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Impedance Matching Techniques in Half Wave Dipoles for Radio Frequency Energy Scavenging

T. Christiana Erinosho[a*], S. Adeniyi Adekola[a,b], K. Akinwale Amusa[a]

[a]Department of Electrical and Electronics Engineering, Federal University of Agriculture, Abeokuta, Nigeria
[b]Department of Electrical and Electronics Engineering, Federal University of Otuoke, Otuoke, Nigeria

Corresponding Author: *toluerinosho@gmail.com*

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ABSTRACT

The main objective of this paper is the design of transmission line impedance matching networks for radio-frequency (RF) energy harvesting Half-wave Dipole (HD) based Rectenna. When the dipole operates in an environment where the ambient impedance is different from the conventional 73 Ω input impedance of the HD, what occurs is a mismatch leading to a loss of sensitive and harvestable RF energy. Proposed here for that purpose are designs of generalized n-element folded HD applicable for impedance matching at 300 Ω and 600 Ω ambient environments. Furthermore, a quarter-wavelength transformer network design is also projected for the realisation of impedance match between the antenna and ambient environment. In the foregoing example, design parameters for the 300 Ω RF environment (ambient free space), two cases of load impedance are considered (73 Ω and 120 Ω) with characteristic impedance of the ambient environment taken as 377 Ω. Although, the idea of quarter-wavelength transformer and the n-element folded dipoles are well known in the literature, however, the use of those ideas in designing rectennas are exploited in this work as an enhancement to the design accuracy of the rectenna.

1. INTRODUCTION

It is necessary to point out, *ab-initio*, that worldwide today; energy recycling is of foremost interest among engineers and environmentalists. There is therefore, a growing interest on how to capture the wasted ambient electromagnetic energy and have it re-used. Wireless Power Transmission (WPT) has been identified as a way of tackling the problem. However, before WPT can be adopted, there ought to be a means of converting RF power to DC power. Wireless systems radiating electromagnetic energy into free space, abound nowadays. However, a significant amount of that energy is wasted. RF energy harvesting is a means of capturing most of that wasted energy.

RECTENNA, a coined name from the combination of a rectifier and an antenna, is an efficient method for the conversion of RF energy into utilizable DC power (William, 1984). Use is made of rectenna, the most important component in RF energy harvesting, to harvest the wasted ambient energy. It is worth pointing out that the rectenna enjoys an essentially endless period of operation time, and does not require restoration distinct from batteries. Not only that, it is fresh for the environment dissimilar to batteries, with no degradation and contamination of the environment. Dipoles and patch antennas have been employed to fabricate narrowband rectennas (Chin *et al.*, 2015 and Yoo *et al.*, 1992). Investigations have been reported on arrays of rectennas within the wideband 2-35 GHz involving arrays of spiral antennas (Takhedmit *et al.*, 2015; Ho and Kai, 2002;Yoo and Kai, 2002; Zbitou *et al.*, 2008; Epp *et al.*, 2000; Vera *et al.*, 2010; Monti *et al.*, 2010 and Yu-Jiun and Kai, 2006). Use of arrays of antennas for rectenna applications has been addressed in the literature (Shinahara and Matsumato, 1998; Yu-Jiun and Kai, 2006).

The aforementioned authors have tackled different aspects of the rectenna problem. What is of interest to us here is impedance matching of the rectenna for optimum power capturing by the harvester. In most applications, rectenna operates at the far-field region, which is usually very large from RF energy source. By virtue of that long distance between the rectenna and RF energy source, the path loss can be considerable large consequent upon which a weak signal is reaching the rectenna. It is then obvious why the rectenna ought to operate at the maximum possible efficiency so as to capture a large amount of RF signal. In order to capture a reasonable amount of energy that is prone to wastage, impedance matching of the rectenna to ambient environment becomes vital.

A variety of rectenna structures have attracted the attention of many investigators in recent years. As examples, the question of arrays of broadband rectennas was tackled by Hagerty *et al.* (2004) while wideband composite rectennas was investigated by Binh *et al.* (2002). Takhedmit *et al.* (2015) proposed an inexpensive rectenna that operates at 2.45 GHz while a hybrid rectenna operating at 35 GHz has been reported in the literature (Ho and Kai, 2002; Yoo and Kai, 2002). Other models of rectennas such as compact rectenna design abound in the literature (Chin *et al.*, 2015; Yoo *et al.*, 1992). Records of experimental investigations on rectennas have also been documented (Young-Ho *et al*, 2002) as well as a host of other works on rectenna designs (Monti *et al*., 2011; Stranner and Kai, 2000).

A rather fascinating contribution to rectenna work is the provision of impedance matching network between the antenna and rectifier (Keyrouz and Visser, 2013; Viser and Vullers, 2010). Reducing the size of the rectenna harvester involves removal of the matching network (Stoopman *et al.*, 2013) in which case, the output impedance of the antenna is made to be identically equal to the input impedance of the rectifier. Interestingly, this is in contra-distinction to what we are proposing in this paper. Our proposal involves a direct matching of the rectenna to the ambient environment. This is because the RF signal reaching the rectenna tends to be very weak, as such, every available incident RF energy ought to be captured for greater efficiency.

The approach adopted here consists of a variety of impedance matching techniques using a folded Half-wave Dipole (HD), generalized n-element folded HD, three-element folded HD and quarter-wavelength impedance matching. These are proposed for rectennas operating at different ambient environments.

The organization of this paper is as follows: section 2 provides the background theory leading to the choice of the dipole diameter as $1 \times 10^{-5}\lambda$ with associated radiated fields as well as its radiation resistance. It is in section 3, that all the designs of interest bordering on impedance matching of HD rectennas are carried out. Also discussed in that section are proposals on two types of impedance matching of HD rectennas. Section 4 provides the concluding remarks.

2. METHODOLOGY

Dipole antenna is the most versatile antenna in wireless applications. A review of its basic theory would be useful in recalling the essential concepts relevant to RF energy scavenging. A typical representation of far-field approximations of a dipole antenna is illustrated in Figure 1. Through a recent numerical evaluation of the dipole characteristics, it has been established that the following parameters are relevant to a better understanding of its radiation mechanism (Adekola *et al*., 2019).

$$(a, f, N_s) = 5\times10^{-6}\lambda, \ f = 550\,MHz, \ N_s = 15 \quad (1)$$

provided a stands for the wire radius, f is the operating frequency and N_s is the number of segments employed in the Method of Moments (MoM), as a computational tool, respectively. By virtue of (1), the diameter of the wire, symbolized by α, assumes a value given as

$$\alpha = 2a = 1x10^{-5}\lambda \quad (2)$$

Results of several numerical evaluations also reported in (Adekola *et al*., 2019), reveal the following specific numerical values at resonance:

$$(l_r, R_{in}) = (0.4889\,\lambda, 72.38\,\Omega) \quad (3)$$

where l_r, R_{in} are the resonant electrical length and input resistance of the dipole, respectively.

From the foregoing, it is reasonable to suggest that (2) is a descriptor of a thin dipole antenna where the diameter tends to zero, consequent upon which the following current can be reasonably assumed to be flowing in the wire:

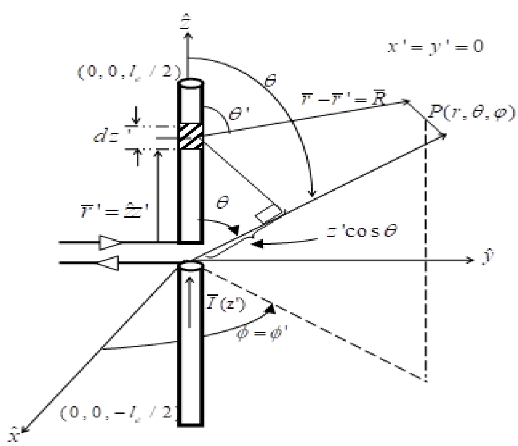


Fig. 1: Diagrammatic representation of far-field approximations for dipole antenna

$$I(x'=0, y'=0, z') = \begin{cases} \hat{z}\,I_n \sin\left[k\left(\frac{l}{2} - z'\right)\right], & 0 \le z' \le {l_e}/{2} \\ \hat{z}\,I_n \sin\left[k\left(\frac{l}{2} + z'\right)\right], & 0 \le z' \le {l_e}/{2} \end{cases} \quad (4)$$

where I_n and l_e are the values of the peak current and electrical length of the wire, respectively. The geometry depicted in Figure 1 describes the far-field diagrammatic representation of the variables.

2.1 Radiated Fields by the Half-wave Dipole

When use is made of the vector magnetic potential, in which (4) is employed as the excitation and by virtue of the usual far-field approximations using Figure 1, one readily obtains, approximately, the E_θ -component of the field radiated by the dipole (Balanis, 2003) as:

$$E_\theta = jI_n\eta \frac{e^{-jkr}}{2\pi r}\left[\frac{\cos\left(\frac{kl_e}{2}\cos\theta\right)-\cos\left(\frac{kl_e}{2}\right)}{\sin\theta}\right] \tag{5}$$

and the H_ϕ is deducible as

$$H_\phi \approx \frac{E_\theta}{\eta} \approx jI_n \frac{e^{-jkr}}{2\pi r}\left[\frac{\cos\left(\frac{kl_e}{2}\cos\theta\right)-\cos\left(\frac{kl_e}{2}\right)}{\sin\theta}\right] \tag{6}$$

2.2 Average Radiated Power Density

Average radiated power density symbolized by W_{av} assumes a form expressed as

$$\bar{W}_{av} = \frac{1}{2}\mathrm{Re}\left[\bar{E}x\bar{H}^*\right] \tag{7}$$

which when simplified further gives

$$\bar{W}_{av} = \frac{1}{2\eta}\left|E_\theta\right|^2 \hat{\mathrm{r}} \tag{8}$$

2.3 Radiated Power

If P_{rad} represents radiated power, one then obtains

$$P_{rad} = \int \bar{W}_{av}.\, d\bar{s} \tag{9}$$

where

$$d\bar{s} = \hat{r}r^2 \sin\theta d\theta d\varphi \tag{9a}$$

Substituting (8) and (9a) in (9) gives

$$P_{rad} = \int_{\varphi=0}^{2\pi}\int_{\theta=0}^{\pi} \frac{1}{2\eta}\left|E_\theta\right|^2 \hat{r}.\left(\hat{r}r^2 \sin\theta d\theta d\varphi\right) \tag{10}$$

It can be easily shown that (10) evaluates to the following value (Balanis, 2003)

$$P_{rad} = 36.56 I_n^{\;2} \tag{11}$$

2.4 Radiation Resistance

It is well-known that the radiated power is expressible in terms of the radiation resistance denoted by R_{rad} , in a form rendered as

$$P_{rad} = \frac{1}{2}I_n^{\;2}R_{rad} \tag{12}$$

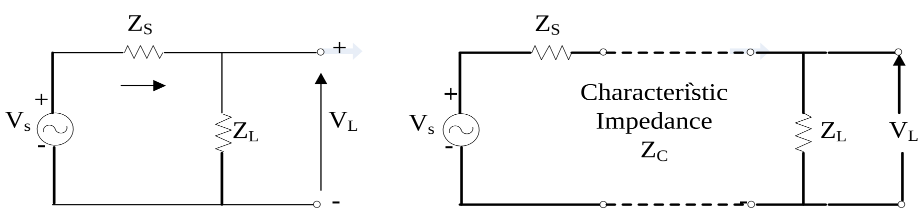
Since (11) and (12) are essentially the same, then that identity provides a value for R_{rad} as

$$R_{rad} = 73\,\Omega \tag{13}$$

One can see at once that the direct integration process carried out here leads us to obtaining the radiation resistance of 73 Ω. Thus, when a conventional half-wave dipole is used for RF energy harvesting, it presents an impedance of 73 Ω to the ambient environment where RF energy is to be harvested. Consequently, for maximum energy capturing, there ought to be impedance matching at the point where the antenna first captures the RF energy. This is addressed in subsequent sections of the paper.

2.5 Design of Impedance Matching for HD Rectennas

Explored here is the design of impedance matching for HD rectennas with applications specific to RF energy harvesting. When designing impedance matching network in electronics, two things ought to be borne in mind namely; maximizing power transferred from the source and minimizing the signal reflected from the load. As depicted in Figures 2(a) and 2(b), what these means are, considerations are given to the output impedance of source as well as the input impedance of the load. When the load impedance is greater than the source impedance, such a situation is termed impedance bridging or bridging for short. In RF energy harvesting, impedance bridging must be avoided as this has deleterious effect on successful RF signal energy harvesting. This is untenable simply as only the load can be varied in RF harvesting via the use of suitable impedance matching circuit.



(a) (b)

Fig. 2: Impedance matching (a) source and load impedance (b) equivalent transmission line with characteristic source and load impedances

If source and load impedances are complex conjugates, we then have

$$Z_s = Z_l{}^* \tag{14}$$

where the asterisk denotes a complex conjugate. In that case, maximum power transfer occurs. Further comments concerning bridging are desirable here. Bridging is inappropriate in RF energy scavenging because it produces a situation whereby power is reflected back to the source in-between the large and small impedances. Reflection produces a standing wave, provided that a reflection occurs at the two ends of the transmission line, thus resulting into additional power loss which can create what is known as frequency-dependent energy loss.

2.5.1 On Application-Specific Rectenna

Examined here is an overview of the impedance matching for rectennas with applications-specific to RF energy scavenging. It has been derived and validated in section two of this paper that the input impedance of the half-wave dipole antenna is approximately 73 Ω. When the dipole is operating in environment whose ambient impedance is different from the 73 Ω impedance, there would be a mismatch leading to a loss of energy. Here we investigate that problem and how to alleviate it. There are two environments that are of interest here namely:

i. The 377 Ω environment (free space)
ii. The 600 Ω environment

The 377 Ω environment is the familiar free space in which the harvester antenna normally operates. Environments for the 600 Ω (not so familiar) can be exemplified as an atmosphere with isolation transformers used for audio distribution system or a telephone line. This is used in isolating signal lines with a characteristic impedance of 600 Ω. This may include radio and telephony circuit. Impedance matching can be carried out in one of the four ways listed as:

i. Folded-dipole impedance matching.
ii. Quarter wavelength impedance matching.
iii. Stub matching.
iv. Matching by balun.

Of the four, the first two are of interest to us and these are considered in what follows.

2.5.1.1 Quantitative Analysis of Folded HD

Illustrated in Figure 3 is the folded dipole geometry made of two dipoles that are parallel to one another. What eventually emerges is a loop connected at both ends of the wire. It is characterized by a separation distance, d_e, much less than the wavelength denoted by λ ($d_e << \lambda$), as well as the length symbolized by l_e ($d_e << l_e$, where $l_e = 0.5\lambda$). Its input impedance is much greater than that of a single dipole and also enjoys a higher bandwidth. It is applicable in a 300 Ω environment which will be demonstrated later in this paper.

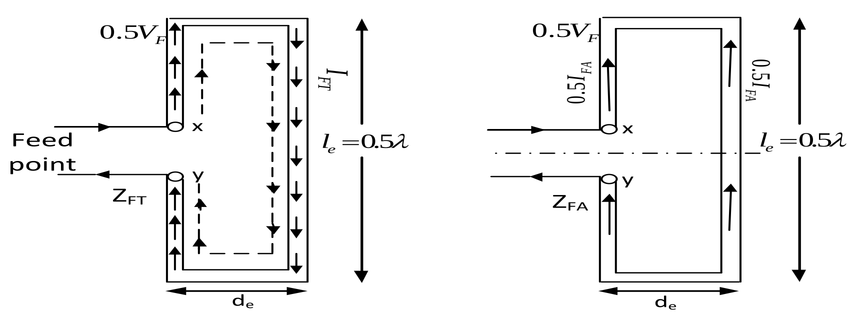


Fig. 3: Equivalent modes of a folded dipole in (a) transmission line mode (b) antenna mode

2.5.1.2 Operational Procedure of the Folded-Dipole

The procedure is two-fold: the transmission line mode and the antenna mode, as depicted in Figure 3. Let us introduce two parameters symbolized by Z_{FT} and Z_{FA} which stand for the impedance of the transmitting mode and the antenna mode, respectively. In the transmission mode, represented by Figure 3(a), the folded dipole does not radiate because the current flowing in the loop is in opposite directions consequent upon which the electromagnetic fields cancel out. That current flowing in the loop is denoted by I_{FT}. On the other hand, for the antenna mode shown in Figure 3(b), the antenna radiates as the current directions in both arms of the loop are the same and shared accordingly such that current of magnitude $0.5I_{FA}$ flows in each of the two arms of the loop. All parameters are clearly displayed in Figure 3(a) and 3(b) for clarity.

2.5.1.3 Input Impedance of the Folded HD

Let the input impedance of the folded HD be denoted by Z_{FD}, we shall derive an appropriate expression for Z_{FD} with reference to Figure 3.

Using Figure 3(a)

$$I_{FT} = \frac{0.5V_F}{Z_{FT}} = \frac{V_F}{2Z_{FT}} \quad (15)$$

and from Figure 3(b),

$$I_{FA} = \frac{0.5V_F}{Z_{FA}} = \frac{V_F}{2Z_{FA}} \quad (16)$$

This has been well articulated elsewhere [23]. The current flowing in a single dipole, symbolized as I_F, is then obtained as

$$I_F = I_{FT} + 0.5 I_{FA} \quad (17)$$

Substitution (15) and (16) in (17) yields

$$I_F = V_F\left(\frac{1}{2Z_{FT}} + \frac{1}{4Z_{FA}}\right) \quad (18)$$

from where the expression for the input impedance, Z_{FD} is arrived at as

$$Z_{FD} = 4Z_{FA}\left[\frac{1}{1 + Z_{FA}/Z_{FT}}\right] \quad (19)$$

since

$$Z_{FT} = jZ_0 \tan\left|\beta(0.5l_e)\right| \quad (20)$$

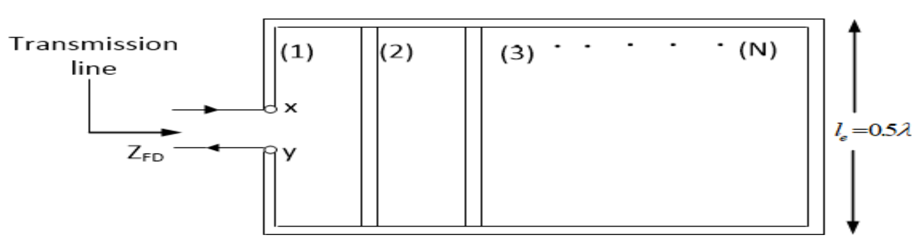
and what is of interest to us is a half-wave dipole, in which case the length, $l_e = \lambda/2$. Further detailed discussions on the above development are available in (Balanis, 2003). Substituting $l_e = \lambda/2$ in (20) gives $Z_{FT} = \infty$ owning to the fact that $\tan(\pi/2) = \infty$, consequent upon which (19) is re-cast in a different form rendered as

$$Z_{FD} = 4Z_{FA} \quad (21)$$

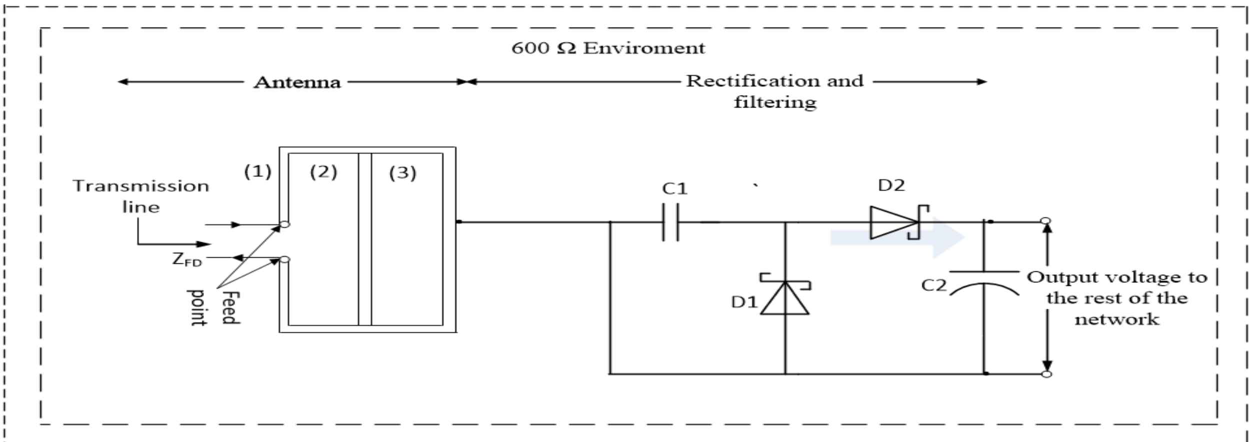
But for half-wave dipole, $Z_{FA} = 73\ \Omega$, thus, (21) evaluates to

$$Z_{FD} \approx 300\ \Omega \quad (22)$$

The ambient free space environment has intrinsic impedance that is approximately equal to 377 Ω, it therefore follows that, the folded dipole whose input impedance is 300 Ω matches that of ambient environment. With that value of impedance, maximum RF energy can be captured by the rectenna for storage and subsequent use by the sensors or for any other applications of interest. This explains the rationale behind the proposal of folded dipole whose input impedance is 300 Ω for rectenna design.



(a)



(b)

Fig. 4: Multi-element folded dipole (a) n-element folded dipole [adapted from (b) impedance matching of a 3-element folded dipole rectenna for the 600 Ω environment

2.5.1.4 Generalized n-element Folded HD

The compact configuration depicted in Figure 4(a) is an n-element folded dipole. By virtue of derivations culminating into (22), expression identified as (21) can be re-cast as

$$Z_{FD} = 2^2 Z_{FA} \qquad (23)$$

from which it can be inferred, without loss of generality, that (23) conforms with a general expression written as

$$Z_{FD} = n^2 Z_{FA} \qquad (24)$$

Equation (24) expresses a generalized expression for the determination of the value of the impedance required to match an n-element folded dipole. By the provision of (24), input impedances of a variety of other environments are easily deducible. Two of such cases are highlighted in the ensuing discussions.

3. RESULTS AND DISCUSSION

3.1 Deduction for 300 Ω and 600 Ω ambient environments

It will be recalled that the situation for the 300 Ω environment through a direct formulation of governing equations has been carried out in section (2.5.1.3) of this paper with associated diagrammatic representation illustrated in Figure 3. Here, we can deduce the same result earlier considered. Thus, if we put n = 2 in (24), we have

$$Z_{FD} = 2^2 Z_{FA} \qquad (25)$$

which readily gives what has been reported in (21) as

$$Z_{FD} = 4 Z_{FA} \qquad (26)$$

and remembering that, (26) then evaluates to

$$Z_{FD} \approx 300\,\Omega \qquad (27)$$

which is the same as (22) obtained earlier.

From what has been described earlier concerning the 600 Ω RF environment, and by virtue of (24), one can realize impedance matching for the 600 Ω RF environment by substituting n = 3 and Z_{FA}= 73 Ω in (24) such that

$$Z_{FD} = 3^2 \; x \; 73 = 657 \; \Omega \qquad (28)$$

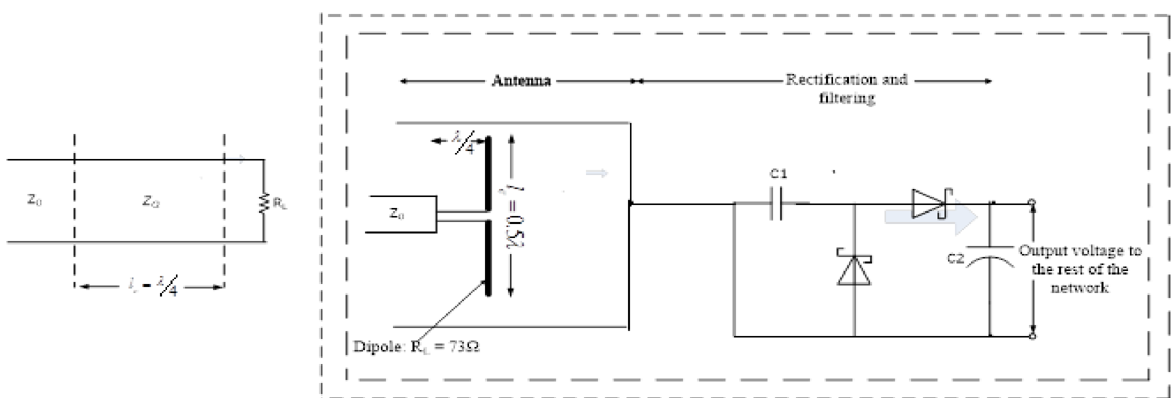
It can be seen at once that (28) with Z_{FD} = 657 Ω clearly matches with the 600 Ω RF environment for maximum energy capture by a 3-element folded dipole. Figure 4(b) portrays the corresponding rectenna with the rectification and filtering circuit. What is of interest now is the quarter-wavelength impedance matching, discussed in the ensuing section.

3.2 Quarter Wavelength Impedance Matching: Quarter-wave Transformer

Simply defined, a quarter-wave transformer is employed in impedance matching so as to minimize the energy reflected when a transmission line is connected to a load. The circuit schematic is depicted in Figure 5(a). It can easily be shown that (Pozar, 2004)

$$Z_Q = \sqrt{Z_0 R_L} \qquad (29)$$

In Figure 5(b), Z_0 = 377 Ω, Z_Q =165.89Ω, R_L = 73Ω, provided Z_Q is the impedance of the quarter-wave transformer, Z_0 is the characteristic impedance of the transmission line while R_L is the load resistance. These parameters are clearly displayed in Figure 5(a) for clarity. Examples are presented in what follows to demonstrate how impedance matching is accomplished.



(a) (b)

Fig. 5: Impedance matching with a quarter-wave transformer (a) a quarter-wave transformer circuit (b) circuit for impedance matching of a dipole rectenna using quarter-wave transformer with rectification and filtering circuit.

3.2.1 377 Ω Environment

Let us examine the following numerical quantities for (Z_0, R_L) such that

$$Z_0 = 377 \, \Omega$$

$$R_L = 73 \; \Omega$$

Consequently, the quarter-wavelength transformer that is required to match the two loads using (29) is determined as:

$$Z_Q = \sqrt{377x73} \quad \cong \quad 165.89\Omega$$

This means a 165.89 Ω quarter-wavelength will enable Z_0 and R_L be matched to avoid reflections. The rectenna implementation of this design is depicted in Figure 5(b) where the rectenna is matched with the ambient environment for maximum RF energy input.

3.2.2 600 Ω Environment

In this case, we have $Z_0 = 600\,\Omega$ and $R_L = 73$ Ω, therefore,

$$Z_Q = \sqrt{600x73} \quad \cong \quad 209.28\Omega$$

Again, a quarter-wavelength of 209.28 Ω will match the two impedances given as 600 Ω and 73 Ω. A rectenna implementation of this design similar to Figure 5(b) can be achieved.

4. CONCLUSION

Highlighted here are designs of transmission line impedance matching networks for RF energy harvesting using HD antennas. When the dipole operates in an environment where the ambient impedance is different from the conventional 73 Ω input impedance of the HD, what occurs is a mismatch leading to a loss of sensitive and harvestable RF energy. Presented here are the designs of antenna-mode folded HD and a generalized n-element folded HD, suitable for applications in $(300\,\Omega, 600\,\Omega)$ ambient environments for input impedance mismatch compensation. Generalized expression for impedance matching network for applications involving HD is derived. Furthermore, for a generalized RF ambient environment, a quarter-wavelength transformer network design is equally projected. Applications of ideas presented in this work in rectenna designs will enhance accuracy of the rectenna and consequently improve the efficiency of the overall system.

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