Printed $H$-plane horn antenna with loaded dielectric-metal composite lens

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Abstract: A printed $H$-plane horn antenna with loaded composite $E$-plane dielectric-metal lens is proposed and analysed in this study. The loaded lens of this horn antenna, which is parallel fed through an Sub-Miniature-A coaxial connector, is integrated on one printed substrate circuit board by utilising the substrate integrated waveguide (SIW) technique. To well illustrate the superiority of the proposed antenna, three types of antennas, namely an SIW $H$-plane horn antenna without any load, an SIW $H$-plane horn antenna only loaded with the dielectric slab, and the proposed SIW $H$-plane horn antenna loaded with both dielectric slab and $E$-plane metal lens (dielectric-metal composite lens), are manufactured, measured and analysed. Compared with the horn antenna only with a loaded dielectric slab, the proposed antenna with dielectric-metal composite lens achieves a higher gain of more than 2 dB in the operating bandwidth. Besides, the antenna has a relatively lower profile and smaller size, which makes the proposed antenna more easily to be fed, integrated and form arrays.

1 Introduction

The $H$-plane horn antenna is one of the most widely used directional antennas, which is constructed by opening a traditional rectangular waveguide along its $H$-plane. It has some obvious advantages such as the simple structure, easy feed, relatively wider bandwidth and higher antenna gain. Manufacturing of the planar rectangular waveguide is now possible thanks to the developing of substrate integrated waveguide (SIW) technique, which has been widely used in the design of microwave circuits [1–5] and antennas [6–10] because of its lower profile and easier integration so as to gradually replace the traditional cumbersome metal waveguide. Recently, the researches on the design of antennas adopting SIW have mainly focused on three aspects: (i) dipole element antennas and their arrays fed by SIW [11–15]; (ii) slot array antennas using SIW techniques [16–20]; and (iii) $H$-plane horn antennas based on SIW discussed in [21] in detail and the corresponding arrays are also presented to improve the antenna gain. Besides, in the aspect of enhancing the gain of horn antenna, loading lens is a commonly employed method, which has been deeply researched in [21–24].

An $H$-plane printed horn antenna with loaded dielectric-metal composite lens is proposed in this paper. The composite lens comprises a dielectric slab and an $E$-plane metal lens composed of a series of metal via-holes. With the delay effects of the loaded dielectric and the speeding impacts of the $E$-plane metal lens, the phase distribution on the antenna aperture can be more uniform so as to enhance the gain accordingly. This process will be achieved by reasonably controlling the length of the loaded dielectric and the distance between the metal lenses. The advantages of the proposed antenna are evidently proved by both simulations and experiments, which match each other well. Among the three types of horn antennas discussed in this paper, the proposed horn antenna with loaded dielectric-metal composite lens achieves the highest gain.

The simulated results are obtained by utilising CST Microwave Studio®. The adopted dielectric substrate has a relative permittivity of 4.4 and a thickness of 5.0 mm, and the typical operating frequency of the antenna is 18 GHz.

2 Antenna structure design

2.1 SIW $H$-plane horn antenna with no load

The structure of the SIW $H$-plane horn antenna without any load is shown in Fig. 1. This antenna is printed on the double-sides of a resin board with the relative permittivity $\varepsilon_r = 4.4$ and the thickness 5.0 mm. It is composed of a feeding part (Part I) and a printed horn part (Part II). The metal via-holes (the solid circles in the right part of Fig. 1) distributed on the horn antenna connect the two metal boards (the dark area in the left part of Fig. 1) located on the both sides of the resin. The Sub-Miniature-A (SMA) coaxial connector is matched to the feeding, two branches of which are, respectively, connected to the inner conductor of the coaxial connector and the flange.

1. The design of Antenna 1 is the base of the latter two types of antennas discussed later in this paper. The key parameters of this $H$-plane horn antenna include the aperture length $W_A = 19$ mm, the aperture height $T_1 = 5$ mm, the waveguide section length $W_5 = 10$ mm and the feeding line width $W_8 = 8$ mm.

2. The waveguide section length $W_7$ is determined by the set operating frequency of the horn antenna, where the working mode in the waveguide is TE$_{10}$ mode, which means that the parameter $a$ must satisfy that $a < \lambda = \frac{\lambda}{2\pi}$, where $\lambda$ is the wavelength for the corresponding set typical operating frequency. The optimised value of $a$ is $a = 1.15\lambda = 19$ mm, where $\lambda$ is the wavelength at 18.2 GHz. The aperture height is the same as the waveguide height, which is determined by the thickness of the resin board, 5.0 mm. The aperture length is set to achieve the optimised $H$-plane horn antenna as $W_6 = 19$ mm. Since the antenna is fed by a balanced microstrip line, the feeding line, on the one hand, connects with the 50 $\Omega$ SMA coaxial connector, and on the other hand, connects with the $H$-plane horn antenna (Part II in the left part of Fig. 1). Therefore, the feeding line is designed to be gradient-structured and the width at the feeding port is $W_8 = 8$ mm. What should be noted is: different from the orthogonal coaxial feeding structure in [21], the antenna in this paper applies the parallel feeding method, because the thickness of the resin board is 5.0 mm, only 0.30$\lambda$. More important, the parallel feeding method is easier to be integrated with the printed transmission line (e.g. microstrip line) in order to solve the problem of being difficult to be integrated for the coaxial line.

Other parameters, such as the distance between via-holes and the diameter of via-holes, are based on the designing concepts proposed in [6]. They are $W_6 = 19$ mm, $W_7 = 10$ mm, $W_8 = 8$ mm, $W_9 = 0.1$ mm, $L_4 = 13$ mm, $L_5 = 12$ mm, $L_7 = 4$ mm, $L_8 = 10$ mm, $D = 0.5$ mm, $d = 0.95$ mm, $T_1 = 5$ mm and $T_2 = 0.1$ mm.
2.2 SIW H-plane horn antenna with loaded dielectric slab

A dielectric slab is loaded on the Antenna 1 discussed in Section 2.1, with the length $W_1 = 41.2$ mm, width $L_1 = 12$ mm and thickness $5.0$ mm. The material of the loaded dielectric slab is the same as the resin board of the antenna. This antenna has three parts (Part I, Part II and Part III in Fig. 2) and is referred to as Antenna 2 in this paper, shown in Fig. 2. The loaded methods are partly drawn lessons from [21, 25].

2.3 SIW H-plane horn antenna with loaded dielectric-metal composite lens

A novel type of E-plane metal lens is loaded on Antenna 2 discussed in Section 2.2, which constitutes Antenna 3 shown in Fig. 3. In Antenna 3, a series of metal via-holes are opened in the dielectric slab, which are connected by the parallel metal strips on the double sides of the dielectric slab in order to form an E-plane metal lens. This lens is composed of nine groups of metal via-holes, as shown in Part III of Fig. 3a. The axis of the middle via-hole strip (labelled 1 in the right part of Fig. 3a) is located in the middle axis of the aperture, and the other eight strips are distributed symmetrically. Where, $W_5 = 11.5$ mm, $W_6 = 16.95$ mm, $W_7 = 14.45$ mm and $W_8 = 13.35$ mm are, respectively the distances between the according strip and the aperture middle axis, and $L_2 = 4.9$ mm, $L_3 = 8$ mm and $L_4 = 10$ mm are the corresponding lengths. This method is inspired by the SIW technique, which is first presented from this paper.

As shown in Fig. 3b, the horizontal and vertical lengths of the loaded dielectric slab are, respectively, $W_1 = 41.2$ mm and $L_1 = 12$ mm. To have a better comparison, the metal lens should be loaded on this slab. The design of lens is based on the geometrical optics as follows in detail:

i. Set the metal grid $KM$ in the middle axis of the aperture. To guarantee the electromagnetic wave to propagate longer distance in the plane formed by two grids, set the length $L_0$ to 2.0 mm, equivalent to a quarter of the waveguide wavelength.

ii. Determine $GI$ and $HJ$. Lengthen the two sides of $H$-plane horn $CD$ and $EF$, the crossover point $X$ and the intersections $G$ and $H$ with the loaded dielectric slab can be confirmed. Then use two long grids $GI$ and $HJ$ to limit the radiation energy from aperture $CE$ between them.

iii. Draw a circle with centre at $X$ and radius equal to $XM$, intersecting the line $CG$ and $EH$ at $L$ and $N$, so as to determine grid $PL$ and $QN$.

iv. There are only three grids within the limits of $GH$, which is due to the refractive index of the E-plane metal lens formed by the grids is proportionate to the factor $\sqrt{1 - (\mu L)^2}$, i.e. the bigger $\lambda/\mu$, the slower $n$ changes and the wider lens bandwidth.

3 Analysis on the working principles of the proposed antennas

The working process of the proposed antennas is as follows. The TEM wave first comes into the balanced microstrip line through the coaxial connector, then spreads into the SIW of the printed H-plane horn, thus finally transforming into TE$_{10}$ mode. High order mode waves caused by this transformation is very small and can be ignored after a proper distance’s spread, because, the electric fields of TEM and TE$_{10}$ waves are both perpendicular to the metal surfaces on the double sides of the dielectric.

The SIW opens along the wider side of the waveguide aperture, which forms the field distribution on the aperture of an $H$-plane horn antenna. This distribution still satisfies the TE$_{10}$ mode, but the phase distribution changes compared to before it opens, which is described in Fig. 4. The electromagnetic wave is assumed as the spherical wave spreading from the centre point $O$ of the $H$-plane horn antenna. After some distances’ spread, it arrives at the points $A$ and $B$, where possess the same phases. However, the phase at point $A$, which satisfies the square-law distribution along the $x$-axis direction as shown in Fig. 4, falls behind the phase at $B$ and can be described by (1). Since the thickness of SIW is relatively small, the phase on the aperture can be considered to distribute uniformly along the $y$-axis. Compared with the uniform distribution, the square-law distribution will lead to a widening of the main lobe and a decrease of the antenna gain, which are disadvantageous to the antenna. To solve this problem, some kind of lens should be added to improve the phase distribution on the aperture

\[
\theta(x) = -K_\phi x^2 \quad (1)
\]

where $\theta(x)$ is the aperture phase distribution function; and $K_\phi$ is the constant in the aperture phase distribution function.

Wang et al. [21] presents a dielectric lens loaded SIW $H$-Plane horn antenna and indicates that the elliptical lens boundary can improve the aperture phase distribution. Here, an $E$-plane metal lens is introduced into the rectangular dielectric slab, which forms a composite dielectric-metal lens. The dielectric slab is a kind of lens that can decrease phase velocity and thus increase the phase delay, while the metal lens can accelerate phase velocity and thus reduce the phase delay. This means it is possible to minimise the phase difference between point $B$ and $A'$, and achieve a more uniform and optimised phase distribution on the aperture by adjusting the parameters reasonably as: $W_2 = 11.5$ mm, $W_3 = 13.35$ mm, $W_4 = 14.45$ mm, $W_5 = 16.95$ mm, $L_2 = 4.9$ mm, $L_3 = 8$ mm, $L_4 = 10$ mm and $L_5 = 2.5$ mm. The analysis above can be expressed and analysed by the following equation:
where $\Delta \varphi$ is the phase delay between two analysed points; $\omega$ is the angular frequency of the electromagnetic wave; $v$ is the phase velocity of the electromagnetic wave and $v = (c/\mu \varepsilon) \sqrt{1 - (\lambda / l)^2}$; and $l$ is the wave front displacement.

### 4 Analysis and simulation

#### 4.1 Analysis of the electromagnetic field amplitude and phase distribution on the aperture

The simulated electromagnetic field amplitude and phase distributions on the aperture of Antennas 1, 2 and 3 are displayed in Fig. 5, where Fig. 5a shows the normalised electromagnetic field amplitude distribution, and Fig. 5b depicts the relative electromagnetic field phase distribution in which the maximum phase value on the aperture is deducted from the phase value.

The simulated result in Fig. 5a indicates that the electromagnetic field amplitude in the middle is bigger than that on the both ends of the aperture, which satisfies the TE$^{10}_{0}$ mode case. However, the amplitude distribution of Antenna 3 is different because of the introduction of the E-plane metal lens. Influenced by the via-hole Array 1 in the right part of Fig. 3a, the electric field amplitude distribution on the aperture is divided into two parts and thus leads to two peak values in the electric field amplitude distribution, which indicates that the radiated electromagnetic field on the aperture can be considered as a two-element array composed of two TE$^{10}_{0}$ mode fields. If the two elements have the same phase distribution, they can produce a directivity 3 dB higher than that produced by only one TE$^{10}_{0}$ mode field in theory due to the interference effect.

It is inferred in Fig. 5b that the electromagnetic field phase distribution on the horn aperture between Antennas 1 and 2 is obviously different. The aperture length of the Antenna 1 is 19 mm, i.e. 1.15$\lambda$_(at 18.2 GHz), belonging to the electronically large size, and the phase of the H-plane follows the square-law distribution (the solid line in Fig. 5b). Introduced a rectangular dielectric slab, the phase distribution of Antenna 2 is still not uniform enough, which can also be considered as the square-law distribution (the dashed line in Fig. 5b), because no optimisation on the boundary of the dielectric slab is proceeded. As for Antenna 3, after loading the E-plane metal lens, the phase distribution appears nearly uniform, what's more, the phase values in Fig. 5b, where the peak value of the field amplitude occurs in Fig 5a, are approximately equal. These results indicate that the
electromagnetic field on the horn aperture of Antenna 3 constitutes a constant-amplitude same-phase two-element array, and thereby produces a higher directivity. This is all contributed to introducing the $E$-plane acceleration lens in Antenna 3, which uniformises the field phase distribution on the horn aperture.

4.2 Simulated results of the proposed antennas

The simulated VSWRs, gains and radiation patterns of the three antennas are provided in Figs. 6 and 7, respectively.

The introduction of the dielectric with a comparatively bigger relative permittivity can limit the absolute bandwidth of the antenna; besides, the dielectric medium inside the aperture with a big relative permittivity and the air outside with a small one fail to obtain a transition process, which probably cause a great reflection of radiated energy, leading to limiting the impedance bandwidth. Therefore, the operating bandwidth of Antenna 1 is only 18–18.5 GHz (relative bandwidth 2.7%) with its VSWR less than 2, similar to the antenna in [21]. Aiming at this, some improvements have been operated on Antennas 2 and 3, where the loaded dielectric slab widths in Antennas 2 and 3 are designed a bit longer than the antenna aperture length to provide a transition which Antenna 1 lacks; nevertheless, the bandwidths of Antennas 2 and 3 are still a bit narrower than the traditional air-filled metallic waveguide $H$-plane horn antennas. However, one of the advantages of Antennas 2 and 3 is that they achieve the operating bandwidths of 14.7–18.5 and 15.2–18.3 GHz (respectively, relative bandwidth 22.9 and 18.5%), both obviously wider than Antenna 1. Moreover, the balanced microstrip line adopted in this paper filters a portion of high mode waves to consequently reduce the energy reflection, because it belongs to TEM mode wave transmission lines. The feeding line is designed to be gradient-structured, which is also beneficial to reducing the energy reflection and thus improve the antenna operating bandwidth.

Within the operating bandwidth, the gains of Antennas 1–3 rise one by one, which indicates that the horn aperture optimisation improves the aperture utilisation factor and accordingly increase the antenna gains. The growth of aperture utilisation factor brings about a problem of the sidelobe level rise in antenna radiation patterns, which is proved by the simulated results in Fig. 7.

What should be noted is, compared with Antenna 2, the gain of Antenna 3 has increased 3 dB (10log2 = 3 dB; the number 2 comes from the two-element array mentioned in Section 4.1) described by the simulated results in Fig. 6b, which proves the correctness of the analysis in Section 4.1.

The radiation efficiency can be obtained directly by simulation and the aperture utilisation factor $\nu$ can be calculated by Formula (3). At the typical operation frequency point 18.3 GHz (Antenna 1), 17.6 GHz (Antenna 2) and 17.6 GHz (Antenna 3), the simulated radiation efficiencies are, respectively, 0.88, 0.74 and 0.98, which indicates:

i. All the radiation frequencies are reasonable high, because the electric size is relatively big which leads better radiation and impedance matching.

ii. The rectangular dielectric slab loaded on Antenna 2 increases its aperture, however, the inhomogeneous distribution of amplitude and phase aperture field (see Fig. 5) results in the reduction of aperture utilisation factor $\nu$.

iii. In the respect of taking full advantage of the aperture, the introduction of the $E$-plane metal lens greatly improves the homogeneity of phase distribution, which causes the aperture utilisation factor increases close to 1.

iv. The introduction of the dielectric inside the horn antenna results in both the raise of aperture field density and the improvement of amplitude distribution homogeneity, which achieves the gain of aperture utilisation factor $\nu$.

$$\nu = \frac{90000 \times 10^{D/10}}{4\pi S F^2}$$

where $D$ is the antenna directivity (dB); $S$ is the antenna physical aperture area (mm$^2$); $F$ is the antenna operation efficiency (GHz).
5 Measured results of the proposed antennas

According to the associated geometrical parameters of antennas in Figs. 1–3, three antenna prototypes (Fig. 8) have been manufactured and measured in an anechoic chamber, using the Agilent E8363B Vector Network Analyser. The experimental results including the VSWRs, gains and radiation patterns of the three antennas are, respectively, as shown in Figs. 8–10.

The measured results indicate that antenna gains of Antennas 1, 2 and 3 rise one by one in the operating bandwidth, which well match the simulated results in Section 4. For example, at 17 GHz, the gain of Antenna 3 is 2.2 dB higher than that of Antenna 2 and 2.8 dB higher than Antenna 1. However, the measured and simulated results still differ a little, whose reasons and analysis are listed below:

i. Measured VSWRs indicate that the operating bandwidth of the Antennas 1, 2 and 3 are 17.97–18.65, 14.66–18.6 and 15.11–18.3 GHz, with the relative bandwidth 3.7, 23.7 and 19.1%, respectively. It is clear that the measured VSWRs values and their bandwidths of the three proposed antennas are consistent with the simulated ones. The small difference is mainly caused by the medium loss, and the difference of SMA coaxial connector and waveguide port feeding between in measurement and simulation.

ii. Compared with the simulated antenna gains, the measured ones decrease by about 2–3 dB, mainly caused by the dielectric medium loss in measurements. What should be noted is that, even though there exists the medium loss in measurements, the trends of the simulated results indicate that the gain of Antenna...
3 still gets obvious increased with the loaded E-plane dielectric-metal composite lens. This is because though the focusing effects of the dielectric slab can balance the loss caused by the dielectric slab, the antenna gain still obtains obviously increased because the loss of the loaded metal lens is much smaller.

iii. Compared with the simulated antenna radiation patterns, the measured ones are almost the same in the mainlobe scope, but differ a lot in the sidelobe. The differences between the measured and simulated results of the antenna radiation patterns are chiefly caused by the influence of fixtures, the shelter from the feed cable, the spot welds when fabricating the antenna and the discreteness of the permittivity of the dielectric slab. The measured front-to-back ratios (FBRs) of the three antennas all exceed 10 dB, which shows their strong directionality.

iv. The simulated result when changing the dielectric loss tangent from 0 to 0.02 is displayed in Fig. 9b, which is an extreme case. It is clear that a significant big decline of the antenna gain emerges which is mainly caused by the dielectric loss at high frequency. It should be emphasised that though the dielectric loss is introduced in the simulations and experiments, the tendency of the both results indicates the gain of the antenna with loaded composite lens increased remarkably. This is because the impact of the focus and loss of dielectric lens on the antenna gain can counteract with each other, however, as a result of a lower metal loss, the contribution to the gain appears evidently.

It can be seen from Fig. 9c, the aperture utilisation factor of Antenna 3 with loaded composite lens has increased significantly up to 0.8 maximum, while Antenna 1 without any load appears around 0.6 which matches the theoretical value ν = 0.64 of common H-plane horn antenna. The aperture utilisation factor of Antenna 2 loaded with dielectric lens performs lower, varied from 0.3 to 0.5, which is due to the non-uniformity of the aperture phase distribution.

6 Conclusion

Three types of printed H-plane horn antennas have been designed, manufactured, measured and analysed in this paper. It is the first time to introduce the SIW techniques into the design of E-plane metal lens, which is successfully loaded in the front end of the H-plane horn antennas to achieve a more than 2 dB gain improvement and have little effect on the operating bandwidth. This type of lens design technique will effectively reduce the antenna profile in its future application.
The operating band of the proposed antennas is decreased to Ku band without adding the thickness by applying the balanced microstrip line feeding structure to the antenna design, which can also greatly broaden the operating bandwidth.

Similar to many other printed antennas, the dielectric medium loss has an influence on the measured radiation patterns and antenna gains. To compensate the decrease of gain caused by the dielectric medium loss, the loaded metal lens proposed in this paper markedly increases the antenna gain and effectively prevents the main lobe of the antenna from being influenced by the dielectric medium loss.

The advantages listed above can be extended to the design of H-plane horn antennas with lower operating frequencies. Beneficial from the advantages, which the printed antennas bring, the antenna size can be greatly decreased and the manufacture will become easier to be achieved, which has a promising application in future.

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8 References

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