Six-Port Interferometric Technique for Accurate W-Band Phase-Noise Measurements

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***Abstract*—An innovative six-port (SP) phase-noise measurement technique for millimeter-wave high-power sources is proposed in this paper. Simulation results using a 94-GHz rectangular waveguide SP circuit model validates the measurement principle for both millimeter-wave oscillators and amplifiers. Phase-noise measurement results of a 100-W extended interaction Klystron amplifier are presented and discussed. Compared to conventional methods, this new method allows accurate low-cost phase-noise measurements.**

***Index Terms*—Frequency conversion, interferometry, millimeter-wave mixers, phase noise, six-port (SP).**

I. Introduction

THE USE of millimeter-wave frequency bands for radio applications was standardized one decade ago [1] and will be intensively exploited by widespread practical applications in the near future. Atmospheric absorption is one of the key factors in choosing a millimeter-wave frequency for a specific application. The W-band (94 GHz) atmospheric absorption average is negligible (0.4 dB/km), as compared to 20 dB/km of V-band (60 GHz) at sea level [2]. Therefore, the W-band represents an attractive candidate for future low Earth orbit satellite communications [2], [3] and imaging radars [4]. The success of these systems critically depends on their ability to reduce the phase noise (which sets the basic limit on many system performance criteria, including jitter, sensitivity, bit error rate, and resolution). At frequencies below 40 GHz, there are various suitable characterization techniques for microwave components, while at W-band, characterization techniques are often inconsistent, subject to inaccuracies or invalid due to high measurement offset. Therefore, it is primordial to develop accurate measurement techniques to characterize millimeter-wave components, including the klystrons, used in high-power transmitters.

Phase noise is one of the most important parameters of microwave sources and its measurement and reduction have drawn the attention of many authors in recent years [5]–[11]. Various techniques have been proposed to perform phase-noise measurements or to obtain control signals in phase-locking systems [5]. In single-channel systems, the output signal of the device-under-test (DUT) can be compared to a reference signal [5] or to its own value at a different instant using a delay line [6]. A dual-channel cross-correlation measurement technique has been proposed in [7] and [8]. The phase noise of an amplifier is measured using a pair of phase-sensitive detectors implemented using conventional multipliers. The authors found that W-band GaAs and YIG tuned multipliers used as phase detectors have an important PM noise, which limits the noise measurement system performances. Therefore, at W-band frequencies, the measurement techniques are not mature, especially because of high measurement noise. An accurate W-band phase detector is the key point of phase-noise measurements.

The interferometry technique has been proposed in single- or dual-channel configurations. As known, a wide range of different apparatuses including such classical examples as the bridge circuit, Dicke's microwave radiometer, and Michelson optical interferometer can be referred to as "null instruments" that are able to make very precise measurements of the relative changes in voltage, effective noise temperature, and optical phase delay, respectively. A microwave application for the X-band interferometric phase-noise measurement technique has been presented in [9] and [10]. Microwave frequency discriminators with interferometric signal processing have proven to be extremely effective to measure the oscillators' phase noise. In addition, a theoretical study of the dual channel interferometer, including an experiment at RF frequencies (100 MHz), have been presented in [11]. The in-phase and quadrature (I/Q) noise detection signals of RF source are generated using 180° couplers and a conventional mixer. The authors propose for microwave bands (up to 40 GHz) the use of Wilkinson power splitters and 90° couplers. The same conclusion has been drawn in our previous work over Ka-band [12], [13]. It is to be noted that the interferometry technique has never been applied to the study of noise phenomena in W-band. Six-port (SP) interferometry has been previously used in various applications such as millimeter-wave receivers [12]–[14], collision avoidance radar sensors [15], [16], and direction of arrival estimation [17].

II. Conventional Measurement Principle

Today's industrial millimeter-wave phase-noise measurement solutions (i.e., Agilent Technologies' E5500 Series) use a

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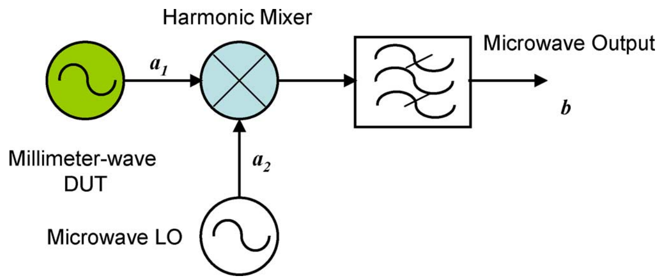

Fig. 1. Simplified block diagram of a millimeter-wave down-converter.

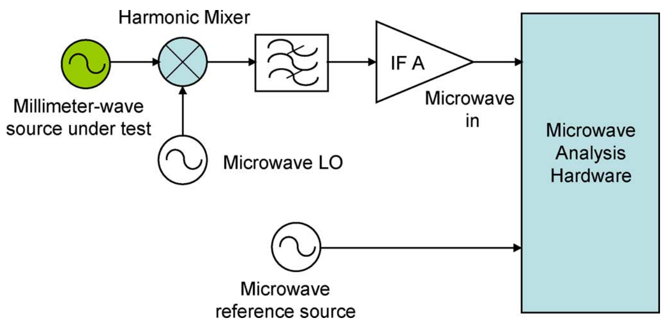

Fig. 2. Simplified block diagram of conventional millimeter-wave one-port phase-noise measurement system.

baseband phase-noise measurement system and microwave/millimeter-wave down-conversion. Unfortunately, mixers are noisy components and reduce the system performances.

The simplified block diagram of a down-converter using harmonic mixers is presented in Fig. 1. A microwave local oscillator (LO), harmonic mixer, and bandpass filter are used to obtain the output microwave signal.

The phase-noise contributions of the millimeter-wave output signals of the DUT and microwave LO can be expressed by the following equations:

$$a_1 = A_1 \cdot \cos\left[\omega \cdot t + \Delta\varphi_{\mathrm{DUT}}(t)\right] \qquad (1)$$
$$a_2 = A_2 \cdot \cos\left[\omega_0 \cdot t + \Delta\varphi_{\mathrm{LO}}(t)\right]. \qquad (2)$$

Due to the harmonic mixing (the microwave LO frequency is multiplied by a factor N during the mixing process), the down-converter's output signal b can be expressed as

$$b = \frac{K}{2} \cdot A_1 \cdot A_2 \cdot \cos\left\{(\omega - N \cdot \omega_0) \cdot t + \left[\Delta\varphi_{\mathrm{DUT}}(t) - N \cdot \Delta\varphi_{\mathrm{LO}}(t)\right]\right\}. \qquad (3)$$

Equation (3) shows an important supplementary noise contribution due to the harmonic mixing. Considering the worst case, the total phase-noise contribution at microwave output can be expressed as

$$\Delta\varphi_{\mathrm{out\,MAX}}(t) = \left|\Delta\varphi_{\mathrm{DUT}}(t)\right| + N \cdot \left|\Delta\varphi_{\mathrm{LO}}(t)\right|. \qquad (4)$$

Fig. 2 shows the simplified block diagram of the conventional millimeter-wave one-port phase-noise measurement system composed of an harmonic mixer down-converter, a microwave reference source, and microwave analysis hardware.

As an example, the Agilent's W-band one-port phase-noise measurement system (for millimeter-wave oscillators measurements) requires an additional harmonic mixer ($\times N$, $N = 14$ in W-band setup) and a low phase-noise microwave oscillator (6 GHz) in order to generate a microwave signal less than 26.5 GHz, the maximum input carrier frequency of the microwave analysis hardware. The down-converted signal at E5500 input (10 GHz in the case of a 94-GHz measurement) has important additional phase-noise components, with a maximal phase aperture according to (4). A second down-conversion, microwave to baseband, is performed inside the E 5500 equipment, increasing baseband phase-noise components. Therefore, the phase-noise contribution of the DUT itself is difficult to be identified with high accuracy.

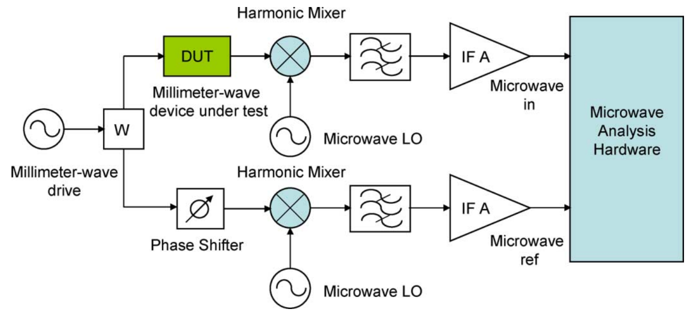

Fig. 3. Simplified block diagram of conventional millimeter-wave two-port phase-noise measurement system.

Fig. 3 shows the block diagram of a conventional millimeter-wave two-port measurement system. As example, Agilent Technologies' W-band two-port measurement system (for millimeter-wave amplifiers) requires two harmonic mixers and two low phase-noise LOs, doubling the additional phase-noise component of (4). As for one-port measurement, a second down-conversion must be performed in order to measure the phase noise near the carrier, increasing the global measurement error. Therefore, accurate millimeter-wave measurements of the DUT are difficult to perform without low phase noise and expensive microwave LOs and high-quality mixers.

Qualitative phase-noise measurements can be obtained by using modulation/demodulation techniques and digital signal processing of demodulated baseband signals. As known, the "diameter" of demodulated clusters is related to both amplitude and phase noise. The demodulated constellation is stabilized using signal processing and the maximum phase and amplitude errors of the input signal are displayed. This technique is currently used in industry for communication klystrons using quadrature phase-shift keying (QPSK) modulated signals. However, a complete noise analysis, i.e., power spectral density in an extended offset range near the carrier (1 Hz–100 MHz), cannot be performed using this technique. In our opinion, the use of the SP interferometry seems to be one of the best approaches for accurate millimeter-wave phase-noise measurement systems, as will be demonstrated below.

III. Interferometric Measurement Principle

As known, the SP is a passive linear component, mostly developed in the 1970s for accurate automated measurements of the complex reflection coefficient in microwave network analysis [18]. The conventional SP circuit has three "q_i" apart points spaced at 120° multiples.

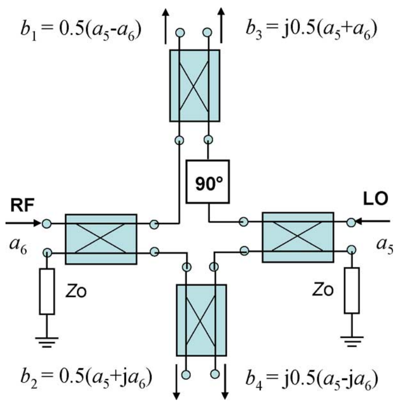

Fig. 4. Simplified block diagram of the SP circuit.

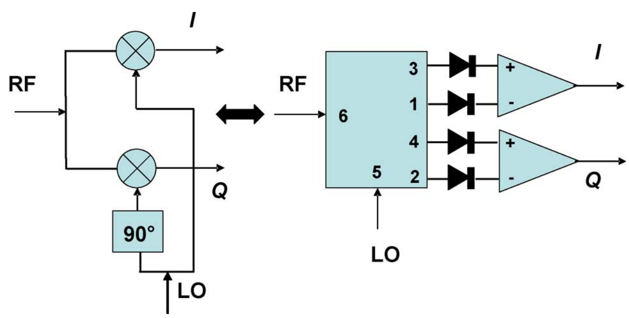

Fig. 5. Equivalence between the conventional I/Q and the SP mixers.

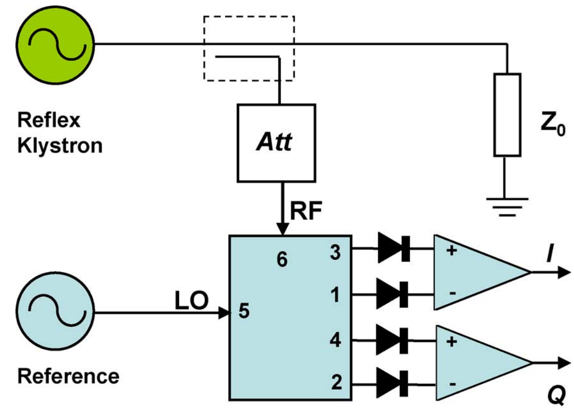

Fig. 6. Basic interferometric measurement setup of a klystron oscillator.

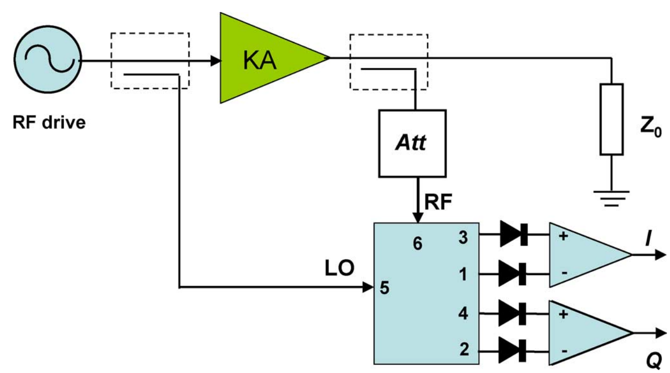

Fig. 7. Basic interferometric measurement setup of a KA.

A new SP architecture having four "q_i" points has been developed for millimeter-wave receivers and radar sensors [14]–[16]. Fig. 4 shows the block diagram of this circuit, composed of four 90° hybrid couplers. The fundamental characteristic of any SP network, which influences its performance in measurement applications, is determined by the position of the "q_i" points in the complex plane. According to SP theory [18], the "q_i" points are complex numbers representing the solution of $|b_i| = 0$ equations. The "q_i" points of the proposed SP circuit are located on a unit circle and spaced by 90° multiples at 0°, 90°, 180°, and 270°.

The equivalence between the conventional I/Q mixer architecture and the SP mixer are presented in Fig. 5 [14].

Power detectors are connected at the SP circuit outputs. Two differential amplifiers generate the I and Q output signals. A baseband phasor is, therefore, obtained using I/Q signals

$$\Gamma(t) = I + jQ = (V_3 - V_1) + j \cdot (V_4 - V_2) \quad (5)$$

$$\Gamma(t) = \alpha(t) \cdot K \cdot a^2 \cdot \exp[j\Delta\varphi(t)]. \quad (6)$$

We note that V_i ($i = 1, \ldots, 4$) are the detected output voltages, $\alpha(t)$ is the amplitude ratio between RF and LO signals, K is a constant, which depends on the power detector characteristics and on the baseband circuit (BBC) gain, a is the amplitude of LO input signal, and $\Delta\varphi(t)$ is the instantaneous phase shift between the RF and LO input signals.

Equation (6) shows that the Γ phasor amplitude is proportional to the amplitude ratio between RF and LO signals [$\alpha(t)$] and its phase is equal to the phase difference between RF and LO input signals [$\Delta\varphi(t)$]. Therefore, this system is capable to perform real-time AM/PM noise measurements.

The proposed one-port device measurement setup uses a W-band reference signal generated by a stable millimeter-wave source injected in the LO port of the SP interferometer, as seen in Fig. 6.

An important category of the DUT is represented by klystron oscillators. This type of klystron is also called a reflex klystron because of the reflex action of the electron beam. As known, the frequency stability of a reflex klystron is in the 10^{-3}–10^{-4} range, as compared to 10^{-6} of a stable millimeter-wave source.

According to (6), the SP also acts as a frequency discriminator when the RF input signals have different frequencies. The Γ phasor rotates in the complex plane with the frequency difference [15]

$$\Delta f = \frac{\Delta\varphi(t)}{2\pi\Delta t}. \quad (7)$$

In (7), the sign of Δf is given by the sense of the Γ phasor rotation (clockwise or counter-clockwise). It is to be noted that, according to (7) and the use of I/Q mixers, relative (Doppler) frequency measurements can be obtained with hertz accuracy at millimeter-wave frequencies.

The proposed two-port device measurement setup is presented in Fig. 7. The LO and RF input signals are obtained using directional couplers and attenuators. As already presented, the baseband I/Q signals are obtained using an SP circuit. The DUT is, in this case, a klystron amplifier (KA).

According to (6) and the equivalence with conventional I/Q mixers, the AM/PM klystron's noise contribution can be measured using baseband hardware. This interferometric technique can perform accurate measurements because the measured baseband phase noise is equal to the DUT's phase noise (6). Compared to conventional measurement systems (please see (3), (4) and related comments), the measurement accuracy using the interferometric principle is demonstrated.

Signal processing of I and Q outputs can be used for both previously discussed measurement techniques. For this purpose, two additional analog-to-digital converters must be connected to the measurement setup outputs.

IV. One-Port Results

Agilent Technologies' Advanced Design System (ADS) simulation of a klystron oscillator (according to the previously proposed measurement setup of Fig. 6) is performed using an SP model based on its S-parameter measurement results [15]. Two different millimeter-wave sources are used for this purpose: a reflex klystron and a stable millimeter-wave source. This source is considered to have a frequency stability of 10^{-6} and the klystron oscillator has a frequency offset and additional phase noise.

A phase-noise modulator is used to simulate the klystron output signal. This device uses Leeson's equation [19] to model the oscillator's phase noise

$$L\{\Delta\omega\} = 10\log\left\{\frac{2FkT}{P_{\rm sig}}\left[1+\left(\frac{\omega_0}{2Q_L\Delta\omega}\right)^2\right]\cdot\left(1+\frac{\omega_c}{|\Delta\omega|}\right)\right\}. \tag{8}$$

It is usual to normalize the mean-square noise voltage to the mean-square carrier voltage and report the ratio L in decibels. This ratio is commonly expressed as "decibels below the carrier per hertz" (dBc/Hz), specified at a particular offset frequency $\Delta\omega$ from the carrier frequency ω_0. The other parameters in the previous equation are the noise factor F, Boltzmann constant k, absolute temperature T, input signal power $P_{\rm sig}$, loaded Q value of the oscillator's resonator Q_L, and corner frequency ω_C. The term containing the corner frequency provides a $1/|\Delta\omega|^3$ behavior at sufficiently smaller offset frequencies.

The parameters used in our simulations are: an output klystron power of 50 dBm, $P_{\rm LO} = -10$ dBm, $F = 50$ dB, an attenuator (Att.) of 40 dB, a directional coupler of 20 dB, a BBC gain of 40 dB, $Q_L = 50$, $f_C = 1$ kHz, and an operating frequency $f_0 = 94$ GHz. In addition, due to the inherent frequency instability, an instantaneous frequency shift (supposedly equal to 0.9 MHz during the ADS envelope analysis of 100 μs) exists between the millimeter-wave sources.

A time frame of 5 μs is presented in Fig. 8. As seen, I/Q periodical signals of 1.1 μs, corresponding to the instantaneous frequency shift, clearly appears. These quadrature signals are also shifted by a quarter of period, corresponding to 90° phase shift. Due to inherent amplitude and phase imbalances of the SP circuit, a small dc offset (around 0.01 V, representing 10% of the signal amplitude) appears.

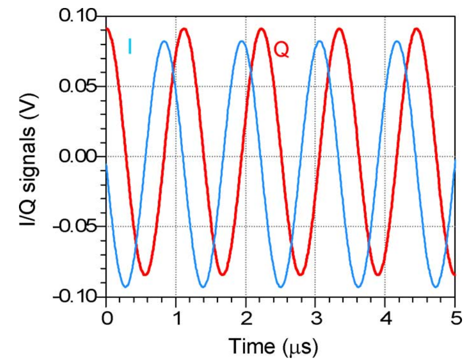

Fig. 8. Sequence of the output I/Q signals in time (oscillator).

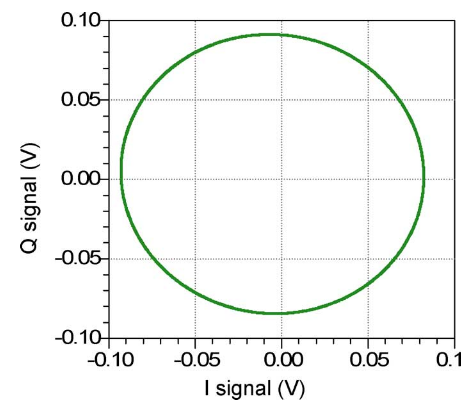

Fig. 9. Diagram of output signals in XY format (oscillator).

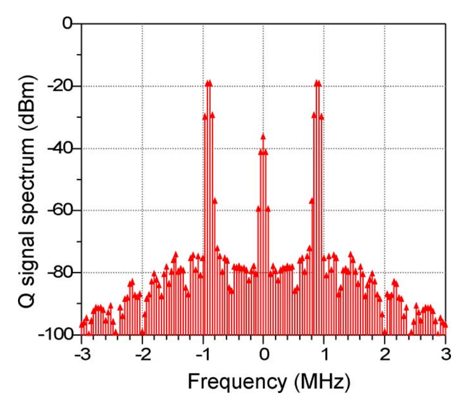

Fig. 10. Simulated Q signal spectrum of klystron oscillator.

The diagram of output signals in XY format (see Fig. 9) shows the circle predicted by (6). The circle radius is proportional to both LO and RF signal amplitudes.

Fig. 10 shows the baseband spectrum of the Q output signal. Similar results are obtained for the I signal. Two spectral lines, spaced 0.9 MHz from the carrier, with additional klystron phase-noise contribution, are clearly identified by measuring the instantaneous frequency shift between the millimeter-wave sources. The above simulation results are in very good agreement with the SP theory and validate the proposed measurement method.

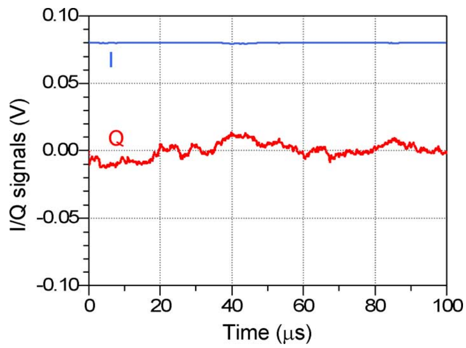

Fig. 11. Sequence of the output I/Q signals in time (amplifier).

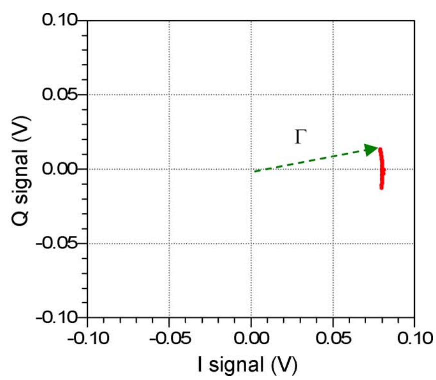

Fig. 12. Diagram of output signals in XY format (amplifier).

V. Two-Port Results

In order to perform two-port ADS system simulations, a KA model composed of a millimeter-wave amplifier and a phase-noise modulator is used (see setup of Fig. 7). Simulation parameters related to added phase-noise and millimeter-wave components are the same as in the previous case of the one-port. In addition, the output power of RF driver is equal to 10 dBm, the KA gain is 40 dB and the second directional coupler has a coupling factor of 20 dB. A time frame of 100 μs is presented in Fig. 11. The I/Q outputs measure the phase noise generated by the phase-noise modulator of the KA model. There is no low-frequency signal because the millimeter-wave input signals have exactly the same frequency.

The increased random amplitude of the Q output versus the I output suggests that a small phase variation is detected (close to the interception point with a horizontal axis). Actually, this result is confirmed by the diagram of output signals in XY format presented in Fig. 12. The phase noise is represented as a movement of the Γ phasor superposed to the circle predicted by (6). In addition, the random magnitude of Γ, the circle radius, represents the amplitude noise contribution.

It is to be noted that the phase shift between millimeter-wave inputs is adjusted in order to obtain the phase of the Γ vector equal to zero the with phase-noise modulator turned off in the KA model. Therefore, for this case, the Q signal represents the DUT's phase-noise contribution and the I signal represents the amplitude noise contribution.

In this diagram, the angular aperture is related to the random phase variation due to peak-to-peak noise and represents a direct measure of the overall klystron's phase noise. In practice, the diagram of the output signal in an XY format can be easily observed using an oscilloscope.

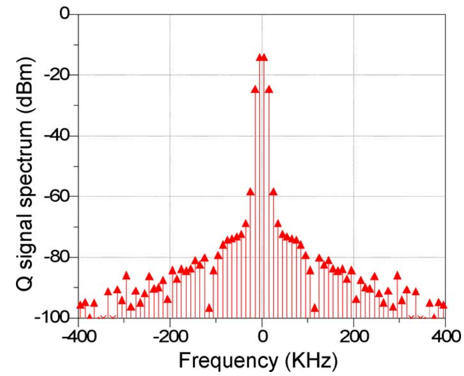

Fig. 13. Simulated Q signal spectrum of KA.

Fig. 14. Phase-noise measurement setup.

The output signal spectrum of Fig. 13 shows the klystron's phase-noise contribution (near the millimeter-wave carrier).

The results presented in Figs. 11–13 are complementary, representing the phase-noise angular aperture and the spectral components of this noise in the vicinity of the millimeter-wave frequency.

Measurements are performed using the 100-W extended interaction klystron (EIK) amplifier of the Poly-Grames Research Center, École Polytechnique of Montréal, Montréal, QC, Canada. The EIK was manufactured by Communications and Power Industries Ltd., Georgetown, ON, Canada.

Fig. 14 shows the experimental setup for two-port measurements. An Anritsu 68177 C signal generator and a millimeter-wave multiplier (X 6) are used to provide the 94-GHz RF drive. A Tektronix TDS 694C digital real time oscilloscope, operating in YT and XY display format, and a Rohde & Schwarz spectrum analyzer model FSIQ 40 are used to capture the output I/Q baseband signals.

As shown in the detailed photograph of Fig. 15, the setup is developed according to the block diagram of Fig. 7. The SP inputs are connected to the RF drive and EIK output using directional couplers and attenuators. The SP outputs are connected to the BBC via millimeter-wave power detectors.

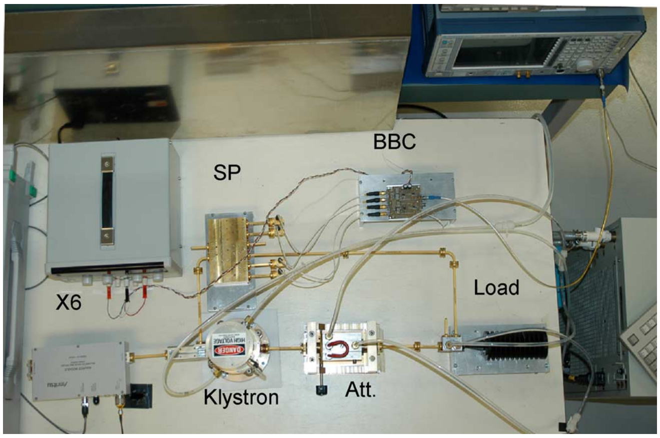

Fig. 15. Phase-noise measurement setup (details).

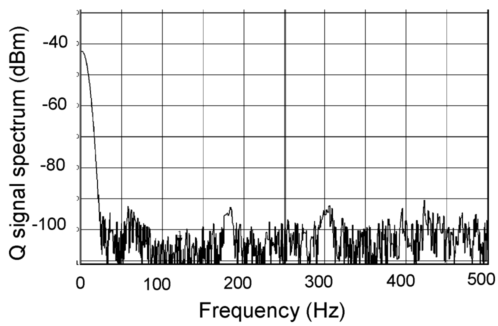

Fig. 16. Measured Q signal spectrum of 100-W KA operating at a reduced power of 50 W.

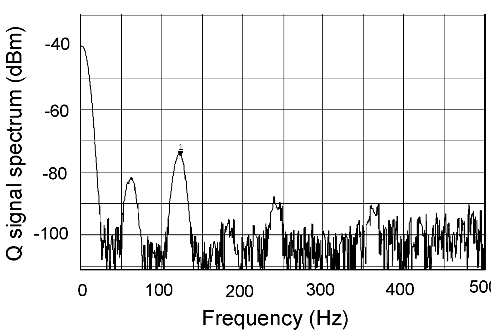

Fig. 17. Measured Q signal spectrum of 100-W KA operating at the maximal power.

Figs. 16 and 17 show typical EIK results of the measured Q signal spectrum at 50 W and at a nominal power of 100 W, respectively. As seen, several lines of the noise spectra occur at 60 Hz main and harmonic frequencies. These harmonics represent one of the most important noise contributions in the output phase-noise spectrum. The peaks on the noise spectrum can be as high as −35 dBc (see Fig. 17, the second harmonic, at 120 Hz from the carrier) and are due to the vibration sensitivity as well as electromagnetic interference.

As is known, the noise floor limits the smallest measurement that can be taken with certainty since no measured amplitude can, on average, be less than the noise floor. As presented, the proposed measurement system contains the SP interferometer system and the Rohde & Schwarz spectrum analyzer (model FSIQ 40). In addition, harmonic mixing at 94 GHz is avoided due to the use of the interferometric approach and, therefore, the noise floor is basically limited by the baseband spectrum analyzer performance [20].

The noise floor of the proposed phase-noise measurement system is measured with millimeter-wave inputs connected to standard WR-12 loads, at a room temperature of 20 °C. The measured noise floor value is approximately −110 dBm at 100 Hz, and −125 dBm at 500 Hz, with the spectrum analyzer resolution bandwidth of 10 Hz, corresponding to −120 and −135 dBm, respectively, at 1-Hz resolution bandwidth. According to the previous comments and results, for Leeson's script, the noise floor is approximately −80 dBc/Hz at 100 Hz, and −95 dBc/Hz at 500 Hz.

As a direct comparison, the specification summary of Agilent Technologies' phase-noise measurement solutions indicates the typical noise floor levels in the 75–110-GHz band, when adding the 11970 series harmonic mixers [21]. These values are −70 dBc/Hz at 100 Hz, and −97 dBc/Hz at 1 kHz, respectively. Therefore, using a linear interpolation of these results, a typical value of −90 dBc/Hz is obtained for the noise floor level at 500 Hz.

We can conclude that the noise floor value of the proposed measurement system is improved, especially because the harmonic mixing is avoided in the down-conversion process. Therefore, this accurate low-cost measurement technique based on SP technology can replace costly millimeter-wave spectrum analyzers with excellent results.

In addition, the measurements using the oscilloscope confirm the simulation results of Fig. 12. Random phase variation having a maximal aperture of around 30° has been observed on the digital oscilloscope screen for the KA operating at maximal power.

As known, in communication systems, the rms phase error must be kept below a specified value, a function of the modulation scheme. As an example, for a phase noise as high as −80 dBc/Hz over a 30-kHz bandwidth, the rms phase error is less than 1.5°, having negligible effects for most modulation schemes [22].

However, only the spectral analysis can show phase-noise spectral components as main harmonics observed in the near-carrier spectrum. These undesired spectral components can be, in principle, reduced by using better quality power supplies and getting rid of the spurious ground loops. Therefore, phase-noise reduction techniques can be implemented to reduce the undesired spectral peaks in order to use such high-power millimeter-wave devices in advanced communication systems.

VI. Conclusion

A new interferometric phase-noise measurement technique of millimeter-wave high-power oscillators and amplifiers is proposed in this paper.

As demonstrated by theory, and validated by ADS system simulations and by measurements of a 100-W extended interaction KA, SP technology allows millimeter-wave interferometric phase-noise measurements with very good accuracy using low-cost equipment.

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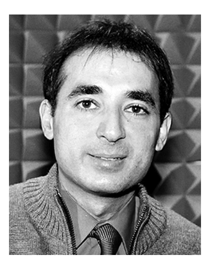

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