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Loss analysis of Flyback in Discontinuous Conduction Mode for sub-mW harvesting systems

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Abstract—The energy harvesting solution permits to autonomously supply low power sensor nodes without chemical battery allowing their massive spreading in various environments. The electrical interface between the harvester and the sensor is a crucial step to maximize the harvested energy and boost its voltage to a minimum value required by the sensor. To achieve input impedance and voltage gain independently, this paper presents a flyback converter in discontinuous mode. Thanks to a proposed flyback model validated experimentally, we have studied the impact of each loss source in order to give some trade-off for designing an efficient sub-mW harvesting interface. We underline the effect of the transformer losses due to the magnetic hysteresis as well as the driving loss impact. Following this method a flyback prototype achieves 71\% power efficiency when harvesting from a microbial fuel cell delivering 90\mu W.

Keywords—flyback design method; energy harvesting; DC-DC converter; maximum power point tracking; microbial fuel cell

I. INTRODUCTION

Harvesting energy in the surrounding environment is an advantageous alternative to conventional batteries for powering autonomously remote sensors in addition to processing in an eco-friendly way. Many researches currently focus on harvesting energy from solar, thermal and vibrational sources scavenged in environments near the sensor. Less analyzed in the literature, the microbial fuel cell (MFC) is an emerging harvesting technology that exploits the waste materials around the sensor. The catalysis properties of bacteria into a couple of redox reactions convert chemical energy from a large range of carbonate substrates such as seafloor sediment or compost into electrical energy [1,2]. This energy source production is relatively robust and low-cost but the generated power is around 100\mu W for cm-scale electrodes and its DC voltage is not sufficient e.g. up to 0.7V to power continuously low-power sensor nodes.

To adapt and store the power generated by the harvester to the sensor, a harvesting interface is required i.e. a DC-DC converter. It has a double objective i) extract the maximum power from the harvester and ii) boost and regulate the voltage in an intermediate energy storage. Then the sensor switches between on and off-states depending on the energy available in this storage. The boost topology is commonly chosen for the harvesting interface [3,4]. However, this architecture suffers from an inherent limitation: the maximal power extraction point and voltage gain cannot be both satisfied in one conversion stage even in discontinuous conduction mode. To overcome this limitation, [4] adds a second stage to adapt the voltage gain independently to the input impedance. But this two-stage conversion topology limits the achievable efficiency.

[5] proposes to use a flyback in discontinuous conduction mode (DCM) to overcome the limitation of the classical boost converter topology and also offer a galvanic isolation adding value in some MFC applications. The work was achieved for an input power of 10mW. This paper proposes to study this topology for sub-mW power range and present a design methodology by analyzing each power loss source especially the transformer. The first section presents briefly the harvester electrical model and explains the flyback operation in DCM to show its ability to adapt its input impedance and voltage gain independently. Then, the trade-off for maximizing the power extraction from a sub-mW power source and optimizing the converter efficiency is given. To help the designer on their choice e.g. transistor sizing, transformer or other components choices, a complete model of the flyback has been introduced and experimentally validated.

II. FLYBACK CONVERTER FOR MFC ENERGY HARVESTING

A. Harvester electrical model and MPPT

The solar, thermal or bio fuel cell harvesters are often modeled by a voltage source \( V_S \) and a series resistance \( R_S \) (Fig. 1) when operating close to their maximum power point MPPT [6]. Identifying these two parameters is a crucial step to determine the impedance value \( R_{IN} \) of the harvesting interface and optimize the power extraction from the harvester. In fact, the power received by the harvesting interface is maximized when \( R_{IN} \) is equal to \( R_S \) and is expressed at MPP as:

\[
\eta_{\text{MEP}} = \frac{V_S^2}{4R_S} \tag{1}
\]

We define the extraction efficiency \( \eta_{\text{MEP}} \) the ratio between the power delivered to the harvesting interface \( P_{IN} \) and the maximum power the MFC can deliver \( P_{MPP} \). \( \eta_{\text{MEP}} \) is equal to unity when the impedance matching is respected. In the case of our MFC prototypes [7], the MPP is of 90\mu W at 0.3V and their static behavior can be modeled with \( V_S=0.6V \) and \( R_S=1k\Omega \).

The power generated by the MFC is not directly useable to power continuously low-power sensor node. A harvesting interface is necessary to boost the harvester output voltage \( V_{IN} \) to a minimum voltage \( V_{OUT} \) required to supply the sensor.
node. The chosen harvesting interface has to present a global efficiency close to unity regarding the low power at stake. This global efficiency includes the extraction efficiency $\eta_{\text{extr}}$ and the PMU conversion efficiency $\eta_{\text{conv}}$ (equation (2), Fig. 1) where the conversion efficiency $\eta_{\text{conv}}$ is the ratio between the power delivered by the harvesting interface $P_{\text{OUT}}$ and $P_{\text{IN}}$.

$$\eta_{\text{extr}} \times \eta_{\text{conv}} = \frac{P_{\text{IN}}}{P_{\text{MPP}}} \times \frac{P_{\text{OUT}}}{P_{\text{MPP}}} = \frac{P_{\text{OUT}}}{P_{\text{IN}}}$$

Moreover, since the MFC performance depends on the environment and $R_S$ varies, the MPP has to be regularly recorded to allow a dynamic impedance matching i.e. to adapt continuously the impedance $R_{\text{IN}}$ of the harvesting interface.

At the output the energy is stored in a capacitance $C_{\text{OUT}}$ and is intermittently delivered to the sensor thanks to a hysteresis comparator shown Fig. 2, making the output voltage oscillate between two values $V_{\text{OUT_MAX}}$ and $V_{\text{OUT_MIN}}$ where $V_{\text{OUT_MIN}}$ is the minimal voltage needed by the sensor. As the switch is opened, energy is getting stored in $C_{\text{OUT}}$ and $V_{\text{OUT}}$ increases. When it reaches $V_{\text{OUT_MAX}}$ the switch closes until $V_{\text{OUT}}$ reaches $V_{\text{OUT_MIN}}$, $C_{\text{OUT}}$ is chosen so that the amount of energy stored in one cycle corresponds to the energy $E_{\text{sensor_cycle}}$ required by one complete cycle of the sensor (equation (4)).

$$C_{\text{OUT}} = \frac{2E_{\text{sensor_cycle}}}{V_{\text{OUT_MAX}}^2 - V_{\text{OUT_MIN}}^2}$$

C. Origin of the flyback power losses

Determining the origin of the power losses in the converter is a very important step. Regarding the working conditions and electrical devices, the conversion efficiency can be disastrous especially with the power level delivered by our MFCs (less than 100μW). The different power losses of the flyback due to the MOSFET and the diode are expressed in Table. 1 considering an ideal transformer with a conversion ratio 1:1. The MOSFET presents an on-state resistance $R_{\text{ON}}$ causing conduction losses during the phase 1 and an internal capacitance $C_{\text{oss}}$ causing switching losses. The diode presents a threshold voltage $V_D$ causing conduction losses during phase 2 and a parasitic capacitance $C_D$. Less studied the transformer introduces non-negligible losses in the flyback especially in sub-nW operation. In the next section, the transformer losses will be modeled in a compact electrical circuit allowing giving some trade-off between all the flyback losses.

| TABLE I. FLYBACK POWER LOSSES |

<table>
<thead>
<tr>
<th>Conduction losses</th>
<th>Switching losses</th>
</tr>
</thead>
<tbody>
<tr>
<td>MOSFET</td>
<td></td>
</tr>
<tr>
<td>$R_{\text{ON}} \frac{V_D^2}{3DR_S^2}$</td>
<td>$\frac{1}{2}C_{\text{oss}} \frac{V_D^2}{2} + V_{\text{OUT}}^2 f$</td>
</tr>
<tr>
<td>Diode</td>
<td></td>
</tr>
<tr>
<td>$\frac{V_DV_S^2}{4V_{\text{OUT}}R_S}$</td>
<td>$\frac{1}{2}C_D \frac{V_S^2}{2} + V_{\text{OUT}}^2 f$</td>
</tr>
</tbody>
</table>
III. FLYBACK DESIGN AND MODELLING

A. Components choices

The MOSFET choice is an important step because it can induce conduction losses with $R_{ON}$ and switching losses with $C_{OSS}$. Reducing one (for instance reducing $R_{ON}$ by increasing the drain-source channel width) generally increases the other (increasing $C_{OSS}$). N-channel MOSFET FDV301N [8] was confirmed to be a good tradeoff for sub-mW operation. Its threshold gate voltage is 1V and it presents a capacitance $C_{OSS}$ of the order of 90pF, a $R_{ON}$ of 3.5Ω and a total gate charge $Q_{G}$ of 150pC when operating with a $V_{GS}$ equal to 1.5V.

The BAT54 diode [9] has been chosen because of its threshold voltage lower than 0.3V and low parasitic capacitance $C_{D}$ of 10pF, thus minimizing the conduction losses in the secondary branch of the flyback as well as the switching losses.

B. Parameters choices

The input capacitance $C_{IN}$ is used to maintain a DC voltage at the input of the flyback. According to equation (5), its value has to be sufficiently important in order to assure a negligible input ripple $\Delta V_{IN}$. We choose to set $\Delta V_{IN}$ equal to 1% of $V_{IN}$.

$$C_{IN} = \frac{V_{IN}}{\Delta V_{IN}} \times \frac{(2 - D)^2}{4R_{SF}} = 100 \times \frac{(2 - D)^2}{4R_{SF}} \tag{5}$$

The output voltage is set to oscillate around 1.8V with a $\Delta V_{OUT}$ of 0.1V.

Supposing the transformer conversion ratio equal to 1, the duty cycle D has to be minimized to keep the flyback in DCM as explained with equation (6), and maximized to avoid an important $L_{1\text{MAX}}$ that may drive the transformer to magnetic field saturation and also induce large conduction losses in the switch. Setting the duty cycle D to 0.5 is a fair tradeoff.

$$\frac{D}{1 - D} \leq \frac{1}{\sqrt{\eta_{\text{conv}}}} \times \frac{V_{OUT}}{V_{IN}} \tag{6}$$

The switching frequency $f$ and the transformer primary inductance $L_1$ offer a certain degree of liberty. To respect the MPP condition given by equation (3), the $\{L_1;f\}$ couple is fixed i.e. increasing $L_1$ means decreasing $f$. Considering the MFC resistance obtained section 2 ($R_{MFC}$=1kΩ), the influence of $L_1$ i.e. $f$ on the conversion efficiency $\eta_{\text{conv}}$ of the flyback converter was evaluated, without considering the driving loss. The result is presented by the blue curve in Fig. 3. When the inductance is too small (i.e. the frequency is high), the switching losses mainly due to the MOSFET parasitic capacitance prevail and highly degrade the flyback efficiency. This result encourages the choice of a frequency close to zero. However a tradeoff has to be made to also avoid a too large transformer in order to limit the circuit to a proper size.

This first study only considers the MOSFET and diode losses. We will now focus on describing the parasitic aspects that can encounter a real transformer and the impact it can have on the conversion efficiency $\eta_{\text{conv}}$.

![Fig. 3. Influence of the transformer primary inductance on the flyback conversion efficiency when working at the MPP with an input voltage of 1.8V and without considering the driving losses.](image)

![Fig. 4. Transformer electrical model.](image)

C. Transformer modelisation

The choice of $L_1$ is crucial. Regarding the variation of the harvester resistance $R_{H}$, it will determine the frequency range according to the MPPT strategy and thus will greatly influence the conversion efficiency $\eta_{\text{conv}}$. In order to understand its impact on the conversion efficiency $\eta_{\text{conv}}$, an electrical model is required. Fig. 4 represents the equivalent electrical circuit of a 1:1 transformer with a primary inductance $L_1$. The copper losses in the primary and secondary side are modeled with $R_1$ and $R_2$, the core losses mainly due to the hysteresis of the magnetic material with $R_p$, the leakage currents with $L_p$, the inter-winding capacitances in the primary and secondary with $C_1$ and $C_2$, and the capacitance between the primary and the secondary with $C_s$.

Two transformers of the same manufacturer [10] that are typically used for energy harvesting applications were characterized in order to compare on the same basis two transformers with different $L_1$ leading to two switching frequencies. The transformers were characterized using a network analyzer with a frequency range of [5Hz, 30MHz]. The values obtained for both of the characterized transformers

<table>
<thead>
<tr>
<th>Ref</th>
<th>Transformer 1</th>
<th>Transformer 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Lp (mH)</td>
<td>78601/9C1</td>
<td>78601/9C1</td>
</tr>
<tr>
<td>Lp (H)</td>
<td>1.8</td>
<td>1.8</td>
</tr>
<tr>
<td>Rp (Ω)</td>
<td>0.35</td>
<td>1.1</td>
</tr>
<tr>
<td>Rp (Ω)</td>
<td>0.35</td>
<td>1.1</td>
</tr>
<tr>
<td>Rp (Ω)</td>
<td>6K</td>
<td>30K</td>
</tr>
<tr>
<td>Rp (Ω)</td>
<td>3</td>
<td>10</td>
</tr>
<tr>
<td>Rp (Ω)</td>
<td>3</td>
<td>10</td>
</tr>
<tr>
<td>Rp (Ω)</td>
<td>28</td>
<td>100</td>
</tr>
</tbody>
</table>

![Table: Transformer parameters](image)
were evaluated thanks to a modeling strategy described in [11].

IV. EXPERIMENTAL VALIDATION

A. Transformer model validation and transformer losses

The flyback was simulated with adding the model of each transformer and compared to the experimental data. The impedance matching was accurately achieved i.e. the extraction efficiency $\eta_{\text{extr}}$ is equal to unity. The conversion efficiencies are represented Fig. 3 and highlight a correct fit between the simulated performances and the ones acquired experimentally. Our flyback electrical models can thus be considered trustable.

Considering the previous results with an ideal transformer (blue curve in Fig. 3), the addition of the transformer parasitic elements considerably degrades the converter conversion efficiency. Then we underline the transformer could be the bottleneck to increase the harvested power efficiency. A previous study on the impact of each transformer parasitic elements revealed that the parallel resistance $R_p$ representing the magnetic loss is inducing the majority of the transformer losses. This explains the greater losses of the 1.8mH transformer presenting a parallel resistance five times smaller (6kΩ) than the 18mH transformer (30kΩ).

B. Driving losses

To assure a self-sufficient process, part of the flyback output power has to be used to supply the MOSFET driver. The power used to supply the sensor is thus $P_{\text{OUT}} - P_G$ where $P_G$ is the power consumed by the MOSFET driver expressed by:

$$P_G = Q_g V_g f$$

(7)

We define the efficiency $\eta_{\text{Supply}}$ expressed by equation (8), as the ratio between the power available to supply the sensor $P_{\text{OUT}} - P_G$ and the maximum power delivered by the MFC $P_{\text{MPP}}$.

$$\eta_{\text{Supply}} = \frac{P_{\text{OUT}} - P_G}{P_{\text{MPP}}} = \frac{\eta_{\text{extr}} \eta_{\text{conv}} P_{\text{MPP}} - P_G}{P_{\text{MPP}}}$$

(8)

Fig. 5 shows the influence of the driving losses on the power efficiency. The blue curve represents the conversion efficiency $\eta_{\text{conv}}$ previously shown in Fig. 3 without considering $P_G$. The yellow curve represents $\eta_{\text{Supply}}$ with ideal transformer that includes this additive driving losses $P_G$ and highlights the available power given to the sensor. The smaller is $L_1$, the higher is $f$ and the higher are the driving losses $P_G$.

The red dots are the supply efficiency $\eta_{\text{Supply}}$ simulated with the two previous characterized transformers models and the green dots are the data acquired experimentally which correctly fit the simulation. Using the 18mH transformer $\eta_{\text{Supply}}$ reaches 71%. The MPPT is respected and $\eta_{\text{extr}}$ is equal to unity. Considering an input power of 90μW, the power useable by the sensor is 64μW. The transformer induces almost 50% of the total losses. Hence to further enhance the flyback performance, the design has to focus on choosing an adapted transformer with very few magnetic losses with a reasonable switching frequency.

V. CONCLUSION

In this paper we analyzed in details the losses of the flyback working in DCM for sub-mW harvesting applications. Thanks to a proposed flyback model validated experimentally, we highlighted the impact of the transformer iron losses on the power efficiency and the need to carefully choose the transformer. By choosing the good tradeoff between the switching frequency and transformer, a prototype was able to transfer 71% of the 90μW maximum power available in the harvester to the sensor with an output voltage of 1.8V.

To go further in the analysis, other transformers will be studied later on.

REFERENCES


