An investigation of novel active phased array components

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Abstract. This paper deals with a novel approach for designing dual-band dipole radiators. The brief explanation of the basic antenna characteristics obtained is given in the paper. The approach proposed is based on the exact analytical solution for the overall input impedance of a system which consists of two ends-fed dipoles. Then, based on this solution the comparison of the analytical and the simulated results is derived and shown in the paper. In the printed model of the system proposed we compensate huge capacitance occurring in the input impedance of a stand-alone ends-fed dipole by decreasing the gap between the dipoles and the ground plane. Besides, the mathematical model of the phase shifter for phased antenna arrays designing, which is designed for improving the linearity of the phase shifting performance is presented below. Thirdly, the crossed dipole-like antenna fed by rectangular waveguide is also proposed in the paper.

1. Introduction
Wireless communication systems as well as large industrial antenna arrays, which are of the great importance in the last few decades, cannot be further developed without deep study of its radiating and feeding elements. Modern tendency in the miniaturization of all kinds of such devices leads investigators to search for new ideas how to achieve desirable radiation characteristics as well as lightweight and low profile of newly designed antennas. These demands are often controversial.

Another subject to point on is that the research into the properties of nontraditional antenna structures is of a great importance today, because it may contribute to discovering of fundamentally new approaches when designing and implementing radiating structures for modern communication systems.

2. Problem statement

2.1. Dual-band ends-fed dipole-like antenna.
In this section, we present a novel dual-band radiator, to design which we use ends-fed technic for excitation of a pair dipole-like radiators (Figure. 1). The design is based on a split power divider [1], which is placed on the bottom surface of the dielectric substrate with permittivity 1.5 and permeability 1. The power divider in tandem with a microstrip lines on the top side of the substrate mentioned form out-of-phase balancing unit (balun). In [2] and [3] authors use kindred approach to excite dual-band structure. However, in contrast to the design mentioned in [2] and [3], where one-side printed balun is used, in this project the balun is formed by both the bottom and the top side components. Also, in this case we use ends-fed excitation technique.
Figure 1. The topology of the dual-band antenna proposed.

As can be seen, the low-frequency dipole-like radiator is meandered along z-axis for reducing the overall structure size.

To explain briefly the approach one should start with theoretical results when deriving a double-dipole structure input impedance. Based on the exact analytical solution for surface current distribution along a linear dipole (along z-axis on Figure 1) when excited at its separated end points [4],

\[ I_z(z) = I_m \sin(\pm kz), \quad (1) \]

the longitudinal component of the electric field induced by one radiator to another can be derived [5]:

\[
E_z = \frac{1}{j4\pi\varepsilon_0\varepsilon_r} \left[ I_m \sin(kl) \left[ \frac{e^{-jkR_l}}{R_l^3} (z-l)(1+jkR_l) - \frac{e^{-jkR_l}}{R_2^3} (z+l)(1+jkR_2) \right] - kI_m \left[ \cos(kl) \left[ \frac{e^{-jkR_l}}{R_l} + \frac{e^{-jkR_2}}{R_2} - 2e^{-jkR_0} \right] \right] \right], \quad (2)
\]

where \( R_l = \sqrt{r^2 + (z-l)^2} \), \( R_2 = \sqrt{r^2 + (z+l)^2} \), \( R_0 = \sqrt{r^2 + z^2} \) — arguments of the free space Green function. The formula in closed form for the input impedance of a stand-alone ends-fed dipole radiator is to be found in [6]. From results in [6] and expression for the mutual impedance the overall input impedance can be derived.

Now, to keep the report brief, we are about to derive equation in closed form for mutual impedance of a system of two ends-fed dipole-like radiators being excited simultaneously with skipping intermediate steps of integral calculations to further formulate the expression for overall input impedance of the system under investigation:

\[
Z_{12} = \frac{C_{ll}^2}{2i} \left[ -e^{ikl_2} \left( Y(s_2^*) - 2Y(s_1^*) + Y(s_1^*) \right) + e^{-ikl_2} \left( Y(s_2) - 2Y(s_1) + Y(s_1^*) \right) \right] - \]

where \( C_{ll} \) is the coupling coefficient.
\[-Ae^{ikl_2} \Gamma(0, ikl_2 + ik(\sqrt{d^2 + (l_1 - l_2)^2} - l_1)) + 2Ae^{ikl_2} \Gamma(0, ikl_2 + ik\sqrt{d^2 + l_2^2}) - \\
-Ae^{ikl_1} \Gamma(0, ikl_2 + ik(\sqrt{d^2 + (l_1 + l_2)^2} + l_1)) - Ae^{ikl_1} \Gamma(0, -ikl_2 + ik(\sqrt{d^2 + (l_1 - l_2)^2} + l_1)) + \\
+2Ae^{-ikl_2} \Gamma(0, -ikl_2 + ik\sqrt{d^2 + l_2^2}) - Ae^{-ikl_1} \Gamma(0, -ikl_2 + ik(\sqrt{d^2 + (l_1 + l_2)^2} - l_1)) + \\
+BG(0, ik(d^2 + l_1^2 - l_1)) - 2\Gamma(0, ikd) + BG(0, ik(\sqrt{d^2 + l_1^2} + l_1)),
\]

\[A = -0.5i\text{ctg}(kl_2)kCI, B = -\frac{iCI_2}{\sin(kl_2)}k,
\]

where \(\Gamma(0, x)\) – incomplete appear gamma function, and \(\Upsilon(\gamma)\) - an auxiliary function, defined as:

\[\Upsilon(\gamma) = \frac{2\sqrt{\gamma} e^{-ikd\sqrt{\gamma}}}{d(\gamma + 1)} + ikC\text{Ci}(kd\sqrt{\gamma}) + kS\text{i}(kd\sqrt{\gamma}).
\]

In (3) \(l_1, l_2, d\) – lengths and spacing of the dipoles, respectively. The integration limits in (3) are following:

\[s_1' = \frac{1 - \sin(\text{arctg}(\frac{-l_2}{d}))}{1 + \sin(\text{arctg}(\frac{l_1 - l_2}{d}))}, \quad s_2' = \frac{1 - \sin(\text{arctg}(\frac{l_1 - l_2}{d}))}{1 + \sin(\text{arctg}(\frac{l_1 - l_2}{d}))};
\]

\[s_1' = \frac{1 + \sin(\text{arctg}(\frac{l_2 + l_1}{d}))}{1 - \sin(\text{arctg}(\frac{l_2 + l_1}{d}))}, \quad s_2' = \frac{1 + \sin(\text{arctg}(\frac{l_1 - l_2}{d}))}{1 - \sin(\text{arctg}(\frac{l_1 - l_2}{d}))};
\]

\[s_1 = \frac{1 - \sin(\text{arctg}(\frac{l_2}{d}))}{1 + \sin(\text{arctg}(\frac{l_2}{d}))}, \quad s_2 = \frac{1 - \sin(\text{arctg}(\frac{l_2 + l_1}{d}))}{1 + \sin(\text{arctg}(\frac{l_2 + l_1}{d}))};
\]

When both the intrinsic and the mutual impedances are formulated, it is possible to derive input reflection coefficient of the system of two ends-fed dipole-like radiators.

2.2. Improved phase shifter

It is known that quarter-wave directional couplers are an elementary base when building wide-band differential phase shifters. The closest such analogue is the Schiffman phase shifter [7]. The phase shifter under investigation was patented in 1987 [8], but so far the results of his research have not been published anywhere.

The strip phase shifter [8] consists of two segments of connected lines, the two adjacent ends of which are connected, and the other two ends are respectively the input and output of the strip phase shifter, characterized in that, in order to reduce distortion of time intervals in a pulsed mode, a circular conductor is additionally connected the entire perimeter with segments of connected lines, which in turn is shown in Figure 2.

Strip phase shifter works as follows. The microwave signal passing from the input to the output undergoes a phase shift determined by the electrical length of the segments of the connected lines. Additionally introduced annular conductor due to electromagnetic coupling with segments of connected lines along the entire perimeter provides additional electromagnetic coupling between segments of
connected lines, as a result of which the linearity of change in phase shift with frequency changes and the distortion of time intervals decreases when operating in pulsed mode.

To draw the topology, it is required to choose coupling coefficients \( k \) in the phase shifter so that at the frequency \( f_0 \) it is provided strictly \(-45^\circ\), and the non-uniformity of the phase shift \( \psi \) in the frequency band would be \( \pm 3^\circ \).

For reference: the phase shift \( \psi \) takes the following values at specific points for any coupling coefficients \( k \):
- at zero frequency \( f/f_0 = 0 \): \( \psi = 0^\circ \);
- at the center frequency \( f/f_0 = 1 \): \( \psi = -45^\circ \);
- on the second harmonic \( f/f_0 = 2 \): \( \psi = -90^\circ \).

Equation transfer function:

\[
S_{21}^{GORB} = S_{13} + \frac{S_{12}^2}{1 - S_{13}},
\]

where \( S_{12} \) and \( S_{13} \) are the voltage transmission coefficients.

\[
S_{12} = j \frac{k \sin \theta}{\sqrt{1 - k^2 \cos \theta + j \sin \theta}},
\]

\[
S_{13} = \frac{\sqrt{1 - k^2 \cos \theta + j \sin \theta}}{\sqrt{1 - k^2}},
\]

where \( \theta \) is the electrical length of the transmission line

\[
\theta = \frac{\pi f}{2 f_0},
\]

where \( f \) is the current frequency, \( f_0 \) is the center frequency.

![Figure 2. The topology of the dual-band antenna proposed.](image)

Required phase shift:

\[
\Psi = -1.25 \theta - \text{argument} \left( S_{21}^{GORB} \right).
\]

The use of the mathematical model is presented in part B of Section “Results”.

2.3. Crossed dipole-like antennas fed by waveguide

It is known that investigation of waveguide-fed radiators is necessary when constructing wireless communication systems. Another important area of using these radiators is radiolocation. The dipole and dipole-like antennas are often used for creating many complex radiators with the appropriate radiation pattern directivity. 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. Then, if a classical center–fed dipole antenna is excited by a rectangular guide operating at the dominant $TE_{10}$ mode, the split coaxial balun is positioned perpendicularly to a broad waveguide wall onto its centerline. The input port of the balun is excited by a coaxial probe placed inside the waveguide. The balun itself is composed of a coaxial line whose outer conductor, at one end, has two opposite open quarter-wave slots. The two sides of the split coaxial line form the electrically balanced nodes with the center conductor connected to one side. However, the use of split coaxial balun with corresponding slots as the dipole support above the guide often leads to unacceptable manufacturing complexity.

In the present study, to simplify radiator fabrication, the ends-fed dipole antenna (EFDA) and the center-end-fed dipole antenna (CEFDA) are used. The whole radiator structure is excited without any baluns through the rectangular waveguide using four coaxial probes only. This is the novel approach related to derivation of such the arrangement and excitation to date.

There are many radiators constructed on a base of turnstile antenna which proposed in [9]. The name ‘turnstile’ is applied to an antenna made up of two dipoles normal to each other and arranged so that their axes intersect at their midpoints. A turnstile antenna can operate in any of several modes, depending on the relative magnitude of the currents at the dipole inputs. When the dipoles of a turnstile antenna are energized by currents of equal magnitude and in phase-quadrature, the spatial field-strength radiation pattern has the omnidirectional (also known as all-directional or non-directional) shape. There are no nulls (or zeros) of radiation, and the maxima occur in directions normal to both dipoles, that is, along the positive and negative $z$-directions. In the plane $xoy$ containing the dipoles, the radiation is linearly polarized, whereas in all other directions it is elliptically polarized. In the $z$-direction, the polarization of the radiation is purely circular. In the two half-spaces separated by the plane $xoy$, the polarization vector rotates in opposite directions. Thus, the radiation field contains all likely states of polarization.

Omnidirectional radiation is produced by horizontally polarized transmitting TV antennas set up on tall masts. Along with the usual cylindrical dipoles, it is convenient in such antennas to employ what are known as batwing dipoles first proposed by B.V. Braude [5]. The batwing element is a number of parallel metal tubes taking up the space within the enclosing rectangular frame. The dipole arms are shorted to the mast for lightning protections at upper points, and the excitation voltage is fed at the lower points. The input impedance is then nearly constant and equal to 140 Ohms over a bandwidth of 7.5%.

Obviously, by adjusting the amplitudes and phases of the currents fed to the crossed dipoles, it is possible to obtain any desired state of polarization. This offers an additional degrees of freedom to effect polarization selectivity in a radio-communication system.

Another way to realize controlled polarization operation at UHF frequencies is the use of waveguide-fed CFDA and CEFDA first described in Russian Patents [4, 10].

To excite the arms of CFDA and CEFDA four probes are used inside the rectangular waveguide (Figure 3; the corresponding Cartesian coordinate system differs from that mentioned above in this subsection). These probes are characterized by corresponding depths inside the waveguide and are positioned perpendicularly to a broad wall of that. Both the ends-fed and center-end-fed dipole antennas are placed parallel to a broad wall of the guide at the distance from that (Figure 3). This distance is equal to one quarter wavelength in free space. To connect the arms of the crossed antennas with corresponding probes four coaxial stubs are used. The filling of the stubs is the material with relative permittivity equal to nearly unity but greater than that.
3. Results

3.1. Dual-band ends-fed dipole-like antenna
For the structure (Figure 1) we simulated return losses and radiation patterns at both central working frequencies. To compensate extremely great capacitance which occurs when exciting ends-fed dipole [6], we decrease the gap between high-frequency dipole and the ground plane. Here we also bring analytical results for input reflection coefficient when the reactive (image) part of the overall input impedance of the system proposed is artificially canceled (Figure 4).

Figure 5 reveals that when a printed model of the structure presented is designed the way the extremely great reactive part of the overall input impedance is compensated by proper setting the crucial antenna sizes it is able to radiate over two separated frequency bands. It seems that the shift of the low central frequency of the simulated model appears due to presence of complicated structure balun, which affects the all system resonance frequencies.

Figure 4. Analytical (blue) and simulated (red) input reflection coefficients (dB) of the antenna proposed.

Figure 5 shows the radiation patterns obtained.
3.2. Improved phase shifter
According to the formulas from the previous section, the mathematical model was calculated. When $k = 0$, we obtain the dependence graph presented in Figure 6. The task in relation to the required values was achieved with $k = 0.68$, the graph with the result is shown in Figure 6.

3.3. Crossed dipole-like antenna
The radiation pattern of the whole aerial has been investigated through full-wave simulation in free space. The corresponding rectangular waveguide [11] has been used at the central frequency 1.75 GHz. The steepest-descent method [12] was then used to determine the coaxial probe location. Figure 7 displays the dependence of the return loss versus frequency. The plots of the principal cuts of 3D radiation pattern at the frequency of 1.75 GHz are plotted in Figure 8. Thus, it can be seen that the proposed whole aerial has the fan–beam radiation pattern. Good agreement between the measured plots (green curves) and simulation-derived results (red markers) can be observed. Some discrepancy is attributed to impedance discontinuities in the test arrangement, the finite accuracy of the optimization procedure, and attainable resolution of the manufacturing.
Figure 7. Measured (green curve) and simulated (red markers) return losses of the whole aerial against frequency (left) and measured (green curve) and simulated (red markers) polar x-y radiation patterns (right).

Figure 8. Measured (green curve) and simulated (red markers) polar z-y radiation patterns.

4. Conclusion
The one of the possible approaches for designing dual-band antennas is presented in the communication. This approach is based on use of a printed balun, which is formed by a split power divider in the ground plane, and the microstrip lines on the top side of the substrate. The analytical and simulated results of the structure designed are compared and examined. From this it follows that when using ends-fed radiators the dual-frequency operating might be achieved.

As a result of the investigation of the phase shifter proposed, it appears to be possible to achieve quite acceptable results, namely: at specific points, the phase shift corresponds to the required values and the non-uniformity of the phase shift in the 1.1÷3.3 GHz frequency band did not exceed ±3°.

A new implementation of the crossed dipole-like antennas has been also proposed. To realize the necessary pattern the four coaxial probes placed inside the rectangular waveguide have been manufactured. The presented investigation paves the way for next generation of the circularly polarized aerials.

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