# Design Methods for Control Class MicroRadio

***This tutorial covers worldwide regulatory issues, and key system, circuit, and firmware design aspects of modern short-range radio systems, particularly those for control applications.***

**Parts 1-2 of 5 Parts**

Farron L. Dacus
RF Architecture Manager
Microchip Technology Inc.
2355 West Chandler Blvd
Chandler, AZ 85224-6199
(480) 792-7017
e-mail: farron.dacus@microchip.com

**MicroRadio or Short-range low cost radio has existed for several decades, primarily in the form of one way control and security class links. With increasingly high levels of integration and processor control, the need becomes stronger to move away from a historic focus on circuit design and towards greater levels of system design. The product designer will find that a basic understanding of these issues is very empowering in terms of properly defining the product and predicting range and reliability of operation. Examples of regulatory influenced design decisions include carrier frequency, when to use ASK and FSK, transmit power averaging, and antenna type. Primarily cost based design issues include receiver topology, type of frequency source technology, synchronization form, level of integration, required baseband processing, and when to step up to two way links or to the ISM bands. System and regulation oriented fundamental issues will be covered in Parts 1-2, with detailed design methodology given in later parts.**

Most readers will be familiar with what we here refer to as "MicroRadio" products such as vehicle keyless entry and garage door opener systems. Bluetooth represents the high end of the product range we seek to identify here by the generic term MicroRadio. The radio links of these historically "control class" systems have typically been of very simple one way form, sometimes so cost constrained as to feature On-Off Keyed (OOK) LC or SAW based transmitters that consist of little more than a single transistor oscillator modulated by keying its power supply, with an encoder chip that can perform key press detection and some form of rudimentary encoding. Receivers for this simple transmitter would usually be an LC or SAW regenerative receiver, a topology that can be implemented with just a few transistors. In recent years digital control has risen to a level often featuring baseline microcontrollers such as the Microchip PIC12C509A or the Microchip KeeLoq® code hopping encoders. Power supplies for portable units now typically consist of one or two lithium coin cell batteries.

This technology is moving towards higher integration levels, brought about by more cost effective and higher frequency CMOS and BiCMOS processes. These increasingly sophisticated products are now poised to advance well beyond control applications and into networked data communications and wireless data acquisition. A major goal of this article is to impart the viewpoint that standard wireless system design techniques and general design rigor are not only applicable but also necessary in the design of MicroRadio systems. We believe these needs will become greater as the complexity of these systems continues to increase, and as a supplier shall seek to provide suitably complete and rigorous applications support.

## Propagation Mathematics and Link Budgeting

The link budget takes the transmitter power, path loss, antenna gains, and receiver sensitivity into account in calculating range. The designer must be aware that this is not an exact calculation, and that one designs links to not show absolute range, but to give desired reliabilities as a function of range and operating conditions. The difference in expected range in an ideal free space analysis and one with the appropriate degrading factors is huge, usually more than an order of magnitude. For MicroRadio, we recommend a second order model that uses a path fade 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 of this level of link budgeting is quite simple. It shall be presented here in a way that is also applicable to certification testing, where field strengths some distance from the device under test are made. We 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. Intuitively, it represents the area where a 100% efficient antenna captures all the energy that would otherwise pass through that area. The maximum effective aperture is related to directivity $D_0$, the maximum directive gain of an antenna on its main lobe, and wavelength $\lambda$, by:

Eq. 1: $$A_{em} = \frac{\lambda^2 D_0}{4\pi}$$

The directivity does not take into account losses due to mismatch and ohmic losses, so achieved effective aperture $A_e = eA_{em}$, where e is total efficiency. For a perfectly isotropic (omnidirectional) antenna with no losses, $D_0 = 1.0$, but the closest practical antennas to such performance are quarterwave whips and similar designs. The quarterwave whip shows a directivity of about 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 given 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 isotropic, and then lump antenna gain variation as a function of position into the standard deviation of path loss.

We use effective aperture to convert rms field strength at the antenna into power delivered to the receiver input:

Eq.2: $$P_{rec} = \frac{{E_{rms}}^2}{\eta} A_e$$

Here $\eta$ is the impedance of free space (377 ohms). Since FCC allowed power levels are given in terms of field strength, this relationship is handy for measuring fundamental and harmonic levels. European regulations are based on the easier to visualize units of Effective Radiated Power

(ERP), or the power that would be radiated from a perfectly isotropic antenna that matches that received on the peak of the main lobe of the actual antenna.

Note from Eq. 1 that $A_e$ is dropping for a given antenna type such as quarter wave whip as the inverse square of frequency. From Eq. 2 we see 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 vs. frequency (with scaled antennas) must be taken back out to calculate 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 given transmit power over a free space link is given by the Friis Transmission Equation. For polarization matched antennas that are aligned on directionality maximums this equation reduces to:

Eq. 3: $$\frac{P_r}{P_t} = \frac{(\frac{\lambda}{4\pi})^2}{R^n} G_{0t} G_{0r}$$

Here $P_r$ is receive power, $P_t$ is transmit power, R is range in meters, n is the path loss exponent (2.0 in free space), $G_{0t}$ is the gain of the transmit antenna and $G_{0r}$ is 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 dB losses in a practical link from ideal), receiver sensitivity S (mW are most convenient), and transmit power $P_t$ (same power units as S). This yields:

Eq. 4: $$R_{max} = \left[ \left( \frac{c}{4\pi f} \right)^2 \left( \frac{DP_t}{S} \right) \right]^{\frac{1}{n}}$$

When converting from ERP to field strength, as is done in comparing U.S. and European regulations, several other relations are handy to have. The power density $S_r$ (watts per square meter) of a uniform plane wave is given in terms of rms E-field strength, free space impedance $\eta$, and effective radiated power $P_{terp}$ as:

Eq. 5: $$Sr = \frac{E^2}{\eta} = \frac{E^2}{120\pi} = \frac{P_{terp}}{4\pi R^2}$$

The last term follows from radiated power and the area of a sphere of radius R. From this equation we may find rms field strength $E_{rms}$ at range R in meters (ideal inverse square propagation) and transmitted isotropic effective radiated power $P_{terp}$ as given below.

Eq. 6: $P_{terp} = 0.03333 R^2 {E_{rms}}^2$

Eq. 7: $E_{rms} = \frac{5.477}{R} \sqrt{P_{terp}}$

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. 2. This data may be expected to remain approximately true for losses in the 300-500 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. It is also true that the received signal strength may be approximately modeled for reliability purposes as log normal, meaning it shows a gaussian distribution over a large number of samples with the unit of measure being in dB terms. 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 the application in question will pay enormous benefits in optimum system design, and in avoiding promising too much to the customer. To use this information in product specification and system design requires us to add 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 dB), with a deliberately selected reliability at the maximum range. A brief example along the lines of a garage door opener may be illustrative.

***Example:*** Assume a transmitter operating at 416 MHz under FCC 15.231 (later reviewed) with a transmit effective radiated power of -15 dBm. Television harmonic interference is assumed to be negligible. The selected receiver shows a noise figure of 8 dB and bandwidth of 60 kHz, and a demodulation and forward error correction combination that needs 12 dB of final signal to noise ratio to achieve the desired bit error rate. The receiver sensitivity may thus 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. We would like to determine effective maximum range for a 95% chance of a successful transmission. From any table of the normalized gaussian distribution we see that 1.65 standard deviations will have an area of 0.9505 under the density curve. To achieve the desired reliability we thus add 1.65 * 7 dB = 11.6 dB to the link losses, giving a total link loss safety margin needed of 21.6 dB, or a degrading factor D=0.00692. Plugging these numbers into Eq. 4 yields a 95% reliable range of 61 meters. Reviewing the graph in Ref. 3 (Rappaport) page 108 shows a 99% reliability for any random range from 0 to 61 meters (the service area). The range of this same link under free space conditions would be predicted at almost 2000 meters, which would of course never hold up in practice.

## Understanding the Regulations

The classical control and security class of MicroRadio is typically in the 300-500 MHz frequency range, at transmit power levels from -30 to +10 dBm, and allowing for "certification". This is a form of approval where the manufacturer has the product tested in an approved laboratory that will confirm the product meets specifications, and assists in submitting a report to regulatory agencies such as the FCC. The agency then grants approval for production, and the end customer never has to deal with any licensing issues. Both European and U.S. rules require that the antenna of the certified equipment be integrated or use an uncommon connector to prevent the user from substituting others. The regulatory requirements that MicroRadio systems must meet have a critical effect on product definition and system design.

## *U.S. Rules*

The Federal Communication Commission (FCC) rules officially govern operation only in the U.S., but in fact are adopted in varying degrees by many other nations in the Americas and the Pacific rim. There are 4 specific FCC rules of high interest to designers of control and security class MicroRadio systems, with interrelationships between the four that must be understood. These four sections are 15.209, 15.231, 15.205, and 15.35. The rules for ISM band radios are given in 15.249 for 1 mW class narrowband systems, and 15.247 for spread spectrum systems up to 1 watt.

***FCC 15.209:*** The so called "general" rule restricts the RF energy that electronic equipment may parasitically emit. The specific level of emissions is 200 uV/meter at 3 meters test range below 960 MHz, and 500 uV/meter above. These field strengths are equivalent to about -49 and -41 dBm ERP, respectively. If used as intentional radiators, these low powers restrict effective range to somewhere from a few feet to a few tens of feet.

***FCC 15.231:*** This is the major authorization for control and security class equipment. For control applications the allowed peak transmit level varies linearly from 260 to 470 MHz as given below:

Eq. 8: $E_{ss}(f) = .041667(f - 260) + 3.75$ ( control operations)

where $E_{ss}$(f) is allowed steady state field strength in rms millivolts/meter at 3 meters for control applications, and f is frequency in MHz. For frequencies above 470 MHz, the allowed power is equal to that at 470 MHz. If we use Eq. 6 to convert to power we note that the transmitted ERP range is -24 to -13 dBm. These increasing FCC allowed power levels are actually in excess of the effects of decreasing antenna aperture with increasing frequency, as shown in Fig. 1 below. The band is about 5 dB better at the high end than the low end, a fact that is apparently not well known, but one whose positive effect may be reduced by the interference of television second harmonics. The allowed harmonic levels are 20 dB below these levels except where they fall in restricted bands (below). Note that TV channel 13 is 210 to 216 MHz, and then allocation skips to channel 14 at 470 to 476 MHz, so direct TV interference is not a problem throughout this range (2nd harmonics can be a problem). The allowed rms steady state field strength for **periodic** operation (such as status reporting) not limited to control is about 8 dB lower and in addition must meet timing restrictions (except in emergencies) of having off time at least 30 times transmit time, with transmissions no more often than once per every 10 seconds. Note also under FCC 15.205 that 260 to 285 MHz falls into a restricted band and may not actually be used.

***FCC 15.205:*** This section documents the “restricted bands” (Table 2) where only spurious emissions are allowed, and where those must meet the general levels of 15.209. Above 1000 MHz averaging may be used (see below).

**Table 1:** Primary Restricted Frequencies under FCC 15.205

| Restricted Frequency Range | Impact |
|---|---|
| 240-285 MHz | No fundamental usage here |
| 322-335.4 MHz | No fundamental usage here |
| 399.9-410 MHz | No fundamental usage here |
| 608-614 MHz | Medical telemetry band, $2^{nd}$ harmonics from 304-307 must meet the general requirement in this segment |
| 960-1240 MHz | Note that 3rd harmonic of 320 to 413.333 MHz must meet general. |
| 1300-1427 MHz | 433.333 to 470 MHz must meet general with $3^{rd}$ harmonic |
| 1435-1626.5 MHz | 287 to 325.3 must meet general with $5^{th}$ harmonic. |

From these restricted bands a list may be formed of frequencies that are desirable to use, as given in Table 2.

**Table 2:** U.S. FCC 15.231 frequencies to **use** to avoid carriers, second, and third harmonics in restricted bands below 1000 MHz.

| Frequency | Comment |
|---|---|
| 285-304 | Stop at 304 to avoid placing $2^{nd}$ in medical band. |
| 307-320 | Stop to avoid $3^{rd}$ in 960-1000, note that 4th of 307-310 is restricted |
| 335.4-399.9 | Stop to avoid direct restricted. |
| 410-470 | Power tops out at 470 MHz. |

From the pure link budget point of view, the most desirable segment from the above table is the 410 to 470 MHz segment. This section may be used if 28 dB of 3rd harmonic rejection is attained. Above 432 MHz also has the virtue of avoiding TV harmonics. If a simple unfiltered loop antenna does not attain this rejection, then an option is to use the reduced segment from 413.33 to 433.33 MHz. This segment is only about one dB below maximum link budget, and places no harmonics below the 6th in restricted bands. But, the 420 to 450 MHz range is an amateur radio band, and there is also some Land Mobile operation from 421 to 430 MHz, and 418 MHz is a popular SAW based frequency that some users might wish to avoid when possible. Based on this potential interference, the segments from 413.33 to 417.9 and 418.1 to 420 MHz would be superior. These segments do, however, fall into the second harmonic of TV channel 12. The 307 to 320 MHz range would be preferred as a band that avoids direct interference and TV harmonics, at the cost of requiring the 4th harmonic of 307 to 310 to meet the general level specification (500 uV, -41 dBm ERP) and suffering about a 3 dB link budget degradation compared to the high end of the band. The sub-range from 310 to 320 MHz is clear of TV harmonics and its own harmonics dodge restricted bands through the 4th. TV harmonics will be covered in some detail later.

***FCC 15.35:*** This section covers the allowed increasing of peak transmit power levels when averaged with off times when using Amplitude Shift Keying (ASK). The use of such averaging is popular in the United States due to an apparent accidental misinterpretation of the rules that has since become accepted as standard practice. The original intent of the rules was probably to allow for averaging to maintain constant energy per bit. However, due to the FCC standard practice of using electric field strength instead of effective radiated power, the rule was in its early usage interpreted to allow for maintaining average field strength, up to a limit of 10X the maximum steady state field strength. Since power is increasing as the square of field strength, this allowed peak power to increase up to 100X, and average power to increase up to 10X. This would appear to be highly useful to the system designer, but note that the decrease in transmit time will result in an increase in signal bandwidth that is exactly proportional to the increase in transmit power, and there is thus no increase in signal to noise ratio if receiver bandwidth tracks. However, there are two exceptions to this generalization discussed in the regulatory system design summary.

The precise mathematics of ASK averaging may be derived in short order. Let us make the following definitions:

$D_c$ = total digital duty cycle (FCC uses the 100 mS segment in the protocol with highest duty cycle). Note that if the bit density is 50% "ones" and if no transmission is made during zeroes, then the duty cycle is down to 0.5 even before any bit shortening of ones is applied.

$D_{c1}$ = duty cycle of each individual "one" counting bit shortening. Note that if the length of a one were cut in half (such as standard Manchester) then $D_{c1}$=0.5 and with 50% ones $D_c$ = 0.25.

$D_{c0}$ = duty cycle of each individual "zero." For simple ASK this could be 0, for standard Manchester it is also 0.5, and for a non standard Manchester it could vary from 0 to 1 depending averaging desired (which also applies to $D_{c1}$).

$D_0$ = logical duty cycle of zeroes in bit stream, typically 0.5

$D_1$ = logical duty cycle of ones in bit stream, typically 0.5

Now for the total duty cycle (fraction of time carrier is transmitted) we may write:

Eq. 12: $D_c = D_0 D_{c0} + D_1 D_{c1}$

Let us define:

$E_{ss}$ = Field Strength steady state = allowed rms field strength at a particular frequency when not averaging.

$E_{pa}$ = Field Strength peak averaging = allowed peak field strength at a particular duty cycle and frequency when averaging. Note that according to FCC convention this "peak" is not the true RF peak. It is the rms carrier strength in volts per meter at the peak of the *envelope*.

We may now write:

Eq. 13: $E_{pa} = \dfrac{E_{ss}}{D_c}$ as the max allowed field peak field strength under the rules, up to a limit of 10X the steady state allowed ($E_{ss}$).

We may substitute Eq. 12 into Eq. 13 and solve for necessary duty cycle within bits given allowed (regulatory) and attainable (hardware limited) field strength levels as:

Eq. 14: $$D_0 D_{c0} + D_1 D_{c1} = \frac{E_{ss}}{E_{pa}} = \frac{AllowedEFieldStrength}{AttainableEFieldStrength}$$

We express this in power terms as:

Eq. 15: $$D_0 D_{c0} + D_1 D_{c1} = \sqrt{\frac{P_{ss}}{P_{pa}}} = \sqrt{\frac{AllowedPowerSteadyState}{AttainablePowerPeak}}$$

The averaging effect may be used to reduce harmonic attenuation requirements above 1000 MHz, where in the restricted bands a level of 500 uV/meter is generally required. The 500 uV/meter level corresponds to 7.5E-8 watts ERP, or –41.2 dBm ERP.

The effect of averaging strongly influences the choice between ASK and FSK in FCC follower countries, which may be quantified as follows. For the case of Non Return to Zero (NRZ) ASK with no pulse shortening, we may write for the average ASK power $P_{ASK}$:

Eq. 16: $$P_{ASK} = 0.5 D_{c1} P_{pa}$$

In this equation $D_{c1}$ = 1.0 if no pulse narrowing is used. If receiver bandwidth is the reciprocal of symbol time $T_s$ without pulse narrowing and $T_{ns}$ with pulse narrowing, and taking into account the 50% duty cycle of simple ASK, then it may be shown that the ratio of signal to noise ratios for ASK transmit power $P_{pa}$ between 0 and 6 dB over the steady state limit $P_{ss}$ and FSK transmit power at $P_{ss}$ is (neglecting the slight difference between narrowband FM and AM):

Eq. 17: $$\frac{SNR_{ASK}}{SNR_{FSK}} = 0.5 D_{c1}{}^2 \frac{P_{pa}}{P_{ss}}$$

This advantage scales from 0 to 3 dB as Ppa advances from 0 dB to 6 dB over $P_{ss}$. Substituting Eq. 15 into Eq. 17 will show that for $P_{pa}$ greater than 6 dB over $P_{ss}$ the advantage for ASK tops out at 3 dB so long as receiver bandwidth is at the reciprocal of symbol time. However, low cost receivers for this class of equipment are often unable to match their noise bandwidth to be only the reciprocal of symbol time, so in practice averaging is often a significant improvement.

## *European Rules*

The reader may have noted that the U.S. rules for control and security operation are not particularly easy to follow, being basically spread over four sections, lacking in some definitions, and not written in a tutorial fashion. However, they're pleasant reading compared to the European rules, which are spread out over many documents, seem to often lack references to essential supporting data, and in general seem to feature considerable disregard for reader convenience. We review these rules to the best of our ability below, and give our interpretations with the warning that we may have made mistakes. Fortunately, there is a fairly high degree of standardization on bands and powers, throughout Europe, with most disagreements coming in the allowed modes and transmit duty cycles. This is achieved under the regional authority of the European Conference of Postal and Telecommunications Administrations (CEPT), which has 43 member nations. The European Telecommunications Standard Institute (ETSI) develops technical standards for CEPT countries.

A significant philosophical difference in the European rules are provisions that go beyond preventing interference to other systems into attempting to guarantee acceptable system performance. Most electronic equipment sold in the European Union must comply with **EMC Directive 89/336/EEC**, and be labeled with the **CE mark**, in conformance with this policy. After April 8, 2000, compliance with these requirements may be self certified by certain procedures (see www.ero.dk). Another important document governing required performance its **ETS 300 683**, "Electromagnetic Compatibility (EMC) Standard for Short Range Devices Operating Between 9 kHz and 25 GHz". This document's performance requirements are centered on interference immunity from both outside electromagnetic fields and disturbances on power supply and control inputs. The document refers to other documents for many measurement details. Designers of finished radio equipment to be marketed in Europe must review this set of documents in some detail.

For control and security applications the most fundamental document is **CEPT ERC Recommendation 70-03E**, downloadable from www.ero.dk. This gives the best general description of allowed applications, frequencies, powers, and other specifications. Further details on test methodology to confirm compliance are given in **ETSI EN 300 220-1**, downloadable from www.etsi.org. EMC compliance is described in **ETSI ETS 300 683**. For ISM (Industrial, Scientific, and Medical) band type operation, which generally includes higher end apps like Bluetooth and wireless LANS, **ETSI 300 328** is in general the applicable document. Europe does not have the 902-928 MHz ISM band, but does use the same 2400-2483.5 MHz ISM band authorized by the FCC, though at a reduced spread spectrum power level of 100 mW (U.S. is up to one watt). Note that up to 10 mW ERP narrowband is authorized in the European 2400 MHz band, though this is not mentioned in ETSI 300 328 with the other 2400 MHz rules. Instead, this is authorized in CEPT recommendation 70-03 under Annex 1. See Table 2 and the footnotes in this annex.

For classic control and security applications similar to FCC 15.231, the European rules allow for use of **433.05-434.79 MHz** under the “Non-specific Short Range Device” rules of Annex 1 under ERC 70-03E. This band segment just misses the 2nd harmonic of TV channel 13, a likely reason for its selection. The approval process is like U.S. certification where individual licenses are not required. The basic rules here are up to **10 mW ERP** (effective radiated power), at less than **10% duty cycle**, with some countries having differing duty cycle limits. This generous transmit power has the capability to provide for an excellent short-range link. Though applications are not specifically limited, the 10% duty cycle limit inherently restricts applications to control, intermittent status reporting, and low-end data acquisition. There is no specific frequency accuracy specification given, but since the band is not channelized the general wide band +/-100 ppm requirement of ETSI 300 220-1 Table 7 (p. 24) should apply. This would appear to be a difficult requirement for low cost SAW based devices to meet, but the apparent practical interpretation of this rule is that it covers temperature and power supply drift, and not set on accuracy, allowing SAW based devices to meet it. The harmonics are not specifically called out, but the general spurious limits of 250 nW below 1000 MHz and 1 uW above 1000 MHz given in ETSI 300 220-1 Table 13 (p. 34) should apply. Note that at full power of 10 mW this is -46 dBc for the 2$^{nd}$ harmonic and -40 dBc for higher harmonics. Since this limit is not set as strictly dBc, for a typical ERP of –10 to –20 dBm attained with a printed loop antenna, the limit to the 2$^{nd}$ is more like –26 to –16 dBc. Note that there are certain band segments where the operating mode parasitics are limited to 4 nW, including 470 to 862 MHz. At –10 dBm ERP this is –44 dBc required spurious suppression in these band segments.

## Broadcast Interference

Though television station frequency allocation skips over the U.S. 260-470 MHz band, they are so powerful that their second harmonics are still a potential interference problem. TV stations 7-13 are allowed power levels up to 316 kW, with harmonics for analog stations limited to -60 dBc or more down (future digital TV is expected to feature much lower harmonics, reported as –110 dBc). Table 4 shows channels and frequency allocations. Note that the segment 285 to 348 MHz avoids TV harmonics, and the segment 324 to 348 MHz avoids TV and FM broadcast harmonics. However, 322 to 335.4 is directly restricted by FCC section 15.205, and there are restricted band harmonic issues that make other segments less than optimum also.

**Table 4:** Television and FM frequency assignments

| Channel | Carrier (MHz) | 2nd Harmonic (MHz) |
|---|---|---|
| 6 | 82-88 | 164-176 |
| 7 | 174-180 | 348-360 |
| 8 | 180-186 | 360-372 |
| 9 | 186-192 | 372-384 |
| 10 | 192-198 | 384-396 |
| 11 | 198-204 | 396-408 |
| 12 | 204-210 | 408-420 |
| 13 | 210-216 | 420-432 |
| 14 | 470-476 | 940-952 |
| FM Stereo | 88-108 | 3rd har = 264-324 |

To gauge the potential problem we refer to a standard handbook such as Ref. 6 for graphs of electric field strength from broadcast stations. A rough approximation is 10 to 90 dBuV at the carrier frequency for distances from the broadcast transmitter from 200 kM down to 5 kM per kW of transmit power. Converting to watts for a 100 kW station with a transmitter tower 100 meters tall that are picked up at the -60 dBc second harmonic by a quarter wave whip at 400 MHz shows a receive power of 1.7E-21 to 1.6E-13 watts spread over the 12 MHz (at the 2nd harmonic) TV bandwidth. Converting to dBm per Hz the spectral density of this interfering source is approximately -225 dBm/Hz at 200 kM up to -146 dBm/Hz at 5 kM. The farther ranges are negligible as interference, but the shorter ranges definitely exceed the -174 dBm/Hz thermal noise floor that receiver sensitivities and link budgets are typically calculated against. With a typical inverse forth power propagation constant in this environment, interference could be expected at up to approximately 20-40 kM range. This is conservative since the receiver antenna height is 9 meters in the graph used. Actual power density may be 10-20 dB lower at the typical 1-2 meters antenna height of MicroRadio receivers, reducing these interference ranges by a factor of 2 to 4, but it does illustrate the potential seriousness of the issue. Some systems are likely suffering interference from this source that significantly reduces range and reliability, without designers being aware of the cause. Narrowband receiver systems can reduce this problem by using frequencies placed at the band edges of the TV harmonics, such as 420 MHz, where energy density is approximately 20 dB lower. Fortunately, improved filtering in newer digital TV stations should reduce the incidence of this interference in the future.

## Regulatory System Design Impacts Summary

***Carrier Frequency:*** The European choices are restricted to 433.05-434.79 at 10% duty cycle (most countries) and a maximum power of +10 dBm among the lower UHF frequencies. If this is not adequate the user may step up to 868-870 MHz with some segments allowing up to +14 dBm at up to 100% duty cycle. The U.S. choice of carrier frequency is a more complicated subject. The segments from 413.33 to 417.9 and 418.1 to 420 MHz are excellent for avoiding placing harmonics through the 5th in restricted bands that require more transmitter filtering circuitry while simultaneously dodging direct interference. They do suffer TV 2nd harmonic interference whose seriousness depends on location, frequency placement, and receiver bandwidth. The segment from 432 to 470 MHz is excellent for allowing near maximum link budget and avoiding TV and FM 2nd and 3rd harmonics, but does require up to -28 dBc harmonic rejection of its own third harmonic, which may be difficult to attain with an unfiltered loop antenna. The segment from 310 to 320 MHz is clear of TV harmonics, and its own harmonics through the 4th dodge restricted bands. It does suffer from the third harmonic of FM broadcasting and its link budget is about 3-5 dB inferior to the higher frequencies. But for the lowest cost applications that cannot pay for transmitter filtering it is an excellent choice, and it is no accident that 315 MHz has been a popular LC and SAW based choice for many years. In general, the 285-470 MHz band available in the U.S. under FCC 15.231 is an under utilized resource with many MicroRadio business opportunities.

***Harmonics:*** By giving up segments of usable frequency to dodge "restricted bands" in the U.S., it is generally possible at typical transmit powers (0 dBm driving a 5% to 10% efficient loop antenna) to satisfy both U.S. and European rules with only about 20-30 dB of harmonic suppression. For example, for U.S. operation at steady state output power in the 413.33 to 433.33 MHz band only 20 dB suppression is needed through the 5th harmonic, and then 26 dB for the 6th. That same transmitter reset to 433.92 MHz for the European market needs at least 22 dB for all harmonics. If the European 0 dBm ERP transmitter uses an efficient quarterwave whip then it needs a minimum of 36 dB second harmonic suppression.

***System Signal to Noise Ratio vs. Size:*** Despite the improved energy per bit possible with averaging in the U.S., the actual link quality remains little affected if receiver bandwidth tracks transmitted spectral occupancy. An exception to this generalization is that ASK averaging without symbol shortening can yield an up to 6 dB improvement over non-averaged ASK and up to 3 dB SNR improvement over FSK. In general, if narrowband receivers can be provided in a given application, much better system design will result due to an equivalent link being maintained with much more spectral efficiency (more channels and less interference) and with smaller transmitters (smaller antennas and batteries). On the other hand, if cost constraints force use of a wideband receiver such as a regenerative, then averaging a shorter (wider bandwidth) pulse of higher power will yield a significant link improvement for that non-ideal case as compared to not averaging.

***ASK vs. FSK:*** In high mobility systems such as cellular angle modulation (FSK or PSK) is generally technically superior due to its improved multipath performance and greater interference immunity. Since Europe allows no ASK friendly averaging effect, European FSK also allows an inherently 3 dB improved link budget by the simple virtue of its 100% duty cycle. For low mobility systems (most MicroRadio applications) where receiver detectors are properly designed to provide a capture effect, FSK has little inherent link advantage. Therefore, in the United States and other FCC countries, the choice between ASK and FSK as far as link budget goes depends on where the transmit power that a particular transmitter, battery, and antenna may attain falls relative to the maximum allowed steady state values. If the max attainable value is less than the steady state limit, then FSK is superior since it will again show a 3 dB link budget improvement. But if the max attainable power falls between the steady state limit and 6 dB above the limit, then ASK shows an advantage because the data duty cycle may be averaged without shortening pulses and widening receiver noise bandwidth. This net link improvement will be 3 dB over FSK if the ones bits may be transmitted at 6 dB above the steady state limit. For transmit peak power levels more than 6 dB above the steady state limit pulse shortening must be used, which widens receiver noise bandwidth. This limits the advantage of ASK to a maximum of 3 dB as compared to FSK if the receiver bandwidth may be controlled to match transmitted spectral occupancy. Low cost receivers often do not allow minimum bandwidth, so in that case ASK with averaging is preferred.

***Phase Noise and Spurious:*** The European rules, specifically ETSI EN 300-220-1 section 6.6, can also be inferred to set a phase noise specification. Table 5 there shows that the spurious noise must meet the general European 250 nW spurious level (-36 dBm) using a 10 KHz spectrum analyzer bandwidth for carriers up to 25 MHz, using a 100 KHz analyzer bandwidth for carriers from 25 MHz to 1000 MHz, and using a 1 MHz analyzer bandwidth for carriers above 1000 MHz. It would not be logical for this to be interpreted as a phase noise requirement when the phase noise offset frequency is still in an allocated band such as 433.05 to 434.79 MHz (it's not a spur in-band), but it makes sense from the regulatory perspective to require integrated phase noise over these spectrum analyzer bandwidths outside the assigned bands to meet these spurious levels. Based on transmitted ERP the inferred phase noise requirement thus becomes:

Eq. 18: $\phi_N(close\ \ in) = -\left(P_{erp}(dBm) + 36 + 10\log(BW)\right)$

At +10 dBm in the 434 MHz band this implies an out of band phase noise of -96 dBc/Hz. If a carrier is set at a nominal frequency of 433.92 MHz, then at 100 ppm maximum error it can get to within 880 KHz of the band edge, which is where this -96 dBc/Hz becomes applicable. The far out transmitted phase noise (which may be helped by the frequency response of the antenna) must meet the 4 nW limit of the European "restricted" segments from 174 to 230 MHz and 470 to 862 MHz. This is given by:

Eq. 19: $\phi_N(far\ \ out) = -\left(P_{erp}(dBm) + 54 + 10\log(BW)\right)$

At +10 dBm and neglecting antenna bandwidth this is a transmitted phase noise of -114 dBc.

It may also be inferred that the restricted band from 470 to 862 MHz with its 4 nW spur spec sets phase noise and spur limits for the second harmonic of the transmitted level, which could come in as low as 867.76 MHz with 100 ppm crystals. This implies that at 2.88 MHz offset from carrier the transmitted phase noise must be at -68 dBc/Hz, and any discrete spurs must be 18 dB below the carrier if the 2nd harmonic is at the 250 nW upper limit. These limits may usually be neglected since the carrier frequency limits turn out to be the worst case.

This same rules section also sets an inferred synthesizer reference (phase detector sampling rate) spurious specification of:

Eq. 20: $SpurSpec = -\left(P_{erp}(dBm) + 36\right)$ for spurs in the 250 nW segments, and

Eq. 21: $SpurSpec = -\left(P_{erp}(dBm) + 54\right)$ for spurs in the 4 nW segments.

The reference spur of -36 dBc for 0 dBm transmitters is not difficult, but at +10 dBm and -46 dBc care is required. The phase noise specifications are not difficult for good quality discrete designs, but may become more troublesome for low Q integrated designs, particularly at +10 dBm ERP. The phase noise of integrated ring oscillators in CMOS is very much inferior to what high Q resonant oscillators deliver. This leads to a need for high loop bandwidths to suppress phase noise, which then tends to worsen synthesizer spurs. A compromise may meet both goals fairly easily at 0 dBm output power, but real care in detailed IC design may be needed to meet these simultaneously at +10 dBm. See Figure 2 for phase noise limits derived from these rules compared to typical performance of an integrated transmitter at two different PLL loop bandwidths. Note that the inferred mask, which is fine in the U.S. also, is from a regulatory perspective. Acceptable phase noise performance for narrowband FSK in terms of phase noise limited signal to noise ratio is a separate technical issue that we shall address in Part 2.

## Conclusion

There is a growing industry awareness of the business opportunity of MicroRadio in general, exemplified at the high end of this product range by the Bluetooth phenomenon and similar cooperative efforts such as HomeRF and the IEEE 802.15 Personal Area Network Work Group. The future winners in this business area are likely to be those that understand not only low cost and low power discrete circuit design, but also the full range of regulatory impact, wireless system design, and integrated circuit design factors that affect the total product definition. As system complexity increases these trade-offs will become even more crucial, requiring a high degree of professionalism in definition and design to be competitive. With this and similar articles we hope to improve the basic information being provided to developers of these products by their component suppliers. Next month's article will continue with a more detailed design tutorial and examples.

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***Acknowledgements:*** The author wishes to express appreciation for the valuable reviews and inputs from the other members of the Microchip RF Products Group, and from Didar Islam, David Saar of Saar Associates, Alan Bensky of Ateret Radio Engineering, Paul True of Linx Technologies, Frank Perkins of RF Monolithics, and Andreas Laute of Melexis.

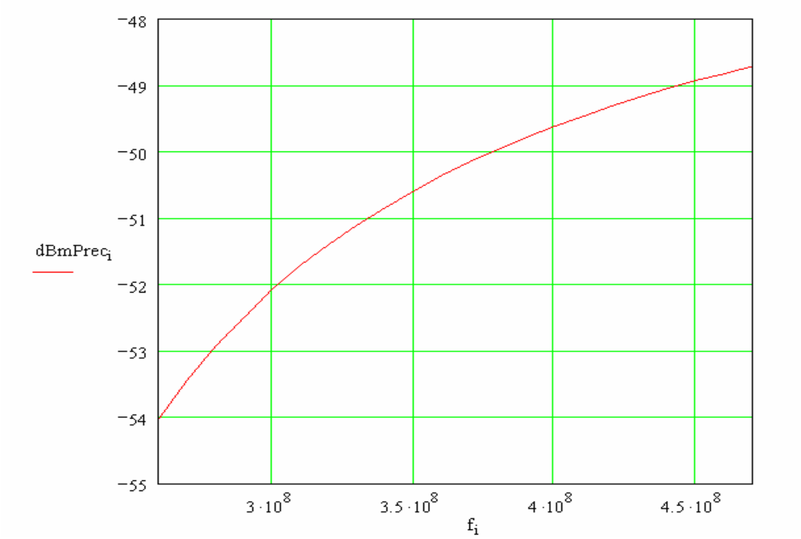


**Figure 1:** Link budget change over frequency when following FCC 15.231 rules. The vertical axis is typical receive power in dBm at 3 meters range when transmit power is at the legal maximum, and the horizontal axis is frequency. An approximately 5 dB improvement in link budget over frequency is present.

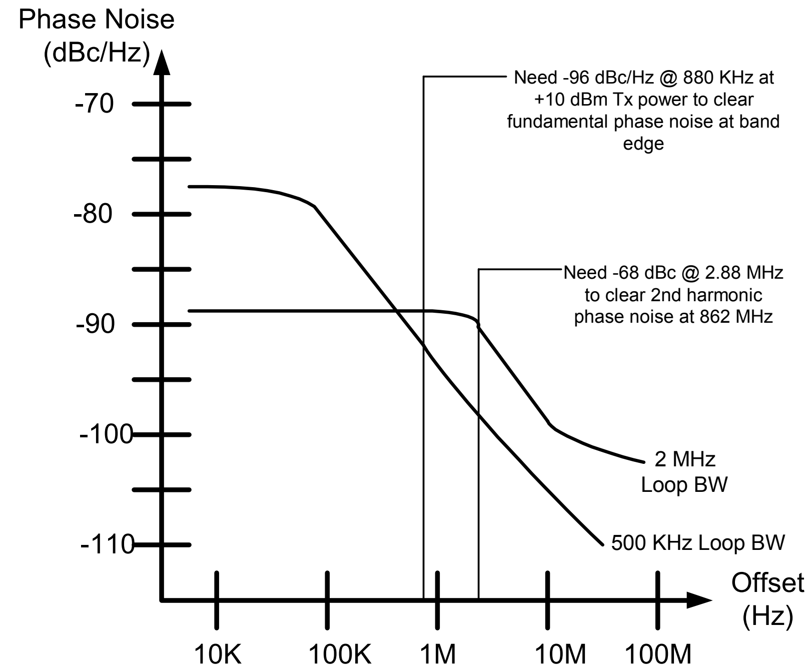


**Figure 2:** Typical integrated transmitter phase noise at two example PLL bandwidths as compared to inferred phase noise requirements from the band edge relative to 250 nW relative to fundamental and for the 2nd harmonic relative to 4 nW. The 2nd harmonic easily clears the -68 dBc/Hz required. However, at full transmit power of +10 dBm erp the phase noise fails by about 3 dB for 500 KHz loop bandwidth and 8 dB for 2 MHz loop bandwidth. The implication is that this 0 dBm transmitter can drive an efficient antenna and pass, but that a 10 dB gain buffer added to take full advantage of the European limit of +10 dBm would cause a regulatory failure.