Closed-Loop Model: An Optimization of Integrated Thin-Film Magnetic Devices

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Abstract

A generic analytical model has been developed to fully describe the flux closure through magnetic inductors. The model was applied to multiple device topologies including solenoidal single return path and dual return path inductors as well as spiral magnetic inductors for a variety of permeabilities and dimensions. The calculated inductance values from the analytical model were compared with simulated results for each of the analyzed device topologies and found to agree within 0.1 nH for the range of typical thin-film magnetic permeabilities ($\sim 10^2 - 10^3$). Furthermore, the model can be used to evaluate behavior in other integrated or discrete magnetic devices with either non-isotropic or isotropic permeability and used to produce more efficient device designs in the future.

Keywords: flux closure, integrated magnetic core, inductors, permeability, thin-film magnets, flux modeling

1. Introduction

Thin film magnetic materials are being increasingly integrated into electronics to boost transduction in inductors [1, 2, 3, 4, 5], transformers [6, 7], micro-electromechanical systems [8, 9, 10], and sensors [11, 12]. Apart from cost effectiveness, the main obstacle standing in the way of proliferation of magnetic integration is the difficulty of understanding and predictably modeling thin-film magnetic behavior. As magnetic devices decrease in size and go from discrete components to integrated forms, the behavior changes dramatically and typical textbook expressions fail to capture it. Various analytical models have been put forth to help predict the behavior of specific device

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topologies [4, 13, 14]. However a more comprehensive model is required to describe a broader range of topologies and explain the change in magnetic flux behavior as devices transition from discrete to integrated forms.

When magnetic materials are converted from bulk to thin film forms for integrated devices, the reluctance of the magnetic core changes, resulting in a different spatial distribution of the flux closure. The main reasons for this change are that compared to bulk magnets, thin-film magnetic materials 1) have lower permeabilities, 2) suffer from larger demagnetization effects, and 3) are often patterned into rectilinear structures with sharp edges that have increased local demagnetization fields. Taking into account all these factors, this paper presents a rather concise and accurate model for magnetic closed-loop devices by considering all the possible paths in which the flux can close. The model is compared against simulations of both integrated and discrete inductors in order to demonstrate its versatility. Additionally, since the model takes into account the different flux paths contributing to the total inductance, a comparison is made in order to identify, as a function of both permeability and design dimensions, which flux path has the greatest contribution to the inductance. Equipped with this information, integrated thin-film magnetic devices can be optimized to maximize both flux flow through the magnetic film and enhancement due to it.

2. Analytical Model

The analytical model is based on an analysis of the total reluctance of the design, considering all the possible ways in which the flux can close. The analysis is similar in methodology to several previous works [13, 14, 15]. While some assumptions had to be made about the device structure and demagnetization factors to simplify analysis, the model was kept as general as possible in order to apply broadly to a wide range of magnetic devices. Assumptions related to structure and demagnetization are explicitly stated as they arise in the derivation below. Closure in the model is assumed to be directly related to isotropic permeability in the magnetic core, which is difficult to achieve experimentally, but has been demonstrated to a certain degree in [16, 17]. Later, the model is generalized for cases of non-isotropic permeability as well. Two solenoid inductor structures with isotropic permeability are described initially, while more topologies and non-isotropic permeability cases will be considered in the discussion section of this paper. The two isotropic inductor structures, illustrated in Fig. 1, are the single return path (SRP) and dual

Figure 1: The two closed-loop magnetic solenoid inductor topologies used for optimizing the analytical model, including either a) only one section or b) two sections through which the flux can return in the magnetic core.

return path (DRP) topologies, which are defined by whether they have one or two sections of magnetic material in which the flux can return. The DRP topology has the advantage that flux generated anywhere in the solenoidal region can close through the nearest magnetic return path, and thereby have a shorter overall flux path and lower reluctance than the SRP topology.

Ampere's law explains that the total magnetomotive force, V_{mmf} , is given by the integral of the magnetic field around a closed loop. This same magnetomotive force can then be understood to be a sum of the drops across each of the different sections of the path.

$$\oint H \cdot dl = V_{mmf} \quad (1)$$

$$\sum_{i=1}^n V_{mmf_i} = V_{mmf} \quad (2)$$

The reluctance of each section of the flux path can therefore be evaluated separately and related back to the equivalent magnetic circuit model to compute the total reluctance and, finally, the total inductance.

For instance, for the SRP design, the circuit model represented in Fig. 2 begins with a flux-forming region where the coil perfectly surrounds the magnetic core. (Although the magnetic potential is not instantaneously created, the model locates it specifically at the beginning of the coil and evaluates the potential drop across this region to be V_{mmf_1} .) The flux then branches off. In one case, the flux can flow entirely through the magnetic curvature and, in another case, the flux can take a shortcut directly across the gap to reach the magnetic core on the other side. After this section, the flux recombines and returns through the magnetic core straight leg section, then once again

splits up to follow either the magnetic curvature or cut across the gap one final time. In parallel with all of this is a reluctance path that describes the flux returning entirely through air.

For each of these sections, an expression for the magnetic field, and subsequently, the flux density and flux are evaluated as a function of permeability and dimensions. The reluctance is the ratio of the magnetomotive force to the flux flowing through the section, $\mathcal{R} = V_{mmf}/\Phi$. The total reluctance of the structure is then evaluated according to the series and parallel circuit models of Fig. 2. Finally, the inductance is equal to the ratio of the square of the number of turns to the total reluctance:

$$L = \frac{N^2}{\mathcal{R}} \quad (3)$$

The reluctance of the flux-forming region is given as

$$\mathcal{R}_{nominal} = \frac{l_m}{\mu_0 \mu_r w_m t_m} \quad (4)$$

where l_m , w_m , and t_m are the length, width, and thickness of the magnetic core, respectively, and μ_0 and μ_r are the free-space and relative permeabilities.

In order to define the flux through the curved core region, the model must account for the variation in the flux density as a function of position. Assuming uniform isotropic permeability, the flux will attempt to travel along the shortest possible path around the curvature, hugging the inner edge of the core, in order to see the smallest reluctance possible, as seen in Fig. 3. Therefore, best designs minimize the radius of curvature. In addition, since the flux is more concentrated in the inner region of the core, it follows that an effective closure width, w_{eff} can be defined beyond which increasing the width dimension of the closure region of the device will result in minimal increase in inductance since very little flux flows near the outer radius of the core. Evaluating this effective width would help in designing area-efficient closed-loop inductors.

To calculate the flux through the curved region, an integral must be evaluated as a function of radius s drawn from the inner edge of the core. The flux density at a given radius, is expressed as:

$$H_{curve} = \frac{V_{mmf_2}}{l_g + \pi s} \quad (5)$$

(a) Single Return Path Reluctance Model

(b) Dual Return Path Reluctance Model

Figure 2: Reluctance models describing each of the a) single and b) dual return path topologies. The total reluctance can be represented as a combination of series and parallel reluctances, where the block $\mathcal{R}_{mag}$ describes the reluctance of the entire magnetic return path and $\mathcal{R}_{air}$ describes the reluctance due to fringing through air.

Figure 3: Distribution of the flux density inside the curved region of the magnetic core. Circles with crosses indicate flux lines into the cross-sectional plane, whereas the curve shows the approximate decrease in flux density as a function of position, $B(s)$, from the inner edge of the core.

$$B_{curve} = \mu_0 \mu_r H_{curve} = \frac{\mu_0 \mu_r V_{mmf_2}}{l_g + \pi s} \quad (6)$$

This expression is then integrated from the inner radius ($s = 0$) to the outer radius based on the width of the core in the closure region ($s = w_m$).

$$\Phi_{curve} = \int_0^{w_m} B_{curve} t_m ds = \frac{\mu_0 \mu_r V_{mmf_2} t_m}{\pi} \ln \left| \frac{l_g + \pi w_m}{l_g} \right| \quad (7)$$

$$\mathcal{R}_{curve} = \frac{V_{mmf_2}}{\Phi_{curve}} = \frac{\pi}{\mu_0 \mu_r V_{mmf_2} t_m} \left(\ln \left| \frac{l_g + \pi w_m}{l_g} \right| \right)^{-1} \quad (8)$$

An expression for the amount of flux that cuts across the gap can be determined once approximate dimensions for the area of the flux region are defined. As the flux crosses the gap, the length of the path is assumed to be the gap length, l_g , while the width can be approximated as half the length of the magnetic core in the flux-forming region, i.e. $l_m/2$, allowing for the flux to cut back across the gap in the other half. (Although this implies the flux travels only an infinitesimal length between these two sections, to first order, this approximation appears to give good results.) The thickness of the flux cross-section area is assumed to be the thickness of the core, t_m . Note that the magnetic field in this case is that in air and the permeability is that of

free space, μ_0 . In order to relate these two values to the magnetic field and flux inside the magnet, demagnetization effects must be taken into account to give an effective permeability based on the demagnetization factor along that dimension [4]. In other words, the magnetic field through air must be related to an equivalent magnetic field inside the magnet using the common expression $H_m = H_{air} - N_d M$. Assuming a demagnetization factor of N_{dw} along the width direction, the expressions for the flux and reluctance for the gap section are given by

$$H_{m_{gap}} = \frac{H_{a_{gap}}}{[1 + N_{dw}(\mu_r - 1)]} \quad (9)$$

$$B_{gap} = \mu_0 \mu_r H_{m_{gap}} = \mu_0 H_{a_{gap}} \quad (10)$$

$$\Phi_{gap} = B_{gap} t_m \frac{l_m}{2} = \frac{\mu_0 \mu_r V_{mmf_2} t_m l_m}{2l_g [1 + N_{dw}(\mu_r - 1)]} \quad (11)$$

$$\mathcal{R}_{gap} = \frac{2l_g [1 + N_{dw}(\mu_r - 1)]}{\mu_0 \mu_r t_m l_m} \quad (12)$$

The reluctance through the straight leg of the core is once again given by

$$\mathcal{R}_{leg} = \frac{l_m}{\mu_0 \mu_r w_m t_m} \quad (13)$$

With these three pieces, the reluctance of the magnetic core branch can be summarized as the combination of two sections where the curvature reluctance and the gap reluctance are in parallel plus a series reluctance of the straight magnetic leg:

$$\mathcal{R}_{mag} = \mathcal{R}_{leg} + \frac{2\mathcal{R}_{curve}\mathcal{R}_{gap}}{\mathcal{R}_{curve} + \mathcal{R}_{gap}} \quad (14)$$

For the reluctance contribution of the fringing flux through air, the flux density is once again a function of position away from the magnetic core. Therefore, a radius r is defined beginning from the edge of the core. The length of the flux path is the sum of the distance the flux travels radially out away from the core, the distance traveled down the length of the core, and the same radial distance traveled back in to the core ($2r + l_m$). Assuming a thin-film magnetic inductor, the cross-sectional area of the flux can be approximated as the integral of the infinitesimal area of the oval-shaped ring

formed by the locus of points at a distance r from the core and infinitesimal width of dr , $(2\pi r + 2w_m + 2t_m)dr$. Once again, in the case of the fringing flux, an effective permeability must be used in order to account for the demagnetization effects at the edge of the magnetic core (with N_{dl} representing the demagnetization factor along the length direction). After integrating the flux out to a maximum effective radius r_{eff} , the final expression for the air-related branch of the reluctance model is

$$\mathcal{R}_{air} = \frac{2[1 + N_{dl}(\mu_r - 1)]}{\mu_0\mu_r} \left(2\pi r_{eff} + (2w_m + 2t_m - \pi l_m) \ln \left| \frac{l_m + 2r_{eff}}{l_m} \right| \right)^{-1} \quad (15)$$

Finally, the total reluctance of the design can be defined as the reluctance of the flux-forming region in series with the parallel combination of the flux through the magnetic closure region(s) and the fringing flux through air. For the SRP design, only one magnetic closure region is used ($\mathcal{R}_{SRP} = \mathcal{R}_{nominal} + \mathcal{R}_{mag} \parallel \mathcal{R}_{air}$), while for the DRP design two magnetic closure regions are used ($\mathcal{R}_{DRP} = \mathcal{R}_{nominal} + \mathcal{R}_{mag} \parallel \mathcal{R}_{air} \parallel \mathcal{R}_{mag}$).

3. Results and Model Accuracy

Verification of the broad applicability of the model was done by comparing calculated results with simulations of both single and dual return path topologies of varied dimensions. Simulations were done using commercial Ansys Maxwell and Ansys HFSS software. For each topology, the width of the magnetic closure regions was varied from zero (meaning no magnetic return path) to the same width as that of the flux-forming region (denoted as w_m and, in this case, equal to 500 μm). Additionally, the isotropic permeability of the magnetic core was varied from $\mu_r = 10$ to $\mu_r = 10^4$. In the case of lower permeability cores (Fig. 4a), the degree of fringing increases because the magnetic core performs little better than air. Although the model predicts an inductance in the correct order of magnitude, its accuracy decreases in this extreme fringing regime. The analytical model performs very well for the range of permeabilities typical of integrated thin-film magnetic cores, as shown in Figs. 4b and 4c. Even when demagnetization effects become more pronounced for shorter closure widths, the incorporation of demagnetization factors into the effective permeabilities for the model give it the flexibility to accurately evaluate the inductance in these cases. The difficulty here comes

from calculating (or simulating) accurate demagnetization factors for given core dimensions. For very high permeabilities, like the $\mu = 10^4$ case in Fig. 4d, the model tends to overestimate the inductance. This inaccuracy is most likely due to underestimating the amount of fringing that occurs across the gap even for high permeabilities.

In an additional demonstration of the model’s flexibility, it was applied to discrete inductors where the thickness of the magnetic core was equal to the width of the core. In this case, the demagnetization effects become less of a consideration ($N_{dw} = N_{dt} = 0.5, N_{dl} = 0$). As a result, as seen in Fig. 5, the model accurately predicts the inductance regardless of permeability or closure width. This implies that perhaps even for thin-film magnetic inductors, the analytical model can become more accurate and robust simply with improved accuracy of the demagnetization factors alone.

4. Discussion

The results of the model give tremendous insight into the behavior of the flux in integrated magnetic inductors as well as the trade-offs involved in their design. For instance, in all cases, the improvement of the DRP over the SRP is evident, but can be seen to increase for higher permeabilities. Both the analytical model and simulation results indicate that the degree of enhancement of inductance in these closure topologies is strongly dependent upon the value of permeability; for low permeabilities, the enhancement is so small that to include a magnetic closure path would be an inefficient consumption of additional chip area.

Deviation of the analytical model from the simulation is understood to be mostly the result of fringing effects for which a number of approximations were made in order to keep the analytical expressions tractable. Nevertheless, the analytical expressions provide a basis for understanding the design and material trade-offs involved in integrated magnetic inductors, knowledge of which can better inform future designs of an entire class of integrated devices whose operation is based on magnetic flux closure.

Since the analytical model is entirely based on the material permeability and dimensions, it can be easily applied to a variety of other device designs. For instance, while the solenoid inductor design is discussed in great detail in this paper, by simply changing the dimensions of the magnetic core, the model can also be used to predict the inductance of the equally popular spiral inductor topology. Spiral inductors have been shown experimentally

Figure 4: Comparison of simulation results using Ansys software and calculated results using the analytical model. For typical thin-film magnetic permeabilities (b, c), the analytical model accurately predicts the inductance. For low and high permeabilities (a, d, respectively), a divergence between the analytical model and simulation results is seen which can be attributed to various fringing effects.

Figure 5: Comparison of simulation results using Ansys software for discrete inductors with equal thickness as width (500 um) and calculated results using the analytical model. Agreement is seen for all permeabilities and dimensions without the need to consider or evaluate specific demagnetization factors other than 0.5, 0.5, and 0 in the width, thickness, and length directions, respectively.

to have large inductance enhancements due to the use of closed-loop magnetic cores [18], the analytical model described here provides a direct prediction of how much. In order to do so, however, the analytical model must tackle the question of non-isotropic permeability; although the model was derived with the assumption of isotropic permeability, it is, in fact, capable of also describing films with non-isotropic permeability. Since each segment of the flux path is accounted for separately, the model is quite robust and can take in different permeabilities for each segment.

The simulated and calculated spiral inductor results shown in Fig. 6 represent a combined example of the analytical model's versatility in terms of both design topologies and isotropic/non-isotropic permeabilities. The inductor shown has two magnetic core segments, each wrapping vertically around part of the spiral inductor. If, for instance, the magnetic material is laminated, this may add the additional complexity that the out-of-plane permeability is much smaller than the in-plane permeability. For this demonstration, the spiral inductor was simulated with in-plane permeability values from 10 to 10^4 and out-of-plane permeability consistently equal to $10\times$ less than the in-plane permeability. The simplistic assumption that the out-of-plane demagnetization factor is one, while zero in other dimensions, gives a remarkably good fit for the simulated data. The only other free parameter in the model is the effective radius for the fringing flux calculation. For this, the same values were used for the spiral inductor topology as for the solenoid topology; the effective radius of fringing flux is seen to decrease by two orders

Figure 6: Demonstration of the analytical model’s accuracy for the spiral inductor topology and non-isotropic permeability. The spiral inductor shown was simulated for permeabilities between 10 and 10^4 with a reduced permeability ($10\times$ less) in the vertical direction. A demagnetization constant of one was used in the vertical dimension and zero in the others. Despite the non-isotropic permeability and the different inductor design, the analytical model is shown to have tremendous accuracy across the simulated range of permeabilities.

of magnitude for each single order of magnitude increase in permeability (i.e. for $\mu_r = 10$, $r_{eff} = 20\mu m$ and for $\mu_r = 100$, $r_{eff} = 0.2\mu m$). Qualitatively, this behavior illustrates how higher permeability magnets tend to guide flux more effectively inside the core, thereby reducing the amount of fringing flux and the distance it travels outside the core.

Furthermore, a note should be made about the applicable frequency range of the analytical model. For very high permeabilities, the penetration depth of the flux inside the magnetic core will be dependent upon the skin depth at the given frequency (Fig. 7). In most other cases with typical values of permeability for thin film magnetic materials ($< 10^3$), the frequency range will be set by the ferromagnetic resonance (FMR) of the material. While the analytical model does not currently include any frequency terms, the influence of FMR and eddy currents on the permeability spectrum can be incorporated into the model using a frequency dependent permeability either analytically [19] or experimentally derived [20, 21, 22].

5. Conclusion

A broadly-applicable analytical model has been derived to describe the flux-closure and inductance of integrated magnetic closed-loop devices. The

Figure 7: Both simulated and calculated results for inductance as a function of frequency for different values of isotropic permeability in the dual return path topology. Calculations were made using the analytical model by replacing the permeability with a frequency-dependent permeability term found in [19]. The eddy current-limited penetration depth for flux becomes detrimental to the closure only for very high permeabilities ($\geq 10^4$), which are not typical for thin film magnetic materials.

analytical model was first optimized for two in-plane flux closure topologies allowing for either dual- or single-sided flux return through a magnetic core. By accounting for the various paths by which the flux can return either through the magnet or fringing through air, the model is able to accurately predict the inductance for a wide range of designs and permeabilities beyond just the solenoid and in-plane isotropic permeability cases.

Agreement between simulation and calculated values using the analytical model is demonstrated for permeabilities across a range of several orders of magnitude and both isotropic and non-isotropic behavior. Finally, the model was demonstrated to also describe spiral inductor topologies, implying its robustness and applicability to an even broader range of possible integrated device designs. Given accurate demagnetization factors for specific device dimensions, the analytical model can be confidently applied to evaluate the inductance or total reluctance due to magnetic flux closure.

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