Evaluation of Exchange-Biased Magnetic Materials for High Density GHz Toroidal Integrated Inductors

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We evaluated the interest of exchanged-biased magnetic materials for integrated toroidal inductors with operating frequency above 1GHz. Exchange-biased magnetic materials enable designers to use closed magnetic cores up to 5GHz, which lead to very high inductance densities. We used simple considerations about DC resistance and gyromagnetic losses in order to estimate the upper limitations of such devices in terms of quality factor. These estimations were used in order to optimize the properties of toroidal magnetic core at 2GHz operating frequency. We used physical parameters based on experimental data for Fe₆₅Co₃₅ ferromagnetic layers coupled with NiMn, NiO or IrMn antiferromagnetic layers. We found that (Fe₆₅Co₃₅/IrMn) multilayers provide the best trade-off for integration and that quality factor limitations can be overcome by the use of thick ($\geq 1\mu\text{m}$) multilayer cores. Also, an important issue for high quality factor is to lower the damping parameter (≤ 0.01). Inductance densities of optimized devices were also calculated and were found to be in most cases higher than for air-core spiral inductors. These results may orientate material developments in order to meet the requirements for future high density microwave inductive devices fabrication.

Key words: integrated inductor, toroidal core, microwave applications, exchange-bias, quality factor, optimization

1. Introduction

Toroidal inductors with magnetic cores are commonly used in a wide range of low frequency applications such as power electronics¹⁾, but this kind of design is seldom used for integrated devices. Air-core spiral inductors are more commonly used because they can be fabricated with integrated technologies more easily. However, the use of magnetic materials in such inductors only provides a moderate increase of inductance²⁾⁻³⁾. Since it provides higher inductance density, toroidal design is interesting for down-sizing inductor footprint. Moreover, since magnetic field is mainly confined in magnetic core, one can expect lower substrate losses and crosstalk than in spiral inductor.

We recently designed and fabricated MHz integrated inductors for high switching frequency DC-DC power conversion and demonstrated high inductance to resistance ratio at MHz frequency⁴⁾. Here, we investigate the opportunities provided by this kind of inductor for applications in the GHz range. In section 2, we will show the advantages of exchanged-biased magnetic materials for integration in microwave toroidal inductor. In section 3, simple considerations will be taken into in order to estimate the upper limit of the quality factor of such inductors. In section 4, the model developed for quality factor estimation will be used in order to optimize the magnetic core of toroidal inductor for an operating frequency of 2GHz. In section 5, we will compute the inductance densities of the optimized inductors. In section 6 we will finally give our conclusion and give orientation of future work.

2. Interest of Exchange-Biased Materials

The main limitation of toroidal inductors in terms of operating frequency comes from the use of closed magnetic circuits which does not allow designers to tune the ferromagnetic resonance frequency of the material with demagnetizing field⁵⁾. Thus, the maximum operating frequency of the closed magnetic circuit is limited by the intrinsic resonance frequency of the material which is limited typically around 1 to 1.5GHz with conventional magnetic films (e.g. CoZrNb, Ni-Fe or Co-Fe alloys). This restricts the operating frequency below 1GHz. Therefore magnetic materials with a high resonance frequency are required. A possibility has been shown in recent work with exchange-bias coupled magnetic films having resonance frequencies exceeding 5GHz⁶⁾⁻⁷⁾. Extensive description of these materials is available in literature⁸⁾⁻⁹⁾. Here we will only focus on simple considerations in order to emphasize on the interesting features of exchange-biased materials for inductor design. These magnetic films are composed of sandwiched ferromagnetic (F) and antiferromagnetic (AF) layers with thicknesses t_F and t_{AF} , respectively. Exchange-bias coupling between F and AF magnetic layers results in an increase of the effective anisotropy field H_{KE} of F layers measured along hard axis. In a rough approximation, H_{KE} can be calculated from (1) with H_{KI} and M_S being the intrinsic anisotropy field and the magnetization of F layers, and J_{EX} the interfacial exchange coupling energy density which mainly depends on the nature and thickness of AF layers.

$$H_{KE} = H_{KI} + \frac{J_{EX}}{t_F \cdot M_S} \quad (1)$$

Fig. 1 Complex permeability μ_{CORE} of exchange-biased magnetic core for various values of H_{KE} . $\alpha=0.02$, $\gamma=1.76 \cdot 10^7 \text{ rad} \cdot \text{s}^{-1} \cdot \text{Oe}^{-1}$, $4\pi M_S=24000 \text{ G}$, $H_{KI}=30 \text{ Oe}$, $t_{AF}=50 \text{ nm}$ and $J_{EX}=0.6 \text{ erg} \cdot \text{cm}^{-2}$.

Complex permeability of ferromagnetic layers μ_{FERRO} along hard axis can then be calculated from (1) and classical Landau-Lifchitz-Gilbert gyromagnetic resonance model in which γ is the gyromagnetic ratio and α is the damping parameter:

$$\mu_{FERRO}(\omega) = 1 + \frac{\gamma \cdot 4\pi M_S \cdot (\gamma \cdot H_{KE} + \gamma \cdot 4\pi M_S + j\omega\alpha)}{(\gamma \cdot H_{KE} + j\omega\alpha) \cdot (\gamma \cdot 4\pi M_S + j\omega\alpha) - \omega^2} \cdot (2)$$

t_{AF} is fixed to the critical thickness for which J_{EX} is maximal. One notes that this value differs with the nature of the AF layers. Therefore t_F can be considered as the tuning parameter for H_{KE} . Also, the filling ratio ρ of the multilayer will depend on the effective anisotropy required as shown below:

$$\rho(H_{KE}) = \frac{t_F}{t_F + t_{AF}} = \frac{J_{EX}}{J_{EX} + (H_{KE} - H_{KI}) \cdot t_{AF} \cdot M_S} \cdot (3)$$

The complex equivalent permeability of an exchange-biased multilayer core μ_{CORE} can then be easily calculated from (2) and (3).

$$\mu_{CORE}(\omega) = 1 + \rho \frac{\gamma \cdot 4\pi M_S \cdot (\gamma \cdot H_{KE} + \gamma \cdot 4\pi M_S + j\omega\alpha)}{(\gamma \cdot H_{KE} + j\omega\alpha) \cdot (\gamma \cdot 4\pi M_S + j\omega\alpha) - \omega^2} \cdot (4)$$

$$\mu_{CORE}(\omega) = \mu_{CORE}' - j\mu_{CORE}''$$

Figure 1 illustrates the frequency profile of μ_{CORE} for various H_{KE} . If H_{KE} is increased by decreasing t_F , the resonance frequency can be shifted toward higher frequencies. Therefore intrinsic magnetic core properties can be finely tuned independently from demagnetizing field, which is very suitable for our purpose.

Moreover, these materials were recently integrated in functional integrated magnetic lines with open and closed magnetic circuits¹⁰⁻¹¹⁾. This indicates these materials being mature for applications in terms of reliability,

Fig. 2 Simplified infinite solenoid design : general view (a) and cross-sectional view (b). t_{WIN} and t_{CORE} are the thicknesses of winding and core, W_{WIN} and W_{CORE} are the widths of winding and core, and S is the spacing between winding turns.

reproducibility and resonance frequency adaptation. Therefore, it is now possible to design functional integrated closed magnetic circuits for applications up to 5GHz and exchange-biased material can be considered as realistic candidates for integration in toroidal inductors. However, a high quality factor has to be demonstrated.

3. Quality Factor Estimation

In this section, our purpose is to take into very simple considerations in order to estimate the upper theoretical limitations on the quality factor Q of toroidal inductors with exchange-biased material magnetic core. We will use a simplified approach. Calculations will be based on the case of the ideal infinite solenoid in which magnetic field is perfectly confined in the core. Proximity effects, substrate losses, stray capacitances and wave propagation are neglected. Eddy currents in core are assumed to be suppressed by suitable lamination with insulation, and magnetic losses are assumed to be perfectly described by classical Landau-Lifchitz-Gilbert gyromagnetic resonance model. Therefore the quality factor obtained can be considered as the ultimate limitation of our devices.

Figure 2 shows the design of an infinite solenoid. From this design inductance L , magnetic losses resistance R_{MAG} and DC resistance R_{DC} can be calculated from the simple formulas below.

$$L = \frac{\mu_0 \cdot \mu_{CORE}' \cdot N^2 \cdot W_{CORE} \cdot t_{CORE}}{L_{SOL}} \cdot (5)$$

$$R_{DC} \geq \frac{2N \cdot W_{CORE}}{W_{WIN} \cdot t_{WIN} \cdot \sigma_{WIN}} \cdot (6)$$

Table 1 Critical thicknesses t_{AF} and corresponding coupling constant J_{EX} of NiMn, NiO and IrMn sandwich-coupled with $Fe_{65}Co_{35}$.

AF	$J_{EX} (erg \cdot cm^{-2})$	$t_{AF} (nm)$
NiMn ¹²⁾	0.9	50
NiO ¹²⁾	0.6	25
IrMn ¹³⁾	0.14	10

Table 2 Optimized toroidal inductors with NiMn, NiO and IrMn AF layers: quality factor Q , optimal effective anisotropy field H_{KE} , filling ratio ρ , ferromagnetic layers thickness t_F and estimated number of layers n .

AF	Q	$H_{KE} (Oe)$	$\rho(\%)$	$t_F (nm)$	n
NiMn	9	230	32	24	15
NiO	6.5	170	17	5	33
IrMn	12	310	53	11	49

Table 3 Optimized toroidal inductors with IrMn AF layers and various damping parameters α and core thicknesses t_{CORE} : quality factor Q , optimal effective anisotropy field H_{KE} , filling ratio ρ , ferromagnetic layers thickness t_F and estimated number of layers n .

$t_{CORE} (\mu m)$	α	Q	$H_{KE} (Oe)$	$\rho(\%)$	$t_F (nm)$	n
0.5	0.03	9.5	360	49	9.5	53
0.5	0.02	12	310	53	11	49
0.5	0.01	18	240	60	15	41
0.5	0.005	26	190	66	20	35
1	0.02	16	400	46	8.5	109
2	0.02	21	520	39	6.5	145

$$R_{MAG} = L \cdot \frac{\mu_{CORE}'' \omega}{\mu_{CORE}'} \quad (7)$$

σ_{WIN} is winding conductivity, N is the number of winding turns and L_{SOL} is the length of the solenoid. We can now apply two design rules. First, the spacing S is assumed to be much smaller than W_{WIN} . Second, t_{WIN} is fixed to twice the skin depth at the operating frequency f_{OP} . We can assume that skin effects are therefore negligible. With these two approximations the upper limit of Q can be estimated from simple expressions.

$$Q \leq \frac{L\omega}{R_{MAG} + R_{DC}} = \left(\frac{1}{Q_{MAG}} + \frac{1}{Q_{DC}} \right)^{-1} \quad (8)$$

$$Q_{MAG} \approx \frac{\mu_{CORE}'}{\mu_{CORE}''} \quad (9)$$

$$Q_{DC} \approx 2 \cdot \mu_{CORE}' t_{CORE} \cdot \sqrt{\mu_0 \cdot \sigma_{WIN} \cdot \pi \cdot f_{OP}} \quad (10)$$

One can see from (9) and (10) that at low frequencies, Q is mainly limited by DC resistance and that at higher frequencies gyromagnetic losses are the main limiting factor. Moreover, if H_{KE} is increased, Q_{MAG} will be enhanced but Q_{DC} will be lowered. Therefore for each case there is an optimal value of H_{KE} which maximizes Q .

Fig. 3 Quality factor optimization of toroidal inductor with (a) NiMn, (b) NiO and (c) IrMn AF layers.

Fig. 4 Quality factor optimization of toroidal inductor with IrMn AF layers. (a) Influence of core thickness t_{CORE} . (b) Influence of damping parameter α .

4. Core Material Optimization

As an example, we will use (8) (9) (10) in order to perform core optimization of a toroidal inductor featuring copper winding ($\sigma_{WIN}=5 \cdot 10^7 S \cdot m^{-1}$), and core based on high magnetization $Fe_{65}Co_{35}$ alloy ($4\pi M_s=24000G$, $H_{KI}=30Oe$, $\gamma=1.76 \cdot 10^7 rad \cdot s^{-1} \cdot Oe^{-1}$ and $\alpha=0.02$) for the F layers. Targeted operating frequency was fixed to $f_{OP}=2GHz$ and we used typical parameters corresponding to NiMn, NiO and IrMn AF layers ¹²⁻¹³⁾ (see Table 1). NiMn provides the strongest coupling strength, NiO is an insulator, and IrMn exhibits the smallest critical thickness t_{AF} . Therefore all of these three materials exhibit special features which make them interesting for integration. Core thickness t_{CORE} is fixed to $500nm$, which can be considered as reasonably thick and realistic for fabrication.

Figure 3 shows that for each case, there is a specific value for H_{KE} which maximizes Q (see Table 2). The best

Table 4 Inductance density L_D of optimized toroidal inductors with IrMn AF layers and various damping parameters α and core thicknesses t_{CORE} .

$t_{CORE}(\mu m)$	α	$L_D(pH \cdot \mu m^{-2})$
0.5	0.03	0.08
0.5	0.02	0.10
0.5	0.01	0.15
0.5	0.005	0.21
1	0.02	0.14
2	0.02	0.19

trade-off is provided by the use of IrMn AF material. Therefore it can be considered in our case as the best candidate for integration. However, one can notice that Q remains moderate ($Q < 12$). This can be improved by the use of thicker core, as shown in Figure 4-a). Higher Q ($Q < 21$) can be obtained with core thickness around $2\mu m$. However, this kind of thick core may be difficult to fabricate because of the large number of layers n required (see Table 3). Moreover, the number of layers will be increased by extra laminations with insulating material which are necessary in order to suppress eddy currents.

Another key issue for Q is the damping parameter α . Figure 4-b) shows the influence of α , i.e. gyromagnetic losses, on Q . We can see that maximal Q can be dramatically changed by variations of α . A lower damping parameter may enable designer to use thinner core with fewer laminations (see Table 3). Therefore damping parameters below 0.01 will be highly suitable for high Q toroidal inductors, but such low values have not been demonstrated yet in exchange coupled multilayers. Moreover, it is known that α is increasing with strong exchange coupling when using thin F layers¹⁴⁾. This might be the main drawback of such materials.

5. Inductance Density Estimation

The inductance density L_D corresponding to each case detailed in Table 3 was evaluated. Again, simple design rules were used. Winding thickness t_{WIN} is twice the skin depth at 2GHz, i.e. around $3\mu m$. Therefore a reasonable dimension for spacing S is also around $3\mu m$, so that winding spacing shape ratio is not superior to 1. In section 3, we assumed S to be much smaller than W_{WIN} . Therefore, we will fix W_{WIN} to four times of S . Inductance density L_D can then be derived from M_S , H_{KE} , ρ , and (5) as below:

$$L = \mu_0 \cdot \left(1 + \rho \frac{4\pi M_S}{H_{KE}}\right) \cdot \frac{t_{CORE}}{(S + W_{WIN})^2} \cdot L_{CORE} \cdot W_{CORE}, \quad (11)$$

$$L_D = \mu_0^2 \cdot \left(1 + \rho \frac{4\pi M_S}{H_{KE}}\right) \cdot \frac{t_{CORE} \cdot \pi \cdot f_{OP} \cdot \sigma_{WIN}}{100}. \quad (12)$$

Equation (12) was applied to the optimized (IrMn/Fe₆₅Co₃₅)-based toroidal inductors, leading to results listed in Table 4. The estimated inductance densities are in most cases higher than those of classical spiral integrated inductor, which typically is around $5nH$ on a $250\mu m \times 250\mu m$ die area, i.e. $0.08pH \cdot \mu m^{-2}$ ²⁾. More

precisely, inductance density is much higher when using thick core ($\geq 1\mu m$) or low damping parameter core (≤ 0.01).

6. Conclusion and Prospects

These preliminary results showed that the design of integrated microwave toroidal inductor with both high inductance density and good quality factor may be possible with high magnetization exchange-biased magnetic core. IrMn was proven to be the AF material which provides the best quality factor at 2GHz. However, one of the two following severe conditions must be fulfilled with first a low damping parameter ($\alpha \leq 0.01$) or second a cumulated core thickness t_{CORE} above $1\mu m$. These indications may orientate further materials developments in order to meet the outstanding requirements for future very high density inductive devices. In future steps full electrodynamics finite element computation of GHz toroidal integrated inductors will be performed in order to compute Q more precisely. Substrate losses and crosstalks, which are crucial for integration in actual microwave circuit environment, will be investigated. Provided that high Q are possible, high magnetization exchange bias based toroidal inductors might offer the best alternative to spiral inductors in terms of global compactness on silicon, including reduced footprint, and much lowered spacing between components.

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