Author: C. John Grebenkemper

Local Oscillator Phase Noise and its Effect on Receiver Performance

All superheterodyne receivers use one or more local oscillators to convert an input frequency to an intermediate frequency before the signal is demodulated. In the ideal receiver, these frequency conversions would not distort the input signal, and all information on the signal could be recovered. In a real-world receiver, both the mixer used for converting the signal's frequency and the local oscillator will distort the signal and limit the receiver's ability to recover the modulation on a signal. Mixer degradations, such as undesired mixing products, can be minimized by proper design in the rest of the receiver. The local oscillator degradations, which are principally random phase variations known as phase noise, cannot be decreased except by improving the performance of the oscillator.

Low oscillator phase noise is a necessity for many receiving systems. The local oscillator phase noise will limit the ultimate signal-to-noise ratio which can be achieved when listening to a frequency modulated (FM) or phase-modulated (PM) signal. The performance of some types of amplitude modulation detectors may be degraded by the local oscillator phase noise. When the receiver is used to monitor phase-shift keyed (PSK) or frequency-shift keyed (FSK) signals, the phase noise may limit the maximum bit error rate which the system can achieve. In FM/FDM (frequency division multiplex) systems, phase noise will often limit the maximum noise power ratio of the receiving system. Phase noise can limit the maximum angular resolution which can be achieved by an interferometric direction-finding receiver. Reciprocal mixing may cause the receiver noise floor to increase when strong signals are near the receiver's tuned frequency; this limits the ability to recover weak signals. All of these effects are due to local oscillator phase noise, and can only be reduced by decreasing the phase noise of the oscillator.

INTRODUCTORY THEORY

A perfect oscillator would be described mathematically by a sinusoidal waveform,

$$V = \cos [\omega_o t].$$

An actual oscillator will exhibit both an amplitude noise modulation, n(t), and a phase noise modulation, $\theta_n(t)$,

$$V = [1 + n(t)] \cos [\omega_o t + \theta_n(t)],$$

where n(t) and $\theta_n(t)$ are random processes. A good local oscillator will exhibit an amplitude-noise modulation power that is much less than the phase-noise modulation power. Furthermore, receiver mixers are usually run at a saturated input power, which will reduce their sensitivity to local oscillator amplitude variations. The net result is that amplitude noise insignificantly contributes to degradations in the receiver performance due to the local oscillator. For this reason, the amplitude noise can usually be ignored.

In its ideal form, the mixer in a receiver multiplies the RF input by the LO input to produce the sum and difference of the two input frequencies. The mixer is usually followed by an IF filter to select the desired IF output frequency. This process is illustrated in Figure 1. The input signal is assumed to be an unmodulated carrier, and the local oscillator is phase-modulated by its phase noise. The output of the frequency converter is at the sum frequency (or difference, depending on the IF filter) of the unmodulated carrier and local oscillator frequencies. The phase noise which was present on the local oscillator has been transferred to the input signal and now appears as a phase modulation of the input carrier. This effect can be extended to a modulated carrier, and results in the addition of an undesired phase noise modulation of the carrier to the desired signal modulation. This phase noise can result in additional noise at the output of the signal demodulator, depending on the type of modulation.

PHASE-NOISE DEFINITIONS

There are a number of ways to measure oscillator phase noise. Table 1 lists some of the more common definitions, along with a brief description on how each term is measured. The single-sideband (SSB) phase noise is the most common measure of oscillator phase instability. It can be directly measured on a spectrum analyzer, providing that the oscillator has low amplitude noise modulation and the spectrum analyzer local oscillators are lower in phase noise than the unit under test. This latter condition is usually

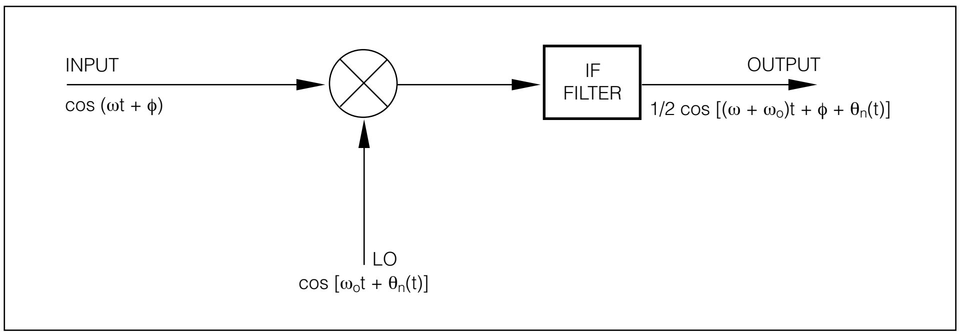

Figure 1. Effect of local oscillator phase noise on a frequency conversion.

Author: C. John Grebenkemper

the limiting factor in the sensitivity of this measurement method. When this measurement is done using an analog spectrum analyzer, the result is usually two to three dB better than what the oscillator is actually achieving. The prime reasons for this error are averaging done in the spectrum analyzer after the log detector and the difference between the resolution bandwidth of the analyzer and its noise bandwidth. Other methods can be used to measure phase noise with greater sensitivity and accuracy at the cost of considerably more complexity in instrumentation.

Incidental frequency modulation (IFM) is often used to specify overall oscillator instability. For IFM to be well defined, it should always be specified with a lower and upper frequency limit. For FM receivers, these limits are normally set to the lower and upper limits of the video pass-band. For other types of receivers the upper limit should be set equal to the IF bandwidth. If no upper limit is set in an IFM specification, then its magnitude tends to become very large. For phase-modulated signals, incidental phase modulation is preferred over IFM, since it provides a better measure of overall oscillator instability for that type of signal.

All of these measures of phase noise can be related to each other by the appropriate mathematical formulas. Table 2 gives the mathematical expressions that relate all of the phase noise definitions given in Table 1. Some of these formulas only apply under special conditions. SSB phase noise can only be converted from the various spectral density measures if the power in the phase fluctuations at frequencies greater than the offset frequency is much less than 1 radian2. The offset frequency at which this condition becomes valid can vary from tens of Hertz

Symbol	Units	Definition
$£(f)$	dBc/Hz	Single-sideband Phase Noise. This is the phase instability of the oscillator measured in the frequency domain. It is the most commonly used measurement of phase noise. A spectrum analyzer can be used to measure it if the oscillator has no amplitude noise modulation and the phase noise of the spectrum analyzers oscillators are less than the measured oscillator. The units of dBc/Hz refer to dB below the carrier measured in a 1-Hz bandwidth.
$S_{\delta f}(f)$	Hz^2/Hz	Spectral Density of the Frequency Fluctuations. This is the power spectral density of a frequency discriminator's output. It can be directly measured by connecting an audio spectrum analyzer to the output of a frequency discriminator whose input is the oscillator under measurement.
$S_{\delta\phi}(f)$	$Radians^2/Hz$	Spectral Density of the Phase Fluctuations. This is the power spectral density of a phase discriminator's output. It can be directly measured by connecting an audio spectrum analyzer to the output of a phase demodulator which has its input connected to the oscillator under test.
$S_y(f)$	1/Hz	Spectral Density of the Fractional Frequency Fluctuations. This is S(f(f) divided by the oscillator frequency squared. The main advantage of this unit of measurement is that it is invariant under frequency multiplication and may therefore be used to judge the relative quality of oscillators at different frequencies.
$\sigma_y(\tau)$		Two-point Allan Variance. This is a time domain measure of oscillator instability. It can be directly measured using a frequency counter to repetitively measure the oscillator frequency over a time period τ. The Allen variance is the expected value of the RMS change in frequency with each sample normalized by the oscillator frequency.
β_f	Hz	Incidental Frequency Modulation. This is a measure of the RMS frequency instability over a band of offset frequencies. It can be calculated by taking the square root of the spectral density of the frequency fluctuations integrated from a lower frequency limit to an upper frequency limit. It can be directly measured by passing the output of a frequency discriminator, whose input is the oscillator under test, through a bandpass filter and determining the RMS frequency variation.
β_ϕ	Radians	Incidental Phase Modulation. This is a measure of the total RMS phase instability over a band of offset frequencies. It can be calculated by taking the square root of the spectral density of the phase fluctuations integrated from a lower frequency limit to an upper frequency limit. It can be directly measured by passing the output of a phase discriminator, whose input is the oscillator under test, through a bandpass filter and determining the RMS phase variation.
f	Hz	Offset Frequency. This is the frequency of the phase or frequency fluctuations. When the oscillator is directly viewed on a spectrum analyzer, this becomes the offset from the carrier frequency.
f_o	Hz	Frequency of Carrier. This is the frequency of the oscillator which is being measured.

Table 1. Phase-noise definitions.

FUNDAMENTAL RELATIONSHIPS

$$S_y(f) = \frac{1}{f_o^2} S_{\delta f}(f)$$

$$S_{\delta\phi}(f) = \frac{1}{f^2} S_{\delta f}(f)$$

$$£(f) = \frac{1}{2} S_{\delta f}(f) \quad \text{if} \int_f^\infty S_{\delta\phi}(f')\,df' \ll 1\ \text{rad}^2$$

$$\delta y^2(\tau) = 2\int_0^\infty Sy(f') \left[\frac{\sin^2(\pi f\tau)}{(\pi f\tau)^2} - \frac{\sin^2(2\pi f\tau)}{(2\pi f\tau)^2}\right] df$$

$$\beta_f = \sqrt{\int_{f_a}^{f_b} S_{\delta f}(f)\,df}$$

$$\beta_\phi = \sqrt{\int_{f_a}^{f_b} S_{\delta\phi}(f)\,df}$$

$$\text{if} \int_{f_a}^\infty S_{\delta\phi}(f)\,df \ll 1\ \text{rad}^2\text{, then}$$

$$\beta_f = \sqrt{2\int_{f_a}^{f_b} f^2\, £(f)\,df}$$

$$\beta_\phi = \sqrt{2\int_{f_a}^{f_b} £(f)\,df}$$

FREQUENCY MULTIPLICATION RULE

$$£_{Mf_o}(f) = £_{f_o}(f) + 20\log(M)$$

Table 2. Phase-noise relationships.

Author: C. John Grebenkemper

to tens of kiloHertz, depending on the quality of the oscillator. The Allan variance can be directly computed from the fractional frequency fluctuations. However, the reverse is not true unless an assumption is made about the power law slope of the spectral density of the fractional frequency fluctuations. The frequency multiplication rule relates the increase in the SSB phase noise to multiplication integer, M. If an oscillator is multiplied in frequency by a factor of ten in an ideal multiplier, the oscillator's SSB phase noise will increase by 20 dB. Similarly, if the oscillator's frequency is divided by ten in an ideal frequency divider, its SSB phase noise will decrease by 20 dB.

LIMIT OF FM SIGNAL-TO-NOISE RATIO

The phase noise of a local oscillator will limit the maximum signal-to-noise ratio that can be achieved with an FM receiver. The oscillator phase noise is transferred to the carrier to which the receiver is tuned and is then demodulated by the FM discriminator. The phase noise results in a constant noise power output from the discriminator. If the phase noise has a power spectral density of, $S_{\delta\phi}(f)$, the output of the discriminator due to the phase noise is $f^2 S_{\delta\phi}(f)$. Figure 2 illustrates a simplified block diagram of an FM receiver. The bandpass filter on the output limits the video bandwidth to that required to pass the signal. The output signal-to-noise ratio is the power in the signal divided by the power in the noise. The power in the signal can be found by,

$$P_s = \lim_{T \to \infty} \frac{1}{T} \int_o^T K_d^2 K_f^2 v^2(t)\, dt,$$

where K_f is the modulator gain constant, K_d is the demodulator gain constant, and v(t) is the instantaneous modulating voltage. If we take the single-sided power spectral density of v(t), which is $S_v(f)$, then this equation becomes,

$$P_s = K_d^2 K_f^2 \int_{f_a}^{f_b} S_v(f)\, df$$

The phase noise of the local oscillator will generate a constant level of noise at the output of the FM receiver. If the RF input to the receiver is sufficiently strong, this source of noise will dominate and therefore limit the maximum signal-to-noise ratio that the receiver can achieve. The power in the noise is found from,

$$P_n = K_d^2 \int_{f_a}^{f_b} S_{\delta\phi}(f)\, df$$

Taking the ratio of these two powers yields a signal-to-noise ratio of,

$$\frac{S}{N} = \frac{K_d^2 \int_{f_a}^{f_b} S_v(f)\, df}{\int_{f_a}^{f_b} S_{\delta\phi}(f)\, df}$$

which simplifies to,

$$\frac{S}{N} = \frac{K_d^2 \int_{f_a}^{f_b} S_v(f)\, df}{\beta_f^2}$$

The local oscillator limited signal-to-noise ratio is equal to the power in the frequency deviation of the signal divided by the incidental frequency modulation squared.

If the FM transmission system uses preemphasis and deemphasis, then the modulator and demodulator gain constants change with frequency. Under this condition, the local oscillator limited sign al-to-noise ratio becomes,

$$\frac{S}{N} = \frac{\int_{f_a}^{f_b} K_f^2(f)\, K_d^2(f)\, S_v(f)\, df}{\int_{f_a}^{f_b} K_d^2(f)\, S_{\delta f}(f)\, df}$$

This latter formula is more difficult to evaluate than the simpler formula, which does not include the effects of preemphasis and deemphasis. However, for most FM transmission systems, the simpler formula will provide an answer which is within a few dB of the correct result. Since local oscillator phase noise performance can vary by this much, it is usually sufficiently accurate to use the simpler formula.

EXAMPLE

What is the local oscillator limited signal-to-noise ratio for an FM signal which has a 5-kHz RMS frequency deviation and a video modulation bandwidth of 300 Hz to 3 kHz? The local oscillator SSB phase noise is a constant -70 dBc/Hz from 100 Hz to 10 kHz.

SOLUTION

The solution requires us to find the modulating power in the signal and the incidental frequency modulation of the local oscillator. The signal has a 5-kHz RMS frequency deviation. The square of this is the modulating power contained in the signal.

P_s $(5\ \text{kHz})^2 = 2.5 \times 10^7\ \text{Hz}^2$

Since this oscillator exhibits a low phase

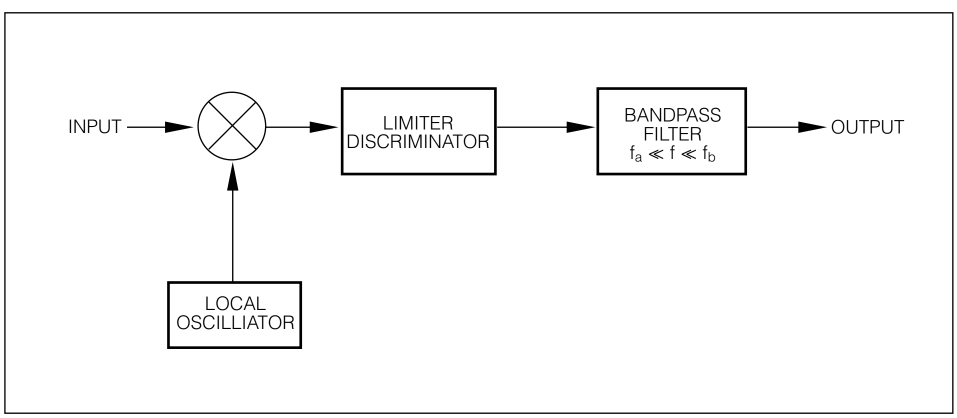

Figure 2. Block diagram of an FM receiver.

Author: C. John Grebenkemper

modulation power, the incidental frequency modulation may be found from,

$$\beta_f = \sqrt{2\int_{f_a}^{f_b} f^2\, £(f)\, df}$$

$$= \sqrt{2\int_{300}^{3000} f^2\, 10^{-7}\, df} = 42 \text{ Hz}$$

The power in the FM demodulator output is the square of the incidental frequency modulation.

$$P_N = \beta_f^2 = 1800 \text{ Hz}^2$$

The ratio of these two numbers yields the local oscillator limited signal-to-noise ratio.

$$\frac{S}{N} = 41 \text{ dB}$$

INCIDENTAL PHASE MODULATION

The local oscillator phase noise can limit the signal-to-noise ratio of a phase-modulated signal to which the receiver is tuned. A simplified block diagram of such a receiver is shown in Figure 3. In this case, the limiting signal-to-noise ratio is determined by the power in the phase modulation divided by the incidental phase modulation squared,

$$\frac{S}{N} = \frac{K_p^2 \int_{f_a}^{f_b} S_v(f)\, df}{\beta_\phi^2}$$

where, K_p is the phase modulator gain constant.

Phase modulation is usually used to transmit digital signals rather than analog signals. For digital signals, the phase of the carrier is shifted in integer multiples of a minimum phase step. For bi-phase shift keying, the phase shift is integer multiples of 180°; for quad-phase shift keying, it is multiples of 90°; and for eight-phase shift keying, it is multiples of 45° Other forms of digital transmission are used, but these usually involve both amplitude and phase-shift keying of the carrier. Local oscillator phase noise will effect the bit error rate performance of a phase-shift keyed digital transmission system. A transmission error will occur any time the local oscillator phase, due to its noise, becomes sufficiently large that the digital phase detection makes an incorrect decision as to the transmission phase. For instance, a QPSK transmission system will make a transmission error if the instantaneous oscillator phase is offset by more than 45° since the phase detector will determine that baud to be in the incorrect quadrant. Digital transmission systems with smaller phase multiples are more sensitive to degradation due to local oscillator phase noise.

The bit error rate degradation due to local oscillator phase noise can only be determined if the probability distribution of the local oscillator phase is known. This cannot be determined uniquely from the measurement of the phase noise without using a detailed model of the oscillator. Furthermore, if the oscillator is within a phase-locked loop, the probability distribution of the phase will be modified by the parameters of the phase-locked loop. For these reasons, it is not practical to attempt to predict the exact degradation in the bit error rate for a non-specific case. However, a rule of thumb can be used to predict the system performance. The rule states that for bit error rates greater than 10^{-6}, the system performance can be maintained to within a few dB of theoretical bit error rates for that modulation type if the incidental phase modulation of the local oscillator is less than one-tenth of the minimum phase step of the phase-shift keyed carrier. The incidental phase modulation should be computed from the natural frequency of the carrier recovery phase-lock loop to one-half of the IF bandwidth. For instance, in a QPSK system, the incidental phase modulation should be less than 9 degrees RMS to meet this rule.

EXAMPLE

A receiver has a local oscillator SSB phase noise given in the table below. What is the incidental phase modulation of the local oscillator integrated from 100 Hz to 1 MHz?

f	(f)
100 Hz	-70 dBc/Hz
1 kHz	-70 dBc/Hz
10 kHz	-70 dBc/Hz
100 kHz	-90 dBc/Hz
1 MHz	-120 dBc/Hz

SOLUTION

If we assume that the integrated power in the phase modulation is much less than 1 radian², then we can evaluate the approximate integral of the SSB phase noise, given in Table 2, to determine the oscillator incidental phase modulation. If the approximation is true, then the resulting answer will be much less than 1 radian. The numerical evaluation of this integral yields,

$$\beta_\phi = \sqrt{2\int_{100\text{ Hz}}^{1\text{ MHz}} £(f)\, df}$$

$$= 0.062 \text{ radians}$$

$$\beta_\phi = 3.6° \text{ RMS}$$

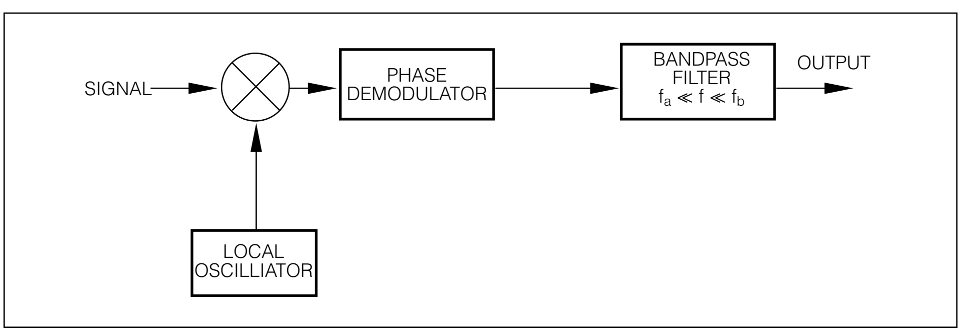

Figure 3. Block diagram of a phase modulation receiver.

The answer is indeed much less than 1 radian, which means that the initial assumption is true. The resulting answer indicates that this local oscillator could be used in a receiving system for an 8-PSK modulation which has a minimum phase shift of 45°.

RECIPROCAL MIXING

Reciprocal mixing will cause the receiving system to lose sensitivity when there is a strong signal near the frequency to which the receiver is tuned. This effect is due to the phase noise of the local oscillator modulating the carrier of the strong signal. The carrier is spread in frequency by the phase noise modulation, which results in a power spectral density that is proportional to the local oscillator's SSB phase noise. When the receiver is tuned to a frequency near the strong carrier, the power density in the strong carrier's noise sidebands may exceed the noise floor of the receiver. If it does exceed the noise floor, then the receiver sensitivity is limited by reciprocal mixing.

This effect is illustrated by Figure 4. The receiver is tuned in the frequency range of three carriers. The strongest carrier is in the center, with a weaker carrier on each side. The local oscillator has an SSB phase noise which decreases with increasing offset frequency. The three carriers will appear at the IF output, and each carrier will have been modulated by the local oscillator phase noise. The phase noise on each carrier will dominate until the noise power is reduced to the noise floor of the receiver. Otherwise, the noise floor is flat across the IF passband. The central carrier is the strongest, and therefore exhibits the strongest phase-noise component. The weaker carrier on the left has a smaller range over which its phase noise dominates. The weaker carrier on the right is nearly masked by the phase noise from the strong carrier. If the receiver was tuned to this carrier, it would achieve a much worse signal-to-noise ratio performance than would be predicted from the receiver's noise figure. This poorer performance is due exclusively to the local oscillator phase noise.

The increase in the noise floor of the receiver can be computed using the following methodology: The receiver noise floor in a one-Hertz bandwidth is the sum of the receiver's noise figure, F, in dB and -174 dBm/Hz,

$$P_n = F - 174 \qquad \text{(dBm/Hz)}$$

The noise generated in the receiver from a nearby carrier is the sum of the carrier power, Pc, in dBm and the SSB phase noise of the local oscillator at an offset frequency equal to the difference between the carrier frequency and the frequency to which the receiver is tuned.

$$P_o = P_c + \pounds(f) \qquad \text{(dBm/Hz)}$$

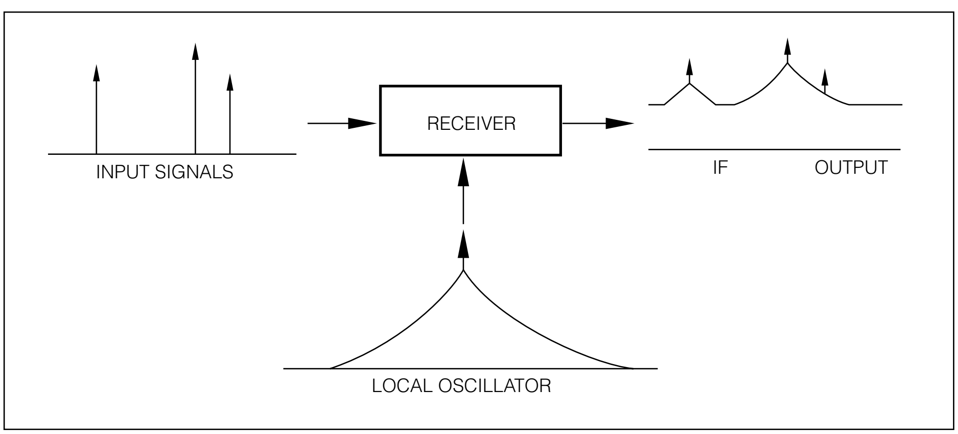

Figure 4. Reciprocal mixing model.

The apparent noise floor of the receiver is the sum of these two powers. To compute this sum, the powers must be converted to absolute power, summed, and then converted back to dBm. If it is necessary to compute the apparent noise floor at different frequencies, this process can be repeated at the desired offset frequencies. The net effect is that the receiver's apparent noise floor decreases as the receiver is tuned away from the carrier until it reaches the underlying noise floor generated by the receiver's front end.

EXAMPLE

A receiver with a 15 dB noise figure is tuned to a carrier with a -20 dBm power level. What is the equivalent receiver noise figure 1 MHz from the carrier when £(1 MHz) = -120 dBc/Hz?

SOLUTION

The noise flow due to the receiver's front end is,

$$P_n = F - 174 = -159 \text{ dBm/Hz}$$

At a 1 MHz offset frequency, the noise power due to the local oscillator phase noise is,

$$P_o = P_o + \pounds(1 \text{ MHz}) = -140 \text{ dBm/Hz}$$

The sum of these two powers at this offset from the carrier is an apparent noise floor of -140 dBm/Hz. The equivalent noise figure of the receiver is the difference between the apparent noise floor and -174 dBm/Hz,

$$F_{eq} = 34 \text{ dB}$$

The receiver noise figure is increased 19 dB when it is tuned 1 MHz away from the -20 dBm carrier.

LOCAL OSCILLATOR SPURIOUS SIGNALS

Besides exhibiting phase noise, local oscillators may also be phase or amplitude modulated by discrete frequencies. These oscillator modulations may produce a different effect than phase noise on the receiver perfor-

Author: C. John Grebenkemper

mance, since they are not generated by a random process. The sources of these discrete frequencies within the receiver are numerous. The power line frequency will often modulate the local oscillator. If the local oscillator is generated using frequency synthesis techniques, then reference frequencies used in the synthesizer will generate spurious signals. Other oscillators and digital frequency dividers in the receiver can generate frequencies which modulate the local oscillator. If the receiver uses a switching power supply, the switching frequency may modulate the local oscillator. While there are other potential sources of modulation, those mentioned above are the most common ones.

Figure 5 shows a plot of SSB phase noise for a local oscillator which is generated using an indirect frequency synthesizer. The peaks in the noise spectrum are generated by the discrete frequency modulation of the local oscillator. Visible in this spectrum are spurious components due to the 60-Hz line frequency, the 30-kHz power supply switching frequency, and the 250-kHz reference frequency of the phase-lock loop synthesizer. The spurious signals are given in units of dBc rather than the dBc/Hz of SSB phase noise. If the SSB phase noise is plotted relative to a different bandwidth, the amplitude of the random phase noise component will change, whereas the discrete spurious signal component will remain constant in amplitude.

Any local oscillator which is generated using a phase-locked loop will always have some spurious signals present in its output. The amplitude and frequency of these spurious modulations may vary as the local oscillator is tuned. Poor layout of the phase-locked loop oscillator circuitry may increase the amplitude and number of these spurious signals. However, even under ideal conditions some of the spurious signals will always be present. It is therefore necessary to define an acceptable level of oscillator spurious modulations.

A spurious specification can be broken down into two regions of interest: inside the video passband and outside the video passband. Any spurious signals present with a modulation rate which is in the video passband of the receiver output must not degrade the incidental frequency modulation or incidental phase-modulation performance required of the receiver. If this condition is not met, the receiver will not meet its desired local oscillator limited signal-to-noise ratio. With some types of demodulators it may be necessary to specify the spurious signals to a lower level at some frequencies. This is particularly true of digital demodulators, which usually contain phase-lock loops that can false-lock to a spurious modulation.

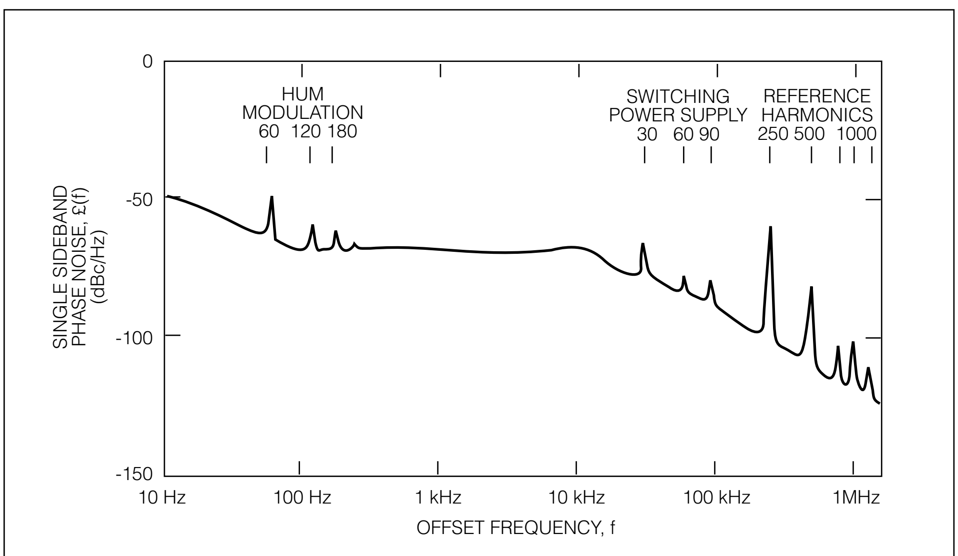

Figure 5. Local oscillator spurious signals.

Outside the video passband, the spurious signals should not degrade the receiver's spurious-free dynamic range. This condition is guaranteed if the spurious signals at offset frequencies greater than the narrowest IF bandwidth are further below the carrier than the spurious-free dynamic range specification. A receiver specified this way will not have any spurious responses from the local oscillator, which occur at power levels less than the input power required to generate intermodulation spurious responses. However, when the power level is sufficiently high, the receiver will have spurious responses that are due to the local oscillator. These spurious responses may be detected as if they were real signals.

A more stringent specification would require that the spurious signals not degrade the reciprocal mixing performance of the receiver. This condition will guarantee that the receiver will never detect any of the spurious signals as a real signal. This condition will be met if no spurious signals can be observed in the SSB phase noise when it is measured with a resolution bandwidth equal to the narrowest IF bandwidth used in the receiver. In effect, the local oscillator SSB phase-noise power in this bandwidth exceeds the spurious-signal power. A specification of this type can be very difficult to meet.

CONCLUSIONS

The local oscillator phase noise will effect the overall performance that can be achieved in a receiving system. Great care should be exercised in determining the desired receiver performance. Once these requirements have been determined, they can be translated into a required level of local oscillator performance. Conversely, if the local oscillator performance is already known, it can be translated into a set of system level performance data.

Overly stringent specifications for the local oscillator performance should be avoided. Improving the oscillator performance of a given design is usually very expensive, both in engineering and production times. Generally, a local oscillator used in an FM receiver does not have to be nearly as low in phase noise as one used in a PSK receiver.

Finally, no data regarding the measurement of phase noise has been presented in this article. The interested reader is referred to the references for more information in this area.

REFERENCES

1. "Time and Frequency: Theory and Fundamentals," Byron E. Blair, Editor, NBS Monograph 140, US GPO, 1974.
2. Floyd M. Gardner, "Phase-lock Techniques," ,John Wiley & Sons, 1979.
3. Peyton Z. Peebles, "Communication System Principles," Addison-Wesley, 1976.
4. Dieter Scherer, "Learn About Low-Noise Design," Part I, Microwaves, April 1979.
5. Dieter Scherer, "Learn About Low-Noise Design," Part II, Microwaves, May 1979.
6. Michael C. Fischer, "Analyze Noise Spectra With Tailored Test Gear," Microwaves, July 1979.
7. "Understanding and Measuring Phase Noise in the Frequency Domain," Hewlett-Packard Applications Note 207.

Vol. 8 No. 6 November/December 1981

WJ Communications, Inc. • 401 River Oaks Parkway • San Jose, CA 95134-1918 • Phone: 1-800-WJ1-4401 • Fax: 408-577-6620 • e-mail: sales@wj.com • Web site: www.wj.com