High gain millimeter wave leaky wave antennas with low side lobe level of the radiation pattern

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Abstract: Two linearly polarized millimeter wave planar leaky wave antennas designed for broadside radiation at 61 GHz with high gain and low sidelobes of radiation patterns are presented. Both antennas have a center-fed single-layer structure composed of two one-dimensional periodic symmetric arrays of metal strips of variable width on a grounded dielectric waveguide, but with different excitation devices for the dielectric waveguide, represented by a microstrip combline and a waveguide-fed linear substrate integrated slotted waveguide array. The antennas exhibit low return loss, antenna gain 26–27 dBi and radiation pattern sidelobe level in both E- and H-planes, not exceeding −20 dB. The antennas may be used, for example, in security sensors.

1 Introduction
In recent years, the use of various devices and radio systems of millimeter wave range has been intensively expanding. These are radio communication systems, (including heterogeneous wireless multiservice networks), car radars, traffic control radars, measuring devices, sensors for security systems. This is mainly due to the significant progress in the field of miniaturization, improving production technology and energy efficiency of various electronic components and devices, including antennas. The electrical and operational characteristics of millimeter-wave devices and systems substantially depend on the antenna characteristics. In many cases, compact, high-gain antennas with low sidelobe level of the radiation pattern are required. To date, there are several types of compact antennas such as planar microstrip antenna arrays, parallel plate slotted antenna arrays [1–4], leaky-wave antennas and arrays (LWA) [1], [5]–[15]. The results of many theoretical and experimental studies [5], [6], [8], [9], [14] show that some types of LWA have good prospects, especially for millimeter wave applications, due to such advantages, as a simple structure and a rather high radiation efficiency, as well as a low-cost printing technology of fabrication. The size of LWA and its radiation properties directly depend on the design and performance of the feeding device for planar dielectric waveguide (DWG), which is a base part of many leaky-wave
structures. In developed so far planar LWA, such devices designed as sectoral horns and mirrors, linear slotted waveguides and microstrip combline arrays \[8\], \[9\], \[10\], \[12\], \[17\]. Despite considerable progress in designing compact and efficient LWA, there are still many possibilities for further improvement of the broadside LWA, such as return loss, sidelobe level (SLL) of the radiation pattern, directivity, radiation efficiency and gain, minimization of the antenna size and weight. This report presents two modified broadside LWA, composed of DWG and one-dimensional periodic metal strip diffraction grating \[12\]–\[15\]. Recently developed antennas like their previous samples also provide low return loss and produce effective linearly polarized broadside radiation, but differ in a reduced level of side lobes of the radiation pattern in both E- and H-planes.

2 Previously developed antennas: design, principles of operation and radiation properties

The principle of operation of LWA, considered in this report, utilizes a well-known effect of surface electromagnetic waves transformation, while propagating along the DWG, into leaky waves, radiating from the structure due to scattering of the surface waves on a periodic diffraction grating \[1\], \[5\]. The antennas operate on the \(-1^{st}\) spatial harmonic of the diffraction field; therefore, they designed in a way to perform as resonant diffraction structures in vicinity of Bragg diffraction of the second order and to form a narrow broadside beam.

2.1 Leaky wave antenna using combline power divider

The center-fed LWA \[12\], \[13\] (Figure 1), developed for centimeter or millimeter wave range, contain a grounded DWG 1, a periodic diffraction grating 2 composed of two subarrays of parallel metal strips (having common central strip), and a feeding device 3 (multichannel power divider) for the DWG.

![Figure 1: Single-layer LWA with a combline power divider.](image)

To feed the antenna, a rectangular input slot in the center of the metal ground plate is used. The feeding device 3 is coplanar to the metal strip grating 2 and composed of the central strip of the grating with two lateral rows of rectangular metal stubs, attached to both edges of the strip, similarly to a printed comb line structure with radiating stubs. The strips and the stubs are periodically positioned along \(OX\) and \(OY\) axis with \(d_x\) and \(d_y\) spacing; the size of the antenna radiating aperture is \(L_x \times L_y\). The main radiation beam direction shown as a thick black arrow on \(OZ\) axis. Period \(d_x\) is to be taken equal to one wavelength of the surface waves, propagating along DWG in \(\pm OX\) directions; period \(d_y\) – equal to one wavelength of waves at the central operational frequency, excited in the transmission line (formed by the central strip with lateral stubs and metal ground plate) and propagating in \(\pm OY\) directions. Therefore, the main beam is normal to the radiating aperture. More
precisely, the transmission line, formed by the central strip with stubs and short-circuited to the ground plate at both ends, performs as an input power divider. According to the above-mentioned spacing $d_y$, at the design frequency one part of the input power directly radiates from the structure along $OZ$ axis, but the other, larger part of the input power, due to the presence of discontinuities (stub), converts into the power of hybrid $TM$-like surface waves, propagating along DWG in $\pm OX$ directions. As the result of their diffraction on the strip grating, the antenna radiates in $OZ$ direction with the electric field linearly polarized in $XOZ$ plane and parallel to $OX$ axis. Since the proposed LWA utilizes a combined corporate-series feed, the antenna operating frequency range limited by the frequency-dependent beam split. The typical SLL of the radiation pattern in both $E (XOZ)$ and $H (YOZ)$ planes is $-13...-15$ dB. It has already been proven that in case of narrow-band applications (1–3) % such LWAs provide a rather high directivity, up to 28–32 dBi, while the total antenna efficiency is 50–75 % at frequencies up to 60–90 GHz [16].

To avoid considerable reflection and return loss at the resonant frequency, the DWG thickness $h$ should be chosen about one quarter wavelength in the media with the same relative dielectric permittivity $\varepsilon_r$ and the width of the central strip of the diffraction grating should be taken close to $d_x$. In such a case, low reflection at the input point of the antenna can be obtained, so that VSWR does not exceed 1.5–2 [12], [13].

2.2 Leaky wave antenna using SIW power divider

To improve the shape of the radiation pattern and especially to reduce the sidelobe level, another type of broadside LWA has been proposed [16]. The antenna design suggests using of a center-fed substrate integrated slotted waveguide (SIW) power divider [3], [17], [18] instead of the comb line to excite DWG.

![Figure 2: Single-layer LWA with SIW power divider](image)

Such solution of the problem illustrates Figure 2. This antenna also contains grounded DWG 1, metal strip grating 2, and alternating phase SIW power divider formed by the central strip 3 of the grating electrically connected to the ground plate by pairs of metal rods 4 (metalized holes in DWG) and short-circuited to the ground plate at both ends.

In compare to the antenna with the same aperture size, using combline excitation device, the antenna with SIW power divider demonstrates practically equal maximum gain, but better SLL ($-15...-17$ dB).

The initial design parameters of the radiating aperture of LWA (Figure 1), such as DWG thickness $h$, strip width $w$ and spacing (period) $d_x$ have been calculated using a simple mathematical model, developed for an infinite periodic lossless structure consisting of a grounded dielectric slab with a 1D-periodic conducting strip grating. Figure 3 [13]. Based on this model, a dispersion equation has been formulated to calculate the complex propagation constant $\beta = \beta_0 - j\alpha$ of the $TM$-type fundamental spatial harmonic (SG) in the structure ($\beta_0$ is the phase coefficient, $\alpha$ is the attenuation coefficient – in
other words, leakage rate) and $-1^{st}$ SG $\beta_1 = \beta - 2\pi/d$, as well as other characteristics, including the frequency dependence of the direction of maximum radiation [1], [12], [13]. To solve the dispersion equation, the expansion of the density of the transverse surface electric current $J_x(x)$ has been used in the boundary condition for the tangential components of the magnetic field vector on infinitely thin lossless metal strips with basis functions (weighted Chebyshev polynomials of the second kind) [14]

![Figure 3: Fragment of a one-dimensionally periodic leaky wave structure](image)

As a result, a dispersion equation can be written in a following well-known form

$$\det G = 0,$$  \hspace{1cm} (1)

where $G$ is a $M \times M$ square matrix ($M$ is the number of basic functions), containing elements

$$G_n = \sum_{n=-N}^{N} K_n J_{\mu n} J_{\mu n}^*,$$  \hspace{1cm} (2)

$$K_n = \frac{\varepsilon}{\eta_n \tan \eta_n h} - \frac{1}{\gamma_n}$$

where $\varepsilon$ is the relative dielectric constant of DWG; $\beta_n = \beta + 2\pi n / d$, $n = -N; N$ is the longitudinal propagation constant of the $n^{th}$ SG in the structure; $2N+1$ is the number of SGs; $\gamma_n = (k_0^2 - \beta_n^2)^{1/2}$ and $\eta_n = (k_0^2 \varepsilon - \beta_n^2)^{1/2}$ are transverse propagation constants of the $n^{th}$ SG over the structure and in the dielectric layer, respectively; $k_0 = 2\pi / \lambda$ is the free space wave number, $\lambda$ – the wavelength; $J_{smm} = \frac{1}{d} \int_{-w/2}^{w/2} J_{sm}(x) \exp(j\beta_n x)dx$ – the $n^{th}$ component of the current $J_x(x)$ decomposition in Fourier series on SG.

In addition, various ways to reduce SLL of the $E$-plane ($XOZ$) radiation pattern have been applied: use of diffraction gratings with non-equidistant strips of non-equal width, use of gratings with equidistant strips of constant width, but with reflecting metal walls, short-circuiting outer edges of the strips to the ground plate at the left and right borders of the DWG.

### 3 Development of leaky wave antennas with reduced sidelobe level of the radiation pattern

To reduce SLL of the $E$-plane radiation pattern in both antennas, a modified (non-exponential) symmetrically decreasing amplitude distribution $A(x)$ of the field along $\pm OX$ directions has been formed. According to a well-known method [6], [8] applied to traveling wave antennas the function $\alpha(x)$ emboding $A(x)$ has been used:

$$\alpha(x) = \frac{1}{2} \left( \frac{\left| A(x) \right|^2}{\int_{-\infty}^{\infty} \left| A(x) \right|^2 + \int_{-\infty}^{\infty} pL(x) dx} + \frac{B}{B_0} \int_{-\infty}^{\infty} \left| A(x) \right|^2 + \int_{-\infty}^{\infty} pL(x) dx \right),$$  \hspace{1cm} (3)
where $P_0=1$ is the input power and $P_L$ is the remaining power at the end of the structure; $L=L_x/2$. For example, to ensure the SLL not more than $-20$ dB, the "cosine on the pedestal" type amplitude distribution has been chosen; for $H$-plane – "cosine" type. After discretization of $a_n(x)$ in accordance to the number of strips $N$ in subarrays and obtaining samples $a_n$, a set of appropriate strip widths $w_n$ ($n=1...N$) has been defined using computer simulation (HFSS) and $S$-matrix parameters calculation of a real finite size structure with three ports (Figure 4).

![Figure 4: Three port model for computer simulation](image)

Then similar procedure has been applied to embedding $A(y)$ via $a_n(y)$ for the combline power divider, resulting in defining set of appropriate stub lengths $w_{pm}$ ($m=1...M$). Worth to notice, that both strip diffraction grating and combline have been kept equidistant ($d_x=\text{const}$ and $d_y=\text{const}$). At first glance, this is not the best solution, since with a constant period of these gratings, a change in the width of the strips or the length of the stubs leads to the change in both the attenuation coefficient and the phase coefficient. In turn, this should lead to a decrease in the directivity factor of the antenna due to non-phase radiation of the periods of the structure. However, the results below show that this does not cause a noticeable decrease in aperture efficiency. At the same time, this approach significantly reduces the amount of calculations as well as simulation time in determining the design parameters of antennas.

With respect to the waveguide-slot power divider, a good result in decreasing SLL of LWA radiation pattern in $H$-plane has been obtained with a constant period of slots, but with a short circuit of the waveguide at a distance of half the wavelength from the centers of the extreme slots (the expected SLL about $-26.4$ dB).

### 4 Computer simulation of new leaky wave antennas

This section presents the results of computer simulations using HFSS program of both new antennas with a reduced SLL of the radiation pattern (Figure 5–7 for LWA1 with combline power divider and Fig. 8–11 for LWA2 with slotted SIW power divider). Both antennas contain DWG made of Rogers RT/Duroid 5880 ($\varepsilon=2.2$, $\tan\delta=0.0009$); the aperture dimensions of LWA1 are $L_x \times L_y = 51 \times 44$ mm$^2$, LWA2 – $L_x \times L_y = 43.2 \times 45.2$ mm$^2$.

The thickness of DWG $h=0.79$ mm. To feed the antennas, a rectangular waveguide $3.6 \times 1.8$ mm$^2$ has been used.
Figure 5: Frequency characteristics of directivity (D), gain (G) and realized gain (GR) of LWA1.

Figure 6: Frequency characteristic of VSWR of LWA1.

Figure 7: Radiation Patterns of LWA1 in E- and H-planes, f=61 GHz.
Figure 8: Frequency characteristics of directivity (D), gain (G) and realized gain (GR) of LWA2

Figure 9: Frequency characteristic of VSWR of LWA2

Fig. 10. Radiation Patterns of LWA1 in $E$- and $H$-planes, $f=61$ GHz
According to the above-mentioned aperture dimensions and simulated values of directivities, the aperture efficiencies of LWA1 and LWA2 are $\nu_1=0.45$ and $\nu_2=0.53$ respectively. The SLL of LWA1 in E-plane is about $-20$ dB, in H-plane $-23$ dB. The SLL of LWA2 in E-plane is about $-21$ dB, in H-plane $-27$ dB ($-30$ dB for the first side lobes).

5 Conclusion

Two new modifications of the already known center-fed linearly polarized LWA for 60.5–61.5 GHz frequency band with low SLL of the radiation pattern have been developed and simulated. Both new antennas have the aperture efficiency of more than 45 %, radiation efficiency more than 93 % with antenna gain 26.5–27 dBi and SLL of the radiation pattern not exceeding $-20$ dB. It is expected that the proposed LWA can be used, for example, in millimeter-wave radio communication systems, wireless computer networks and security sensors.

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