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MEASUREMENT OF OSCILLATOR PHASE AND AMPLITUDE NOISE AT W-BAND

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1. INTRODUCTION

This paper describes a technique used to measure oscillator noise sidebands at 94GHz. It is extendible to other frequencies in the millimetre-wave band. Oscillator noise sidebands are particularly significant for Continuous Wave (CW) radars, where the sidebands of the transmit/receive leakage can degrade the system noise figure. Measuring oscillator noise is also a particular challenge at higher millimetre-wave frequencies where components are relatively narrow-band and measuring systems relatively insensitive.

Phase noise at a few hundred kilohertz from carrier can be measured using spectrum analysers, but further from the carrier analysers are not sufficiently sensitive, and close to the carrier the noise can be hard to measure due to the high drift rate of the carrier. Amplitude modulation (AM) noise sidebands are also almost always well below the noise floor of a spectrum analyser. The phase noise, however, can to be cancelled in the final application by careful adjustment of the local oscillator path lengths, so that the, inherently much lower, AM noise levels can in fact limit system sensitivity, and the sensitivity of the test rig to AM and Frequency Modulated (FM) noise can be measured with calibration signals.

Phase noise is most sensitively measured by beating together two oscillators, and measuring the spectrum of the resulting beat signal. This is a more sensitive scheme than using the harmonic mixers conventionally used with spectrum analysers. By performing multiple measurements with multiple oscillators, the noise levels of the individual oscillators can be deduced.

2. BACKGROUND

2.1 FMCW Radars as a Good Reference Case

References 1 and 2 describe the theory of the effects of transmitter sensitivity on the noise sidebands of CW radars. One of the simplest ways to understand the ultimate significance of noise sideband is with respect to an FMCW radar.

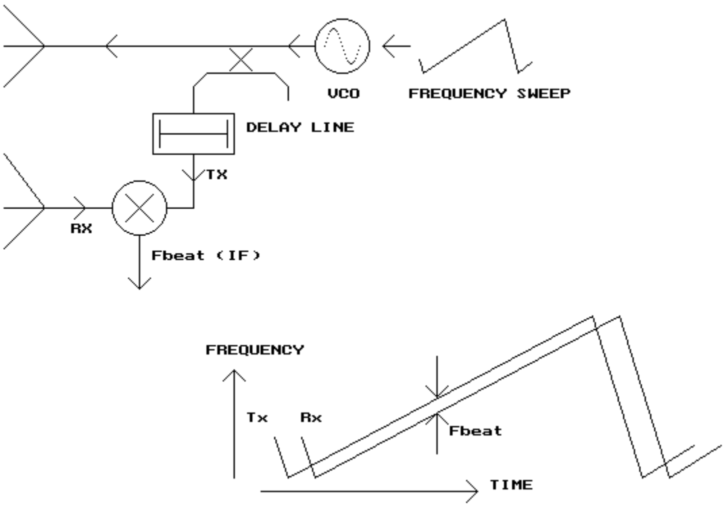

Figure 1: Outline of an FMCW Radar

Figure 1 shows the operation of a simple linear FMCW radar. A voltage controlled oscillator is modulated to produce a linear ramp of frequency against time. Part, usually most, of this signal is transmitted and another part is used to provide the local oscillator signal to the receiver mixer.

The received signal is then mixed with the sample of the transmitted signal to produce a beat frequency. It can be seen from the timing diagram that when the signal is received from a single target the output of the mixer is a beat frequency which is proportional to the time delay between the transmitted and received signals, i.e. to the range to the target.

It will be noted that a delay line[3] is shown in the path feeding the local oscillator signal to the mixer. As is implied below, this can be used to minimise the detection of FM noise on any signals leaking directly between the transmitter and the receiver.

In many systems, of course, the transmit and receive paths are connected to a common antenna through a circulator.

If there is more than one target present, then each produces its own signal at the mixer output with a frequency proportional to its range. The total spectrum of frequencies thus represents the ranges and received powers corresponding to the variety of targets which are present. It is nowadays relatively simple to analyse such a spectrum using a fast Fourier transform (FFT) processor to provide a series of signals representative of the strength of the reflectors in each range bin.

2.2 Significance of Transmitter Noise Sidebands

For a CW or FMCW radar, if the transmitter power is P_t and the attenuation of the leakage path between transmitter and receiver is L_p then the detected AM noise level in the receiver is

$$P_{am} = P_{t.}L_pL_mn_{am.}(4\cos^2\phi) \qquad (Eqn.\ 1$$

where

L_m is the conversion loss of the mixer

n_{am} is the transmitter's single sideband AM noise level w.r.t. to the transmitter carrier power,

ϕ is the relative phase between the leakage signal and the local oscillator,

and the noise is measured in 1Hz bandwidth and the noise is measured at an IF frequency which is the same as the distance from the carrier at which the AM noise is measured.

The degradation of the receiver noise figure due to the AM noise is then

$$\delta_{AM} = [P_t\, L_p n_{am}\,(4\cos^2\phi)]/(kTN) \qquad (Eqn.\ 2$$

where, as usual

k is Boltzmann's constant

T is the absolute temperature of the receiver and

N is the receiver noise figure.

It can be shown that, although the first equation included a term for the mixer loss, this latter equation is valid whether or not there is an RF preamplifier, provided that the path loss and the noise figure are both measured at the same point.

On average over all phases $\cos^2\phi = 0.5$ and, as expected

$$\delta_{AM0} = 2P_{t.}\, L_p n_{am.}/(kTN) \qquad (Eqn.\ 3$$

The corresponding equations for FM noise replace the cosine' function by a 'sine' to account for the fact that AM and FM signals are detected optimally at different (quadrature) phases of the local oscillator. The corresponding equation for the detected FM noise power is:

$$P_{fm} = P_{t.}\, L_p L_m n_{fm.}(4\sin^2\phi)\sin^2(\omega_m\delta\tau/2) \qquad (Eqn.\ 4$$

where the additional term "$\sin^2(\omega_m\delta\tau/2)$" reflects the correlation between transmitter and local oscillator noise sidebands, for which.

ω_m is the radian frequency of the IF, i.e. also the distance from carrier, at which the noise is being measured and

$\delta\tau$ is the differential time delay between the path from the master oscillator via the leakage path to the Radio Frequency input of the homodyne mixer and path from the master oscillator to the Local Oscillator port of the mixer.

This correlation can be understood by realising that 'noise' level is not affected by the absolute value of the carrier frequency, but only by changes in the frequency between the different times taken for the master oscillator signal to get to the two ports of the mixer. The degradation of the noise figure due to FM noise is thus

$$\delta_{FM0} = 2P_{t.}L_{p}/n_{fm.}\sin^2(\omega_m\delta\tau/2)/(kTN) \qquad (Eqn.\ 5$$

Similar equations also apply to the exciter noise sidebands of pulse-Doppler radars, although in such systems, usually:

a) the dynamic range requirements are less severe than for CW systems

b) there is generally also a significant, uncorrelated, noise source from the transmitter and

c) the ranges are usually such that most of the FM noise is effectively decorrelated.

Hence this problem of oscillator noise far from carrier is generally most significant for CW systems such as automotive radars, missile seekers or dedicated Low Probability of Intercept (LPI) radars.

Note that although the FM noise typically reduces at greater distances from carrier, the correlation also degrades for greater frequencies. In fact if the FM noise reduces at 20dB/decade away from the carrier, corresponding in fact to a constant frequency deviation per unit bandwidth, the detected noise floor is flat out to the frequency at which there is no longer any correlation. Note also that in the particular case where the mean detected FM and AM noise powers are the same then the total detected transmitter noise sideband power becomes independent of the relative phases of the LO and RF signals.

2.3 Importance of AM Noise Sidebands

The significance of the AM noise, and hence the need to be able to measure it, can be seen if specific numbers are substituted into equations 2, 3 and 5.

If we take a W-band FMCW radar, as this is the subject of the noise measuring system, we may assume that, in round numbers:

P_t = +10dBm

$L.$ = 20dB

$n_{am.}$ = -160dBc/Hz at 1MHz from carrier

N =-10dB

then δ_{AM0} = 0.5, i.e. the noise figure is increased by about 2dB. Note that, typically, this is significant, but not dominant.

For these systems, $\delta\tau$ would typically be of the order of 0.1ns a differential path length of about 2cm) then, at this same 1MHz from carrier, the cancellation term is

$$\sin^2(\omega_m\ \delta\tau/2) \approx 2.5 \times 10^{-8}\text{, or 76dB}$$

and

$n_{fm.}$ = -100dBc/Hz at 1MHz from carrier

so

δ_{FM0} = -23dB, or a degradation of only 0.5% in the noise figure.

Note that the oscillator FM noise has much less effect on the radar's sensitivity than does the AM noise, hence the importance of being able to measure the latter with a reasonable degree of accuracy. Note that, at least for the system analysis, a high degree of accuracy is not usually necessary, since if the oscillator noise level has more than a second-order effect on the overall noise figure the radar, the radar is probably badly designed,.

Note that the figures used above are different from those given in reference 1 since those therein were more appropriate to a microwave LPI radar.

3. MEASUREMENT TECHNIQUES

FM noise at distances between a few tens of kilohertz and a megahertz or so from the carrier can usually be measured directly on a spectrum analyser. Closer to carrier, particularly at millimetric frequencies, the instability of the carrier prevents measurements being made and further away the noise levels are generally below the analyser's noise floor. At millimetric frequencies the sensitivity issue is exacerbated by the relative insensitivity of the analysers, but made easier by the relatively high noise levels of the oscillators.

3.1 A.M. Noise

By contrast, the AM noise level is virtually impossible to measure by conventional techniques. However, the basis of the technique described to measure it is the realization that if the AM noise level is significant, it should be measurable by building a complete homodyne receiver (as in figure 1) and detecting the degradation of its sensitivity due to the leakage noise sidebands. By this technique, if the degradation of the noise figure cannot be detected at the maximum power level which can be put into the receiver, then it cannot be problem in a real system.

However, the configuration can be simplified if the homodyne mixer, with its two ports and the need to match the phases, can be replaced by single detector: since there is then no longer any difference between 'RF' and 'LO' paths, the system should have no sensitivity to FM noise and will have no sensitivity to the relative phases of the signals. It is, however, necessary to check the sensitivity of the system to Frequency Modulation since the principle relies very heavily on suppression of FM noise to measure AM noise components with power levels 60dB or more below the FM noise.

Since it is novel, this paper will concentrate on the measurement of AM noise.

Figure 2 shows the complete block diagram of the measurement rig.

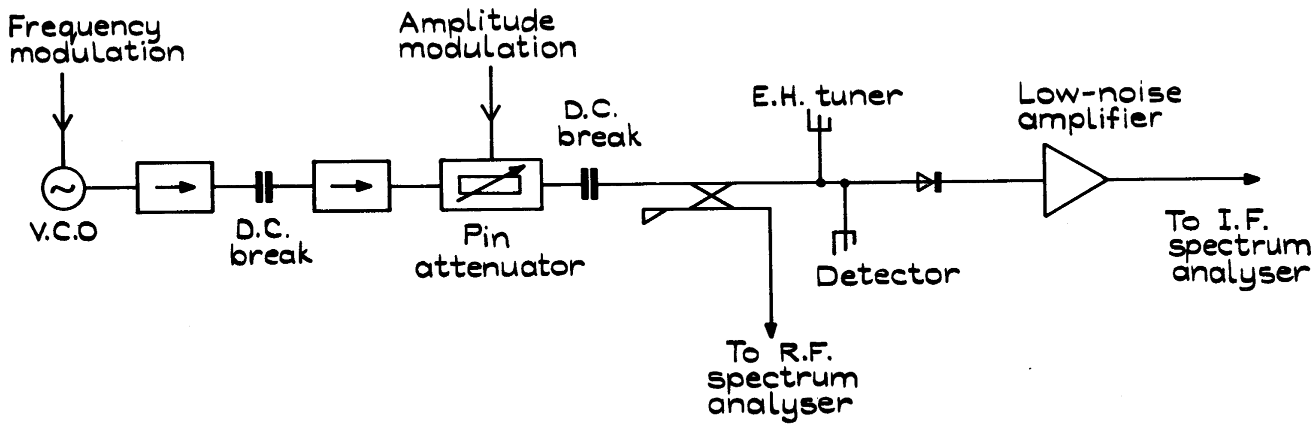

Figure 2: AM Noise Measurement Rig

It relies on the oscillator being able to be frequency modulated in order to measure the sensitivity of the rig to FM noise. It should be possible to measure the FM sensitivity with a separate oscillator from that being measured, if one is sure that the sensitivity will not be significantly enhanced when another oscillator is used.

The principle is, as implied above, to detect the signal and measure the variations in the DC level to estimate the oscillator's A.M noise level. To obtain adequate sensitivity, the low noise amplifier should have a noise figure comparable to that used in an actual radar.

The system is calibrated for a given oscillator, i.e. for a given input power level, by applying deliberate AM and FM modulation to the oscillator's output, measuring the applied modulation depth on a millimetre-wave spectrum analyser and measuring the detected signal, after amplification, on an IF spectrum analyser.

The calibration relies on the system having a high enough dynamic range to be able to produce modulation levels which can be detected on the RF spectrum analyser without overloading the IF amplifiers. Since the IF is only a few megahertz at most, good linearity can be achieved by applying negative feedback to the amplifier. Clearly, its limiting behaviour must be characterized to ensure that it is still operating in its linear range when the deliberate amplitude modulation is being applied. Of course, it is not necessary actually to be able to detect the effects of FM on the IF, provided that if nothing is seen then one still has an adequately-sensitive upper limit to the sensitivity.

The DC breaks are needed to prevent earth currents from the modulation signals entering the IF amplifier directly, and also to prevent earth currents from the deliberate amplitude modulation entering the oscillator and causing spurious frequency modulation. The presence of such inadvertent FM is easily detected as an imbalance in the AM noise sidebands of the signal.

The rig was built in WG27 waveguide, in which the d.c. breaks were easily built using sellotape as an insulator, ignoring the alignment dowels and using nylon fixing screws. The mismatch due to this was negligible compared to that of the pin attenuator.

The E-H tuner is used because the detector and the PIN attenuator are both bad matches to the waveguide and the sensitivity of the detector is unpredictable unless the tuner is used. It is optimised to maximise the sensitivity to AM. In doing so it also minimises the sensitivity to FM, which is also a good thing.

With a nominal oscillator power of +10dBm, the power into the detector was 2.6dBm. The sensitivity of the detector itself was about 170mV/mW into a load of 10kΩ. The noise figure of the IF amplifier at this impedance was about 1dB and the noise floor was equivalent to an AM noise level of -161dBc/Hzs.s.s.b at 1MHz from carrier. Since changes in the receiver noise level of half a decibel could be detected, the ultimate noise floor of the system was below -170dBc/Hz.

The rig was usable between about 60kHz and 4MHz from carrier, over which range the ratio of AM sensitivity to FM sensitivity was between 78.5dB, at the lower limit and 54.9dB at the upper end. The characterization was performed with applied modulation depths in the order of -30dBc and-50dBc.

Figure 3 shows the measured AM noise levels of a Philips varactor tuned gunn oscillator.

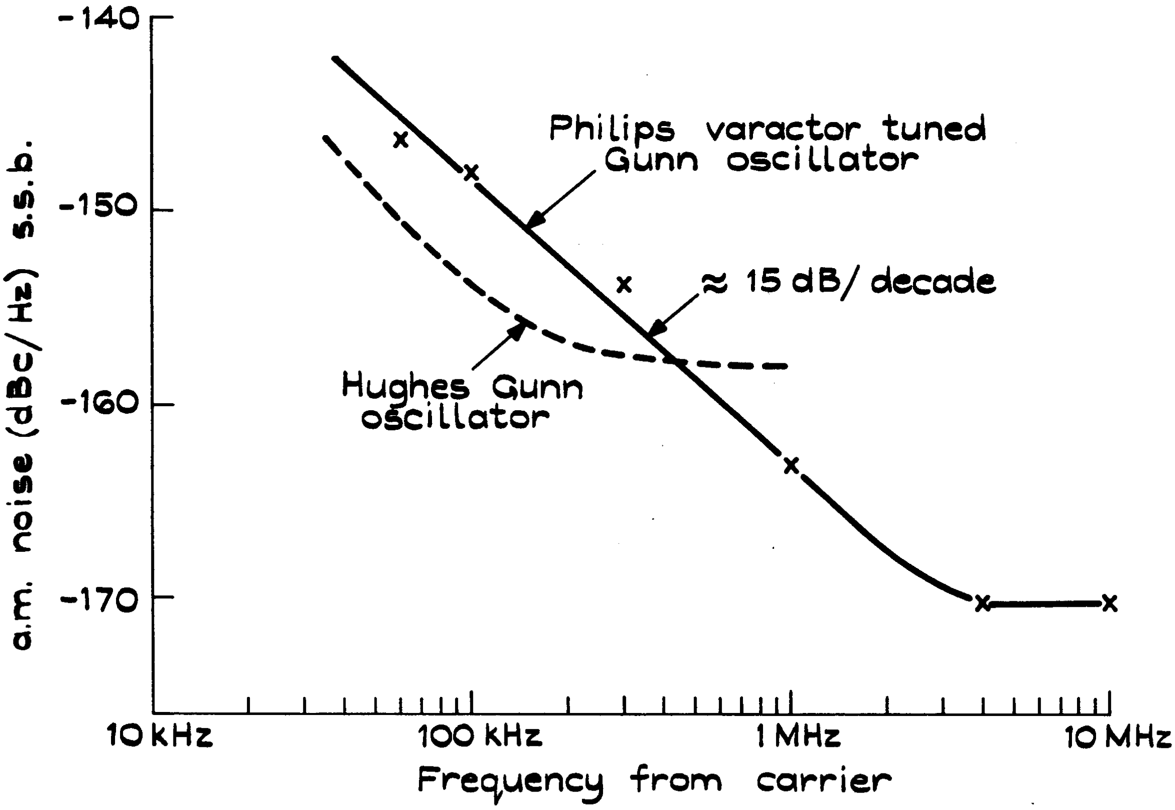

Figure 3: Measured AM Noise Levels

The dotted curved is published (catalogue) data for a Hughes oscillator. Note that both show a significant fall-off with frequency at lower offsets, typical of the 1/*f* noise associated with gallium arsenide devices of 1980's vintage. The noise floor of -170dBc/Hz is lower than most published data, but is typical of many careful measurements - the published data may just reflect the limited sensitivity of other measuring systems.

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3.2 FM Noise

Our preferred method of measuring FM noise is to beat together two millimetre-wave oscillators. This may appear to introduce an extra source of uncertainty in the noise of the second oscillator, however:

a) this is more sensitive that using the conventional harmonic mixer, because of the greater sensitivity of a conventional mixer

b) at millimetric frequencies, the far-from carrier FM noise of the multiplied-up spectrum analyser oscillator is in fact comparable with that of the oscillators being measured and

c) whereas a single measurement is ambiguous, if three oscillators are used the three pairwise comparisons which are available allow the noise levels of the individual oscillators to be resolved.

Figure 4 shows the FM noise levels of the three oscillators used.

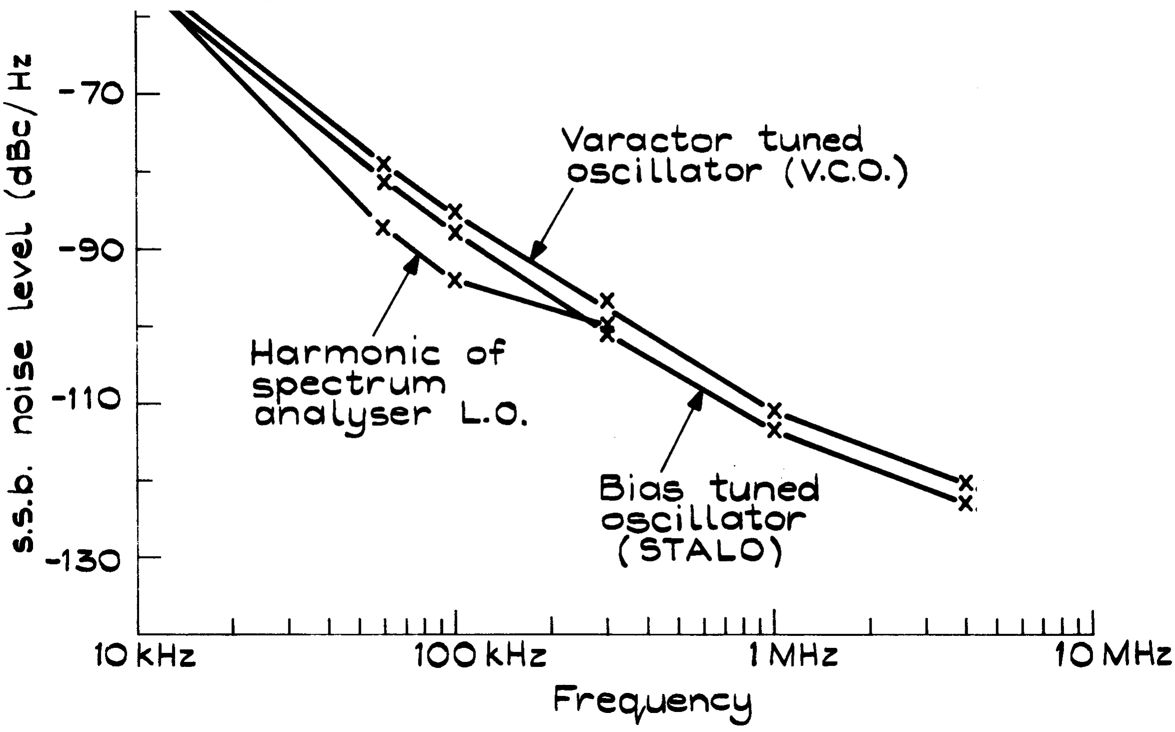

Figure 4: Measured FM Noise Levels

Note that two millimetre-wave oscillators were used and the additional measurements were made by mixing the signal from these with a multiplied-up spectrum analyser. The measurements with the spectrum analyser LO were made with a harmonic mixer and were unreliable at more than 300kHz from carrier, due to the relatively poor sensitivity of this arrangement. The total noise power measured by the single measurement which was available further from the carrier was then partitioned between the two oscillators assuming that the relative noise levels of the other two oscillators were the same at higher offsets as they had been at those where the complete three-way measurement had been possible. Of course, a production system could use three oscillators to resolve the ambiguities over the whole frequency range of interest, and then use this one calibrated oscillator as the 'LO' for all future measurements.

The following may also be noted:

a) as was asserted above, the multiplied-up spectrum analyser is not significantly quieter than the VCOs.

b) at 1MHz from carrier the FM noise of the VCO was about -111dBc/Hz, which would limit the sensitivity of the AM noise measurements to -175dBc/Hz. This is just over 10dB below the actual AM noise level, showing both that the AM noise floor measurement is valid, but also that the precaution of measuring the FM-to-AM conversion in the AM measurement rig was wise.

For reference, the spectrum analyser was an HP 8566B, and it should be noted that more modern spectrum analysers, with commercial millimetre-wave harmonic mixers, may well be significantly more sensitive.

4. FUTURE DEVELOPMENTS

The operation of the AM measurement system can be automated if the calibration is well characterized. The noise floor of the system is in fact very stable for a modern transistor amplifier, and the other calibrations can be performed as an algorithm which can be automated with modern test gear which can be connected to a computer. Clearly, more professional d.c. breaks can be constructed, but whether they would actually provide any significant benefit is an open question.

The FM measurement technique may now be superseded by better spectrum analysers, but it may still be of value for calibration.

5. REFERENCES

1: Stove, A. G., 'Linear FMCW Radar Techniques,' IEE Proc F, Radars Signal Process., 1992, **139**, (5), pp343-350

2: Stove, A. G., 'Erratum to 'Linear FMCW Radar Techniques, ' IEE Proc F, Radars Signal Process., 1992, **140**, (2), p137

3: European Patent EP 0138 253 B1, 'Noise Reduction in CW Radar Systems.'