The linearly polarized waveguide-fed dipole-like antenna

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The modified waveguide-fed dipole-like antenna is designed using the induced current method. To excite the antenna arms, the two equal in magnitude and 180° and out-of-phase excitations are created using the standard rectangular waveguide. Since the area around the radiating arms is free of any conductors (exception for a finite ground plane), the antenna radiation is of high polarization purity. The stand-alone antenna operating near 1.7 GHz was fabricated and tested to verify the design validity. Both the far-field pattern and the return loss level show good agreement with the designed parameters. Such antenna is optimal for using as a building block in the linearly polarized assemblies as well as the stand-alone radiators.

\textbf{Keywords:} dipole-like antenna; waveguide feed; linear polarization; induced current method

1. \textbf{Introduction}

The dipole antenna is one of the most simple radiators with the appropriate radiation pattern directivity. Then, if the dipole antenna is excited by a rectangular waveguide operating at the dominant $TE_{10}$ mode, the split coaxial balun, commonly, is positioned perpendicularly to a broad guide wall onto its centerline.\textsuperscript{[1]} The input port of the balun is excited by a coaxial probe placed inside the waveguide at the distance of $\lambda_g/4$ away from its bottom, where $\lambda_g$ is the guide wavelength. Waveguides are the traditional feeding structures for stand-alone and phased array antennas due to their low losses at high frequencies that results in high radiation efficiency. As a rule, such waveguide-fed antennas and arrays were designed using traditional antenna/array synthesis methods.\textsuperscript{[2]}

In the algorithms, the most arduous component for the design is a split coaxial balun that simultaneously plays the roles of a lossless impedance matching network and dipole support. The balun itself is composed of a coaxial line whose outer conductor, at one end (radiating end), has two opposite open quarter-wave slots. The two sides of the radiating end form the electrically balanced terminals with the center conductor connected to one side. Such balun construction was a successful discovery in the field of classical center-fed dipole excitation over the VHF and UHF bands.\textsuperscript{[3–6]} The split coaxial baluns designed for symmetric wire dipole antennas were earlier briefly reviewed.\textsuperscript{[7,8]} However, the split coaxial balun disposed as the dipole support above

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the rectangular waveguide decreases the polarization purity of HF radiation and leads to high manufacturing complexity.

In the present study, to increase radiation polarization purity and simplify antenna fabrication, the ends-fed dipole radiator (EFDR) is proposed. The radiator is excited without a balun and through the rectangular waveguide using two coaxial probes only. The first and second probes are placed inside the waveguide at the distances of $3\lambda_g/4$ and $\lambda_g/4$ away from the bottom, respectively. The probes are offset from the narrow waveguide wall by distance $x_1$, as shown in Figure 1. The advantages of this construction are high antenna efficiency, considerable mechanical strength, and great accuracy achievable at the design and realization stages.

This study is aimed at two main things. First, it is principle to obtain the closed-form expression for the radiation resistance of the EFDR placed at a different height above the ground surface. This resistance is found using the earlier established current distribution along the radiator.[9] Second, the design procedure and the corresponding optimization algorithm are developed for the probe-fed dipole-like antenna. The adopted synthesis technique, based on the desired specifications of radiation properties and complex input impedance, is used to determine the geometrical parameters of the proposed antenna mounted on the metallic planar surface. Up to now, the investigations of this type of antenna are very scarce. Thus, the theoretical evaluation of the EFDR above the ground is presented in Section 2 and the numerical results given in Section 3 are used to verify the analysis. Finally, the derivation of the closed-form expression obtained for the EFDR radiation resistance above the ground is reported in the Supplementary Materials.

2. EFDR input impedance

As it is shown in Figures 1 and 1S, the EFDR contains two collinear cylindrical wires of radii $a_w$. The wires are electrically thin and the relation $a_w \ll \lambda$, where $\lambda$ is the free-space wavelength, is fulfilled. The internal ends of the wires are directed one to the other with a small gap, whereas the opposite-phase excitations are conducted to
external wire ends. Both wires, each of length $l$, are centered and oriented along the $z$ axis. The EFDR itself radiates in the homogeneous outside space characterized by permittivity $\varepsilon$ and permeability $\mu$ with the corresponding wave number $k = \omega \sqrt{\varepsilon \mu}$ and intrinsic impedance $\eta = \sqrt{\mu/\varepsilon}$.

To analyze the antenna performance near an infinite conductive plane, virtual sources (images) will be introduced to account for the reflections.[8] These are imaginary sources, which, when being combined with the real sources, form an equivalent system. To begin the discussion, let us assume the elementary field sources being located at a height $h$ above the ground. From Figures 1 and 1S, one can conclude that the real $R_{in}$ and imaginary $X_{in}$ parts of input impedance $Z_{in} = R_{in} + jX_{in}$ can be expressed by the induced emf method using the results related to the stand-alone EFDR, as was earlier reported.[9] To match the antenna, the input impedance should be matched. According to Ref. [8], the input impedance can be written as

$$Z_{in} = Z_{11} + (I_2/I_1)Z_{12},$$

where $Z_{11} = R_{11} + jX_{11}$ represents the input impedance of EFDR radiating in an infinite medium, $Z_{12} = R_{12} + jX_{12}$ is the mutual impedance between two identical EFDRs arranged in the side-by-side configuration, $I_2/I_1$ is the current ratio in accordance to the image theory. The side-by-side configuration was described in detail, as related to two classical center-fed dipoles.[8] As for a horizontal electric radiator (i.e. for a dipole or for an EFDR), the current in the image must flow in the direction opposite to that of the actual radiator current, and this ratio is equal to $-1$. Therefore, $Z_{in} = Z_{11} - Z_{12}$. Referring to the results obtained in Ref. [9] for the EFDR radiation in an infinite medium, the following expressions can be written for a free space

$$R_{in} = \frac{R_{11} - R_{12}}{\sin^2(x)}, \quad X_{in} = \frac{X_{11} - X_{12}}{\sin^2(x)},$$

$$R_{11} = R(d = ka_w), \quad X_{11} = X(d = ka_w),$$

$$R_{12} = R(d = 2kh), \quad X_{12} = X(d = 2kh),$$

where

$$R = 60\{\sin^2(x)\left(\sin^2\left(\sqrt{4x^2 + d^2}\right) - \sin^2(d)\right)$$

$$+ 3\text{Ci}(d) + 0.5\text{Ci}\left(\sqrt{4x^2 + d^2} + 2x\right)$$

$$+ 0.5\text{Ci}\left(\sqrt{4x^2 + d^2} - 2x\right)$$

$$- 2\text{Ci}\left(\sqrt{x^2 + d^2} + x\right) - 2\text{Ci}\left(\sqrt{x^2 + d^2} - x\right)\},$$

$$X = 60\{\sin^2(x)\left(\cos^2\left(\sqrt{4x^2 + d^2}\right) - \cos^2(d)\right)$$

$$- 3\text{Si}(d) - 0.5\text{Si}\left(\sqrt{4x^2 + d^2} + 2x\right)$$

$$- 0.5\text{Si}\left(\sqrt{4x^2 + d^2} - 2x\right)$$

$$+ 2\text{Si}\left(\sqrt{x^2 + d^2} + x\right) + 2\text{Si}\left(\sqrt{x^2 + d^2} - x\right)\}.$$
\[
\sin'(x) = \sin(x)/x, \quad \cos'(x) = \cos(x)/x, \quad x = kl,
\]

\(\text{Si}(x)\) and \(\text{Ci}(x)\) are the sine and cosine integrals, respectively.

If radius \(a_w\) is defined as \(a_w = l/10^3\), then the result of (2), as given by the induced emf method (for \(R_{in}\) only), is equal to that (denoted by \(R_{in}^{hem}\)) given by the integration over a closed hemisphere in the far-zone [10–14] (Supplementary Materials)

\[
R_{in}^{hem} = \frac{60}{\sin^2(\alpha)} \left\{ 1.5\gamma + \ln(2\sqrt{\alpha^2}) + \frac{\text{Ci}(4\alpha)}{2} \right. \\
- 2\text{Ci}(2\alpha) - 3\text{Ci}(\beta) \\
+ 2 \left[ \text{Ci}\left(\sqrt{\alpha^2 + \beta^2} - \alpha \right) + \text{Ci}\left(\sqrt{\alpha^2 + \beta^2} + \alpha \right) \right] \\
- \frac{1}{2} \left[ \text{Ci}\left(\sqrt{4\alpha^2 + \beta^2} - 2\alpha \right) + \text{Ci}\left(\sqrt{4\alpha^2 + \beta^2} + 2\alpha \right) \right] \\
+ s^2 \left[ \sin^2(2\alpha) + \sin^2(\beta) - \sin^2(4\alpha^2 + \beta^2) - 1 \right] \right\}
\]

where \(\beta = 2kh\). When \(l/\lambda = 0.43575\) and \(h/\lambda = 0.25\), it is obtained

\[
R_{in} = R_{in}^{hem} = 1515 \, \Omega
\]

In case of coincidence, the quantities may be considered as mutual verification of the calculation. In addition, this value will be used as an appropriate initial parameter in the full-wave simulation performed below.

3. Design procedure, numerical, and experimental results

In order to verify the design procedure, a set of antennas with a different size and operating frequency has been designed and analyzed using the WIPL-D package.[11] The results show, for all tested cases, a good agreement with the design objectives, both in the far-field pattern and the frequency response. As an example, the antenna operating at 1.7 GHz is described in detail.

First, the length difference between the output probes which excite the far ends of the radiator is equal to \(\lambda_g/2\), as shown in Figure 2S. The condition means that the required 180° phase difference between the excitations (i.e. between the adjacent magnetic loops of the dominant mode in the rectangular waveguide) is automatically satisfied. Additionally, this leads to the condition for the overall radiator length \(2l = \lambda_g/2\).

Second, if the radiator is located above the guide (i.e. above the ground) at height \(h = \lambda/4\), the coaxial stubs containing the air filling play the role of quarter-wave transformers. Since the mutual impedance \(Z_{12}\) between two radiators is the function of distance \(2h\) only (i.e. the induced emf method cannot take into account the wire radius \(a_w\) in the mutual impedance,[8]) the overall imaginary part \(X_{in}\) of the input impedance (2) must be zero when choosing the appropriate wire radius \(a_w\) in the expression for imaginary part \(X_{11}\) of impedance \(Z_{11}\). As a result, the overall input impedance \(Z_{in}\) becomes purely resistive: \(Z_{in} = R_{in} = R_{11} - R_{12}\). Under the appropriate selection of wire radius \(a_w\), this leads to the following condition for the waveguide broad wall width:

\[
a/\lambda = 0.61.
\]
Such an approach paves the way to implement both output coaxial probes which are placed inside waveguide \( x_1 \) mm away from its narrow wall (Figure 1S). To continue the design of the antenna located in free space and operating at 1.7 GHz, the standard WR-430 waveguide (\( a = 109.2 \) mm, \( b = 54.61 \) mm) with the 2.03 mm wall thickness is considered. The corresponding wavelengths and dimensions are (in mm)

\[
\begin{align*}
\lambda &= \frac{a}{0.61} = 179, & \hat{\lambda}_g &= 312, \\
l &= \frac{\hat{\lambda}_g}{4} = 78, & h &= \frac{\lambda}{4} \approx 45
\end{align*}
\]

Since \( l/\lambda = 0.43575 \) and \( h/\lambda = 0.25 \), from (5), the real part \( R_{in} \) of the EFDR above the ground is equal to 1515 \( \Omega \). Whereas the imaginary part \( X_{in} \) is minimized (\( |X_{in}| \approx 0.001 \Omega \)) through the appropriate choice of wire radius \( a_w^* \): \( a_w^* = 1.361 \) mm. This means that \( X_{11} \) [i.e. \( X(d = ka_w^*) \)] is equal to \( X_{12} \) [i.e. \( X(d = 2kh) \)]. Since the radiator is connected to the waveguide through two coaxial probes, each probe is terminated by resistance \( R_{in}/2 = 757.5 \) \( \Omega \). Then, the resistances are transformed to the appropriate value by the air-filled coaxial quarter-wave transformer of geometrical length \( h \) (the corresponding electrical length is equal to 90°) and characteristic impedance \( W_h \). For example, if the appropriate value is equal to 9 \( \Omega \), then \( W_h = \sqrt{9R_{in}/2} = 82.6 \) \( \Omega \). This requires that the coaxial probe length \( l_p \) inside the waveguide and the distance \( x_1 \) from the narrow wall (Figure 1S) should be found numerically using one of the 3D EM optimizers. In the considered case, the optimization by WIPL-D [11] leads to the following final dimensions (Figures 1S and 2S): \( l_p = 37.8 \) mm and \( x_1 = 6.4 \) mm. Since the length difference between the probes is equal to \( \hat{\lambda}_g/2 \) and each probe is surrounded by the adjacent magnetic loops of the dominant mode, the required phase difference 180° between the EFDR excitations is automatically satisfied. If the radius \( r_{in} \) of the inner conductor inside the air-filled quarter-wave transformer is equal to the wire radius (i.e. \( r_{in} = a_w^* = 1.36 \) mm), then the inner radius \( r_{out} \) of the outer cylindrical conductor is expressed as

\[
r_{out} = r_{in} \exp\left(\frac{82.6}{60}\right) = 5.4 \text{ mm}
\]

Note, that the distance between the waveguide flange and nearby coaxial probe may be selected arbitrarily.

Finally, the comparison of the simulated and measured parameters of the antenna with an extended ground plane is shown in Figures 2–4. The simulated results were obtained by the full-wave analysis of the complete antenna. The dependence of the calculated and measured return losses at the waveguide input on frequency is shown in Figure 2. The impedance matching was measured using an Agilent N5241A Performance Network Analyzer (PNA–X). The experimental results show that the input impedance bandwidth is 1.8% for the level \( S_{11} \leq -10 \) dB in the range from 1.68 to 1.71 GHz. The slight difference between the numerical and measured values may be due to the round corners of the waveguide, as well as the discontinuity of the coaxial stubs, fabrication imperfection and manufacturing interconnections. These features are not taken into account at simulation. As it is evident from Figure 2, the impedance bandwidth of the antenna is relatively small. This is caused by the sharp frequency variation of EFDR input impedance in contrast to that of the classical center-fed dipole.

The antenna radiation patterns were measured at the indoor far-field laboratory. The dedicated test set-up is arranged inside the anechoic chamber. Namely, the tested antenna is placed above the rotating mount (pedestal) and standard transmitting antenna, like a pyramidal horn, is excited by E8257D PSG Analog Signal Generator.
The calculated and measured $E$- and $H$-plane radiation patterns are shown in Figures 3 and 4. The measured cross-polarization levels are $-13$ dB ($E$-plane cut) and $-17$ dB ($H$-plane cut) below the main lobe in the upper hemisphere. It is seen that the measured radiation patterns are in good agreement with the simulated ones.

The antenna gain is measured using the Russian standard horn antenna for power transmission, where the classical Friis Transmission Equation can be applied for the gain calculation [8]:

$$\frac{P_r}{P_t} = |S_{21}|^2 = \left(1 - |\Gamma_r|^2\right) \left(1 - |\Gamma_t|^2\right) \left(\frac{\lambda}{4\pi R}\right)^2 G_t G_r$$

where $P_r$ is the received power, $P_t$ is the transmitted power, $G_r = 11.25$ dB ($G_r = 13.3$ dimensionless) is the gain of lossless receiving pyramidal horn antenna, $G_t$ is the
unknown gain of above lossless transmitting dipole-like antenna, $\lambda = 179$ mm is the free-space wavelength, $R = 1500$ mm is the separation distance between the two transmitting and receiving antennas, and $|S_{21}|^2 = 0.0029 (-25.4 \text{ dB})$ is the measured scattering parameter between part 2 and part 1 of the vector network analyzer. The measured reflection coefficients at the terminals of the receiving pyramidal horn antenna and transmitting dipole-like antenna are $|\Gamma_r| = 0.2$ (the corresponding VSWR = 1.5) and $|\Gamma_t| = 0.05623$, the corresponding return loss is equal to $-25 \text{ dB}$ at $f = 1.7$ GHz in accordance with Figure 2. The test fixture is organized so that the antennas are aligned for maximum radiation between them and are polarization matched. Thus, from (8), one can find $G_t = 2.527$ (dimensionless) or $4.025 \text{ dB}$.

4. Conclusions

In the present study, a new implementation of the waveguide-fed dipole-like antenna is proposed. To realize the required pattern, an EFDR has been described. Such radiation properties allow the radiator to be a qualitative stand-alone module or a suitable candidate for phased arrays. The attractive antenna features are the compact size, when designed for the UHF band, and simpler manufacture. The antenna parameters are promising for new applications in wireless communication systems.

Disclosure statement

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Supplemental data

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