Dimensioning a Retro-Directive Array for Communications via a Stratospheric Platform

John Thornton

High Altitude Platforms–craft maintaining stations in the stratosphere at altitudes of around 20 km–have been proposed as a means of supporting wireless telecommunications. They could exploit the best aspects of both terrestrial and satellite systems and support efficient frequency re-use plans. For solar powered platforms the power available for the downlink amplifiers may be minimal, particularly at night and/or higher latitudes. This paper discusses a novel type of link based on a modulated retro-directive transponder carried by the HAP. Relying chiefly on the ground station infrastructure, this would substantially reduce power consumption on the platform. We investigate the efficiency of the transponder aperture as a function of its area by developing general models for losses in the transmission lines which interconnect antenna pairs in the retro-directive array.

I. INTRODUCTION

Stratospheric or High Altitude Platforms (HAPs) are increasingly being cited [1], [2] as playing an important role in future communications systems and applications. They have the potential to exploit many of the best aspects of terrestrial and satellite based systems, while offering advantageous propagation characteristics. Platforms based on airships [3], [4], solar powered un-piloted aircraft [5], [6], and conventional piloted aircraft [7] have all been mooted and are at various stages of development. While the conventional piloted aircraft approach would seem the least technologically challenging, being essentially a conventional aircraft flown in shifts of a few hours duration, the solar powered solutions appear the most attractive from considerations of endurance and hence running costs and of environmental impact. Such platforms are unlikely to be fully developed for several years, but their potential for supporting high capacity wireless services does appear to be attractive.

A disadvantage of the solar powered HAP is its reliance on fuel cells which would be charged during daylight hours to power the platform and its payload by night. The power budget afforded by this cycle of charge and discharge using current solar cell and fuel cell technology may be marginal, and particularly so at increasing latitudes during the winter months [2]. While advances in technology should improve matters, a HAP communications payload will always be allocated only the electrical power which remains from the share consumed by the motors and control systems which keep the platform aloft (for aircraft) or on-station (for airships). The majority of the available power is likely to be consumed by these propulsion and stability systems [8]. The prime power consumer in the

---

Manuscript received Aug. 9, 2001; revised Nov. 27, 2001.

John Thornton (phone: +44 1904 433220, e-mail: jt21@ohm.york.ac.uk) is with the Department of Electronics, University of York, UK.

payload would be the RF power amplifiers (PAs), which, along with the antennas, are among the most critical components in the communications payload. In circumstances where the supply of power is marginal, the PAs would be run at reduced power or be shut down completely, implying reduced quality or complete loss of service. The backhaul down link–a high data rate link connecting all the down link data to a terrestrial network via a hub (ground-station)–has been identified as the most susceptible to marginal link budgets such as would be encountered during severe rain events [9]. To some extent, this can be obviated with spatial ground-station diversity [10], but cannot mitigate against a loss of DC power on the HAP. This work examines an alternative type of link which may have a place in circumventing the problem of link loss where RF power is lacking.

The pioneers of satellite communications suggested a link via a reflector as a means of facilitating over-horizon communications, where the reflector might be passive [11] or include amplifiers [12]. Indeed, reference to reflective links via satellite can be found in textbooks of the 1960s–[13] includes an interesting chapter entitled "Passive space communication systems." However, satellite communications evolved in a very different direction, based on separate uplink and downlink frequencies transmitted via traveling wave tube amplifiers and high gain antennas mounted on a stabilized space vehicle. The reflective link was overlooked in light of such developments. However, the very much reduced path length to a HAP compared to a satellite leads to new possibilities for a link via a reflector, which can now be examined in a fresh light.

II. WIRELESS COMMUNICATIONS LINKS TO MODULATED SCATTERERS

Point-to-point radio links undergo free space loss as an inverse function of distance squared ($1/r^2$). In radar, the link, by reflection from a target of interest, undergoes a further $1/r^2$ path loss leading to the $1/r^4$ term in the monostatic radar equation:

$$P_r = \frac{P_t G_r G_t \sigma \lambda^2}{(4\pi)^3 r^4}, \quad (1)$$

where P_r is the power received by a target of radar cross section (RCS) σ which is at a distance r from the transceiver which transmits power P_t. G_r and G_t are the gains of the receive and transmit antennas, respectively. A link of this type is usually intended to establish the presence, location, speed, size and type, etc., of a scatterer and is not usually regarded as a "communication" link. However, where the scatterer is replaced by a cooperative target, or transponder, data can be transferred from the target to the interrogating transceiver. The transponder may be "active," i.e., it contains DC and RF power sources [12], [14], or be "passive" [11], [15], where DC power consumption is minimal or zero. In the latter case, a fraction of the incident energy is re-directed towards the interrogator. This type of link is useful where the transponder is required to consume little power and be interrogated from any angle, such as an RF identification (RFID) tag. The transponder must have a close to omni-directional response which implies an antenna without directive gain. Such a link also suffers the $1/r^4$ term of the radar equation and requires excessive transmit power for operation at long range. However, if the RCS of the transponder can be sufficiently increased, the link budget can be made operable at increasing distances. Such an increase in RCS would normally be associated with an increase in the directivity of the transponder's antenna since

$$\sigma = \frac{\lambda^2}{4\pi} G^2, \quad (2)$$

and the increase in directivity would lead to a reduced angular response. The implication of this approach is that the transponder would have to be carefully aligned with the interrogating antenna to avoid pointing losses. For this reason, this technique is not usually adopted. However, if the transponder's antenna is replaced by a retro-directive aperture, the alignment difficulty does not occur. A retro-directive aperture is one which reflects energy primarily back in the direction of the source and can be formed, for example, from a trihedral corner reflector whose RCS is large over a wide range of incidence angles. However, this structure is not convenient for the RCS modulation which is necessary to establish the data link in the reflected spectrum. A more convenient structure is the modulated retro-array–a variant of the Van Atta microwave retro-reflector [16]. This structure comprises any number of antenna pairs which are joined by equal length transmission lines. By placing modulators (or switches) in the transmission lines, the RCS can be modulated, and the information in the modulation products is reflected only in the direction of the interrogating signal. Such a transponder has been proposed for applications where RF power is unavailable at the target of interest and has also found application in range finding systems [17] for vehicle control and collision avoidance. It can be fabricated from low-cost printed circuits [18] and requires only a modest power supply to drive the modulation circuits.

III. DIMENSIONING A RETRO-DIRECTIVE TRANSPONDER FOR A HAP LINK

Returning to our consideration of the power-limited HAP

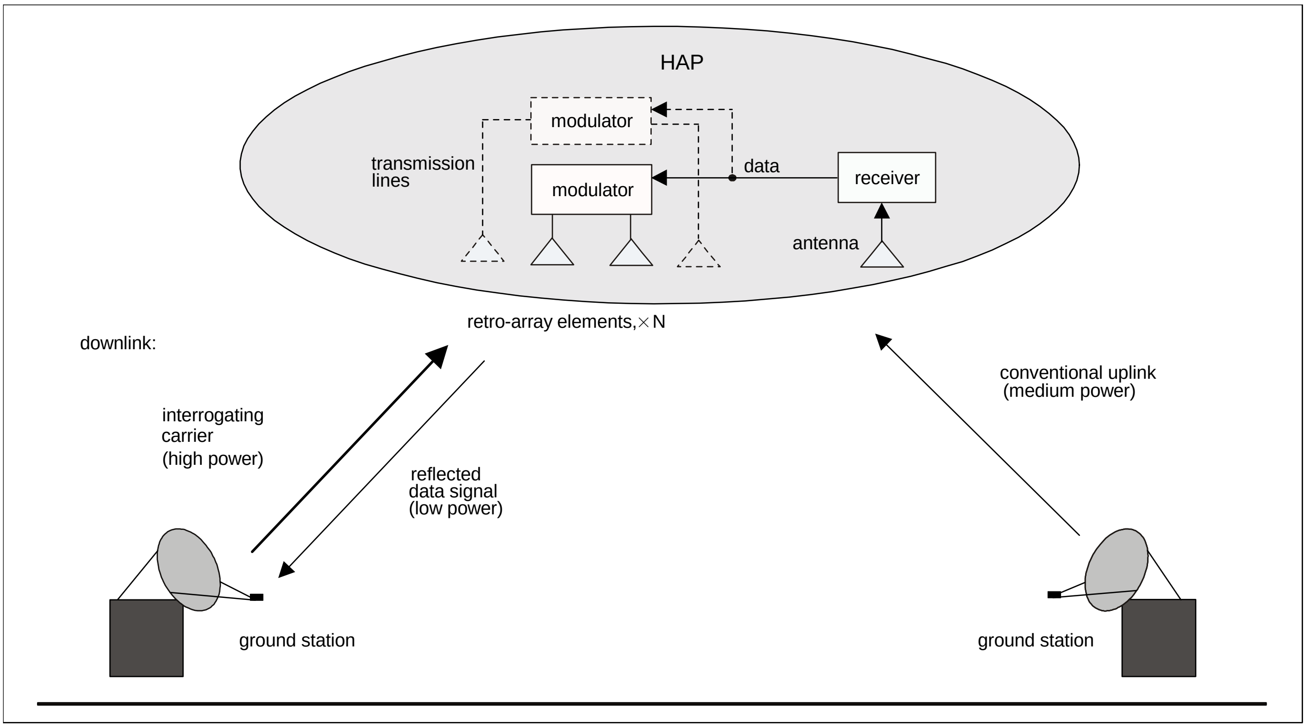

Fig. 1. HAP communications link via “passive” transponder.

downlink, we suggest a novel type of link where a retro-directive passive transponder is placed on the HAP and illuminated from a ground station where the information in the reflected spectrum would be demodulated. The geometry is illustrated in Fig. 1. As discussed in the introduction, it is interesting to trace the heredity of this philosophy to satellite applications proposed in the 1960s [14], [19], when active transponders (i.e., including amplifiers) were being suggested.

Since the link suffers $1/r^4$ free space loss, means of increasing the link budget are sought which include maximizing the following parameters within practicable limits:

- transmit antenna gain
- receive antenna gain
- transmit power
- transponder RCS.

The first three of these parameters are functions of the ground station and ought not to present severe restrictions beyond those of cost. (A similar philosophy has been proposed, albeit in a more extreme form, to provide *all* a platform’s power needs, including propulsion, via multiple microwave links from the ground [20].)

Since the transponder would be placed on the HAP, its size and weight would be expected to be limiting factors; this paper will consider the expected RF efficiency of the transponder as its size increases. At this stage the frequency of operation should be discussed. Proposed HAP carrier frequencies include the bands close to 2 GHz allocated to 3rd generation mobile services and the LMDS/MVDS bands of 28 GHz and 38 GHz discussed above [2]. An additional band, specifically for HAPs, is close to 48 GHz, where 600 MHz of bandwidth is allocated. The choice of carrier frequency would dictate the type of technology adopted for the transponder. Refs. [17] and [18] discussed fabrication techniques based on 16 element arrays of printed antennas with microstrip interconnects and silicon switching networks. These operated at either 2.5 GHz or 9.4 GHz, and the technique could be scaled to the millimeter-wave bands. However, as we consider increasing the size considerably, the losses in the interconnecting lines place an increasing restriction on RF efficiency. Beyond 30 GHz or so, waveguide might be a better solution.

A further complication in choosing the carrier frequency arises from the apparent fluidity of the international regulations–a platform below a height of 20 km may be treated as part of a terrestrial system and can be allowed a greater choice of bands. In addition, the link under discussion has no RF transmitter on the HAP and appears to fall outside the conventional classifications. For these reasons, the choice of carrier frequency will be deferred and a link budget will be presented as a function of aperture area only.

Link budget

The HAP downlink backhaul ground station might be similar in specification to a typical satellite earth station, but with some important differences such as a lesser antenna gain and transmit power. An unusual feature is that the transmitted bandwidth is that of the carrier only since no data is present in the outgoing signal. This bandwidth is that of the RF source alone, which is governed by its phase noise. Also, the RF power amplifier(s) may be driven fully into saturation without the usual corollary of intermodulation distortion, which is again due to the absence of modulation (data) in the transmitted carrier.

To present an elementary link budget which is frequency independent, we can plot the received power or signal-to-noise ratio at the ground as a function of the reflector aperture length *L*. For a square array with uniform element spacing, the total number of elements is given by

$$n = \left(\frac{L}{a} \right)^2 , \tag{3}$$

where *a* is the element separation. Since the total aperture gain is the product of the number of elements and the gain of each element (*ng*), the RCS for the aperture from (2) is

$$\sigma = \frac{L^4 g^2}{4\pi\, s^4 \lambda^2} , \tag{4}$$

where $a = s\,\lambda$. Substituting (4) into (1) yields a term for received power P_r which is independent of wavelength λ, hence,

$$P_r = 10\log_{10}\left(\frac{\mathrm{L}}{4\pi s r} \right)^4 g^2 + 2G_{tr} + P_t , \tag{5}$$

where powers P_r and P_t are here expressed in dBW, and the transmit/receive antenna gain G_{tr} in dBi. The ratio of received signal power to noise can be shown by subtracting from (5) the thermal noise power (in dBW). Hence the signal-to-noise ratio is shown as a function of the reflector length in Fig. 2, where a noise temperature of 290 K and noise bandwidths of 1, 10 and 100 MHz have been used by way of illustration. Element separation *a* is 0.5 λ and the element directivity is 3 dBi, which is typical of a single printed antenna.

In Fig. 2, we have used $G_t = G_r = 35$ dBi and $P_t = 10$ W. To some extent the figures are arbitrary, but they have been chosen to be significantly less than for a typical satellite earth hub where we may have $G_t = G_r = 45$ dBi and $P_t = 50$ W [21]. The distance to the reflect-array is 30 km (equating to a ground distance of 22 km and a HAP height of 20 km). The derivation in Fig. 2 is straightforward and intended only to initiate a first order estimate of the required reflector dimensions–loss terms have not yet been included. The actual efficiency of the modulated reflect-array would be rather less than unity and dependent on the carrier frequency and means of construction, as discussed below. The dimensions will also be constrained by the payload capability of the platform. For example, the European *HeliNet* [5], [8], [10] solar powered HAP proposes an aperture area of between 1 and 2 m^2 for its broadband payload, while it is anticipated that future airship-based platforms would support several square meters of aperture.

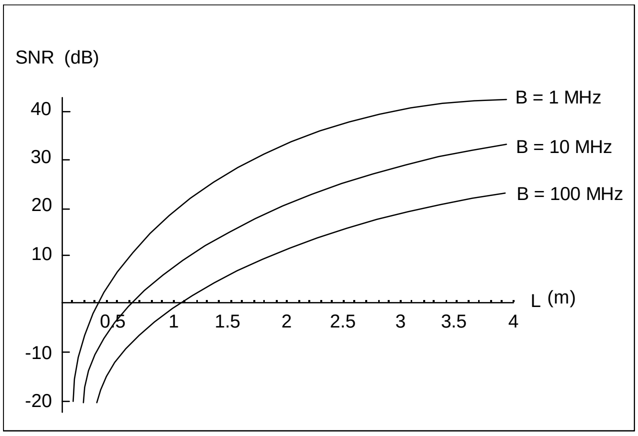

Fig. 2. Received SNR from square, lossless reflect-array of side length *L* at distance of 30 km, for ground EIRP = 45 dBW.

IV. LOSSES AS A FUNCTION OF ARRAY SIZE

Since the RCS of an aperture is proportional to the square of the ratio of area to wavelength [22], we can rewrite (2) in terms of the antenna element effective area a_{el} and the transmission line efficiency η, and we have

$$\sigma = \frac{4\pi}{\lambda^2} \eta \left(\sum a_{el} \right)^2 , \tag{6}$$

where the summation is over all elements. For a Van Atta array, the equal line lengths imply that η is a single parameter which is the product of line length and loss per unit length–in this case the derivation is trivial. However, for planar realizations of the retro-array, the equal line length condition is inconvenient since it is difficult to physically route the lines on the circuit board. An alternative geometry is the unequal-line-length retro-array, where line lengths are allowed to differ by an integer multiple of wavelengths. Indeed, the circuits reported in [17] and [18] have employed this principle. In this case, since phase-conjugation of the reflected signal occurs at a single frequency, the structure is a narrow-band variant [23] of the Van Atta array. A fortunate consequence of this approach is that transmission line lengths and associated losses are now minimized and the array is hence more efficient. To calculate the efficiency, the effective area contributed by each antenna element pair must be weighted by the corresponding efficiency of their interconnecting line, and the overall RCS calculated from the summation of these weighted values. Models for estimating minimum line lengths are presented for two scenarios: crossing lines and non-crossing lines.

Non-crossing geometry

Figure 3 shows the innermost two concentric antenna groups within a square array of $n\times n$ antenna elements spaced by a. In this case the transmission lines are a planar type such as microstrip and not allowed to cross. This type of construction is very convenient for integrating with both the modulation circuitry and printed antennas, such as discussed in [17] and [18] for a 16-element scale model.

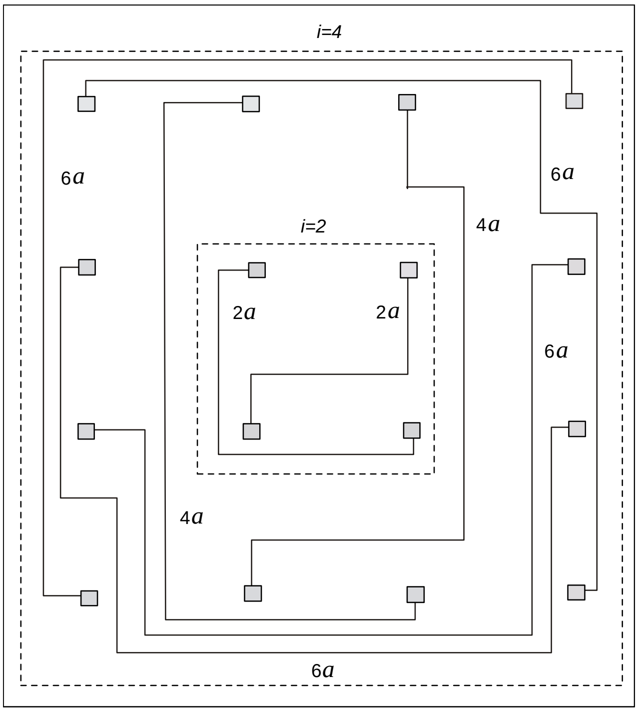

Fig. 3. Non-crossing lines: model for minimum line lengths.

The interconnecting lines as drawn do not represent a practical circuit, but illustrate a model for deriving *minimum* line lengths. Thus, the inner group i=2 has 2 lines of length 2a, and the next group i=4 has 4 lines of length 6a and 2 lines of length 4a. Using i as a counter for each concentric group, there are i lines of length $(2i–2)\,a$ and $(i–2)$ lines of length $(2i–4)\,a$, where $i = 2,4,6,\ldots,n$. Modifying (6) we have

$$\sigma = \frac{4\pi}{\lambda^2}\left(\sum\nolimits_j^{N/2} 2a_{el}\sqrt{\eta j}\right)^2, \qquad (7)$$

where each antenna pair j is considered and hence the summation is to halve the total number of elements. We can further extend this argument from an antenna pair to an antenna concentric group i, where the effective area contributed by the group is found by weighting with the appropriate transmission line loss, which is linearly related to the line length. Hence,

$$\begin{aligned} Ai = {} & 2a_{el}\, i\sqrt{10^{-\frac{(2i-2)al_{pm}}{10}}} \\ & + 2a_{el}(i-2)\sqrt{10^{-\frac{(2i-4)al_{pm}}{10}}} \end{aligned} \qquad (8)$$

is the effective area contributed by the i-th group after losses have been considered, where l_{pm} is the loss factor in the transmission line, in dB/m. The RCS for the array can now be calculated by summing over all concentric groups i, hence,

$$\sigma = \frac{4\pi}{\lambda^2}\left(\sum_{\substack{i=2\\ step2}}^{n} Ai\right)^2, \qquad (9)$$

which can be evaluated by selecting a value for l_{pm} and the array size n. The former is a function of the frequency, as well as the transmission line type and material loss, and is often accurately derived or measured for a range of media such as microstrip or waveguide. To illustrate the effect of increasing loss, (9) has been evaluated as a function of array length L, where $n = L/a$ and $a = 0.5\ \lambda$ at a 2 GHz carrier frequency, as shown in Fig. 4.

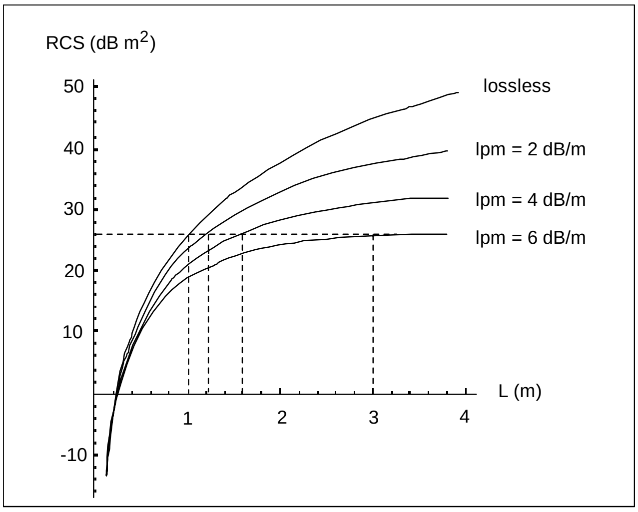

Fig. 4. Maximum RCS at 2 GHz for $a = 0.5\ \lambda$ for non-crossing line routing.

In Fig. 4, the effect of increasing transmission line loss is apparent as a slowing of the rate of RCS increase with array length. The results are for the minimum line lengths and hence maximum RCS and so represent an upper bound on RCS for planar, non-crossing transmission lines. These might typically be microstrip, where loss at 2 GHz might be as low as 2 dB/m. However, where narrower track widths are used to alleviate circuit congestion, losses tend to increase rapidly [24]. The results are useful in showing by how much the dimensions of a lossy retro-directive transponder would need to be increased, compared to the lossless case, to achieve a given link budget. For example, the RCS highlighted in Fig. 4 is given by a lossless array of length 1m, which needs to be increased to at least 1.2, 1.6 or 3.0 m for line losses, respectively, of 2, 4, or 6 dB/m.

Crossing geometry

If transmission lines are allowed to cross, e.g., by using coaxial lines, an alternative derivation of minimum line lengths can be found. It is stressed that the line lengths must still be adjusted so that they differ by an integer number of wavelengths, so that phase conjugation occurs at the design frequency; thus these models for minimum lengths represent an optimistic case. In a similar approach to the above, where each concentric antenna group i is considered, there are 2 lines between the diagonals of length

$$l_{diag} = a\sqrt{2(i-1)^2} \tag{10}$$

and from the symmetry of the square matrix, there are groups i, j of 4 lines of length

$$l_{ij} = a\sqrt{(i-1)^2 + \ (i-j)^2}\ , \tag{11}$$

where j takes values from 3 to $i - 1$ in steps of 2. Hence, the RCS for the co-ax fed array can be expressed as

$$\sigma = \frac{4pi}{\lambda^2}\left(\sum_{\substack{i=2\\ step2}}^{n}\left[4a_{el}\left(10^{a\sqrt{2(i-1)^2}\cdot\frac{lpm}{10}}\right)^{\frac{1}{2}} + \sum_{\substack{j=3\\ step2}}^{i-1} 8a_{el}\left(10^{a\sqrt{(i-1)^2+(i-j)^2}\cdot\frac{lpm}{10}}\right)^{\frac{1}{2}}\right]\right)^2 . \tag{12}$$

Evaluation of (12) is shown in Fig. 5 as a function of array length and compared with the results for the non-crossing geometry.

In Fig. 5, the reduced path lengths in the non-crossing geometry lead to a quantifiable increase in array RCS for a given loss factor. In practice, a planar medium such as microstrip tends to exhibit more loss than coaxial transmission line, which would lead to a further advantage in terms of efficiency in the latter case. There are clearly many variable parameters, not least including the RF frequency and the chosen means of construction, but the above method is intended as a first estimate of the severity of transmission line loss where passive transponders of high RCS are sought.

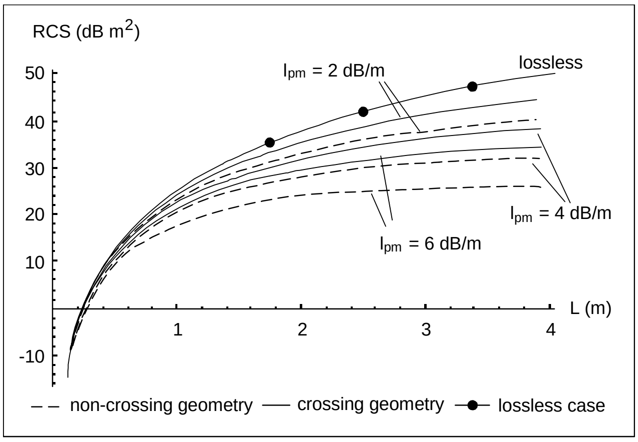

Fig. 5. Comparison of array RCS for crossing and non-crossing geometries at 2 GHz.

Link budget for lossy reflectors

Finally, the effect of the transmission line loss on reducing the predicted SNR values shown in Fig. 2 can be derived. Unlike the derivation of Fig. 2, a carrier wavelength must be chosen, since this scales the lengths of the lines which give rise to the loss. The SNR value will be reduced by the difference between the lossless aperture and the lossy reflector, as illustrated in Figs. 4 and 5, for a given line loss l_{pm}. Choosing again a frequency of 2 GHz, the modified SNR values for a 10 MHz signal bandwidth are plotted in Fig. 6 for nominal l_{pm} values of 2 dB/m and 4 dB/m for both the crossing and non-crossing geometries.

The results in Fig. 6 will of course scale linearly on changing the EIRP or bandwidth, etc. Where line loss is severe, the link budget improves very slowly above a certain aperture length

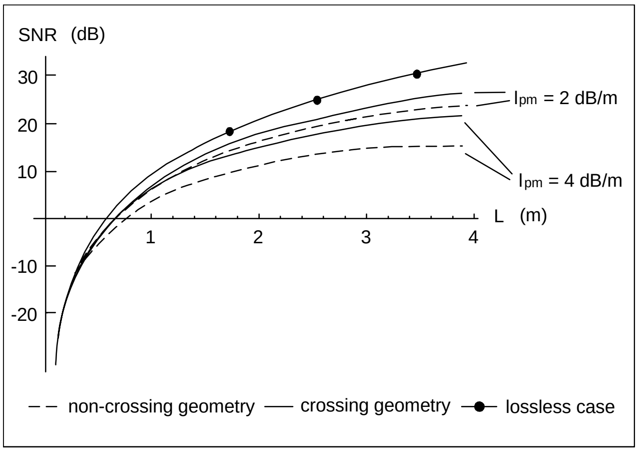

Fig. 6. Received SNR from square, lossy reflect-arrays of side length L at distance of 30 km, for ground EIRP = 45 dBW and 10 MHz noise bandwidth at 2 GHz carrier.

(e.g., 3 m in Fig. 6) to the extent that the increase in fabrication complexity may be considered pointless. It is hence very important to choose materials which minimize transmission line loss.

V. CONCLUSIONS

The novel type of link described in this paper is different from the traditional view of a High Altitude Platform downlink implementation: the conventional requirement for antenna alignment and power amplification is removed by using a retro-directive array with a sufficiently large aperture, which acts as a passive transponder. The emphasis on communications infrastructure is placed on the ground station, where power and weight constraints are much more relaxed than on the platform. We summarize the properties of the link as follows:

***Advantages*:**

- The HAP transponder requires no RF power amplifiers.
- The HAP transponder need not be steered or stabilized.
- The majority of the RF infrastructure is placed at the ground station.
- The RF power can be increased during adverse weather.
- The ground station power amplifiers can be driven into saturation without intermodulation distortion occurring.

***Disadvantages*:**

- The ground station requires higher RF power than a conventional point-to-point link; hence it is relatively expensive.
- The transponder is difficult to make efficient and large at higher frequencies.
- The atmospheric loss is squared compared to a one-way link.

The economic viability of the HAP concept for delivery of telecommunications remains to be proven. Limitations of solar power sources and uncertainties over stability and station keeping present obstacles to communications systems designers. The proposed technique could substantially mitigate these limitations, at least in securing a downlink to dedicated ground stations for critical data such as command and control, or a backhaul for services such as news gathering. To maximize the link budget, increasing transponder aperture size is required, and estimates for the aperture efficiency have been presented. These use models for transmission line minimum length so as to place an upper limit on efficiency for both planar construction techniques and transmission lines such as co-axial cables which are allowed to cross. Losses in the latter case are substantially reduced, particularly for large apertures.

REFERENCES

[1] G.M. Djuknic, J. Freidenfelds, and Y. Okunev, “Establishing Wireless Communications Services via High-Altitude Aeronautical Platforms: A Concept Whose Time Has Come?,” *IEEE Comm. Mag.*, Sept. 1997, pp. 128-135.

[2] T.C. Tozer and D. Grace, “High-Altitude Platforms for Wireless Communications,” *IEE Electronics and Comm. Eng. J.*, June 2001, pp. 127-137.

[3] SkyStation, see: www.skystation.com.

[4] Airship Technologies Group, see: www.atgplc.com.

[5] HeliNet Project, see: www.helinet.polito.it.

[6] Aerovironment, see: www.aerovironment.com.

[7] N.J. Colella, J.N. Martin, and I.F. Akyildiz, “The HALO Network,” *IEEE Comm. Mag.*, June 2000, pp. 142-148.

[8] D. Grace, J. Thornton, T. Konefal, C. Spillard, and T.C. Tozer, “Broadband Communications from High Altitude Platforms – The HeliNet Solution,” *invited paper, WPMC01*, Aalborg, Denmark, Sept. 2001.

[9] T. Konefal, P.A. Watson, A.K. Shukla, and A. Akram, “Prediction of Monthly and Annual Availabilities on 10-50 GHz Satellite-Earth and Aircraft-to-Aircraft Links,” *IEE Proc., Microwaves, Antennas and Propagation*, vol. 147, no. 2, Apr. 2000, pp. 122-127.

[10] J. Thornton, D. Grace, C. Spillard, T. Konefal, and T. Tozer, “Broadband Communications from a High Altitude Platform– The European *HeliNet* Programme,” *IEE Electronics and Comm. Eng. J.*, June 2001.

[11] J.L. Ryerson, “Passive Satellite Communication,” *Proc. IRE*, vol. 48, no. 6, 1960, pp. 613-619.

[12] R.C. Hansen, “Communications Satellites Using Arrays,” *Proc. IRE*, vol. 49, no. 6, 1961, pp. 1066-1074 (also correction pp. 1340-1341).

[13] G.N. Krassner and J.V. Michaels, *Introduction to Space Communication Systems*, McGraw-Hill, 1964, pp. 25-28, 149-150, and 239.

[14] R.C. Hansen, *Microwave Scanning Antennas Vol 3-Self-Phased Arrays*, Academic Press, 1966, pp. 371-372.

[15] R.J. King, *Microwave Homodyne Systems*, Peter Peregrinus, 1978, pp. 266-272.

[16] E.D. Sharp and M.A. Diab “Van Atta Reflector Array,” *IRE Trans. on Antennas and Propagation*, vol. AP-8, July, 1960, pp. 436-438.

[17] J. Thornton and D.J. Edwards, “Range Measurement Using Modulated Retro-Reflectors in FM Radar System,” *IEEE Microwave and Guided Wave Lett.*, vol. 10, no. 9, Sept. 2000, pp. 380-382.

[18] J. Thornton and D.J. Edwards, “Modulating Retroreflector as a Passive Radar Transponder,” *Electronics Lett.*, vol. 34, no. 19, Sept. 1998, p. 1880.

[19] E.L. Gruenberg and C.M. Johnson, “Quasi-Passive Satellite Relay Communications System,” *IEEE Trans. Antennas and Propagation*, vol. 12, 1964, pp. 215-223.

[20] J.J. Schlesak, A. Alden, and T. Ohno, “A Microwave Powered High Altitude Platform,” *IEEE MTT-S Int’l Microwave Symp*, New York, May 1988.

[21] *Earthstation Technology Handbook*, 5th revision, June 1999,

INTELSAT Application Support and Training, 3400 International Drive, NW, Washington DC, USA.

[22] C.A. Balanis, *Advanced Engineering Electromagnetics*, Wiley, 1989, p. 594.

[23] U.K. Patent Application Nos. 9907117.7 and 9912840.7 and International Application No. PCT/GB00/01136, “Transponder.”

[24] T. Edwards, *Foundations for Microstrip Circuit Design*, Wiley, 1991.

John Thornton is a physics graduate of the University of York, UK, and also obtained an MSc. in microwave physics from the University of Portsmouth, UK, in 1995. He has previously held research positions at the Rutherford Appleton Laboratory, UK, developing sub-millimeter wave solid state sources and receivers. From 1997 he worked at the University of Oxford’s Department of Engineering Science, on projects including passive radar transponders, array antennas and superconducting filters. In 2000 he returned to York to join the Department of Electronics and is part of a team studying future communications applications including stratospheric platforms. He maintains an interest in self-phasing array antennas and is in the final stages of completing a Doctoral thesis with the UK’s Open University (Materials Engineering Department).