AN1539

An IF Communication Circuit Tutorial

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ABSTRACT

This article is intended to be a tutorial on the use of IF communication integrated circuits. The ISM band channel bandwidths and the Motorola MC13156 are used within this article as a platform for discussion. An examination of the devices topology is provided along with a discussion of the classical parameters critical to the proper operation of any typical IF device. The parameters reviewed are impedance matching the mixer, selecting the quad tank and filters and concluding with a overview of bit error rate testing for digital applications. Upon completion, the reader will have a better understanding of IF communications basics and will be able to specify the support components necessary for proper operation of these devices.

BACKGROUND ON ISM BAND

The instrumentation, scientific, medical (ISM) band targets three bands of frequencies. These bands are: 902–928 MHz, 2400–2483.5 MHz and 5725–5850 MHz. Spread spectrum techniques can be used within these bands to minimize adjacent channel interference. Systems operating with radiated power greater than 0.75 μW and up to a maximum of 1.0 W are required to implement either a frequency hopping scheme or a direct sequence scheme to lower the spectral power density on a particular channel. Systems with radiated power levels equal to or below 0.75 μW are not required to implement spread spectrum. The frequency hopping channels associated with this band must have a bandwidth of 500 kHz or less. The RF carrier modulation technique is usually conventional fm. For the purposes of this article we will concentrate on the 902–928 MHz band, as this band will furnish the carrier which is down–converted later in the article.

The two recommended methods of implementing the spreading function mentioned above operate as follows. The direct sequence scheme uses a pseudo–random data stream of at least 127 bits that is combined with the data to be modulated. The combination of the relatively low data rate baseband data with a high speed pseudo–random bit stream forces the spectrum of the baseband signal to occupy more bandwidth. By forcing this spreading, the power spectral density of the signal at any discrete frequency is reduced, thus limiting this signals propensity to interfere with other users. The frequency–hopping scheme requires that the modulator hop over at least 50 channels within 20 seconds. This

corresponds to a park time of 0.4 seconds/channel. This scheme reduces the likelihood of co–channel interference by intentionally limiting the amount of continuous time a carrier remains on channel. The spectral power density (SPD) measurement criteria for either scheme is the same — 8.0 dBm/3.0 kHz bandwidth. This is computed as follows:

$$SPD_{(dBm)} = 10 \log \left[\frac{(1 \text{ Watt}) (3 \text{ kHz})}{(500 \text{ kHz}) (1 \text{ mWatt})} \right]$$

Intuitively it may be obvious that spread spectrum systems would exhibit a better signal to noise ratio than their non–spread counterparts. This improvement in S/N ratio is termed processing gain. Systems operating within this band must exhibit 10 dB processing gain over systems not implementing a spreading function. For direct sequence systems, the processing gain is related to the pseudo–random code or chip rate. For frequency–hopping systems, the gain is related to the hopping rate and the number of channels being hopped.

Allocation of these bands by the FCC has fostered new product opportunities. These products include and are not limited to high frequency cordless phones, high speed wireless lans (>9600 bps), point to point data terminals, and wireless telemetry.

A FUNCTIONAL OVERVIEW OF IF ICs
FOR COMMUNICATIONS

In conventional radio, RF carriers conveying intelligence by way of some modulation technique are down–converted from the high frequency carrier to some other intermediate frequency through a process called mixing. Following the mixing process, the baseband signal is recovered through some type of demodulation scheme. We will be focusing our discussion on the IF portion of the receiver. By way of example, the Motorola MC13156 will be used as practical alternative to a purely theoretical discussion. This split IF device is targeted at the "IF strip" or portion of the radio. This integrated device contains all the blocks necessary to completely recover an original baseband signal from an incoming higher frequency carrier. The functional blocks associated with this device are common to all IF communication type integrated circuits, and as such provide a convenient and widely applicable vehicle for discussion.

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THE MIXER

The function of the mixer is to multiply the incoming RF carrier with a signal supplied by the local oscillator. The mixing process exploits the non–linear properties within the mixing device (i.e. a diode) to produce a whole host of product terms. As an example, consider the terms produced by a second order non–linear device with the following inputs:

$$F_{(lo)} = \cos \omega_{(lo)}t \text{ and } F_{(rf)} = A \cos \omega_{(rf)}t$$

The output spectrum is:

$$A0 + A \cos \omega_{(lo)}t + aA \cos \omega_{(rf)}t + b \cos^2 2 \omega_{(lo)}t + 2bA \cos \omega_{(lo)}t \cos \omega_{(rf)}t + bA\cos^2 2 \omega_{(rf)}t$$

By trigonometric expansion, the term $2bA \cos \omega_{(lo)}t \cos \omega_{(rf)}t$ expands to:

$$bA \cos (\omega_{(lo)} + \omega_{(rf)}) t + bA \cos (\omega_{(lo)} - \omega_{(rf)}) t$$

The desired multiplication product is the difference product which becomes the intermediate frequency. The products produced from this multiplication are largely dependant on the type of mixer employed. The most common topology used in current integrated circuits is the double balanced configuration. The balance term is used to describe the degree of isolation between the local oscillator and rf input and between these two inputs and the IF port. Double balanced mixers perform better in terms of suppressing leakage between these ports than do their single balanced counterparts. Analysis of the double balanced mixer also shows that all but the sum and difference frequencies are effectively cancelled out. This eases the task of passband filtering the resultant IF difference frequency. Critical parameters for the mixer include its conversion gain and input and output impedances and third order intercept. Higher conversion gain is an asset that eases filter selection (a higher loss can be tolerated will still deliver enough gain to run the limiter). The mixer input impedance needs to be matched to the filter that is band–pass filtering the signal that has been mixed down from a down–converter. In the case of the ISM band, the rf carrier would be in the 902 – 928 MHz range. The down–converter's mixer would convert this frequency to some frequency less than 200 MHz, the upper frequency limit of the device. Third order intercept is a measure of how well the distortion products in the device are suppressed. The higher the third order intercept point (specified in dBm) the greater the suppression of these distortion products.

THE LOCAL OSCILLATOR

The local oscillator, as mentioned earlier, is the other half of the mixer's drive. The LO within the MC13156 consists of a bipolar transistor pinned out to be configured as a Colpitts type oscillator. This internal transistor can also serve as a buffer for an external oscillator signal. The critical parameters associated with most IC forms of on chip LO's are stability and required drive. For discrete oscillators not using a crystal as their reference, external passive components are used to set the LO center frequency. Some degree of temperature stability can be achieved using components with complementary temperature coefficients. For very stable requirements, a crystal is recommended. In most cases, tuning of a crystal can be accomplished by using a series tunable inductor to "pull the crystal on frequency."

THE IF AMPLIFIER

The primary purpose of the IF amp is to amplify the output of the mixer. This output, which has been band pass filtered (the difference frequency produced by the mixer being the center frequency of the filter), requires amplification to ensure adequate drive to the limiter. The IF amp also supplies a current to the input of the summing amps of the RSSI circuit. Critical parameters associated with the IF are its input and output impedance and its gain. Input and output parameters are important because it is necessary to match these parameters to the band–pass filter impedances to minimize mismatch loss and subsequent loss in overall gain. Historically, the two most common IF frequencies are 455 kHz and 10.7 MHz. Both the 10.7 MHz and the 455 kHz ceramic filters used in IFs are easily matched to the MC13156 using external resistors when appropriate. Proper filter impedance matching to the IF amplifier is essential for these devices to operate to specification.

RSSI BLOCK

The received signal strength indicator is a summing network which takes the output currents from the IF amp and the limiter and produces a current that can be used to indicate the receive signal strength. Most IF devices are configured such that the current produced is a log function of the received signal. Critical parameters of the RSSI section include dynamic range measured in dB and its slope measured either in µV/dB or µA/dB. Non–linearities in RSSI dynamic range can usually be traced back to either improper filter matching or selection which results in excess insertion loss. Devices specifying RSSI slope in µA/dB use current sources at their outputs and thus require a resistor to ground to develop a indication (voltage) of the magnitude of the received signal.

THE LIMITER

The limiter takes its input from the IF amplifier and removes any amplitude variations from the IF frequency. It is necessary to remove these variations because the detector in addition to responding to frequency changes will also attempt to demodulate amplitude variations in the IF. These variations decrease sensitivity and introduce distortion in the recovered audio or bit errors in the recovered data. Important parameters for the limiter section include its limiting gain and input impedance.

THE DETECTOR

Most integrated IF circuits employ a quadrature detector for recovering the original baseband information (data/voice). This circuit is favored because the frequency dependant components, the inductor, capacitor, and the damping resistor are external to the device. The part of the circuit that is implemented internal to the device is the four quadrant multiplier. This multiplier is very economical to embed in silicon and is easily interfaced to the remaining external detector components. The quadrature detector operates as follows: First the incoming IF frequency is split into two parts. Second one of the parts is phase shifted by 90 degrees. Third the two parts are multiplied together and the result is the recovered baseband information. A more detailed discussion of the operation of the quadrature detector is given later.

THE DATA SLICER

The function of the data slicer is to reconstitute the data shape of the recovered baseband information. This information is usually heavily filtered at the transmitter and receiver IF to conserve spectrum and restrict noise. As a result of this filtering, the information is sinewave in nature and is not directly suitable for use in a digital system. Upon initial inspection, one may conclude that the slicer is just a comparator, however this is not the case. Incorporated within the slicer block is an auto–threshold mechanism that automatically tracks the dc value of the incoming data stream. Variation in the data stream's dc value is caused by the continuous change in the averaged duty cycle of the recovered data. Long strings of 1's and 0's are the cause of the extremes in this value. It is essential that the point of comparison (decision point) be tightly controlled throughout the recovery, or this dc variation may cause bit errors due to misinterpretation of the current data level.

MIXER OVERVIEW

MIXER INJECTION

Before discussing high/low side injection, a review of image frequency is in order. Recall that the function of the mixer in this case is to down–convert the incoming rf signal to 10.7 MHz. With an input frequency of 144.45 MHz, the mixer requires that the LO frequency be either 155.15 MHz (high side injection) or 133.75 MHz (low side injection). The choice of running either high or low side injection is largely a function of the adjacent spectrum in the area of the receiver's rf frequency. As an example, consider the case where the LO is

high side injected at 155.15 MHz. An image frequency now exits at 165.85 MHz. This frequency, if present from another transmitter at sufficient power level, will mix to produce its own 10.7 MHz difference product and corrupt/distort the intended signal. In a perfect world, proper planning could be exercised when selecting the LO frequency as to place the image in a band that is relatively quiet in a particular area. Since this is impractical, heavy premixer bandpass filtering should be and is employed to attenuate the image.

LAB MEASUREMENT OF MIXER INPUT IMPEDANCE

The network analyzer is a convenient tool that can be used to measure the input impedance of the mixer. Generally the procedure for measuring Z_{in} is as follows:

— set up the analyzer and calibrate out parasitics. These include interconnecting cable and pcb inductance and capacitance. This process is usually automated on newer analyzers. The process of calibration essentially tells the analyzer what a short, open and 50 ohms load looks like on the pcb. With the short, open and load conditions now established in the system, the IF device can be inserted into the pcb and measured.

— The analyzer is setup to sweep a set of frequencies around the incoming RF frequency, and a plot of the input impedance is displayed.

Since some integrated IF devices use a differential input type mixer, care must be taken to properly terminate the other input while measuring the device. On most integrated devices, the mixer's differential input impedance will track, eliminating the need to perform the second measurement. See Figure 1 for a termination example.
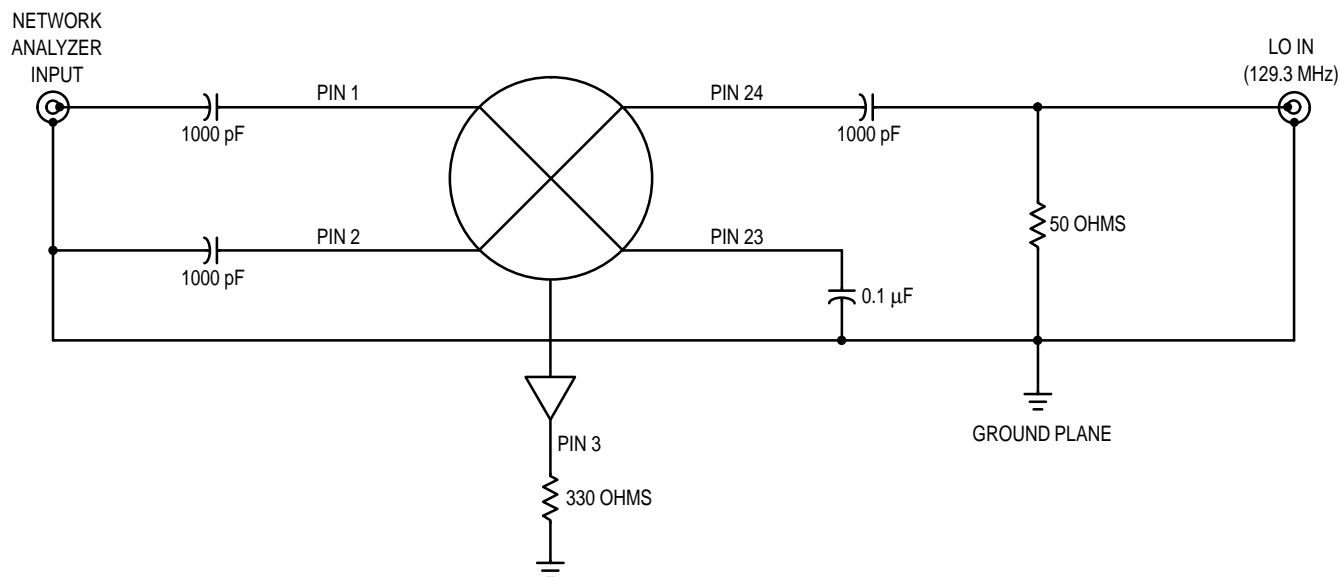

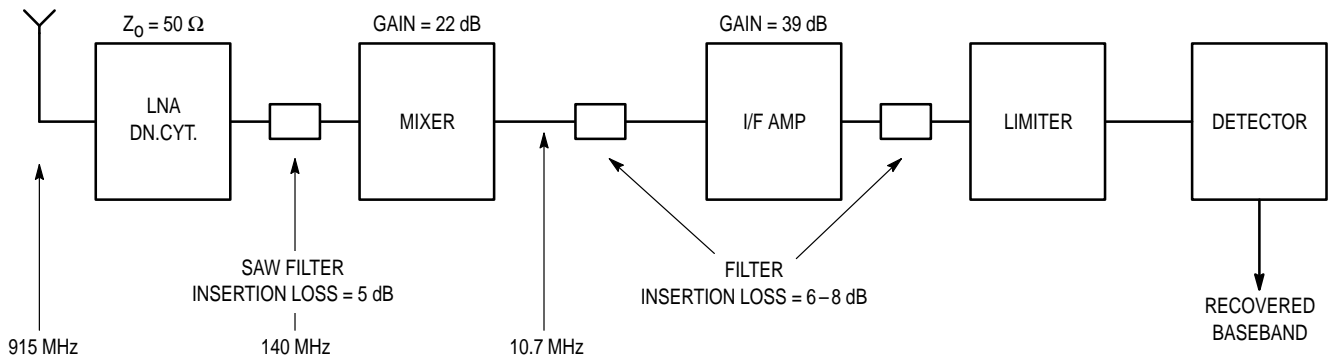**Figure 1. MC13156 Mixer Impedance Measurement Circuit**

Figure 2. IF Block Diagram

IMPEDANCE MISMATCH EFFECTS ON SENSITIVITY

To evaluate the effects of impedance mismatch on overall system sensitivity, consider the block diagram in Figure 2.

For discussion, let the 3.0 dB limiting sensitivity of the device be 6.0 μV or −91.4 dBm. This power level must be maintained into the input of the mixer in order for the limiter to maintain constant envelope amplitude into the detector. This number is specified with a 6.0 dB insertion loss assumed for both filters. The actual power delivered to the limiter input can be calculated by summing all the losses and gains in the IF system measured in dB. In Figure 2, assuming a −91.4 dBm input power to the MC13156, the limiter input power is:

$$V_{(lim)} = -91.4 \text{ dBm} + 22 \text{ dB} - 6 \text{ dB} + 39 \text{ dBm} - 6 \text{ dB} = -42.4 \text{ dBm}$$

The measured single–ended mixer input impedance is 82−j280 ohms. With no matching network employed, let's evaluate the effects of the mismatch in a system with a driving and transmission line impedance of 50 ohms. The load reflection coefficient is:

$$R = \frac{(Zl - Zo)}{(Zl + Zo)}$$

$$R = \frac{(82 - j280 - 50)}{(82 - j280 + 50)}$$

$$R = 0.91 \text{ at an angle of } -18.64 \text{ degrees}$$

To compute the mismatch loss (ML) in dB we use the following formula, where r is the magnitude of R:

$$ML = -10 \log (1 - |r^2|)$$
$$= -10 \log (1 - 0.83) = 7.69 \text{ dB}$$

The 3.0 dB limiting sensitivity has been degraded by 7.69 dB! Because of the mismatch our new 3.0 dB limiting sensitivity is −83.71 dBm or 14.58 μV! The sensitivity in voltage is found by solving the following:

$$V = \sqrt{10^{\frac{dBm}{10}} (50) (.001)}$$

DESIGNING THE MATCH

The measured series input impedance of the mixer at 144.45 MHz is 82–j280 ohms. This corresponds to an equivalent parallel impedance of 1038 ohms in parallel with 3.62 pF. The objective is to match this impedance to the output impedance of the SAW (surface acoustic wave) bandpass filter, which is being fed by a low noise amp (LNA) and down–converter. Let the output impedance of the filter be 50 ohms. See Figure 3a below.
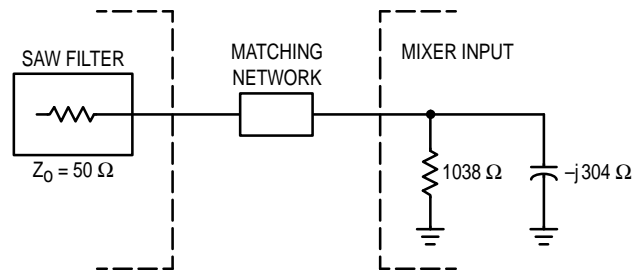

Figure 3a. Parallel Equivalent Circuit

To match the equivalent parallel input impedance, we need to parallel resonate out the 3.62 pF (−j304 Ω) capacitor with a parallel inductor and then match the remaining 1038 resistive portion to the 50 ohm source. This is done as follows:

$$L = \frac{1}{([(2)(\pi)(144.45 \times 10^6)]^2 \ (3.62 \times 10^{-12})} = 334 \text{ nH}$$

Now setting (as above) the Qs equal;

$$Qs = Qp = \sqrt{\frac{1038}{50} - 1} = 4.44$$

$$Xs = XC = 50(4.44) = 222 \ \Omega$$

$$Xp = XL = \frac{1038}{4.44} = 233 \ \Omega$$

thus C = 4.96 pF and L = 256 nH. The complete circuit is shown in Figure 4.

Notice that we have two inductors parallel. These can be replaced by 1 inductor with the value (256)(334)/(256)+(334) = 0.144 μH. The reader is encouraged to examine the applications section of the MC13156 data sheet and note the variable inductor value of 0.1 μH used in the mixer input circuit.
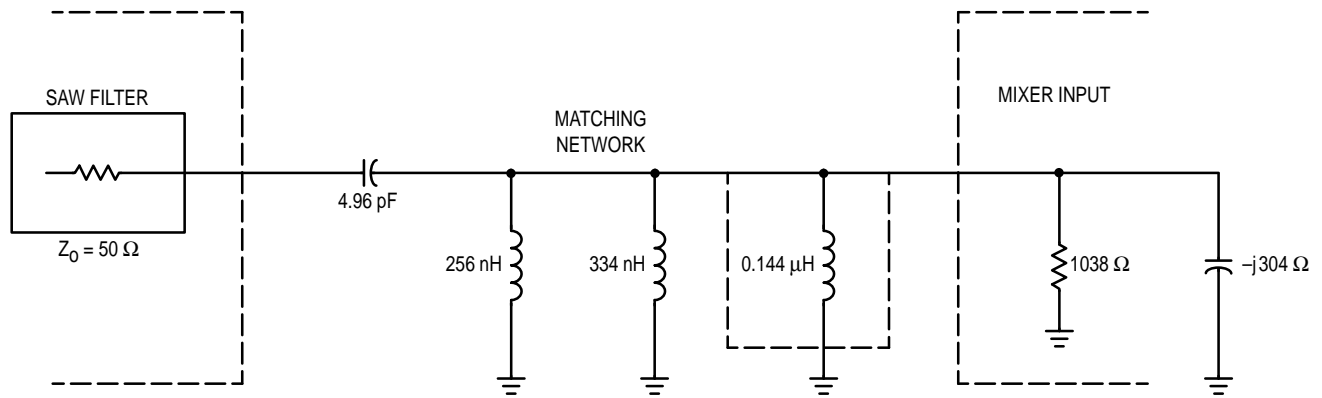
Figure 4. Parallel Equivalent Circuit

IF DEVICE SUPPORT COMPONENT SELECTION

IF FILTER SELECTION IN CONVENTIONAL Fm

For conventional radio (i.e. analog fm), the bandwidth of the IF section for very low distortion was easily determined. An estimation of the required bandwidth and thus the filter bandwidth can be obtained by using Carson's Rule:

$$Bw = 2(fm + \Delta Fc)$$

where Fm = the maximum modulation frequency and Δ Fc = peak carrier deviation (a function of the modulation index chosen.)

As an example, a 1200 Hz sine wave feeding a transmitter with a modulation index (M) of 0.5, would cause a carrier deviation (Δ Fc) = (Fm x M) = 600 Hz. Therefore, the required IF bandwidth is 3600 Hz.

To recover a square wave of the same frequency, the required bandwidth is subject to the harmonic content of the modulating waveform. To recover the square wave such that it exhibits minimal degradation in its rise and fall time, requires the recovery of at least the 7th harmonic, in our case 8400 Hz. Maintaining our modulation index of 0.5 gives a new IF required bandwidth of 2 x (8400 + 4200) = 25.2 kHz! Thus an analog system requires a bandwidth greater than 20 times the fundamental modulation frequency to reproduce the digital baseband information with little distortion.

IF FILTER SELECTION IN A GFSK ENVIRONMENT

To conserve bandwidth, digital systems filter the baseband digital signal to eliminate extraneous spectrum in the data. By filtering the data at the modulator and using a comparator to detect the zero crossing of the data at the receiver, we can reconstitute the baseband data. This comparator on integrated IF devices is called a data slicer.

A theoretical analysis of the derivation of the required bandwidth of a digital IF is beyond the scope of this article. As an alternative, a set of empirically derived values, which yield acceptable bit error performance (10^{-5}) when used in a Gaussian Filtered Shift Keying (GFSK) system, are listed. The required IF bandwidth is a function of the baseband data rate.

The values that follow are all a function of the baseband data rate (BDR).

Baseband data rate	=	100 Kb/s nrz (50 kHz effective)
IF bandwidth	=	BDR = 100 kHz
Baseband filter bandwidth	=	0.50 x BDR = 50 kHz
Carrier Deviation	=	0.32 x BDR = 32 kHz
Modulation Index	=	0.64

Note: that the 100 kHz required bandwidth is well within the ISM 500 kHz channel bandwidth.

This device, as mentioned earlier, is termed a "split" IF part. This comes from the fact that the single difference product can be filtered in two places, after the mixer and after the IF amp. As it is sometimes difficult to find a single filter device with the right attenuation characteristics, splitting the overall attenuation requirement between two filters usually results in a more timely and economical solution.

COMPOSITE BANDWIDTH EVALUATION

Since multiple filters are being used in this split IF, it becomes necessary to evaluate the combined effects of these filters on the overall bandwidth of the IF. An approximation of the 3.0 dB IF circuit bandwidth given the 3.0 dB bandwidth of the two filters is given by:

$$\text{Composite Bandwidth}_{(3\ dB)} = \frac{1}{\sqrt{\dfrac{1}{bw1^2} + \dfrac{1}{bw2^2}}}$$

As mentioned earlier, the required IF bandwidth in our 100 kb/s digital application was 100 kHz. Thus the composite bandwidth of the two filters must be at least 100 kHz. Filters with a spec'ed 3.0 dB bandwidth of 150 kHz will suffice, as their combined bandwidth is approximately 106 kHz.

FILTER INSERTION LOSS EFFECTS

There are two losses associated with the IF filters, insertion loss and mismatch loss. As shown earlier with mixer mismatch loss, the effect of additional loss was to decrease the receiver's limiting sensitivity on a dB by dB basis. The same decrease applies to insertion loss. All filters have a specified insertion loss. As long as the filter matching parameters are met, the insertion loss specified will be valid. Insertion loss increases as the match departs from ideal. The mismatch loss can be evaluated in a similar vain as was the mismatch loss of the mixer.

NOISE BANDWIDTH EFFECT ON SENSITIVITY

Thermal noise present in any IF circuit will degrade device performance. This reduction in performance manifests itself as a higher level signal requirement (more signal power) to maintain a given signal to noise ratio on a channel. The input noise power (Np) in an IF system of bandwidth B is calculated as:

$$Np = kTB$$

where   k = 1.38 x 10^{-23} (Boltzmans constant)
T = temperature in degrees Kelvin (°C + 273)
B = bandwidth of the system.

In our discussion above, assuming a 20°C ambient, when the analog bandwidth was 3120 Hz, the noise power is 1.26 x 10^{-17} Watts or −139 dBm. This is the noise floor. In the digital IF example where the composite bandwidth was 106 kHz, the new noise floor is 4.28 x 10^{-16} Watts or −123 dBm. The specified 3.0 dB limiting sensitivity on the MC13156 is −91.4 dBm; therefore, in a matched system with no other noise sources present, the noise power component in the IF bandwidth would be negligible. However, where this number becomes significant is when you add other system factors into the sensitivity equation. These factors, to name a few, include the receiver noise figure (that is, the amount of noise added to the system from internal amplifiers etc.), fade margin in all systems using other than line of sight transmission, and in digital systems, the required carrier to noise ratio for a given bit error rate using GFSK modulation. When these and other factors are considered, the required signal power at the receive antenna increases dramatically. Thus the noise power needs to be evaluated in concert with these other system factors before full realization of its effect on sensitivity can be understood. These factors can easily force the minimum received signal power requirement into the −75 dBm range.

FM DEMODULATION USING THE QUADRATURE TANK

With an overview of the quadrature detector having been given earlier, it is now possible to review the theory of operation behind the detector/quadrature tank. Recall that the carrier is split into two parts. One part goes directly to the four quadrant multiplier, while the other part is routed to the quad tank. With zero carrier deviation, the resulting IF center frequency is 10.7 MHz. As the carrier undergoes deviation (a phase shift), the IF signal will also deviate away from 10.7 MHz. The quad tank reacts to this deviation by producing yet an additional phase shift in the signal supplied to it. This phase shift can be as much as 90 degrees, and is proportional to the carrier deviation. Assuming a 90 degree initial phase shift, if one normalizes the instantaneous IF center frequency plus deviation as sine(∞t). The input signals to the multiplier are:

$V_{in(a)} = \sin(\omega t)$

$V_{in(b)} = \sin(\omega t + \frac{\pi}{2} (k)\Delta\omega) = \cos(\omega t - (k)\Delta\omega)$

since $\sin(a)\cos(b) = 0.5(\sin(a+b) + \sin(a-b))$

where a = ωt and b = ωt − (k)$\Delta\omega$

$V_{out} = 0.5(\sin(\omega t + \omega t - (k)\Delta\omega) + \sin(\omega t - \omega t + (k) \Delta\omega))$

The result of this function is an output at 2ωt and a low frequency component given by sin((k)$\Delta\omega$). This low frequency component is the recovered baseband audio or analog comparator input data.

SELECTING/DESIGNING THE QUAD TANK

The selection of the quad tank is easy once a decision is made as to what the IF center frequency will be. In our discussion, 10.7 MHz has been used and thus our tank's center frequency will match this selection. In some cases, the quad tank can be purchased as an assembly, with the cap and inductor in the same package. When the integrated tank is not available, a shielded inductor and shunt capacitor are used. The higher the unloaded inductor Q, the better, as a lower loaded Q value is always attainable using a shunt resistor.

DE–QUING FOR INCREASED BANDWIDTH

The bandwidth of the quad tank is usually set so that the tank has minimal effect on the overall IF. In our discussion, the IF bandwidth requirement was found to be 100 kHz. Setting the quad tank bandwidth at 1.5 times the IF bandwidth will in most cases satisfy this requirement. Thus our quad tank bandwidth will be 150 kHz. For example, consider the following:

Quad tank center frequency = 10.7 MHz
Capacitor value   = 150 pF

We need to solve for the inductance value which will cause resonance at 10.7 MHz.

$$L = \frac{1}{(2\pi 10.7 \times 10^6)^2 (150 \times 10^{-12})} = 1.47 \ \mu H$$

Note: This inductor is tunable to permit adjustment of the tank's resonant frequency to match the IF center frequency.

For this example, assume that the inductor has an unloaded Q of 100. Since the parallel equivalent resistance of the inductor (Rp) = (Q) x (Xl) and Xl = 99 ohms, Rp = 9900 ohms. Because the quad tank Q is related to the IF circuit Q, we need to determine the value of the IF Q before proceeding.

$$Q_{(I/F)} = \frac{I/F \ Center \ Freq.}{Composite \ Bandwidth} - \frac{10.7 \ MHz}{106 \ kHz} = 101$$

Since a good approximation of the quad tank bandwidth is 1.5 times the IF bandwidth, it follows that its Q should decrease. To "de–Q" the network we install a shunt resistor whose value is determined by:

$$Q_{(Tank)} = \frac{Tank \ Center \ Freq.}{(1.5)(I/F \ Bandwidth)} = \frac{10.7 \ MHz}{150 \ kHz} = 71.3$$

$Q_{(Tank)} = \frac{Rp'}{XL}$, where XL = 99 Ω, and Rp' is Rp II Rext

$Rp' = Q_{(Tank)} \times (XL) = 7.1 \ K\Omega$

using the conductance values of Rp' and Rp

$$Rext = \frac{1}{1.408 \times 10^{-4} - 1.01 \times 10^{-4}} = 25.1 \ K\Omega$$

Thus installing a 25.1 kohm resistor in parallel with the inductor with its existing Rp of 9.9 kohms will set the Q of the tank to approximately 71.

POST DETECTION FILTERING

Probably the most widespread post detection filter is the deemphasis filter used in broadcast FM. This filter, which is used to compensate for the pre–emphasis filter in the transmitter, equalizes the noise power between the low and high frequency components in the recovered audio from the detector. This filtering is mandated by the FCC. In a digital and analog receiver, and in particular the IF section, the recovered data that comes out of the detector is also accompanied by a 2 x IF component frequency discussed earlier. This component is usually easily filtered as it is much higher in frequency than the recovered baseband data. In our discussion, this 21.4 MHz component can be adequately attenuated by a single pole filter whose cut off frequency is set at 150 kHz, or 1.5 x the baseband data rate.

BIT ERROR TESTING

Figure 5 shows a block diagram of a typical bit error rate test (bert) setup. This setup would be applicable to the evaluation of any GFSK IF receiver device. The bert is supplied a clock from an external source in this example; however, it can optionally be used with its own internal clock. The generator produces a repeating data pattern called a frame word that is filtered before being applied to the rf modulator. The filter in this example, as mentioned earlier, removes the harmonic content of the modulating data. The MC13156 down–converts the rf signal and recovers the baseband 100 kb/s data. The recovered data is compared to the modulated frame word and errors are tabulated for display. Error rates of 10^{-4} are easily obtained with IF bandwidths in the $100-110$ kHz range.

FUNCTION GENERATOR

BIT ERROR RATE TESTER

RF GENERATOR
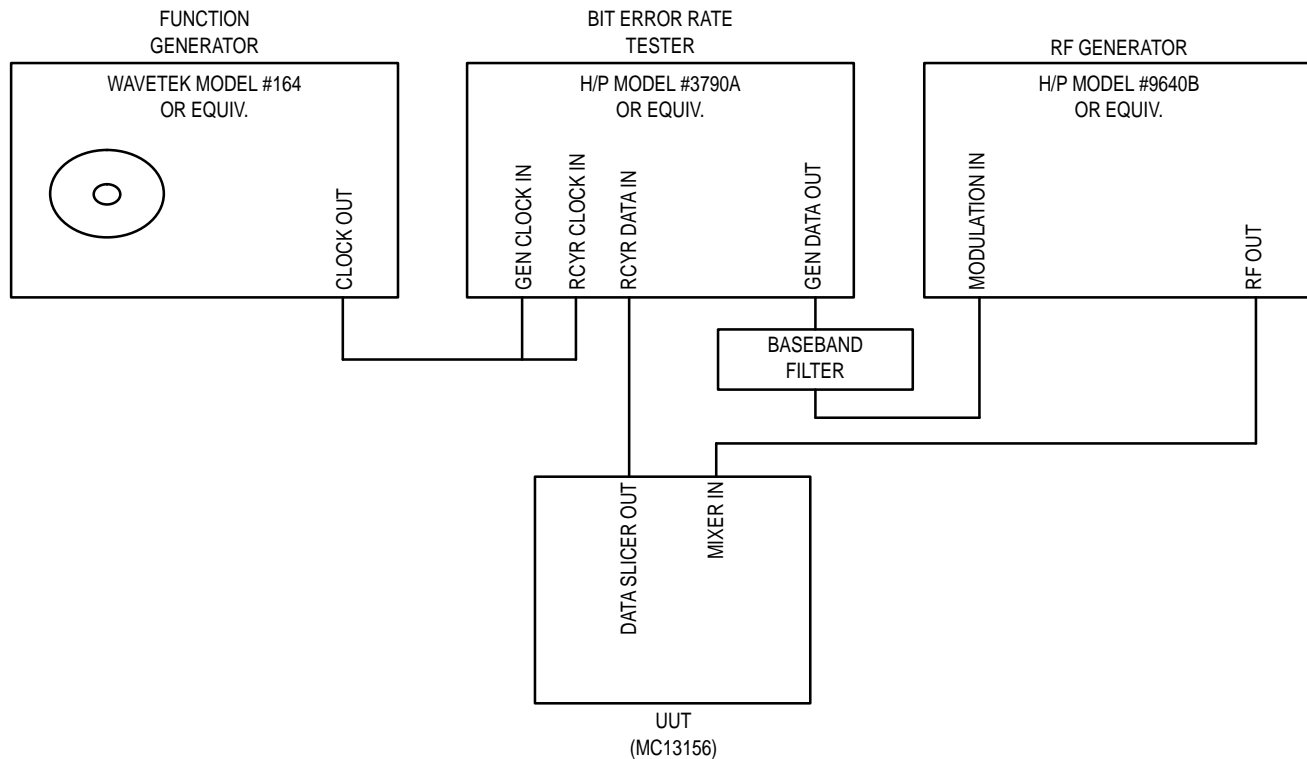
Figure 5. MC13156 Bit Error Rate Test Setup

SUMMARY

The use of integrated IF devices in wireless applications is so widespread that a basic understanding of the proper application of these devices is essential to both the novice and experienced designer. It is my sincere hope that this overview, having defined some of the basic attributes of FM IF strips, will serve as a cornerstone through which a broader understanding of wireless design can be secured.

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ACKNOWLEDGEMENT

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