Two Way UHF Wireless Transceiver Path Loss Predictions With Communication Integrity Expectations

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30 August 2002

Abstract

This article examines the free space path loss of two commercially available transceivers operating in the Ultra High Frequency (UHF) band of 902 through 928 Megahertz (MHz). The well established Friis Transmission Formula is utilized to combine transmitted power, antenna gains of both the transmitting and receiving units, wavelength of operation and physical separation of the two stations (sender and receiver) to provide received power available at the receiving unit input terminals. Further, power reduction due to impedance mismatch is discussed along with predictions of excess signal attenuation due to walls, office partitions, floors and general building configurations. These predictions of excess signal attenuation constants are extracted from an excellent article of the Institute of Electrical and Electronics Engineers (IEEE) Transactions on Antennas and Propagation (Vol. 40, No. 2, Feb 1992). Excess signal attenuation constants presented here are the results of actual experimental testing at the mid band frequency of interest (914 MHz). The present article offers basic antenna theory as a means of understanding the Friis transmission formula. Antenna properties are discussed as contributing factors to establishing a desired communications link. An analogy to optical signal transmission is offered along with an interpretation of Receiver Operating Characteristics (ROC) which relate the probability of successful message reception to a defined signal to noise (S/N) environment. This article does not discuss the security of spread spectrum technology utilized by the subject tranceivers. This concern is addressed elsewhere in other literature. A compendium of references for Radio Frequency (RF) signal loss calculations and ROC predictions accompany this article.

Background

Recognition Source LLC is the commercial manufacturer of a high technology security door lock system, that employs an Ultra High Frequency (UHF) radio link between an electronic lock mechanism at the point of intended entry and a remote transceiver which verifies a digital code sent to it by the transceiver at the door. In essence, a person wishing to gain access to a facility will present a coded card (a key) to the door lock transceiver. This unit is referred to as the WAPM (Wireless Access Point Module). The WAPM interrogates a remote Wireless Panel Interface Module (WPIM) which cycles through an executive program loop and monitors the frequency band of 902-928 MHz. This radio frequency (RF) link is of a spread spectrum nature and the digital baseband information is of a Minimum Shift Keying (MSK) nature. Minimum Shift Keying may be considered a form of continuous phase FSK (Frequency Shift Keying) [1,2] with baseband information presented to varactors which accomplish the necessary frequency modulation. The choice of modulation selected here is judicious as the WPIM should be capable of multiple interrogations [2]. Upon verification of a valid interrogation code the WPIM transmits the necessary modulated code back to the WAPM to affect delatching of the door lock. In the event that a WAPM interrogation goes unacknowledged by the companion WPIM the WAPM retransmits a new interrogation on the same channel pending a randomized time delay. The manufacturer of this system has taken every effort to assure that system operation is transparent to the operator and that a communications link of high integrity is established. A unique feature of the system is its "3 DBM linking process." At the time of installation the system establishes the correct base line of RF power output necessary for a quality communications circuit. From this level the system automatically increases power by a factor of two assuring an enhanced margin for an "over the air" handshake. This paper intends to underscore the efficacy of RF data transmission as provided by Wyreless Access products offered by Recognition Source LLC.

Introduction

In order to understand and appreciate the nature of electromagnetic wave propagation (in this instance radio frequency waves) it is necessary to introduce the concept of an isotropic radiator. Simply stated, an isotropic radiator emits equal amounts of power into all of three dimensional space. Often this type of power source is referred to as a "point" source. Imagine the sun at the center of our solar system. In many ways this is a typical model of an isotropic emitter. The total energy radiated by such a generator may be found by the time integral of the total power radiated; that is, the average power multiplied by time. Clearly, the total power must be integrated (summed) over all space (all directions). As a matter of convenience we imagine that the point source is located at the center of a closed Gaussian surface. In other words, we envision the isotropic radiator is surrounded by a fictitious "bubble" known as the Poincare sphere. Again, the source is at the center of this sphere. It is well known that the area of a sphere is $4\pi r^2$. Here, just as a circle contains 2π radians (360 degrees) the sphere contains 4π steradians and r is the distance from the sphere center to the sphere itself.

Knowing the power emitted by an isotropic radiator it is a simple matter to calculate the power density (S) at the surface of the sphere:

$$S = \frac{P}{4\pi r^2} \qquad \text{EQ. 1}$$

"S" is in watts per square meter if "r" is measured in meters. "P" is in watts. Actual antennas designed for RF applications are never isotropic. In fact, they exhibit a "radiation pattern." In some directions they provide strong power densities and in other directions they may not radiate at all. Often times an antenna is quoted as having "gain." But this term can be misleading since it is not uncommon to think of an amplifier as having gain. An amplifier is an active system and consequently more power exits its output port than the power supplied to the input port. An antenna on the other hand is a passive device. It can not radiate more power than what it accepts at its excitation terminals. However, consider an observer receiving a signal of a specific power density at some distance from an isotropic source. If the observer is told that the transmitter is 100.0 watts then the observer associates that power level with the received power density. Now, if the transmitting antenna were reconfigured so that the same 100.0 watts were directed not into the Poincare sphere but totally into the hemisphere that faces the receiver, the observer believes that the transmitter power was doubled. in other words, power "wasted" into the backside hemisphere has been redirected (or "folded") into the forward hemisphere to augment the existing radiation. In fact, the initial 100.0 watts has been radiated through 2π steradians (not 4π steradians represented by the Poincare sphere). The radiating antenna in this case is said to have an algebraic gain of 2.0.

For antenna analysis and design work it is convenient to speak of a "solid angle" through which power is directed. In Fig. 1 the increment of the solid angle ($d\Omega$) is shown

$$d\Omega = \sin\theta \, d\theta \, d\phi$$

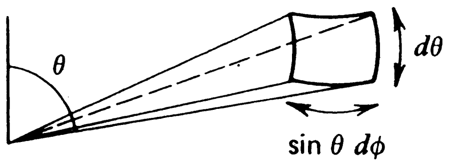


Fig. 1

θ is measured in the vertical plane (top to bottom yields π radians). ϕ is measured in the horizontal plane (circle yields 2π radians).

It is clear that θ (theta) exists over π radians and ϕ (phi) exists over 2π radians. Thus

$$\Omega = \iint \sin\theta d\theta d\phi$$

$$= 2(-\cos\theta)\Big|_0^{\pi/2} \phi\Big|_0^{2\pi} = 4\pi \; steradians$$

Antennas

Antennas are characterized by the following:

A) Pattern
B) Gain
C) Efficiency
D) Impedance
E) Aperture

The simplest of antennas is a half wave wire dipole with two terminals at the midpoint of the wire. This antenna has the electric field pattern depicted in Fig. 2. The power density pattern is shown in Fig. 3. The wire runs vertically in both figures and these patterns should be regarded as symmetric when rotated about the vertical axis. This simple structure possesses an algebraic gain of 1.64.

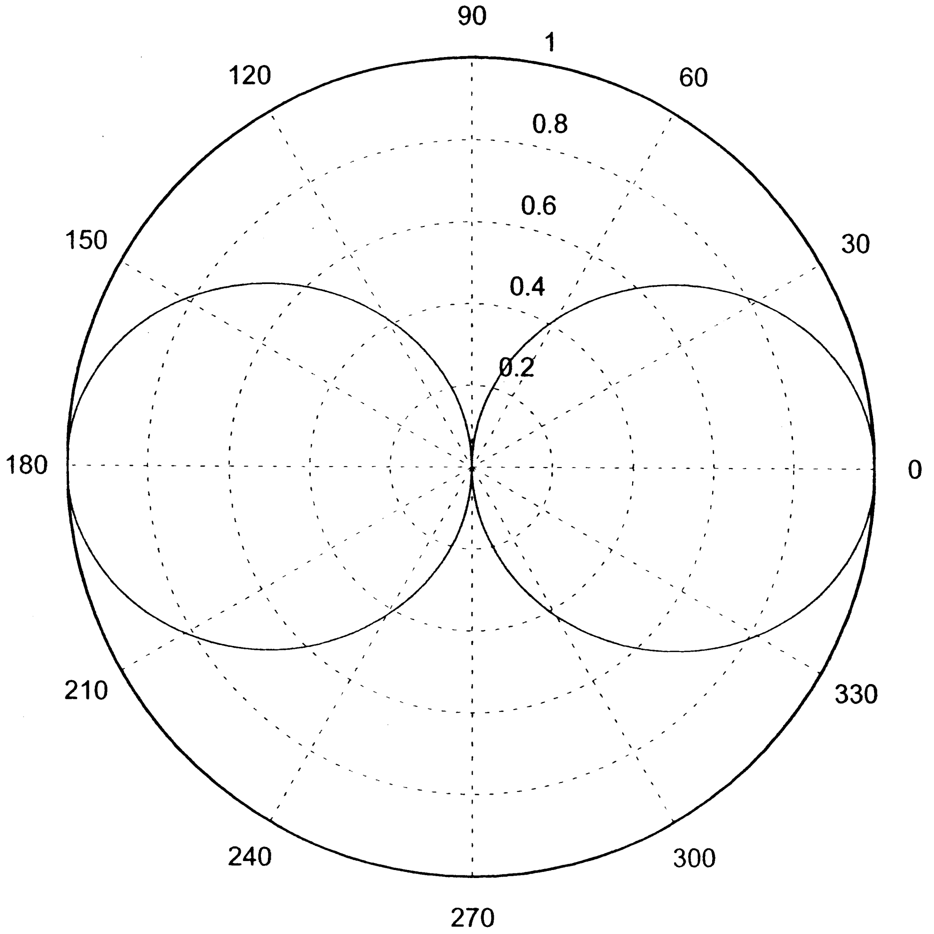
Since power density levels may vary over orders of magnitude it is appropriate to express power density (and power) ratios on a logarithmic scale. Consider two power density levels at S_2 and S_1. We define their ratio in decibels:

$$dB = 10 \log_{10} \frac{S_2}{S_1} \qquad \text{EQ. 2}$$

If S_1 were the power density level due to an isotropic radiator then the "gain" of the halfwave dipole would be:

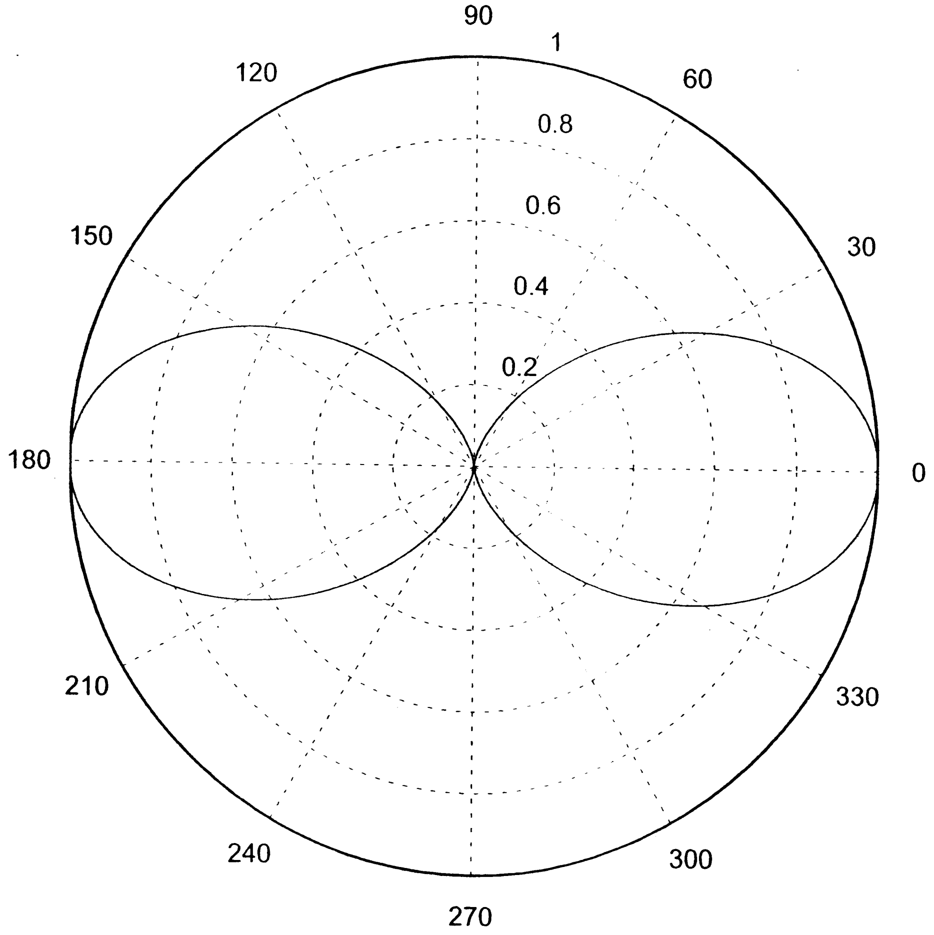
$$10 \log_{10} \frac{1.64}{1.00} = 2.15 dBi \qquad \text{EQ. 3}$$

This is read: "2.15 decibels above isotropic." It is noted that the pattern and gain are inextricably linked. It should also be noted that for this example the antenna is "dead" at the top and bottom. That is, it exhibits "nulls." Maximum power density occurs at the center of the "main lobes." The power density falls to one half of this maximum at 34 degrees up and 34 degrees down. The half wave dipole exhibits a 78 degree half power beam width (HPBW).



Electric Field Pattern, Half Wave Dipole

Fig. 2



Power Density Pattern, Half Wave Dipole

Fig. 3

Physically small antennas used at 902-928 MHz typically have ohmic resistive losses that are many orders of magnitude below their so called "radiation resistance." Thus, we may regard these radiators as essentially 100.0 percent efficient. Efficiency here is understood in the conventional sense: total power radiated divided by the total power supplied to the antenna. We do not expect these transmitting antennas to heat up. With an efficiency of 100 percent the terms "directivity" and "gain" are used interchangeably.

All antennas have an electrical impedance. We recognize that currents in conductors give rise to magnetic fields with the consequence of inductive reactance. Also, electric fields produced give rise to capacitive reactance. However, these phenomena are only responsible for static near field effects. While they are crucial to antenna operation, they are not contributors to the launched electromagnetic wave (EMW) which is a "far field" observation. In fact, the local near field amplitudes and the EMW amplitude equilibrate at approximately one-sixth of a wavelength from an elementary radiator. This would be about 2.2 inches at 900.0 MHz. Distances of several wavelengths from the antenna clearly represent the far field (EMW) zone. Properly designed antennas are usually adjusted in length so that the electrical terminals exhibit pure real (resistive) impedance with zero reactance. For example, if a dipole is "cut" to a length which is exactly a half wavelength according to EQ. 4. we find a terminal impedance of 73+j44 ohms (+j44 is an inductively reactive component):

$$c = f\lambda \qquad \text{EQ. 4}$$

The speed of light is "c." The frequency is "f" and the free space wavelength is "λ." A "wavelength" on any antenna structure is shorter that the wavelength in free space; a consequence of the wave velocity being slower on the radiator than free space. If the antenna is shortened by 5.0 percent then its impedance is real at about 70.0 ohms. If the antenna is shortened further it then exhibits a lower real component accompanied by a capacitive (negative reactance) term.

The real component of the impedance is referred as to the antenna's radiation resistance if the connection occurs at the position of maximum current (often this location is termed the "current loop"). This radiation resistance (R) accounts for the electrical power accepted and radiated by the antenna:

$$Power\ (in\ watts) = i_{RMS}^2 R \qquad \text{EQ. 5}$$

Here, i_{RMS} is the current fed to the antenna and R is the radiation resistance of the radiator. As a reminder, all power density values integrated over the Poincare sphere must equal the power given in EQ. 5. This is a consequence of 100 percent efficiency.

Up to this point the discussion has focused on the transmitting antenna. But antennas possess a reciprocal nature. To put it a different way: a good transmitting antenna makes for a good receiving antenna. The question arises as to how much of the wave power is captured by a receiving antenna? The question suggests that the antennas must possess an "aperture." The validity of this suggestion comes about from a receiving antenna immersed in a flux of the electromagnetic wave. Understandably, if the wave's power density is in watts per square meter and the output terminals of the antenna supplies power (in watts) to a load, the antenna must in fact have an equivalent "area" or properly, aperture. Some antennas have a physical aperture. A good example is the "horn" antenna. However, the physical aperture is usually not the "effective" electrical aperture. This is because the EMW produces currents on the antenna structure that are not uniformly distributed. Numerous textbooks on the subject of antennas [3, 4, 5, 6, 7] relate the gain (G) of an antenna to its effective aperture (Ae):

$$G = \frac{4\pi}{\lambda^2} Ae \qquad \text{EQ. 6}$$

It is seen that high gain antennas have large apertures. The effective aperture of the halfwave dipole example may be calculated:

$$\frac{G\lambda^2}{4\pi} = \frac{1.64\,\lambda^2}{12.566} = 0.13\lambda^2$$

This value of effective aperture for the halfwave dipole describes a rectangle which is approximately $0.25\lambda \times 0.5\lambda$.

It is readily seen that the intercepted power of a receiving antenna maybe calculated by a product of terms (EQ. 7):

$$P_R = P_T G_T \frac{1}{4\pi r^2} Aer \qquad \text{EQ. 7}$$

Here, P_R=received power, P_T=transmitted power, G_T=gain of the transmitting antenna, r=distance of separation and Aer=effective aperture of the receiving antenna.

Using EQ. 6:

$$P_R = P_T G_T \frac{1}{4\pi r^2} \frac{\lambda^2}{4\pi} G_R$$

or

$$P_R = P_T \frac{G_T G_R \lambda^2}{(4\pi r)^2} \qquad \text{EQ. 8}$$

G_R is the gain of the receiving antenna.

The Friis Transmission Formula

In the previous section the received power was seen to be the products of transmitted power, transmitter antenna gain, receiving antenna gain and the square of the wavelength divided by the square of 4πr. This is the Friis Transmission Formula in algebraic form. It is convenient to re-represent this relationship in logarithmic (decibel) format:

$$P_R(dBm) = P_T(dBm) + G_T(dB) + G_R(dB) - 20\log r(km) - 20\log f(MHz) - 32.44 \qquad \text{EQ. 9}$$

Here dBm refers to decibels above one milliwatt with r expressed in kilometers and f in Megahertz. If we wish to express r in feet EQ. 9 can be rewritten:

$$P_R(dBm) = P_T(dBm) + G_T(dB) + G_R(dB) - 20\log r(ft) - 20\log f(MHz) + 37.88 \qquad \text{EQ. 9A}$$

All forms of the Friis Transmission Formula do not account for antenna misalignment or impedance mismatches which may occur between an antenna and associated electronics. The former concern deals with antennas that are radiating or "looking" in a direction which is off to the side of a main power lobe of the pattern. This issue, however, is addressed through a gain reduction factor interpreted directly from the antenna power density pattern. The second concern (impedance mismatch effects) is examined from a voltage division view point. A properly designed antenna interface system exhibits little to no effect of impedance mismatch. Maximum power is transferred when the source impedance and the load impedance are complex conjugates. That is, the real impedance components are equal and all reactance is canceled. From the perspective of the transmitter or receiver, the power reduction factor (q) for mismatched real impedance components can be expressed as:

$$q = \frac{4R_A R_T}{R_T^2 + 2R_T R_A + R_A^2} \qquad \text{EQ. 10}$$

R_A is the antenna radiation resistance and R_T is the source resistance of the transmitter; q may be any value between zero and one. Expressed in decibels:

$$q(dB) = 6 + 10\log R_A + 10\log R_T - 20\log(R_A + R_T) \qquad \text{EQ. 11}$$

It is seen immediately that even in a mismatched case of $2R_A=R_T$ or $R_A=2R_T$ the incurred loss is only about 0.5 dB.

System RF Power Budget Example

The RF receiver portion of Wyreless Access products is quoted as having a sensitivity corresponding to a level of -90.0 dBm. This means that its detector can discern an input signal level of only 1.0 picowatt up against its own internally generated noise. This does not mean to say that with an input power level of 1.0 picowatt the detector will not make an error of discrimination between the desired signal and noise components. Higher levels of the desired signal are necessary to raise the probability of accurate discrimination on a bit by bit basis to an acceptable level. This matter is discussed in a later section on Receiver Operating Characteristics (ROC).

Consider the typical RF transmitter power level of 0.23 watts. Some of this power is distributed amongst spurious emissions which are significantly below (orders of magnitude) the desired in band signal. Thus, it is not unreasonable to take the in band power level at 0.23 watts. Actual antenna pattern measurements (E. F. Electronics Co. Report, Date 3-14-02) demonstrated that the current WAPM antenna possesses gains between 4.57 dBi and 3.22 dBi. The variation in gain quoted here depends on antenna orientation to the main lobes which are quite wide (≈80 degrees). The same report examines the present whip antenna utilized by the WPIM with gains between 3.87 dBi and 1.7 dBi. Again, the gain variation is dependent on antenna orientation to the main lobes which are also quite wide (between 80 and 110 degrees). A remote antenna accessory is available to effectively place radiation from the WPIM closer to the WAPM. Maximum gains for this novel "c" antenna extend between 2.62 dBi and 3.61 dBi. For this antenna half power beamwidths provide coverage of 80 degrees and beyond (orientation dependent).

Assuming a physical separation between the WAPM and the WPIM of 200.0 feet we utilize EQ. 9A to determine the input power to the receiver. With minimum antenna gains (3.22 dBi and 1.7 dBi) the system RF power budget is calculated:

Transmit Power	0.23 watts	=	+23.6 dBm
Xmit Ant Gain	G_T	=	+3.22 dBi
Rcvr Ant Gain	G_R	=	+1.70 dBi
Distance	-20log(200)	=	-46.02 dB
Frequency	-20log(914)	=	-59.22 dB
Constant	+37.88	=	+37.88
	Total		-38.84 dBm

Notice this is 51.16 decibels above -90.0 dBm (receiver sensitivity at noise floor). -38.84 dBm represents a power level of 0.13 microwatts or more than one hundred thousand times the power necessary for the receiver to discern the presence of an intended signal. To put this into perspective a normal voice band communication channel (telephone) is only required to maintain noise margins of about 30 to 35 decibels. Higher margins (up to 50 dB) may be required to guard against displaced channel (sideband) components competing for detection. This issue is addressed under the section of "Dynamic Channel Switching." The 30 to 35 decibel value stated for voice communication is comparable to the separation between left and right channels on an FM stereo broadcast system. In many cases no more than 25.0 decibels separation is achieved.

A consideration, however, which must be addressed deals with noise injected into the receiver by the antenna. When an antenna is connected to a receiver it is common practice to represent it by a resistor which is "matched" to the receiver input impedance. The antenna temperature represents the effective noise of the ambients which include atmospheric phenomena, Galactic noise and man made interference. Since every installation of Wyreless Access products is unique, man made noise can only be ascertained through survey. However, naturally occurring noise is predictable, and at the ultra high frequency of interest it has been shown that for the intended application this phenomena is negligible:

$$\text{noise power} = N = kTBG \qquad \text{EQ. 12}$$

Here k=Boltzmann's constant=1.38×10^{-23} joules per degree Kelvin; T=temperature in degree Kelvin; B=bandwidth in Hertz; and G=antenna gain. It is convenient to calculate EQ. 12 in decibel format. In the 900 Mhz frequency region noise from the unobstructed sun is comparable to aggregate Galactic contributions. Using the black body radiation of the earth, T is taken to be 310 degrees Kelvin [1]:

Boltzmann'sConstant (dBm/K/Hz) = -198.60

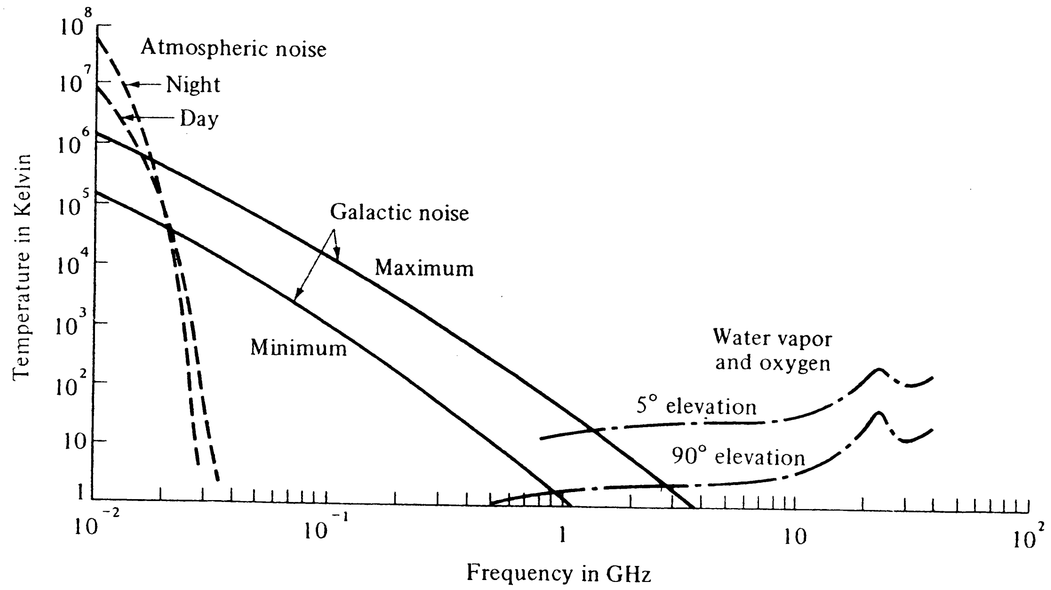
Total Noise Temperature (dB above 1° Kelvin) = +24.90

Bandwidth (dB above one Hertz) = $10\log(928-902)(10^6)$ = +74.15

G_{MAX} = +4.57

Total Antenna Noise Power (dBm) = -95.0 dBm

Notice that this is a worst case calculation using the highest antenna gain. Yet this noise contribution is well below (-5.0 dB) the receiver noise floor. Galactic sky noise temperature is significantly lower than the 310 degrees Kelvin value used in the above calculation [8]:



Average Sky Noise Temperature From [8]

Fig. 4

Excess Attenuation

In the foregoing sections, the Friis Transmission Formula predicted only free space path loss. In other words, the receiving and transmitting antennas faced one another unobstructed. It is well understood that if the antennas are not "boresighted" (i.e., face off with some angle less than optimum) receiver power is reduced. Further, it is assumed that both antennas launch and receive identical wave polarization. Notwithstanding these observations, it must be realized that the antennas used in this system maintain like polarization and are relatively broad beamed so that any signal degradation which does occur from misalignment will only be on the order of a few decibels. A more important concern is "excess attenuation" provided by obstructions that may exist between the antennas. For the intended separation in space between the WAPM and the WPIM units excess attenuation is identical to that experienced by 900 MHz wireless telephones. Moreover, many of these telephones utilize lower power levels ($\approx$100.0 mW) and maintain exceptionally fine communication. They too experience excess attenuation due to obstructions and this subject has been addressed in the technical literature. Of all the documents available on the subject, a particular journal article has been underscored as quite definitive. Here we refer to the Institute of Electrical and Electronics Engineers (IEEE) Transactions on Antennas and Propagation (Vol. 40, No. 2, Feb 1992) [9]. The article is entitled: "914 Mhz Path Loss Prediction for Indoor Wireless Communications in Multifloor Buildings." The measurement frequency of this article corresponds to the frequency of interest. The authors (Seidel and Rappaport) of the cited article propose three path loss models. The first model suggests that the mean path loss ($\overline{PL}$) is simply an exponential function dependent on receiver transmitter separation. Excess attenuation is absorbed into an exponent (n) different than 2.0 as put forth in the Friis Transmission Formula. In other words:

$$\overline{PL} = K\left(\frac{d}{d_0}\right)^n \qquad \text{EQ. 13}$$

Here, K is a constant, d_0 is a reference distance (1.0 meter) and d is the actual separation between receiver and transmitter. This suggestion modifies the Friis Transmission Formula as follows:

$$P_R(dBm) = P_T(dBm) + G_T(dB) + G_R(dB) - n\text{x}10\log R(ft) - 20\log f(MHz) + 37.88 \qquad \text{EQ. 14}$$

Experimental test data indicates that "n" could be between 1.81 and 5.04. The wide range of "n" comes about because of multiple obstructions and through floor attenuation of many floors. Also, the reason for an "n" value below 2.0 must be regarded as a multipath reflection contribution which provide a stronger signal than predicted by the Friis calculation. This approach is not recommended for our analysis since specific field contour plots of actual signal strengths are peculiar to the test configuration.

The second model offered by the cited article addresses interfloor attenuation through "Floor Attenuation Factors" used to modify the first approach. This model separates the same floor attenuation (with "n" between 2.76 and 3.27) from the Floor Attenuation Factor. Thus, total attenuation is a combination of the two.

The third model more appropriately addresses the excess attenuation offered by individual obstructions. A distinction is made between soft (fabric covered plastic) partitions and hard (concrete/plasterboard) walls. Total excess attenuation is then calculated by knowing the number ("p") of "soft" obstructions in the path and the number ("q") of "hard" obstructions in the path:

$$\text{Total EA} = p \times AF(\text{soft}) + q \times AF(\text{hard}) \qquad \text{EQ. 15}$$

Here, EA refers to excess attenuation and AF refers to attenuation factor in decibels. Thus, total EA must be subtracted from EQ. 9A giving the total power received:

$$P_R(dBm) = P_T(dBm) + G_T(dBm) + G_R(dB) - 20\log r(ft) - 20\log f(MHz) + 37.88 - EA(dB) \qquad \text{EQ. 15}$$

The cited article examines four buildings: a) a grocery store (one floor); b) a retail store (one floor); c) a five floor office building; and d) a four story office building.

Single soft obstructions yielded EA values between 0.92 dB and 1.57 dB with a nominal value of 1.39 dB. Hard obstructions yielded EA values between 1.99 dB and 2.45 dB with a nominal value of 2.38 dB. Referring back to the section "System RF Power Budget Example" we can now modify the received power by knowing the EA values. Considering two hard obstructions and one soft obstruction:

-38.84 dBm

-2(2.45) (worst case)

<u>-1(1.57) (worst case)</u>

-45.31 dBm

This calculated value yields a 44.7 dB margin above the noise floor. Again, an excellent signal to noise (S/N) ratio is achieved.

For specific details of experimental procedure used and data compilation methods used by Seidel and Rappaport the reader is referred to this excellent article.

<u>An Optical Analogy for Path Loss</u>

Visible light is electromagnetic radiation. Therefore an analogy is possible. Consider a 25 watt commercially available incandescent light bulb which operates from 120 volt AC power. A modest survey of readily available lamps reveal luminous flux outputs between 180 lumens and 235 lumens. The sad truth about incandescent bulbs is that they possess very low efficiencies. Most of the supplied electrical power goes into the production of heat. From the Handbook of Engineering Fundamentals [10]:

1 watt = 680 lumens

Thus, for a nominal 25 watt lamp of 200 lumens:

$$P = \frac{200}{680} = 0.29 \ \textit{watts (optical output)}$$

This optical power is quite close to the RF optical output power supplied by a Wyreless Products transmitter. If this lamp is illuminated at a distance of 200 feet from an observer its operation is easily discernible even in a moderately lit atmosphere. The bulb is approximately an isotropic source. For a human observer at 200 feet (61 m) we have:

Approximate pupil radius = 0.0011 m

Then,

$$P_{25} = \frac{0.29}{4\pi(61)^2} \text{x} \pi \text{x} (0.0011)^2 = 2.36\text{x}10^{-11} \ watts$$

or in dBm:

$$P_{25} = 10 \log \frac{2.36\text{x}10^{-11}}{10^{-3}} = -76.3 \, dBm$$

Now, it must be realized that the human eye is quite sensitive and a natural question to put forth should be directed at electrical detection. The Friis Transmission Formula is applicable even though RF and optical frequencies are widely separated. A small silicon photodiode has a very small signal capture area and consequently it must be prefaced by a lensing system. Consider such a system with an effective circular aperture of only one inch (1/39.37 m) radius. For an optical wavelength in the red band (630 nanometers) we have [3]:

$$G_R = \frac{4\pi}{\lambda^2} Ae \quad , Ae = effective\ aperture\ (area)$$

$$= \frac{4\pi}{(630\text{x}10^{-9})^2} \pi \left(\frac{1}{39.37}\right)^2$$

$$= 6.4\text{x}10^{10}$$

or in decibels:

$$G_R = 10 \log 6.4\text{x}10^{10} = +108 \, dB$$

As expected, the short optical wavelength results in the lens/diode receiver achieving a very large gain. This is offset in the Friis Transmission Formula by a large attenuating factor related to the high frequency.

$$\frac{c}{\lambda} = f, \ where\ c = speed\ of\ light$$

or

$$f = \frac{3\text{x}10^8}{630\text{x}10^{19}} = 4.76\text{x}10^{14} \ Hz$$

$$= 4.76\text{x}10^8 \ MHz$$

At a separation of 200 feet we use the Friis Transmission Formula:

P_T(dBm)	=	+24.6dBm	(0.29 watts)
G_T	=	0.0	(isotropic)
G_R	=	+108.07	
-20log200	=	-46.02	
$-20\log 4.76\times10^8$	=	-173.55	
+37.88	=	+<u>37.88</u>	
Total		-49.02 dBm	(Received Optical Power)

A typical silicon photodetector (PIN diode) has nominal characteristics as follows [11]:

0.5 A/W	Responsivity
2×10^{-9}A	Dark Current (zero optical signal)

The optical power delivered to the diode from the 25 watt bulb is:

$$-49.02\text{dBm} = 1.25\times10^{-8} \text{ watts}$$

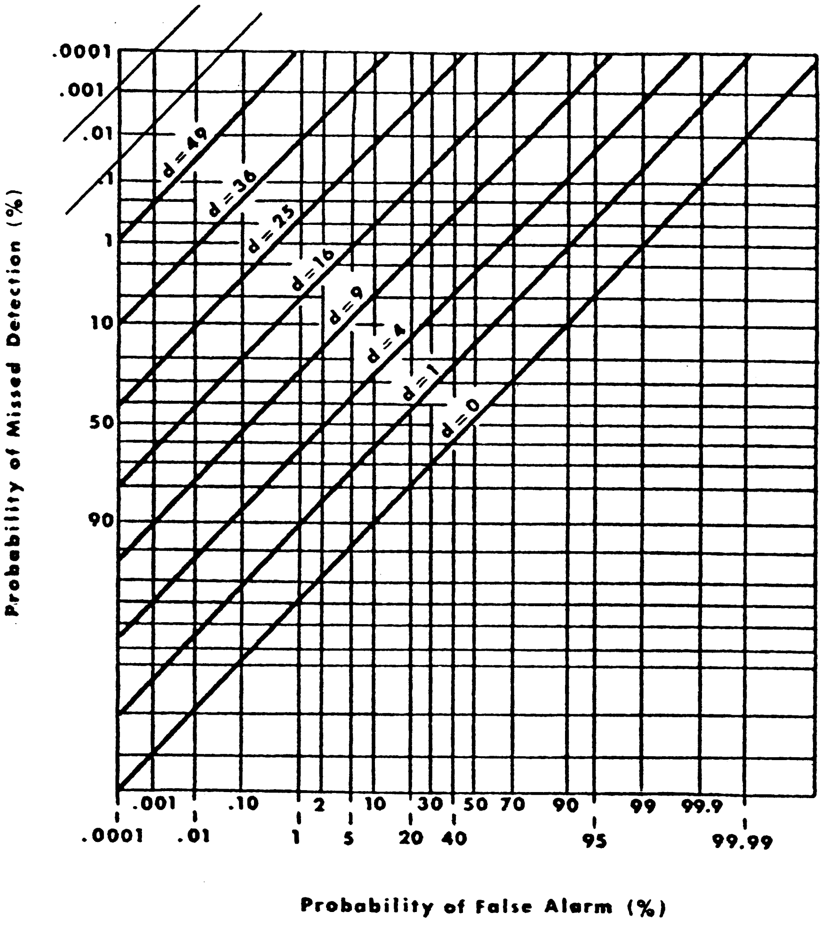
This corresponds to a current shift of 6.25 nanoamps or an increase of only 5 dB above the zero optical signal level. The importance of this analogy is that the -49.02 dBm signal is easily detected yet this power level is approximately 10 dB below the received RF power for the aforementioned Wyreless Acess products.

Receiver Operating Characteristics (ROC)

All digital receivers employ a detector which essentially establishes a decision threshold for determining whether a "1" or "0" has been received. In properly designed systems this threshold is set such that the missed detection probability is comparable to the false alarm probability [12]. Obviously, the signal to noise (S/N) ratio will establish the ambiguity of the detector's decision. This observation is virtually independent of the detector type. Peterson and Birdsall [13] give a convenient graphical method of determining the probability of missed detection and the probability of false alarms predicated on the signal to noise ratio (Fig. 5).

Both missed detection and false alarms are of concern since either situation results in a failure to establish a viable hand shake between the WPIM and the WAPM. These so called "false alarms" refer to the detector of either unit supplying an erroneous bit somewhere in the code stream. Either type of failure is of a fail safe nature and results in reinterrogation. In this graph, the family of lines represent corresponding S/N ratios. The highest S/N ratios are in the upper left hand corner with d=49:

$$10 \log 49 = +16.9 dB$$



$$d = \frac{S}{N} \text{ algebraic ratio of powers}$$

Receiver Operating Characteristics, Probability [12, 13]

Fig. 5

In Fig. 5 each line is spaced uniformly with respect to the entire family and therefore it is possible to construct two additional lines beyond d=49. Beyond the line labeled d=49 one can draw:

	d=64	or	10log64	=	+18.0 dB
and	d=81	or	10log81	=	+19.0 dB

Since the probability of missed detection is set equal to the probability of a false alarm, the single bit error rate (BER) can be predicted. The integrated spread spectrum transceiver utilized within Wyreless Access products specifies the following bit error rates without the RF link:

10^{-5}	for	19.0 S/N
10^{-4}	for	17.5 S/N
10^{-3}	for	16.5 S/N

When these figures are compared to data from Fig. 5 it is realized that excess noise within the transceiver has been accounted for:

	BER Fig. 5 Predic.	BER Transceiver Spec.
19.0dB = 19.0 dB	5×10^{-6}	10^{-5}
18.0 dB ≈ 17.5 dB	5×10^{-5}	10^{-4}
16.5 dB ≈ 16.9 dB	5×10^{-4}	10^{-3}

All S/N values quoted here are extremely modest. Based on prior RF received power calculations one concludes that a 19.0 dB S/N ratio can be established.

At present, Wyreless Data Products utilize a total of 70 bytes (560 bits) for data transmission:

Preamble:	22 bytes
Header:	1 byte
Utility Data:	9 bytes
Card Data:	33 bytes (max)
Trailer:	5 bytes

The probability of a message being received correctly can be calculated:

Assuming +19 dB S/N:

$$1-10^{-5} = 0.99999$$

$$(0.99999)^{560} = 0.9944 = 99.44 \text{ percent}$$

This, however, is <u>not</u> the system availability seen by the operator. Parity bit checking is accomplished electronically and a failure to receive a proper message results in an immediate retransmission (retry). In total, the system will accomplish up to three such transmissions before a failure is reported. Retries are totally transparent to the operator and result in enhanced system availability. It is thus possible to calculate the number of usage attempts or messages (N) necessary to achieve a 50 percent probability of a single failure (non-recognition of a valid challenge):

$$(1 - (1 - 0.9944)^3)^N = 0.5$$

$$N \log(1 - (1 - 0.9944)^3) = \log 0.5$$

$$N \approx 3.947 \times 10^6 \text{ messages}$$

To appreciate the significance of this number N consider the system challenged every 12 seconds, every minute of every day. In order to witness a 50 percent probability of a non-recognized *single* valid challenge an observation of a year and a half must elapse!

<u>Dynamic Channel Switching</u>

One of the most difficult problems encountered in electronic communication via RF is the situation where deterministic interference occurs on the channel in use. Unlike random noise "deterministic" signals have properties similar (center frequency and bandwidth) to the desired. In the receive mode the detector becomes confused over discriminating between interfering and intended signals. Of all devices similar in RF emissions to Wyreless Access products one needs to examine the effects of cordless telephones. The reason for this concern is that an individual could carry one of these operating devices in close physical proximity to a WPIM or WAPM module. Experiments have been conducted on typical cordless telephones to determine their power levels and spectrum signatures.

Modern cordless telephones operating in the 900 MHz region utilize a power management technique that raises transmitter power in response to weak signal situations. A sample of these telephones include the GE Model 2-91155TB handset and the Bell South MH9026BK handset. The former's center frequency was noted to be around 922 MHz with output powers between +5.0 and +17.0 dBm. The Bell South handset exhibited a central frequency of 917 MHz with output power levels between 0.0 dBm and 10.0 dBm. If these findings are contrasted against Wyreless Access products it is immediately observed that although the Wyreless Access device uses a higher power level (+20.0 to +23.0 dBm) a significant power margin does not exist. Examining the worst case (GE telephone with +17.0 dBm) yields a mere 3.0 to 6.0 dB margin. This situation does not lend itself to successful mitigation by increasing transmitted power. Rather, Wyreless Access products afford the elegant solution of Dynamic Channel Switching. Beginning at approximately 903 MHz and extending to approximately 926 MHz there exist 15 frequencies which are decomposed into 3 equally spaced channel intervals. Thus, if any interval is occupied by an interference source the system automatically selects an unused interval establishing the desired communication circuit. The effectiveness of this technique is well understood from a power-bandwidth view point. The cordless telephone emission is such that a frequency displacement of just 3 MHz from peak power frequencies shows attenuated sideband power to levels 40 dB below the maximum state emissions. Through Dynamic Channel Switching frequency spacing will exceed 3 MHz affording favorable power margins of up to 60 dB.

Summary

The analyses put forth for RF signal loss and Receiver Operating Characteristics indicate that the presented system affords reliable and transparent operation to the user. Signal margins are sufficient so that transceiver separations of 200 ft with complex obstructions provide intended performance. Moving obstructions have not been discussed because the bit transmission rate (16 microsecond per bit) is so rapid that even a vehicle moving at 60 miles per hour can be regarded as stationary. Further, such moving objects will impart doppler frequency shifts on the order of 160 Hertz which are properly accepted and processed by the wide band receivers of both the WPIM and WAPM. Severe electromagnetic environments are properly addressed by site surveys [14].

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Biography

Vincent P. McGinn was born in New York City and attended New York University (presently Polytechnic University of New York). Here he was awarded the bachelors and masters degrees in electrical engineering. Upon being commissioned in the United States Air Force he was assigned to numerous electrical engineering positions in flight controls for aircraft such as the F-4, B-52, A-7 and KC-135. He was also involved in solving electromagnetic compatibility problems at the NORAD combat operations center. At Wright Patterson AFB he developed lightning assessment tests for fly by wire aircraft, served as world wide avionics officer for the F-15 and headed up the avionics analysis of the Soviet MIG-25 Foxbat. In 1972 he attended the University of Southern California and is a graduate of the Aerospace System Safety program. Later McGinn was assigned to develop RAS/RAM structures for low observable aircraft at Rockwell International. He has had numerous teaching positions at several universities and was awarded an M.S. (Astronomy) and Ph.D. (Electrical Engineering) from the Pennsylvania State University. Dr. McGinn is a commercial helicopter pilot with FAA airframe mechanic certification. He is a registered professional engineer in the states of Illinois and Ohio. He holds several domestic and foreign patents on microwave and optical systems. He is also a member of the IEEE (senior member), OSA, SAE, Tau Beta Pi, Eta Kappa Nu and the national military honor society, Scabbard and Blade.

Brian Hall received his undergraduate degree in Electrical Engineering from Northern Illinois University in 2000. During that period, he developed and taught courses at NIU in hybrid circuit manufacturing techniques, which included: wire bonding, thick film printing, soldering and reflow, and laser trimming. He fabricated and tested electroluminescent panels on ceramic and polymer substrates. He reengineered, designed and constructed RF transceivers using multi-layer ceramic technology and he modeled, fabricated and tested high performance filters for satellite communications systems. He is currently developing a one-kilowatt high efficiency power supply for high-end audio amplifiers while completing his master's degree in electrical engineering with an emphasis on solid-state design.