Lunar Bread Crumbs: Propagation model, Link budgets and modulation

I. Astronaut to Lander link

A link budget [1] is a formula that relates the carrier-to-noise, Γ, to the communication link parameters. It is a preliminary tool to estimate the performance of any communication system. In order to calculate a link budget we need path loss models, noise power levels, shadow and fade margins etc.

A. Large scale path loss model

Path loss measured in dB , is defined as the difference between effective transmitted power and received power. Path loss prediction

$$L_{dB} = G_{T,dB} + G_{R,dB} + 20\log h_T h_R - 40\log d \tag{1}$$

Total fade loss margin

$$M_{\text{Total}} = M_{\text{AvgFadeDepth}} + M_{CI} \tag{2}$$

Assumptions:

- S-band(2.4GHz)
- base antenna height=5m, astronaut antenna height=2m
- a terrain similar to Metero Crater, Arizona, whoe environmental parameters are similar to the lunar surface
- conductivity and relative dielectric constant valuse are set to: $\sigma = 10^{-4}$ S/m, μ_r = 3.

Fade depth statistics are calculated using the following assumptions [1]:
Average Fade Depth = $6.38\,$dB
Standard Deviation = 5.26 dB

The additional power needed for the desired coverage (CI,confidence interval):

- std. dev. = 68% CI
- std. dev. = 95% CI
- std. dev. = 99% CI

We select a 95% CI in our calculation, the total fade loss margin is calculated to be:

$$\begin{aligned} M_{\text{Total}} &= M_{\text{AvgFadeDepth}} + M_{CI} \\ &= M_{\text{AvgFadeDepth}} + 2 \times std.dev. \\ &= 6.38 + 2 \times 5.26 \\ &= 16.9\,dB \end{aligned} \tag{3}$$

Path loss prediction, when $d = 2000$ m , $G_T = 0$ dB, $G_R = 0$ dB

$$\begin{aligned} L_{dB} &= G_{T,dB} + G_{R,dB} + 20\log h_T h_R - 40\log d \\ &= -112.04\,dB \end{aligned} \tag{4}$$

$$\begin{aligned} L_{dB} - M_{\text{Total}} &= -112.04 - 16.9 \\ &= -128.94\,dB \end{aligned} \tag{5}$$

II. Small scale fading

Due to the presence of scatterers and reflecting objects, signals reaching the receiver will undergo multi-path fading. If there is line-of-sight (LOS) between transmitter and receiver, the fading is called Rician. Otherwise for non line-of-sight (NLOS) communications, the fading is known as Rayleigh fading. We will consider NLOS wave propagation over the lunar surface. The envelope of the NLOS signal has the following Rayleigh distribution. Also shown isits power has exponential distribution, which we use to calculate the cumulative distribution function (CDF) as shown in equation 6.

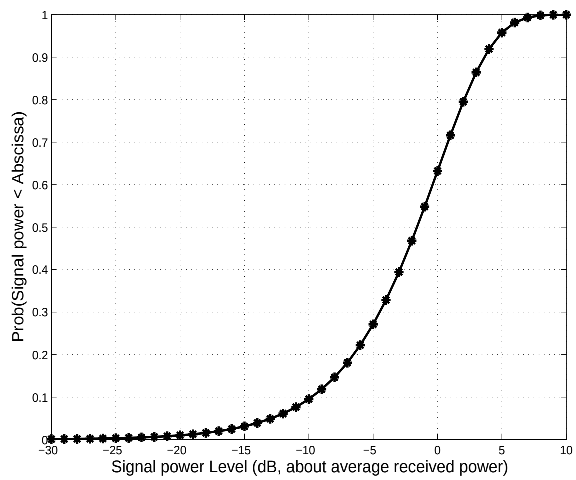

Fig. 1: Theoretical CDF plot of received power

$$
\begin{aligned}
f(y) &= \frac{y}{\sigma^2}\exp\left(-\frac{y}{2\sigma_y^2}\right)(0 \le y \le \infty) \\
&= 0\,(y < 0)
\end{aligned} \tag{6}
$$

Received power at a distance d, $P_R(d)$ has an exponential distribution given below in 7:

$$
f_{P_R}(p) = \frac{1}{P_R(d)}\exp\Big(-p/P_R(d)\Big) \tag{7}
$$

The power delay profile provides the received power as a function of excess delay [2]. We use typical hilly terrain model [1] with total excess delay of $\tau_{\max} = 20\,\mu s$.

A. Cumulative distribution function

The CDF of path loss gives how the actual path loss deviates from the average large-scale path loss. Here we calculate the probability of path loss exceeding a certain threshold. That is given below

$$
\begin{aligned}
p(P_R(d) < P_{\text{thresh}}) &= \int_0^{P_{\text{thresh}}} f_{P_R}(p)dp \\
&= 1 - \exp\left(-\frac{P_{\text{thresh}}}{P_R(d)}\right)
\end{aligned} \tag{8}
$$

CDF of path loss is plotted in Fig. (1) we can see from this plot how small scale fading is severe with respect to the average received power. However CDF is same due to the threshold applied relative to the received power.

III. Link budget

In the link budget calculations we use the measured prototype antenna gains and transmit power levels compatible with OSHA guidelines. The measured prototype antenna which will be used for both the transmitter and reciever is a quarter-wave monopole. It was fabricated using a highly conductive copper plate whose area exceeded the area of the near-field region of the radiating element.. The measured antennahas an average gain under dBi over its azimuth; however, given the uncertainty in the testing set-up (the antenna was mounted at a slant) and the ability to make a new prototype in the next phase, we are confident that a gain of 0dBi can be utilized.

We use transmit power level of 0.3162 W and system noise temperature at Lander as 250 K. Complete link budget for astronaut to lander link is given in Table I. We have a carrier-to-noise ratio of 6.3 dB available at Lander for demodulation purposes.

TABLE I: Link budget for Astronaut to Lander link

Received Power at Lander terminal	
Astronaut transmit power	-5 dBW
Transmit antenna gain	0 dB
Fading margin (FM)	-10 dB
Shadow margin (M_{shad})	-16.9 dB
Maximum path loss ($PL(d = 2000\,\text{m})$)	-112.04 dB
Miscellaneous losses	-3.0 dB
Receiving antenna gain	0 dB
Received power	-141.94 dBW
Lander terminal noise power	
Boltzman's constant	-228.6 dBW/K/Hz
System noise temperature, 250 K	23.98 dBK
Receiver noise Bandwidth, 137.315 kHz	51.38 dBHz
Noise power, N	-153.24
Lander terminal carrier-to-noise ratio (C/N)	6.3 dB

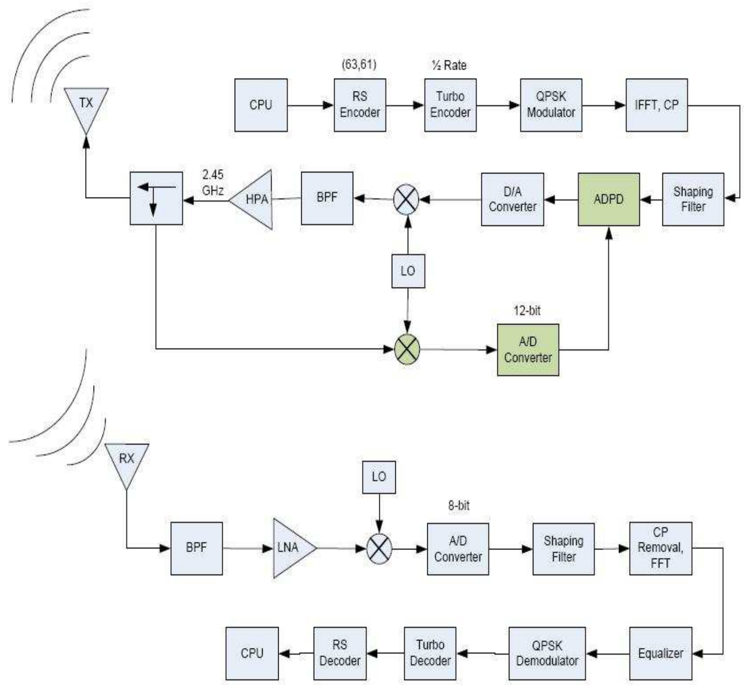

Fig. 2: Communication link between astronaut and lunar lander

IV. Digital modulation and coding

The block diagram of the link between the astronaut and the lunar lander is shown in Fig.(2). Here the binary data is encoded with $(63, 61)$ Reed-solomon (RS) code first followed by turbo coding with with a $R_{\text{c, turbo}} = 1/2$ rate. QPSK ($b = 2$ bits/symbol) is used to map the binary data to symbols and then OFDM is used as block modulation scheme to over come the multi-path channel inter-symbol-interference(ISI). The RS coder in the link can increase bit rate by 20% [3] and turbo code is used as an inner error correction code in the link to achieve good coding gains.

We use RRC filters at the transmitter and receivers to limit the noise bandwidth (B_n) to symbol rate (R_s). We choose RRC filters with a roll off factor $\alpha = 0.45$ which results in a channel bandwidth of $R_s(1 + \alpha) = 1.649$ MHz.

Final Computations: Let N_{cp} be the length of cyclic prefix in the OFDM modulation, then $N_{cp} = \tau_{\max}/T_s \approx 23$. We use 9 subcarriers for pilot data for channel estimation. 16 subcarriers are nulled for adjacent channel isolation and $N = 208$ subcarriers for QPSK data symbol. We have an overhead of 28 samples out of $N_T = 256$ total samples per OFDM symbol.

The bit rate for the link for OFDM-QPSK modulation with turbo coding is

$$\begin{aligned} R_b &= N/N_T \times R_{\text{c,turbo}} \times b \times R_s \\ &= 256/284 \times 1/2 \times 2 \times 1.649\,\text{Mbps} \\ &= 0.9243\,\text{Mbps} \end{aligned} \tag{9}$$

With outer RS coding we get a final bit rate for the astronaut to lander link : $R_b = 1.2 \times 0.9243 = 1.1092\,\text{Mbps}$

V. Backscatter communication with passive RFID

A. Power-Up link budget

Received power at RF tag is given by [4]

$$P_t = \frac{P_T G_T G_t \lambda^2 X p_\tau}{(4\pi r)^2 \Theta B F} \tag{10}$$

where P_T : Power transmitted by the reader, G_T: reader transmit antenna gain, G_t: tag gain, X : polarization mismatch p_τ : power transmission coefficient, r : reader to tag distance Θ: RF tag antenna's on object gain penalty, B: path-blockage loss, and F: is the power up fade margin.

from the above equation we can write the range of tag for $P_T = 0$dBW, $G_T = 0$ dB, $G_t = 0$ dB, $X = 0$ dB , $p_\tau = 0$ dB, $\Theta = 0$ dB, $B = 0$ dBW, $F = 0$ dB and $P_t = -20$ dBm

$$\begin{aligned} r &\leq \frac{\lambda}{4\pi}\sqrt{\frac{P_T G_T G_t X p_\tau}{P_t}} \\ &\leq 2.1\text{m} \end{aligned} \tag{11}$$

Hence based on link budget, the maximum distance between astronaut and tag can't exceed 2.1 m.

B. Co-located backscatter link budget

The received backscatter power P_R [4] is

$$P_R = \frac{P_T G_T^2 G_t^2 \lambda^4 M}{(4\pi r)^4 \Theta^2 F_\alpha} \tag{12}$$

where $r = 2$ m , $F_\alpha = 0$ dB and M is the modulation factor given below.

$$M = \frac{1}{4}|\Gamma_A - \Gamma_B|^2 \tag{13}$$

where reflection coefficient for state A is

$$\Gamma_A = \frac{Z^A_{\text{RFIC}} - Z^*_{\text{ant}}}{Z^A_{\text{RFIC}} + Z^*_{\text{ant}}} \tag{14}$$

where the input impedance for state A and B are

$$\begin{aligned} Z^A_{\text{RFIC}} &= R_L + jX_A \\ &= R_L - \frac{jwL}{w^2LC - 1} \end{aligned} \tag{15}$$

$$\begin{aligned} Z^B_{\text{RFIC}} &= R_L + jX_B \\ &= R_L + jwL \end{aligned} \tag{16}$$

where R_L, X_A, X_B is the load resistance, and reactance for state A and B respectively. We choose inductance $L = 51.6$ nH and capacitance $C = 1.3$ pF. After substituting above values we get $M = 0.0024$ and received power at tag reader $P_R = -123.8$ dBW.

C. Modulation for RFID tag reader

The block diagram of monostatic RFID reader system is shown in Fig. (4). The RFID reader uses differential phase shift keying (PSK) and a unique code sequence on each tag to demodulate the backscattered modulated data. Each RF tag sends its unique code sequence $\{C_n\}$ with negligible cross correlation between different codes, at a period N_s with phase modulated data $b_{\lfloor m/N_c \rfloor} = \cos(wt + \theta_{\lfloor m/N_s \rfloor})$. The phase change needed to send digital signal is achieved by toggling between inductance and capacitance load in the RF tag. During each code sequence of N_s symbols one modulated symbol is transmitted by the RF tag. That is

$$s(t) = C_n \cos(wt + \theta_{\lfloor m/N_s \rfloor}) \tag{17}$$

and the received signal is

$$r(t) = \sum_{l=0}^{L-1} C_n g(t - \tau_l) \cos\left(wt + \theta_{\lfloor m/N_s \rfloor} - \phi_l\right) + \sum_{k=0}^{N_{cl}-1} C_k h(t - \tau_{cl,k}) \cos\left(wt + \theta_{\lfloor k/N_s \rfloor} - \phi_{cl,k}\right) + n(t) \tag{18}$$

where the carrier phase is ϕ, and $g(t)$, $h(t)$ are channel responses of the tag of interest and clutter. We show that use of a DPSK demodulator and spread spectrum will result in a negligible clutter effect. The received signal during each code chip period is multiplied by $\cos(2\pi f_c t)$ and $\sin(2\pi f_c t)$, then integrated over a frame period of T_f. Here T_f is chosen such that all the multi-path and clutter components are received within that period. After integration we can write the received signal as

$$y_I(t) = \sum_{l=0}^{L-1} C_n g(t - \tau_l) \cos\left(\theta_{\lfloor m/N_s \rfloor} - \phi_l\right) + \sum_{k=0}^{N_{\mathrm{cl}}-1} h(t - \tau_{\mathrm{cl},k}) C_k \cos\left(\theta_{\lfloor k/N_s \rfloor} - \phi_{\mathrm{cl},k}\right) + n^{'}(t) \tag{19}$$

$$y_Q(t) = \sum_{l=0}^{L-1} C_n g(t - \tau_l) \sin\left(\theta_{\lfloor m/N_s \rfloor} - \phi_l\right) + \sum_{k=0}^{N_{\mathrm{cl}}-1} h(t - \tau_{\mathrm{cl},k}) C_k \sin\left(\theta_{\lfloor k/N_s \rfloor} - \phi_{\mathrm{cl},k}\right) + n^{''}(t) \tag{20}$$

or alternatively

$$y(t) = \sum_{l=0}^{L-1} C_n g(t - \tau_l) \exp\left\{ j(\theta_{\lfloor m/N_s \rfloor} - \phi_l) \right\} + \sum_{k=0}^{N_{\mathrm{cl}}-1} h(t - \tau_{\mathrm{cl},k}) C_k \exp\left\{ j(\theta_{\lfloor k/N_s \rfloor} - \phi_{\mathrm{cl},k}) \right\} + z(t) \tag{21}$$

Now we multiply $y(t)$ by C_n and and sum over N_s chips and obtain mth sample as

$$y_m = N_s g_m \exp\left\{ j(\theta_m - \hat{\phi}) \right\} + h_m \sum_{k=0}^{N_{\mathrm{cl}}-1} \exp\left\{ (\theta_k - \hat{\phi}) \right\} \sum_{n=1}^{N_s} C_k C_n + z_m \tag{22}$$

where it is assumed that $\phi_i - \phi_j \approx \hat{\phi}, \forall i, j$. As the sequence C_n has negligible cross correlation, the clutter component will be zero. Now we multiply y_m and y_{m-1}^* we get

$$\begin{aligned} y_m y_{m-1}^* = N_s^2 g_m g_{m-1}^* \exp\left\{ j(\theta_m - \theta_{m-1}) \right\} + N_s g_m \exp\left\{ j(\theta_m - \hat{\phi}) \right\} z_{m-1}^* \\ N_s g_{m-1} \exp\left\{ -j(\theta_{m-1} - \hat{\phi}) \right\} z_m + z_m z_{m-1}^* \end{aligned} \tag{23}$$

This is in the familiar form of the DPSK signal metric [5] and its error probability for DPSK demodulation in the tag reader is given below

$$P_b = \frac{1}{2} \exp\left(- \frac{P_R T_S N_S}{N_0} \right) \tag{24}$$

We plot the performance of RF tag reader in Fig. 3) and we can see that for bit rate of 12, we get bit error rate of 10^{-4} at a distance 3.4 m between Astronaut and tag.

REFERENCES

[1] G. L. Stuber, *Principles of mobile communication*. Boston, MA: Kluwer academic, 2001.
[2] T. S. Rappaport, *Wireless communications*. NJ: Prentice Hall, 2002.
[3] Diatta, D. D. Geest, and B. Geller, “Reed solomon turbo codes for high data rate transmission,” in *IEEE VTC’04*, pp. 1023–1027, May 2004.
[4] J. Griffin and G. Durgin, “Complete link budgets for backscatter-radio and rfid systems,” *IEEE Antennas and propagation mag.*, vol. 51, pp. 11–25, Apr. 2009.
[5] J. G. Proakis, *Digital communications*. NY: McGraw Hill, 2000.

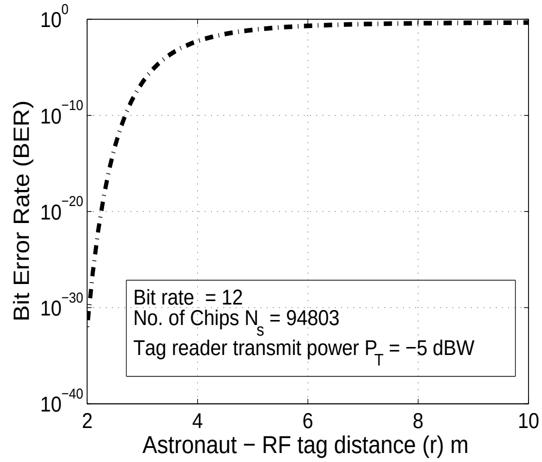

Fig. 3: RF tag reader performance with DPSK demodulation

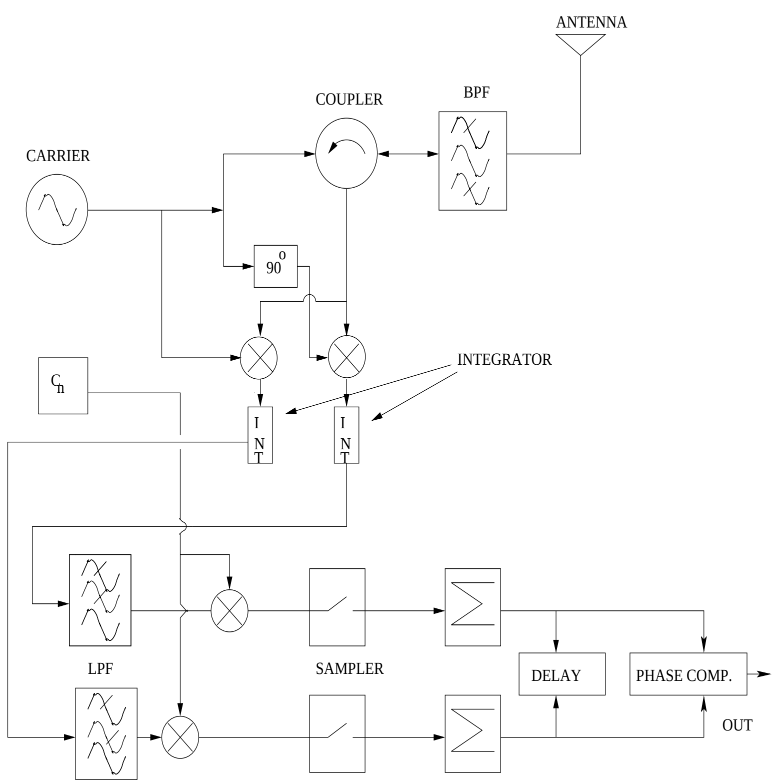

Fig. 4: Block diagram of monostatic RFID reader