

GEORGE PONTIS

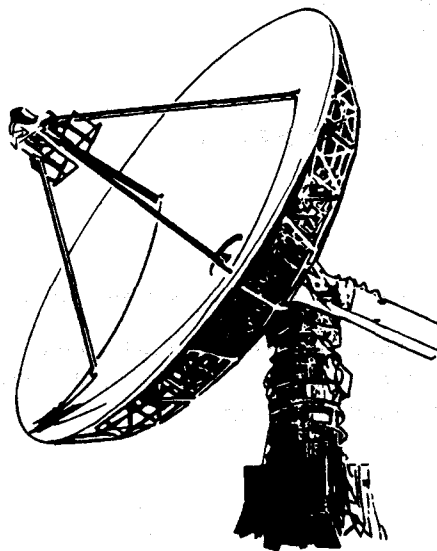
Generation of Low Phase Noise Microwave Signals

**Dieter Scherer
September 1981**

**RF & Microwave
Measurement
Symposium
and
Exhibition**

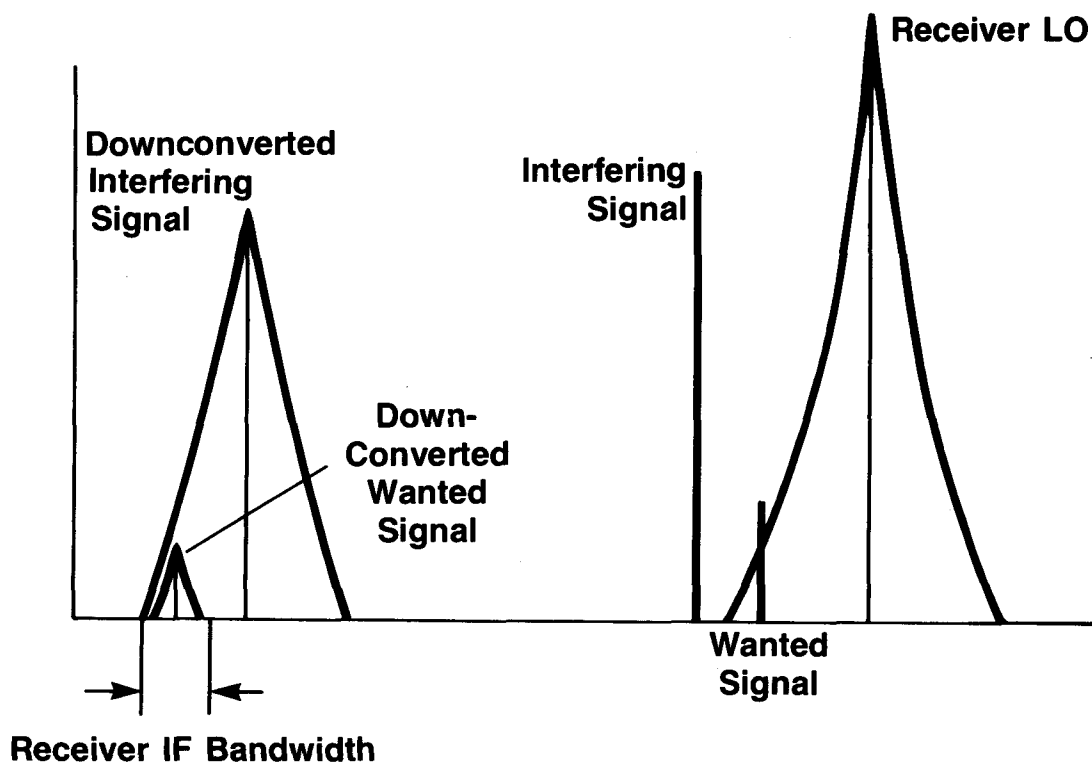


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1. Why are Low Phase Noise Microwave Signals needed?

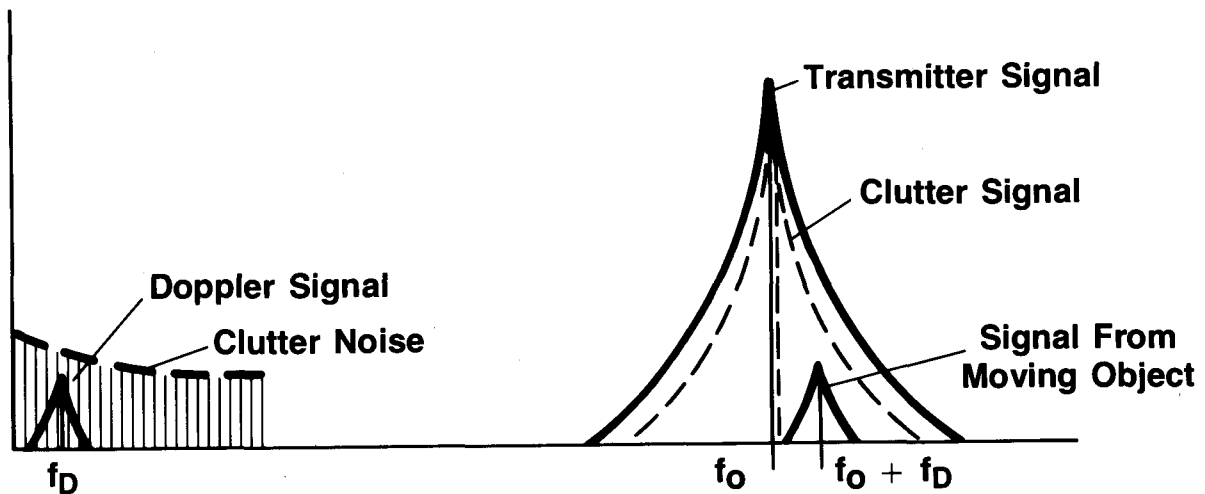
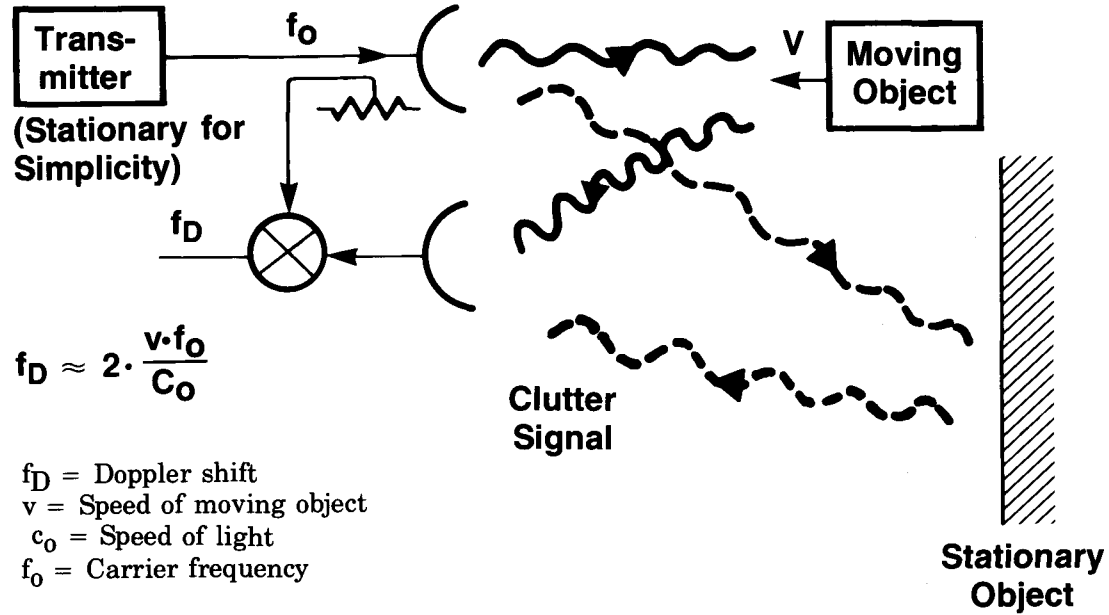
Example 1: Effect of Local Oscillator Phase Noise on Receiver Selectivity in Multi Signal Environment



The receiver tries to see a weak wanted signal in the presence of a strong interfering signal.

The downconversion process transfers the noise sidebands of the receiver LO to both the wanted and interfering IF signal. Sideband noise of the interfering IF signal falling into the IF bandwidth submerges the wanted IF signal.

Example 2: Effect of Carrier Phase Noise in a Doppler Radar System



The receiver of the Doppler Radar System detects the weak frequency shifted return signal from a moving object as well as a strong, unwanted return signal at f_0 from a large stationary object (clutter signal).

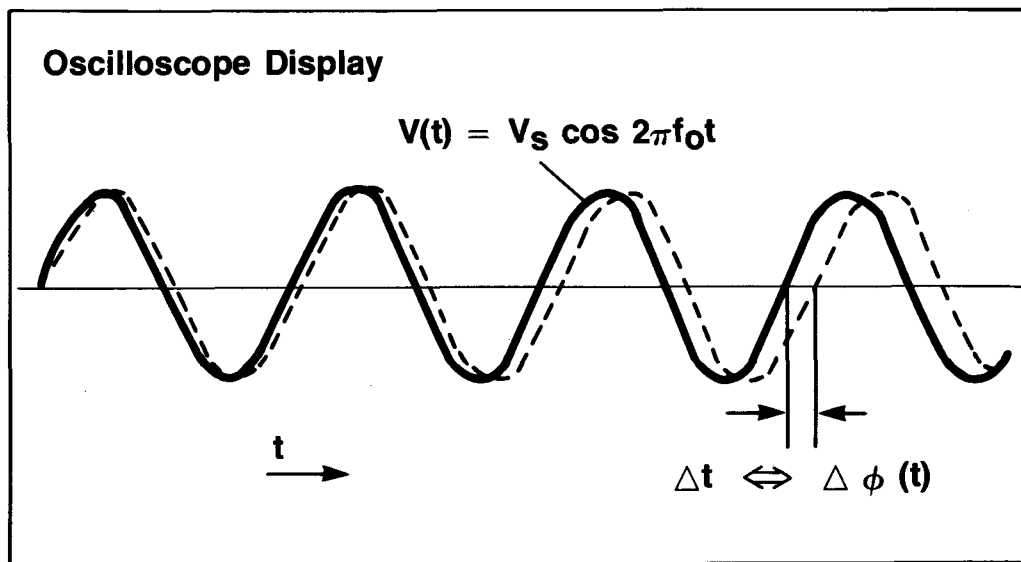
The delay of the clutter return decorrelates the phase noise of the clutter signal from the transmitter signal. The resulting clutter noise at the IF port may exceed the weak Doppler signal.

2. Basic Representations and Definition of Phase Noise

In the Time Domain:

v_t = Signal with random phase fluctuation $\Delta \phi (t)$

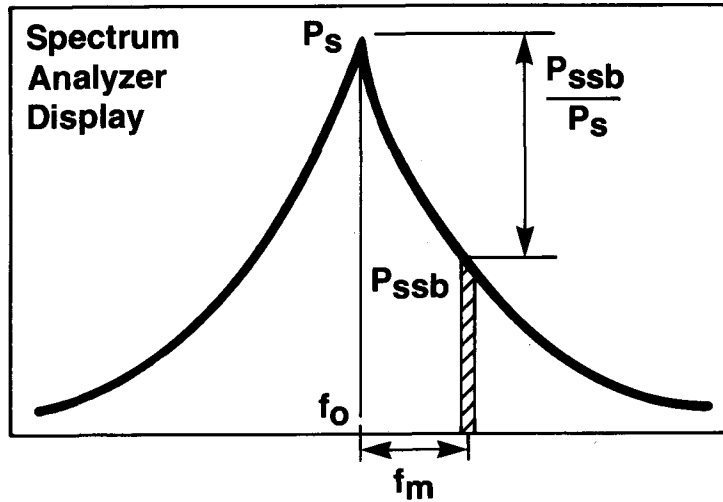
$$v(t) = V_S \cos [2\pi f_0 t + \Delta \phi (t)]$$



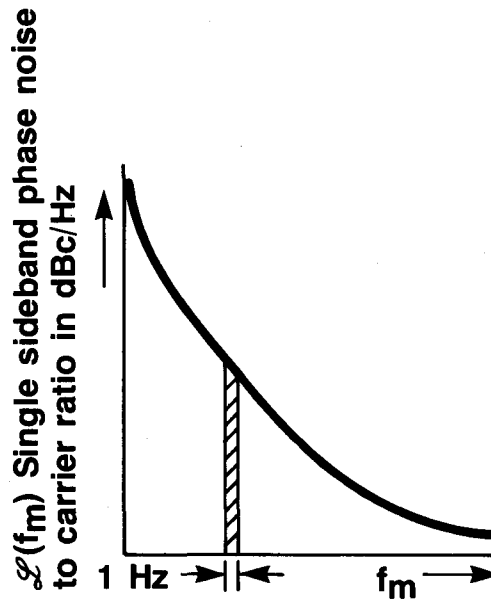
Frequency and phase are related by

$$f(t) = \frac{1}{2\pi} \frac{d\phi(t)}{dt}$$

In the Frequency Domain:



$$\mathcal{L}(f_m) = \frac{P_{ssb} \text{ (per 1 Hz)}}{P_s}$$



$\mathcal{L}(f_m)$ describes the RF power spectrum and is defined as the ratio of the single sideband power of phase noise in a 1 Hz bandwidth (f_m Hz away from the carrier frequency) to the total signal power.

3. Key Relations of $\mathcal{L}(f_m)$, $S_{\Delta\phi}(f_m)$, $S_{\Delta f}(f_m)$

For $\Delta\phi_{rms} \ll 1$

$$\mathcal{L}(f_m) = \frac{1}{2} \Delta\phi_{rms}^2 = \frac{1}{2} \left(\frac{\Delta f_{rms}}{f_m} \right)^2 = \frac{P_{ssb}}{P_s}$$

Spectral Density of Phase Fluctuations

$$S_{\Delta\phi}(f_m) = \Delta\phi_{rms}^2 = 2 \mathcal{L}(f_m)$$

Spectral Density of Frequency Fluctuations

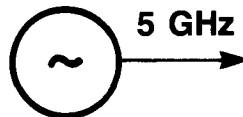
$$S_{\Delta f}(f_m) = \Delta f_{rms}^2(f_m) = f_m^2 S_{\Delta\phi}(f_m) = 2 f_m^2 \mathcal{L}(f_m)$$

4. Three Basic Approaches

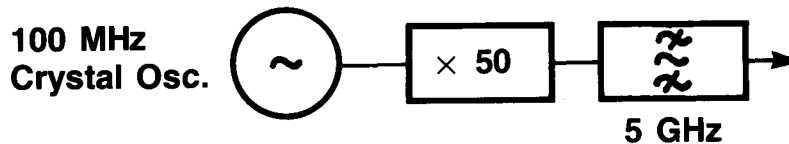
to obtain a "quiet" (low phase noise) microwave signal are considered in the following pages.

A. Build a Quiet Microwave Oscillator

e.g., to build a 5 GHz oscillator, use a bipolar transistor and a high Q cavity resonator

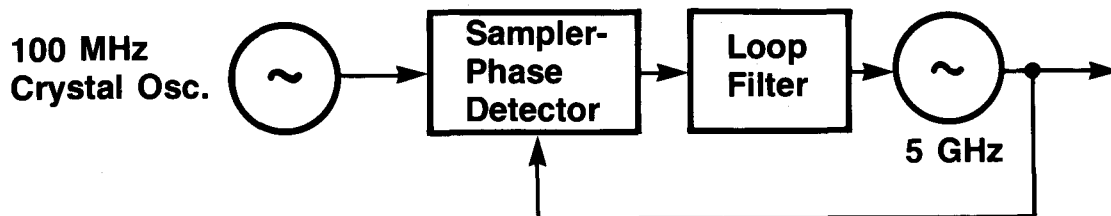


B. Multiply a Low Phase Noise HF or VHF Source



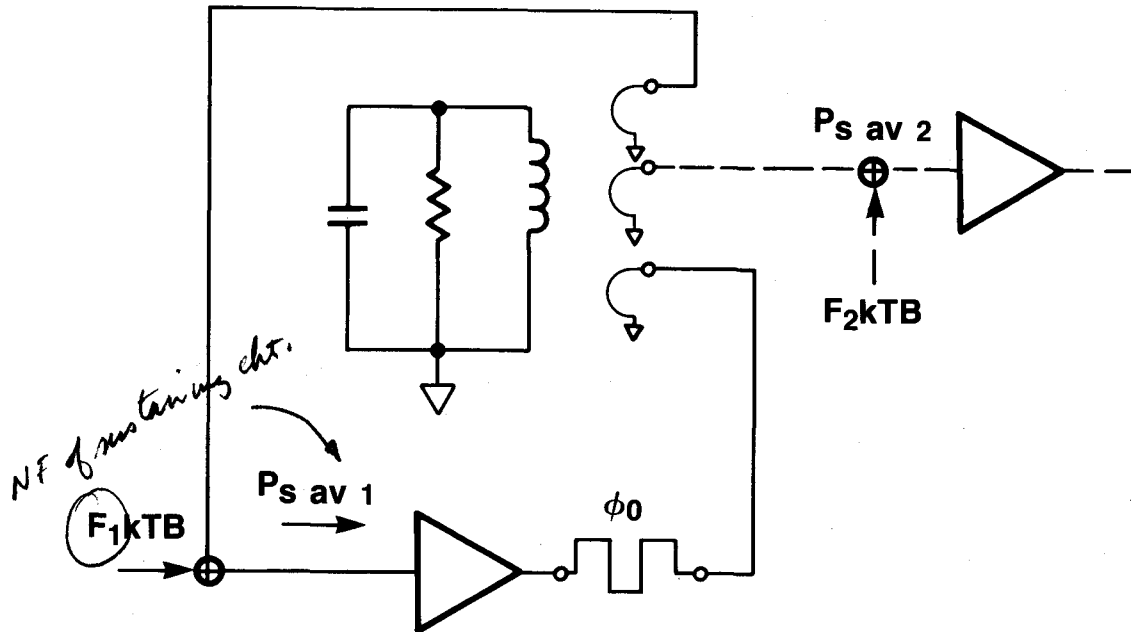
e.g., with a 100 MHz crystal oscillator drive a step recovery diode multiplier generating the 5 GHz signal.

C. Phaselock the Microwave Oscillator to the Quiet HF or VHF Source



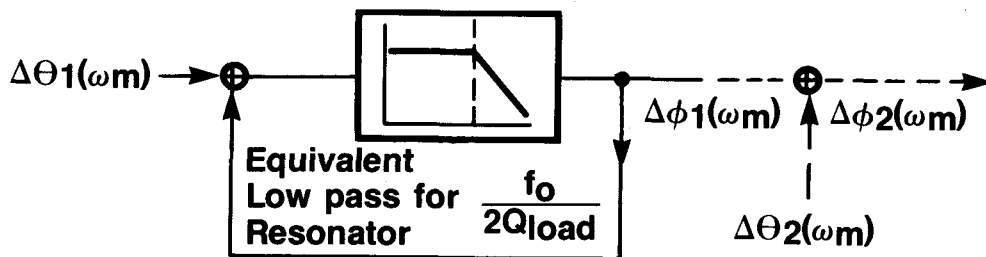
e.g., combine the performance advantages of a 100 MHz crystal oscillator and the 5 GHz microwave oscillator by phase locking the two sources via a sampling phaselock loop.

4.1 Phase Noise of Free Running Oscillators



Equivalent Phase Feedback Loop:

$$L(\omega_m) = \frac{1}{1 + \frac{j\omega_m 2Q_{load}}{\omega_0}}$$



The closed loop response of the phase feedback loop due to the perturbation $\Delta\theta_1(\omega_m)$ is

$$\Delta\phi_1(\omega_m) = \frac{\omega_0}{j2\omega_m Q_{load}} \times \Delta\theta_1(\omega_m)$$

ex. output

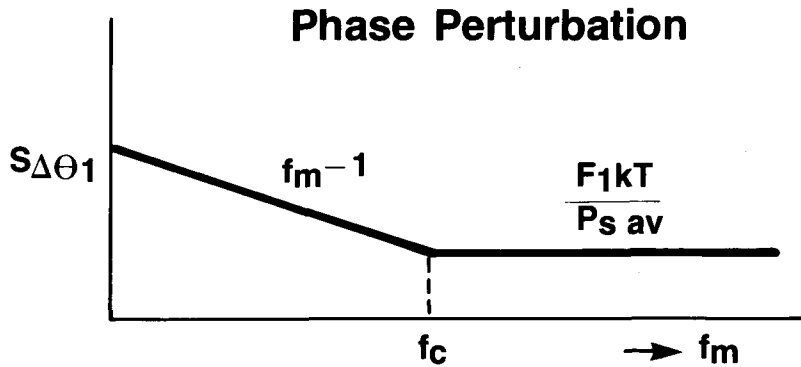
phase perturbation in input

The phase perturbation, $\Delta\theta_1(\omega_m)$, or its power spectral density, $S_{\Delta\theta_1}$, describes both the additive phase noise of the amplifier, $\frac{F_1 kT}{P_{s av 1}}$, and empirically the $\frac{1}{f}$ noise effect by f_c . (See also 5.1 and 5.2)

Power spectral density

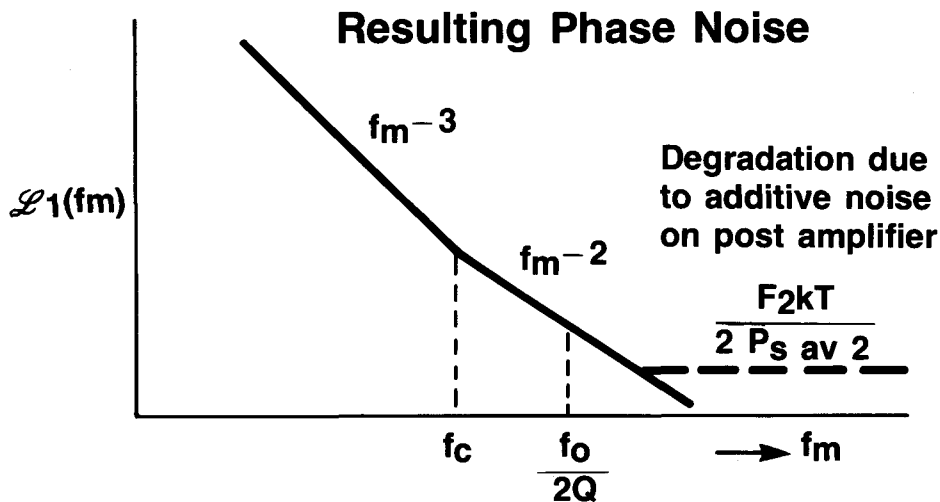
$$S_{\Delta\theta_1} = \Delta\theta_1^2(\omega_m)_{rms} = \frac{F_1 kT}{P_{s av 1}} \left(1 + \frac{f_c}{f_m}\right)$$

available noise power per Hz. (circled $F_1 kT$)
available signal power total (circled $P_{s av 1}$)
1/8 corner (arrow pointing to f_c)



The resulting phase noise of the signal coupled out of the resonator is

$$\mathcal{L}_1(f_m) = \frac{1}{2} S_{\Delta\phi_1} = \frac{1}{2} \frac{1}{f_m^2} \left(\frac{f_0}{2Q_{load}}\right)^2 \times S_{\Delta\theta_1}$$



Additive noise of the following amplifier establishes the phase noise floor, $\frac{F_2 kT}{2 P_{s av 2}}$

$$L(f_m) = \frac{F_i k T}{2 P_{\text{avail}} f_m^2} \left(\frac{f_0}{2Q} \right)^2 \left(1 + \frac{f_c}{f_m} \right)$$

4.2 Optimization of Phase Noise in Oscillators

- **Maximize Q_{loaded}**
 - Use resonator with maximal unloaded Q
 - Maximize reactive energy stored in resonator
 - Limit signal without degrading Q

- **Minimize phase perturbation**
 - Use device with low noise figure
 - Maximize signal level
 - Choose device with low flicker noise
 - Minimize effect of flicker noise

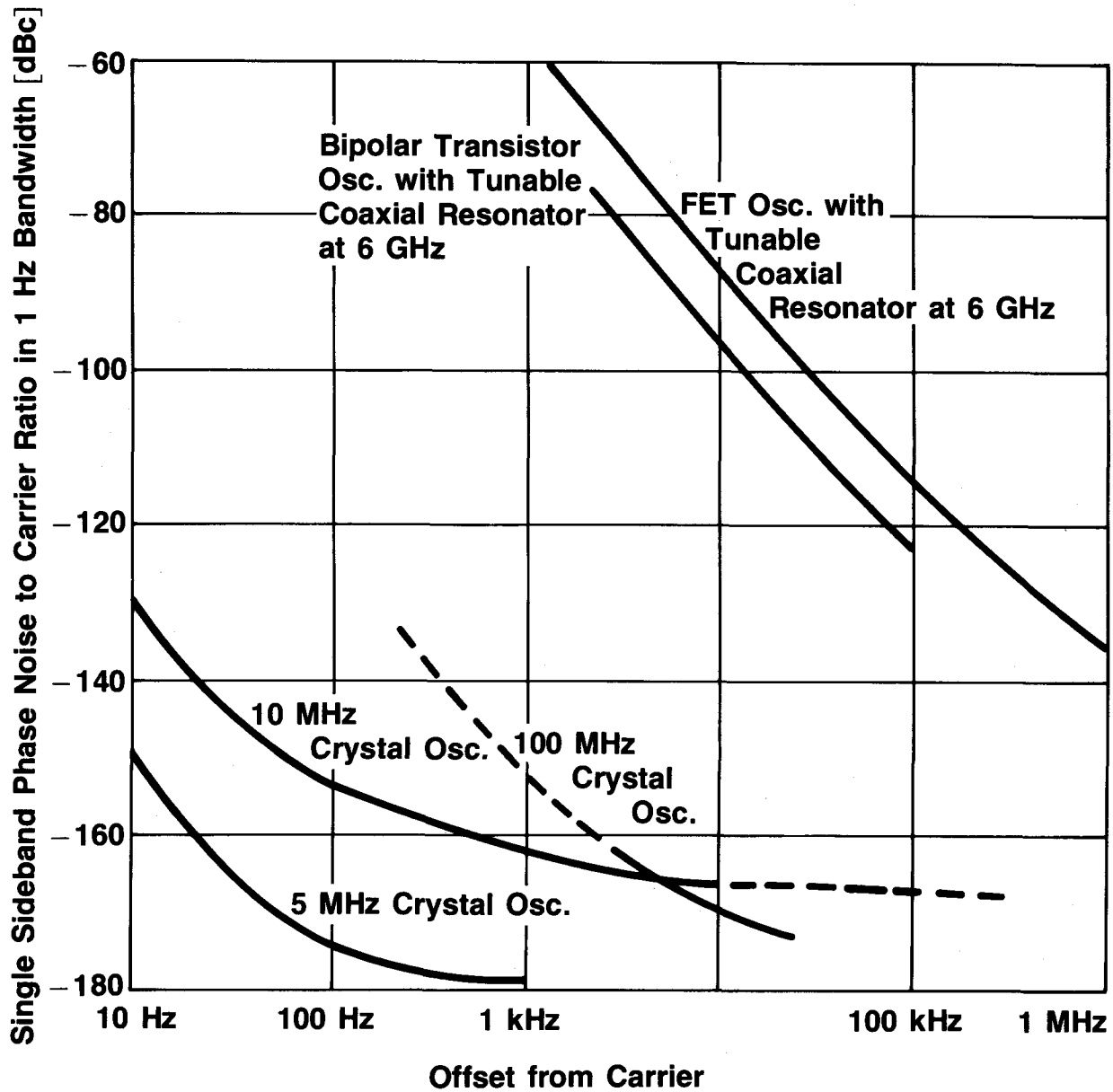
- **Couple output signal from resonator**

- **Set low-noise floor**
 - Keep output signal sufficiently high
 - Use low NF post amplifier

4.3 EXAMPLES OF HIGH Q RESONATORS

Resonator	Frequency	Unloaded Q	Temp. Coeff.	Major Limitation
Quartz Crystal, 3rd Overtone	10 MHz	1.5×10^6	$35 \times 10^{-9}/^{\circ}\text{C}^2$	applicable power
Quartz Crystal, 7th Overtone	100 MHz	90,000	$35 \times 10^{-9}/^{\circ}\text{C}^2$	applicable power, spurious modes, microphonic phase modulation
Quartz Crystal, 11th Overtone	280 MHz	40,000	$35 \times 10^{-9}/^{\circ}\text{C}^2$	applicable power, spurious modes, microphonic phase modulation
SAW Resonator on Quartz	280 MHz	18,000	$35 \times 10^{-9}/^{\circ}\text{C}^2$	applicable power
SAW Resonator on Quartz	800 MHz	10,000	$35 \times 10^{-9}/^{\circ}\text{C}^2$	applicable power
Dielectric Cyl. Waveguide (TE _{01λ}) (Ba Ti ₄ O ₉) <i>Barrin titanite 6433</i>	4.9 GHz	8,300	<u>2ppm/°C</u>	
Cylind. Waveguide (TE ₀₁₁)	>2 GHz	20,000 to 50,000 <i>Copper plated</i>	1.5 ppm/°C (Invar)	size (for L, S Band), microphonic phase modulation

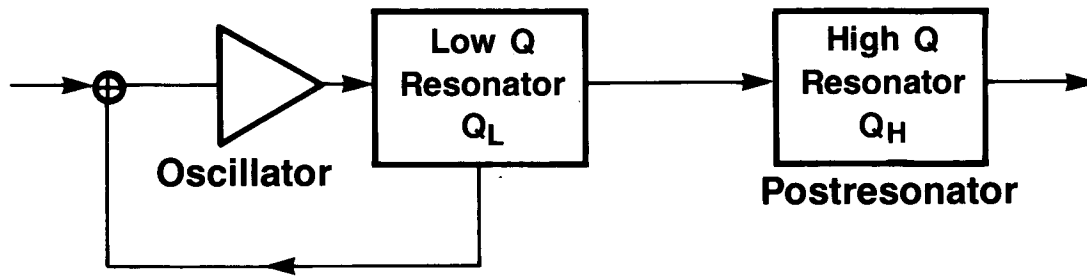
4.4 Examples of Low Phase Noise Oscillators



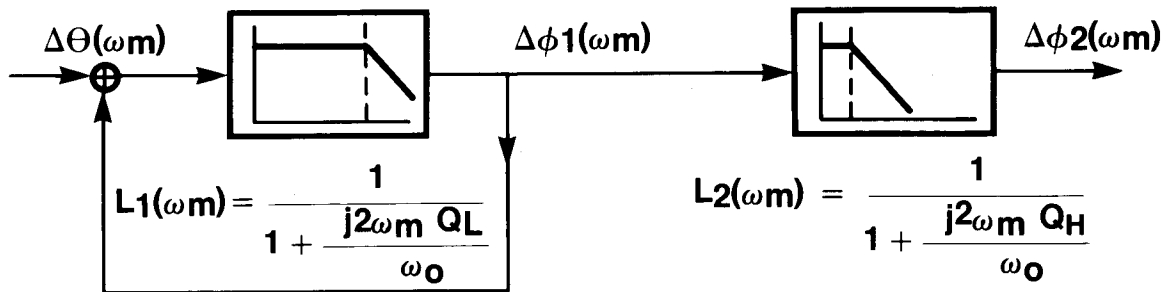
With the exception of the two tunable microwave oscillators, these examples of fixed oscillators give an indication of state of the art performance regarding low phase noise.

The two tunable oscillators at 6 GHz are shown to demonstrate the performance difference between GaAs FET's and bipolar transistors observed in otherwise comparable circuits.

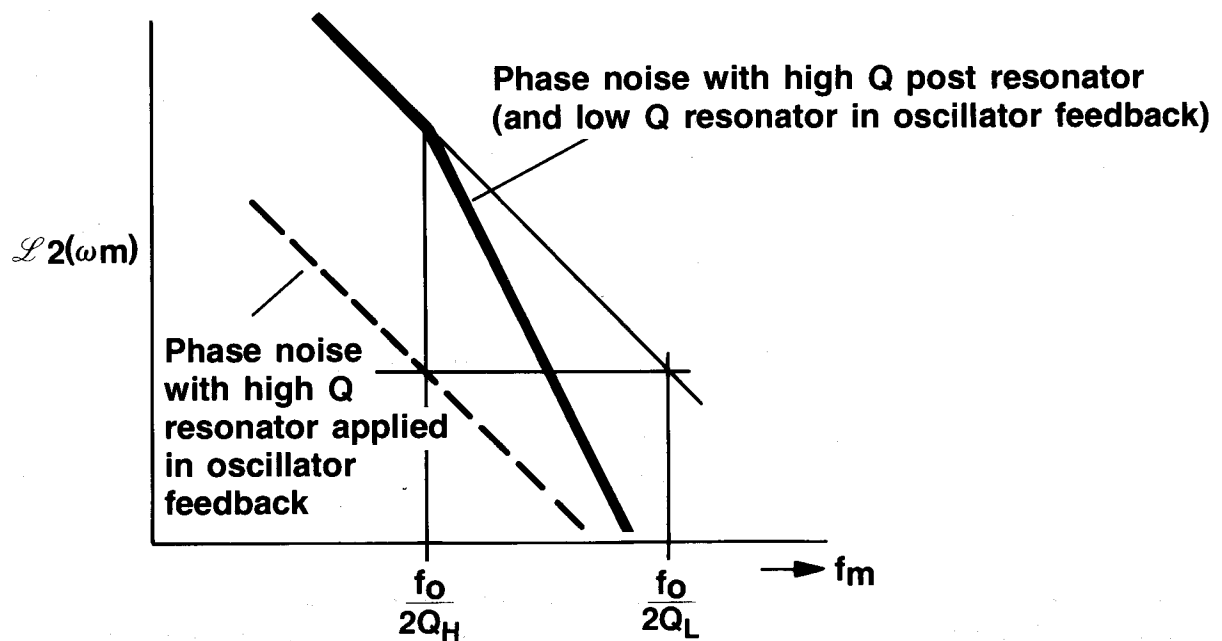
4.5 High Q Resonator Employed as Postresonator



Equivalent Circuit:

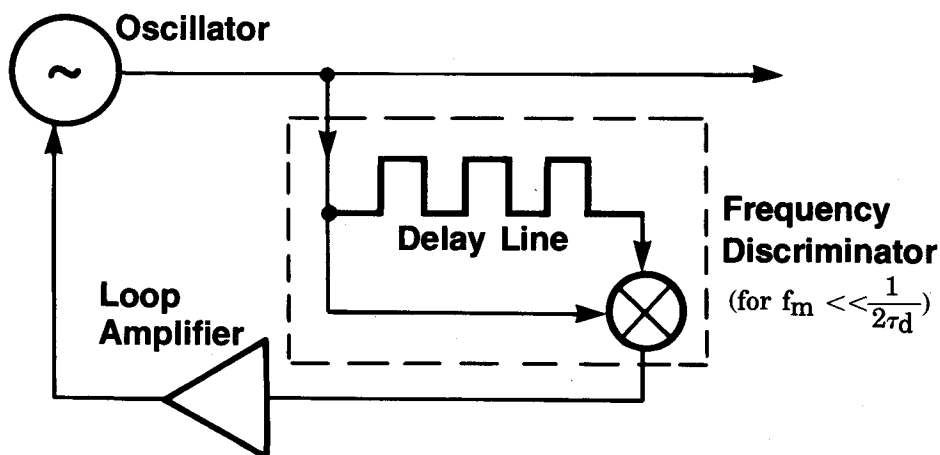


$$\Delta\phi_2(\omega_m) = \Delta\phi_1(\omega_m) \times L_2(\omega_m) = \Delta\theta \times \frac{\omega_0}{j2\omega_m Q_L} \times \frac{1}{1 + \frac{j2\omega_m Q_H}{\omega_0}}$$

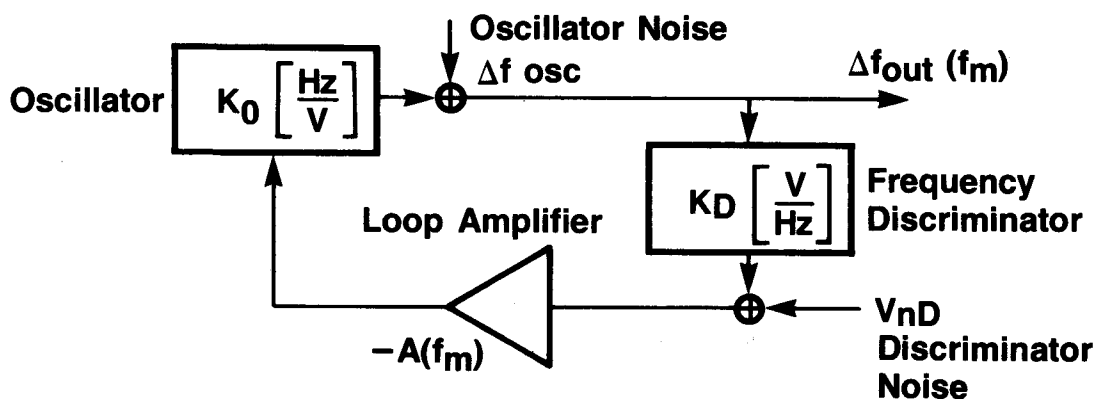


The application of a high Q postresonator lowers phase noise of an oscillator signal outside the bandwidth of the resonator. The above graph also suggests that lower phase noise can be obtained by applying the high Q resonator in the oscillator feedback.

4.6 Stabilization of Oscillator with Frequency Discriminator



Equivalent frequency feedback loop:



Reduction of oscillator noise (Δf_{osc} or $\Delta \theta_{\text{osc}}$):

$$\frac{\Delta f_{\text{out}}}{\Delta f_{\text{osc}}} = \frac{\Delta \phi_{\text{out}}}{\Delta \phi_{\text{osc}}} = \frac{1}{1 + K_0 K_D A(f_m)}$$

Frequency fluctuation caused by discriminator noise (V_{nD})

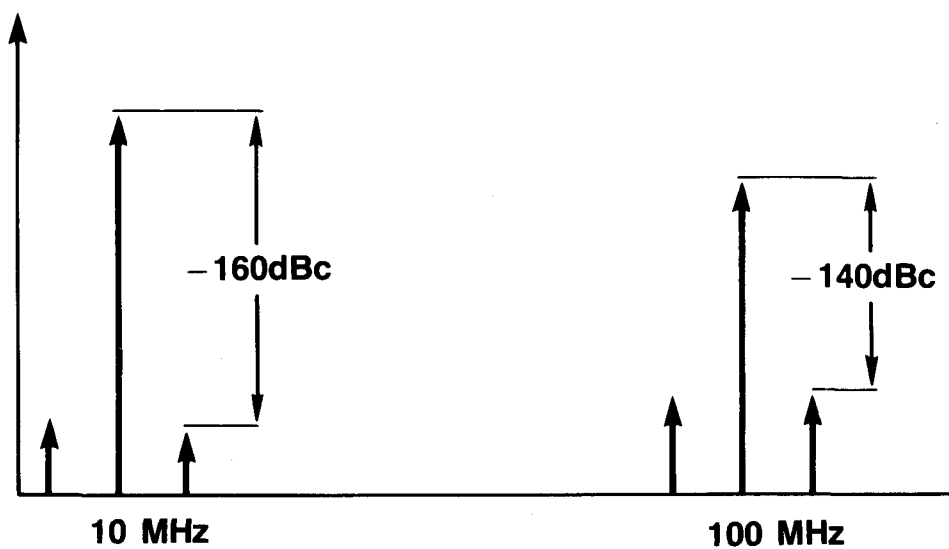
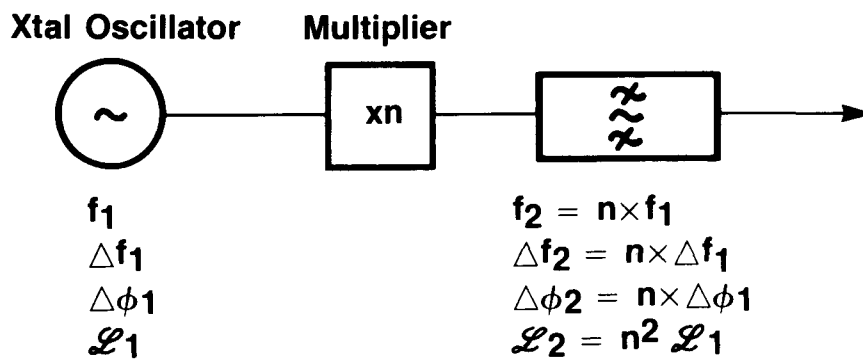
$$\Delta f_{\text{out}} (f_m) = V_{nD} \times \frac{1}{K_D} \times \frac{1}{\frac{1}{K_0 K_D A} + 1}$$

Phase noise caused by discriminator noise (obtained by integrating $\Delta f_{\text{out}} (f_m)$):

$$\Delta \phi_{\text{out}} (f_m) = V_{nD} \times \frac{1}{f_m K_D} \times \frac{1}{\frac{1}{K_0 K_D A} + 1}$$

The limitations of this system lie in the discriminator noise, gain, and bandwidth. The frequency discriminator typically consists of a delay line or microwave resonator and a mixer operating as a phase detector.

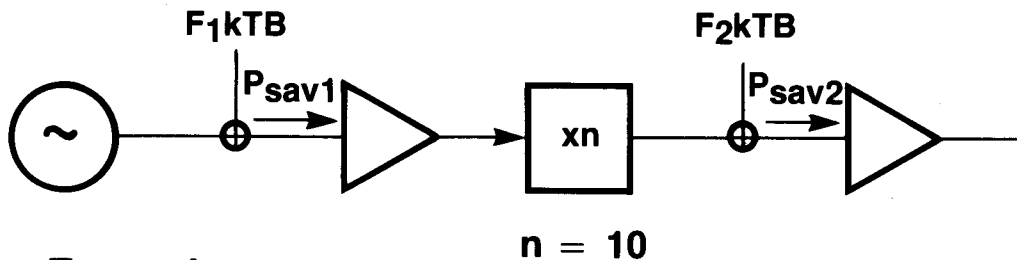
Multiplication of a Stable HF or VHF Source



Theoretical degradation of phase noise (with ideal multiplier)

$$\frac{\mathcal{L}_2}{\mathcal{L}_1} = 20 \lg n$$

5.1 Phase Noise Caused by Additive Noise in Amplifiers

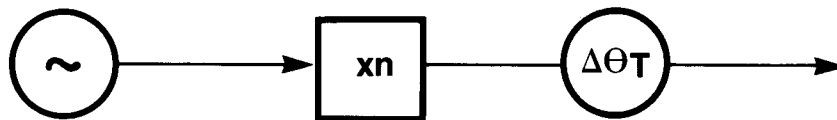


Example:

$P_{sav1} = +16 \text{ dBm}$
 $F_1 = 5 \text{ dB}$

$P_{sav2} = 0 \text{ dBm}$
 $F_2 = 5 \text{ dB}$

Equivalent Circuit



Equivalent phase noise generator

$$S_{\Delta\theta_T} = S_{\Delta\theta_1} + S_{\Delta\theta_2} = n^2 \frac{F_1 kT}{P_{sav1}} + \frac{F_2 kT}{P_{sav2}}$$

Example:

$S_{\Delta\theta_1} = -174 \text{ dBm} + 5 \text{ dB} - 16 \text{ dBm} + 20 \text{ dB} = -165 \text{ dB}$

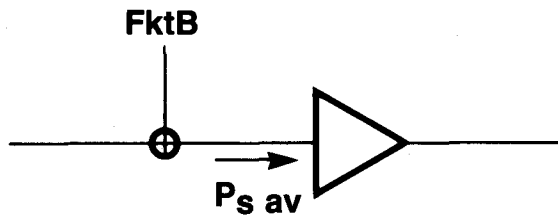
$S_{\Delta\theta_2} = -174 \text{ dBm} + 5 \text{ dB} - 0 \text{ dBm} = -169 \text{ dB}$

$S_{\Delta\theta_T} = -163.5 \text{ dB}$

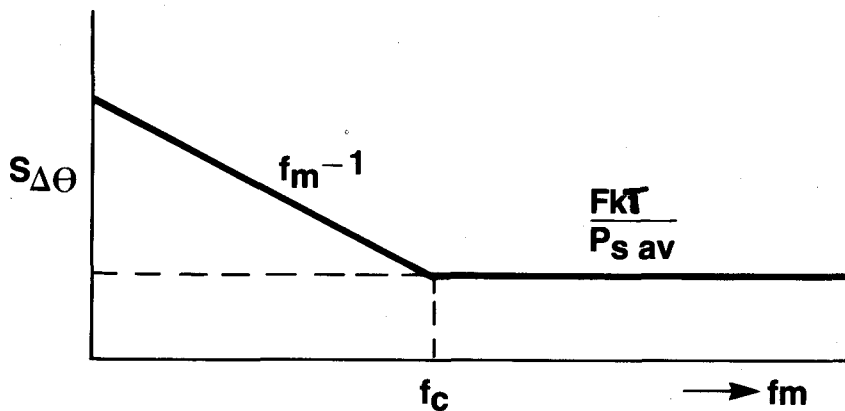
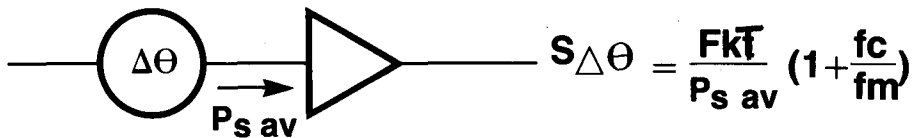
$\mathcal{L}_T = \frac{1}{2} S_{\Delta\theta_T} = -166.5 \text{ dBc}$

Obviously, high signal levels and low noise figure amplifiers on both ends of the multiplier minimize the effect of additive noise.

5.2 Phase Noise Caused by Multiplicative ($\frac{1}{f}$) Noise in Amplifiers

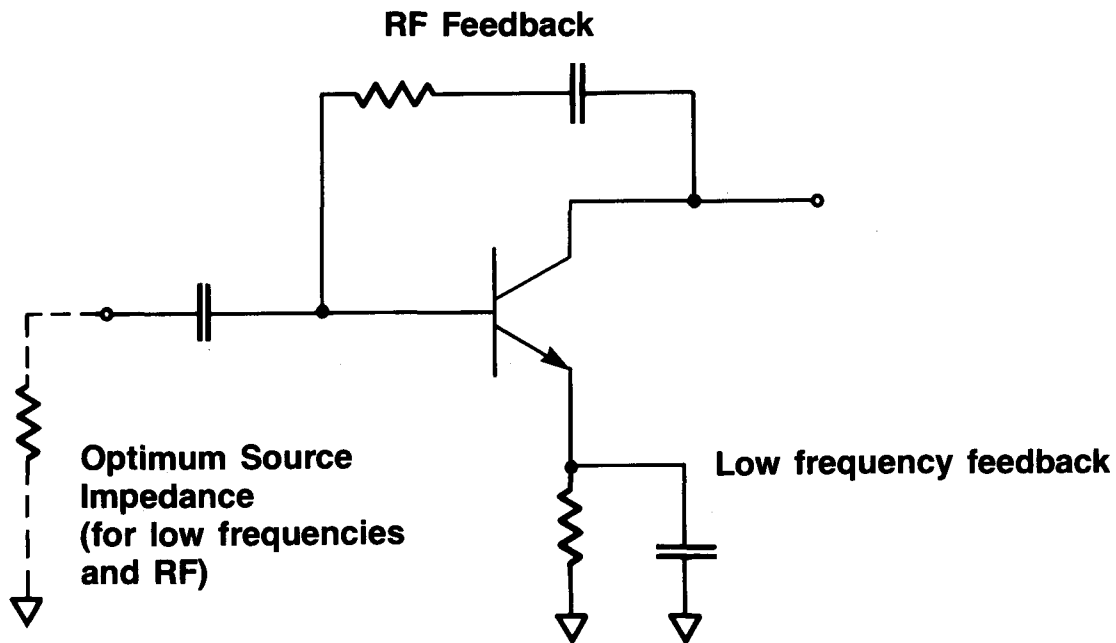


Empirical characterization of phase noise caused by ($\frac{1}{f}$) noise:



The corner frequency, f_c , is very device dependent. It can range from 1 kHz (low frequency bipolar transistor) to 100 MHz (GaAs FET). The $\frac{1}{f}$ type of increase in phase noise close to the carrier is caused by low frequency device noise modulating the phase of the passing RF signal by modulating the transconductance and the input and output impedances of the amplifier.

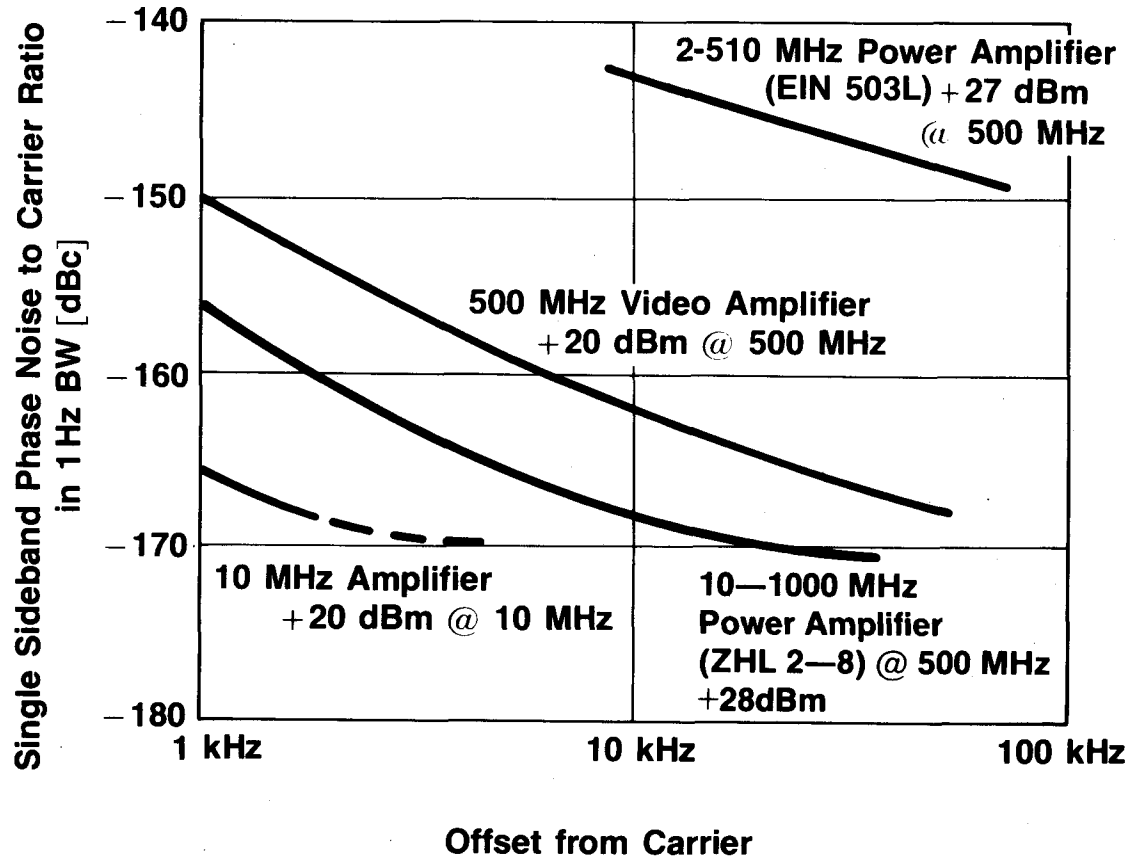
5.3 Reduction of $\frac{1}{f}$ Related Phase Noise in Amplifiers



The effect of $\frac{1}{f}$ noise can be reduced by

- negative feedback at low frequencies (bias stabilization)
- negative feedback for the RF signal
- designing the RF amplifier for low noise figure also at low frequencies (optimum source impedance)
- choosing a device with low $\frac{1}{f}$ noise
- choosing a device with good linearity

5.4 Some Phase Noise Measurements on Power Amplifiers



The above examples do not represent performance limits. Only the 10 MHz amplifier is a design where consideration was given to low phase noise performance by the choice of the active device (2N5943) and by RF and low frequency feedback.

5.5 Methods of Frequency Multiplication

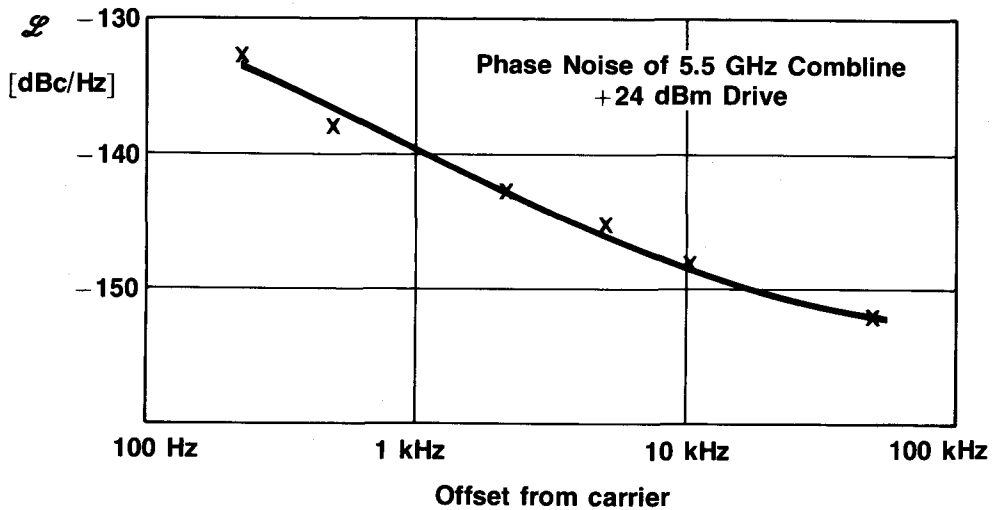
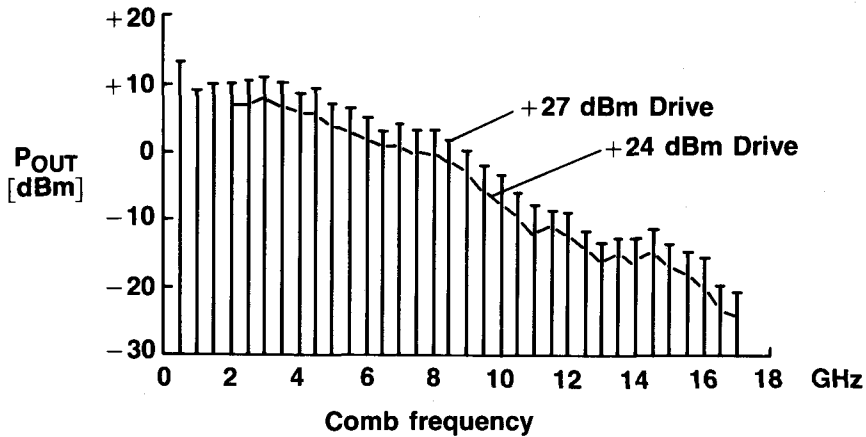
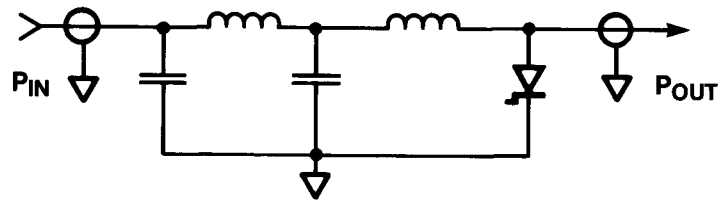
Basic Methods of Harmonic Generation	Nonlinear Device	Example	* \mathcal{L} [dBc/Hz] @ 1 kHz Offset
Nonlinear Reactance: Varactor Multiplier	Varactor		
Comb Generation through Snap Action	Step Recovery Diode (Snap diode)	500 MHz \times 11 (11th Harmonic selected)	-160
Nonlinear Conductance:			
Diode Multiplier	Schottky Barrier Diode	10 MHz \times 2 Schottky diode Doubler	< -169
Transistor Multiplier <i>See ref. 1</i>	Bipolar Transistor FET Dual gate FET (Si and GaAs)	1 MHz \times 35 Transistor (current mode switch)	-172

*Referred to input signal.

Frequency multipliers can be grouped according to the two basic principles of operation: harmonic generation by nonlinear reactance (capacitance) and harmonic generation by nonlinear conductance.

Multipliers with nonlinear capacitance typically resemble more efficient, simpler circuits. But they generate significantly more phase noise than multipliers based on nonlinear conductance, like a Schottky barrier diode doubler. Higher AM to PM conversion efficiency as well as charge fluctuations due to fluctuations of minority carrier lifetimes are speculated to cause this difference.

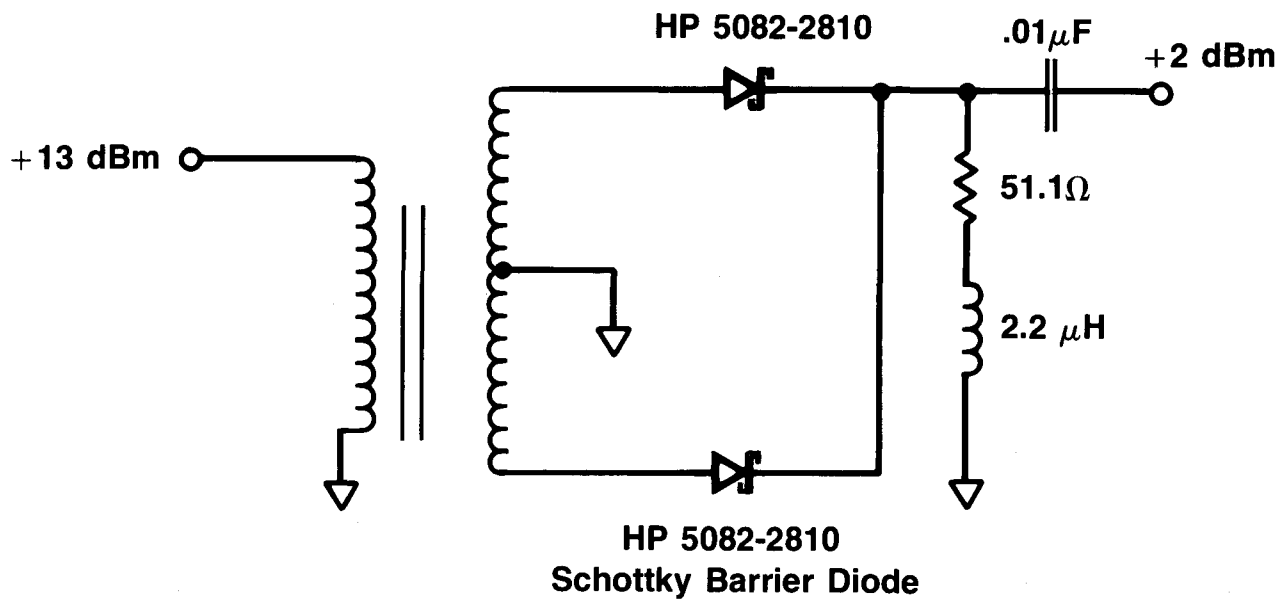
5.6 Example: 500 MHz Step Recovery Diode Pulse Generator



Phase noise measured on the above step recovery diode pulse generator (HP 33004A) proved to be quite sensitive to the drive level. Optimum results were obtained at +24 dBm. More phase noise is generated with increased drive levels, especially as the pulse peak starts to drive the diode into breakdown, and with decreased drive levels where the comb line signal becomes low enough to be degraded by additive noise.

Although the impedance matching network at the input of the multiplier circuit is optimum at 500 MHz, frequency deviations of $\pm 10\%$ can be tolerated with only slightly reduced conversion performance.

5.7 Example: 10 MHz Schottky Barrier Diode Doubler

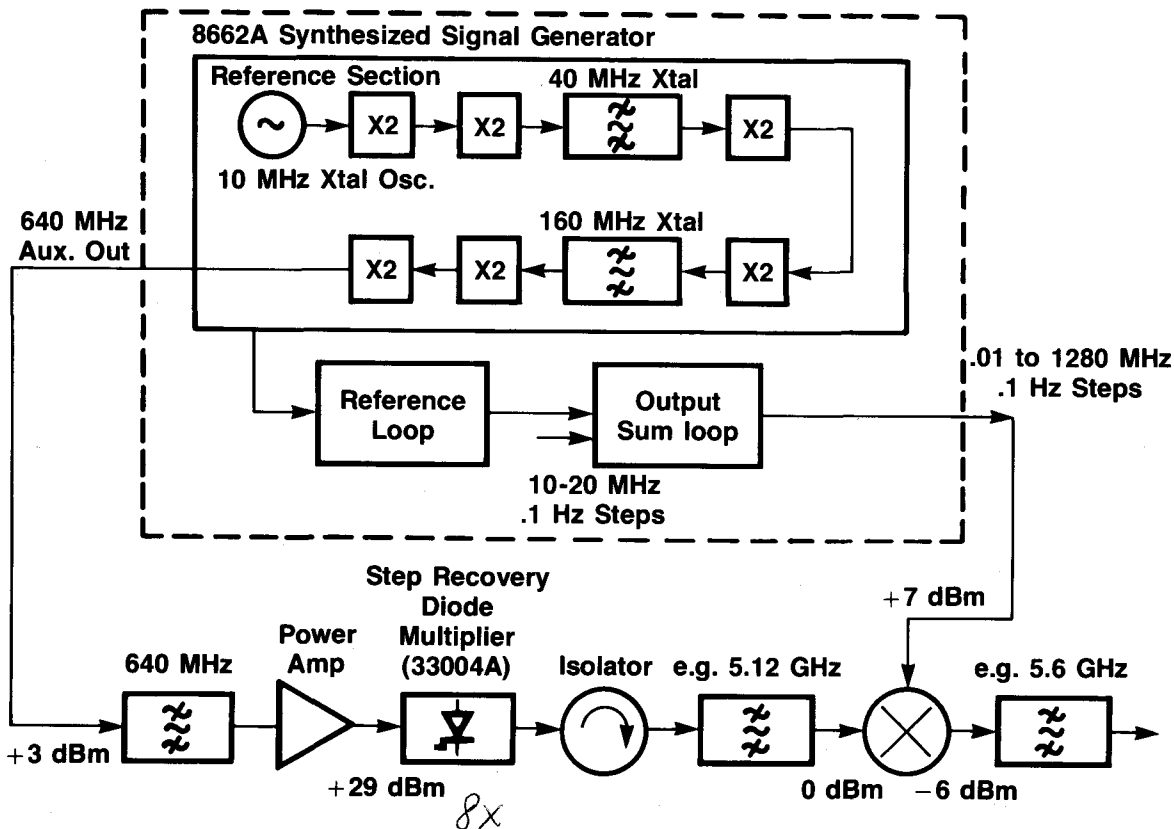


@ 20 MHz

\mathcal{L} @ 1 kHz Offset < -163 dBc

The low flicker noise and the virtual absence of minority carrier storage make the Schottky barrier diode a good choice as a nonlinear component for low phase noise frequency multiplication.

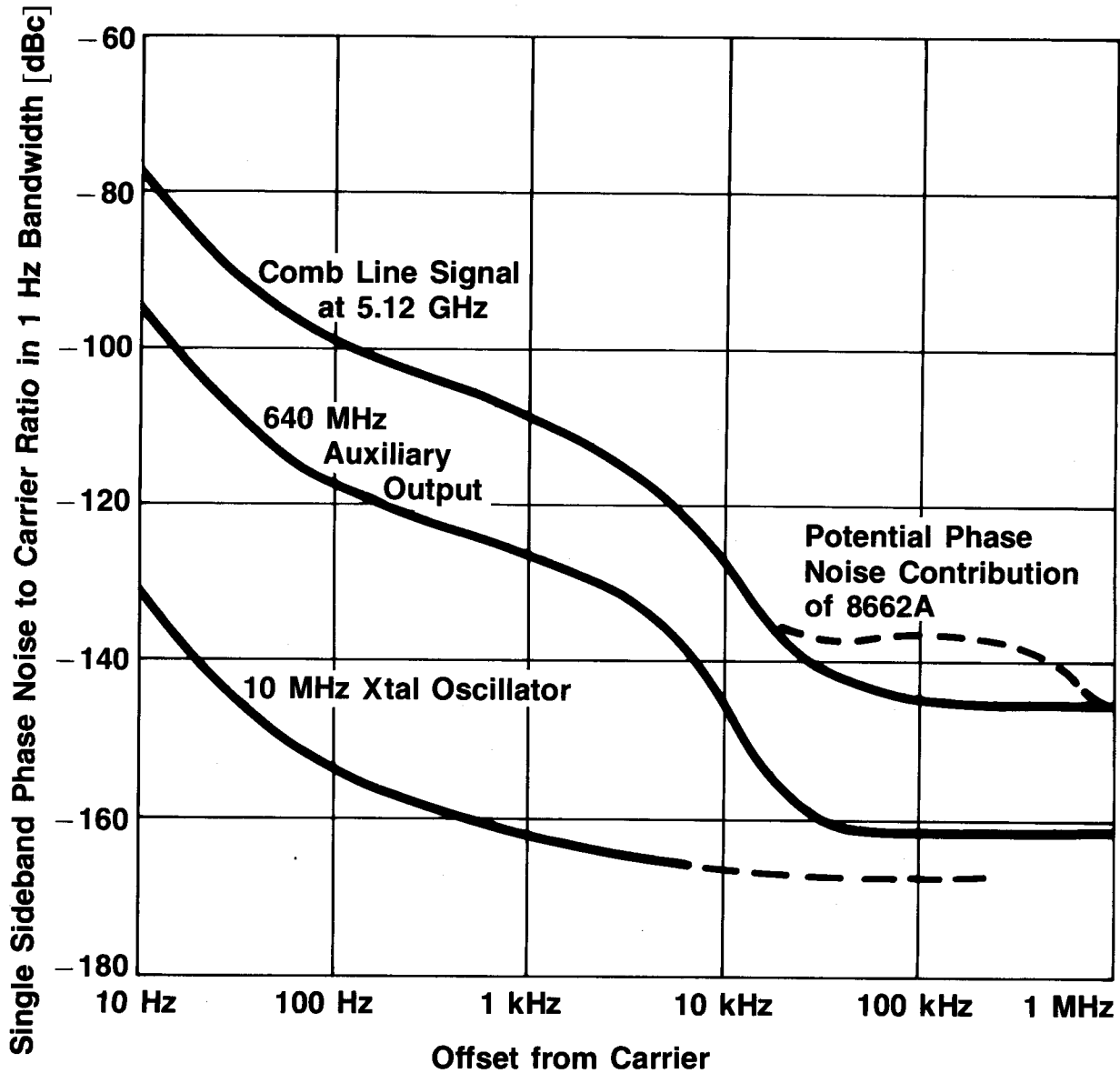
5.8 Example of a Low Phase Noise Multiplication Scheme for Microwave Signal Generation



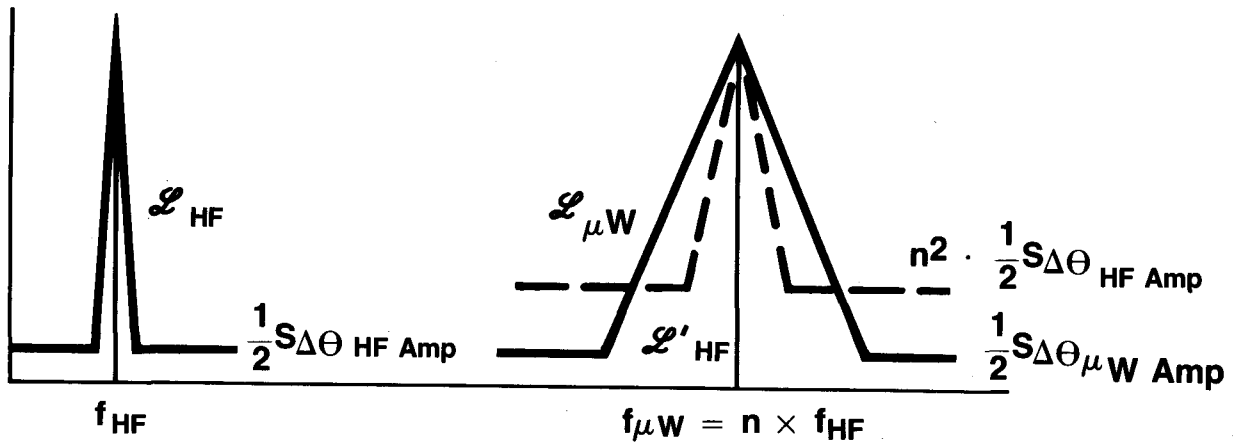
The 8662A Synthesized Signal Generator is used here to generate a very stable and quiet microwave signal. The reference section of this low phase noise RF synthesizer doubles the 10 MHz Xtal signal six times. Monolithic crystal filters are employed at 40 and 160 MHz to reduce sideband noise. The auxiliary output signal at 640 MHz drives a step recovery diode multiplier generating 640 MHz comb lines useful up to 18 GHz.

To obtain a continuous synthesized microwave signal (500 to 1000 MHz bandwidth), one comb line is singled out by a bandpass filter and mixed with the standard .01 to 1280 MHz output of the 8662A.

Phase Noise of a Synthesized Microwave Signal Using the 8662A Synthesized Signal Generator



6. Principal Comparison Between Frequency Multiplication of HF-Source and Fundamental Microwave Oscillator



	Phase Noise of HF-Source, multiplied @ $f_{\mu W}$	Phase Noise of Microwave Oscillator
Close-in phase noise	$\frac{1}{2} \cdot \frac{1}{f_m^2} \left(\frac{f_{\mu W}}{2Q_{HF}} \right)^2 S_{\Delta\theta_{HF}}$	$\frac{1}{2} \cdot \frac{1}{f_m^2} \left(\frac{f_{\mu W}}{2Q_{\mu W}} \right)^2 S_{\Delta\theta_{\mu W}}$
Noise floor	$n^2 \cdot \frac{1}{2} S_{\Delta\theta_{HF \text{ Amp}}}$	$\frac{1}{2} S_{\Delta\theta_{\mu W \text{ Amp}}}$

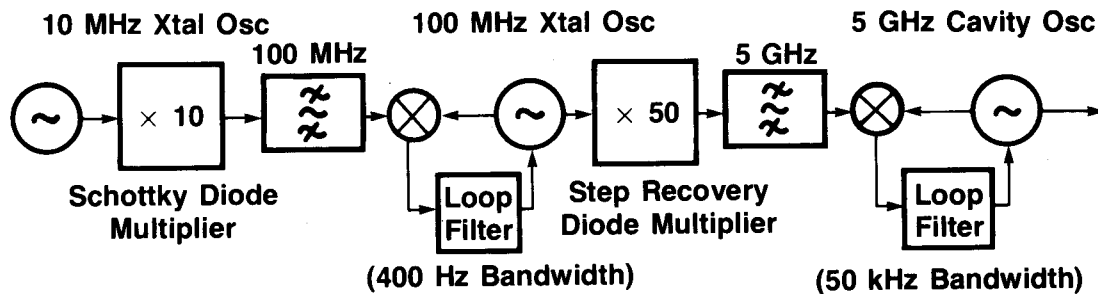
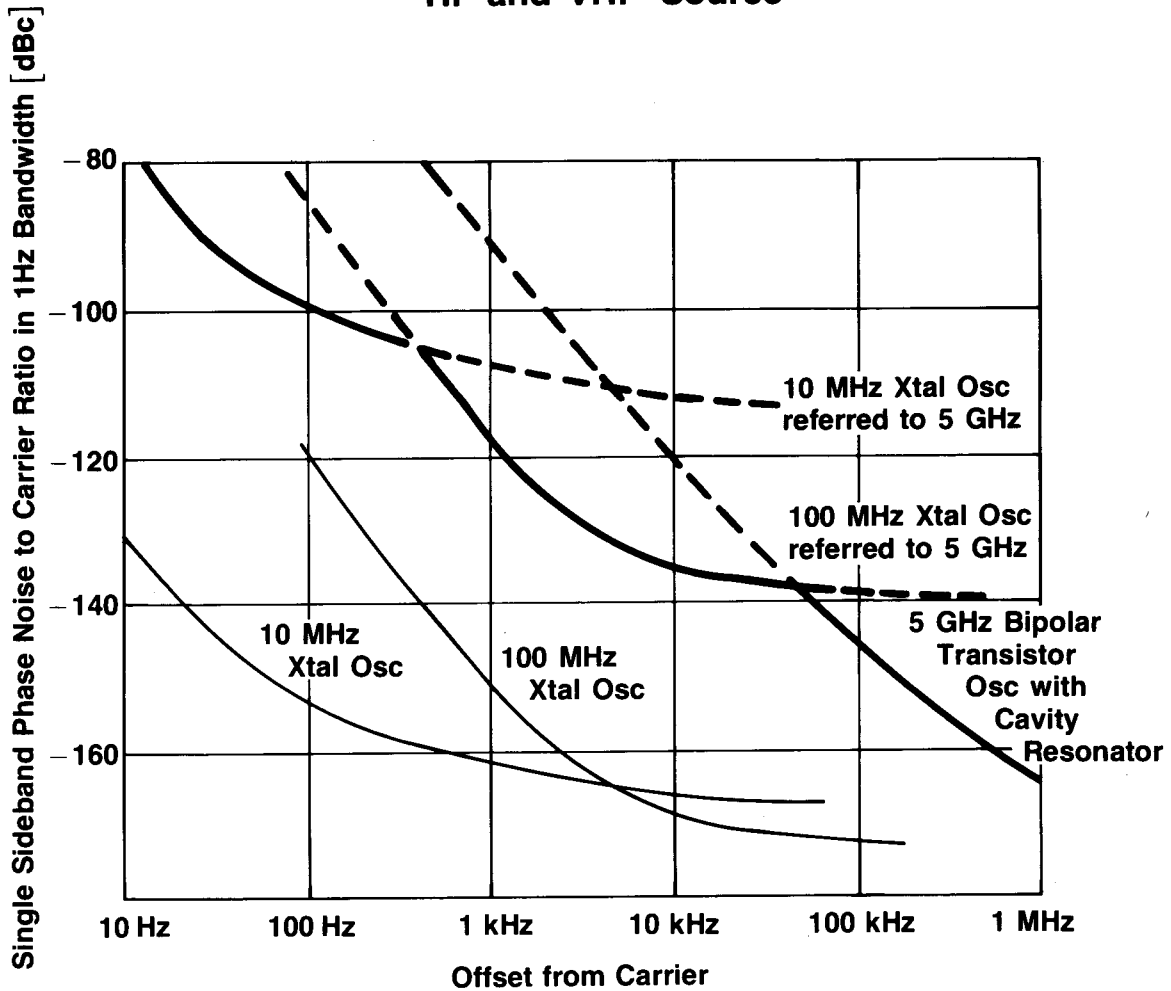
Close-in phase noise ratio
$$\frac{\mathcal{L}'_{HF}}{\mathcal{L}_{\mu W}} = \left(\frac{Q_{\mu W}}{Q_{HF}} \right)^2 \cdot \frac{S_{\Delta\theta_{HF}}}{S_{\Delta\theta_{\mu W}}}$$

Noise floor ratio
$$\frac{\mathcal{L}'_{HF}}{\mathcal{L}_{\mu W}} = n^2 \cdot \frac{S_{\Delta\theta_{HF \text{ AMP}}}}{S_{\Delta\theta_{\mu W \text{ AMP}}}}$$

With identical Q at f_{HF} and $f_{\mu W}$ and neglecting device differences, the multiplied HF source and the microwave oscillator would have comparable close-in noise. However, resonators with much higher Q are available at HF (e.g., 2×10^6 for 10 MHz Xtal) and more likely the active device of the HF oscillator causes less phase perturbation due to lower flicker noise. The multiplied HF source therefore will have lower close-in phase noise than the fundamental microwave oscillator.

The noise floor of each source is established by additive noise in the amplifier following the oscillator and can be assumed comparable. Since the noise floor of the HF source is additionally multiplied by n , the fundamental microwave source yields a lower noise floor.

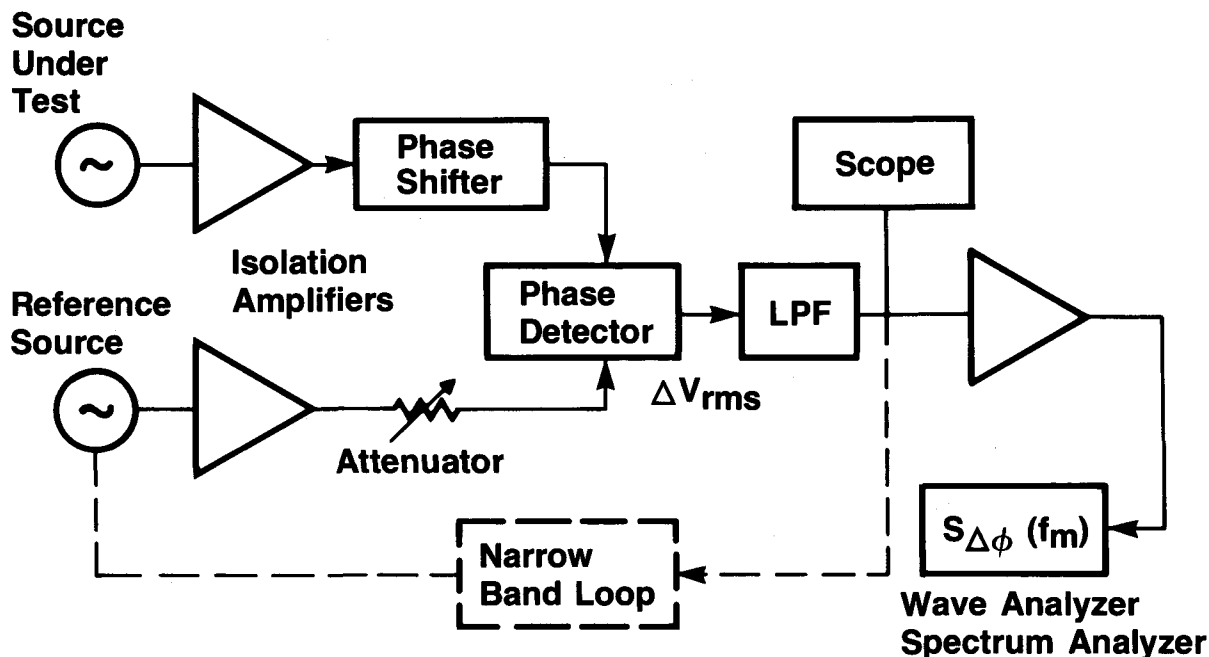
6.1 Microwave Oscillator Phase-Locked to HF and VHF Source



The above system is a paper design of a low phase noise S-band source. It combines the optimum segments of phase noise sidebands of 3 sources, a 10 MHz crystal oscillator, a 100 MHz (7th overtone) crystal oscillator, and a 5 GHz bipolar transistor oscillator with a cavity resonator.

The multipliers linking the oscillators are chosen such that their phase noise contribution is negligible. This requires a Schottky barrier diode multiplier in the first link and an efficient step recovery diode multiplier in the second link.

7.1 Phase Noise Measurement with 2 Sources and Phase Detector



$$\Delta\phi_{\text{rms}} = \frac{1}{k_{\phi}} \Delta V_{\text{rms}}$$

k_{ϕ} = Phase detector constant
 = V_B peak for sinusoidal beat signal

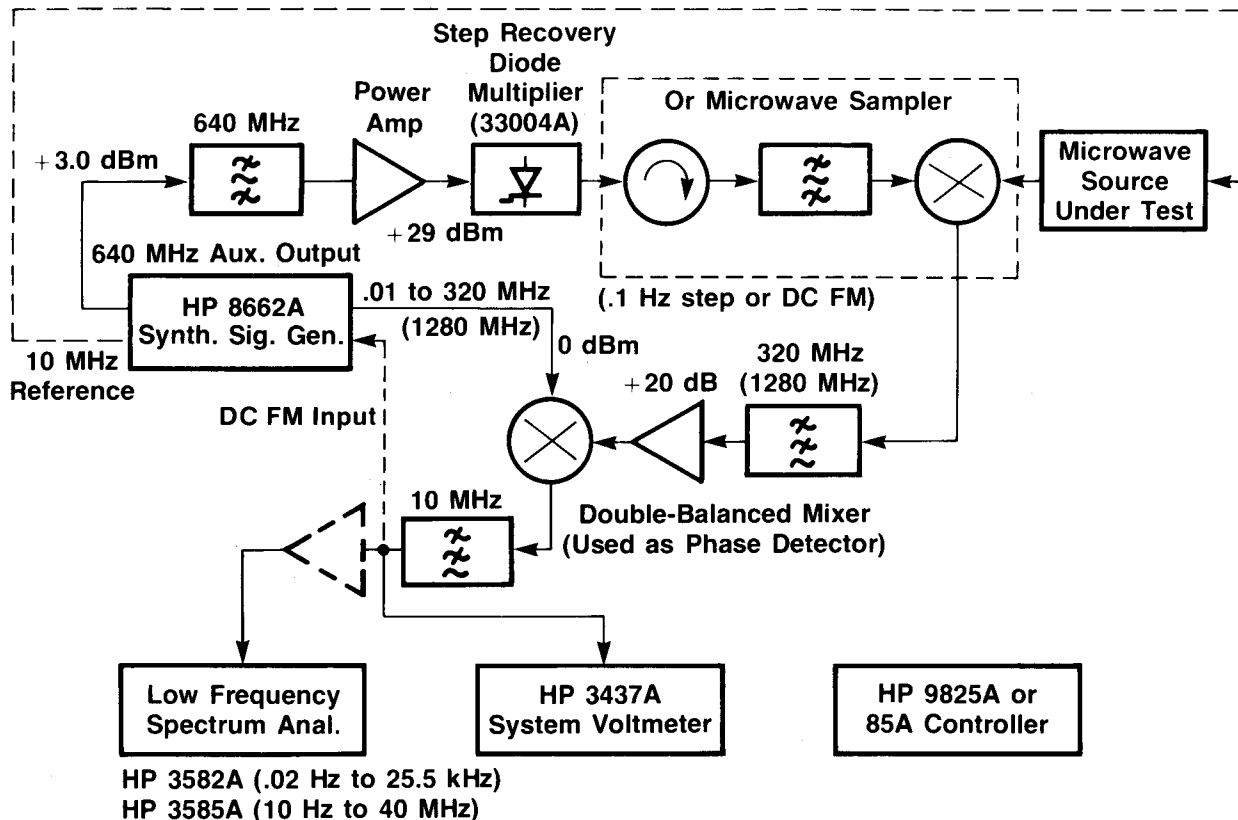
for double-balanced mixer in linear operation

$$\mathcal{L}(f_m) = \frac{1}{2} S_{\Delta\phi}(f_m) = \frac{1}{4} \frac{(\Delta V_{\text{rms}})^2 (1 \text{ Hz})}{V_B^2_{\text{rms}}}$$

A double-balanced mixer is used here as a phase detector. It requires both signals to be in phase quadrature steady statewise. Most sources would drift out of quadrature during the period of measurement and therefore have to be phaselocked in a narrow band phaselock loop. The amplifiers should prevent injection locking.

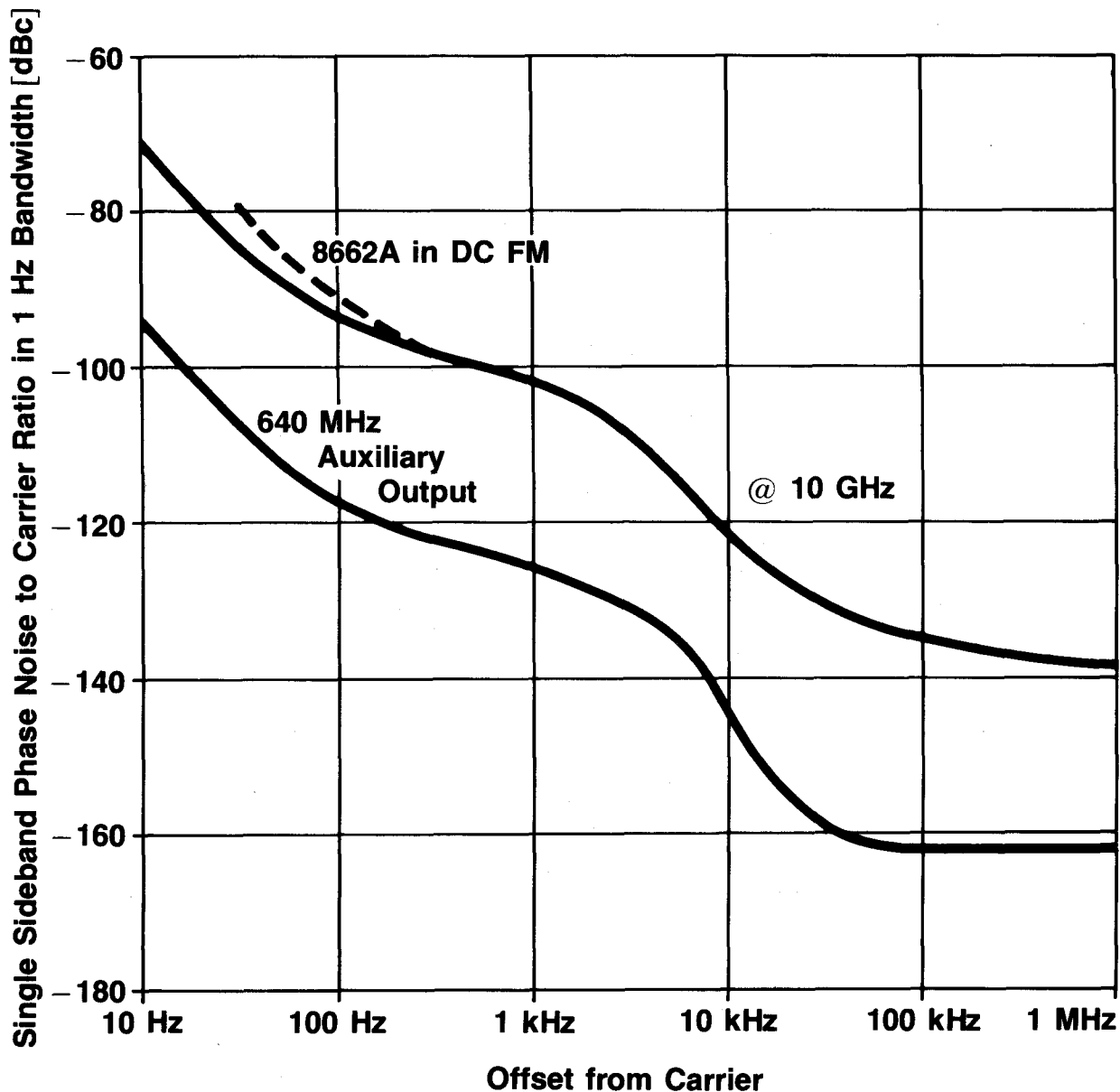
A noise floor as low as -172 dBc at 1 kHz offset has been measured with this setup using a high level mixer and a low noise post amplifier.

Automatic Phase Noise Test System for Microwave Sources



The HP 8662A provides both a very quiet 640 MHz auxiliary reference output to down convert the microwave source under test and a .01 to 1280 MHz RF signal for phase comparison with the IF of the microwave down converter. The microwave filter and mixer can be replaced by a broadband microwave sampler. The 8662A performs automatic calibration by a programmed frequency and level offset. Quadrature setting on the mixer-phase detector is controlled by probing the beat signal with a programmable voltmeter and stopping the phase advance of the 8662A when the beat signal voltage is sufficiently close to 0. This assumes that the 8662A and the source under test are driven by the same reference oscillator. With no common reference, the 8662A can be phaselocked in a narrow bandwidth to the microwave source (e.g., free-running oscillator) via its DC FM port.

System Noise Floor of Automatic Test Set at 10 GHz



GLOSSARY OF SYMBOLS

B	Bandwidth
FKTB	Available noise power in bandwidth B
f₀	Carrier frequency
f_c	Corner frequency of flicker noise
f_m	Fourier frequency (sideband-, offset-, modulation, baseband-frequency)
f(t)	Instantaneous frequency
Δf(t)	Instantaneous frequency fluctuation
$\mathcal{L}(f_m)$	Ratio of single sideband phase noise to total signal power in a 1 Hz bandwidth f_m Hertz from the carrier.
P_s	Signal power
P_{ssb}	Power of single sideband
P_{s av}	Available signal power
Q_{load}	Quality factor of loaded resonator
S_{Δf}(f_m)	Spectral density of frequency fluctuations
S_{Δθ}(f_m)	Spectral density of phase perturbation
S_{Δφ}(f_m)	Spectral density of phase noise
t	Time
v(t)	Instantaneous voltage
V_s	Peak voltage of sinusoidal signal
Δθ(t)	Instantaneous fluctuation of phase perturbation
Δφ(t)	Instantaneous phase fluctuation
ω	Angular frequency

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