

About This Application Note

This application note is intended to provide detailed application information for the MICRF001 QwikRadio™ Receiver IC. All aspects of device application, including operating modes, frequency selection, data rate selection, and external component selection, will be discussed in-depth. Important aspects of system design, and design trade-offs, are also discussed.

Micrel appreciates that prospective customers reflect a broad range of expertise and experience. Some readers will already be well-versed in developing and using remote-control wireless data links, while others may need further guidance. This is further complicated by the fact that governmental regulation of systems employing devices like the MICRF001 vary from country to country.

This application note contains additional information for prospective users who need further guidance. A glossary is provided at the end of this application note. Also provided is a subject-matter bibliography, which identifies related application notes, and other material which may be useful, depending on background and interest.

MICRF001 UHF Receiver Overview

The MICRF001 was designed as a cost-competitive solution to existing superregenerative receivers and superheterodyne receivers, for remote-control type wireless data links. Further, due to topological improvements and higher levels of integration, this IC is more easily applied than any before it, and the user does not need to be an "RF expert."

The MICRF001 is a complete superheterodyne OOK receiver IC, intended for use in the UHF frequency band from 300MHz to 440MHz. The IC incorporates complete UHF down conversion and data demodulation functions on the same IC and provides data output (logic levels) compatible with a wide variety of data decoder ICs. Micrel's proprietary enhancements to the superheterodyne topology lower manufacturing costs and speed time-to-market.

The MICRF001 is data format independent and may be easily incorporated into both existing and new applications. Further, the MICRF001 is completely compatible with either SAW-based or LC transmitters, without any transmitter modifications whatsoever. Range performance of the device is typically 100 meters, depending on data format, data rate, and antenna type.

Patent-pending design techniques provide features unmatched by any previous solution, namely:

- Complete elimination of expensive and bulky SAW resonators and coils
- Automatic receive frequency tuning and alignment anywhere in 300MHz to 440MHz band

- Complete compatibility with both LC and SAW transmitters
- Programmable, fully integrated (demodulator) baseband filtering
- Virtual elimination of LO reradiation
- Low-cost, high-reliability CMOS technology
- Lowest parts count in the industry

Applying the MICRF001

The MICRF001, because of its high level of functional integration, requires only one ceramic resonator and two capacitors to form a complete UHF OOK receiver for wireless applications. Neither of the capacitors are high-precision components, and the ceramic resonator need only exhibit a modest $\pm 0.5\%$ initial accuracy. Range performance with a quarter-wave antenna is typically 100 meters open-field. To complete the design process, the user need only perform three simple calculations:

- Compute the frequency of the ceramic resonator
- Determine the slicing-level time constant and compute C_{TH}
- Determine the AGC time constant and compute C_{AGC}

Superregenerative vs. Superheterodyne

Superregenerative (SR) Receivers

In most radio applications, the receiver is selectivity-limited. In other words, the minimum signal recoverable by the receiver is limited not by the thermal noise of the receiver (which sets the receiver "sensitivity"), but other "noise" located in the "ether," for example from other transmitters, like FM radio stations. Until the advent of the MICRF001, the simplest and most cost-effective UHF radio receiver for wireless remote-control applications was the SR receiver^{1,2}. SR receivers have been around for many decades, have an RF bandwidth of several MHz, and demonstrate a typical range of about 100 meters. SR receivers are basically homodyne (direct-to-baseband) receivers, elegant in their simplicity, but analytically difficult to understand due to their inherent nonlinearity. Development and optimization of these receivers is generally an empirical process.

SR receivers exhibit a number of drawbacks, limiting their application and ease-of-use, namely:

- Reradiate RF noise (called "regen noise"), for which regulatory limits usually exist^{3,4}
- Require manual frequency tuning
- "Regen noise" crosstalk limits how closely SR receivers can be collocated

- Require managing an extensive bill-of-materials of discrete components
- Cannot easily take advantage of IC integration

To be fair, SR receivers have several important advantages, namely:

- No local oscillator (LO), with its associated complexity
- Wide RF bandwidth allows operation with LC transmitters
- Easily tuned to any frequency of choice
- Readily designed for low power (e.g., battery) operation

Superheterodyne (SH) Receivers

In recent years, SH receivers⁵ have made in-roads on SR receivers because (1) the SH is more analytically understandable, (2) selectivity improvements can improve the range performance, (3) SH circuits can be integrated into an IC, and (4) SH receivers can be easily electronically tuned. Still, the lowest cost solution has traditionally been the SR, *especially where the user is willing to manually tune the receive frequency.*

While SH selectivity, and hence range, can be improved, this improvement comes at a sizable cost. Specifically, the LO must be very accurate, and for this reason is generally derived from either a crystal or SAW resonator. Such devices are far more expensive than LC tank circuits. Further, the high receiver LO accuracy also requires that the transmit frequency be highly accurate. So here again, the transmitter frequency must be crystal or SAW resonator-based. *LC-based transmitters simply will not do!*

The one manufacturing advantage to SH receivers is the fact that the receivers and transmitters do not require tuning of the transmit and receive frequencies, due to their requirement for accurate resonators. An additional advantage is that the demodulator can be integrated along with the down converter to help lower costs.

Disadvantages include (1) LO reradiation back through the antenna, which can be minimized through IC integration, and (2) operating frequency is difficult and costly to customize.

MICRF001 Differences from Standard SH Receivers

If one should list the best features of both the SR and SH receiver that should be incorporated into a new receiver, the list would be:

- No manual tuning of the receiver
- Compatible with both LC and SAW/crystal transmitters, without modification
- No expensive crystals, SAWs, or coils required
- Operating frequency is easily customized
- No SR “regen” noise to reradiate
- Integration to minimize LO reradiation to within regulatory limits
- Integrated demodulator to lower costs

- A very small bill-of-materials to manage
- Cost/range comparable to SR receiver
- Simple, parasitic insensitive layout

All of these features have been incorporated into the MICRF001, yielding a receiver that is simple to apply, easy and inexpensive to manufacture, and requires no manual tuning whatsoever.

The first four features have been accommodated through a patent-pending architectural change to the basic SH topology. This allows Micrel to employ the basic SH down conversion approach without the accuracy requirements generally imposed. All that is required is to attach a ceramic resonator to the MICRF001, in the range 2.0MHz to 3.5MHz, based on the desired system transmit frequency. Ceramic resonators are significantly less expensive than crystal or SAW resonators, and are easily and inexpensively customized to any frequency of choice.

The remaining features are accommodated by completely integrating all functions of the receiver on an IC, including the demodulator. Demodulator bandwidth is programmable in four discrete values, which gives designers sufficient flexibility without complicating the development process.

Finally, if the user chooses, the MICRF001 can be easily converted to a standard SH receiver, where the ceramic resonator would then be replaced with a crystal. In this application, the transmitter frequency must also exhibit the same crystal or SAW-resonator accuracy.

MICRF001 Theory of Operation

The block diagram in Figure 1 illustrates the basic structure of the MICRF001. Identified in the figure are the three principal functional blocks of the IC, namely (1) SH UHF down converter, (2) OOK demodulator, and (3) reference and control. Also shown in the figure are two capacitors (C_{TH} , C_{AGC}) and one timing component (CR), usually a ceramic resonator. With the exception of a supply decoupling capacitor, these are all the external components needed with the MICRF001 to construct a complete UHF receiver.

An example of sweep operation would be where the MICRF001 must operate with LC-based transmitters, whose transmit frequency may vary as much as $\pm 0.5\%$ over initial tolerance, aging, and temperature. In this (patent-pending) mode, the LO frequency is varied in a prescribed fashion which results in down conversion of all signals in a band 2% to 3% around the transmit frequency.

A range penalty will occur in installations where there exists a competing signal of sufficient strength in this small frequency band of several percent. (This penalty also exists with SR type receivers, as their RF bandwidth is also generally 2% to 3%. So any application for a SR receiver is also an application for the MICRF001.)

With the exception of a supply decoupling capacitor, these are all the external components needed with the MICRF001 to construct a complete UHF receiver.

External Control Signals and Mode Selection

Three control inputs are shown in Figure 1: SEL0, SEL1, and SWEN. Through these logic inputs the user can control the

operating mode and programmable functions of the IC. These inputs are CMOS compatible, and are pulled-up on the IC. The inputs SEL0, SEL1 control the demodulator filter bandwidth in four binary steps from approximately 0.6kHz to 4.8kHz, and the user must select the bandwidth appropriate to their needs.

The SWEN pin allows the device to be configured in either its normal (sweep) operating mode, or in standard (fixed) SH receiver mode. Sweep operation is selected when SWEN is high, and is the default mode for the IC.

For applications where the transmit frequency is accurately set for other reasons (e.g., applications where a SAW transmitter is used for its mechanical stability), the user may choose to configure the MICRF001 as a standard SH receiver (fixed mode), mitigating the aforementioned problem of a competing close-in signal. This can be accomplished by grounding the SWEN pin. Doing so forces the on-chip LO frequency to a fixed value. In such a case, the ceramic resonator would be replaced with a crystal. Generally, however, the MICRF001 can be operated with a ceramic resonator adequately, no matter whether the transmitter is LC or SAW.

Slicing Level and the C_{TH} Capacitor

Extraction of the dc value of the demodulated signal for purposes of logic-level data slicing is accomplished by external capacitor C_{TH} and the on-chip switched-cap “resistor” R_{SC} , indicated in Figure 1. The effective resistance of R_{SC} varies in the same way as the demodulator filter bandwidth, in four binary steps, from approximately 1600k Ω to 200k Ω . Once the filter bandwidth is selected, this “resistance” is determined; then the value of capacitor C_{TH} is easily calculated, once the slicing-level time constant is determined. Values vary somewhat with decoder type, but typical slicing-level time constants range 5ms to 50ms. Optimization of the C_{TH} value will be required to maximize range, and is dis-

cussed later in “Applying the MICRF001: Selecting the Slicing Level Time Constant and C_{TH} Capacitor.”

AGC Function and the C_{AGC} Capacitor

The signal path features automatic gain control (AGC) to increase input dynamic range. An external capacitor, C_{AGC} , must be applied to set the AGC attack and decay time-constants. With the addition of only a capacitor, the *ratio* of decay-to-attack time-constant is fixed at 10:1 (i.e., the attack time constant is 1/10th the decay time constant), and this ratio cannot be changed by the user. However, the attack time constant is selectable by the user through the value of capacitor C_{AGC} .

By adding resistance from the C_{AGC} pin to VDDBB or VSSBB in parallel with the C_{AGC} capacitor, the *ratio* of decay-to-attack time-constant may be varied. See “Applying the MICRF001: Selecting Demodulator Filter Bandwidth” and Figure 5c.

Reference Oscillator and External Timing Element

All timing and tuning operations on the MICRF001 are derived from the reference oscillator function. This function is a single-pin Colpitts-type oscillator. The user may handle this pin in one of three possible ways:

- Connect a ceramic resonator
- Connect a crystal
- Drive this pin with an external timing signal

The third approach is attractive for further lowering system cost if an accurate reference signal exists elsewhere in the system (e.g., a reference clock from a crystal or ceramic resonator-based microprocessor), and flexibility exists in the choice of system transmit frequency. The user should ac couple this signal into the REFOSC pin, and resistively divide (or otherwise limit) the signal to approximately 0.5Vpp. A sinusoid is preferred, and sharp transitions on this signal should be avoided to the extent possible.

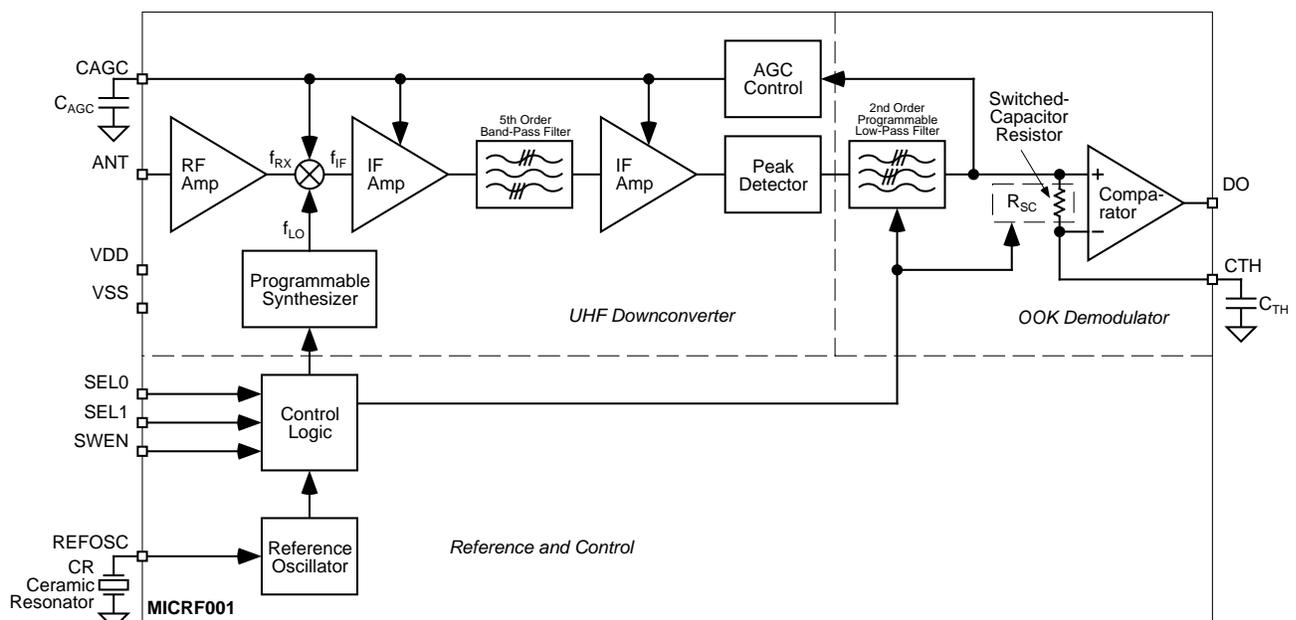


Figure 1. MICRF001 Simplified Block Diagram

Applying the MICRF001

User Selected Parameters

Once it has been determined that the MICRF001 meets such other system requirements as cost, power, range, etc., the user only needs to determine five parameters in order to implement a wireless data link using the MICRF001. These are as follows:

Operating mode, dependent on transmitter frequency accuracy and stability used in the system, selected by SWEN control pin,

Transmit frequency (f_{TX}) of the system, which determines the specific frequency of the timing element or signal (f_T),

Demodulator Filter Bandwidth, dependent on minimum data pulse width, selected by SEL0 and SEL1 control pins.

Slicing Level Time Constant, dependent on data burst period, preamble length, etc., selected by the value of capacitor CTH.

AGC Time Constant, dependent on system decode time, selected by the value of capacitor C_{AGC} .

Other aspects of the system design are assumed to have been determined, like data rate, whether transmissions include a startup preamble, and whether data is transmitted continuously or in packetized "bursts." (Selection of such issues should not be treated lightly, as they can impact range if not optimized.)

Selecting the Timing Reference

Frequency and Accuracy

As with any SH radio, the LO frequency must be separated from the received frequency by such an amount that the resultant IF (f_{IF}) falls into the bandwidth of the BPF. So it is important to understand the various relationships between the timing frequency (f_T) and frequency characteristics of the elements in the RF and IF path. f_{IF} and f_T are shown in Figure 1.

Lets begin with the relationship of the timing signal frequency f_T to the BPF frequency characteristics. Bandwidth of the BPF is $1\text{MHz} \pm 200\text{kHz}$, and is not a strong function of f_T . However, the center frequency of the BPF (f_{IF}) does vary with f_T as described by the following equation:

$$(1) \quad f_{IF} = f_T \left(\frac{M + \alpha}{390} \right) 2.00$$

where:

f_{IF} and f_T are in MHz, $M=128$, and $\alpha = 1$ for sweep- and $\alpha = 2$ for fixed-modes of operation.

The design equations for each of the modes, fixed and sweep, can now be written. First consider fixed mode.

Fixed Mode

For fixed mode operation, the following equations apply,

$$(2) \quad f_{IF} = f_{TX} - f_{LO}$$

$$(3) \quad f_T = \frac{f_{LO}}{M + 2}$$

assuming low-side mixing*. Substituting Equation 3 into Equation 1, and the result into Equation 2 yields:

$$(4) \quad f_{LO} = f_{TX} \left(1 + \frac{2.00}{390} \right)^{-1}$$

where all frequencies are in MHz.

So, once the transmit frequency f_{TX} is defined, and the mode is defined as fixed, the procedure to compute the timing frequency is simply to compute f_{LO} from Equation 4, and then solve for f_T using Equation 3.

Fixed-Mode Example

Given $f_{TX} = 315\text{MHz} \pm 200\text{kHz}$ (initial tolerance, age, temp. variation)

From Equation 4:

$$f_{LO} = 315 \times \left(1 + \frac{2.00}{390} \right)^{-1} = 313.393\text{MHz}$$

From Equation 3:

$$f_T = \frac{313.193}{130} = 2.4092\text{MHz}$$

(Use this crystal frequency)

Hence, if f_{TX} is exactly 315MHz, and the f_T is exactly 2.4092MHz, then the f_{LO} based on f_T will result in an f_{IF} that is exactly in the center of the BPF passband. (This can be shown by comparing the results of Equations 1 and 2).

Now lets consider the frequency variations of the transmitter and receiver. Assume that the crystal initial tolerance, aging, and temp variation were, collectively, $\pm 250\text{ppm}$, which is fairly standard for crystals. From Equation 3 note that the f_{LO} will vary by $\pm 250\text{ppm}$ as well, or $\pm 78\text{kHz}$.

Maximum frequency misalignment occurs when the transmit frequency is high (low) and correspondingly the receive frequency is low (high) due to inaccuracies in the timing elements of the system. This maximum error between transmitter and receiver ($200\text{kHz} + 78\text{kHz} = 278\text{kHz}$) must be less than one-half the minimum BPF bandwidth, 400kHz, to guarantee that the resultant IF falls in the BPF passband. Since this requirement is met, system performance will not be impaired even for worst-case system timing errors.

Transmit Frequency f_{TX}	Crystal Frequency f_T
315MHz	2.4107MHz
418MHz	3.1990MHz
433.92MHz	3.3208MHz

Table 1. Crystal Frequencies (f_T) for Several Common SAW Transmitter Frequencies Fixed Mode, Low-Side Mixing

Table 1 illustrates appropriate crystal frequencies for f_T , for fixed mode operation and some common SAW-based Transmitter frequencies. Low-side mixing is assumed.

* For high-side mixing, Equation 2 becomes $f_{IF} = f_{LO} - f_{TX}$. Rederiving f_{LO} would result in the (+) sign in Equation 4 being changed to a (-) sign. The f_T derived in this case, from Equation 3 would work equally as well as the f_T derived from Equations 3 and 4 for low-side mixing. This is another way of saying that there are always two equally valid LO frequencies for a superheterodyne receiver. The reader can use either method without penalty.

Sweep Mode

In sweep mode, the LO is “swept” across a band of frequency in the vicinity of the transmit frequency. The minimum and maximum values of f_{LO} during a sweep are given as

$$(5) \quad f_{LO(\min)} = f_T \times M$$

$$(6) \quad f_{LO(\max)} = f_T \times (M + 2)$$

$$(7) \quad f_{LO(\text{avg})} = f_T \times (M + 1)$$

The sweep range is simply the difference between $f_{LO(\max)}$ and $f_{LO(\min)}$:

$$(8) \quad \Delta f_{SW} = f_{LO(\max)} - f_{LO(\min)} = 2f_T$$

In contrast to fixed mode, $f_{LO(\text{avg})}$ should be set equal to f_{TX} , so that the LO frequency will sweep *symmetrically* around f_{TX} . So using Equation 7 compute f_T for sweep mode. Since $f_{TX} = f_{LO(\text{avg})}$, from Equation 7:

$$(9) \quad f_T = \frac{f_{TX}}{M + 1}$$

The best way to explain what is happening in this mode is through a frequency plan as shown in Figure 2. Key points on this diagram are the transmit frequency, f_{TX} , and the minimum and maximum LO frequencies, $f_{LO(\min)}$, and $f_{LO(\max)}$. As the LO is swept across $f_{LO(\min)}$ to $f_{LO(\max)}$, identified as the *sweep band*, all signals in this band are swept into the IF passband. Coverage is further extended on each side of this sweep band to f_{MAX} and f_{MIN} by an amount equal to $f_{IF} + f_{BP/2}$, due to the BPF bandwidth. This is identified as the *coverage band*. The two values f_{MAX} and f_{MIN} are given by

$$(10) \quad f_{MAX} = f_{LO(\max)} + f_{IF} + f_{BP/2}$$

$$(11) \quad f_{MIN} = f_{LO(\min)} - f_{IF} - f_{BP/2}$$

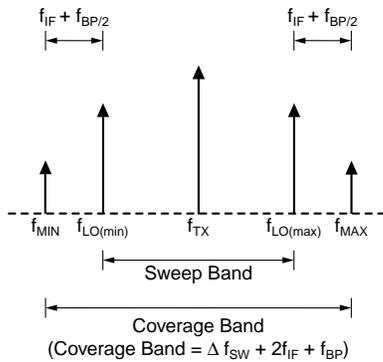


Figure 2. Sweep (SWP) Mode Frequency Plan

Once the transmit frequency f_{TX} is defined, the user computes f_T from Equation 9, and then connects a ceramic resonator equal to f_T . And that's it!

Equations 1, 6, 7, and 8 are only necessary for determining the accuracy requirements of the ceramic resonator and transmit frequency to assure system operation over worst-case frequency error. For purposes of performing worst-case analysis, Δf_{SW} is symmetrical around f_{TX} , and never less than 80% of the value computed from Equation 8.

Sweep-Mode Example:

Given $f_{TX} = 387\text{MHz} \pm 0.5\%$ (init. tol. = $\pm 0.15\%$, age = $\pm 0.15\%$, temp. variation = $\pm 0.2\%$).

From Equation (9):

$$f_T = \frac{387}{129} = 3.00\text{MHz}$$

(a standard value resonator)

At this point, the process of determining the ceramic resonator frequency is complete!

Now lets look at our frequency error budget. The composite difference between the transmit and receive frequencies must be less than the minimum value of $f_{MAX} - f_{TX}$, which may be written:

$$(12) \quad \text{Min. } (f_{MAX} - f_{TX}) = f_{TX} + \frac{0.8\Delta f_{SW}}{2} + f_{IF} + \frac{f_{BP/2}}{2} - f_{TX} = 0.8f_T + f_{IF} + \frac{f_{BP/2}}{2}$$

after some algebraic reduction. The term in brackets is the minimum value of Equation 10, given device tolerances.

$f_{LO(\max)}$ comes from Equation 6, Δf_{SW} from Equation 8, f_{IF} comes from Equation 1, and the minimum of $f_{BP/2}$ is 400kHz. Using Equation 12,

$$\text{Min. } (f_{MAX} - f_{TX}) = [2.4 + 2.23 + 0.4] = 5.03\text{MHz}$$

The (minimum) coverage band is simply twice this value, or 10.06MHz, for a total (minimum) RF bandwidth of 2.6% (i.e., $2 \times 1.3\%$).

This coverage band, nominally centered at f_{TX} , can shift (worst-case) by the total error of the ceramic resonator. Assume for the ceramic resonator an initial tolerance = $\pm 0.3\%$, age = $\pm 0.3\%$, temperature variation = $\pm 0.3\%$. Summing these terms yields a total shift of the band is $\pm 0.9\%$.

Now assume that the (MICRF001) coverage band center frequency moves its maximum amount, namely 0.9% in one direction. Since the sweep coverage is no less than $\pm 1.3\%$, the transmitter frequency can be allowed to vary by as much as $\pm 0.4\%$, or about $\pm 1.5\text{MHz}$ over initial tolerance, age, and temperature. Such accuracy in LC-type transmitters is readily achievable.

This example is based on full temperature range operation (-40°C to $+85^\circ\text{C}$) and 10 year aging of the ceramic resonator. In less stringent applications, and depending on the frequency inaccuracies of the transmitter, it may be possible to guarantee system performance at maximum transmit/receive frequency misalignment with a less expensive $\pm 0.5\%$ initial tolerance ceramic resonator. However, $\pm 0.3\%$ ceramic resonators are available at little additional cost.

Selecting Demodulator Filter Bandwidth

The user can generally improve system range performance modestly by proper selection of the demodulator (baseband) filter bandwidth. The user can select from four options, controlled via the SEL0 and SEL1 (logic) inputs. Filter band-

width does vary somewhat with the timing frequency f_T , and may be calculated from the equation:

$$(13) f_{DEM} = \frac{0.003f_T}{N}$$

where N is an internal (binary) divisor that varies with filter selection, according to the following Table 2:

SEL0	SEL1	N
1	1	2
0	1	4
1	0	8
0	0	16

Table 2. Filter Divisor N vs. SEL0, SEL1.

Table 2 illustrates demodulator filter bandwidths as a function of SEL0/SEL1 input control pins and the transmit frequency, f_{TX} Equation 13, and sweep-mode operation Equation 9. This chart can be used equally well for both operating modes of the MICRF001, as the small errors between modes (sweep or fixed) can be neglected.

		Transmit Frequency f_{TX}			
SEL0	SEL1	315MHz	387MHz	418MHz	433.92MHz
1	1	3.6kHz	4.5kHz	4.8kHz	5.0kHz
0	1	1.8kHz	2.2kHz	2.4kHz	2.5kHz
1	0	0.9kHz	1.12kHz	1.2kHz	1.25kHz
0	0	0.45kHz	0.56kHz	0.60kHz	0.63kHz

Table 3. Demodulator Bandwidth vs. Transmit Frequency and SEL0/SEL1.

As N increases, the demodulator filter bandwidth decreases. The appropriate filter bandwidth should be selected not based on data rate, but on the minimum pulsewidth in the data pattern, since even for low data rates the pulse duty cycle within a bit time may be small, requiring more bandwidth to preserve pulse fidelity.

A good example of this is a data rate of 100b/s, where a bit time is 10ms. Generally, a filter on the order of 100Hz would be adequate for noise filtering. If, however, the data is pulse-width-modulated (PWMed) such that a logic-1 is 90% pulse width, and a logic-0 is 10% pulse width, then a filter bandwidth of 10X, or about 1kHz would be required to adequately preserve the 10% pulse width. This wider filter bandwidth allows more noise to pass to the comparator, lowering the signal-to-noise ratio and range performance. *The reader should conclude from this example that near-50% PWM coding schemes are preferred for these types of wireless data links, and deviation from this will reduce system range, usually on the order of 30% or so, as a rule-of-thumb. Manchester coding is an example of a 50% scheme. 33/66% PWM is also commonly used.*

From data communication theory, it turns out that the optimum demodulator bandwidth is related to the minimum pulse width in the data stream by the relationship

$$(14) f_{DEM(opt)} = \frac{0.65}{\text{Minimum Pulse Width}}$$

which provides sufficient bandwidth to recover the full pulse amplitude. [Equation 14 is derivable from analyzing the Fourier (spectral) components of the data pulse.] The user should employ Equation 14 to compute the appropriate demodulator bandwidth for his particular data coding scheme. Then use Equation 13 if Table 3 is not adequate.

Example: Selecting a Demodulator Filter Bandwidth

Given: $f_{TX} = 387\text{MHz}$, data rate = 1000b/s, with 33/66% pulse-width modulation. This is illustrated in Figure 3 for a data-1 followed by a data-0. The minimum pulse width (PW_{min}) in this case is 333 μs . Using Equation 14, the optimum $f_{DEM} = 1.95\text{kHz}$. Thus, for $f_{TX} = 387\text{MHz}$, select the filter associated with $N = 4$, or 2.25kHz as the closest value.

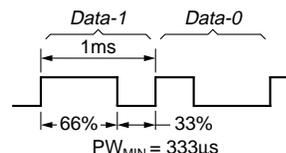


Figure 3. 1kb/s Data with 33/66% PWM

Selecting the Slicing-Level Time Constant and C_{TH} Capacitor

The output of the MICRF001 demodulator must be converted to logic levels to be compatible with available (digital) decoders. This operation is accomplished by the comparator, as shown in Figure 1. The comparator is simply a comparator, whose slicing threshold is a voltage developed on the C_{TH} pin. This voltage is developed by simply putting the demodulated data signal through an RC lowpass filter ($R_{SC}-C_{TH}$ in Figure 1), which extracts the average voltage of the data signal. It is then easy to compare the data signal against its average (dc) value via a comparator to construct a logic-level data signal.

The slicing level time constant selection is influenced heavily by the user's choice of coding scheme, and by the particular choice of decoder device to be connected to the MICRF001 output data pin. This section simply illustrates how to go about selecting the appropriate time constant and value of C_{TH} capacitor generically.

A typical data link waveform for MICRF001 applications, is illustrated in Figure 4, superimposed with the average value for three different $R_{SC}-C_{TH}$ (slicing level) time constants.

Constituent elements of this waveform include: (1) preamble, (2) delimiter, (3) data, and (4) dead time. Only the data portion of the example waveform is common to all systems. Some systems may not have preambles, some may not have delimiters, and some may not have a dead time period. The particular type of encoder/decoder pair selected for the system usually determines what this waveform will actually look like.

Simply stated, the slicing level must provide a dc voltage indicative of the average value of the data link waveform, suitable for accurate slicing of the data signal into logic levels.

Usually a broad range of time constant values exist that will work for a system, and system range performance is insen-

sitive to reasonable time constant variations, once the time constant is optimized. (A $\pm 20\%$ X7R ceramic capacitor is generally adequate.)

The slicing level time constant must obviously be in between the two extremes of zero time constant and infinite time constant. If the time constant is too short (e.g., zero), the voltage on C_{TH} would exactly follow the data signal, and no slicing would occur. For a time constant that is too long (e.g., infinite), the voltage on C_{TH} would never move from zero, and again no slicing will occur. The usual range of the $R_{SC} \cdot C_{TH}$ product is 5ms to 50ms. *If the value of C_{TH} is too small or too large, the decoding process can be completely obliterated—that is, no range at all. So if everything else appears set up correctly for the system to work, and the range is very poor or zero, check to be sure the value of C_{TH} is reasonable.* Also check that the correct value of R_{SC} is being used in the calculation, especially as R_{SC} varies with filter selection, as described subsequently.

Assuming that the preamble is provided for system initialization, then it remains to select a time constant such that a sufficient dc voltage is developed over the preamble portion, for slicing the data portion of the waveform (time t_0 in the example of Figure 4). Figure 4 illustrates a simulation of slicing level vs. time constant for 0.5ms, v(4); 5ms, v(2); and 50ms, v(3); time constants.

From Figure 4 one can easily see that a time constant of 0.5ms is too small, and even 5ms is not as good with the exception of the first data burst, where v(2) exceeds v(3) as a 50ms time constant. But as the time constants increase, decode response times also increase, and the system decode speed becomes sluggish. Secondly, if the time constant is smallish, as in v(2), the output pulses from the comparator function will exhibit some pulse width distortion, which may impact range performance.

The goal should be to use the largest time constant possible that still meets the maximum system decode time constraints.

As a rule-of-thumb, the time constant should be significantly longer than a bit time, say five times this value. For example, if the data rate is 1kb/s, and the slicing level time constant is 1ms, the extracted dc voltage will nearly

follow the data bit, so data slicing does not occur very effectively.

Certain other cases, however, may dictate how large the slicing level time constant can be, for example if the data waveform does not include a preamble, or the preamble as well must be preserved for the decoder to operate properly. Then the time constant must be small enough that a dc voltage is built up almost instantaneously from the first bit. Alternatively, if the system is allowed to miss decoding the first data “burst” then the time constant needn't be so small. This case effectively treats the first data “burst” as a preamble.

Now that guidelines have been established for choosing a slicing-level time constant, the next logical question is how the user sets the time constant. This time constant is expressed as a product of capacitance and resistance. Specifically, a capacitor is attached to the C_{TH} pin, denoted as C_{TH} . The resistance is provided on-chip as a “switched-capacitor” resistor (R_{SC} in Figure 1). The resistance value is a function of divisor N and timing frequency f_T (see “Selecting Demodulator Filter Bandwidth”), according to the equation:

$$(15) \quad R_{SC} = \frac{N}{f_T (3 \times 10^{-12})}$$

For example, if $f_T = 3.00\text{MHz}$, and $N = 2$, then $R_{SC} = 202\text{k}\Omega$. The time constant is then simply the product of R_{SC} and C_{TH} .

C_{TH} Calculation Example:

Assume that a 4ms time constant is desired, with $f_{TX} = 387\text{MHz}$ (i.e., $f_T = 3.00\text{MHz}$), and $N = 4$.

From Equation 15:

$$R_{SC} \times C_{TH} = 404\Omega \times C_{TH} = 4\text{ms}$$

$$\text{thus } C_{TH} = 9.9\text{nF}$$

To verify optimization of the C_{TH} capacitor value, a dual-trace oscilloscope is recommended. With one probe set on the C_{TH} pin and the other on DO pin (data output), one can easily see waveforms like shown in Figure 4. By “playing” with the value of C_{TH} and looking at the oscilloscope, the user can quickly optimize the C_{TH} value.

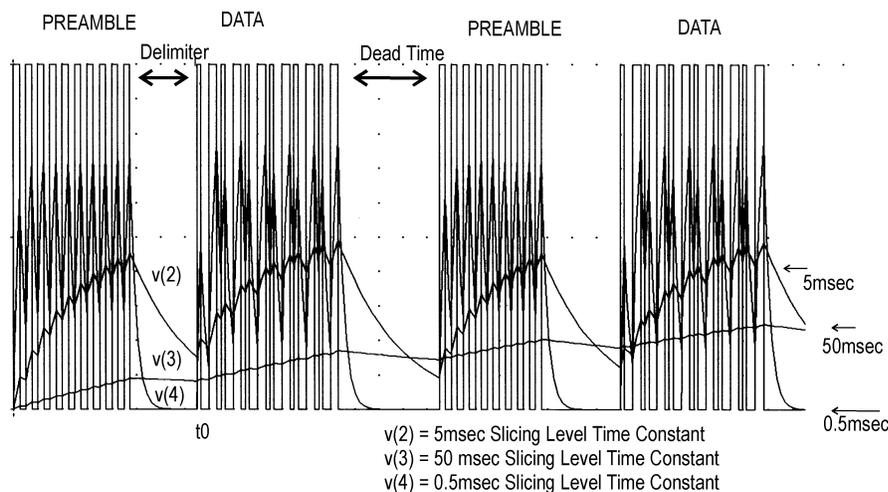


Figure 4.

Even after doing all of the above, the value of C_{TH} may not be optimum (i.e., maximum range), as variations will exist depending on particulars of decoder choice, existence of preamble, etc. So the final optimization of C_{TH} comes by performing range tests. The way this is accomplished at Micrel is to determine the range using a C_{TH} value as determined above. The value is then changed by $2\times$ and $1/2\times$, and the range test is repeated. This process is continued until the value of C_{TH} is determined which provides maximum range.

Selecting the C_{AGC} Capacitor

The signal path has automatic gain control (AGC) to increase input dynamic range. An external capacitor C_{AGC} must be applied to set the AGC attack and decay time-constants. The ratio of attack-to-decay time-constant is fixed at 10:1 if the user simply attaches a capacitor to the C_{AGC} pin. (Attaching resistance from the C_{AGC} pin to V_{DDBB} or V_{SSBB} in parallel to the C_{AGC} capacitor will allow one to adjust the attack-decay ratio. However an integrating capacitor must always be used.)

The signal on this pin is current-based, with an attack current of $15\mu\text{A}$, and a decay current of $1.5\mu\text{A}$. Generally, a 1.5V swing on this pin modulates the IF gain from zero to full, so using the equation for current into a capacitor ($I/C = dV/dt$), and desiring to set a dt -attack time-constant, compute the value of C_{AGC} using the equation

$$(16) \quad C_{AGC} = \Delta t (1 \times 10^{-5})$$

For example, for a 3ms attack time-constant, $dt = 3\text{ms}$, and so $C_{AGC} = 30\text{nF}$. The resulting decay time constant will be 30ms.

The AGC time-constants should be as long as possible to minimize “ripple” of the AGC control voltage, but short enough to meet any system decode response time requirement. As a goal, select C_{AGC} so that the AGC control voltage ripple does not exceed 10mVpp during transmission receptions (range may be impacted as the control voltage ripple increases).

C_{AGC} Accuracy Requirements

Whether the value of C_{AGC} must be critically determined by the user depends on the application.

For ac-powered applications, where the receiver will be powered up once, the value of C_{AGC} is not critical. In this application, the AGC circuit will integrate a voltage on C_{AGC} representing the local “ether” noise, centering the dynamic range of the system around this noise power level. Once this level is acquired, it need not change. So using a larger capacitor does not impose a penalty; in fact it will reduce the control voltage ripple, as discussed earlier.

Finally, systems which require that data decoding at the receiver begin rapidly may also benefit from optimizing C_{AGC} . Such an example is in systems which use the Motorola encoder/decoder chip set. This chip set provides no preamble, expects the decoding process to begin immediately, and requires two good successive code words, without false transitions in the dead time (Figure 4) between code words, to indicate a successful decode. If the AGC attack time is set sufficiently fast, then as the first code word comes through, the AGC voltage can acquire its new level, which will act as

a “squelch” on the “ether” noise, minimizing false transitions in the dead time, which will reduce system decode time. Experimenting with the value of C_{AGC} in such systems may improve range performance.

I/O Pin Interface Circuitry

Interface circuitry for the various I/O pins of the MICRF001 is shown in Figures 5a through 5f. Specific information regarding each of these circuits is discussed in the following subparagraphs. Not shown are ESD protection diodes which are applied to all I/O pins.

ANT Pin

ANT input circuitry is shown in Figure 5a. The ANT pin is internally ac-coupled via a 3pF capacitor, to an RF N-MOSFET transistor. Impedance on this pin to V_{SS} is quite high at low frequencies, and decreases as frequency increases.

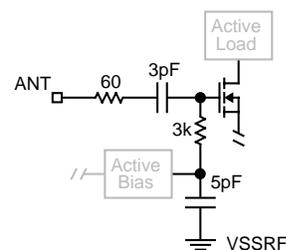


Figure 5a. Antenna Input

In the UHF frequency range, the device input can be modeled as $3\text{k}\Omega$ in parallel with 2pF (pin capacitance) shunt to V_{SSRF} .

C_{TH} Pin

Figure 5b illustrates the C_{TH} output pin circuitry. C_{TH} pin is driven from a P-MOSFET source-follower biased with approximately $20\mu\text{A}$ of current. Transmission gates TG1 and TG2 isolate the 3.3pF capacitor. Internal control signals PHI1/PHI2 are related in a manner such that the impedance across the transmission gates looks like a “resistance.” The dc potential on the C_{TH} pin is approximately 2.2V, fundamentally determined by the V_{GS} of the two P-MOSFET source-followers shown.

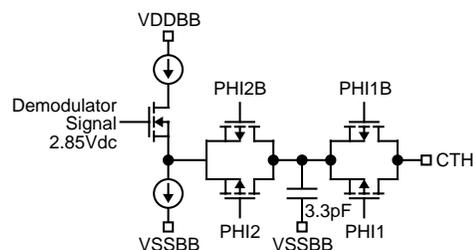


Figure 5b. Slicing-Threshold Input

C_{AGC} Pin

Figure 5c illustrates the AGC-pin interface circuitry. The AGC control voltage is developed as an integrated current into a capacitor C_{AGC} . The attacking current is nominally $15\mu\text{A}$, while the decaying current is a 1/10th scaling of this, approximately $1.5\mu\text{A}$. Signal gain of the RF/IF strip inside the IC diminishes as the voltage on C_{AGC} decreases. By simply adding a capacitor to C_{AGC} pin, the attack/decay time constant ratio is fixed at 1:10. Further discussion on setting the

attack time constant is found in AN-01, section 6.5. Modification of the attack/decay ratio is possible by adding resistance from C_{AGC} pin either to VDDBB or VSSBB, as desired.

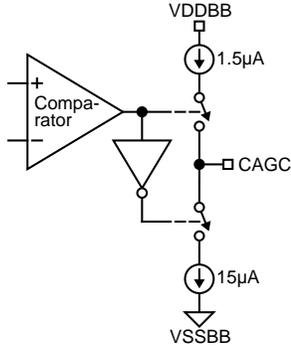


Figure 5c. AGC Capacitor Connection

DO Pin

The output stage for the data comparator (DO pin) is shown in Figure 5d. The output is a 10µA push-10µA pull, switched-current stage. Such an output stage is capable of driving CMOS-type loads.

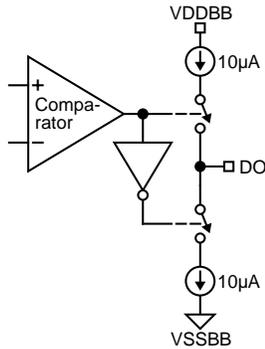


Figure 5d. Digital Data Output

REFOSC Pin

The reference oscillator input circuit is shown in Figure 5.5. Input impedance is quite high (200kΩ). This is a Colpitts oscillator, with internal 30pF capacitors. This input is intended to work with standard ceramic resonators, connected from this pin to VSSBB. The resonators should not contain integral capacitors, since these capacitors are contained inside the IC.

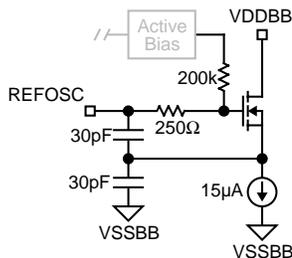


Figure 5e. Reference Oscillator

Externally applied signals should be ac-coupled. The nominal dc bias voltage on this pin is 1.4V.

SELO, SEL1 and SWEN Pins

Control input circuitry is shown in Figure 5f. The standard input is a logic inverter constructed with minimum geometry

MOSFETs (Q2, Q3). P-MOSFET Q1 is a large channel length device which functions essentially as a “weak” pullup to VDDBB. Typical pullup current is 5µA, leading to an impedance to the VDDBB supply of typically 1MΩ.

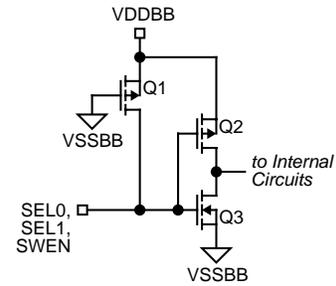


Figure 5f. Typical Control Input

MICRF001 External Component Selection

Encoder/Decoder Pairs

The MICRF001 was designed specifically for 33/66% PWM coding schemes that also include a preamble for “wake-up.” A good example is the Microchip Technology, Inc., KEELOQ product.

The MICRF001 can also be used with other standard decoders, like the Holtek or Motorola chip sets. The Motorola chip set employs a coding scheme with narrow pulses (<10% of data rate) that requires larger-than-necessary demodulator bandwidth for a given baud rate, which may modestly impact system range. Otherwise performance is quite acceptable.

Encoder Part No.	Decoder Part No.	Manufacturer
HCS3XX	PIC16C5X with 93C46	Microchip Technology (KEELOQ)
HT12E	HT12D	Holtek
MC145025	MIC145027	Motorola

Table 4. Encoder/Decoder ICs

The Holtek chip set provides 33/66% PWM like the Microchip Technology chip set. However it provides no preamble, and data transmissions are alternately “bursted” with 12 bits of data interspersed with an equivalent amount of “dead” time, with no activity. The first burst must be treated as a preamble, generating the slicing level for the MICRF001 demodulator. Since burst and dead times are the same length, a slicing level can only be extracted if the average value of the burst is > 50%. This means that of the 12 bits transmitted, more of the bits must be 66% than 33% width. Micrel recommends using this chip set for low performance applications which require only 4- to 6-bit addressing, which allows the user to set the remaining bits for 66% width.

Ceramic Resonators

Ceramic resonators can be purchased with and without integral capacitors. The MICRF001 is designed to use ceramic resonators without integral capacitors. Two recommended vendors are:

Part No.	Manufacturer
CSAx.xxxMG	Murata-Erie
ZTA-x.xxxMG	ECS International

where x.xxx signifies frequency (e.g., 3.000 for 3.000MHz). As of the time of this printing, ceramic resonators from the above companies are available through DIGIKEY (800-344-4539), an electronic parts distributor. A very thorough presentation of ceramic resonators is available from Murata-Erie, Catalog P-050B ("Piezoelectric Devices")⁶. Murata-Erie may be contacted at 800-831-9172. Other vendors of ceramic resonators include Panasonic and Toko, although Micrel has no experience with either company's resonator products.

Ceramic resonators are available in several standard values in the range from 2.00MHz to 3.50MHz, appropriate for the MICRF001. Typical six- σ limit on initial accuracy is $\pm 0.5\%$, aging is $\pm 0.3\%$, and temp variation is $\pm 0.3\%$. Devices can be purchased to tighter initial tolerances with a modest additional cost. Ceramic resonators can also be customized (like crystals) generally to 3-decimal accuracy, but at a higher cost than for standard values, and usually with some minimum purchase requirements.

If possible, try to use a standard ceramic resonator value from the vendor, and use Equation 9 to compute f_{TX} for the transmitter [i.e., select the standard ceramic resonator value, then let its value determine the transmitter frequency (f_{TX})]. The transmitter (presumably LC-type) can then be tuned to f_{TX} . Otherwise, if the user defines f_{TX} first, there will be an additional fractional error term to deal with associated with rounding the ceramic resonator value to 3-decimal places.

Finally, if the user chooses to use a ceramic resonator from a vendor other than Murata, it is recommended that operation of the particular resonator with the MICRF001 be certified. This can be done using the technique described in "Further Test Suggestions" to view the MICRF001 LO frequency. With the resonator under test attached, put the MICRF001 into fixed mode, and calculate the expected LO frequency (f_{LO}) using Equation 3. Then view the LO by monitoring the power supply as suggested in "Further Test Suggestions" with a spectrum analyzer. Any shift in the reference frequency f_T will show up as a shift in f_{LO} from its expected value.

PCB Layout Considerations

Layout is extremely easy and fairly noncritical for the IC and associated capacitors and ceramic resonator. The only pin requiring any special RF-handling at all is the ANT pin. The PCB can be single-sided to reduce cost, and no RF-shielding is required. The supply pins VDDRF and VDDBB may be tied together at the IC, and a good quality decoupling capacitor placed from this point to the VSSRF pin. Micrel's experience is that larger capacitors provide better performance, in the range 1 μ F to 10 μ F. The user should experiment with different size capacitors in the most cost sensitive applications to achieve the necessary range at the smallest capacitor value. (Depending on supply noise, a good high-frequency (e.g., 0.001 μ F) ceramic capacitor placed in parallel may improve performance).

Stray capacitance on the REFOSC pin should be kept to a minimum. Such capacitance will tend to lower the timing frequency f_T slightly. For example, 10pF of stray capacitance will lower the timing frequency by about 0.5%. Parasitic capacitance is minimized by keeping the PCB trace to the ceramic resonator short, and not placing metal on the opposite-side of the PCB from the REFOSC pin and trace. It is preferred that the two grounds, VDDRF and VDDBB be kept separate back to the common power supply connection (i.e., keep VSSBB currents from flowing through the VSSRF ground return). Location of the capacitors C_{TH} and C_{AGC} , as well as the ceramic resonator, are not very critical except that they should be taken to VSSBB rather than VSSRF.

Antenna Design Considerations⁷

Three types of antennas are normally associated with MICRF001 applications, (1) quarter-wave monopole, (2) helical (coil), and (3) pcb loop. Of these antenna types, the quarter-wave monopole is preferred, giving the best range performance. Typical SR and SH receivers employing quarter-wave monopole antennas generally provide about 100 meters open field range. The MICRF001 is designed to be connected to a quarter-wave monopole antenna at its input ANT pin, with no further coupling network, and provide this similar range. Of course the monopole can be a free-standing "wire," or a metal trace built directly into the pcb. The antenna need not be a quarter-wavelength long, but shorter lengths will reduce range.

Although the other two antenna types, helical and pcb loop, are less efficient antennas than the monopole, they provide cost and/or ergonomic advantages. The pcb loop antenna is a very inexpensive structure to produce, being built into the pcb. The only requirement is that there be enough "room" on the pcb to build a sufficiently effective antenna. This antenna has the further advantage that it can be hidden from the end-user. The primary disadvantage is that this is usually the least efficient of the three antenna types being discussed, with a typical range of 30 meters. Loop antennas are more commonly found in transmitters than receivers.

Where the antenna must be hidden, but still be rather efficient, a good choice is the helical antenna. The structure is simply a "coil" of wire, inexpensive, and easy to fabricate. Typical range for this type of antenna is 60 meters. (Both helical and pcb loop antennas are commonly found in transmitter designs, which usually requires a small unit with the antenna hidden from the user.)

Another aspect of antenna selection, aside from its effectiveness at "picking-up" signal from the "ether," is its directionality. Every type of antenna has its own signature of energy distribution in the 3-dimensional space that surrounds it. This signature is generally composed of "peaks" and "nulls"; if the transmit antenna "peak" is pointed at the receiving antenna's "peak," then range is maximized. However optimizing this for a particular application or installation is generally not practical. All one can do in these cases is choose the antenna that meets the other constraints, like cost and form-factor, then test to determine if the system provides adequate range in all the required directions and orientations. *Unfortunately much of this work is empirical.*

The MICRF001 is designed to work with any of the three antenna types described above without any special coupling circuits required. However Micrel does recommend some sort of dc path be provided from the ANT pin to VSSRF. This could be provided by an LC filter attached from the ANT pin to VSSRF, or simply a resistor. (See “Additional ANT (Antenna RF Filtering”). The ANT input is internally ac-coupled, so no external biasing is required. The input circuit impedance is very high (i.e., a FET gate), shunted by approximately 2pF of parasitic capacitance, giving an input ac-impedance of about 200Ω at 400MHz. Specialized 50Ω antennas may also be connected directly to the MICRF001.

The antenna should be connected as closely as possible to the IC pin, minimizing parasitic inductance between the antenna pad and the pin. (Use a >25mil trace, <200mil long, if possible). The antenna can be located away from the pin of the device, but this requires using a transmission line from the antenna to the IC input pin and an impedance matching coupling circuit. Micrel does not recommend such an approach for users who lack the necessary RF expertise and test equipment to design transmission line matching networks. For further information regarding different antenna structures, and locating the antenna away from the IC, see “Application Note 23: MICRF001 Antenna Design Tutorial.”

Additional Antenna Input RF Filtering

An optional discrete inductor-capacitor (LC) filter can be added to the ANT pin if necessary to provide further selectivity, and protection from equipments that generate strong EM fields (e.g., electric motors) located in the vicinity of the receiver. The L can be built into the pcb foil pattern to reduce cost (depending on trace width and thickness, pcb trace inductance is approximately 15nH/inch of track. If preferred, a small SMT inductor may be used instead). A capacitor, C, is placed in parallel with L to form a resonant tank at the antenna pad. The resonant frequency is given as

$$(17) f_{RES} = \frac{1}{2\pi\sqrt{LC_T}}$$

where $C_T = C + C_P$, and C_P is the parasitic capacitance attached to the ANT pin. The parasitic capacitance of the IC itself is approximately 2pF; more may be attributed to the pcb traces and any metal in the vicinity. Frequency f_{RES} should be

set equal to the transmit frequency f_{TX} . The input impedance of the IC results in a Q of the LC tank in the vicinity of 8 for a 3dB bandwidth of >40MHz for the filter. Thus the filter does not have to be tuned if a $\pm 10\%$ COG or NPO type cap is used. Use small value capacitors for C, which exhibit self-resonant frequencies above 500MHz. Parallel capacitors to achieve the required amount of capacitance if necessary. Alternatively use surface-mount capacitors, which exhibit very good high frequency performance¹³.

Do not expect this LC filter to improve range much, if any. This filter provides protection against input overload from machinery (motors, etc.) located near the receiver assembly. Also, although the input filter is not strictly required, Micrel recommends that at least some dc path to VSSRF be provided to protect the input circuit from damage due to large local EM fields, like lightning. Examples are the LC filter, or simply a resistor to VSSRF.

Certification and optimization of the LC filter during the development phase may require a network analyzer or vector voltmeter. Such certification should be done on the production pcb, to account for construction parasitics. With the LC filter in place, use the network analyzer or vector voltmeter to measure the bandpass response of this filter connected to the ANT pin. The C can be adjusted incrementally until the bandpass characteristic is centered at the transmit frequency f_{TX} .

A less expensive alternative to network analyzers is first to determine the expected value of C from Equation 17, then use a variable capacitor for C which straddles this value, tuning C until range is maximized. Determine the value of the tuned C, and mimic this value with fixed components.

Alternatively, simply step-in different values of capacitors until range is optimized. This hand-tweaking approach is less accurate, but gets around the need for expensive RF test equipment like network analyzers.

If this technique is performed correctly, the resulting system range should be about the same, whether or not the LC filter is attached to the ANT pin. This should be verified as a cross-check.

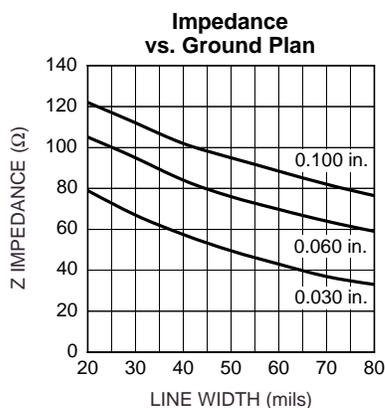


Figure 6.

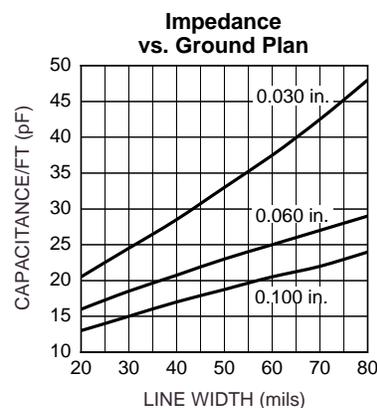


Figure 7.

Constructing a PCB Inductor

Using Figures 6 and 7, one can determine the inductance of a metal track on a PCB¹⁰, assuming the metal to be 1-oz (1.5-mil) thick copper. These figures normally are used to determine the characteristic impedance of a (straight) metal track over a ground plane (i.e., microstrip transmission line), as functions of dielectric thickness between track and ground plane, and track width.

For example, from Figure 6, for a 0.060-inch thick pcb, with a 20-mil wide metal trace over a ground plane, the characteristic impedance, Z , is about 105 Ω .

Now, from Figure 7, the capacitance C_0 per inch for this 105 Ω transmission line is about 16pF/ft. Using the equation

$$(18) \quad L_0 = Z^2 \times C_0$$

one can easily compute L_0 , the inductance per foot of this 105 Ω transmission line, as 14.7nH/inch. So a 1-inch track of metal 20 mils wide will deliver 14.7nH of inductance. This track should be fairly straight; turns in the track will modify the inductance somewhat. The track should also be kept away from other metal, and the track should terminate into a low inductance ground plane. Secondary error terms like skin-effect can be neglected up to 500MHz.

Range Testing

Users experienced in range testing these types of wireless systems know all-too-well the orientation dependence of such systems. For less experienced users, range testing will be an acquired skill. Generally the user will choose a range on which to perform his relative evaluations, realizing that measurements vary as the range locale varies. If the range is not "open," proximity effects will usually result in a diminished range measurement. In addition, obstructions between the transmitter and receiver (walls, etc.) will also diminish range. (This doesn't mean that the user needs to run out to an open field to do each range test. Optimize your system in your immediate locale, then go out to the "open" field to make the final measurements.)

Systems such as these are usually range tested by mounting the receiver a prescribed distance above the ground (Micrel uses 4 feet). A tester walks away from the receiver holding a transmitter that he periodically actuates to verify decodes. At a distance where decoding fails, the tester reorients the transmitter until decodes again come through, then continues walking away from the receiver, until decoding again fails.

This process is repeated until no transmitter orientation can be found for which the system successfully decodes. As unscientific as it sounds, this is the methodology generally used for range testing. For users without previous range testing experience, the first few range tests will seem uncontrolled and cumbersome, but eventually the user will get a "feel" for range testing.

A reminder—Lots of little "hidden" things can go wrong during evaluation and testing that may elicit unrepeatable results. For example, forgetting to turn-off other competing signal sources, like RF generators and other SR receivers that are being tested in the laboratory (which reradiate regenerative noise) will corrupt range mea-

surements. Don't forget to disconnect any scope probes from the test board as well. For example, connecting a scope probe to the C_{TH} pin will introduce a small offset error which may affect range measurements. So before you make a conclusion, always double-check your setup.

Transmitter Options

Transmitters for MICRF001 applications are fairly simple structures, perhaps deceptively simple. On the surface, all that is required to construct a transmitter is a single-transistor oscillator that can be turned on and off by a data signal delivered from a (digital) encoder like those discussed in "MICRF001 External Components Selection: Encoder/Decoder Pairs." The three most common types of transmitters for MICRF001 applications are LC, SAW, and crystal oscillators. Design requirements such as cost, mechanical stability, packaging, and tuning will influence the choice among these three.

Unfortunately, however, things are never quite so simple. Designing circuits for UHF operation requires knowledge of device operation at these frequencies. For, example, above its self-resonant frequency, a capacitor is no longer a capacitor, but takes on inductive qualities. Circuit layout becomes critical. The transmitter development problem is further complicated by regulatory issues. Parameters such as transmit power, harmonic radiation, and data characteristics are regulated, and must meet certain criteria, which further vary across regulatory agencies. These agencies generally require that product be tested and qualified prior to sale.

Different levels of solution exist, depending on what the designer is willing to tolerate. For designers who possess the circuit expertise, oscillator designs can be found in the literature^{11,12}, and numerous products exist which can be used as a starting-point for development. Examples are:

- Part number KESTX01/02 transmitter IC from GEC Plessey Semiconductor (408) 438-2900, based on a crystal oscillator
- SAW transmitter application schematic in specification for RF Monolithics (800) 704-6079 SAW device, part number RO2103A
- TX99K LC-transmitter from Ming Microsystems (800) 473-6606, with accompanying schematic. This unit can also be purchased from Ming for resale, if desired.

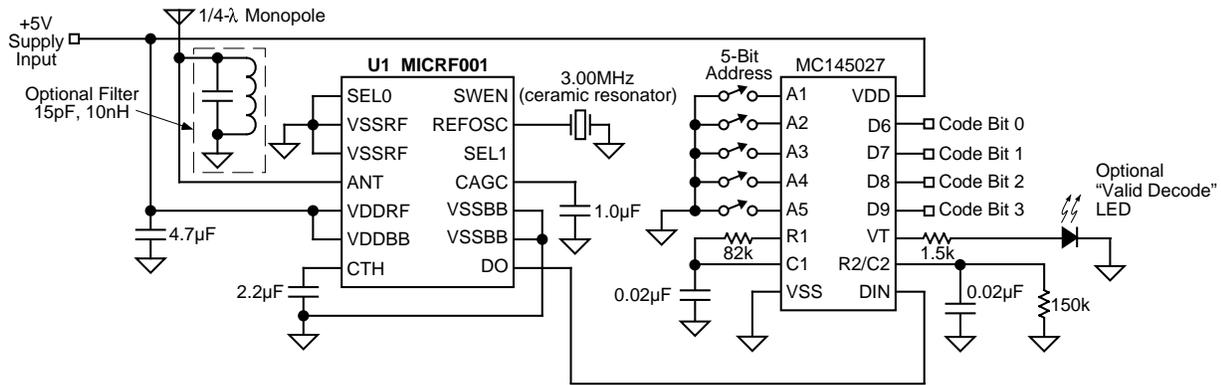
It must be understood that the final design, whatever it might be, must comply with all applicable regulations.

For the less adventurous, complete off-the-shelf, plug-in solutions are available. Two examples:

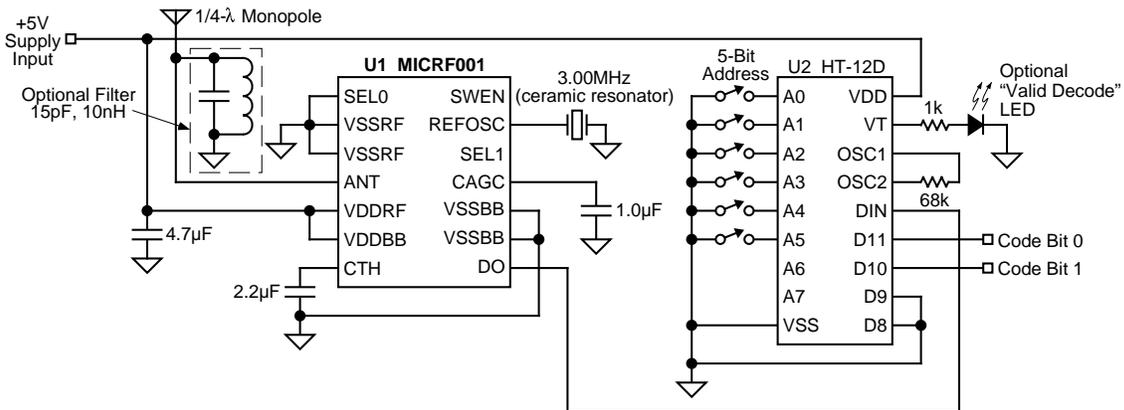
- Part number AM-TX1-xxx, from RF Solutions + 44 (0)1273 488 880 (UK)
- Part number TXM-xxx-LC, from Linx Technologies (541) 471-6256

These modules may meet the appropriate regulatory constraints. Even then, however, the final assembly probably will require certification by the regulatory body.

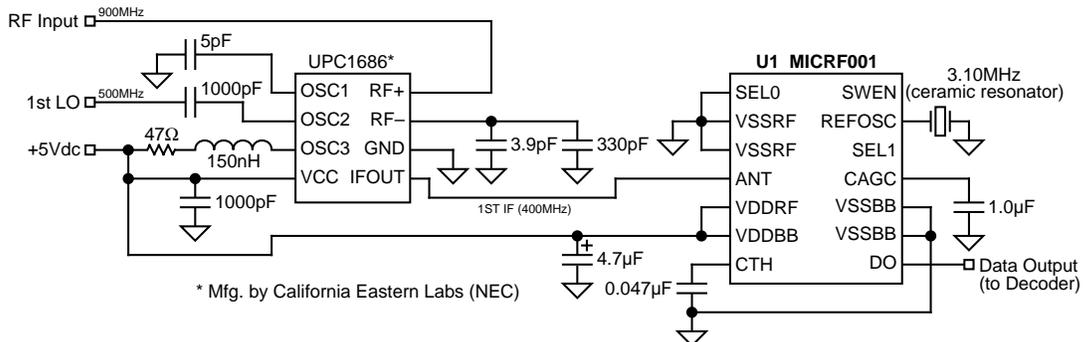
One final comment about designing and testing transmitters. At least in the United States, transmitters being developed for



**Figure 8b. MICRF001 Receiver/MC145027 Decoder
387MHz Operating Frequency
140bps Operation**



**Figure 8c. MICRF001 Receiver/Holtek Decoder
387MHz Operating Frequency
800bps Operation**



* Mfg. by California Eastern Labs (NEC)

Figure 9. 900MHz OOK Data Modern Application

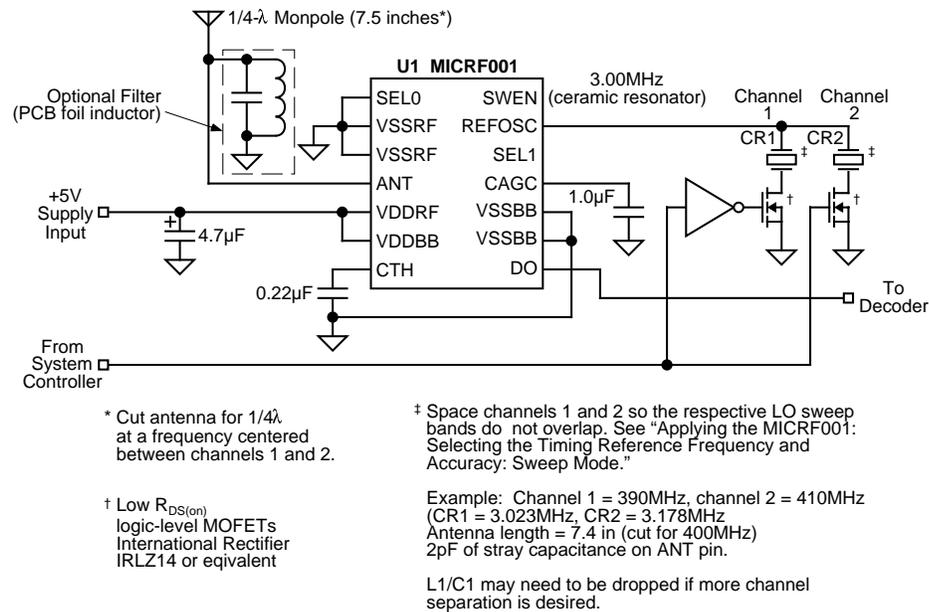


Figure 10. Dual-Channel MICRF001 Receiver

Glossary

Δf_{sw}

Symbol for the sweep frequency range used in this treatise. This is the range of frequencies over which the LO is "swept" within the MICRF001.

BPF

Band pass filter, which provides the frequency selectivity for superheterodyne type receiver.

FCC Part 15

Regulatory document for the United States for frequency band 260MHz to 470MHz.

f_{IF}

Symbol for intermediate frequency used in this treatise.

f_{LO}

Symbol for local oscillator frequency used in this treatise

f_{TX}

Symbol for transmit frequency (also called the RF input signal at the receiver) used in this treatise.

IF

Intermediate-frequency, generated in superheterodyne type receivers by "mixing" of the input RF signal with the local oscillator.

LC-transmitter

Simple, inexpensive transmitter applicable for FCC Part 15 applications. Transmit frequency is set by an inductor-capacitor (LC) resonant circuit. Lacks good initial frequency accuracy and stability.

LO

Local oscillator, a stable reference frequency "mixed" with the input RF signal to generate the output IF signal.

OOK (on-off-keyed modulation)

A type of RF carrier modulation appropriate for wireless data communications, similar to AM (Amplitude Modulation). Two logic states are conveyed, one signified by presence of a carrier, the other signified by absence of carrier. Alternative names are ASK (amplitude-shift-key), CPCA (carrier-present-carrier-absent).

pcb

Printed circuit board.

RF

Radio frequency, usually used in association with wireless systems.

SAW-transmitter

Transmitter whose frequency is based on the accuracy of a surface acoustic wave (SAW) device. Such transmitters have very good initial accuracy, temperature stability, and mechanical stability.

SH receiver

Superheterodyne receiver

SR receiver

Superregenerative receiver.

UHF

Ultra high frequency, which includes the band 300MHz to 440MHz.

Bibliography and Reference

Superregenerative Receivers

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