Closed-Form Approximations for Link Loss in an UWB Radio System Using Small Antennas

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Abstract:
A critical need in the evaluation of an UWB radio system is the calculation of the energy link loss between the source at the transmit antenna and the receiver load. While the rigorous calculation of link loss in a wideband pulsed system requires a full transient electromagnetic solution for the transmit and receive antennas, we show in this paper that accurate approximations for link loss can be obtained for the special cases of electrically small dipole or loop antennas, with gaussian or gaussian doublet (monocycle) generator waveforms. We also consider the error involved with applying the much simpler narrowband Friis transmission formula. It is found that the use of the basic Friis formula can result in link loss errors of more than 60 dB for an UWB system having severely (impedance) mismatched antennas, but may give results correct to within a few dB for well-matched narrowband antennas, or if the formula is augmented with an impedance mismatch correction factor. It appears that the dominant limitation of the Friis formula, when applied to UWB systems, is the broadband effect of mismatch between the transmit/receive antennas and their source or load impedances. Numerical examples are presented for electrically short dipoles, resonant dipoles, and broadband lossy dipoles, for both gaussian and monocycle input pulse waveforms.

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Introduction:

Ultra wideband (UWB) radio systems rely upon the radiation and propagation of baseband transient pulses. As described in [1]-[2], there are many features of UWB radio (e.g., the utilization of under-used spectrum segments, mitigation of indoor fading and multipath effects, low power densities, and high levels of multi-user scaling) that have led to intense interest in this new technology. A critical need in the design and evaluation of an UWB radio is the calculation of the energy link loss between the transmitting source and the receiver – a task made difficult by the fact that the absence of a sinusoidal carrier precludes the use of the Friis formula. The rigorous calculation of UWB link loss requires a complete transient electromagnetic solution (using numerical finite difference or integral equation techniques) for the transmit and receive antennas to account for the effects of impedance mismatch over a wide bandwidth, pulse distortion effects, and the effects of frequency dependent antenna gains and spreading factors. In this paper, however, we show that accurate approximations for link loss can be made for the special cases of electrically small dipole or loop antennas, with gaussian or gaussian doublet (monocycle) generator waveforms. We also find that the Friis formula may give reasonably good results when the antennas are relatively narrowband.

We first summarize the calculation of UWB energy transmission based on the rigorous electromagnetic analysis of transient radiation and reception, including the effects of generator and receiver impedances, for an arbitrary input waveform. Next we derive closed-form approximations for the link loss in a UWB radio system using electrically small dipole or loop antennas, for either gaussian or monocycle input waveforms (the two UWB radio excitations most commonly used in practice). Numerical examples are presented for three types of antennas (an electrically short dipole, a resonant dipole, and a broadband lossy dipole), for both gaussian and monocycle input pulse waveforms.

We also consider the much simpler technique of applying the narrowband Friis transmission formula, and compare with rigorous calculations and approximate closed-form results. It is found that the use of the basic Friis formula can result in link loss errors of more than 60 dB for an UWB system having severely (impedance) mismatched antennas, but may give results correct to within a few dB for well-matched narrowband antennas, or by augmenting the formula with an impedance mismatch correction factor. We conclude that the dominant limitation of the Friis formula when applied to UWB systems is not the frequency dependence of the spreading factor or antenna gain terms, but the broadband effect of mismatch between the transmit/receive antennas and their source or load impedance. Pulse distortion effects also limit the accuracy of the Friis approximation, but to a much lesser degree.

Link Loss Based on Rigorous Electromagnetic Analysis:

We assume a canonical UWB radio configuration like that shown in Figure 1, where the transmit antenna is driven with a voltage source $V_G(\omega)$ having an internal impedance $Z_G(\omega)$, and the receive antenna is terminated with load impedance $Z_L(\omega)$, and has a terminal voltage $V_L(\omega)$. The input impedance of the transmit and receive antennas are $Z_T(\omega)$ and $Z_R(\omega)$, respectively. The antennas are separated by a distance *r*, assumed to be large enough so that each antenna is in the far field region of the other over the operating bandwidth.

Let $H_{LG}(\omega)$ be the voltage transfer function that relates the receive antenna load voltage to the generator voltage at the transmit antenna [3]-[5]:

$$V_L(\omega) = H_{LG}(\omega) V_G(\omega) e^{-j\omega r/c}, \qquad (1)$$

where c is the speed of light. Note that the exponential factor representing the time delay between the transmit and receive antenna has been separated from the transfer function. Although not explicitly shown, it should be understood that this transfer function is dependent on range as well as the elevation and azimuth angles at each antenna.

The time domain voltage waveform at the receive antenna is then found as,

$$v_L(t') = \frac{1}{2\pi} \int_{BW} H_{LG}(\omega) V_G(\omega) e^{j\omega t'} d\omega, \qquad (2)$$

where $t' = t - r/c$ is the retarded time variable.

The following energy quantities can also be defined. The energy delivered to the transmit antenna is given by,

$$W_{in} = \frac{1}{2\pi} \int_{BW} \frac{|V_G(\omega)|^2 R_T(\omega)}{|Z_T(\omega) + Z_G(\omega)|^2} d\omega, \qquad (3)$$

where $R_T(\omega)$ is the real part of $Z_T(\omega)$**.** The energy received by the load at the receive antenna is given by,

$$W_{rec} = \frac{1}{2\pi} \int_{BW} \frac{|V_L(\omega)|^2}{Z_L^*(\omega)} d\omega. \qquad (4)$$

The integrations in (2)-(4) are over the bandwidth of *–B* to *B* Hertz, where *B* is the effective bandwidth of the generator waveform.

To calculate link loss for a specific set of antennas and a given generator waveform, the transfer function of (1) is first computed over a range of frequencies that cover the system bandwidth (as determined by the spectrum of the generator waveform). This can be done using a numerical electromagnetic analysis (e.g., moment method or finite difference technique), as described in [3]-[5]. Next, the input energy is computed using (3), then the received energy using (4). The link loss is defined as the ratio of these two quantities. Note that this calculation includes polarization mismatch, propagation losses, antenna efficiency, impedance mismatches, and waveform distortion effects.

For the results that follow, we define a gaussian generator waveform as,

$$v_G(t) = V_0 e^{-t^2/2T^2}, \tag{5a}$$

and a monocycle (gaussian doublet) generator waveform as,

$$v_G(t) = V_0 \frac{t}{T} e^{-t^2/2T^2}. \tag{5b}$$

Note that the gaussian pulse has non-zero DC content, although this does not contribute to either the input energy or receive energy.

Closed-Form Approximations for UWB Link Loss for Short Dipoles:

Using reasonable approximations it is possible to derive closed-form expressions for the link loss of a UWB radio system using electrically small dipoles or loops, and either a gaussian pulse or a monocycle generator waveform. These results appear to be the only special cases that can be expressed in closed form, and are therefore useful for showing the dependence of waveform shape, receiver impedance, and gain factors in more general situations. In the results to follow, we assume that both transmit and receive antennas are identical, are polarization matched, and are oriented so that each is in the main beam of the other.

The input impedance of an electrically short lossless dipole of half-length $h = L/2$ and radius a can be approximated as [3], [6]:

$$Z_{in}(\omega) = R_{in}(\omega) + jX_{in}(\omega) \simeq \alpha\omega^2 - j/\omega C_0, \tag{6}$$

where $\alpha = \eta_0 h^2 / 6\pi c^2$, $C_0 = \dfrac{-h}{120c\left[1 + \ln\dfrac{a}{h}\right]}$, c is the speed of light, and $\eta_0 = 377\,\Omega$ is the impedance of free space. This approximation is accurate for frequencies up to where the dipole length is less than λ/20. Over this range the input resistance is less than 0.5 Ω, while the input reactance is at least several thousand ohms.

The input energy of (3), for the gaussian generator voltage of (5a), can be evaluated as,

$$W_{in} = V_0^2 T^2 \alpha C_0^2 \int_{-\infty}^{\infty} \omega^4 e^{-\omega^2 T^2} d\omega = \frac{3\sqrt{\pi} V_0^2 \alpha C_0^2}{4T^3} = \frac{3\sqrt{\pi} V_0^2 h^2 \eta_0 C_0^2}{24\pi c^2 T^3}, \tag{7a}$$

while for the monocycle generator voltage of (5b), the input energy is,

$$W_{in} = V_0^2 T^4 \alpha C_0^2 \int_{-\infty}^{\infty} \omega^6 e^{-\omega^2 T^2} d\omega = \frac{15\sqrt{\pi} V_0^2 \alpha C_0^2}{8T^3} = \frac{5\sqrt{\pi} V_0^2 h^2 \eta_0 C_0^2}{16\pi c^2 T^3}. \tag{7b}$$

Observe that these input energies do not depend on the source resistance, R_G. From [3], the transfer function defined in (1) for a UWB radio using short dipole antennas can be written as,

$$H_{LG}(\omega) = \frac{-j\omega\mu_0 h^2 Z_L(\omega)}{4\pi r\left[Z_G(\omega) + Z_{in}(\omega)\right]\left[Z_L(\omega) + Z_{in}(\omega)\right]}, \tag{8}$$

where $Z_{in}(\omega) = Z_T(\omega) = Z_R(\omega)$ is the input impedance of the transmit and receive dipoles (assumed to be identical). Thus, in (8) we can ignore R_{in} and Z_G in the denominator (it is generally desired to use relatively small values of R_G to maximize power transfer, while X_G should be small to minimize resonance effects). There are then two cases of practical interest for the load resistance, depending on whether $R_L << 1/\omega C_0$, or $R_L >> 1/\omega C_0$. For the first case, R_L can be ignored in the denominator of (8), and the transfer function can be approximated as,

$$H_{LG}(\omega) \simeq \frac{j\omega^3 C_0^2 \mu_0 h^2 R_L}{4\pi r}. \quad \text{(small } R_L\text{)} \tag{9}$$

Then the receive energy of (4) can be evaluated as

$$W_{rec} = \frac{V_0^2 T^2 C_0^4 \mu_0^2 h^4 R_L}{16\pi^2 r^2} \int_{-\infty}^{\infty} \omega^6 e^{-\omega^2 T^2} d\omega = \frac{15\sqrt{\pi} V_0^2 C_0^4 \mu_0^2 h^4 R_L}{128\pi^2 T^5 r^2}, \tag{10}$$

and the resulting energy link loss is,

$$L_{link} = \frac{W_{rec}}{W_{in}} = \frac{15\eta_0 R_L}{16\pi}\left(\frac{C_0 h}{Tr}\right)^2. \quad \text{(gaussian, small } R_L\text{)} \tag{11}$$

When $R_L >> 1/\omega C_0$, X_{in} can be ignored in the denominator of (8), and the transfer function can be approximated as,

$$H_{LG}(\omega) \simeq \frac{\omega^2 C_0 \mu_0 h^2}{4\pi r}. \text{ (large } R_L\text{)} \tag{12}$$

The receive energy of (4) can then be evaluated as,

$$W_{rec} = \frac{V_0^2 T^2 C_0^2 \mu_0^2 h^4}{16\pi^2 r^2 R_L} \int_{-\infty}^{\infty} \omega^4 e^{-\omega^2 T^2} d\omega = \frac{3\sqrt{\pi} V_0^2 C_0^2 \mu_0^2 h^4}{64\pi^2 T^3 r^2 R_L}, \tag{13}$$

and the energy link loss is,

$$L_{link} = \frac{W_{rec}}{W_{in}} = \frac{3\eta_0 h^2}{8\pi r^2 R_L}. \qquad \text{(gaussian, large } R_L\text{)} \tag{14}$$

Figure 2 shows a comparison of the closed-form results of equations (11) and (14) with rigorous data from a moment method solution [7] for the short dipole example used above. For these parameters, it is seen that the "small R_L" result of (11) works well for R_L up to about 1000 Ω, while the "large R_L" form works well down to about 20,000 Ω. In between there is a transition region where a closed-form result is not feasible. Interestingly, it appears that minimum link loss occurs in this region.

Results for the monocycle waveform of (5b) can be similarly derived. For small R_L, the receive energy of (4) is evaluated with the transfer function of (9) to give,

$$W_{rec} = \frac{V_0^2 T^4 C_0^4 \mu_0^2 h^4 R_L}{16\pi^2 r^2} \int_{-\infty}^{\infty} \omega^8 e^{-\omega^2 T^2} d\omega = \frac{105\sqrt{\pi} V_0^2 C_0^4 \mu_0^2 h^4 R_L}{256\pi^2 T^5 r^2}, \tag{15}$$

and the resulting energy link loss is,

$$L_{link} = \frac{W_{rec}}{W_{in}} = \frac{21\eta_0 R_L}{16\pi} \left(\frac{C_0 h}{Tr}\right)^2. \qquad \text{(monocycle, small } R_L\text{)} \tag{16}$$

For large R_L, the receive energy of (4) is evaluated with the transfer function of (12) to give,

$$W_{rec} = \frac{V_0^2 T^4 C_0^2 \mu_0^2 h^4}{16\pi^2 r^2 R_L} \int_{-\infty}^{\infty} \omega^6 e^{-\omega^2 T^2} d\omega = \frac{15\sqrt{\pi} V_0^2 C_0^2 \mu_0^2 h^4}{128\pi^2 T^3 r^2 R_L}. \tag{17}$$

Then the energy link loss is,

$$L_{link} = \frac{W_{rec}}{W_{in}} = \frac{3\eta_0 h^2}{8\pi r^2 R_L}. \qquad \text{(monocycle, large } R_L\text{)} \tag{18}$$

Figure 3 shows a comparison of the closed-form results of equations (16) and (18) with rigorous data from a moment method solution [7] for the short dipole example used above. For these parameters, it is seen that the "small R_L" result of (16) works well for R_L up to about 1000 Ω, while the "large R_L" form works well down to about 20,000 Ω. Again, the optimum link loss occurs between these values.

Link Loss Using the Narrowband Friis Transmission Formula:

The Friis link equation that applies to CW radio systems is given by [6],

$$P_r(\omega) = P_t(\omega)\frac{G_t(\omega)G_r(\omega)\lambda^2}{(4\pi r)^2}, \tag{19}$$

where P_r and P_t are the received and transmitted powers, G_t and G_r are the transmit and receive antenna gains, and λ is the wavelength at the operating frequency. Note that this result does not include propagation losses, polarization mismatch, or impedance mismatch at either the transmit or receive antenna. Also note that the Friis formula, since it applies only to CW (sinusoidal) signals, does not account for pulse distortion effects at either antenna, or even the type of waveform used at the generator.

If the transmitted signal consists of digital data at a bit rate R_b bits/s, then the energy per bit on transmit and receive is $E_{bt} = P_t / R_b$ and $E_{br} = P_r / R_b$. Then (19) can be written in terms of the transmit and receive bit energies as

$$E_{br}(\omega) = E_{bt}(\omega)\frac{G_t(\omega)G_r(\omega)\lambda^2}{(4\pi r)^2}. \tag{20}$$

The frequency dependence of each term is explicitly shown in (19)-(20). Note that the factor $(r/\lambda)^2$ has a frequency dependence of 6 db per octave, but this is reduced to a maximum error of 3 dB at either end of the octave for a single frequency chosen at midband. Similarly, the frequency variation of the antenna gains is typically small over a wide frequency range for many practical antenna elements. An electrically short dipole antenna, for example, has a gain of about 1.8 dB for all frequencies below resonance. The effect of impedance mismatch can be included (at a particular frequency, ω) by multiplying (20) by the factor $\left(1-\left|\Gamma(\omega)\right|^2\right)$, where $\Gamma(\omega)$ is the reflection coefficient at the receive antenna given by

$$\Gamma(\omega) = \frac{Z_R - Z_L}{Z_R + Z_L}. \tag{21}$$

Note that the effect of mismatch at the generator is not included – this is because we have chosen to use W_{in}, the energy delivered to the transmit antenna, as opposed to the energy available from the generator.

Examples and Discussion:

To compare specific numerical results, we consider the link loss for three different transmit/receive antenna pairs. We choose $T = 4.42\times10^{-10}$ s for both the gaussian pulse and the monocycle waveforms of (5), resulting in a 10 dB bandwidth of 550 MHz for the gaussian pulse, and a 10 dB bandwidth of 70 MHz to 790 MHz for the monocycle pulse. The gaussian waveform contains power at very low frequencies (and DC), which is not radiated by any of the antennas considered here. The parameters for each of the three antennas are given below:

An electrically short dipole: Dipole length = 1.0 cm, dipole radius = 0.02 cm, $Z_L = Z_G = 50\,\Omega$. The 10 dB bandwidth for the magnitude of the resulting transfer function is 10.2 GHz to 18.9 GHz. This element is severely mismatched over the bandwidth of either input signal.

A resonant dipole: Dipole length = 30.0 cm, dipole radius = 0.02 cm, $Z_L = Z_G = 72\,\Omega$. The 10 dB bandwidth for the magnitude of the resulting transfer function is 410 MHz to 580 MHz. This is a relatively narrowband element, but is well-matched to the source and load impedances at its resonant frequency of 500 MHz.

A lossy resonant dipole: Dipole length = 30.0 cm, dipole radius = 0.02 cm, dipole conductivity = 100 S/m, $Z_L = Z_G = 800\,\Omega$. The 10 dB bandwidth for the magnitude of the resulting transfer function is 190 MHz to 990 MHz. This is a broadband element, and is reasonably well-matched to the source and load impedances over the bandwidth of the input signals. Due to the lossy loading, the efficiency of this element is about 10%.

A plot of the transfer function magnitude (as defined in (1)) versus frequency for transmit/receive pairs of these antennas is shown in Figure 4. The resulting energy link losses are shown in Table 1.

Table 1. Normalized (r = 1) Energy Link Loss for Various Antennas and Excitations

Antennas	Gaussian Rigorous (Eqs 1-4, 5a)	Monocycle Rigorous (Eqs 1-4, 5b)	Midband Frequency	Midband Friis (Eq 20)	Midband Friis and Z-Mismatch
Short Dipoles	-85.5	-84.0 dB	430 MHz	-20.8 dB	-87.0 dB
Resonant Dipoles	-23.9	-23.9 dB	500 MHz	-22.1 dB	-22.4 dB
Lossy Dipoles	-43.1	-41.8 dB	500 MHz	-22.1 dB	-22.3 dB

The first two columns of data refer to the rigorous calculation of link loss using the full electromagnetic solution summarized by equations (1)-(4), for the gaussian and monocycle input pulses. These solutions include essentially all relevant effects, including impedance mismatch, pulse distortion, and frequency variation of gain and propagation factors. Observe that the link loss differs by a few dB for the two different input pulses when broadband elements are used (short dipoles or lossy dipoles). In contrast, waveform shape has little effect on link loss when the antennas are relatively narrowband (resonant dipoles), since the relatively narrow portion of the input spectrum that is passed by the antennas results in an essentially sinusoidal waveform.

The remaining three columns present data associated with the Friis formula of (20). The midband frequency is the frequency at which the calculation is performed, and has been selected to be at the maximum response of the associated transfer function (for the resonant and lossy dipoles), or near the midband of the input waveform bandwidth (for the short dipoles). The gain for each antenna was assumed constant at 1.8 dB. Note that using the basic Friis formula of (20), without impedance mismatch correction, gives an error of more than 60 dB when the antennas are severely mismatch (short dipoles), but gives results within a few dB of the correct result for narrow band matched antennas (the resonant dipoles). If the efficiency of the lossy dipoles is included in the Friis calculation (10% efficiency, or 20 dB loss for combined transmit and receive antennas), reasonable results (-42.3 dB) are also obtained for this case.

We conclude that for narrowband antennas, the Friis formula can give results within about 1 dB for UWB systems (of course, it is generally undesirable to use such narrowband antennas for a wideband system). For broadband elements, application of the Friis formula with the impedance mismatch factor can produce results that are accurate to about 3 dB. More complicated elements, such as arrays or traveling wave antennas, will likely lead to different conclusions.

Closed-Form Approximations for UWB Link Loss for Small Loops:

Closed-form approximations can also be derived for electrically small loops with gaussian or monocycle excitations. Since the procedure is the same as used above for electrically short dipoles, only the key results are presented here.

Consider two circular wire loop transmit and receive antennas having loop radius *a*, and wire radius *b*. For frequencies where $a < 0.03\lambda$ the input impedance of the loop can be approximated as [6],

$$Z_{in}(\omega) = R_{in}(\omega) + jX_{in}(\omega) = \beta\omega^4 + j\omega L_0, \tag{22}$$

where $\beta = \pi\eta_0 a^4 / 6c^4$, and $L_0 = \mu_0 a\left[\ln\frac{8a}{b} - 2\right]$ is the loops self-inductance (the wire self-inductance can also be included, if desired.

Then the input energy for the gaussian generator voltage of (5a) can be evaluated as,

$$W_{in} = \frac{V_0^2 T^4 \beta}{L_0^2}\int_{-\infty}^{\infty} \omega^2 e^{-\omega^2 T^2} d\omega = \frac{\sqrt{\pi} V_0^2 \beta}{4TL_0^2}, \tag{23a}$$

while for the monocycle generator voltage of (5b), the input energy is,

$$W_{in} = \frac{V_0^2 T^4 \beta}{L_0^2} \int_{-\infty}^{\infty} \omega^4 e^{-\omega^2 T^2} d\omega = \frac{3\sqrt{\pi} V_0^2 \beta}{4 T L_0^2}. \qquad (23b)$$

We assume that $\omega L_0 >> R_G$, and consider two cases of receiver load resistance. For $R_L << \omega L_0$, the transfer function of (8) can be approximated as,

$$H_{LG}(\omega) \simeq \frac{-j\pi\omega\eta_0 a^4 R_L}{4c^3 L_0^2 r}, \qquad \text{(small } R_L\text{)} \qquad (24)$$

while for $R_L >> \omega L_0$ the transfer function reduces to,

$$H_{LG}(\omega) \simeq \frac{\pi\omega^2 \eta_0 a^4}{4c^3 L_0 r}. \qquad \text{(large } R_L\text{)} \qquad (25)$$

Using these results in (4) gives the link loss for gaussian pulses as,

$$L_{link} = \frac{W_{rec}}{W_{in}} = \frac{3\pi\eta_0 a^4 R_L}{8c^2 L_0^2 r^2}, \qquad \text{(gaussian, small } R_L\text{)} \qquad (26)$$

and,

$$L_{link} = \frac{W_{rec}}{W_{in}} = \frac{9\pi\eta_0 a^4}{8c^2 T^2 r^2 R_L}. \qquad \text{(gaussian, large } R_L\text{)} \qquad (27)$$

The resulting link loss for the monocycle waveform is,

$$L_{link} = \frac{W_{rec}}{W_{in}} = \frac{3\pi\eta_0 a^4 R_L}{8c^2 L_0^2 r^2}, \qquad \text{(monocycle, small } R_L\text{)} \qquad (28)$$

and,

$$L_{link} = \frac{W_{rec}}{W_{in}} = \frac{15\pi\eta_0 a^4}{16c^2 T^2 r^2 R_L}. \qquad \text{(monocycle, large } R_L\text{)} \qquad (29)$$

Conclusion:
Closed-form approximations for the energy link loss in a UWB radio system using electrically small dipole or loop antennas have been presented for gaussian and gaussian monocycle excitations. The utility and limitations of the Friis formula has also been discussed, and examples presented for various types of antennas. The accessibility of these results should be useful for systems engineers working with UWB radio technology.

In a general sense, the essential problem with short pulse radio transmission that differentiates it from a CW (or narrowband) system is the distortion introduced by practical transmit and receive antennas. These antennas, which form the interface between plane waves and circuitry at both the transmitter and receiver, are a direct cause of pulse distortion in a UWB radio system. Fundamentally, this is due to non-TEM (reactive) fields in the near zone of each antenna, which lead to the impedance mismatch terms noted above, as well as the radiation mechanism itself. In principle, it is possible to use pure TEM mode antennas (infinite biconical and TEM horns, for example) to achieve distortionless pulse transmission and reception, but this is of limited practicality because of the large sizes required for such antennas to avoid end reflections.

References:

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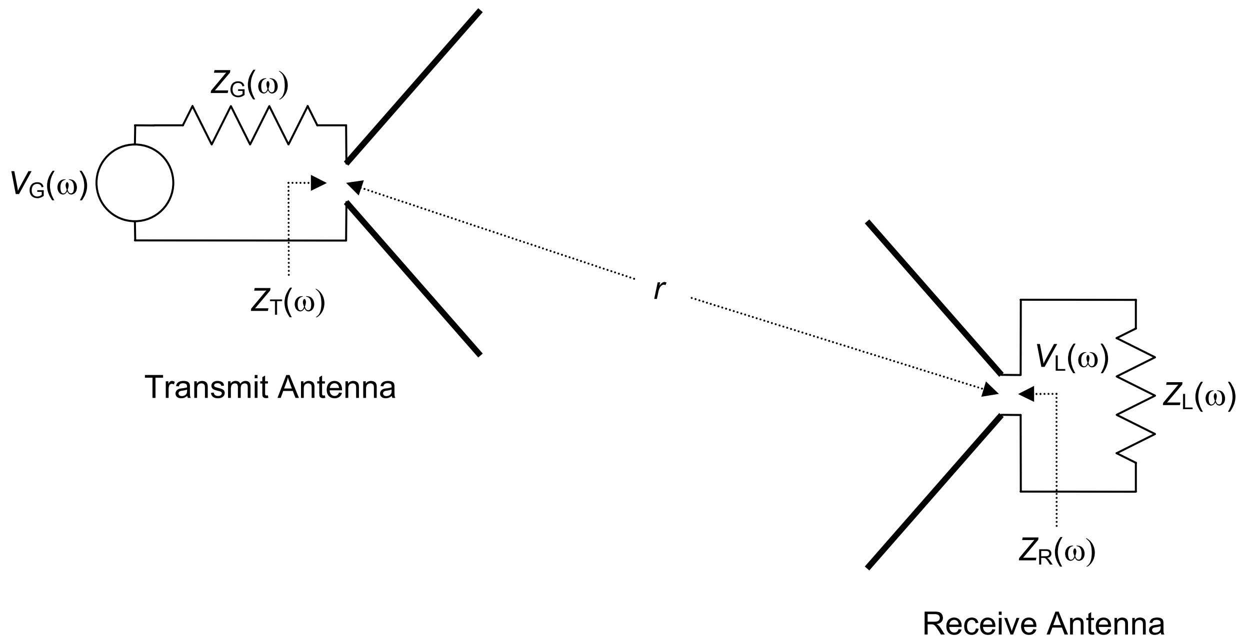
Figure Captions:

Figure 1. Frequency domain model of transmit and receive antennas for a UWB radio system.

Figure 2. Comparison of closed-form versus exact link loss (multiplied by r^2) for an UWB system using two electrically short dipoles with a gaussian generator waveform, versus receive load resistance. Dipole length = 1.0 cm, dipole radius = 0.02 cm, $Z_G = 50\,\Omega$, $T = 4.42\times10^{-10}$ s.

Figure 3. Comparison of closed-form versus exact link loss (multiplied by r^2) for an UWB system using two electrically short dipoles with a monocycle generator waveform, versus receive load resistance. Dipole length = 1.0 cm, dipole radius = 0.02 cm, $Z_G = 50\,\Omega$, $T = 4.42\times10^{-10}$ s.

Figure 4. Transfer function magnitudes versus frequency for an UWB radio system using three different transmit/receive antennas. (normalized by r)



$Z_G(\omega)$
$V_G(\omega)$
$Z_T(\omega)$
Transmit Antenna
r
$V_L(\omega)$
$Z_L(\omega)$
$Z_R(\omega)$
Receive Antenna

Figure 1.

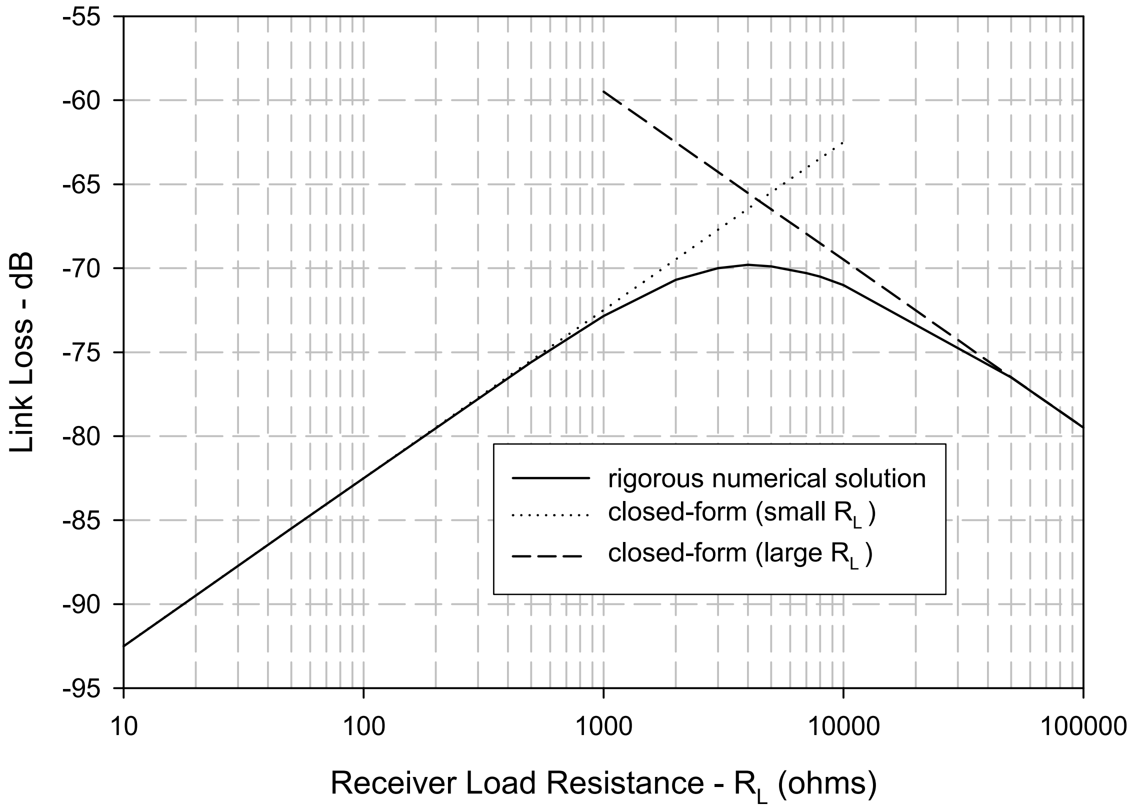
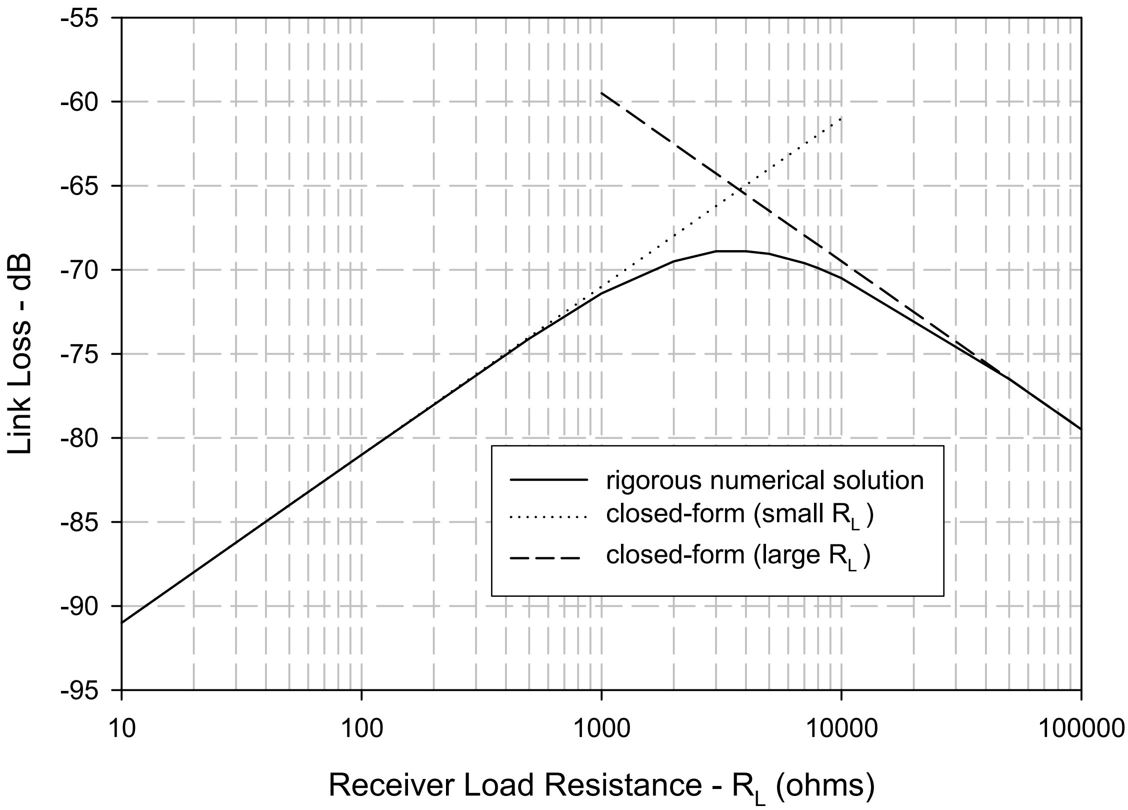

-55
-60
-65
-70
-75
-80
-85
-90
-95
Link Loss - dB
10
100
1000
10000
100000
Receiver Load Resistance - R_L (ohms)
rigorous numerical solution
closed-form (small R_L)
closed-form (large R_L)

Figure 2.



-55
-60
-65
-70
-75
-80
-85
-90
-95
Link Loss - dB
10
100
1000
10000
100000
Receiver Load Resistance - R_L (ohms)
rigorous numerical solution
closed-form (small R_L)
closed-form (large R_L)

Figure 3.

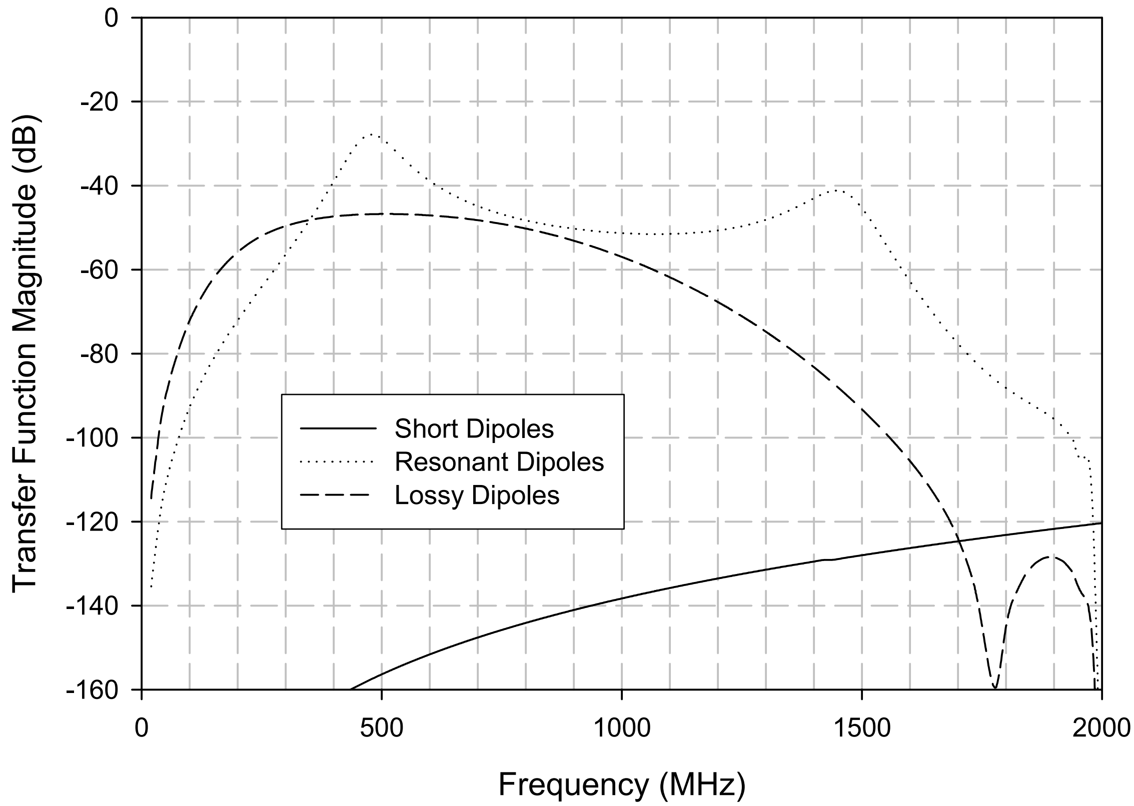

0
-20
-40
-60
-80
-100
-120
-140
-160
0
500
1000
1500
2000
Transfer Function Magnitude (dB)
Frequency (MHz)
Short Dipoles
Resonant Dipoles
Lossy Dipoles

Figure 4.