Design Of Short-Range Radio Systems

The first installment of this four-part series offers design techniques and regulatory issues in advanced short-range radio systems.

Short-range radios, also known as microradios, have existed for several decades, primarily in the form of one-way control and security-class links. With an increase in integration and processor control, the design of these radios has become more of a system than a circuit issue. The design of these microradios is further complicated by regulatory issues, such as which carrier frequency to use, choice of modulation scheme, whether to use transmit-power averaging, and the type of antenna for a particular design. In addition, cost issues include receiver (Rx) topology, frequency-source topology, synchronization format, level of integration, baseband processing, and when to step up to two-way links or to move to industrial-scientific-medical (ISM) frequency bands. Part 1 of this four-part series will review the basics of radio-wave propagation, while Parts 2 and 3 will cover regulatory- as well as system-oriented issues and design methodologies, respectively.

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Microradios are commonly associated with consumer applications such as remote-keyless-entry (RKE) devices for automobiles and garage-door opening systems. Bluetooth represents the high end of the product range identified by the generic term "microradio." These "control-class" applications have historically been one-way systems, sometimes so cost constrained as to feature on-off-keyed (OOK) inductive-capacitive (LC) or surface-acoustic-wave (SAW)-based transmitters (Txs) that consist of little more than a single transistor oscillator that is modulated by keying its power supply with an encoder chip that can perform key-press detection and some form of rudimentary encoding. Low-cost Rxs have included an LC or SAW regenerative Rx, a topology that can be implemented with only a few transistors. Digital control has been added in recent years, to a level often featuring baseline microcontrollers such as the Microchip PIC12C509A or the Microchip KEELOQ [code-hopping encoders from Microchip Technology, Inc. (Chandler, AZ)]. Power supplies for portable-radio units now typically consist of one or two lithium (Li) coin-cell batteries.

With the availability of higher-frequency, cost-effective complementary-metal-oxide-semiconductor (CMOS) and bipolar-CMOS (BiCMOS) semiconductor processes, microradio technology is moving toward higher levels of integration. With more powerful digital control, these radio systems are poised to move beyond control applications and into network data communications and wireless data acquisition

(DAQ). As the complexity of these short-distance radio systems increases, engineers must apply standard wireless-system design techniques such as the use of a link budget.

A link budget considers Tx power, path loss, antenna gain, and Rx sensitivity when calculating radio range. The link budget is not meant as an exact calculation, but to provide desired reliabilities as a function of range and operating conditions. The difference between radio range under ideal free-space conditions and in an environment with more realistic signal degradation can be more than an order of magnitude. A suggested approach is the use of a second-order model that uses a path fade which is higher than inverse square, and the assumption of a log-normal probability distribution of signal strength with standard deviations ranging from 4 to 16 dB as a function of environment.

The mathematics for this level of link budget is simple, and will be presented here in a way that is also applicable to certification testing, where analyses are made of field strengths some distance from the device under test (DUT). A derivation can start with the effective aperture of the receive antenna, which is the ratio of the power delivered to the load to the incident RF power density. Effective aperture can be thought of as the area where a 100-percent efficient antenna captures all of the energy that would otherwise pass through the same area without the antenna. The maximum effective aperture is related to directivity, D_0, the maximum directive gain of an antenna on its main lobe and wavelength, λ, by:

$$A_{em} = \lambda^2 D_0 / 4\pi \qquad (1)$$

The directivity does not take into account losses due to mismatch and ohmic losses, so the effective aperture, A_e, is equal to eA_{em}, where e is the total efficiency. For a perfectly isotropic (omnidirectional) antenna without losses, $D_0 = 1$. The closest practical antennas to this performance are quarter-wave whips and similar designs. A quarter-wave whip shows a directivity of approximately 1.7 and efficiency losses exclusive of matching of generally less than 1 dB. The gain of the antenna varies as a function of relative orientation which, for mobile terminals, is not well-controlled and must be viewed statistically. An acceptable practice for a particular microradio application is to measure path loss at various antenna orientations and positions relative to the human body that are appropriate for that application, come up with an average loss relative to an isotropic antenna, and then lump antenna-gain variation as a function of position into the standard deviation of path loss.

Effective aperture can be used to convert the root-mean-square (RMS) field strength at the antenna into power delivered to the Rx input:

$$P_{rec} = (E_{rms}^{2} / \eta)A_e \qquad (2)$$

where:

η = the impedance of free space (377 Ω).

E_{rms} = the RMS field strength at the antenna, and

P_{rec} = the power delivered to the Rx input.

Since the power levels permitted by the Federal Communications Commission (FCC) are presented in terms of field strength, this relationship is handy for measuring fundamental and harmonic signal levels. European regulations are based on units of effective radiated power (ERP), or the power that would be radiated from a perfectly isotropic antenna which matches that received on the peak of the actual antenna's main lobe.

Note from Eq. 1 that A_e is dropping for a particular antenna type such as quarter-wave whip as the inverse square of frequency. From Eq. 2, it can be seen that if electric field is constant over frequency with A_{em} dropping over frequency, then P_{rec} must be declining with the inverse square of frequency. This is usually referred to as increasing path loss with frequency, a somewhat confusing choice of terminology, since this loss occurs even if power density is frequency independent. What is actually physically happening is that the ability to gather the power density is declining over frequency if directivity (receive-antenna type) is held constant. It is as if a smaller lens is being used to focus sunlight. This fact must be accounted for in regulatory harmonic measurements—the "free" 6-dB/octave drop due to the increase in free space path loss versus frequency (with scaled antennas) must be taken back out to calculate the field strength of harmonics correctly. The only way to hold constant or increase A_e with increasing frequency is to introduce a larger and directional antenna.

Receive power for a particular transmit power over a free space link is provided by the Friis Transmission Equation. For polarization-matched antennas that are aligned on directionality maximums this equation reduces to:

$$P_r / P_t = \left\{\left[(\lambda / 4\pi)^2\right]\right. R^n G_{0t} G_{0r} \qquad (3)$$

where:

P_r = receive power,

P_t = transmit power,

R = range (in meters),

n = the path-loss exponent (2 in free space),

G_{0t} = the gain of the transmit antenna, and

G_{0r} = the gain of the receive antenna.

These gains are the same as directivity multiplied by efficiency loss. For practical link calculations, it is helpful to massage Eq. 3 into a form giving range as a function of degrading factor "D" (the linear form of all decibel losses in a practical link from ideal), Rx sensitivity S (milliwatts are most convenient), and transmit power Pt (the same power units as S). These manipulations yield:

$$R_{max} = \left[(c / 4\pi f)^2 (DP_t / S)\right]^{1/n} \quad (4)$$

When converting from ERP to field strength, as is done in comparing US and European regulations, several other relations come in handy. The power density, S_r (in watts per square meter) of a uniform plane wave is provided in terms of RMS electric-field strength; free-space impedance, η; and effective radiated power, P_{terp}, as:

$$S_r = E^2 / \eta = E^2 / 120\pi = P_{terp} / 4\pi R^2 \quad (5)$$

The last term follows from radiated power and the area of a sphere of radius R. From this equation, it is possible to find RMS field strength, E_{RMS}, at range R in meters (ideal inverse-square propagation) and transmitted isotropic effective radiated power, P_{terp}, as:

$$P_{terp} = 0.03333 R^2 E_{rms}{}^2 \quad (6)$$

$$E_{rms} = (5.477 / R) \left(P_{terp}\right)^{0.5} \quad (7)$$

This basic compliance-oriented physics flows directly into link budgeting by taking degrading factors into account as shown in Eq. 4. An excellent source of raw data specifically for the 900 MHz ISM band is ref. 1. These data may be expected to remain approximately true for losses in the 300-to-500-MHz range, normally used for control and security applications. Depending on environment (such as indoor or outdoor, building type, range, operation between floors, etc.), the path-loss exponent changes from 2.0 for free space to a range from 1.8 to 5.0.

Safety Margin

It is also true that the received signal strength may be approximately modeled for reliability purposes as log normal, meaning that it shows a Gaussian distribution (in decibels) over a large number of samples. The standard deviation of this signal-strength variation will typically vary from 4 to 16 dB over a wide range of operating conditions. A few days of engineering time invested in properly modeling the link statistics for a particular application will pay enormous benefits in optimum system design. To use this information in product specification and system design requires the addition of a safety margin to the link budget to provide the desired reliability. This safety margin is most conveniently specified as a number of standard deviations in the statistical variation of path loss (in decibels), with a deliberately selected reliability at the maximum range. A brief example along the lines of a garage-door opener may be illustrative.

For example, assume that a Tx is operating at 416 MHz under FCC 15.231 rules (which will be reviewed in Part 2) with a transmit ERP of −15 dBm. Television harmonic interference is assumed negligible. The selected Rx shows a noise figure of 8 dB, a bandwidth of 60 kHz, and a demodulation and forward-error-correction (FEC) combination that requires 12 dB of final signal-to-noise ratio (SNR) to achieve the desired bit-error rate (BER). The Rx sensitivity is calculated to be −106 dBm. The mean transmit-power degradation due to antenna orientation and body absorption is experimentally determined to be −10 dB. Experimentation also shows that under the desired operating conditions, the link displays a path-loss exponent of 2.5 and a standard deviation in signal strength of 7 dB.

Desired Reliability

It is desirable to determine effective maximum range for a 95-percent chance of a successful transmission. From any table of a normalized Gaussian distribution, it can be seen that 1.65 standard deviations will have an area of 0.9505 under the density curve. In order to achieve the desired reliability, 1.65 × 7.00 dB = 11.60 dB is added to the link losses, providing a total required link-loss safety margin of 21.6 dB, or a degrading factor D = 0.00692. Plugging these numbers into Eq. 4 yields a 95-percent reliable range of 61 m. Reviewing the graph on p. 108 of ref. 2 shows a 99-percent reliability for any random range from 0 to 61 m (the service area). The range of this same link under free-space conditions can be calculated at approximately 2000 m, a range which would never hold up in practice.

Next month, this three-part article series will continue with an examination of regulations for short-range radio systems in the US and in Europe. MRF

REFERENCES

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FOR FURTHER READING

Constantine Balanis, *Antenna Theory,* Harper & Row, New York, 1982.