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MULTIBAND ASYMMETRIC BICONICAL DIPOLE ANTENNA WITH DISTRIBUTED SURFACE IMPEDANCE AND ARBITRARY EXCITATION

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A numerical-analytical solution of a problem concerning the current distribution and input characteristics of asymmetric biconical dipole with distributed surface impedance and arbitrary excitation and derived in the thin-wire approximation. Solution correctness is confirmed by satisfactory agreement of numerical and experimental results from literary sources. Numerical results are given for the input characteristics of the dipole in the case of its asymmetric excitation by a point source.
Keywords: *Biconical dipole; Distributed surface impedance; Asymmetric excitation; Current distribution; Input characteristics*
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INTRODUCTION

Among modern communication systems (both mobile and stationary), the leading place is occupied by multi-frequency (multi-channel) structures. The main functional element of such devices are antennas, which differ from each other in their design features, for example, they can be single-element structures (mono-frequency) or multi-element (multi-frequency) structures [1]. At the same time, the operation of a multi-frequency antenna, implemented by expanding the operating band (broadband antenna), in turn, can lead to a significant weakening of its noise-immune properties. Typically, designers take the path of combining several antennas operating at different frequencies into one structure [2-8]. This approach significantly complicates the design of the antenna device and is a difficult factor to overcome on the path to its miniaturization.

In recent decades, a large number of publications have appeared devoted to multi-band printed antennas directly integrated into communication devices, for example [9-19] and references therein. In this case, the electrodynamic characteristics of the antennas are obtained using commercial programs such as ANSYS HFSS, Feko and others. However, in this case, the calculation and further optimization of antennas in order to achieve the specified characteristics lead to a sequential search of a large number of options and, consequently, large amounts of computer resources and computational time.

The use of dipoles with an asymmetric excitation point, i.e. with an arbitrary position along their length, for creating multi-band antennas has been repeatedly proposed by researchers in various publications [5, 7, 20-22]. However, in these literary sources only perfectly conducting dipoles were considered. Another solution is to use a dipole antenna with asymmetric excitation and distributed surface impedance, directly integrated into the body of the communication device [23, 24]. In this case, the frequency response of the antenna may have several resonances that prevent the radiation (receiving) of electromagnetic waves outside the resonant frequency bands.

On the other hand, one of the additional parameters for obtaining the specified electrodynamic characteristics of antennas in the form of a cylindrical dipole can be a change in the radius of the cross section of the dipole along its length. In the case of a linear increase in the radius of the vibrator from the excitation point of the antenna to its ends (biconical dipole), this antenna resonates at a smaller geometric length, and is also more broadband compared to a dipole of constant radius (see, for example, [25–30] and references in them). However, all of them are devoted to calculating the electrodynamic characteristics of perfectly conducting dipoles excited at the geometric center by a concentrated electromotive force (EMF). Also, as is known, to analyze receiving antennas it is necessary to know the current in the scattering dipole excited by the incident electromagnetic wave [31].

The purpose of this paper is to study a multiband antenna for communication systems based on an asymmetric biconical dipole with a distributed surface impedance and arbitrary excitation. Thus, we will combine in one design all the advantages of asymmetric excitation, biconical geometry and the presence of a distributed surface impedance. The antenna can operate in several frequency bands. The antenna characteristics will be modeled by using a numerical-analytical method, known as the generalized method of induced EMF, proposed by the authors earlier in [23, 24].

PROBLEM FORMULATION AND SOLUTION OF THE INITIAL INTEGRAL EQUATION

Let us limit ourselves by the linear law of the radius change along the dipole (Fig. 1), which, in its turn, is rather good approximation and for another dependences $r(s)$, for example, exponential one, at small angles ψ. Let the dipole with distributed internal linear impedance z_i of the $2L$ length and the $r(s)$ variable radius, located in free space, be

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excited by the electrical field $E_{0s}(s)$ of the impressed sources (tangential component). The monochromatic fields and currents depend on time t as $e^{i\omega t}$ ($\omega = 2\pi f$ is the circular frequency, f is the frequency, measured in Hz). At this the dipole stays electrically thin ($kr(s) << 1$, $r(s) << 2L$, $k = 2\pi/\lambda$, λ is the wavelength in free space). Then the integral-differential equation relatively to the $J(s)$ current for the impedance boundary condition on the dipole surface [31]:

$$\left(\frac{\mathrm{d}^2}{\mathrm{d}s^2} + k^2\right)\int_{-L}^{L} J(s')\frac{e^{-ik\tilde{R}(s,s')}}{\tilde{R}(s,s')}\mathrm{d}s' = -\frac{i\omega}{\cos\psi}\left[E_{0s}(s) - z_i J(s)\right] . \tag{1}$$

Here $\tilde{R}(s,s') = \sqrt{(s-s')^2 + r^2(s)}$, $\psi = (\psi_1 + \psi_2)/2$. Note that at $r(s) = const = r_0$, equation (1) transforms into an equation for the current in an impedance dipole of constant radius with a quasi-one-dimensional core $\tilde{R}(s,s') = R(s,s') = \sqrt{(s-s')^2 + r_0^2}$.

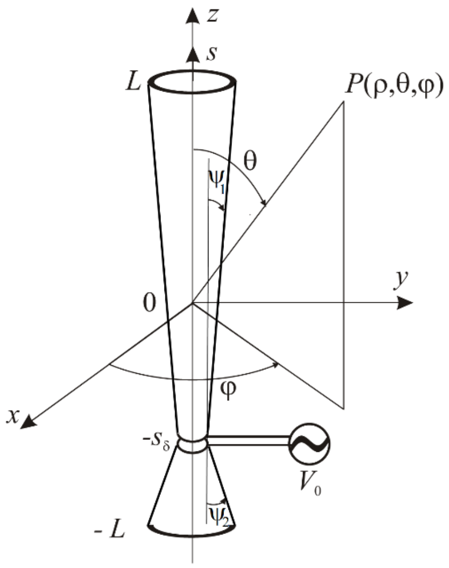

Figure 1. The dipole geometry and accepted designations

The excitation fields of extraneous sources E_{0s} can be divided into symmetric and antisymmetric components relative to the dipole geometrical center. These components marked by the indices "s" and "a" are presented as $E_{0s}(s) = E_{0s}^{s}(s) + E_{0s}^{a}(s)$. Quite naturally, the dipole currents can also have also two components $J(s) = J^{s}(s) + J^{a}(s)$. The initial equation (1) relative the dipole current obtained by using the boundary conditions $J^{s,a}(\pm L) = 0$ has the following form

$$\left(\frac{\mathrm{d}^2}{\mathrm{d}s^2} + k^2\right)\int_{-L}^{L} [J^{s}(s') + J^{a}(s')]\frac{e^{-ik\tilde{R}(s,s')}}{\tilde{R}(s,s')}\mathrm{d}s' = -\frac{i\omega}{\cos\psi}[E_{0s}^{s}(s) + E_{0s}^{a}(s)] + \frac{i\omega}{\cos\psi} z_i [J^{s}(s) + J^{a}(s)], \tag{2}$$

where $E_{0s}^{s,a}(s)$ are the projections of extraneous source fields on the dipole axis, s and s' are the local coordinates related to the dipole axis and surface.

The equation (2) can be represented by the system of two independent integral equations, relative to the unknown currents $J^{s}(s)$ and $J^{a}(s)$

$$\begin{aligned}
&\left(\frac{\mathrm{d}^2}{\mathrm{d}s^2} + k^2\right)\int_{-L}^{L} J^{s}(s')\frac{e^{-ik\tilde{R}(s,s')}}{\tilde{R}(s,s')}\mathrm{d}s' = -\frac{i\omega}{\cos\psi}[E_{0s}^{s}(s) + z_i J^{s}(s)],\\
&\left(\frac{\mathrm{d}^2}{\mathrm{d}s^2} + k^2\right)\int_{-L}^{L} J^{a}(s')\frac{e^{-ik\tilde{R}(s,s')}}{\tilde{R}(s,s')}\mathrm{d}s' = -\frac{i\omega}{\cos\psi}[E_{0s}^{a}(s) + z_i J^{a}(s)].
\end{aligned} \tag{3}$$

The dipole currents can be presented as product of the unknown complex amplitudes $J_n^{s,a}$ and weight functions $f_n^{s,a}(s')$ ($n = 0,1$) as

$$J^{s,a}(s') = J_0^{s,a} f_0^{s,a}(s') + J_1^{s,a} f_1^{s,a}(s'), \quad f_n^{s,a}(\pm L) = 0 . \tag{4}$$

The solution of the equations system (3) can be obtained by the generalized method of induced EMF [23, 24]. To do so, let us multiply the left- and right- hand parts of the equations (8) by the functions $f_n^s(s)$ and $f_n^a(s)$, and integrate the resulting expressions over the vibrator length. Thus, the following algebraic equations system (SLAE) is obtained

$$\begin{cases} J_0^s Z_{00}^{s\Sigma} + J_1^s Z_{01}^{s\Sigma} = -i\omega/(2k\cos\psi)E_0^s, \\ J_0^s Z_{10}^{s\Sigma} + J_1^s Z_{11}^{s\Sigma} = -i\omega/(2k\cos\psi)E_1^s, \end{cases} \quad \begin{cases} J_0^a Z_{00}^{a\Sigma} + J_1^a Z_{01}^{a\Sigma} = -i\omega/(2k\cos\psi)E_0^a, \\ J_0^a Z_{10}^{a\Sigma} + J_1^a Z_{11}^{a\Sigma} = -i\omega/(2k\cos\psi)E_1^a, \end{cases} \tag{5}$$

where ($m = 0,1; n = 0,1$),

$$Z_{mn}^{s,a} = \frac{1}{2k}\left\{-\frac{\mathrm{d}f_m^{s,a}(s)}{\mathrm{d}s}A_n^{s,a}(s)\Big|_{-L}^{L} + \int_{-L}^{L}\left[\frac{\mathrm{d}^2 f_m^{s,a}(s)}{\mathrm{d}s^2} + k^2 f_m^{s,a}(s)\right]A_n^{s,a}(s)\mathrm{d}s\right\}, \tag{6a}$$

$$\tilde{Z}_{mn}^{s,a} = -\frac{i\omega}{2k}\int_{-L}^{L} f_m^{s,a}(s) f_n^{s,a}(s) z_i\,\mathrm{d}s, \quad Z_{mn}^{(s,a)\Sigma} = Z_{mn}^{s,a} + \tilde{Z}_{mn}^{s,a}, \tag{6b}$$

$$A_n^{s,a}(s) = \int_{-L}^{L} f_n^{s,a}(s')\frac{e^{-ik\tilde{R}(s,s')}}{\tilde{R}(s,s')}\mathrm{d}s', E_m^{s,a} = \int_{-L}^{L} f_m^{s,a}(s)E_{0s}^{s,a}(s)\mathrm{d}s. \tag{6c}$$

Then the dipole current can be written as $J(s) = J^s(s) + J^a(s),$ where

$$J^{s,a}(s) = -\frac{i\omega}{2k\cos\psi}\left[\frac{E_0^{s,a}Z_{11}^{(s,a)\Sigma} - E_1^{s,a}Z_{01}^{(s,a)\Sigma}}{Z_{00}^{(s,a)\Sigma}Z_{11}^{(s,a)\Sigma} - Z_{10}^{(s,a)\Sigma}Z_{01}^{(s,a)\Sigma}} f_0^{s,a}(s) + \frac{E_1^{s,a}Z_{00}^{(s,a)\Sigma} - E_0^{s,a}Z_{10}^{(s,a)\Sigma}}{Z_{00}^{(s,a)\Sigma}Z_{11}^{(s,a)\Sigma} - Z_{10}^{(s,a)\Sigma}Z_{01}^{(s,a)\Sigma}} f_1^{s,a}(s)\right]. \tag{7}$$

Note that the approximate numerical-analytical solution (7) of the integral equation (1) is valid for any excitation field $E_{0s}^{s,a}(s)$ of the dipole (δ -generator, plane wave incidence at an angle to the longitudinal axis of the dipole, etc.), as well as for the complex distributed impedance of the dipole.

DIPOLE EXCITED IN AN ARBITRARY POINT

Let the dipole be excited in the point $s = -s_\delta$ by the voltage δ -generator with amplitude V_0, as shown in Fig. 1. Then

$$\begin{aligned} &E_{0s}(s) = V_0\delta(s + s_\delta) = E_{0s}^s(s) + E_{0s}^a(s), \\ &E_{0s}^{s(a)}(s) = (V_0/2)\left[\delta(s + s_\delta) + (-)\delta(s - s_\delta)\right], \end{aligned} \tag{8}$$

where δ is the Dirac delta function. In this case, the dipole current defined by (7) can be written as $J(s) = J^s(s) + J^a(s)$,

$$J^{s,a}(s) = -\frac{i\omega}{2k\cos\psi}V_0\left[\frac{\tilde{E}_0^{s,a}Z_{11}^{(s,a)\Sigma} - \tilde{E}_1^{s,a}Z_{01}^{(s,a)\Sigma}}{Z_{00}^{(s,a)\Sigma}Z_{11}^{(s,a)\Sigma} - Z_{10}^{(s,a)\Sigma}Z_{01}^{(s,a)\Sigma}} f_0^{s,a}(s) + \frac{\tilde{E}_1^{s,a}Z_{00}^{(s,a)\Sigma} - \tilde{E}_0^{s,a}Z_{10}^{(s,a)\Sigma}}{Z_{00}^{(s,a)\Sigma}Z_{11}^{(s,a)\Sigma} - Z_{10}^{(s,a)\Sigma}Z_{01}^{(s,a)\Sigma}} f_1^{s,a}(s)\right], \tag{9}$$

where r_δ and r_L are the radiuses of the dipole in point $s = -s_\delta$ and on its end.

Let us choose the functions $f_0^{s,a}(s)$ obtained after substituting the expressions (8) into the general solution of the equation for the current by the averaging method [23] in the form:

$$\begin{aligned} &f_0^s(s) = \cos\tilde{k}s_\delta \sin\tilde{k}L\cos\tilde{k}s - (1/2)\cos\tilde{k}L(\sin\tilde{k}\,|s - s_\delta| + \sin\tilde{k}\,|s + s_\delta|), \\ &f_0^a(s) = \sin\tilde{k}s_\delta \cos\tilde{k}L\sin\tilde{k}s + (1/2)\sin\tilde{k}L(\sin\tilde{k}\,|s - s_\delta| - \sin\tilde{k}\,|s + s_\delta|), \end{aligned} \tag{10}$$

where $\tilde{k} = k - \dfrac{i\bar{Z}_S[3/2 - r_\delta/(2r_L)]}{2r_L\cos\psi\ln(2L/r_L)}$, $\bar{Z}_S = \bar{R}_S + i\bar{X}_S = \dfrac{2\pi r_L z_i}{Z_0}$ is the distributed surface impedance, normalized on the free space wave impedance, $Z_0 = 120\pi$ Ohm. The current distribution function $f_1^s(s)$ is defined in [32], and the function $f_1^a(s)$ can be found as solution of the integral equation (1), obtained for the case $z_i = 0$ and $\psi = 0$ [20]. These functions can be written as:

$$f_1^s(s) = \cos\tilde{k}s - \cos\tilde{k}L\,, \tag{11a}$$

$$f_1^a(s) = \sin ks - (s/L)\sin kL\,. \tag{11b}$$

The coefficients $Z_{mn}^{s,a}$ $(m = 0,1; n = 0,1)$ in the formulas (9) can be obtained from formulas (6) and $\tilde{E}_0^s = \cos\tilde{k}s_\delta \sin\tilde{k}\left(L - |s_\delta|\right)$, $\tilde{E}_1^s = \cos\tilde{k}s_\delta - \cos\tilde{k}L$, $\tilde{E}_0^a = -\sin\tilde{k}|s_\delta|\sin\tilde{k}\left(L - |s_\delta|\right)$, $\tilde{E}_1^a = \sin ks_\delta - \left(s_\delta/L\right)\sin kL$.

The input impedance $Z_{in} = R_{in} + iX_{in}$ can be presented as:

$$Z_{in}[\mathrm{Ohm}] = \frac{60i}{J_0^s f_0^s(s_\delta) + J_1^s f_1^s(s_\delta) + J_0^a f_0^a(s_\delta) + J_1^a f_1^a(s_\delta)}\,, \tag{12}$$

Then, the voltage standing wave ratio (VSWR) in the antenna feeder with the wave impedance W is equal to:

$$\mathrm{VSWR} = \left(1 + |S_{11}|\right)/\left(1 - |S_{11}|\right), \tag{13}$$

where $S_{11} = \left(Z_{in} - \mathrm{W}\right)/\left(Z_{in} + \mathrm{W}\right)$ is the reflection coefficient in the feeder.

NUMERICAL RESULTS

As an example, let us present the input characteristics (modulus of the reflection coefficient $|S_{11}|$ in the supply feeder line with wave resistance W =50 Ohm) of an asymmetric biconical ideally conducting dipole with dimensions $2L$ =138 mm, r_δ =1 mm, r_L =3 mm (Fig. 2). This choice of the dipole length is due to the condition of the first resonance at the frequency f =0.9 GHz (GSM 900 (880÷960 MHz)). Note that for a regular dipole with a radius r =2 mm, its length would be equal to $2L$ =156 mm. As can be seen from the graphs, an asymmetric biconical dipole (with smaller geometric length than a regular dipole) is also resonant at two or more frequencies. Moreover, this tendency will increase with increasing the angle ψ. Also, if the dipole has a distributed surface impedance of the inductive type, then its resonant length will be even less [24].

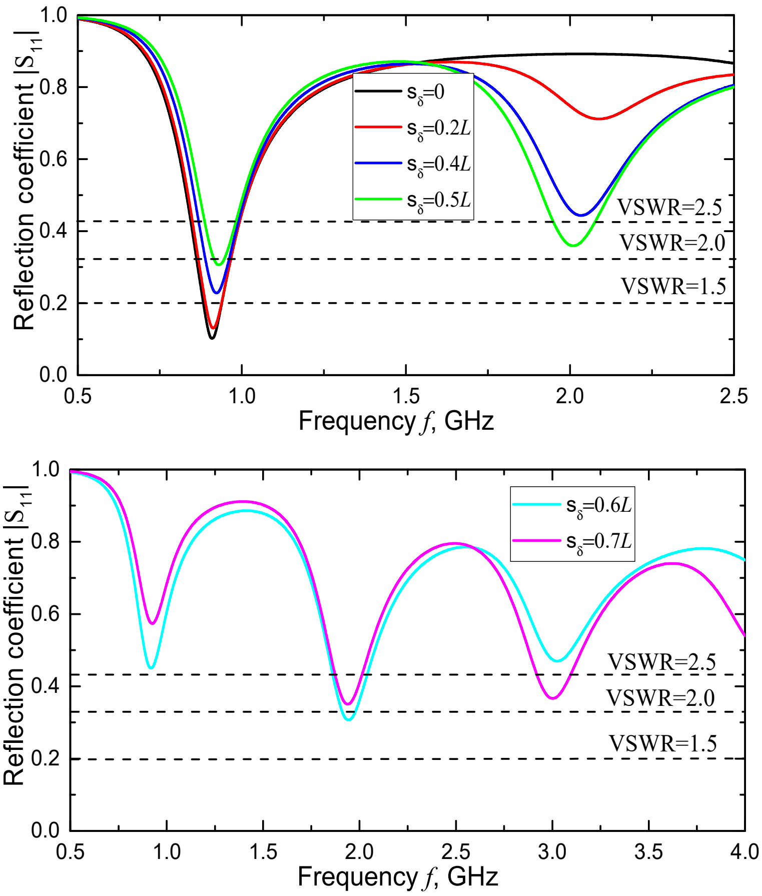

Figure 2. The dependences of the modulus of the reflection coefficient in the supply feeder line versus frequency for different positions of the excitation point s_δ at $2L$ =138 mm, r_δ =1 mm, r_L =3 mm

CONCLUSION

A numerical-analytical solution of a problem concerning a current distribution and input characteristics of the asymmetric biconical dipole with distributed surface impedance and arbitrary excitation derived in the thin-wire approximation. The solution was carried out by the generalized method of induced EMF. The term numerical implies the numerical calculation of some integral terms. Otherwise, approximate analytical expressions are obtained. Solution correctness is confirmed by satisfactory agreement of numerical and original experimental results and well-known literary sources. The characteristic property of the antenna is the possibility of resonant tuning to the selected frequencies (depending on the geometric and electro-physical parameters of the dipole), which does not decrease the noise-resistant properties as compared with broadband antennas. Numerical results are given for the input characteristics of the dipole in the case of its asymmetric excitation by a point source. Analysis of electrodynamic characteristics of the proposed dipole antenna has proved the possibility of practical applications of this antenna for multiband portable radio stations, electronic gadgets and base stations and other antenna systems.

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БАГАТОСМУГОВА АСИМЕТРИЧНА БІКОНІЧНА ДИПОЛЬНА АНТЕНА З РОЗПОДІЛЕНИМ ПОВЕРХНЕВИМ ІМПЕДАНСОМ І ДОВІЛЬНИМ ЗБУДЖЕННЯМ

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Чисельно-аналітичний розв'язок задачі про струморозподілення та вхідні характеристики біконічного асиметричного диполя з розподіленим поверхневим імпедансом і довільним збудженням, отримані в наближенні тонкого дроту. Правильність рішення підтверджується задовільним узгодженням чисельних і експериментальних результатів з літературних джерел. Наведено численні результати для вхідних характеристик диполя при його асиметричному збудженні точковим джерелом. Аналіз електродинамічних характеристик запропонованої дипольної антени довів можливість практичного застосування цієї антени для багатодіапазонних портативних радіостанцій, електронних гаджетів і базових станцій.

Ключові слова: *біконічний диполь; розподілений поверхневий імпеданс; асиметричне збудження; розподіл струму; вхідні характеристики*