Towards a Link Budget for Ultra WideBand (UWB) Systems

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A Link Budget for UWB Systems?

A link power budget is an important calculation for any radio communications system. It is a relatively simple calculation that is often done without much thought being given to the assumptions that stand behind it. However, the ultra wideband occupied by UWB systems and the pulse nature of UWB complicate the calculation of received signal power. The commonly used Friis transmission formula may give misleading or incorrect results when applied to UWB systems.

The Friis Transmission Formula or "path loss" formulas used for most communication systems link design predict that the received signal power will decrease with the square of increasing frequency. With narrowband systems, this change in received power over the signal bandwidth is usually ignored as it has little effect. UWB can occupy octave or even decade bandwidths so the frequency dependence of Friis appears as a filter where $|H| \propto 1/f^2$. This will distort the frequency spectrum of UWB pulses and thus distort the pulse shape.

Path Loss in UWB Systems

Path loss is usually thought of as a property of the path that the radio waves traverse, but "path loss" actually includes assumptions about antennas. It is necessary to look at the development of Friis to understand its limitations and to develop an alternative that is more useful for UWB.

Friis is based on the inverse spreading law. Consider a transmitter that radiates power equally in all directions. The power radiated is the Effective Isotropic Radiated Power ($EIRP$). If the transmitter is in free space and it is surrounded by a sphere of radius d, then the flux density is just the power radiated divided by the area of the sphere:

$$F = \frac{EIRP}{4\pi d^2} \qquad watts/m^2 \tag{1}$$

Each time the distance, d, from the transmitter is doubled, the flux is reduced by a factor of four. The flux is dependent only on the $EIRP$ and the distance from the transmitter. There is no frequency dependence.

Knowledge of the power at the terminals of receiver antenna is what is desired. The received power, P_r, is obtained by multiplying the effective collecting area of the antenna, i.e. the effective aperture A_e, by the flux density.

$$P_r = \frac{EIRP}{4\pi d^2} A_e \qquad watts \tag{2}$$

The gain of an antenna is related to its effective aperture:

$$G = \frac{4\pi}{\lambda^2} A_e \tag{3}$$

Solving for the effective aperture and substituting the result into (2) results in the familiar Friis Transmission Formula:

$$P_r = \frac{EIRP}{4\pi d^2}\left(\frac{\lambda^2}{4\pi}\right)G_r = EIRP\left(\frac{\lambda}{4\pi d}\right)^2 G_r \tag{4}$$

The squared factor is referred to as the "path loss."

$$Path\ Loss = \left(\frac{\lambda}{4\pi d}\right)^2 \tag{5}$$

However, this can be misleading since it appears that the "path loss" is a function of frequency where the inverse spreading loss is only a function of distance. The frequency dependent term in the path loss is a result of an assumption that was made about the antenna. When (3) was substituted into (4), it was implicitly assumed that the antenna had a constant gain. This means that the effective aperture decreases with frequency. This assumption introduces the frequency dependent term in the path loss. If the antenna is assumed to be constant aperture instead, then (2) defines the received power and there is no frequency dependence.

The Effective Isotropic Radiated Power ($EIRP$) also assumes a constant gain antenna. $EIRP$ can be written as:

$$EIRP = P_t G_t \tag{6}$$

Where P_t is the transmitter power and G_t is the transmitter antenna gain. If a constant aperture antenna is assumed, the a flux density will be a function of wavelength:

$$F = P_t\left(\frac{4\pi A_{et}}{\lambda^2}\right)\left(\frac{1}{4\pi d^2}\right) = \frac{P_t A_{et}}{\lambda^2 d^2} \qquad watts/m^2 \tag{7}$$

For a constant aperture transmitter antenna, the flux density increases with increasing frequency or decreasing wavelength.

There are four possible combinations of constant gain/constant aperture antennas since there is an antenna at each end of the path.

Constant gain transmit/constant gain receive (Friis):

$$P_r = P_t G_t \left(\frac{\lambda}{4\pi d}\right)^2 G_r \qquad (8a)$$

Constant gain transmit/constant aperture receive:

$$P_r = P_t G_t \left(\frac{1}{4\pi d^2}\right) A_{er} \qquad (8b)$$

Constant aperture transmit/constant gain receive:

$$P_r = P_t \left(\frac{4\pi A_{et}}{\lambda^2}\right)\left(\frac{1}{4\pi d^2}\right)\left(\frac{\lambda^2}{4\pi}\right) G_r = P_t A_{et} \left(\frac{1}{4\pi d^2}\right) G_r \qquad (8c)$$

Constant aperture transmit/constant aperture receive:

$$P_r = \frac{P_t (A_{et})(A_{er})}{(\lambda d)^2} \qquad (8d)$$

Systems that employ contain gain transmit/constant gain receive antennas or constant aperture transmit/constant aperture receive antennas will experience a received power that is a function of frequency/wavelength. The received power for constant aperture transmit/constant aperture receive systems will increase by the square of the frequency while the received power for constant gain transmit/constant gain receive will decrease as the square of the frequency.

With constant aperture illumination, horn and reflector antennas are constant effective aperture antennas. Antennas such as the log periodic and the four square (?) that have a constant pattern with frequency are constant gain antennas.

It is possible to extend the expressions in (8) to include an arbitrary path loss exponent, n. For the constant gain transmit/constant aperture receive case (8b) becomes:

$$P_r = P_t G_t \left(\frac{1}{4\pi d_o^2}\right)\left(\frac{d_o}{d}\right)^n A_{er}$$

$$d_o = 1\, meter \qquad d:\ meters$$

Friis does not tell the whole story. This analysis suggests that is will be difficult (but not impossible) to separate the antenna from the path. The received power in a UWB system that uses one constant gain and one constant aperture antenna will be frequency independent.

Transmitter Power

UWB transmits high amplitude pulses that only last for nanoseconds. The result is a high peak to average power. The question is: what is the transmitted power in the link budget calculation? In addition, some UWB systems detect single pulses rather than many pulses averaged over time. For these reasons, it may make more sense to talk about transmitter energy per pulse rather than power in much the same way that energy per bit is calculated in digital systems.

$$E_b = PT_b \tag{9}$$

P is the transmitted power and T_b is the bit rate. This calculation assumes that each bit is represented by a square pulse. UWB pulses are not square so it will be necessary to calculate the energy is the equivalent square pulse:

$$E_p = \int_{pulse\,length} |x(t)|^2 dt \tag{10}$$

E_p is the energy in a single pulse and $x(t)$ is the pulse. Since most UWB pulses have shapes that can be approximated by simple functions, it should be relatively easy to evaluate the expression in (10).

Receiver

UWB receivers tend to fall into two categories. The first are threshold receivers. These receivers produce an output when the signal level exceeds a preset threshold. These receivers usually employ a tunnel diode with some kind of adjustable bias that results in a constant false alarm rate. I believe that the receiver used by MultiSpectral Solutions is of this type. The sensitivity of this type of receiver is related to the false alarm rate. Constant false alarm rate receivers are covered in the radar literature.

The second type of receiver is a correlator followed by an integrate and dump. A block diagram for the correlator receiver is shown in Figure 1. This receiver consists of a local pulse generator, a multiplier, and an integrate and dump followed by some kind of decision circuit. It can be thought of in several different ways. From one perspective it is a direction conversion receiver with the "local oscillator" being the pulse generator. If the locally generated pulse meets the following criteria:

$$x_{local}(t) = x_{in}(-t) \tag{11}$$

The receiver is a matched filter and the S/N ratio is maximized.

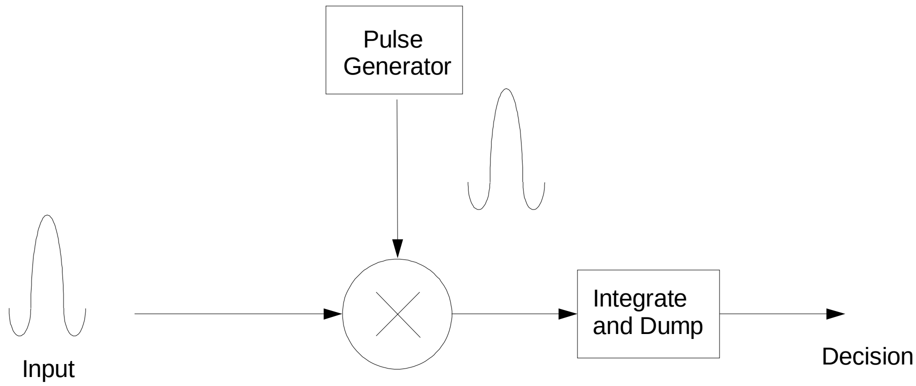

Figure 1: Correlator receiver

Unfortunately, the input pulse will be corrupted by the channel. The input will be the transmitted pulse plus sum of all the multipath components. Since the channel is not known precisely and generating a time inverted version of an arbitrary pulse is not possible, the correlator receiver will always be suboptimum. The energy received will be some faction of the energy available:

$$E_r = \eta E_a$$

Where E_r is the energy received and E_a is the energy available. η will always be less than one. Including η is a link budget calculation may be open to debate. Normally link power budgets are referred to the antenna input terminals and the demodulation process is not included. However, UWB is somewhat unique. The entire received energy is available at the antenna terminals but it cannot be captured in its entirety. η may be considered in much the same way implementation loss is added to communications systems in order to compensate for less than optimum conversion of C/N to effective S/N in the demodulator.

Noise

It is normally assumed that the noise in the system is Additive White Gaussian Noise. The front end of a UWB receiver must admit the entire bandwidth of the UWB pulse. This means that a significant number of interferers, both UWB and narrowband, will be admitted. While the correlation process will select out the desired signal, the correlator must contend with multiple signals. A roughly analogous situation may be a CDMA receiver that must select the desired CDMA signal from among many.

Since the correlator receiver is a linear system, noise performance in the presence of AWGN only can be determined by conventional methods. The noise performance of this type of receiver should be in the literature somewhere. It is probably worth finding this analysis and adapting it to UWB receivers.

An additional question relates to the possibility of using coherent averaging to increase the post detection S/N. The radar literature may be a point of departure to explore this issue.

Propagation Issues

UWB is unique in its ability to resolve small differences in time. These time differences can be translated into spacial differences. Ranging and position location are widely studied applications for UWB. John McCorkle from Extreme Spectrum has some dramatic pictures created with UWB. The showed the ability of UWB to resolve the top and the bottom of telephone poles. This is done by observing the time difference of arrival.

John pointed out that it is not possible to get such a picture from frequency domain measurements done with wideband network analyzer sweeps. The problem is that the sweep is done at discrete frequencies with a fairly long dwell period. The result is that it is not possible to resolve the reflections from individual objects. While the network analyzer will give a broadband complex frequency response, it is not a true chirp measurement. The result is that the network analyzer sweeps are non causal. It is not clear if this is a limitation when doing network analyzer propagation measurement. A number of researchers are making measurements in this fashion but has the technique been rigorously validated.

John McCorkle also has some observations about antennas. He had measurement data for multipath reflections from a three sided corner reflector. He suggested that the observed phase shifts in the multipath reflections were due to off axis antenna phase shifts. Antenna gain, polarization purity, and phase shift change with angle off bore sight. Gain, polarization, and phase shift are three parameters that should be explored and their effect on propagation noted if our effort to measure propagation effects independent of the antennas is to be successful.

Results presented by the Bob Scholtz at MURI meeting suggest that there may be some things to consider in choosing parameters to describe propagation phenomena. Scholtz and Moe Win did extensive measurements (see: Impulse Radio, Robert.A. Scholtz and Moe Z Win, IEEE PIMRC '97, Helsinki). Win went to Time Domain in Huntsville for six weeks to do measurements. It appears that much of this data has never been published.

Scholtz showed some of this data at the MURI meeting and made some interesting observations. The line of sight (LOS) data looks like you would expect. The first ray is the strongest and that is followed by numerous multipath components. The multipaths appear almost as a continuum. The 20 GHz bandwidth of the scope used for the measurements allows very fine time resolution of the multipath components.

The Scholtz/Win measurement system employed an HP54121T 20GHz digitizing scope. This is a classic sampling scope so it does not have a real time mode. It requires a repetitive wave form.

Scholtz gave an interesting interpretation to the non-LOS data. The first multipath components were not the strongest ones and he suggested that signal level builds up as the various multipath components "leak" into the room. The signal level reaches a maximum and then the multipath components appear to die off exponentially.

Should new parameters be proposed to characterize the non-LOS path?