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Biconical Horn Antennas for Near Uniform Coverage in Indoor Areas at Mm-Wave Frequencies

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Abstract—A method is presented for designing and dimensioning biconical horn antennas in such way that the level of received power does not depend strongly on the separation distance between a centralized base station and a remote radio terminal within an indoor pico-cell. Results of cell-coverage measurements at 58 GHz are presented using these antennas within eight different indoor environments. The measurement results show that, using biconical horn antennas, an overall uniform coverage can be achieved for both line-of-sight and obstructed line-of-sight topologies.

I. INTRODUCTION

INDOOR radio LAN's operating in the millimeter (mm) frequency range may offer a large information transport capacity and sharply defined cell boundaries. The use of mm-wave frequencies enables the creation of high traffic/user density cells, because at mm-wave frequencies the boundaries of an indoor pico-cell (i.e., cell radius < 100 m) are formed by walls and floors consisting of "hard" materials, like concrete and steel, through which mm-waves cannot propagate [1], [2]. If individual rooms have such walls, then frequency reuse based on a one-cell-per-room scenario is possible on one and the same floor (without causing significant interference to neighboring cells). If, on the other hand, only the superstructure of the building consists of "hard" materials, whereas individual rooms have walls made of "soft" materials like plasterboard, sheet rock or gypsum which give only low attenuation, then frequency reuse based on a one-cell-per-floor scenario is possible. Radio coverage within a pico-cell can be controlled and optimized by appropriately dimensioning the applied antennas. In this paper, the achievement of near uniform coverage in an indoor pico-cell by compensating path loss by antenna gain is treated. The principle is illustrated schematically in Fig. 1.

II. METHODOLOGY

For the treatment of the coverage problem in a multipath medium, we will consider an indoor radio link consisting of a base station (BS), a remote radio terminal (RRT) and the multipath channel in between. The height difference between the remote antenna and the base station antenna is denoted by the parameter $\Delta h$, while the variable $r$ represents the separation distance between the two antennas, as depicted in Fig. 1. The antennas are identical biconical horn antennas with a geometry according to Fig. 2. They consist of a radial section, i.e., the spacing between the lower antenna part and upper antenna part with spacing distance $a$ and diameter $b$, a circular waveguide with interior diameter $d$, and a biconical horn. The horn has an aperture width $A$ and a length $L$, which is measured from the (virtual) horn apex in the radial section to the center of the aperture. The circular waveguide contains a polarizer which transforms the incoming linear $TE_{01}$ mode into two perpendicular-directed $TE_{11}$ modes $90^\circ$ out of phase, thus resulting in a circularly polarized wave. Since the radial line is excited with a circularly polarized wave, the biconical horn exhibits an omnidirectional radiation characteristic in the azimuth plane. The
polarization of the launched wave is elliptical for $1/2\lambda < a < \lambda$ and vertical for $a < 1/2\lambda$, where $\lambda$ is the wavelength in free space [3]. For a good omnidirectional radiation pattern, only one polarization state should be considered. Suppression of the modes which support the vertical electric field component for $1/2\lambda < a < \lambda$ is mechanically difficult [4]. This problem will not arise if $a < 1/2\lambda$ is taken, because the radial section will support only one mode in this case. This mode will generate a vertically polarized radiation pattern.

The radiation pattern in the elevation plane is determined by the horn dimensions $A$ and $L$. The propagation path length $l$ from the horn apex to the horn aperture increases towards the horn edges, so the aperture plane is not an equiphasal plane. The phase variation in the aperture plane is given by $\exp(-j2\pi(l - L)/\lambda))$. This is similar to the aperture phase distribution of an E-plane sectorial horn antenna with aperture width $A$ and length $L$ in the E-plane. The radiation pattern in the elevation plane can therefore be calculated by the methods described for sectorial horns excited with the corresponding mode, which is the TE$_{01}$ mode in our case [5]. The elevation electric field pattern (magnitude), defined as $|F(\theta, s)| = |E(\theta, s)|/|E(0, s)|$ with $\theta = \arcsin(\Delta h/r)$ can be expressed as [6]

$$|F(\theta, s)| = \frac{1 + \cos(\theta)}{2} \times \left[\frac{\sqrt{C(r_2) - C(r_1)}^2 + [S(r_2) - S(r_1)]^2}{4C^2(2\sqrt{s}) + S^2(2\sqrt{s})}\right],$$  

(1a)

where $C(x)$ and $S(x)$ are the Fresnel integrals defined in [5] and

$$r_1 = 2\sqrt{s} \left[1 - \frac{1}{4s} \frac{A}{\lambda} \sin \theta \right],$$

$$r_2 = 2\sqrt{s} \left[1 + \frac{1}{4s} \frac{A}{\lambda} \sin \theta \right].$$

(1b)

with parameter $s$ being the phase error on the antenna aperture, which can be expressed as

$$s = \frac{1}{8} \frac{A^2}{\lambda} \frac{1}{L/\lambda}.$$  

(1c)

If only the line-of-sight (LOS) ray is considered, then the relative power/distance relationship can be obtained by noting that

$$\frac{P_r(r, \theta, s)}{P_t} \sim \frac{D^2(s)}{|F(\theta, s)|^4} \left(\frac{\lambda}{4\pi r}\right)^2,$$  

(2)

in which $P_r$ represents the received power, $P_t$ is the transmitted power and $D(s)$ represents the antenna directivity.

Fig. 3 shows curves of relative received power against the product of the antenna-dimension parameter $\lambda/A$ and antenna-distance/height-difference ratio $r/\Delta h$ for different values of $s$ relative to the no phase error case (i.e., $s = 0$, the aperture plane is an equiphasal plane). The factor $(1 + \cos \theta)/2$, which appears in (1a), is not included in Fig. 3, since it has a negligible effect for most situations. These plots are universal coverage curves from which values of $A$ and $L$ can be derived for a certain coverage. It appears from Fig. 3 that a coverage is achieved with only about 3 dB variation in received power versus distance in the region $0.6 < (\lambda/A)(r/\Delta h) < 5$, if we introduce a phase error $s = 0.32$ on the antenna aperture. This implies for most indoor cells with a maximum distance/height ratio requirement $(r/\Delta h)_{max} = 30$ that $\lambda/A = 1/6$, resulting in a minimum distance/height ratio $(r/\Delta h)_{min} = 3.6$. Note that $(r/\Delta h)_{min}$ can be decreased by increasing $\lambda/A$ at the expense of $(r/\Delta h)_{max}$.

The coverage curves of Fig. 3 are derived using the LOS ray only. The reflected rays shown in Fig. 1 will contribute to the uniform coverage in the same manner as the direct ray if the total ray length is substituted for separation distance and if additional losses due to reflections are taken into account. For small separation distances, the reflected rays are likely to dominate the direct ray, hence the additional power of reflected rays will thus contribute to a better overall coverage performance.

III. ANTENNA DIMENSIONING

For experimental verification, we constructed a pair of identical horn antennas having a geometry as depicted in Fig. 2. The antennas are designed to operate within a 2 GHz bandwidth centered around 58 GHz ($\lambda = 5.2$ mm). We took $\lambda/A = 1/6$ and $s = 0.32$ for near uniform coverage up to $(r/\Delta h)_{max} = 30$ as determined earlier, resulting in $A = 31.2$ mm and $L = 73.1$ mm.
For \( a = 2 \) mm, the standing wave ratio of both constructed antennas could be tuned by the screw (see Fig. 2) to an average value of SWR = 1.4 over the considered bandwidth. Radiation patterns were measured in an anechoic room. A 2 mm radial section spacing distance resulted in a vertically polarized radiation pattern with at least 25 dB suppression of the horizontally polarized field components. Furthermore, a maximum deviation of about 2 dB from an omnidirectional radiation pattern was found. A typical measured radiation pattern in an elevation plane is depicted in Fig. 4, together with a theoretical pattern according to (1a)-(1c). Good agreement between theoretical and measured results is shown for elevation angles ranging from \(-60^\circ\) to \(60^\circ\) (and \(120^\circ\) to \(240^\circ\)). Outside this range, horn edge diffraction becomes dominant [8]. These diffraction effects, which are not included in (1), have a small favorable effect on the coverage in the service area close by the base station. The antenna directivity \( D(s) \) can be determined by pattern integration which results for both measured and theoretical pattern in \( D(0.32) = 9 \) dB. The 3 dB beamwidth of both patterns is \(9^\circ\). The good agreement between theoretical and measured results indicates that biconical horn antennas can be designed with clearly defined characteristics for the use in propagation measurements.

IV. PROPAGATION MEASUREMENTS

In order to verify the theory presented in the previous section, power versus distance measurements in various buildings of the Eindhoven University of Technology were performed. The measurements were based on a frequency step sounding technique as described in [9]. The measurement environments can be considered as a single room only, because mm-waves are severely attenuated by most inner walls. In the following, these rooms are denoted A to H. The diagram in Fig. 5 shows the approximate dimensions of each room and measured reflection coefficients of dominant wall material for each room, i.e., wood for A and G, rock wool for B, concrete for C, D and H and metal for E and G. The values of the reflection coefficients are obtained by measuring the ratio of the reflected ray amplitude at perpendicular incidence and the reflected ray amplitude obtained by covering the wall under consideration with a metal plate. The depicted values in Fig. 5 are dB values of the associated ratios of reflected ray power and incident ray power. A more detailed description of the measurement environments is given in [9].

The measurements have been carried out at approximately 20 randomly chosen positions for the remote radio terminal in each room. The biconical horn antenna on the remote station was located 1.4 m above the floor. The biconical horn on the base station was elevated to 3 m. The base station was placed in the center of the room. Both antennas were leveled horizontally at every measurement position. The measured transfer function at each position of the remote antenna \( H(f) \) can be used to determine the normalized received power by

\[
\frac{P_r}{P_t} = \frac{1}{f_{\text{max}} - f_{\text{min}}} \int_{f_{\text{min}}}^{f_{\text{max}}} |H(f)|^2 df, \tag{3}
\]

with \( f_{\text{min}} = 57 \) GHz and \( f_{\text{max}} = 59 \) GHz. A scatter plot of the normalized received power versus separation distance on a log scale for Room D is shown in Fig. 6 for both LOS and obstructed LOS (OBS) situations.

Fig. 6 shows irregular variations of received power values. This is mainly caused by the differences of power contributions associated with reflected rays; especially in Room D, reflected rays appear and disappear suddenly while moving the remote radio terminal through the room. This is a result of incidental
reflections from a number of metallic cabinets which are spread throughout the room. In addition, the nonideal radiation patterns in the azimuth plane of both antennas (about 2 dB deviation from omidirectional) contribute to the spread in received power. Furthermore, multipath effects as a result of constructive and destructive addition of phasor terms associated with some dominating rays of roughly equal intensity may contribute to the spread in received power. However, deep nulls as encountered in conventional narrow-band indoor radio links are absent, because of the frequency diversity which is inherent to the large bandwidth.

The lines in Fig. 6 are linear fits based on a minimum mean square error criterion. If the following linear relationship is assumed,

\[ 10 \log_{10}(P_r/P_t) \sim -10\alpha \log_{10}(r), \]

where the parameter \( \alpha \) expresses the decay exponent of the received power versus distance due to the combination of path loss and antenna gain compensation, then the slope of the regression line gives the experimental value of \(-\alpha\).

The decay rate \( \alpha \) of all measurement sets is determined and listed in the diagram in Fig. 7 for every environment, together with the average and standard deviation of the normalized received power values (\( P_r/P_t \) ratios are averaged, not dB values).

The diagram in Fig. 7 shows decay exponent values close to zero for LOS situations, except for Room G. This exception can be explained by the fact that in this room: 1) the direct ray does not provide uniform coverage because \( r < 6 \) m for all measurement positions, and 2) the reflected rays cannot dominate the direct ray because the walls of this room are made of wood, which has a low \(-13\) dB reflectivity. The same effect, although less serious, can be observed for Room H with concrete walls having \(-2\) dB reflectivity. This effect does not appear in Room F since this room has metal walls, so individual reflected rays dominate the direct ray because their path lengths fall in the uniform coverage range of \( r > 6 \) m. The conjecture that the reflected rays dominate the direct ray is confirmed by the fact that the measured values for LOS and OBS situations are almost the same.

By comparing the average and standard deviation of power values in Fig. 7 for different rooms, it is obvious that rays reflecting against the metal objects which are present in Room D contribute to the average power level, and therefore improve coverage. However, they also contribute to the spread in received power. The presence of regular metallic structures, like the smooth metallic walls of Room E, results in a relatively high average power value with a relatively small standard deviation.

In all environments except Room G, equal or only slightly lower power values are obtained for \( r < 6 \) m, in comparison to the average of measured values in the uniform coverage range (\( r > 6 \) m). In rooms with many metal objects and/or metal walls (D and E), no difference in average power values for either region was found, while only slightly lower values are measured for \( r < 6 \) m in rooms without strongly reflecting surfaces. The power contributions of reflected rays are responsible for this, as explained earlier.

Not only in the \( r < 6 \) m region but also in the \( r > 6 \) m region, remarkably small differences (of only a few dB's) between LOS and OBS situations are measured. This demonstrates the significance of reflected rays in maintaining coverage in both regions under OBS conditions. This indicates that the theoretical lower bound of the uniform coverage region, as derived for the direct ray, is a less relevant design constraint if the antennas are to be designed for operation within a highly reflective room. It is clear that for such rooms, \( (r/\Delta h)_{\text{min}} = 3.6 \) may be taken as a lower bound for a near symmetric coverage if only the direct ray is considered and if a uniform coverage range of about 50 m is the design objective. In order to achieve uniform coverage in small and low reflective indoor environments as Room G, however, it would be more appropriate to choose a lower value for \( (r/\Delta h)_{\text{min}} \), resulting in smaller antenna directivity and smaller antenna dimensions.

V. CONCLUSIONS

A method has been derived for designing and dimensioning biconical horn antennas that gives near uniform coverage within a pico-cell in the sense that the level of received power does not depend strongly on the position of a remote radio terminal within that pico-cell.

Using a set of properly dimensioned biconical horn antennas, cell coverage in the 57–59 GHz band has been measured within a number of indoor pico-cells. Results obtained under both LOS and OBS conditions have been presented separately.

For all but one of the measured indoor areas, the absolute value of the power decay rate exponent was found to be less than one. This is relatively small compared to power-distance relationships commonly found in the UHF band where (near) omnidirectional antennas are used [10], [11]. It is demonstrated...
that, even though the antennas mainly radiate in the horizontal direction, good coverage is still achieved just beneath the base station antenna (separation distance smaller than 6 m). Reflected rays are responsible for maintaining coverage in this region. Uniform coverage throughout a pico-cell is a highly desirable feature of a radio network if fair access and equality in performance for all remotes is a design objective. The principle of antenna gain compensation to achieve this has the additional advantage that the radio link with the largest separation distance between base station and remote radio-station, being normally the worst case situation, experiences full antenna gain of both antennas.

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