Experimental Characterization of Wearable Devices for RF Energy Harvesting in WBANs

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Abstract—In this paper, we identify the spectrum opportunities for radio frequency (RF) energy harvesting through power density measurements from 350 MHz to 3 GHz. The field trials have been performed in Covilhã by using the NARDA-SMR spectrum analyser with a measuring antenna. Based on the identification of the most promising opportunities, a dual-band printed antenna operating at GSM bands (900/1800) is proposed, with gains of 1.8 and 2.06 dBi, and efficiency varying from 77.6 to 82%, for the highest and lowest operating frequency bands, respectively. Guidelines for the design of RF energy harvesting circuits and choice of textile materials for a wearable antenna are also discussed. Besides, we address the development and experimental characterization of three different prototypes of a five-stage Dickson voltage multiplier (with and without match impedance circuit) responsible for harvesting the RF energy. All the three prototypes (1, 2 and 3) can power supply the sensor node for RF received powers of 2 dBm, -3 dBm and -4 dBm, and conversion efficiencies of 6, 18 and 20%, respectively.

I. INTRODUCTION

Future improvements in radio frequency (RF) energy harvesting technology will facilitate the creation of a network with no need of dedicated transmitters, as a reliable source of wireless energy power [1]. This can be accomplished by enabling the capture of electromagnetic energy from multiple available ambient RF energy sources, such as mobile base stations, TV and radio transmitters, microwave radios, and mobile phones. Moreover, since wireless body area network (WBAN) nodes are battery operated, energy recharging is a possibility, avoiding the need of battery replacement. However, the service lifetime of the electronic components could be a major concern if there is no possibility to collect enough energy to generate the voltage needed to drive the sensor node. Medium access control (MAC) and routing protocols also play an important role in the network performance [2]. As a consequence, choosing the best opportunities poses a high effect on the overall network performance, as well as on the energy consumption. In the context of WBANs, electromagnetic RF energy harvesting is accomplished by using wearable antennas that allow for power supplying the sensor nodes [3]. Ubiquitously available RF sources, operating at different bands, are therefore exploited for RF electromagnetic energy harvesting purposes.

In this work, we have identified the opportunistic radio frequency bands for RF energy harvesting. Moreover, based on power density measurements, we have been able to identify the best spectrum opportunities that may be considered in order to conceive multiband antennas for electromagnetic energy harvesting. Besides, we present the guidelines for the development of wearable flexible antennas and design of circuits to harvest RF energy. The RF energy harvesting system to be developed in the context of the PROENERGY-WSN project [4] consists of an impedance matching circuit, rectifier and energy storage.

The remainder of the paper is organized as follows. Section II addresses the spectrum opportunities based on the field trials that were held in Covilhã, Portugal. Section III describes an efficient dual-band antenna for collecting RF energy. The guidelines for the choice of textile materials to be used in future wearable antennas are also presented. Section IV describes the fundamentals of RF energy circuits. Section V presents the simulation and experimental results of the three prototypes (with and without match impedance circuit) of the 5-stage Dickson voltage multiplier followed by the discussion of the results. Finally, conclusions are drawn in Section VII.

II. INDOOR AND OUTDOOR SPECTRUM OPPORTUNITIES

In order to seek the best spectrum opportunities for RF energy harvesting, we have conducted several field trial measurements in Covilhã, Portugal, in both indoor and outdoor environments, by using the NARDA-SMR spectrum analyser with a measuring antenna.

A. Average Received Power

By analysing the power density measurements in 36 different locations, we intend to find the best frequencies for RF energy harvesting. Besides, the identified spectrum opportunities are being considered to conceive multi-band antennas. The location for the measurements is shown in Fig. 1. To determine the received power, \( P_r \), of the spectrum analyser, we multiply the power density, \( P_d \), by the effective receiving area of the antenna, \( A_e \), and gain, \( G \), as follows:

\[
P_r = \frac{|E|^2}{120 \cdot \pi}
\]

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where $E$ is the electric field and $\lambda$ is the wavelength.

\[
P_r[\text{dBm}] = 10 \cdot \log \left( \frac{P_a \cdot \lambda^2 G}{4 \cdot \pi} \right) + 30 \tag{2}
\]

To choose the best frequency bands for electromagnetic energy harvesting, we have determined the average of each of the $n$ values of the received power, $P_r [\text{W}]$ in linear units, in five different locations, where $n$ is the number of measurements taken, for each frequency. The average received power, in $\text{dBm}$, is given as follows:

\[
\overline{P_r[\text{dBm}]} = 10 \cdot \log \left( \frac{\sum_{i=1}^{n} P_{ri}[\text{W}]}{n} \right) + 30 \tag{3}
\]

### B. Indoor Opportunities

Figure 2 presents the indoor spectrum opportunities for the higher education institution in Covilhã, the set of frequencies with high energy available for harvesting comprises the range from 934 to 960 MHz (GSM900), 1854 to 1892 MHz (GSM1800), 2116 to 2160 MHz (UMTS), 2359 MHz (amateur, SAP/SAB applications, video), and 2404 to 2468 MHz (Wi-Fi).

### C. Outdoor Opportunities

The location of public places in the outdoor scenario for the field trial results are identified in Fig. 1, for the locations numbers 8, 9, 12, 13, 14, 21 and 22. The corresponding values of the average received power are shown in Fig. 3. The set of frequencies with more energy available for harvesting are in the range from 79 to 96 MHz (mobile/radio broadcast stations), 391 MHz (emergency broadcast stations), 750 to 759 MHz (digital television broadcast stations), 935 to 960 MHz (GSM 900 broadcast stations), 1854 to 1870 MHz (GSM 1800 broadcast stations) and 2115 to 2160 MHz (UMTS broadcast stations).

### III. ANTENNA FOR RF ENERGY HARVESTING

#### A. Dual-band antenna

In the previous section we have identified the GSM (900/1800) as the most promising bands for RF energy harvesting. In this section, we propose a dual-band antenna suitable for possible implementation within clothes for body worn applications based on spectrum opportunities, as shown in Fig. 4. Table I presents the corresponding antenna dimensions [5].

#### Table I. Proposed Dual-band Antenna Dimensions

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Dimension [mm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L$, $L_{gd}$, $L_f$, $L_{fs}$</td>
<td>120, 100, 78, 30</td>
</tr>
<tr>
<td>$L_{m1}$, $L_{m2}$, $L_{m3}$, $\text{gap}$</td>
<td>21, 43, 31, 3</td>
</tr>
<tr>
<td>$W$, $W_f$, $W_{m1}$, $W_{m2}$, $W_{m3}$</td>
<td>80, 1.5, 12, 5, 8</td>
</tr>
</tbody>
</table>

The Cordura® fabric type was considered for the dielectric substrate, presenting a permittivity, $\varepsilon_r$, of 1.9, loss tangent, $\tan \delta$, of 0.0098 and relative height of 0.5 mm. For the conductive sections of the antenna an electrotextile (Zelt), with an electric conductivity $1.75 \times 10^5 \text{ S/m}$ was considered. The variation of the...
return loss obtained from numerical simulations and measurements is presented in Fig. 5 for the proposed dual-band antenna, considering a return loss $S_{11} < -10$ dB.

The lowest frequency band considered is from 820 MHz to 1000 MHz, covering the entire E-GSM band, while the highest considered band is in the range 1690 MHz to 1930 MHz, covering the entire DSC1800 band. It is worth noting, however, that the radiation pattern for the dual-band antenna suffers a deformation at 1800 MHz. The obtained radiation pattern in the YZ and XZ planes for the proposed antenna (where dimensions are presented in Fig. 4), is based on numerical simulation and is shown in Fig. 6.

![Simulated and measured return loss for the proposed dual-band antenna.](image)

Figure 5. Simulated and measured return loss for the proposed dual-band antenna.

![Simulated radiation pattern for the dual-band antenna in the YZ plane (dashed) and XZ plane (blue solid), at a) 900 MHz and b) 1800 MHz.](image)

Figure 6. Simulated radiation pattern for the dual-band antenna in the YZ plane (dashed) and XZ plane (blue solid), at a) 900 MHz and b) 1800 MHz.

The gains for the dual-band antenna are about 1.8 dBi and 2.06 dBi, with a radiation efficiency of 82% and 77.6% for the lowest and highest operating frequency bands, respectively.

**B. Guidelines for the development of wearable flexible antennas**

In the near future, people might contain upon himself a wide range of devices and sensors embedded in clothing. Thus, the new generation of garments will be able to perform a continuous monitoring in various fields, such as health and sport [6], communicating the data generated to remote receivers. Wearable antennas are the link that integrates cloths in the communication system, making the integration of electronic devices more discrete and less intrusive.

Specific requirements for wearable antennas are the planar structure and flexible materials. Textiles, being universally used and easily available, are appropriate materials to design wearable antennas. However, some characteristics of the materials highly influence the performance of the antenna, e.g., the dielectric constant and the loss tangent, for a specific frequency, the thickness, the moisture content in a specific environment (Regain), as well as the geometrical and mechanical stability of the fabric. In general, ordinary textiles present a very low dielectric constant, in the range between 1 and 2 (as they are very porous). However, textile materials are constantly exchanging water molecules with the environment which might increase their dielectric constant. Therefore, fabrics for dielectrics should present a low Regain. For the conductive components of planar antennas, specific electrical conductive textiles have been successfully used, and are available on the market.

The choice of the textile materials for the antenna design impacts on the performance, where the construction of the antenna assumes a particular importance because textiles are highly deformable materials. In [7], the authors have shown that the assembly of the conductive patch with the dielectric substrate is critical, and the geometrical dimensions should remain stable when connecting these two components. The mechanical stabilization of both materials is thus essential to preserve the desired characteristics of the antenna. Therefore, is important, that at least one of the components, the dielectric substrate or the patch, is made of highly rigid materials as for instance a rigid woven made of high tenacity fibres [7]. Furthermore, the cut of the materials must be very accurate. As the conductive material is thinner than the dielectric, its cut is critical. Therefore it is advisable to use a woven as patch rather than knit. In the woven one can follow the direction of the yarns, which facilitates the cutting accuracy [7]. Also, for the very narrow parts, the cut should be made in the transverse direction of the fabric avoiding the fraying of the yarns.

When choosing the connection technique between the various layers one needs to be very careful because it should not affect the electrical properties of the patch, particularly its electrical resistance. The authors from [7] have shown that adhesive sheets enable to achieve very promising results.

**IV. RF ENERGY HARVESTING CIRCUITS**

One of the main factors that influence RF energy harvesting is the path loss. The Friis free-space equation relates the received power, $P_r$, at a distance, $d$, with the transmitted power, $P_t$, as follows:

$$P_r = P_t \cdot G_t \cdot G_r \cdot \left(\frac{\lambda}{4\pi d}\right)^2$$

(4)

where, $G_t$ and $G_r$ are the antenna gains of the transmitter and receiver, respectively. Based on equation (4), we can observe that, the received power depends on the frequency (the higher the frequency is the lower the received power is), and decreases with the square of the distance (path-loss exponent equal to 2). To conceive an RF energy efficient rectifier which enables to rectify and amplify the input voltage (corresponding to $P_r$), the Cockcroft–Walton and Dickson voltage multiplier circuits can be considered, as shown in Figs. 6 a) and b). This work only addresses the Dickson voltage multiplier, as shown in Fig. 7, since, according to [8], both topologies have a similar performance.
In order to have a self-sustainable WBAN, energy efficient harvesting and management techniques must be considered.

However, one of the drawbacks of wireless sensor devices is the finite battery capacity, as well as the voltage generated by the RF energy harvesting circuit, since it may be insufficient to drive the mote(s) (at least 1.8V is needed for the IRIS mote). By means of simulation, we have addressed the main aspects and parameters that influence the performance of the Dickson voltage multiplier, e.g., the choice of diodes, the number of stages and the load impedance. We have used the Advanced Design System (ADS) [9] from Agilent and varied the RF received power, $P_{RF}$, from -50 dBm to 20 dBm. The considered centre frequency is equal 945 MHz since, in this work, we have identified the GSM 900 band as the most promising one for RF energy harvesting. The Dickson voltage multiplier presented in Fig. 7 b) is mainly formed by diodes and capacitors in parallel (instead of being in series, as for the Cockcroft–Walton voltage multiplier). As the input peak voltage from the antenna signal is usually much lower than the diode forward conduction voltage [8], diodes with low turn-on voltage have been considered. Therefore, we considered HSMS-2850 Schottky diodes from Avago Technologies optimized for low power applications. The number of rectifier stages has a major impact on the output voltage of the Dickson voltage multiplier. According to [10], the output open circuit voltage (OCV) of an N-stage Dickson voltage multiplier is given as follows:

$$V_{out} = 2 \cdot N \cdot (V_{in} - V_f)$$  \hspace{1cm} (5)

where $N$ is the number of stages, $V_{in}$ is the input voltage amplitude and $V_f$ is the forward conduction voltage of diodes.

As shown in equation (5), the output voltage is directly proportional to the number of stages. However, practical restrictions (e.g., conversion efficiency) limit the number of permissible stages. The conversion efficiency, $\eta_c$, of a Dickson voltage multiplier is responsible for providing a representation of the overall performance of the circuit, and defines the relationship between the output DC power, $P_{DC}$, and the RF received power, $P_{RF}$, as follows:

$$\eta_c[\%] = \frac{P_{DC}}{P_{RF}} \cdot 100 \hspace{1cm} (6)$$

where, the output DC power is given by:

$$P_{DC} = V_{DC} \cdot I_{DC} \hspace{1cm} (7)$$

To show the impact of number of stages on conversion efficiency and output voltage in terms of simulation, we have also used the ADS software.

V. SIMULATION AND EXPERIMENTAL RESULTS

A. With no Impedance Matching Circuit

Figures 8 and 9 analyse the impact of the number of stages (3, 5 or 8) on the output voltage and conversion efficiency of the Dickson voltage multiplier, by assuming a load impedance of 100 kΩ. These simulation results were obtained through a harmonic balanced analysis (i.e., a frequency domain method) that evaluates the steady state solution of a nonlinear circuit.

![Figure 8](image8.png)

Figure 8. Impact of the number of stages on the output voltage for an N-stage Dickson voltage multiplier with a load impedance of 100 kΩ.

![Figure 9](image9.png)

Figure 9. Impact of the number of stages on the conversion efficiency for an N-stage Dickson voltage multiplier with a load impedance of 100 kΩ.

By analysing Fig. 8, we conclude that, by increasing the number of stages, we increase the output voltage. The saturation is theoretically obtained by multiplying the number of stages by the reverse breakdown voltage (i.e., 3.8V for HSMS-2850). By considering the number of stages equal to 3, 5 or 8, the maximum output voltage obtained by simulation is approximately 11, 18 and 29 V, respectively, which corresponds to RF received powers of 10, 12 and 15 dBm.

Figure 9 presents the effect of the RF received power on the conversion efficiency. By adding more stages, the peak of the conversion efficiency curve shifts toward the higher received power region, similarly to the results from [11]. As a consequence, we have chosen the 5-stage Dickson voltage multiplier as the best circuit for RF energy harvesting for...
WBANs. This is explained by the fact that more than 5 stages will not bring substantial improvement for the power levels considered, due to energy losses along the chain [8]. Moreover, since the wireless sensor nodes need at least 1.8 V for operation (i.e., approximately -10 dBm RF input power from simulated results), the 5-stage Dickson voltage multiplier is the one which presents the best performance in terms of conversion efficiency. Based on the previous conclusions, we have decided to conceive a 5-stage Dickson voltage multiplier Prototype 1 in a printed circuit board (PCB) fabricated with two layers by using a FR-4 epoxy glass substrate. The performance was evaluated by using the E8361C PNA Microwave Network Analyser and the Agilent E4433B Signal Generator.

By comparing the simulation (Sim.) and the experimental (Exp.) results, we conclude that there are an average deviation of 59% between the simulation and experimental results for the output voltage, whereas the experimental saturation value occurs at the RF input received power of 16 dBm. Besides, there is an average deviation of 75% between the simulation and experimental results for the conversion efficiency. The maximum conversion efficiency is approximately 8%. By analysing the experimental and simulation results we may conclude that the deviations for the output voltage and conversion efficiency can be explained by the impedance mismatching between the antenna and the 5-stage Dickson voltage multiplier prototype, as shown in Fig. 11 from [12]. The prototype measured impedance is $Z_{in} = 262.49 + j401.65 \Omega$ at 945 MHz. In Fig. 11 from [12], it is noticeable that the measured return loss coefficient, $S_{11}$, is extremely high. As so, it will induce significant amount of reflection. Therefore, there is a need to design a match impedance circuit to cancel the losses due to reflection. One solution is to use a single stub matching circuit for improving both the output voltage and conversion efficiency, like in [13], as discussed in next section.

B. With Impedance Matching Circuit

In [12], the impact of the load impedance on the conversion efficiency of the 5-stage Dickson voltage multiplier is discussed by analysing its dependence on the RF received power through simulations. Based on the simulation results, we are able to conclude that the optimal conversion efficiency is achieved when the load impedance is 50 kΩ. If the resistive load value is too low or too high, the conversion efficiency significantly decreases.

In WBAN the nodes equivalent impedance is different for each radio operation state (RX, TX and SLEEP). According to [12], the value of the impedance from Mica2 mote in the deep sleep state is 100 kΩ. This value is considered as the impedance reference load for the IRIS motes.

The match impedance circuit is designed for the aforementioned impedance reference load of 100 kΩ. Two prototypes (2 and 3) were developed after testing Prototype 1. Prototypes 2 and 3 are fabricated in the same special RF substrate (i.e., RO4003 from Roger). Prototype 2 is composed by a 5-stage Dickson voltage multiplier alone, whereas Prototype 3 includes a 5-stage Dickson voltage multiplier along with the match impedance circuit as shown in Fig. 10. The development of Prototype 3 was developed with a match impedance circuit that takes into account the measured impedance from Prototype 2, i.e., $Z_{in} = 130 + j10.5 \Omega$ at 945 MHz. Thus, Prototype 3 presents a measured impedance of $Z_{in} = 34.9 + j1.66 \Omega$ at 945 MHz.

![Prototype 3 constituted by impedance matching and five stages Dickson voltage multiplier circuit.](image)

Fig. 11 presents the comparison between the simulated and experimental results (of the three prototypes) for the output voltage of the 5-stage Dickson voltage multiplier as a function of the RF received power. In the envisaged scenario we consider two levels for the output voltage: the minimum voltage to power supply the sensor node (i.e., 1.8 V) and the advised voltage to power supply the sensor node (i.e., 3 V), which are presented in Fig. 11 by orange lines.

![Impact of the RF received power on the output voltage for the 5-stage Dickson voltage multiplier with a load impedance of 100 kΩ.](image)

By observing Fig. 11 we conclude that Prototype 1 presents an average deviation from the simulated values of around 59%, whereas the saturation zone occurs at an RF received power of 16 dBm. Both prototypes (2 and 3) present a similar behaviour, but Prototype 3 shows better results compared to Prototype 2 because it benefits from the impedance matching circuit. By comparing the results obtained for Prototypes 2 and 3, the average deviation in terms of output voltage is approximately 24%. In Prototype 3, the saturation zone is not achieved, since previous tests have shown that to increase the RF received power (in the signal generator) to levels higher than 0 dBm may cause burning of the electronic components from the prototype. This is the reason why we present limited experimental values for Prototype 3. In turn, for Prototype 2 we achieved the saturation zone at 15 dBm, since the match impedance circuit is not considered.

Fig. 12 shows the impact of the RF received power on the conversion efficiency for the three prototypes, with a load impedance of 100 kΩ. By analysing Fig. 12, on the one hand,
we verify that Prototype 1 presents an average deviation between the simulated and experimental results (in terms of conversion efficiency) of around 75%, leading to a maximum conversion efficiency of 8% (for an RF received power of 3 dBm). On the other hand, Prototypes 2 and 3 present an average deviation of 40%, matching a maximum conversion efficiency of 22% (for an RF received power of 0 dBm).

The experimental results are within the simulated results, but we expected that Prototype 3 would show higher conversion efficiency results, since it considers an impedance matching circuit, compared to Prototype 2. By comparing the results between Prototype 1 and Prototype 3, the latter one shows an increase of around 15% in the conversion efficiency when considering the highest peak value for each curve. This increase in the conversion efficiency from Prototype 3 is explained by use of a specific substrate for RF applications and by the adaptation of the length of the transmission microstrip lines that allowed optimizing the match impedance.

![Figure 12](image)

Figure 12. Impact of the RF received power on the conversion efficiency ($\eta_0$) for a 5-stage Dickson voltage multiplier with a load impedance of 100 k\(\Omega\).

The deviations between the simulation and experimental results are explained by the PCB manual manufacturing techniques that we employed to develop each one of the prototypes or by not knowing the real parameter values from the diodes used in the prototypes. Also, in the simulations we consider the values supplied by the manufacturer which could induce some error in the simulation results presented.

By considering the aforementioned scenario in which the advised voltage to power supply a sensor node is considered, i.e., 3 V, Prototype 1 can power supply the sensor node for a RF received power of 2 dBm, with a conversion efficiency $\eta_0=6\%$. With Prototype 2 the sensor node can be power supplied for a RF received power of -3 dBm, with $\eta_0=18\%$. Finally, Prototype 3 is capable of supplying an output voltage of 3 V for an RF received power of -4 dBm, with $\eta_0=20\%$.

Combining the results and the spectral opportunities identified in [12], it is possible that the levels of received power harvested from the environment are not enough to fulfill the goal of power supplying a WBAN. The data collected for the maximum levels of received power are around -27 dBm, which are insufficient to generate an output voltage of 1.8 V.

VI. CONCLUSIONS

In this paper we have identified the spectrum opportunities for RF energy harvesting to power supply the wireless sensor nodes in real indoor/outdoor scenarios. The set of indoor/outdoor most promising frequency bands are 79 to 96 MHz (mobile/radio broadcast stations), 391 MHz (emergency broadcast stations), 750 to 758 MHz (digital television broadcast stations), 935 to 960 MHz (GSM 900 broadcast stations), 1855 to 1868 MHz (GSM 1800 broadcast stations) and 2115 to 2160 MHz (UMTS broadcast stations). For the GSM (900/1800) frequency bands, a dual-band printed antenna has been proposed, with gains of the order of 1.8-2.06 dB and 77.6-82% efficiency. The design of RF energy harvesting circuits and the choice of textile materials for a wearable antenna have also been analysed. We have also simulated the behaviour of a 5-stage Dickson voltage multiplier for power supplying an IRIS mote. Three prototypes (1, 2 and 3) for the 5-stage Dickson voltage multiplier were proposed and experimentally characterized. If we consider the minimum voltage to power supply the sensor node (i.e., 1.8 V), all prototypes (1, 2 and 3) can power supply the necessary voltage to a sensor node work for an RF received power of -1 dBm, -7 dBm and -8 dBm, for conversion efficiencies of 5%, 16% and 17%, respectively. If the advised voltage to power supply the sensor node (i.e., 3 V) is considered, all prototypes (1, 2 and 3) are able to power supply the sensor node for an RF received power of 2, -3 and -4 dBm, with conversion efficiencies of 6, 18 and 20%, respectively. Future work includes improving the process of manufacturing the printed circuit boards in order to achieve improved impedance matching whilst reaching values for the conversion efficiency closer to the simulated ones.

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