Using Array Signal Processing to Improve Rural Area Coverage in Future Personal Satellite Communications Services

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Abstract[2]

We address wide-area coverage for hand-held portables using an on-board signal processing satellite in geostationary (GEO) orbit at minimum payload cost. A GEO link, being power-limited, differs from terrestrial mobile communications systems that use direct radiating antennas by requiring a focal-fed parabolic reflector, which acts as a fixed analog beamformer. We propose to cascade a high-gain parabolic reflector and a digital maximum signal-to-interference-plus-noise ratio (SINR) multi-beamformer for spatial multi-user interference suppression to achieve coverage extension. This involves on-board antenna array signal processing of multiple feeds. It is shown that nearly uniform coverage of an area the size of Canada can be provided using only five parbolic reflector apertures. Compared to conventional focussed multibeam system, link margins are typically increased from a minimum requirement of 5.25 dB to 7.5 dB in border areas. The more uniform coverage reduces dynamic power control requirements.

1 Introduction

Narrow-band adaptive antenna arrays have been shown to potentially increase capacity of terrestrial mobile communications systems employing code division multiple access (CDMA) [1]. In this letter, we apply multi-beam adaptive array processing to parabolic reflector antenna feeds to improve coverage in mobile satellite communications employing hand-held portable terminals. Although GEO-orbit satellites are able to carry a large signal processing payload, their large 40,000 km orbital radius requires sophisticated link design [2]. For example, in [3] North American continental coverage is achieved with a single GEO satellite employing 4-6 spatially isolated beams formed from apertures located in the *feed plane* of a parabolic reflector antenna. Unfortunately, allocating spatially non-overlapping beams results in non-uniform coverage, which makes it difficult to support hand-held terminals.

Coverage can be improved by adding more feed elements to the parabolic reflector. However, the increase in the number of up/down converters significantly increases the satellite payload and antenna complexity. We consider only the uplink (handset-to-satellite) since system performance is limited by low-power hand-held mobile transmitters that must reach a GEO-orbit satellite.

In the following, we show that coverage can be improved over conventional isolated-beam systems by introducing adaptive signal array processing to a parabolic multi-feed reflector antenna. Unlike terrestrial systems, the environment here is noise-limited, which precludes the use of simple antennas; in addition to interference suppression capability, plane wave focusing is required for sufficient signal gain. We calculate the received complex electromagnetic signal received from the feed plane of the parabolic reflector antenna. As in [1], CDMA with PN-code filtering is used to estimate each mobile's array response. However, the approach to beamforming is extended from [1] to the case where (1) the link is noise-limited and (2) users are not power controlled prior to beamforming. Rather than capacity enhancement that has been investigated elsewhere, beamforming is applied to the problem of obtaining maximum wide-area coverage.

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2 Uplink Communication Receiver

2.1 Antenna Design and Beam Coverage

Unlike terrestrial systems that assume simple antenna elements, a parabolic reflector is used, which is a distributed aperture with focusing, i.e., analog beamforming has already occurred prior to downconversion and digitization. Let D_1 be the reflector diameter, F the focal length of the paraboloid. When a feed is displaced slightly from the reflector focus, the effective aperture distribution experiences phase variation, which in turn gives rise to a small beam pattern distortion. The feeds themselves may be rotationally symmentric horn or microstrip antennas.

Feed placements and their respective complex electric field radiation pattern can be determined using [4]. As an example, suppose that coverage of major urban centers in Canada is to be provided using five spot beams, where the 3 dB below maximum gain contours are shown in Figure 1. The number of feeds is limited by the available physical area for the apertures near enough to the focal point to capture a significant portion of the received power.

To take account of the parabolic reflector, we must calculate the received complex electric field received at a feed from a user located on the Earth via [4]

$$E(\theta_i, \phi_i) = \int_0^1 \int_0^{2\pi} sE_f(s, \zeta) \exp\left[-j\frac{\pi}{\lambda} D_1 s\mu \cos(\zeta - \varrho)\right] \, ds d\zeta \tag{1}$$

where θ_i, ϕ_i are the incident angles upon the antenna with the z-axis and x-axis, respectively, $\varrho = \arctan\frac{v-v_m}{u-u_m}$, $\mu = \sqrt{(u-u_m)^2 + (v-v_m)^2}$, $u = \sin\theta_i \cos\phi_i$ and $v = \sin\theta_i \sin\phi_i$ are the xy coordinates. The calculation of the effective aperture distribution, $E_f(s, \zeta)$, can be found in [5].

The gain in the direction of maximum radiation, or beam center, is given by $G_{max} = \left(\frac{4\pi}{\lambda^2}\right) A\eta$ where λ is the wavelength, A is the aperture area, and η is the aperture efficiency. Once the electric field for a given feed has been determined, its gain pattern is

$$G(\theta_i, \phi_i) = G_{max} |E_n(\theta_i, \phi_i)|^2 \tag{2}$$

where E_n is $E(\theta_i, \phi_i)$ normalized to unity at the beam center.

2.2 CDMA Bit Stream

In the following, we adopt an up-link bit-stream organization based on [6]: the information bits are convolutionally encoded using a rate 1/3 code of constraint length 9. The data symbols are then multiplied by a user-designated long code (4.9152 Mhz), and the signal is then binary phase-shift keying (BPSK) modulated. The users are assumed to be transmitting asynchronously. In our conventional five-beam arrangement used for comparison purposes, each feed projects one spot beam and supports users within its own 3 dB beamwidth. The received signal $r(t)$ can be expressed as

$$r(t) = \sqrt{2P_1} b_1(t) c_1(t) + \sum_{i=2}^{M} \sqrt{2P_i} b_i c_i(t - \tau_i) e^{j\theta_i} + n(t) \tag{3}$$

where M is the number of users, and $b_i(t)$ and $c_i(t)$ are, respectively, the data symbol and the long code for the i^{th} user. The time and phase delays for the i^{th} user are given by τ_i and θ_i, respectively. Power received from the i^{th} user by the feed allocated to the signal's coverage area is given by P_i, and $n(t)$ is white Gaussian noise with zero mean and a spectral density of N_o.

Without loss of generality, $\tau_1 = \theta_1 = 0$ for user $i = 1$ if we assume perfect self-synchronization. Let T_s denote the data symbol period. For simplicity, $b_i(t)$ and $c_i(t)$ are rectangular pulses with bit and chip durations of T_s and T_c, respectively, where $T_s \gg T_c$. All user signals are assumed to be mutually uncorrelated and uncorrelated with $n(t)$. The long codes have a period of $2^{42} - 1$ chips [6], with negligible cross-correlation. The received despread symbol $z_1(t = T_s)$ can be written as

$$z_1(T_s) = Re\left\{\frac{1}{T_s}\int_0^{T_s} \sqrt{2P_1}b_1(t)\,dt + MAI + N(T_s)\right\} \tag{4}$$

where the multiple access interference (MAI) and noise are, respectively,

$$MAI = \frac{1}{T_s}\sum_{i=2}^{M}\sqrt{2P_i}\int_0^{T_s} b_i(t-\tau_i)c_i(t-\tau_i)c_1(t)e^{j\theta_i}\,dt \tag{5}$$

$$N(T_s) = \frac{1}{T_s}\int_0^{T_s} n(t)c_1(t)\,dt \tag{6}$$

The desired signal term in (4) reduces to $\sqrt{2P_1}b_1(T_s)$ and the signal power P_s is given by $2P_1$. Since the number of users is large, we approximate the MAI term in (4) by a Gaussian distribution. The real part of MAI is zero-mean, since data symbols and the spreading codes are independent of one another and $E(b_i(t)) = 0$. Assuming BPSK modulation, mutually independent user signals as well as CDMA correlation results of [7], it can be shown that the MAI and noise terms can be simplified to $E(Re\{MAI\}^2) = \frac{2T_c}{3T_s}\sum_{i=2}^{M} P_i$, and $E(Re\{N(T_s)\}^2) = \frac{1}{2T_s^2}\int_0^{T_s} 2N_o c_1^2(t)\,dt = \frac{N_o}{T_s}$ The signal-to-interference-noise ratio (SINR) for this receiver is therefore given by

$$SINR = \frac{2P_1T_s}{N_o + \frac{2(\mathrm{VA})}{3W_s}\sum_{i=2}^{M} P_i} \tag{7}$$

where $W_s \equiv 1/T_c$ is the CDMA system bandwidth, and VA represents the voice activity factor, which is typically 0.35 to 0.40 [8].

2.3 Satellite Link Performance Analysis

Determing the SINR requires that the power received from all users and feeds be calculated. The MAI term can be approximated by its average value according to the following monte-carlo procedure: M users are randomly distributed across the coverage area. With the knowledge of the coordinates of each user, (7) is calculated as the sample average of 40 runs, each with a different random seed. The performance of thirty-six representative locations for the desired user was chosen for detailed analysis. The total number of users in the system was chosen to be 625, i.e., 125 users per beam to ensure that the minimum required link margin is maintained for all users within any of the five beams [9]. The received carrier power P_1 from the user at feed (n) is $P_1^{(n)} = \frac{P_uG_uG^{(n)}}{L_{fs}L_{at}}$ where P_u and G_u are the transmitted power and the antenna handset gain of the terrestrial user, respectively. The satellite antenna gain for the incident signal, $G^{(n)}$, is calculated as the n^{th} component of (2), and L_{at} is the atmospheric loss. The free space loss is given by $L_{fs} = (4\pi d/\lambda)^2$ where d is the distance between the user and the satellite. All handsets are assumed to be equipped with stub antennas of 3 dBi gain [2] and transmitting at 600 mW. The receiver noise spectral density $N_\circ$ is given by $N_\circ = KT_i$ (W/Hz) where K is the Boltzman constant, and T_i is the on-board satellite with noise temperature 500° K.

The receivers implement soft-decision Viterbi decoding for a rate 1/3 code of constraint length 9. In a white Gaussian noise channel, to ensure a maximum probability of error of 10^{-3}, a minimum received $E_b/N_\circ$ of 2.1 dB for a coherent receiver is required [10]. This translates to a minimum required SINR at the receiver of 0.34 dB, where the 2/3 factor in the denominator of (7) is taken into account. Through detailed analysis and simulation of the receiver, it can be shown that the extra loss in SINR due to a Rician fading channel with K=10 dB is 1.60 dB. Allowing an extra safety margin of at least 3.3 dB, the overall minimum required SINR at the receiver becomes 5.25 dB, which shows that for a maximum of 625 active users in the coverage area (roughly 125 users per beam), all users within any of the five beams will have an SINR equal or above this minimum requirement. If the number of active users exceeds 625, some users on the beam borders will have to operate at a SINR below 5.25 dB.

3 Satellite Beamforming

In contrast to simple antennas, the phase induced by an incident signal on the feed of a parabolic reflector antenna does not correspond to direction of arrival. At the feed placement corresponding to the antenna focal point, the satellite receives a real-valued electric field. At other feed locations, a complex-valued electric field is received. Due to GEO satellite geometry, the imaginary part of the signal is much smaller than the real part but increases away from the focus. Improvement in discrimination based on complex array response is therefore offset by a loss in peak focusing gain, a phenomenon that has no counterpart in the case of simple antennas.

In the beamforming receiver, we incorporate all antenna feed inputs into an L-element signal vector, which can be written as:

$$x(t) = \sum_{i=1}^{M} \sqrt{2P_i} c_i(t-\tau_i) b_i(t-\tau_i) e^{j\theta_i} A_i + n_v(t) \tag{8}$$

where $\sqrt{P_i} = \sqrt{P_i^{(1)} + P_i^{(2)} + ... + P_i^{(L)}}$ is the power received from the i-th user and $A_i = (\frac{\sqrt{P_i^{(1)}} A_i^{(1)}}{\sqrt{P_i}}, \frac{\sqrt{P_i^{(2)}} A_i^{(2)}}{\sqrt{P_i}}, ..., \frac{\sqrt{P_i^{(L)}} A_i^{(L)}}{\sqrt{P_i}})^T$ is the $L \times 1$ complex unit *array response vector*. The thermal noise is expressed by vector $n_v(t)$ with covariance matrix $E(n_v(t)n_v^\dagger(t)) = 2N_o I_L$, where N_o is the noise spectral density, I_L is the $L \times L$ identity matrix, and † refers to conjugate-transpose. The received signal notation is now in a format that corresponds to that of more studied terrestrial antenna arrays. For this vector case, the SINR expression (7) generalizes to

$$SINR = \frac{2P_1 T_s |w_1^\dagger A_1|^2}{(w_1^\dagger w_1) N_o + \frac{2(\mathrm{VA})}{3W_s} \sum_{i=2}^{M} P_i |w_1^\dagger A_i|^2} \tag{9}$$

3.1 Determination and Estimation of Optimum Weights

The input vector to reference user no. 1's beamforming network $v_1 = (v_{11}, v_{12}, ..., v_{1L})^T$ is given by

$$\begin{aligned} v_1 &= \sqrt{2P_1}b_1(T_s)A_1 + \sum_{i=2}^{M}\frac{1}{T_s}\int_0^{T_s}\sqrt{2P_i}b_i(t-\tau_i)c_i(t-\tau_i)c_1(t)e^{-j\theta_i}A_i dt + \frac{1}{T_s}\int_0^{T_s} n_v(t)c_1(t)dt \quad (10) \\ &= m_1A_1 + (NI)_1 \end{aligned}$$

where $m_1 = \sqrt{2P_1}b_1(T_s)$, A_1 is user no. 1's array response vector and $(NI)_1$ represents the MAI and thermal noise terms. The MAI can be approximated by a multivariate Gaussian distribution with zero mean and covariance matrix R_{NI}, the beamformer output can be written as $y_1 = w_1^\dagger m_1 A_1 + w_1^\dagger (NI)_1$ and the SINR is given by

$$SINR = \frac{E|w_1^\dagger m_1 A_1|^2}{E|w_1^\dagger (NI)_1|^2} \quad (11)$$

It is well-known that (11) is maximized for $w_1 = (R_{NI})^{-1}A_1$. To estimate A_1 and $(NI)_1$, we modify the method of [11] to account for the lack of power control, and to account for the fact that the received power at each feed from each user is different. In our case, the pre-despreading covariance matrix R_{xx}, assuming statistically independent data symbols, PN chips and noise samples is

$$R_{xx} = 2(P_1A_1A_1^\dagger + \sum_{i=2}^{M} P_iA_iA_i^\dagger + N_oI_L) \quad (12)$$

where I_L represents an L by L identity matrix. We remark that (12) differs from [11] Eq.(7), mainly due to the presence of the unknown constants $P_1, P_2, \ldots P_M$, which affect the estimation of the MAI covariance matrix.

It can be shown that the post-despreading covariance matrix $R_{v_1v_1}$, for desired user no. 1 can be reduced to

$$R_{v_1v_1} = 2(P_1A_1A_1^\dagger + \frac{2T_c}{3T_s}\sum_{i=2}^{M} P_iA_iA_i^\dagger + \frac{1}{T_s}I_LN_o) \quad (13)$$

Assuming sufficient data to obtain accurate covariance estimates $\hat{R}_{xx}$ and $\hat{R}_{v_1v_1}$, of R_{xx} and $R_{v_1v_1}$, respectively, (12) and (13) can be combined into a matrix H to eliminate the MAI:

$$H \equiv T_s\hat{R}_{v_1v_1} - \frac{2}{3}T_c\hat{R}_{xx} = 2P_1A_1A_1^\dagger(T_s - \frac{2}{3}T_c) + 2N_o(1 - \frac{2}{3}T_c)I_L \quad (14)$$

The right-hand side of Eq. (14) is of the form of a rank-one matrix in additive white noise, and A_1 can be estimated as the largest eigenvalue eigenvector. To maximize (11) the matrix R_{NI} must be known. From Eq. (13), $R_{NI} = R_{v_1v_1} - 2P_1A_1A_1^\dagger$ Since $R_{v_1v_1}$ and A_1 can be estimated, the only unknown term is P_1. If an accurate estimate for N_o were available, P_1 can be solved for using Eq. (14). However, a prior estimate for N_o is not necessary: the estimate H in Eq. (14) is of the form $H = KA_1{A_1}^\dagger + \beta I_L$ where β and K are unknown constants, and we need to estimate $KA_1{A_1}^\dagger$. Performing an eigen-decomposition on H, $H = V\Lambda_H V^\dagger$ where Λ_H is a diagonal matrix with eigenvalues arranged in decreasing order, i.e.,

$$\Lambda_H = \mathrm{diag}(\lambda_{H1}, \lambda_{H2}, \ldots, \lambda_{HL}) = \mathrm{diag}(K, 0, \ldots, 0) + \mathrm{diag}(\beta, \beta, \ldots, \beta)$$

If we form the estimate

$$\hat{K} = \lambda_{H1} - \frac{\sum_{i=2}^{L} \lambda_{Hi}}{L-1} \tag{15}$$

then from Eqs. (14) and (15) R_{NI} can be calculated as

$$R_{NI} = R_{v_1 v_1} - \frac{\hat{K}}{T_s - \frac{2}{3}T_c} A_1 {A_1}^{\dagger} \tag{16}$$

In the above we have averaged the $L-1$ noise subspace eigenvalues of H to form our noise estimate. We note that it is not necessary to calculate all L eigenvalues per user. The noise subspace can be deflated to one dimension as proposed in [12].

3.2 Sub-optimum Receiver

As an alternative to the maximum SINR solution, w_1 can be instead set equal to A_1. In this case, from (9) the SINR becomes

$$SINR = \frac{2P_1T_s}{N_o + \frac{2(\mathrm{VA})}{3W_s} \sum_{i=2}^{M} P_i |A_1^{\dagger} A_i|^2} \tag{17}$$

since $|A_1^{\dagger} A_1| = 1$. This *sub-optimum* beamforming receiver minimizes the thermal noise term in (9), and will nonetheless reduce the MAI and increase the received desired signal power. This corresponds to the optimum beamformer if perfect power control is assumed and users are uniformly distributed around the coverage area.

4 Results

Due to higher population density concentrated in suburban areas, we compared a spatially-uniform user distribution with a non-uniform user distribution where 2/3 of the users were assumed to be concentrated in eight selected major suburban areas. The rest were uniformly distributed in the coverage area. To determine the SINR, the received complex electric field is calculated from all users to all five feeds using (1) and (2). Performance averaged over the selected types of user locations are summarized in Tables 1 and 2, while the thirty-six individual link analyses are presented in [9]. Note that optimum beamforming has produced a sizable improvement in the received SINR even despite the parabolic reflector's beam focusing. In contrast, in the conventional isolated system, users located near or outside the 3 dB beam edge do not meet the required 5.25 dB link SINR. Overall, the improvement varied from 1 dB to 2.4 dB. Typically, the loss in SINR for the suboptimum versus the optimum receiver was small, ranging from 0.1 to 0.5 dB. It is also noteworthy that more uniform coverage across the entire wide area is obtainable using beamforming. From Table 1 it can be seen that the average uplink SINR variation among the three user location classes by adding beamforming is reduced significantly from 3.6 to 1.8 dB

In conclusion, an array signal processing system for enhanced personal satellite communications employing parabolic antenna beamforming has been presented. The improved performance resulted in a reduction of required transmitted terminal power by as much as 2.5 dB, or 55%, in areas of poor coverage. This enables hand-held terminals to communicate at practical power levels of 600 mW throughout the entire coverage area. This approach also has been extended to future Ka-band systems [13]. In this case, the potential satellite payload reduction could be larger due to the smaller beam size.

User Type	Average SINR (dB)		
	No Beamforming	Optimum BF	Suboptimum BF
Inside 3dB contour Mean (Std. Dev.)	8.07 (0.51)	8.37 (0.41)	8.00 (0.46)
On beam edge Mean (Std. Dev.)	6.02 (0.54)	7.31 (0.82)	7.03 (0.87)
Outside 3dB contour Mean (Std. Dev.)	4.43 (0.25)	6.50 (0.85)	6.26 (0.80)

Table 1: Average SINR comparison for the case of uniformly distributed users.

User Type	Average SINR (dB)		
	No Beamforming	Optimum BF	Suboptimum BF
Inside 3dB contour Mean (Std. Dev.)	7.97 (0.57)	8.34 (0.56)	7.90 (0.54)
On beam edge Mean (Std. Dev.)	5.88 (0.63)	7.13 (1.03)	6.82 (1.03)
Outside 3dB contour Mean (Std. Dev.)	4.37 (0.75)	6.43 (1.50)	6.16 (1.33)

Table 2: Average SINR comparison for the case of 2/3 suburban-concentrated users.

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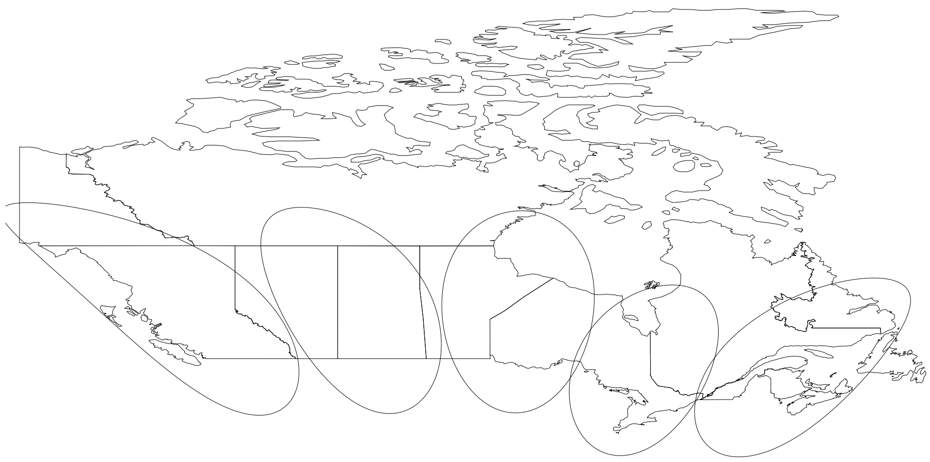

Figure 1: Beam Footprints on the Coverage Area

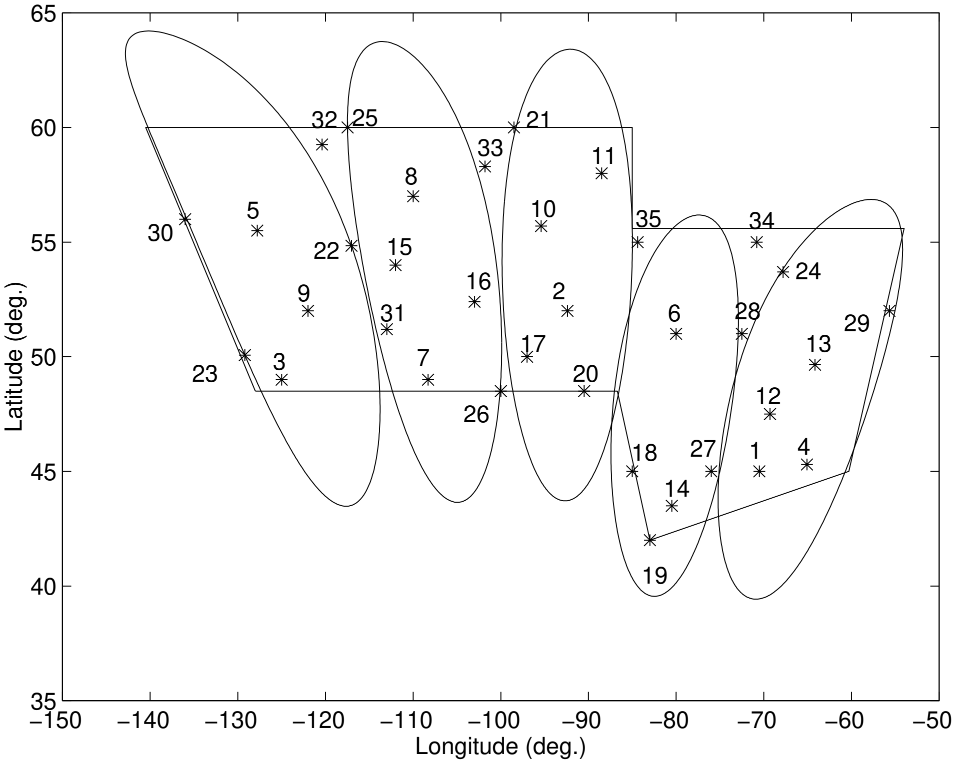

Figure 2: Locations for the Desired User in the Coverage Area

List of Figure Captions

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