ANALYSIS AND PREDICTION OF PHASE NOISE IN RESONATORS AND OSCILLATORS

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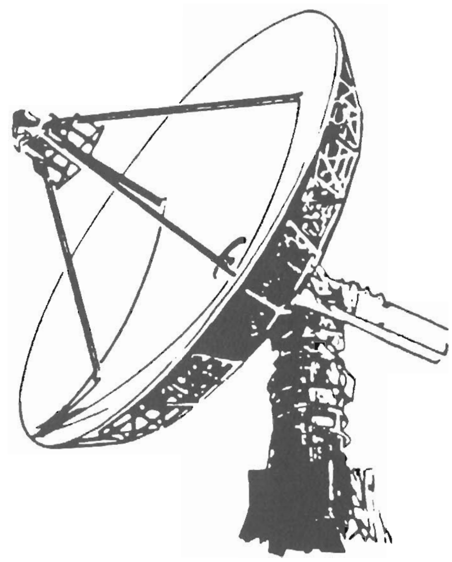

HEWLETT PACKARD

ANALYSIS AND PREDICTION OF PHASE NOISE IN OSCILLATORS AND RESONATORS

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- NEGATIVE RESISTANCE MODEL:
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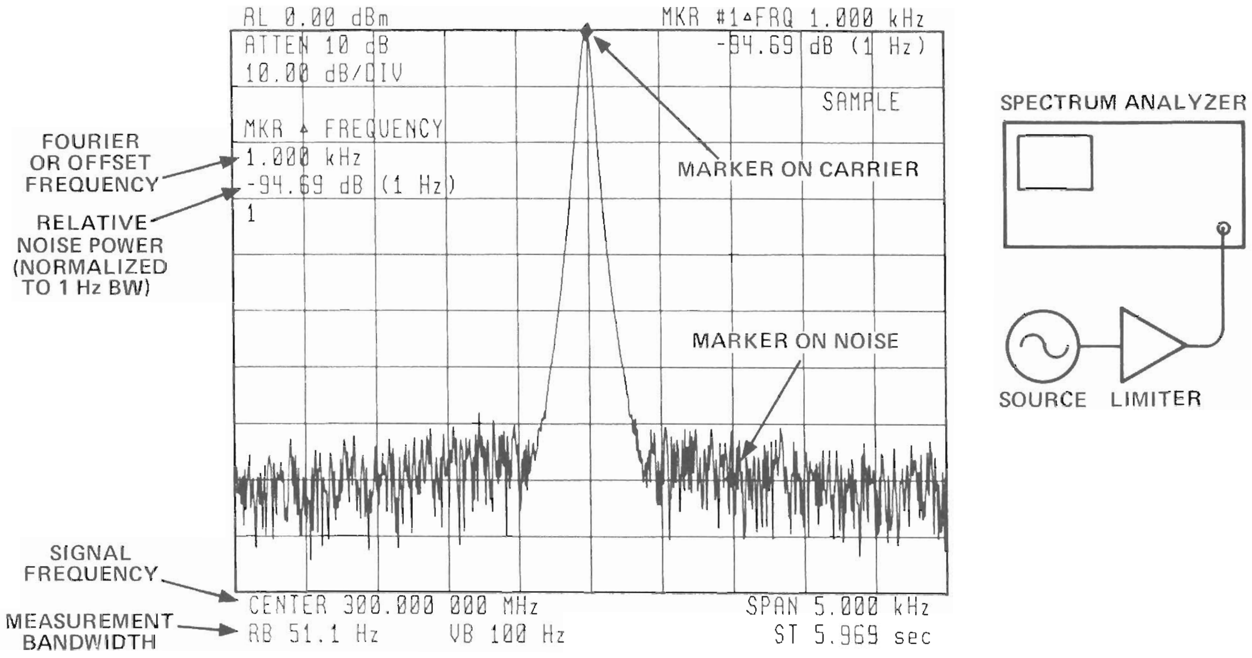
PHASE NOISE — WHAT IS IT?

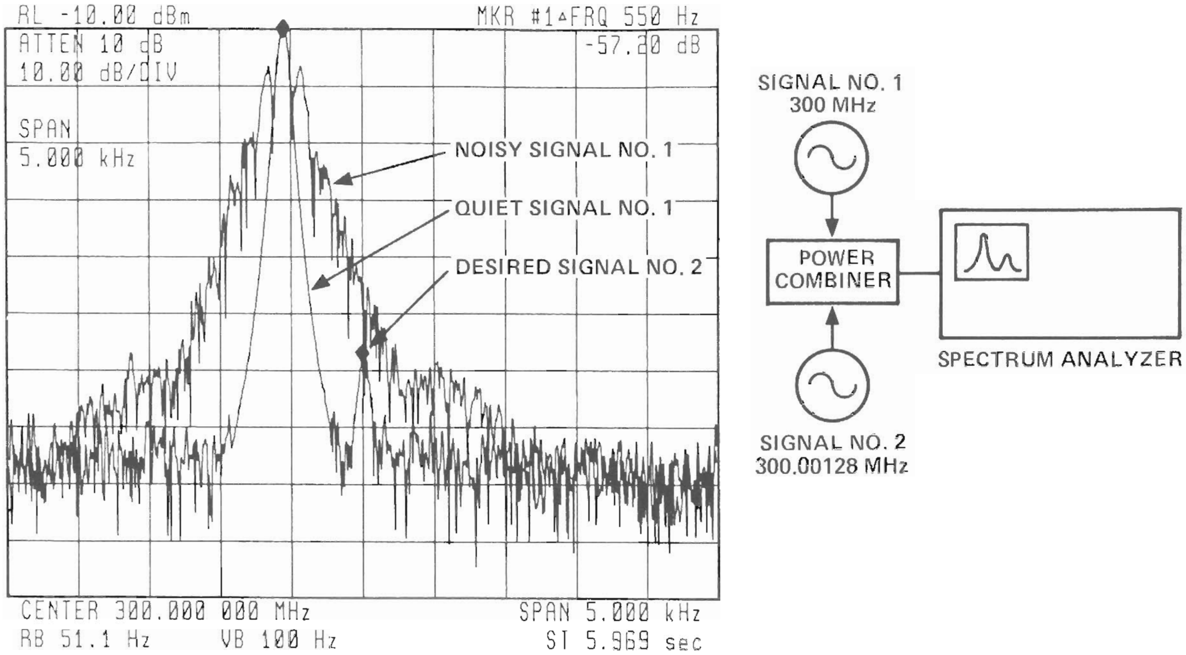


Phase noise is a measurement of the uncertainty in the phase of a signal. It is measured as the ratio of noise power in quadrature (90° phase) with the carrier signal to the power of the carrier signal. This is opposed to AM noise which is noise in phase with the carrier signal. Either noise is measured at a given offset frequency from the carrier, normalized to a 1 Hz bandwidth. Two measurements of phase noise are common: $S\Delta\phi(f)$, the spectral density of phase fluctuations; and $\mathcal{L}(f)$, the single sideband phase noise. $S\Delta\phi(f)$ is twice $\mathcal{L}(f)$ since $S\Delta\phi(f)$ is related to total phase change (both sidebands) and $\mathcal{L}(f)$ is a single sideband measurement as would be seen on a spectrum analyzer.

$\mathcal{L}(f)$ can be measured on a spectrum analyzer by comparing the noise to the carrier. First the carrier level is measured and marker set to peak. Then marker delta is selected and marker noise selected. Another trace is taken, this time in sample mode and the marker set to the offset frequency desired. This method will automatically take care of the corrections necessary for accurate noise measurement. These include effects of log amplifier upon noise, noise (impulse) bandwidth vs gaussian bandwidth correction, peak detection vs rms noise and scaling of measurement bandwidth to 1 Hz normalization bandwidth. These are covered in HP Application note 150-4.

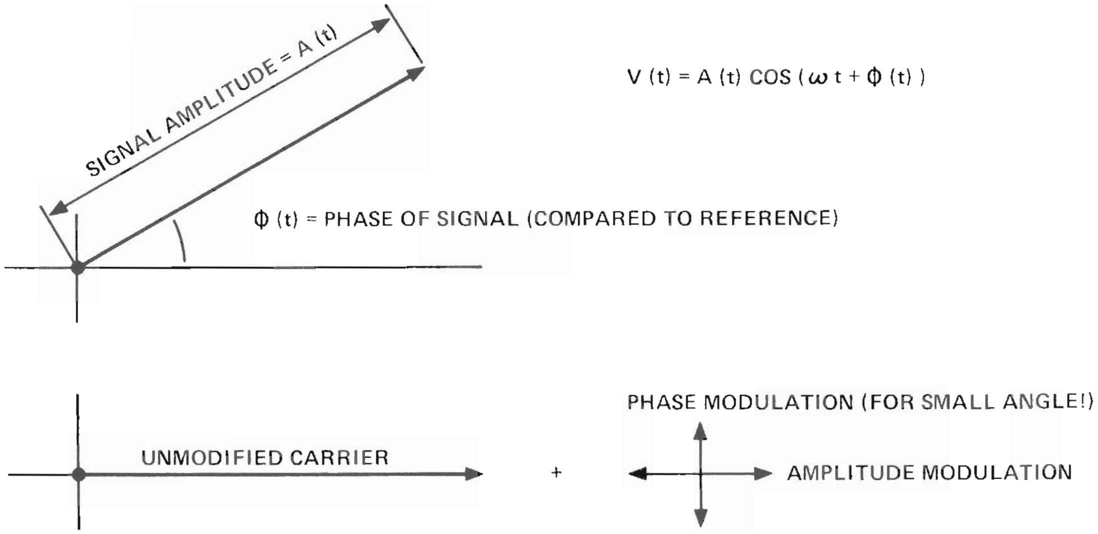
The phase noise of a signal can only be measured by a system that has equal or better noise performance. Any phase noise of a receiver will be convolved with the input signal and smear the phase noise out to a level at least as high as the local oscillator's noise. Therefore most very quiet low frequency sources can't be measured on a spectrum analyzer where an equivalently quiet reference is multiplied up by a factor of up to 300 times in generating the local oscillator signal, so methods are used involving direct mixing with two identical sources or a known very good source as reference.

WHY IS GOOD PHASE NOISE DESIRABLE?



Why be concerned about phase noise? There are as many reasons as there are oscillators. The basic problem shows up as a noisy signal masking a desired signal. The noisy signal may be a strong signal adjacent to a very weak desired signal, or it might be the local oscillator of a receiver that smears the strong adjacent signal over the weak signal. Phase noise limits the dynamic range available in any system. This may show up as higher error rate in a disc memory, lack of definition in an ultrasound imaging system, loss of radar sensitivity at low doppler shifts or even the fact that you can't hear a distant radio station because a strong local station has been mixed on top of it due to a poor local oscillator in your radio.

PHASOR REPRESENTATION OF SIGNAL



To understand phase noise it is convenient to use phasor diagrams. The phasor diagram can be looked at as a snapshot in time of the amplitude and phase of a signal compared to a perfect reference of the same frequency as the nominal signal frequency. It is analogous to the output of a mixer if the signal frequency were mixed down to dc, assuming both in phase and quadrature outputs are available. Another analogy often used is a spinning vector (the signal) and strobe light triggered by the perfect reference.

As the lower diagram shows, a small signal added to the unmodulated carrier can cause either AM, PM or both modulation types. This is **only** valid for small disturbing signals, since large angles require J Bessel functions to specify the sideband and harmonic levels. For small x, $J_o(x)$ is approximated by 1, $J_1(x) = x/2$, and all other $J_n(x)$ Bessel functions are approximately zero.

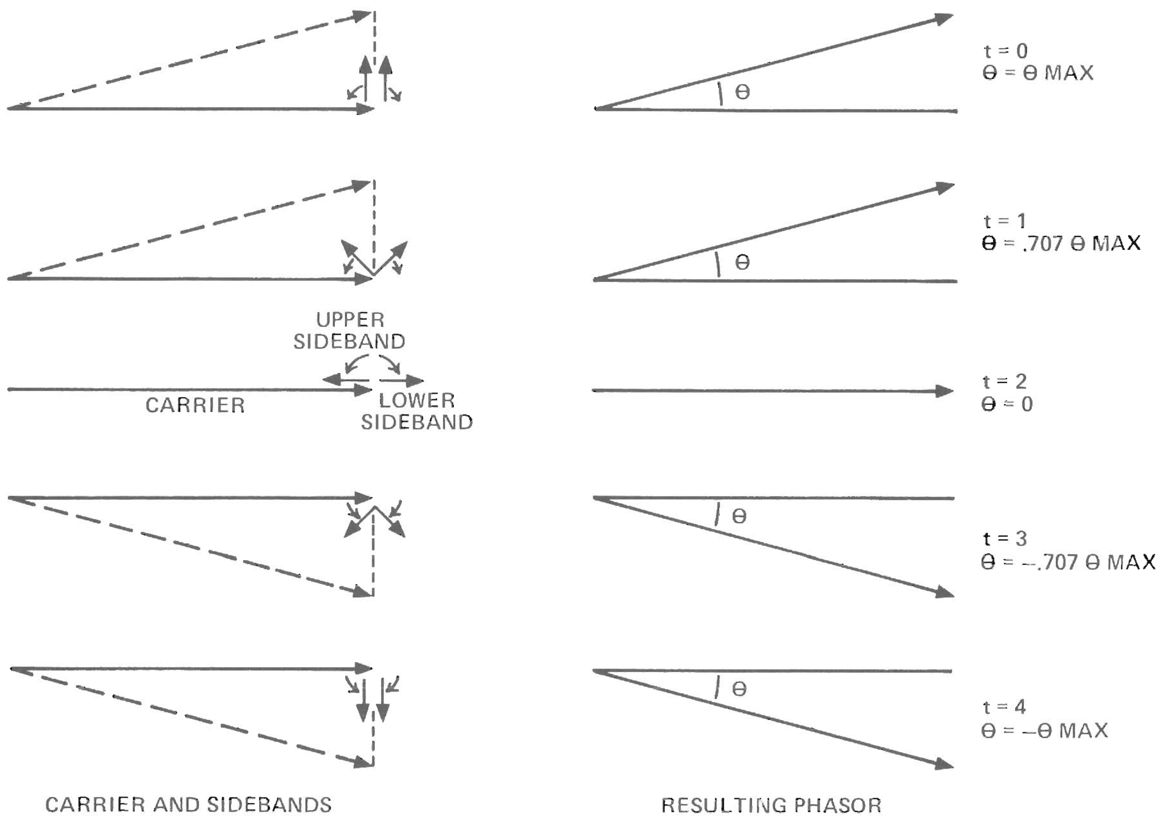
$$
\begin{aligned}
V(t) &= \mathrm{COS}\,(W_o t + x\,\mathrm{SIN}\,W_m t) \\
&= J_o(x)\,\mathrm{COS}\,W_o t - J_1(x)\,[\mathrm{COS}\,(W_o - W_m)t - \mathrm{COS}\,(W_o + W_m)t] \\
&\quad + J_2(x)\,[\mathrm{COS}\,(W_o - 2W_m)t + \mathrm{COS}\,(W_o + 2W_m)t] \\
&\quad - J_3(x)\,[\mathrm{COS}\,(W_o - 3W_m)t - \mathrm{COS}\,(W_o - 3W_m)t] \\
&\quad + J_4(x)\ldots
\end{aligned}
$$

for $x \ll 1$

$$J_o(x) = 1 - (x/2)^n$$

$$J_n(x) = (1/n!)\,(x/2)^n \quad \text{for } n > 0$$

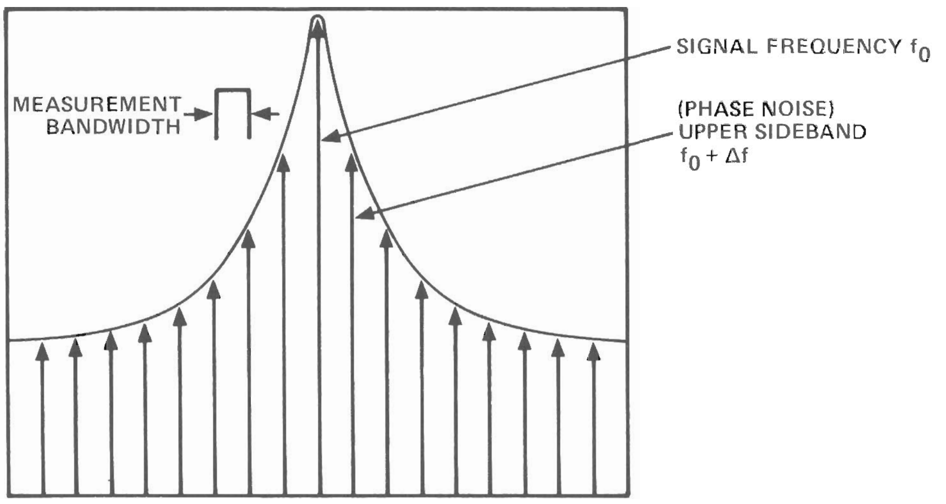
PHASOR DIAGRAM OF PHASE MODULATION



As can be seen, pure phase modulation results when both upper and lower sidebands are equal in amplitude and phased so as to add when in quadrature with the carrier signal and to cancel when in phase. This is true only for narrow angles because the carrier length must be maintained constant or AM will result. This holds well for phase noise lower than -40 dBc to -60 dBc in the bandwidth of measurement. Above these levels more energy appears in the harmonics of the offset frequency. It is important to remember that phase and amplitude modulation are measured in a circular coordinate system, while the addition of sidebands is done in a rectangular coordinate system. Approximations break down at high modulation levels.

$\mathcal{L}$ (f) corresponds to the relative level on one sideband while $S\Delta\phi(f)$ corresponds to the total phase angle change which includes both sidebands.

PHASE NOISE APPROXIMATED BY DISCRETE SIGNALS

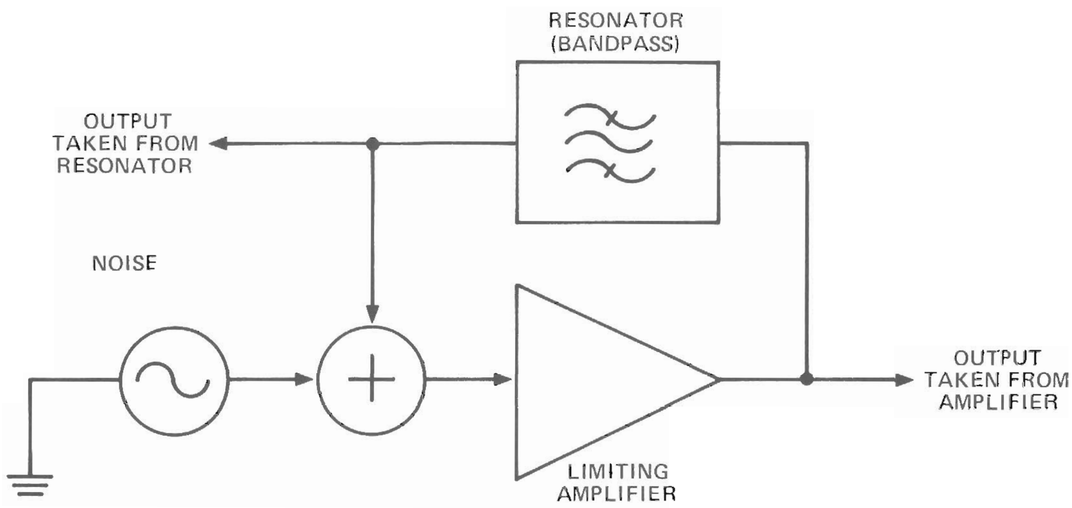


For purposes of analysis we replace the average noise power in each measurement bandwidth with a discrete constant power signal. Noise is **not** predictable as to its exact value, but we can estimate a time average with good confidence. If the bandwidth of measurement included many such discrete signals equally spaced in frequency, the signals would be indistinguishable from a constant noise power density. The substitution of discrete signals allows separate analysis of each frequency.

Real signals do not have a single frequency carrier. The concept of a carrier signal with noise sidebands is an artifact of the measurement bandwidths conveniently used. If a carrier is examined with a small enough bandwidth, the signal can be resolved into a power spectral density in a given operating bandwidth. The carrier is exactly that power derived from multiplication of the power spectral density multiplied by the impluse bandwidth. For a single pole bandpass structure, the impulse bandwidth is $\pi/2$ times the 3 dB bandwidth.

Most oscillators and signals have operating bandwidths much smaller than the .1 Hz to 10 Hz minimum bandwidths available with typical instrumentation. The carrier concept makes very good sense in this case since phase noise will be low enough (typically) such that narrow band FM approximations hold.

LEESON'S MODEL OF OSCILLATOR PHASE NOISE



OSCILLATOR BLOCK DIAGRAM

A standard model for phase noise in oscillators is that of D. B. Leeson, based on feedback theory. Several basic assumptions are made:

1. The amplifier has high gain and limits at a level corresponding to the output power level.

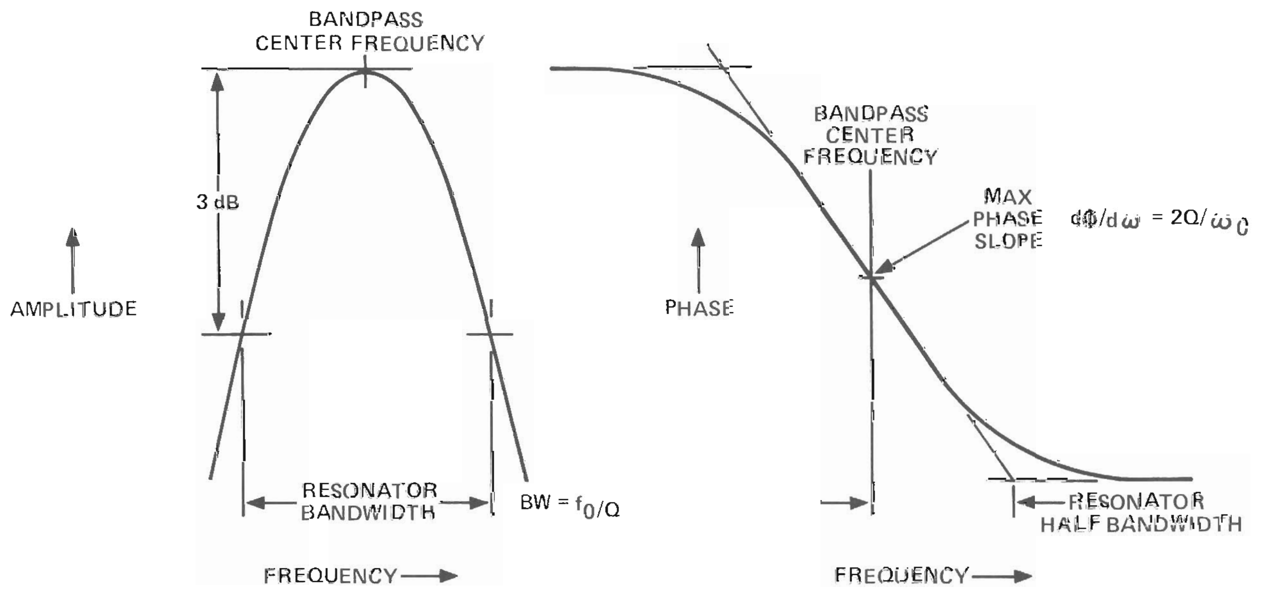
2. The resonator is a bandpass type structure with center at the frequency of oscillation.

3. The noise source corresponds to both the noise figure of the amplifier, and any other additional noise sources.

4. The limiting action of the amplifier removes the AM component of the noise.

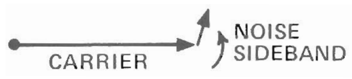
5. The circuit oscillates at zero dB loop gain and zero degrees phase (or multiple of 360°) around the loop.

RESONATOR CHARACTERISTICS

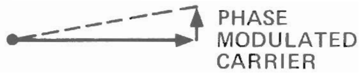


The resonator amplitude versus frequency characteristics follow the single pole model, with 3 dB bandwidth related to Q as above. The important characteristic for phase noise considerations is that of phase shift versus frequency. Inside the resonator bandwidth the phase shift with frequency approaches a constant with maximum slope at the center frequency. Outside that bandwidth, no feedback signal is available through the filter.

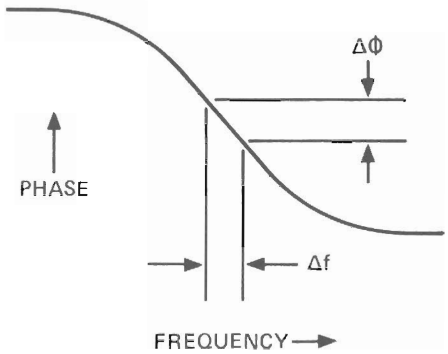
GENERATION OF PHASE NOISE



SIGNAL AT
AMPLIFIER INPUT



SIGNAL AT
RESONATOR INPUT



PHASE INDUCED FREQUENCY CHANGE

Consider what happens when the output of the resonator is added to a noise source at a specific offset frequency from the carrier. Both AM and PM components are present in equal powers (considering a single sideband at a time). Once the signal passes through the limiter, assuming no AM to PM conversion, the AM is removed leaving only phase modulation and reducing sideband level by 3 dB.

Phase modulation at a given offset means that the total phase around the loop oscillates back and forth at a rate equal to the offset frequency. The requirement that the oscillator operates with zero degrees phase around the loop causes the oscillator frequency to shift so as to counteract the phase change due to noise. This action converts noise in phase to noise in frequency.

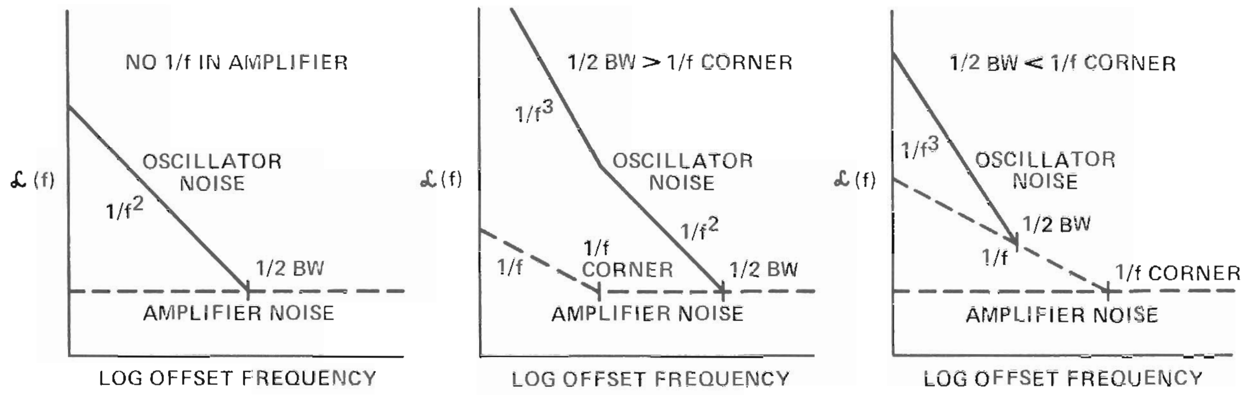
However, if the offset frequency is greater than one half the bandwidth, the bandpass character of the resonator removes the phase modulation before the oscillator responds. Thus, the key offset frequency is the half bandwidth.

For small deviation frequencies (similar to the phase moduation approximation) the sideband amplitude is given by the narrowband FM approximation:

$$\frac{\text{Single Sideband Amplitude}}{\text{Carrier Amplitude}} = \frac{\Delta f}{2f_m} = \frac{\text{Deviation Frequency}}{\text{Twice The Modulation Frequency}}$$

Converting to power gives a $1/f^2$ relationship of sideband level versus offset frequency assuming constant phase modulation with frequency.

THREE CASES OF OSCILLATOR PHASE NOISE



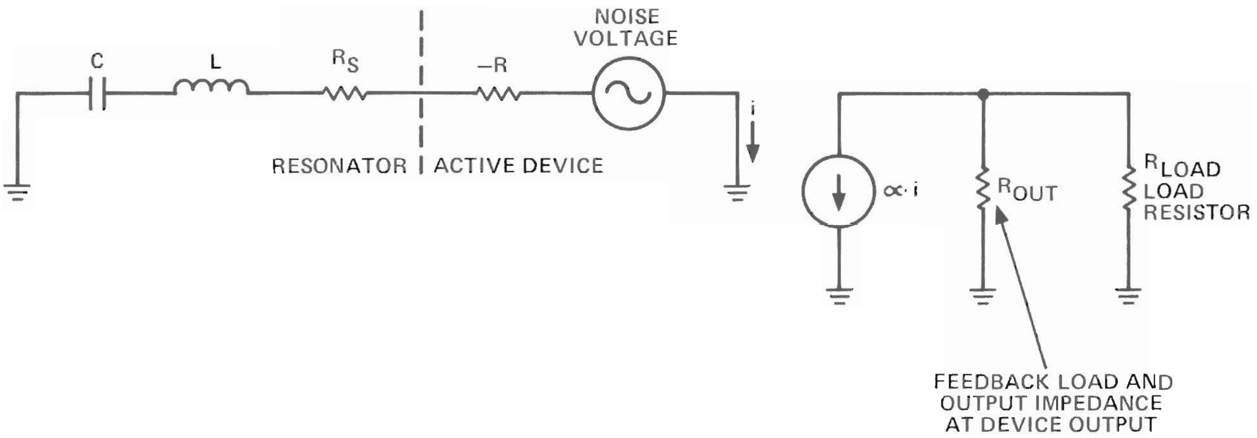
The phase noise of an oscillator depends upon the noise of the open-loop amplifier and upon the half bandwidth of the resonator. If the amplifier has no 1/f noise region, the oscillator will have $1/f^2$ noise below the half bandwidth. All active devices have some sort of 1/f region, it seems.

If the 1/f corner frequency is low, the oscillator will have $1/f^2$ noise slope until that corner frequency is reached. This is the case with many LC oscillators.

Crystal oscillators often have narrow bandwidths and one could have a lower bandwidth than the 1/f corner of a typical device, giving a region of 1/f noise and then $1/f^3$ noise as offset frequency decreases.

The 1/f region might be due to either amplifier or resonator. In many cases the noise of the resonator dominates; especially in the case of a crystal or SAW device.

NEGATIVE RESISTANCE OSCILLATOR SERIES CASE



W.P. Robbins, in his book **"Phase Noise in Signal Sources"**, talked of another way to look at oscillators: "the output of an oscillator is just amplified noise in a very narrow band." Negative resistance provides the power necessary to amplify the input noise up to the output power level.

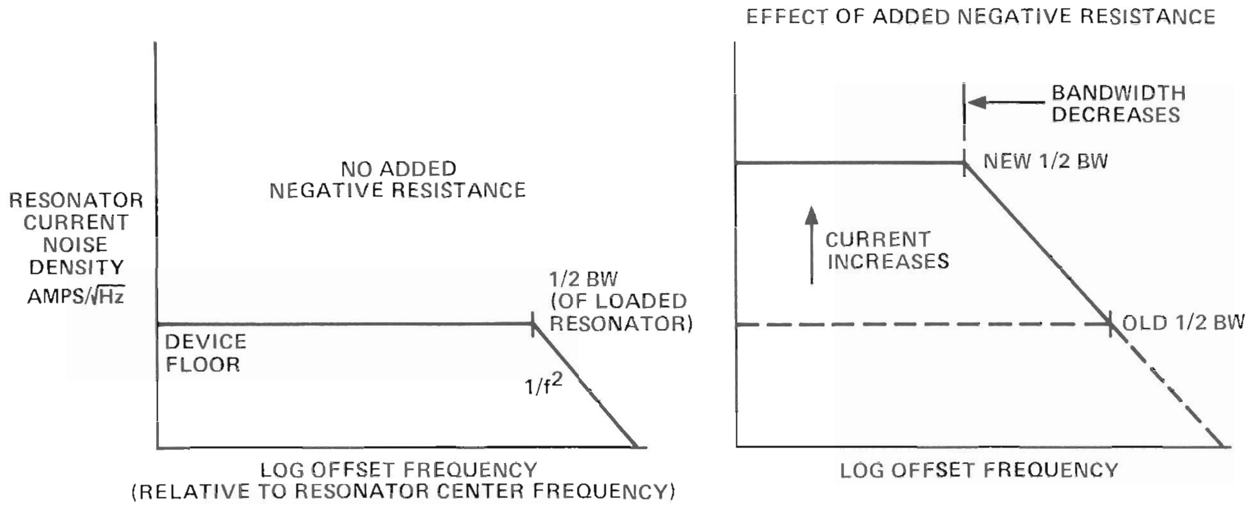
All oscillators may be modeled using a negative resistance or conductance model, with a few changes in perspective. The resonator is separated out from the sustaining stage. That active device, with associated feedback, provides the negative resistance necessary to cause oscillation. Conventional wisdom has the loss completely canceled by the negative resistance, allowing the signal to continue at a constant power level.

Consider the series oscillator circuit. The resonator has a loss associated with it of magnitude Rs. The active device input impedance (as an amplifier) might also be lumped into this loss as well. Assume for the moment that no negative resistance is present. Under these conditions, the resonator loaded Q may not be much lower than under unloaded conditions. A noise voltage density appears in series with the resonator. That voltage and the resonator series resistance have power determined by the device noise figure and thermal noise in the resistance.

$$-174 \text{ dBm/Hz} + \text{Noise Figure (dB)} = E_n^2/R_s = \text{Power Spectral Density (in the Resonator)}$$

A portion of the resonator current flows through the output load and another part is fed back to generate the negative resistance.

NOISE AND EFFECT OF ADDING NEGATIVE RESISTANCE



Assume we still have not added any negative resistance. The noise voltage density is constant with frequency, for constant noise figure. The noise current density will also be constant inside the resonator bandwidth. Part of this current density will appear across the output load. The total output power is related to the square of the current density multiplied by the bandwidth. If we assume half the resonator current flows through the output load, (matched case) the output power will be one fourth the resonator power.

Next consider what happens as we increase the magnitude of the negative resistance. As the series resistance is canceled by the negative resistance, the current density increases, since the noise voltage density is constant. The resonator bandwidth also changes. As the total resistance drops, the bandwidth goes down at the same rate that the current increases.

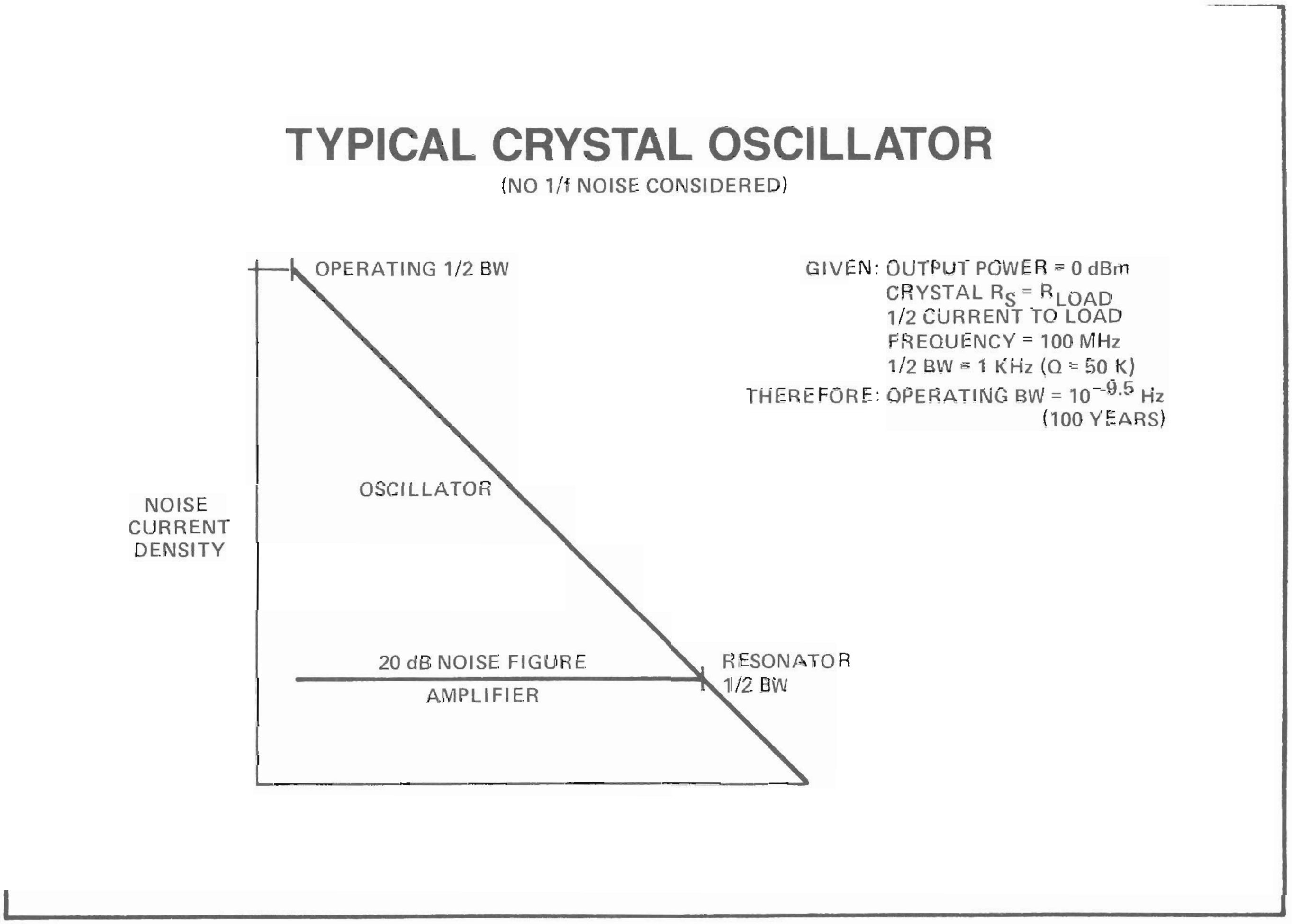
$I = E_n/R_{total}$ $\quad$ $Q = L/R_{total}$ $\quad$ $BW = f_o/Q$

Since the output power is related to the square of the current but only directly to the bandwidth, the output power goes up directly as the resistance drops. Again, limiting action removes the AM noise by constantly changing the negative resistance. This results in a loss of one half of the input power.

$$P_{out} = (1/8)\ I_n^2\ R_{load}\ BW$$

$$P_{out} = (1/8)\ (E_n/R_{total})^2\ R_{load}\ (f_o\ R_{total}/wL)$$

$$P_{out} = (1/8)\ FkTBW\ (R_{load}/R_{total})$$



If the resonator resistance were completely canceled, the power would rise to infinity. This is obviously not the case since the output power is finite and determined by oscillator limiting. The limiting action controls the amount of negative resistance applied to the resonator. As oscillations are building up, the bandwidth will not be much less than the resonator bandwidth. Only when the oscillator output power closely approaches the desired limiting value does the bandwidth get very small.

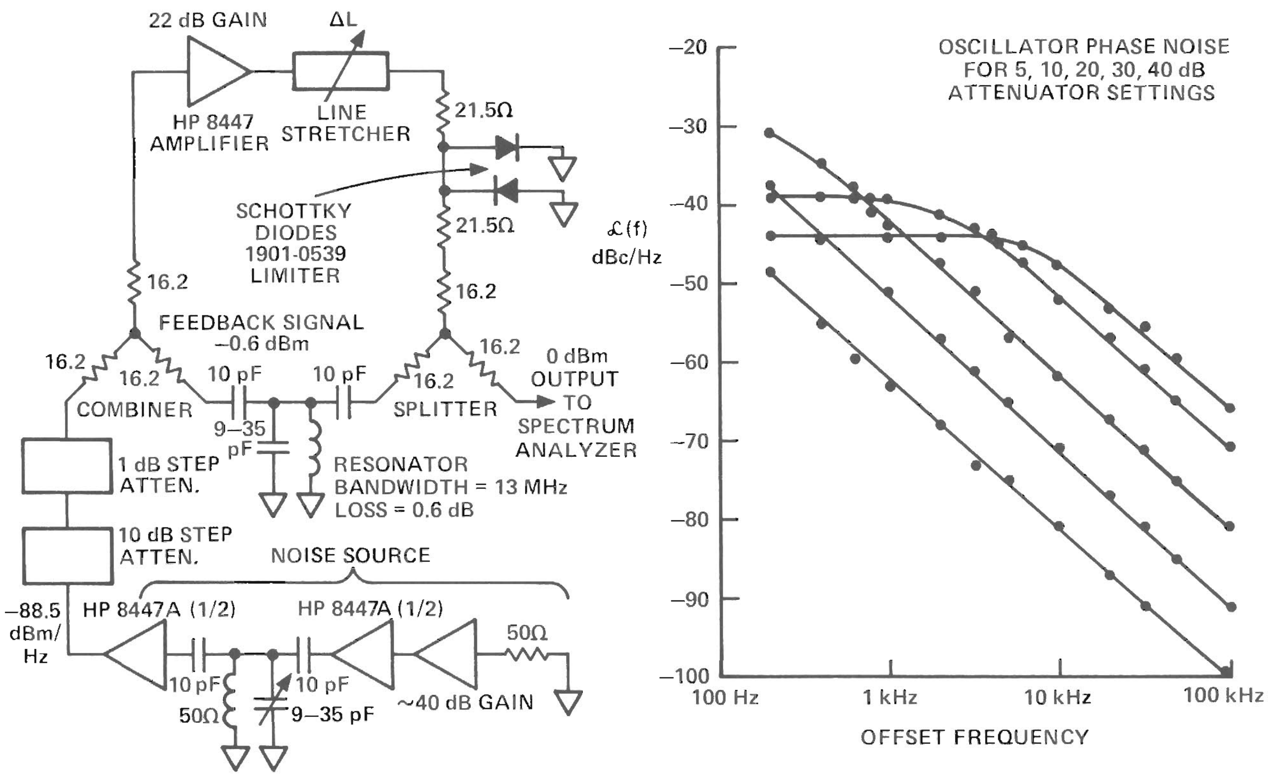
A typical crystal oscillator might have output power of zero dBm (1 mw). Assume the other parameters take on values as indicated above. A twenty dB noise figure for the transistor would imply -154 dBm/Hz noise power spectral density inside resonator bandwidth. The power at the output is reduced by six dB if current splits equally between output and feedback, and further reduced by three dB because we only consider phase noise after limiting occurs and AM noise is removed. A bandwidth of 2 kHz would have noise (impulse) bandwidth a factor of $\pi/2$ higher, giving total output power of -128 dBm.

The power must be increased by a factor of $10^{12.8}$ by canceling loss with the negative resistance. The operating bandwidth will be reduced at the same time to the incredibly small value of $10^{-9.5}$ Hz, or a time constant of 100 years.

The bandwidth might as well be infinitely small, except that it does allow some insight into what happens in the oscillator to analyze the circuit in this way. As the noise level fluctuates so does the bandwidth, providing a constant output power determined by the limiting mechanism.

$$-154 \text{ dBm/Hz} - 6\text{dB} - 3\text{dB} + 10 \log (2\text{kHz} \times 2\pi) = -128\text{dBm}$$

DEMONSTRATION OF OPERATING BANDWIDTH



A model indicating such incredibly small bandwidths is difficult to accept. Normal oscillators, even the noisiest, would have prohibitively low bandwidths to allow verification of the model. But by adding a noise source to an otherwise quiet oscillator, the concept of a very small operating bandwidth can be verified.

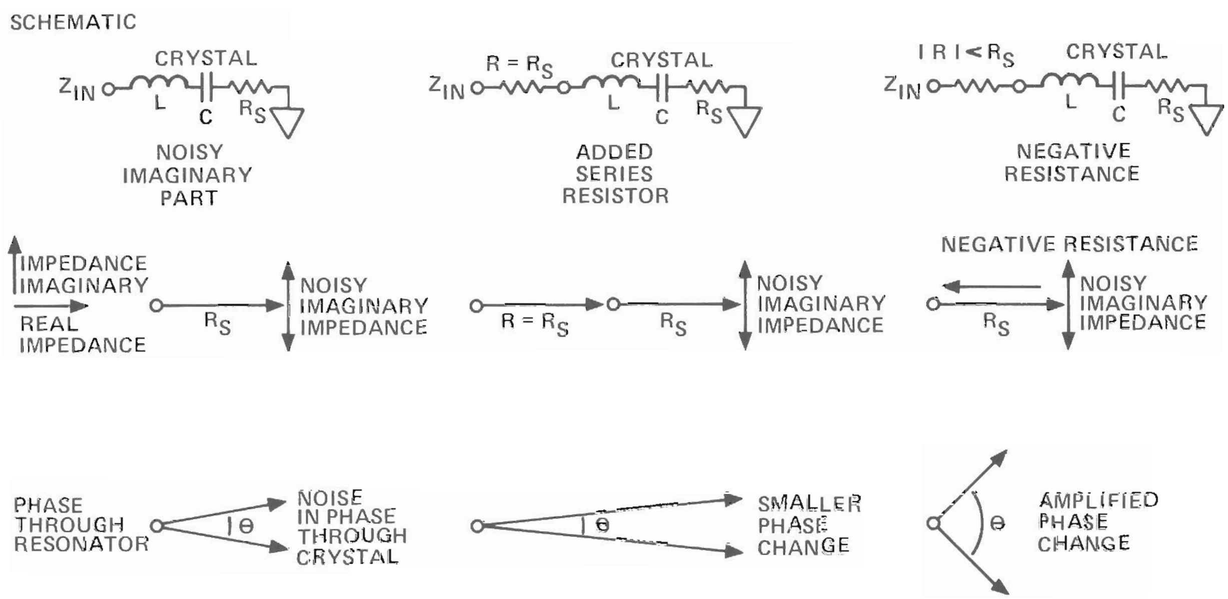
The oscillator consists of a power amplifier of 22 dB gain driving a limiter. The limiter output is split two ways providing output and feedback signals. The feedback signal passes through a resonator with 13 MHz bandwidth and 0.6 dB loss, where it is combined with the excess noise source. A line stretcher is used to set the oscillator phase to give oscillations at the resonator center frequency.

The phase noise was measured for attenuations of 5, 10, 20, 30 and 40 dB, and is plotted above. Calculation of the expected bandwidth gives close agreement. Actual measured bandwidths for 5 and 10 dB attenuation were 14.6 kHz and 4.2 kHz. Calculated bandwidths assumed the input noise was increased to −0.6 dBm by oscillator action, giving expected bandwidths of 17.1 kHz and 5.4 kHz. The difference between actual and predicted behavior corresponds to .7 and 1 dB noise, which is quite accurate.

-88.5 dBm/Hz $- 5$dB $+0.6$ dB -6 dBm -3 dB $+10 \log (13 \text{ MHz } \pi/2) = -28.8$ dB

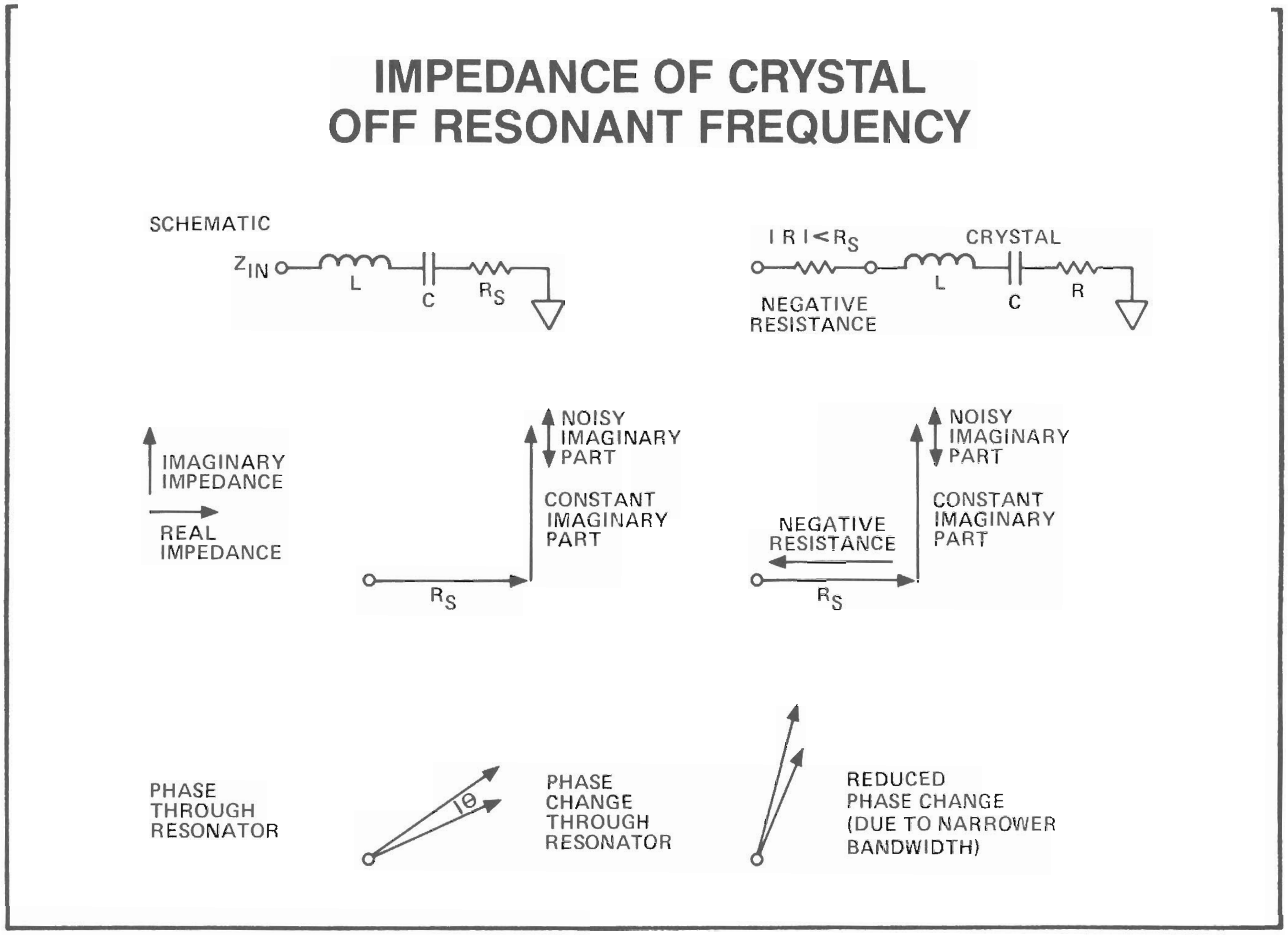
Operating Bandwidth = 13 MHz X 1.32 X 10^{-3} = 17.1 kHz

IMPEDANCE OF CRYSTAL AT RESONANT FREQUENCY



The very fact that we operate with a finite loss in the resonator helps to understand exactly what happens in the oscillator. If we consider a resonator with some internal phase noise (frequency uncertainty), there exists a real part of impedance and noisy imaginary part of impedance at the center of the passband. The noisy imaginary part of impedance will cause a phase shift in any signal passing through that resonator as shown above.

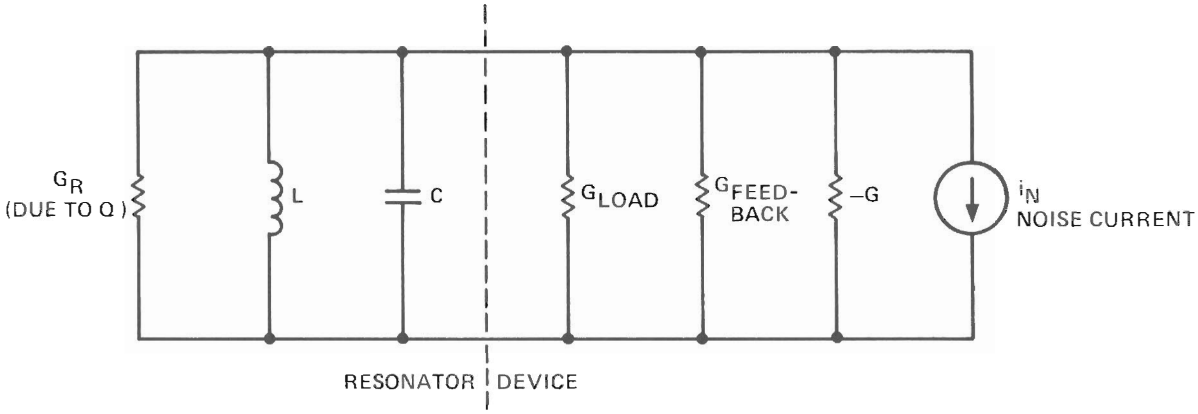
If we add a series resistor, the effect of the noisy imaginary part is reduced, and the phase noise is lower. However if we add a negative resistance in series, we change our reference point to a point much closer to the noisy imaginary part of impedance. The same noisy imaginary part has much wider phase deviation with negative resistance added. This demonstrates the effect of increasing noise with less cancellation by the negative resistance.



A similar action occurs for signals off resonance. For the case with no added negative resistance, the imaginary part of impedance has a constant offset and a second noisy part, comparable to that at resonance. When negative resistance is added, the angle change seen off resonance drops significantly, since both extremes are very close to 90^o phase. This shows the effect of narrowing of the bandwidth that occurs as resonator loss is canceled.

The generation of $1/f^2$ phase noise outside the operating bandwidth has already been shown. If a resonator or amplifier has additional phase noise as shown here, the oscillator frequency will be modulated with 1/f frequency noise. The 1/f character of frequency deviation will generate $1/f^3$ phase noise slope in the oscillator, just as a constant frequency deviation with offset frequency was seen to generate $1/f^2$ phase noise slope.

NEGATIVE RESISTANCE OSCILLATOR: PARALLEL CASE



The negative conductance oscillator serves as a model for the parallel resonator circuit. Oscillation is modeled again as amplified noise in a very narrow bandwidth, with limiting action being the control of the negative conductance.

Consider the circuit with zero negative conductance. The parallel LC circuit has loss and associated thermal noise. That noise and noise associated with the noise figure of the active device are used to derive the noise current. The power spectral density of current equals Fkt, of which half is in the phase noise and half in amplitude noise direction. Only that signal in phase with the amplitude direction contributes to output power.

In the parallel case the output appears as voltage across the load. As the negative conductance increases in magnitude, the parallel conductance decreases, approaching zero. The output voltage density increases directly with decrease of conductance, meaning power spectral density increases as voltage squared. The bandwidth decreases directly as conductance drops. Since the output power is derived from voltage squared and bandwidth directly, it increases directly as total conductance drops.

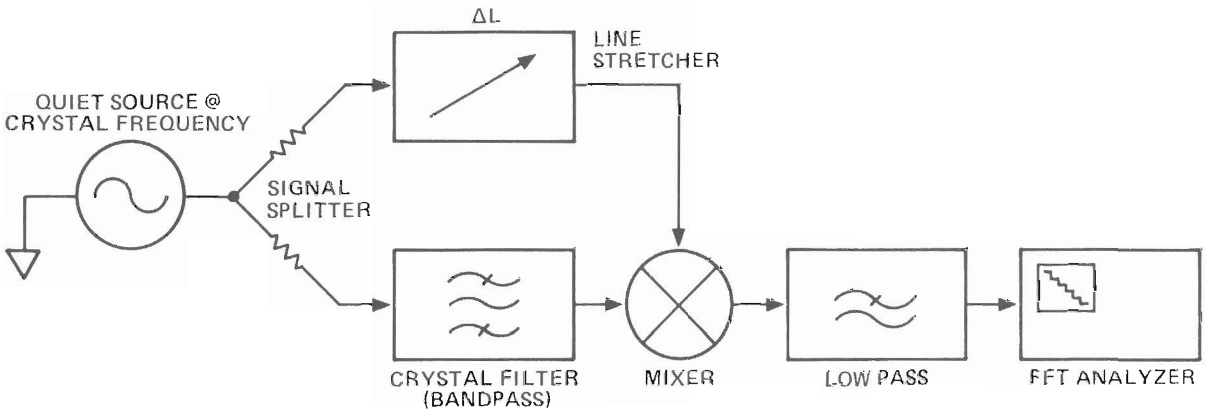
$$G_{total} = G_r + G_{load} + G_{feedback} + (-G)$$

$$V_{out} = i_n \quad G_{total}$$

$$BW = f_o/Q$$

$$Q = 1/\omega L \; G_{total}$$

SINGLE CRYSTAL DISCRIMINATOR — PHASE NOISE TEST

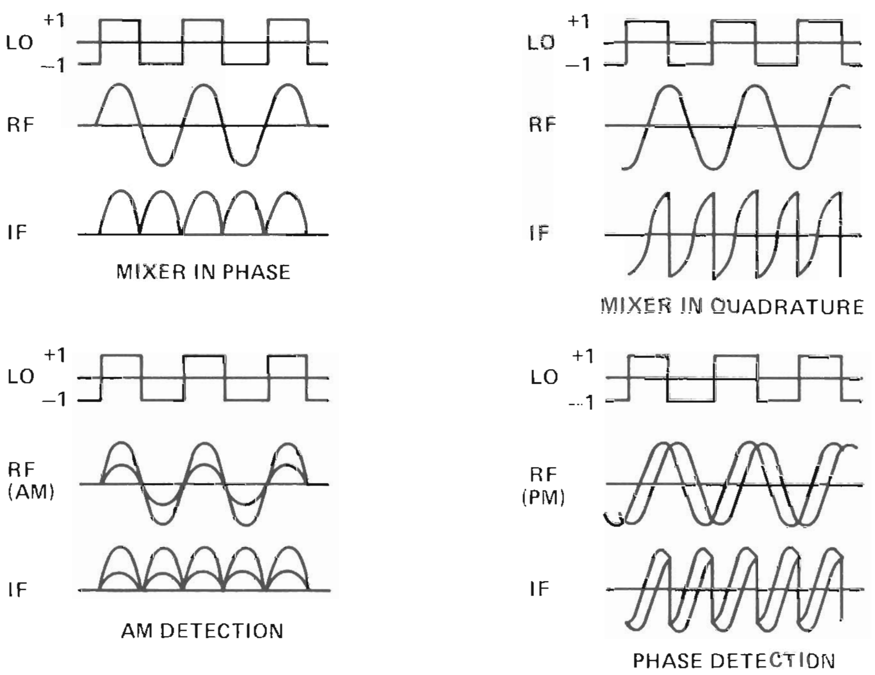


The crystal resonator may often be the dominant source of noise in an oscillator. The noise expected from the transistor circuitry may be 10 to 30 dB lower than that actually seen in the oscillator. As crystals are swapped in and out of the oscillator the phase noise is seen to be related to the crystal rather than the transistor. The noise of a given crystal is repeated in different oscillators.

A system to measure the phase noise of a single crystal resonator was constructed as above. The circuit has many of the attributes of a discrimination. The signal from a very quiet source is split into two paths. One signal passes through a line stretcher which is used to achieve 90° phase (quadrature) between LO and RF signals at the mixer. The output of the line stretcher serves as LO for the mixer.

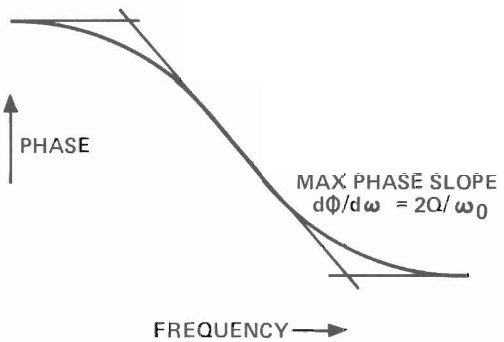
The second signal path passes through a crystal filter. Any noise in phase of the resonator impedance will cause a change in the phase of the signal at the output. The mixer will detect the change in phase and it will be amplified and observed on the spectrum analyzer. The carrier signal will be mixed down to DC and offset frequencies will be mapped to baseband.

MIXER AS A PHASE DETECTOR

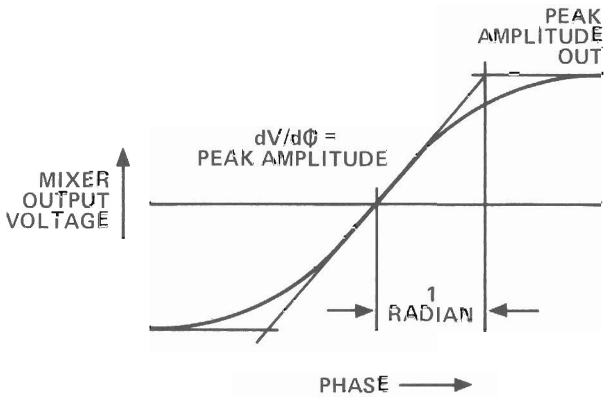


To review how a mixer operates as a phase detector, consider the two cases of LO and RF phase above. For the case of both signals in phase, the output appears similar to a full wave rectified signal. Any amplitude change appears as a shift in the average output. Small phase changes have little effect. In the second case, that of quadrature, the average output (low passed) is zero volts. Any amplitude changes tend to cancel out, but phase changes shift the output up or down.

PHASE NOISE TEST SYSTEM DETAILS



RESONATOR: PHASE VS FREQUENCY



MIXER OUTPUT VS PHASE

The test system has several factors that limit its performance (dynamic range). Phase noise of the source will be detected and mask device noise. As shown above, the maximum phase slope $d\phi/dw$ is the inverse of the half bandwidth, and the mixer phase detection sensitivity is equal to the peak output amplitude.

$d\phi/df = 2Q/f_o = 1/(\text{half BW})$ (this is the resonator group delay)

$dV/d\phi$ = Peak Output Amplitude

By narrow band FM approximation:

$$\frac{\text{Phase Noise Amplitude}}{\text{Carrier Amplitude}} = \frac{\Delta f}{2f_m} = \frac{\text{Deviation Frequency}}{\text{Modulation Frequency}}$$

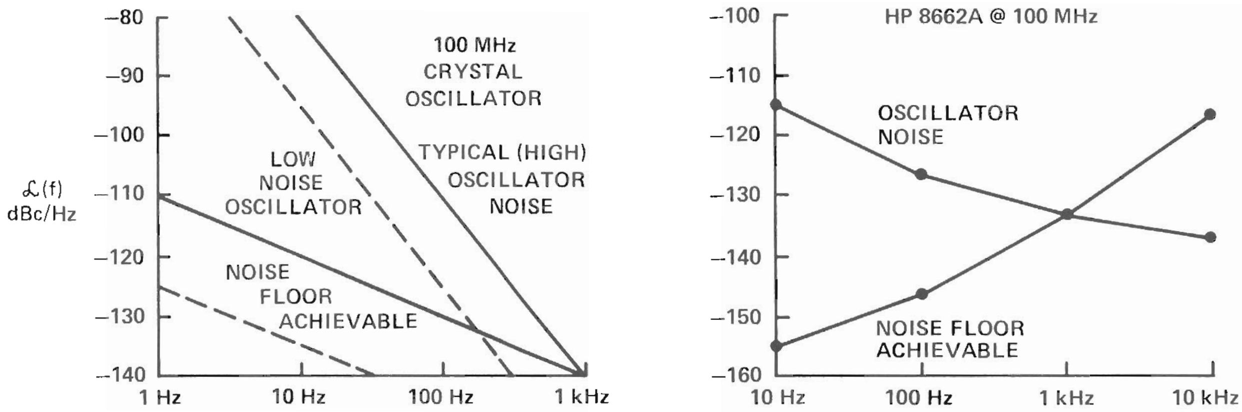
$\Delta\phi = \Delta f/(\text{half BW})$ and $\Delta f = 2f_m$ Source Phase Noise

ΔV = Peak Amplitude $2f_m$ Source Phase Noise/(half BW)

The mixer presents resonator phase noise from both sidebands (upper and lower) at the IF output. Both sidebands are assumed correlated and to add as voltage. Therefore the lowest detectable resonator noise (due to source masking) is given below:

SSB resonator noise floor $= \Delta V$/(twice Peak Output Amplitude)
$= f_m$ Source Phase Noise/(half BW)

EXAMPLE NOISE FLOOR CALCULATIONS



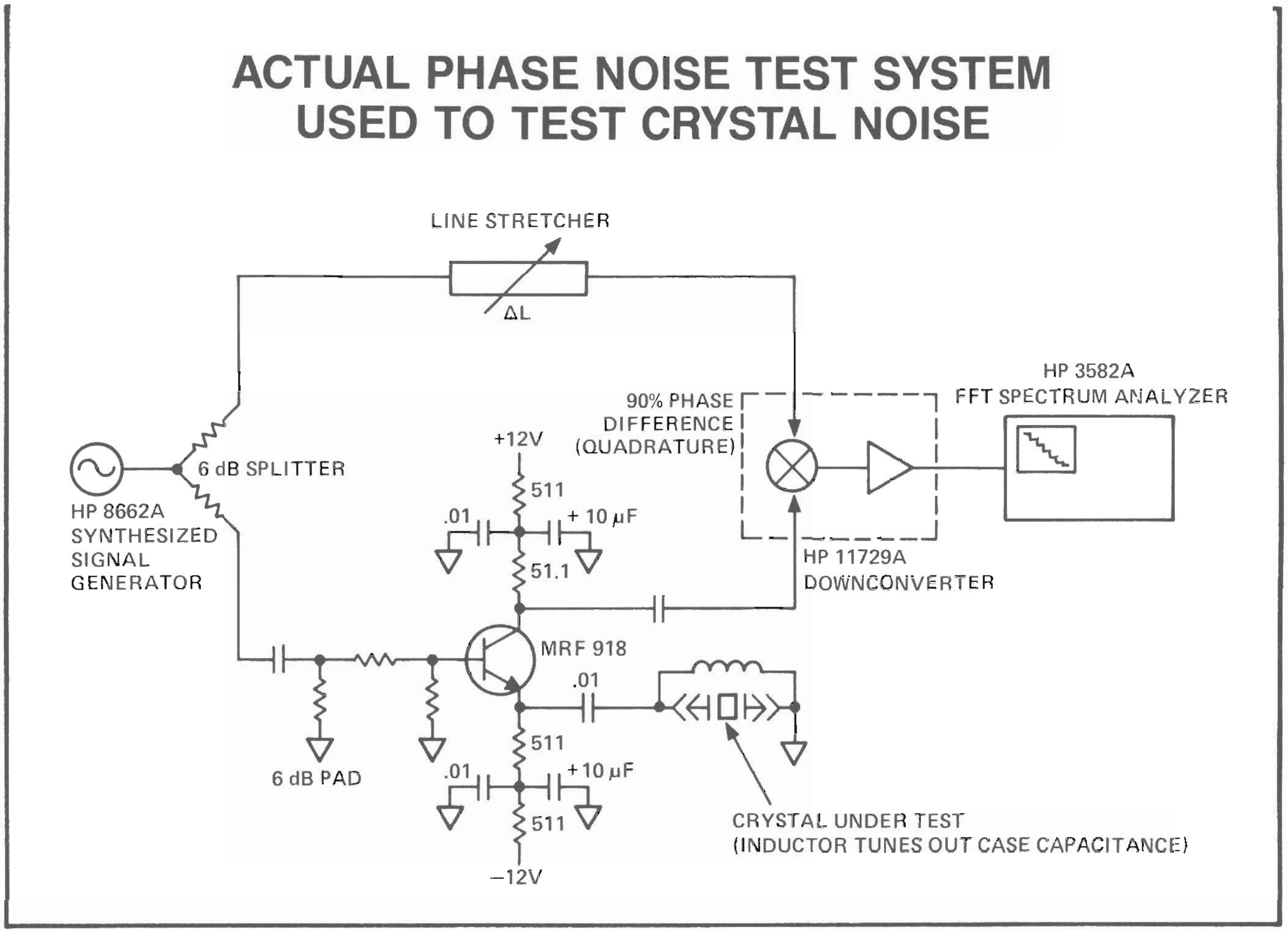
The phase noise of the source is folded around the half bandwidth and pulled down at 20 dB per decade inside that half bandwidth. Outside the bandwidth of the resonator the sideband amplitude will be rolled off due to the bandpass character of the resonator.

Another contributor to noise floor is the low frequency noise of the spectrum analzyer or preamp at the output of the mixer. The amplifier input noise relative to the peak output of the mixer sets the double sideband noise detection floor. An amplifier with 3 nV/ Hz input noise and mixer output of .1 volt would give a noise floor:

$$20 \log (3\ nV/0.1V \times 2) = \text{SSB Noise Floor} = -156\ dBc/\ Hz$$

Several examples of discriminator noise calculations are shown above. The 100 MHz oscillator will give a phase noise floor comparable to the crystal noise floor in that oscillator. To measure a quieter crystal, a better source such as an HP 8662A is needed.

Previous systems for testing crystal phase noise have used two crystals to avoid the problem of discriminating the source phase noise, but with a very clean source (HP 8662A) a single crystal measurement is possible. This avoids confusion and makes correlation of noise with other parameters easier.

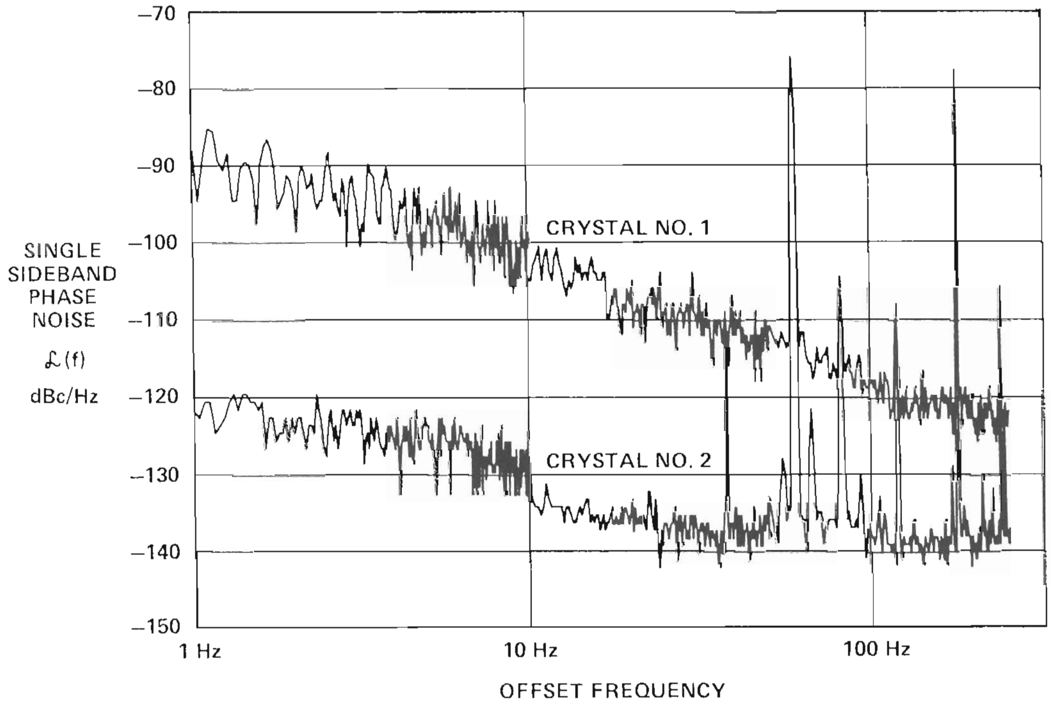


One important thing to remember is often neglected. The crystal should be presented with an impedance that won't degrade the Q, or else phase noise will also be altered. The active device is biased to always operate in a class A condition to preserve a low driving impedance. The input voltage to the crystal filter network is reduced by a 6 dB attenuator and presented to the base. This same voltage is present at the emitter, across the series resistance of the crystal. Enough current must be available through the biasing source to drive the crystal.

The crystal also has the case capacitance tuned out by a parallel inductor. This removes the effects of parallel resonance so close to series resonance.

The HP 11729A down converter and HP 3582A FFT spectrum analyzer allow noise measurements down to low offsets. The dc output and a low-noise amplifier (DC) built with an OP-27 Op Amp allowed measurement down to -130 dBc/Hz at 1 Hz.

NOISY AND QUIET CRYSTALS IN FREQUENCY DOMAIN

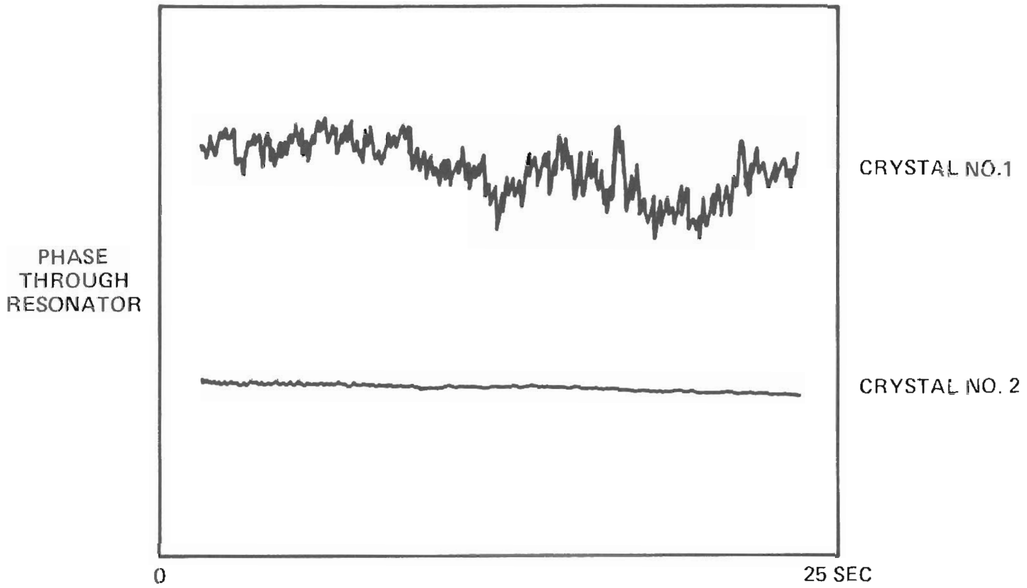


The phase noise of quiet and noisy crystals covered quite a large range. The noisy crystals were typically characterized by noise slopes greater than 10 dB per decade (corresponding to 1/f noise). Sometimes the sinc function shape was visible in noisy plots indicating a "pop" or noise burst.

A definite quieting with warm-up was also observed. The noise of even the best crystals was worst when first tested. This fires speculation of contamination that is shaken off or evaporated off.

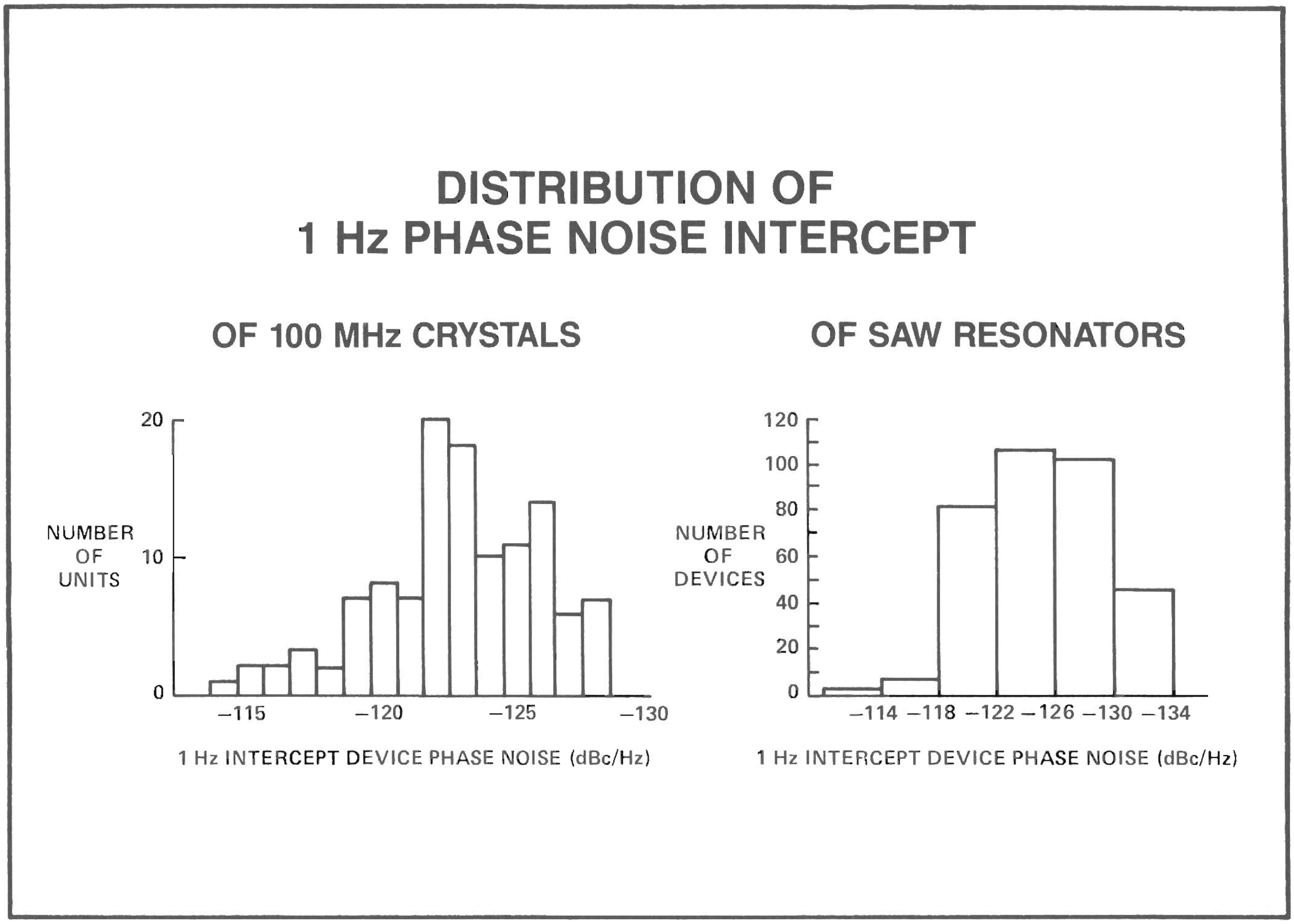
Another effect seen was that of ambient temperature changing the crystal frequency. The phase noise below tens of Hz offset often shows $1/f^2$ behavior, which is removed by good thermal insulation. If a random fluctuation of temperature caused random frequency change, one would expect $1/f^2$ phase noise using narrow band FM approximations as before.

NOISY AND QUIET CRYSTALS IN TIME DOMAIN



Plots of the input time domain record were made of crystals at the extremes of low and high noise. The quiet crystal showed a very low noise level and constant drift during warm-up. The noisiest crystal had periods of what looked like oscillations or "popcorn noise".

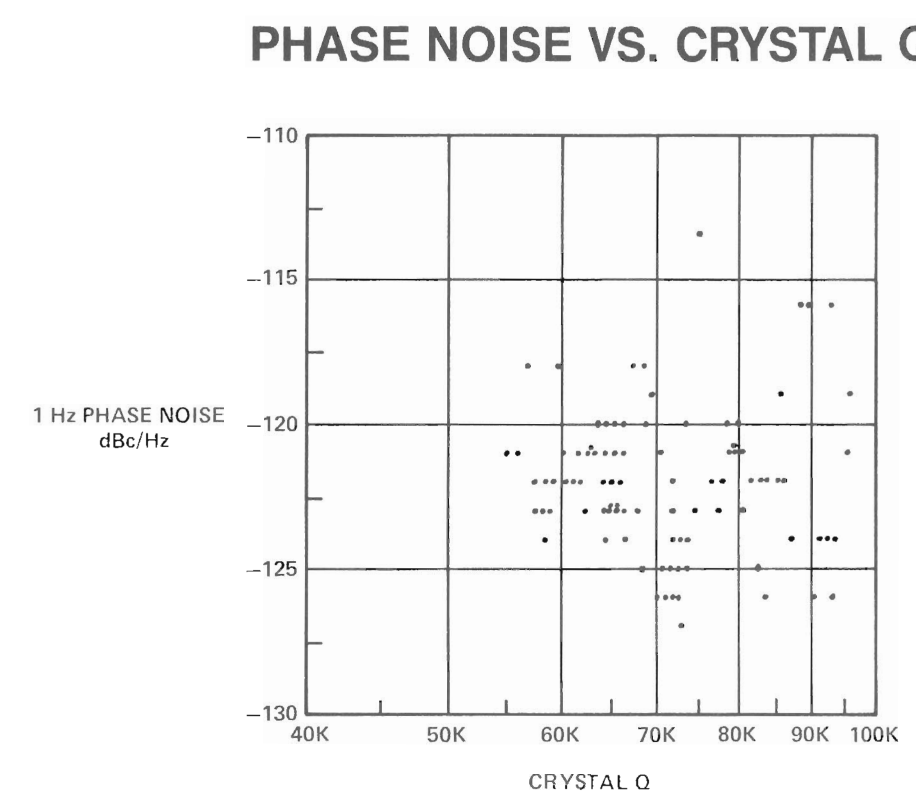
The graph and phase noise chart show data for 100 MHz crystals, where most measurements were taken, but other crystals were also measured, with similar noise levels. Several crystals were measured on many overtones, and it was observed that if one overtone had popcorn noise, it was likely to be noisy at all overtones measured. The exact cause of the noisy behavior remains to be found, but speculation has centered upon the cleanliness of the metalization of the electrical contacts.



From a sample of several hundred 100 MHz crystals, those crystals showing phase noise with a slope of 1/f or lower were plotted on a histogram. The distribution was centered around -120 to -125 dBc/ Hz. The crystals with burst noise were very unpredictable in noise level and were removed from this analysis.

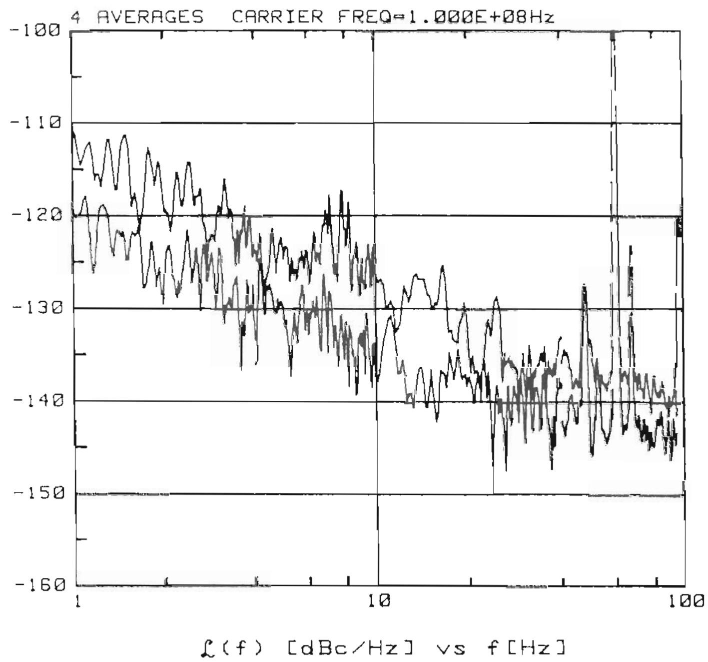
This distribution is remarkably similar to that reported by Bob Bray and Scott Elliot, reprinted here adjacent to the crystal distribution. The histogram reports the 1 Hz phase noise intercept of SAW devices.

The measurement was done over a 10 Hz to 100 Hz offset frequency range, with 1 Hz intercepts scaled by a 1/f slope. The temperature variations gave higher phase noise interepts than those extrapolated from the 10 Hz to 100 Hz offset frequency phase noise.



The crystal phase noise was plotted against many variables to look for a correlation between some easier-to-measure parameter and phase noise. One often-reported link was between Q and phase noise, but not much is seen here to indicated a strong link. One possibility seen was the connection between R_S change versus drive level and higher phase noise. This test was not done on the 100 MHz crystals, but on several 20 MHz crystals without conclusive proof. More investigation needs to be done.

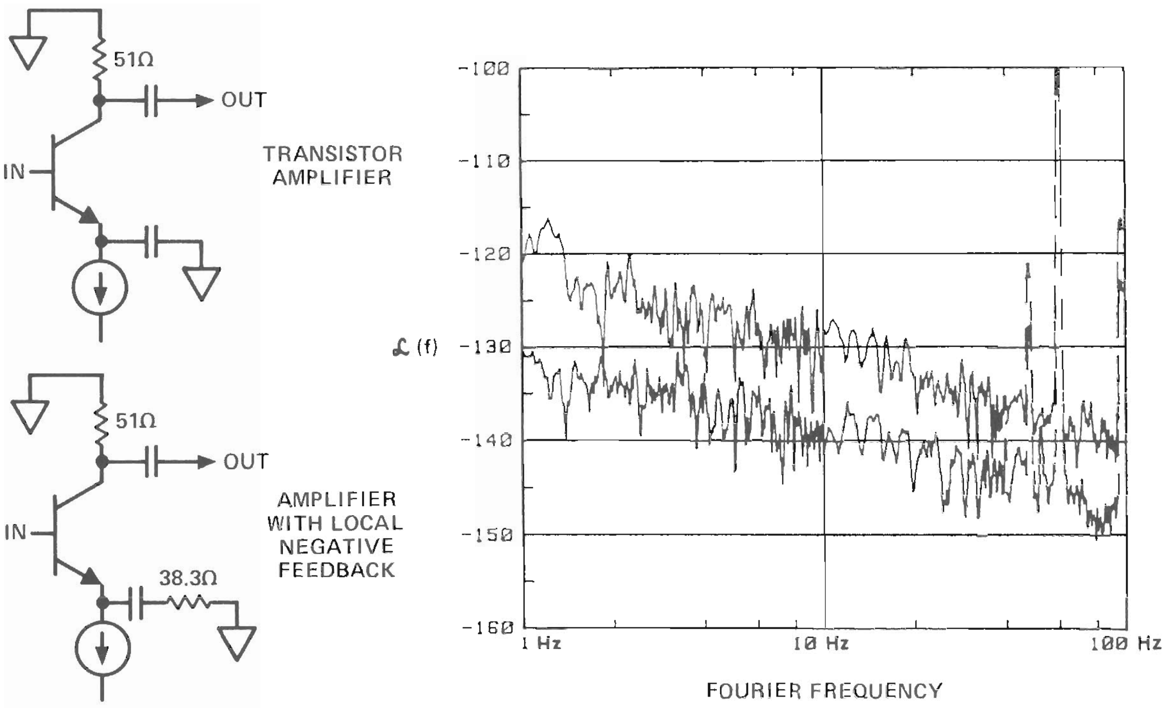
CRYSTAL PHASE NOISE @ 100 MHz #14



Earlier on, it was indicated that adding resistance in series with the crystal should reduce the noise level and increase the bandwidth. The plot of phase noise above shows two tests of crystal phase noise. On the lower trace a series resistor equal to crystal series resistance was added. The noise shows a drop of about 6 dB as predicted by the model.

If the noise of the crystal is reduced too far, we begin to see the noise floor (thermal or noise figure) dominating phase noise.

TRANSISTOR PHASE NOISE — WITH AND WITHOUT FEEDBACK



Transistors also have similar phase noise problems. The transistor used in the crystal test fixture was bypassed directly to ground at the emitter and its phase noise measured. A resistance equal to crystal series resistance was added in series with the bypass capacitor and noise was seen to drop to a value equivalent to the measurement floor. This indicates that the crystal noise does in fact dominate among phase noise contributors. The crystal impedance provides enough local negative feedback to reduce device phase noise.

PREDICTED AND ACTUAL OSCILLATOR PHASE NOISE

CRYSTAL	CONDITION	PHASE NOISE (dBc/Hz) @ OFFSET FREQUENCY					
		10 Hz	20 Hz	30 Hz	50 Hz	70 Hz	100 Hz
NO. 1	PREDICTED	–101	–109	–114	–120	–124	–128
	ACTUAL	–101	–109	–115	–122	–125	–129
NO. 2	PREDICTED	–102	–110	–114	–120	–124	–128
	ACTUAL	–101	–111	–117	–124	–125	–130
NO. 3	PREDICTED	–99	–108	–112	–119	–123	–127
	ACTUAL	–98	–108	–114	–122	–126	–129
NO. 4	PREDICTED	–92	–101	–106	–113	–117	–122
	ACTUAL	–88	–97	–104	–113	–116	–121
NO. 5	PREDICTED	–93	–102	–108	–115	–119	–124
	ACTUAL	–95	–103	–109	–117	–121	–126
NO. 6	PREDICTED	–99	–107	–112	–118	–122	–127
	ACTUAL	–97	–106	–112	–118	–124	–130
NO. 13 $R_S = 37.2\Omega$	Actual, No Added Resistor	–42	–52	–58	–67	–71	–76
	10Ω SERIES	–42	–52	–58	–67	–71	–76
	20Ω SERIES	–42	–52	–58	–66	–71	–76
	25Ω SERIES	–39	–50	–55	–63	–68	–71
NO. 17	Actual, No Added Resistor	–52	–61	–66	–74	–79	–81
	10Ω SERIES	–51	–61	–66	–75	–78	–81
	20Ω SERIES	–47	–57	–63	–68	–74	–78
	28Ω SERIES	–44	–52	–57	–64	–68	–71

Several examples of crystal phase noise measurements, predicted oscillator noise and achieved oscillator noise appear above. The noise of an oscillator is generally predictable from crystal noise and bandwidth measurements.

One interesting aspect of the data shows that for small amounts of added resistance, no degradation of oscillator noise occurred close to the carrier. A rise in noise near the half bandwidth is expected. For series resistance comparable to the crystal internal series resistance, some degradation of noise is seen.

Prediction of noise assumes that the crystal is measured in an environment similar to that it will see in an oscillator. The driving impedances should be similar, to assure bandwidths of test and oscillator resonator are identical. Given the crystal phase noise at an offset frequency, the oscillator noise is predicted by increasing that noise by 20 dB per decade as offset frequency decreases inside the resonator half bandwidth.

HOW TO MINIMIZE PHASE NOISE

- MAXIMIZE THE Q (WITH CONSTANT DEVICE NOISE).
- MAXIMIZE THE SIGNAL POWER VS NOISE POWER.
- DRIVE THE CRYSTAL (SERIES) WITH A CONTROLLED, LOW IMPEDANCE.
- DRIVE THE CRYSTAL WITH A LINEAR AMPLIFIER (DON'T LIMIT IN THE STAGE THAT DRIVES THE RESONATOR).
- USE A QUIET RESONATOR, THE OSCILLATOR FREQUENCY WILL FOLLOW THE RESONATOR FREQUENCY FLUCTUATIONS.
- CAREFULLY CONTROL LIMITING MECHANISM SO AS TO NOT INTRODUCE AM NOISE.
- USE A QUIET DEVICE.
- TAKE OUTPUT POWER AFTER THE RESONATOR IF POSSIBLE TO FILTER NOISE AT LARGE OFFSET FREQUENCIES.
- OPTIMIZE THE NOISE WHERE IT IS NEEDED, ESPECIALLY CONSIDER CLOSE IN NOISE VS. LARGE OFFSET NOISE REQUIREMENTS.
- IN A FILTER APPLICATION, CONSIDER REDUCING RESONATOR Q IF WIDER BANDWIDTH IS ACCEPTABLE.

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