**Communication Techniques, Inc.**

**Source and Synthesizer Phase Noise Requirements for QAM Radio Applications**
by William Reuter, Senior Design Engineer, Synthesizer Group, CTI

QAM modulation is widely used in digital microwave radios for achieving high data transmission rates (up to 155 Mbits/sec) over relatively narrow bandwidths (see Table 1). Digital radio systems designed using such modulation schemes must balance the effects of phase noise from local oscillators with the demodulator parameters in determining overall performance. In a M-QAM system each symbol transmitted contains k bits, where $2^k = M$ and k is typically 4 to 8. In contrast, BPSK transmits 1 bit per symbol and QPSK transmits 2 bits per symbol while requiring the same RF bandwidth for a given symbol rate. The M states in a high order QAM symbol constellation (see Fig. 1) are more closely spaced than in BPSK or QPSK and therefore require lower noise relative to the average carrier power to eliminate errors. The rms phase noise of the local oscillators — after being filtered by the demodulator — must be low enough to not cause bit errors and not significantly degrade the fade margin of the link. Additionally the local oscillators must exhibit low microphonics since only a small phase deviation could cause an error. It is useful to be able to answer the questions "How should phase noise be characterized?" and "What phase noise is really necessary?" in order to properly specify the phase noise of local oscillators to meet system requirements without overspecifying. This article describes how basic demodulator parameters interact with phase noise to affect performance and provides a method to determine what LO phase noise is required for typical QAM systems. Although for simplicity this analysis assumes M = 16, 64, 256 or 1024, the results for similar systems with M = 32, 128 or 512 can usually be estimated by interpolation.

**Table 1. Selected Parameters of Some M-QAM Systems**

| M = | 4 (=QPSK) | 16 | 64 | 256 | 1024 |
|---|---|---|---|---|---|
| Number of bits per symbol = k = theoretical spectral efficiency in bits/sec/Hz | 2 | 4 | 6 | 8 | 10 |
| Practical Spectral Efficiency in bits/sec/Hz (approx) | 1.66 | 3.33 | 5.00 | 6.66 | 8.33 |
| Min. Symbol Rate in Msym/sec (no FEC) to transmit: | | | | | |
| 1.544 Mbits/sec (DS1) | 0.77 | 0.39 | 0.26 | 0.19 | 0.15 |
| 2.048 Mbits/sec (E1) | 1.02 | 0.51 | 0.34 | 0.26 | 0.20 |
| 6.312 Mbits/sec (DS2) | 3.16 | 1.58 | 1.05 | 0.79 | 0.63 |
| 8.448 Mbits/sec (E2) | 4.22 | 2.11 | 1.41 | 1.06 | 0.84 |
| 12.352 Mbits/sec (8 DS1) | 6.18 | 3.09 | 2.06 | 1.54 | 1.24 |
| 24.704 Mbits/sec (16 DS1) | 12.4 | 6.18 | 4.12 | 3.09 | 2.47 |
| 34.368 Mbits/sec (E3) | 17.2 | 8.59 | 5.73 | 4.30 | 3.44 |
| 44.736 Mbits/sec (DS3) | 22.4 | 11.2 | 7.46 | 5.59 | 4.47 |
| 90 Mbits/sec (~2 DS3) | 45.0 | 22.5 | 15.0 | 11.3 | 9.00 |
| 155.52 Mbits/sec (STS-3) | 77.8 | 38.9 | 25.9 | 19.4 | 15.6 |
| S/N for $10^{-3}$ BER (dB) | 9.8 | 16.5 | 22.6 | 28.4 | 34.3 |
| for $10^{-6}$ BER | 13.6 | 20.4 | 26.6 | 32.6 | 38.5 |
| for $10^{-7}$ BER | 14.3 | 21.2 | 27.4 | 33.4 | 39.4 |
| for $10^{-8}$ BER | 15.0 | 21.9 | 28.1 | 34.1 | 40.1 |
| for $10^{-9}$ BER | 15.6 | 22.5 | 28.7 | 34.7 | 40.7 |
| for $10^{-10}$ BER | 16.1 | 23.0 | 29.2 | 35.2 | 41.2 |
| $E_b/N_0$ = S/N minus (dB) | 3.0 | 6.0 | 7.8 | 9.0 | 10.0 |
| Peak Power to Average power (dB) | 0 | 2.6 | 3.7 | 4.2 | 4.5 |
| Average Power relative to $d^2$ power (dB) | 3.0 | 10.0 | 16.2 | 22.3 | 28.3 |
| Smallest phase deviation for error (deg.) | 45.0 | 16.9 | 7.7 | 3.7 | 1.8 |

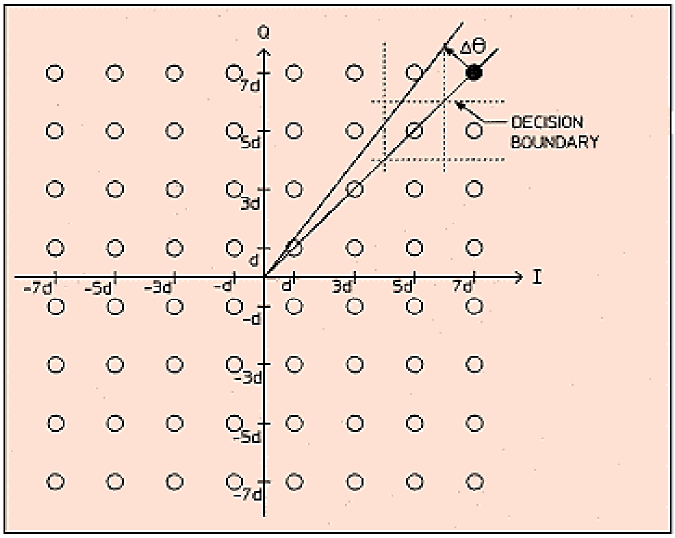


*Figure 1. 64 QAM Constellation*

### QAM Block Diagram

A block diagram of a typical QAM radio link is shown in Figure 2. The local oscillators used for the final up conversion and the first down conversion usually are the largest phase noise contributors because they are at the highest frequencies. The phase noise added by the LOs used in lower frequency up and down conversion (and by the modem LOs) is typically negligible.

To obtain coherent demodulation, a carrier recovery PLL is used to phase lock the demodulator local oscillator VCO to the IF frequency. The quadrature outputs of this VCO separate the I and Q components of the signal in the conversion to baseband. Each component then goes through a matched Nyquist lowpass filter which has a bandwidth of $F_S/2$ where $F_S$ is the symbol rate. The demodulator usually includes an adaptive equalizer which contains tapped delay lines with tap coefficients that are dynamically adjusted to minimize intersymbol interference (ISI). The decision slicer determines which of the M states is nearest to the received vector and computes the difference in phase between the received vector and that state. This phase error signal is fed to a loop amp which outputs a tuning signal to the VCO, closing the PLL. All of these demodulator functions can be implemented digitally.

Forward Error Correction (FEC) algorithms are used to correct infrequent bit errors at a cost of a fractional increase in the number of bits that need to be transmitted. Interleaving spreads out the bits making it easier to correct a burst of errors.

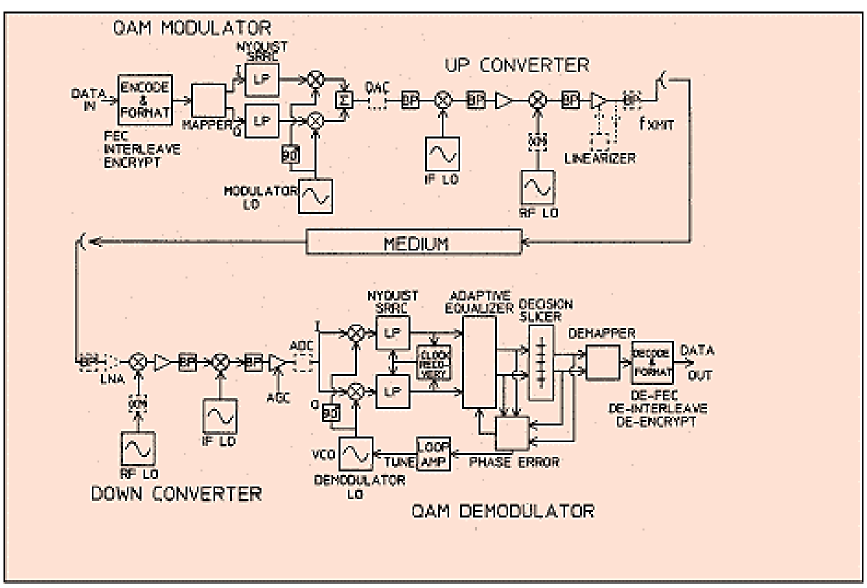


*Figure 2. Typical QAM System Block Diagram*

## Bit Error Rate vs. S/N

The theoretical bit error rate for a matched Nyquist filter QAM receiver for M = 4, 16, 64, 256 and 1024 assuming equiprobable symbols and no FEC is

$$P_e\,(S/N)= \frac{1-\left[1-2*\left(1-\frac{1}{\sqrt{M}}\right)*Q\left(\sqrt{\frac{3*(S/N)}{(M-1)}}\right)\right]^2}{\log_2 M} \quad \text{where } Q(x)=\frac{\int_x^{\infty} e^{\frac{t^2}{2}}dt}{\sqrt{2\pi}}$$

S/N = average signal power / noise power in two-sided Nyquist bandwidth

$$F_S = \frac{E_b}{N_b} * k$$

These BER curves are shown in Figure 3 and are evaluated for some BERs in Table 1. Note that there is only a 2.5-2.7 dB difference in S/N between a BER of $10^{-6}$ and a BER of $10^{-10}$. System performance requirements specify availability as the percentage of time the BER must be below a given threshold such as $10^{-6}$ or $10^{-3}$. $10^{-6}$ is a typical BER threshold for data transmitted on microwave links and is used in the CCIR Rec. 592-2, 1994 F Series, Part 1 definition of Degraded Minutes. $10^{-6}$ is assumed to be the threshold in this analysis and $(S/N)_6$ is defined by $P_e\,((S/N)_6) = 10^{-6}$. FEC typically lowers the S/N needed to obtain a given BER by 2-4 dB relative to the unFEC'd curves.

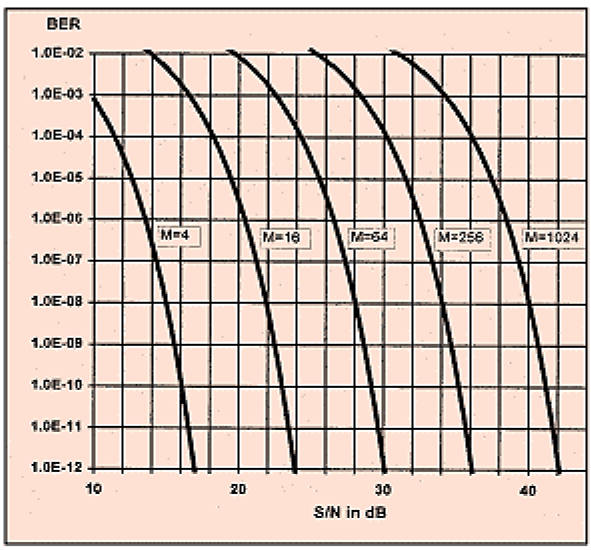


*Figure 3. BER vs. S/N for M-QAM*

## Residual S/N and Fade Margin

Under normal (unfaded) conditions on an LOS link, the ratio of the received signal level (RSL) to receiver thermal noise C/N = RSL / (kTBF) is 30-65 dB higher than $(S/N)_6$. A high normal C/N is used to provide margin for signal fading which may intermittently occur primarily because of multiple path effects (and rain above 10 GHz). However, a low residual BER remains (usually < $10^{-10}$) at high C/N due to residual noise from equipment imperfections and external effects. Local oscillator phase noise is one source of the residual noise. Other potential contributors include amplitude and group delay distortion, power amplifier intermodulation products, quantization errors, carrier recovery PLL noise, clock recovery loop jitter and residual ISI from Nyquist filters. External factors may include multipath dispersion. Many of the noise contributions are reduced by the adaptive equalizer. The accumulation of these noise sources tends toward a Gaussian pdf and produces an effective $(S/N)_{res}$ which corresponds to the residual BER observed at high RSL. For this analysis $(S/N)_{res}$ is assumed to be independent of RSL which is usually a reasonable approximation except at very low input levels. If a simplifying assumption is made that the components making up $(S/N)_{res}$ are independent additive Gaussian, then the inverse S/Ns add powerwise and $(S/N)_{res} = \frac{1}{\Sigma(N/S)_i}$ where $(N/S)_i$ = the effective noise to signal ratio due to the ith source.

Fade margin is the difference between the unfaded RSL and the RSL at which the BER is at the defined threshold (= $10^{-6}$). On the occurrence of a deep fade due to heavy rain or multipath, the residual noise adding to receiver thermal noise affects performance as a degradation of the fade margin. If $(C/N)_6$ is defined as the C/N required to obtain a BER=$10^{-6}$ in the presence of $(S/N)_{res}$ then D, the degradation of the fade margin, is the ratio of $(C/N)_6$ to $(S/N)_6$. This is also known as the implementation loss. If it is assumed that the receiver front end thermal noise and the residual noise are additive Gaussian, then $(S/N)_6 = \frac{1}{(N/C)_6 + (N/S)_{res}}$ and the degradation in fade margin due to the residual S/N is $D = \frac{(C/N)_6}{(S/N)_6} = \frac{1}{1-\frac{1}{Y}}$ (= $-10 * \log_{10}(1-\frac{1}{Y})$ in dB) where $Y = \frac{(S/N)_{res}}{(S/N)_6}$ is the residual S/N level relative to $(S/N)_6$. For example, if $(S/N)_6$ = 26.6 dB for 64 QAM and $(S/N)_{res}$ = 32.6 dB, then Y = 6 dB, the fade margin degradation is 1.26 dB and $(C/N)_6$ = 27.86 dB. The curve of D vs. Y, plotted in Figure 4, is useful in showing the change in fade margin due to a change in residual noise. Note that the curve is symmetrical, that is, Y and D can be interchanged to determine Y for D > 3 dB.

A degradation in fade margin Di could be defined for each noise source individually by substituting $(S/N)_i$ for $(S/N)_{res}$. The overall D is calculated by finding $(S/N)_{res}$ as the inverse of the $(N/S)_i$ summation and then referring to Fig. 4.

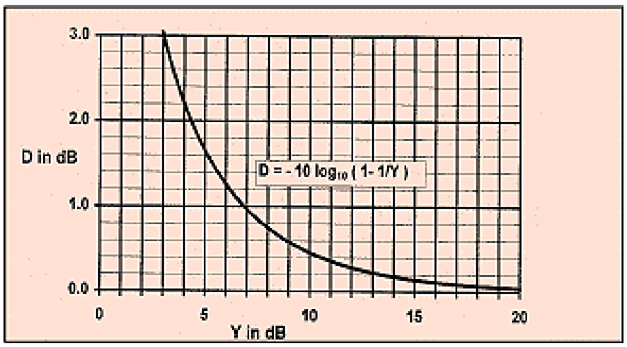


*Figure 4. Degradation in Margin D due to Y - $Y=(S/N)_{res}/(S/N)_6$*

**Carrier Recovery PLL** Local oscillator phase noise is usually a major part of residual noise. Its effect on performance depends on the noise it produces at the output of the demodulator. In general, the mean square phase noise at the demodulator output due to the phase noise of a local oscillator is given by $\sigma_\theta^2 = \int_0^\infty S_\theta(f)\,|W(f)|^2 df$ where $S_\theta(f)$ = power spectral density of the phase noise = 2* $\mathcal{L}(f)$ (provided the integral of $\mathcal{L}$ from f to infinity is <<1 radian$^2$)

$\mathcal{L}(f)$ = the single sideband (SSB) phase noise power/Hz to carrier power ratio vs. offset frequency

W(f) = the transfer function from local oscillator phase to demodulator output phase

The demodulator is effectively a multi-octave bandpass filter to phase variations since

- the Nyquist LP filter rejects frequency components above $F_S/2$
- the carrier recovery PLL acts as an HP filter by rejecting low frequency components

The purpose of the carrier recovery PLL is to track out low frequency variations in the carrier frequency and to maintain phase coherence. The IQ mixers, low pass filters, symbol decision slicer and phase error estimator form a type of phase detector with low pass filter. The phase error is fed to a loop amp where the gain and breakpoint frequencies are set. The output of the loop amp is used to tune the VCO. The phase lock loop model in Figure 5 can be used to estimate the behavior of the carrier recovery PLL with respect to phase deviations. $\theta_R$ represents the phase error (relative to the transmitted constellation point) of the received IF signal and $\theta_D$ represents the phase error at the demodulator output. The dashed box approximates the effect of the IQ mixers, $F_S/2$ lowpass filters, adaptive equalizer, decision slicer and phase error calculator. The open loop gain is $GL(f) = G_a(f) * G_{LP}(f) * \frac{K_V}{jf} \sim B_0 * \frac{(jf+Fz)}{(jf)^2} * e^{-j2\pi f Td} * |G_{LP}(f)|$ where $G_a(f) = K_a * \frac{(jf+Fz)}{(jf)}$ (type II 2nd order PLL assumed) $B_0 = K_a * K_V$ with $K_a$ in V/radian and $K_V$ in Hz/V $\zeta = 0.5 * \sqrt{\frac{B_0}{F_z}}$ is the damping factor and $G_{LP}(f) \sim |G_{LP}(f)| * e^{-j2\pi f Td}$ where $|G_{LP}(f)|$ is the Nyquist filter amplitude response and the phase response is modeled for simplicity by a time delay Td.

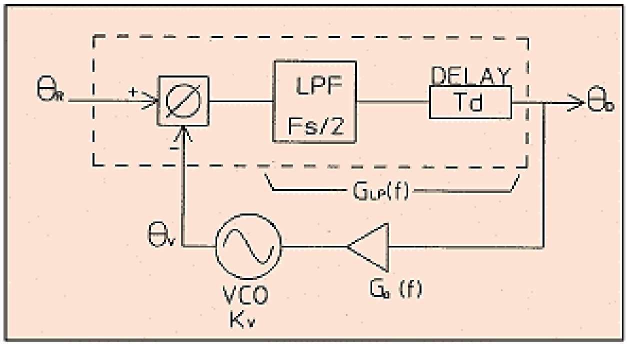


*Figure 5. Model for Carrier Recovery PLL*

The transfer function of most interest is the magnitude response of the output to phase variations on the input signal, which for practical purposes defines |W(f)|:

$$|W(f)| = |\Theta_D(f)/\Theta_R(f)| = \left|\frac{1}{1+GL(F)}\right| * |G_{LP}(f)|$$

The low frequency part of this function is plotted in Figure 6 for a damping factor of 0.7 ($Fz=B_0/2$) which is commonly used in carrier recovery PLLs. This transfer function, denoted the suppression function, provides rejection of input phase noise that increases at a rate of 40 dB per decade as the offset frequency decreases below Fz. Frequencies above $F_S/2$ are rolled off by $|G_{LP}(f)|$.

$B_0$ is the zero dB crossing of the Bode plot of open loop gain (for $Fz < B_0$) and is used here as a convenient measure of the PLL bandwidth. By making $B_0$ higher, the rejection of LO phase noise and microphonics is increased and acquisition is faster. However as $B_0$ is increased 1) the delay in the Nyquist filter and other circuits may cause peaking and 2) noise may increase due to other sources such as the loop amplifier. The delay depends on the symbol rate, the number of the tapped delay lines and the configuration. Figure 6 includes the suppression functions for $B_0$ = 0.02 to 0.08 times 1/Td. For $B_0$ >= 0.04 x (1/Td) the PLL significantly elevates the phase noise out to ~ 4 x $B_0$. The integrated phase noise may increase as $B_0$ is increased if the local oscillator noise is not rolling off fast enough. Note that these curves all have a -6 dB point at about $B_0/2$ = Fz and are essentially identical below Fz. The effects of the adaptive equalizers and other digital processing circuits may be difficult to model exactly but the overall demodulator suppression function can often be determined by measurement.

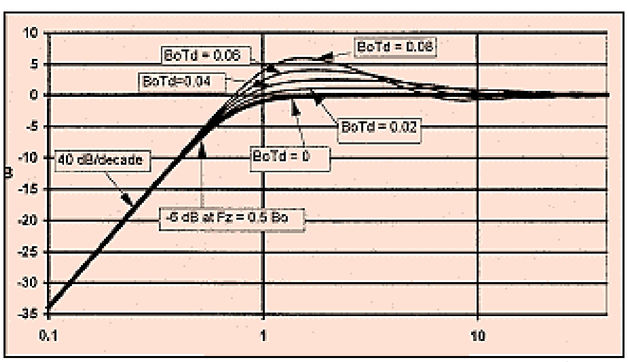


*Figure 6. PLL Supression Function |W(f)| - Type II 2nd Order $\zeta$ = 0.7 with Delay: BoTd = 0, 0.02, 0.04, 0.06 & 0.08*

Higher symbol rate demodulators have less time delay since the symbol interval is shorter and the delay is a multiple of the symbol interval. Therefore they permit higher PLL bandwidths, making high symbol rate digital radios usually more tolerant of close in phase noise than low symbol rate radios. Typical carrier recovery PLL bandwidths for symbol rates from 2 to 35 Msym/sec QAM modems are 10 to 100 kHz.

This carrier recovery method is known as decision directed. Other PLL schemes for coherent carrier recovery such as times-n and Costas loop are useful for certain QAM constellations. In general they have a high pass response to phase variations similar to the decision directed PLL. In some applications, noncoherent demodulation is used in star type QAM constellations that employ differential modulation for simplicity or to suppress dispersion effects for low symbol rates. Since the change in phase symbol-to-symbol is the detected parameter, the weighting function has the form of sin ($\pi f/F_S$). This function has high suppression at low offsets and has a narrower bandpass than the weighting function for carrier recovery PLL demodulation.

**Carrier Recovery PLL PM Suppression Measurements on a QAM System**

In experimental tests performed at Communications Techniques, Inc., the suppression function was measured at 64 QAM and 256 QAM on an RF loopback system operating at 30 and 40 Mbits/sec respectively (see Fig.7). The setup used FEC/encoding per ITU-T (J.83), adaptive equalization and a standard pseudo-random bit stream per CCITT Rec O.151. Sinusoidal phase modulation was injected on one of the local oscillators and its level increased until a BER threshold was reached. The suppression function was obtained by repeating this for modulating frequencies from 1 to 500 kHz. The resulting suppression functions for 64 and 256 QAM at ~5 Msym/sec are plotted in Figure 8 for different carrier recovery PLL parameter settings. The slope of the rejection curve is nearly 40 dB/decade and increased peaking is seen at higher PLL bandwidths, consistent with the model above.

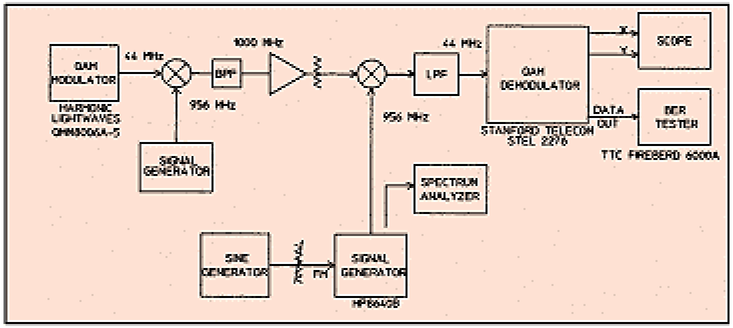

*Figure 7. QAM Setup for PM Supression Test*

Figure 9 vividly demonstrates the phase noise rejection properties of the carrier recovery PLL. Even with a 2000 degree peak phase deviation at 1 kHz, the 256 QAM system operates with $<10^{-10}$ BER. The carrier recovery PLL suppresses the 2000 degree deviation down to a level that eliminates crossings of the decision boundaries. The spectrum analyzer shows the gross spreading due to the large amount of phase deviation while the constellation displays a stable error free pattern.

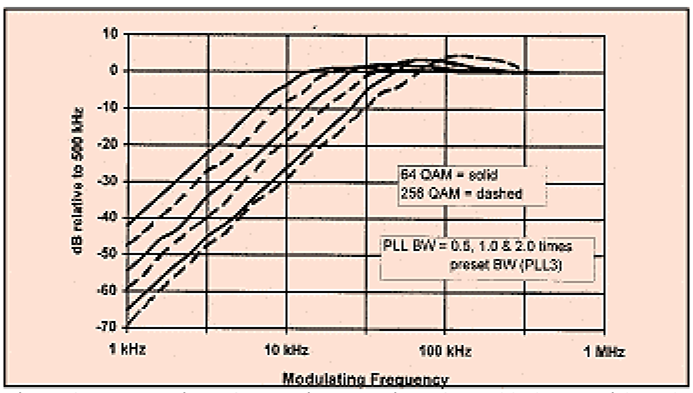

*Figure 8. Measured Pm Supression Function - STEL 2276 Demod 64 & 256 QAM at 5 MSym/Sec with Different PLL Bandwidths*

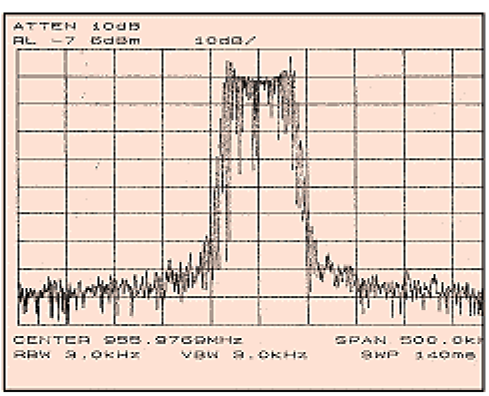

*Figure 9a.*

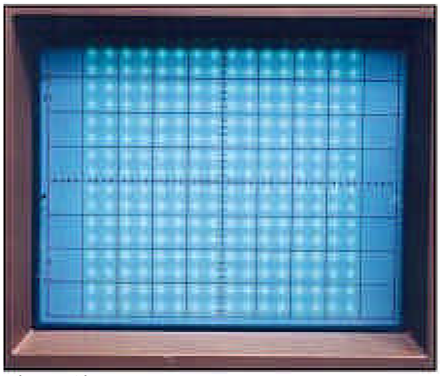

*Figure 9b.*

## Integrated Phase Noise

The SSB phase noise of a phase locked DRO (PDRO) at 12.3 GHz is shown in Figure 10 along with weighting due to a typical W(f) for a QAM demodulator with $B_0$ = 30 kHz, $\zeta$= 0.7, Td=0 and symbol rate = 10 Msym/sec. The weighted phase noise shows the effect of the carrier recovery PLL and the Nyquist $F_S/2$ filter. Integration of this curve on both sides of the carrier (called double sideband or DSB) results in an rms phase noise of 0.0022 radian. Defining $(S/N)_p = 20 \times \log_{10} (1/\sigma_\theta)$ per Kucar and Feher then $(S/N)_p$ = 53.1 dB. This is plenty of margin above $(S/N)_6$ for 16, 64 and 256 QAM.

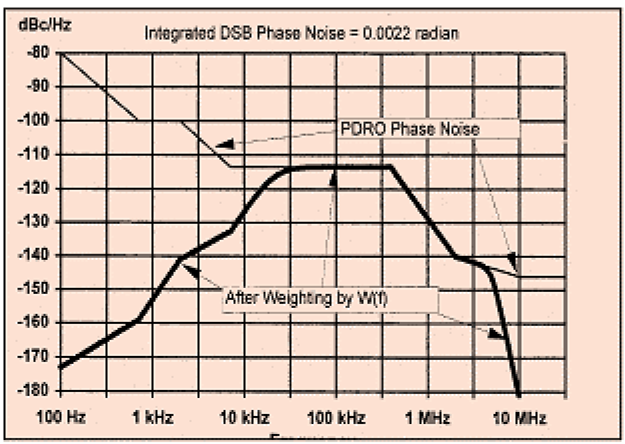


*Figure 10. PDRO Phase Noise at 12.3 GHz, Before and After Demod. Weighting*

Figure 11 shows the phase noise of a CTI PTPS synthesizer at 10 GHz with 875 kHz steps. It also shows the results of weighting with parameters $B_0$ = 25, 50 and 100 kHz, $\zeta$= 0.7, Td = 0 and $F_S$ =10 Msym/sec. The advantage of a high carrier recovery PLL bandwidth is clearly substantial when the phase noise curve is rolling off. The DSB integrated phase noise is declining at about 6-7 dB per octave as $B_0$ increases from 25 to 100 kHz. The margin of $(S/N)_p$ over $(S/N)_6$ for 64 QAM is 12.4, 19.7 and 25.5 dB for $B_0$ = 25, 50 and 100 kHz respectively. From Figure 4, $B_0$ = 50 and 100 kHz would provide low degradation but with a 25 kHz $B_0$ the degradation may be marginal depending on other noise sources. For 16 QAM the 25 kHz $B_0$ should be acceptable. For 256 QAM the $(S/N)_p$ to $(S/N)_6$ margins are 6.4, 13.7 and 19.5 dB. Therefore a 100 kHz $B_0$ is satisfactory and a 50 kHz $B_0$ may be marginal.

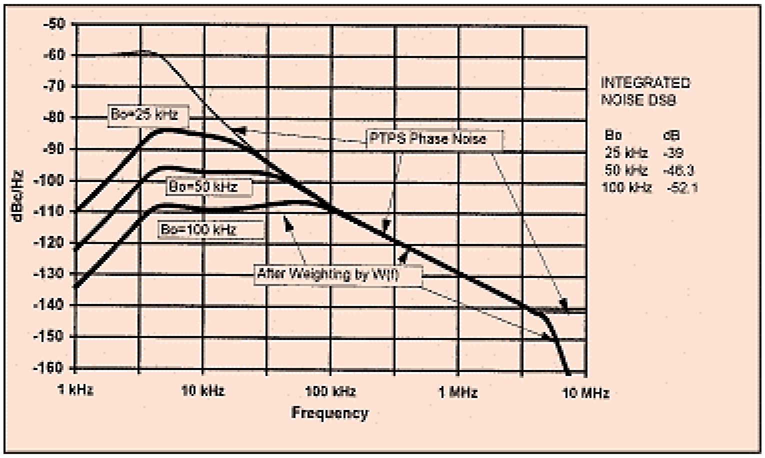


*Figure 11. PTPS Phase Noise at 10 GHz - Before and After Demod. Weighting*

**Integrating Bode Segments**

One way of simplifying the integration is to approximate the LO phase noise and W(f) using Bode segments. Then the series of segments can be integrated using the integrals shown in Table 2:

**Table 2.**

| | Segment Formula | Integral over the segment | |
|---|---|---|---|
| A: | $K1 * \frac{F1^n}{f^n}$ | $\frac{K1*F1}{n-1} * \left(1 - \left(\frac{F1}{F2}\right)^{n-1}\right)$ | n>0 and n#1 |
| B: | $K1 * \frac{F1}{f}$ | $K1 * F1 * \log_{10}\left(\frac{F2}{F1}\right) * 2.3$ | |
| C: | $K1$ | $K1 * (F2 - F1)$ | |
| D: | $K2 * \frac{f^n}{F2^{-n}}$ | $\frac{K2*F2}{n+1} * \left(1 - \left(\frac{F1}{F2}\right)^{n+1}\right)$ | n>0 |

where F1 and F2 are the left and right endpoints of the Bode segment
K1 is $\measuredangle(F1)$
K2 is $\measuredangle(F2)$

An example of the integration of a SSB Bode phase noise plot is as follows:

| Bode Segment (dBc/Hz) | | Calculation | Result |
|---|---|---|---|
| -120 at 1 kHz to -110 at 10 kHz | D | $10^{(-110+40-3)/10} * 0.99$ | $0.05 \times 10^{-6}$ |
| -110 at 10 kHz to -120 at 100 kHz | B | $10^{(-110+40)/10} * 1 * 2.3$ | $0.23 \times 10^{-6}$ |
| -120 at 100 kHz to -120 at 1 MHz | C | $10^{(-120+59.5)/10}$ | $0.89 \times 10^{-6}$ |
| -120 at 1 MHz to -140 at 10 MHz | A | $10^{(-120+60)/10} * 0.9$ | $0.9 \times 10^{-6}$ |
| -140 at 10 MHz to -140 at 20 MHz | C | $10^{(-140+70)/10}$ | $0.1 \times 10^{-6}$ |
| | | | $2.17 \times 10^{-6}$ |

The SSB integrated phase noise is $10 * \log_{10}(2.17 \times 10^{-6}) = -56.6$ dBc which is -53.6 dBc DSB.

**Characterizing Integrated LO Phase Noise Using the Start Frequency of Integration**

A useful simplification is to approximate W(f) with a brick wall filter from F1 to F2 = $F_S/2$ where F1 is chosen as approximately Fz = $0.5*B_0$ for $\zeta = 0.7$. This will result in some error depending on Td and the shape of $S_\theta$ (f). However, it facilitates a useful characterization of the phase noise of a local oscillator for carrier recovery PLL applications. Given a phase noise curve for the local oscillator, define IPN(F1) as follows:

$$\mathrm{IPN(F1)} = \int_{F1}^{F2} S_\theta(f)\,df = 2 \cdot \int_{F1}^{F2} \mathcal{L}(f)\,df$$

This is the DSB integration of phase noise from a variable start frequency F1 to a fixed stop frequency F2. The stop frequency is ideally $F_S/2$ although if the far out noise is low enough, the exact stop frequency isn't critical. Giving the integral as a function of start frequency allows modem designers to optimize the carrier recovery PLL bandwidth. For example, for the PDRO above a plot of the integral from F1 to F2 as a function of F1 where F2 is 1 and 25 MHz is shown in Figure 12. This curve rises very slowly as F1 decreases from 100 kHz and therefore the carrier recovery PLL bandwidth is not critical. Figure 13 gives IPN(F1) for the 10 GHz PTPS synthesizer with F2 = 25 MHz. It also shows in dashed lines the levels needed to obtain 10 dB margin over $(S/N)_6$ for 16, 64 and 256 QAM. The 10 dB margin is a reference for reasonable performance but usually some extra design margin is added. This graph gives a visual method of trading off design margin versus carrier recovery PLL bandwidth. If the dependence of other system residual noise on carrier PLL bandwidth is known, an optimum PLL bandwidth could be determined. Note that if the 10 GHz PTPS were multiplied by 4 for the 38 GHz radio band, the integrated noise curve would rise 12 dB ($=20*\log_{10}(4)$). Operation at 16 and 64 QAM would still be satisfactory with a practical carrier recovery PLL bandwidth.

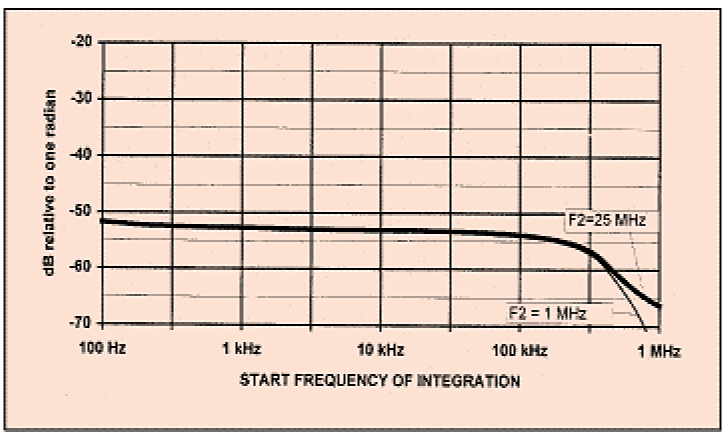


*Figure 12. 12.3 GHz PDRO Integrated DSB Phase Noise - Integrated from Start Frequency to 1 MHz and 25 MHz*

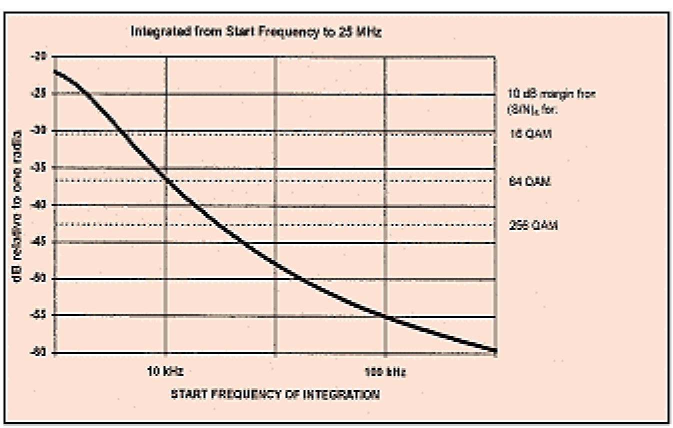


*Figure 13. 10 GHz PTPS Synthesizer Integrated DSB Phase Noise - Integrated from Start Frequency to 25 MHz*

## Microphonics

Microphonic local oscillators can cause major problems in high M-QAM systems. This is because the close spacing of the symbol points and the high bit rate tend to result in a large number of errors when a microphonic disturbance occurs. If synchronization is lost, it may take considerable time to resync and realign the adaptive filters during which significant data would be lost. Local Oscillators that have low microphonic susceptibility are a necessity. Phase-locked DROs have a proven track record in QAM radios and are known generally to be low in microphonics. Coaxial Resonator Oscillators (CROs) are also good design type because most of the electromagnetic fields are contained within a metalized solid block of ceramic. YIG based oscillators have a history of being microphonic and therefore may not be preferred for QAM applications. Crystal oscillators may need to be vibration isolated since they have a fractional frequency sensitivity on the order of $10^{-9}$/g. Wide PLL bandwidths increase the suppression of VCO frequency changes caused by mechanical vibrations. Careful mechanical design of the oscillator is needed to minimize the frequency response to vibration. A wide carrier recovery loop PLL bandwidth will also help in suppressing microphonic effects and should be considered when optimizing the demodulator.

## CTI Sources and Synthesizers for Digital Radio

Most QAM digital microwave radios employ phase locked DROs for up and down conversion because of their intrinsic low noise, reliability and simplicity. CTI produces a range of PDROs that are operating in digital radio and LMDS applications from 4 to 40 GHz. Figure 10 shows the phase noise of a typical 12.3 GHz unit with a 100 MHz reference. These units are ideal for fixed frequency QAM radios due to their low microphonics, compact size and cost effectiveness. They can be provided with or without an internal reference.

CTI's synthesizer solutions for QAM radios provide the ability to cover a band of frequencies with a single unit. This avoids the logistics problems associated with small quantities of a large number of frequencies. Also the frequency programmability allows inventorying units for spares and fast turnaround. These synthesizers are based on CTI's proven CRO and DRO capabilities and avoid the microphonic problems inherent in many YIG based oscillator designs. Two series of units are available:

The PTPS series synthesizers covers bands up to 6% wide between 1 and 20 GHz with step sizes from 10 kHz at 1 GHz and from 0.5 MHz at 20 GHz. Typical phase noise and integrated phase noise are shown in Figures 11 and 13 for a 9.5-10 GHz unit with 875 kHz steps. These synthesizer units work best with relatively high carrier recovery PLL bandwidths due to the rise in phase noise from 100 kHz to 10 kHz.

For lower noise QAM applications, CTI offers the DRS series. The DRS units are available with smaller steps than the PTPS and have low phase noise above 10 kHz. Typical phase noise and integrated phase noise are shown in Figures 14 and 15 for a 7-7.2 GHz unit with 125 kHz steps and for a 13-13.4 GHz unit with 500 kHz steps. Also shown are curves for frequency multiplication of the 13-13.4 GHz unit by 2 and 3 for the 28 GHz and 38 GHz bands. The stop frequency of integration is 25 MHz which corresponds to a symbol rate up to 50 Msym/sec. The DRS can work with fairly low carrier recovery PLL bandwidths.

Each series is available in a dual synthesizer package with independently programmable outputs to accommodate transmit/receive frequency pairs.

## Considerations in Specifying Integrated LO Phase Noise

Allocating a budget for integrated LO phase noise can approach a point of diminishing returns. From Figure 4 there is a ~1.0 dB degradation in fade margin due to one LO that is 7 dB better than $(S/N)_6$ (or two LOs that are each 10 dB better than $(S/N)_6$). Specifying an LO with 5 dB lower noise decreases the degradation from 1.0 to 0.28 dB. An additional 5 dB reduction in phase noise yields only a 0.19 dB decrease in degradation to 0.09 dB. These increments are not large when compared to total fade margins of 30-65 dB. Therefore, there is only marginal benefit in overspecifying LO phase noise since the difference may be too small to be noticeable.

In arriving at a more exact estimate of fade margin and its degradation, the improvement due to FEC as well as other sources of residual noise should be taken into account. Alternatively, the FEC improvement can be utilized as a design margin to the degree it is not reduced by other sources of implementation loss.

**Step-by-Step Procedure**

An approach for selecting and specifying a PDRO or a synthesizer for use in a M-QAM system based on a given minimum fade margin spec and a residual BER spec is as follows (assuming the BER threshold is $10^{-6}$):

1. From the BER curve for the QAM system determine $(S/N)_6$ = the S/N needed for a BER = $10^{-6}$. Note that the BER curve for a given system depends on M, the constellation, FEC and other factors.
2. From the the unfaded signal level and fade margin requirement determine $(C/N)_{fade}$ at maximum fade. Calculate the degradation allowed at maximum fade = D = the difference between $(C/N)_{fade}$ and $(S/N)_6$.
3. From Figure 4 determine $(S/N)_{resmin}$ = the minimum residual S/N to meet the fade margin requirement. From the BER curve check that $(S/N)_{resmin}$ will meet the residual BER specification.
4. Determine a specification for $(S/N)_{res\text{-}spec} > (S/N)_{resmin}$. The corresponding spec for the fade margin degradation Dspec can be found from Figure 4.
5. From $(S/N)_{res\text{-}spec} = 1 / [\, (N/S)_1 + (N/S)_2 + ....]$ allocate a budget for $(N/S)_i$ for each contributor to the total degradation including the phase noise of the LOs and allowance for margin.
6. From the DSB integrated phase noise vs. start frequency of a prospect LO determine the minimum Fz of the demodulator carrier recovery PLL needed to meet the budgeted (N/S) for that LO. Take into account the 20 $\log_{10}(N)$ noise increase if the LO is to multiplied up.
7. Optimize Fz as required to minimize N/S for the LO phase noise, microphonics and N/S from other sources, taking into account peaking that may occur due to delay.

For example:

1. Assuming theoretical 64 QAM and then $(S/N)_6$ = 26.6 dB;
2. Assume a spec fade of 50 dB results in a $(C/N)_{fade}$ of 29.5 dB. D=29.5-26.6 = 2.9 dB;
3. From Figure 4 $(S/N)_{resmin}$ = 26.6+3.1 = 29.7 dB;
4. Set $(S/N)_{res\text{-}spec}$ = 30 dB; then Y = 3.4 dB and $D_{spec}$ = 2.7 dB;
5. Assume the budget for local oscillator 1 is $(N/S)_{LO1}$ = -30-10 = -40 dB;
6. Assuming a 38 GHz radio and that a DRS Synthesizer at 13 GHz with a X3 multiplier is to be used, then Fz min ~8 kHz;
7. After evaluating the trade-offs of decreasing phase noise and microphonics vs. negatives of increasing PLL bandwidth, set the carrier recovery PLL Fz to 20 kHz.

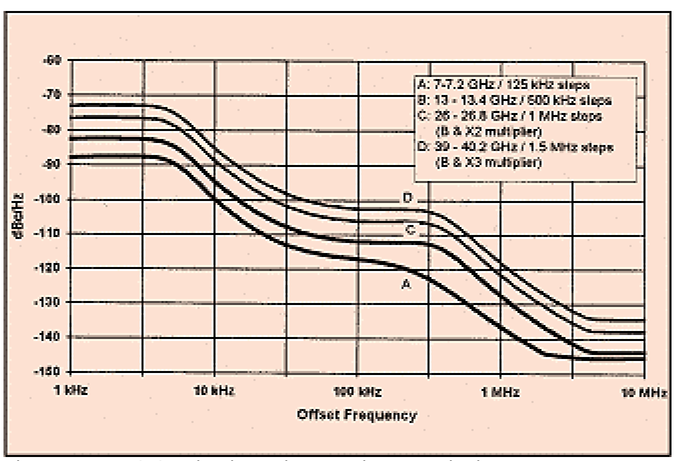

*Figure 14. DRS Synthesizer Phase Noise – Typical*

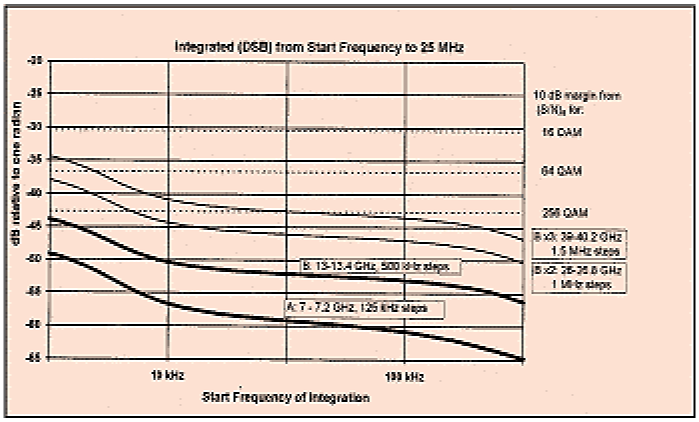

*Figure 15. DRS Synthesizer Typical Integrated Phase Noise - Integrated (DSB) from Start Frequency to 25 NHz*

## Summary

The M-QAM systems used in digital microwave radio to achieve high data rates require high S/N due to the close spacing of states in the I-Q symbol constellation. Low rms phase noise integrated over the effective passband of the demodulator is needed from the local oscillators. Tests on 64 and 256 QAM systems demonstrate that high suppression to close-in phase noise is provided by a carrier recovery PLL. The carrier recovery PLL bandwidth is key in determining the effect of local oscillator phase noise on BER. Characterization of a local oscillator's phase noise in terms of the integral from a variable start frequency to one half the symbol rate is useful in determining the carrier recovery PLL bandwidth needed to achieve rms phase noise objectives. When the S/N due to integrated phase noise is 10 dB better than the S/N needed for BER=$10^{-6}$ there is about a 0.5 dB degradation in fade margin relative to a noiseless oscillator and reductions in phase noise have a diminishing effect on system performance. Practical results show that PDRO sources and synthesizers of the type produced by CTI meet or exceed the requirements of M-QAM operation in modern digital radio systems.

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