Electrically Small Spiral PIFA for Deep Implantable Devices

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ABSTRACT In this paper, a miniaturized implantable circularly polarized spiral Planar Inverted-F Antenna (SPIFA) in the UHF (600-800 MHz) band is presented. This antenna is intended for deep implantable devices such as leadless pacemakers and deep brain stimulation (DBS), which facilitates the reception of RF power from an external transmitter. The antenna is electrically small, with a volume of $\pi \times 5 \, \tilde{\text{mm}} \times 5 \, \tilde{\text{mm}} \times 3.2 \, \tilde{\text{mm}}$ and a diameter of 0.022λ. The performance of the proposed antenna in terms of reflection coefficient, realized gain and axial ratio are assessed when accounting for the effects of operating in different types of human body tissues, different biocompatible materials and different thicknesses and depths of the implanted antenna. Finally, the antenna is prototyped and measured in free space, a phantom model, in a cow’s fat and muscle tissues to validate the simulation results, indicating good agreements. A realized gain around $-20 \, \text{dBm}$ is achieved when operating in 50 mm depth in cow’s muscle tissue while having electrically very small dimensions compared to implantable antennas reported in the literature.

INDEX TERMS Implantable antennas, PIFAs, spiral antennas, circular polarization, equivalent circuits.

I. INTRODUCTION

Wireless technologies and communication are nowadays involved in every aspect of human life, and includes areas such as telemedicine and implants [1]. Throughout the past few decades, implantable devices which are implanted inside the human body for transmitting power have been developed at a rapid pace [2], [3]. The two prevalent methods of power transfer for these devices are low frequency inductive links and radio-frequency (RF) links. Unlike the inductive links, RF links are able to transfer not only power but also high data rate signals [4], [5]. Such RF-link approach is already applied in a variety of applications, including blood-glucose monitoring [6], [7], pacemakers [8]–[10], and retinal implants [11], [12], whereas newer implantable medical devices continues to evolve.

In this paper, the RF link approach is selected for power transmission to an implantable antenna. While the authors in [13], [14] proposed wideband antennas which are more suitable for data transmission and communication, bandwidth is of less importance in this work as it is applied for power transfer. In most medical studies, the frequency band is selected based on the licensed frequencies (such as MICS and ISM) [13]–[19]. This method of selection may not be optimal for the application in question [4], [20]. In applications where the relative orientation of the transmitter and the receiver might change, especially in implantable cases with body movements, circularly polarized (CP) antennas are usually chosen. This is due to their orientation and polarization independence between the transmitter and the receiver [13], [14], [17]–[19], [21], [22].
Due to the limitations of space in the body, compact antennas with small volumes are the optimal choice for implantable devices [22]. Patch antennas which satisfy this requirement can be selected for this purpose. However, their large expected size in the UHF band will be unsuitable for implantable medical applications. To alleviate this, antenna miniaturization techniques can be used, such as meandered line, spiral line [16], [23]–[25] and slot [14], [18], [26]–[28], as well as the addition of shorting pins [13], [28]–[30]. These methods add distributed inductance into the antenna impedance by increasing their current paths, or is also equivalent to electrically lengthening the antenna dimension. Another effective miniaturization technique is by applying the PIFA topology, which is an easy-to-fabricate technique for the printed antennas [13], [27].

In implantable applications, the implanted device is covered by a biocompatible material to prevent direct contact between the antenna and the body tissues [13], [14]. However, the performance of the enclosure which is located in the near-field region of the body tissues can be easily affected. It is indeed more accurate to assume that the antenna is operating in a lossy body medium, which then requires a different definition for radiation patterns [31]. However, in this case, the conventional method of analysis is used to characterize the antenna. This is to ensure that other factors which potentially impact its performance such as the electrical properties of the body tissues and depth of the implant can be also considered and studied.

Based on the Chu-Wheeler limitation, it is evident that by decreasing the dimensions of the antenna, its quality factor increases. This restrains most of the radiated power into the near field, resulting in the improper radiation of the antenna [32], [33]. To avoid this, methods to ensure that the quality factor can be reduced for proper far-field antenna radiation must be introduced [34]. One such method is to increase the loss of the substrate in implantable antennas.

In this paper, a circularly polarized spiral PIFA (SPIFA) with a volume of $\pi \times 5\ \text{mm} \times 5\ \text{mm} \times 3.2\ \text{mm}$ is proposed in the UHF band, to supply power to the implantable devices located in deep human tissues. Finally, a rectifier is designed and implemented behind the antenna to produce a rectenna. An initial work [35] studied the details of different rectifier structures and also a wireless charging system for such applications. In this paper, a new, electrically small (approximately $\frac{\lambda}{4}$) and efficient antenna is introduced for deep implantable application. The contributions of this paper are as follows:

- The matching is performed using its lossy substrate while maintaining a satisfactory efficiency for such antenna.
- This antenna is one of the few designs (besides [16], [23]) which is capable of operating in different environments (free space, muscle, fat) with small variations in realized gain, while simultaneously maintaining its bandwidth and resonant frequency. On the contrary, recent work in [10], [13], [17]–[19], [26], [28]–[30], designed and optimized their antennas in a lossy (in-body) environment and are almost unable to operate when surrounded by free space.

The effects of different environments in which the antenna operates, such as fat and muscle phantoms, and when integrated within two types of biocompatible capsules are investigated. Besides that, the effects of the depth of the implant are also studied. Since the main goal of the antenna is to wirelessly receive power and charge the battery of the leadless pacemaker, the detailed study of Specific Absorption Rate (SAR) is considered beyond the scope of this manuscript. This is due to its low expected received power and non-radiating functionality. Nonetheless, SAR study can be considered in future work, especially in situations where the proposed antenna is being used as a transmit antenna to channel data from the pacemaker.

The proposed antenna characteristics; frequency band, bandwidth, miniaturization, and polarization. Section III gives an argument on the limitations of small antennas and equivalent circuit of the proposed antenna. In Section IV, the performance of the proposed antenna inside the different body tissues of human body, the effect of different antenna depths in body tissues and biocompatible enclosing materials over the antenna are evaluated. Section V presents the measurement results and a comparison table with implantable antennas available in literature. Finally, Section VI outlines the concluding remarks from this work.

II. DESIGN OF ANTENNA

Implantable antennas should be as small as possible when operating inside the body, without affecting the surrounding tissues. One of the best options for implantable antennas is the planar structure, which occupies a relatively small space. In designing conventional planar antennas, their operating frequency and substrate properties are the main determinant of the overall dimensions and the performance of the antenna. With increasing permittivity ($\varepsilon_r$) of the substrate, the antenna performance can deteriorate, despite being smaller in size [36]. The proposed antenna is a spiral PIFA printed on the FR4 substrate (with a relative permittivity, $\varepsilon_r = 4.3$, and loss tangent, $\tan\delta = 0.025$), as depicted in Fig. 1. Another FR4 substrate is placed below the antenna ground to mount the rectifier circuit. To investigate the possible effects of the rectifier, these components are also included in the final antenna model, as shown in Fig. 1(b). Fig. 1(c) illustrates the overall antenna and the rectifier structure with the substrates.
hidden. Its overall volume is about $\pi \times 5 \, \text{rm} \times 5 \, \text{rm} \times 3.2 \, \text{rm}$.

**A. FREQUENCY BAND**

For implantable antennas, their operation in high frequency bands will typically result in the increase of the absorbed electromagnetic energy by the body tissues, and the increase of the propagation path loss. At the lower frequencies, on the other hand, a compact and implant-sized antenna is electrically too small to radiate; hence, its radiation and total efficiency would not be acceptable. The authors in [20] demonstrated that the optimal frequency band for implanted medical devices should be chosen based on the antenna efficiency and the losses caused by the surrounding tissues. As the depth of the implant increases, the absorbed power by the tissues also increase. Thus, the depth of the implant has a considerable impact on the selections of the frequency band. The UHF band from 600 to 800 MHz is a proper option for implants with 50 mm of depth into the tissues [4], [20]. Two other factors to be considered when selecting the operating frequency in medical environments are the electromagnetic compatibility (EMC) and the electromagnetic interference (EMI). However, these aspects are not expected to pose serious problems when applied in a controlled indoor environment. In this work, the operating frequency of the proposed SPIFA in free space is tuned to be operating at around 700 MHz, and can be tuned to any intended MICS/ISM bands if needed by modifications of the physical parameters of the SPIFA.

**B. BANDWIDTH**

There is a fundamental tradeoff between the antenna bandwidth (BW) and efficiency in electrically small antennas (ESA). As the BW increases, the antenna faces significant efficiency reduction [37], making it unsuitable to be implanted deep into the tissues. In power transfer applications, on the other hand, the BW is not much of importance; therefore, its reduction can be tolerated as the focus on increasing efficiency is placed during the design stage. This is the main reason that the proposed antenna is designed with a 10 MHz bandwidth, which is sufficient for power transfer purposes.

**C. MINIATURIZATION**

One type of printed antennas with quasi omnidirectional pattern is the printed T-strip antennas (PTSA), which resonates at $\frac{\lambda_g}{2}$ (where $\lambda_g = \frac{\lambda_0}{\sqrt{\epsilon_r}}$ is the guided wavelength). To enable radiation at 710 MHz, the conventional antenna length is about 130 mm, which is too long to be inserted in the body, see Fig. 2(a). By applying the shorting pin technique, the length of the PTSA is roughly halved, as illustrated in Fig. 2(b). However, this new antenna length is still too large to suit this application. To increase antenna compactness, one method is to wind the strip into a spiral structure. The final compact structure is a spiral printed inverted-F antenna, with the largest diameter of 10 mm. In addition to that, by winding the strip of the PIFA into a spiral form, the resonant frequency does not change significantly, while its size is reduced by the order of 13, as shown in Fig. 3. Table 1 summarizes the detailed dimensions of the antenna. The Archimedean spiral shape in the $xy$ plane is obtained using following equations:

$$x(t) = 0.23 \cos (t)(20 - (t/2)) \quad (1a)$$
$$y(t) = 0.23 \sin (t)(20 - (t/2)) \quad (1b)$$

where $t$ is an auxiliary parameter.

**D. CIRCULAR POLARIZATION**

As illustrated in Fig. 4, two perpendicular current components, $I_z$ and $I_\phi$, are present in the structure. The equivalent current along the $z$-direction is obtained from the shorting and the feeding pins, which are related as $I_z(\text{equivalent}) = |I_z| e^{j4/\pi}$. Due to the narrow strip between the feeding and the shorting pins, a small phase difference is introduced in the

![Figure 1. Simulated SPIFA structure: (a) Front. (b) Back. (c) Side with hidden substrates.](image-url)
FIGURE 2. Printed T-strip antennas (a) Without the short pin. (b) With the short pin.

FIGURE 3. Simulated reflection coefficients of the three antennas designed on FR4, Case 1: Printed T-strip antenna (130 mm). Case 2: PIFA with the shorting pin (64 mm). Case 3: SPIFA (10 mm).

TABLE 1. SPIFA Dimension.

| Parameter                          | Value (mm) |
|------------------------------------|------------|
| Track Width                        | 0.5        |
| Small Radius                       | 1.5        |
| Big Radius                         | 5          |
| Antenna Substrate Height           | 2.2        |
| Rectifier Substrate Height         | 0.8        |
| Distance from Feed to Short Pin    | 3          |

FIGURE 4. Surface current distribution at 710 MHz.

FIGURE 5. AR at 710 MHz at the $\theta = 90^\circ$ direction.

FIGURE 6. LH and RH patterns of the proposed antenna in the azimuth and elevation plane at 710 MHz.

The E-field is studied in the yz plane when its propagation is y-directed. The two perpendicular components of the electric field, $E_z$ and $E_\phi$ in yz plane are denoted as $E_z$ and $E_\phi$. As shown in Fig. 5, the phase difference between the two E-field components is approximately $90^\circ$ at $\theta = 90^\circ$ (towards the endfire direction), whereas their amplitude difference is less than 3 dB at 710 MHz. The proposed antenna patterns in the elevation and azimuth planes at 710 MHz are shown in Fig. 6, indicating right-handed circular polarization (RHCP).

An alternative method in analyzing the circular polarization behavior of this structure is explained as follows, which is based on the comparison of the E-field and surface current at each moment valid for omni-directional antennas, as described in [38]. The E-field on the radiating plane mainly contributes to the vertically-polarized waves at the end-fire direction of the antenna. Similarly, the current surface on the radiating plane introduces the horizontally-polarized waves at the horizontal xy-plane of the antenna. In Fig. 7, the electric field (right) and the surface current distribution (left) of the antenna are presented at different times. At $t = 0$, the electric field is maximum, and the vertically-polarized wave radiates, whereas at $t = T/4$, the surface current increases and the horizontally-polarized wave starts radiating. At $t = T/2$, the E-field reaches its maximum value in the reverse direction, and thus the antenna radiates vertically but towards the reverse direction relative to $t = 0$. Finally, at $t = 3T/4$, the surface current again reaches its maximum but in a reverse direction compared to when $t = T/4$. It also can be seen that there is always a $90^\circ$ phase difference between the vertical and horizontal waves. Therefore, the $90^\circ$ phase difference...
between two orthogonal components inherently exists. The AR beamwidth of the antenna is at least 147° and 149° respectively in the φ = 0° and φ = 90° planes, which is sufficiently wide.

III. OPERATING PRINCIPLES OF THE SPIFA

As the largest dimension of the proposed antenna is less than $\frac{\lambda}{4}$, this antenna is considered an ESA. However, to the authors’ best knowledge, the operation and performance of such lossy ESAs has not been well studied in literature. This section attempts to explain its limitations using the Chu-Wheeler’s principles and proposing an equivalent circuit to explain the effect of the lossy substrate on the ESA.

A. ELECTRICALLY SMALL ANTENNAS’ LIMITATIONS

An antenna is defined to be electrically small when its largest dimension, $a$, is smaller than the value given by (2) [32].

$$ka = \frac{2\pi}{\lambda} a \leq \frac{1}{2}$$

where $k$ is the medium’s propagation constant. One of the main characteristics of ESAs is their high radiation quality factor ($Q_{rad}$). Hence, much of the antenna’s power is stored in its near field, whereas a small portion of its power is radiated. Also, the BW is inversely proportional to the quality factor. The fundamental limitation of ESAs, which is the trade-off between their $Q_{rad}$ and dimensions was first reported in [32], [33], [39] and more comprehensively in [40]. Equation (3) approximates the relation between $Q_{rad}$ and the largest dimension for a lossless circularly polarized antenna [40].

$$Q_{rad} = \frac{1}{2} \left( \frac{1}{k^3 a^3} + \frac{2}{ka} \right)$$

In this case, largest dimension of the proposed antenna is 10 mm, and the wavelength at 700 MHz is 420 mm. Based on the criterion mentioned in (2) and (3), the proposed SPIFA ($ka \approx 0.15$) is an ESA and the minimum achievable value of $Q_{rad}$ of the lossless antenna is 155.

Another limitation of ESAs is their very low radiation resistance, which leads to very poor impedance matching and radiation efficiency. Ref. [34] and [37] illustrate the effects of radiation efficiency on the ($Q - ka$) fundamental limits. A method to increase the radiation efficiency is to increase the radiation level, thereby expanding the current as much as possible [41]. On the contrary, such method cannot be applied in the proposed antenna due to its limited volume. Increasing the input resistance by adding to antenna’s loss will lower its quality factor and improves its BW, while further decreasing the radiation efficiency. The relation between the antenna size and the quality factor for different values of efficiency is depicted in Fig. 8, similar to levels reported in [34]. To summarize, a lossy antenna’s quality factor is directly proportional to radiation efficiency ($Q \propto \eta$) [37].

In printed antennas, the loss of the substrate adds to the input resistance, decreases the efficiency of the antenna, while decreasing their $Q$. Based on Fig. 8, as the radiation efficiency decreases with a lossy substrate, its $Q$ decreases. To investigate the effect of loss, the proposed antenna is simulated using two types of substrates: a lossy substrate and a lossless substrate. The reflection coefficients for both cases in Fig. 9, indicate that the antenna is matched when using
the lossy substrate. On the contrary, almost all fed power is reflected back into the antenna when using the lossless substrate. Thus, the impedance matching and the radiation efficiency, which are influenced by the level of loss in the substrate, need to be traded-off with care. The calculated total efficiency for the lossless and lossy SPIFA indicates that the latter is about 7 dB better than its lossless counterpart.

The proposed antenna’s bandwidth is approximately 10 MHz at 700 MHz, with an approximate quality factor of 70. Fig. 8 studies the quality factor of the SPIFA versus electrical size. It can be seen that the antenna is close to the fundamental limitation for lossy ESAs.

### B. EQUIVALENT CIRCUIT

Printed Archimedean spiral strip can be modeled as a transmission line wrapped into a spiral shape. The characteristic impedance and propagation constant of the spiral transmission line model depend on the arm width, the substrate thickness, and the spacing within the spiral structure [42]. The equivalent circuit for the proposed antenna is shown in Fig. 10. Observing from the feed point located on the spiral structure, two parallel line sections can be observed. Firstly, a transmission line terminated with the shorting pin, and secondly, the spiral arm located on the other side. The equivalent circuit of each transmission line section includes an inductor and a capacitor. The parallel resistors represent the effect of the lossy substrate [43]. In the first section denoted in Fig. 10, the shorting pin is modeled as an inductor, which increases in value with the decrease of its diameter [21]. However, the diameter of the shorting pin is limited by the width of the spiral arm. Thus, TL1 ($L_1$, $C_1$) is terminated with a relatively large inductance value ($L_2$). The second transmission line (TL2) in the equivalent circuit can be modeled as a periodic circuit, with each of its unit cells composed of two parts. The first part, which consists of an inductor ($L_2$), a capacitor ($C_2$) and a resistor ($R$) represents a simple lumped element model for a transmission line. The second part of this transmission line, which resembles a T-shaped section is modelled using two inductors ($L$). A capacitor ($C$) is then introduced in between them to account for the bent transmission line in the spiral structure [42]. The values of the parallel capacitors ($C_2$) for these transmission lines are estimated using the microstrip capacitance model. The radiation resistance ($R_r$), which is proportional to the radiation efficiency ($\eta_{\text{rad}}$) can be calculated using the following relations:

\[
R_r + R_L = 50 \, \Omega 
\]

\[
\eta_{\text{rad}} = \frac{R_r}{R_r + R_L} 
\]

where $R_L$ in the proposed antenna represents the loss of the substrate. With $\eta_{\text{rad}} = -29$ dB, the radiation resistance in the equivalent circuit is calculated to be about 0.06 $\Omega$, which is small and negligible. The final equivalent circuit model is then simulated in Advanced Design System (ADS) software to validate its accuracy. With six periodic LC sections as transmission lines and minor tunings, similar reflection coefficients are obtained from the equivalent circuit model and from the 3D full-wave simulations. Note that the six-line section is sized based on the wavelength and the transmission line model to result in reflection coefficients with satisfactory accuracy. Each transmission line section is less than one tenth of a wavelength, $\lambda/10$ [44]. Besides that, the equivalent circuit values are also affected by the gap between lines. Its initial values are first estimated based on the method presented in [45]. Then, the values of the equivalent model (T-shape) [42] are obtained by tuning their values until an acceptable reflection coefficient is attained. An example of the tuning of a resistance value in this equivalent circuit is shown in Fig. 11. On the other hand, the optimized values of the inductors and capacitors of the equivalent circuit are summarized in Table 2. Note that this equivalent circuit is modeled in free-space only, without considering the on-body case. The high dielectric properties of the body affected the effective wavelength of the medium and increased loss. This directly impacts the resonant frequency, thus affecting the values of $L$, $C$, and $R$ in the equivalent circuit.

In the lossless case, the antenna input resistance only contains radiation resistance, which is much less than 50 $\Omega$. This leads to a high mismatch at the target frequency. Simulation results of the equivalent circuit also showed a significant increase in mismatch when the resistors in the equivalent circuit are omitted. On the contrary, for the lossy antenna, the resistance caused by the lossy substrate results in satisfactory impedance matching at the desired resonant frequency, as illustrated in Fig. 11(a). This interesting phenomenon will be explained in more detail in the next subsection. Meanwhile, to provide more understanding of this behavior, five values of $R$ are selected to study the effects of the loss tangent. It is noticed that the higher the loss tangent, the smaller the $R$.

### TABLE 2. Values of inductors and capacitors in the equivalent circuit.

| Parameters | Values (mm) |
|-----------|-------------|
| $L$       | 0.45 nH     |
| $C$       | 0.76 nF     |
| $L_{TL}$  | 0.67 nH     |
| $C_{TL}$  | 145.25 pF   |
| $L_1$     | 5 nH        |
| $C_1$     | 180 pF      |
| $L_2$     | 40 nH       |
C. OPTIMUM SUBSTRATE LOSS

The effect of a lossy substrate on the performance of an electrically small antenna can be further analyzed using its equivalent circuit. This is due to its similarity with the 3D full wave simulations, as observed from the previous sub-section. Since the proposed antenna operates in the receiving mode, changes in its total efficiency are more significant than its radiation efficiency. The total efficiency of the antenna consists of two components, namely the radiation efficiency ($\eta_{\text{rad}}$) and the mismatch loss $(1 - |S_{11}|^2)$, according to (5).

$$\eta_{\text{tot}} = \eta_{\text{tot}} \times (1 - |S_{11}|^2) \quad (5)$$

On one hand, with the increase in the lossy element, the antenna reflection coefficient is directly improved, but on the other hand, the radiation efficiency of the antenna is reduced. To determine which component is more dominant, the $S_{11}$, $\eta_{\text{rad}}$ and $\eta_{\text{tot}}$ are calculated in terms of loss factor, $R$, in Fig. 12(a), (b) and (c), at three different operating frequencies. The radiation efficiency is obtained through the ratio of $P_{\text{in}}$ and $P_{\text{Loss}}$ to the input power, whereas $P_{\text{Loss}}$ is calculated using the resistors model. As expected, the $(1 - |S_{11}|^2)$ factor increases with the value of $R$, while at the same time, $\eta_{\text{rad}}$ decreases. The multiplication of the two illustrates that there is a compromise between $\eta_{\text{rad}}$ and the $(1 - |S_{11}|^2)$ term in this circuit. Therefore, an optimal $R$ can be found at each frequency, which results in the highest total efficiency. However, this loss factor is hardly controllable, as it depends mainly on the substrate’s loss tangent which is more or less uniform in practice. Such mechanism of optimizing $R$ to result in the best total efficiency is effective.

IV. ANTENNA IN-BODY OPERATION

In implantable applications, body tissues are placed in the near field region of the antenna, which could impact the antenna operation [46]. The different electrical properties of
the different types of tissues and their thicknesses may affect the antenna differently. Due to the limited bandwidth of this antenna, it is important to consider all possible situations that may affect its operating frequency. In this section, the effects of skin, fat and muscle on the antenna operation are investigated. In addition to that, another study is conducted on the antenna performance when different materials are used to encapsulate it.

A. BIOCOMPATIBLE MATERIALS

To prevent a direct contact between the antenna and the body tissues, it is necessary to encapsulate the antenna using biocompatible materials. This material ideally must not influence the characteristics of the antenna significantly. Teflon \((\varepsilon_r\text{-Teflon} \approx 2.1)\) and Alumina \((\varepsilon_r\text{-Alumina} \approx 9.9)\) are two of such materials well known for this purpose [13], [15], [26]. To evaluate their effects, the antenna is encapsulated using both materials before being inserted into a single-layered cube of fat tissue dimensioned at 50 mm \(\times\) 50 mm \(\times\) 50 mm. The effects in encapsulating the antenna using the two materials are illustrated in Fig. 13. Compared to the antenna without cover assessed in free space, the operating frequency of the antenna shifts downwards when it is integrated with Teflon and is inserted into the fat tissue. On the other hand, a larger downward shift is observed when the antenna is encapsulated with Alumina prior to the insertion into the fat tissue. Nonetheless, the structure still operates with acceptable AR when integrated with Teflon. This is unlike the case encapsulation using Alumina, where the antenna AR degrades to exceed 3 dB.

The differences in AR caused by the different encapsulating materials can be explained by the theory governing the incident waves at the interface of two materials [47]. When the relative permittivity of two environments is distinct, the vertical component of the transmission coefficient will differ dramatically from its horizontal component. Due to this significant amplitude difference between the two perpendicular electric field components, the AR of the circularly-polarized wave will then deteriorate. Similarly, in the case of the proposed antenna, there are considerable differences in dielectric constant between Alumina as the antenna encapsulating material, and the FR4 substrate \((\varepsilon_r\text{-FR4} \approx 4.3)\) and the fat tissue \((\varepsilon_r\text{-Fat} \approx 5.5)\) where the antenna is operating in.

B. ANTENNA PERFORMANCE IN FAT TISSUE

To investigate the effects of the human body tissues on the antenna characteristics, the designed antenna encapsulated in Teflon is evaluated when inserted into a 50 mm \(\times\) 50 mm \(\times\) 50 mm human body phantom, modeled as: i) a single-layered fat tissue phantom, and ii) a two-layered skin and fat tissue placed on top of each other. A 5 mm-thick skin \((\varepsilon_r\text{-Skin} \approx 42.5)\) layer encloses the fat tissue. In both evaluations, the antenna is inserted with a depth of \(h = 40\) mm from the top of the phantom, as shown in Fig. 14. The dielectric values of the Duke model from IT'IS [48] have been used in simulations.

The AR (in the azimuth plane defined based on coordinates shown in Fig. 1) and \(S_{11}\) when the antenna is evaluated inside these two phantoms are illustrated in Fig. 15. The antenna pattern is observed to be omnidirectional, with a maximum gain is on the horizontal plane (defined in coordinate shown in Fig. 1). The antenna patterns in the azimuth and the elevation planes are depicted in Fig. 16, indicating that the effect of the skin tissue is negligible on the antenna performance. This is due to the size of the antenna near-field region, which is calculated to be less than 1 mm, based on [22], based on the frequency and the largest dimension which is 10 mm. Hence, only the nearest tissues to the antenna can impact its characteristics, and the skin tissue, not being in the near-field, is of no major contribution to the antenna performance.

The axial ratios of this antenna in the azimuth plane, and the reflection coefficients when located with different depths
FIGURE 16. Simulated 2D far field patterns of the SPIFA when located in the single-layered and two-layered phantoms at 680 MHz in a) the azimuth and b) the elevation planes.

FIGURE 17. Simulated reflection coefficient and AR of the antenna when located in a single-layered fat tissue and located with different \( h \).

C. ANTENNA PERFORMANCE IN MUSCLE TISSUE

In this section, the effect of the muscle tissues on the implantable antenna characteristics is studied. To do this, a three-layered cube containing 5 mm-thick skin, 5 mm-thick fat and 40 mm-thick muscle is considered, as shown in Fig. 18. The \( S_{11} \) and AR when the antenna is placed at a depth of 40 mm into the muscle tissue is presented in Fig. 19. It is observed that the resulting resonant frequency is lower than when placed in the fat layer, as presented in Fig. 13. This is obvious, as the muscle tissue is higher in dielectric permittivity compared to that of fat. Another observation is that when located in the muscle tissue, the circular polarization of the antenna deteriorates severely. This is caused by the significant difference between the dielectric permittivity of the muscle tissue \( (\epsilon_r_{\text{Muscle}} \approx 55) \) and the proposed antenna, which is made of FR4 and Teflon capsule \( (\epsilon_r_{\text{Teflon}} \approx 2.4) \). A possible method to improve AR in this situation is the use of a capsule with a higher dielectric constant. Table 3 shows that as the dielectric permittivity of the capsule increases, the AR reduces towards the acceptable level of 3 dB. Due to this reason, Alumina which features a high dielectric permittivity is implemented as the antenna encapsulating material next. A gradual improvement in the AR performance relative to the Teflon encapsulation is observed, as illustrated in Fig. 19.

Besides that, the antenna patterns shown in Fig. 20 also indicate gain improvement in the elevation plane in comparison to the antenna encapsulated with Teflon. To summarize, the proposed antenna with the Alumina encapsulation is operational at 500 MHz with an AR of 13 dB inside the muscle tissue.

It is worth noting that the orientation of the proposed antenna in the \( \phi \) direction (in which the antenna has a symmetry and omni-directional pattern) does not affect the radiation pattern. The orientation of the antenna in the \( \theta \) direction also does not considerably affect its radiation pattern. For a leadless pacemaker application, the device, and consequently the antenna, is embedded and fixed in the right ventricle of the heart. This limits the movement or rotation of the antenna. Nonetheless, the predicted worst case may happen when the antenna is rotated in the \( \theta \) direction. However, this will only result in a slightly higher loss in the link budget, as the broadside of the SPIFA will be directed towards the on-body transmitting antenna. These observations are validated via simulations results obtained by rotating the antenna in the \( \theta \) direction in the muscle tissue. To study the effects of the
surrounding tissues in a more realistic manner, the antenna is simulated as shown in Fig. 21 with the skin and fat located on one side of the operating antenna. The resulting azimuth and elevation patterns are illustrated in Fig. 21. Further simulations also confirmed that it is the hosting tissue that affects the antenna performance. It can be concluded that the effects of the antenna orientation and the tissues outside the antenna encapsulation are negligible on its radiation pattern and reflection coefficient.

V. MEASUREMENT RESULTS

The fabricated antenna which is fed using a UFL connector is illustrated in Fig. 23. The antenna is covered with a capsule made of Teflon which can be biocompatible and easily available, and then inserted into the body tissue. A hole is drilled at the bottom of the capsule to access the SPIFA during measurements, as shown in Fig. 23.

A. MEASUREMENT OF REFLECTION COEFFICIENT

Measurements of reflection coefficient and the resonant frequencies of the encapsulated SPIFA are performed under three scenarios. First, it is measured inside a phantom model which mimics the electrical properties of muscle tissue at the operating frequency [49], while being connected to a Vector Network Analyzer (VNA) at the other end. Secondly, measurements were performed with the encapsulated SPIFA inserted into a piece of realistic fat tissue obtained from a cow. Finally, the third setup involves evaluating the SPIFA when being inserted inside a cow’s muscle tissue. All setups are illustrated in Fig. 24.

The simulated and measured reflection coefficients of the proposed antenna in free space and with the SPIFA operating in the phantom, fat, and muscle tissues are compared in Fig. 25. The obtained results indicate that the bandwidth produced by the proposed SPIFA remains almost unchanged when it is located inside different tissues. Despite that, the resonant frequency of the encapsulated SPIFA shifted slightly downwards due to the relatively higher dielectric constant in the phantom, fat and muscle tissues in comparison to free
FIGURE 25. Measured reflection coefficients of the SPIFA encapsulated in Teflon when evaluated in free space, in the phantom, in the fat tissue and in the muscle tissue and simulated reflection coefficient of SPIFA in free space.

The changes of the resonant frequency of the proposed antenna in different environments are still less than 100 MHz. On the contrary, in most work [29], a slight change in the implant depth or changing the environment of the antenna disrupts its operation or shifts its frequency significantly. It is worth noting that, as mentioned in section I and as can be seen in Fig. 25, the antenna can operate in different environments resulting in the antenna to be applied for other applications as well.

B. MEASUREMENT OF POLARIZATION

The polarization of the SPIFA is measured next in a laboratory environment. A linearly polarized Log Periodic Dipole Array (LPDA) is connected to a signal generator and is used to transmit RF power towards the encapsulated SPIFA located inside the fat tissue. In turn, the SPIFA is connected to a spectrum analyzer, as shown in Fig. 26. Measurements were performed in the fat tissue at six different frequencies around the resonant frequency, each with two orthogonal polarization [26]. The difference between the received signal levels (in dB) in the vertical and the horizontal polarization at every operating frequency is summarized in Table 4. It can be seen that the maximum difference in free space and fat is only 4.1 dB, which indicates that the received signal power is almost independent of the direction of the transmitter. This behavior of the SPIFA with Teflon encapsulation is also prevalent in CP antennas. Given that the antenna’s capsule, which is inserted into the muscle, is made of Teflon, the results of the structure when assessed in the muscle tissue are also in good agreement with the simulation results. Note that the listed values in Table 4 are the difference between the received signal levels when the transmitter is oriented vertically and horizontally. Therefore, they are independent of the amount of reflected power ($S_{11}$) or the amount of delivered power to the structure at these frequencies.

C. RADIATION PATTERN MEASUREMENT

Measurements of the antenna radiation pattern for such electrically small antennas need to be performed when it is connected to measurement equipment using a cable. Despite performing this experimental examination inside an anechoic chamber, the simulated and measured patterns clearly indicate that the effect of the cable is prevalent. This is due to the very low antenna gain, and the radiation from the cable is of similar magnitude relative to that of the antenna, which considerably affects measurement accuracy. Thus, the result from this setup in 27 is compared with the simulation result obtained from an exact measurement setup modeled with the cable to account for its effects. This is in contrast to the results in Fig. 6, where the simulated pattern is obtained without any cable connected to the antenna. Nonetheless, in practice, the rectifier circuit is placed directly behind the antenna structure, as shown in Fig. 30, which eliminates the need for the cable when operating as a complete system using the proposed antenna.

D. MEASUREMENT OF GAIN

The gain of an antenna can be measured using Friis transmission equation as in (6) when the receiver and the transmitter antenna are spaced with a fixed distance [21].

$$\frac{P_r}{P_t} = (1 - |\Gamma_t|^2)(1 - |\Gamma_r|^2)\left(\frac{\lambda}{4\pi d}\right)^2 G_t G_r |\rho_t - \rho_r|^2$$  

$P_r$: Receiving power  
$P_t$: Transmitting power

![Experimental setup to measure polarization.](image)

TABLE 4. Relative difference between received signal levels for transmitter oriented vertically and horizontally.

| Frequency (MHz) | Amplitude Difference in Free Space | Amplitude Difference in Fat with Teflon Capsule (dB) | Amplitude Difference in Muscle with Teflon Capsule (dB) |
|----------------|----------------------------------|-----------------------------------------------|-----------------------------------------------|
| 580            | -                                | -                                             | 23.3                                          |
| 590            | -                                | -                                             | 21.7                                          |
| 600            | -                                | -                                             | 20.6                                          |
| 610            | -                                | -                                             | 19.9                                          |
| 620            | -                                | 3.3                                           | 19.1                                          |
| 630            | -                                | 3.3                                           | 22.8                                          |
| 640            | 3.7                              | 3.2                                           | -                                             |
| 650            | 3.4                              | 2.8                                           | -                                             |
| 660            | 3.6                              | 2.9                                           | -                                             |
| 670            | 3.7                              | 3.1                                           | -                                             |
| 680            | 3.9                              | 3.8                                           | -                                             |
| 690            | 4.1                              | -                                             | -                                             |
$G_r$: Receiving antenna gain
$\Gamma$: Reflection coefficient (for the transmitter ($t$) and for the receiver ($r$)).
$G_t$: Transmitter antenna gain
$\rho$: Polarization vector for the transmitter ($t$) and for the receiver ($r$)).

This equation is valid when the distance between the transmitter and the receiver, $d$, is more than $\frac{2D^2}{\lambda}$, where $D$ is the largest dimension of antennas. This is to ensure that they are in each other’s far field region. The known gain of the LPDA transmitter is 5.6 dB, whereas the transmitted power from a signal generator is 0 dBm. The receiver and transmitter antenna is located at 2 m distance from each other. The antenna with the capsule is located 50 mm inside the muscle. The amount of received power measured by a spectrum analyzer is $-57$ dBm. To measure the received power of the SPIFA, the receiver needs to be positioned in such a way that the effect of the cable on the received power is negligible. PLF is 0 dB when the antennas are operating in the same polarization, and $-3$ dB when the polarization of one antenna is linear and the other one is circular.

Table 5 summarizes all parameters from (6). The obtained gain of the receiving SPIFA ($G_r$) in the muscle tissue is $-20$ dB. The simulated gain of the antenna in this same condition is $-20$ dB. The improved gain when the antenna is operating in the muscle tissue relative to that of in free space ($-27$ dB) can be explained as follows. The presence of the lossy muscle around the antenna reduces the loss contributed by the substrate and facilitated the antenna matching with less substrate loss. This can be observed form the power budget listed in Table 6. Simulations indicated that the power loss in the substrate decreased from $0.488$ W (out of $0.5$ W) in free space to $0.325$ W when operating in the muscle, indicating $0.154$ W of power loss. In other words, the presence of the muscle tissue contributed to the resistance, $R$, in Fig. 10 to Fig. 12, and reduced the contribution of the substrate loss in the antenna matching.

Table 7 compares the properties and performance of the proposed SPIFA with other reported implantable antennas in literature. It is evident that the proposed antenna features the smallest dimension with an improved antenna gain, considering the deeper location of the antenna inside the body relative to previous investigations. Results also show that the SPIFA can be operated in body with satisfactory performance.
RF transmitting devices operating close to the human body. In this evaluation, the proposed antenna is used as the receiver and is located inside the human body, whereas a transmitter is located at a distance of 20 cm from the human body. Based on the IEEE C95.1 2005 [50], the maximum allowed received power is calculated to be less than $-20$ dBm at 655 MHz when the antenna is placed at a depth of 50 mm in the three-layered phantom. For the designed rectifier, measurements indicated that the maximum efficiency of the rectifier at $-20$ dBm RF input power is about 40%.

Next, the transmit power from LPDA is varied to validate the performance of the rectenna in this work. With a fixed distance of 20 cm between the transmitter and the receiver, the output power and the efficiency of the rectenna can be calculated from the measured voltages. The measured DC output voltage and the total power transmission efficiency at 655 MHz are depicted in Fig. 31. It is noticed that when the transmitted power is 10 dBm, the rectifier indicated its highest efficiency, with a resulting output voltage of 0.25 V.

**VI. LIMITATIONS OF THE STUDY**

In this study, the AR of the proposed antenna when it is implanted in the muscle tissue degrades to exceed 3 dB due to the significant difference between the permittivity of the FR4 substrate, Teflon capsule, and the surrounding environment (inside the muscle tissue). One way to improve the AR is to apply a biocompatible material with higher dielectric constant as the capsule instead of Teflon, as proposed in Table 3. However, for this study, such materials were not available to be evaluated experimentally. The radiation pattern of the proposed antenna cannot be independently measured with satisfactory accuracy due to the radiation from the cable connected...
to the antenna during measurements. The radiation of the cable is comparable to the antenna gain, which considerably affects the accuracy of the pattern measurements. This issue is a measurement limitation and is beyond control in practice. However, in real implementations, the rectifier circuit can be placed directly behind the antenna structure, as shown in Fig. 30, which eliminates the need for the cable when operating as a complete system with the proposed antenna.

VII. CONCLUSION

This paper presents a compact and implantable circularly polarized UHF antenna at 655 MHz. A significant reduction in the antenna’s dimensions was achieved using the Archimedean spiral structure in combination with a shorting polarized UHF antenna at 655 MHz. A significant reduction in Fig. 30, which eliminates the need for the cable when connected to the proposed SPIFA antenna.

The proposed antenna features improved radiation efficiency even when applied deep into the human tissue (50 mm depth, −20 dB) in comparison to literature. Since the antenna is applied for an implantable device, the tradeoff between the size, the efficiency, and the bandwidth are significant limitations of the antenna design. The principles of the antenna design, operation, and its size and matching limitations are first discussed. This is followed by the modeling of an equivalent circuit for the antenna. Next, the effects of different human body tissues, its depth inside the tissues and the implementation of biocompatible materials on the antenna performance are investigated in terms of axial ratio and reflection coefficient. To validate simulation results, the proposed antenna performance is measured in free space, in a phantom with electrical properties similar to muscle, and then in a cow’s fat and muscle tissues. Finally, the antenna operation is also demonstrated for wireless power transfer by integration with a rectifier. Results indicate good agreements between simulations and measurements, validating the potential of the proposed antenna for application in implantable devices with satisfactory performance.

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