Batch-Type Millimeter-Size Transformers for Miniaturized Power Applications

Farid Amalou, Etienne L. Bornand, and Martin A. M. Gijs

***Abstract*—We have developed a novel batch-type technology for making three-dimensional (3-D) millimeter-sized transformers for ultrasmall low-power (0.1 to 1 Watt) applications. The technology uses the 3-D micropatterning of ferrite wafers by powder blasting to form the magnetic cores of the inductive devices, and combines these cores with electrical windings made by flex-foil printed circuit board technology. Microfabrication and assembly of the parts can be done in a batch process on a wafer/foil level, opening the way to further size reduction of the components. We have measured the inductive and resistive properties of our devices as a function of frequency and device geometry. The results clearly show the high potential of our technology for power applications in which small-size is important.**

***Index Terms*—Batch process, ferrite, inductance, miniaturization, planar transformers, powder blasting, power electronics, printed circuit board (PCB) technology.**

I. Introduction

MINIATURIZATION of inductive devices for signal transformation, sensitive magnetic field detection, or switch-mode power supply applications is an increasingly important subject today. Often these devices determine, to an important extent, the size of the complete application in which they are employed. Especially power electronics has been a field of strong miniaturization, thanks to the increasing frequencies (0.1 to 1 MHz) of inductive device operation. These high frequencies necessitate the use of high-resistivity magnetic core materials, such as ferrites, for reducing eddy current losses in the inductors or transformers. During recent years, a large research effort has been developed in the field of so-called planar magnetics, where one integrates flat three-dimensional (3-D) ferrite cores with Cu windings realized in planar printed circuit board (PCB) technology [1]–[3]. This device geometry has lead to devices with reduced height and enhanced power density in a vast range of power (few Watt to many kW). However, for special ultrasmall low-power applications like mobile electronics or space electronics, such advanced solutions do not exist, as individual component sizes prevent an easy assembly and handling of the devices.

On the other hand, in the field of microsystems, several miniaturized inductive applications, realized in a batch-type technology, have been demonstrated. Park *et al.* [4] reported on 0.5 cm size inductors based on a 15 μm thick electroplated $Ni_{80}Fe_{20}$ and $Ni_{50}Fe_{50}$ magnetic cores. These structures are realized by a number of sputter deposition, electroplating, photoresist spinning and patterning steps. Inductance values of 0.5 μH were reported at 10 kHz. Löchel *et al.* [5] have used thick resist technology, sputtering, and electroplating/etching methods to fabricate coils of a few mm size based on NiFe core material. Microtransformers of 5 mm size integrated with diodes on a Si wafer were reported by Mino *et al.* [6], [7]. These transformers are based on an amorphous magnetic core prepared by sputtering, in the form of three separate layers of CoZrRe, each 5 μm thick, with 0.1-μm SiO_2 spacer layers, to reduce eddy current losses. Recently, we have proposed a new technique for the realization of inexpensive and flat transformer devices with relatively high inductance [8], [9], based on the combination of micropatterned amorphous magnetic cores of very high permeability ($\mu_r \cong 10^5$) with electrical windings realized in PCB technology. All these microsystem power transformation solutions are based on integrated and hence, potentially cheap batch-fabrication processes. However, as the microfabrication technologies used are essentially two-dimensional, cross-sections of the ferromagnetic core materials are small and, due to the use of conducting magnetic materials, high-frequency operation is disabled.

In this paper, we propose a novel method for the batch-fabrication of 3-D inductors and transformers, with which we open the way to economically feasible ultra-miniaturized ($\leq$1 mm) low-power ($<$ 1 Watt) applications. Our devices are geometrically similar to larger size planar magnetic components, and consist of two half magnetic ferrite cores and a printed circuit board or flex foil carrying the electrical windings around the core. The 3-D ferrite cores are microstructured out of a 1-mm-thick ferrite wafer using a newly developed batch-type micropowder blasting process [10]. In this way, we can realize many cores in parallel and assemble them at wafer-level with the electrical winding patterns; hereafter, complete devices are separated. We have measured the electrical properties of our devices such as main inductance, leakage inductance, and resistance, as a function of frequency and device geometry. The results clearly show the high potential of our technology for ultra-small low-power applications.

II. Transformer and Inductor Design

As mentioned previously, our devices are geometrically similar to larger planar magnetic components. The electrical windings are elaborated using standard PCB technology on flex foils. Two designs were considered for the coil patterns containing 12 and eight turns, respectively, corresponding to track widths of 60 and 80 μm, respectively. Double-sided PCB technology with

Manuscript received April 14, 2000; revised March 27, 2001.

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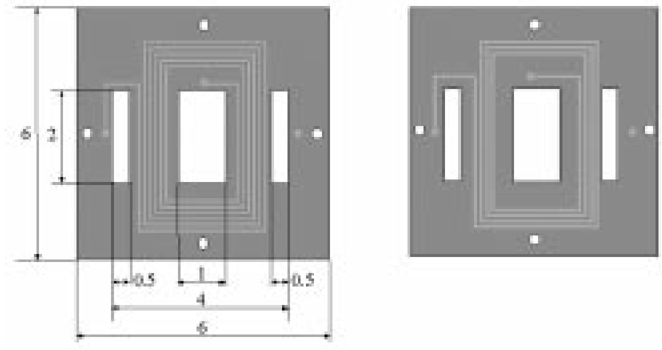
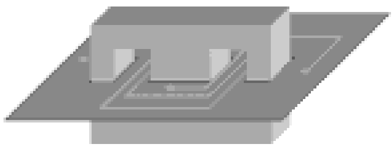
Publisher Item Identifier S 0018-9464(01)05163-9.

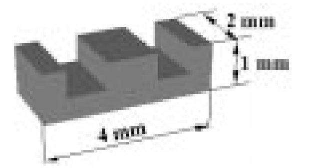


(a)

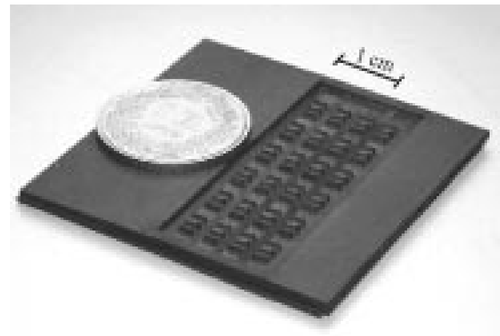


(b)

Fig. 1. Schematic view (a) of a top view of the coils containing, respectively, 12 and 8 turns, and (b) an assembly of a coil with two E-cores. Dimensions are in mm.



(a)



(b)

Fig. 2. Representation of ferrite structuring. (a) Shape and dimensions of the E-cores and (b) photography of an array of identical E-cores realized on a ferrite wafer.

via holes is used. Thus, for each of the windings, each side of the flex-foil holds half of the number of turns. Fig. 1(a) shows a top view of the designs. General dimensions (in mm) are also given in this figure.

Ferrite cores are well suited for a variety of magnetic circuit applications, and their relatively high permeability factors make them especially useful for high inductance values with a minimum number of turns, resulting in smaller component size. Fig. 1(b) shows a perspective view of the electrical windings integrated with two ferrite E-cores. The geometry is the same for the inductors (containing one coil of windings) and transformers (containing two sets of windings). A simple relation allows the calculation of the inductance L of a ferrite core

$$L = A_L N^2 \tag{1}$$

where A_L is the inductance factor and N the number of turns. The A_L factor contains the geometrical and magnetic properties of the ferrite core through the relation

$$A_L = \frac{\mu_0 \mu_e}{\sum \frac{l}{A}}, \qquad \text{with } \mu_e = \frac{\mu_i}{1 + \frac{l_g \mu_i}{l_e}} \tag{2}$$

where A and l are, respectively, the effective area and the path length of the core, and μ_e is the effective permeability introduced to take into account the influence of an eventual air gap in a closed magnetic circuit. In the denominator, the summation is performed over all parts of the magnetic circuit. In the second expression, μ_i represents the initial permeability of the material, l_g the air gap length, and l_e the effective length of the magnetic circuit. One can deduce from (2) that scaling down the quantities l and A with the same factor does not affect the value of the inductance factor, and expect the same magnetic performance as for large scale devices. One observes that the influence of an air gap on the effective permeability will be more important when reducing scale, leading to lower A_L values. Using (1) and (2), one can obtain typical values of 22 μH and 10 μH, respectively, for 12 and 8 turns and for the dimensions of the ferrite E-core shown in Fig. 2(a). These values where calculated taking into account a parasitic gap of 14 μm (will be discussed later at paragraph 4.3) generated from the superposition of two ferrite E-cores.

III. Fabrication Process

For the realization of the windings we used a standard PCB process on polyimide flexible substrates. For the 3-D microfabrication of the ferrite E-cores, we have used a very innovative approach, namely, mechanical micro-erosion by powder blasting. The micropowder blasting technique is a batch process, which allows three dimensional patterning with high precision. The powder, consisting of 30-μm size alumina particles (Al_2O_3) is ejected in a compressed air flux through a rubber tube to a nozzle. This approach allows patterning of a large variety of brittle materials, such as ferrite wafers. We used the 3F3 ferrite of Philips for our experiments [11], which is particularly well suited for power applications. An uniform exposure of a ferrite wafer is obtained using an x axis translation stage for the substrate table, and a y axis translation stage for the nozzle. Patterning is performed by applying a 0.5-mm-thick laser cut stainless steel mask containing the desired structures on the ferrite wafer. The mask is held in intense contact with the wafer using a hard magnet biased substrate table and allows masking selected ferrite wafer areas against erosion. More details about this technique are given in [10].

As indicated in Fig. 2(a), typical dimensions that we realized using this approach are in the millimeter range. Scanning the surface of a ferrite wafer along the x and y directions and using the appropriate metallic masks, we obtain an array of very small and identical ferrite E-cores, as shown in Fig. 2(b). Finally, the E-cores and the windings are assembled all together. Fig. 3 shows an assembled mini-transformer.

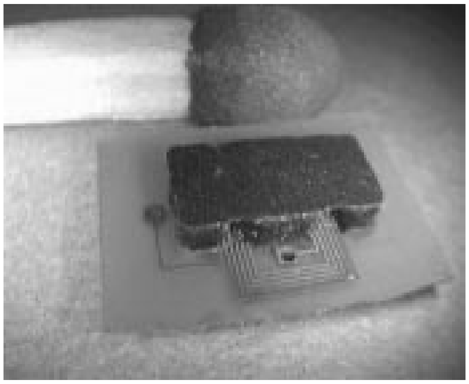


Fig. 3. Assembled mini-transformer.

Our approach clearly demonstrates the potential of this technique for batch processing of magnetic devices of very small dimensions, opening the way to even further size reduction of such components.

IV. Experimental Results

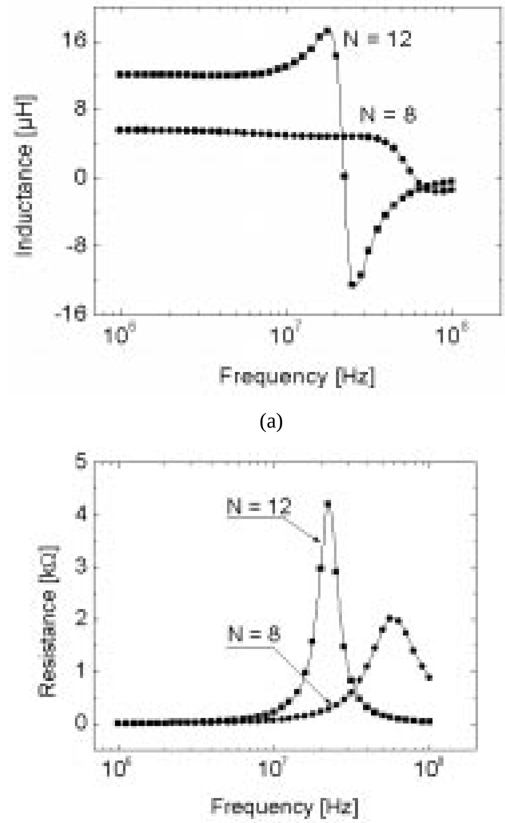
In order to cover a wide domain of frequencies, the electrical characterization of the mini-transformers have been performed with a hp 4194A "Impedance/Gain-Phase Analyzer" and a hp 4291A "RF Impedance/Material Analyzer." With these two instruments, the frequency was varied from 10^4 up to 10^8 Hz. The inductance, resistance and gain of the devices were recorded using sinusoidal signal levels of 0.5 V rms. During measurements of the electrical properties, the ferrite cores were clamped mechanically, resulting in a minimum air gap.

A. Main Inductance and Resistance

The conditions for the main inductance and resistance measurement were the following: one coil at a time was mounted with two ferrite E-cores and the measurements of the inductance and resistance were performed simultaneously. The results are shown in Fig. 4(a) and (b) respectively. Measurements indicate that the inductance of the 12-turns coil is about 12.2 μH, while it is 5.5 μH for the eight-turns coil, over more than two decades of frequencies, starting from the lowest frequency used. In Fig. 4(a) and (b), only measurements for frequencies above 1 MHz are shown. Negative inductance values in Fig. 4(a) indicate a capacitive element given by $C = -L^{-1}\omega^{-2}$. Indeed, at high frequencies, the device is seen by the impedance analyzer as a capacitance. The plot of Fig. 4(a) clearly shows the resonant transition between inductive and capacitive behavior for the 12-turns coil device. The resonance frequency is about 22.5 MHz.

The inductance of a spiral coil depends on the geometry of the winding and is proportional to the square of the number of turns, N^2 [12]. As the geometrical factors of the 12-turns and eight-turns coils are about the same, the inductance ratio between the two devices should be 2.25. An experimental value of 2.22 is obtained.

The resistances, as shown by Fig. 4(b), are constant over about three decades of frequencies, starting from 10^4 Hz (data not shown), with values of 3.6 Ω for the 12-turns coil and 1.3 Ω for the eight-turns coil. Then, they increase for frequencies greater than 1 MHz and reach maximum values of 4.2 and 2 kΩ, respectively, for 12 and eight turns at the respective resonance frequencies.



(a)

(b)

Fig. 4. Frequency dependence of (a) the inductance and (b) the resistance of the inductive devices.

B. Leakage Inductance

The simplified equivalent circuit of the transformer is represented in Fig. 5. U_{source} and R_{source} are, respectively, the voltage and resistance of the source and C_p the capacitance of the primary. R_w is the total winding resistance and $L_{l,p}$ the total leakage inductance, i.e., the primary inductance with the secondary shorted. $L_{m,p}$ is the open circuit inductance and R_{loss} the shunt loss resistance, representing the core losses. As shown by the figure, it is possible to represent entirely the transformer with its primary circuit. C_s' and R_L' are the secondary capacitance and the load resistance, both referred to the primary. They are related to the true secondary capacitance C_s and the true load resistance R_L by (3) and (4). Finally, N_p and N_s are the number of turns of the windings of the primary, respectively, the secondary

$$C_s' = \left(\frac{N_s}{N_p}\right)^2 C_s \tag{3}$$

$$R_L' = \left(\frac{N_s}{N_p}\right)^{-2} R_L. \tag{4}$$

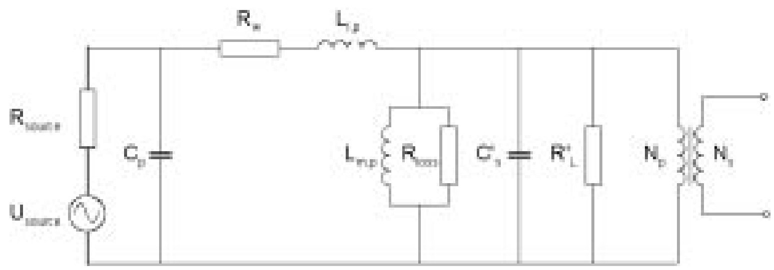


Fig. 5. Simplified equivalent circuit of the transformer.

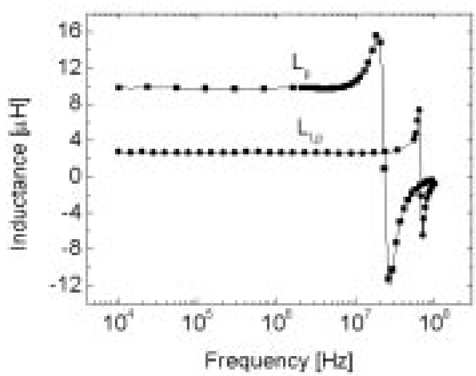


Fig. 6. Dependences of the inductances L_p and $L_{l,\,p}$ of the primary circuit of the mini-transformer versus the frequency.

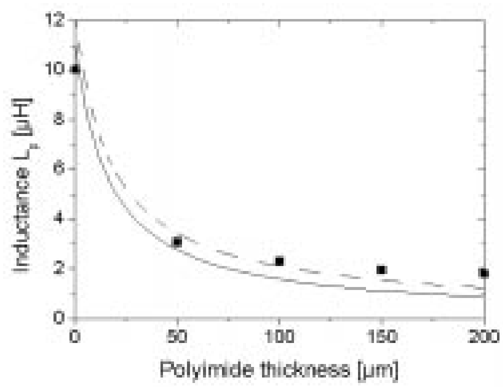


Fig. 7. Inductance L_p of the primary circuit versus air gap at 100 kHz for mini-transformers: measured values are represented by squares. Computed values without and with fringe correction factor are represented by the solid and dashed lines, respectively.

Thus, when the secondary is short circuited ($R'_L = 0$), the resonance frequency $f_{r_{l,\,p}}$ is given by

$$f_{r_{l,\,p}} = \frac{1}{\sqrt{L_{l,p}C_p}} \tag{5}$$

while, when it is open, the resonance frequency f_r is given by

$$f_r = \frac{1}{\sqrt{L_p(C_p + C'_s)}} \tag{6}$$

where the total measured inductance $L_p \cong L_{m,\,p} + L_{l,\,p}$.

The inductances L_p and $L_{l,\,p}$ of the primary circuit are reported in Fig. 6. The value of L_p is about 9.9 μH over more than two decades of frequencies and the resonance frequency is 23.4 MHz. The leakage inductance $L_{l,\,p}$ of about 2.7 μH over more than three decades of frequencies gives an acceptable main to leakage ratio of 2.7. The resonance frequency is about 66.4 MHz. Introducing these values in (5), (6), and (3), one gets $C_p = 85$ nF and $C_s = 220$ nF. We can note that, due to device assembly and measurement inaccuracies, the values of the capacitances are indicative and give at most the order of magnitude, i.e., both are in the order of 100 nF.

C. Influence of the Air Gaps

The gap lengths l_g between the two E-cores have been varied by adding interpolated layers of 50 μm thick polyimide films between them. As in the magnetic circuit, the gap does intervene two times, we have

$$l_g = 2(t + p) \tag{7}$$

where t is the thickness of the stack of polyimide layers and p the parasitic gap, depending on the roughness[1] R_z of the contact surfaces of the ferrite E-cores.

In the most simple theoretical approach, the main inductance of the primary circuit is given by

$$L_p = \frac{\mu_0 N_p^2 A_c}{l_g + \dfrac{l_m}{\mu_r}} \tag{8}$$

where

- μ_0 permeability of the free space;
- μ_r relative permeability of the ferrite;
- A_c cross-sectional area of the inner core leg;
- l_m magnetic path length.

$A_c = 2\ \text{mm}^2$, and $l_m = 6.5$ mm. In Fig. 7, the experimental (isolated points) and theoretical values (solid line) of L_p are reported versus t, for a frequency of 100 kHz. Measurement[2] of R_z gave a value of about 19 μm. As the two contact surfaces can overlap, a value for p of 14 μm has been considered for the computation.

The match between the experimental and theoretical curves is improved (dashed line) by introducing F, a correction factor with which (8) must be multiplied, given by [13]

$$F = 1 + \frac{l_g}{\sqrt{A_c}} \ln \frac{2G}{l_g} \tag{9}$$

[1]The maximum individual peak to valley height, cf. ISO 4287/1.

[2]The roughness has been measured with an instrument (noncontact measurement) of UBM Messtechnik GmbH.

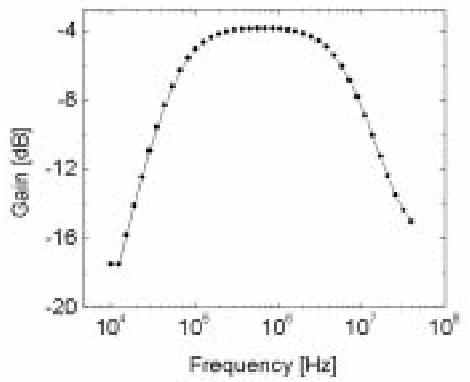


Fig. 8. Frequency dependence of transformation gain of the mini-transfomer.

where G is the vertical dimension of the core window (which is 1 mm in our case). The F factor expresses the fact that the flux does present fringing effect around the gaps.

D. Transformer Gain Measurements

The gain $g_{[\mathrm{dB}]}$ of the transformer is given by

$$g_{[\mathrm{dB}]} = 20 \log \frac{V_s}{V_p} = 20 \log \frac{N_s}{N_p} \tag{10}$$

where V_s and V_p are the secondary and primary voltages. As shown by Fig. 8, where the dependence of $g_{[\mathrm{dB}]}$ versus the frequency is reported, the gain of the mini-transformer, which is about -3.8 dB between 200 kHz and 2 MHz, does correspond approximately to the theoretical value of -3.5 dB given by the 8 : 12 ratio.

V. Conclusion

Using an innovative technique, i.e., the 3-D machining of brittle ferrites wafers by powder blasting, a novel mini-transformer has been realized. The wafer level process of both E-cores and coils open the way for further size reduction of the device's dimensions, with complete batch technology, including the assembly. The electrical characterizations that have been done clearly show that the mini-transformer is well suited for low power applications at frequencies in the order of 0.1 to 1 MHz.

Acknowledgment

The authors would like to thank M. Hermanjat and P. Vosseler, Department of Electrical Engineering, Swiss Federal Institute of Technology, Lausanne (EPFL) for their assistance in the realization of the flex foil PCBs.

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