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Broadband Microstrip Antenna with a Hairpin Bandpass Filter

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Abstract—A new design for a broadband microstrip antenna is proposed, the working band of which is broadened using a bandpass filter based on two hairpin resonators electromagnetically coupled with a half-wave radiating resonator with a rectangular strip conductor. It is shown that the operating frequency range of this antenna is significantly extended as compared with that of analogous antennas, while its directivity and polarization diagrams remain highly stable within this range, which makes this design promising for applications. Operability of the antenna is demonstrated using its operating prototype. The measured characteristics of the antenna prototype agree well with the calculated data.

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Microstrip antennas (MSAs) are widely used in communication and radar systems and special microwave radio equipment [1]. They have certain advantages over classical antennas, including a high degree of miniaturization, manufacturability, and ability to radiate linearly, circularly, and elliptically polarized waves [2]. In the conventional MSAs, the relative working bandwidth is fairly small, since in such antennas the radiating elements are microstrip resonators with a relatively high Q factor. The high Q factor of radiating resonators is ensured by implementation on dielectric substrates with high permittivity ε and, accordingly, weak coupling with free space. The working band of MSAs is extended by different matching units, e.g., wave impedance transformers or filters [3, 4].

However, many available antenna matching units are complex and difficult to manufacture and tune. In addition, broadband antennas with such matching devices have serious drawbacks: one is the large difference between the antenna directivity diagrams measured at different operating frequencies caused by spurious radiation of matching devices elements; another is the presence of narrow nonradiative frequency ranges in the broad working band of the antenna, despite the sufficiently low level of return from the antenna input over the entire working band, according to the frequency dependence of return loss $S_{11}(f)$. In these narrow ranges, there are spurious high-Q resonances in the matching devices, including traveling wave resonances, which effectively absorb the incident electromagnetic wave power and prevent its radiation. Therefore, it would be incorrect to judge antenna broadbandness, as many authors do, only by the low return loss level in the frequency dependence $S_{11}(f)$.

It is not necessary to screen the matching device to reduce it's spurious radiation. Radiation can be prevented by wrapping regular strip conductors, e.g., in the form of a hairpin [5]. In a hairpin resonator, the ends of a strip conductor are close to one another at the first oscillation mode and their rf potentials have opposite signs [6]; thus, such a resonator is nearly nonradiative. As a result, wrapping a strip conductor of the matching resonator in a hairpin makes it possible to stabilize the antenna directivity diagram over the entire working band [7]. In view of the aforesaid, it is important to develop and study broadband MSAs without spurious resonances at operating frequencies and with a directivity diagram stable over the entire working band.

This Letter presents the design of a broadband antenna with the rectangular conductor of a radiating element with dimensions of $L_a \times W_a$ (Fig. 1). The antenna working band is significantly extended using a bandpass filter comprising two half-wave resonators of different lengths L_{r1} and L_{r2}, in which strip conductors with width w_r are wrapped in the form of hairpins with an inner gap S_r. The filter is electromagnetically coupled both with a rectangular strip element $l_c \times w_c$ in size connected to a coaxial 50-Ω input connector through a hole in the substrate and to the rectangular radiating element. Since the hairpin resonators are interdigi-

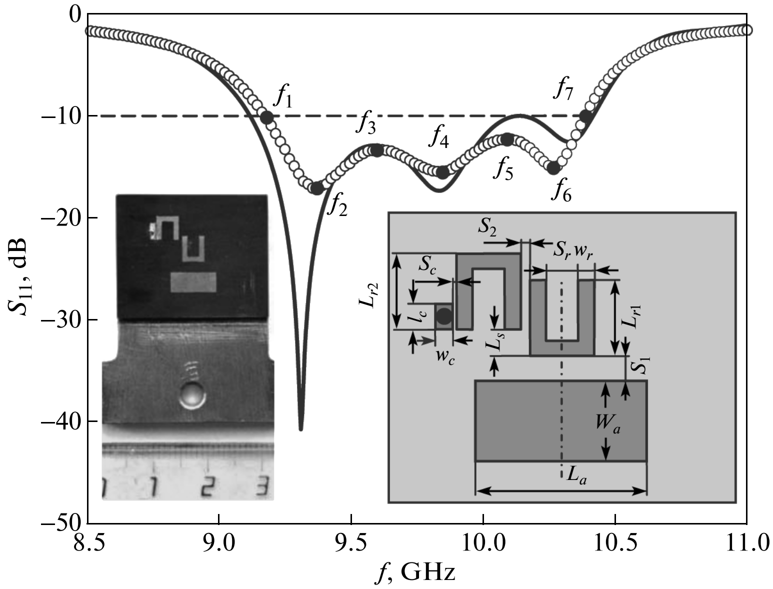

Fig. 1. Design of microstrip antenna with hairpin resonators and frequency dependences of return loss. Line indicates calculation and dots indicate experiment. Inset: photograph of antenna prototype.

talized, their inductive and capacitive interactions are added [8] and the unity coupling value is tuned by the gap S_2 and the parallel shift of the conductors relative to one another by a value L_s. As is known, there are rf current antinodes in the middle of both the strip conductor of the radiating element, which is a half-wave resonator, and each of the hairpin resonators; therefore, the radiating element is coupled mainly inductively with the nearest resonator. Obviously, this coupling can be tuned by both the gap S_1 and the length of the coupling region of the strip conductors, which depends on both their shift L_s and inner gap S_r of the resonators. Note that the input strip conductor connected to the coaxial connector is coupled mainly capacitively with the nearest hairpin resonator and the coupling is tuned by the gap S_c.

A parametrical synthesis of the antenna was carried out by numerical electrodynamic analysis of a 3D model of the studied microstrip structure implemented on a 2-mm-thick substrate made of a FLAN material with the permittivity $\varepsilon = 2.8$ traditionally used in microwave electronics. The design parameters of the strip conductors were chosen so that the maxima of return loss from the antenna input in the working band Δf observed in dependence $S_{11}(f)$ were at a level of −10 dB and the antenna had the maximum frequency bandwidth at a level of −10 dB, which was the antenna working band. For definiteness, the center of this working band was chosen in the X range, which is the most acceptable for satellite communication systems and on-board radars. As a result, the relative working bandwidth of the synthesized antenna at a level of −10 dB was $\Delta f/f_0 = 13.2\%$ and the center of this band was $f_0 = 9.77$ GHz.

The design parameters of the strip conductors of the synthesized antenna are (Fig. 1) $L_a = 8.1$ mm, $W_a = 3.6$ mm, $S_1 = 1.8$ mm, $S_2 = 0.51$ mm, $L_{r1} = 4.7$ mm, $L_{r2} = 4.27$ mm, $w_r = 1$ mm, $S_r = 2$ mm, $S_c = 0.11$ mm, $L_s = 3.93$ mm, $w_c = 1$ mm, and $l_c = 3.1$ mm. According to these conductor dimensions, we fabricated a working prototype of the antenna (see photograph in inset to Fig. 1). The measured frequency dependence of the return loss $S_{11}(f)$ of the prototype ($\Delta f/f_0 = 12.4\%$ at $f_0 = 9.79$ GHz) indicated by dots in Fig. 1 agrees well with the calculated data indicated by a solid line. As expected, the frequency dependence $S_{11}(f)$ represents a three-resonator design comprising two hairpin resonators of the filter and one regular radiating resonator. Note that we substituted the real dimensions of the conductor topology, which were measured on the fabricated device, in the calculated model. Comparison of the investigated structure with an analogous antenna designed on one hairpin resonator [7] shows that adding one more hairpin resonator leads to extension of the antenna working band by a factor of 1.5.

The directivity and polarization diagrams, which are the most important characteristics of antennas, were studied at seven characteristic frequencies indicated by black dots in Fig. 1. These frequencies correspond to the working band edges $f_1 = 9.18$ GHz and $f_7 = 10.39$ GHz, the minima $f_2 = 9.34$ GHz, $f_4 = 9.86$ GHz, and $f_6 = 10.27$ GHz, and the maxima $f_3 = 9.59$ GHz and $f_5 = 10.08$ GHz of the return loss. The amplitude and polarization angular dependences of the studied MSA were measured using a special broadband measuring antenna operating in the receiving mode. Circuits for measuring the directivity diagrams

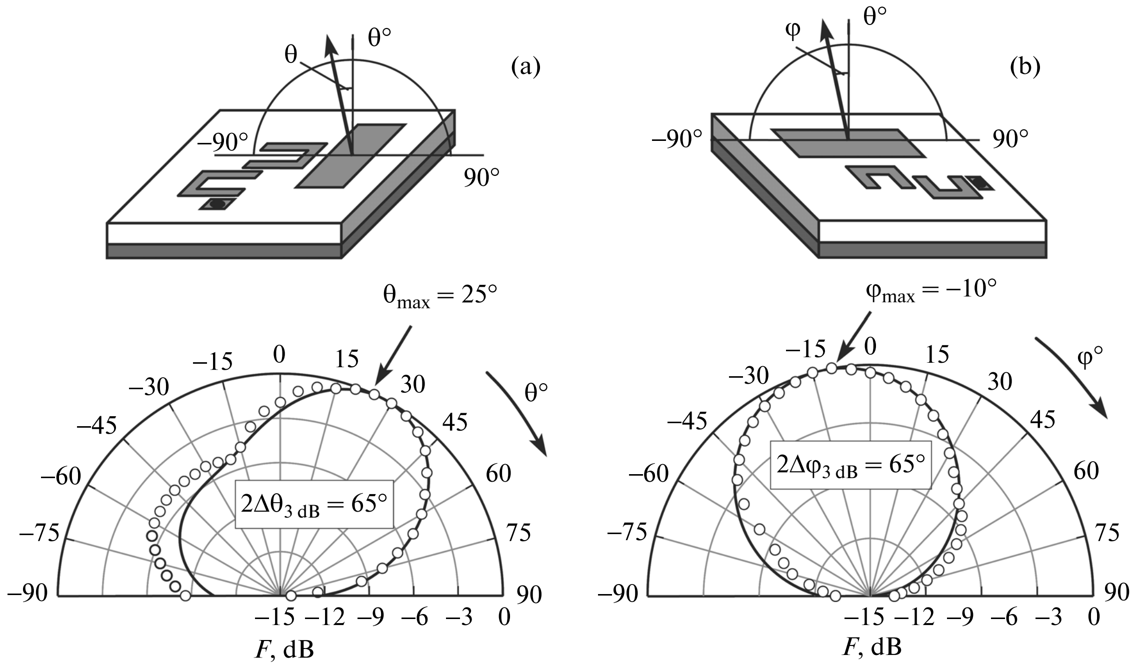

Fig. 2. Measuring circuits and measured amplitude directivity patterns of antenna in (a) orthogonal and (b) parallel planes obtained at frequency $f_4 = 9.86$ GHz. Line indicates calculation and dots indicate experiment.

in the two orthogonal planes of the antenna are shown in Figs. 2a and 2b. For definiteness, the plane perpendicular to the long axis of the antenna strip conductor is called orthogonal, and the plane parallel to this axis, parallel. Dots in Figs. 2a and 2b indicate the measured amplitude angular dependences $F(\theta)$ and $F(\varphi)$; and lines indicate the calculated data, which are also indicative of good agreement between the theory and experiment. The dependences $F(\theta)$ and $F(\varphi)$ in the orthogonal and parallel antenna planes were measured near the center of the antenna working band at the frequency $f_4 = 9.86$ GHz. The angular widths of the antenna radiation lobe measured at a level of –3 dB from the maximum are $2\Delta\theta = 65°$ and $2\Delta\varphi = 80°$ in the orthogonal and parallel planes, respectively. The antenna maximum radiation directions determined as centers of the radiation angular widths are $\theta_{max} = 25°$ and $\varphi_{max} = -5°$. The measured amplitude angular diagrams showed that the directional properties of the antenna are fairly stable at all characteristic frequencies f_1–f_7, i.e., over the entire working band, both in the orthogonal and parallel planes. The maximum radiation direction changes merely within $\theta_{max} = -5°$–$25°$ and $\varphi_{max} = -30°$–$5°$ and the radiation lobe width changes within $2\Delta\theta = 60°$–$80°$ and $2\Delta\varphi = 70°$–$80°$.

It is important that the experimentally observed minor deviation of the maximum antenna radiation direction from the substrate plane normal is related to the presence of strip conductors of the matching filter located near the strip conductor of the radiating resonator from only one side (Fig. 1). Such a design asymmetry not only leads to minor deviation of the radiation direction but also has weak dispersion in the antenna working band, which manifests itself as a minor variation in deviation angles θ_{max} and φ_{max} and angular widths $\Delta\theta$ and $\Delta\varphi$ of the antenna radiation lobe in the two planes.

The circuit for measuring polarization patterns of this antenna and the calculated patterns are shown in Fig. 3. For simplicity, we present the dependences on angle ψ obtained only at the limiting working frequencies $f_1 = 9.18$ GHz and $f_7 = 10.39$ GHz and near the center of the working band $f_4 = 9.86$ GHz. Like in the amplitude directivity patterns, orientation ψ_0 of the antenna polarization plane was determined as the center of the angular width of the radiation lobe measured at a level of –3 dB. The investigations showed that over the entire antenna working band, the position of the polarization plane changes within the narrow angular range $\psi_0 = -5°$–$8°$ and the mean value is $\psi_0 \approx 1°$.

The analysis of our results showed that the antenna ellipticity coefficient, which can be easily determined from the signals measured in the maxima and minima of the polarization patterns, is no more than 0.12 over the entire working band. Therefore, the investigated antenna can radiate (receive) electromagnetic waves that are polarized almost linearly.

Thus, we studied the new design of a microstrip antenna whose working band is significantly extended by a hairpin bandpass filter, which is integrated on one substrate with the radiating element. The antenna is high-performance, simple, and manufacturable. In addition to the broad working band, advantages are the high stability of the directivity and polarization dia-

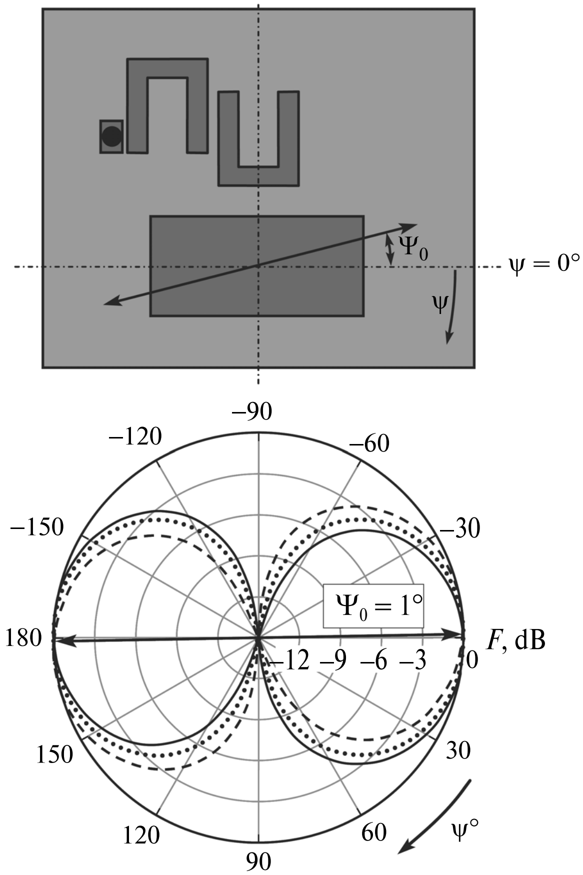

Fig. 3. Circuit for measuring polarization patterns of investigated MSA and polarization dependences calculated at frequencies f_1 = 9.18 GHz (solid line), f_4 = 9.86 GHz (dots), and f_7 = 10.39 GHz (dashed line).

grams over the entire working band and the small ellipticity coefficient, which indicates the ability of the antenna to receive and radiate almost linearly polarized electromagnetic waves. The operability of the antenna was demonstrated by an operating prototype implemented on a 2-mm-thick and 28 × 22 mm in size FLAN substrate (ε = 2.8). The measured characteristics of the prototype agree well with the results of electrodynamic calculation of the microstrip 3D model. The antenna's relative working bandwidth is $\Delta f/f_0$ = 12.4% at the center frequency f_0 = 9.79 GHz. All the above-mentioned factors make the antenna promising for application in communication and radar systems and special radio equipment.

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