LOW NOISE FREQUENCY SYNTHESIS

F. L. Walls
Time and Frequency Division
National Bureau of Standards
Boulder, Colorado 80303

C. M. Felton
Felton Electronics Design
P.O. Box 4013
Boulder, Colorado 80306

Abstract

This paper reviews the various definitions of phase noise and changes in the phase noise of a signal under noiseless multiplication, division, and translation. Next the phase noise in selected non-cryogenic rf and microwave oscillators is reviewed. Using a systems approach one can synthesize a microwave signal where the close in phase noise is controlled by a low frequency crystal oscillator while the high frequency phase noise is controlled by a microwave source. This approach yields a phase noise performance that is superior to that possible with a single source. Finally the phase noise of various amplifiers, multipliers, and dividers is compared. The phase noise of dividers while generally inferior to that of the best multipliers, is often sufficient for most applications. Special note should be made that the phase noise quoted in the literature for some dividers is perhaps pessimistic and that the phase noise of the system will often be seriously affected if placed under vibrational or thermal stress.

Introduction

This paper is to reviews the various aspects which govern the synthesis of signals with low phase noise. The synthesis of such signals from the rf region to the microwave region and beyond requires both low noise local oscillators and processing techniques. We therefore briefly review the model of a signal, definitions of phase noise, and the theoretical change of phase noise under noiseless multiplication, division, and translation. Next we review the phase noise of selected non-cryogenic rf and microwave sources. In the rf region the low phase noise sources are virtually all controlled by quartz crystal devices. At present the best available quartz devices have phase noise close to the carrier (typically Fourier frequencies from about 0.01 Hz to the half bandwidth of the crystal resonators) which varies approximately as $S_\varphi(f) = K/(Q^4f^3) = \nu_0^4/(K^3f^3)$ where Q is the quality factor of the crystal, K is a constant for a given level of crystal technology, and ν_0 is the oscillation frequency. For the best quartz crystals currently available, $K = Q\nu_0$ is of order 1.2 x 10^{13}. The phase noise far from the carrier generally depends only on the oscillator technology and crystal drive.

The phase noise of microwave (and lumped circuit LC) oscillators is usually dominated by the noise in the sustaining circuit at a level determined by the quality factor of the resonator. The phase noise of selected microwave oscillators is compared to that obtained from multiplying various quartz controlled oscillators to the same frequency. It is shown that the lowest phase noise microwave reference signals have the phase noise within about 10-100 kHz of the carrier controlled by several selected quartz crystal devices while further out the noise is determined by a microwave source.

The phase noise of frequency multipliers and dividers is also explored. Both odd and even-order multipliers offer phase noise which is usually lower than that of available sources. Even-order multipliers, however, tend to be more sensitive to input amplitude and environmental effects than odd-order ones. The phase noise of several selected frequency dividers including microwave types is compared. In general the phase noise of presently available frequency dividers is inferior to that of good frequency multipliers although it is sufficient for most applications. There is a wide disparity between the phase noise quoted in the literature for emitter coupled logic families and that presented here. Special attention must be paid to the environment under which the system has to operate. Vibration, thermal stress, or electrical noise can easily lead to seriously degraded performance.

Signal Model

The signal can be expressed as

$$V(t) = (V_0 + \epsilon(t)) \sin[2\pi\nu_0 t + \phi(t)], \qquad (1)$$

where V_0 is the nominal peak output voltage, and ν_0 is the nominal frequency of the signal. The time variations of amplitude have been incorporated into $\varepsilon(t)$ and the time variations of the actual frequency, $\nu(t)$, have been incorporated into $\varphi(t)$. (Complex waveforms such as a square wave can be expressed as the sum of several terms such as given in equation 1.) The actual frequency can now be expressed as

$$\nu(t) = \nu_0 + d\,\frac{[\phi(t)]}{2\pi dt}\,. \qquad (2)$$

The fractional frequency deviation is defined as

$$y(t) = \frac{\nu(t)-\nu_0}{\nu_0} = \frac{d[\phi(t)]}{2\pi\nu_0 dt} \quad (3)$$

Power spectral analysis of the output signal V(t) combines the power in the carrier ν_0 with the power in $\varphi(t)$ and $\varepsilon(t)$ and therefore is not a good method to characterize $\varepsilon(t)$ or $\varphi(t)$. Since in many precise signals, understanding the variations in $\varphi(t)$ or y(t) is of primary importance, we will confine the following discussion to frequency domain measures of y(t), neglecting $\varepsilon(t)$ except in cases where it sets limits on the measurements of y(t).

Definitions of Phase Noise

Spectral (Fourier) analysis of y(t) is often expressed in terms of $S_\varphi(f)$, the spectral density of phase fluctuations in units of rad^2/Hz of measurement bandwidth centered at Fourier frequency f from the carrier ν_0. Intuitively $S_\varphi(f)$ can be understood as the mean squared phase deviation in a measurement bandwidth BW centered at Fourier frequency f from the carrier as shown in equation 4.

$$S_\varphi(f) = \varphi^2(f)/BW \quad rad^2/Hz \quad (4)$$

In practice the measurement bandwidth must be small compared to f especially when $S_\varphi(f)$ is changing rapidly with f. $S_\varphi(f)$ is uniquely related to the spectral density of fractional frequency fluctuations, $S_y(f)$, as

$$S_\varphi(f) = \nu^2/f^2 \; S_y(f). \quad (5)$$

It should be noted that these are single-sided spectral density measures containing the phase or frequency fluctuations from both sides of the carrier. Other measures encountered are $\mathcal{L}(f)$, dBc/Hz, and $S_{\Delta\nu}(f)$. These are related by [1]:

$$S_{\Delta\nu} = \nu_0^2 \, S_y(f) \quad Hz^2/Hz$$

$$\mathcal{L}(f) == \tfrac{1}{2}S_\varphi(f) \qquad f_1 < |f| < \infty \quad (6)$$

$$\text{for} \int_{f_1}^{\infty} S_\varphi(f)\, df << 1 \; rad^2$$

$$dBc/Hz = 10 \log \mathcal{L}(f).$$

$\mathcal{L}(f)$ and dBc/Hz are single sideband measures of phase noise which are not defined for large phase excursion and are therefore measurement system dependent. A more accurate specification of single sideband phase noise would be 1/2 $S_\varphi(f)$, which is always well defined [1].

When the frequency of a signal is changed by a fractional amount N using any combination of perfect multipliers or dividers, $S_\varphi(f)$ is changed by N^2. The multiplication or division process can be modeled as phase amplification or attenuation. Therefore scaling the frequency by N also scales the phase fluctuation by N as shown in equation 7, where $S_\varphi(f,\nu_2)$ is the phase noise of the signal at carrier frequency ν_2, and $S_\varphi(f,\nu_1)$ is the original phase noise at carrier frequency ν_1. This is in contrast to heterodyning the signal against a reference frequency (frequency translation), where the output signal contains the phase noise of both sources as shown in equation 8.

$$S_\phi(f,\nu_2) = \left(\frac{\nu_2}{\nu_1}\right)^2 S_\phi(f,\nu_1) + S_\phi^N(f) \quad \text{mult/div} \quad (7)$$

$$S_\varphi(f,\nu_2) = N^2 S_\varphi(f,\nu_1) + M^2 S_\varphi(f,\nu_{ref}) + S_\varphi^N \quad \text{translation}, \quad (8)$$

where $\nu_2 = N\nu_1 + M\nu_{ref}$, $S_\varphi(f,\nu_{ref})$ is the phase noise of the reference signal, and $S_\varphi^N(f)$ is the equivalent extra noise added by the synthesis electronics. The linewidth of the signal can change significantly due to the frequency the synthesis process. The linewidth of the signal, $2f_1$, can be roughly defined as[2]

$$<\phi^2> = \int_{f_1}^{\infty} S_\phi(f)df = \frac{1}{\sqrt{2}} \, rad^2 \, . \quad (9)$$

The fractional power of the signal within a Fourier frequency f_c of the carrier is given by [2]

$$P_c = e^{-\phi^2_{f_c}} \, . \quad (10)$$

An analysis of equations 9 and 10 shows that the linewidth of a signal grows relatively slowly as the frequency is multiplied until the phase modulation due the broadband noise (noise pedestal) approaches 1 rad^2, at which point the linewidth can increase many decades for small changes in N. The reverse is true for division [2]. The frequency at which this abrupt change in linewidth occurs is sometimes referred to the "collapse frequency."

Phase Noise of Available Sources

In order to assess the limits of achievable frequency synthesis it is helpful to review the phase noise of nominally available sources. The most common sources used for precision frequency synthesis are controlled by quartz acoustic devices. Figure 1 shows the

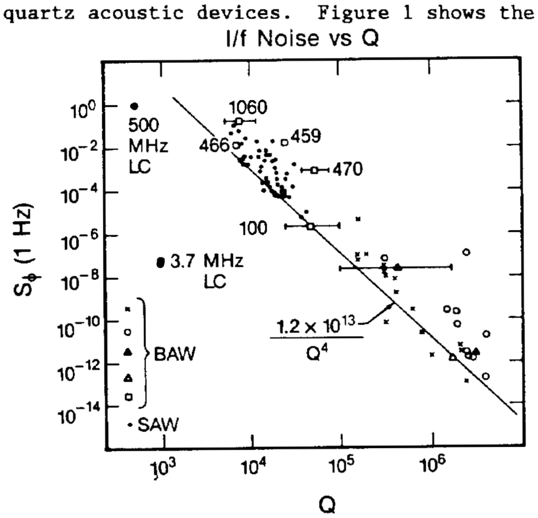

Figure 1. Flicker of frequency levels of a number of quartz resonator controlled oscillators as a function of unloaded Q factor. Adapted from [3-5].

nominally best achieved phase noise 1 Hz from the carrier for such devices measured at the oscillation frequency [3,4]. This Figure includes both bulk wave devices (BAW) and surface acoustic wave devices (SAW). Measurements on many quartz devices show that $S_\varphi(f) \sim K/(Q^4 f^3)$ where K depends on the acoustic losses within the quartz and not on the noise in the electronics. Q is the unloaded quality factor. The value of $S_\varphi(1\ Hz)$ is characteristic of the flicker frequency level achieved in these oscillators and not the broadband phase noise which is nearly independent of frequency. The flicker performance typically dominates other effects for Fourier frequencies of approximately 0.01 Hz to the half bandwidth of the resonator, $\nu_0/2Q$. At lower Fourier frequencies, drift and temperature effects often dominate. For the best quartz resonators $Q\nu_0 \sim K = 1.2 \times 10^{13}$. The result of this is that close-in phase noise of the best quartz devices scales as $S_\varphi(f) = \nu_0^4/(K^3 f^3)$. This is in contrast to LC oscillators where the phase noise is due to the phase noise in the electronics [3,5]. See Figure 1.

Figure 2 shows a more complete description of the phase noise in three different oscillators under

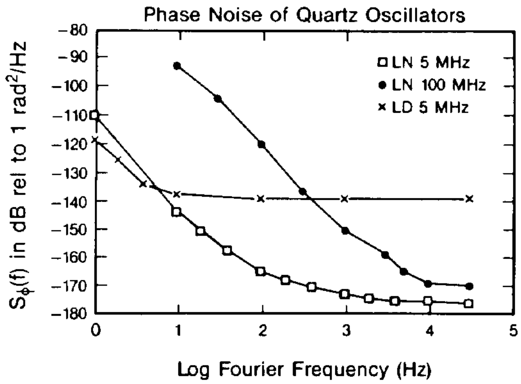

Figure 2. Phase noise of three selected BAW controlled oscillators as a function of Fourier frequency offset from the carrier.

benign laboratory conditions. Near the carrier the phase noise is usually dominated by flicker frequency at a level given by figure 1. At larger Fourier frequencies, the phase noise is determined by the added noise of the electronics and the signal level in the resonator. Figure 3 shows the phase noise of the oscillators of figure 2 multiplied to X-band, assuming a perfect multiplier chain. Figure 3 shows that a composite oscillator system would provide better phase noise over a wide region of Fourier frequencies than a single oscillator. This could be accomplished by phase locking the LN 100 MHz oscillator to the LN 5 MHz oscillator at a unity bandwidth of approximately 400 Hz which is in turn locked to the LD 5 MHz oscillator at a bandwidth of approximately 10 Hz [6]. The oscillators chosen for this example are only used to illustrate the system approach to providing a reference frequency and may not be optimum for all situations. Specifically, SAW oscillators at several hundred MHz have very low phase noise and could be used to reduce the wideband phase noise even further [3,7]. One could also consider using a quartz post filter to reduce the wideband phase noise.

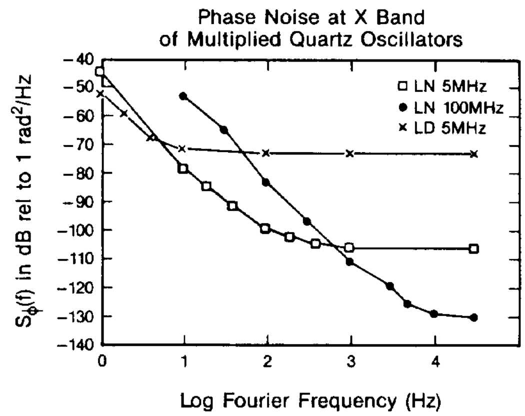

Figure 3. Phase noise of the oscillators displayed in Figure 2 multiplied to 10 GHz.

Under field conditions the performance is often compromised by orders of magnitude due to the vibration sensitivity which for quartz devices is typically of order $dy = 2 \times 10^{-9}/g$ acceleration. (For a review see references 8 and 9.) If the vibrations are severe enough, the power in the carrier can be lost as described above [2]. Microwave oscillators also have a relatively high vibration sensitivity which in some cases is worse than that of quartz oscillators. Recently, selected units showing something of an order of magnitude improvement have been tested. Also it has been possible to reduce this sensitivity for Fourier frequencies up to several hundred Hz using compensation techniques [8]. This adds considerable complexity to the oscillator. Temperature variations generally affect the long-term frequency of the oscillator and not the phase noise for frequencies above about 0.1 Hz.

Figure 4 shows a comparison of the phase noise from an optimum combination of the multiplied quartz oscillators shown in figures 2 and 3 along with several microwave oscillators at X-band [10]. Below Fourier frequencies of some tens of kHz, the multiplied quartz oscillators generally have the lowest phase noise and are often used to phase lock microwave sources. With modern solid state amplifiers it is possible to raise the power level of these sources to at least watts without serious compromise of the phase noise. Figure 5 shows the phase noise of several GaAs FET amplifiers along with that traditionally measured in Si bipolar amplifiers up to about 1 GHz [11]. Also included is the typical phase noise of a double balanced mixer using Schottky diodes.

The wideband phase noise of the available microwave oscillators is nearly universally dominated by the added phase noise of the sustaining circuit, $S_\varphi(f,Amp)$ [3,12], and the loaded resonator quality factor, Q_L as shown in equation 11 . This relationship is easily derived from the phase shift around the loop required for oscillation, and the induced fractional change in oscillation frequency due to small phase fluctuations, $\Delta\varphi$, given in equation 12.

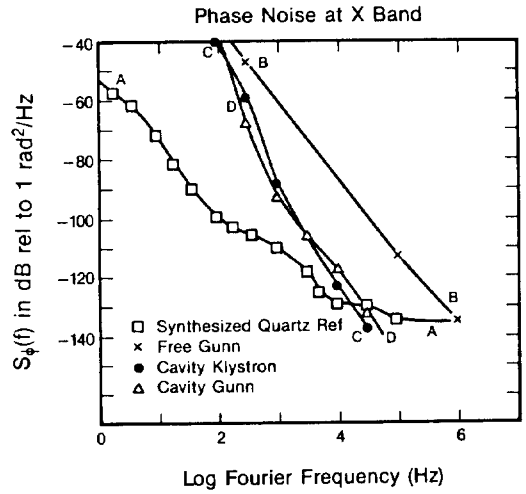

Figure 4. Curve A is the theoretical phase noise performance of a signal optimally synthesized from the multiplied quartz controlled oscillators from figure 3. Curve B is the phase noise of a free running Gunn oscillator [10], Curve C is phase noise of a klystron, injection locked to the reflected signal from a transmission cavity with an unloaded Q of 50,000 [10]. Curve D is the phase noise of a Gunn oscillator, injection locked to the reflected signal from a transmission cavity with an unloaded Q of 50,000 [10].

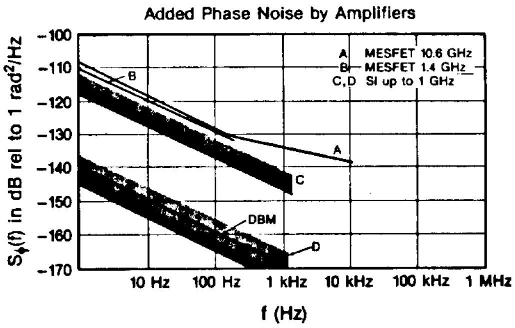

Figure 5. Curve A is the phase noise added to the signal from a MESFET amplifier operating at 10.6 GHz. Curve B is the phase noise of a MESFET amplifier operating at 1.4 GHz [11]. Curve C is the phase noise performance of a generic silicon bipolar amplifier for frequencies up to about 1 GHz with the emitter resistance capacitively bypassed. Curve D is the performance of generic silicon amplifiers with some unbypassed emitter resistance. The nominal value required is the reciprocal of the transconductance [14,16].

$$S_y = 1/(2Q_L)^2 \, S_\varphi(f,\mathrm{Amp}) \qquad (11)$$

$$\varphi = n2\pi \qquad \Delta\varphi = 2Q_L dy$$

$$n = 0,1,2\ldots \qquad (12)$$

Very close to the carrier the phase noise is often dominated by the frequency drift of the resonator. We know of no data which shows that the phase noise of quality L-C (including cavity) oscillators at all Fourier frequencies beyond a few Hz is not dominated by the sustaining circuit [3,5]. Figure 1 shows data from two different LC oscillators where the phase noise is from 25 to 86 dB less than that expected from a quartz oscillator of similar Q factor [5]. Therefore improvements in the phase noise of the sustaining circuits and or improvements in loaded quality factors would directly translate into better phase noise of the oscillators. This is in contrast to the crystal controlled oscillators where the phase noise within the half bandwidth of the resonators is usually dominated by noise in the quartz resonator [3,4].

Figure 6 shows one method of reducing the wideband phase noise of a source by using a reference cavity.

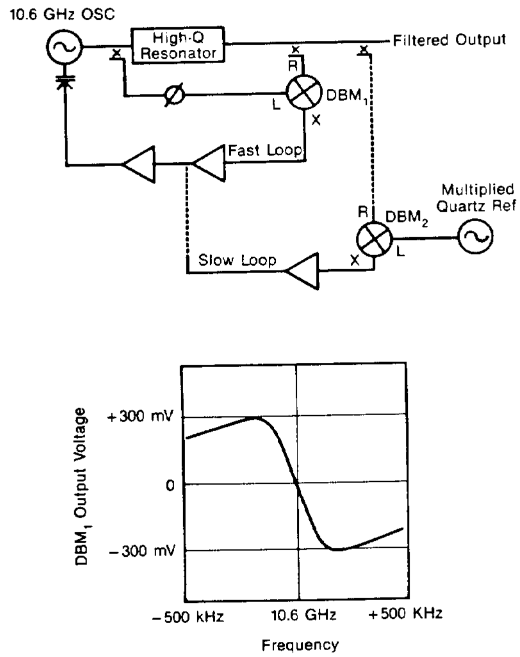

Figure 6. One possible scheme for locking an oscillator to the center of an external cavity with a fast loop in order to improve the phase noise out to Fourier frequencies of order several MHz. The low noise performance in primarily due to the availability of ultra low noise double balanced mixers. Also shown is a method of phase locking the system to a multiplied quartz reference signal. See also [13].

The key practical point here is that the phase noise of the locked system is limited by the expression in equation 11 with the phase noise now given by that of the double balanced mixer. Considerable improvement in phase noise is possible over that available from traditional microwave oscillators using a reference cavity of similar Q factor, since the phase noise of a good microwave double balanced mixer is typically some 30 dB better than that of presently available microwave amplifiers. Figure 7 shows the expected performance with a loop bandwidth of about 2 MHz. Truly extraordinary performance might be expected if the present revolution in high temperature superconductivity lead to high Q room temperature cavities [5,13].

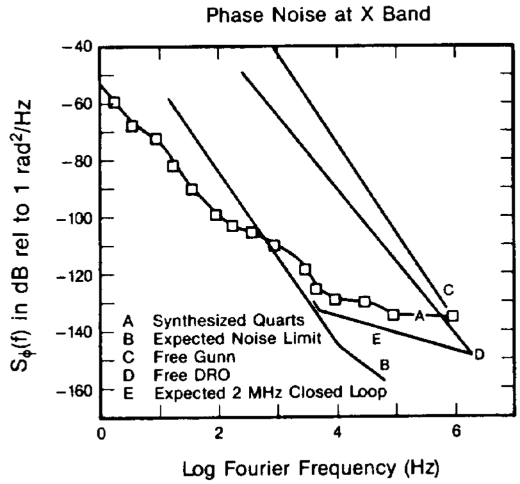

Figure 7. Curve A shows the theoretical phase noise performance of a signal optimally synthesized from the quartz oscillators of figure 3. Curve B shows the theoretical noise floor of the scheme shown in figure 6 A for a reference cavity Q factor of 20,000, a noise in DBM_1 of $S_\varphi(f) = 10^{-14}/f$, and a phase sensitivity of 0.5 V/rad. Curve C shows the generic performance of a Gunn oscillator, and curve D shows the performance of a DRO oscillator. Curve E shows the approximate phase noise expected from the 10.6 GHz oscillator locked to an external reference cavity using the scheme shown in figure 6a with a loop bandwidth of 2 MHz and the assumptions used in calculating curve B.

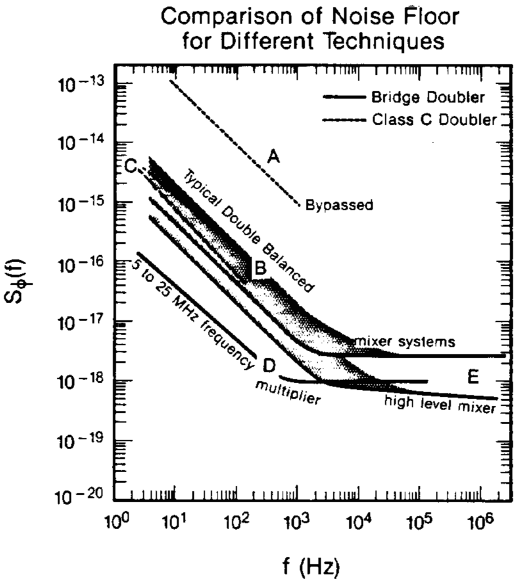

Figure 8. Curve A is the phase noise of a traditional class c doubler with capacitive bypassing of the emitter resister [14]. Curve B is phase noise of the same doubler with 34 ohms of unbypassed emitter resistance. Curve C is the phase noise of a full wave bridge doubler using Schottky diodes. Curve D is the phase noise of a 5 to 25 MHz multiplier using emitter coupled pairs [15]. For comparisons the phase noise in typical double balanced mixers is shown in curve E.

Phase Noise of Selected Multipliers, and Dividers,

The added phase noise due to frequency multiplication can be spectacularly low. Figure 8 shows the phase noise that was measured on a traditional class c doubler with and without emitter bypassing [14], a fullwave doubler using Schottky diodes, and a quintupler using emitter coupled pairs. The traditional even-order multipliers are particularly sensitive to phase variations with amplitude and environment since the zero crossings of the output are not directly associated with those of the input. In the odd-order multipliers such as the emitter coupled pair examined by Baugh [15], the zero crossings of the output are closely tied to that of the input and these devices show relatively low sensitivity to input power level and circuit parameters. In all these devices particular care must be paid to providing sufficient unbypassed emitter impedance in order to suppress the flicker phase modulation in the active junction [14,16].

The phase noise of frequency dividers and digital circuits in general is not very well documented. We have measured the noise in several families of emitter coupled logic (ECL) and have obtained numbers which are generally 6 to 20 dB better than that in the literature [17-20]. Figure 9 shows the phase noise that we measured for several families of divide by 20 circuits versus input frequency. In all cases the phase noise was referred to the output frequency. We found that special care was needed in order to cancel the phase noise of the driving source. Specifically in one divide by 10 circuit where the

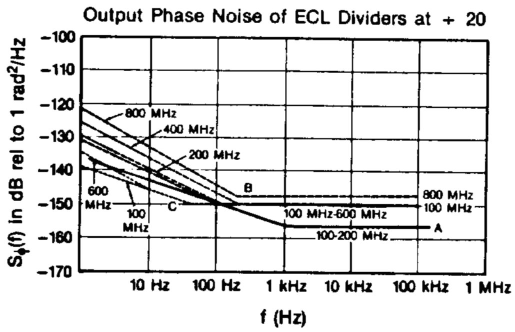

Figure 9. Shows the output phase noise which we measured for 3 different configurations of ECL logic. Curve A was obtained using the scheme shown in divider 1 for both channels of figure 10. Curve B was obtained using the scheme shown in divider 2 for both channels of figure 10. Curve C was obtained using the scheme shown in divider 3 for both channels of figure 10. In order to achieve these results it was necessary to trim the level of the low noise power supplies, to use a very low noise frequency synthesizer for the source, and to adjust the count of the dividers to obtain ~ 90° phase shift between the outputs. A low noise buffer amplifier to translate the logic levels to +10 to +13 dBm drive for the mixers was also necessary.

waveform was not symmetric, we found that the source noise dominated the measurements. Adding a divide-by-2 circuit improved the measured noise by 10 to 30 dB, which is much more than the 6 dB expected from the division process. Originally the measured phase noise was very sensitive to the bias voltage. After adding the divide-by-2 circuit the phase noise was independent of the bias voltage as long as it was within the manufacturer specifications. We used a symmetric push-pull buffer to translate the digital signal to a form appropriate to drive the double balanced mixer as shown in figure 10. These buffer amplifiers have a small signal gain of about 6 dB and typically drive a 50 ohm load at +13 dBm. The buffers are followed by 3 dB pads in order to reduce the standing wave ratio on the cables leading to the mixer. This also greatly reduces the variations in the dc output of the mixer due to changes in the signal amplitude or temperature variations. Typical mixer sensitivities were 0.3 to 1 V/rad at the zero crossing. Figure 11 compares the quoted phase noise for several different divider types available in the literature. We suspect some of the data in the literature were limited by the source noise and/or the output circuit used to drive the mixer. The phase noise of our buffers is so low as not to contribute to the noise.

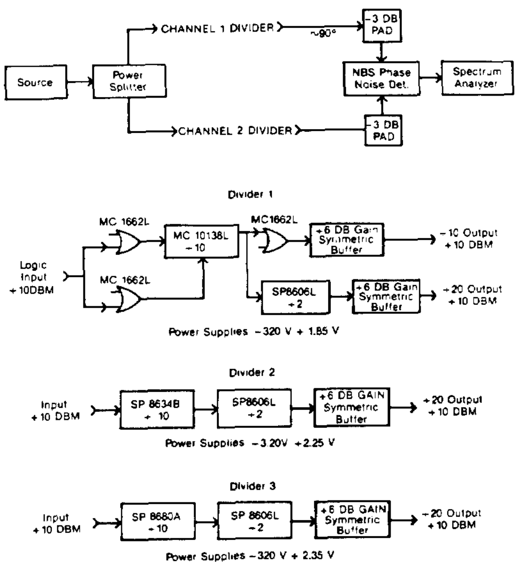

Figure 10. Shows the block diagram of the measurement set up for the phase noise measurements shown in Figure 9. (Certain commercial devices are identified in this paper in order to adequately specify the experimental procedure. Such identification does not imply recommendation or endorsement by the National Bureau of Standards, nor does it imply that the devices identified are necessarily the best available for the purpose).

Stone [21] has brought to our attention a type of regenerative divider which is self starting and should have very low noise, perhaps limited only by the available amplifiers. The general concept shown in figure 12 is similar to that used in divider circuits 20 years ago. We have not had the time to measure the phase noise in such circuits but have demonstrated that it can be used to divide by 2, 4, 6, 8, and 10 with a suitable emitter coupled pair multiplier in the feedback loop. Obviously other division factors are possible using even order multiplication within the loop. This design should operate to frequencies in excess of 40 GHz using present technology.

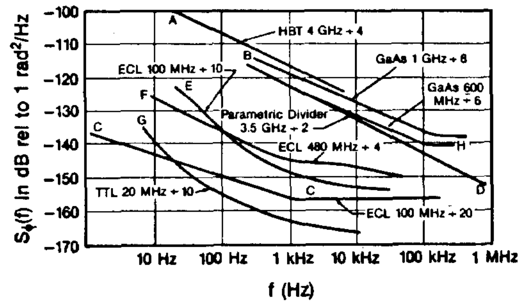

Figure 11. This figure shows the output phase noise of several types of dividers. Curve A shows the phase noise of a divider constructed using GaAs Hetrojunction bipolar transistors. Curves B and H show the output phase noise of a GaAs MESFET based divider for in input frequency of 1 GHz and 500 MHz [17]. Curve C shows the phase noise of the silicon ECL divider 1 of figure 10. Curve D shows the phase noise of a parametric divider that can operate as high as 18 GHz [20]. Curve E and F show the phase noise of two types of silicon ECL dividers from [17]. Curve G shows the phase noise in a TTL divider from [17].

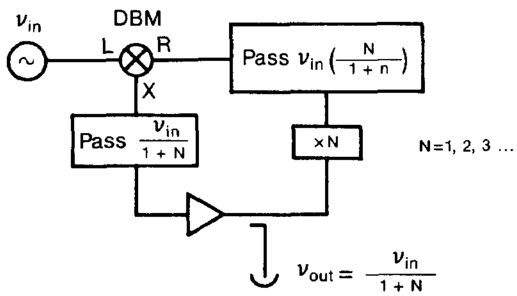

Figure 12. Block diagram of a self starting regenerative divider [21]. This scheme should operate with presently available technology to at least 40 GHz.

Conclusions

We have shown that in general a composite system is capable of synthesizing reference signals with better phase noise than that of any single source. In such a system a multiplied low frequency quartz controlled oscillator generally controls the phase noise close to the carrier while the wideband noise is determined by a higher frequency quartz oscillator and/or by a microwave oscillator. Frequency multipliers generally have lower phase noise than present dividers, although, the noise in several types of dividers is sufficient for many applications. The phase noise of quartz controlled oscillators is determined by the acoustic losses in the quartz and therefore is not likely to improve significantly. The phase noise of LC and cavity controlled oscillators is due to the phase noise of the sustaining amplifier and is likely to improve significantly. In some applications environmental effects can significantly degrade performance.

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