A simple, wireless powering scheme for MEMS devices

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ABSTRACT

With recent developments in micromachining technology, fabrication of discrete microdevices is maturing; consequently, system integration is becoming an ever more important issue. One obstacle to such systems is the diverse power requirements of microdevices, especially actuators. Since some types of actuators exhibit relatively high voltage or power requirements, it is not feasible to integrate power supplies on-chip, and it is often inconvenient for the MEMS system to be tethered to interconnects for purposes of supplying power. On-chip wireless power sources can be implemented to circumvent this problem. Here, a simple wireless powering scheme, which utilizes a transformer with an air gap in its core, is demonstrated. The transformer secondary is fabricated on-chip and is detachable from the transformer. Experiments and simulations are performed to maximize the coupling between the primary and secondary. Coupling coefficient close to 0.8 was obtained. Frequency properties of the transformer were studied. In the case of the thin-film secondaries demonstrated here, the transformer operates at frequencies less than a few MHz. Usably high voltage (223.4 Vpp) and high power delivered to a load (4.5 W rms) were obtained from the secondary to demonstrate the transformer capabilities.

Keywords: MEMS power supply, on-chip powering, inductive coupling, integrated inductors

1. INTRODUCTION

As a relatively new research area, microelectromechanical systems (MEMS) have received ever-growing attention. Although much work has been done on discrete devices, relatively few complete systems on a chip have been demonstrated. Such systems will realize enhanced functionality and should lead to greater acceptance. However, most MEMS devices, especially microactuators, have special requirements for their power supplies; some devices require high voltage, while some require high input power. Consequently, in some cases more than one power supply may be necessary to drive such a chip, thus increasing the number of interconnects, which is undesirable for packaging. Also, it is desirable to develop a method to release MEMS systems from power supply tethering in order to give them some degree of autonomy, and allow implantation of devices in biomedical or other applications. This paper explores the feasibility of inductively powering microsystems to decrease their overall size, increase their available energy budget, and free them from mechanical and electrical contacts.

One method for on-chip powering is to use electrochemical cells, such as batteries, but this suffers from limited storage capacity and lifetime. Therefore, a regenerating power supply will be more desirable under some circumstances. Some other schemes that generate electricity on site have already been prototyped. For example, Williams and Yates demonstrated a microelectric generator to utilize kinetic energy from movements and vibration, having dimensions of around 5 mm x 5 mm x 1 mm and generating 1 µW at 70 Hz, and 0.1 mW at 300 Hz. Qu et al. demonstrated a 16 mm x 20 mm x 0.05 mm thermopile module that yields a thermoelectric output of 8.4 mV/K and is capable of generating a voltage of 0.25 V at a temperature difference of 30 K. The above-mentioned strategies cannot produce a usable amount of power for many MEMS devices. As an alternative, a hydrogenated amorphous silicon (α-Si:H) solar cell array has been demonstrated as an on-chip power source for electrostatic MEMS. It is made up of a series-connected array of 100 single solar cells (a single cell of a triple layer of p-i-n / p-i-n / p-i-n a-Si:H) occupying a total array area of 1 cm², producing an open circuit voltage of 200 volts, with a conversion efficiency less than 5%. While this method can provide enough voltage for electrostatic devices, its current sourcing ability is limited. Besides, it requires an external light source that may not always be available.

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In comparison, inductively coupling power onto the chip combines the advantages of no physical connections, no moving parts, high reliability and efficiency, IC fabrication compatibility and ease of packaging. Here we report a prototype power supply based on a transformer secondary coil that is microfabricated on a silicon chip. The transformer primary is wound around a ferrite core incorporating a variable air gap. When the secondary is inserted into the air gap, energy is coupled from the primary side into the secondary side. For such a strategy, there is the necessity of an inductive power coil, but many systems allow such a coil to be included, leaving the MEMS system free of interconnect wiring. Our design for test, and ultimate inclusion in a blood monitoring system, is a C-core design. Others have proposed different core designs for implantable devices \(^7\),\(^8\), which offer lower coupling coefficients, and therefore less power delivered to a load than the design described here. We have achieved coupling of up to 4.5 W\(_{\text{rms}}\) of power to a load through our secondary coil.

2. THEORY

Transformers operate by the principle of electromagnetic induction. When alternating voltage is applied on a coil, a changing magnetic field common to both coils is produced in the magnetic path formed by the core. In turn, an alternating voltage of the same frequency is induced in the other coil, and electrical energy is transferred from the input circuit to the output.

The proposed detachable integrated transformer differs from a conventional one in that the secondary is not physically attached to the rest of the transformer, but is instead integrated on a silicon chip. Since the energy transfer is achieved through the magnetic field in the air gap, it is important to relate the transformer performance to the coupling between the primary and secondary. For transformers of imperfect coupling, with schematic as shown in Fig. 1, the voltage drop across each terminal can be expressed as:

\[ V_1 = j\omega L_1 I_1 - j\omega M I_2, \]
\[ V_2 = j\omega M I_1 - j\omega L_2 I_2 = Z_L I_2, \]

where \( L_1, L_2, \) and \( M \) are the primary inductance, secondary inductance and mutual inductance, respectively, and \( Z_L \) is the load impedance across the secondary terminals. The degree of coupling between the primary and secondary is defined as the coupling coefficient, \( k \), which is related to \( L_1, L_2 \) and \( M \) by:

\[ k = \frac{M}{\sqrt{L_1 L_2}}. \]

The coupling coefficient is one of the most important figures of merit for transformer performance. When one coil (for instance, the secondary) is short circuited, \( V_1 \) can be written as:

\[ V_1 = j\omega L_1 I_1 - j\omega M \left( \frac{M}{L_2} \right) I_1 = j\omega \left( 1 - k^2 \right) L_1 I_1. \]

Note that the impedance looking into the primary of an ideal transformer with its secondary shortcircuited is zero. For a non-ideal transformer, the remaining impedance with the secondary shorted is equivalent to an inductor of value \((1-k^2)L_1\). The equivalent inductor \((1-k^2)L_1\) is called the primary leakage inductance because this term does not contribute to...
magnetic flux in the secondary, but rather subtracts from the possible impressed voltage on a load. (Similarly, the same relationship can be obtained for the secondary side.) We use this relationship below in measurements to obtain the coupling coefficient.

With a load $Z_L$ on the secondary side, voltage $V_1$ and current $I_1$ are related by,

$$V_1 = j\omega L_1 + \frac{\omega^2 M^2}{Z_L + j\omega L_2} I_1,$$

which can be rewritten as

$$V_1 = j\omega (1 - k^2) L_1 I_1 + \left(\frac{j\omega k^2 L_1 \cdot k^2 \frac{L_1}{L_2} Z_L}{j\omega k^2 L_4 + k^2 \frac{L_4}{L_2} Z_L}\right) I_1.$$  \hspace{1cm} (6)

Therefore, the impedance on the secondary side can be reflected onto the primary side multiplied by a factor of $k^2 \frac{L_1}{L_2}$, and it is in parallel with the primary magnetizing inductance $k^2 L_4$. For ideal transformers, $k^2 \frac{L_1}{L_2}$ can be reduced to

$$\left(\frac{N_1}{N_2}\right)^2.$$  \hspace{1cm} (7)

Based on the above derivation, an equivalent circuit model of the transformer is shown in Fig. 2 with parasitic effects taken into consideration. Both primary and secondary coils have parasitic resistance and capacitance, and the parasitics on the secondary side are treated as loads. The secondary of the transformer is a spiral inductor fabricated by planar processing on an oxidized silicon wafer. Significant parasitic effects are introduced into the secondary side because of limitations imposed by the processing techniques. There exists relatively large series resistance caused by the small metal thickness, and also non-negligible capacitance shunting the inductor due to the close proximity of the metal lines and fields through the oxide to the substrate. The series resistance reduces the energy transfer efficiency into the

![Transformer equivalent circuit](image)
load, and parasitic capacitance together with the inductance forms an equivalent tank circuit with a self-resonant frequency \( f = \frac{1}{2\pi \sqrt{L C}} \), imposing an upper limit on the operating frequency of the transformer, beyond which the secondary no longer functions as an inductor, but rather as a capacitor. Parasitic resistance and capacitance of the primary are much less important compared with those on the secondary side. The core loss resistance, \( R_{\text{CORE}} \), in parallel with the primary accounts for the energy losses in the core.

The elements in the circuit have different importance for transformers of different configurations. In this paper, the major concern is to improve the coupling coefficient and find the optimal operating frequency range. These issues will be addressed in the following sections.

3. COIL FABRICATION AND CHARACTERIZATION

The transformer is composed of a wire-wound primary, a ferrite core and a microcoil on a silicon wafer as the secondary. The core is made up of two pieces so as to provide an adjustable air gap. Figure 3 shows the fixture for supporting the core and coil during test. The fabricated secondary is mounted on a printed-circuit board for easy handling. The secondary is a square spiral microcoil with average side length of 12 mm, with geometry as in Fig. 4(a). Processing involves e-beam evaporation of a layer of copper on an oxidized silicon wafer, and patterning by liftoff. The copper coil is 1.2 microns thick. The coil has 10 turns and the metal traces are 80 \( \mu \)m wide on a 100 \( \mu \)m pitch. The physical and electrical parameters of the core and the secondary are summarized in Table 1.

![Fig 3. Micro-transformer setup for measurement.](image)

Table 1. Physical parameters of the transformer core and secondary.

<table>
<thead>
<tr>
<th>Core</th>
<th>Secondary</th>
</tr>
</thead>
<tbody>
<tr>
<td>Material</td>
<td>Philips 3F3</td>
</tr>
<tr>
<td>Side length</td>
<td>50 x 40 mm(^2)</td>
</tr>
<tr>
<td>Cross sectional area</td>
<td>8 x 8 mm(^2)</td>
</tr>
<tr>
<td>Air gap width</td>
<td>0 - 10 mm</td>
</tr>
<tr>
<td>Specific power loss 100 °C, 400 kHz, 50 mT</td>
<td>&lt;150 mW/cm(^3)</td>
</tr>
<tr>
<td>Inductance</td>
<td>3.4 ( \mu )H</td>
</tr>
<tr>
<td>Relative permeability at 25°C</td>
<td>2000</td>
</tr>
</tbody>
</table>

The coil impedance was measured using an Agilent 4294A impedance analyzer over the frequency range from 40 kHz to 110 MHz. Curve fitting of the impedance versus frequency characteristic was performed using a 3-element
model, as shown in Fig. 4(b), to extract the values of the lumped elements. A comparison of measured coil impedance with calculated values from the model is given in Fig. 4(c) with \( L = 3.4 \, \mu H \), \( R = 95.5 \, \Omega \), \( C = 67 \, \text{pF} \); it can be seen that the measured values and calculations agree well with each other. The self-resonant frequency of the secondary coil is around 10 MHz. These parameters will be used below in predicting the frequency response of the transformer.

Fig. 4. Secondary coil impedance measurement and curve fitting using lumped element model (corresponding to elements in (b), \( L = 3.4 \, \mu H \), \( R = 95.5 \, \Omega \), \( C = 67 \, \text{pF} \)).

### 4. TRANSFORMER MEASUREMENT

For reasons mentioned earlier, the coupling coefficient is an important figure of merit to evaluate the design of a transformer. The coupling coefficient is obtained in the following way, referring to Fig. 2: Open circuit the primary and measure total secondary inductance, \( L' = L_2 \); short circuit the primary and measure secondary inductance, which is now the leakage inductance, \( L'' \). The difference in coil inductance between the two measurements is the coil magnetizing inductance, which is equal to \( L_{2,m} = k^2 L_2 \). Therefore \( k = \sqrt{\frac{L_{2,m}}{L_2}} = \frac{L' - L''}{L'} \). (For an ideal transformer, this should agree with measuring \( k \) from the primary side. Our measurements could not be performed on the primary since the equations are not valid for large \( Z_L \), which is the case for our coils with series resistance of 95.5 ohms.) Strictly speaking, only when the coil parasitic effects, such as those due to capacitance, have a negligible effect on the total reactance can accurate inductance values be measured. Therefore, our measurements are inevitably subject to some errors since the susceptance due to parasitic capacitance becomes significant at high frequency. Nevertheless, our measured \( k \) values are meaningful at lower frequencies, and they serve the purpose of comparison between different transformer designs.

The first step in optimizing transformer design is to examine to what degree the relative positions of the two coils will affect their coupling. Since the primary coil is free to move along the core, placement of the primary was investigated, as shown in Fig. 5(a). The data were taken for transformers with an air gap of 2 mm.
Figure 5 suggests that coupling coefficient is frequency dependent, which is due to a variety of effects. At moderately high frequencies, eddy currents limit B field penetration of conductors in the windings which reduces leakage energy, effectively increasing $k$. At higher frequencies, parasitic capacitances short circuit the secondary, so the effective coupling decreases. The data at lower frequencies, however, are accurate since the capacitive susceptance at lower frequencies is low enough to be neglected. The coupling coefficients taken at the lower end of the curves are 0.80, 0.75 and 0.41 for 8-turns split primary, primary on the small arm and primary on the large arm, respectively. These results demonstrate that the coupling coefficient can be increased greatly by decreasing the physical separation between...
the primary and secondary in spite of the use of a ferrite core of high permeability. This design is not applicable for implanted MEMS systems, but is practical for other systems, such as a lab-on-a-chip, currently under development at Notre Dame. (The secondary coil self-resonant frequency is lower than that in Fig. 4(c), because that measurement used an air-core, while here the coil is inserted in the air gap of a ferrite core, whose increased inductance leads to a reduced self-resonant frequency.)

To verify that the coupling coefficients are reasonable, the ratios of open circuit secondary voltage to impressed primary voltage were measured and plotted over the frequency range from 50 kHz to 3 MHz, shown in Fig. 5(c), and compared to simulations using the above k values. The agreement of the measured and simulated data suggests that the k values extracted from the coil measurements are valid in spite of the assumptions discussed above.

The coil inductances are used in the simulations in Fig. 5(c) instead of coil turn numbers because the relationship 

\[ L = \frac{N^2}{\mathcal{R}} \]  

(where N is the turn number, and \( \mathcal{R} \) is the reluctance of the magnetic path) is valid only for perfect coupling, and in non-ideal transformers such as these, i.e. for \( k < 1 \), the square root of the inductance ratio no longer equals the turns ratio. The measured primary inductances were 13.7 \( \mu \)H, 6.3 \( \mu \)H and 3.8 \( \mu \)H for the 8 turns primary on the large arm, on the small arm and split primary, respectively. One might think that the primary inductance is independent of the coil’s position within the magnetic path, but for a non-ideal inductor, flux leakage plays an important role in reluctance drops around the magnetic loop. In this case, placing the coil across the air gap has two main effects contributing to increasing the voltage ratio: it (a) generates a maximal field strength across the air gap, increasing the k value, as seen in Fig. 5(b), and (b) decreases the primary inductance, which is equivalent to reducing the number of primary turns in an ideal transformer. Neither of these effects would occur were it not for the presence of flux leakage in this non-ideal transformer.

The next step taken to maximize magnetic coupling is to investigate the dependence of coupling coefficient on the turn numbers of the primary coils. Figure 6 gives the experimental results for this with a 2 mm air gap and 10-turn secondary. It can be seen that the coupling coefficient of the 5-turn primary is noticeably less than that of primaries having more turns, and the small differences in the coupling coefficients of 8-, 10- and 20-turn primaries are likely due to the lack of precision in creating the hand-wound primary coils. Coils of only a few turns have reduced coupling coefficient because for low turn numbers, the generated magnetic flux is not effectively constrained within the core, and flux leakage is more serious.

Fig. 6. Coupling coefficients for transformers with split primaries of various turn numbers.
The above measurements are all performed with a 2 mm air gap, and it is useful to examine the behavior of coupling coefficients at different air gaps. Figures 7 (a) and (b) show the coupling coefficients of transformers with 10-turn and 20-turn primaries with air gaps of 2 to 5 mm. As expected, the coupling coefficient drops as the air gap widens. The coupling coefficients decrease at approximately the same rate for transformers with 10-turn and 20-turn primaries. It also shows that a 10-turn primary can confine magnetic flux to the core as well as a 20-turn primary.

In the foregoing discussion, it is mentioned that transformers cannot operate at frequencies exceeding the self-resonant frequency of any of its coils. An exploration of transformer performance beyond its normal frequency range reveals some interesting features. Figure 8 shows that the voltage ratio remains constant up to 1 MHz, then rises rapidly and peaks between 9 ~ 10 MHz, decreasing sharply at higher frequencies. This phenomenon can be explained as follows: At low frequencies, the open circuit voltage ratio can be expressed by the relationship

\[ \frac{V_2}{V_1} = k \sqrt{\frac{L_2}{L_1}} \]

which is determined by the transformer characteristics; at higher frequencies, other factors such as parasitic capacitance begin

![Fig. 7. Coupling coefficients of transformers with (a) 10-turn and (b) 20-turn primary at different air gap widths.](image)

![Fig. 8. The ratio of open-circuit secondary voltage to impressed primary voltage as function of frequency.](image)
to dominate, and the secondary voltage is strongly influenced by resonance between the parasitic capacitance and the leakage inductance. Using the coil equivalent circuit in Fig. 4(b) with the coupling coefficient, the transformer voltage ratio as a function of frequency is simulated using Micro-Cap 6™ and the results are shown in Fig. 9. The amplification seen in Fig. 9 results from a series resonance of internal secondary parasitic capacitance and leakage inductance.

The above discussion establishes that in order to maximize magnetic coupling, the primary should be split across the air gap and there is a lower limit of turn number for the driving coil; due to the parasitic capacitance inherent to integrated inductors and within the transformer, the operating frequency of the proposed transformer should not exceed a few MHz.

Since the project goal is to develop an on-chip powering scheme accommodating the special power requirements of MEMS devices, it is relevant to demonstrate the versatility of our prototype transformer. For the transformer of 10-turn primary and a 2 mm air gap, 223.4 Vpp was obtained for an impressed primary voltage of 253.1 Vpp at 744.9 kHz, as shown in Fig. 10. Such a voltage is appropriate for the operation of such MEMS devices as electrohydrodynamic microfluidic pumps, electrostatic actuators, and many other devices, especially if integrated with a simple half- or full-wave rectifier bridge, with or without filtering or regulation. In addition, we have demonstrated 4.5 Wrms of power delivered to a 100 Ω load. It can be concluded that the prototype transformer is capable of generating usably high voltage and high power for applications in MEMS systems.
5. SUMMARY

In this paper, a transformer with a detachable on-chip secondary is proposed to act as an wireless power source for MEMS devices, and an operational transformer is demonstrated. Experiments and simulations have been carried out to characterize the integrated coil, maximize the magnetic coupling, and determine the transformer frequency properties. A maximum coupling coefficient of 0.8 has been achieved for an air gap of 2 mm, and the measurements have been verified by simulations. Parasitic capacitances on chip and in the transformer limit the transformer operation frequency to a few MHz.

For the time being, all the measurements were done with high-resistance, thin-film secondary coils, yet the transformer has already demonstrated high voltage and high power capability. With the adoption of more advanced processes, e.g. deep reactive ion etching and electroplating, it is projected that the transformer performance can be further improved through significant decrease in internal resistance, as well as possible back-side placement to conserve chip real estate.

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