Very high frequency and microwave interferometric phase and amplitude noise measurements

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The interferometric technique allows close-to-the-carrier measurements of both phase and amplitude noise, improving the instrument noise floor by 10–25 dB as compared to the traditional method based on a saturated mixer. Principles and basic equations describing the noise measurement system are given, together with design strategies suitable to microwave and very high frequency bands. Two prototypes, operating at 9 GHz and 100 MHz are discussed in detail. The relevant features of these prototypes are the capability to operate in a wide power range, below 0 dBm and above 20 dBm, and low noise floor. The latter is about -180 dB rad^2/Hz (white) and 150 dB rad^2/Hz (flicker) at 1 Hz Fourier frequency, at carrier power from 9 to 15 dBm. © *1999 American Institute of Physics.* [S0034-6748(99)02301-1]

I. INTRODUCTION

The interferometric method for measuring close-to-the-carrier phase and amplitude noise was proposed by Sann[1] about 30 years ago as a means to characterize microwave power amplifiers. Meanwhile, a quite similar technique was used by Horn[2] to extend the dynamic range of traditional HF spectrum analyzers, up to 40 MHz, when used to measure noise or low power sidebands in the vicinity of the carrier. But it was only recently that the interferometric technique revived the attention of the time and frequency metrology community, after some impressive results published by the University of Western Australia. In fact, it has been demonstrated[3] that for the X band a phase modulation (PM) noise measurement system (NMS), based on this method, shows a noise floor 10–25 dB lower than that of traditional NMSs based on a saturated mixer. Hence, the interferometric method makes it possible to measure the noise of some devices, such as phase shifters and isolators, that could not be measured with the saturated mixer. In addition, the interferometric method allows the detection of amplitude modulation (AM) noise. Finally, it makes the instant value of PM or AM noise available in real time, allowing dynamic cancellation of noise.[4]

After studying the interferometric method, we built two NMSs, operating at 9 GHz and 100 MHz, respectively. The latter, which is thought to be new, was able to be constructed after solving certain problems that are not present in the other band.

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II. TRADITIONAL PHASE NOISE MEASUREMENT METHOD

When a pure sinusoidal signal $s_i(t)$ of frequency ν_c passes through a device under test (DUT), the latter adds its internally generated noise. Then, the DUT output signal can be represented as

$$s_o(t)=\sqrt{2R_0P_o}[1+\alpha(t)]\cos[2\pi\nu_c t+\varphi(t)], \tag{1}$$

where R_0 is the impedance and P_o the power at the output of the DUT; $\varphi(t)$ and $\alpha(t)$ are, respectively, the PM noise and the relative AM noise generated by the DUT. The physical quantity of major interest is the power spectrum density (PSD) $S_\varphi(f)$ of $\varphi(t)$, as a function of the Fourier frequency f. $S_\alpha(f)$ is also of interest.

The traditional scheme for measuring $S_\varphi(f)$, shown in Fig. 1, is based on a double balanced mixer (DBM) saturated at both the inputs, used as a phase detector. Setting the phase shift γ so that the two input signals are in quadrature, the mixer provides a voltage $V\propto\varphi$. Introducing the power gain $K_\varphi=V^2/\varphi^2$, the measurement of $S_V(f)$, usually obtained by means of a fast Fourier transform (FFT) analyzer, gives $S_\varphi(f)$ according to $S_\varphi(f)=S_V(f)/K_\varphi$. The value of K_φ depends on the mixer and its driving power and must be determined experimentally; in most practical cases $\sqrt{K_\varphi}$ spans in the 0.1–0.3 V/rad range ($K_\varphi=-20$ to -10 dBV^2/rad^2), the highest value occurring when the mixer is driven by a relatively high power, say 15 dBm. If the driving power decreases, K_φ decreases rapidly and the mixer becomes unusable; for most DBMs, the minimum power turns out to be in the milliwatt range. For this reason some DUTs which need to operate at low power, like quartz resonators, cannot be measured with this method without inserting an amplifier; unfortunately, in most cases of interest, the noise contribution of this amplifier is higher than the DUT noise.

The white noise limit $S_{\varphi 0}$ of the above method comes from the voltage noise of the operational amplifier (OPA). In fact, low noise operational amplifiers show a noise floor of the order of 1 nV/$\sqrt{\mathrm{Hz}}$ ($S_V = -180$ dBV^2/Hz); cascading this amplifier to a high gain mixer ($K_\varphi = -10$ dBV^2/rad^2), the expected noise floor is $S_{\varphi 0} = -170$ dB rad^2/Hz; the latter is close to the noise floor of some commercially available instruments. It should be remarked that the above operational amplifier shows poor noise performance because it is misused. In fact, the amplifier noise is at its minimum when the input is loaded by a resistance $R_b = \sqrt{S_V(f)/S_I(f)}$, which is in the k$\Omega$ range; the latter is far from the mixer output impedance, which is typically 50 Ω.

As regards to the flicker noise, we could not find in the literature any theoretical estimate for the scheme of Fig. 1. In our experience, the flicker limit for a well constructed machine operating in good conditions would be about -140 dB rad^2/Hz at 1 Hz from the carrier, due to the noise of the mixer.

III. BASICS OF THE INTERFEROMETRIC METHOD

The scheme of the interferometric NMS is shown in Fig. 2. The hybrid coupler makes the vector addition, i.e., the *interference*, of its input signals. Thus, setting ℓ' and γ' equal to the DUT attenuation and phase shift, respectively, all the carrier power goes to the Σ output, and the carrier is suppressed at the Δ output. The DUT noise sidebands, which are *not* suppressed by the interference mechanism, are amplified by the low noise amplifier (LNA) and downconverted to base band by the mixer. Setting the detection phase γ'' to 90° or 0°, the mixer downconverts PM or AM noise, respectively. The attenuation ℓ'' must be set so that the LO input of the mixer is driven with appropriate pumping power.

The conversion gain of the NMS, i.e., $K_\varphi = V^2/\varphi^2$ or $K_\alpha = V^2/\alpha^2$ can be derived as follows. Letting $N_\varphi(\nu)$ and $N_\alpha(\nu)$ the PSD of the quadrature and in-phase DUT output noise around the carrier frequency ν_c, the PSD at the Δ output of the hybrid is

$$S_\Delta(\nu) = \frac{N_\varphi(\nu) + N_\alpha(\nu)}{2\ell_h}, \tag{2}$$

where ℓ_h is the loss of the hybrid. Let us now introduce the LNA gain g_a, the mixer output impedance R_0 and the mixer loss ℓ_m; the latter also includes the 3 dB intrinsic loss, due to the fact that the mixer makes the sum and the difference of its input frequencies, and consequently the input power is split into two bands. Assuming that—as it happens in reality—the DUT noise sidebands are symmetrical with respect to the carrier, so that $N_\varphi(\nu_c - f) = N_\varphi(\nu_c + f)$ and $N_\alpha(\nu_c - f) = N_\alpha(\nu_c + f)$, the DUT noise downconverted to base band turns out to be

$$S_V(f) = \frac{2R_0 g_a}{\ell_h \ell_m}\left[N_\varphi(\nu_c + f)\sin^2\gamma'' + N_\alpha(\nu_c + f)\cos^2\gamma''\right]. \tag{3}$$

Still assuming the sideband symmetry, PM and AM noise are related to N_φ and N_α by

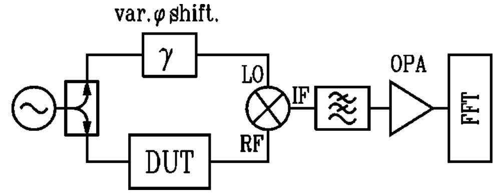

FIG. 1. Scheme of the traditional phase noise measurement system based on a saturated mixer.

$$S_\varphi(f) = 2\,\frac{N_\varphi(\nu_c + f)}{P_o}$$

and

$$S_\alpha(f) = 2\,\frac{N_\alpha(\nu_c + f)}{P_o}.$$

Combining the above with Eq. (3), we obtain

$$K_\varphi = \frac{R_0 g_a P_o}{\ell_h \ell_m} \tag{4}$$

for $\gamma'' = 90°$, and

$$K_\alpha = \frac{R_0 g_a P_o}{\ell_h \ell_m} \tag{5}$$

for $\gamma'' = 0°$.

The following example gives a picture of the reality. Assuming $\ell_h = 0.5$ dB and $\ell_m = 6$ dB, typical values for hybrids and mixers, choosing a $g_a = 37$ dB amplifier and setting the driving power so that $P_o = 15$ dBm, K_φ turns out to be 32 dBV^2/rad^2. The latter is 42 dB higher than the gain of a traditional NMS in similar conditions.

The expected white noise floor of the NMS can be derived from the LNA equivalent input noise $S_{\Delta 0} = F_a k_B T_0$ in the absence of DUT noise; F_a is the amplifier noise figure, $k_B = 1.38 \times 10^{-23}$ W/Hz the Boltzmann constant, and $T_0 = 290$ K the absolute reference temperature. It is assumed that the temperature of the interferometer is close to T_0. In this condition, the noise at the mixer output is

$$S_{V0}(f) = 2\,\frac{R_0 g_a}{\ell_m}\,F_a k_B T_0. \tag{6}$$

Combining the latter with Eq. (4), under the assumption that $F_a k_B T_0$ gives equal contributions to AM and PM, we obtain the PM noise floor of the NMS:

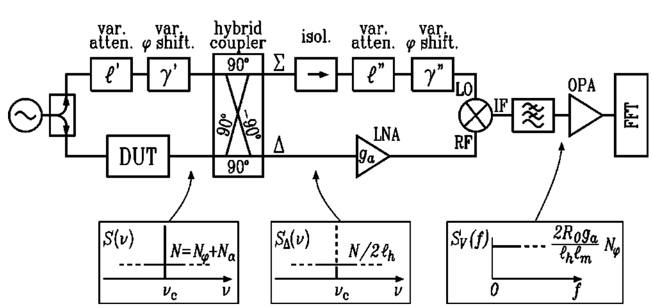

FIG. 2. Scheme of the interferometric phase noise measurement system.

$$S_{\varphi 0}(f) = 2\ell_h \frac{F_a k_B T_0}{P_o}. \tag{7}$$

The same development yields the AM noise floor

$$S_{\alpha 0}(f) = 2\ell_h \frac{F_a k_B T_0}{P_o}. \tag{8}$$

It is worth mentioning that $S_{\varphi 0}$ and $S_{\alpha 0}$ decrease as the DUT carrier output power P_o increases. In principle, P_o is limited by the maximum power of the passive components (hybrid, phase shifter, etc.), which could handle 30 dBm or more. Considering a realistic situation in which, for example, $P_o = 15$ dBm and $F_a = 2$ dB, we obtain $S_{\varphi 0} = -185$ dB rad^2/Hz. The latter is about 15 dB lower than the floor of a saturated mixer NMS in similar conditions.

A really important point which makes the interferometer attractive is that the flicker noise contribution of the LNA can be made negligible. In fact, recent studies[5,6] show that close-to-the-carrier flicker of amplifiers is due to the dc-bias flicker, up-converted by nonlinearity. Obviously, the radio-frequency flicker noise of an amplifier vanishes if the carrier power is sufficiently small to keep the device in its full linear regime; this is ensured by the carrier suppression mechanism in a properly designed NMS. The flicker noise contribution of the mixer can also be made negligible by a proper choice of g_a. Minor flicker contributions still remain, mainly due to the variable attenuator and the phase shifter.

Finally, in our opinion, the term ''carrier suppression'' should not be used as a synonym of ''interferometric'' because it is ambiguous. In fact, in both cases the carrier is suppressed, by multiplication or by vector addition, and the term ''carrier suppression'' is used also for the saturated mixer NMSs.[7]

IV. MICROWAVE AND VHF DESIGN STRATEGIES

The residual carrier power at the Δ output of the hybrid is a critical parameter because it could make the LNA flicker. Unfortunately, a suppression specification can hardly be drawn because the flicker behavior of commercially available amplifiers is not documented. Thus, we can only give some hints that come from experience:

(1) Best results are obtained with g_a in the 30–40 dB range. Lower values make the noise of the mixer and low frequency amplifier to be critical, while higher values make the carrier suppression constraint too difficult to meet.

(2) The residual carrier power $g_a P_r$ at the amplifier output must be much lower than the maximum amplifier power P_m; the latter is usually specified as the ''1 dB compression level.'' A margin $P_m/(g_a P_r)$ of 35–40 dB or more is needed, depending on the amplifier.

(3) Basically, two parameters are responsible for the radio-frequency flicker of an amplifier, namely dc-bias flicker and nonlinearity. Although we have no information about the former, we can infer the latter from the harmonic intercept power, which is always specified for commercially available amplifiers. As a result, amplifiers showing the highest intermodulation intercept power tend to be the best ones.

In our experience, carrier suppression of the order of 60–80 dB can be necessary, depending on the carrier power and the amplifier dynamic range. Let us consider an example in which the amplifier shows $g_a = 40$ dB, $P_m = 15$ dBm and needs a power margin $P_m/(g_a P_r) = 35$ dB for full linearity. In this condition, the residual carrier P_r at the amplifier input must be less than -60 dBm; consequently, if the DUT output power is $P_o = 15$ dBm a carrier suppression of 75 dB must be ensured.

It should be noted that a carrier suppression of 80 dB implies that γ' and ℓ' are set within 100 μrad and 8.7×10^{-4} dB, respectively, to their nominal values.

A. Microwave design

Phase matching is the greatest technical difficulty at microwave frequencies. In fact, because the wavelength inside the cables is about 25 mm at 9 GHz, matching the phases within 100 μrad for 80 dB carrier suppression is equivalent to matching the electrical lengths within 0.4 μm. Obviously phase matching must be stable at that level for the duration of the experiment, say half an hour. Some commercially available phase shifters are adequate to do so, after a really patient adjustment.

A further difficulty related to the short wavelength comes from mechanical vibrations. In fact, at 10 GHz a noise floor of -180 dB rad^2/Hz corresponds to electrical length fluctuations of 4×10^{-12} m/$\sqrt{\text{Hz}}$. In our experiments sufficient stability is obtained by fixing all the parts onto an antivibrating table of the same type of those commonly used for optics; moreover, all the cables connecting the NMS to the external world are secured to the table.

Microwave amplifiers show a wide bandwidth, often more than 10 GHz. Noise integrated over such a wide band can push the amplifier out of linearity. A bandpass filter can be needed.

Finally, microwave hybrids and mixers show poor isolation, typically of the order of 20 dB. The obvious consequence is an unwanted feedback of the amplified signal through the mixer and the hybrid. In order to prevent oscillation or measurement alteration, isolation must be increased by inserting some ferrite isolators; the best configuration must be determined experimentally.

B. VHF design

Phase matching, phase stability, and a sufficient damping of mechanical vibrations are much easier to achieve than in the X band because the wavelength is 100 times longer. The adoption of semirigid cables, SMA type connectors, and the antivibrating table ensures sufficient stability. In spite of this, for a series of reasons the design turns out to be more difficult than in the X band.

Eliminating the harmonics at frequencies the multiple of ν_c is a critical point because the carrier suppression mechanism has no effect on them. As almost all the components show a bandwidth of 2–3 decades, these unwanted signals would be present in the entire circuit, pushing the LNA out of linearity and making it flicker. The only known solution

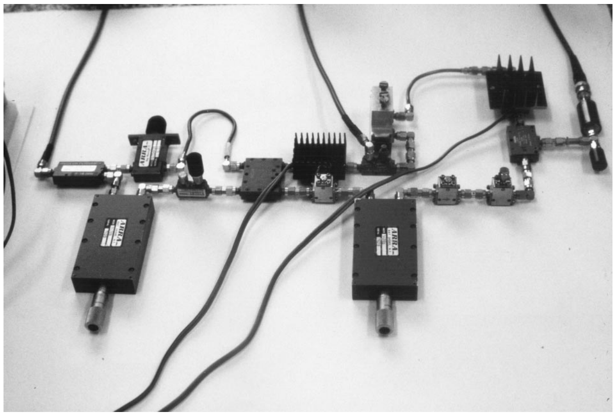

FIG. 3. Microwave prototype of interferometric noise measurement system.

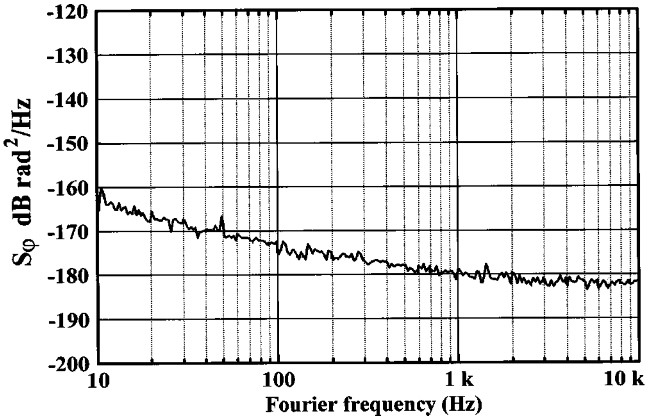

FIG. 5. Instrument phase noise $S_{\varphi 0}(f)$ of the microwave NMS prototype measured at $\nu_c = 9.13$ GHz and DUT power $P_o = 14$ dBm.

consists of inserting low Q bandpass filters in certain points of the circuit.

In our experience, the most difficult problem in the realization of the VHF interferometer arises from the variable phase shifters. We tried some microwave devices, available in our laboratory, consisting of a transmission line whose length can be varied by means of a micrometer. Apart from the small delay range (0.1–1 ns), that can be extended with a set of calibrated cables, these phase shifters turned out to be scarcely useful because of their high flicker noise; the same devices worked successfully in our microwave experiments. We guess that this anomalous behavior could be due to the parasitic capacitance in parallel with nonperfect contacts, which behaves as a short circuit at 9 GHz and takes in acoustic noise when used at 100 MHz. Anyway, we designed a phase shifter specific for this application, based on a LC network with variable capacitances.

Variable attenuators suitable for the VHF band are generally based on potentiometers and for this reason they tend to generate more flicker noise than the microwave ones, based on movable absorbing surfaces. We are still searching for a more satisfactory solution, consisting either of better potentiometers or a different physical principle.

It should be mentioned that the flicker performance of variable attenuators and phase shifters is usually not documented in the device specification, and consequently the possibility to find low noise devices relies upon experience and a pinch of good luck.

Ferrite isolators are not available for the VHF band and must be replaced by active isolators. Although noise is not critical at the Σ output of the hybrid, where the isolators are to be placed, it is really important to drive both the active isolator and the mixer at the appropriate power level.

The presence of electromagnetic pollution can be a relevant problem at some frequencies. In fact, in highly populated areas of Europe and the USA—well covered by FM broadcastings—the electromagnetic field is often of the order of $+100$ dB μV/m in the 88–108 MHz band. Solutions are strongly dependent on the local pollution, as well as the design.

In principle, the above solutions can also be used at lower frequencies of great interest for high spectral purity applications, such as 5–10 MHz.

V. IMPLEMENTATION AND RESULTS

Figures 3 and 4 show our prototypes, working at $\nu_c = 9.13$ GHz and $\nu_c = 100$ MHz. The contrast between the straightforward world of microwaves and the more complicated world of VHF, where many custom parts must be handmade, is evident even at first sight.

Because our main interest is the characterization of the NMS, in both cases the DUT is replaced by a short cable; operating with this ''null DUT,'' we inserted the variable elements ℓ' and γ' in the opposite arms of the interferometer because this makes the adjustment easier.

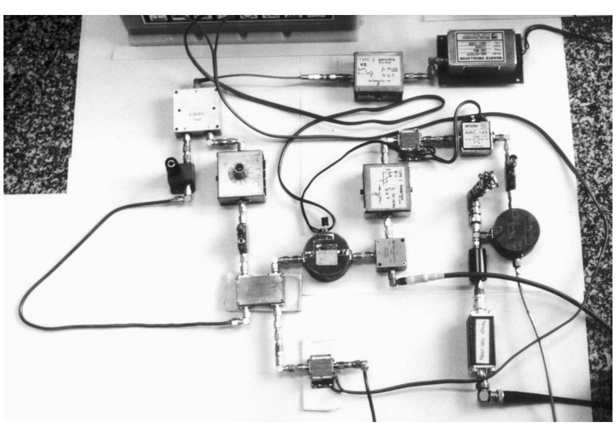

FIG. 4. VHF prototype of interferometric noise measurement system.

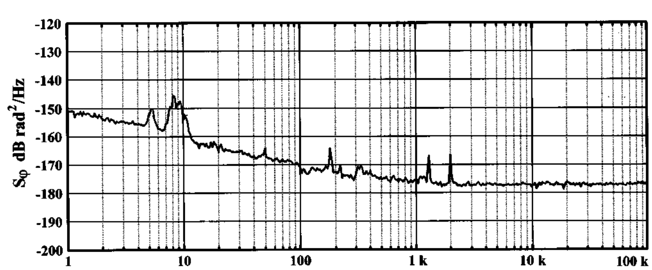

FIG. 6. Instrument phase noise $S_{\varphi 0}(f)$ of the VHF NMS prototype measured at $\nu_c = 100$ MHz and DUT power $P_o = 9$ dBm.

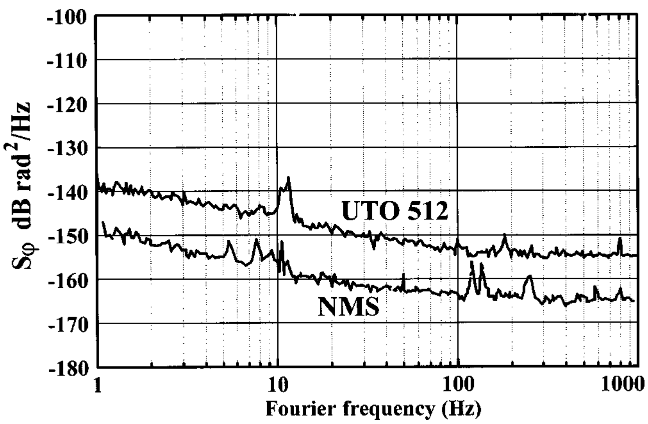

FIG. 7. Phase noise $S_\varphi(f)$ of a UTO 512 amplifier measured at $\nu_c = 100$ MHz in weak signal conditions. The actual output power is $P_o = 0$ dBm, while the 1 dB compression power is $P_m = +8$ dBm. The instrument noise of the NMS is also reported.

Let us first consider the microwave NMS. The amplifier consists of two cascaded modules with a directional coupler and dielectric filter in between; the coupler allows the monitoring of the residual carrier, while the filter limits the bandwidth to 100 MHz in order to prevent the second stage to flicker. The relevant characteristics of the amplifier are $g_a = 41.5$ dB and $F_a = 2$ dB. Losses of the hybrid and the double balanced mixer are $\ell_h = 1$ dB and $\ell_m = 8.5$ dB, respectively. Working with a DUT power $P_o = 14$ dBm, the calculated gain is $K_\varphi = 33.9$ dBV^2/rad^2, which is close to the measured value $K_\varphi = 34.3$ dBV^2/rad^2. The latter was obtained by injecting a suitable modulation in the DUT path and measuring the corresponding voltage at the mixer output. A similar technique was used to tune γ''. In this case we injected a modulation perpendicular to the desired direction of detection and we nulled the output amplitude adjusting γ''.

The VHF prototype operates at $\nu_c = 100$ MHz, with a DUT power $P_o = 9$ dBm. In this prototype, the amplifier consists of 3 cascaded modules showing an overall gain $g_a = 42$ dB and a noise figure $F_a = 3.5$ dB; the latter can be improved to $F_a \approx 1$ dB by replacing the available amplifier with a low noise one. Two low Q bandpass filters, inserted at the output of the driving oscillator and at the output of the first amplifier stage, reduce the harmonic distortion to about 80 dB below the carrier. In order to monitor the residual carrier and the harmonic distortion, the amplifier includes two test points, at the output of the first module and at the output of the last one. The hybrid coupler is a 90° lumped element device showing a loss $\ell_h = 1$ dB; the loss of the double balanced mixer is $\ell_m = 6$ dB. Working with $P_o = 8$ dBm, the calculated gain is $K_\varphi = 31$ dBV^2/rad^2, close to the measured value $K_\varphi = 32.3$ dBV^2/rad^2. The measurement of the latter, as well as the tuning of γ'' was done with the same methods as explained above.

The measured noise floor of the microwave NMS is shown in Fig. 5. The white noise component turns out to be $S_{\varphi 0} = -183$ dB rad^2/Hz, close to the calculated value. In addition, a flicker component $S_{\varphi 0} = -162$ dB rad^2/Hz at $f = 1$ Hz is present. The measured instrument noise of the 100 MHz NMS, shown in Fig. 6, consists of a flicker component $S_{\varphi 0} = -150$ dB rad^2/Hz at $f = 1$ Hz plus a white component $S_{\varphi 0} = -177$ dB rad^2/Hz; the latter is close to the calculated value and is higher than the floor of the microwave NMS because of the lower value of P_0. We believe that the higher flicker noise of the VHF prototype, as compared to that of the microwave one, is due to the attenuator.

We wish to point out that the above experiment was done in laboratory conditions, where a shielded chamber was not available. Moreover, we did not perform any signal treatment to hide the effects of disturbance, i.e., the residuals of the mains, electromagnetic pollution, and acoustic vibrations. These unwanted signals were eliminated by proper mechanical assembly and by carefully choosing the best path for cables, ground, and power supply. By observing Figs. 5 and 6, one can recognize a resonance in the 10 Hz region, due to the antivibrating table, plus some spectral lines in the 200 Hz–2 kHz region, due to mechanical resonance of the semirigid cables of the interferometer, and some residuals of the mains, at 50 Hz and multiples.

Although our interest is focused on the NMS itself, we measured some components for demonstrative purposes only. Figure 7 shows the phase noise of a commercial amplifier (UTO 512, manufactured by Avantek) working in weak signal conditions, $P_o = 0$ dBm; in fact, the 1 dB compression level of this amplifier is $P_m = +8$ dBm. In these conditions, the gain of the NMS is $K_\varphi = 24$ dBV^2/rad^2 and the instrument noise floor is $S_{\varphi 0} = -165$ dB rad^2/Hz. This measurement would not have been possible with the traditional scheme of Fig. 1.

Figure 8 reports the noise of some microwave phase shifters manufactured by Arra, measured at 100 MHz. Flicker PM noise of the 9428 B and 3428 A are close to the instrument noise floor. The available 9428D is an old component and its internal contacts could have been stressed. Finally, it should be pointed out that these phase shifters are not designed to operate in the VHF band.

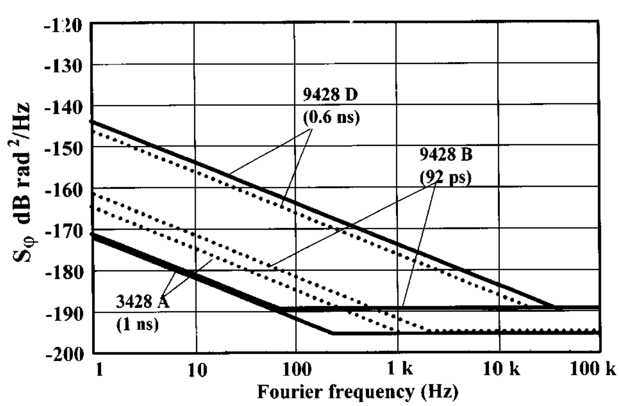

FIG. 8. Phase and amplitude noise of some microwave phase shifters measured at $\nu_c = 100$ MHz. Solid line: phase noise $S_\varphi(f)$; dotted line: amplitude noise $S_\alpha(f)$.

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