Abstract:

This report covers a major, three-year work package (WP 3.2) for the CAPANINA programme. The remit for the research was very broad in scope, as it concerned antenna technologies for high altitude platform communications. In this context, particular challenges were encountered by the high carrier frequencies involved, the required high data rates for broadband services to users 'on-the-move', and the consequent need for highly directive antenna beams to track their targets.

Attention was paid to antennas for a HAP payload and antennas for a high speed vehicle. For the HAP antenna, where multi-beam functionality is particularly important, the spherical lens antenna was developed. A two-layer dielectric lens produced an efficiency comparable with a conventional reflector antenna but with a reduced height profile and provided the potential for multiple scanned beams over a very wide angle. For the train antenna, a mechanically steered small and light weight array antenna was identified as the most promising and prototypes were constructed and evaluated. The highly directive Strip-Slot-Foam-Inverted Patch antenna enabled implementation of an array with a reduced number of elements; thus decreasing feeding network complexity. A 64-element prototype was tested and the measured results compared well with the theoretical computations.

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EXECUTIVE SUMMARY

This report covers a major, three-year work package (WP 3.2) for the CAPANINA programme. The remit for the research was very broad in scope, as it concerned antenna technologies for high altitude platform communications. In this context, particular challenges were encountered by the high carrier frequencies involved, the required high data rates for broadband services to users 'on-the-move', and the consequent need for highly directive antenna beams to track their targets.

Thus the theme of scanning (or steerable) antennas underpins the enquiries which have been reported. In chapter 2, possible approaches are considered for communicating with the ground from high aerial vehicles and the other ground-end of the link, particularly on a moving vehicle such as a fast train, is examined as well. The work builds to some extent on the earlier work carried out by the HeliNet project under the 5th Framework Programme. Having defined the scenario, Chapter 3 deals with candidate solutions for mechanically steered antennas, phased array antennas, spherical lens antennas and describes the prior art. For the train antenna, a mechanically steered small and light weight array antenna was identified as the most promising. For the HAP antenna, where multi-beam functionality is particularly important, the spherical lens antenna was selected for further work.

Chapter 4 describes the experimental work carried out on the two preferred approaches selected in Chapter 3 and covers the work carried out on the Broadband Printed Array Antenna at CSEM and on the Hemispherical Lens Prototype at UOY. In both cases there is the assumption that beam steering is primarily a mechanical operation.

In chapter 5 the practical constraints for antenna steering for a train are examined and Chapter 6 lists the available electrical and mechanical steering technologies currently available.

Having dealt with train aspects in Chapter 5, Chapter 7 concentrates on a HAP cellular network at the system level and considers some antenna payload issues, for which the spherical lens antenna was selected as the most appropriate.

To summarize, there have been broadly two areas of investigation:

(i) Antenna(s) for HAP payload

(ii) Antenna(s) for high speed vehicle.

The requirements for both share much common ground.

For the HAP payload antennas for the millimetre-wave bands, near-term solutions would be provided by dedicated beam antennas. Medium term solutions are offered by multi-beam lens antennas, where a lens aperture is shared by a number of feeds, thus minimising payload mass. Long-term solutions may be offered by digital beam forming or "smart antennas". While all techniques have been considered during the course of WP3.2, and experimental data gathered for aperture and lens antennas, the long term solution (digital beam forming) has not been explored in practice under WP3.2: rather, the signal processing aspects of this technology lie under the remit of WP3.3

For the vehicular antenna, a similar technology road-map has been identified, i.e. near term (single beam mechatronic antenna), medium term (multi-beam mechatronic lens antenna), and again the long term solution offered by Digital Beam Forming (DBF), which is considered elsewhere (WP3.3).
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<th>Description</th>
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<tbody>
<tr>
<td>AF</td>
<td>Array Factor</td>
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<tr>
<td>AR</td>
<td>Axial Ratio</td>
</tr>
<tr>
<td>BW</td>
<td>Bandwidth</td>
</tr>
<tr>
<td>COTS</td>
<td>Commercial Off-The-Shelf</td>
</tr>
<tr>
<td>CP</td>
<td>Circular Polarisation</td>
</tr>
<tr>
<td>DBF</td>
<td>Digital Beam Forming</td>
</tr>
<tr>
<td>EIRP</td>
<td>Effective Isotropic Radiated Power</td>
</tr>
<tr>
<td>FSL</td>
<td>Free Space Loss</td>
</tr>
<tr>
<td>HAP</td>
<td>High Altitude Platform</td>
</tr>
<tr>
<td>LOS</td>
<td>Line Of Sight</td>
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<tr>
<td>GLL</td>
<td>Grating Lobe Level</td>
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<td>MLL</td>
<td>Main Lobe Level</td>
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<tr>
<td>RX</td>
<td>Receiver</td>
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<tr>
<td>SLL</td>
<td>Side Lobe Level</td>
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<tr>
<td>SOTA</td>
<td>State Of The Art</td>
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<tr>
<td>SSSFIP</td>
<td>Strip-Slot-Foam-Inverted Patch</td>
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<td>TX</td>
<td>Transmitter</td>
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1 Introduction

Today’s users of broadband services are generally dependent upon either cable or satellite technologies for the connection to their home or office. Cable can provide excellent capability, particularly fibre, but is only commercially viable in high usage areas. Satellite provides widespread geographic coverage but has limited capability and changing capacity. Communications via High Altitude Platforms (HAPs) will develop broadband capability from aerial platforms to deliver cost effective solutions providing a viable alternative to cable and satellite, with the potential to reach rural, urban and travelling users.

High Altitude Platforms (HAPs) are airships or planes, operating in the stratosphere, at altitudes of typically 17 - 22km (around 75,000 ft). At this altitude (which is well above commercial aircraft height), they can maintain a quasi-stationary position, and support payloads to deliver a range of services: principally Communications, and Remote Sensing.

Communications services include broadband, 3G, and emergency communications, as well as broadcast services. A HAP can provide the best features of both terrestrial masts (which are then not required) and satellite services (which would be highly expensive). In particular, HAPs permit rapid deployment, and highly efficient use of the radio spectrum (largely through intensive frequency re-use).

New types of antennas needed to provide these communications, in particular antennas for the HAP payload and antennas for ground stations of which one of the most difficult to provide will be for a high speed vehicle such as a train. The technology developed aims to support data rates of up to 120 Mbit/s to fixed and moving users anywhere within a HAPs 60 km coverage area. In ITU-R [29], the frequency band allocated for HAPs applications is defined to be 27.5-31.3 GHz.

HAPs offer an effective way of exploiting mm-wave spectrum by supporting multi-cell architectures. The viability of such systems is largely determined by the overall system data capacity, which is in turn governed by the properties of the antenna payload which serves the cells on the ground. In previous studies, "ideal" antenna beams have been used to model carrier-to-interference ratio. In such cases, dedicated aperture antennas such as lens antennas may be used to produce the required beam shapes which are in general asymmetric. This approach leads to one antenna for each cell and thus a bulky payload. Alternative ways are needed to minimise the HAP payload and to keep the train antenna as small as possible.
2 Possible Approaches for the RF, EM and Mechatronics Aspects of the Train and HAP Antennas

There are elements of the antenna specification that are essentially the same for every antenna type.

2.1 System Requirements and Link budgets

The link budget can be defined to provide differing levels of service, the biggest factor in service level being the permitted duration of rain outages. A link budget for 28GHz with a HAP at 17km altitude and a radial distance of 30km from the sub-platform point can be seen in Table 1. This does not make any assumptions about the transmit power or antenna gains, but does suggest three different modulation schemes requiring different signal to noise ratios. In addition three different levels of service are defined 99.99%, 99.90% and 99.0% availability, the result at the bottom of the table are 9 different numbers representing the required sum of transmit power (dBm) plus HAP antenna gain (minus feed losses) plus customer antenna gain (minus feed losses). It is envisaged that different combinations of antennas and powers will result in different levels of service provision.
### Transmitter (HAP)

<table>
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<td>Required Amplifier Input Back-off</td>
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<td>0</td>
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<td>Power per backhaul carrier (dBm)</td>
<td>26.0</td>
<td>26.0</td>
<td>29.0</td>
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<td>Antenna gain (dBi)</td>
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<td>Antenna feed loss (dB)</td>
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<tr>
<td>HAP EIRP (dBm)</td>
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<td>29.0</td>
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### Receiver (Ground Station)

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<td>Thermal noise density (dBm/Hz)</td>
<td>-173.8</td>
</tr>
<tr>
<td>Receiver noise figure (dB)</td>
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<tr>
<td>Receiver noise density (dBm/Hz)</td>
<td>-168.8</td>
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<td>Receiver interference noise density (dBm/Hz)</td>
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### Link Parameters

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<td>Platform Height (km)</td>
<td>17.0</td>
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<tr>
<td>LOS Distance (km)</td>
<td>34.48</td>
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<td>FSPL (dB)</td>
<td>152.1</td>
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<tr>
<td>Edge of cell and antenna beam losses</td>
<td>5.0</td>
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<tr>
<td>Clear air losses (dB)</td>
<td>157.8</td>
</tr>
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</table>

### Table 1 Link Budget for 28GHz

| Received margin clear air (dB)   | -55.6 | -50.1 | -39.3 |
| Minimum required transmit power clear air (dBm) | 81.6 | 76.1 | 68.3 |

| Rain Attenuation Zone K 99.99% (dB) | 32.5 |
| Rain Attenuation Zone K 99.90% (dB) | 12.3 |
| Rain Attenuation Zone K 99.00% (dB) | 3.3  |

| Received margin 99.99% (dB) | -88.0 | -82.6 | -71.8 |
| Received margin 99.90% (dB) | -67.8 | -62.4 | -51.6 |
| Received margin 99.00% (dB) | -58.8 | -53.4 | -42.6 |

| Minimum required transmit power (dBm) + Antenna Gains 99.99% | 114.0 | 108.6 | 100.8 |
| Minimum required transmit power (dBm) + Antenna Gains 99.90% | 93.8  | 88.4  | 80.6  |
| Minimum required transmit power (dBm) + Antenna Gains 99.00% | 84.8  | 79.4  | 71.6  |
2.2 Geometry

The geometry of the train antenna can be flexible with the one constraint that it needs to be commensurate with installation on top of a high speed train and hence should have a reasonably low profile.

2.3 Required Scan Range

The sweep angle for the train antenna should be 150°. The figure is based on a HAP elevation angle is 30° (i.e. above the horizon), combined with the attitude of the train (in terms of maximum levels of track camber etc.) which has been estimated at 15°, leading to an effective elevation angle of 15°. It is possible that there might have to be a reduced link specification at these extremes of the coverage area for example a reduced data rate or more coding.

2.4 Coverage Area

Within the HeliNet project the coverage area was defined as a 60 km diameter circle. This was based on a maximum elevation angle of the HAP from the ground user of approximately 30°.

3 Candidate Solutions and prior art

It is intended that this section will contain many subsections each dealing with a different alternative solution exploring solutions previously reported in the literature as well a new or hybrid solutions.

The antennas could be anything from a full phased array through mixed electrically and mechanically steered to a single aperture antenna that is entirely mechanically steered, the aim is to explore the solution space and set out:

- advantages and disadvantages
- flexibility
- complexity
- cost
- power requirements
- performance
- geometry
- steering rate

3.1 Scenario

The scenario of interest is that of a terminal mounted on a moving vehicle (i.e. a high speed train) which communicates with a terminal / transponder on an aerial platform i.e. a HAP. This scenario is not too dissimilar with satellite communications to vehicles. It is thus of much commercial interest and a mainstream topic in communications research and technology.

It is worth listing some of the characteristics and requirements of this scenario:

- Scanning to low elevation angles (a figure of 15° has recently been mooted).
- ...thus, about 150° conical scan - almost a full hemisphere.
- any loss in the antenna gain associated with scanning to a low elevation angle is particularly disadvantageous - it is here where maximum gain is need to counter the free-space and rain losses associated with the path length.
The speed of scanning over large angles is relatively slow - this is associated with the change in location of the vehicle from place to place.

The speed of scanning over smaller angles may need to be rapid - this is associated with changes in attitude of the vehicle.

Both transmit and receive functionalities are required.

A multi-beam function would allow for communication with multiple platforms.

Constraints on the terminal: The terminal carried by the vehicle (train) is probably not too constrained by mass, or area, as we assume it would be roof mounted. It is probably not too constrained by power. It may have some constraint on physical height, and probably a medium constraint on cost.

### 3.2 Candidate technologies

The scope here is very large. A few candidates are listed.

#### 3.2.1 Mechanically steered antenna

The antenna might be a reflector, lens-horn, or array type. This would be one of the simplest solutions, and would have good RF performance. The flexibility of the RF feed, and a possible requirement for a rotary RF joint, could be an issue. It could be advantageous to include a frequency-converter at the feed (as with domestic satellite TV receiver) - this would have to be powered, and would add to the steered mass. The antenna would most likely require housing with an RF transparent cover (radome) to isolate it from wind stresses, rain etc.

This is not a very elegant solution, although it may be a relatively cheap one. Disadvantages would include the inability to produce independently steered, multiple beams over a wide angle. High-speed scanning to maintain track as the vehicle attitude changes may be problematic. Some isolation from pitch, roll and vibration may be needed. A narrow-angle multi-beam facility should be feasible for reflector with multiple feeds, and this could assist with high-speed tracking, in a similar way to monopulse radar. This multi-beam facility would require multiple feeds, further adding to the mass which is steered.

There may be commercially available products which a close to meeting our requirements, e.g. ship mounted satellite terminals, although the majority of these are likely to be for Ku band.

#### 3.2.1.1 Mass of typical antennas

In this section some typical antennas for 28 GHz are shown as these can provide a first estimate of typical dimensions and mass.
Lens-corrected horn.

A group of such items are illustrated in Figure 2, which are manufactured by Flann in the UK.

![Lens-corrected horn antennas](image)

**Figure 2 Lens-corrected horn antennas**

The 150 mm lens aperture yields two orthogonal half power beamwidths of approximately 5° and 6°. The gain is about 30 dBi. The aperture diameter is 160 mm and the horn length is about 250 mm. The mass is 740 g. A reflector antenna would give comparable performance for similar mass and volume.

**Plastic dual-reflector antenna.**

We have (on loan) a compact and lightweight reflector antenna manufactured by ERA Ltd. to meet the demands of a mass-producible and cost-effective antenna for broadband applications at 28 GHz.

![Plastic dual-reflector antenna](image)

**Figure 3 Plastic dual-reflector antenna.**

Measurements of the antenna shown indicate a gain of around 27 dBi. The aperture measures 140 mm x 100 mm, and its weight is approximately 80 g. It has been developed for a steerable base station antenna for terrestrial applications. Due to its compactness and small weight, it would present a modest load to mechanical actuators. This item may not be commercially available, or at least difficult to obtain.

### 3.2.2 Mechanically steered small high gain array

This solution consists of a relatively small high gain array and a high precision fully mechanical beam steering system.

The advantages are the antenna’s low profile, low cost, small size (~20×20cm), light weight (<1Kg) and construction with fully passive components. According to the results reported by Japanese researchers, a gain of more than 29 dBi has been achieved using a three-layer 16×16-element array [25]. The potential disadvantage of this approach is that there is no fast wide angle electrical beam steering. This can be compensated by a rapid and high precision mechanical steering system.
One scenario of the mechanical system that can be imagined is a two-axis steering system driven by the desired elevation and azimuth angles. The train speed, pitch/roll/vibration and banking angle in the turns should be considered as important constraints in the mechanical design. The approximate dimensions of the mechanical part should fill in a box of dimensions 25×25×22 cm (Figure 4).

Appending to Figure 5, the scan angle speed can be computed as

$$\theta = \tan^{-1}\left(\frac{300 \text{ km}}{3600 \text{ s}} \cdot \frac{1}{20 \text{ km}}\right) = 0.24^\circ / \text{s}$$

The required pointing angle precision can be estimated by taking an example of a square horn whose beam shape is approximated using function

$$\left[\frac{\pi d}{\lambda} \cdot \sin(\phi) \right]^2$$

where \(d\) is side length and \(\phi\) is a half of beamwidth angle. Assuming a figure of 5 dB for antenna pointing losses relative to the bore sight based on the link budget in section 2.1, one can easily evaluate the pointing angle precision for the two cases listed in Table 2.
3.2.3 Phased array/ digital beam forming: circuit level constraints

The term "smart antenna" is often used in this context. In terms of complexity and ambition, this solution is at the opposite extreme compared to the mechanically steered dish. Some recent programmes in this field are reviewed.

The main advantages include: conformality of the array, electronic beam scanning, rapid scanning over wide angles, multiple beam forming, free space combining of transmitted power. The main disadvantage is probably the very large number of antenna and active elements which are required to achieve a high gain aperture. Added to this, a conventional approach requires sub-half wavelength element spacing. This leads to very complex RF circuitry which may be close to being physically non-implementable. Less conventional approaches could include the use of widely spaced element, sub-arrays etc. In any case, a very large number of elements are required, which will be extremely costly. A planar array also exhibits scanning loss which inevitably reduces the aperture gain where it is most needed.

These effects, along with the constraints imposed by the physical circuit and the inter-element spacing, are discussed below along with some examples and derivations from first principles.

3.2.3.1 Array factor

The array factor $F$ is the interference pattern for the electric field. It is a function of element spacing $a$ (in the simplest case this is the same in both cartesian axes i.e. $a = a_x = a_y$). The maximum is given by the number of elements.

The normalized two dimensional array factor is:

$$F(\theta,\phi) = \frac{1}{M} \frac{\sin \left( \frac{M \cdot u_x}{2} \right)}{\sin \left( \frac{u_x}{2} \right)} \cdot \frac{1}{N} \frac{\sin \left( \frac{N \cdot u_y}{2} \right)}{\sin \left( \frac{u_y}{2} \right)}$$

where $u_x = k \cdot a_x \cdot \sin \theta \cdot \cos \phi + \delta_x$

and $u_y = k \cdot a_y \cdot \sin \theta \cdot \sin \phi + \delta_y$

$k$ is the wave number $\frac{2\pi}{\lambda}$ and $a_x$ and $a_y$ are the element spacings along the $x$ and $y$ axes respectively.

Similarly, $\delta_x$ and $\delta_y$ are the relative phase shifts between elements.

Below is the array factor for $4 \times 4$ elements where $a = a_x = a_y$ (in wavelengths).
3.2.3.2 Directivity

On scaling for absolute directivity $D$ we use

$$D(\theta, \phi) = \frac{4\pi P(\theta, \phi)}{P_{\text{rad}}}$$

where $P_{\text{rad}}$ is the total power radiated over all space:

$$P_{\text{rad}} = \int_{\theta = 0}^{\pi} \int_{\phi = 0}^{2\pi} P(\theta, \phi).\sin \theta \, d\theta \, d\phi$$

Directivity is shown in Fig. 7. It is here assumed that the elements radiate into half space only. Evidently the directivity is maximized for a spacing of about $0.75 \lambda$. As the spacing increases $D$ tends to oscillate above and below the value associated with the number of elements ($= M.N$). Thus, for 16 elements, this mean value for $D$ is about 12 dB, plus 3 dB from the half-omni elements, leads to 15 dB.

3.2.3.3 Implications of wide element spacing

Figure 6 Array factor for 4 x 4 rectangular array as a function of element spacing.

Clearly for increasing $a$ (e.g. from $a<0.1$ to $a=1$) the main lobe becomes more narrow. For $a>1$ grating lobes appear. We have not yet considered scaling the pattern for absolute directivity, nor the effect of beam scanning.
Wide spacing ($a > \lambda$) leads to grating lobes. It has been argued that grating lobes do not contribute to interference because the area on ground and hence the number of users within the beam does not change. This argument is only approximately true, because the beam directivity (a measure of total radiated power - i.e. the spatial integral of the radiation pattern) is seen to vary quite a lot with element spacing.

![Figure 8 Loss in Directivity associated with grating lobes, 8 x 8 element array](image)

**3.2.3.4 Effect of antenna element radiation pattern**

Thus far the antenna elements have been assumed to be omni-directional (or half-omni for half space). For practical, directive elements, the radiation pattern is the product of the array factor and the elemental pattern. The directivity of the element can be exploited to suppress grating lobes. The maximum possible element directivity is that associated with the maximum physical size. The radiation pattern for the element may be approximated by the “sine” function for the element aperture:

$$P_{\text{element}} = \left( \frac{\sin(a \sin \theta)}{a \sin \theta} \right)^2$$

(The element radius could be smaller than dimension $a$, but not larger.)

The product of such an element pattern and the array factor is shown in Fig 9, where the suppression of the grating lobes is evident. In practice, the element pattern might have a broader beam (it cannot have a narrower beam) and so the sidelobe suppression may not be as great an effect as that illustrated.
3.2.3.5 Effect of beam scanning

For the 4 x 4 array with $a = 0.5$ wavelength spacing, with directive elements, the effect of scanning is illustrated below. For small scan angles e.g. $< 5^\circ$ the beam shape is not much affected. For increased scan angles the grating lobes of the array factor move into the element pattern and hence become visible.
The effect of grating lobes can be suppressed by using interleaved arrays because the effective sub-array spacing is reduced.

3.2.3.6 Sub-array interleaving

The effect of grating lobes can be suppressed by using interleaved arrays [51].

4-element primary array coefficients: 1, 1.74, 1.74, 1

primary array spacing: 0.68 \( \lambda \)

4-element secondary array coefficients: 1, 1.66, 1.66, 1

secondary array spacing: 1.71 \( \lambda \)

The secondary array spacing is less than the primary array length of 2.73 \( \lambda \) which is only physically possible if the primary arrays are interleaved.
The arrays reported are one-dimensional (linear) arrays. The following array patterns are not scaled for directivity.

![Primary array pattern](image1)

![Secondary array pattern](image2)

![Combined patterns (boresight conditions)](image3)

![Combined array patterns with scanned beam](image4)

**Figure 12 Array patterns**

The following patterns use a secondary array spacing of 2.73 \( \lambda \) i.e. the primary array length. This shows the case for a non-interleaved array, hence the grating lobes are increased compared to the above (interleaved) case.

![Secondary pattern](image5)

**Figure 13 Secondary pattern**
The literature reports on an experimental array prototype. The title's reference to "millimetre-wave" appears somewhat spurious, as the prototype operates at X-band, claiming to be a scale model of a 60 GHz system. The prototype uses printed type antennas. A scanning beam was not implemented, but two fixed-beam versions, for a non-scanned and scanned beam respectively. The scan angle of the latter is about 17°.

While the inter-leaving principle is interesting, this paper doesn't appear to shed much light on implementation issues, particularly for two-axis scanning or operation at frequencies above X-band.

3.2.3.7 Discussion: phased array techniques

Having reviewed a number of techniques which attempt to reduce the complexities of array topologies for smart antennas, it was felt that none offer the required functionality in terms of efficiency, scan performance and cost. Often, when a technique is put forward to mitigate some disadvantage (e.g. use of sub-arrays to reduce number of active components), some other disadvantage is made apparent (e.g. reduced scan angle or appearance of grating lobes).

An array antenna variant was later developed, successfully, by CSEM, but as this was mechanically steered we do not refer to it as a 'smart' (or beamforming) antenna.

3.2.4 Array variants and combined electronic and mechanical steering

The disadvantages of a conventional planar array can be mitigated in various ways if we allow one antenna axis to be mechanically steered, as illustrated in Figure 15. Most likely this would be the azimuth axis. If the array boresight is then tilted to say 37.5°, it is require to scan in elevation over a 75° range to meet the requirement (15° elevation to vertical.) This approach mitigates against scanning loss and, if the array does not scan in azimuth, far fewer active elements (phase shifters) are required. This solution however does not permit electronic (rapid) scan or multiple-beam forming across azimuth planes.
A type of array which could implement the single-axis electrically steered concept of Figure 15 is shown in Figure 16.

3.2.5 Recent projects in phased arrays and Digital Beam Forming (DBF)

In this section we list some of the features of recent and ongoing European projects in phased array antennas for communications. Their objectives are 2-way satellite communications for moving terminals, using 20/30 GHz band. These projects have similar aims to the antennas research work packages of CAPANINA and so it is particularly useful to investigate their approaches, outputs and costs.

3.2.5.1 SUITED project

This IST Framework 5 project ran from 2000 for 30 months. There were 11 partners. The value was 9 M euro, the EU contribution was 4 M euro. TTI of Spain were the lead partners in the array design and fabrication.

For more information see Microwave Journal, January 2004 Vol.47 No. 1.
Technical summary, SUITED antenna terminal:

<table>
<thead>
<tr>
<th>frequencies</th>
<th>20 GHz receive</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>30 GHz transmit</td>
</tr>
<tr>
<td>array design</td>
<td>sub-arrays</td>
</tr>
<tr>
<td></td>
<td>(may be inclined)</td>
</tr>
<tr>
<td></td>
<td>microstrip</td>
</tr>
<tr>
<td></td>
<td>8 element sub-sub array.</td>
</tr>
<tr>
<td></td>
<td>4 x 8 linear sub-array</td>
</tr>
<tr>
<td></td>
<td>12 rows</td>
</tr>
<tr>
<td></td>
<td>total no. radiating elements = 384</td>
</tr>
<tr>
<td></td>
<td>no. phase shifters = 12</td>
</tr>
<tr>
<td>phase shifter loss</td>
<td>20 dB</td>
</tr>
<tr>
<td>scanning</td>
<td>mechanical azimuth scan.</td>
</tr>
<tr>
<td></td>
<td>electronic elevation scan</td>
</tr>
<tr>
<td></td>
<td>elevation scan specification 32° - 48°</td>
</tr>
<tr>
<td>grating lobes</td>
<td>approx - 10 dB</td>
</tr>
<tr>
<td>receive gain, 20 GHz</td>
<td>40 dB claimed</td>
</tr>
<tr>
<td></td>
<td>G/T for aperture estimated at 5 dB/K</td>
</tr>
<tr>
<td>dimensions (mm)</td>
<td>receive</td>
</tr>
<tr>
<td></td>
<td>300 x 200 x 80</td>
</tr>
<tr>
<td></td>
<td>transmit</td>
</tr>
<tr>
<td></td>
<td>240 x 120 x 80</td>
</tr>
</tbody>
</table>

3.2.5.2 SANTANA

This all-German consortium of 4 organisations including DLR and Astrium, funded by the German government, have developed digital beam forming arrays for 20/30 GHz. An application they have emphasized is satellite communications to aircraft during flight, and hence very rapid wide-angle scanning is required.

The project ran for about 3 years, and finished at the end of 2003. The project delivered a 16 element module which was intended as a demonstrator for a follow-on phase SANTANA II. The whole budget allocated until the end of SANTANA II is approx. 4.5 MEuro and comprises around 350 person months (this info. courtesy of DLR).

Technical summary, SANTANA antenna terminal:

<table>
<thead>
<tr>
<th>frequencies</th>
<th>20 GHz receive</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>30 GHz transmit</td>
</tr>
<tr>
<td>array design</td>
<td>4 x 4 element module</td>
</tr>
<tr>
<td></td>
<td>intended as a plug-in module for part of a much bigger array</td>
</tr>
<tr>
<td>scanning</td>
<td>DBF</td>
</tr>
<tr>
<td></td>
<td>(no mechanical scan)</td>
</tr>
<tr>
<td>grating lobes</td>
<td>suppressed by sub-wavelength spacing</td>
</tr>
<tr>
<td>gain</td>
<td>receiver specification:</td>
</tr>
<tr>
<td></td>
<td>for 5 Mbps: 26 dBi</td>
</tr>
<tr>
<td></td>
<td>for 155 Mbps: 41 dBi</td>
</tr>
<tr>
<td></td>
<td>this clearly cannot be met by the 4 x 4 prototype, which would offer approx 16 dBi</td>
</tr>
</tbody>
</table>

Separate transmit and receive arrays were fabricated. The approach was to use microstrip radiating elements fed by dielectric-loaded circular waveguide. MMIC amplification and frequency conversion stages were employed at each element. The MMIC components were mounted on a water-cooled substrate. For both arrays the physical size of the antenna was dominated by the IF circuits and associated electronics.
3.2.6 The spherical lens antenna

Due to its symmetry, a spherical lens can employ multiple feeds to produce multiple beams without any scanning loss. For beam steering, the feed position must be changed. This could be achieved either by mechanically moving a single feed, or by switching between multiple feeds which are spaced around the outside of the lens.

A classic type of spherical lens is the Luneburg lens, which has its focus exactly on the lens surface. The dielectric constant varies radially between 1 at the surface and 2 at the centre. This type of lens is rather difficult to fabricate, and is usually approximated by a series of concentric shells.

A constant-dielectric spherical lens is obviously much easier to fabricate, but suffers from non-exact focussing, reflection loss, and therefore a somewhat limited aperture efficiency. A lens comprising two dielectric shells can be a good compromise.

A very useful technique is to use a combination of a hemispherical lens with a reflective ground plane. This has the same effective aperture as a full spherical lens, but occupies half the height (see Figure 17).

![Diagram of Luneburg lens and hemispherical lens.](image)

Figure 17 Luneburg lens and hemispherical lens.
The technique illustrated has many advantages for a scanning, multi-beam antenna placed on a train:

- The lens does not have to be steered. The moving part (the feed) has much less mass than the lens.
- The physical height of the antenna is approximately half that of a circular dish of equivalent gain.
- Multiple feeds can be employed for multiple beams
- There is no scan loss.
- It should be within our capabilities and budget to build a prototype.

These advantages have lead to the Luneburg lens approach being used for a satellite communications antenna for use on aircraft. The latter point (no scanning loss) is not strictly true for the hemispherical lens. For a zenith-directed beam, the feed blocks the aperture and results in reduced gain. This, however, is less important for the zenith beam because here some loss can be tolerated (Figure 1).

There are of course some apparent disadvantages:

- Mechanical scanning over a spherical surface is not trivial.
- Each beam requires its own independent steering mechanism. This looks quite complicated and difficult to achieve in practice. In some cases a feed could block the field of view of the other feed / beam.
- The routing of a flexible RF cable, waveguide, or use of a rotary RF joint is needed (for each feed).
- Rapid scanning, to cope with pitch/roll/vibration, may be difficult.

Many of these disadvantages are also associated with any mechanically steered solution, but are here ameliorated by the fact that it is only the feed - a relatively small and light weight item - which is steered, as opposed to the antenna in its entirety. There is scope to consider some novel approaches which could improve the functionality and performance of a hemispherical multi-beam lens antenna:

- Use a multiple element feed array for electronic elevation scan. Then, only azimuth scan is mechanical. A disadvantage is increased aperture blockage for high elevation angles
- Use of a feed cluster for a multiple-beam electronic scan over a narrow angular range. This could accommodate rapid beam tracking to cope with vehicle pitch/roll/vibration.

![Figure 18 Hemispherical lens with elevation scanning feed cluster.](image)
Hemispherical lens for customer premises equipment antenna.

It is worth digressing slightly to point out that variants of the hemispherical lens antenna could be an economically viable solution for the customer antenna. In this case a relatively narrow range of steering angles is needed (~20°) and this could either be achieved with mechanical steering or with a multi-beam feed.

A disadvantage of the geometries illustrated above could be the physical height required to house the primary feed and steering jig for high elevation angle operation. A cover (radome) will be required. It would be ideal if this cover could contribute constructively to the focussing effect of a multi-shell lens system, where the focus lies inside the outer shell.

---

**Figure 19 Hemispherical antenna variants with limited scan.**

**Figure 20 Sketch of a possible multi-shell lens configuration incorporating radome.**
3.3 Recommendations following the survey of candidate technologies

A number of concepts for steerable antennas have been discussed. While emphasis has been placed on an antenna for a ground vehicle the arguments are generally applicable also to an antenna for the HAP.

The aim of Work Package 3.2 was to develop a prototype antenna which shows the key features required for a steerable antenna terminal for CAPANINA.

While many techniques and options could be further investigated, two contrasting solution families emerged following this study phase of the project, these being:

- direct radiating phased array with digital beam forming.
- mechanically steered antenna

3.3.1 Phased array and DBF

This offers:

**advantages:**
- Ultimate beam agility and "smart" capability, in theory.
- Free space power combining for transmit array.

**disadvantages:**
- Cost: Certainly beyond the financial scope of CAPANINA.
- Complexity: very difficult to implement a full scale prototype.
- Scan loss: increases at low elevation angles, where loss can least be tolerated.
- Variant with mechanical steering largely negates the advantages of DBF.

Recommendation:

In terms of fabricating and performing measurements, the phased array solution was considered too ambitious compared to the resource available to the project, and is similarly a very expensive solution when we consider that CAPANINA’s aims are cost effective broadband. The experiences of other projects (e.g. SANTANA) corroborate this conclusion, where much more resource has been deployed to develop a 16 element phased array prototype. The intention is not in any way to criticize these other EU projects in smart antennas, rather, it is to question the value of attempting to replicate or advance on their work with a small fraction of their resource.

Nevertheless, the DBF solution has remained an area of research interest, particularly under WP3.3 where beamforming methods have been studied in great detail. In terms of pursuing an experimental programme of research under WP32 however, other more pragmatic solutions were found preferable.

3.3.2 Mechatronic Solution

This offers:

**advantages:**
- Cost: full scale prototype within the financial scope of CAPANINA.
- Complexity: moderate
- Scan loss: negligible.

**disadvantages:**
- Complex mechanical drives and RF feed routing
- Non-proven technology
3.3.3 Recommendations

Two approaches were identified as being particularly suitable.

For the train antenna, a mechanically steered small array antenna was identified as the most promising for the following design and experimental phase. Its inherent light weight and relative ease of fabrication makes it a very cost effective solution and one which is easily integrated with practical mechatronic and control systems.

For the HAP antenna the spherical lens antenna was considered the most promising solution to take forward into a phase of further research, including that of design, fabrication and measurement.
4 Experimental Investigations of Preferred Solutions

4.1 Broadband Printed Array Antenna

The main objective of the antenna design in this section is targeted at a ground mobile-user requirement.

4.1.1 Introduction

Microstrip antennas are used in a wide range of applications thanks to their thin profile, light weight, low cost and ease of integration with other RF devices. In array configuration, they can achieve high gain and become an alternative to parabolic reflectors for millimetre wave applications, such as communications via HAPs.

However, microstrip antennas operate only over a narrow bandwidth due to their resonant nature. Future HAPs subscriber antennas require typically 30 dBi gain and a frequency bandwidth of 13% around 29 GHz. Two techniques commonly used to improve the bandwidth are to add parasitic patches which are gap-coupled with the main resonator [30][31][32][33] or to use strip slot coupled multilayer structure [34]. The former reported an achieved operating bandwidth (Voltage Standing Wave Ratio VSWR = 2) up to 14%, while the latter increased the bandwidth to 19.5% with a stripline-fed multilayer antenna.

When using microstrip configuration at high frequency, some important issues need to be addressed with care. The first issue is microstrip loss which becomes important at high frequency due to dielectric materials. Special low-loss substrate should be considered. In addition, attention should also be paid when designing a microstrip feeding network of arrays having a large number of elements. Strip lines and/or multilayer structures can help to minimize the length of microstrip feeding lines thus to reduce loss. The second issue is the required high gain with respect to the small size of the antenna. Techniques already known for the gain enhancement are such as multiple superstrates [22], high permittivity superstrate [23] [24] as well as coupling and shielding [25]. The gain G and directivity D of an antenna is related by radiation efficiency η as $G = \eta \cdot D$, where η includes the conduction and dielectric efficiency which depend on the geometry of the antenna and on the substrate material. It is observed that directivity increases with increase in substrate thickness while efficiency has an inverse trend. So care needs to be taken to ensure the antenna efficiency is maintained when the gain is used as a global parameter to be optimised on the antenna performance.

The third issue comes with the question how to provide large bandwidth with a printed antenna. To address this problem, solutions found from literatures mentioned stacked patch [26] and slot excitation. The purity of the circular polarisation (CP) within very large bandwidth is another issue. Till now, sequential rotation feeding network [27] or polarisation transformer [28] has been employed to get good AR of CP within a relatively large bandwidth. Last but not least, from the point-of-view of fabrication, substrate planarity should be guaranteed because phase error very sensitive at high frequency applications.

In this work, a strip slot coupled solution has been proposed because the feeding network is located on a different layer than the patches enabling the routing to be optimized separately. In addition, we considered an array composed of a small number of elements to further reduce feeding network complexity while maintaining the performance required by HAPs application.

4.1.2 Techniques for second lobe reduction

In the design procedure of any array, array factor (AF) is an important element which is, in general, related to the number of elements, the array geometry, the excitation phase and magnitude, as well as the element spacing. For a N-element linear array of uniform spacing and uniform power distribution, the AF can be written as
Report on steerable antenna architectures and critical RF circuits performance  CAP-D24-WP32-UOY-PUB-01

\[ AF \equiv \left[ \frac{\sin \left( \frac{N}{2} kdcos\theta + \beta \right)}{kdcos\theta + \beta} \right] \]

Where

\[ N = \text{number of elements in the array} \]
\[ d = \text{element spacing in the array} \]
\[ \theta = \text{angle between observing direction and array direction} \]
\[ \beta = \text{phase excitation of the elements in the array} \]
\[ k = \text{wave number defined as equal to} \frac{2\pi}{\lambda} \]

In a uniform amplitude and phase array, the element spacing should be less than \( \lambda \) to prevent unwanted grating lobes in broadside radiation region [39]. With the high-order mode SSFIP of size \( \lambda \), it is impossible to satisfy this spacing condition. Here we will discuss three possibilities of reducing the sidelobes level.

### 4.1.3 Non-uniformly spaced array

In general, the sidelobes of array antennas with uniform amplitude excitation can be reduced by using non-uniform spacing between elements. In reference [42], an iteration method based on the solution of a set of linear simultaneous equations at each iteration is proposed to obtain the non-uniform array element positions. The position of the \( n^{th} \) element from the origin is written as:

\[ d_n = \left( \frac{n}{2} + \varepsilon_n \right) d \]

where \( d \) is the spacing between elements of the referential uniform array, and \( \varepsilon_n \) the shifted distance with respect to its uniform array location.

In Figure 21, the red curve shows resulted side lobe level (SLL) of a 24-element linear array by using one set of element positions proposed in [42]. The SLL of the same array with uniform spacing is also plotted (in blue) for comparison. This technique results in 11 dB SLL reduction when assuming the element spacing of the referential uniform array \( d = 0.5\lambda \).

![Figure 21 Non-uniform spacing applied to a regular array in which \( d = 0.5\lambda \).](image-url)
In our case, the spacing of array elements is much bigger due to the large size of the highly directive array element. Consequently, grating lobes related to the array geometry are unavoidable, as can be observed on the blue curve in Figure 22. The application of the same set of nonuniform positions reduces the grating lobe level (GLL) but arises at the same time the SLL associated to the grating lobes. This result is shown by the red curve in Figure 22. For more satisfactory SLL/GLL reduction, new codes need to be implemented by taking into account the large element spacing.

**Figure 22** Non-uniform spacing applied to a special array in which \( d = 2\lambda \).

### 4.1.4 Non-uniform power distribution

Non uniform power distribution can lower SLL (Dolph-Tschebyscheff power distribution) or even totally kill the second lobes (Binominal power distribution). Under the condition that element spacing is equal or less than \( 0.5\lambda \), Dolph-Tschebyscheff power distribution can be designed using Tschebyscheff polynomials to provide any desired SLL. When required \( SLL = -\infty dB \), the Tschebyscheff array becomes a Binominal array which has no side lobes.

Applying the condition of element spacing \( d = 2\lambda \), Figure 23 shows comparison of uniform, Dolph-Tschebyscheff and binomial power distribution: with \( SLL = -25 \) dB for Dolph-Tschebyscheff. It is apparent that the grating lobes are not modified by power distribution. In addition, the main beam width and the side lobe level are compromised.
4.1.5 **Specific single element radiation**

For any arrays of identical elements, its far field is expressed as the product of the field of a single element and the array factor (AF). The directional characteristics of the radiating elements themselves give an additional degree of freedom for working towards the desired radiation pattern. In Figure 24 is an example showing how a special radiation pattern of the array element, which has nulls at the place where the grating lobes occur in the AF function, can contribute to lower the side lobe level. The special radiation pattern can be worked out mainly through changing patch geometry.

**Figure 24** Combination of AF and special element pattern
4.1.6 Single high gain broadband antenna design

Features of SSFIP antennas operating at fundamental resonance have been well documented in [34], [35] and [36]. A full-wavelength SSFIP providing two off-axis main lobes at 30° angles with respect to broadside has also been reported [37]. Here we present a novel design of a higher order mode SSFIP with etched patch which exhibits high directivity in broadside direction.

As presented in Figure 25, the patch (top-layer) of the high-order mode SSFIP is coupled with the microstrip feed line (bottom-layer) through the slot (mid-layer). The patch is of size $\lambda$ at operating frequency. The patch has been etched in a fractal-like shape to provide a specific radiation pattern as well as broadband impedance matching. The antenna in Figure 25 has been simulated using commercial electromagnetic Method of Moment (MoM) software [38]. The substrate is Rogers RT/Duroid 5880 with thickness $h = 0.254$ mm and dielectric constant $\varepsilon_r = 2.2$. The overall size of the antenna is $13.8 \text{ mm} \times 16.5 \text{ mm} \times 1.4 \text{ mm}$. The radiation pattern was studied to enhance grating lobe suppression when integrated as an element into an array.

![Figure 25 Different layers of a SSFIP antenna](image)

Figure 26 depicts impedance matching bandwidth obtained by simulation. As it can be seen in Figure 26, the operating frequency for a VSWR = 2 ranges from 27.0 to 33.1 GHz, meaning 20.3% of impedance bandwidth. An example of radiation pattern in the E-plane ($\phi = 90^\circ$) and H-plane ($\phi = 0^\circ$) is presented in Figure 27.

Table 3 summarizes the gain and half-power beamwidth at four simulated frequencies. The gain flatness remains less than 1 dB in the whole frequency range. The simulated cross-polarization is about 75 dB lower than the co-polarization meaning that the high order mode is properly excited.

![Figure 26 Single element antenna: impedance bandwidth (VSWR) as a function of frequency.](image)
<table>
<thead>
<tr>
<th>Operation frequency [GHz]</th>
<th>Gain [dBi]</th>
<th>E-plane HPBW [°]</th>
<th>H-plane HPBW [°]</th>
</tr>
</thead>
<tbody>
<tr>
<td>27.5</td>
<td>10.1</td>
<td>32.8</td>
<td>30.4</td>
</tr>
<tr>
<td>28.5</td>
<td>10.8</td>
<td>30.8</td>
<td>29.6</td>
</tr>
<tr>
<td>29.5</td>
<td>10.4</td>
<td>29.2</td>
<td>30.0</td>
</tr>
<tr>
<td>30.5</td>
<td>9.9</td>
<td>27.8</td>
<td>30.8</td>
</tr>
</tbody>
</table>

Table 3 Main simulated characteristics of the single antenna

Figure 27  Co-polarization radiation pattern in E-plane ($\phi = 90^\circ$) and H-plane ($\phi = 0^\circ$) at 28.5 GHz

The prototype has been fabricated by two PCB manufacturers. It is important to mention that the performance of the antenna is sensitive to the thickness of the spacer placed in between the patch and the ground plane containing the slot (see Figure 25). To be able to compare the simulated and the measured results, the spacer thickness in the simulation was adjusted to be the same as that of the prototype. The result of the comparison is shown in Figure 28. The difference between the measured and simulated results is due to the uncertainty of dielectric loss at high frequency.

Figure 28  Measured and simulated impedance matching for adjusted model
4.1.7 Array performance

In a uniform amplitude and phase array, the element spacing should be less than $\lambda$ to prevent unwanted grating lobes in broadside radiation region [39]. With the high-order mode SSFIP of size $\lambda$, it is impossible to satisfy this spacing condition. To minimize the grating lobes, the inter-element spacing of $1.8\lambda$ is adjusted to place the H and E-plane grating lobes at the radiation direction where the single element antenna has a null in the radiation pattern.

Due to hardware limitations, only 4- and 16-element arrays were simulated using the commercial EM software [38]. In Figure 29, impedance matching bandwidth versus frequency range of the 16-element array is presented. The broadband behaviour of single element remains stable in the array configuration. Simulated cross-polarization is about 23 dB lower than the co-polarization. The improved radiation gain of the single element and the $2\times2$ subarray compared with a normal square patch are shown in Figure 30 and Figure 31. It can be observed that with our special shaped single element radiation pattern, 6 dB of energy in the grating lobes are moved into the main lobe. Other simulated performance of the single element and the sub-arrays are summarized as well in Table 4.

![Figure 29 16-Element array: impedance bandwidth ($S_{11}$ magnitude) as a function of frequency](image1)

![Figure 30 Simulated gain of the single element compared to a normal patch](image2)
Figure 31  Simulated gain of the $2 \times 2$ subarray compared to a normal square patch

<table>
<thead>
<tr>
<th>Simulation results</th>
<th>Single element</th>
<th>4-element array</th>
<th>16-element array</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency range</td>
<td>27.1 – 33 GHz</td>
<td>26.3 – 32.9 GHz</td>
<td>26.7 – 32.4 GHz</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>19.6%</td>
<td>21.8%</td>
<td>19.3%</td>
</tr>
<tr>
<td>HPBW</td>
<td>31°</td>
<td>15.4°</td>
<td>7.6°</td>
</tr>
<tr>
<td>Gain</td>
<td>10.7 dBi</td>
<td>17.5 dBi</td>
<td>23.2 dBi</td>
</tr>
<tr>
<td></td>
<td>(5 dB better than a rectangular square patch)</td>
<td>(&gt;5.5 dB better than omnidir.-element array of same aperture)</td>
<td>(&gt;5.1 dB better than omnidir.-element array of same aperture)</td>
</tr>
<tr>
<td>Gain flatness</td>
<td>~ 2 dB</td>
<td>&lt; 1 dB</td>
<td>&lt; 1 dB</td>
</tr>
<tr>
<td>MLL/SLL</td>
<td>-</td>
<td>~ 11 dB</td>
<td></td>
</tr>
</tbody>
</table>

Table 4  Simulation results of single element and sub-arrays
Figure 32 Predicted radiation of 64- and 256-element array based on single element pattern

4.1.8 Measurement results

A 64-element prototype (Figure 33) has been fabricated for verification. The measured impedance matching presented in Figure 34 exhibits relatively good broadband performance. The theoretical radiation pattern of a general case, e.g. array of $x$-element spaced by distance $d$, has been computed by applying the planar array theory on a simulated single element. The theoretical results of a 64-element array are compared with the measured curves in Figure 35, Figure 36, Figure 37 and Figure 38. Measured radiation patterns show excellent agreement with theoretical prediction in both E- and H-plane. Electrical features of the array remain stable through the whole frequency range. In Figure 35 it is shown that the E-plane grating lobes are minimized to 15.7 dB lower compared to the main lobe at 27.5 GHz. The grating lobes level for other frequencies can be found both in Figure 37 and in Table 5. The E- and H-plane cross-polarizations are at about -23 dB compared to the co-polarization within the whole operation frequency range. Measured cross-polarizations are presented in Figure 36 and Figure 38. The second lobes of value -11.7 dB at 30.5 GHz are due to the second lobes of the single element at that frequency. The front-to-back ratio is 13 dB for the worst case. This ratio can be further enhanced to 30 dB by absorbing material placed on the backside of the array (Figure 39).

The gain estimation of the 64-element array has been made by comparing with a commercial broadband Horn antenna of +/-1 dB uncertainty [40]. The values and comparison with the simulations are summarized in Table 5. The measured gains are 3.5 to 5.8 dB lower than computed values depending on operating frequencies. A way that could improve the theoretical prediction is to include the loss of the feeding network into MATLAB simulation. At high frequency, any uncertainties concerning the dielectric material characteristics or metallic surface conductivity impact the gain predicted by the simulations.
Figure 33  64-Element prototype

Figure 34  Measured VSWR of 64-element with respect to frequency
Figure 35  Comparison of normalized co-polarization radiation patterns of the 64-element array at 27.5 GHz

Figure 36  Measured normalized co- and cross-polarization at 27.5 GHz
Figure 37 Comparison of normalized E-plane co-polarization radiations of the 64-element array

Figure 38 Measured normalized E-plane co- and cross-polarization at 28.5 and 29.5 GHz
--- | --- | --- | --- | --- | ---  
27.5  |  28.2  |  24.7  |  3.4  |  3.6  |  -15.7  
28.5  |  28.8  |  24.9  |  3.6  |  3.7  |  -14.2  
29.5  |  28.4  |  22.6  |  3.6  |  3.6  |  -12.3  
30.5  |  28.1  |  22.4  |  3.0  |  3.4  |  -11.7  

Table 5  Performance of 64-element array

Figure 39  Enhanced E-plane backward radiation of the 64-element array at 27.5 GHz
4.1.9 Polarizer

The present polarizer has the function of converting a wave from linear polarization to circular polarization. It is also called polarization transformer. The basic approach of a polarization transformer is to make an array of structures which appear to be predominantly inductive to one polarization and predominantly capacitive to the orthogonal polarization [41]. The polarizer is built with four-layer meander line arrays which produce two field components with differential phase shift of 90° form the incident wave and generate circular polarization. Using a polarizer to get the circular polarization from the linear polarized array has several advantages:

- High polarization purity (low axial ratio) over a large frequency bandwidth
- Low dissipation loss
- Good VSWR for impedance matching
- Compact size at Ka band
- Capability of handling high power (15.5 W/cm² at 9 GHz reported in [41])
- Low complexity
- Low cost

A polarizer, to be placed in front of the array antenna, has been formed from multiplayer foam with a thin flexible dielectric substrate for the etched conductors and has been rescaled for 30 GHz application. A side view of the polarizer section and the layouts are shown in Figure 40 and Figure 41.

![Figure 40 Four-layer meander line polarizer](image-url)
4.1.10 Loss consideration

4.1.10.1 Loss in the antenna

Loss in a printed antenna is distributed in the form of dissipation in the conductors and dielectric, and surface wave. Antenna radiation efficiency $\eta$ is primarily related to the substrate thickness and permittivity. Though, cautions need to be made through the design of array at high frequency. For example, low-permittivity low-loss substrate was employed to widen the impedance bandwidth and reduce the surface wave excitation. Metal thickness of 25 $\mu$m has been chosen for compromising between loss and spurious radiation. For the design considered, loss in the microstrip feed line was estimated to be from 10 to 10.5 dB/m at 28.5 GHz.

The radiation efficiency $\eta$, can be estimated using the maximum directivity $D_0$ and the gain $G_0$ of the antenna as: $G_0 = \eta \cdot D_0$. Theoretically, the maximum directivity of any antenna can be computed from its maximum effective aperture [39]:

$$A_{\text{eff}} = \frac{\lambda^2}{4\pi} D_0$$

Based upon the aperture size of the antennas and the simulated gain, the radiation efficiency at 28.5 GHz of the single and array antennas was estimated and summarized in Table 6.

<table>
<thead>
<tr>
<th></th>
<th>Single element</th>
<th>2×2 array</th>
<th>4×4 array</th>
<th>8×8 array</th>
<th>16×16 array</th>
</tr>
</thead>
<tbody>
<tr>
<td>Efficiency</td>
<td>88.6%</td>
<td>59.2%</td>
<td>36.6%</td>
<td>23.2%</td>
<td>10.2%</td>
</tr>
</tbody>
</table>

Table 6 Estimated efficiency at 28.5 GHz of the single element and array antenna

The single element efficiency is lower than a microstrip line fed patch. When using a strip-slot coupled feeding, as the substrate thickness increases, the surface wave power increases, and consequently the efficiency decreases. It can be observed that the radiation efficiency decreases rapidly as array element number increases due to increasing loss in the microstrip feed lines.
4.1.10.2 Loss in the polarizer

The loss in the polarizer comes from three sources: loss in the meander lines, loss in the spacers and loss in the substrate. For the particular polarizer design here, the loss in the passband is approximately [43]:

\[
L = \frac{12.5}{\omega} \left( \frac{1}{Q_{\text{meander}}} + \frac{1}{Q_{\text{spacer}}} + \frac{1}{Q_{\text{substrate}}} \right) \text{ dB}
\]

where \( Q \) is the unloaded quality factor related to different sources and \( \omega \) is the bandwidth.

The loss in copper meander lines is a function of the skin depth, the free-space wavelength, the metal resistivity and the meander lines dimensions. The loss at 10 GHz has been calculated in [41] as to be approximately 0.02 dB. It can be easily scaled to the frequency of 30 GHz by considering \( Q \propto \frac{1}{(f)^{1/2}} \) and we get the loss due to the meander lines at 30 GHz is approximately 0.035 dB.

The loss in the spacer and the substrate related to the dielectric properties is estimated as [43]

\[
\text{total loss in dB} = \frac{27.3 \varepsilon_r \tan(\delta)}{\lambda_0} \sum h_i
\]

where \( \lambda_0 \) is the wavelength in air, \( \varepsilon_r \) and \( \tan(\delta) \) are respectively the relative permittivity and the dissipation factor of the material and \( h_i \) is the thickness of the \( i^{th} \) substrate. The loss \( \tan(\delta) \) of the spacer becomes more important at 30 GHz, in the order of 0.01, while the \( \varepsilon_r \) remains stable at 1.04. Thus the loss dissipated in the three quarter wavelength spacers is about 0.2 dB.

The relative permittivity of flexible substrate is about 3.1 and its dissipation factor \( (\tan(\delta)) \) is as low as 0.0009 (tested at 10 GHz). With the thickness of 50µm, the loss in the substrate is negligible.

\( \text{(The test result of loss tangent is available at up to 10 GHz from the substrate manufacturer. However in the cost estimation, a margin has been taken in considering higher frequency application)} \)

In summary, the total energy dissipated in the polarizer at 30 GHz is estimated to be 0.24 dB.

4.1.11 Conclusion for printed array antenna

A broadband high-order mode SSFIP antenna with high directivity in the broadside direction has been developed. The highly directive SSFIP antenna enables implementation of an array with a reduced number of elements; thus decreasing feeding network complexity. Subarrays of 4, 16 and 64 elements were studied by simulation. A 64-element prototype has been tested and the measured results have been compared with the theoretical computations and shown to be in good agreement.
4.2 Hemispherical lens prototype.

4.2.1 Constant index lens

A basic hemispherical lens prototype was been constructed and investigated experimentally at the University of York.

This first prototype consisted of a single layer of polyethylene, of dielectric constant 2.3

The antenna is illustrated in Figure 42. The feedhorn illustrated is not optimized for this application. The lens diameter is only 160 mm which is somewhat less than anticipated for the final requirement.

The purpose of this first prototype was to investigate scan performance.

![Figure 42 Measurement jig for proof-of-concept (this one is set up for about 65° elevation angle).](image)

The measurement facility was an anechoic chamber where the antenna under test was placed on a motorized turntable and acted as the receiver. The carrier frequency used was 28 GHz. A harmonic mixer was used to downconvert the carrier to 1.7 GHz so that a coaxial cable could be used to connect to a spectrum analyzer, external to the chamber, without incurring excessive loss. WR75 waveguide was routed into the chamber to provide the local oscillator signal at around 13 GHz. Data capture was automated under control of a personal computer running LabView which interfaced with the turntable controller via a serial port, and with the spectrum analyzer via a GPIB interface.

It should be noted that, due to the limited length of the chamber (2.7 m between transmit and receive antennas) the measurement was not that of the far field. Nevertheless the modal expansion analysis can readily accommodate near-field effects and thus the presented results compare measurement with theory at the 2.7 m distance (not far field). This has the effect of sidelobe growth compared to the far field patterns.
Figure 43 Radiation patterns for 4 elevation angles.

Figure 43 shows measured radiation patterns for the reference antenna (a lens-corrected horn type) and 4 cases of elevation angle for the single-shell hemispherical lens. The main features of this hemispherical lens prototype are:

- inferior gain compared to the reference antenna (between 1.1 and 2.9 dB)
- good wide angle scan properties
- a significant pattern degradation for 90° elevation (when the aperture is blocked)

The feature of aperture blockage is due to the hemisphere lens’s use of a ground plane reflector. Thus, for a 90° (or similar) elevation angle, the primary feed blocks the aperture. This leads to gain reduction and increased sidelobe levels.

As has been discussed, the zenith gain reduction is not particularly problematic for the HAP geometry, since here the link length is at a minimum. Of greater importance is that the scan loss at low elevation angles should be minimized. Here, the advantages of the hemispherical lens are apparent, as highlighted in Figure 44. The normalized measured scan loss of the hemispherical lens is shown (4 points) compared with the free space loss for a HAP height of 20 km. Also shown is an approximation of scan loss for a planar antenna (e.g. a planar array) based on the effective aperture area.
Figure 44 shows how the maximum scan loss of the hemispherical lens coincides with the minimum free space loss and improves at lower elevation angles. In contrast, the trend for a planar array is that the scan loss increases at lower elevation angles.

(For a spherical lens there would be no reflector plane and no aperture blockage and a plot of the type in Figure 44 would, theoretically, show all points at 0 dB scan loss.)

4.2.2 Primary feeds

A better primary feed was then developed, being a scalar circular waveguide feed of the design reported by [15] whereby three concentric grooves of quarter-wave depth were machined into the waveguide flange. This yields a broader primary feed beamwidth than that of the pyramid horn, thus better illuminating the lens, and also offers better cross-polar performance. Using this feed, the measured gain was approximately 30 dBi, representing an aperture efficiency of approximately 40%. The measured radiation patterns are compared in

Figure 45. The scalar feed can also be used in combination with a dielectric waveguide polarizer so as to yield circular polarisation.
Having established that the desired wide scan properties are achievable, and investigating experimentally the effect of the primary feed on inherent efficiency of single-layer hemisphere lenses, the advantages of multi-shell lenses are next discussed.

4.2.3 Development of a multi-layer lens

The rationale of this section of the research programme was to investigate the performance gains associated with a multi-layer lens compared to a single layer lens, so as to quantify trade-offs between lens antenna cost and performance. A key objective was to find a solution to a major problem of the Luneburg lens antenna i.e. the complexity of the fabrication problem and the associated performance degradation which can easily be introduced.

4.2.3.1 Analysis of single-shell and two-shell lens designs

A mathematical analysis of the radiation properties of concentric dielectric spherical shells is given by [16] where a modal expansion method is used to derive the field scattered when the source is a short dipole. This technique has been used to investigate the radiation patterns of various combinations of shell radius and dielectric constant. Taking these as input parameters, the scattering matrices and resulting far field patterns, which are the sum of the source and scattered fields, are computed using approximately 250 lines of Mathematica code. The maximum directivity is then computed from the far field pattern. The number of terms needed in the modal expansions is proportional to the maximum dimension of the antenna system. For an 8 wavelength outer radius 60 terms are adequate. While [18] reported on an optimisation of dielectric constant, of particular interest to the present study is to choose from a small list of possible materials which have low loss, known $\varepsilon_r$, are machinable and obtainable.

The focal distance is important as this should be minimized (within limits dictated by the phase centres of practical primary feeds) so as to minimize the physical height of the antenna. The geometry is shown in Figure 46, where $f$ is the distance of the primary feed from the lens centre. A second feed in phase quadrature with the first can be arranged to synthesize a directive primary feed pattern. The materials considered are listed in [20].
As a reference point against which to compare a two-shell design, the theoretical properties of single-shell spherical lenses are firstly analysed.

Figure 47 shows the sensitivity of an 8 wavelength radius spherical lens to the location of the feed (the theoretical directivity of a uniform 8 \( \lambda \) aperture is 34.7 dBi). The trend that lower \( \varepsilon_r \) materials exhibit a greater focal distance is expected, as reported by [19], although, to digress, the “paraxial focus” of [19] which is given by \( F = \frac{\sqrt{\varepsilon_r}}{2\lambda} \) leads to a somewhat larger value of the focal length than that which maximizes directivity as derived by the modal expansion method.

The need to move the feed away from the lens edge leads to an increase in the required height for the antenna installation. A minimum feed-lens separation of 0.2 \( \lambda \) has been used so as to accommodate a practical primary feed, where the phase centre lies, typically, up to this distance inside the feed aperture [15]. The maximum directivity \( D \) for the three materials is shown in Table 8 along with the feed position at which this was achieved. The last column shows the case for a fixed 8.2 \( \lambda \) feed location and where the lens diameter is reduced accordingly so that the relative focal position is maintained. This allows a comparison of performance where the total height is fixed.
Figure 47. Spherical lens computed directivity $D$ versus feed location for three materials.

Table 8. Directivity and focus points of spherical lenses.

<table>
<thead>
<tr>
<th></th>
<th>$D$ (dBi) for 8 $\lambda$ lens radius</th>
<th>feed location ($\lambda$)</th>
<th>$D$ for fixed 8.2 $\lambda$ feed location</th>
</tr>
</thead>
<tbody>
<tr>
<td>polyethylene</td>
<td>31.9</td>
<td>9.2</td>
<td>31.3</td>
</tr>
<tr>
<td>Rexolite</td>
<td>32.7</td>
<td>9.0</td>
<td>32.0</td>
</tr>
<tr>
<td>fused silica</td>
<td>29.1</td>
<td>8.2</td>
<td>29.1</td>
</tr>
</tbody>
</table>

A two-shell lens with feed fixed at 8.2 $\lambda$ was next investigated, where the outer radius $r_2$ was fixed at 8 $\lambda$ and $r_1$ allowed to vary. Materials were chosen from Table 8 and always with $\varepsilon_{r1} > \varepsilon_{r2}$. (Dielectric loss is readily handled in the modal analysis but was found to be small for the materials chosen.) To summarize the effects observed:

- there is no advantage adding an outer layer to a fused silica core.
- there is minimal advantage in using the low-constant foam as an outer layer. The layer would need to be thin (~ 1 $\lambda$) and hence difficult to fabricate.
- The best results were obtained using a Rexolite inner core of radius 4.2 $\lambda$ and a polyethylene outer core. This yielded 33.46 dBi which represents 76% aperture efficiency (albeit with an idealized feed model).

To sum up, given a fixed antenna height, a 2-shell lens of 8$\lambda$ outer diameter offers a 1.46 dB gain improvement over a single shell lens. Whether this gain is justified would depend on manufacturing costs and the constraints on terminal size.

Lenses of greater diameter were next investigated: for a feed at 11.5 $\lambda$, the directivity of a single Rexolite sphere is maximized at 34.4 dBi for a radius of 10.35 $\lambda$ while a 2-shell Rexolite/polyethylene lens of outer radius 11.1 $\lambda$ offers 36.1 dBi, being a 1.7 dB improvement.

These dimensions would represent an outer lens diameter of 236 mm at 28 GHz - about the largest size which was considered suitable for machining at the University of York Physics and Electronics Mechanical Workshops. This antenna prototype was then fabricated.
4.2.3.2 Two-layer lens performance

The above discussed two-layer hemisphere lens was fabricated. The core radius was 56.7 mm while the outer diameter was 235.5 mm. The inner, concave surface of the outer polyethylene layer was machined first, but this later experienced some mechanical stress relief on machining the outer profile and which initially resulted in the inner Rexolite core protruding very slightly. When placed on the 350 mm square ground plane, a small air gap (about 0.3 mm) was initially present. Measured gain was 34.2 dBi. (This includes, being mindful of the near field distance, a correction factor of +0.57 dBi which was computed by evaluating (1) as a function of \( r \).) The components are shown in Figure 48.

![Components of two-layer lens (left: polyethylene shell, right: Rexolite core.)](image)

To eliminate the air gap at the ground plane, the flat (lower) surfaces of both layers were lapped using a fine abrasive. A small residual air gap remained at the lower edge of the interface between the two dielectric layers, due to the stress distortion of the polyethylene. Nevertheless, following this modification the measured gain was 35.1 dBi ± 0.4 dB which represents an aperture efficiency in the region 68 %. This efficiency is comparable with a conventional reflector antenna and suggests that dielectric loss is not too significant a factor. The measured radiation pattern is compared to the theory, taking into account the 2.7 m measurement distance, in Figure 49 (a), while the theoretical far field patterns are shown in Figure 49 (b).
4.2.4 Mechanical feed steering concepts

A number of layouts were considered. One tentative concept is illustrated in Figure 50 which shows two independently steerable feeds. This arrangement would allow the antenna to communicate simultaneously with two different vehicles or user groups, which would facilitate handover between cells and also allow for diversity techniques. The sizes of the moving components should be minimized so as to minimize both the aperture blockage and the height of the installation.
A design of a steering mechanism of the type illustrated in Figure 50 was carried out by CSEM. However, construction of a prototype at UOY could not be carried out within the project’s resource constraints. The seeking of external funds to develop a prototype has been tackled as a project exploitation task.

### 4.2.5 Multi-beam Lens Antenna Conclusions

Hemisphere lens antennas have been identified as a promising solution for a multi-beam, wide-scanning antenna for communications to moving vehicles via satellite or High Altitude Platform and using Ka band. The use of a hemisphere with a ground plane yields an effective aperture size the equivalent of a full sphere. The radiation properties of such structures (which may use multiple concentric dielectric layers) has been described as a superposition of two regions of that of a spherical lens, namely those arising from the real and virtual sources in the plane of observation. The spherical lens patterns may be computed using an established modal expansion technique, which accounts for the boundaries at each concentric layer. The hemisphere radiation theory agrees well with measurements of a single layer polyethylene lens antenna at 28 GHz. This antenna also exhibits the desired wide scanning properties, but a mediocre aperture efficiency of 40%.

To improve upon this efficiency, the properties of two-layer lenses have been investigated, with an emphasis on a constrained antenna total height and therefore limited by a fixed primary feed location. Also, a few readily available dielectric materials have been chosen, rather than attempting to optimize a Luneburg-like radial variation of dielectric constant. Within this framework, two-layer lenses based on a Rexolite ($\varepsilon_r = 2.53$) inner core and polyethylene ($\varepsilon_r = 2.28$) outer layer have been put forward as a good solution. Measurements of a two-layer prototype of 236 mm outer diameter indicate a gain of 35.1 dBi at 28 GHz, which represents an aperture efficiency of 68%. This efficiency is comparable with a conventional reflector antenna while the 2-layer lens structure offers a reduced height profile and the potential to use multiple scanned beams over a very wide angle.

### 4.2.5.1 Acknowledgment

Thanks to Brent Wilkinson and Rob Easton at the Electronics and Physics mechanical workshop at The University of York for their meticulous construction of the antenna components.
5 Mechanical Constraints Related To Train Application

Establishing and maintaining a continuous RF link from a moving train to a stationary balloon, or satellite for that sake, is a demanding task. The main reasons come from the loss of line of sight (LoS) due to the alternating landscape and swinging train tracks as is found in much of Europe. With the high speed tilting train, this feature adds another difficulty for the antenna steering algorithms.

However, why shouldn’t such a link still be feasible and reliable?

Adding a steering antenna on a train is feasible and has been tested by different companies as the example shown hereby mounted on a French TGV train (Figure 51).

One can establish a preliminary list of difficulties to overcome such as:

* Lateral and vertical vibration due to train bogey oscillations and rail swing
* Rotations due to the rail curves
* Tilt motion due to new generation tilt trains and their COG compensation.
* Local obstructions from tunnels, mountains, narrow valleys, high buildings etc

From the literature one can get some information to help settle the requirements for the LOS control algorithms of a mobile (azimuth / elevation type) antenna.

Typical measurements (from a Swedish X2000 high-speed train) show oscillations up to some 10 Hz, which also interact with the human body attitude control, something every train commuter has experienced (Figure 52).

Track curve rotations for a high-speed train are up to 4°/s for curve radii of 1 to 1.25 km at 185 km/h or 440 to 800 metres at 125 to 145 km/h velocities. 

![Figure 51 Steerable antenna mounted on a TGV roof (http://telecom.esa.int/telecom/www)](http://telecom.esa.int/telecom/www)
Furthermore the high-speed tilting trains, allowing a significant increase in travel speed on the old sinuous rail tracks of the last century, add additional difficulties by their extra tilt angles and higher tilt velocities, as illustrated in Figure 53.

However, on the plus side, the number of unknown degrees of freedom (DoF) for a direct LoS link are limited because, to communicate in relay to a sequence of high-flying balloons, trains on ground only move in a two-dimensional plane (approximately with respect to the HAPs. Furthermore, the train paths are well known and can be “learned” with fuzzy logic type algorithms; this is also valid for most obstacles. Similarly, there will be a gradual change in travelling speed of the trains because of their limited acceleration and deceleration capabilities. This allows the dynamics of beam steering control and maintenance to be separated into two categories:
* Low speed (seconds or tens of seconds timeframe) for track and relay position tracking algorithms, including relay hand-over due to known obstacles
* High speed (fraction of a second to some 10-20 Hz timeframe) tilt / rotation motions due to train oscillations, track swings etc

If the low speed components are highly predictable, so are also, at least partly, the high-speed components.

Most regular train commuters have noticed that the train does similar swings and lurches at the same location day after day. This is mainly due to imperfections of the rail tracks, and this behaviour could be “learned” by the appropriate algorithms and stored locally. As most trains travel over the same path week after week this accumulated knowledge can be efficiently reused.

To overcome the loss of communication due to tunnels and major obstacles, one obvious solution is to use two or more terminals on a given carriage composition. Today most train compositions are made of predetermined groups formed typically by 3 to 9 wagons of typically of 25-metres length each. Hence the two receiver antennas could be situated some 75 to 225 metres apart allowing for a minimum obstruction length without full link loss. For longer tunnels, as the tunnel walls are close to the train structure, other types of antenna and WiFi transmission could be considered, possibly with one or two fixed stations at each tunnel entrance. Similar systems are readily used today for the mobile phone network in the tunnels in several countries.

<table>
<thead>
<tr>
<th>Train velocity [km/h]</th>
<th>Time in seconds to travel a train length in metres</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>75</td>
</tr>
<tr>
<td>50</td>
<td>14</td>
</tr>
<tr>
<td>100</td>
<td>28</td>
</tr>
<tr>
<td>150</td>
<td>42</td>
</tr>
<tr>
<td>200</td>
<td>56</td>
</tr>
<tr>
<td>250</td>
<td>69</td>
</tr>
<tr>
<td>300</td>
<td>83</td>
</tr>
<tr>
<td>350</td>
<td>97</td>
</tr>
<tr>
<td>400</td>
<td>111</td>
</tr>
</tbody>
</table>

Table 9 Time in seconds to travel the distance of a train length for estimation of tunnel obstruction times
6 State-of-the-Art of Electrical and Mechanical Steering Technologies

6.1 State-of-the-Art of Mechanical Steering System

A review of potential tip-tilt steering systems was made of products available on the market that fulfil the requirements for the steerable antenna mechanism. The principle technical requirements used to evaluate the mechanism are given in the table below (Table 10). The payload considered had a total mobile mass of 1kg, including interfaces.

<table>
<thead>
<tr>
<th>Item</th>
<th>Requirements</th>
</tr>
</thead>
<tbody>
<tr>
<td>Scan angles</td>
<td></td>
</tr>
<tr>
<td>Azimuth</td>
<td>±180°</td>
</tr>
<tr>
<td>Elevation</td>
<td>0-90° (15°-90°)</td>
</tr>
<tr>
<td>Angular speed</td>
<td>0.24°/sec</td>
</tr>
</tbody>
</table>

Table 10 Resume of Tip-tilt mechanism requirements

6.1.1 Ball Aerospace

The Ball Aerospace two-axis control gimbals [44] are destined for a space application market and are designed for the high performance market. The cost issue is considered as an expensive solution with respect to a low-cost train application that is envisioned in the scope of this project.

Figure 54 Two-axis control gimbal
6.1.2 Directed Perceptions Pan-Tilt Unit

The Directed Perceptions product range is destined primarily for use on closed circuit surveillance systems to mount a video camera. This solution is considered as low-cost yet meets all technical requirements for a train antenna application. It can be used with a direct mount or with nodal brackets if the centre of rotation of the payload needs to be aligned with the rotation axes (see Figure 55). The pan-tilt unit is delivered with an integrated controller which easily interfaces to a PC host computer (see Figure 56).

![Figure 55 Nodal version with brackets aligns payload in pan and tilt rotation axes](image.png)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>EMS 704</th>
<th>EMS 706</th>
<th>EMS 708</th>
<th>EMS 712</th>
</tr>
</thead>
<tbody>
<tr>
<td>Yoke Diameter (in.)</td>
<td>4</td>
<td>6.7</td>
<td>8.5</td>
<td>12</td>
</tr>
<tr>
<td>Weight, payload (lb)</td>
<td>2.2</td>
<td>23</td>
<td>9.9</td>
<td>22</td>
</tr>
<tr>
<td>Weight, gimbal (lb)</td>
<td>11.8</td>
<td>26</td>
<td>29</td>
<td>41</td>
</tr>
<tr>
<td>Material</td>
<td>Titanium</td>
<td>Titanium</td>
<td>Titanium</td>
<td>Beryllium</td>
</tr>
<tr>
<td>Resonant frequency (Hz)</td>
<td>75</td>
<td>60</td>
<td>100</td>
<td>50</td>
</tr>
<tr>
<td>Range of travel (deg)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Elevation</td>
<td>80</td>
<td>60</td>
<td>30</td>
<td>74</td>
</tr>
<tr>
<td>Azimuth</td>
<td>160</td>
<td>180</td>
<td>Continuous</td>
<td>44</td>
</tr>
<tr>
<td>Accuracy (arcmin)</td>
<td>1200</td>
<td>96</td>
<td>96</td>
<td>25</td>
</tr>
<tr>
<td>Operating voltage (V dc)</td>
<td>28 ±4</td>
<td>28 ±4</td>
<td>28 ±4</td>
<td>28 ±4</td>
</tr>
<tr>
<td>Standard data interface</td>
<td>Analog</td>
<td>Analog</td>
<td>Analog</td>
<td>Analog</td>
</tr>
<tr>
<td>Operating temperature range (°C)</td>
<td>-20 to 40</td>
<td>-13 to 35</td>
<td>-19 to 38</td>
<td>-20 to 40</td>
</tr>
</tbody>
</table>

Table 11 Ball Aerospace control gimbals product range
Pan-Tilt Performance:

* Maximum Rated Payload: 2.7kg
* Maximum Speed: over 300 degrees/second
* Acceleration/Deceleration: Trapezoidal, on-the-fly speed and position changes
* Resolution: 3.086 arc minutes (0.051428°)
* Tilt Range (approx): 78° range (31° minimum up and 47° down) with option of 80° down (111° range)
* Pan Range (approx): ±159° (318° range) with option of ±180° (360° range)

6.1.3 QuickSet International

Quickset [45] is the leading supplier of rugged all-weather antenna positioning systems for mobile or fixed mount applications. Their products are used in space, marine, military and surveillance applications. Their smallest range products are positioners to handle payloads of up to 10 kg. Even though this particular product range is very versatile, it is considered as oversized for a steerable antenna application on a train.

| Medium Positioners to handle payloads up to 20 lbs. (10 kg) |
|---|---|---|
| **QPT-15XD** | 15 lb capacity, 24 VAC fixed speed models, ± 217.5° pan, ± 90° tilt. | **QPT-20XD** | 20 lb capacity, 12 & 24 VDC variable speed, ± 217.5° pan, ± 90° tilt. | **QPT-20IC** | With integrated microprocessor control. |

Table 12 Products available that meet requirements
Controllers

QuickSet offers a full suite of pan and tilt controllers for precise and smooth control of pan and tilts and can easily be desktop or rack mounted. Controllers range from basic pushbutton/joystick to full RS 232 & RS 422 PC Computer Controlled models.

QuickTrac
Models for 12 & 24 VDC output, 12 & 24 VDC input, 115 & 220 VAC input, position feedback meters, autocan, proportional joystick operation, remote joystick, rack mount.

QuickComm
Models for 12, 24, & 115 VDC output, 12 & 24 VDC and 115 & 220 VAC input, position feedback meters, RS-232/422, presets, autocan, proportional joystick operation, rack mount.

Analog Controls
Models for 12, 24, & 115 VDC output, 24 & 115 VAC output, 12 & 24 VDC input, 115 & 220 VAC input, position feedback meters, autocan, remote joystick, pushbutton or joystick.

Table 13 Control interface to a host computer

<table>
<thead>
<tr>
<th>KEY FEATURES:</th>
</tr>
</thead>
<tbody>
<tr>
<td>• Weatherproof, suitable for Inverted Operation</td>
</tr>
<tr>
<td>• 435° pan rotation</td>
</tr>
<tr>
<td>• 20 pound load (12 &amp; 24 VDC), 15 pound load (24 VAC)</td>
</tr>
<tr>
<td>• All metal gearing for improved durability</td>
</tr>
<tr>
<td>• Aluminum housing with powder coat finish for white &amp; black models</td>
</tr>
<tr>
<td>• Improved torsional rigidity of both housing and table</td>
</tr>
<tr>
<td>• Improved corrosion resistance</td>
</tr>
<tr>
<td>• Single bolt cover access for internal limit switch adjustments</td>
</tr>
<tr>
<td>• Optional heater is available</td>
</tr>
<tr>
<td>• Position feedback encoders or potentiometers available in certain models</td>
</tr>
</tbody>
</table>

Figure 57 Product features

6.1.4 AvL Technologies

The AVL designs [46] are based on a patented Roto-Lok® drive system that utilizes highly-reliable aircraft control cables in a redundant configuration to provide a zero-backlash, lightweight, very stiff and accurate antenna positioner. This high-tech performance is achieved using low-tech components - by simply wrapping the cable around the driver capstan several times before wrapping the larger driver drum.

The AvL product range is in the high-performance range (higher cost) as the Quickset International products. It was therefore also excluded as a potential solution for the current application.
6.1.5 CSEM custom-design antenna gimbal

Another approach considered was a custom, CSEM designed antenna gimbal as shown in Figure 59 and Figure 60 for a two-axis steering mechanism. This solution would have required a full development in order to meet all the requirements. A number of critical tasks would have been required to perform a detailed design study in order to size and select the various components. A dynamic analysis to size the motors and gearing system would be necessary, integrating adequate encoders, electronics and rack development. These tasks were considered but would have required extensive resources from the project and it was discarded. Dedicated closed-loop control software with a tracking algorithm would have also been a critical point to add to the development. For the reasons stated above, the drive assembly needed to be a commercially available product with the effort emphasized in the integration of
the chosen antenna to the system and to develop the tracking algorithm and associated communication requirements (antenna error signal input/drive commands).

![Antenna Gimbal Diagram](image)

**Figure 59** CSEM custom-design drive system

![Antenna Gimbal Diagram](image)

**Figure 60** CSEM custom-design antenna gimbal solution

### 6.2 Testbed of Mechanical Steering System

#### 6.2.1 Antenna gimbal design based on Commercial Off-The-Shelf modules

*Payload mounting at intersection of rotation axes*

In order to perform tests with an antenna, a support structure will be designed and built to interface the antenna to the gimbal drive units. The gimbal drive units are based on the Directed Perceptions Pan-Tilt Unit. For a nodal configuration (centre of antenna intersect centre of axes rotation) of the device, the baseline unit (left-below) can be modified with a custom support.
The configuration above was not necessary for our application and a direct mechanical support to the pan-tilt unit was manufactured in order to perform the tests with a tracking signal (see Figure 62).

**Payload mounting at pan rotation axis**

The more direct and simple mounting approach is the pan-tilt configuration that can easily support the antenna.

The following information defines the standard unit that was used for the tests:

- **Model number:** PTU-D46-70
- **Power supply product code:** 239 AC/DC Power supply 220VAC/24VDC
- **C-Programmers Interface:** PTU-CPI

### 6.2.2 Signal tracking algorithm implementation

The signal tracking system was implemented based on power level detection. The receiver mechanism is firstly initialized to point to a direction where it gets maximum of RF power level. The initialization follows a spiral trajectory referred to the Pan and the Tilt angles as presented in Figure 63. The system continues oscillating around the initial maximum point, measuring power level in the neighbourhood of the centre and moving to a new maximum point if any change in power level is detected. This is illustrated by the schema bloc presented in Figure 64. The radiation pattern of the transmitting antenna needs to be provided for the system for the initialization period and for defining the decision threshold (Figure 66).
6.2.3 Tracking algorithm testbed

A testbed has been built for demonstrating the tracking algorithm performance at 27.5 GHz.

As a first step, two commercial large-bandwidth horn antennas (Figure 65) were used for the emission and the reception. By definition, the antenna radiation pattern is measured at far-field region where radiation is independent on the distance between TX and RX antennas. The distance where the far-field region begins is defined in [39]:

\[ R = \frac{2D^2}{\lambda} \]

where D is the largest linear dimension of the antenna.
It's easy to show that the far-field region starts at about 3 metres from the horn antenna, 4 metres from
the 64-element array, and 16 metres from the 256-element array. It is possible to make the test at
shorter distance in near-field region (for example 1-2 metres) under the condition that the radiation
pattern referred to should be measured at that distance. In the present work, the transmitter and the
receiver were placed at 4 metres apart.

As presented on the block diagram in Figure 66, the transmitter consists of a signal generator and the
horn antenna providing a beam of 37° width (HPBW) at 27.5 GHz. The antenna is interfaced to the
generator through coaxial cable. At the receiver, on a mobile platform, another horn antenna of the
same performance catches microwave energy in air and sends it back to a power detector which
performs RF-to-DC conversion. The resultant DC signal is amplified before being sent to the unit where
the trajectory is computed. The algorithm has been run on a PC using a data acquisition card (DAQ)
from National Instruments. It has been observed that the tracking quality is good as long as the
transmitter speed is at the same range as the sample speed of the PC (sample frequency 10 Hz). In
future research work, the geometry of the volume used in the algorithm can be further optimised. In
addition, with additional investment on software (LabVIEW Real-Time) and hardware (new acquisition
card), more rapid tracking should be achievable (at sample frequency 1KHz).

Figure 65  ETS Lindgren model 3116 antenna and interface
Figure 66  Block diagram of tracking algorithm testbed

Figure 67  Control unit with added trajectory perturbation

Figure 68  The 64-element subarray and the polarizer mounted on the Pan-Tilt Unit
6.3 Electrical beam steering: possibility of array built with directive elements

The objective of this section is to evaluate the characteristics and limitations of electrically steered antenna arrays built with directive element, with special emphasis on the possibility of steering direction and the gain that can be achieved at that direction.

Here we consider a \( M \times N \)-element planar array in \( xy \)-plane with uniform power distribution and uniform element spacing. The normalized total array factor \( AF \) is a combination of the AF in the \( x \) and \( y \)-directions [39]:

\[
AF = AF_x \cdot AF_y = \sum_{m=1}^{M} e^{i(m-1)(kd_x \sin \theta \cos \phi + \beta_x)} \sum_{n=1}^{N} e^{i(n-1)(kd_y \sin \theta \sin \phi + \beta_y)}
\]

where \( \theta \) and \( \phi \) are respectively elevation of azimuth angle of reference point. \( \beta \) is the phase difference between the elements of the array. If we want to adjust the main beam of the array to a direction of \( \theta = \theta_0 \) and \( \phi = \phi_0 \), the phase difference between the elements need to be

\[
\beta_x = -kd_x \sin \theta_0 \cos \phi_0 \\
\beta_y = -kd_y \sin \theta_0 \sin \phi_0
\]

The total radiation of the array at a certain direction is obtained by multiplying the single element radiation at reference point by the resulted AF at that direction. The above described formulas were implemented using MATLAB. Simulations have been run for the upper-sphere above the array aperture. Achievable array gain as a function of steering angle is presented in Figure 69.

![Figure 69 Radiation of a 4-element array using directive element](image-url)
6.4 Conclusions on mechatronic array antenna

A tracking algorithm has been developed using the maximum finding method. The algorithm was tested on a COTS pan-and-tilt system with commercial horn antennas.

When mounted on a train, the system has to establish the communication link the first time without prior knowledge of the position of the target (cold start). This approach requires an iterative search of the sky to locate the other unit (opposite receiver/transmitter). CSEM has bread-boarded an azimuth elevation antenna terminal and tested a few algorithms with success for such iterative scans, a first step towards a complete mobile antenna link to be mounted and tested on a moving train.
7 Properties of Spherical Lens Antennas for High Altitude Platform Payload

This self-contained chapter discusses the potential benefits of using multi-beam spherical lens antenna for the HAP payload antennas in the context of the spot-beam architecture for a cellular communications network.

7.1 Introduction

HAPs may exploit millimetre-wave spectrum by supporting multi-cell architectures. The overall data capacity is governed by the properties of the antennas which serve the cells. Optimum coverage is expected when each antenna radiation pattern is tailored to its respective cell, but this leads to a bulky payload. In contrast, this section shows how a payload based on multi-beam spherical lenses, which have wide-scan properties but also offer circular beams, can give comparable performance while offering a much more compact payload. Models for spherical lens radiation patterns are presented which include the effect of the primary feed beamwidth. This allows a comparison of aperture diameter, sidelobe level, and the resulting co-channel interference, from which may be estimated the performance and relative mass of different antenna payload types.

The ITU have allocated the 47/48 GHz band [6] for usage by HAPs worldwide, and the 27/31 GHz band has also been allocated to about 40 countries [7]. At these frequencies it is difficult and expensive to produce electronically re-configurable antennas based on phased-array principles and digital beam forming (DBF). For example, DBF presents challenges associated with: the close spacing of radiating elements, the requirement for a large number of elements, the need to minimize sidelobe levels, the cost and complexity of RF active circuits, and the problem of scanning loss. The latter point is particularly detrimental for HAPs communications because links to users at low elevation angles, where scanning loss for a horizontal planar array are worst, experience also the longest free space and atmospheric losses. Nevertheless a number of programmes are developing DBF antennas for Ka band e.g. [8] which reports a 16-element module, and the Japanese HAPs programme [9] reporting a similar prototype for 28 GHz while conceding that horn antennas stabilized by mechanical gimbals are a more pragmatic solution for 48 GHz.

The more conservative solution for the HAP antenna payload is a group of aperture antennas, where each antenna serves a single quasi-stationary cell. This approach has been studied in some depth [4], [5], [10] where the important relationship between the radiation patterns and the CIR levels on the ground has been quantified for various channel re-use schemes. Ultimately, the economic viability of a HAP cellular network may be determined by the extent to which it can maximize the re-use of spectrum. This paper explores the properties of multi-beam antennas based on spherical lenses, since these can yield a much more compact HAP antenna payload compared to the use of single-beam aperture antennas. An estimate of payload mass compared to CIR coverage levels is introduced.

7.2 Dedicated beams for the HAP antenna payload

It is worth listing the pros and cons of using dedicated beams, i.e. one antenna per cell. Advantages include:

- each antenna pattern can be tailored to the shape required by each cell.
- asymmetric beams yield equal size, circular cells, which aids tessellation of cells and maximizes coverage [11].
- low side-lobe levels are feasible e.g. -40 dB.

Disadvantages include:
• A bulky, massive payload, which requires mechanical stabilisation.
• A non-reconfigurable network.

These disadvantages can to some extent be overcome with multiple-beam antennas which allow for the payload mass and volume to be reduced, but at the expense of eroding the advantages associated with dedicated beams because it becomes more difficult to tailor each beam to each cell. Multi-beam aperture antennas typically use a cluster of primary feeds to illuminate a secondary aperture which might be a prime-focus reflector or lens, but the angular range of scanning is limited. For a HAP payload a wide conical scan is needed e.g. 150° for users at elevation angles as low as 15°. In the following sections we explore the potential to use multiple-beam, wide-scanning antennas based on spherical lenses, and quantify the CIR values which could be expected.

7.3 Properties of Spherical Lens Antennas

In the context of HAPs communications, the very useful property of antennas using spherical lenses is their ability to form multiple beams over a wide range of angles without inducing any scanning loss. A number of spherical lens variants are well know, and some are briefly reviewed below.

A classic type of spherical lens is the Luneburg lens [12], which focuses an incident plane wave at a point on the lens surface. The spherical geometry allows for multiple beams to be produced from multiple feeds [14], and it is this property which is attractive for a HAP payload. The Luneberg lens requires a radial variation of dielectric constant:

$$\varepsilon_r = 2 - \left(\frac{r}{R}\right)^2$$

(1)

where \(r\) is radius within the lens, and \(R\) is the outer radius. Thus the dielectric constant varies from 1 at the lens edge to 2 at the centre and there is no reflection loss due to abrupt transitions. This leads to a very good aperture efficiency, but it somewhat problematic to manufacture. In practice such a lens is often fabricated from a set of concentric shells and this inevitably degrades the efficiency: a useful treatise is given by [17]. However, for limited bandwidth applications a two-shell design can give almost equivalent performance [18].

A single shell spherical lens has no single focus, but exhibits a paraxial focus when a small proportion of the lens is illuminated. The paraxial focus may lie inside or outside the lens outer radius. A low dielectric spherical lens can give good performance - e.g. [19] reports on a Teflon lens (\(\varepsilon_r = 2.08\)) for wide-scanning automotive radar at 77 GHz - and is of course much more simple to manufacture than a multi-shell lens.

Using the Luneberg lens aperture distribution given by [12], where the primary feed pattern is also taken into account, the far field radiation pattern can be computed. The geometry is illustrated in Figure 70 for a generalized lens where the focus may lie inside or outside the lens radius \(a\). However, for the remainder of the analysis we will assume a conventional Luneburg lens where \(f = a\).
A convenient model for the primary feed power pattern $P_{PF}$ as a function of angle $\theta''$ is:

$$P_{PF} = \cos[\theta'']^n$$  \hspace{1cm} (2)

where

$$\theta'' = \arcsin\frac{r}{f}$$  \hspace{1cm} (3)

then from (2) and [12] we obtain for the radial power distribution $P(r)$ in the aperture plane:

$$P(r) = \frac{\cos\left[\arcsin\frac{r}{f}\right]^n}{\sqrt{f^2 - r^2}}$$  \hspace{1cm} (4)

The power pattern $P(\theta)$ of (4) may be derived in general by the transform to the far field. However, for $n=1$ the $r$ dependence of $P(r)$ disappears and we have a uniform aperture. This of course yields the highest aperture directivity for a given diameter $2a$ and also serves as a useful shortcut for deriving the radiation patterns for each HAP payload antenna because the standard formula for a uniform aperture may be used [13]. Where a more severe primary feed roll-off is applied i.e. by increasing $n$, the lens aperture plane experiences a tapered distribution, leading to lower side-lobes but also a reduced directivity; in such a case the lens diameter must be increased to recover the required beamwidth for each cell (this is explored in detail later). Radiation patterns from (3), for a lens diameter of 10 wavelengths, are shown in Figure 71. When the $n=1$ model is used to generate CIR in a multi-cell network as described below, this type of radiation pattern gives the lowest CIR values due to the high side-lobe levels.
7.4 Power Contours and Co-channel Interference

In contrast to earlier studies where each antenna beam is optimized, the use of multiple-beam spherical lens antennas leads to each beam of a given cluster having the same beamwidth and also having circular symmetry. Taking the layout of 121 cells shown in Figure 72, where cells are arranged in 6 concentric hexagonal rings, we can specify an antenna beamwidth associated with each ring based on the mean required beamwidth. This is a modification of the methodology of [5] and leads to non-circular cellular power footprints but allows the use of a single spherical lens aperture for each ring and thus a much more compact antenna payload. Elements of such a payload are illustrated in Figure 73, which shows how the innermost group of cells are served by the smallest spherical lens with a cluster of appropriately spaced feeds, and the outermost group of lens are similarly served by the largest spherical lens. The lens diameters indicated are approximations based on uniform aperture illumination and a 28 GHz carrier frequency and thus serve as a benchmark estimate for minimum antenna dimensions at this frequency.

Figure 73. Concept for antenna payload: multi-beam Luneburg lens for each cell group

Figure 74 shows power contours for a co-channel cell group where case (a) is for dedicated beams which are asymmetric, and case (b) is for circular symmetric beams which could be generated using multi-beam spherical lenses. The difference in these two methods is apparent when we observe how the circular beams lead to non-circular cell power contours. The HAP payload in the former case comprises...
121 individual antennas, while in the latter case it comprises 6 spherical lenses (one for each hexagonal ring), plus a single antenna for the centre cell. The lenses require multiple feeds, these typically being waveguide horns. In the outer ring the cell azimuth spacing is typically 10.1° and the lens diameter for 28 GHz would be approximately 120 mm. This leads to a feed horn spacing of about 21 mm which is adequate for a small horn fed by circular waveguide or dielectric-loaded horn.

(a) dedicated beams. (b) Multiple beam spherical lenses

Figure 74. Power contours for 1 of 3 channels (contour spacing is 1.5 dB).

The co-channel CIR values associated with the two different payload types of Figure 74 (a) and (b) are presented as cumulative distributions in Figure 75. Comparing the dedicated, asymmetric beam case with the multi-beam spherical lens case the CIR difference is typically between 5 - 7 dB across the coverage area. This trend is expected because the dedicated beams model uses a flat sidelobe floor at -40 dB, while the circular beams model uses the theoretical sidelobe structure for uniformly illuminated circular apertures (Figure 71) whose mean relative level is somewhat greater than -40 dB. Hence we are able to present a comparison of CIR performance for (i) an established payload model using dedicated beams and (ii) a more compact, less massive payload based on a group of minimum size spherical lenses. The term "minimum size" refers to the assumption that the aperture is uniformly illuminated and is thus, for each cell group, the smallest lens aperture which can be used to generate the required beamwidths.

Figure 75. CIR coverage distribution.
7.5 Reducing the spherical lens sidelobe level

In the multi-beam spherical lens case the CIR values can be improved by using bigger lenses with tapered aperture distributions since this leads to lower sidelobe levels. In this section a recipe is sought from which we may estimate the dimensions of a multi-beam spherical lens payload which would deliver CIR values equivalent to the dedicated beam payload. A tapered aperture distribution implies that a directive primary feed is used. This is also equivalent to \( n > 1 \) in (4). The effect of the widening of the secondary beam is shown in Figure 2.

If the primary feed pattern is given by (3) then, for a given primary feed HPBW \( \theta_p \), since

\[
\cos\left(0.5\theta_p\right)^n = \frac{1}{2}
\]

we have

\[
n = \log_{\cos\left(0.5\theta_p\right)}\left(\frac{1}{2}\right)
\]

and thus the primary feed pattern is described by a single parameter \( n \) which is derived from its HPBW. We may then relate the aperture field distribution to the far field by employing the integral:

\[
E(\theta) = \frac{1}{r} A(r) r J_0 \left( k a \sin(\theta) r \right) dr
\]

where \( A(r) \) is the radial aperture field distribution which is given by \( A(r) = \sqrt{P(r)} \) from (4). We may then evaluate (7) for a given combination of aperture radius \( a \) and primary feed HPBW so as to derive the HPBW of the secondary field. One approach is to plot the main lobe patterns and in each case search for the half-power points using a computer algorithm. This procedure lead to the results summarized in Figure 76 where the primary feed HPBW\'s chosen are 30°, 45°, 60° and 90° and these parameters give rise to the discrete data points which are shown.

![Figure 76. Relationship between primary feed and secondary HPBW for Luneburg lens.](image)

(a) Secondary HPBW versus primary HPBW for radii in wavelengths as labelled

(b) Beamwidth widening factor compared to uniform aperture.

In Figure 76(a) can be seen the expected trends that a larger aperture gives rise to a narrower secondary beam, while a narrower primary feed beam gives rise to a wider secondary beam. The latter effect is also
associated with a reduction in sidelobe levels. Figure 76(b) shows the increase in secondary beamwidth, relative to the uniform aperture \((n=1)\), as a function of aperture radius \(a\) and for the four cases of primary HPBW chosen. Here we find a very useful result: for a given primary feed, the widening of the secondary beam is by an approximately constant factor for the range of aperture radii chosen.

Considering the HAP antenna payload, we now have a recipe for estimating the amount by which the antenna beamwidths would need to be increased when the lens aperture experiences an amplitude taper brought about by a generic primary feed. The beam widening must be countered by a commensurate and linear increase in the lens radius so as to recover the beamwidth which is required by each cell group. The calculated scaling factors are listed in Table 14 and summarized as an extrapolated function of primary feed HPBW in Figure 77.

<table>
<thead>
<tr>
<th>primary feed HPBW (°)</th>
<th>beam scaling factor</th>
<th>first sidelobe level relative to peak gain (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>uniform aperture</td>
<td>1</td>
<td>-17</td>
</tr>
<tr>
<td>90</td>
<td>1.06</td>
<td>-19</td>
</tr>
<tr>
<td>60</td>
<td>1.22</td>
<td>-24</td>
</tr>
<tr>
<td>45</td>
<td>1.43</td>
<td>-30</td>
</tr>
<tr>
<td>30</td>
<td>1.87</td>
<td>-44</td>
</tr>
</tbody>
</table>

Table 14. Summary of Luneburg lens beam widening and sidelobe level reduction for various primary feeds.

The derivation of CIR for the co-channel cell group now requires (7) to be evaluated so as to derive the power footprints of each cell. (The integration involved in (7) is much more computationally intensive than the previously reported radiation pattern models and thus leads to much longer computer run times.) Some results for computed CIR distributions are shown in Figure 78 for multi-beam antennas using various primary feed beamwidths and where the aperture diameter is scaled accordingly so as to maintain the required secondary beamwidth.

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Figure 77. Increase in secondary beamwidth, compared to that of uniform aperture, as a function of primary beamwidth.

Figure 78 CIR distributions for dedicated and multi-beam payloads.
7.6 Relative payload mass

To close the analysis we present an estimate of the total mass for the HAP antenna payloads which have been discussed. For the dedicated beam payload, a working estimate for total lens mass can be derived as follows: Using the mass of our experimental asymmetric beam lens antenna [11] as a benchmark, and noting that, to a good approximation, there are straightforward relationships between half power beamwidth (HPBW), mass, and lens diameter $D$:

$$\text{HPBW} \propto \frac{1}{D} \quad \text{and} \quad \text{mass} \propto D^3$$

hence by scaling to the 466 g. mass of the experimental polyethylene lens whose HPBW in the narrow plane$^1$ is $5^\circ$ we obtain an estimate for each lens mass in the dedicated beam payload:

$$\text{mass} = 466 \left( \frac{5}{\text{HPBW\ degrees}} \right)^3 \quad (\text{g})$$

Using the iterative method of [5] to derive HPBW for each cell in the 121 cell network yields a total lens mass of 28.8 kg. The figure is somewhat tentative, and has been derived for a HAP at a height of 20 km above a 60 km diameter service area. Clearly, if these parameters are altered, the antenna beamwidths (which are tailored to each cell subtended angle) are altered and thus the lens masses also change. The mass of the primary feeds has not been considered in this analysis.

The estimate for the mass of the multi-beam spherical lens payload can be derived from the diameter of each lens and (assuming a single shell lens design) the material specific gravity. Again assuming a polyethylene material, whose density is 947 kg / m$^3$, and uniformly illuminated apertures for the minimum size payload case, a mass of 1.93 kg is derived for the group of 6 spherical lenses. The figure would be less for multi-shell or Luneburg type lenses. Again, the mass of the primary feeds has not been considered, as it is possible that these could be very similar for the two payload cases and their number would also be the same.

The trade-offs between use of a dedicated beam payload assuming -40 dB sidelobe floors, and multi-beam spherical lenses assuming close-to-uniform aperture illumination, are summarized in Table 15. It is stressed that the masses presented are estimates based on extrapolation from those of representative components. Nevertheless we can see that, subject to these various assumptions, the multi-beam lens approach leads to approximately 93 % saving in the mass of the dielectric lenses at the expense of a CIR degradation typically of 6 dB.

<table>
<thead>
<tr>
<th></th>
<th>number of primary feeds</th>
<th>number of lenses</th>
<th>estimate of total lens mass (kg) (for 28 GHz)</th>
<th>CIR (dB) (minimum / maximum)</th>
</tr>
</thead>
<tbody>
<tr>
<td>dedicated beams</td>
<td>121</td>
<td>121</td>
<td>28.8</td>
<td>12 / 24</td>
</tr>
<tr>
<td>multi-beam spherical lenses:</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>(i) minimum aperture size</td>
<td>121</td>
<td>6</td>
<td>1.93</td>
<td>6 / 18</td>
</tr>
<tr>
<td>(ii) scaled aperture with 60° primary feeds</td>
<td>121</td>
<td>6</td>
<td>3.50</td>
<td>12 / 29</td>
</tr>
</tbody>
</table>

Table 15. Comparison of payload properties and CIR values for dedicated beams and multi-beam spherical lenses serving 121 cell network

$^1$ The experimental asymmetric beam lens antenna has a circular cross section whose diameter is dictated by the narrow (elevation) beamwidth of the cell. The beamwidth in the azimuth direction is broadened by modifying the lens surface [11].
7.7 Conclusions for HAP Spherical Lens Antennas

HAPs offer an effective way of exploiting mm-wave spectrum by supporting multi-cell architectures. The viability of such systems is largely determined by the overall system data capacity, which is in turn governed by the properties of the antenna payload which serves the cells on the ground. In previous studies, “ideal” antenna beams have been used to model carrier-to-interference ratio. In such cases, dedicated aperture antennas such as lens antennas may be used to produce the required beam shapes which are in general asymmetric. This approach leads to one antenna for each cell and thus a bulky payload.

Where multi-beam antennas are used the mass and volume of the HAP payload can be much reduced. A convenient approach is to employ spherical lens antennas using multiple feeds. This work has considered the properties of such antennas and shown how for a 121 cell network the payload may be reduced to just 6 multi-beam antennas. Disadvantages of this antenna type is that radiation patterns are in general of circular cross-section, which leads to non-circular cell footprints, and a compromise beamwidth must be chosen for a given cell group.

The resulting CIR levels have been computed for spherical lens antennas where the aperture distribution is uniform, this being the smallest antenna but with the highest sidelobes. The technique was then extended to consider the effect of the primary feed beamwidth, since this may be chosen to taper the aperture distribution and thus reduce sidelobe levels, but must be accompanied by an increase in lens diameter so as to maintain the required directivity for the cell group. A generalized recipe was found whereby the aperture scaling term was derived for a range of primary feed beamwidths. This allowed direct comparison of the relative size of the antenna payload for given CIR levels. The trade-off is one where the smallest antennas have the highest sidelobes and hence lowest CIR, while increasing the antenna size leads to increased CIR levels.

The benchmark payload, using 121 dedicated antennas where each is assumed to have a mean sidelobe floor at -40 dB, offers CIR levels between 12 dB and 18 dB and was shown to have a total dielectric lens mass estimated at 28.8 kg. Compared to this, a payload of 6 multi-beam spherical lenses of minimum possible size would offer a mass reduction of at least 93 % (assuming single shell, not Luneburg lens) but a CIR degradation of around 6 dB. It was shown that the CIR levels offered by the dedicated antenna payload may also be obtained by the multi-beam spherical lens payload if each lens aperture diameter is increased by a factor of 1.22 in conjunction with primary feeds of 60° half-power beamwidth. The total lens mass in this case would be approximately 88 % less than that of the benchmark payload.

While operational HAP communications services have yet to be deployed in practice, it has been shown that a group of multi-beam spherical lenses could offer a very practical and compact antenna payload. This would support multi-cell architectures with adequate levels of co-channel interference and with a very considerable mass saving compared to a payload of a type which uses one lens antenna for each cell.
8 Final Conclusions

In this report various candidate solutions have been considered across a wide spectrum but two widely differing clusters have emerged. In one cluster, the more straightforward, conservative or near-term solutions would be offered by proven antenna technologies such as microwave apertures (waveguide horns), lenses and reflectors, integrated into either the HAP payload for serving cellular architectures, or a mechanical control system for deployment on a high-speed train. Such solutions would be characterized by excellent electromagnetic performance, but perhaps rather unwieldy mechanical systems to implement the steering functionality. In the opposite cluster would be the ‘smart’ or digital beam forming (DBF) antennas: here, no moving parts would be needed, and the smart antenna would theoretically offer some astounding advantages: multi-beam apertures offering null steering for interference cancellation and very rapid beam scanning. Due to project resource limitations, DBF techniques were not explored experimentally but were investigated theoretically in Workpackage 3.3.

The array antenna developed at CSEM, primarily considered for vehicular communication, was implemented using Strip-slot-foam Inverted patch technology - a multi-layer substrate approach which overcomes the bandwidth limitations of single-layer printed antennas. This yielded a lightweight antenna easily driven by servo motors, and several approaches to gimbals were considered. Tracking algorithms were developed and a laboratory testbed so as to demonstrate a representative, integrated antenna with a control system.

For the HAP antenna, dielectric hemisphere lenses were developed to validate a family of antenna types based on lens technology, at York. A notable and novel result was the development of a two-layer lens antenna, this being loosely inspired by the Luneburg antenna but offering a simpler and more reliable route to construction using well-behaved, low-loss materials. The antenna offered a gain of 35.4 dBi at 28 GHz, being 68 % aperture efficient. While a steering mechanism for this antenna was beyond the project's resources, a number of concepts have been discussed and sketched in some detail.

In the final chapter, the application of spherical lens antennas for a multi-beam HAP payload for cellular communications, where these are shared by clusters of primary feeds, was theoretically investigated at the system-level. Here, the benchmark was the ‘dedicated beam’ payload model from HeliNet, where elliptic beam antennas are dedicated to each cell on the ground. The lens antenna solutions put forward have slightly less advantageous beam shapes, but can very substantially save on payload volume and mass due to the shared apertures.
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