Noise Calculations and Experimental Results of Varactor Tunable Oscillators with Significantly Reduced Phase Noise

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Abstract—**The single-sideband phase noise of varactor tunable GaAs MESFET oscillators is investigated. Two oscillator circuits with different microstrip resonator circuits were designed and fabricated. Using a resonator consisting of coupled microstrip lines instead of a single microstrip line, which is a planar monolithically integrable structur, phase noise is reduced significantly because the quality factor is higher for the coupled resonator. The phase noise is calculated using a nonlinear time domain method, which solves the Langevin equations, describing the deterministic and stochastic behavior of an oscillator by pertubation methods. Calculated and measured phase noise agree within the accuracy of measurements. The very low phase noise of 95 dBc/Hz at 100 kHz offset frequency is achieved.**

I. Introduction

SINGLE SIDEBAND phase noise of planar GaAs MESFET oscillators with oscillation frequencies in the range of 10 to 20 GHz amounts to from −52 to about −83 dBc/Hz at an offset frequency of 100 kHz [1]. The phase noise of varactor tunable oscillators is high compared with single-frequency oscillators. It mainly depends on the noise behavior of the active devices as well as on the quality factor of the passive resonator structures.

The quality factor of planar integrated resonator structures using microstrip lines or coplanar waveguides is low compared to other resonators. On the other hand, high-Q resonators like DR- or YIG-devices are not monolithically integrable together with MMIC's and require much area. The significant noise contribution in case of microwave oscillators is the $1/f$-noise of the GaAs-MESFET and the varactor diode, which is upconverted to the oscillation frequency. The baseband noise of these devices was measured. In case of the GaAs-MESFET, the voltage dependent behavior of the $1/f$-noise is modeled using a voltage controlled noise voltage source at the gate terminal of the transistor. In the case of the varactor diode, no baseband noise above the noise floor of the measurement system was measured. We calculated and measured the single-sideband phase noise of two types of planar integrated varactor tunable oscillators with different resonator circuits at the gate terminal. The phase noise is calculated using a nonlinear time domain method [2]. The noise sources of the oscillator are

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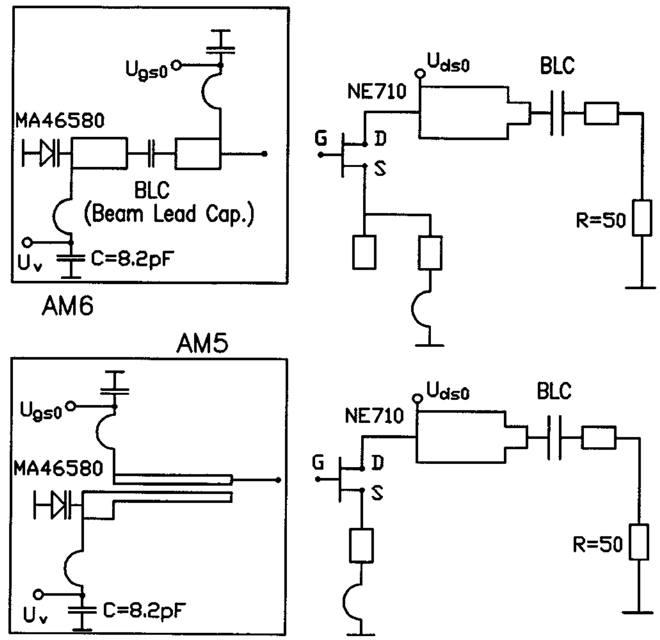

Fig. 1. The oscillator circuit with two different tunable resonator circuits at the gate terminal of the GaAs-MESFET.

$1/f^{\alpha}$-noise sources and white noise sources, like thermal and shot noise sources. The upconversion of the baseband noise to the harmonics and the modulation of the noise sources at the steady state are taken into account. Very low phase noise of −95 dBc/Hz is obtained by an oscillator with a coupled microstrip line resonator compared to a single microstrip line resonator.

II. Oscillator Design

The voltage controlled oscillators are designed with respect to maximum tuning range. In case of an oscillator circuit in common source configuration, maximum tuning bandwidth is obtained with a varactor at the gate circuit [3]. The frequency determining network at the gate of the GaAs-MESFET forms a series resonant circuit. The oscillator circuit with capacitive feedback at the source terminal is shown in Fig. 1, including two different tunable resonator circuits. One of the oscillators (called AM5) is designed with a tunable resonator, consisting of a coupled microstrip line that is terminated by a varactor diode. In the case of the other oscillator (called AM6), the varactor is coupled to the GaAs-MESFET using a single

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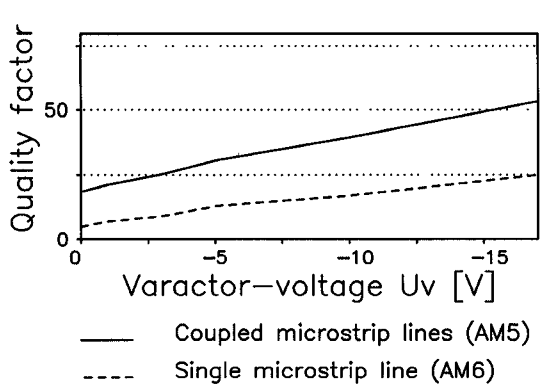


Fig. 2. The quality factors of the series resonance circuits at the gate terminal of both oscillators.

microstrip line. The advantage of coupled microstrip lines instead of a single microstrip line is a higher input impedance phase slope and therefore a higher quality factor. The quality factor of both resonator circuits in dependence of the varactor voltage is depicted in Fig. 2. For the calculation of the quality factor, the circuit's input impedance in the proximity of the resonance frequency is modeled by a series resonant circuit. The calculated losses of both resonators are equal. The inductance of the resonant circuit is higher in case of the coupled microstrip resonator. The quality factor of the resonator circuit with coupled microstrip lines is twice as high as the quality factor of the single microstrip line resonator.

Another advantage is that it yields dc isolation of varactor and GaAs-MESFET gate without an additional capacitance, and it is suitable for monolithic integration.

Both oscillators were fabricated using hybrid thin film technology on semi-insulating GaAs-substrate. The microstrip lines are connected to the transistor pads without using bond wires [4].

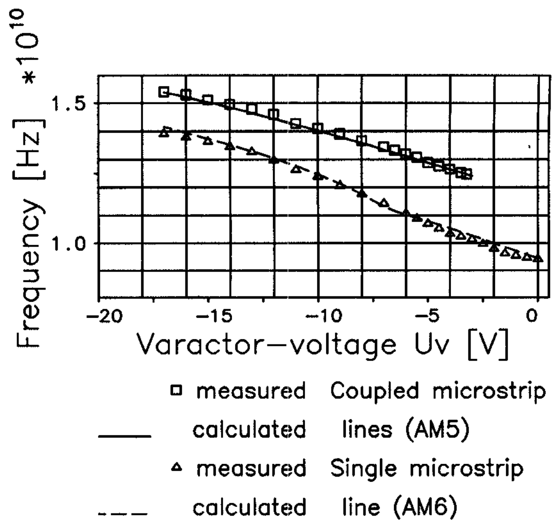


Fig. 3. The tuning characteristic of both oscillators.

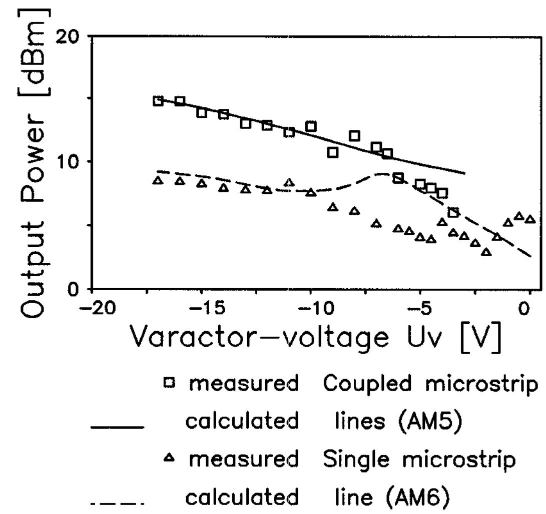


Fig. 4. Calculated and measured output power of both oscillators.

III. Modeling of the Unperturbed Oscillator

Tuning characteristics and output power are computed using a combined frequency and time domain method [5]. The oscillator circuit is divided into a linear and a nonlinear subcircuit. The linear subcircuit is described by a hybrid matrix in the frequency domain. The nonlinear subcircuit is represented by a set of first-order nonlinear differential equations in the time domain. The current voltage characteristics, the nonlinear capacitance, and the small signal equivalent circuit of the varactor diode are modeled according to [6]. The nonlinear current source of the GaAs-MESFET is modeled according to [7]. For the large signal equivalent circuit a modified SPICE-model is used [8]. Fig. 3 shows the tuning characteristic and Fig. 4 the output power of both oscillators. The tuning range of the oscillator with coupled microstrip line resonator amounts to nearly 3 GHz. In case of the oscillator with the single microstrip line resonator, a tuning range of 4.6 GHz is achieved. The output power of the oscillator AM6, measured with a HP71210C spectrum analyser, is between 4 and 9 dBm. Up to 5 dB higher output power is achieved by the oscillator with coupled microstrip lines.

IV. Low Frequency Noise of the Devices

For determination of the oscillators phase noise, the voltage-dependent low-frequency noise of the GaAs MESFET and the varactor diode must be investigated. This low-frequency noise is upconverted to the oscillation frequency. Voltage-controlled oscillators usually exhibit higher phase noise compared to fixed-frequency oscillators. One reason is the limitation of the resonator quality factor imposed by the requirement of a sufficiently high tuning range. An additional reason for higher phase noise of tunable oscillators is the noise originating from the series resistance of the varactor diode, which causes an increase of the frequency-independent white noise spectrum. The question whether the GaAs-MESFET or the varactor diode generates the dominant part of the frequency dependent $1/f^{\alpha}$-noise has to be answered.

The baseband noise of both devices was measured and noise sources were modeled in dependence of the gate-source- and source-drain-voltage for the GaAs-MESFET and the tuning voltage for the varactor, respectivly. The measurement was done using the HP3048 phase noise measurement system. The

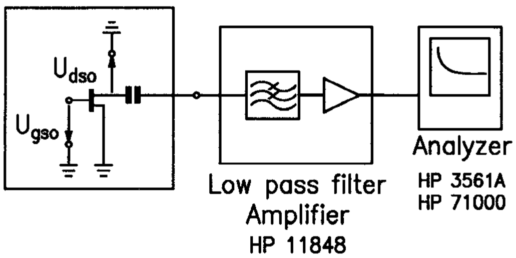


Fig. 5. Baseband noise measurement setup.

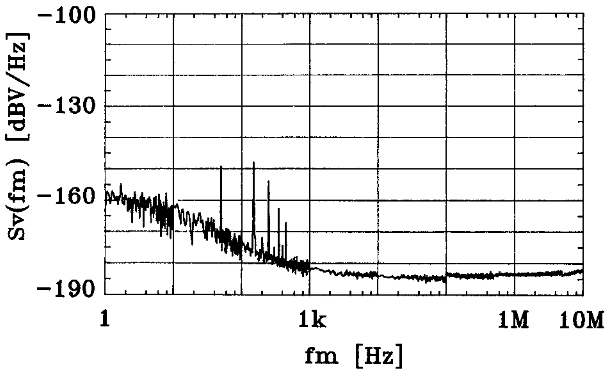


Fig. 6. Noise floor of the baseband measurement setup.

baseband noise measurement setup consists of the HP11848, including a low-pass filter and a low-noise amplifier. In the frequency range from 1 Hz to 100 kHz, the signal analyzer HP3561 is used. The measured and displayed signal quantity is the spectral power density of the voltage fluctuations

$$S_v(f) = 10 \cdot \log \frac{u_{\rm rms}^2}{1{\rm V}^2} [{\rm dBv}] \tag{1}$$

A value of -120 dBv corresponds to a noise voltage of 1 $\mu{\rm V}/\sqrt{\rm Hz}$. The frequency range of measurement is extended up to 10 MHz, using a spectrum analyzer HP71000. Fig. 5 shows the measurement setup. The noise floor of the baseband measurement system is depicted in Fig. 6. The spectral density is below -160 dBv up to an offset frequency of 1 kHz and about -180 dBv in the upper frequency range.

The low-frequency noise of the GaAs-MESFET was measured at several bias voltages. The source terminal of the GaAs-MESFET was grounded. At the drain terminal, the noise power was measured via a 100-μF coupling capacitor. The baseband noise measured at one bias point is shown in Fig. 7. $S_v(f)$ shows $1/f^\alpha$-behavior up to a cutoff frequency f_c, which is defined as intercept point between the $1/f^\alpha$-dependence domain and the white noise floor. The exponent α indicates the slope of the baseband noise. It is obtained as an average value of the slope of the spectral density $S_v(f)$ measured between 1 kHz and 100 kHz at various bias voltages. In the frequency range above 1 kHz, a decrease of 13 dBv/decade is measured. The average value of α amounts to 1.3. Because the intercept point f_c is above 10 MHz, the white noise floor does not occur in the baseband noise measurement results. The voltage dependent $1/f^\alpha$-noise is modeled using a voltage controlled

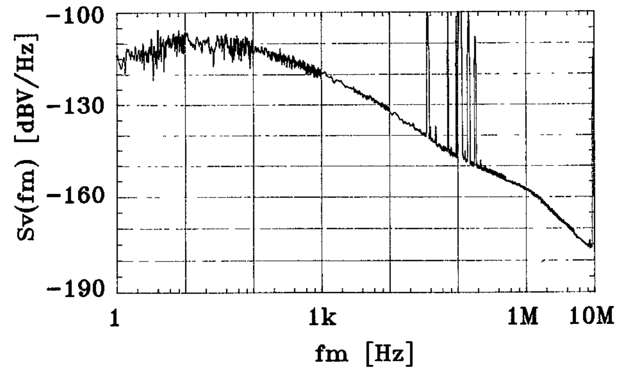


Fig. 7. Spectral density of the voltage fluctuations of the GaAs-MESFET at the bias voltages $U_{\rm ds0} = 3$ V, $I_{\rm ds} = 30$ mA, $U_{\rm gs0} = -0.7$ V.

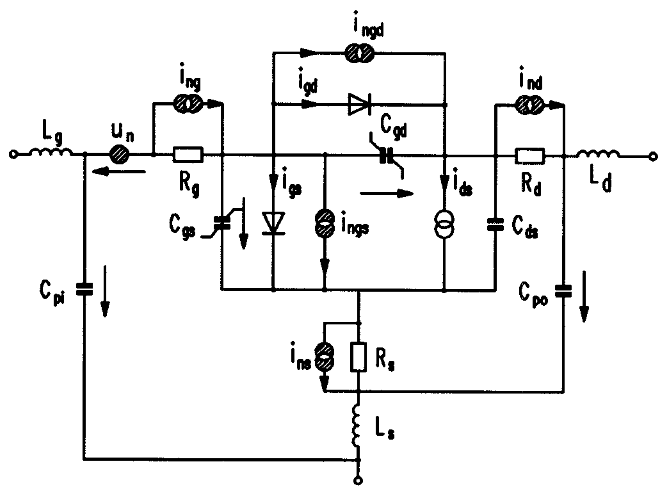


Fig. 8. Large-signal model of the GaAs-MESFET including the noise sources.

noise voltage source u_n at the gate terminal of the GaAs-MESFET. The model of the transistor including the noise sources is depicted in Fig. 8. For modeling the baseband noise the measured spectral power density $S_v(f)$ at a frequency of 10 kHz is taken into account. The voltage dependent behavior of $S_v(f)$ at a frequency of 10 kHz is shown in Fig. 9. The effective value u_n of the noise voltage is calculated in dependence on the measured spectral density, considering the transconductance and the drain-source conductance of the GaAs-MESFET:

For $U_{\rm ds0} \leq U_{{\rm ds}r}$ and $U_{\rm gs0} \geq U_{{\rm gs}r}$:

$$\begin{aligned} u_n = u_{n1} = {} & B_1 \cdot \tanh \\ & \times \left[U_{\rm ds0} \cdot B_2(1.0 + B_9(B_{10} - U_{\rm gs0})^2)\right] \\ & \times \left[B_4 + B_3(B_5 - B_8 U_{\rm ds0})^{(-2.0 B_6 (U_{\rm gs0} - B_7))}\right]. \end{aligned} \tag{2}$$

For $U_{\rm ds0} \leq U_{{\rm ds}r}$ and $U_{\rm gs0} < U_{{\rm gs}r}$

$$\begin{aligned} u_n = u_{n2} = {} & u_{n1}(U_{\rm ds0}, U_{{\rm gs}r}) \\ & + \frac{\delta u_{n1}}{\delta U_{\rm gs0}}\Big|_{(U_{\rm gs0} = U_{{\rm gs}r})} \cdot (U_{\rm gs0} - U_{{\rm gs}r}). \end{aligned} \tag{3}$$

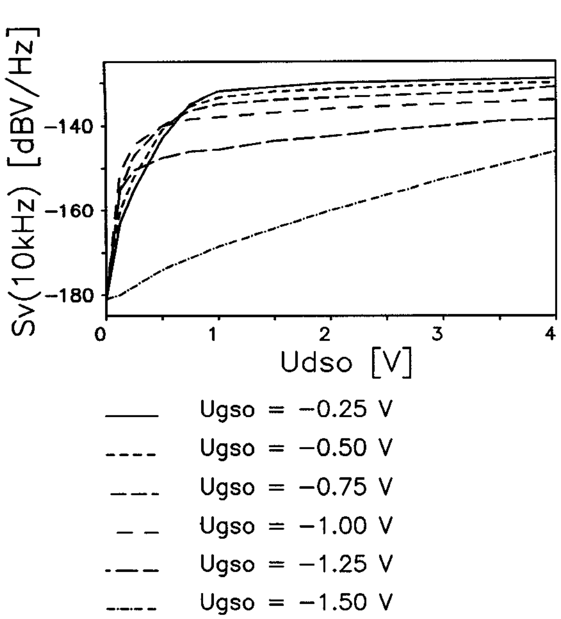


Fig. 9. The spectral density of the voltage fluctuations of the GaAs-MESFET in dependence on the bias voltages.

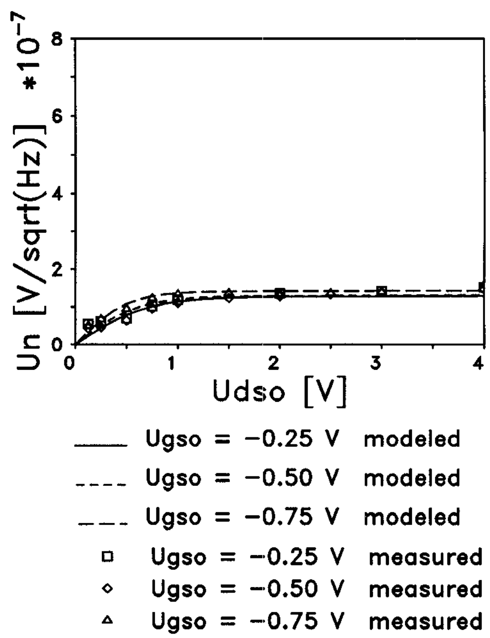
In case of not measurable bias points with $U_{\mathrm{ds0}} > U_{\mathrm{ds}r}$ the noise voltage source u_n is fixed at the value of u_n at $U_{\mathrm{ds0}} = U_{\mathrm{ds}r}$. The approximation coefficients are:

$$
\begin{array}{llll}
B_1 & = 0.58904 \cdot 10^{-7} \frac{V}{\sqrt{\mathrm{Hz}}} & B_2 & = 1.25763\ \mathrm{V}^{-1} \\
B_3 & = 0.29881 & B_4 & = 2.16383 \\
B_5 & = 27089 & B_6 & = 2.41680\ \mathrm{V}^{-1} \\
B_7 & = -0.76953\ \mathrm{V} & B_8 & = 1.54153\ \mathrm{V}^{-1} \\
B_9 & = 0.60428\ \mathrm{V}^{-2} & B_{10} & = 0.228108\ \mathrm{V} \\
U_{\mathrm{gs}r} & = -1.18500\ \mathrm{V} & U_{\mathrm{ds}r} & = 5.365380\ \mathrm{V}
\end{array}
$$

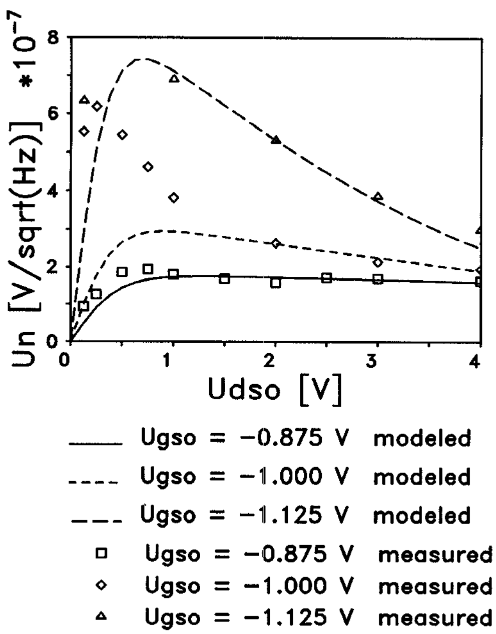
The voltage dependence of the noise voltage source is depicted in Fig. 10. In case of bias points with conductance above 0.1 mS, measured and approximated data agree within 4%.

The hyperabrupt varactor diode used in the oscillator circuits is reverse biased up to a voltage of −17 V. $1/f$-noise in reverse-biased diodes has been investigated in [9], being a surface noise, which is voltage independent and small compared to the white noise. For sufficiently large reverse bias, noisy leakage currents in excess of the saturation current occur and are modeled by a voltage dependent $1/f$-noise current generator across the diode junction. Low-frequency noise in reverse-biased pn-junction diodes caused by fluctuations of the carrier mobility are supposed [10]. Voltage noise spectra of pn-junction diodes were measured in the frequency range from 0.1 Hz up to 250 kHz at different bias voltages [11]. At voltages below 0.43 V, where generation-recombination processes dominate, a spectral density of voltage fluctuations $S_f \propto 1/f^{\alpha}$ of about $5 \cdot 10^{-14}$ V^2/Hz at a frequency of 1 Hz were measured. As well, a voltage dependent change of the exponent α from 0.65 to 1.0 was observed.

The low-frequency noise of the varactor diode was measured at several reverse bias voltages, using a coupling capacitor of 100 μF. The spectral density S_v at a bias voltage $U_v = -1$ V is plotted in Fig. 11. None of the measured baseband noise data show neither a spectral density above the noise floor of the baseband measurement system in Fig. 6, nor



(a)



(b)

Fig. 10. Calculated and measured $1/f^{\alpha}$-noise voltage source u_n.

a decrease proportional to $1/f^{\alpha}$-noise. So, the assumption is possible, that in case of the used GaAs-MESFET and varactor diode, the upconverted $1/f$-noise of the oscillator is generated by the $1/f$-noise of the GaAs-MESFET. The model of the varactor diode is depicted in Fig. 12. The noise sources are the shot noise of the diode (i_{nv}) and the thermal noise of the resistors.

V. Phase Noise Calculation

The determination of the single-sideband phase noise of the oscillators is done using the method described in [2]. This calculation uses pertubation methods to solve the Langevin equations that describe the deterministic and stochastic behav-

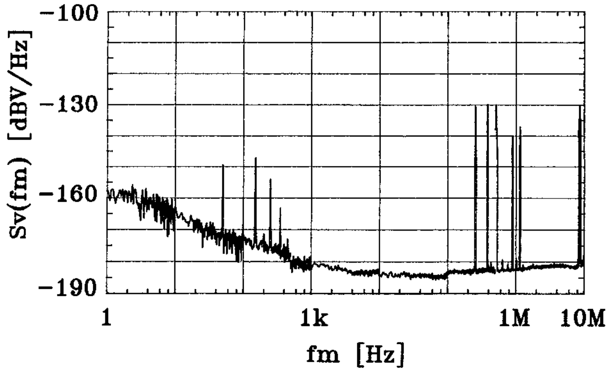


Fig. 11. Baseband noise of the varactor diode at a varactor voltage $U_v = -1$ V.

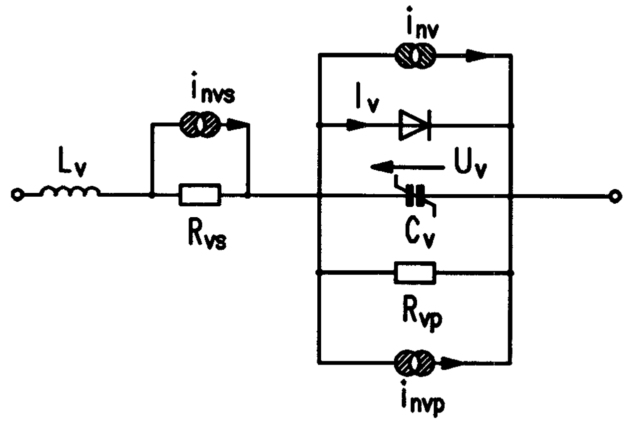


Fig. 12. Large signal model of the varactor diode including the noise sources.

ior of an oscillator. The oscillator circuit is described by a lumped element model together with the white- and $1/f^{\alpha}$-noise sources:

$$\dot{\mathbf{x}} = f(\mathbf{x}(t), \xi(t), y_1(t), \ldots, y_M(t)) \tag{4}$$

The components of the vector $\mathbf{x}$ are the state variables of the system. The vector ξ describes white gaussian noise sources with correlation functions

$$<\xi_i(t)\xi_j(t')> = \Gamma_{ij}\delta(t - t') \tag{5}$$

and $y_1(t)$ to $y_M(t)$ represents M $1/f^{\alpha}$-noise sources. For the GaAs-MESFET, the $1/f^{\alpha}$-noise source is characterized by the autocorrelation spectrum

$$C_{y_1 y_1} = \frac{c_1(U_{\mathrm{gs0}}, U_{\mathrm{ds0}})}{|2\pi f|^{\alpha_1}} \tag{6}$$

whereby $c_1(U_{\mathrm{gs0}}, U_{\mathrm{ds0}})$ is the modeled low-frequency noise spectral density of the GaAs MESFET at 10 kHz. In case of the varactor diode, the $1/f^{\alpha}$-noise source is characterized by the autocorrelation spectrum

$$C_{y_2 y_2} = \frac{c_2}{|2\pi f|^{\alpha_2}} \tag{7}$$

whereby c_2 is a constant value concerning to the voltage fluctuations spectral density of the varactor diode at 10 kHz.

The unperturbed steady state of an oscillator corresponds to a stable limit cycle in the phase space. Noise sources are small compared to the signal amplitudes and cause deviations from the limit cycle. In the time domain, the normal form equations of motion describing the noisy oscillator circuit are given by the linearization of the system equations (4) with respect to the noise sources

$$\dot{\mathbf{x}} = f(\mathbf{x}) + \mathbf{G}(\mathbf{x})\xi(t) + \sum_{l=1}^{M} \mathbf{g}^l(\mathbf{x}) y_l(t) \tag{8}$$

$$G_{ij} = \left. \frac{df_i(\mathbf{x}, \xi, y_1, \ldots, y_M)}{d\xi_j} \right|_{\substack{\xi = 0 \\ y_1 = \cdots = y_M = 0}} \tag{9}$$

$$g_i^l = \left. \frac{df_i(\mathbf{x}, \xi, y_1, \ldots, y_M)}{dy_j} \right|_{\substack{\xi = 0 \\ y_1 = \cdots = y_M = 0}} \tag{10}$$

The matrix $\mathbf{G}(\mathbf{x})$ and the vector $\mathbf{g}(\mathbf{x})$ are determined by the circuit topology of the lumped element oscillator model and the location of the noise sources in the active and passive devices; they depend on the state variables of the oscillator. Thus, the feedback of the oscillators state onto the noise sources is considered in the calculation, which causes a nonlinear modulation of the noise sources.

Previous to the determination of the phase noise of an oscillator the unperturbed steady state solution $\mathbf{x}^0(t)$ with the oscillation frequency f_0 must be calculated. To avoid numerical instabilities while calculating the phase noise spectrum at a small frequency deviation $f_m = f - f_0$, a consistent pertubation theory for nonself-adjoint systems is used [12]. The set of differential equations, (4), is linearized around the steady state of the system. The solutions of the linearized set of equations lead to a separation in randomly phase-shifted unperturbed solutions and amplitude deviations. The random phase deviations of the unperturbed steady state constitute the phase noise, which is the dominant contribution of noise in oscillators. Introducing the effective spectral densities of the noise sources, a correlation spectrum of the oscillator noise is calculated.

The single-sideband phase noise $L(f_m)$ is defined as the ratio of the noise power in a bandwidth of 1 Hz at a frequency deviation f_m from the oscillation frequency f_0 and the signal power at the oscillation frequency f_0. Applying the described method [2], the calculation of $L(f_m)$ results in:

$$L(f_m) = \frac{\Delta f_{3\,\mathrm{dB}}}{\pi f_m^2} + \frac{\omega_0^2 \cdot |g_{1,0}^1|^2}{(2\pi f_m)^{2+\alpha_1}} + \frac{\omega_0^2 \cdot |g_{1,0}^2|^2}{(2\pi f_m)^{2+\alpha_2}} \tag{11}$$

$\Delta f_{3\ \mathrm{dB}}$ denotes the 3-dB bandwidth of a Lorentzian line, generated by the random motion of the phase, and corresponds to the phase noise due to the white noise sources. The coefficients $g_{1,0}^1$ and $g_{1,0}^2$ only depend on the steady state solution of (8) and the given coefficients of vector $\mathbf{g}^l$, (10). They are a measure for the coupling of the lth $1/f^{\alpha}$-noise source to the phase noise and therefore determine how much of this low-frequency noise is upconverted to the oscillation frequency f_0 [2], [12]. Therefore, mixing and upconversion of the baseband noise due to the nonlinearities of the oscillator circuit are considered.

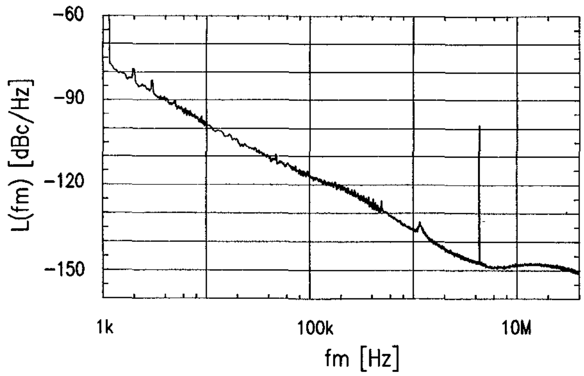


Fig. 13. Noise floor of the phase noise measurement system.

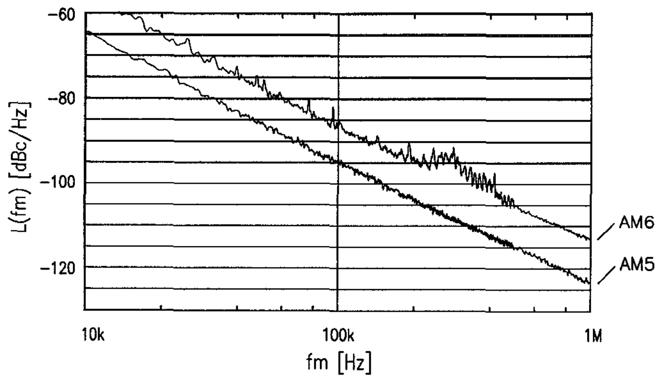


Fig. 14. Single-sideband phase noise of both oscillators at a varactor voltage of −4 V.

The resonator properties of a complex oscillator circuit are modeled by the lumped element equivalent circuit, considering the losses of the microstrip lines and of other circuit elements. In case of a basic oscillator model, a van der Pol oscillator, the single-sideband phase noise caused by white noise sources can be derived analytically [12]. The resonator quality factor Q and the output power P_0 can be included in the equation for $L(f_m)$. The dependence of $L(f_m)$ on Q and on P_0 is equal to Leesons [8]:

$$L(f_m) = \frac{\omega_0^2 \cdot kTM}{4\pi^2 f_m^2 \cdot P_0 Q^2} \tag{12}$$

The noise measure M is introduced in [12] to characterize the active device, where $(M-1)$ is the ratio between the spectral density of the noise of the active device and the noise of the losses.

Certainly, using the above described numerical method, parameters are calculated in dependence on the large signal model of the oscillator. Therefore the quality factor cannot be included in the equations for $L(f_m)$, but it is taken into account in the differential equations describing the equivalent circuit. In particular, the feedback of the oscillation onto the noise sources, which results in multiplicative noise and the modulation of the noise sources due to the nonlinearities of the active devices are considered in the calculation. In Leesons formula, the oscillator circuit noise is characterized exclusively by the noise figure of the active device, which is measured under small signal conditions.

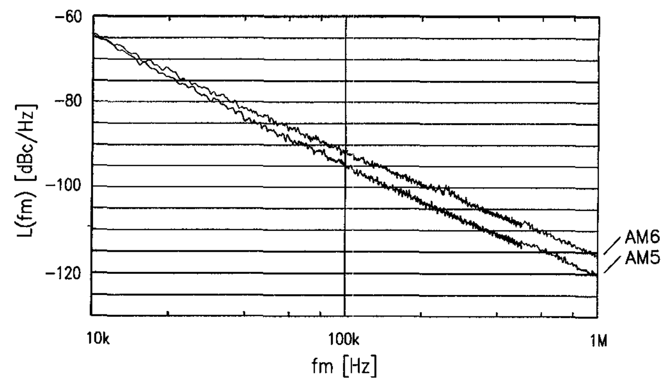


Fig. 15. Single-sideband phase noise of both oscillators at a varactor voltage of −17 V.

VI. Results

The single-sideband phase noise of the oscillators was measured with a HP3048 system using the frequency discriminator method [14]. The time delay of the discriminator is about 60 ns. The oscillator signal is downconverted to the frequency range of the phase detector, which is between 5 MHz and 1.6 GHz, using the carrier noise test set HP11729C. Maximum input frequency of the measurement setup is about 18 GHz, therefore a phase noise measurement at higher harmonics is not possible. The noise floor of the measurement system, measured with the synthesizer HP8662, is shown in Fig. 13.

Measured single-sideband phase noise of both oscillators is shown in Figs. 14 and 15, respectively. At a varactor voltage of −4 V, the single-sideband phase noise of the coupled microstrip line oscillator AM5 is about 8 dB below the phase noise of the oscillator with the single microstrip line resonator in the whole measured frequency range. Therefore, the lower phase noise occurs in the offset frequency range with dominating upconverted low frequency noise as well as in the range where white noise sources determine the phase noise. At a varactor voltage of −17 V, different phase noise occurs mainly in the range where white noise sources dominate. The single-sideband phase noise of the oscillator AM5 is about −95 dBc/Hz at a varactor voltage of −4 V, likewise −17 V. In contrast to this, single-sideband phase noise of the oscillator AM6 changes by variation of the tuning voltage. Calculated and measured phase noise of the oscillator with single microstrip line resonator (AM6) is depicted in the Figs. 16 and 17, respectively. Phase noise data agree within the accuracy of measurements. Deviations at offset frequencies above 3 MHz are due to the length of the delay line and the related measurement sensitivity. The measured voltage fluctuations at the output of the frequency discriminator in dependence on the frequency fluctuations are given by:

$$\Delta V(f_m) = \left(K_\phi 2\pi\tau_d \frac{\sin(\pi f_m \tau_d)}{(\pi f_m \tau_d)} \right) \Delta f(f_m) \tag{13}$$

To avoid having to compensate for the $\sin(x)/x$ response, measurements are made at offset frequencies (f_m) less than $1/2\pi\tau_d$. Using the delay line with $\tau_d = 60$ ns, measurements up to 2.6 MHz are possible without compensation. Calculated and

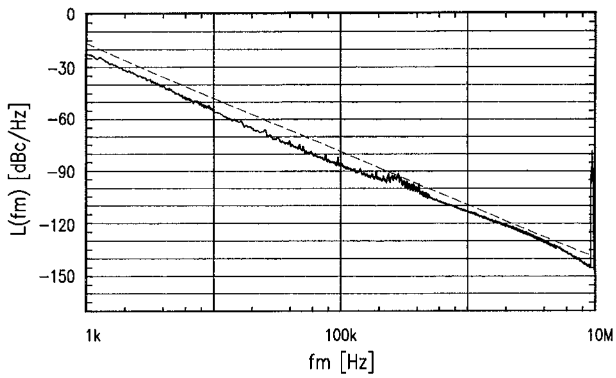


Fig. 16. Calculated (- — -) and measured single-sideband phase noise of the single microstrip line oscillator AM6 at a varactor voltage of −4 V.

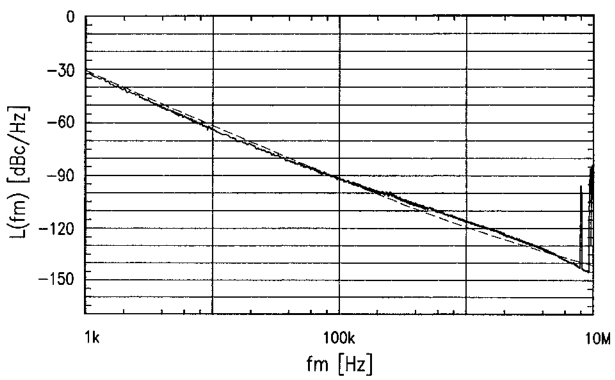


Fig. 17. Calculated (- — -) and measured single-sideband phase noise of the single microstrip line oscillator AM6 at a varactor voltage of −17 V.

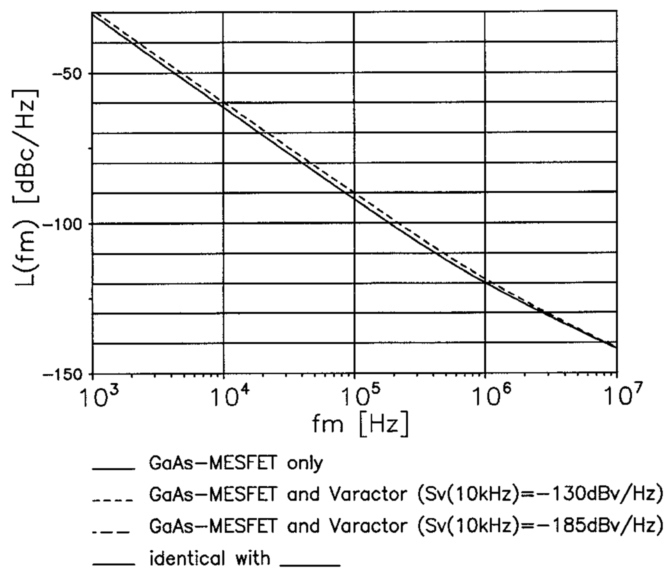


Fig. 18. Calculated single-sideband phase noise of the single microstrip line oscillator AM6 at a varactor voltage of −17 V with $1/f$-noise source of the varactor diode.

measured phase noise of the oscillator with coupled microstrip lines agree within the accuracy of measurements, too.

The phase noise results of oscillator AM6 calculated with Leesons formula differ from the numerical calculated value about an average of 4 dB. The calculation is based on measured noise figure data and oscillator output power as well as calculated unloaded resonator quality factor at varactor bias voltages of −4 and −17 V, respectively. At $U_v = -17$ V, the noise figure is about $F = 3.67$ dB, taking the admittance at the transistor input port into account. The oscillators output power comes to $P = 7.1$ mW and the resonators quality factor $Q = 25$. The single-sideband phase noise calculated using Leesons formula results in $L(f_m = 10\text{ kHz}) = -57$ dBc/Hz compared to a numerical calculated $L(f_m) = -61$ dBc/Hz and a measured $L(f_m) = -64$ dBc/Hz. Furthermore, the numerical calculation shows the right tendency by changes of the phase noise at different varactor voltages and different transistor bias operating points.

For investigation of the sensitivity of the phase noise on the varactor noise compared to the noise of the MESFET, an additional assumed $1/f^\alpha$-current noise source is included in the circuit parallel to the shot noise source of the varactor diode. The coefficient $\alpha = 1$ is chosen for calculation. Two different constant values of the spectral density of voltage fluctuations $S_v(f)$ are taken into account for the investigation, respectivly. One value is about −185 dBv/Hz at f =10 kHz, which is identical with the baseband measurement results. The second value is assumed to −130 dBv/Hz, which is in the range measured by Kleinpenning [10], and which corresponds to the maximum baseband noise of the GaAs-MESFET. The calculations were done at the two varactor voltages $U_v = -4$ and $U_v = -17$ V, respectively.

Additional phase noise due to the $1/f^\alpha$-noise source of the varactor diode only occurs in case of a varactor voltage $U_v = -17$ V. The calculated results of the oscillator AM6 are depicted in Fig. 18. Taking the measured spectral density $S_v(f) = -185$ dBv/Hz into account, there is no additional phase noise due to the $1/f^\alpha$-noise of the varactor diode. Both phase noise curves are identical. In case of the $1/f^\alpha$-noise source of the varactor diode with an assumed spectral density $S_v = -130$ dBv/Hz, the phase noise at an offset frequency of 100 kHz is about 2 dB higher. Due to the different coefficients α of both $1/f^\alpha$-noise sources, the slopes of the phase noise curves are slightly different. In the range of offset frequencies above 1 MHz, where white noise sources are dominant, both phase noise curves agree. Although the voltage fluctuations spectral density of the varactors $1/f^\alpha$-noise was taken in the maximum range of the GaAs-MESFET's $1/f^\alpha$-noise, the phase noise increase is only about 2 dB. Since the increase in phase noise is below 3 dB, this indicates that in case of spectral densities of the $1/f^\alpha$-noise sources of GaAs-MESFET and varactor diode in the same range, the sensitivity of the phase noise on the varactor $1/f^\alpha$-noise is lower than in case of the GaAs-MESFET.

VII. Conclusion

Signal and phase noise properties of two planar integrated tunable GaAs MESFET oscillators with different resonator circuits at the gate terminal of the transistor are calculated using nonlinear methods. The phase noise is calculated in the time domain using perturbation methods. The single-sideband phase noise of a varactor tunable microwave oscillator is reduced significantly to a value of −95 dBc/Hz, using a coupled microstrip line resonator instead of a single microstrip line resonator at the gate terminal of the transistor. Using a proper design, this resonator exhibits a quality factor twice as high as the quality factor of a single microstrip line resonator. Measured output power of the oscillator is about 12 dBm. In spite of the higher quality factor of this resonator circuit compared to a single microstrip line, a tuning bandwidth of more than 20% is achieved.

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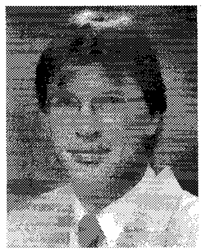

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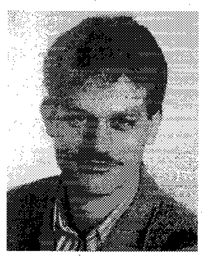

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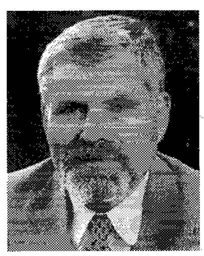

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