Soft Magnetic Materials and
Devices on Energy Applications

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1 Abstract

The fast development of wireless communication system in recent years has been driving the development of the power devices from different aspects, especially the miniaturized volume and renewable power supply, and etc. In this work, we studied the high frequency magnetic properties of the soft magnetic material -- FeCoB/Al$_2$O$_3$/FeCoB structures with varied Al$_2$O$_3$ thickness (2nm to 15nm), which would be applied in to the integrated inductors. Optimized Al$_2$O$_3$ thickness was found to achieve low coercive field and high permeability while maintaining high saturation field and low magnetic loss. Three types of on Si integrated solenoid inductors employing the FeCoB/Al$_2$O$_3$ multilayer structure were designed with the same area but different core configurations. A maximum inductance of 60 nH was achieved on a two-sided core inductor. The magnetic core was able to increase the inductance by a factor of 3.6 ~6.7, compared with the air core structures.

Vibration energy harvesting technologies have been utilized to serve as the renewable power supply for the wireless sensors. In this work, two generations of vibration energy harvesting devices based on high permeability magnetic material were designed and tested. The strong magnetic coupling between the magnetic material and the bias magnetic field leads to magnetic flux reversal and maximized flux change in the magnetic material during vibration. An output power of 74mW and a working bandwidth of 10Hz were obtained at an acceleration of 0.57g (g=9.8m/s$^2$) for the 1$^{st}$ generation design, at 54Hz. An output voltage of 2.52 V and a power density of 20.84 mW/cm$^3$ were
demonstrated by the 2\textsuperscript{nd} generation design at 42 Hz, with a half peak working bandwidth of 6 Hz.
2 Acknowledgement

I would like to take this opportunity to express my appreciation for my advisor Dr. Nian Sun, who has provided me the opportunity to pursue research on magnetic material and energy applications. He has been constantly guiding, supporting and encouraging me, without which I could not have been writing this dissertation. I sincerely appreciate all of his advice during my four-year PhD study, because it helped me master the techniques of solving different kinds of engineering problems and also made me grow.

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1 Introduction

Magnetic materials are traditionally classified by their volume magnetic susceptibility, $\chi$, which represents the relationship between the magnetization $M$ and the magnetic field $H$ strength, $M = \chi H$. The first type is diamagnetic, for which $\chi$ is small and negative $\chi \approx -10^{-5}$ and its magnetic response opposes the applied magnetic field. Examples of diamagnetics are copper, silver, gold, and etc. Superconductors form another special group of diamagnetics whose $\chi \approx -1$. The second group has small and positive susceptibility, $\approx 10^{-3} \sim 10^{-5}$ called paramagnets. It has weak magnetization but aligned parallel with the bias magnetic field. Typical paramagnetic materials are aluminum, platinum and manganese. The most widely recognized magnetic materials are the ferromagnetic materials, whose susceptibility is positive and much greater than 1, $\chi \approx 50$ to 10000. Examples of ferromagnetic material are iron, cobalt, nickel and several rare earth metals. $^{[1,1]}$

Based on the coercivity, ferromagnetic materials are divided in to two groups. One is called “hard” magnetic material, with coercivity above $10 \text{ kA} \text{ m}^{-1}$; the other group is call “soft” magnetic material, whose coercivity is below $1 \text{ kA} \text{ m}^{-1}$.

Ferromagnetic materials are widely applied in different aspects of everyday life and work, such as permanent magnets, electrical motors, magnetic recording, power generation, energy harvesting, and inductors.
1.1 Basic Magnetic Characteristics

The mostly desired essential characteristics for all soft magnetic materials are high permeability, low coercivity, high saturation magnetization and low magnetic loss.

1.2 Hysteresis Loop and Coercivity

In an external magnetic field $H$, the magnetic material gets magnetized and shows a finite spontaneous magnetization $M$. In the MKS unit, total magnetic flux $B=\mu_0 H+M$, Figure 1.1 shows a typical $H$ vs. $M$ magnetization hysteresis curve of a magnetic material. For a

![Hysteresis Loop Diagram](image)

Fig. 1.1 A typical magnetization hysteresis loop of magnetic materials.
soft magnetic material, \( M \) follows \( H \) readily; with a high relative permeability \( \mu=\frac{B}{\mu_0 H} \).

Soft magnetic materials differ from hard magnetic materials for their much smaller coercivity values.

### 1.3 Anisotropy

The dependence of magnetic properties on a preferred direction is called magnetic anisotropy. Different types of anisotropy include: magnetocrystalline anisotropy, shape anisotropy and magnetoelastic anisotropy.

Magnetocrystalline anisotropy depends on the crystallographic orientation of the sample in the magnetic field. The magnetization reaches saturation in different fields.

Shape anisotropy is due to the shape of a mineral grain. A magnetized body produces demagnetizing field which acts in opposition to the applied magnetic field.

Magnetoelastic anisotropy arises from the strain dependence of the anisotropy constants. A uniaxial stress can produce a unique uniaxial anisotropy.

The anisotropy constant \( K \) is defined as the volume density of anisotropy energy \( E_a \). The anisotropy field is the magnetic field needed to rotate the magnetization direction in the hard direction. \( H_k \) of a magnetic film can be read on the hysteresis loop along the hard axis, in Fig. 1.2.

### 1.4 Saturation Magnetization

The saturation magnetization is defined as the volume density of maximum induced magnetic moment, which can be shown on the hysteresis loop. The saturation
magnetization $4\pi Ms$ of the magnetic thin film can be read directly from the hysteresis loop measured along the out-of-plane direction, shown in Fig. 1.2.

Fig. 1.2 A typical hysteresis loop of the magnetic thin film.

### 1.5 Permeability

The magnetic permeability describes the relation between magnetic field and flux, $B=\mu H$, and $\mu=\mu_0\mu_r$, where $\mu_r$ is the relative magnetic permeability. Also, the relative magnetic permeability is related to the susceptibility by $\chi=\mu_r-1$. With the values of $H_k$
and $4\pi M_s$, which can be read from the hysteresis loops, the permeability of a magnetic thin film could be evaluated by 

$$\mu = \frac{4\pi M_s}{H_k} + 1.$$ 

1.6 Soft Magnetic Material

Fig. 1.3 Chronological summary of major developments of soft magnetic materials.
Soft magnetic materials are economically and technologically the most important of all magnetic materials, and they have been used to perform a wide variety of magnetic functions. Some applications demand high permeability; others emphasize low energy loss at high frequencies. \[1,2\] Figure 1.3 shows a chronological summary of major developments of soft magnetic materials. \[1,3\]

From the beginning of the 19\textsuperscript{th} century, research has been focused on the developing of higher permeability $\mu$, saturation magnetization $M_s$ and lower coercivity $H_c$. The advent of rapid solidification technology (RST) in the 1970s and 1980s provided metallurgists a route to novel compositions and microstructures.\[1,4\] Amorphous metals (also called metallic glasses) produced by RST were arguably the most important development in soft magnetic materials. Four major families of soft magnetic materials are: electrical steels, FeNi and FeCo alloys, ferrites, amorphous metals, and the typical properties are shown in Table 1.1 and 1.2.\[1,5\]

**Table 1.1 Major families of soft magnetic materials with typical properties.**

<table>
<thead>
<tr>
<th>Category</th>
<th>$B_s$ (T)</th>
<th>$\rho$ ($\mu\Omega$ m)</th>
<th>$H_{c,30}$ (Oe)</th>
<th>Typical Core Loss (W/kg) @ 50/60 Hz</th>
<th>Applications, Notes</th>
</tr>
</thead>
<tbody>
<tr>
<td>A. Steels</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>lamination (low C)</td>
<td>2.1-2.2</td>
<td>0.4</td>
<td></td>
<td>2.0 (60/1.0)</td>
<td>inexpensive fractional hp motors</td>
</tr>
<tr>
<td>non-oriented (2% Si)</td>
<td>2.0-2.1</td>
<td>0.35</td>
<td></td>
<td>2.7 (60/1.0)</td>
<td>high efficiency motors</td>
</tr>
<tr>
<td>convent. grain oriented (CGO M-4)</td>
<td>2.0</td>
<td>0.48</td>
<td>5000</td>
<td>0.9 (60/1.5)</td>
<td>50/60 Hz distribution transformers</td>
</tr>
<tr>
<td>high grain oriented (HGO)</td>
<td>2.0</td>
<td>0.45</td>
<td></td>
<td>1.2 (60/1.7)</td>
<td>50/60 Hz D15’s high design $B_{max}$</td>
</tr>
<tr>
<td>B. Fe3Ni(Co), 40-50Ni</td>
<td>[1.6]</td>
<td>[0.48]</td>
<td>[150,000]</td>
<td>[110] (50k/0.3)</td>
<td>high value used as thin ribbon</td>
</tr>
<tr>
<td>77-80Ni (square permalloy)</td>
<td>[1.1]</td>
<td>[0.55]</td>
<td>[150,000]</td>
<td>[40] (50k/0.2)</td>
<td>highest $\mu$, lowest core loss of any metallic material</td>
</tr>
<tr>
<td>79Ni-12Co (4-7 Mo permalloy, supermalloy)</td>
<td>[0.8]</td>
<td>[0.38]</td>
<td>$10^6$</td>
<td>[33] (50k/0.2)</td>
<td>highest $B_c$, commerical soft magnetic material</td>
</tr>
<tr>
<td>9PCo-2Ni (permendur, superendur)</td>
<td>[2.3]</td>
<td>[0.35]</td>
<td>[50,000]</td>
<td>[2.2] (60/2.0)</td>
<td></td>
</tr>
<tr>
<td>C. Ferrites</td>
<td>[0.5]</td>
<td>[2 \times 10^8]</td>
<td>[600]</td>
<td>[35] (30k/0.2)</td>
<td>power supply inductors, transformers</td>
</tr>
<tr>
<td>MnZn</td>
<td>[0.35]</td>
<td>[10^9]</td>
<td>[4000]</td>
<td></td>
<td>MHz applications</td>
</tr>
</tbody>
</table>
Reading heads employing soft magnetic materials are widely used in magnetic recording. The development in soft magnetic material has resulted in reduced size and improved efficiencies of power-handling electrical devices, such as motors, generators, inductors, transformers and other transducers.
REFERENCES


2 Experimental Methods

The self-made soft magnetic films used in this research were deposited by DC or RF magnetron sputtering, which is a physical vapor deposition (PVD) technique.

The magnetostatic properties of these magnetic films were measured with the vibrating sample magnetometer (VSM).

2.1 Physical Vapor Deposition (PVD)

Physical vapor deposition denotes the vacuum deposition processes, such as evaporation, sputtering, ion-plating, and ion-assisted sputtering. During the deposition, the coating material switches into a vapor transport phase, which does not generally rely on chemical reactions but by physical mechanism. In the current semiconductor industry, PVD technology is entirely based on physical sputtering.

During the magnetron sputtering process, the substrate is placed between two electrodes in a low-pressure, \(~10^{-7}\) Torr, vacuum chamber. The electrodes are driven by an RF power source, which generates plasma and ionized the rare gas (such as argon) between the electrodes. A DC bias voltage is used to drive the gas ions towards the surface of the cathode, which contains a target material, and knock off atoms from the surface. The free atoms then condense on the substrate surface to form a thin film. A
strong magnetic field is applied to contain the plasma near the surface of the target to increase the deposition rate. Figure 2.1 shows the schematic of magnetron sputtering. Figure 2.2 is the picture of the PVD system used to make all the self-made magnetic materials in this work.

![Schematic of magnetron sputtering](image)

**Fig. 2.1.** Schematic of magnetron sputtering.
2.2 Vibrating Sample Magnetometer (VSM)

A vibrating sample magnetometer (VSM) is an instrument that is used to measure the magnetostatic properties of magnetic films. As shown in Fig. 2.3[^2^], a VSM pick-up coil transducer, a lock-in amplifier, a gaussmeter and a controller. The electromagnet pair provides the magnetizing DC field. The piezoelectric oscillator physically vibrates the sample in the magnetic field at a fixed sinusoidal frequency. The motion of the sample
relative to the magnetic field modulates the magnetic flux, which is generally consists of

![Fig. 2.3 A schematic picture of VSM.](image)

an electromagnet pair, a piezoelectric mechanical oscillator, picked up by the detection coil. The detection coil generates the signal voltage due to the changing flux emanating from the vibrating sample. The lock-in amplifier is used to measure the voltage relative to the piezoelectric oscillation. The gaussmeter measures the applied DC magnetic field. The output is usually the magnetic moment M or flux B as a function of the field H. The picture of a VSM system is shown in Fig. 2.4. The entire system also consists of the power regulator and a computer interface.
Fig. 2.4 A picture of the VSM system.
2.3 Ferromagnetic Resonance (FMR) Detection

When a bias static magnetic field \( H \) is present, the magnetization vector \( M \) keeps the damped precession motion around \( H \) due to the magnetic torque until it gets aligned to the same direction because of the existence of the damping. The Gilbert form of damping is used to describe the relaxation mechanism. This equation takes the form:

\[
\frac{1}{\gamma} \frac{\partial M}{\partial t} = -(M \times H_{\text{eff}}) + \frac{\alpha}{\gamma M} (M \times \frac{\partial M}{\partial t}),
\]  

(2.1)

where \( \gamma \) is the gyromagnetic ratio and \( \alpha \) represents the Gilbert damping parameter.

Fig. 2.5 The processional motion of the moment \( M \) around the static magnetic field \( H \), before and after the dynamic component \( h \) applied.
If the magnetic field contains a dynamic component $h$ whose frequency equals the intrinsic frequency of the damping, the energy provided by the dynamic $h$ compensates the damping loss during the precession. In this case, the magnetization vector $M$ keeps rotating around $H$, as shown in Fig. 2.5. [2,3]

In the FMR measurement, the microwave signal is usually used as the dynamic magnetic field source. A complete testing system includes an electro-magnet pair, an AC coil, a lock-in amplifier, a coplanar waveguide, a crystal signal detector and a computer. The electro-magnet is able to generate large bias magnetic field, up to Tesla range. The AC coil provides a small AC magnetic field, and it is usually controlled by the output end of lock-in amplifier applying an AC current. The magnetic thin film is placed in the center area of the coplanar waveguide, where the film interfaces with the microwave signal. The output microwave signal at the other end is received by the detector. The

![Fig. 2.6 The FMR signal on (a) the absorbed power vs. static field plot and (b) $dP/dH$ vs. field plot.](image)

Fig. 2.6 The FMR signal on (a) the absorbed power vs. static field plot and (b) $dP/dH$ vs. field plot.

effect of the magnetic film could be seen by drawing the absorption curve of the microwave, as shown in Fig. 2.6 (a). Also, by doing the derivative of the absorbed energy
over the magnitude of the magnetic field, the linewidth could be easily read from the plot, shown in Fig. 2.6(b).

2.4 High Frequency Permeability Measurement

The complex permeability of magnetic materials, $\mu = \mu' - j\mu''$, is a very important magnetic characteristic under AC magnetic field. In order to meet the requirement of the miniaturization of electronics, the working frequency of magnetic materials has been increasing. At low frequency ranges, the permeability could be measurement by

![Image](image.png)

**Fig. 2.7 The permeability measurement system consists of a network analyzer and a coplanar waveguide.**
wrapping coils around the magnetic material. However, when it comes to MHz range, the parasitic capacitor due to the coils and the increased loss in the wires make the old method invalid.

New high-frequency permeameters have been developed to extract the frequency profile of the complex permeability of thin films at high frequencies. The testing system consists of a transmission line and a network analyzer, as shown in Fig. 2.7. Only the reflection and transmission parameter can be used to extract the permeability in the frequency domain. A pick-up coil working as a signal sensor is able to increase the sensitivity in low-frequency ranges. Figure 2.3 shows a typical permeability testing platform.

After getting the s-parameters from the network analyzer, the permeability can be simply obtain by

\[
\mu_r = \frac{Z_0}{j(k_f l \mu_0 \omega_0)} \left( \frac{1 + S_{11H_{bias}} - S_{21H_{bias}}}{1 - S_{11H_{bias}}} - \frac{1 + S_{11H_{ref}} - S_{21H_{ref}}}{1 - S_{11H_{ref}}} \right),
\]

(2.2)

where \(l, t, \mu_0, \omega_0, \) and \(Z_0\) are the sample length, sample thickness, permeability constant, angular frequency, and characteristic impedance of the fixture respectively. A scaling factor \(k\) (0.1 ≤ \(k\) ≤ 1) is determined from curve fitting and extrapolation of the relative permeability back to \(\mu = \frac{4\pi M_s}{H_0} + 1\) at zero bias and frequency ~ 0 Hz.
REFERENCES


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3 Soft Magnetic Thin Film:

FeCoB/Al$_2$O$_3$/FeCoB Structure with Varied Al$_2$O$_3$ Thickness

In this work, we studied the effects of the Al$_2$O$_3$ layer thickness on the RF magnetic properties of FeCoB(100nm)/Al$_2$O$_3$/FeCoB(100nm) sandwich structure, by measuring the magnetization hysteresis loop, FMR linewidth and permeability spectrum of each sample under different conditions.

3.1 Introduction

Magnetic/insulator multilayer films have attracted considerable interest. They show significantly reduced eddy current loss, and reduced out of plane anisotropy compared to single layer metallic magnetic films. A typical sandwich system is composed of two ferromagnetic layers with a non-magnetic interlayer in between. Numerous experiments have been done to understand the interlayer interaction of ferromagnetic layers through intermediate non-magnetic insulating layers, which in many instances determine various magnetic properties of the metallic magnetic films. The magnetic/insulator multilayer structures provide great opportunities for RF magnetic devices such as integrated magnetic inductors and transformers. Amorphous Fe$_{70}$Co$_{30}$B
films have been reported to have a high magnetization, low coercivity, high permeability and low magnetic loss at high frequency. Laminated FeCoB/insulator multilayer constitutes great core materials for integrated magnetic inductors. High magnetization\[3.1, 3.2\], low coercivity and large magnetic/insulator thickness ratio are preferred for achieving high effective permeability in the magnetic/insulator multilayer core. However, the thickness of the insulator layer is directly linked to the domain states of the multilayer\[3.3\] and is limited by the interlayer exchange coupling\[3.4, 3.5, 3.6\], causing high coercivity and large RF loss tangent. In addition, the insulator in metallic magnetic/insulator multilayer need to be thick enough so that it is pin hole free.

### 3.2 Experimental

The samples were prepared using the PVD (Physical Vapor Deposition) system; by DC sputtering the Fe\(_{70}\)Co\(_{30}\) target and RF sputtering the B and Al\(_2\)O\(_3\) targets at room temperature onto Si substrates to make the sandwich structure FeCoB(100nm)/Al\(_2\)O\(_3\)/FeCoB(100nm), and single layer FeCoB(200nm). The base pressure of the sputter system was 10\(^{-8}\) Torr. The thickness of the alumina layer was varied from 2nm to 15nm. Deposition rate of each layer was determined from the sputtering time vs. film thickness plot, which was evaluated by Dektak profilometer and NanoSpec optical spectrometer. A postannealing process was performed in a magnetic field of about 200Oe at 300\(^\circ\)C for 5 hours in the same PVD system vacuum chamber. Magnetization curve of the samples before and after the magnetic anneal were measured by a Vibrating Sample magnetometer (VSM). The ferromagnetic resonance absorption signals were tested with a field-swept ferromagnetic resonance testing system.
Permeability spectrum was measured with a microwave network analyzer.

Fig. 3.1 (a) Sandwich structure schematic (b) Coercivity of non-annealed sandwich structures decreases as the insulating alumina thickness increases, indicating reduced exchange coupling between adjacent FeCoB layers.

3.3 Hysteresis Loops

Figure 3.1(b) shows the coercivity of the FeCoB(100nm)/Al$_2$O$_3$/FeCoB(100nm) trilayers as a function of Al$_2$O$_3$ thickness before magnetic annealing. The schematic of the sandwich structure is shown in inset (a) of Fig. 3.1. The coercive fields drastically decreased from 5Oe to 0.5Oe as the alumina thickness ($d_A$) increased from 2nm to 6nm, and remained below 0.5Oe up to a $d_A$ of 15nm. Higher coercive fields at lower thicknesses (below 6nm) indicate the existence of interlayer coupling between the
ferromagnetic layers. As the Al₂O₃ thickness is increased from 2nm to 6nm, the interlayer coupling degrades, resulting in a significant decrease in the in-plane coercive field of the sandwich structure.

Figure 3.2(a) displays the saturation field of magnetization curves as a function of Al₂O₃ thickness from 2nm to 15nm. Higher saturation fields at lower $d_A$ indicate a stronger interlayer coupling for the trilayer structure. In the antiferromagnetic coupled region, a typical magnetization hysteresis loop for $d_A=2$nm is shown in inset (b) of Fig. 3.2. With further increase of $d_A$, from 6nm to 15nm, the interlayer coupling transits to the
ferromagnetic coupled region, for which the magnetization curve is shown in inset (c) of Fig. 3.2. The conclusion of antiferromagnetic coupling from Fig. 3.2 agrees with that observed from Fig. 3.1.

After annealing, the coercive field along the easy axis still follows the same trend as shown in Fig. 3.1(a). Somewhat differently, it drops below 0.5Oe in the case of \( d_A = 3\text{nm} \) and the magnetization curve is shown in Fig. 3.3(a). Out of plane magnetic hysteresis loops of all samples are nearly identical both before and after annealing, as shown in Fig. 3.3(b), indicating a large saturation magnetization, \( 4\pi M_s \), value over 1.5Tesla. It could be

![Hysteresis Loop Diagram](image)

**Fig. 3.3** Hysteresis loops of (a) sandwich structure with Al\(_2\)O\(_3\) thickness of 3nm after annealing, and (b) all samples in large field scale, after magnetic anneal. A coercive field as low as 0.5Oe was obtained for the trilayer FeCoB(100nm)/ Al\(_2\)O\(_3\)(3nm)/ FeCoB(100nm). High \( 4\pi M_s \) value of over 1.5 Tesla was obtained for all samples.
concluded that the critical $\text{Al}_2\text{O}_3$ thickness at which the antiferromagnetic coupling transfers to ferromagnetic coupling for annealed trilayer film is between 3nm to 6nm.

### 3.4 FMR Signals

The ferromagnetic resonance (FMR) reflects the dynamic process of magnetization and the motion of magnetic dipoles can be described by the following form of the Landau-Lifshitz equation:

$$\frac{dM}{dt} = \gamma (M \times H) - \frac{\alpha}{|M|} M \times \frac{dM}{dt}, \quad (3.1)$$

where $M$ and $H$ represent the magnetization and the magnetic field respectively, $\gamma$ is the gyromagnetic constant and $\alpha$ is the damping factor. The FMR linewidth of the annealed single layer FeCoB(200nm) film and all sandwich structures were measured. The sandwich structure of FeCoB(100nm)/$\text{Al}_2\text{O}_3$/FeCoB(100nm) is expected to solve two significant problems which exist in single layer films \cite{3,7}. One problem is the complicated magnetic domain structure which appears typically in single layer films. In order to form closure magnetic domain configuration, local magnetization in magnetic domains in particular on the edge of the film orients along the edge, not always parallel to the easy direction. The magnetic coupling between the neighboring FeCoB films can eliminate this domain structure. Another problem with single layer films is the larger magnetic loss due to the higher thickness. The inserted insulator spacer divides the thicker single layer into two thinner ones, in which the eddy current loss is decreased. FMR signal of the sample with $d_A=3$nm is shown in Fig. 3.4, in which the linewidth is 29Oe, compared to 50Oe of a single FeCoB (200nm) layer, at 8.5GHz.
Fig. 3.4 FMR signal along easy axis of annealed 200nm single layer FeCoB film and sandwich layer with Al₂O₃ of 3nm, at 8.5GHz.

In order to further study the interlayer coupling and magnetic loss within the sandwich structures, FMR signals were measured along the easy axis at 7GHz, 8GHz, 9GHz and 10GHz, respectively, as shown in Fig. 3.5. The acoustic mode appears at a certain value of magnetic field \( H \), which is determined by two equations

\[
f_{\text{FMR,HA}}^+ = \gamma \sqrt{(H - H_a)(H + 4\pi M_s)} \quad (\text{hard axis}) \tag{3.2}
\]

\[
f_{\text{FMR,EA}}^+ = \gamma \sqrt{(H + H_a)(H + 4\pi M_s + H_a)} \quad (\text{easy axis}) \tag{3.3}
\]

where \( M_s \) is the saturation magnetization of each FeCoB layer. \( H \) is the applied magnetic field. \( H_a \) is the in-plane uniaxial anisotropy field, induced by the magnetic annealing. With equation (3.2) and (3.3), \( H_a \) and \( 4\pi M_s \) were calculated from the FMR field along
Fig. 3.5 FMR signals along the easy axis for all samples at (a) 7GHz, (b) 8GHz, (c) 9 GHz and (d) 10 GHz. Optical modes are marked with arrows.
easy and hard axis for each sample at 9GHz, shown in Table 3.1. Equation (3.2) and (3.3) also indicate that as frequency increases the resonance peak moves to higher magnetic field. The main FMR peak is the acoustic resonance mode, referring to the in-phase precession between the neighboring ferromagnetic layers.\textsuperscript{[3,8]} Optical mode could also be predicted by these two equations\textsuperscript{[3,9]}

\[ J_{\text{eff}} = -2J_{\text{inter}} / dM \]  

\[ J_{\text{eff}} = -2J_{\text{inter}} / dM \]  

Where \( J_{\text{inter}} \) is the interlayer exchange coupling constant, and \( d \) is the thickness of a single ferromagnetic layer. The optical modes were observed for sandwich structures with Al\(_2\)O\(_3\) thickness below 6nm, and it represents out-of-phase precession.\textsuperscript{[3,10]} Equation (3.2) and (3.3) gave a straightforward relationship

\[ H_{\text{optical}} = H_{\text{acoustic}} + J_{\text{eff}}. \]  

The interlayer exchange coupling is determined by the term \( J_{\text{eff}} \). A positive \( J_{\text{eff}} \) indicates an antiferromagnetic coupling whereas a negative \( J_{\text{eff}} \) represents a ferromagnetic coupling.\textsuperscript{[3,11]} In Fig. 3.5, all optical modes are located at a higher field than their acoustic signals, showing positive \( J_{\text{eff}} \) and antiferromagnetic coupling. The existence of optical modes in sandwich structures with Al\(_2\)O\(_3\) thickness below 6nm also indicates that the interlayer antiferromagnetic coupling only happens in trilayers with thinner insulating layers. The interlayer exchange coupling coefficient \( J_{\text{eff}} \) for each sample, and the distance between the
acoustic mode and the optical mode are shown in Table 3.1. All positive values correspond to antiferromagnetic coupling, which agree with earlier conclusions. It is interesting to notice that the optical mode is much weaker than its acoustic mode. Although it does not decrease monotonically with the Al₂O₃ thickness, it tends to get closer to the acoustic peak in thicker Al₂O₃ cases, indicating weaker coupling in samples

Table 3.1 FMR Resonance Field along Easy and Hard Axis at 9GHz, In-plane Anisotropy, $4\pi Ms$ and Exchange Coupling Coefficient for Each Sample

<table>
<thead>
<tr>
<th>Al₂O₃ (nm)</th>
<th>$H_{EA}$ (Oe)</th>
<th>$H_{HA}$ (Oe)</th>
<th>$H_a$ (Oe)</th>
<th>$4\pi Ms$ (Tesla)</th>
<th>$J_{eff}$ (Oe)</th>
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<tr>
<td>2</td>
<td>589.8</td>
<td>618.7</td>
<td>14</td>
<td>1.65</td>
<td>64.6</td>
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<td>3</td>
<td>609</td>
<td>629.4</td>
<td>10.4</td>
<td>1.61</td>
<td>57.3</td>
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<td>4</td>
<td>609.7</td>
<td>633.6</td>
<td>12.2</td>
<td>1.6</td>
<td>83.7</td>
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<td>6</td>
<td>611.1</td>
<td>631.9</td>
<td>10.6</td>
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</tr>
<tr>
<td>10</td>
<td>611.1</td>
<td>629.3</td>
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<tr>
<td>15</td>
<td>608.1</td>
<td>628.8</td>
<td>10.5</td>
<td>1.61</td>
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</table>
with thicker insulator layers. This also means that the ferromagnetic coupling switches to antiferromagnetic coupling faster in field when the insulating layer gets thicker.

The measured FMR linewidth of each sample is pretty close, but slightly increases as the Al\textsubscript{2}O\textsubscript{3} thickness decreases. At 9GHz, the linewidth of the sandwich with Al\textsubscript{2}O\textsubscript{3} of 15nm is 33.2Oe, while that with Al\textsubscript{2}O\textsubscript{3} of 2nm is 34.8Oe.

As seen in all curves of Fig. 3.5, there is a 2nd FMR signal located on the lower field side of the main peak, which is a standing spin wave. It also shifts to upper field positions at higher frequency. The spin wave mode could be described as

$$H_n = H - \frac{2\pi^2 A n^2}{M_s d^2},$$

(3.8)

where $H_n$ is the required static field for exciting a particular spin-wave mode with order number $n$. $A$ is the exchange constant, related to the exchange coupling \[^{3.12}\]. The uniform processing mode corresponds to the case of $n=0$. The standing spin wave modes observed in Fig. 3.5 are most likely the case of $n=1$, which are the fist modes to the lower field side of the main peaks. $d$ is the thickness of the film. From equation (3.8), it is clear that the standing spin wave moves to higher field location with the uniform mode when the frequency increases. The standing spin wave is due to the coupling of the surface spin-wave modes of two adjacent magnetic layers, when the degeneracy of the FMR signals is lifted \[^{3.13}\].

### 3.5 Permeability

Figure 3.6 exhibits the change of permeability of the trilayers as a function of frequency. The real part of permeability is maintained above 600 for all samples, up to about 1GHz. The imaginary part indicate a zero field resonance frequency around
Fig. 3.6 Permeability spectrum measured under zero fields for all samples.

1.5GHz to 1.7GHz. The resonance frequency in each inset is close and floats around 1.5 GHz and it does not linearly depend on the thickness of the insulator layer. This matches with the observation of close resonance field of each sample at a fixed frequency from Fig. 3.5 and Table 3.1. The slight variation in the resonance frequency and field is very
likely related to the perturbation on the deposition environment of each sample, such as
the temperature and pressure. Also, it is interesting to notice a higher order mode in the
3nm case, which is due to spin wave resonance along the thickness direction \(^{[3,14]}\). This
means that the spin wave resonance is observed in both frequency-swept and field-swept
FMR measurements, therefore mutually confirmed.
3.6 Conclusion

Magnetization hysteresis loops, FMR curves and permeability spectrum were measured for a series of FeCoB(100nm)/Al₂O₃/FeCoB(100nm) sandwich structures with varied Al₂O₃ thickness. The coercive force dropped below 0.5Oe when the Al₂O₃ thickness increases above 3nm, indicating the transition of interlayer antiferromagnetic coupling to ferromagnetic coupling. The appearance of the optical modes in the FMR curves for the samples with thin Al₂O₃ layers also identified the antiferromagnetic coupling. The FMR linewidth of the sandwich structure was found to be much lower than that of a 200nm single layer FeCoB film. The permeability spectrums were tested to show high permeability of above 600 up to from 1.5GHz to 1.7GHz. The low coercive force, high $4\pi M_s$, narrow FMR linewidth and high permeability at high frequency region, make FeCoB(100nm)/Al₂O₃/FeCoB(100nm) structure a good candidate of RF device applications, which require soft magnetic films with high saturation magnetization and low magnetic loss.
REFERENCES


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4 Integrated Inductors with FeCoB/Al$_2$O$_3$ Multilayer Films

In this work, three different types of inductors were theoretically studied, fabricated and tested. A mathematical model was set up to calculate the inductance, resistance and quality factor. IC micro-machining technologies such as lithography, electrical plating, and lift-off were used to perform the fabrication. The FeCoB multilayer structure with optimized Al$_2$O$_3$ insulator thickness, which was discussed in Chapter 3, was deposited as the inductor core material. Vector network analyzer based RF probing system was used to implement the testing.

4.1 Introduction

Passive elements such as inductors, capacitors and transformers play a critical role in today’s wireless communication applications. They have been applied in areas such as oscillators, low-power converters and filters. The miniaturization and reliability of IC fabrication technology encourage the embedding of these passive components directly into the Si substrate. One challenge comes from the magnetic components that suffer from rapidly increased eddy current and magnetic loss at high frequencies. However, the micromachining technology is able to improve the control over the thickness and dimensions of the magnetic cores that optimize the magnetic properties at high frequency.

Table 4.1 shows the performance summary of the on-Si integrated RF inductors.
Table 4.1 Performance summary of the various bulk/surface micromachined RF inductors fabricated on silicon.

<table>
<thead>
<tr>
<th>Feature</th>
<th>Bulk micromachining</th>
<th>Spiral on porous silicon</th>
<th>Surface micromachining</th>
<th>Spiral on polyimide</th>
<th>Suspended spiral</th>
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<td>Yoon [19]</td>
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<td>Rogers [22]</td>
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<td>Self resonant frequency (GHz)</td>
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defined by conventional $Q = \frac{\Im(1/Y_n)/\Re(1/Y_n)}{\Re(1/Y_n)}$

$Q$ increased abruptly before peak $Q$

susceptible to Cu contamination since Cu deposited on silicon without any diffusion barrier

subject to substrate type and resistivity (µ= silicon, 0.0072Ω-cm) for the porous process

expected value

using the fabrication technology developed in this work

Si/SiGe

4.1.1 Quality Factor

The efficiency of an inductor is characterized by its quality factor $Q$. The fundamental definition of the quality factor is

$$Q = \frac{\omega L_s}{R_s},$$

where $L_s$ is the overall inductance and $R_s$ the conductor loss or serial resistance. An equivalent energy model is usually used to describe the efficiency, as shown in Fig. 4.1.

[4.4]
Fig. 4.1 Equivalent energy model representing the energy storage and loss mechanisms in a monolithic inductor. Note that $C_o = C_p + C_s$.

As a result, we have

$$Q = 2\pi \frac{\left| \text{Peak Magnetic Energy} - \text{Peak Electric Energy} \right|}{\text{Energy Loss in One Oscillation Cycle}}$$

$$= \frac{\omega L_s}{R_s} \times \frac{R_p}{R_p + \left[ \left( \frac{\omega L_s}{R_s} \right)^2 + 1 \right] \cdot R_s} \times \left( 1 - \frac{R_s^2 C_0}{L_s} - \omega^2 L_s C_0 \right),$$

(4.2)

where $\omega L_s / R_s$ accounts for the magnetic energy stored and the ohmic loss in the conductor. The second term stands for the loss factor due to the silicon substrate. The last term is the self–resonance factor due to the increase in the peak electric energy with frequency and the vanishing of $Q$ at the self-resonance frequency.
4.1.2 Self-resonance Frequency

The impedance of an inductor is a complex function of the frequency, which consists of the real part (resistance) and the imaginary part (reactance). Because of the parasitic capacitance between neighboring turns, the reactance of an inductor is determined by the combined effect of the capacitance and the inductance. As frequency increases, the reactance becomes zero at some frequency point, and remains negative afterwards. This critical frequency value where the reactance is zero is defined as the self-resonance frequency \( f_{SR} \). As shown in Fig. 4.2, the quality factor drops below zero when the reactance of the inductor becomes capacitive.

---

![Graph showing the quality factor as a function of frequency](image)

**Fig. 4.2** The quality factor as a function of frequency.
4.1.3 Different Types of Integrated Inductor

The discrete inductors are usually made by wrapped coils, and soldered onto PCB board, while an on-chip integrated inductor is fabricated on Si substrate by IC micro-fabrication technologies. For a magnetically enhanced on-chip inductor, the numbers of coil and winding geometries have been investigated including spiral, toroidal, solenoid and meander structures. [4.5]

Spiral Structure

One of the earlier applications of the spiral-type inductive components was integrated magnetic recording heads, which have been investigated since the 1970s. [4.6][4.7]

Figure 4.3 shows a top view and a side view of the spiral inductors. Figure 4.4 shows the equivalent circuit. The quality factor could be calculated as

\[
Q = 2\pi \cdot \frac{|E_{p\text{-magnetic}} - E_{p\text{-electric}}|}{E_{loss}} = \frac{\omega L_0}{R_0} \cdot f_{\text{sub,}} \cdot f_{\text{LC}} \tag{4.3}
\]

A spiral inductor fabricated on top of a 90nm CMOS process is shown in Fig 4.5. [4.3] A converter serving spiral structure based inductor was simulated to demonstrate a high inductance of 1.2µH, which consists of copper windings sandwiched between two laminated NiFe layers shielding the magnetic flux induced from the inductor windings. [4.8] A planar micromachined spiral inductor at low (Hz – kHz) frequencies with a
magnetic core was claimed to have an inductance of 2.2 $\mu$H/mm$^2$, four to five times greater than its air-core reference.\cite{4.9}

Fig. 4.3 (a) Top view and (b) side view of a 3.5-turn square spiral inductor.

Fig. 4.4 The equivalent circuit of a spiral inductor.
Fig. 4.5 Optical microscope images of copper/polyimide based 4-turn elongated spiral inductor with magnetic material fabricated in a 90 nm CMOS process.

The magnetically sandwiched spiral inductors suffer from high copper loss, because the magnetic field generated by the top and bottom magnetic layers crosses the coils. Due to the large air gap between the neighbor turns, the number of turns needs to be very large in order to obtain high inductance, as a result, leading to large dimension. However, it does have high bulk energy \( W_{max} = \frac{1}{2} \cdot L \cdot I_{max}^2 \) and does not have via problems, which is brought in by the fabrication process.
Toroidal Structure

An integrated toroidal inductor is fabricated on a silicon wafer by using a multilevel metallization technique to realize the coils, as shown in Fig. 4.6.\textsuperscript{[4,10]} In one reported toroidal structure, a wrapped coil was wound around a 30 µm thick micro-machined NiFe bar, which can produce a closed magnetic circuit as well as minimized leakage flux. A 4mm by 1mm by 130 µm dimension is able to achieve 0.4 to 0.1 µH at 1 kHz to 1 MHz.\textsuperscript{[4,11]}

The drawbacks of toroidal inductors include the complexity of via process, the high serial resistance, fast saturation, low bulk energy.

\begin{figure}
\centering
\includegraphics[width=0.5\textwidth]{toroidal_structure.png}
\caption{3-D illustration of the on-chip integrated toroidal inductor. The circled section represents a unit turn.}
\end{figure}
Meander Structure

A meander inductor is shown in Fig. 4.7. The three-dimensional meander magnetic core (NiFe) is wrapped by the planner meander shape coil to form flux loop. An inductance of 30 nH/mm² was achieved at 5 MHz with 30 turns and a dimension of 4mm by 1mm. The quality factor of a meander inductor could be expressed as

$$Q = \frac{\omega \mu_0 \mu_r N A_c A_w}{2(W+L) \rho l_c},$$

(4.4)

Where $A_w$ is the cross section area of conductor, $2(W+L)$ is the length of one meander coil turn, and $\rho$ is the resistivity of conductor material.

Fig. 4.7 Schematic diagrams of (a) the meander type integrated inductor with multilevel magnetic core and (b) the more conventional solenoid-bar type inductor. The structure of the two inductor schemes is analogous.
The meander inductor also has three layers. As a result, the via process of the magnetic material adds to the complexity, which also introduces the parasitic air gaps. Besides, it suffers from the low bulk energy, too. However, because of the single electrical layer, it does not have high serial resistance problem.

**Solenoid Structure**

The electroplated solenoid integrated inductor has several advantages over its counterparts. It is more compact because only the bottom conductor lines occupy areas on the substrate. A relatively small increase in the area or number of turns leads to high inductance. A typical layout is shown in Fig. 4.8. An electrically tunable integrated radio frequency inductor based on a planar solenoid with a thin-film NiFe core was reported to achieve 85%, 35% and 20% at 0.1, 1 and 2 GHz, respectively, for inductances in the range of 1 to 150 nH. The calculation of an ideal solenoid inductor can be written as

\[
L = \frac{\mu_0 \mu_r N^2 A_c}{l_c}, \tag{4.5}
\]

where \(A_c\) is the cross-sectional area, \(l_c\) is the length of the closed magnetic core. The quality factor is

\[
Q = \frac{\omega L}{R} = \frac{\omega \mu_0 \mu_r N A_c A_w}{2 W \rho l_c}, \tag{4.6}
\]

where \(A_w\) is the cross-sectional area of the conductor, \(2W\) is the length of the coil per turn and \(r\) is the resistivity of the metal conductor.
4.1.4 The Magnetic Core

From the material point of view, the magnetic core materials need to exhibit low coercivity, high magnetization, and low magnetic loss. The magnetic loss can be defined as the ratio of the imaginary and real part of the permeability, $\tan\delta_\mu = \mu''/\mu'$. The magnetic thin films are often patterned to reduce eddy current, and the shape-induced magnetic anisotropy would also increase the resonance frequency.\[4.15\] [4.16] Other than that, reducing the thickness of the films or using laminations are also able to increase the resonance frequency of the inductors while reducing the magnetic loss. A 15% tuning range and a quality factor between 5 and 11 up to 5 GHz were reported of an on-chip inductor that used a NiFe film. The eddy currents and ferromagnetic resonance are suppressed in the permalloy film using patterning and lamination in order to enable operation in the RF range.\[4.17\]
From the fabrication point of view, the magnetic material should be fully compatible with CMOS processing. As a result, they need to come with matured deposition and etching techniques and high-temperature stability. To develop a technique to prepare magnetic films that is fully compatible with standard CMOS technology processing is challenging. Besides FeCoB, two commonly used soft magnetic materials are Co-4.5%Ta4.0%Zr and Ni-20%Fe. They could be prepared using magnetron sputtering deposition. Fig. 4.9 shows a cross-section view of an integrated inductor.\cite{4.18}

![Cross-sectional SEM image of inductors integrated on an 130 nm CMOS process with 6 metal levels. Two levels of CoZrTa magnetic material were deposited around the inductor wires using magnetic vias to complete the magnetic circuit.](image)

\textbf{Fig. 4.9} Cross sectional SEM image of inductors integrated on an 130 nm CMOS process with 6 metal levels. Two levels of CoZrTa magnetic material were deposited around the inductor wires using magnetic vias to complete the magnetic circuit.
4.2 Prototype and Layout Design

As discussed earlier, the solenoid type inductors have the advantages of compactness and high magnetic energy density. In order to study the effect of the number of turns and the magnetic core shape on the inductance and quality factor, three types of inductors were designed, fabricated and tested.

Type I design is shown in Fig. 4.10. This is a very conventional structure, in which coils are wrapped around a rectangular magnetic core. The picture on the right is

Fig. 4.10 Inductor design type I. Left top is a schematic top view, where the blue pads indicate the vias. Left bottom is the 3-D schematic. The red rectangular on the top is the top layer metal consisting the top layer of the coil, while the blue rectangular is the bottom layer. The top and bottom metal are connected by vias, which are in green and yellow. The right picture is the 6-layer layout design done with Cadence.
Fig. 4.11 The cross sectional view of inductor type I, shown in Fig. 4.10, along (a) the transverse direction and (b) longitudinal direction. The yellow parts indicate the metal, the dark blue parts are the magnetic cores and the light blue area is the polyimide. The grey part stands for the Si substrates.

the 6-layer layout design of a 10-turn type I inductor for the mask fabrication, and it was drawn with the Cadence layout tool. The big rectangular loop along the contour is
designed for the ground signal connection purpose during testing. It also serves as the shield for each single device, so that the testing signal is not affected by the devices around it. The cross sectional views are shown in Fig. 4.11.

Type II design is shown in Fig. 4.12, in which two separated cores are applied and the coils distribute from one side to the other side, so that the cores generate opposite magnetization and tend to have a closed flux loop.

![Fig. 4.12 The schematic drawn of inductor type II, the top view (left top), 3-D view (left bottom) and a 6-layer layout of a 10-turn structure (right).](image)

Type III design is shown in Fig. 4.13. Two separate cores are also applied in the design. However, they are connected on the sides, with two triangle pieces. Thus, the core makes a loop shape, configuring a closed flux loop. Certainly, the coils need to be designed so that the magnetizations on two sides have opposite directions.
Fig. 4.11 The schematic pictures of inductor type III, the top view (left top), 3-D view (left bottom), and a 6-layer layout of a 10-turn device.

Five different numbers of turns were designed for each inductor type -- 6, 8, 10, 12, 14, respectively. Air core structures were also included into the layout, in order to study the core effect. Parameters like the core dimensions, coil width and gaps are shown in Table 4.2. The core dimension for Type I is 2 mm by 2.4 mm; That for Type II and III is 1mm by 2.4mm for each side. Open structures, like the one shown in Fig. 4.14 are especially designed for the testing purpose, which will be discussed in the following sections. Some testing structures were designed on the layout to monitor the fabrication quality, as shown in Fig. 4.15.
Fig. 2.14 The layout design of an 8-turn inductor Type I (upper) and its open structure with only the contour ground structure and the testing pads (lower). The cross arrays in the open area are the aligning marks.
Fig. 4.13 Layout design of the testing structures, including the via resistance testing bars, polyimide step monitor, precision monitor and sheet resistance testing pads.
4.3 Theoretical Model

4.3.1 Inductance

The inductance of an inductor has two components, self $L_{\text{self}}$ and mutual inductance $M$.

\[ L = L_{\text{self}} + M. \]  \hspace{1cm} (4.7)

Self Inductance

An alternating magnetic field can be induced when an alternating current is flowing along a conductor, according to Ampere’s law. The self inductance of a rectangular conductor is written as

\[ L_{\text{self}} = 2 \cdot l \cdot \left\{ \ln \left( \frac{2 \cdot l}{w + t} \right) + 0.5 + \frac{w + t}{3 \cdot l} \right\}, \]  \hspace{1cm} (4.8)

where $L_{\text{self}}$ denotes the conductor inductance in nH, $l$, $w$ and $t$ represent the length, width and thickness of the conductor in cm, respectively.

By following equation 4.8, the self inductance of a rectangular conductor could be calculated. Figure 4.16 shows the calculated results for 1 µm thick rectangular conductors with different length and width.\textsuperscript{[4.20]} The dimension of the conductor does have effects on the self inductance, because the inductance is mainly determined by the outer magnetic flux generated by the conductor. The inductance of a single loop in nanohenries is given by\textsuperscript{[4.21]}
Fig. 4.16 Self inductance value for a rectangular conductor versus its length and width (the thickness is fixed at 1µm).

\[ L = 4\pi a \left[ \ln \left( \frac{8\pi a}{\omega} \right) - 2 \right], \]  

(4.9)

where \( a \) is the radius and \( w \) is the width of the strip, both in centimeters.

The AC self inductance of a solenoid type inductor is given by \(^{4,25}\)

\[ L_{\text{self (ac)}} = L_{\text{(dc)}} \frac{s}{s \frac{\sinh \frac{s}{\delta_c} + \sin \frac{s}{\delta_c}}{\cosh \frac{s}{\delta_c} + \cos \frac{s}{\delta_c}}}, \]  

(4.10)

where \( s \) the thickness of a single lamination of the core and \( \delta_c \) is the skin depth of the core, and
\[ L_{(dc)} = \mu_c N^2 A/l_c, \quad (4.11) \]

which is the DC self inductance, with \( N \) the number of turns, \( \mu_c \) and \( l_c \) the equivalent permeability and length of the core, respectively.

**Mutual Inductance**

Mutual inductance \( M_{12} \) may be defined as the inductive influence of one coil or circuit upon another, defined by

\[ M_{12} = \frac{d\Psi_{12}}{dl_1}, \quad (4.12) \]

where \( \Psi_{12} \) is the flux generated by circuit 1, \( I_1 \) is the current flows in circuit 1. The mutual inductance in nanohenry between two parallel conductors is

\[ M = 2 \cdot l \cdot Q, \quad (4.13) \]

where \( l \) is the length of the conductor in centimeter and \( Q \) is a geometry coefficient, defined by \[^{[4,22]}\]

\[ Q = \ln \left[ \frac{l}{GMD} + \sqrt{1 + \frac{l^2}{GMD^2}} - \sqrt{1 + \frac{GMD^2}{l^2} + \frac{GMD}{l}} \right], \quad (4.14) \]

where GMD is approximately the average length of the conductors.
One inductor could always be considered as two half parts. Each half has mutual inductance with respect to the other half. The total mutual inductance is the sum of them if they are connected in series along the same direction. Otherwise, the mutual inductance due to the two halves tends to cancel each other.\(^{[4,23]}\)

The AC mutual inductance of a solenoid type inductor was derived as\(^{[4,24]}\)

\[
M_{(ac)} = R_{w(ac)} \frac{A}{2\pi f} \left[ \frac{e^{2A} - e^{-2A} - 2\sin(2A)}{e^{2A} + e^{-2A} - 2\cos(2A)} + 2 \frac{N_{L}^{2} - 1}{3} \cdot \frac{e^{A} - e^{-A} + 2\sin(A)}{e^{A} + e^{-A} + 2\cos(A)} \right],
\]

(4.15)

where \(R_{w(ac)}\) is the AC winding resistance and \(A\) is a dimensionless quantity, which depends on the winding conductor geometry. For a strip wire of width \(a\) and height \(b\), the quantity \(A\) is written as\(^{[4,25]}\)

\[
A = \frac{b}{\delta} \sqrt{\frac{a}{p}},
\]

(4.16)

where \(\delta\) is the skin depth of the winding metal, \(p\) is the winding pitch or the distance between the centers of two adjacent conductor lines.

The total AC inductance of a solenoid type inductor is \(L_{ac} = L_{self(ac)} + M_{(ac)}\).

4.3.2 Resistance and Skin Effect

The frequency dependence of the resistance of a conductor results from the non-uniform distribution of the current density through the cross section. As the frequency
increases from DC to higher range, the current density tends to be higher near the skin, which is so called skin effect. Eddy current could be generated when a conductor is placed in an alternating magnetic field. According to Lenz’s law, conductors tend to generate eddy current in order to resist the external magnetic field. The effective distribution depth of the current from the skin, defined as skin depth or depth of penetration, could be written as

\[ \delta = \frac{1}{\sqrt{\pi f \mu_c \sigma_c}}, \]

(4.17)

where \( f \) is the frequency, \( \mu_c \) is the permeability of the core material, \( \sigma_c \) is the conductivity of the conductor. Generally, the skin depth decreases as the frequency increases, and the resistance increases accordingly. The resistance due to the skin effect is

\[ R = \frac{\rho \cdot l}{w \cdot \delta (1 - e^{-t/\delta})}, \]

(4.18)

where \( \rho \) and \( l \) represent the resistivity and length of the conductor.

The AC winding resistance of a solenoid type inductor was derived to be

\[ R_W = R_{W(d)} A \cdot \left[ \frac{e^{2A} - e^{-2A} + 2 \sin(2A)}{e^{2A} + e^{-2A} - 2 \cos(2A)} + 2 \cdot \frac{N_l^2 - 1}{3} \cdot \frac{e^A - e^{-A} + 2 \sin(A)}{e^A + e^{-A} - 2 \cos(A)} \right]. \]

(4.19)
where A is a dimensionless quantity, which depends on the winding conductor geometry. The first and second terms in equation 4.15 represent the skin effect and the proximity effect contributions to the winding AC resistance, respectively.

### 4.3.3 Lamination Core Loss

For a given magnetic material and frequency, the skin depth of the core \( \delta_c \) can be calculated with equation 4.13, and the effective resistance \( R_e \) from eddy current loss can be approximated as \(^{[4.29]}\)

\[
R_e \approx \frac{\pi^2}{3} L_0 \mu_c \sigma_c f^2 t^2 \quad \text{for} \quad \frac{t}{\delta_c} < 0.5, \tag{4.20}
\]

where \( L_0 \) is the DC inductance, and \( t \) is the thickness of the lamination. The skin depth of the FeCoB at 45 MHz is around 1.1 \( \mu \)m, calculated by equation 4.13. Generally, 1.1 \( \mu \)m thick laminations should be sufficient to prevent the substantial eddy current losses in the 20 MHz to 50 MHz range. However, from equation 4.14, the effective resistance is proportional to the square of the lamination thickness when it is smaller than 550 nm. Hence, the loss can be significantly reduced by further reducing the lamination thickness. \(^{[4.30]}\)

The equivalent AC resistance of the core inside a solenoid type inductor is given by \(^{[4.25]}\)

\[
R_c = \omega L_{(dc)} \frac{\delta_c}{s} \sinh \frac{s}{\delta_c} - \sin \frac{s}{\delta_c} \left( \cosh \frac{s}{\delta_c} + \cos \frac{s}{\delta_c} \right), \tag{4.21}
\]
The total AC resistance $R_{ac}$ of an integrated inductor is the sum of the winding resistance $R_w$ and core equivalent series resistance $R_c$, $R_{ac} = R_w + R_c$.

### 4.3.4 Parasitic Effects

**Substrate Parasitics**

The electromagnetic field generated by the on-chip integrated inductors interfaces with the Si substrate, and induces substrate parasitic loss. For the solenoid type integrated inductors, the substrate parasitic capacitor is formed between the bottom coils and the Si substrate, with the dialectical layer in between.

**Parasitic Capacitance between Lines**

The parasitic capacitance also exists between each neighboring coil lines, both on the top layer and bottom layer. The capacitance between two top conductor lines $C_t$ and that between two bottom lines $C_b$ are described as:

$$
C_t = C_b = \frac{\varepsilon A}{d} = \frac{\varepsilon (w \cdot b)}{s},
$$

where $\varepsilon$ is the dielectric constant of air, $b$ is the thickness and $w$ is the length of the conductor line.

### 4.3.5 Quality Factor of the Solenoid type Inductors

An inductor could be equivalent to a circuit, shown in Fig. 4.17. The equivalent series resistance $R_s$ and reactance $X_s$ are
\[ R_s = \frac{R_{ac}}{(1 - \omega^2 L_{ac} C)^2 + (\omega CR_{ac})^2}, \tag{4.23} \]

\[ X_s = \frac{\omega L_{ac}(1 - \omega^2 L_{ac} C - \frac{C R_{ac}^2}{L_{ac}})}{(1 - \omega^2 L_{ac} C)^2 + (\omega CR_{ac})^2}. \tag{4.24} \]

Fig. 4.17 Equivalent circuit of an integrated inductor.

Hence, the equivalent series inductance \( L_s \) is \( L_s = \frac{X_s}{\omega} \). The quality factor could be

\[ Q = \frac{X_s}{R_s}. \tag{4.25} \]

### 4.4 On-chip Fabrication

#### Step 1: PI1 formation

Polyimide layer-1 formation on the silicon substrate, shown in Fig.4.18. The adhesion promoter VM-652 was pre-coated on the silicon surface. The polyimide PI2611 was spin coated with a Laurel WS-400 Lite series spin processor. Because the polyimide is usually very thick, a multi-step spin speed is usually used to form a uniform coating. The 3-step spin speeds were set to be 500 rpm for 30 seconds, 1000 rpm for 10 seconds
and 2000 rpm for 30 seconds. 20 minutes of softbake on the hot plate was taken afterwards at 150°. The curing process of PI2600 was performed in a nitrogen gas flowing furnace tube. A slow heating up process is very important, so that the polyimide surface does not crack. The cured polyimide layer was measured to be 10 µm.

**Step 2: Deposit bottom coil seedlayer**

A seedlayer of the bottom metal was deposited with PVD, shown in Fig.4.19. The seedlayer consists of a 20nm Cr layer followed by a 100nm Cu layer. The Cr layer is able to improve the adhesion between the Cu and polyimide surfaces.

**Step 3: Photo bottom coil**

A thick layer of photoresist AZ P4620 was spin coated and patterned above the seedlayer, as shown in Fig.4.20. The exposure time is 60 seconds, and the developing took 2 minutes with the diluted developer AZ400K (1:4 DI water).
Fig. 4.20 A thick photoresist layer was patterned above the seedlayer.

Step 4: Electrical plating of the bottom coil

A commercialized electrolyte was used for the electrical plating process. The current density was set to be 0.14 mA/mm². The coil thickness was measured with the Dektek profilometer, which was around 6 µm. Acetone was then applied on the wafer to strip the photoresist layer. This step is shown in Fig.4.21.

Fig. 4.21 Electric plating of the bottom coils and the removal of PR.
Step 5: Etch coil seedlayer

The wet etching of the seedlayer requires two different solutions. In order to remove the Cu seedlayer on the top, the diluted Ammonium hydroxide with a drop of hydrogen peroxide (50:1:300 DI water) was used. To remove the Cr seedlayer, an alkaline solution was made, so that the Cu layer was not affected. This alkaline solution had 3 units of potassium ferricyanide, 2 units of sodium hydroxide and 100 units of DI water. After the removal of the seedlayer, the wafer looks like that shown in Fig.4.22.

![Fig. 4.22 The schematic view of the wafer (upper) and a zoom-in picture of the real wafer (lower) after the Cr/Cu seedlayer removed.](image)

Step 6: Polymide layer-2 patterning

A second layer of polyimide (PI4100) was spin coated and patterned above the bottom coil to form a 10μm thick dielectric layer, as shown in Fig.4.23. A lithography
process was performed in order to open the via openings. Because of the thickness of the polyimide, the developing process might need to be repeated for two to three times.

Fig. 4.23 Polyimide layer-2 coating and patterning.

Step 7: PR patterning for magnetic core

As discussed in Chapter 3, the multilayer structure of FeCoB/Al$_2$O$_3$ serves as a good magnetic core with high permeability and low loss. Due to the difficulty of the wet etching of this multilayer structure, a lift-off patterning method was used for the magnetic core. As a result, a thick photoresist layer was spin coated and patterned above PI2 layer, as shown in Fig.4.24.

Fig. 4.24 Photoresist layer coating and patterning for the magnetic lift-off process.
Step 8: Magnetic deposition and lift-off

According to the study in Chapter 3, the critical thickness of Al₂O₃ is 6 nm, above which the anti-ferromagnetic coupling disappears. A 4 µm thick multi-layer was deposited with PVD, which is almost the limit of the PVD deposition. The film starts to peel off above this thickness due to the high induced stress. The composition of the core is Cr (20nm) / [FeCoB (500nm)+Al₂O₃ (6nm)] × 8. The lift-off was done with acetone and ultrasound applied. This step and the zoom-in pictures on the wafer are shown in Fig. 4.25.

Fig. 4.25 Schematic view of the lift-off process (upper), and pictures of the lift-off patterned cores (lower).
Step 9: PI3 coating and patterning

A 10 µm layer of polyimide HD 4100 was coated and pattern above the magnetic layer. Thus, a dielectric layer was formed underneath the top metal layer, while leaving the vias openings, as shown in Fig. 4.26.

![Polyimide layer-3](image)

Fig. 4.26 Polyimide layer-3 was spin coated above the magnetic layer, and patterned to have the via openings.

Step 10: Deposit top coil seedlayer

Similar with step 2, the seedlayer of top metal was deposited with PVD, as shown in Fig. 4.27(a). The only difference was that a low power re-sputtering step was taken after the regular Cu deposition. This process could make a continuous seedlayer formed on the side wall of the via openings. The N\textsubscript{2} plasma hit on the Cu atoms at the via bottom, which then get reflected and re-deposited on the side wall, as shown in Fig. 4.27(b).
Fig. 4.27 (a) A seedlayer (Cr/Cu) was deposited above PI3 layer. (b) The re-sputter process is able to re-distribute the seedlayer in the via openings during the deposition.

**Step 11: Photo top metal**

In this step two different masks were used to pattern the thick photoresis. Because the via openings are around 30μm deep and 30 μm to 50 μm wide, the high aspect ratio added to the difficulty of cleaning the photoresist inside the openings. As a result, one mask that only patterns the via openings were used for lithography before applying the top coil mask. Thus, photoresist inside the via openings was exposed for a longer time and pre-developed before the coil structure got exposed. This step is shown in Fig.4.28.

**Step 12: Plate top coil and etch top coil seedlayer**

This step is exactly the same with the electrical plating and seedlayer removing process of the bottom coil, as shown in Fig.4.29. Some pictures of the fabricated devices are shown in Fig.4.30.
Fig. 4.28 An 8 µm thick photoresist layer was coated and patterned above the seedlayer.

Fig. 4.29 After the top metal layer plated, the photoresist was stripped with acetone. The seedlayer was removed as discussed earlier.
Fig. 4.30 Picture of the on-chip inductors, Type I, type II and Type III respectively.
4.5 On Wafer Measuring

The testing frequency range of the inductors located between 300 kHz and 500 MHz. At the high frequency range, the signal wavelength is comparable with the dimension of the IC component. As a result, the current and voltage do not remain unchanged everywhere on the chip and they need to be treated as propagating microwaves. The characterization of the on-chip integrated inductors requires the microwave testing system.

4.5.1 Testing Platform

The testing system used in the work consists of a vector network analyzer, a probe station, and two RF GSG probes. The parasitic effects induced by the connection cables could be minimized with this on-chip testing method.

Vector network analyzer

The vector network analyzer performs measurements based on the reflection method. The measured scattering matrix (S-parameters) provides a complete description of the networks. It relates the voltage waves incident on the ports to those reflected from the ports. Once the scattering parameters of the network are known, conversion to other matrix parameters can be performed if necessary. The scattering matrix is defined as

\[
\begin{bmatrix} V_1^- \\ V_2^- \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} V_1^+ \\ V_2^+ \end{bmatrix}, \]

(4.26)
where \( V_n^- \) represents the amplitude of the voltage wave reflected from port \( n \) and \( V_n^+ \) represents the amplitude of the one incident on port \( n \). A specific element the scattering matrix can be determined as

\[
S_{ij} = \frac{V_i^-}{V_j^+} \bigg|_{V_k^+ = 0 \ for \ k \neq j}.
\]  

(4.27)

**Probes and probe stations**

The RF probes perform the physical contact to the device under test, as shown in Fig. 4.31(a). In order to control the position of the probe tips, they are mounted on to a probe station by screwed on to the arms of two micropositioners, as shown in Fig.4.31 (b).

![Fig. 4.31 (a) The GSG probes are mounted on the arms of two micropositioners which can be controlled by the (b) probe station. Meanwhile, they are connected to the two ports of the vector network analyzer through the low loss coaxial cables.](image)
The station enables a high accuracy alignment as well as a tight control of the pressure applied to the probes. Meanwhile, the GSG probes are connected to the two port of the network analyzer through the low loss coaxial cables, also shown in Fig. 4.13(a).

### 4.5.2 Calibration

The commercialized on-chip calibration substrate was used in order to calibrate the VNA directly at the probe tip interface. Different calibration sets could be used, such as SOLT (Short-Open-Load-Through), LRL (Line-Reflection-Line), and LRM (Line-Reflection-Match). The SOLT calibration method was applied in this work.

**SOLT calibration**

The SOLT calibration procedure goes through a few steps.

**Step 1: Select the Right Option on the VNA**

Select the two-port calibration option under the calibration menu of the VNA.

**Step 2: Measuring the Short Standard**

This step is shown on the left top of Fig. 4.32. After the probe tips touching the pad of the standard, simply click the “short” soft button on the screen, confirming the executing of the short calibration.

**Step 3~5: Measuring the Open and Load Standard**

Similar with step 2, these steps are done by contacting the probe tips to the standards and confirming it in the VNA.
Fig. 4.32 The SOLT calibration procedure includes four steps: Measuring the short, open, load and through standard.

After the calibration, the testing system is ready to be used for on-chip inductor testing.
4.5.3 Parameter Extraction

In order to extract the scattering parameters from the VNA, an equivalent circuit model needs to be set up. There are three possible setups for measuring an inductor with a VNA, as shown in Fig. 4.33.\(^{[4,20]}\)

![Image of equivalent circuit models](Image95x428to519x585.png)

Fig. 4.33 Three possible configurations for measuring a two-port device \(Z_L\) with a VNA. They are two-port measurement with the device placed in series with the ports, two-port measurement with the device placed in parallel with the ports, and one-port measurement, respectively from left to right.

**Π Model of Two-port Measurement with the Device Placed in Series**

The Π model shown in Fig. 4.34 is the electric circuit most often used to calculate the scattering parameters of a two-port measurement of integrated inductors for RF application.
Fig. 4.34 Setup for the calculation of the scattering parameters of a two-port measurement with the $\Pi$ model.

De-embedding Technique

Because the probe tips are not able to be built small enough to directly contact the on-chip device, the contact pads are usually designed to fit the probe tip dimensions, as shown in Fig. 4.14. In order to eliminate the parasitic impedance that brought by the testing pads, the de-embedding technique was developed.

First, the complex scattering parameters got with the device under testing (DUT) need to be converted into the admittance matrix $[Y]$, using the following equations:

$$Y_{11} = \frac{1}{Z_0} \frac{(1-S_{11})(1+S_{22})+S_{12}S_{21}}{(1+S_{11})(1+S_{22})-S_{12}S_{21}},$$  \hspace{1cm} (4.28)
where $Z_0$ is 50 Ohm. Second, measure the S parameters of the open structure with only the testing pads but not the DUT, as shown in the lower picture of Fig. 4.14. Thus, the accurate admittance matrix is

$$[Y_{\text{de-embed}}]=[Y_{\text{DUT}}]-[Y_{\text{PAD}}].$$

(4.32)

The de-embedded S-parameters could be calculated by

$$S_{11} = \frac{(Y_0-Y_{11})(Y_0+Y_{22})+Y_{12}Y_{21}}{(Y_0+Y_{11})(Y_0+Y_{22})-Y_{12}Y_{21}},$$

(4.33)

$$S_{12} = \frac{-2Y_{12}Y_0}{(Y_0+Y_{11})(Y_0+Y_{22})-Y_{12}Y_{21}},$$

(4.34)

$$S_{21} = \frac{-2Y_{21}Y_0}{(Y_0+Y_{11})(Y_0+Y_{22})-Y_{12}Y_{21}},$$

(4.35)

$$S_{22} = \frac{(Y_0+Y_{11})(Y_0-Y_{22})+Y_{12}Y_{21}}{(Y_0+Y_{11})(Y_0+Y_{22})-Y_{12}Y_{21}},$$

(4.36)

where $Y_0=1/Z_0$. 

$$Y_{12} = \frac{1}{Z_0} \frac{-S_{12}}{(1+S_{11})(1+S_{22})-S_{12}S_{21}},$$

(4.29)

$$Y_{21} = \frac{1}{Z_0} \frac{-S_{21}}{(1+S_{11})(1+S_{22})-S_{12}S_{21}},$$

(4.30)

$$Y_{22} = \frac{1}{Z_0} \frac{(1+S_{11})(1-S_{22})+S_{12}S_{21}}{(1+S_{11})(1+S_{22})-S_{12}S_{21}},$$

(4.31)
**Inductor Parameters**

The inductance, resistance and quality factor of the inductor could be calculated with the de-embedded admittance matrix, as the following

\[
L = \frac{1}{\omega} \cdot \text{Im} \left( \frac{-1}{Y_{12}} \right), \quad (4.37)
\]

\[
R_s = \text{Re} \left( \frac{-1}{Y_{12}} \right), \quad (4.38)
\]

\[
Q = -\frac{\text{Im}(Y_{11})}{\text{Re}(Y_{11})}, \quad (4.39)
\]

**4.6 Result and Analysis**

Measurement was done on 25 sets of on-chip integrated inductors. The dimensions are shown in Table 4.2. As the number of turns gets increased, the width of coil decreases in order to maintain a constant value of the gap. After the on-chip testing with the testing platform discussed earlier, the de-embedding was done and the measured inductance and quality factors were extracted from the admittance parameters. These results were plotted and going to be discussed in this section.
Table 4.2 The Design Matrix of the on-Chip Integrated Inductors

<table>
<thead>
<tr>
<th># of turns on single core</th>
<th>width of single magnetic core</th>
<th>length of single magnetic film</th>
<th>width of Cu wire</th>
<th>gap between each two Cu wire turns</th>
</tr>
</thead>
<tbody>
<tr>
<td>6</td>
<td>2000</td>
<td>2400</td>
<td>322</td>
<td>20</td>
</tr>
<tr>
<td>8</td>
<td>2000</td>
<td>2400</td>
<td>246</td>
<td>20</td>
</tr>
<tr>
<td>10</td>
<td>2000</td>
<td>2400</td>
<td>198</td>
<td>20</td>
</tr>
<tr>
<td>12</td>
<td>2000</td>
<td>2400</td>
<td>164</td>
<td>20</td>
</tr>
<tr>
<td>14</td>
<td>2000</td>
<td>2400</td>
<td>140</td>
<td>20</td>
</tr>
</tbody>
</table>

Type I (including air-core structure)

<table>
<thead>
<tr>
<th>6</th>
<th>1000</th>
<th>2400</th>
<th>322</th>
<th>20</th>
</tr>
</thead>
<tbody>
<tr>
<td>8</td>
<td>1000</td>
<td>2400</td>
<td>246</td>
<td>20</td>
</tr>
<tr>
<td>10</td>
<td>1000</td>
<td>2400</td>
<td>198</td>
<td>20</td>
</tr>
<tr>
<td>12</td>
<td>1000</td>
<td>2400</td>
<td>164</td>
<td>20</td>
</tr>
<tr>
<td>14</td>
<td>1000</td>
<td>2400</td>
<td>140</td>
<td>20</td>
</tr>
</tbody>
</table>

Type II&III (including air-core structure)
4.6.1 Number-of-Turn Dependency

Figure 4.35 ~ 4.37 show the frequency dependence of inductance and quality factor for three different designs, with five different numbers of turns, from 6 to 14. For inductors of Type I, as the number of turns $N$ increases, the inductance value was measured to range from 13 nH to 53 nH, around 35MHz. It can be seen that the inductance is proportional to $N^2$, as described in equation (4.5). The inductance of Type II increases from 14 nH to 57 nH as $N$ increases and that of Type III ranges from 15 nH to 60 nH. It could be concluded that the design of Type III generally provides the highest inductance, with the indistinct advantage.

Inductors with lower number of turns show higher quality factors, and it is related to the lower resistance loss and lower winding capacitance. The measured peak quality factor of Type I is around 8 for 6 turns of winding; the peak quality factor of Type II is 6.7 and that of Type III is 6.3. At low frequencies, the quality factor of each type is very close for different $N$ values, which indicates that the $N$ dependency of the quality factor is not very obvious at frequency range much lower than the peak region. As frequency goes up, the quality factor curves start to spread, indicating the high frequency loss of the inductors, resulting from the core and parasitic effects.

Certainly, by magnetically annealing the cores at high temperature, the magnetic property could be dramatically improved, and so does the inductance and quality factor.
Fig. 4.35 The measured inductance and quality factor of inductor type I, for different numbers of turns.
Fig. 4.36 The measured inductance and quality factor of inductor type II, for different numbers of turns.
Fig. 4.37 The measured inductance and quality factor of inductor type III, for different numbers of turns.
4.6.2 Core Structure Dependency

For a solenoid type inductor with the magnetic core of 2mm ×2.4mm in area, the inductance is close when the number of turns is the same for the three types of cores, shown in Fig. 4.38. Type III does not show a very obvious advantage on it. The only difference between Type II and Type II is the connection between the two-sided magnetic bars. Theoretically, Type III with a closed core loop tends to keep the magnetic flux inside the magnetic material, generating less leakage. However, the magnetic material was not annealed before performing this measurement. The magnetic domains inside the magnetic material are very likely with disordered magnetization directions. Compared with the rectangular cores, the triangle shape magnetic pieces connecting the two-sided bars do not have a high permeability, due to the lower shape anisotropy. Hence, they do little effect on keeping a condensed magnetic loop. This lousy magnetic property could be improved by doing the magnetic high temperature annealing of the devices. However, the above discussion about the shape anisotropy does explain the fact that the Type I inductor shows lower inductance than Type II and III. The rectangular core of Type I has a lower length to width ratio (2.4/2) than that of Type II and III (2.4/1), which means a smaller shape anisotropy, thus a lower permeability along the axial direction. As a result, for a fixed number of turns, inductor Type I shows lower inductance than the other two types.

The comparison of the inductance between the magnetic core and air core devices can be seen very clearly from each plot of Fig. 4.38. As the number of turns increases, the inductance of Type I is increased with a magnetic core by a factor of 3 ~ 6.75, compared
with that of an air core structure. The inductance of Type II and III is increased by a factor of $3.75 \sim 6.35$. 

(a) 

(b)
Fig. 4.38 The inductance plots for different core types, with the same numbers of turns. (a) N=6; (b) N=8; (c) N=10; (d) N=12; (e) N=14.

The effect of the magnetic core on the quality factor could be seen from Fig. 4.39. Inductor Type I with a magnetic core of shows higher quality factor for cases of N=6, 8, 10, 12, respectively, but the advantage gets eliminated as the number of turns increases. It even gets lower than that of Type III with a magnetic core when N equals 14. Hence, the magnetic core in Type I design tends to have higher loss for large number of turns. It could be related to the eddy current that is generated inside the core due to the alternative magnetic field. For a certain type of magnetic material, a larger cross section usually shows higher eddy current induced loss, as discussed in Chapter 3.
The loss generated purely from the winding coil and the substrate could be seen from the quality factor plots of the air core inductors. The inductors of Type I design have a much higher quality factor over the other two types without the magnetic core. Although the total area, the wire width and gap are the same for the same number of turns, the two-sided solenoid structure for Type II and III has a double winding turns, leading to higher electrical loss.
Fig. 4.39 The quality factor plots for different core types, with the same numbers of turns.

(a) N=6; (b) N=8; (c) N=10 (d) N=12; (e) N=14.
4.6.3 Conclusion

For a solenoid type inductor, the performance could be optimized by improving the core structure and adding winding turns. Add more winding turns is the most straightforward way of increasing the inductance. However, the thinner conductor line dimension would probably add to the difficulty of fabrication process. By dividing the magnetic core into halves and winding the coils on two sides and in series, the inductance could be increased. Joining the separated halves at two ends is able to improve the inductance further due to a closed magnetic loop inside the core material.

The influence on the quality factor was also discussed. The loss comes from the magnetic core, the inductor winding and the substrate. For large number of turns, the two-sided magnetic core shows lower loss than the single piece design, due to the reduced eddy current loss.

Moreover, a magnetic annealing is very likely able to improve the magnetic properties of the magnetic core, hence, reducing the overall loss.
REFERENCES


4.4 http://holst.stanford.edu/~CPYue


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5 Soft Magnetic Material Applied

Vibration Energy Harvesting

5.1 Introduction

Recent years, the application of mobile electronics and wireless sensors have been developing rapidly in several fields, including environmental, industrial and medical. The traditional power supply system, including batteries and wired sources, are suffering from several drawbacks. Batteries need frequent recharging or replacement, which is costly and impossible especially for the wireless networks with thousands of physically embedded nodes [5.1]. And the power supply cable seriously limits the flexibility. As a result, self-renewable power supply is becoming a critical issue in the applications of mobile electronics and wireless sensors.

Energy harvesting, sometimes defined as “power scavenging” or “energy extraction”, is the way to convert the ambient to usable electrical energy. The critical part might be smart materials or structures. And the collected energy is generally to drive the wireless electric devices or charge the embedded batteries.
5.1.1 Available Energy Sources

There are different types of available ambient energy, such as solar, wind, vibration, thermal and RF wave, etc. Table 5.1 shows most of the available ambient energy sources and their performance.\textsuperscript{[5.2]}

### Table 5.1 Comparison of different types of available ambient energy sources and their performance

<table>
<thead>
<tr>
<th>Source</th>
<th>Power density(^\text{1-year lifetime}) (μW/cm(^2))</th>
<th>Power density(^\text{10-year lifetime}) (μW/cm(^2))</th>
<th>Source of information</th>
</tr>
</thead>
<tbody>
<tr>
<td>Scavenged power sources</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Solar (outdoors)</td>
<td>15,000—direct sun</td>
<td>15,000—direct sun</td>
<td>Commonly available</td>
</tr>
<tr>
<td></td>
<td>150—cloudy day</td>
<td>150—cloudy day</td>
<td></td>
</tr>
<tr>
<td>Acoustic noise</td>
<td>0.003 at 75 dB</td>
<td>0.003 at 75 dB</td>
<td>Theory</td>
</tr>
<tr>
<td></td>
<td>0.96 at 100 dB</td>
<td>0.96 at 100 dB</td>
<td></td>
</tr>
<tr>
<td>Daily temp. variation</td>
<td>10</td>
<td>10</td>
<td>Theory</td>
</tr>
<tr>
<td>Temperature gradient</td>
<td>15 at 10°C gradient</td>
<td>15 at 10°C gradient</td>
<td>Roundy (2003)</td>
</tr>
<tr>
<td>Shoe inserts</td>
<td>330</td>
<td>330</td>
<td>Sturier (1996)</td>
</tr>
</tbody>
</table>

### Solar Energy

Solar energy is the most well known and widely applied source, since the solar cells are able to generate excellent power density with high efficiency. During peak hours, the power density of solar radiation on the Earth’s surface is approximately 100mW/cm\(^2\).\textsuperscript{[5.3]} And the silicon solar cells are a relatively mature technology with efficiencies ranging from 12% to 25%. However, the advantage could be easily limited when no sunlight present because the device is deeply embedded or when the panel is obscured by rain, dirt or snow.
Vibration Energy

Vibration energy harvesting has great potential simply because the vibration energy source is easily accessible. The available vibrating source is almost everywhere from vibrating equipment, machinery and structures like bridge, working microwave, running automobile, aircraft wing, etc. Hence, the vibration energy might be the most available energy source for the wireless sensors compared with the other choices. The current technology is focusing on conversion of the vibration energy from the vibrating surface into electric energy. And the output power could be as high as 63 mW from a device with a total volume of 75cm$^3$. Vibration energy harvesting is not only able to provide high output electric power with high efficiency, but also reduce the unnecessary noise or mechanical vibration which might potentially cause structure damage.

Thermal Energy

Power generation based on thermoelectric effects has been known since the Seebeck effect was discovered in the 1800s.$^{[5,4]}$ The Seebeck effect is defined as, a temperature difference across a conductor generates an electrical potential. It converts thermal energy into electrical energy. Figure 5.1 illuminates the power generation mode of thermoelectric material$^{[5,5]}$. 
Generally, a number of thermoelectric elements can be connected electrically in parallel and/or in series shaping a thermoelectric generator (TEG) device.\textsuperscript{[5,6]} Identifying materials with a high thermoelectric figure of merit \(Z(=S^2\sigma/k)\) has been proven an extremely challenging task. The solid state energy calls for materials with high electrical conductivity \(\sigma\), high Seebeck coefficient \(S\) and low thermal conductivity \(k\).\textsuperscript{[5,7]} Development of new concepts and theories about electron and phonon transport speeded up the research on thermoelectric material after 30 years of slow progress. The \(Z\) value of \(\mathrm{AgPb}_m\mathrm{SbTe}_{2+m}\) is over 1.5 around 700K. Figure 5.2 shows the figure of merit \(ZT\) as a function of temperature for several bulk thermoelectric materials\textsuperscript{[5,5]}.
Fig. 5.2 Figure of merit $ZT$ as a function of temperature for several bulk thermoelectric materials.

RF Energy

Radio frequency (RF) radiation has been applied in the information exchange of ID and credit cards. It seems like only very small amount of electromagnetic radiation energy was involved in this process. However, most of the electrical equipments, including the cell phones, GPS, broadcast radio, TV and microwave ovens, are generating the RF energy everywhere all the time, which makes the RF energy a very ample and accessible source. The problem is collecting all these disparate sources and converting them in useful energy. The conversion is based on a rectifying antenna (rectenna), constructed with a Schottky diode located between the antenna dipoles.\textsuperscript{[5,8]} It was
reported that 2.5mW power from X-band source horn antenna with 1W 10GHz radiation to a receiver horn antenna for driving piezo-sensor nodes over a distance of 0.61m.\textsuperscript{[5,9]} However, the energy provided by the RF energy generator is still very low and with a low efficiency.
### 5.1.2 Vibration Energy Harvesting Mechanisms

Table 5.2 Comparison of several key figures of merit for different vibrating energy harvesting mechanisms

<table>
<thead>
<tr>
<th>Mechanisms/Products</th>
<th>$f_{center}$ (Hz)</th>
<th>$a$ (g)</th>
<th>$P_{max}$ (mW)</th>
<th>$R_{output}$ ($\Omega$)</th>
<th>$V_{total}$ (cm$^3$)</th>
<th>Power Density (mW/cm$^3$)</th>
<th>HPBW* (Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Magnetoelectric micro-cantilever$^{[5.10]}$</td>
<td>52</td>
<td>0.06</td>
<td>0.046</td>
<td>4k</td>
<td>0.15</td>
<td>0.31</td>
<td>0.5</td>
</tr>
<tr>
<td>Electrostatic$^{[5.11]}$</td>
<td>50</td>
<td>0.91</td>
<td>1.052</td>
<td>-</td>
<td>1.8</td>
<td>0.58</td>
<td>-</td>
</tr>
<tr>
<td>Magnetoelectric composite beam$^{[5.12]}$</td>
<td>40</td>
<td>1</td>
<td>-</td>
<td>3M</td>
<td>-</td>
<td>0.4</td>
<td>-</td>
</tr>
<tr>
<td>Volture Piezo Energy Harvester-V25W$^{[5.13]}$</td>
<td>30</td>
<td>1.1</td>
<td>6.5</td>
<td>-</td>
<td>0.98</td>
<td>6.63</td>
<td>-</td>
</tr>
<tr>
<td>Magnetostrictive cantilever$^{[5.15]}$</td>
<td>58.1</td>
<td>1</td>
<td>0.2</td>
<td>~80</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Perpetuum PMG37$^{[5.16]}$</td>
<td>21.9</td>
<td>1</td>
<td>92</td>
<td>-</td>
<td>130.67</td>
<td>0.7</td>
<td>12.9</td>
</tr>
<tr>
<td>KCF technologies VPH300$^{[5.17]}$</td>
<td>360</td>
<td>0.239</td>
<td>4.1</td>
<td>-</td>
<td>208.86</td>
<td>0.0196</td>
<td>5</td>
</tr>
<tr>
<td>This work: High-µ beam with bias fields</td>
<td>54</td>
<td>0.57</td>
<td>74</td>
<td>1</td>
<td>68.96</td>
<td>1.07</td>
<td>10</td>
</tr>
</tbody>
</table>

*Power density, in unit of mW/g cm$^3$. 

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There are three main transduction mechanisms employed to harvest vibration energy. They are electromagnetic (inductive), piezoelectric and electrostatic (capacitive).\cite{5.18} Table 2 shows the comparison of several key figures of merit for different vibrating energy harvesting mechanisms.

**Piezoelectric generators**

The piezoelectric effect was discovered in 1880. If certain crystals were subjected to mechanical strain, they became electrically polarized and the degree of polarization was proportional to the applied strain. Conversely, these materials deform when exposed to an electric field.\cite{5.18} Moreover, this effect is anisotropic. The piezoelectric generators converts the vibration induced deforming in the piezoelectric material into electrical energy.

The first US patent on energy harvesting was issued to a piezoelectric generator whose output was rectified and stored into a capacitor.\cite{5.19} Studies have shown that an average gait walking human of weight 68kg, produces 67 W of energy at the heel of the shoe\cite{5.20}. A shoe heel mounted piezoelectric energy harvester was claimed to generate an average power of 8mW.\cite{5.21} The cantilever model based structure is used to harvest the vibration energy at a particular frequency by forming a resonator, since the resonator absorbs the maximum energy from the source when the system reaches sympathetic vibration. An electromagnetic micro-cantilever device with a volume of 0.15cm$^3$ was reported to generate a maximum power 46µW at 52Hz and an acceleration of 0.59m/s$^2$.\cite{5.22} The piezoelectric mechanism is well combined with MEMS technology.
One MEMS power generator with transverse mode thin piezoelectric film is proved to provide\textsuperscript{[5.23]} A micro-generator was demonstrated to\textsuperscript{[5.24]} A piezoelectric cantilever harvesters were reported to achieve a maximum output power of 790µW with a tip mass weighing 10g at the acceleration 9m/s\textsuperscript{2} and frequency 72Hz, without a coil or magnet. \textsuperscript{[5.25]}

**Electrostatic generators**

A capacitor is initially charged with a voltage source V, creating equal but opposite charge Q on the plates, so that it is charged with a certain amount of energy \( E=0.5QV=0.5CV^2=0.5Q^2/C \). When the source is disconnected, the energy remains. If the two plates have moved relatively to each other, the capacitance is changed, simply because \( C=\varepsilon A/d \), where \( \varepsilon \) is permittivity of the medium in between the plates and d is the distance. As a result, the relative movement of the plates results in the change in the stored energy. The electrostatic generators could be easily classified into three types based on the relative movement of the plates: \textsuperscript{[5.26]} In-plane overlap varying, shown in Fig. 5.3, in-plane gap closing, shown in Fig. 5.4, and out-of plane gap closing, shown in Fig. 5.5 \textsuperscript{[5.27]}. It was found that in-plane gap closing mode generates the highest power with an optimized design producing 100µW/cm\textsuperscript{3} and the in-plane overlap varying mode generates the lowest power. The MEMS technology is well integrated in the capacitive generator. Optimized electrostatic energy harvesting devices showed an output power 1mW under vibration amplitude of 90µm at 50Hz. \textsuperscript{[5.11]}

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Fig. 5.3 In-plane overlap varying.

Fig. 5.4 In-plane gap closing.

Fig. 5.5 Out-of-plane gap closing.
Electromagnetic generators

The Faraday effects tell us that the electric current can be induced if the magnetic flux is changing inside the conducting loop. In the real case, a conducting coil is usually used, and the relative movement between the coil and the permanent magnet provides the flux change. The open circuit voltage depends on the change rate of the flux and the number of the loops in the coil. A micro-electromagnetic energy harvester was proved to generate an output power of 400 $\mu$W, shown in Fig. 5.6\textsuperscript{5,28}. Another electromagnetic generator based on four moving magnets and a fixed coil demonstrated a peak output power of 3.9 mW and an average power of 157 $\mu$W.\textsuperscript{5,29} An electromagnetic micro-

Fig. 5.6 Electromagnetic energy harvesting mechanism.
cantilever device with a volume of 0.15 cm$^3$ was reported to generate a maximum power 46 $\mu$W at 52Hz and an acceleration of 0.59 m/s$^2$.[5,10] A planar coil based structure was investigated, with a polyimide film as the spring, a NdFeB magnet. The coil was made from a 1.5 $\mu$m thick aluminum layer.[5,30]

5.1.3 Key Materials of Vibration Energy Harvesting Devices

Most of the vibration energy harvesters have applied smart materials, except for those based on the relative movement between the magnets and the coil. Generally, smart materials employed by vibration energy harvesters insist three types: piezoelectric, magnetostrictive and high permeability material. The first two are traditional and widely applied. They could not only work independently as the key material in the device, but also combined to make a double layer or sandwich structure. The last was proved lately to provide a high magnetic flux change rate with a fixed magnet pair, so as to generate very high output power and provide very broad working bandwidth. Table 5.3 shows a comparison between the energy harvesting behaviors of these materials.
Table 5.3 Comparison between different materials applied in vibration energy harvesting system

<table>
<thead>
<tr>
<th>Type</th>
<th>Advantages</th>
<th>Disadvantages</th>
</tr>
</thead>
<tbody>
<tr>
<td>Piezoelectric</td>
<td>Direct conversion between mechanical and electrical energy, combine with MEMS technology, small size</td>
<td>Brittle, eddy current, suffer from degraded polarization after a certain time of service, narrow bandwidth</td>
</tr>
<tr>
<td>Magnetostrictive</td>
<td>Mechanically reliable</td>
<td>Bias magnetic field and coil required, low efficient, narrow bandwidth, narrow bandwidth</td>
</tr>
<tr>
<td>Piezoelectric + magnetostrictive</td>
<td>Able to harvest both vibration and AC magnetic energy, strong coupling, high output</td>
<td>Two phases part from the adhesive after a certain time of deforming, large bias magnetic field generally required, narrow bandwidth</td>
</tr>
<tr>
<td>High Permeability</td>
<td>Mechanically reliable, broadband, very high output</td>
<td>Magnets and coil required</td>
</tr>
</tbody>
</table>
**Piezoelectric Material**

Piezoelectric material has advantages like easy operation and direct conversion between mechanical energy and electric energy. As a result it is the most widely used material in the vibration energy harvesting application. And it is widely available in many forms including single crystal (e. g. quarts), piezoceramic (e. g. lead zirconate titanate or PZT), thin film (e. g. sputtered zinc oxide), screen printable thick-films based upon piezoceramic powders and polymeric materials such as polyvinylidenefluoride (PVDF).

\[ d = \frac{\text{strain}}{\text{electric field}} \text{m/V} \text{, or } d = \frac{\text{short circuit charge density}}{\text{stress}} \text{C/N}. \]

The piezoelectric strain constant, \( d \), is defined as \( d = \frac{\text{strain}}{\text{electric field}} \text{m/V} \), or \( d = \frac{\text{short circuit charge density}}{\text{stress}} \text{C/N} \). The piezoelectric effect has two working modes: transverse mode \( d_{31} \) and longitudinal mode \( d_{33} \). A power density of 70 \( \mu \text{W/cm}^3 \) has been demonstrated with the PZT bimorph.\(^{[5,31]}\)

**Magnetostrictive Material**

Magnetostriction is defined as the change of the physical dimension of a material in response to the bias magnetic field or its magnetization change. The magnetostriction coefficient \( \lambda \) is to describe this effect and it saturates with \( \lambda_s \). Typical magnetostrictive materials applied in vibration energy harvesting systems are Terfenol-D (Tb0.3Dy0.7Fe1.9-2), crystalline alloy (ETREMA Products) and Metglas 2605 (Fe81B13.5Si3.5C2) metallic amorphous. Table 5.4 shows the magnetostriction constants of different soft magnetic materials.
Conversely, inverse-magnetostriction also known as Villari Effect is defined as the change in permeability subject to the mechanical stress on the material. This effect is often applied in some vibration energy harvesters only with the existence of a bias DC magnetic field to make sure the net magnetization in the material in one direction. The Teffnol bar based harvester is able to generate an induced voltage of 150 V at 1 kHz, with 0.25 Ω resistances and a bias field of 200 Oe. A cantilever of adhesive bonded six layers of magnetostrictive Metglas 2605SC ribbon in a solenoid was claimed to generate a maximum output power of 900µW at 1g.

**Composite of Piezoelectric and Magnetostrictive Material**

The combination of piezoelectric and magnetostrictive material basically lay off the picking coil for the magnetostrictive phase, since the mechanical deform transfers from
An electromagnetic micro-cantilever device with a volume of 0.15cm$^3$ was reported to generate a maximum power 46µW at 52Hz and an acceleration of 0.59m/s$^2$. A sandwich structure consisting of two piezoelectric layers and one magnetostrictive layer, or named as magnetoelastic (ME) sensor structure, was able to generate the maximum power 10.8mW at 60Hz and 0.1g acceleration with a total device volume 113cm$^3$, due to the vibration induced magnetization change in the magnetostrictive layer. A magnetoelectric energy harvester with a beam consisting of both magnetostrictive and piezoelectric phases was demonstrated to generate an output voltage of 4V under a vibration acceleration of 0.05g at 20Hz and an AC magnetic field amplitude of 20e.

**High permeability material**

High permeability material also known as soft magnetic material is easily magnetized by a bias magnetic field. The magnetic flux density could be fairly high compared with the other magnetic materials, when external magnetic field is applied. The electromagnetic harvesting mechanism basically requires high flux change inside the solenoid according to Faraday’s law, described as

\[ V_{induced} = N \frac{d\Phi}{dt} = N \cdot \mu \frac{dH}{dt}, \quad (5.1) \]

where $N$ is the number of coil turns, $\Phi$ is the magnetic flux of each turn, $t$ is time, $\mu$ is the effective permeability of the core. High permeability material applied as the core can dramatically increase the induced voltage.

Some commercially available high permeability materials are Metglas 2605 S3A with $\mu_r=35000$ and saturation induction 1.41T, Metglas 2714A with $\mu_r=1,000,000$ after
annealing and saturation induction 0.57T and MuShield with a maximum permeability of 300,000 and a saturation magnetization of 7500 Gauss.

5.2 Wideband Vibration Energy Harvester with High Permeability Material

5.2.1 Introduction

The conventional vibration energy harvesters are designed as linear resonant structures with narrow operating bandwidths, making them impractical when applied in a real world environment which provides vibration sources with a wide frequency spectrum. As a result, efforts have been made to explore non-linear mechanism that leads to wide working bandwidth of vibration energy harvesters.

In this section, a schematic model of a novel wideband vibration energy harvester would be discussed. The employing of the high permeability material generates a strong magnetic coupling, leading to non-linear effect.
5.2.2 Basic Mechanism

Fig. 5.7 The section view of the schematic design of the vibration energy harvesting device. Dimension of each part is: 4.4cm ×3.2cm×4cm for the solenoid, 1.25cm×2.2cm×1.5cm for the magnet pair including the gap in between, 1.3cm×1.5cm×2.5cm for the mounting frame on one side and 0.5cm×1.5cm×0.6cm on the other.

In this work, an alternative vibration energy harvesting model is set up and experimentally verified, which is based on the strong magnetostatic coupling between the cantilever beam and a bias magnet pair. The inhomogeneous bias magnetic field enables the highly permeable beam to experience complete magnetic flux reversal twice in one vibration period, which leads to maximized magnetic flux change rate in the solenoid. At the same time, the magnetic potential energy makes the cantilever vibrate in a wider potential well than in the simple harmonic case, which allows a wide working bandwidth of the harvester. Schematic design of the vibration energy harvester is shown in Fig. 5.7.
The key component of this energy harvester is a high permeability (high-$\mu$) single layer beam, with one end fixed and the other end vibrating inside a solenoid. Two identical rectangular hard magnets are placed in close proximity to the free end of the high-$\mu$ beam outside the solenoid to induce a non-homogeneous bias field which acts to induce magnetization reversal. The bias magnets are aligned in parallel with each other, and parallel to the beam. As shown in Fig. 5.8, magnets form a closed flux with the magnetizations antiparallel to each other. When the free end of the cantilever passes through the spatially inhomogeneous fringing magnetic field generated by the bias magnet pair, the magnetization of the magnetic cantilever would be reversed by the

---

**Fig. 5.8** Magnet pair with antiparallel magnetic moment provides closed magnetic field lines, making sure the maximum magnetic flux change, from $\Phi$ to $-\Phi$ during the vibration.
antiparallel bias magnetic field by 180°. The magnetization reversal in the magnetic

cantilever leads to maximized flux change in the solenoid, resulting in an induced voltage
which varies at the same frequency of the mechanical vibrating source. However, when
the magnetization of the two bias magnets in parallel, as shown in Fig. 5.9, the repulsing
magnetic field would keep the beam magnetized towards the same direction all the time.
As a result, the magnetic flux change will be smaller in this case compared to that in the
antiparallel case in Fig. 5.8, leading to a lower output voltage and an induced AC voltage
with a frequency that is double of that of the mechanical vibration frequency.

Fig. 5.9 Magnet pair with parallel magnetic moment provides repelling magnetic field lines,
in which case the magnetic flux changes from $\Phi$ to 0 and back to $\Phi$ during the vibration.
5.2.3 Prototype and Testing System

The harvester, including the MµShield® cantilever, the solenoid and a SmCo magnet pair, is seated on a vibrating stage, as shown in Fig. 5.10. The stage is driven by an audio power amplifier connecting to a lock-in amplifier. The mechanical movement of the stage is monitored by an accelerometer. Voltage output of the harvester in time domain is monitored by a digital oscilloscope. Total volume of the energy harvester is 68.96cm$^3$, including the solenoid with the beam inside, the magnet pair with the gap in between and the mounting structure. The coil resistance of the solenoid is 1 Ohm, the

Fig. 5.10. Prototype of the wideband high permeability vibration energy harvester.
inductive impedance is 2.86 Ohm at 54 Hz, which is obtained by measurement. A lead tip mass is attached on the free end to adjust the vibration amplitude and the intrinsic frequency, which weighs 0.5 g.

5.2.4 Theoretical Model

The total induced voltage across the coil could be predicted by doing integral along the solenoid, shown in 5.11, because the magnetic field magnitude varies along the beam. According to Faraday’s law the open circuit voltage can be expressed by:

$$V_{\text{open}} = \frac{d\phi(t)}{dt} = \frac{d}{dt} \int \mu_0 \left[ H[x, y(x)] + M[x, y(x, t)] \right] \cdot A \cdot dN = \frac{d}{dt} \int \mu_0 M[x, y(x, t)] \cdot A \cdot dN,$$

(5.2)

![Fig. 5.11. The Schematic of the calculation of induced voltage across the solenoid.](image)
where $M$ is the magnetization in the beam, $A$ is the cross section area of the beam and $dN$ the number of loops in the infinitesimal length unit of the solenoid $dx$, shown in Fig. 5.11, and

$$dN = \frac{N_L}{d_w} dx. \quad (5.3)$$

$N_L$ is the number of coil loop layers in the solenoid and $d_w$ is the copper wire diameter of the coil. The dimension of the beam is $4.6\text{cm} \times 0.8\text{cm} \times 0.0254\text{cm}$, with a length: width: thickness ratio 181:31.5:1, which makes sure that the length direction is the magnetic easy axis. It has a maximum permeability of 300,000 and a saturation magnetization of 7500 Gauss. The magnetic hysteresis loop $M(H)$ along the length direction of the beam

![Fig. 5.12 Hysteresis loop of the MuShield beam with the dimension of 4.6cm × 0.8cm × 0.0254cm.](image)
was measured, shown in Fig. 5.12, in order to calculate the magnetization $M[x,y]$ at an arbitrary point on the beam in the non-uniform magnetic field. Vibration amplitude of an arbitrary point on the beam $y(x,t)$ is got by combining solutions of Euler–Bernoulli beam equation $y(x)$ and the equation of motion for the cantilever $y(t)$. According to the Euler-Bernoulli equation and boundary conditions, the maximum deflection of the cantilever is

$$y(x) = \frac{3Lx^2-x^3}{2L^3},$$

[5.34], as shown in Fig. 5.13. By combing the time dependent term $a(t)$, the beam shape function at time $t$ is

$$y(x,t) = a(t)\left(\frac{3Lx^2-x^3}{2L^3}\right),$$

(5.4)
where $L$ is the length of the beam and $a(t)$ is the amplitude at free end, which is determined by the following equation of motion,

$$m_{\text{eff}} \frac{d^2 a(t)}{dt^2} = - \frac{du}{da} - b\dot{a}(t) + F_{\text{drive}}, \quad (5.5)$$

$m_{\text{eff}}$ is the effective mass of the beam plus tip mass at the free end, which is $0.74\text{g}$. The first term on the right hand side $- \frac{du}{da}$ is the total force due to the total potential energy,

$$\frac{du}{da} = \frac{d(U_{\text{magnetic}} + U_{\text{elastic}})}{da}, \quad (5.6)$$

where the elastic potential energy $U_{\text{elastic}} = \frac{k}{2}a^2$, with $k$ the elasticity coefficient which is experimentally determined to be $77\text{N/m}$, and $U_{\text{magnetic}}$ the magnetic potential due to the bias magnet pair

$$U_{\text{magnetic}} = \int_0^L -B[x, y(t)] \cdot d\vec{m}, \quad (5.7)$$

in which $\vec{m}$ is the magnetic moment in the beam. Size of the identical SmCo hard magnet is $2.2\text{cm} \times 1.3\text{cm} \times 0.2\text{cm}$, providing a fringing magnetic field $\sim 500\text{Oe}$ at the free end of cantilever and $\sim 10\text{Oe}$ in the middle of the solenoid. The distribution of magnetic field is obtained with discrete spatial magnetic field measurements and fitted with simulation.

The second term $-b\dot{a}$ in equation (5.5) is the mechanical damping term, with $b$ the damping constant which is experimentally determined to be $0.0024\text{Ns/m}$. The third term $F_{\text{drive}}$ stands for the vibration driving force applied on the fixed end, which is a sine wave input.

When a load resistance is connected across the coil, the vector voltage across each element has the relationship $\dot{V}_{\text{open}} = \dot{V}_{\text{load}} + \dot{V}_{\text{coil res}} + \dot{V}_{\text{coil ind}}$. The solenoid is
equivalent to a resistance in series with an inductance. The output power is optimized when the impedance is matched, 
\[ X_L = 2\pi fL = X_C = \frac{1}{2\pi fC} \] 
and \( R_{load} = R_{coil} \), which could be done by inserting a capacitance and adjusting the load resistance. In this way, the maximum output power is

\[
P_{max} = \left( \frac{V_{open}}{2} \right)^2 \frac{1}{R_{load}} = \frac{1}{4R_{load}} \left( A\mu_0 N_b \frac{dM}{dx} \right)^2 \left( \int \frac{dM}{dt} \frac{x}{(x,y,t)} \right) dx^2. \tag{5.8}
\]

Equation (5.8) indicates that at a particular frequency, the output power depends on the change rate of magnetization in the beam.

### 5.2.5 Results and Discussion

Figure 5.14 shows calculated and measured results of the open circuit voltage in two cases. When the magnet pair is set to have antiparallel magnetization, the energy harvester shows a high open circuit voltage with a peak value 544mV at a vibration frequency of 54Hz and acceleration of 0.57g. As expected, the output voltage with a doubled frequency 108Hz with a significantly lowered peak value of 8mV was observed for the case when the two bias magnets are arranged with parallel magnetization. It is interesting to note that the mechanical vibration source is a sine wave signal, while the output voltage is not, but with narrow peaks with a full width at half maximum of 1ms. This is related to the nonuniform magnetic field spatial distribution, leading to the approximate square wave time varying magnetic flux, shown in Fig. 5.15. The free end amplitude dependent flux is also plotted. It is clear that the flux in the beam is reversed immediately while passing the nonstable equilibrium position in the middle. The sharp
Fig. 5.14 Measured and calculated results of the open circuit voltage for the energy harvesting device at the mechanical vibration frequency 54 Hz, acceleration 0.57 g (g=9.8 m/s²).

drop and rise in flux is the reason for large induced voltage.

Figure 5.16 shows the frequency response of the harvester. The maximum measured output power is 74mW on a 1Ω load and with a time average value 5mW at an acceleration 0.57g corresponding to a maximum power density of 1.07 mW/cm³ or 1.88 mW/g cm³.
Fig. 5.15 Normalized magnetic flux as a function of time and free end amplitude, at vibration frequency of 54 Hz, acceleration 0.57 g.

Fig. 5.16 Measured and calculated frequency response of the energy harvester.
5.2.6 Nonlinear Effect

From Fig. 5.16, we can see that the working bandwidth is about 10Hz, 18.5% of the central frequency, compared with 2.1Hz\textsuperscript{[5.14]} or 3.5% of the central frequency, for a typical linear oscillator harvester. The major reason for the large bandwidth is that the magnetic coupling is not linear to the displacement of the oscillator, so that the nonlinear effect provides the system a wider working bandwidth. \textsuperscript{[5.35]} As shown in Fig. 5.17,

Fig. 5.17 Elastic potential energy, magnetic potential energy and total potential energy of the oscillation system as functions of free end displacement of the beam.

compared with the elastic potential energy, the magnetic potential energy curve has double potential wells closely connected, resulting in a wider well. As long as the
acceleration is larger than its threshold value, or when the cantilever is supplied with enough energy to get over the barrier in the middle, it can vibrate between two shallow potential wells, realizing a relatively wide oscillation region at a particular driving frequency. However, if the oscillator is not supplied with enough energy or the damping is so strong that it is not able to get over the potential barrier, the oscillation is limited in one well, instead of two. In this case, the behavior of the oscillator is more like that of a linear one, with narrow working bandwidth.

Both the calculated and measured curves in Fig. 5.16 exhibit unsymmetrical peaks about the central frequency. This is because the performance of the oscillator at lower band differs from that at higher band. At lower band, the oscillation behavior is dominated by nonlinear effect. The output power decreases slowly as the frequency reduced, simply due to the mismatch between the intrinsic and driving source frequency. Even if at a frequency as low as 30 Hz, oscillation between two potential wells was still observed. However, in higher frequency range than 54 Hz, the performance is dominated by linear effect and oscillator was observed trapped in one well, leading to one sided narrow peaks in the voltage signal. This is because in higher frequency range, larger dynamic speed results in larger damping force, so that the cantilever does not have enough energy to climb over the potential barrier in the middle. This single well oscillation dramatically decreases the harvesting efficiency because the cantilever beam can hardly reach flux reversal.
5.2.7 Advantage of High Permeability Vibration Energy Harvesting

Table 5.2 shows the figures of merit for vibrating energy harvesters with different types of working mechanisms and materials, including magnetoelectric, electrostatic, piezoelectric, magnetoelectric sensor based, magnetostrictive and high permeability material based devices. Among all these different mechanisms and products, this high-µ energy harvester, which is based on the magnetic interaction between the vibrating high-µ beam and the bias hard magnets, generates relatively high power density and wide working bandwidth. Besides, the single layer metallic high-µ beam does have advantages from the material point of view. First of all, it is mechanically more robust compared with most of the piezoelectric materials or glue bonded multilayer materials. Secondly, the single layer metallic high-µ beam does not have problems of degraded performance after a certain working time.

5.2.8 Summary

In summary, an energy harvesting platform was theoretically studied and tested, which incorporated a vibrating high permeability cantilever, a solenoid and a bias magnet pair. Interaction between the high permeability magnetic beam and the bias magnets leads to complete flux reversal of the high permeability beam, which generates a maximum power of 74 mW and a high power density of 1.07 mW/cm³ at 54 Hz at an acceleration of 0.57g. The inhomogeneous magnetic field leads to a non-linear magnetic force on the high permeability beam, resulting in a nonlinear oscillator with a wide working bandwidth of 10 Hz or 18.5% of the operating frequency. This alternative
energy harvesting platform using high permeability magnetic materials provides great opportunities for energy harvesters that have high power output and wide working bandwidth.

5.3 High Output Vibration Energy Harvester with High Permeability Material

5.3.1 Introduction

Vibration energy harvesting technologies have been developing rapidly, and have shown great potential in many different applications. However, these applications have been severely limited by the amount of energy that can be harvested. Achieving high power and high power density vibration energy harvesters, which has attracted a lot of recent interests, will enable a much wider range of applications for vibration energy harvesters [5.36, 5.37]. So far, piezoelectrics based vibration energy harvesters which have received the most amount attention, have demonstrated much higher power density than their magnetic based counterparts, even though most piezoelectric based energy harvesters show a narrow bandwidth or a limited operating frequency range of 2~5% of the center operating frequency. For example, a piezoelectric bare beam based vibration energy harvester can generate a power of 6.63 mW/cm³[5.13]. Nevertheless, theoretically, the magnetostatic energy density $\frac{1}{2}\mu H^2$ in high permeability magnetic materials is $10^5$–$10^6$ times that of the electrostatic energy density $\frac{1}{2}\varepsilon E^2$ in piezoelectrics [5.38]. Vibration energy harvesters with hard or soft magnetic materials have been studied and tested [5.39, 5.40, 5.41]. However, the full potential of achieving high power density in
vibration energy harvesters with high permeability magnetic materials has not yet been realized.

This section discusses about an alternative design of a vibration energy harvester, which was demonstrated to significantly increase the output power density, by utilizing the strong magnetostatic interaction between solenoids with high permeability magnetic materials and vibrating hard magnet pair. There are two main reasons for improved output power density. First, more layers of high permeability materials served as the induction core than the first generation design, discussed in section 5.2. Besides, the magnetostatic energy on both sides of the magnets was utilized by placing one solenoid on each side.

5.3.2 Basic Mechanism

The schematic design of the energy harvester is shown in Fig. 5.18. Two identical solenoids with high permeability/insulator multilayer cores were placed at two sides of a vibrating hard magnet pair with anti-parallel magnetization. The key components of this energy harvester are the two identical solenoids with a high permeability (high-µ) MuShield® core inside which are placed at two sides of the magnet pair and fixed on a vibrating stage. The magnet pair is placed with the two magnets having anti-parallel magnetization and supported by a regular circular cross-section spring with its bottom fixed on the surface. When the magnet pair moves up and down with respect to the vibrating stage, the magnetic field inside each solenoid changes the direction periodically. The magnetostatic coupling between the solenoids and the time varying inhomogeneous bias magnetic field results in a non-linear oscillation and a complete magnetic flux
Fig. 5.18 The schematic design and working mechanism of the high power vibration energy harvester. (a) The magnet pair moves to the top. (b) The magnet pair moves to the bottom.

reversal in the solenoids. The presence of highly permeable cores dramatically increases the magnitude of magnetic flux inside the coils. Thus, a large induced voltage is generated on both sides of the solenoid, and the output voltage can be doubled with serial connection of the two solenoids.

5.3.3 Theoretical Analysis

The mass of the hard magnet pair, the stiffness of the supporting spring and the magnetostatic coupling between the solenoids and hard magnet pair together determine
the resonance vibration frequency and the output voltage of the energy harvester. The equivalent spring-mass system becomes a non-linear oscillation system due to the magnetostatic coupling between the solenoids and the hard magnet pair. This non-linear effect can be explained from the potential energy point of view, as explained in section 5.2. The elastic potential energy of a spring-mass system is a well-known linear one, with only one minimum value, which happens when the mass passes the equilibrium position in the middle. In contrast, the magnetostatic potential energy has two identical minimum values due to the coupling between the magnets and solenoids, which appear when the magnets move a short distance up or down from the equilibrium position in the middle. As a result, the superposition of two different types of potential energy make a non-linear one, leading to a wider oscillation frequency range, as has been shown in Fig. 5.17.

The total induced voltage equals the integral over the whole solenoid, because the magnetic field varies along the axis. The open circuit voltage can be expressed by:

\[
V = 2 \frac{d\phi(t)}{dt} = 2 \frac{d\int [H[x, y(t)] + 4\pi M[x, y(x, t)] \cdot A \cdot dN]}{dt} = 2 \frac{d\int 4\pi M[x, y(x, t)] \cdot A \cdot dN}{dt}, \quad (5.9)
\]

where \( A \) is the total cross section area of the multilayer cores, \( dN \) is the number of loops in the infinitesimal length unit of the solenoid, and

\[
dN = N_L \cdot \frac{dx}{dW}. \quad (5.10)
\]
$N_L$ is the number of loop layers and $d_w$ is the copper wire diameter. Hence, the maximum output power, which happens when the load impedance equals the conjugate of the output impedance of the solenoid coil, is

$$P_{\text{max}} = \left( \frac{V}{2} \right)^2 = \frac{16}{R_{\text{coil}}} \left( A' \pi \frac{N_L}{d_w} \right) \left( \int_0^L \frac{dM[x, y(x, t)]}{dt} \right) dx,$$

(5.11)

where $S$ is the number of layers in each core and $A'$ is the cross section area of one layer. Equation (5.11) indicates that the output power increases as the resonance frequency increases, if all other parameters are kept constant. Certainly, for the same source power, amplitude would be reduced if the frequency gets higher. Moreover, at a particular frequency, the output power depends on the total magnetic flux change in the solenoid in one oscillation period, which is directly related to the permeability of the magnetic cores. The solenoid with soft magnetic MuShield® core, which has a high permeability, has a great potential for generating a high voltage output. Meanwhile, the multilayer structure of MuShield® material is expected to generate a much larger flux change than a single layer. The application of a single layer was discussed in the earlier section.

### 5.3.4 Prototype and Testing System

The entire device is powered by a vibrating stage, which is driven by an audio power amplifier. Voltage output of the harvester in time domain is monitored by a digital oscilloscope. The device picture is shown in Fig. 5.19. The SmCo hard magnet has a dimension of 2.2cm × 1.3cm × 0.2cm. Each solenoid core is a 28-layer high permeability MuShield material, with a dimension of 2cm × 2cm × .002 inches. Total volume of the energy harvester is 6.44cm × 3.25cm ×1.4cm = 29.3 cm$^3$, including the solenoids, the
magnet pair and the gap in between. The coil resistance of each solenoid is 1.3 Ohm.

Fig. 5.19 Structure of the vibration energy harvester. Dimension of each component is: 2 × 2.5 × 1 cm³ for the solenoids, 1.25 × 2.2 × 1.5 cm³ for the magnetic pair, including the gap in between.

5.3.5 Results and Discussion

Figure 5.20 indicates the measured open circuit voltage of the energy harvester with different springs at resonance frequencies of each. For spring #1, with resonance frequency 27 Hz, the peak voltage is 1.18 V for acceleration of 2g (g=9.8 m/s²); spring #2, with resonance frequency 33 Hz generated maximum voltage of 1.64 V for acceleration of 3g; spring #3, brought up peak voltage 2.52 V, with intrinsic frequency 42 Hz and acceleration of 5 g. Increasing acceleration values were applied to maintain the same source vibration amplitude. The maximum output power on a 2.6 Ohm load is 133.88 mW, 258.62 mW and 610.62 mW, respectively, as shown in Fig. 5.21.
Considering the total practical volume of the device 29.3 cm$^3$, this device has good performance with the maximum power density, 20.84mW/ cm$^3$ at 42 Hz with spring #3. Q factor of the harvester at 42 Hz was 16, which was obtained from the decay curve of output voltage when turning off the source $^{[5,10]}$. Almost the entire device damping is generated from the mechanical collision between the spring supported magnets and the solenoid holder. That means a much lower input force or acceleration is needed and

![Fig. 5.20 Measured results of the open circuit voltage for the energy harvesting device with three different springs at respective resonance frequencies: spring #1 at 27 Hz; spring #2 at 33 Hz and spring #3 at 42 Hz.](image)

much higher Q factor could be reached by using better crafting technique. A simple relation between frequency and power can be derived from Eq. (5.11), $P_{\text{max}} \sim \left( \frac{\Delta M}{\Delta T} \right)^2 \sim$
$f^2$, if all other parameters kept constant, where $\Delta M$ is the flux change per period and $\Delta T$

![Graph showing output power vs. frequency for different springs.](image)

**Fig. 5.21** Measured maximum output power of the harvester with three different springs, at resonance frequency of each.

the period. In fact, the measured results agree with the parabolic curve fitting, as shown in Fig. 5.21. Clearly this vibration energy harvester design can accommodate different vibrating frequencies of the environment by changing the spring that is connected to the hard magnet pair. If the vibration amplitude of the testing stage kept the same, the output power and power density are proportional to the second power of the vibration frequency. Hence, if this $P_{\text{max}} \sim f^2$ can be extrapolated to higher frequencies, much higher output power density can be achieved on condition of constant amplitude. It is noticeable that large working bandwidth could still be obtained at high frequency due to the non-linear
effect. These exciting data prove a promising future of the high-permeability material based energy harvesting mechanism.

In order to study the frequency spectrum with it, testing data were collected at different source frequency. Indicated in Fig. 5.22, output power shows sagging rise before 42 Hz and rapid decline afterwards. The major reason for the asymmetrical curve is the non-linear oscillation with increasing mechanical damping as the ascending frequency \(^{[5,37]}\). The half-power bandwidth of the device with spring #3 was measured to be 6 Hz, ~15% of the central frequency, which is much higher than the typical 2~5% bandwidth of typical piezoelectric cantilever based energy harvesters \(^{[5,13]}\).

![Figure 5.22](image.png)

Fig. 5.22 Measured output power spectrum of the harvester with spring #3. Maximum output is 610.62 mW, obtained at 42 Hz, corresponding to a volume density of 20.84 mW/cm³. This curve shows a half-peak working bandwidth of 6 Hz.
5.3.6 Advantages of the 2\textsuperscript{nd} Generation High Permeability Vibration Energy Harvester and Summary

Compared to the previous vibration energy harvester design based on a vibrating high-\(\mu\) beam and a stationary bias hard magnet pair, this new generation device has a new structure that utilizes a vibrating hard magnet pair and a stationary solenoid pair with thick multilayer high-\(\mu\) core materials. The multilayer high permeability solenoids core leads to significantly increased flux change in the solenoid within one period without increasing the total volume of the device. In addition, the two solenoids at both sides of the vibrating magnets make full use of the spatially inhomogeneous bias magnetic fields at both sides of the magnets, leading to doubled power output, and a dramatically enhanced power density by \(\sim 20\times\) over the previous energy harvester design with high-\(\mu\) materials.

In summary, we investigated a 2\textsuperscript{nd} generation of magnetic based vibration energy harvesters, achieving a high power density larger than 20mW/cm\(^3\) with acceleration of 5 g, which is over 3 times of the best power density data reported for vibration energy harvesters \[^{[5,13]}\]. Unlike the piezoelectric cantilever based vibration energy harvesters that suffer from narrow bandwidth and brittle cantilever and degraded polarization after a certain time of service, the new generation vibration energy harvester design based on multilayer high permeability magnetic material exhibits giant power density, high output power and large bandwidth, which provides great opportunities for practical compact energy harvesters.
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6 Conclusion and Future Work

The basic magnetic properties of soft magnetic materials were discussed in this work, including the low coercive field and high magnetization. The widely used experimental methods of magnetic films were introduced. All the FeCoB films related to this work were deposited with the physical vapor deposition system. Some typical magnetic material characterization methods were explained. The vibrating sample magnetometer (VSM) is used to detect and plot the quasi-static magnetic field response of the magnetic material. The ferromagnetic resonance curve shows the high frequency magnetic loss.

A series of sandwich structure of FeCoB(100nm)/Al₂O₃/FeCoB(100nm) was made with the PVD system and studied, in order to find the optimized thickness of the Al₂O₃ insulating layer. It was found that the anti-ferromagnetic coupling exists when the Al₂O₃ thickness is lower than 6 nm, resulting a higher coercive field and FMR linewidth. As a result, a minimum thickness of 6 nm of Al₂O₃ should be achieved when applying the FeCoB multilayer structure for high frequency energy applications.

The magnetron sputter deposition technology is compatible with the on-chip micro-fabrication method. On Si chip integrated inductors applying the FeCoB/Al₂O₃ multilayer core structure were designed, fabricated and tested. Three different types of solenoid inductors were designed, with the same area but different shape of cores. The theoretical model was discussed in order to analyze the inductance, resistance and quality factor. The on-chip micro-fabrication process was designed as discussed. The testing
technique which requires a vector network analyzer and a probe station was introduced, as well as the de-embedding and parameter extracting method. After the measurement, the inductance and quality factor curves were plotted. A maximum inductance of 60 nH was obtained at low frequency. It was found that the inductance increases with the number of turns when keeping the other variables unchanged. The magnetic loss gets lower when the core was divided into two separate parts. However, the electrical loss increases because the total number of turns doubles for the two-sided design. A comparison between the inductors with and without the magnetic cores was done by plotting their inductance or quality factor curves together. The inductance was increased by a factor of 3.6 ~ 6.7 with the magnetic cores. The magnetic loss due to the core is very high, so the magnetic annealing process is very necessary to improve the magnetic core, which will be done in the future. Also, the on chip integrated transformer could be studied, which shares the same fabrication process.

The soft magnetic materials also play an important role in serving some other energy related applications, such as the vibration energy harvesting system. Two generations of high permeability material applied vibration energy harvesting devices were designed, made and tested. The strong magnetic coupling between the magnets and the high permeability material is able to generate a large induced voltage across the solenoid, when they have relative vibration. This coupling also generates the magnetic potential energy of the oscillation system, which enables the so called non-linear effect, leading to a wide working bandwidth of the devices. The 1st generation of vibration energy harvester was able to generate 74mW on a 1Ω load and with a time average value 5mW at an acceleration 0.57g corresponding to a maximum power density of 1.07
mW/cm$^3$ or 1.88 mW/g cm$^3$. Meanwhile, the device had a working bandwidth as wide as 10 Hz/54 Hz. The design of the 2$^{\text{nd}}$ generation was improved, so that the magnetic energy on both sides of the magnets could be utilized. This device was proved to be able to generate 610.62 mW at 42 Hz, corresponding to a volume density of 20.84 mW/cm$^3$ with a half-peak working bandwidth of 6 Hz. The energy loss in this design mainly came from the mechanical dissipation between the parts of the device. The efficiency could be significantly improved in the future if the joint parts are re-designed and made with more professional machines.

In summary, the application of soft magnetic materials with high permeability was done on two different energy related topics: on chip integrated inductors and vibration energy harvesting. The decent testing results of all devices proved the feasibility and potential.