APPLICATION NOTE

The Link Budget and Fade Margin

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The Link Budget and Fade Margin

1. Introduction

When planning a long road trip to a remote destination, one of the first considerations is the fuel requirement. One considers the storage capacity and rate of consumption to calculate the fuel required to not only reach the destination, but to also arrive with some level of reserve or margin of safety; accounting for the unforeseeable.

A very similar process is involved in planning an RF (radio frequency) telemetry link. One begins with the output power capacity of the transmitter and sums the system gains and losses to determine the level of power actually delivered to the receiver. To ensure a reliable link, the level of power available to the receiver should be in excess of that required for a minimum level of performance.

An account of all the various gains and losses between the transmitter and the receiver is referred to as the *link budget*. The system factors involved in this accounting are illustrated in FIGURE 1-1.
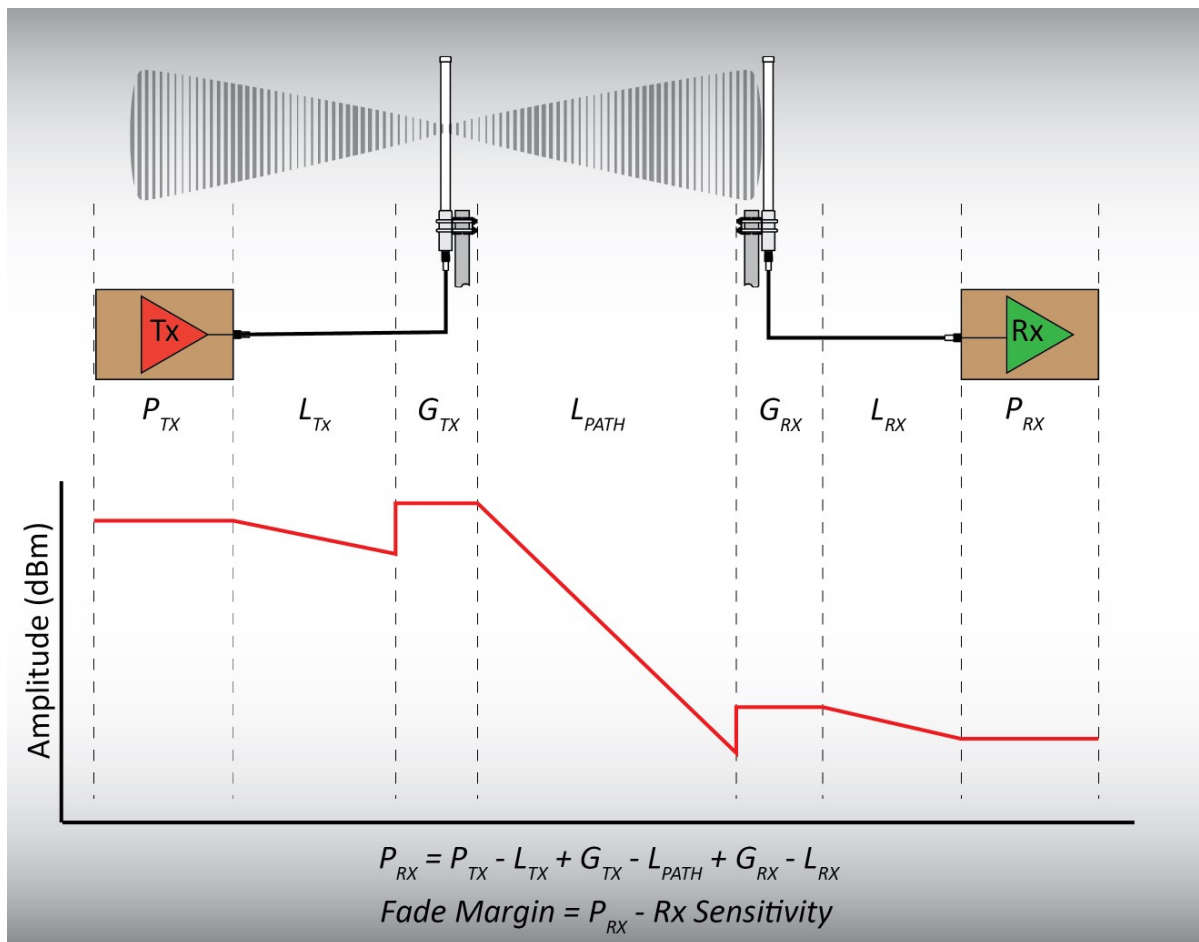

$$P_{RX} = P_{TX} - L_{TX} + G_{TX} - L_{PATH} + G_{RX} - L_{RX}$$

$$Fade\ Margin = P_{RX} - Rx\ Sensitivity$$

FIGURE 1-1. System Gain-Loss Profile for a Link Budget

Where:

P_{TX} = the transmit power in dBm.

L_{TX} = the total system loss in dB at the transmitter.

G_{TX} = the antenna gain in dBi at the transmitter.

L_{PATH} = the total propagation losses in dB between the transmit and receive
  antennas.

G_{RX} = the antenna gain in dBi at the receiver.

L_{RX} = the total system loss in dB at the receiver.

P_{RX} = the receive power in dBm.

The level of received power in excess of that required for a specified minimum
level of system performance is referred to as the *fade margin*. So called,
because it provides a margin of safety in the event of a temporary *attenuation*
or *fading* of the received signal power. The minimum required received power
level used for the link budget can be totally arbitrary—owing to the designer's
knowledge and experience—but is most often tied to the receiver's sensitivity.
Simply put, the receiver's *sensitivity* specifies the minimum RF input power
required to produce a useable output signal. Typical values for receiver
sensitivity fall within the range of –90 to –120 dBm.

Given the following link description:

- RF320 radio (both ends): RF output power = 5 watts (37 dBm);
  sensitivity = 0.25 μV (–119 dBm) for 12 dB SINAD (signal-to-noise-
  and-distortion ratio); operating frequency = 170 MHz.

- Omnidirectional antenna (both ends) is an FG1683: gain = 3 dBd
  (5.15 dBi); VSWR (voltage standing wave ratio) < 2:1.

- The elevation of the transmitting antenna is 30 meters.

- Transmission line at transmitter: pn 31332 Surge Suppression Kit
  (loss ≈ 0.5 dB); 100 ft. of Belden® 9914 coaxial cable (loss ≈ 1.7 dB
  @ 170 MHz); miscellaneous connector loss ≈ 0.5 dB.

- The elevation of the receiving antenna is 10 meters.

- Transmission line at receiver: pn 31332 Surge Suppression Kit (loss
  ≈ 0.5 dB); 50 ft. of Belden 9914 coaxial cable (loss ≈ 0.85 dB @ 170
  MHz); miscellaneous connector loss ≈ 0.5 dB.

- The path of propagation is an unobstructed line of sight (LOS) over a
  smooth earth.

- The distance between the transmitting and receiving antennas is 20
  miles (32.2 km).

The link budget input parameters can now be derived.

2.   Transmit Power

The RF320 RF output power is specified in units of watts. The following equation is used to convert power in watts to power in dBm:

$$P_{dBm} = 10 \log(P_{Watts}) + 30 \qquad\qquad (2\text{-}1)$$

The RF output power of the example radio is 5 watts.

Therefore:

$$P_{dBm} = 10 \log(5 \text{ watts}) + 30$$

$P_{TX} = 37 \text{ dBm}$

3.   System Loss

System loss is the sum of the total insertion loss in the transmission line plus any loss due to an impedance mismatch with the antenna.

Except for the case where an antenna is mated directly to a transceiver's antenna connector, there will likely exist some combination of coaxial cables, surge suppressors, and possibly even bandpass filters used to connect the transceiver to the antenna. Collectively, these devices comprise what is termed the *transmission line*.

Each device in the transmission line that does not produce a signal gain (amplifier) will exhibit some degree of signal loss; a decrease in the signal level at its output relative to its input is known as *insertion loss*. The datasheet for a given device will usually list the insertion loss, and, as with all things "RF", the value will be frequency dependent. In most cases, the dominant loss will be attributed to the insertion loss of the relatively longer run of cable connecting a tower-mounted antenna. The cable manufacturer's datasheet will typically specify cable loss in a table listing attenuation in dB per unit of length versus frequency as shown in TABLE 3-1.

TABLE 3-1. Attenuation Specification for Belden 9914	
Nom. Attenuation:	
Freq. (MHz)	Attenuation (dB/100 ft)
5	0.4
10	0.5
50	1.0
100	1.4
200	1.8
400	2.6
700	3.6
900	4.1
1000	4.4
1500	5.5
1800	6.1
2000	6.5
2500	7.5
3000	8.3
4000	9.9

Interpolating from TABLE 3-1, 100 ft. of 9914 cable will exhibit a loss of approximately 1.7 dB at an operating frequency of 170 Mhz. One should note that this is the loss of the coaxial cable only and does not include the additional loss of any connectors used to construct a finished cable assembly. For the purposes of a link budget, the insertion loss for connectors can generally be ignored, but for the sake of completeness, an acceptable approximation is 0.1 dB per connector, feed-thru, or adapter.

Because maximum power is transferred to the antenna when the output impedance of the transceiver is matched to the input impedances of the transmission line and antenna, a mismatch of impedances will result in a relative loss of radiated power referred to as *mismatch loss*. It is likely that each device in the transmission line will exhibit some small deviation from the standard 50 Ω characteristic impedance, and the net effect is the aggregate of these cascading mismatches. However, for the purposes of a link budget, the small effects of transmission line devices are negligible and the mismatch specification of the antenna can be assumed to be the dominant factor. The most commonly quoted indicator of impedance mismatch on an antenna datasheet is VSWR and is typically specified as some maximum value over the operational bandwidth; for example, VSWR < 2:1. The following equation is used to convert VSWR to mismatch loss (*ML*) in dB (use only the left-hand value of the ratio).

$$ML = -10 \log\left\{ 1 - \left[\frac{VSWR - 1}{VSWR + 1} \right]^2 \right\} \qquad (3\text{-}1)$$

Therefore, a VSWR of 2:1 equates to a mismatch loss of 0.511 db.

Based on the preceding information, we can now calculate the system loss for both ends of the link.

L_{TX} = surge kit (–0.5) + cable (–1.7) + connectors (–0.5) + mismatch (–0.511) ≈ –3.2 dB

L_{RX} = surge kit (–0.5) + cable (–0.85) + connectors (–0.5) + mismatch (–0.511) ≈ –2.35 dB

4.   Antenna Gain

By convention, antenna gain figures used in a link budget are expressed in units of dBi; gain relative to a theoretical isotopic radiator. It is not uncommon for a manufacturer's datasheet to express antenna gain in units of dBd, gain relative to an actual dipole antenna. Relative to an isotropic radiator, a standard half wave vertical dipole antenna will exhibit an intrinsic gain of 2.15 dB in the horizontal. The following equation is used to convert gain in units of dBd to units of dBi:

$$dBi = dBd + 2.15 \qquad\qquad (4\text{-}1)$$

The FG1683 antenna in the example has a specified gain of 3 dBd. Therefore, to calculate the gain in dBi of the transmitting antenna and the receiving antenna:

G_{TX} = G_{RX} = 3 + 2.15

G_{TX} = G_{RX} = 5.15 dBi

5.   Path Loss

In most cases, path loss is the principal contributor to loss in the link budget. It is the sum of free space loss plus additional losses induced by the interaction of the EM (electromagnetic) wavefront with the terrain and/or obstructions along the path of propagation.

5.1   Line-of-Sight Path of Propagation

For the majority of RF telemetry links, the primary mode of EM wave propagation is said to be line of sight, a direct, unobstructed path between the transmitting and receiving antennas. Therefore, for a single antenna, the maximum path of propagation is limited by the distance to the RF horizon. As illustrated in FIGURE 5-1, the RF horizon differs from the optical horizon.
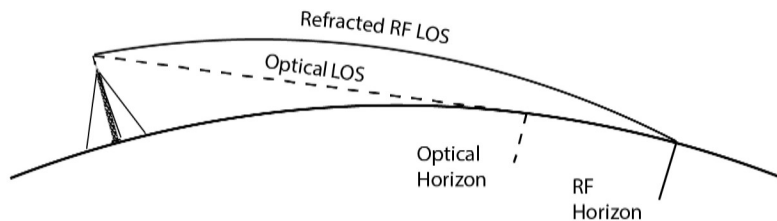

FIGURE 5-1. Optical Horizon vs. RF Horizon

The optical horizon derives from an optical LOS which is a straight, direct path of slant-range distance from the antenna (or eyeball) to a point tangent to the earth's surface. An RF LOS follows a curved path that is initially parallel to the earth's surface but is progressively bent toward the surface due to the refractive properties of the atmosphere. Therefore, the distanced to the RF horizon will be somewhat ($\approx$7%) greater than the distance to the optical horizon.

For a standard atmosphere (standard refraction = k = 1.33) over a smooth earth, the distance to the RF horizon is related to the height of the antenna as follows:

$$d_{HOR} = 4.124\sqrt{h} \tag{5-1}$$

Where:

d_{HOR} = distance in kilometers to the RF horizon

h = the antenna height in meters above a smooth earth

For h in feet and d in statute miles:

$$d_{HOR} = 1.414\sqrt{h} \tag{5-2}$$

For an RF telemetry link, the maximum line-of-sight path distance is equal to the sum of the RF horizon distance for both the transmitting and receiving antennas:

$$LOS_{MAX} = 4.124\sqrt{h_{TX}} + 4.124\sqrt{h_{RX}} \tag{5-3}$$

Where:

LOS_{MAX} = the maximum line-of-sight path distance in kilometers

h_{TX} = height of the transmitting antenna in meters above a smooth earth

h_{RX} = height of the receiving antenna in meters above a smooth earth

NOTE	For practical, non-smooth earth applications, all heights should be relative to a common reference elevation such as mean sea level (MSL).

Therefore:

$$LOS_{MAX} = 4.124\sqrt{30} + 4.124\sqrt{10}$$

$$LOS_{MAX} = 35.6 \text{ km}$$

For a link to be considered as having a line-of-sight path of propagation, the distance between the transmitting and receiving antennas must be equal to or less than the maximum line-of-sight path distance:

$$d_{PATH} \leq LOS_{MAX} \tag{5-4}$$

Since 32 km (the distance between the transmitting and receiving antennas) is less than the maximum allowable 35.6 km, this link qualifies as a LOS path of propagation.

5.2   Free Space Propagation Model

As an EM wave propagates in free space, the power density per unit area decreases in proportion to the frequency and the square of the distance traveled. These facts give rise to the classic free space loss equation:

$$FSL_{dB} = 32.45 + 20 \log(d) + 20 \log(f) \tag{5-5}$$

Where:

FSL_{dB} = free space loss in dB

d = distance in kilometers

f = frequency in megahertz

For distance in statute miles:

$$FSL_{dB} = 36.58 + 20 \log(d_{miles}) + 20 \log(f) \tag{5-6}$$

Therefore, for a distance of 32.2 km and an operating frequency of 170 MHz:

FSL_{dB} = 32.45 + 20 log (32.2) + 20 log (170)

FSL_{dB} = 107.2 dB

While free space loss alone is often used in link budget calculations, it is important to understand that in this context, the term "free space" is meant literally; no atmosphere and no reflective surfaces or obstructions of any type. This does not represent a realistic environment for earth-based telemetry links and, for many path scenarios, the use of free space loss alone will not result in a realistic link budget.

5.3   2-Ray Multipath Propagation Model

In the more typical terrestrial setting, the EM wave must propagate through nonhomogeneous atmosphere over a path of often mixed terrain and uneven topography. Additionally, system design constraints may require that a link be established over a path containing unavoidable manmade or natural obstructions. Many of these non-free-space elements in the physical environment can cause the propagating wavefront to be absorbed, scattered, refracted, reflected, or diffracted. For a discussion on line-of-sight obstructions, see *Line of Sight Obstruction* app note.

The net effect of these propagation mechanisms on the signal level reaching the receiving antenna can be difficult to evaluate analytically and requires a level of detailed information typically not available without a precise survey of the intended path. For these reasons, a commercial software application using established empirical propagation models linked to a GIS database is the recommended tool for predicting losses for an obstructed line-of sight path of propagation. A discussion of these tools and their application is beyond the scope of this document.

For the purposes of this exercise, a standard atmosphere and an unobstructed line-of-sight (LOS) path over a smooth earth will be assumed.

NOTE
In this context, the term *smooth earth* refers to the surface of an idealized spherical earth of constant radius.

For even the most benign of LOS scenarios, the signal captured by the receiving antenna is seldom the result of the direct path of propagation alone. In all but the very shortest of path lengths, there will likely be an indirect or reflected signal arriving at the receiving antenna coincident with the direct signal. The resultant signal delivered to the connected receiver will be the vector sum of the relative amplitude and phase of the direct and reflected wavefronts.

For an unobstructed LOS path over relatively flat terrain, the primary source of reflections is the earth's surface. The effect of the ground reflected wavefront on the received signal is largely dependent on the distance between the transmitting and receiving antennas, the relative height of the antennas, and the reflective properties of the earth's surface. The path geometry for a 2-ray propagation model is illustrated in FIGURE 5-2.
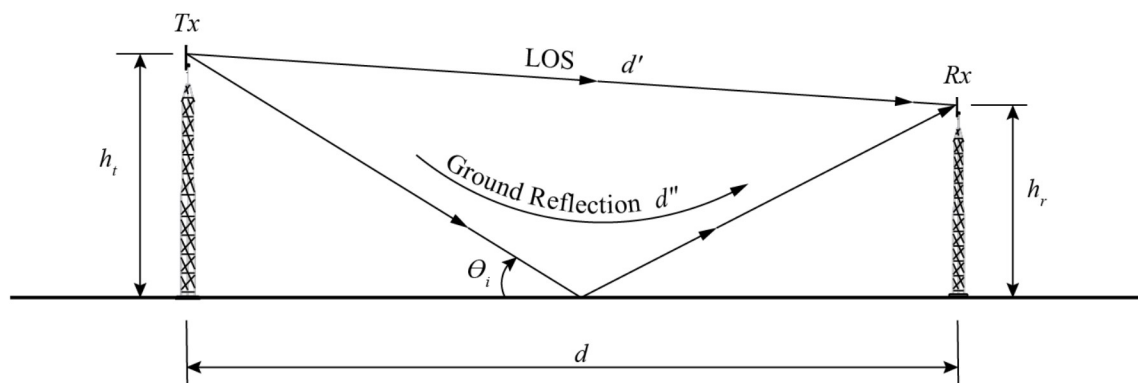

FIGURE 5-2. Path Geometry for 2-Ray Propagation Model

The reflected wavefront will arrive at the receiving antenna differing in phase and, possibly, amplitude relative to the direct wavefront.

The difference in phase is due to a combination of the increased time required to travel the greater distance (d'') and any phase shift induced into the reflected wavefront at the point of reflection. For a horizontally polarized wavefront, there will always be a 180° phase shift induced at the point of reflection. For a vertically polarized wavefront, a phase shift in the reflected wavefront only occurs for an angle of incidence (Θ_i) of less than 30°. The phase shift in a

ground-reflected, vertically polarized wavefront can approach 180° for very small, *grazing angles* of incidence.

Any decrease or attenuation in the amplitude of the reflected wavefront will be partially due to the longer path, but will be largely dependent on the electrical properties of the reflecting surface, the E-field polarity of the incident wave, and the angle of incidence. For a highly conductive surface material such as steel or sea water, there will be little to no attention of the reflected wavefront. For poorly conductive surfaces such as soil, rock, wood, and fresh water, the amplitude of the reflected wave can vary greatly.

The reflected wavefront will interfere with the direct wavefront either constructively or destructively. *Constructive interference* occurs when the wavefronts arrive more or less in phase ($\theta_{diff} < \pm 90°$). A 0° phase shift with a small difference in amplitude can result in as much as a 6 dB gain in received signal strength relative to the direct wavefront alone. Conversely, *destructive interference* occurs when the wavefronts arrive more or less out of phase ($\theta_{diff} > \pm 90°$). With a phase difference of 180° and a small difference in amplitude, the wavefronts will cancel out, resulting in a null in the received signal level.

It should be clear from FIGURE 5-2 that the factors affecting path loss are greatly dependent on the relative heights of the antennas and the path distance.

When the difference in antenna heights is large and the path distance is less than a *critical distance* (d_c), the direct and reflected wavefronts will alternate between constructive and destructive interference for successive values of d (1, 2, 3, and so forth). However, the path loss will, on average, follow the square-of-the-distance behavior defined by the free space loss equation.

When the path distance is equal to or greater than the critical distance, the relative antenna heights become very small compared to the path distance, and the angle of incidence will approach 0°. For this path geometry, the phase shift contributable to a difference in path lengths becomes very small, and the phase shift induced in the reflected wave approaches 180° for both vertical and horizontal polarization. Under these conditions, the power density per unit area will decrease in proportion to the fourth-power of the distance, and the path loss can be calculated using the following equation:

$$PL_{2Ray} = 120 - 20\log(h_{TX}h_{RX}) + 40\log(d) \qquad (5\text{-}7)$$

Where:

PL_{2Ray} = 2-ray path loss in dB

h_{TX} = height of the transmitting antenna in meters

h_{RX} = height of the receiving antenna in meters

d = distance between antennas in kilometers

The critical distance is calculated as follows:

$$d_c = \frac{4\pi h_{TX} h_{RX}}{\lambda} \qquad (5\text{-}8)$$

Where:

d_c = critical distance in meters

h_{TX} = height of the transmitting antenna in meters

h_{RX} = height of the receiving antenna in meters

λ = wavelength of the propagating EM wave; 1.76 meters @ 170 MHz.

Therefore:

$$d_c = \frac{4 \cdot \pi \cdot 30 \cdot 10}{1.76}$$

d_c = 2.14 kilometers

For $d < d_c$: calculate path loss using the free space propagation model (Eq. 5-5).

For $d \geq d_c$: calculate path loss using the 2-ray propagation model (Eq. 5-7).

Because the distance between antennas is 32.2 kilometers, this example requires the 2-ray propagation model.

Therefore:

$$L_{PATH} = PL_{2Ray} = 120 - 20\log(30 \cdot 10) + 40\log(32.2)$$

L_{PATH} = 130.8 dB

6.   Received Signal Level

Having derived all of the input parameters to the link budget, we can now calculate the power level arriving at the receiver's input.

From FIGURE 1-1:

$$P_{RX} = P_{TX} - L_{TX} + G_{TX} - L_{PATH} + G_{RX} - L_{RX}$$

Therefore:

P_{RX} = 37 dBm – 3.2 dB + 5.15 dBi – 130.8 dB + 5.15 dBi – 2.35 dB = –89 dBm

7. Fade Margin

By subtracting the receiver's specified sensitivity from the calculated received signal level, we can determine the extent to which transient path losses or signal fading can be tolerated before system performance is impacted.

As has been noted, the receiver's sensitivity specifies the minimum RF input power required to produce a useable output signal. A point of consternation for the designer is that transceiver manufacturers will often define and quantify "a useable output signal" in differing and sometimes ambiguous ways.

Two common methods of specifying receiver sensitivity are:

- The minimum input signal level required to limit the number of errors in the received digital data stream to a maximum Bit Error Rate (BER). A typical specification would be: –103 dBm for 1 x 10^{-4} BER— meaning, one bit error for every ten thousand bits received.

- The minimum input signal level required to produce a minimum SINAD ratio in the demodulated audio. SINAD is the ratio, in dB, of (Signal + Noise + Distortion) to (Noise + Distortion) and is an expression of audio quality for voice communications. A typical specification would be: 0.28 µV for 12 dB SINAD. A somewhat subjective industry standard specifies a SINAD ratio of 12 dB as the minimum required for intelligible voice communications.

The sensitivity specification for the RF320 series of radios states the RF input level in units of microvolts (µV). For link budget calculations, it is convenient to convert units of voltage to units of power. For a 50 Ω system (the standard for the telecommunications industry), the following equation can be used to convert volts to power in dBm:

$$P_{dBm} = 10 \log \left[\frac{(V \cdot 10^{-6})^2}{50} \right] + 30 \qquad (7\text{-}1)$$

Where:

P_{dBm} = power in dBm

V = rms voltage in microvolts

Therefore:

Rx Sensitivity at 0.25 uV for 12 dB SINAD

$$Rx\ Sensitivity\ (P_{dBm}) = 10 \log \left[\frac{(0.25 \cdot 10^{-6})^2}{50} \right] + 30$$

Rx Sensitivity = –119 dBm

We can now calculate the fade margin for the link. From FIGURE 1-1:

Fade Margin = P_{RX} – Rx Sensitivity = (–89 dBm) – (–119 dBm) = 30 dB

Where P_{RX} was derived in Section 6, *Received Signal Level (p. 10)*, and *Rx Sensitivity* was derived earlier in this section.

8.   Conclusions

A link budget provides a quick, simplistic assessment of a link's viability and should only be used as a design tool. The calculations herein provide only theoretical approximations and do not account for all of the myriad, real-world variables that can affect system performance. All link budgets should be verified via observed measurements before committing to an installation. Analysis is no substitute for empirical data.

If a highly reliable, mission critical RF telemetry link is required, the design goal should be for a minimum fade margin of 20 to 30 dB. If the link budget calculations or on-site measurements indicate a fade margin of less than 10 dB, one should exercise all possible options to improve upon this figure. Some possible options are:

- Use an antenna with a higher gain specification on one or both ends of the link. One should be cognizant of any FCC regulations that may put limits on the maximum radiated power for given transmitter site.

- Increase the antenna elevation at one or both ends of the link. If path obstructions or multipath interference is suspected, even a small increase (or decrease) of one-half wavelength could make a significant difference in received signal level. Any increase in system losses due to a longer transmission line are usually more than offset by the decrease in path loss.

- Add a repeater site to the path. By far, the largest factor in a link budget is path loss.

A quick web search will reveal numerous online calculators and software products for calculating path loss and link budgets. Some are better than others. At the very least, the exercise outlined in this application note should impart an awareness—if not a complete understanding—of some of the factors involved in designing an RF telemetry link.

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