  \hspace{1cm} (11)

\[ L_s = \text{Im} \left( \frac{Y_{11}}{\omega \det Y} \right) \]  \hspace{1cm} (12)

\[ M = \text{Im} \left( \frac{Y_{12}}{\omega \det Y} \right) \]  \hspace{1cm} (13)

A comparison between the extracted compact model and the fabricated transformer is shown in Figure 8.

V. CONCLUSIONS

A methodology for designing unique spiral transformers has been outlined, which uses simple first principle inductance calculations to reduce the design cycle time by reducing the number of iterations that must be performed in an EM simulator. A compact model that demonstrates excellent bandwidth is extracted and compared to measurements from a fabricated inductor; it shows a maximum deviation of 10% from DC to 20 GHz.

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REFERENCES


