# GRIDDED CIRCULAR PATCH ANTENNAS

**G. Clasen[1] and R. J. Langley[1]**
[1] Electronic Engineering Laboratory
University of Kent
Canterbury, Kent CT2 7NT, United Kingdom

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**ABSTRACT:** *The performance of circular microstrip patch antennas is presented where the conducting patch and ground plane are constructed from a grid. Improved cross polarization, mode suppression, and bandwidth are possible, but are accompanied by lower gain and reduced front-to-back ratios. A computer simulation reveals the current structure. The density and shape of the grid pattern are important parameters. Such antennas are can be manufactured within glass laminates.* © 1999 John Wiley & Sons, Inc. Microwave Opt Technol Lett 21: 311–313, 1999.

**Key words:** *printed antennas; microstrip patch antennas; gridded patches*

## 1. INTRODUCTION

Microstrip antennas [1] consist of a printed conducting shape situated above a conducting ground plane. These conductors are usually continuous (solid). Rectangular printed antennas where the patches are made from a conducting grid (mesh) have been reported previously [2]. Improved bandwidth, but lower gain were the main observations. There will be applications, such as GPS and tolling, where it is desirable to print patch antennas within the glass laminate of a windscreen/rear screen. In the manufacturing process, it is not possible to use a solid silver patch or ground plane as this distorts the heating profile during lamination, causing structural distortions; hence, a gridded conducting structure must be used. This letter discusses the performance of a circular disk microstrip patch antennas where the solid conducting printed patch and the ground plane are replaced by conducting grids of appropriate geometry for the mode of patch excitation. Performance is compared with an equivalent conventional patch in all cases. For this study, the patches were initially printed on RT Duroid substrate, 0.78 mm thick, with a relative permittivity of 2.33 so that an assessment of the effects of meshing the antenna could be made. In later work, the antennas were screen printed on glass. The performance of conventional patches with glass subtrates and superstrates has been discussed elsewhere [4, 5].

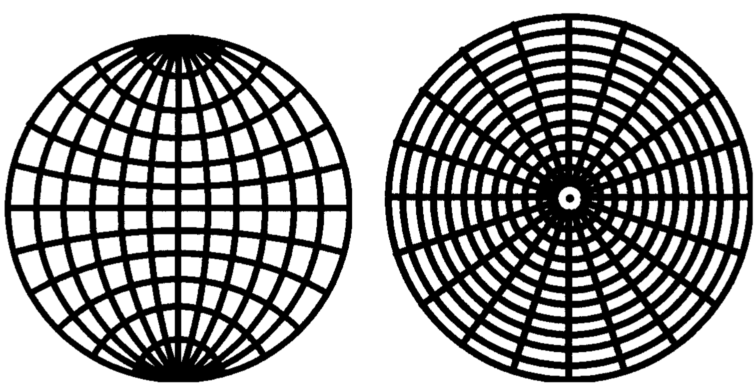
Figure 1 shows two gridded patches. The gridded patch in Figure 1(a) operates in the $TM_{11}$-mode; the grid line geometry is in the direction of the currents which flow on the patch, and is optimum and particular to this mode. Grid patterns have been designed for other modes of excitation [3], such as the $TM_{21}$ shown in Figure 1(b). An advantage is that gridding the patch suppresses unwanted mode propagation since the line geometry is designed for a given model of operation.

## 2. MEASUREMENTS

Experimental comparisons were made among four patch combinations: a standard circular patch and a grid patch over a solid conductor ground plane, and solid and gridded patches over a gridded ground plane. The gridded ground plane had the same conductor geometry as the patch directly beneath it, with an additional polar grid extending over the remaining ground plane beyond the patch diameter. In the examples shown in Figure 1, the standard patch was 75 mm in diameter with a ground plane extending to 130 mm, while the gridded patch was 60 mm over a 110 mm diameter ground plane. Both patches resonated at 1.48 GHz. Hence, a gridded patch is smaller for a given resonant frequency. Doubling the density of the grid lines on the same patch increases the resonant frequency from 1.48 to 1.69 GHz.

Figure 2 shows the 45° plane radiation patterns for a gridded patch and a solid patch; the gridded patch has twice the density of the lines shown in Figure 1(b). Gridding the patch while retaining a solid ground plane reduces the gain by 1.7 dB, but improves the cross polarization very significantly, from −16 to −24 dB. The bandwidth remains substantially the same. The gain decreases by a further 2 dB for the less densely gridded patch of Figure 1(b), but the cross polarization again improves.

Gridding the ground plane as well as the patch has several effects. In Figure 3, the full radiation patterns are plotted in the principle planes for a grid patch over both solid and gridded ground planes. Using a grid ground plane now improves the gain by 2 dB, but the front-to-back ratio degrades due to the partially transparent conductor, from 14 dB for the solid ground plane to 7 dB for a low-density gridded ground plane in the *H*-plane and 10 dB in the *E*-plane. The beamwidth increases by about 10% when a gridded ground plane is used, but more significantly, the bandwidth improves from 1 to 2.2%. However, the cross polarization increases to −9 dB, significantly higher than the standard patch.



(a) (b)

**Figure 1** Gridded patch antennas. (a) $TM_{11}$-mode. (b) $TM_{01}$-mode

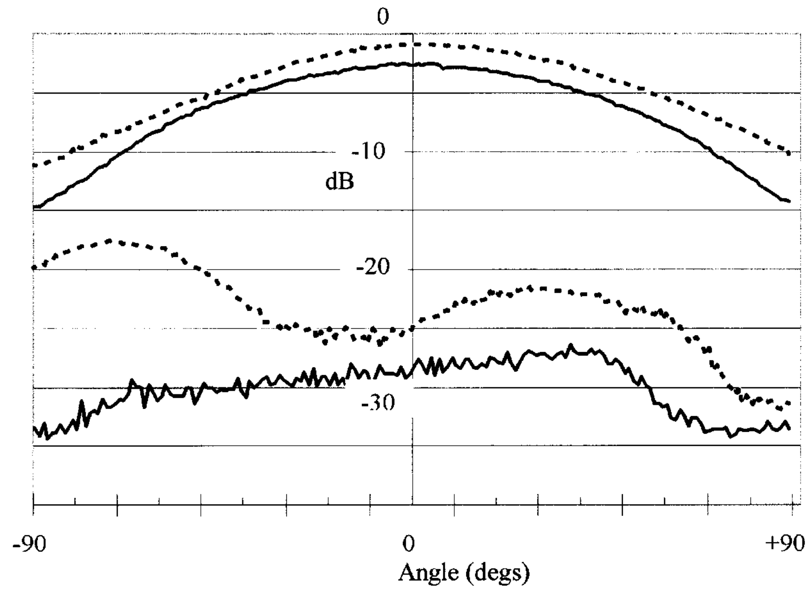



**Figure 2** 45° plane copolar and cross-polar radiation patterns for patches over a solid conductor ground plane. ---- solid patch. —— grid patch

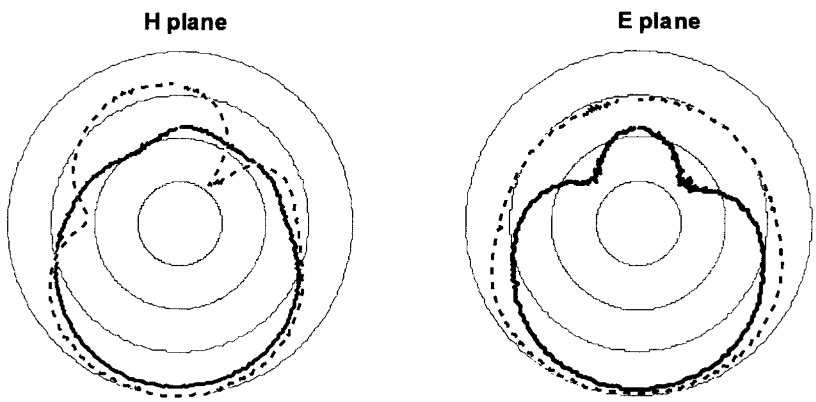



**Figure 3** *E*- and *H*-plane radiation patterns for gridded patch over different ground planes. —— solid ground plane. ---- gridded ground plane

## 3. SIMULATION

Figure 4 plots the current distribution on the gridded patch obtained from a simulation on HP Momentum software. For a conventional patch, the peak currents are toward the edge of the patch, and thus the aperture radiates from edges only. The peak currents for the grid patch are at the center, and the radiating fields emanate from all lines, not just those at the edge. Consequently, the radiating field amplitude tapers across the patch, and so gives rise to a wider beamwdith. An improved cross-polarization performance results due to currents being forced to follow the line geometry. It is very simple to show how changing the grid line direction degrades the cross polarization.

## 4. CONCLUSIONS

Preliminary results are presented for gridded circular patches. The advantages of using a grid patch with a solid ground plane are better cross polarization and unwanted mode suppression, but the antenna gain reduces as the density of the grid lines reduces. There are advantages in using a gridded patch together with a gridded ground plane at the expense of

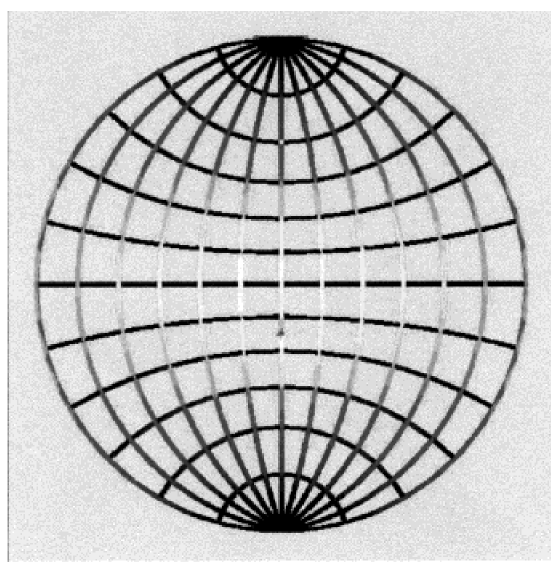


**Figure 4** Currents flowing on gridded patch

cross polarization and the front-to-back radiation levels, namely, the entire antenna can be printed into a glass laminate, a degree of optical transparency is therefore possible, and the bandwidth improves significantly. There is also an improvement in the gain, although it is still marginally lower than a standard patch antenna. These results indicate that there will be many applications where patch antennas can usefully be integrated into glass structures, particularly for use in mobile communications and telematics.

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# RADIATION FROM A WEDGE-TYPE NONRADIATIVE DIELECTRIC WAVEGUIDE RADIATOR

**J. A. G. Malherbe,**[1]
[1] College of Engineering
University of Pretoria
Pretoria 0002, South Africa

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**ABSTRACT:** *The radiation properties from an open-ended wedge-type nonradiative dielectric waveguide are measured and compared to calculated values. The analysis is based on the assumption of a quadratic phase function for the field in the radiating aperture, as is used for the analysis of electromagnetic horns. It is found that the theoretical approach is valid.* © 1999 John Wiley & Sons, Inc. Microwave Opt Technol Lett 21: 313–315, 1999.

**Key words:** *nonradiative dielectric waveguide; antennas*

## 1. INTRODUCTION

The nonradiative dielectric (NRD) waveguide has been identified as a very useful structure as a guiding medium for microwave and millimeter-wave frequencies. This is due to the fact that, not only can circuit components such as couplers, filters, etc., be fabricated in an NRD waveguide, but antennas can also be integrated into circuits. A number of papers have recently described radiating elements that are, in effect, open-ended NRD waveguide structures that are unterminated [1–4]. They offer advantages of simplicity of construction, together with the integrated nature inherent in the structure. The radiation pattern of the open-ended NRD waveguide radiators is fixed by the structure of the waveguide, and sharper radiation patterns can only be obtained by arranging elements in an array [5–7].

In this paper, the radiation properties of a type of NRD waveguide that tapers as a wedge are examined; this structure makes it possible to increase the size of the radiating aperture without the need for a complex physical structure, thereby increasing the gain of the element. A theoretical analysis is developed along the lines described in [4] for a rectangular cylindrical cross section of the dielectric portion of the waveguide, but a quadratic phase term is introduced to describe the phase properties of the fields in the radiating aperture. The radiation pattern so obtained is compared to measured data.

## 2. ANALYSIS

The physical construction of the antenna is shown in Figure 1. It consists of a slab of dielectric material sandwiched between two conducting plates, as for a conventional NRD waveguide, but the slab is wedge shaped, with the feed at the narrow dimension, and the wide dimension forming the radiating aperture. The feed width is determined by the network in which the wedge is to be employed as a radiator.

The phase front of the fundamental NRD waveguide mode in the wedge is assumed to be cylindrical at the aperture, and the introduction of a quadratic phase error in the illuminating function will therefore describe the radiating fields, as is conventionally done for the analysis of horn antennas [8]. Due to the phase error, the fields across the aperture will vary as

$$E_y(y) = E_y(o)e^{-j(\beta_g/2L)y^2} \tag{1}$$

where $\beta_g$ is the propagation constant of the wave in the aperture, and $L$ is the "length" of the wedge, as shown in Figure 1. For a wide wedge, the total variation of the fields in the aperture can be approximated by the product of the phase error function and the conventional NRD waveguide modes as described in [4]. In the dielectric (region I), the fields are then given by

$$E_{y_1} = \frac{\left(k^2\epsilon_r - \beta_y^2\right)}{j\omega\mu_0\epsilon_0\epsilon_r}\cos(\pi x/a)\cos(\beta_y y)e^{-j\beta_g z}e^{-j(\beta_g/2L)y^2} \tag{2}$$

$$H_{x_1} = \frac{j\beta_g}{\mu_0}\cos(\pi x/a)\cos(\beta_y y)e^{-j\beta_g z}e^{-j(\beta_g/2L)y^2} \tag{3}$$

while the fields in region II are given by

$$E_{y_2} = \cos(\beta_y b/2)\frac{\left(k^2 + \alpha_y^2\right)}{j\omega\mu_0\epsilon_0} \times \cos(\pi x/a)e^{-\alpha_y(|y|-b/2)}e^{-j\beta_g z}e^{-j(\beta_g/2L)y^2} \tag{4}$$

$$H_{x_2} = j\cos(\beta_y b/2)\frac{\beta_g}{\mu_0} \times \cos(\pi x/a)e^{-\alpha_y(|y|-b/2)}e^{-j\beta_g z}e^{-j(\beta_g/2L)y^2}. \tag{5}$$