Low Phase Noise GaN HEMT Oscillator Design
based on High-Q resonators

Mikael Hörberg
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Photos of designed and analyzed oscillators

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Abstract

The thesis considers the design and optimization of oscillators targeting low phase noise, given boundary conditions from the technology. Crucial technology figures are power capability, RF noise figure, low-frequency noise and the quality factor (Q-factor) of the resonator. Parameters that can be optimized from a design perspective are the resonator coupling, bias point and waveforms. The technology used in this study is GaN-HEMT, due to its low RF noise figure, high power capability and good DC to RF efficiency.

The focus has been on the resonator coupling which is an essential part of the oscillator design. Strong coupling with high power transfer to the resonator improves the phase noise. Contradictory it will also decrease the loaded Q-factor of the resonator. The optimum coupling factor is found to be between $\beta=1/2$ and $\beta=1$, defined as the ratio of power dissipated in the resonator compared to the total power delivered by the active device.

Several designs in various resonator technologies have been investigated. For example, an oscillator based on an aluminum cavity connected to a GaN-MMIC reflection amplifier has a phase noise of $-145$ dBc/Hz at 100 kHz offset from a 9.9 GHz carrier. The analysis of the coupling’s effect to the cavity shows the optimum phase noise occurs for $\beta$ close to unity, which is equivalent to an open loop gain close to 0 dB. A MMIC oscillator based on the same reflection amplifier and a quasi-lumped on-chip-resonator has a phase noise of $-106$ dBc/Hz at 100 kHz offset from a 15 GHz carrier, which clearly shows that the phase noise scales with the Q-factor of the resonator.

A reflection amplifier with an electronically controlled gain is also designed for control of the resonator coupling. Varactors in the termination network perform a gain adjustment without changing the bias point of the active transistor. The phase noise of a cavity oscillator based on this reflection amplifier is $-136$ dBc/Hz at 100 kHz offset from 8.5 GHz. A similar oscillator with a mechanically tuned cavity has about 3 dB better phase noise. Despite a small degradation in phase noise, the simplicity facilitated with electronic tuning motivates this design for practical applications.

High-Q tunable elements are key components for frequency control. This work reports an ohmic cantilever radio frequency electromechanical system (RF-MEMS) integrated on a PCB forming a tunable ground plane inside a cavity. Vertical and horizontal positions of the MEMSs are investigated for trade-offs between tuning-range, frequency resolution and phase noise. Placing the PCB at 1 mm depth from the cavity wall, 5 % tunability around 10 GHz is reached, with 100 kHz phase noise ranging from $-140$ dBc/Hz to $-129$ dBc/Hz. Placing the PCB deeper into the cavity, at 2.5 mm, the tuning range can be increased to 12.3 %, with 100 kHz phase noise varying from $-133$ dBc/Hz to $-123$ dBc/Hz.

A varactor-tuned cavity oscillator has been implemented using the same PCB. It presents a tuning range of 1.6 %. The optimum phase noise at 100 kHz is ranging from $-111$ dBc/Hz to $-118$ dBc/Hz. At 1 MHz offset the phase noise is varying from $-138$ dBc to $-146$ dBc, versus the tuning-range.

GaN-HEMT devices from different commercial vendors have been used for the designs. For modeling purposes, low-frequency noise is measured for all devices. A special high-voltage and current low-frequency noise test setup was developed and used for benchmarking of different GaN HEMTs versus other technologies, e.g., GaAs InGaP HBTs and GaAs-HEMTs.

Keywords: GaN HEMT, MMIC, MEMS, phase noise, low-frequency noise, resonator coupling, cavity resonator, reflection oscillator
List of publications

Appended papers


Other papers


Thesis


As part of the authors’s doctoral studies, some of the work presented in this thesis has previously been published in [g]. Figures, tables and text from [g] may therefore be fully or partly reproduced in this thesis.
# Notations and abbreviations

## Notations

<table>
<thead>
<tr>
<th>Notation</th>
<th>Description</th>
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<tbody>
<tr>
<td>$\beta$</td>
<td>Coupling factor</td>
</tr>
<tr>
<td>$C$</td>
<td>Capacitance</td>
</tr>
<tr>
<td>$f_{1/f^2}$</td>
<td>Flicker noise corner frequency</td>
</tr>
<tr>
<td>$f_m$</td>
<td>Offset frequency</td>
</tr>
<tr>
<td>$f_0$</td>
<td>Carrier center frequency</td>
</tr>
<tr>
<td>$F$</td>
<td>Noise Figure</td>
</tr>
<tr>
<td>$g_m$</td>
<td>Transistor transconductance</td>
</tr>
<tr>
<td>$I_c$</td>
<td>Collector current</td>
</tr>
<tr>
<td>$I_d$</td>
<td>Drain current</td>
</tr>
<tr>
<td>$k$</td>
<td>Boltzmann’s constant</td>
</tr>
<tr>
<td>$L$</td>
<td>Inductance</td>
</tr>
<tr>
<td>$\cal{L}(f_m)$</td>
<td>Single sideband phase noise at offset $f_m$</td>
</tr>
<tr>
<td>$P_{AVO}$</td>
<td>Available RF power</td>
</tr>
<tr>
<td>$P_{RF}$</td>
<td>RF-power</td>
</tr>
<tr>
<td>$Q_0$</td>
<td>Unloaded Quality-factor</td>
</tr>
<tr>
<td>$Q_L$</td>
<td>Loaded Quality-factor</td>
</tr>
<tr>
<td>$R$</td>
<td>Resistance</td>
</tr>
<tr>
<td>$T$</td>
<td>Temperature in Kelvin</td>
</tr>
<tr>
<td>$V_c$</td>
<td>Collector voltage</td>
</tr>
<tr>
<td>$V_d$</td>
<td>Drain voltage</td>
</tr>
<tr>
<td>$\omega$</td>
<td>Angular frequency (rad/s)</td>
</tr>
<tr>
<td>$\Gamma_R,\Gamma_A$</td>
<td>Reflection coefficient to resonator, amplifier</td>
</tr>
<tr>
<td>$q$</td>
<td>Elementary charge</td>
</tr>
<tr>
<td>$Y_{in}$</td>
<td>Input admittance</td>
</tr>
<tr>
<td>$Z_{in}$</td>
<td>Input impedance</td>
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## Abbreviations

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
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<tbody>
<tr>
<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
</tr>
<tr>
<td>BJT</td>
<td>Bipolar Junction Transistor</td>
</tr>
<tr>
<td>CMOS</td>
<td>Complementary Metal Oxide Semiconductor</td>
</tr>
<tr>
<td>CW</td>
<td>Continuous Wave</td>
</tr>
<tr>
<td>DUT</td>
<td>Design Under Test</td>
</tr>
<tr>
<td>DR</td>
<td>Dielectric Resonator</td>
</tr>
<tr>
<td>DRO</td>
<td>Dielectric Resonator Oscillator</td>
</tr>
<tr>
<td>EVM</td>
<td>Error Vector Magnitude</td>
</tr>
<tr>
<td>FET</td>
<td>Field Effect Transistor</td>
</tr>
<tr>
<td>FFT</td>
<td>Fast Fourier Transform</td>
</tr>
<tr>
<td>FOM</td>
<td>Figure of Merit</td>
</tr>
<tr>
<td>GaAs</td>
<td>Gallium Arsenide</td>
</tr>
<tr>
<td>GaN</td>
<td>Gallium Nitride</td>
</tr>
<tr>
<td>HBT</td>
<td>Heterojunction Bipolar Transistor</td>
</tr>
<tr>
<td>HEMT</td>
<td>High Electron Mobility Transistor</td>
</tr>
<tr>
<td>HFSS</td>
<td>High Frequency Structure Simulator, Software tool from Ansys Inc.</td>
</tr>
<tr>
<td>InGaP</td>
<td>Indium Gallium Phosphide</td>
</tr>
<tr>
<td>LFN</td>
<td>Low Frequency Noise</td>
</tr>
<tr>
<td>LTE</td>
<td>Long Term Evolution</td>
</tr>
<tr>
<td>LTI</td>
<td>Linear Time Invariant</td>
</tr>
<tr>
<td>LTV</td>
<td>Linear Time Variant</td>
</tr>
<tr>
<td>MEMS</td>
<td>Microelectromechanical System</td>
</tr>
<tr>
<td>MIMO</td>
<td>Multiple Input Multiple Output</td>
</tr>
<tr>
<td>PLL</td>
<td>Phase Locked Loop</td>
</tr>
<tr>
<td>QAM</td>
<td>Quadrature Amplitude Modulation</td>
</tr>
<tr>
<td>SiC</td>
<td>Silicon Carbide</td>
</tr>
<tr>
<td>SiGe</td>
<td>Silicon Germanium</td>
</tr>
<tr>
<td>SIW</td>
<td>Substrate Integrated Waveguide</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal to Noise Ratio</td>
</tr>
<tr>
<td>SSB</td>
<td>Single Side Band</td>
</tr>
<tr>
<td>VCO</td>
<td>Voltage Controlled Oscillator</td>
</tr>
<tr>
<td>YIG</td>
<td>Yttrium-Iron-Garnet, ferromagnetic resonator</td>
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</tbody>
</table>
Contents

ABSTRACT .......................................................................................................................... iii

LIST OF PUBLICATIONS ................................................................................................. v

NOTATIONS AND ABBREVIATIONS ............................................................................. vii

1. INTRODUCTION ........................................................................................................... 1

2. CHALLENGES IN OSCILLATOR DESIGN .................................................................. 5
   2.1 GENERAL REQUIREMENTS ON OSCILLATORS .................................................. 5
   2.2 SYSTEM ASPECTS ................................................................................................. 7
   2.3 FUNDAMENTALS OF OSCILLATOR DESIGN ....................................................... 8
       2.3.1 Oscillator representation ............................................................................... 8
       2.3.2 Active part including noise generation ....................................................... 11
       2.3.3 Resonator ..................................................................................................... 15
   2.4 PHASE NOISE MODELS ....................................................................................... 17
       2.4.1 A linear theory of phase noise ..................................................................... 17
       2.4.2 Leeson’s equation ....................................................................................... 19
       2.4.3 Everard’s model .......................................................................................... 20
   2.5 RESONATOR COUPLING .................................................................................... 21
       2.5.1 A theory of resonator coupling ................................................................... 21
       2.5.2 Resonator coupling design impact ............................................................... 25
   2.6 FREQUENCY TUNING ......................................................................................... 26
       2.6.1 Varactor based tuning ................................................................................. 28
       2.6.2 MEMS based tuning .................................................................................... 30
   2.7 BENCHMARKING OSCILLATORS ......................................................................... 33
       2.7.1 Figures of merit ............................................................................................ 33

3. COMPONENT CHARACTERIZATION .......................................................................... 39
   3.1 ACTIVE DEVICE CHARACTERIZATION ............................................................. 39
   3.2 FLICKER NOISE CHARACTERIZATION ............................................................... 40
   3.3 CAVITY RESONATORS AND MODES .................................................................. 45
       3.3.1 General theory of rectangular waveguide resonators .................................. 45
       3.3.2 Tuning with an electronically controlled wall ............................................. 48
       3.3.3 Analog cavity tuning with varactors ............................................................. 50
       3.3.4 Varactor device characterization ................................................................. 51
   3.4 MEMS DEVICE PROPERTIES ............................................................................. 52
       3.4.1 Measured electrostatic behavior ................................................................. 52
       3.4.2 RF-MEMS model ........................................................................................ 54
       3.4.3 Electromechanic behavior .......................................................................... 55
       3.4.4 Potential improvements of used RF-MEMS ............................................... 56
3.5 Tuning Components and Technology Comparison ................................................................. 57

4. Oscillator Design and Characterization ................................................................. 59
  4.1 Experimental Setup ........................................................................................................ 59
  4.2 Cavity Based Oscillators .................................................................................................. 60
    4.2.1 Fixed frequency ......................................................................................................... 60
    4.2.2 Tunable frequency .................................................................................................... 62
  4.3 Flexible Hybrid Oscillator ............................................................................................. 66
  4.4 MMIC Based Oscillators ................................................................................................. 68
    4.4.1 Integrated oscillator .................................................................................................. 68
    4.4.2 Reflection amplifier with adjustable gain .................................................................. 70

5. Conclusions .......................................................................................................................... 73

Summary of Appended Papers ......................................................................................... 75

Acknowledgements ............................................................................................................ 77

References ............................................................................................................................ 79
Chapter 1.

Introduction

Ever since I was a child, I have been interested in radio equipment. It started in the early 1980’s, when I was around ten years old, and my father bought me an electronic hobby kit for simple experiments. The kit provided possibilities of building an AM-transmitter with a few tens meter range in the medium wave band using an ordinary transistor radio. Fascinated by the radio technology, and unaware about radio regulations, I tried to improve the range, transmit quality and to enable other frequencies. By that time, I had very little knowledge about limitations of performance in radio communication. Nevertheless, these were my first steps towards a career as radio developer at Ericsson and a Ph.D. in oscillator design at the Microwave Electronics Laboratory at Chalmers.

Evolution of the oscillator

In the early days of microwave engineering, the invention of the oscillator made it possible to build a signal source to verify Maxwell’s equations. Heinrich Hertz was a pioneer in this area, and with the construction of a spark gap generator, he could generate electromagnetic waves. The radiated signal was broad and covered from hundreds of Hz to several hundreds of MHz [1]. Hertz’s pioneering work was followed by several similar experiments elaborating with Maxwell’s theory. The oscillator circuits were studied in details, and models of the behavior were developed. The knowledge that the oscillator had to consist of a frequency selective resonator and an amplifying part to compensate for its loss was established as well as feedback theories for stability analysis and fulfillment of the oscillation criteria [2], [3], [4].

Historically, the parameters measured for oscillators were the frequency of oscillation, power and noise. Early experiments were based on trial and error, without deeper insight into the theory. Later the popular classic papers of how to synthesize ideal oscillators were written as from Colpitts [5], [6] and Clapp [7], [8]. An early paper to describe the noise properties was presented by Leeson in 1966 [9]. He defined many of the key parameters in oscillator design, such as the importance of the power coupled to the resonator, the amplifier noise figure, the quality factor of the resonator and the noise behavior versus oscillation frequency and offset frequency. Other important contributions were Kurokawa’s analysis of frequency stability [10], [11] and Rizzoli’s detailed analysis of noise sources and their contribution to phase noise [12]. Driven by needs from computer aided design (CAD) applications, transistor models were developed, e.g., Gummel-Poon for the bipolar transistors [13] and various models of field effect transistor (FET and JFET etc.) with noise properties [14], [15], [16], [17].
A different approach to models for general noise theory was presented by Lee and Hajimiri [18]. They considered the time-varying behavior of the current waveform in the transistor and formed a theory based on noise impulse functions of the oscillating periodic signal.

The improved transistor models and theory of noise generation together with better simulators, as microwave harmonic-balance for frequency domain analysis and different time-domain simulators, have been vital steps towards improved oscillators.

**Oscillators and phase noise in communication systems**

The ever-increased need of data capacity, as well as better coverage in modern communication systems, force the radio parts to have the cutting edge performance. The capacity entails a higher symbol rate, which involves a larger modulation bandwidth, or in reality with spectral bandwidth limitations, an improved spectral efficiency with a more advanced modulation format.

One of the main limitations of the spectral efficiency is spectral purity of the local oscillator (LO) used for frequency translation [19], [20]. Even for high-frequency systems, e.g. emerging standards as 5G, and millimeter-wave point-to-point communication, e.g., the E-band, the LO signal is generated at a relatively low frequency and subsequently multiplied, using frequency multipliers. Frequency multiplication will degrade near carrier phase noise as well as far carrier noise that is limited by the signal to noise ratio of the oscillator.

According to linear noise theory, e.g., Leeson’s equation [9], near carrier phase noise as well as far carrier noise floor benefit from a high power process such as wide bandgap GaN High Electron Mobility transistors (GaN-HEMTs). This is an emerging technology for power electronics due to its high breakdown voltage, high power capability, and good efficiency. A drawback with GaN-HEMTs is the high amount of flicker noise that up-convert to near carrier phase noise. However, in wideband systems, near carrier phase noise is less critical for capacity compared to far carrier noise. Nevertheless, near carrier noise must be sufficiently good to avoid phase slipping issues and problems in the clock recovery. These issues may be addressed by using a high-Q resonator so that the oscillator reaches good near-carrier phase noise, despite relatively high flicker noise.

The resonator coupling is essential for minimizing the phase noise [21]. Other critical issues are frequency tuning for PLL-locking and techniques for wide frequency reconfiguration. Any tuning or reconfiguration will inevitably degrade the spectral purity of the oscillator signal. High-Q tuning elements, as RF-MEMS switches [22] for digital control combined with weakly connected solid state varactors for analog tuning, is an interesting technology choice.

**Contribution of this work**

This thesis reports on GaN-HEMT oscillators where the coupling between the active device and the passive resonator is carefully optimized for best phase noise given the properties of the used components. All designs are based on AlGaN/GaN-HEMTs with very similar properties while the passive resonators are varied as well as the coupling methods. Moderate-Q integrated resonators, as well as high-Q cavity resonators, have been investigated. Further, trade-offs between tunability and phase noise are investigated.
Different tuning elements, e.g., MEMS and semiconductor varactors have been used. In all designs, the phase noise is analyzed carefully and benchmarked versus the theoretical noise floor [21], [23].

Some examples of designed oscillators are reported in Paper [A-G]. Paper [C] shows an aluminum cavity connected to a GaN-HEMT MMIC reflection amplifier. The coupling to the cavity is adjusted by changing the position of the cavity along a microstrip line, the optimum phase noise is showing a state-of-the-art performance for GaN-HEMT oscillators of -145 dBc/Hz at 100 kHz offset from a 9.9 GHz carrier. Paper [E] reports on the same MMIC-amplifier integrated with an on-chip-resonator using a coupling network based on a quasi-lumped impedance transformation. The phase noise is -106 dBc/Hz at 100 kHz from a 15 GHz oscillation frequency. Roughly 40 dB difference in phase noise can be explained by a factor 100 difference in Q-factor.

Mechanically tuning of the oscillator coupling as in Paper [C] is efficient but not production friendly. An alternative is to control the coupling factor by adjusting the amplifier gain. Paper [D] reports on an oscillator based on a reflection amplifier with electronic gain control implemented with varactors in the terminating network. A phase noise of -136 dBc/Hz at 100 kHz offset from an 8.5 GHz oscillation frequency is measured for this circuit, which is 3 dB worse compared to a mechanically tuned oscillator in the same process, demonstrating that electronic tuning is feasible without drastic degradation in performance.

The thesis also investigates tunable and reconfigurable cavity oscillators. Paper [B] reports on a MEMS-tuned cavity oscillator, where digital RF-MEMS-switches are embedded in the cavity resonator to create an electronically tunable ground plane. The concept is further investigated in Paper [A] where several cavity configurations are analyzed. Trade-offs between digital tuning range, frequency resolution and quality-factor of the resonator are discussed. This is performed by changing the position of the MEMSs in the horizontal and vertical position inside the cavity. An optimum performance of -123 dBc/Hz to -133 dBc/Hz with 12.3 % tunability around 10 GHz is reached. The RF-MEMS tuning can be combined with an analog tuning to enable PLL-locking. A method to couple a varactor inside the cavity with the same building practice is presented in Paper [G].

Paper [F] demonstrates a hybrid resonator with a lumped LC-tank connected to a single GaN-HEMT device, with a tuned input and output network. It is flexibly designed to study the up-converted phase noise from a bare die transistor device, the influence of operating point and flicker noise are discussed. It is found that far carrier noise improves with increased power while near carrier noise has an optimum for rather low DC power and degrades slightly for increased DC current associated with higher flicker noise.

**Thesis Outline**

The thesis is organized as follows. Chapter 2 presents background theory, design challenges, phase noise models and methods for benchmarking performance. Chapter 3 describes component and sub-block characterization used in the oscillator designs in this study, and Chapter 4 presents and discusses the complete oscillator designs investigated in this work. Finally, Chapter 5 concludes the thesis and discusses some future challenges.
Chapter 2.

Challenges in oscillator design

This chapter presents typical oscillator properties and a general circuit representation. Details about the individual building blocks are discussed with a focus on important trade-offs made by the designer, e.g., choice of active device and resonator. Linear phase noise models and optimization methods to improve phase noise are presented. In particular, the resonator coupling, and issues regarding different frequency tuning techniques are discussed. Finally, it demonstrates ways to benchmark oscillators and shows an overview of state-of-the-art published oscillators.

2.1 General requirements on oscillators
An oscillator is typically specified by a set of parameters, of which most commonly used are listed below with short description [1].

1. **Frequency range.** It is probably the most fundamental property of an oscillator. It defines the tunable range of oscillation frequencies. Typical oscillators have less than one octave tuning in frequency. Otherwise, several oscillator cores have to be combined and be provided with dividers or multipliers. The tuning is necessary for phase-locking to external references, and further, a wide tuning is often desired for, e.g., selection of oscillation sub-band.

2. **Output power.** A sufficient power level must be supplied to connected mixers in a system. Otherwise, external buffer amplifiers are needed. Typical drive levels needed for diode mixers are less than $<10$ dBm. Active mixers require lower level.

3. **Phase noise.** The purity of the signal is a fundamental property. Various noise sources in and outside the active part modulate the oscillator, resulting in energy and spectral distortions on both sidebands around the carrier. The noise contributes to amplitude modulation (AM) and frequency modulation (FM). Due to signal power compression, the active parts practically suppress the AM-noise, and the remaining part consists of FM-noise, which time integral is phase noise. For simplicity, the single-sideband (SSB) phase noise is used and described by power spectral density normalized to 1 Hz bandwidth.
Chapter 2 Challenges in oscillator design

4. **Harmonic suppression.** Harmonic suppression is important to avoid high frequency RF-leakage in distribution networks of signal power from the local oscillator to different consumers, e.g., mixers. Mixers for I and Q-modulation are sensitive for the phase balance, which can be disturbed by improper harmonic termination.

5. **Out-of-band spuriouses.** Out-of-band spuriouses not related to harmonics are often specified in regular requirements on spectral purity. A synthesizer in a phase-locked-system can generate digitally modulated spuriouses.

6. **Frequency pulling.** The oscillator must be insensitive to load variations, in particular, for reactive loads. A common remedy is to use extra buffer amplifiers. Pulling is often measured by a load change that exhibits non-unity VSWR overall phase variations. Pulling robustness is needed if several oscillators are combined in a system with potential RF-leakage and risks of frequency injection locking.

7. **Frequency pushing.** The oscillator must be robust to variations in power supply voltage and must have low impact on the frequency.

8. **Tuning characteristics.** This parameter describes the oscillation frequency dependence versus the tuning voltage. Ideally, a linear dependency is desired. At least, the oscillator must have a proper oscillation over the whole tuning range, especially for 0 V to enhance a proper start-up condition. A monotonic and continuous behavior is important in a phase-locked system.

   Large tuning sensitivity increases the up-conversion of modulation noise, which sets a strict requirement on the noise level on the tuning circuitry.

9. **Power consumptions and efficiency.** These parameters are always of interest, especially in multi-radio systems with several oscillators.

   The short-term behavior of the signal purity is usually defined as phase noise in the spectrum domain as Figure 2-1 (a), and the long-term frequency deviation is normally defined in the time domain as Figure 2-1 (b). The long-term drift is seldom of interest for a voltage controlled oscillator as it normally is locked to a long-time stable reference.

   In communication systems, the phase noise within the modulation bandwidth close to carrier affects the error-vector-magnitude (EVM) of modulated signals, while the far-out noise disturbs the selectivity by the reciprocal mixing of unwanted signals to the channel.
2.2 System aspects

The increased need for data capacity, e.g., in the microwave backhaul, demands an improved spectral efficiency. The available channel bandwidth limits the maximum number of symbols per second transmitted on a microwave carrier. The spectral efficiency is improved by coding each symbol by a larger number of bits as is illustrated in constellation plots in Figure 2-2 for modulation format of 2 bits and 10 bits, respectively.

The more complex the modulation scheme, the more sensitive the system is to imperfections in the radio transport, as rain, multi-path fading, etc. Furthermore, the spectral purity degrades due to imperfections such as noise in the receiver and nonlinearities in the transmitter. The receiver requires a certain signal-to-noise ratio (SNR) to minimize failures in signal detection. In particular, the phase noise in the frequency conversion of the radio chain is crucial and the signal sources are not allowed to fluctuate in frequency over time. Figure 2-4 illustrates the impact of phase noise on the measured capacity in Mbit/s for a radio link as a function of the constellation format of 2 to 11 bits, corresponding to 4-QAM to 2048-QAM, and channel bandwidth.
Figure 2-4 Measured channel capacity on a radio link in Mbits/s, plotted versus constellation format and channel bandwidth. The guided lines show different phase noise requirements at 100 kHz offset that must be fulfilled. The plot is valid for a constant SNR for the signal detection [24].

In reality, phase noise limits the available modulation for a certain bandwidth. Figure 2-4 presents boundaries indicating what modulation format that can be used for a given phase noise, signal bandwidth and capacity. The contours of the phase noise requirement indicate that performance for low offset frequency as 100 kHz is less dominating for a broadband signal. For broadband signals, the far-out noise floor is more dominant and must be kept low.

Other demands are on flexibility and reconfigurability for radio units to handle multiband, multistandard, etc. The trends go for software defined radios with generic hardware, which has to be cost effective, have low weight and size. This sets new requirements on the tuning technology in the frequency dependent parts, as for the oscillators. Tuning technologies that do not severely degrade the phase noise are necessary.

2.3 Fundamentals of oscillator design

2.3.1 Oscillator representation

An oscillator generates a periodic output signal by converting DC-power to RF-power. The oscillator contains an amplifier with a specific feedback path that at start up allows noise to grow and create a wanted signal. The noise is shaped by a highly filtering and selective positive feedback at the particularly wanted frequency, and it is amplified until it reaches saturation. In the closed loop, the gain will saturate to unity at a steady state, and the active device will only compensate for the losses in the feedback path. Signals not fulfilling the oscillation condition will be suppressed. If the oscillator suffers a transient disturbance, as pulling from a load variation, or as pushing from a change in bias supply, it will correct itself to the frequency fulfilling the oscillation condition.

The oscillator can be described either as a positive feedback system or a negative resistance (i.e., reflection amplifier) connected to a passive resonator. There are no strict differences between the two representations, and they can be used arbitrarily for representation of an oscillator. In the following chapters, there are some general characteristics of each type discussed.
Feedback oscillators

Figure 2-5 shows a positive feedback system.

\[
Y(j\omega) = X(j\omega) \frac{A(j\omega)}{1 - A(j\omega)H(j\omega)} \tag{2-1}
\]

The output signal in Figure 2-5 can be expressed as

The denominator should equal zero to cause an oscillation, and following conditions must be fulfilled, the so-called Barkhausen criterion

\[
|A(j\omega)H(j\omega)| > 1 \tag{2-2}
\]

\[
\angle A(j\omega)H(j\omega) = n360^\circ \tag{2-3}
\]

Negative resistance oscillators

Figure 2-6 shows the case of a negative resistance oscillator. In this respect, the amplifying active part is described as a one-port, with a reflection gain. The coupled resonator is also a one-port.

\[
|\Gamma_R\Gamma_A| > 1 \tag{2-4}
\]

\[
\angle \Gamma_R\Gamma_A = n360^\circ \tag{2-5}
\]

Topology

A negative resistance can be illustrated from a simple FET-model as in Figure 2-7 (a). In Figure 2-7 (b), the gate terminal is grounded, and a negative resistance appears on the drain terminal, which is derived in (2-6) to (2-8).
Chapter 2 Challenges in oscillator design

Figure 2-7 (a) FET-model. (b) FET with grounded gate.

Following expression can be extracted from Figure 2-7 (b)

\[
V_{in} = I_1 \left( \frac{R_{ds}}{j \omega R_{ds} C_{ds} + 1} \right) - V_{gs} = (I_{in} - g_m V_{gs}) \left( \frac{R_{ds}}{j \omega R_{ds} C_{ds} + 1} \right) - V_{gs}
\]  

(2-6)

\[
V_{in} = I_{in} \left( 1 + \frac{g_m}{j \omega C_{gs}} \right) \left( \frac{R_{ds}}{j \omega R_{ds} C_{ds} + 1} \right) + \frac{I_{in}}{j \omega C_{gs}}
\]  

(2-7)

and if \( R_{ds} \gg 1/\omega C_{ds} \), the expression is simplified to

\[
Z_{in} = \frac{V_{in}}{I_{in}} = \frac{1}{j \omega C_{gs}} + \frac{1}{j \omega C_{ds}} - \frac{g_m}{\omega^2 C_{gs} C_{ds}}
\]  

(2-8)

which shows a negative resistance combined with a serial coupling of \( C_{gs} \) and \( C_{ds} \).

As previously mentioned there is an analogy of seeing the oscillator as a feedback type or as a reflection type. Figure 2-8 demonstrates an example, where different groundings are used and, consequently, the view of the feedback paths is changed. The bias networks are not shown. Figure 2-8 (a) shows a feedback path of an 180-degree \( \pi \)-filter connecting the amplified signal back to input in a common emitter configuration. Figure 2-8 (b-c) shows the same design, but with the ground reference moved to the base side. Figure 2-8 (c) is equivalent with a common-base Colpitts oscillator. Finally, Figure 2-8 (d) shows a break-up of the resonator formed by an inductor connected to the capacitance network, principally defined by \( C_1 \) and \( C_2 \) in serial [25]. The latter network is similar to the transistor circuit in Figure 2-7 (b), and it will have a negative resistance into the collector, according to (2-8). If the intrinsic capacitances are neglected, the real part of the impedance is negative as a function of the transistor transconductance, real\( (Z_{in}) = -g_m/(C_1 C_2 \omega^2) \), which enables oscillation.
2.3.2 Active part including noise generation

**Noise sources in semiconductor devices**

The active device has a central role for the phase noise around the carrier. The device noise floor and the low-frequency noise (LFN) will due to up-conversion in the oscillation process create phase noise. This can be minimized by choosing a proper device technology.

There are two main types of transistors, the field effect transistor (FET) and the bipolar transistor (BJT). The first type is characterized of a conductive channel controlled by an electric field, while the latter type is bipolar and has two PN-junctions of charge carriers of electrons and holes. An example of FET is the high electron mobility transistor (HEMT). Common substrate technologies for HEMTs are GaN and GaAs. Examples of bipolar transistors are the heterojunction bipolar transistors (HBT) that commonly are processed in materials like GaAs, InGaP or SiGe. Typically, HEMTs have lower high-frequency noise level, but higher noise at low frequencies than the bipolar transistors.

Without going into details of semiconductor physics, following parts of the chapter will describe how the noise sources impact on phase noise in the oscillator application. Most focus is on the internal noise figures of the transistors. Other effects as the power capability of creating a better signal-to-noise-ratio, the drain efficiency and the linearity, are discussed for the oscillator designs in Chapter 4.

The noise sources with the most impact on oscillator phase noise coming from the active parts are thermal noise, shot noise, flicker noise and generation-recombination (GR) noise.

**Thermal noise**

Thermal noise described as white Gaussian noise has no bias dependence. The resistance of the circuit and its equivalent noise temperature determines the level of the noise floor. Thermal noise was investigated by Nyquist and Johnson [26], [27]. The noise is a distribution of voltages and currents in a network due to the thermal electron agitation, called Brownian motion. It can be modeled as
\[
\frac{<i_{\text{th}}^2>}{\Delta f} = \frac{4kT}{R}
\]

where \( R \) is the circuit resistance, and \( T \) the physical temperature or an equivalent noise temperature and \( \Delta f \) the bandwidth.

**Shot noise**

The discrete nature of charge flow can generate shot noise, because of the emission of charge carriers across a potential barrier, and it is therefore contributed in bipolar devices. It is a white and frequency independent noise, which increases with bias current. It can be modeled as,

\[
\frac{<i_{n,s}^2>}{\Delta f} = 2qI_{DC}
\]

where \( I_{DC} \) is the bias current, and \( q \) is the elementary charge.

**Flicker noise**

The conductance of semiconductors and metals can fluctuate in time and produce flicker noise. The reason for this is variation in the number of charge carriers and mobility in the presence of a DC-current. It is thus dependent on the bias current and the frequency. Models of flicker noise are presented in [16] and [17]. The flicker noise can be mathematically expressed as

\[
\frac{<i_{fl}^2>}{\Delta f} = \frac{KfI_{DC}^{A_f}}{f^{F_{fe}}}
\]

where the flicker noise coefficients \( K_f, A_f, \) and \( F_{fe} \), are used to curve fit measured data, while \( f \) is the frequency.

**Generation-Recombination (GR)-noise**

Fluctuations in the number of free carriers cause this kind of noise, where the carrier concentration typically can vary over several orders of magnitude. The noise may appear in transitions between the conduction band and other energy levels in the energy gap, in the conduction band or valence band, etc. GR-noise depends on bias current and frequency as

\[
\frac{<i_{n,gr}^2>}{\Delta f} = \frac{K_bI_{DC}^{A_b}}{1 + \left(\frac{f}{f_b}\right)^2}
\]

where \( I_{DC} \) is the bias current, \( K_b \) and \( A_b \) are fitting parameters. Frequency \( f \) is related to frequency \( f_b \) when GR-centers are activated [28].

The noise sources occur differently in a HEMT or a BJT. Figure 2-9 shows a noise model of a HEMT with a representation of their main noise sources. The gate resistance and drain resistance contribute most to thermal noise. Flicker noise and GR-noise appear across the channel, and they are represented as noise sources on the gate and the drain.
Thermal noise:

\[ \langle i_{nRg}^2 \rangle = \frac{4kT}{R_G} \Delta f \]

\[ \langle i_{nRd}^2 \rangle = \frac{4kT}{R_d} \Delta f \]

\[ \langle i_{nRgd}^2 \rangle = \frac{4kT}{R_{gd}} \Delta f \]

Flicker noise and GR-noise:

\[ \langle i_{nG}^2 \rangle = \frac{K_1 f^{A_{1f}}}{f^{E_{1f}}} \Delta f + \frac{K_{1b} f^{A_{1b}}}{1 + \left( \frac{f}{f_{1b}} \right)^2} \Delta f \]

\[ \langle i_{nd}^2 \rangle = \frac{K_2 f^{A_{2f}}}{f^{E_{2f}}} \Delta f + \frac{K_{2b} f^{A_{2b}}}{1 + \left( \frac{f}{f_{2b}} \right)^2} \Delta f \]

Figure 2-9 (a) Noise model of a HEMT with corresponding main noise sources [29]. (b) Expressions of the noise sources.

For a BJT, thermal noise appears due to the base resistance. Shot noise appears in the junction between the collector and the base. Figure 2-10 shows a noise model of a BJT with their corresponding main noise sources.
Challenges in oscillator design

Thermal noise:

\[ < i_{nrb}^2 >= \frac{4kT}{r_b} \Delta f \]

Shot noise:

\[ < i_C^2 >= 2qI_C \Delta f \]

Shot noise, flicker noise and GR noise:

\[ < i_B^2 >= 2qI_B \Delta f + \frac{K_{1f}I_B^{A_{1f}}}{f_{f_i}^2} \Delta f + \frac{K_{1b}I_B^{A_{1b}}}{1 + \left( \frac{f}{f_b} \right)^2} \Delta f \]

(b)

Figure 2-10 (a) Noise model of a BJT with corresponding main noise sources [30]. (b) Expressions of the noise sources.

Flicker noise and GR-noise have a dominant low-frequency (LF) behavior. LFN-measurement is important in device characterization as it can reveal imperfections in the semiconductor process. Some imperfections can cause traps and surface states for the electrons, which lead to LFN. LFN-measurement is commonly used for benchmarking the technology. The phase noise close to the carrier in oscillators is dependent on LFN which is up-converted in the oscillation process.

The amount of LFN varies with transistor technology. A commonly used parameter is the flicker noise corner-frequency, \( f_c \), which defines the frequency where the LFN has the same magnitude as the white noise floor. It should be mentioned that there are uncertainties to estimate the device noise floor as the measurement noise floor also contributes. Devices with high RF-noise figure can by this definition get a lower flicker noise corner-frequency. This argues to handle this parameter carefully. The flicker noise corner-frequency is illustrated in Figure 2-11. Vertical structures like HBTs generally have lower LF-noise than FETs and HEMTs which are more exposed in the surface. Traps and surface states for the electrons are more frequent in laterally structured devices. For example, InGaP/GaAs HBTs have LF corner-frequencies of about 20 kHz, while in the case of GaN-HEMT it can be about 1 MHz. The lower estimated flicker noise corner frequency for bipolars can also be due to the higher shot noise in these devices. Paper [H] presents detailed LFN measurements comparing different transistor technologies.

Figure 2-11. Low-frequency noise and the device noise floor combined with the measurement system noise floor.

The choice of the active part depends on where the noise performance is most critical to optimize. A trade-off between low flicker-noise and low noise floor compared to the power of the carrier is often necessary. Figure 2-12 shows an illustration of phase noise profiles
for typically sized and biased GaN-HEMTs and silicon BJTs, respectively, with a resonator in PCB-technology (Q≈100). GaN devices used in this study are beneficial for their low noise floor far out, and their high power efficiency and power capability perform an excellent signal-to-noise ratio. The drawbacks of the higher flicker noise compared to other technologies will become less important for broadband signals that are expected in future communication systems. The targeting application for broadband communication in this study identifies the noise floor as most critical [19]. In this perspective GaN-technology would be a good choice.

Figure 2-12 Typical phase noise profiles with frequency slope stated for GaN-HEMT with high flicker noise compared to silicon BJTs with less flicker noise but with higher noise floor with a resonator on PCB.

2.3.3 Resonator

Q-factor of the resonator

Passive circuits forming the resonator in an oscillator are often characterized and benchmarked regarding a quality factor (Q-factor). In a resonator, the energy alternates between electrically and magnetically stored energy. The Q-factor of a resonator can be expressed as the ratio between stored energy and the energy loss per cycle, i.e., average power loss,

\[ Q_{def} = 2\pi \frac{\text{Stored energy}}{\text{Energy dissipation per cycle}} = \omega \frac{\text{Stored energy}}{\text{Average power loss}} \]

The magnetic energy \( W_m \) is maximized when the electrical energy \( W_e \) is zero and vice versa, i.e., \( W_{\text{stored}} = W_{e,\text{peak}} = W_{m,\text{peak}} \). Measurement of the Q-factor according to the general definition is quite impractical, and it is often necessary to use some of the following derived methods. The first method is using the 3 dB bandwidth of the frequency response of the power transmission, or the power reflection at the center frequency \( \omega_0 \).

\[ Q = \frac{\omega_0}{\Delta\omega_{3dB}} \quad (2-13) \]

Another relation is the phase slope of the input impedance. For a parallel resonator,

\[ Q = -\frac{\omega_0}{2} \left. \frac{\partial \Phi}{\partial \omega} \right|_{\omega=\omega_0} \quad (2-14) \]

For a serial resonator the phase slope shifts sign, and the quality factor can be expressed
\[ Q = \frac{\omega_0}{2} \frac{\partial \phi(Z_s)}{\partial \omega} \bigg|_{\omega = \omega_0} \]  

(2-15)

If an electrical model of the resonator is used, the Q-factor of the resonator can be calculated directly from the component values,

\[ Q = \frac{\omega_0 L_s}{R_s} = \frac{1}{\omega_0 C_s R_s} \]  

(2-16)

\[ Q = \frac{R_p}{\omega_0 L_p} = \frac{\omega_0 C_p R_p}{1} \]  

(2-17)

(a) Serial resonance  
(b) Parallel resonance

Figure 2-13 (a) Schematics of a serial resonance circuit and (b) a parallel resonance circuit.

**Loaded Quality factor**

When resonators are not connected or very weakly connected to a load, they are characterized by the unloaded Q-factor, \( Q_0 \). In reality, this metric is often impractical, as resonators always have to be connected to some device injecting power into it, or to couple out a wanted signal from it. In the case of an oscillator, this means that it must connect to an active device, thus be loaded by its internal resistance. This load network can also be characterized by an external Q-factor, \( Q_{\text{ext}} \). The total loaded Q-factor, \( Q_L \), for the network is then calculated as

\[ \frac{1}{Q_L} = \frac{1}{Q_0} + \frac{1}{Q_{\text{ext}}} \]  

(2-18)

**Design impact**

The unloaded quality factor, \( Q_0 \), of a resonator is of importance for effective filtering with high frequency selectivity, that will shape the noise around the carrier. Dependent on the building practice, different types of resonators are used, e.g., dielectric resonators as used in DROs, metal cavities, lumped components, distributed transmission lines as resonant stubs and crystals. The technology choice is affected by the Q-factor needed, power handling of the resonator, frequency tunability, etc.

<table>
<thead>
<tr>
<th>Example of resonator</th>
<th>Unloaded Q-factor (( Q_0 )) Typical values</th>
<th>Frequency (GHz) Typical values</th>
</tr>
</thead>
<tbody>
<tr>
<td>DR (dielectric resonator)</td>
<td>10 k-25 k</td>
<td>&lt;50</td>
</tr>
<tr>
<td>YIG (yttrium-iron-garnet)</td>
<td>1 k-10 k</td>
<td>&lt;50</td>
</tr>
<tr>
<td>Cavity aluminum (without silver plating)</td>
<td>3000</td>
<td>&lt;40</td>
</tr>
<tr>
<td>Microstrip on low loss PCB</td>
<td>100</td>
<td>&lt;40</td>
</tr>
<tr>
<td>Lumped LC</td>
<td>100-300</td>
<td>&lt;10</td>
</tr>
<tr>
<td>MMIC (III-V-substrate)</td>
<td>40</td>
<td>&lt;100</td>
</tr>
</tbody>
</table>
There are some considerations to design an oscillator and to utilize the resonator and the Q-factor in an optimum way.

- The phase condition for the closed loop, \( n \times 360^\circ \), should occur for maximum phase slope \( (\partial \Phi(Z)/\partial \omega) \). This is further discussed when tunability is added in Chapter 2.6.

- The coupling of the resonator to the active device will decrease the total loaded Q-factor, \( Q_L \), with a negative impact on phase noise, which is contradictory to that high power coupled to the resonator is beneficial. An optimum coupling exists which is discussed in Chapter 2.5.

### 2.4 Phase noise models

Modeling phase noise has been continuing for several decades and is still not a completely understood research area. The complexity depends on the nonlinear way the noise from different sources is up-converted around the carrier, which differs from the linear noise figures in example amplifiers. Besides the linearity aspects, there are considerations about modeling the noise sources as stationary and time-invariant sources, or as time-variant meaning that they will depend on the current and voltage trajectory waveform across the transistor. The latter approach implies a cyclo-stationary model. Limiting to the linear theory, the two most used models are the linear time-invariant model (LTI) by Leeson [9], and the linear time-variant model (LTV) by Hajimiri [18]. In this study, we have focused on the time-invariant linear models only for the study of optimizing the resonator coupling, power handling and noise figures of the active device. The limitation is that the up-conversion of flicker noise will not be accurately modeled [31], as it is in a cyclo-stationary process where the noise level is dependent of the current waveform across the transistor. Phase noise far out is less dominated by flicker noise and will be predicted fairly well by the time-invariant models originated from Leeson.

#### 2.4.1 A linear theory of phase noise

A general feedback oscillator system can be described as [25].

![Figure 2-14 General oscillator feedback system](image)

The transfer function can be written as

\[
\frac{Y(j\omega)}{X(j\omega)} = \frac{H(j\omega)}{1 - H(j\omega)}
\]  

(2-19)

The Taylor expansion around the oscillation frequency \( \omega_0 \) can be expressed as
where at resonance $H(j\omega_0) = 1$ and for small offset frequencies $\Delta \omega$, $|\Delta \omega \frac{dH}{d\omega}| \ll 1$, the following expression can be written

\[
\frac{Y}{X}(\omega_0 + \Delta \omega) \approx \frac{-1}{\Delta \omega \frac{dH}{d\omega}} \tag{2-21}
\]

The power ratio can be expressed as

\[
\left| \frac{Y}{X(\omega_0 + \Delta \omega)} \right|^2 = \frac{1}{(\Delta \omega)^2 \left| \frac{dH}{d\omega} \right|^2} \tag{2-22}
\]

The forward function in the loop can be assumed on the form

\[
H(\omega) = |H|e^{j\emptyset} \tag{2-23}
\]

This is equal to the impedance transfer function for a simple LC-network and following expression can be written

\[
\frac{dH}{d\omega} = \left[ \frac{d|H|}{d\omega} + j|H| \frac{d\emptyset}{d\omega} \right] e^{j\emptyset} \tag{2-24}
\]

The squared absolute value can be expressed as

\[
\left| \frac{dH}{d\omega} \right|^2 = \left| \frac{d|H|}{d\omega} \right|^2 + \left| \frac{d\emptyset}{d\omega} \right|^2 |H|^2 \tag{2-25}
\]

For standard LC-oscillators $|d|H|/d\omega|^2$ is much smaller than $|d\emptyset/d\omega|^2$. Furthermore, $|H|$ is close to unity at stationary oscillation. By using these simplifications and inserting (2-25) in (2-22), the ratio $Y$ and $X$ can be reformulated as

\[
\left| \frac{Y}{X} (\Delta \omega) \right|^2 = \frac{\omega_0^2}{4(\Delta \omega)^2} \frac{1}{\omega_0^2 \frac{d\emptyset}{d\omega}^2} = \frac{1}{4Q_L^2} \left( \frac{\omega_0}{\Delta \omega} \right)^2 \tag{2-26}
\]

The definition of the loaded Q-factor according to (2-15) is inserted. Under the assumption that the noise power at input before the active device is $|X(j\omega)|^2 = kT$, the filtered noise expression becomes

\[
|Y(\Delta \omega)|^2 = \frac{kT}{4Q_L^2} \left( \frac{\omega_0}{\Delta \omega} \right)^2 \tag{2-27}
\]

The expression can be normalized to the carrier signal power built up in the same node, i.e., at the node for the power before the active device, $P_{in}$. If a single-sideband signal is considered, the noise level can be reduced to half of the value, and the noise-to-signal power ratio for offset frequency $f_m$ can be expressed as

\[
\mathcal{L}(f_m) = \frac{|Y(\Delta \omega)|^2}{2P_{in}} = \frac{kT}{2P_{in}} \left( \frac{f_0}{2Q_L f_m} \right)^2 \tag{2-28}
\]
2.4.2 Leeson’s equation

Leeson used the linear phase noise theory and incorporated additional parameters for flicker noise corner $f_{1/f^3}$, the noise floor, and an empirical noise figure $F$ to (2-28), and expressed the phase noise as [9]

$$L(f_m) = 10 \log_{10} \left[ \frac{FkT}{2P_{in}} \left[ 1 + \left( \frac{f_0}{2Q_Lf_m} \right)^2 \right] \left[ 1 + \frac{f_{1/f^3}}{f_m} \right] \right] \quad (2-29)$$

which shows the normalized single-sideband for an offset frequency $f_m$. Leeson originally formulated an expression with respects to a feedback oscillator where the power $P_{in}$ denotes the power at the input of the amplifier after the loss in the resonator. At optimum resonator coupling, $Q_L/Q_0 = 1/2$, which later will be derived, the power dissipated in the resonator $P_r$ is half of the total power dissipated in the loop $P_{RF}$, or in other words, twice the power transmitted by the resonator thus $P_r = 2P_{in}$. The power at the resonator input is equal to the output of the amplifier, which is equal to the input power of the amplifier multiplied by its power gain, thus $P_{out} = G_A P_{in} = 4P_{in} = 2P_r$. The power gain is calculated from the gain needed for oscillation, $G_A = 1/|1 - Q_L/Q_0|^2$. The discussed power level in the feedback loop is illustrated in Figure 2-15 [32].

$$P_r = Q_d/Q_d P_{RF}$$

At optimum:

- $Q_d/Q_0 = 1/2$
- $P_r = P_{RF}/2$
- $P_{in} = P_r/2$
- $G_A = 4$
- $P_{out} = 4P_{in} = 2P_r = P_{RF} = P_{AVO}$

(P$_r$ = dissipated power in the resonator)
(P$_{RF}$ = total dissipated power in the loop)

![Figure 2-15 Power levels in the loop](Image)

Using power dissipated in the resonator, the single-sideband phase noise can be expressed as

$$L(f_m) = 10 \log_{10} \left[ \frac{FkT}{P_r} \left[ 1 + \left( \frac{f_0}{2Q_Lf_m} \right)^2 \right] \left[ 1 + \frac{f_{1/f^3}}{f_m} \right] \right] \quad (2-30)$$

Alternatively, if the power at the amplifier output is used, the same single-sideband phase noise can be expressed as

$$L(f_m) = 10 \log_{10} \left[ \frac{2FkT}{P_{out}} \left[ 1 + \left( \frac{f_0}{2Q_Lf_m} \right)^2 \right] \left[ 1 + \frac{f_{1/f^3}}{f_m} \right] \right] \quad (2-31)$$

Figure 2-16 (a-b) illustrate the empirical model for different cases: for a high $Q_L$-factor, a low $Q_L$-factor, compared to the $f_{1/f^3}$-factor.
Figure 2-16 (a) Phase noise behavior, where the resonator bandwidth due to high Q falls inside the flicker noise corner, $f_{1/f}$. (b) Phase noise behavior, where the bandwidth due to low Q falls outside the flicker noise corner.

The single-sideband phase noise floor given by $L_{floor} = \frac{FkT}{2P_{in}}$ is discussed in [33]. The noise level of -177 dBc/Hz for a 0 dBm signal has been verified in experiments [34]. The reason for half of the thermal noise energy relatively the carrier is explained by the definition of single-sideband [32]. Within the resonator bandwidth $f_m = \frac{f_0}{2Q_L}$, the amplitude noise is suppressed by amplitude compression, but the far-out noise floor is contributed by both phase and amplitude noise in equilibrium [35].

### 2.4.3 Everard’s model

Everard expressed the phase noise in a closed form as a function of coupling factor regarding the quote $Q_L/Q_0$ for feedback type oscillators. For the optimum coupling condition, the minimum phase noise is expressed both for feedback type and for a reflection type or a negative resistance type oscillator.

#### Feedback type oscillator

A phase noise model according to [36], [23] is described as

$$L(f_m) = 10\log_{10} \left[ A \frac{FkT}{8(Q_0)^2(Q_L/Q_0)^2(1 - Q_L/Q_0)^N P(f_0/f_m)^2} \right]$$

(2-32)

where for the cases 1-3:

1. $N=1$ and $A=1$ if $P=P_{RF}$ and $R_{OUT}=0$, i.e. $P$ is power dissipated in both resonator and the internal resistance $R_{IN}$, in Figure 2-17.

2. $N=1$ and $A=2$ if $P=P_{RF}$ and $R_{OUT}=R_{IN}$, i.e. $P$ is power dissipated in both resonator and the internal resistances, $R_{OUT}$, and $R_{IN}$ in Figure 2-17.

3. $N=2$ and $A=1$ if $P=P_{AVO}$ and $R_{OUT}=R_{IN}$, i.e. $P$ is the available power from the amplifier at the output terminated in a matched load in Figure 2-17.

Case 2 can be used with $P=P_{RF}$, $R_{OUT}=R_{IN}$ for the optimum resonator coupling of $Q_L/Q_0=1/2$ which is derived in 2.5.1.
The quote is corresponding to $P_r = P_{RF}/2$ (power dissipated only in the resonator), and it can be evaluated to an expression for minimum phase noise for a feedback oscillator as

$$\mathcal{L}(f_m) = 10 \log_{10} \left[ \frac{FkT \left( \frac{f_0}{f_m} \right)^2}{Q_0^2 P_r} \right]$$

which is the same as stated in (2-28) for $Q_0 = 2Q_L$ and $P_r = 2P_{in} = P_{AVO}/2$.

**Negative resistance type oscillator**

The minimum phase noise for a reflection type oscillator is in [21] derived to

$$\mathcal{L}(f_m) = 10 \log_{10} \left[ \frac{FkT \left( \frac{f_0}{f_m} \right)^2}{2Q_0^2 P_r} \right]$$

where $P_r$ is representing the power dissipated only in the resonator.

### 2.5 Resonator coupling

Physical boundaries as the resonator Q-factor, the device noise factor and power capability are important figures as earlier discussed. Important is also to find a proper coupling between the resonator and the active device. This chapter shows the derivation of the optimum coupling in detail from [8], [37], [32], [38]. The theory assumes only thermal noise with Q-multiplication effect which is valid close to the phase noise skirts. Other optimums are found if flicker-noise is considered, and furthermore, a different optimum is found when minimizing the far out noise. Also, the system noise figure, $F$, is assumed to be constant (versus different impedance) during this optimization.

#### 2.5.1 A theory of resonator coupling

**Feedback oscillator**

Figure 2-17 depicts a model of a feedback type oscillator.
If $R_{in} = R_{out} = Z_0$ is assumed, the following expression can be written for the loaded $Q$-factor

$$Q_L = \frac{\omega L}{R_{Loss} + R_{in} + R_{out}} = \frac{\omega L}{R_{Loss} + 2Z_0} = Q_0 \left( \frac{1}{1 + \frac{2Z_0}{R_{Loss}}} \right)$$

(2-35)

The ratio to the unloaded $Q$-factor is calculated by defining a coupling factor $\beta$ as

$$\frac{Q_L}{Q_0} = \left( \frac{1}{1 + \frac{2Z_0}{R_{Loss}}} \right) = \frac{1}{1 + \beta}$$

(2-36)

The rms-voltage over the resonator can be expressed as

$$V_r = V_s \frac{R_{Loss}}{R_{Loss} + 2Z_0} = V_s \frac{Q_L}{Q_0}$$

(2-37)

Maximum available power from the amplifier at output can be expressed as

$$P_{avo} = \frac{V_s^2}{2 \cdot 2Z_0}$$

(2-38)

and the power dissipated inside the resonator is formulated when inserting (2-37) and (2-38) as

$$P_r = \frac{V_r^2}{R_{Loss}} = \frac{V_s^2 Q_L^2}{Q_0^2 R_{Loss}} = \frac{P_{avo} 4Z_0 (Q_L/Q_0)^2}{Q_L} = 2P_{avo} \left( \frac{Q_0}{Q_L} - 1 \right) \left( \frac{Q_L}{Q_0} \right)^2$$

(2-39)

If (2-39) is inserted in (2-32) for case 3, the best phase noise is achieved when maximum power is dissipated in the resonator. The optimum in terms of $Q_L/Q_0$ is when [21]

$$\frac{Q_L}{Q_0} = \frac{1}{2}$$

(2-40)

or if considering the insertion loss expression,

$$S_{21} = (1 - \frac{Q_L}{Q_0})$$

(2-41)

An interpretation of this is that half of the available power is dissipated in the resonator, one quarter is reflected, and one quarter is transmitted ($=|S_{21}|^2$) which was illustrated in
Figure 2-15. The best phase noise according to Everard’s model (2-32) for the feedback case with $R_{in}=R_{out}$, can be summarized to

$$\mathcal{L}(f_m) = \frac{2FkT}{P_{av0}Q_0^2} \left( \frac{f_0}{f_m} \right)^2$$  \hspace{1cm} (2-42)

or equivalently expressed in the dissipated power in the resonator

$$P_r = \frac{V_r^2}{R_{loss}} = \frac{P_{av0}}{2}$$  \hspace{1cm} (2-43)

The phase noise is calculated to

$$\mathcal{L}(f_m) = \frac{FkT}{P_r Q_0^2} \left( \frac{f_0}{f_m} \right)^2$$  \hspace{1cm} (2-44)

which is the same expression as given in (2-33).

**Negative resistance oscillator**

Figure 2-18 depicts a model of a reflection type oscillator.

The amplifier is modeled with an internal resistance $R_{in}$, and a voltage noise source. The loss resistance in the cavity is described by $R_S$. The ratio of loaded and unloaded quality factor according to (2-36) can be expressed as

$$\frac{Q_L}{Q_0} = \frac{R_s}{R_s + R_{in}} = \frac{1}{1 + \frac{R_{in}}{R_s}} = 1 + \beta$$  \hspace{1cm} (2-45)

An expression of the phase noise using (2-32) with $P = P_{RF}, R_{OUT}=0$, i.e., case 1 for power dissipated in both the resonator and in the internal resistance in the amplifier $R_{in}$, is stated
Chapter 2 Challenges in oscillator design

below. The circuit may be seen as a one-port with one port grounded, according to [21], [39],

$$
\mathcal{L}(f_m) = \frac{FkT}{8(Q_0)^2(Q_L/Q_0)^2(1 - Q_L/Q_0)P_{RF}} \left(\frac{f_0}{f_m}\right)^2
$$

(2-46)

The equation simplifies when inserting (2-36) in (2-46) to

$$
\mathcal{L}(f_m) = \frac{FkT}{8(Q_0)^2\beta/(1 + \beta)^3P_{RF}} \left(\frac{f_0}{f_m}\right)^2
$$

(2-47)

It can be shown with differentiation that (2-47) will have a minimum value for

$$
\beta = \frac{1}{2}
$$

(2-48)

or equivalently,

$$
\frac{Q_L}{Q_0} = \frac{2}{3}
$$

(2-49)

This means that the power dissipated in the resonator is $P_r = 2/3P_{RF}$. If the expression is inserted in (2-47), it will evaluate the optimum phase noise value as [37], [21],

$$
\mathcal{L}(f_m) = \frac{9FkT}{16P_rQ_0^2} \left(\frac{f_0}{f_m}\right)^2
$$

(2-50)

Equation (2-50) has been derived from the analytical expression, and it is quite similar to (2-34) using the approximate derivation in [21]. It can be worth noting that the power dissipated inside the resonator, $P_r$, is only half in the negative resistance oscillator compared to the feedback oscillator for a given phase noise performance (2-44). Therefore, it is beneficial to use reflection type oscillators where the resonators are the bottleneck for power capability, as for power limited crystal resonators or for resonators tuned with power-limited varactors that will be discussed in 2.6.1.

If the power delivered from the amplifier $P_{RF}$ is considered, the optimum value of $\beta$ is used, and the efficiency is stated $P_{RF} = \eta P_{DC}$, following expression of phase noise can be written

$$
\mathcal{L}(f_m) = \frac{FkT}{8(Q_0)^2\beta/(1 + \beta)^3 \eta P_{DC}} \left(\frac{f_0}{f_m}\right)^2
$$

(2-51)

which for optimum coupling according to (2-48) simplifies to

$$
\mathcal{L}(f_m) = \frac{27}{32 \eta P_{DC}Q_0^2} \left(\frac{f_0}{f_m}\right)^2
$$

(2-52)

Equation (2-52) can be used to calculate the bound on phase noise for available $P_{DC}$, and $Q_0$ from the technology.
2.5.2 Resonator coupling design impact

In this study, two possible methods to match the resonator coupling for a reflection oscillator are considered: by changing the impedance coupling to the resonator, or by changing the reflection gain of the amplifier. The following designs exemplify the two methods.

A cavity oscillator with varied resonator coupling by changing the position of the cavity from an excitation microstrip line is demonstrated in Paper [C]. The corresponding resonator resistance, $R_S$, and the amplifier gain for the optimum bias point $-R_N$ at the best phase noise shows the ratio $R_S/|R_N|$ is close to unity, which is further discussed in 4.2.1.

The different approach to adjust the resonator coupling $R_S/|R_N|$ by changing the gain represented by $-R_N$ of the active device is reported in Paper [D]. A reflection amplifier is used where the gain can be tuned by a passive change using varactors in its termination network without affecting the bias level of the active transistor. The oscillator with the electronic gain control is found to give about the same result as a mechanically tuned coupling factor, which is further discussed in 4.4.2.

There are published results for reflection oscillators where the optimum loss resistance in the resonator is showed for one third of the absolute value of the negative resistance from the amplifier, [40], [41] and Paper [F]. That optimum can still be applied as a good design rule. The optimum resonator coupling which is derived in (2-48) for a reflection oscillator is $\beta = 1/2$, [23]. It corresponds to the ratio of the dissipated power in the resonator to the power in the loop ($P_{RF}$) as $Q_L/Q_0 = 1/(1 + \beta) = 2/3$. The open loop gain is important to consider for the optimum coupling. It is found in measurement that high loop gain compression shall be avoided, due to the non-linear up-conversion of flicker noise and thermal noise, and consequently, a $\beta$ close to unity, or $Q_L/Q_0 = 1/2$ is preferable. Generally, $\beta = 1/2$ to $\beta = 1$ should be considered dependent on the design condition, as due to flicker noise content, or the needed margin for start-up oscillation, etc.

The theoretical effect on phase noise for different coupling expressed in $Q_L/Q_0$ according to (2-46) is shown in Figure 2-19. Two different definitions of power are used for calculating the $Q_L/Q_0$, as the power dissipated in the loop or as available power. The usage of the available power requires a matched load, which is not necessary if dissipated power is used.

![Figure 2-19 Minimum phase noise versus Q_L/Q_0 with power defined as the dissipated power in the resonator, and defined as maximum available power for a feedback type oscillator.](image-url)
It can be noticed from Figure 2-19 that the deviation in phase noise between coupling factor $\beta = 1/2$, ($Q_L/Q_0 = 2/3$) to $\beta = 1$, ($Q_L/Q_0 = 1/2$), for constant dissipated power is about 0.7 dB for this linear model. The different coupling factors corresponds to a variation in $S_{21}$ according to (2-41) to about 3.5 dB. In reality, when incorporating the non-linear up-conversion of noise, the difference in phase noise can be significantly larger. In Table 4-2 in 4.2.1, a phase noise variation of 5 dB is shown due to a variation of about 3.5 dB in the open loop gain.

**Impact of phase condition**

The phase condition of $n \cdot 360^\circ$ is desired to be fulfilled for the frequency where the maximum phase slope occurs, thus at $\max(\partial \Phi(Z)/\partial \omega)$ or equivalently at maximum Q-factor. If a phase error is introduced in the loop, the frequency is changed to meet the phase condition. For a feedback oscillator the phase error can be written as

$$\Delta \Phi_{\text{error}} + \angle (S_{21}(\Delta f)) = 0 \quad (2-53)$$

and generally the transfer function versus offset $\Delta f$ is rolled-off as

$$S_{21}(\Delta f) = \frac{S_{21}(0)}{1 + j2Q_L \frac{\Delta f}{f_r}} \quad (2-54)$$

If considering the argument of the expression of (2-54) used in (2-53), the following expression can be stated

$$\tan(\Delta \Phi_{\text{error}}) = 2Q_L \Delta f / f_r \quad (2-55)$$

where the changed oscillation frequency $f_0 = f_r + \Delta f$ to fulfill the condition from (2-53). If no phase error, $\Delta \Phi_{\text{error}} = 0$, the oscillation frequency is equal to the original oscillation frequency $f_r$.

The voltage gain in the loop can be written as [42]

$$\frac{1}{G_v} = |S_{21}(\Delta f)| = S_{21}(0) \cos^2(\Delta \Phi_{\text{error}}) \quad (2-56)$$

The phase noise expression can be rewritten with power gain and the phase error in the open loop, $\Delta \Phi_{\text{error}}$, consequently showing the degradation in phase noise as [42],[23],

$$L(f_m, \Delta \Phi_{\text{error}}) \propto \frac{L(f_m,0)}{\cos^4(\Delta \Phi_{\text{error}})} \quad (2-57)$$

### 2.6 Frequency tuning

Frequency tuning can be achieved by changing the equivalent inductance or capacitance values that define the resonance frequency

$$f_0 = \frac{1}{2\pi\sqrt{LC}} \quad (2-58)$$
A common method uses varactors connected as part of the capacitance $C$, connected single or switched from a capacitor bank for generating several sub-bands. These elements can be integrated with the resonator circuitry to directly affect the resonance condition. Alternatively, they can be coupled externally to the resonator to change the load of the resonator or to change the phase condition.

The tuning elements can be implemented as semiconductor devices with junction capacitance or formed as digital switches of PIN-diodes with connected lumped capacitors. They can alternatively be implemented in micro-machined structures as MEMS-varactors or RF-MEMS switches, etc.

The advantage of using switched elements in combination with analog tuning within the sub-band is that the analog elements can be weakly coupled to enhance a narrow band tuning and still keep a high loaded Q-factor. The performance of the digital switches are thus important, and they are preferably implemented as low-loss switches example in RF-MEMS technology. Tuning with semiconductor varactors is described in 2.6.1 and tuning with RF-MEMS-switches in 2.6.2.

A frequency tuning by a change in the phase condition, i.e., a change in the group delay $\Delta t_g$ can be written as

$$
\Delta f = \frac{1}{360 \Delta t_g} \frac{\theta}{\Delta t_g}
$$

and the Q as

$$
Q = \pi f_0 t_g
$$

The phase shift in the signal path does not degrade the Q-factor of the resonator. A disadvantage is the phase condition may not be fulfilled for the frequency corresponding to maximum phase slope of the impedance. The introduced phase error degrades the performance according to (2-57), and therefore this method is often limited to narrow-band tuning.

A wideband resonator often used in instruments is the yttrium-iron-garnet, (YIG), resonator. It consists of a high-Q ferrite of $Y_2Fe_2(FeO_4)_3$, and normally it has a spherical form that can be widely tuned by varying the DC-bias of an applied magnetic field [1]. Due to the change in the ferromagnetic resonance, the frequency can be tuned over several octaves. YIGs are commonly manufactured for center frequencies from about 500 MHz to 50 GHz, and the unloaded Q-factor is normally greater than 1000. The fabrication is quite difficult with a complex polishing and aligning process of the ferrite sphere to the coupling loop. Other drawbacks are the high DC-current consumption needed for tuning and their normally bulky solution (volume of several cm$^3$) with difficulties for miniaturization and integration compared to solid state semiconductor tuning elements. However, more compact solutions using planar YIG-structures have recently demonstrated successful results [43].

As previously discussed, the resonator coupling is essential for an efficient usage of the resonator to maintain a high Q-factor as well as keeping a high signal-to-noise ratio. This task becomes even more challenging for tunable resonators, where the loaded Q-factor has to remain high over a certain bandwidth with an optimum coupling factor [44].
impact on the phase noise is that all tuning circuitry itself will cause modulation noise that will degrade the phase noise. The effect of modulation noise is empirically modeled by adding an extra term to Leeson’s equation [45],

\[ \mathcal{L}(\Delta f) = 10\log_{10} \left[ \frac{FkT}{2P_s} \left[ 1 + \left( \frac{f_0}{2Q_L \Delta f} \right)^2 \right] \left[ 1 + \frac{\Delta f_{1/3}}{\Delta f} \right] + \frac{K_0^2 V_m^2}{8\Delta f^2} \right] \] (2-61)

where besides the previously defined parameters, \( K_0 \) represents tuning sensitivity (Hz/\( V \)) and \( V_m \) the voltage noise on the varactor.

### 2.6.1 Varactor based tuning

A varactor may be incorporated in the capacitance value in (2-58) to direct change the resonance frequency. An example of a varactor tuned serial \( LC \)-resonator for a negative resistance oscillator is studied in this section. The negative resistance, \(-R_N\) represents the transistor gain, and \(R_s\) represents the resonator loss. For a steady state \(-R_N + R_s = 0\). A tuning capacitor, i.e., varactor, is added in serial with the resonator as in Figure 2-20, and the varactor loss resistance \( R_V \) is incorporated in the total loss of the resonator when dimensioning the coupling factor.

![Figure 2-20 Schematic of a reflection oscillator with series varactor [32].](image)

The oscillation frequency is expressed by

\[ f_0 = \frac{1}{2\pi} \sqrt{\frac{C_s + C_v}{L C_s C_v}} = \frac{1}{2\pi} \sqrt{\frac{P_s Q_s}{L C_v P_v Q_v}} \] (2-62)

where \( P_s \) is the total power dissipated in the resonator, and \( P_v \) is the power dissipated in the varactor. The quality-factor for the complete resonator, i.e., the loss in the resonator compared to the total capacitance in the resonator, is given by,

\[ Q_s = \frac{C_s + C_v}{\omega_0 R_s C_v C_s} \] (2-63)

The quality factor of the varactor, i.e., loss in the varactor compared to its capacitance is given by

\[ Q_v = \frac{1}{\omega_0 R_v C_v} \] (2-64)

The ratio \( P_s/P_v = R_s I_{res}^2/R_v I_{res}^2 = R_s/R_v \) due to the same current in the serial resonator. The tuning sensitivity with respects to the varactor capacitance is set by [32]
\[
\frac{\partial f}{\partial C_v} = -\frac{f_0}{2C_v} \frac{P_vQ_v}{P_s Q_s}
\]  

(2-65)

which indicates that a wider tuning bandwidth for a given coupling \(P_s/P_v\) and \(f_0\) is achieved with a lower varactor capacitance with a higher quality factor. The maximum tuning range is given when \(P_vQ_v/P_s Q_s \rightarrow 1\), that is when \(C_v/C_s \rightarrow 0\) for the varactor \(C_v\) in serial with the capacitance \(C_s\) of the resonator. The relative tuning sensitivity will also be maximized to

\[
\frac{\partial f}{f_0} \frac{\partial C_v}{C_v} = -\frac{1}{2}
\]  

(2-66)

**Power handling in varactors**

The Q-factor of the resonator is limiting the noise performance and tuning-range of the oscillator. Further, the voltage handling capability of the varactor itself is limiting a wideband tuning. The varactor is normally modeled as a voltage-controlled capacitor in serial with a loss resistance, and if \(V_R\) is representing the rms-voltage across the loss resistance in Figure 2-20, the dissipated power is expressed by

\[
P_v = \frac{V_R^2}{R_v}
\]

(2-67)

and the rms-voltage across the varactor capacitor, \(V_c\), can be expressed as

\[
V_c = Q_v V_R
\]

(2-68)

The power dissipated in the varactor is equal to

\[
P_v = \frac{V_c^2}{Q_v^2 R_v}
\]

(2-69)

As the phase noise is minimized by maximizing \(1/\rho_{res}Q_{\text{L, res}}\) in the resonator, the term \(V_c^2/R_v\) will be a figure-of-merit for a good varactor [23]. This term sets the maximum power handling of the varactor in the resonator. Equation (2-50) expresses the corresponding minimum achievable phase noise as

\[
\mathcal{L}(f_m) = \frac{9FkT R_v}{16V_c^2} \left(\frac{f_0}{f_m}\right)^2
\]

(2-70)

In this case, the varactor contributes with all loss and power dissipation of the resonator, and the losses of other parts are neglected. The minimum phase noise given by (2-70) is valid for a reflection type oscillator. This level would be degraded twice for a feedback type oscillator (2-44). A feedback oscillator needs twice the dissipated power in the resonator to achieve the same performance as for a reflection oscillator.

Another effect to consider is that the voltage swing across the varactors increases with a stronger coupling to the resonator, besides for a higher Q-factor and increased input power level. The voltage swing can exceed the reverse-bias of the varactor in a case of junction capacitance based varactors. This effect starts to degrade the Q-factor in a large-signal condition at a rather low power level and must be considered [36]. Figure 2-21 (a) shows
the large-signal Q-factor degradation discussed in Paper [G], which demonstrates a varactor tuned cavity oscillator at 10 GHz. Figure 2-21 (b) shows the corresponding voltage swing across one single varactor in an anti-serial pair for different power $P_{in}$ in the cavity. Bias level of -2 V is applied on the varactors. For higher power level and thus higher voltage swing, current leakage in the forward region will cause the Q-factor degradation. For low power, the voltage swing scales by $\propto Q\sqrt{P_{in}}$, until Q degrades at higher $P_{in}$.

As the power handling of the varactor can be a bottleneck for the phase noise of wideband varactors, it is beneficial to use two anti-serial varactors to divide the voltage swing. Another solution is to use weakly coupled varactors for a narrowband tuning. They can be used together with high power capable switches e.g. MEMS-switches for sub-band selection. This solution is proposed in Paper [A].

Generally, VCO tuning characteristics are nonlinear. The $C_v(V_v)$ characteristic is on the form

$$C_v(V_v) = \frac{C_{v0}}{(1 + \frac{V_v}{\varphi})^\gamma} \quad (2-71)$$

where $\varphi$ is the built-in potential, $C_{v0}$ the capacitance at zero bias and $\gamma$ is the varactor junction sensitivity. For abrupt varactors $\gamma = 0.5$, and for hyperabrupt varactors $1 \leq \gamma \leq 2$. As the $C_v(V_v)$ from (2-71) has a nonlinear behavior, normally two anti-serial varactors are used in designs to suppress harmonics from tuned capacitance [32]. Also, usage of hyperabrupt varactors will give a more linear frequency tuning, e.g., for $\gamma = 2$, a completely linear tuning will occur which is seen if using (2-71) in (2-58). This case is beneficial to suppress non-desired up-converted spurious.

### 2.6.2 MEMS based tuning

During the recent decades, extensive efforts have been invested in developing micro-machined structures, as RF-MEMS. They have become good candidates as tuning elements due to their high power capability and low loss [22]. However, drawbacks are their slow switching speed and that the moving mechanical structures can be influenced by vibrations. This causes inevitably electroacoustic coupling. Sensitivity for vibrations around 10-20 kHz is found in common MEMS with typical modal mass and spring constant for the moving structure [22], [46]. Rather high voltage is needed for MEMS types using
electrostatic actuation, around 10 V-30 V. This high voltage are also beneficial for the power capacity, due to the capable RF-swing before the switch will self-actuate. For a pull-in voltage of 1-30 V, the RF power required for self-actuation in a 50-ohm system corresponds to 0.02-18 W. Another application of low voltage MEMS-bridges are power-limiters, due to the protection capability and survival of high power.

The MEMS-switches can be categorized in several ways, regarding actuation mechanism, movement type, contact type, the circuit configuration and geometric type. Short descriptions of each category are discussed in following sections.

Common mechanisms to apply the force to the necessary movement of the membranes are listed below.

- **Electrostatic.** A bottom electrode beneath a membrane, which is connected to the RF-path, creates an attraction by an electrostatic force when a voltage is applied. The membrane bends to the other position, counterbalancing the spring force of the membrane. It reverts to its original position without the applied voltage. Figure 2-22 (a-c) show principal drawings.

- **Thermal.** A piece of material that changes shape with temperature causes the movement of the membrane.

- **Magnetostatic.** There are two different methods. A Lorentz force actuates patterned coils on the membrane by an externally applied magnetic field. The other method uses a ferromagnetic coating deposited on the membranes, which interacts with an externally applied magnetic field.

- **Piezoelectric.** A material that changes shape due to piezoelectric phenomenon causes the movement of the membrane.

The movement direction can be

- **Vertical**

- **Lateral**

The connection type in the RF-path can be

- **Metal-to-metal.** An ohmic connection is established between the membrane and the disrupted transmission line. The bottom electrode is still isolated. Figure 2-22 (a) shows a principal drawing.

- **Capacitive.** The distance between the gaps of the membranes and the disrupted transmission line is changed when the membrane is attracted to the other position. Figure 2-22 (b) shows a principal drawing.

- **Varactor.** The height of the membrane to the transmission line varies analogously by the voltage forming a controllable capacitance. Figure 2-22 (c) shows a principal drawing.
The membrane connecting to the RF-path can be

- **Series**, shown in Figure 2-22 (a).
- **Shunt**, e.g., connecting a grounded pad shown in Figure 2-22 (b).

The membrane topology can be

- **Cantilever type.** No residual stress remains within the membrane, as there is freedom for the tip to move which diminish static forces. The membrane is fastened only along one side shown in Figure 2-22 (a).
- **Fixed-fixed-beams.** The beams are fastened on two opposite sides and they will bend down in the middle in the presence of an actuation force shown in Figure 2-22 (b).
- **Circular diaphragms or membrane.** This is a production friendly shape.

![Typical MEMS-topologies](image)

(a) Ohmic cantilever. (b) Shunt capacitive. (c) Varactor with airgap separation to bottom electrode.

Figure 2-23 Cantilever MEMS from this study. (a) Chip photo. The chip size is 2970 x 1615 um and the MEMS contains three switch elements. (b) Chip photo. Cantilever in open position for one switch element.

Figure 2-23 (a-b) shows chip photo of the RF-MEMS-switches used in this study. They are designed by RF Microtech and fabricated at the FBK foundry.

RF-MEMSs have demonstrated good performance due to their high power handling and low loss. The power capacity is needed for a broad tuning range as the tuning elements need to be strongly coupled to the resonator. Further, a high power capacity in the resonator is needed for good phase noise performance. The low loss entails a high Q-factor, which
not will be suffered from large-signal degradation due to the high power capacity. However, the side-effects of electroacoustic vibration and some reliability issues are continuously improved.

2.7 Benchmarking oscillators

Benchmarking oscillators in literature are often a challenging task as several performance parameters are involved. These parameters can involve output power, power consumption, center frequency, offset frequency, tunability, etc. It is, therefore, a difficult task to define a single figure-of-merit for benchmark an oscillator. The figure could either, relate the performance to the physical boundaries set by the process technology as the Q-factor, and the device noise floor; or alternatively be dependent on some typical application.

According to Leeson’s equation (2-29), the noise floor is measured relative the RF-power in the resonator, and therefore increasing the power is beneficial. The power can be increased by increasing the current density in the transistor, or by scaling the transistor size for the same current density. Drawbacks are that the close-in flicker noise, as well as shot noise for a bipolar device are increased with higher drain current, which will counterbalance the benefit of the increased signal power. The far out noise floor of HEMTs is improved at increased power.

2.7.1 Figures of merit

The power can be increased by combining several oscillators locked to the same reference. In general, the carrier power can be increased by \(N^2\), if the output signals from \(N\) identical oscillators are added coherently in phase. Their noise power is added by the factor \(N\), as the noise sources are uncorrelated. The noise to power ratio is improved by the factor \(N\) [25]. This is advantageous e.g., in MIMO systems using several oscillators [47].

The benefit of the increased output power compared to the noise floor implies that benchmarking oscillators might be measured with a DC-power normalized figure-of-merit. The figure-of-merit is further normalized with center frequency, and frequency offset for a fair comparison.

A commonly used key parameter for benchmarking is as previously suggested the power-normalized figure-of-merit, hereinafter referred to simply as \(FOM\). DC-power normalized figures-of-merit must be handled with care. An example is that low power consuming oscillators can still reach high merit figures, although having low or moderate absolute phase noise performance. For a radio system, the absolute phase noise performance is the major important merit figure to increase the data capacity. However, if DC-power-normalized \(FOM\) is used together with the absolute phase noise performance to benchmark oscillators, it becomes a complementary measure of the power efficiency of the oscillator, and grades the design quality. The DC to RF-power is important to maximize, and the RF-power in the resonator is beneficial for the phase noise. The maximum \(FOM\) will be shown to be scaled by the unloaded Q-factor of the resonator and the noise floor within the \(f_0/2Q\) bandwidth.
To benchmark the oscillator performance versus different technologies, $FOM$, is calculated under the assumption of no flicker noise. This figure was defined by Wagemans [48], and it was further refined in [49]. It is defined as

$$ FOM = -\mathcal{L}_{\text{meas}}(f_m) + 20 \log_{10} \left( \frac{f_0}{f_m} \right) - 10 \log_{10} \left( \frac{P_{\text{DC}}}{1\text{mW}} \right) $$

(2-72)

where $\mathcal{L}_{\text{meas}}(f_m)$ is the measured phase noise at offset frequency $f_m$, $f_0$ is the oscillation frequency and $P_{\text{DC}}$ (mW) the DC-power consumption.

In Leeson’s equation, this is used as a base for scaling the effect of resonator power, and Q-factor, and to relate the figures to a physical noise floor. The fitting parameter, $F_{\text{eff}}$, can there be extracted from (2-52) as

$$ F_{\text{eff}} = \mathcal{L}_{\text{meas}}(f_m) - 20 \log_{10} \left( \frac{f_0}{f_m} \right) - 10 \log_{10} \left( \frac{27kT}{32} \right) - 30 $$

$$ + 10 \log_{10} \left( \frac{P_{\text{DC}}}{1\text{mW}} \right) + 20 \log_{10} (Q_0) $$

(2-73)

$F_{\text{eff}}$ can further be expressed with the definition of $FOM$ in (2-72) as

$$ F_{\text{eff}} = 174.6 + 20 \log_{10} (Q_0) - FOM. $$

(2-74)

In case of no flicker noise and no RF-noise from the active device, and 100 % DC to RF conversion efficiency, meaning that $F_{\text{eff}} = 0$ dB in (2-74), $FOM_{\text{max}}$ is found to be linearly related to the unloaded Q-factor of the resonator. Normally $F_{\text{eff}} > 0$ dB, and $FOM$ will be reduced. Obvious reasons for $F_{\text{eff}} > 0$ dB in (2-74) are the finite efficiency and the effect of non-ideal resonator coupling. If the effects of finite efficiency and non-ideal resonator coupling are considered, an operating noise figure can be defined by using the expression in (2-46) as

$$ F_{\text{op}} = 173.9 + 10 \log_{10} \left( \frac{8\eta\beta}{(1 + \beta)^3} \right) + 20 \log_{10} (Q_0) - FOM $$

(2-75)

It results that $F_{\text{op}} = F_{\text{eff}}$ in the ideal case $\eta = 100\%$ and ideal coupling $\beta = 1/2$.

The best performance that can be achieved at $F_{\text{eff}} = 0$ dB in (2-73) for given process parameters, $Q_0$ and power capability, $P_{\text{DC}}$ (mW) can be expressed as

$$ \mathcal{L}_{\text{min}}(f_m) = -174.6 - 10 \log_{10} (P_{\text{DC}}) - 20 \log_{10} (Q_0) + 20 \log_{10} \left( \frac{f_0}{f_m} \right) $$

(2-76)

This equation can be used to show trend lines for the phase noise performance in the used technology.

Table 2-2 presents process parameters as $Q_0$ and practical power capability. Calculated trend lines of the phase noise at 100 kHz for a 10 GHz oscillation frequency are shown. The power capability values are calculated from an averaged DC-consumption of the considered MMIC-designs among those from [50]-[65]. The reported results on DC-consumption in this calculation is used to take care of the practical issues for good bias
level. For example the choice of transistor size and bias condition to manage the needed cooling etc., influence on the typically used devices. Typical Q-factors for the respective process technology are found in textbooks as [51]. Phase noise trend lines and published results together with designs from this work are shown in Figure 2-24. The phase noise trend lines are based on calculations from (2-76). For mature processes like CMOS, the results are fairly well saturated at the phase noise trend line, but for GaN-HEMT there is still a significant discrepancy between reported values and the theoretical limit. Some oscillators designed in this work in hybrid technology, as for the cavity oscillators, are also included in Figure 2-24.

Table 2-2 Phase noise limit at 100 kHz offset from a 10 GHz oscillation frequency, calculated from process parameters for MMICs. The given bias levels are averaged values from the published oscillators, [50-65]

<table>
<thead>
<tr>
<th></th>
<th>GaN-HEMT,</th>
<th>InGaP-HBT</th>
<th>CMOS 90 nm</th>
<th>SiGe-HBT</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_c / V_d$ (V)</td>
<td>30</td>
<td>5</td>
<td>1</td>
<td>3</td>
</tr>
<tr>
<td>$I_c / I_d$ (mA)</td>
<td>120</td>
<td>90</td>
<td>9</td>
<td>25</td>
</tr>
<tr>
<td>Q_0</td>
<td>40</td>
<td>40</td>
<td>10</td>
<td>10</td>
</tr>
<tr>
<td>Phase noise@1 MHz (dBc/Hz), $f_{center}$=10 GHz</td>
<td>-142</td>
<td>-133</td>
<td>-104</td>
<td>-113</td>
</tr>
</tbody>
</table>

Figure 2-24 Phase noise versus oscillation frequency for oscillators in different MMIC technologies, compared to GaN HEMT oscillators designed in this work. The trend lines are valid for typical MMIC technology parameters, presented in Table 2-2. Note that two oscillators designed in this work are designed in hybrid technology and consequently have significantly better Q-factor.

Table 2-3 shows a comparison to other state-of-art fixed frequency oscillators with high-Q-resonators in different technologies at X-band.
Table 2-3 State-of-the-art published High-Q oscillators for fixed frequency

<table>
<thead>
<tr>
<th>Ref</th>
<th>( \frac{\Delta}{100 \text{ k}} ) (dBc/Hz)</th>
<th>( f_0 ) (GHz)</th>
<th>( P_{dc} ) (mW)</th>
<th>FOM@100 KHz</th>
<th>Active device technology</th>
<th>Resonator technology</th>
</tr>
</thead>
<tbody>
<tr>
<td>HMC-C200 8.0-8.3 GHz [Hittite]</td>
<td>-140</td>
<td>8.2</td>
<td>700(^{**})</td>
<td>209</td>
<td>N/A</td>
<td>DRO</td>
</tr>
<tr>
<td>[66], fig 9</td>
<td>-132(^{(*)})</td>
<td>9.2</td>
<td>N/A</td>
<td>N/A</td>
<td>GaAs pHEMT</td>
<td>Silver coated cavity</td>
</tr>
<tr>
<td>DRO-10.600-FR (PSI)</td>
<td>-137</td>
<td>10.2</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>DRO</td>
</tr>
<tr>
<td>[67]</td>
<td>-118</td>
<td>10.6</td>
<td>54(^{(***)})</td>
<td>201</td>
<td>GaAs HEMT</td>
<td>DRO</td>
</tr>
<tr>
<td>[67]</td>
<td>-123</td>
<td>10.6</td>
<td>46(^{(***)})</td>
<td>207</td>
<td>GaAs pHEMT</td>
<td>DRO</td>
</tr>
<tr>
<td>[67]</td>
<td>-135</td>
<td>10.6</td>
<td>N/A</td>
<td>N/A</td>
<td>SiGe HBT</td>
<td>DRO</td>
</tr>
<tr>
<td>[68]</td>
<td>-149</td>
<td>7.61</td>
<td>400</td>
<td>221</td>
<td>GaAs InGaP HBT</td>
<td>DRO</td>
</tr>
<tr>
<td>[68]</td>
<td>-142</td>
<td>7.61</td>
<td>100</td>
<td>220</td>
<td>GaAs InGaP HBT</td>
<td>DRO</td>
</tr>
<tr>
<td>Paper [C]</td>
<td>-145</td>
<td>9.9</td>
<td>200</td>
<td>221</td>
<td>GaN HEMT</td>
<td>Aluminum cavity</td>
</tr>
</tbody>
</table>

\(^{(*)}\) The plot in [66] is used to estimate the phase noise. \(^{(**)}\) A buffer amplifier is included, but output power is comparable with other output power in this work. \(^{(***)}\) calculated from power density and transistor size.

**Figure of merit for tunable oscillators**

The power-normalized figure of merit, \( FOM \), is physical in the sense of the phase noise scaling with DC-power consumption, loaded Q-factor, oscillation- and offset frequency. A figure-of-merit value also including the tunability is more difficult to define. A definition, commonly used by MMIC designers, is to normalize to 10 % tuning, which is considered to be a typical tuning range for MMIC VCOs [69],

\[
FOM_T = -L_{meas}(f_m) + 20\log_{10}\left(\frac{f_0}{f_m}\right) - 10\log_{10}\left(\frac{P_{DC}}{1\text{mW}}\right) + 20\log_{10}(TR/10)
\]  \tag{2-77}

or expressed as

\[
FOM_T = FOM + 20\log_{10}(TR/10)
\]  \tag{2-78}

where the factor \( TR \) is the percentage relative tuning ratio. From (2-78) it is clear that the \( FOM_T \) will be degraded from the \( FOM \)-value if tuning range lower than 10 % is achieved.

Another figure-of-merit is using the absolute tuning bandwidth without the DC-power consumption as [44]

\[
FOM'_T = -L_{meas}(f_m) + 20\log_{10}\left(\frac{BW}{f_m}\right)
\]  \tag{2-79}

A benefit of this definition is that normalizing to the center frequency is not needed explicitly, as it is already considered when the absolute tuning frequency \( BW = TR \cdot f_0/100 \) is used, instead of relative tuning as in (2-77).

However, a complication of the figure-of-merit in (2-77) and (2-79) is that they do not asymptotic converge to the figure-of-merit of a fixed frequency oscillator as in (2-72) when the tuning bandwidth decreases to zero.

Table 2-4 presents a comparison between tunable oscillators technologies.
Table 2-4 State-of-the-art published high-Q tunable oscillators

<table>
<thead>
<tr>
<th>Ref</th>
<th>Phase noise @ 10 kHz (dBC/Hz)</th>
<th>Phase noise @ 100 kHz (dBC/Hz)</th>
<th>Phase noise @ 1MHz (dBC/Hz)</th>
<th>fc (GHz)</th>
<th>Tuning-ratio (TR) (%)</th>
<th>$Q_0$</th>
<th>FOM</th>
<th>FOM&lt;sub&gt;T&lt;/sub&gt; (*)</th>
<th>Technology</th>
</tr>
</thead>
<tbody>
<tr>
<td>[70]</td>
<td>-135</td>
<td>-</td>
<td>9.84665-9.8472</td>
<td>0.0056</td>
<td>10k-22k</td>
<td>-</td>
<td>-</td>
<td>DRO, SiGe</td>
<td></td>
</tr>
<tr>
<td>[71]</td>
<td>-100</td>
<td>-129</td>
<td>-149.5</td>
<td>8.2</td>
<td>1.28</td>
<td>510 load</td>
<td>211.7</td>
<td>194</td>
<td>microstrip, SiGe</td>
</tr>
<tr>
<td>[72]</td>
<td>-121.7</td>
<td>-133@50k</td>
<td>13.3</td>
<td>0.165</td>
<td>13k</td>
<td>-</td>
<td>-</td>
<td>DRO, GaAs pHEMT</td>
<td></td>
</tr>
<tr>
<td>[73]</td>
<td>-78</td>
<td>&lt;-100</td>
<td>-123</td>
<td>11.4033-11.5565</td>
<td>1.34</td>
<td>-</td>
<td>-</td>
<td>SiW, SiGe</td>
<td></td>
</tr>
<tr>
<td>[74]</td>
<td>-134.5 (at 500k)</td>
<td>-3.475</td>
<td>6</td>
<td>167.1</td>
<td>162.7</td>
<td>MEMS-varactor, microstrip, SiGe</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>[75]</td>
<td>-95.6</td>
<td>-125.1</td>
<td>11.16</td>
<td>4.1</td>
<td>300</td>
<td>193</td>
<td>185</td>
<td>SiW, GaAs HEMT</td>
<td></td>
</tr>
<tr>
<td>[76]</td>
<td>-122</td>
<td>12.2</td>
<td>2.5 mechanical</td>
<td>189.1</td>
<td>177</td>
<td>SIW, HJ-FET</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>[43]</td>
<td>-100</td>
<td>-130</td>
<td>2.4</td>
<td>67</td>
<td>-</td>
<td>YIG</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>[77]</td>
<td>-132.7</td>
<td>5.3</td>
<td>0.094</td>
<td>Vt-DRO,</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>[78]</td>
<td>-108.4</td>
<td>-129.4</td>
<td>-152.4</td>
<td>23.12</td>
<td>3.8</td>
<td>214.5</td>
<td>206</td>
<td>Möbius strip SiGe HBT</td>
<td></td>
</tr>
<tr>
<td>[79]</td>
<td>-109</td>
<td>22.1</td>
<td>20.6</td>
<td>181</td>
<td>187</td>
<td>SiGe Integrated</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>[80]</td>
<td>-117</td>
<td>9.5</td>
<td>4.8</td>
<td>180.9</td>
<td>174.5</td>
<td>SiW, pHEMT</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Paper [A]</td>
<td>&lt;-129</td>
<td>&lt;-158</td>
<td>5</td>
<td>500</td>
<td>202</td>
<td>196</td>
<td>MEMS (@1mm)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Paper [A]</td>
<td>&lt;-123</td>
<td>&lt;-153</td>
<td>10</td>
<td>12.3</td>
<td>500</td>
<td>196</td>
<td>198</td>
<td>MEMS (@2.5 mm)</td>
<td></td>
</tr>
<tr>
<td>Paper [C]</td>
<td>-144</td>
<td>&lt;-165</td>
<td>9.9</td>
<td>-</td>
<td>3800</td>
<td>227</td>
<td>-</td>
<td>Cavity fixed freq</td>
<td></td>
</tr>
<tr>
<td>Paper [G]</td>
<td>-118</td>
<td>-146</td>
<td>10</td>
<td>1.6</td>
<td>400</td>
<td>201</td>
<td>185</td>
<td>Cavity with diode varactors</td>
<td></td>
</tr>
</tbody>
</table>

(*) FOM<sub>T</sub> = -20log$\Delta f$+10log(f<sub>0</sub>/\Delta f)-10log(P<sub>DC</sub>/1mW)+20log(TR/10) or

FOM<sub>T</sub> = FOM+20log(TR/10)
Chapter 3.

Component characterization

This chapter discusses component and device characterization of sub-blocks used throughout in the oscillator study.

It initially starts to describe the used active device regarding DC-characteristic for a bare die and DC and gain behavior for the same device integrated in a MMIC-amplifier. It further presents a developed low-frequency noise measurement setup for high bias condition suitable for GaN-HEMTs. The measured characteristics adapt the simulation models for the oscillator for more accurate noise prediction. Further, the chapter describes a high-Q resonator of an aluminum cavity with flexibility to adjust the coupling to an excitation microstrip line. The chapter discusses the cavity modes in detail, and finally, it reports the characterization of used tuning elements of MEMS and varactors and discusses a proposal of how to efficiently integrate them in the cavity to enhance tunability.

3.1 Active device characterization

The used reflection amplifier is based on a GaN-HEMT, 8x50 um gate periphery and 0.25 um gate length processed in TriQuint’s 3MI-process. DC-characteristics versus model for a bare die are shown in Figure 3-1 (a-b).

![Figure 3-1 IV-plots simulated (red square) and measured (blue circle) for the 8x50 um HEMT used in the design (a) I_d versus V_d, (b) I_d versus V_g.](image)

Figure 3-2 (a-c) show bias characteristics I_d versus V_d, I_d versus V_g and reflection gain, respectively, for the MMIC reflection amplifier based on the 8x50 um HEMT.
The different IV-characteristic for the complete MMIC compared to the bare die depends on a source resistance added in the MMIC for bias regulation and for increasing the gain, as is described in Paper [C].

3.2 Flicker noise characterization

Verification of the low-frequency noise (LFN) spectra will put some specific demands on the instrumentation setup. The bias tee must have a low cutoff frequency, and it is necessary to consider the input impedance of the low noise amplifiers not to affect the DUT, etc. Two measurement methods are analyzed in the study, and they are discussed in the following sections.

Current-noise amplification

The first method is using a transimpedance preamplifier system, based on a commercial amplifier (SR570) [81]. For low bias level, <5 mA, the internal bias supply in SR570 has a satisfying noise performance. A higher bias level will require an external bias-tee. That may limit the lowest cutoff frequency, because of the practical size of the inductors and capacitors used for the filtering. Also, large capacitors are often of the electrolyte type, which themselves contribute to flicker noise, and therefore the system noise level must be investigated carefully in the setup for clarifying its limitations. Figure 3-3 illustrates the setup, which is described in detail in [82].
Figure 3-3. LF-noise setup with current-noise amplification and external bias.

The transimpedance amplifier has low input impedance, around 1 Ω. The bias-tee components, 150 mH, and 22 mF are chosen to be parallel resonant at 3 Hz according to (3-1) for not short-circuiting the LF-noise from the DUT. On the other hand, they must make a good conduction for the current noise to the low impedance current amplifier.

\[
f_{res} = \frac{1}{2\pi \sqrt{0.15 \cdot 0.022}} = 3 \text{ Hz}
\]  

(3-1)

The output of the amplifier is connected to a dynamic signal analyzer, Agilent 35670A, which measures the voltage of the signal in time domain and calculates the FFT for frequencies up to 100 kHz.

One disadvantage of the current amplifier method is the settling time to charge the large capacitor in the DC-block when changing bias which makes these measurements for different bias very time-consuming.

**Voltage-noise amplification**

The other method is based on a high-impedance voltage amplifier, SR560 [83]. The bias to the device is connected through a load resistor, which in the setup is chosen to 100 Ω. This value is suitable for the studied sizes of devices. As there is a lower cutoff-frequency for the bias components in the signal path for this setup, it can measure noise almost down to DC. That makes the method very attractive. However, the load resistor must be chosen carefully, preferably less than the transistor channel resistance, \( R_{ds} \), to ensure that the external load dominates for the total resistance and that it is low enough with an acceptable low thermal noise contribution. The current noise is calculated by considering the channel resistance \( R_{ds} \) in parallel with the external load resistor which is seen in Figure 3-4.
Figure 3-4. LFN setup with a voltage amplification with SR560 on the drain and with SR570 on the gate, essentially for biasing purpose.

The SR560 measures the voltage noise terminated by the load resistance, which is the load resistor, $R_{\text{load}}$, in parallel with $R_{ds}$. The current noise is calculated by normalizing the voltage noise with the total load impedance. The used voltage gain of the SR560 is about 40 dB (100 V/V), and its noise floor at the input is about $4 \text{nV/} \sqrt{\text{Hz}}$. The transimpedance amplifier SR570 measures the gate noise simultaneously, but main reason in this measurement is to use its internal supply to bias the gate.

$$< I_{\text{drain noise}} > = \frac{< V_{\text{meas noise 35670}} >}{(R_{ds}/R_{\text{load}}) \cdot G_{\text{voltage gain SR560}}}$$

The same dynamic signal analyzer, Agilent 35670A, samples the output signal from the low noise amplifiers. A description of a similar setup is found in [84].

**Test bed verification**

The system noise level for the different setups is investigated by measuring a BJT transistor, 2N2222, because it has a relatively low LFN. Low bias (<8 mA, 3 V) makes it convenient to compare the different setups without introducing too much LFN from the setup itself. Different bias supplies have been used to control their noise impact in the voltage-noise measurement setup; a battery and a parameter analyzer, Agilent 4156. In the current-noise measurement setup, the internal supply in SR570 is used. Figure 3-5 (a) shows the results of this investigation. Figure 3-5 (b) shows the result of a GaN-HEMT device measured with the two different methods.
The thermal noise floor of the load resistor (100 Ω) used in the SR560-setup is calculated as

\[ \langle i_{n,100}^2 \rangle = \frac{4kT}{R} = 1.6 \cdot 10^{-22} \text{ (A}^2/\text{Hz)} \]  

(3-3)

This is about 10 dB lower than the lowest device noise floor measured in Figure 3-5, and therefore it is considered to have a negligible effect on the system noise floor in the studied frequency range.

The study concludes that the noise from the power supply affects more than the measurement method itself, and therefore, the voltage-noise amplification method is used for further LFN measurements. Besides, the voltage-noise method provides LFN measurement almost down to DC, which is shown in Figure 3-5 (b). Paper [H] presents a detailed comparison between different devices in GaN measured by this setup. Later, a compliance verification with a commercial system, E4727A, from Keysight Technologies, was performed, which showed similar results.

LFN at 1 kHz, 10 kHz and 100 kHz normalized to DC power is a relevant benchmark parameter for oscillator applications. Table 3-1 present measured LFN for GaN-HEMTs compared to other III-V material (InGaP HBT and GaAs pHEMT).
Table 3-1 Measured devices with voltage amplification method reported in Paper [H]

<table>
<thead>
<tr>
<th>Device</th>
<th>Size</th>
<th>$I_{d}/W_{b}$ (nm)</th>
<th>$F_{fe}$</th>
<th>$I_{d}$ (mA)</th>
<th>$V_{d}$ (V)</th>
<th>$V_{L}I_{c}$ (A/Hz$^{*}$V$^{-1}$) $\times 10^{-15}$</th>
<th>$V_{L}I_{c}$ (A/Hz$^{*}$V$^{-1}$) $\times 10^{-18}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>GaN HEMT 1</td>
<td>2x75</td>
<td>250</td>
<td>1.5</td>
<td>19.3</td>
<td>10</td>
<td>0.426</td>
<td>2.21</td>
</tr>
<tr>
<td>GaN HEMT 2</td>
<td>4x75</td>
<td>250</td>
<td>1.3</td>
<td>27.4</td>
<td>10</td>
<td>0.227</td>
<td>0.831</td>
</tr>
<tr>
<td>GaN HEMT 3</td>
<td>2x50</td>
<td>250</td>
<td>1.3</td>
<td>20.0</td>
<td>10</td>
<td>0.675</td>
<td>3.37</td>
</tr>
<tr>
<td>GaN HEMT 4</td>
<td>4x50</td>
<td>250</td>
<td>1.3</td>
<td>38.9</td>
<td>10</td>
<td>1.53</td>
<td>3.92</td>
</tr>
<tr>
<td>GaN HEMT 5</td>
<td>8x50</td>
<td>250</td>
<td>1.3</td>
<td>81.0</td>
<td>10</td>
<td>3.10</td>
<td>3.83</td>
</tr>
<tr>
<td>GaN HEMT 6</td>
<td>4x50</td>
<td>250</td>
<td>1.5</td>
<td>4.73</td>
<td>10</td>
<td>0.450</td>
<td>9.51</td>
</tr>
<tr>
<td>GaN HEMT 7</td>
<td>8x50</td>
<td>250</td>
<td>1.5</td>
<td>52.6</td>
<td>10</td>
<td>2.79</td>
<td>5.31</td>
</tr>
<tr>
<td>InGaP HBT 1</td>
<td>4x20</td>
<td>1000</td>
<td>1.2</td>
<td>9</td>
<td>3</td>
<td>0.007</td>
<td>0.389</td>
</tr>
<tr>
<td>InGaP HBT 2</td>
<td>2x20</td>
<td>1000</td>
<td>1.2</td>
<td>9</td>
<td>3</td>
<td>0.004095</td>
<td>0.152</td>
</tr>
<tr>
<td>GaAs pHEMT 1</td>
<td>2x20</td>
<td>150</td>
<td>1</td>
<td>12.6</td>
<td>3</td>
<td>0.142</td>
<td>3.75</td>
</tr>
<tr>
<td>GaAs pHEMT 2</td>
<td>4x20</td>
<td>150</td>
<td>1</td>
<td>24.8</td>
<td>3</td>
<td>0.187</td>
<td>2.51</td>
</tr>
<tr>
<td>GaAs pHEMT 3</td>
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<td>5.55</td>
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<td>3.86</td>
</tr>
<tr>
<td>GaAs pHEMT 4</td>
<td>4x20</td>
<td>100</td>
<td>1</td>
<td>11.2</td>
<td>3</td>
<td>0.181</td>
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<td>GaAs pHEMT 5</td>
<td>2x40</td>
<td>100</td>
<td>1</td>
<td>11.5</td>
<td>3</td>
<td>0.183</td>
<td>5.32</td>
</tr>
<tr>
<td>GaAs pHEMT 6</td>
<td>4x40</td>
<td>100</td>
<td>1</td>
<td>24.9</td>
<td>3</td>
<td>0.265</td>
<td>3.55</td>
</tr>
</tbody>
</table>
The parameters in Table 3-1 are denoted by \( L_g \) for the gate length, \( W_b \) base width, \( F_{fe} \) for the frequency slope \( (f_{Fe}^2) \), and \( LFN \) \( (A^2/Hz) \) at 1 kHz, @ 10 kHz, and @ 100 kHz for the measured LFN at the respective offset frequencies. It is obvious that the flicker noise at <100 kHz is higher for GaN-HEMTs than for InGaP HBTs. GaN-HEMTs are however better if the power-normalized noise is considered. The active devices for almost all designed oscillators in this study are based on the GaN-HEMT 8x50 (G3), which is shaded in Table 3-1. Figure 3-6 (a) shows the measured LFN for this used device versus frequency compared to simulation model, and Figure 3-6 (b) shows the measured LFN at 100 kHz versus bias.

![Figure 3-6 Measured low-frequency noise of (a) measured LFN (A^2/Hz) versus frequency. Fixed \( V_g=10 \) V for different current compared to simulation. (b) Measured LFN (A^2/Hz) at 100 kHz versus bias.](image)

### 3.3 Cavity resonators and modes

The resonator is as previously mentioned a critical component for the oscillator phase noise performance close to the carrier. Some oscillators designed in this study are based on external high-Q resonators implemented as aluminum cavities. The following chapters discuss the theory and the characterizing of the used cavity resonators.

#### 3.3.1 General theory of rectangular waveguide resonators

The propagation property of a waveguide depends on its dimensions. The phase propagation constant of a rectangular metal waveguide can be expressed

\[
\beta = \sqrt{k^2 - k_c^2}
\]

where

\[
k = \omega \sqrt{\mu \varepsilon}
\]

and

\[
k_c = \sqrt{\left(\frac{m \pi}{a}\right)^2 + \left(\frac{n \pi}{b}\right)^2}
\]

In (3-6) \( a \) and \( b \) are the side lengths of the rectangular cross section. The permittivity \( \varepsilon \) and the permeability \( \mu \) define the medium properties, \( \omega \) the used angular frequency, and \( m \) and \( n \) define the mode number. The waveguide wavelength along \( z \)-axis can be written as
\[ \lambda_g = \frac{2\pi}{\beta} = \frac{2\pi}{\sqrt{\omega^2 \mu \varepsilon - \left( \frac{m\pi}{a} \right)^2 - \left( \frac{n\pi}{b} \right)^2}} \]  

(3-7)

A cavity resonator can be realized by short-circuiting the waveguide along the z-axis, at \( z = 0 \) and at \( z = d \), respectively. The resonances will occur during the following condition,

\[ l \frac{\lambda_g}{2} = d, \quad l = 1, 2, 3, \ldots \]  

(3-8)

Figure 3-7 depicts the E-field for mode \( l=1 \), and 2, and \( m=1 \).

![Figure 3-7 TE\text{mnl}-mode shown in a resonator for l=1 and 2, and m=1](image)

By inserting (3-7) in (3-8), the resonance frequencies can be written as [85]

\[ f_{mnl} = \frac{c_0}{2\pi\sqrt{\mu_r \varepsilon_r}} \sqrt{\left( \frac{m\pi}{a} \right)^2 + \left( \frac{n\pi}{b} \right)^2 + \left( \frac{l\pi}{d} \right)^2} \]  

(3-9)

The dominant mode at the lowest frequency will be the TE\text{101}-mode \( (m = 1, l = 1) \), which evaluates to \( a=d=\lambda_g/2 \). It is common to set \( b=\lambda_g/4 \) to assure maximum margin to other resonant modes.

The E-field in y-direction for this mode, \( E_y(x,z) \), is creating a standing wave inside the cavity, which can be expressed as [85]

\[ E_y = E_0 \sin \frac{\pi x}{a} \sin \frac{\pi z}{d} \]  

(3-10)

Figure 3-8 illustrates the \( E_y \)-field-distribution for the TE\text{101}-mode.
The stored electric energy at resonance is calculated by

\[ W_e = \frac{\varepsilon}{4} \int_V E_y E_y^* dV = \frac{\varepsilon abd}{16} E_0^2 \]  

(3-11)

which at resonance is the same as the stored magnetic energy, \( W_m = W_e \).

The contribution of losses in the cavity is power dissipation in the cavity wall, dielectric loss in the substrate and radiated power. The power loss due to finite conductivity in the wall is expressed by [85],

\[ P_c = \frac{R_s}{2} \int_{\text{walls}} |H_t|^2 dS = \frac{R_s E_0^2 \lambda^2}{8 \eta^2} \left( \frac{ab}{d^2} + \frac{bd}{a^2} + \frac{a}{2d} + \frac{d}{2a} \right) \]  

(3-12)

where the surface resistivity \( R_s = \sqrt{\omega \mu_0 / 2 \sigma} \), and \( \eta^2 = \mu / \varepsilon \), and \( H_t \) is the tangential magnetic field at the wall’s surface.

Q-factor due to the conductive loss is calculated by

\[ Q_c = \frac{2 \omega_0 W_e}{P_c} \]  

(3-13)

The dielectric loss is expressed by

\[ P_d = \frac{1}{2} \int_V \vec{J} \cdot \vec{E}^* dV = \frac{\omega \varepsilon''}{2} \int_V |\vec{E}|^2 dV = \frac{abd \omega \varepsilon'' |E_0|^2}{8} \]  

(3-14)

where \( \varepsilon = \varepsilon' - j \varepsilon'' \), for the effective dielectric constant including substrate loss. The Q-factor due to substrate loss is estimated for \( \varepsilon'' \ll \varepsilon' \) to

\[ Q_d = \frac{2 \omega W_e}{P_d} = \frac{\varepsilon'}{\varepsilon''} = \frac{1}{\tan \delta} \]  

(3-15)
The total unloaded Q-factor for the cavity can be calculated to

\[ Q_0 = \left( \frac{1}{Q_c} + \frac{1}{Q_d} \right)^{-1} \]  

(3-16)

In this study, the lowest resonant mode, TE\textsubscript{101}-mode, for a rectangular metal cavity is considered for simplicity. That also makes a large frequency separation to other undesired modes.

The dielectric loss and the conductive loss in the cavity walls limit the Q-factor of the resonator. Therefore, this study uses an air-filled cavity in aluminum with \( Q_0 \) measured to 3800 which is further reported in Paper [C].

Similar calculations appear for dielectric resonators, which might support other more peculiar modes with several tens of thousands in Q-factor. The Whispery-Gallery mode is an example of an azimuthal propagation of the wave around the resonator [86], [87], etc.

Other popular modes for dielectric resonators are the evanescent modes where the field will cut off at the boundary of the high permittivity of a dielectric to the surrounded low dielectric air. This mode has a decreased height of the resonator, \( L \), and appears lower than \( 2L/\lambda_g < 1 \), which means that the resonator is reactively terminated at the ends. That mode is beneficial for having more dense volume, larger spurious-free region, and for being easier coupled with improved feasibility for monolithic integration.

The frequency tuning of high-Q resonators as metal cavities or ceramic resonators is non-trivial. A consequence of their high Q-factor is that they will have very limited coupling possibilities of tuning elements and accessibility to the resonator structure. In particular, if external loading or phase shifting techniques are used outside the resonator, these will have limited effect on the frequency tuning. A more efficient method is to directly disturb the EM-field inside the cavity as proposals presented with tunable ground plane using RF-MEMS switches as in Paper [A] and Paper [B], or by using GaAs-varactors connected to field probes inside the cavity as demonstrated in Paper [G].

Furthermore, the tuning elements require very high Q-factors to have minor degradation of the total Q-factor of the resonator. Also, the power (or voltage) capability can be a severe problem due to high voltage and current swing in high-Q resonators across the terminals of the tuning elements. The conversion of modulation noise must be considered, or in the case of electromechanical tuning as for MEMS-technology, the vibration sensitivity should be handled with care.

### 3.3.2 Tuning with an electronically controlled wall

In principle, the resonance frequency can be tuned by changing the distance between the short-circuiting planes in the z-direction of the cavity resonator. Paper [A] analyzes a method to do that by using RF-MEMS-switches on a PCB intruded in the cavity close to the wall as is illustrated in Figure 3-9. This method aims to digitally tune the frequency to
specific sub-bands, and the analog fine-tuning within the sub-band can be achieved by weakly coupled varactors.

![Image](a) ![Image](b)

Figure 3-9. E-field contribution with an electrically tunable wall inside the cavity. (a) RF-MEMSs in open state. (b) RF-MEMSs in a closed state.

Two effects mainly contribute to the frequency change. The electrical volume is contracted when the switches are in a closed state, due to the short-circuiting by the switched ground plane. The change in volume $-\Delta V_b$, that is obscured by the switches, creates the frequency change as [85],

$$\frac{\omega - \omega_0}{\omega_0} = -\frac{2\Delta V_b}{V_0} \quad (3-17)$$

The other effect is when switches are in open state, and they, besides, forming the original volume $V_0$ also add part of an increased volume $\Delta V_s$ beneath the switch elements with other medium with dielectric constant $\varepsilon_r$ as [85]

$$\frac{\omega - \omega_0}{\omega_0} = -\frac{(\varepsilon_r - 1)\Delta V_s}{2V_0} \quad (3-18)$$

The new oscillation frequency is denoted $\omega$, and the original resonance frequency is denoted $\omega_0$ corresponding to the volume $V_0$. Figure 3-10 (a) and (b), respectively, show HFSS-simulations with RF-MEMS switches in open state and closed state for a three-row configuration of RF-MEMS in parallel. The RF-field penetrates through the RF-MEMS switches in the open state, but will be shielded by the switches in the closed state.

![Image](a) ![Image](b)

Figure 3-10 HFSS-simulations of the E-field. (a) MEMSs are in open state. (b) MEMSs are in closed state.
3.3.3 Analog cavity tuning with varactors

Tuning technologies performed with varactors integrated on chip-resonator circuitry for example in MMIC-solutions as Colpitts oscillators are successful for a large tuning-range [69], but the processes have often poor Q-factors for the resonators. External high-Q resonators, example DROs, make the varactor coupling difficult to achieve for a broad tuning range, consequently due to their high Q-factors as previously mentioned. The discussed method to change a load capacitance outside the resonator is demonstrated in [72], [88], [71], or to change the phase condition with a drawback to not utilize the optimum Q-factor of the resonator is showed in [70] and [89]. These techniques provide a tuning range of only a few tenths of a percent. More attractive is to embed the varactors inside the resonators, as solid-state varactors integrated with SIW-resonators demonstrated in [90], or by using high-Q-MEMS varactors in SIW as in [91] and [92], or for lumped LC-resonators [74].

Paper [G] demonstrates an aluminum cavity with embedded bare die GaAs varactor diodes inside a cavity. They are exposed directly of the RF-field for most efficient coupling. The varactors align with the E-field in the waveguide. Their capacitance $C_{\text{var}}$ increases the electrical length of the waveguide, which is short-circuited at the end by a lid, and the resonance frequency will decrease. This is illustrated in Figure 3-11, and despite the model is developed for a two-conductor TEM-transmission line, it can principally model the behavior at resonance for a waveguide, in particular for the TE$_{101}$ mode with short-circuited condition for a $\lambda_g/2$-resonance. The $C_{\text{var}}$ is aligned with the E-field, transversally to the propagation direction.

![Figure 3-11](image)

Figure 3-11 An equivalent circuit of the terminated waveguide with an included varactor.

Two-port S-parameter measurements of the designed cavity have been performed for different setups. Figure 3-12 (a) and (b) show the reflection coefficient and unloaded Q-factor, respectively, for intrusion depth of the assembled position of the varactors of 1 mm. Figure 3-13 (a) and (b) show the reflection coefficient and unloaded Q-factor, respectively, for an intrusion depth of 2.5 mm. The larger intrusion depth gives a stronger coupling of the varactor, which increases the tuning range from 1.6 % to 2.0 %. The deviation to the compared HFSS-simulation depends on parasitic and tolerances for the different assembly alternatives. The total lower Q-factor of the cavity at higher frequencies, despite the better Q-factor internally for the varactor at more reversed bias, is explained by the cavity itself degrades the Q at a larger disturbance from its natural resonance frequency.
Figure 3-12. Varactors assembled at 1 mm depth in the cavity at different reversed voltages. (a) Measured reflection coefficient compared to HFSS-simulation in dotted black trace. (b) Measured Q-factor compared to HFSS-simulation in dotted black traces.

Figure 3-13. Varactors assembled at 2.5 mm depth in the cavity at different reversed voltages. (a) Measured reflection coefficient compared to simulation in dotted black trace. (b) Measured Q-factor compared to simulation in dotted black traces.

3.3.4 Varactor device characterization

The used varactors are manufactured by MA/Com of type MA46H146 bare-die and flip-chip mounted in GaAs-technology with abrupt $\gamma=0.5$ doping profile. They are characterized by a resonant Deloach-structure [93], shown in Figure 3-14 (a) and (b), and the measured transmission is shown in Figure 3-14 (c). Measurement extracts $R_s$ and $C_j$ to 2 ohm and 40 fF, respectively, for a reversed varactor voltage of -24 V. The unloaded $Q_0$ for the varactor, defined as $Q_0 = \text{imag}(Z_{var}) / \text{real}(Z_{var}) = 1/\omega C_j R_s = 200$ at 10 GHz.

![Varactor model diagram]

(a)
Figure 3-14 (a) Schematic of the test-structure. (b) Photo of the PCB test-circuit. (c) Measured transmission compared to simulation on the test-structure for different varactor voltages.

3.4 MEMS device properties

The used RF-MEMSs in this study are aimed for a digital sub-band selection for multichannel oscillators with moderate requirements on switching speed. The switches are characterized during electrostatic conditions by S-parameters to define a small-signal model.

However, in the application of the oscillator, the electroacoustic phenomena due to the electrodynamic behavior affect the phase noise. Particularly sensitiveness is at offset frequencies corresponding to the mechanical resonances.

3.4.1 Measured electrostatic behavior

The RF-MEMS is a mechanical system where an electrostatic force pulls down a membrane described as [22], [46], [94],

\[ F_e = -\frac{1}{2} \varepsilon_0 A \frac{V_{act}^2}{d^2} \]  

(3-19)

where \( A \) is the membrane area, \( d \) the distance between the membrane and the actuator, \( \varepsilon_0 \) the dielectric constant and \( V_{act} \) the actuation voltage. The electric force \( F_e \) counteracts a mechanic force \( F_m \) which is described as

\[ F_m = k \Delta z \]  

(3-20)

where \( k \) defines the spring constant of the membrane and \( \Delta z \) the displacement towards the actuator. For small displacements, a stable condition is established for \( F_m = F_e \), which makes a controllable capacitor, thus a varactor. The electric force, \( F_e \), increases by larger displacement and will counterbalance the increased mechanical force, \( F_m \). The needed electric force (or applied voltage \( V_{act} \)) to create a displacement \( \Delta z \) can be expressed as

\[ V_{act} = \sqrt{\frac{2k}{\varepsilon_0 A}} (d_0 - \Delta z)^2 \Delta z \]  

(3-21)
Equation (3-21) tells that there exists two possible displacements $\Delta z$ for the same actuation voltage $V_{act}$ lower than the certain voltage, $V_{pull-in}$. At $V_{pull-in}$ the both states coincide and instability makes the system collapse. Figure 3-15 illustrates this behavior.

![Figure 3-15](image)

Figure 3-15 Calculated actuation voltage versus the membrane displacement according to (3-21) for the used RF-MEMS. The maximum gap for unbiased state, $d_0 = 3 \text{ um}$. 

The membrane will then snap to the bottom position. This appears when $\Delta z = d_0 / 3$, [22], where $d_0$ is the maximum gap in the unbiased state. Typical values for the RF-MEMS used in this study are $A=0.170 \times 0.11 \text{ mm}$, $d_0=3 \mu\text{m}$ and the needed pull-in voltage, $V_{pull-in}=20 \text{ V}$. This evaluates to a spring constant of about $k=8 \text{ N/m}$. The spring constant for a cantilever MEMS can also be calculated as [22],

$$k_c = 2Ew \left( \frac{l}{t} \right)^3 \frac{1 - \frac{x}{l}}{3 - 4 \left( \frac{x}{l} \right)^3 + \left( \frac{x}{l} \right)^4}$$

(3-22)

where $l$ is the total length of MEMS-membrane (170 um), and $w$ is the width (110 um), $x$ is the distance from the anchor edge to where the actuator force starts due to the bottom electrode (about 20 um, seen in Figure 3-16 (b)) and $t$ the thickness of the MEMS-membrane (~2 um). The Young’s modulus for gold $E = 80 \cdot 10^9 \text{ N/m}^2$. That evaluates to $k_c=8.5 \text{ N/m}$, which is approximately the same as previous calculation.

![Figure 3-16](image)

Figure 3-16 Photo of the MEMS-die element. (a) The size of membrane: 170 um x 110 um. (b) A MEMS-switch where the cantilever membrane has accidently been bent up. The actuator electrode is about 130 um x 110 um and starts about 20 um from the anchor edge to the left.
A low spring constant is necessary to decrease the needed actuation voltage, but will also result in a low restoring force and therefore may suffer from stiction problems. For high-reliability applications, the spring constant should not be designed lower than 10 N/m [22].

3.4.2 RF-MEMS model

Reflection and isolation measurements for different actuation voltage are shown in Figure 3-17 (a,b), where the pull-in voltage is seen where the first switch elements snap-in, resulting in an isolation degradation in the insertion loss graph (at about 20 V-30 V). Further increased voltage forces more switches to close. The intermediate state causes an undesired stub which changes reflection phase rapidly versus frequency and reduces the isolation.

![Reflection and Isolation Measurements](image)

Figure 3-17 (a) Reflection coefficient for different actuation voltages. (b) Isolation for different actuation voltages. (c) Photo of the PCB test-circuit. (d) Chip photo of the MEMS, size 2970 um x 1615 um.

RF-model of the MEMS

![RF-Model](image)

Figure 3-18 Model of the used MEMS.

The components in the model in Figure 3-18 are extracted for $L_{par}$ and $C_{par}$ to 0.3 nH and 22 fF, respectively, in open state. In the closed state, the line inductance, $L_{par}$ is increased...
to 0.4 nH due to a prolonged length through the switch region. The ohmic contact resistance is estimated to 1 ohm in the closed state and infinity in open state, and the included bond wires are estimated to 1 nH. The parasitic capacitance of the transmission line to the control wire are found negligible, and more, those lines have a high impedance to protect further RF-propagation.

For a two-port measurement of a serial MEMS-switch, the total parasitic capacitance in the open state can be approximated from the isolation as

$$|S_{21}|^2 = 4\omega^2 C_{par,tot}^2 Z_0^2$$  \hspace{1cm} (3-23)

From Figure 3-17 the measured $S_{21}$=-27 dB at 10 GHz in open state, and by using (3-23) it can estimate $C_{par,tot}$=7 fF, or about 21 fF per switch element which are in serial. That agrees well with $C_{par}$ in the model in Figure 3-18.

The on-resistance in closed state can be calculated at low frequency as

$$S_{21} = 1 - \frac{R_{sw, on, tot}}{2Z_0}$$  \hspace{1cm} (3-24)

if the total impedance of MEMS-switch is $Z_s = R_{sw} + j\omega L$, and $\omega L \ll R_{sw}$.

RF-MEMS used in varactor application will, due to the beam instability at $1/3$ of the beam displacement at the pull-in voltage, have limited capacitance ratio. Normally the capacitance ratio measures to about 1:1.25 for cantilever beams, and 1:1.4 for fixed-fixed beams. For a three-plate design, this ratio can be increased to about 1:2. However, techniques as dual-air-gap can increase the ratio further. An example in [95] shows a capacitance ratio of 1:3.

### 3.4.3 Electromechanic behavior

The dynamic response of the mechanical system can be described as

$$m \frac{d^2 x}{dt^2} + b \frac{dx}{dt} + kx = f_{ext}$$  \hspace{1cm} (3-25)

where $x$ is the MEMS cantilever displacement, $b$ the damping coefficient, $k$ the spring constant and $f_{ext}$ an externally applied force. This expression is useful to determine the switching speed as well as potential electroacoustic resonances.

A corresponding transfer function which models the frequency response can be written as

$$\frac{X(j\omega)}{F(j\omega)} = \frac{1}{k \left(1 - \frac{1}{\omega/\omega_0}^2 + j\omega/(Q\omega_0)\right)}$$  \hspace{1cm} (3-26)

where the resonance frequency $\omega_0 = \sqrt{k/m_{mod}}$, and $Q = k/(\omega_0 b)$ is the quality factor of the resonant beam and $m_{mod}$ its modal mass about 0.35 to 0.45 times the static mass. The response of (3-26) is further discussed and analyzed in [22]. If assuming $m_0 = \rho t A$, and for the used switches in this study, $\rho = 19300$ kg/m$^3$ (gold), membrane thickness
56  Chapter 3 Component characterization

t = 3 \mu m, and A = 0.170 \times 0.11 \text{ mm}, evaluates to \( m_0 = 1.0 \times 10^{-9} \text{ kg} \). By using the earlier estimated spring constant of 8 N/m from the pull-in voltage measurement, electroacoustic resonance can be calculated to \( \omega_0 = 137 \times 10^3 \text{ rad/s}, \text{ or } 21 \text{ kHz} \).

The mechanical resonance is more obvious in the open state than in the closed state. Figure 3-19 (a) shows the noise hump appearing at slightly above 10 kHz in the open state of the MEMS with 12.3 \% tuning running at a center frequency of 10 GHz. Compared to the phase noise for the closed state in Figure 3-19 (b) the resonance has diminished.

![Figure 3-19 Phase noise of a MEMS-tuned cavity oscillator with a three-row MEMS configuration measured at different bias level versus offset frequency as reported in Paper [A]. (a) All switches are in the open state. (b) All switches are in the closed state.](image)

3.4.4 Potential improvements of used RF-MEMS

This study has found some issues and suggested improvements for the used RF-MEMS and they are discussed below.

**Stiction**

The stiction issues and fastened MEMS-membranes became a severe problem during the analysis. As several switch elements were biased in parallel with a considerable spread in pull-in voltage, between 20 V to 90 V, the highest voltage was used to enable all elements to the closed state. That implied a risk for causing bonding effect and damaging the membrane for those with lower pull-in voltage. Non-conductive capacitive MEMSs, which probably are more robust for permanent stiction, are in that perspective preferred.

**Resonances**

Electroacoustic resonance was found for the MEMS in particular in the open state, due to lack of mechanical stability for the membrane in this position. A redesign of another top electrode to fix the membrane for the open position would partly solve this problem.

More holes in the membrane and therefore a decreased modal mass would decrease the switching time or increase the resonance frequency and probably make it less sensitive for acoustic vibrations. The effect of the squeeze film damping is also reduced with holes in the membrane which would further decrease the switching time.
Reliability
A capsulated RF-MEMS would be preferred to avoid contamination of glue and solder paste in the assembly processes.

The MEMS-switch could be designed to cover an analog tuning as a varactor. In this case, a combined technology could enhance for both PLL-locking and digital sub-band control.

3.5 Tuning components and technology comparison
Different high-Q tuning techniques compared to methods used in this study are presented in Table 3-2, regarding tuning range and unloaded Q-factor. In oscillator applications, other aspects as power handling, power loss, or vibration sensitivity must also be considered. Semiconductor varactor tuning has previously been discussed with drawbacks of large-signal degradation of Q-factor when strong coupling and high RF-power are applied. RF-MEMS varactors are robust for high power, but suffers from poor vibration robustness and may have non-continuous tuning characteristic due to the pull-in voltage.

Table 3-2 Comparison of different high-Q tuning techniques

<table>
<thead>
<tr>
<th>Ref</th>
<th>fc (GHz)</th>
<th>Tuning-ratio (TR) (%)</th>
<th>Q&lt;sub&gt;0&lt;/sub&gt;</th>
<th>Technology</th>
</tr>
</thead>
<tbody>
<tr>
<td>[96]</td>
<td>32.5</td>
<td>10</td>
<td></td>
<td>Ferroelectric</td>
</tr>
<tr>
<td>[43]</td>
<td>3</td>
<td>67</td>
<td></td>
<td>YIG</td>
</tr>
<tr>
<td>[97]</td>
<td>1</td>
<td>28</td>
<td>65</td>
<td>Varactor diode</td>
</tr>
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<td>12.5</td>
<td>1.23</td>
<td>150</td>
<td>SIW and switchable varactor diode</td>
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<tr>
<td>[91]</td>
<td>3.45</td>
<td>90</td>
<td>300-650</td>
<td>SIW and 3D MEMS varactor/switch</td>
</tr>
<tr>
<td>[92]</td>
<td>4.8</td>
<td>33</td>
<td>300-500</td>
<td>SIW and MEMS varactor/switch</td>
</tr>
<tr>
<td>[99]</td>
<td>6.425</td>
<td>5.4</td>
<td>140-240</td>
<td>SIW and MEMS varactor/switch</td>
</tr>
<tr>
<td>[100], [74]</td>
<td>3.475</td>
<td>6</td>
<td></td>
<td>MEMS varactor</td>
</tr>
<tr>
<td>[101]</td>
<td>23</td>
<td>5</td>
<td>750-1450</td>
<td>MEMS cantilever</td>
</tr>
<tr>
<td>Paper [A]</td>
<td>10.065</td>
<td>2.1</td>
<td>700-1115</td>
<td>MEMS digital (2 steps), one row@1 mm</td>
</tr>
<tr>
<td>Paper [A]</td>
<td>9.915</td>
<td>8.1</td>
<td>400-500</td>
<td>MEMS digital (2 steps), one row@2.5 mm</td>
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<tr>
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<td>MEMS digital (8 steps), three rows@1 mm</td>
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<tr>
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<td>12.3</td>
<td>250-826</td>
<td>MEMS digital (8 steps), three rows@2.5 mm</td>
</tr>
<tr>
<td>Paper [G]</td>
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<td>1.5</td>
<td>300-650</td>
<td>Varactor diode, two rows@1 mm depth</td>
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<tr>
<td>Paper [G]</td>
<td>10.4</td>
<td>2</td>
<td>300-400</td>
<td>Varactor diode, two rows@2.5 mm depth</td>
</tr>
</tbody>
</table>
Chapter 4.

Oscillator Design and Characterization

This section describes the oscillators designed and analyzed in this study. Considerable efforts are spent on how to optimize the phase noise given the discussed boundary conditions. Critical conditions for the active device are, as earlier mentioned, the power handling capacity and RF noise figure. Therefore, GaN has followed as a good choice, which has been the used technology for all designs. Furthermore, a high-quality resonator is needed, which in the first design is implemented with a cavity and discussed in Chapter 4.2.1. A design with a hybrid technology implemented with a discrete LC-network is shown in Chapter 4.3. Besides the importance of a high unloaded quality-factor, the resonator has to be optimally coupled to take advantage of its potential, which is studied throughout all oscillator designed. A MMIC-oscillator achieves the optimum coupling by utilizing an impedance transformed quasi-lumped resonator to match the gain of the amplifier as is shown in Chapter 4.4.1. A different approach to match the power coupling to the resonator is by adjusting the gain of the amplifier. Chapter 4.4.2 demonstrates an electronically adjustable reflection amplifier.

A digitally tunable cavity oscillator using an electronically movable ground plane of MEMS-switches with minor degradation of the performance is discussed in Chapter 4.2.2. Analog tuning by using an additional tuning element of semiconductor varactors is discussed in Chapter 4.2.2.

4.1 Experimental setup

Phase noise measurements have been performed with the PLL-method, using a commercial signal source analyzer from Rohde & Schwarz, FSUP50 [102]. Figure 4-1 shows a block diagram of the PLL-method. Only free-running oscillators are tested in this study.

![Block diagram of the PLL-method](image)

Figure 4-1 A block diagram is showing the used PLL-method [103].
The noise floor in the instruments may limit when measuring high-performance oscillators, in particular at a high carrier frequency. Table 4-1 shows the typical phase noise performance for the used FSUP50 for different offset and carrier frequencies. An important test condition in this study is at 100 kHz offset frequency at 10 GHz carrier frequency. The instrument performance at this condition is not sufficient to measure the cavity oscillator reported in Paper [C].

Table 4-1 FSUP50 phase noise sensitivity versus input frequency and offset [102].

<table>
<thead>
<tr>
<th>Offset frequency</th>
<th>Input frequency and Input level +10 dBm, Typical values R&amp;S®FSUP50 (dBc/Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>10 MHz</td>
</tr>
<tr>
<td>1 kHz</td>
<td>-161</td>
</tr>
<tr>
<td>10 kHz</td>
<td>-168</td>
</tr>
<tr>
<td>100 kHz</td>
<td>-170</td>
</tr>
<tr>
<td>1 MHz</td>
<td>-175</td>
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<tr>
<td>10 MHz</td>
<td>-179</td>
</tr>
<tr>
<td>30 MHz</td>
<td></td>
</tr>
</tbody>
</table>

To improve the noise floor in the measurement system, two nearly identical oscillators are used as is illustrated in Figure 4-2. One of these is slightly shifted in frequency, and the phase noise measurement is performed on the down converted and significant lower IF-frequency, which in our study is around 30 MHz.

![Diagram](image)

Figure 4-2 Setup of mixed measurement.

The effect of combining to equal and uncorrelated oscillators is 3 dB extra noise contribution. The measured noise figure on the down-converted signal has to be afterward compensated.

### 4.2 Cavity based oscillators

A cavity based oscillator is using a testfixture containing a MMIC reflection amplifier on a PCB for connecting the resonator. The resonator is varied in different designs, as implementation for fixed frequency, or enhancing tunability by RF-MEMS-switches or by varactors.

#### 4.2.1 Fixed frequency

The resonator for fixed frequency is simply an empty aluminum cavity dimensioned for the TE$_{101}$-mode. The cavity is placed on top of an exciting microstrip line connected to a
reflection amplifier. The amplifier is a MMIC-amplifier designed in a 0.25-μm GaN-HEMT process at TriQuint 3MI process. Figure 4-3 (a) shows a photo of the oscillator with the cavity and Figure 4-3 (b) shows a schematic.

Figure 4-3 (a) Picture of the cavity oscillator. (b) Schematic.

The setup is very flexible to adjust the cavity coupling by changing the perpendicular position to the microstrip line, and the phase condition by sliding the cavity along the strip. The oscillation frequency is about 10 GHz.

Table 4-2 shows the phase noise performance at 100 kHz and 400 kHz, respectively, for different cavity positions and open loop gain. The best performance is achieved for the weakest coupling with an open loop gain close to unity. Reason is that high gain will force the amplifier in deep compression, which is not beneficial for up-converted flicker noise and thermal noise.

Table 4-2 Measured phase noise versus coupling of the resonator

| Cavity position offset (mm) | $R_s$/loss (Ω)/(dB) | Optimum phase noise performance @100 kHz (dBc/Hz) @bias level ($V_d/V_g$) | Phase noise @400 kHz (dBc/Hz) | FOM$^{(1)}$ @400 kHz (dB) | $F_{op}^{(2)}$ @400 kHz (dB) | Gain (dB) Reflection Amp @ opt for 100 kHz bias ($V_d/V_g$) | $|R_N|^{(3)}$ (Ω) | $R_s$/$|R_N|$ | Open loop gain$^{(4)}$ (dB) |
|---------------------------|--------------------|---------------------------------|------------------------------|--------------------------|--------------------------|--------------------------|------------------|------------------|---------------------|
| 0.6                       | 8.1/2.8            | -139@4/-1.9                    | -152.5                      | 219                       | 15                       | 6.8                      | 18.7             | 0.4              | 4                    |
| 3.4                       | 14/5               | -141.5@4/-1.8                  | -155.5                      | 223                       | 10                       | 7.0                      | 19.1             | 0.7              | 2                    |
| 3.9                       | 19.2/7             | -141.5@5/-1.6                  | -156                        | 221                       | 9                        | 8                        | 21.6             | 0.89             | 1                    |
| 5.9                       | 20/7.4             | -144.5@5/-1.8                  | -160                        | 227                       | 7                        | 7.9                      | 21.3             | 0.94             | 0.5                 |

$FOM=\text{-}\mathcal{L}(\Delta f)+20\log(f_0/\Delta f)-10\log(P_{DC}/1\text{mW})$

$F_{op}=173.9+10\log[8\eta\beta/(1+\beta)]+20\log(Q_0)-\text{FOM}$

The negative resistance is the real part after phase compensation to a serial resonator in 50 ohm system.

$G_{loop}=20\log([-R_{n}-Z_0]/(R_s+Z_0)+20\log([-R_n]+Z_0)/(-R_n+Z_0)]$

The phase noise for the optimally coupled resonator is shown versus bias at 100 kHz offset in Figure 4-4 (a) and versus offset frequency for optimum bias in Figure 4-4 (b). Optimum $FOM$ at 400 kHz offset is measured to 227 dB. Under the same optimum condition, the power was +5 dBm, the power consumption $P_{DC}$=200 mW, and the flat noise floor -165 dBc/Hz.
The phase noise of this oscillator is below the noise floor of the instrument used for measurement. Therefore, the setup proposed in Chapter 4.1 with two equally oscillators for down-converting the signal was used. Once the best achievable performance of the two equally tuned and biased oscillators is determined, one of these oscillators is kept fixed at this condition as a reference oscillator. Different test conditions have been applied to the other oscillator. The measured results are compensated for the noise floor of the reference oscillator. Figure 4-5 shows a photo of the test setup.

The design is further analyzed in Paper [C].

4.2.2 Tunable frequency

The frequency tunability for the resonator on the first design is implemented by a tunable ground plane inside the cavity of an intruded PCB of MEMS-switches. In the second design, semiconductor varactors coupled to the E-field are assembled on the PCB.

Figure 4-6 shows a sketch of the complete setup of the MEMS-tuned oscillator, with the intruded PCB illustrating the positions of the MEMS-switches and a schematic of the setup. The setup is further demonstrated in Paper [B].
Figure 4-6 Cross section sketch of the setup of a cavity oscillator with RF-MEMS-switches forming an electronically tunable wall.

Different assembly alternatives of the switches, for a one-row configuration and a three-row configuration are shown in Figure 4-7 (a-b), respectively. Figure 4-7 (c) shows a photo of the completely assembled PCB.

Figure 4-7 Photo of the assembled switches and PCB. (a) One-row configuration. (b) Three-row configuration. (c) A layout of the PCB. The cavity opening is 20.7 mm x 10.35 mm.

The positions of the MEMS-switches, by varying the intrusion depth of the PCB (1 mm and 2.5 mm) and the configuration on it (one row or three rows) are analyzed with trade-offs for tuning-range, frequency resolution, Q-factor and phase noise. The oscillation frequency is about 10 GHz. Figure 4-8 shows a summary of the studied setups with frequency tuning versus the bit position of the MEMS-states.

Table 4-3 shows a more detailed analysis of the performance.
Figure 4-8 Summary of the studied MEMS-configurations versus oscillation frequencies. State “0” equals open MEMS-state, and “1” equals closed MEMS-state, for the assembled positions, respectively. State “x” denotes a not assembled position.

Table 4-3 Tuning range, Phase noise and Q-factor, versus simulations for different MEMS-setup.

<table>
<thead>
<tr>
<th>Setup</th>
<th>State</th>
<th>Measurement</th>
<th>Simulation</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>Q_0</td>
<td>Freq (GHz)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>One-row MEMS @1 mm depth.</td>
<td>x0x</td>
<td>1115</td>
<td>9.96</td>
</tr>
<tr>
<td></td>
<td>x1x</td>
<td>700</td>
<td>10.17</td>
</tr>
<tr>
<td>One-row MEMS @2.5 mm depth.</td>
<td>x0x</td>
<td>400</td>
<td>9.51</td>
</tr>
<tr>
<td></td>
<td>x1x</td>
<td>500</td>
<td>10.32</td>
</tr>
<tr>
<td>Three-row MEMS @1 mm depth.</td>
<td>000</td>
<td>1050</td>
<td>9.84</td>
</tr>
<tr>
<td></td>
<td>001</td>
<td>700</td>
<td>10.07</td>
</tr>
<tr>
<td></td>
<td>011</td>
<td>700</td>
<td>10.22</td>
</tr>
<tr>
<td></td>
<td>111</td>
<td>1123(*)</td>
<td>10.33</td>
</tr>
<tr>
<td>Three-row MEMS @2.5 mm depth.</td>
<td>000</td>
<td>300</td>
<td>9.42</td>
</tr>
<tr>
<td></td>
<td>001</td>
<td>250</td>
<td>10.05</td>
</tr>
<tr>
<td></td>
<td>011</td>
<td>350</td>
<td>10.43</td>
</tr>
<tr>
<td></td>
<td>111</td>
<td>826(*)</td>
<td>10.65</td>
</tr>
<tr>
<td>Empty PCB @1 mm depth.</td>
<td></td>
<td>1817</td>
<td>9.998</td>
</tr>
<tr>
<td>Empty PCB @2.5 mm depth.</td>
<td></td>
<td>1602</td>
<td>9.925</td>
</tr>
</tbody>
</table>

(*) Lower on-resistance measured than assumed in the model.  
(**) Too low loop gain to measure with the used amplifier.  
(***) Tuning range is defined as 2(f_{max}-f_{min})/(f_{max}+f_{min})

From Table 4-3 it is obvious that the tunability increases by the intrusion depth of the PCB, and consequently degrades the phase noise performance. The tuning also compresses, when the number of MEMSs set in the closed state are increased. This is probably due to more distorted RF-field, which complicates the positioning of the MEMS to achieve a
uniform frequency resolution versus the states. However, the concept of using MEMS-switches to tune the ground plane inside the cavity works in general with good result. They can be used for digital control of the resonance frequency over a quite broad range, 12\% with phase noise performance in the range of -133 dBc/Hz to -123 dBc/Hz at 100 kHz offset frequency. $FOM$ and $FOM_T$ are 196 dB and 198 dB, respectively. These results are further analyzed and reported in Paper [A].

Another oscillator is based on a varactor-tuned cavity which is analyzed in Paper [G]. The varactors are assembled with similar techniques as reported in Paper [A], on a PCB intruded inside the cavity. Figure 4-9 shows a picture of the complete setup and including sub-blocks.

![Figure 4-9 Picture of the oscillator for a varactor-tuned cavity.](image)

The varactors are aligned with the E-field, and their capacitance increases the electrical length of the waveguide, which is short-circuited at the end by a lid. A model of the tuned cavity was discussed in Chapter 3.3.3.

![Figure 4-10. Varactor cavity at 1 mm depth. Phase noise (dBc/Hz) versus varactor voltage (-20 V to -0.1 V) for optimum bias $V_d/V_f=11.5$ V/-1.8 V. Support lines for -20 dB/dec and -30 dB/dec are added.](image)
Figure 4-11. Varactor cavity at 1 mm depth. (a) Phase noise at 100 kHz and 1 MHz offset frequency, respectively, versus frequency for optimum bias $V_d/V_g=11.5$ V/-1.8 V. (b) Frequency and output power versus varactor voltage.

Figure 4-10 shows the phase noise versus offset frequency. The varactor voltage is swept from 0 to -20 V. Figure 4-11 (a) shows phase noise versus oscillation frequency at offset frequencies of 100 kHz and 1 MHz, at the bias point $V_d/V_g=11.5$ V/-1.8 V. For offset frequencies lower than 200 kHz the flicker noise contributes to a -30 dB/decade slope. At offset frequencies in the range 200 kHz to 1 MHz thus for the -20 dB/decade region, the phase noise variation is minor affected versus varactor voltage and tuning sensitivity. This indicates that the modulation noise is less dominating, which otherwise should have strong effect in this region. However, the change in varactor bias and tuning frequency influence on the total Q-factor due to the large-signal degradation, which contributes to the overall variation in the phase noise. Figure 4-11 (b) shows the tuning frequency and the output power versus varactor voltage for the optimum bias, $V_d/V_g=11.5$ V/-1.8 V. The best phase noise is measured at the most reversed varactor voltage, thus at less large-signal degradation of Q due to most margin for RF-voltage across the varactors and most power capability.

The optimum phase noise at 100 kHz offset and 1 MHz offset, respectively, measures -118 dBc/Hz and -146 dBc/Hz. The measured small signal $Q_0=400$, which can be scaled to the $Q_0$-factor of the fixed frequency cavity oscillator in Paper [C] with expected 20 dB lower phase noise. A deviation in further degradation of 7 dB from this value indicates that the phase noise is not primarily limited by the unloaded $Q_0$ of the cavity but rather by modulation noise proportional to the tuning sensitivity and large-signal effects. Previously, Chapter 2.6.1 discussed the power dependent Q-factor due to the RF-voltage limitation across the varactors. Further, the measured oscillator has higher bias current and gain for loss compensation at optimum, which introduces more flicker noise, in particular at 100 kHz offset.

### 4.3 Flexible hybrid oscillator

An oscillator was developed as a testbed to use LF-noise characterized devices in an oscillator setup. The oscillation frequency is 1 GHz, and phase noise is studied versus offset frequency and bias level.

A hybrid oscillator based on GaN-HEMT device and a lumped element resonator implemented on a low loss PCB has been designed. A photo and a schematic are shown in Figure 4-12 (a) and Figure 4-12 (b), respectively. The testbed is made flexible so that a bare die transistor device can be close connected between two pieces of PCB, one with the
termination network on the gate side, and one with different resonator topologies on the drain side. \( R_{\text{stab}} \) is for stabilization and suppresses out-of-band resonances.

![Figure 4-12](image_url) (a) Photo of the hybrid oscillator. (b) Schematic.

![Figure 4-13](image_url) (a) Measured phase noise in dBC/Hz versus bias level for different offset frequency. (a) 100 kHz offset frequency. (b) 1 MHz offset frequency.

Figure 4-13 (a) and (b), respectively, show measured phase noise at 100 kHz and 1 MHz. It is found that at 100 kHz offset, the performance is limited by flicker noise, and stays essentially constant if power is increased. Thus, the benefits with higher power level counterbalance the increased flicker noise at higher current. At 1 MHz, the increased signal power due to higher bias level is beneficial for the phase noise.

![Figure 4-14](image_url) (a) Measured phase noise is shown. Support lines for -20 dB/decade and -30 dB/decade are added. (b) Measured LF noise is shown of the used 8x50 um GaN at \( V_{\text{GS}} = 10 \) V for different \( I_{\text{ds}} \) versus simulation.
Figure 4-14 (a) shows measured phase noise for all swept bias level, and Figure 4-14 (b) shows measured LF-noise, fitted to a simulation model, with $k_f=6\times10^{-10}$, $A_f=0.87$, and $F_{fe}=1.5$ according to (2-11). The noise floor for this component is quite low, compared to the thermal noise current of a 50 $\Omega$ resistor that measures $3 \cdot 10^{-22}$ A$^2$/Hz. Optimum phase noise is measured for an open loop-gain around 1.5 dB, and the equivalent loss resistance in the serial resonator is about $R_S=11\ \Omega$ or -4 dB, and $Q=49$. The measured reflection gain for the transistor at optimum bias $V_d/V_g=5V/-2V$ is about 5.5 dB, or a corresponding $|R_N|=15\ \Omega$ and $R_S/|R_N|=0.7$.

At 1 MHz offset frequency, a phase noise of -150 dBc/Hz is measured from 1 GHz oscillation frequency, and a power-normalized figure of merit (FOM) of 186 dB is reached. The noise figure $F=21\ dB$, out of which 2-3 dB is dependent on a none-optimum coupling, and 2 dB due to a DC to RF-efficiency of 65 %. The noise figure compensated for the finite efficiency and none-optimum coupling is 17 dB. The results are further analyzed in Paper [F].

4.4 MMIC based oscillators

This work presents two MMICs, one integrated oscillator discussed in Chapter 4.4.1, and one reflection amplifier with electronic gain control intended for an external resonator discussed in Chapter 4.4.2.

4.4.1 Integrated oscillator

The design uses the same active transistor as the previously presented reflection amplifier for the cavity oscillators and is integrated to a 15 GHz on-chip resonator. The resonator is implemented as a parallel resonator, that is transformed by a quarter-wavelength transformer and adjusted to fulfill the phase condition. Figure 4-15 (a) and (b), respectively, show a chip photo and a schematic of the oscillator.

![Chip photo of the oscillator](image)

![Schematic](image)

Figure 4-15 (a) Chip photo of the oscillator, size 2.0x1.0 mm. (b) Schematic.

The transformer gives flexibility to match the coupling factor, or the impedance, to the active device. Matching the gain of the amplifier, or equivalently the negative resistance $R_N$, to the equivalent serial resistance $R_S$ of the resonator according to (2-45) and (2-47),
is difficult by scaling the width of an open quarter-wavelength microstrip stub. Schematic of such resonator is depicted in Figure 4-16 (a), and calculation of achievable $R_s$ is

$$R_s \approx \frac{c_0}{4W} \sqrt{\frac{\pi \mu_0}{\sigma \varepsilon_r f}} \quad (4-1)$$

where $c_0$ is the free space velocity, $\varepsilon_r$ relative permittivity, $\mu_0$ permeability, $W$ is the width of the microstrip line, $\sigma$ the metal conductivity, and $f$ the frequency. If connected to a lumped parallel resonator as a quarter-wavelength transformer shown in Figure 4-16 (c), the dimensions will be more feasible for the lumped resonator as

$$L = \frac{Z_c^2}{R_s \omega_0 Q} \quad (4-2)$$

$$C = \frac{Z_c^2 Q}{R_s \omega_0} \quad (4-3)$$

where $L$ and $C$ are the corresponding lumped resonator components and $Q$-factor to achieve a certain $R_s$ using a transformer with characteristic impedance $Z_c$. Another topology to match the coupling can be achieved by a coupled $\lambda/2$ resonator as in Figure 4-16 (b). The separation between lines in the coupler-resonator can be used to set the desired coupling factor. However, the topology is quite area consuming.

Phase noise results from the oscillator with the transformed resonator is shown in Figure 4-17 (a-b), versus offset frequency for all measured bias, and at a fixed offset of 100 kHz versus bias, respectively. The optimum phase noise measures -106 dBc/Hz at 100 kHz for an oscillation frequency of 15 GHz with a corresponding $\text{FOM}=191$ dB. The results are further reported in Paper [E].
4.4.2 Reflection amplifier with adjustable gain

An oscillator with the feature to optimize the resonator coupling by adjusting the gain of the reflection amplifier is reported in Paper [D]. The two different methods by optimizing the gain in comparison to a change in the coupled impedance of the resonator was discussed earlier in Chapter 2.5.2. The amplifier is made for 8.5 GHz, with an adjustable gain between 0 dB to +7 dB, and it is designed and fabricated in a 0.25-μm GaN-HEMT process at UMS, GH25-10 process.

A connected cavity locks the frequency to 8.5 GHz, and for different varactor voltages, the reflection gain is changed along the black line in Figure 4-19 (a). Figure 4-19 (b) shows the corresponding phase variation at the fixed frequency of 8.5 GHz, which is less than +/-10 deg, and have minor impact on the resonator phase condition.
In Figure 4-20 (a), the starting position is a strong coupling (low $R_s$) which means that the amplifier gain $-R_N$, has to be lowered by a higher varactor voltage, compared to the situation in Figure 4-20 (b). In Figure 4-20 (b) the starting position of the cavity is weak coupling ($R_s \approx |R_N|$), and the gain increases with a lower varactor voltage. In both cases the optimum phase noise performance appears at the same bias level, showing that the gain tuning has a minor impact on the bias level of the active transistor.

The minimum phase noise reached for optimum gain setting is -136 dBc/Hz, which can be compared to -139 dBc/Hz that was obtained with the mechanical tuning of the resonator impedance, i.e., by moving the cavity. Despite the slight degradation due to the electronically controlled circuitry, the design is motivated as it easily can be made adaptive to compensate for process and temperature variations. The loop-gain can also be adjusted to ensure a proper start-up condition, and then be optimized for the stationary condition. FOM at 100 kHz for an electronically gain control and manual gain control measures to 212 dB, and 215 dB, respectively.
Chapter 5.

Conclusions

The thesis has demonstrated optimization of oscillators targeting low phase noise. Choice of technology for the active device and the resonator is important, but as crucial is to utilize and combine the included parts to form a well-designed oscillator. Parameters as the power capacity, RF noise floor, and flicker noise for the amplifying part and the unloaded quality-factor for the resonator circuitry set boundaries for the theoretically achievable performance.

In this work, an effective noise figure is defined to measure the performance against the theoretical noise level. It has been shown that the performance in the 20 dB region is limited by the Q-factor and the DC-power efficiency. This means that the commonly used power-normalized figure-of-merit figure, $FOM$, is only limited by the unloaded quality-factor of the resonator and the thermal noise floor. To reach the noise floor, the resonator must be optimally coupled to the active device, which means a trade-off between power coupled to the resonator and loaded Q-factor. Experimentally, the optimal coupling factor is found to be $\beta=1$, in contrast to $\beta=1/2$ which is predicted by a small-signal linear model. One reason for the deviation is that the model does not consider non-linear noise up-conversion.

The thesis reports on several state-of-the-art oscillators. A cavity based oscillator shows an excellent phase noise of -145 dBc/Hz at 100 kHz from a 9.9 GHz oscillation frequency. A MMIC oscillator demonstrates a phase noise of -106 dBc/Hz at 100 kHz from a 15 GHz carrier. The two oscillators present an effective noise figure of about 16 dB above the theoretical noise floor. A substantial contribution to the effective noise figure is the finite DC to RF efficiency, which for the cavity oscillator and MMIC oscillator measures around 6 % and 8 %. Compensation for the efficiency shows operational noise figures for the cavity oscillator and MMIC-oscillator of 4 dB and 5 dB, respectively, extracted in the $1/f^2$ region.

One of the factors most critical for the operational noise figure is the resonator coupling factor, which may be tuned in different ways, e.g., mechanically or by changing the bias condition to control the gain of the reflection amplifier. The thesis also reports on a reflection amplifier with separate electronic gain control. It demonstrates electronic tuning with phase noise within 3 dB from the performance of a mechanically tuned cavity oscillator based on the same technology. The electronic tuning function can be used for compensation of assembly tolerances and temperature spread.
Beside low-phase-noise fixed frequency oscillators, different tuning technologies are also considered. It is investigated how the tuning elements limit the performance of tunable oscillators. High-Q tuning elements with good power capacity are for this reason, RF-MEMSs. We have demonstrated digital tuning by using switches forming an electronically tunable ground plane inside a cavity resonator. The position of the MEMSs on a PCB intruded at different depth in the cavity are investigated regarding the tuning range, frequency resolution, Q-factor, and phase noise. The design demonstrates a 12.3% digital tuning with phase noise varying from -133 dBc/Hz to -123 dBc/Hz at 100 kHz offset frequency around an oscillation frequency of 10 GHz.

For analogue tuning, solid state varactors are investigated. The setup is very similar as for the MEMS setup, with GaAs varactors assembled on the PCB. The varactors are aligned with the E-field of the TE\textsubscript{101} resonant mode inside the cavity to form an efficient coupling for varied positions. The coupling is chosen for or a compromise between tunability and noise performance. A tuning range of 1.6% around a 10 GHz oscillation frequency is recorded with a phase noise down to -118 dBc/Hz at 100 kHz offset. A problem with the varactors are their finite power capability limiting the large-signal Q-factor. Measurements of the cavity for low excitation power show significantly higher $Q_0$ than for the power levels valid in the resonator under oscillation condition. The large-signal Q-degradation is identified as the main reason why the measured phase noise is worse compared to simulations based on small-signal Q-factor. This motivates usage of high power capacity varactors as SiC-varactors for tuning of high Q-resonators.

The RF-MEMS tuning has shown some issues regarding electroacoustic conversion, i.e., microphony. In particular, for MEMS-switches in the open state with no forces applied on the membrane, mechanical resonances can appear and modulate the signal. For this reason, MEMS-based varactors are not preferable, despite their extremely high power capability. A remedy to suppress the resonances is to prevent the membrane from motion in the static states. A secondary electrode, which could fix the membrane in the open state, would be beneficial. Further investigation also includes some reliability issues regarding stitching and bonding effect of the membranes.

This work has demonstrated optimization and analysis of oscillators. Parameters to influence by design are discussed to optimally utilize the performance of the including parts and achieve performance close theoretical bounds. By combining different technologies, excellent phase noise and good tunability are demonstrated. The thesis has shown methods and strategies to build optimized oscillators, which are needed to meet the demands of cutting edge technology in future communication systems.
Summary of appended papers

Paper [A]
Analysis of a MEMS Tuned Cavity Oscillator on X-band
In this paper, a cavity oscillator on X-band with an electronically tunable wall of RF MEMS-switches are analyzed in detail regarding the position of the MEMSs’ for best tuning range, phase noise, and digital resolution.
My contribution is design and implementation of the oscillator, verification, and analysis of the result. I have been the first author of the paper.

Paper [B]
RF-MEMS Tuned GaN HEMT based Cavity Oscillator for X-band
In this paper, a cavity oscillator on X-band with an electronically tunable wall of RF MEMS-switches shows a digital tuning capability of 5 %, maintaining good phase noise of about -140 dBc/Hz at 100 kHz offset. A novel building practice of an aluminum cavity with an intruded PCB of MEMS switches coupled to a microstrip line connected to a GaN-HEMT reflection amplifier is demonstrated.
My contribution is design of the oscillator with MEMS-tuning, verification, and analysis. I have been the first author of the paper.

Paper [C]
Phase Noise Analysis of an X-Band Ultra-low Phase Noise GaN HEMT based Cavity Oscillator
In this paper, a state-of-the-art cavity oscillator for fixed frequency on X-band is presented. It reports an analysis to find the best resonator coupling, by changing the position of the cavity.
My contribution is design of the oscillator, verification, and analysis of the results. I have been first author of the paper.

Paper [D]
A GaN HEMT X-band Cavity Oscillator with Electronic Gain Control
This paper presents a novel method to change the gain of a MMIC-reflection amplifier in GaN-HEMT, without degrading the Q-factor for a connected resonator. By an electronic change of the gain, the optimum coupling to the resonator can always be maintained, which is essential for a well-designed oscillator.
My contribution is MMIC-design, verification, and analysis of the oscillator. I have been the first author of the paper.
Paper [E]
Low phase noise power-efficient MMIC GaN-HEMT Oscillator at 15 GHz based on a Quasi-lumped on-chip resonator
This paper reports on a fixed frequency MMIC oscillator, with an integrated resonator of lumped LC, which is optimally coupled by an impedance transformation to the active transistor. It shows a state-of-the-art performance of a fixed frequency oscillator in GaN-HEMT.
My contribution is verification and analysis of the design. I have been the first author of the paper.

Paper [F]
Phase noise analysis of a tuned-input/tuned-output oscillator based on a GaN HEMT device
This paper reports on a hybrid oscillator, with a GaN-HEMT device connected to a lumped LC-resonator on a PCB. By changing the resonator impedance, an optimal coupling is found, and the optimum bias for best phase noise is analyzed for different frequency offset to show the effect of flicker noise. The GaN-HEMT device is characterized regarding flicker noise.
My contribution is designing the oscillator, measuring, and analyzing the results. I have been the first author of the paper.

Paper [G]
An X-band varactor-tuned cavity oscillator
This paper reports on a varactor tuned cavity oscillator. An embedded PCB with mounted varactors inside the cavity demonstrates a tuning capacitance efficiently coupled to the RF-field. This method enhances a rather large tuning-range and maintaining good Q-factor of the resonator.
My contribution is designing the oscillator, measuring and analyzing the results. I have been the first author of the paper.

Paper [H]
Low-Frequency Noise Measurements - A Technology Benchmark with Target on Oscillator Applications
This paper reports on low-frequency noise measurements of different devices, as GaN-HEMT, InGaP HBT, and GaAs pHEMT with high bias level in the saturated region for targeting oscillator application.
My contribution is developing the measurement setup.
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Paper [A]
Analysis of a MEMS Tuned Cavity Oscillator on X-band

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accepted for to *IEEE Transactions on Microwave Theory and Techniques*, 2017.
Analysis of a MEMS Tuned Cavity Oscillator on X-band

Mikael Hörberg, Thomas Emanuelsson, Herbert Zirath, Fellow, IEEE, and Dan Kuylenstierna

Abstract—This paper reports on the analysis of a radio frequency micro electromechanical systems (RF-MEMS) tuned cavity oscillator on X-band based on a GaN-HEMT MMIC reflection amplifier. The RF-MEMS-switches are mounted on a low loss PCB intruded in an aluminum cavity that is coupled to a microstrip line connected to the reflection amplifier. The paper investigates the influence of the number of switches as well as their positions with respect to phase noise and tuning range. Vertical and horizontal positions of the switches are varied with target on optimum trade-off between phase noise and total tuning range. For a three-row MEMS-configuration at 1 mm depth from the end cavity wall, a tuning range of 4.9% is measured. The center frequencies are ranging from 9.84 GHz to 10.33 GHz with measured phase noise of -140 dBc/Hz to -129 dBc/Hz at 100 kHz offset. A similar three-row MEMS setup at 2.5 mm depth provides a tuning range of 12.3 % with measured phase noise of -133 dBc/Hz to -123 dBc/Hz at 100 kHz offset.

Index terms—Cavity, GaN HEMT, oscillator, phase noise, radio frequency micro electromechanical systems (RF-MEMS).

I. INTRODUCTION

UNABILITY and reconfigurability are key features in modern communication systems using generic and common broadband radio products. This puts requirements on the components to have a high bandwidth, a low weight, a low volume and a low cost without sacrificing capacity.

Designing the frequency generation circuitry can be a challenging task. To achieve high tunability in oscillators, the unloaded quality factor (Q-factor) of the resonator must remain high over a large bandwidth. A solution is to divide the bandwidth in several configurable narrow sub-bands, which for instance can be done by switching a capacitor bank. A drawback is that the switch elements usually degrade the performance. Tunable ferroelectric components can be used as demonstrated in [1], or yttrium-iron-garnet (YIG) resonators as in [2] to achieve broadband tuning. YIGs have drawbacks in terms of size and dc-power. Varactor diodes for tuning are reported in [3] and [4], but they have rather high loss and are nonlinear. Substrate integrated wave guides (SIWs) and cavity resonators with varactor diodes are demonstrated in [5] but often have rather moderate Q-factors.

In recent years micro machined technology, e.g. RF-MEMS, has shown good progress [6]. MEMS tuning can be separated into MEMS-varactors and MEMS-switches [7]. Mainly, the former has the advantage of continuous analogue tuning while the latter is restricted to discrete states. MEMS-varactors can also maintain high Q-factor over a large bandwidth [8]. SIW cavities designed for evanescent modes with MEMS-switches combined varactors have demonstrated good results with large tuning range and high unloaded-Q, 1.9 - 5.0 GHz with Q from 650-300 in [9], and 4.0-5.6 GHz with Q from 300-500 in [10], respectively. A similar evanescent mode substrate integrated resonator with RF-MEMS switches and varactors is also demonstrated in [11], at 6.25-6.60 GHz with Q 140-240.

A drawback though with MEMS-varactors is that they are sensitive to vibrations and significantly prone to electroacoustic resonance, known as microphony [12], [13]. In this perspective, digital MEMS-switches may be a better alternative as they can have well defined open and closed states and thus being less sensitive to microphony. In reality MEMS-varactors also often have discontinuities in their tuning characteristics [8], which further favors the choice of MEMS-switches. The frequency gaps between the digital switch states can be covered by a small tuning varactor in parallel with the switches. A low-phase noise X-band reconfigurable oscillator based on MEMS-switches was reported in [14].

This paper reports on analysis of reconfigurable cavity oscillators based on MEMS-switches. A thorough investigation of the MEMS-switches horizontal and vertical positions in the cavity is presented with focus on trade-offs between phase noise, tuning capability and frequency resolution. The analysis covers the oscillator reported in [14] as well as variants thereof.

The paper is outlined as follows. Section II presents essential theory for oscillator phase noise and its relation to cavity quality factor. Section III describes the oscillator implementation with cavity and PCB design aspects. Section IV discusses the MEMS technology, characterization and modelling of the MEMS-switches. In Section V the oscillator phase noise performance and tuning range are reported for Zirath are also employed at Ericsson AB, Lindholmspiren 11, SE-41756, Gothenburg, Sweden.

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different configurations. Measurements compared to simulations of the cavity with respect to tuning range and quality factor are discussed together with a comparison to other published results. Section VI discusses general mechanical behavior of the MEMS and some side effect as already mentioned, microphony and yield, as well as potential improvements. Finally the paper is concluded in Section VII.

II. THEORY

Leeson’s equation can be used to express oscillator phase noise as [15], [16],

\[
L(\Delta f) = 10 \log \left[ \frac{\Delta f}{2 \gamma} \left[ 1 + \left( \frac{f_0}{2 Q_0 \Delta f} \right)^2 \right] \left[ 1 + \frac{\Delta f_{1/f}}{\Delta f} \right] + \frac{K_2 V_m^2}{8 \Delta f^2} \right]
\] (1)

where \( k \) is Boltzman’s constant, \( T \) the temperature in Kelvin, \( P_r \) the RF-power dissipated in the resonator, and \( Q_c \) its loaded quality factor. \( F \) and \( \Delta f_{1/f} \) are fitting parameters for the noise level of the active device. The last term represents modulation noise, where \( k \) is the tuning sensitivity (Hz/V), and \( V_m \) the voltage noise on the varactor [16]. From (1) it is intuitively understood that maximizing \( P_r Q_c^2 \) gives the best phase noise near carrier. This motivates a high power capacity reflection amplifier in GaN with an external high quality-factor resonator, that is optimally coupled to the amplifier [17].

Resonant modes of a rectangular cavity are described by

\[
f_{mn} = \frac{c_0}{2\pi \sqrt{\mu_r \epsilon_r}} \sqrt{\left( \frac{m\pi}{a} \right)^2 + \left( \frac{n\pi}{b} \right)^2 + \left( \frac{l\pi}{d} \right)^2}
\] (2)

where \( a, b, \) and \( d \) are the width, depth and length of the cavity. The lowest resonant mode is TE\(_{101}\)-mode, setting \( a=d=\lambda/2 \), while \( b=\lambda/4 \) to assure a maximum margin to other resonant modes. Fig. 1 shows the E-field distribution following a half-sine wave in horizontal and vertical direction for the TE\(_{101}\)-mode.

Fig. 1. E-field distribution for the TE\(_{101}\)-mode in the cavity.

Frequency tuning can be achieved by changing the width or length of the cavity by introducing an electrically movable wall consisting of an array of RF-MEMS-switches. Fig. 2(a) and Fig. 2(b), respectively, show a cross section of the cavity with switches in open state and in closed state, thus short-circuiting the E-field.

The intrusion depth \( \Delta t \) in Fig. 2 affects the coupling in vertical direction and sets the tuning range. As the E-field follows a half sine wave distribution also in the horizontal direction for a TE\(_{101}\)-mode, the strongest coupling is at the middle position.

The frequency change for a small perturbation in shape can be estimated as [18],

\[
\frac{\omega - \omega_0}{\omega_0} = -\frac{2 \Delta V_b}{V_0}
\] (3)

where \( \Delta V_b \) is the volume occupied by the RF-MEMS-switches in the closed position. In the open state of the RF-MEMS, theRF-MEMS die itself, and its carrier substrate causes a dielectric change due to the volume of the substrate that will be exposed by the RF-field. This also gives an approximate additional frequency change in the open state as [18],

\[
\frac{\omega - \omega_0}{\omega_0} = -(\epsilon_r - 1) \Delta V_c
\] (4)

where \( \Delta V_c \) is the volume of the exposed RF-MEMS and carrier substrate and \( \epsilon_r \) its corresponding relative dielectric constant.

The total Q-factor when degrading the unloaded Q-factor of the cavity by intruding a substrate with a loss tangent tan\( \delta \), occupying a portion \( \eta \) of the total volume, can be estimated as

\[
\frac{1}{Q_{\text{cap}}} = \eta \tan \delta + (1-\eta) \frac{1}{Q_0}
\] (5)

where \( Q_0 \) is the Q-factor of the air filled cavity. Note that (5) gives an upper limit on the achievable Q-factor. In reality the PCB for the MEMS-switches includes also metal patterns that will introduce losses, and radiation due to RF-leakage.

III. MODULE DESIGN

A. MMIC technology and reflection amplifier

The amplifier is designed in Triquint’s 0.25-\( \mu \)m 3MI GaN HEMT process. The HEMT has a gate periphery of 8x50-\( \mu \)m. A nonlinear transistor model is extracted from large-signal S-parameter data [19]. Low frequency noise is also measured and implemented in the model [20], [21].

In the reflection amplifier the source is degenerated with a parallel LC-network used to peak its gain. The reflection gain appears at the gate while the drain is inductively terminated...
Fig. 3(a) and Fig. 3(b), respectively, show chip photo of the reflection amplifier and schematics of it. Fig. 4 shows the amplifier characterized on a PCB for different $I_d$ and $V_d$. The gain peaks at 10 GHz.

![Schematics](image)

**Fig. 3.** (a) Chip photo of the reflection amplifier. Size 1.0x1.0mm. (b) Schematics.

![Gain and Resistance](image)

**Fig. 4.** Measured gain and corresponding negative resistance with phase compensation for serial resonance in 50 ohm system of the reflection amplifier at different bias.

### B. Cavity and PCB

The cavity resonator is fabricated in two blocks of aluminum, one deep open waveguide and one shallow shortened piece, shown in Fig. 6. The two parts are clamped together with a low loss PCB, Rogers’ 5870 intruded between. The PCB is 4-layer, plated with silver and gold on top and bottom. MEMS-switches are mounted on the PCB. They are exposed to the RF-field inside the cavity area. A minimum amount of metal in this region is used for only short routing, which facilitates the RF-field to penetrate through the substrate and to be bounded by the whole cavity volume for the resonance condition. On the border of the two cavity pieces, there are rigid bars of through-vias on the PCB to keep the ground plane at the cavity walls intact. Bias lines for the MEMS-switches are routed in the mid metal layers, and shielded from the RF-feeding by top and bottom metallization to minimize the exposure of the RF-field to them [14]. The MEMS-die includes high impedance bias lines for further RF-choke.

The cavity is placed on top of an exciting microstrip line on a 4 mm wide, and 0.38 mm thick substrate, placed in a 1 mm deep trench in a brass plate. This plate also forms the ground plane at the bottom of the cavity [22]. The cavity wall is thinned where it is crossing the trench for minimizing unwanted coupling to the exciting microstrip line beneath the wall. A cross section sketch, the layout of the PCB and a schematic of the setup are shown in Fig. 5.

![Cross section sketch](image)

**Fig. 5.** Cross section sketch of the test fixture, a layout of the PCB, and a schematic of the setup. The dimensions $a$, $b$ and $d$ define the volume of the cavity.

The setup has flexibility to adjust the phase condition by sliding the cavity along the microstrip line. A perpendicular displacement can be used to optimize the coupling and loop gain. The reflection amplifier and the cavity are mounted on separate PCBs and carriers in order to allow individual characterization.

Fig. 6 shows the two cavity pieces. The lid has a ground collar around the opening to increase the pressure on the through-vias of the PCB which forms a well-defined ground connection.

![Two cavity pieces](image)

**Fig. 6.** Photo of the two cavity pieces. The size of the cavity opening is 20.7x10.35x20.7 mm. The depth of two variants of shortened pieces are 1 mm and 2.5 mm, respectively.

The resonator is designed to operate in the TE$_{101}$-mode, thus setting $a$, $b$, and $d$ for the width, depth and length of the cavity to $a=d=\lambda/2=20.7$ mm while $b=\lambda/4$ according to (2).

The substrate has a total height of 800 μm, tanδ=0.0017, and $\varepsilon_r=2.2$, and its thickness compared to the air-filled height of the cavity sets the ratio $\eta=0.8$ mm/20.7 mm=3.9 % of the volume. The air-filled cavity is simulated and measured to $Q=3800$. As predicted by (5) the Q-factor degrades slightly when a PCB with finite tanδ is intruded into the cavity, a Q-value of 3100 is then reached which is also confirmed by HFSS-simulations. The measured Q-factor of the cavity with an intruded PCB without MEMS is somewhat lower, 1700, which is due to the limiting conductivity and that the RF-leakage of the substrate vias in the intruded PCB degrades the performance. A simpler via setup was used in the simulator,
and imperfections in surface roughness of the PCB causing worse ground connection was not accurately modeled. For an assembled substrate, the losses in the mounted MEMS’s will further degrade the total Q-factor, which is measured to 500-1000 for all MEMS-states.

Different variants of the PCB, with different mounting alternatives as MEMS-switches mounted in one row, or in three rows are shown in the photos in Fig. 7(a) and Fig. 7(b), respectively. Fig. 7(c) shows the layout of the PCB for the MEMS-mounting.

The MEMS-switches are aligned with the E-field in the cavity. In the nonconductive state, i.e. open state, the RF-field can penetrate through the substrate to be bounded by the whole cavity volume. This makes the electric dimensions of the cavity larger, and the resonance frequency lower. Fig. 8(a) and (c) show E-field plots for the open state, for a one-row MEMS and a three-row MEMS setup, respectively. Fig. 8(b) and (d) show E-field-plots for the conductive state, i.e. closed state, for the one-row MEMS and the three-row MEMS. The MEMS-switches are grounding the E-field at this position, and the electric dimensions are shrunk. The tuning range is dependent on the distance from the cavity wall to the switching element, i.e. the intrusion depth, as well as the MEMS-switch position on the PCB itself. The E-field has the highest magnitude in the middle of the substrate, and a change of ground plane in this position gives more effect. Two MEMS-switches in series are used to improve the isolation in the open state. This is also beneficial due to dimensional reasons to avoid undesired metal for feeding the MEMS-switches. The total length of λc/4 is used across the width of the cavity wall, which meets the length of two MEMS’s.

Ideally several rows of MEMS-switches improve the tuning range, and they can be placed to have uniform resolution giving a broad cumulative tuning. In this way the loaded Q-factor can be maintained high over a broad tuning range as the MEMS-switches are found to have a good Q-factor.

Fig. 7. (a) Photo of the mounted MEMS in one-row configuration. (b) Photo of a three-row configuration. (c) Layout of the PCB for the mounted MEMS-switches intruded in the cavity.

Fig. 8. HFSS-simulation of the magnitude (color scale) and vector of the E-field at 2.5 mm depth. (a) One-row MEMS in open state, f=9.65 GHz. (b) One-row MEMS in closed state, f=10.32 GHz. (c) Three-row MEMS with all in open state, f=9.32 GHz. (d) Three-row MEMS with all in closed state, f=10.66 GHz.

C. Integration of oscillator test bench

Fig. 9 shows the assembled oscillator setup [14]. The cavity board is terminated on the opposite side of the connecting amplifier. This will avoid out-of-band resonance. This port is also beneficial to feed out RF-power, through an isolator to suppress load-pulling from the instruments. A flat-cable connects the cavity intruded MEMS-substrate to another board for biasing and digital control of the MEMS-states. Bias for gate and drain to the MMIC-amplifier are connected by coaxial cables to the power supply.

Fig. 9. Photo of the assembly of the MEMS-controlled cavity oscillator.

IV. MEMS TECHNOLOGY

The used switch is an ohmic cantilever RF MEMS-switch designed by RF Microtech, and fabricated at the FBK foundry [23]. They are built up of three switching elements in series and processed on low loss quartz substrates of 500 µm thick. The switching membrane is of gold, and has a size of 170 µm x 110 µm, which is suspended 3 µm above the interrupted line and anchored to the substrate on one side. It is pulled down by applying a voltage on the electrode. All bottom electrodes are connected to a common bias pad, and the cantilever anchor sides are connected to a dc-return to one of the RF-ports. This is shown in a chip photo in Fig. 10(a). Fig. 10(b) shows a sketch of a cross section of one stage switching element. The control voltage for the electrode is insensitive to polarity, and the applied dc-voltage to set the closed state is about 70 V. Without control voltage the cantilever springs to the open-position.
In this study the MEMS’s are characterized by a two-port small signal measurement using a VNA E8361A from Keysight. A simple model of the MEMS is shown in Fig. 11, where the extracted values of the parasitics in open state i.e., the switch element overlap capacitance $C_{par}$ and line inductance $L_{par}$ are determined to be 22 fF and 0.3 nH, respectively. In closed state, the line inductance $L_{sw}$ is slightly higher due to extended length through the switch region, and measures to about 0.4 nH. The chip photo Fig. 10(c) and Fig. 10(d) shows a minor overlap area for the ohmic contact region (about 20 μm), which assumes small variations of these parasitics due to state position. The switching resistance $R_{sw}$ is calculated to about 1 Ω in the closed state and infinity in the open state. A bond wire $L_{bond}$=1 nH is included in the model, and the parasitic capacitances for the actuator electrodes, $C_{par,act}$ and $C_{ipar,act}$ were estimated to about 10 fF.

![Fig. 11. Model of a three-stage MEMS.](image)

Fig. 10.(a) Chip photo of the MEMS. The size is 2970 x 1615 μm. (b) Sketch of the cross section of one stage switching element. (c) One switching element in open state. (d) The switching membrane has been bent up.

Fig. 12(a) shows the measured reflection coefficient and Fig. 12(b) shows the transmission coefficient for different applied electrode voltages, versus a simulation for closed and open state. It can be noticed that some intermediate states occur when the electrode voltage is swept, which turns out like undesired stubs with large phase change versus frequency seen in Fig. 12(a), or as reduced isolation in Fig. 12(b). This is due to the fact that all three switching elements on the die do not switch at the same voltage. With an actuator voltage lower than the MEMS pull-in voltage, the distance from the membrane to the disrupted line is stable and controlled by the actuator voltage until it collapses and will snap to the bottom position [24].

![Fig. 12. (a) Polar plot of measured and simulated reflection coefficient at different bias voltages, for frequency 6 GHz-11 GHz. (b) Measured and simulated magnitude of insertion loss at different bias voltages.](image)

where $A$ is the membrane area, $d$ the distance between the membrane and the actuator, $\varepsilon_0$ the dielectric constant, and $V$ the actuation voltage. The electric force $F_e$, counteracts a mechanic force $F_m$, described as

$$F_m = k\Delta z$$

where $k$ defines the spring constant of the membrane and $\Delta z$ the displacement towards the actuator. For small displacements, a stable condition is established for $F_m = F_e$, with characteristics of a varactor. The electric force, $F_e$ increases by larger displacement and there is a maximum displacement where instability makes the system collapse. The membrane will then snap to the bottom position. This appears when $\Delta z = d_0/3$ [7], where $d_0$ is the maximum gap in the unbiased state. Typical values for the RF-MEMS used in this study are $A=0.170 \times 0.111 \text{ mm}$, $d_0=3 \mu\text{m}$, and the required pull-in voltage is slightly above $V=20 V$. This evaluates to a spring constant of about $k=8 \text{ N/m}$.

V. RESULTS AND ANALYSIS

Fig. 13 presents the frequencies covered by all different configurations considered in this study. The number of used MEMS-switches, their positions on the PCB, the intrusion depths of the PCB in the cavity and the MEMS-states are varied. The phase noise measured with a signal source analyzer FSUP50 from Rhode & Schwarz is reported and analyzed for each setup in Section A and Section B. In Section C the cavity is analyzed standalone for different configurations by small-signal measurements versus simulations.
Fig. 13. Measured oscillation frequency for different configurations and activation of MEMS-switches.

A. Phase noise measurements of the oscillator with one-row MEMS tuning

1) 1 mm depth

The cavity with one row of MEMS-tuning at an intrusion depth of 1 mm in the cavity was placed at the optimum resonator coupling close to the center of the excitation microstrip line. The MEMS-switches were switched from open to closed state, measuring oscillation frequencies at 9.96 GHz and 10.17 GHz, respectively. Defining tuning range as $2(f_{max} - f_{min})/(f_{max} + f_{min})$, this corresponds to 2.1 %. Measured phase noise at 100 kHz offset frequency versus bias is shown for open state in Fig. 14(a) and for closed state in Fig. 14(b). An optimum phase noise of -140 dBc/Hz@100 kHz is achieved for both states, at a quite low bias level of $V_g=4.6$ V. The phase noise is almost invariant to variations in $V_g$, and more power in the resonator at a higher bias level also counterbalances the increased flicker noise at a higher drain current. Fig. 15(a) and (b) show the phase noise at 1 MHz offset for open and closed state, respectively, with optimum at slightly higher bias due to less impact of flicker noise for this offset.

Fig. 16(a) shows the phase noise versus offset frequency for all bias points in open state, and Fig. 16(b) shows the phase noise for all bias points in closed state. 

Fig. 16. Measured phase noise versus offset frequency for one-row MEMS at 1 mm depth for all bias states. (a) MEMS in open state. (b) MEMS in closed state. The $V_g/V_d$ is swept from -2/4 V to -1.4/12 V.

2) 2.5 mm depth

The cavity with one row of MEMS-tuning at an intrusion depth of 2.5 mm was measured. The measured oscillation frequencies of the two states are 9.51 GHz and 10.31 GHz, respectively, corresponding to a tuning range of 8.1 %. The phase noise results at 100 kHz for different bias are shown in Fig. 17(a-b), while Fig. 18(a-b) shows the phase noise at 1 MHz. Fig. 19(a-b) shows phase noise versus offset frequency for open and closed state, respectively.

Fig. 17. Measured phase noise at 100 kHz versus bias for one-row MEMS at 2.5 mm depth. (a) MEMS-state open. (b) MEMS-state closed.

Fig. 18. Measured phase noise at 1 MHz versus bias for one-row MEMS at 2.5 mm depth. (a) MEMS-state open. (b) MEMS-state closed.

Fig. 19. Measured phase noise versus offset frequency for one-row MEMS at 2.5 mm depth for all bias states. (a) MEMS-state open. (b) MEMS-state closed. The $V_g/V_d$ is swept from -2/4 V to -1.4/12 V.
Electroacoustic noise from the MEMS-membranes are seen for the open states slightly above 10 kHz offset for both alternative intrusion depths [13], [12]. At an intrusion depth of 2.5 mm shown in Fig. 19(a) the impact is worse due to the stronger MEMS-coupling in this case.

B. Phase noise measurements of the oscillator with three-row MEMS tuning

1) 1 mm depth

The cavity with a three-row MEMS-tuning at an intrusion depth of 1 mm was measured. In Fig. 20(a-d), the measured phase noise at 100 kHz at MEMS-state, 000, 001, 011 and 111 are shown. The notation is as ‘0’ equals open state, and ‘1’ equals closed state, and the bit-positions correspond to the physical positions of the rows for the MEMS. The corresponding oscillation frequencies are, 9.84 GHz, 10.07 GHz, 10.22 GHz and 10.33 GHz. This brings about a tuning range of 5 %, and the minimum phase noise ranges from -140 dBc/Hz to -129 dBc/Hz over the states. A worse phase noise performance is seen at the intermediate states, probably due to less uniform RF-field distribution between MEMS set in different states, and resulting in more RF-leakage. The optimum bias also changes due to phase changes with frequency. The bias level has to compensate for the phase condition, and the oscillations may not appear for the maximum loaded Q [25]. Phase tuning with the bias level might also be disadvantageous due to other conditions with respect to flicker noise, etc. Fig. 21(a-d) shows the phase noise for 1 MHz offset for all measured MEMS states.

Measured phase noise versus offset frequency for all bias points for the MEMS-states above are shown in Fig. 22(a-d).

Fig. 20. Measured phase noise at 100 kHz versus bias for three-row MEMS at 1 mm depth . (a) MEMS state 000. (b) MEMS-state 001. (c) MEMS-state 011. (d) MEMS-state 111.

Fig. 21 Measured phase noise at 1 MHz versus bias for three-row MEMS at 1 mm depth. (a) MEMS state 000. (b) MEMS-state 001. (c) MEMS-state 011. (d) MEMS-state 111.

Fig. 22. Measured phase noise versus offset frequency for a three-row MEMS at 1 mm depth for all bias states. (a) MEMS state 000. (b) MEMS-state 001. (c) MEMS-state 011. (d) MEMS-state 111. The $V_s/V_d$ is swept from -2/4 V to -1.4/12 V.

In Fig. 22(a-d) it is seen that the electroacoustic resonance around 10 kHz diminishes by the number of switches set in closed state. This further manifests that the open state with open positions of the cantilever membranes is more sensitive to vibrations.

2) 2.5 mm depth

In Fig. 23(a-c), the measured phase noise at 100 kHz offset at MEMS-state 000, 001 and 111 is shown. The corresponding oscillation frequencies are 9.42 GHz, 10.05 GHz and 10.65 GHz. This generates a tuning range of 12.3 %, and the optimum phase noise ranges from -133 dBc/Hz to -123 dBc/Hz over the states. The state 011 could not be measured due to insufficient loop gain. Also in this setup the optimum bias differs due to phase changes with frequency, and the bias level compensates for the phase condition. Fig. 24(a-c) shows the phase noise for 1 MHz offset for all measured MEMS-states.

Measured phase noise versus offset frequency for the MEMS-states are shown in Fig. 25(a-c).
Fig. 23. Measured phase noise at 100 kHz for three-row MEMS at 2.5 mm depth. (a) MEMS-state 000. (b) MEMS-state 001. (c) MEMS-state 111.

Fig. 24 Measured phase noise at 1 MHz for three-row MEMS at 2.5 mm depth. (a) MEMS-state 000. (b) MEMS-state 001. (c) MEMS-state 111.

Fig. 25. Measured phase noise versus offset frequency for three-row MEMS at 2.5 mm depth for all bias states. (a) MEMS-state 000. (b) MEMS-state 001. (c) MEMS-state 111. The $V_g/V_d$ is swept from -2/4 V to -1.4/12 V.

Similar to the previous setup of 1 mm intrusion depth, the electroacoustic behavior disappears when all MEMSs are in closed state, and the phase noise is worse at the intermediate states.

C. Measurements, simulations and analysis of the resonator

Measurements and HFSS-simulations of the resonator in a one-row configuration of MEMS at an intrusion depth of 1 mm and 2.5 mm, respectively, have been performed and the results are shown in Fig. 26(a-b). The reflection coefficients are higher in the simulations, especially at the 2.5 mm intrusion depth, probably due to more cavity loss, lower Q and less coupling in reality. For the stronger coupling at the deep position the capacitive parasitic is stronger, which shifts down the resonance peak for the open state compared to simulations.

Fig. 26. Measured versus simulated magnitude of the reflection coefficient of the resonator with one-row MEMS. Markers are set on the measured peak. (a) MEMS in open state (x00) and closed state (x11) at 1 mm depth, respectively. (b) MEMS in open state (x00) and closed state (x11) at 2.5 mm depth, respectively. Notation x=not mounted.

Measurements compared to simulations of a three-row configuration of MEMS at a depth of 1 mm and at a depth of 2.5 mm are shown in Fig. 27(a) and (b), respectively. As there are three rows of MEMS’s, they can be set in eight different states. Due to the symmetry of MEMS-positions from the layout picture shown in Fig. 7(c), some states coincide, and therefore only six frequencies are unique, of which four are presented in Fig. 27(a) and (b).
Fig. 27. Measured versus simulated magnitude of the reflection coefficient of the resonator with a three-row MEMS. Markers are set on the measured peaks. (a) At a depth of 1 mm. State notation (xxx), 0=open state, 1=close state. (b) At a depth of 2.5 mm.

The measured reflection coefficients are lower in the intermediate states, especially for the stronger MEMS-coupling at an intrusion depth of 2.5 mm. This is probably caused by the non-uniform RF-field around the MEMS’s set in different states. A mismatch in peak frequency for the lowest frequency at 2.5 mm depth is probably due to uncertain parasitic around the MEMS in the model.

Table I shows a summary of the measured and simulated Q-factor, resonance frequency and measured optimum phase noise for all evaluated MEMS-configurations.

<table>
<thead>
<tr>
<th>Setup</th>
<th>State</th>
<th>Measurement</th>
<th>Simulation</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>Qs</td>
<td>Freq</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>(GHz)</td>
</tr>
<tr>
<td>One-row MEMS @1 mm depth.</td>
<td>x0x</td>
<td>1115</td>
<td>9.96</td>
</tr>
<tr>
<td></td>
<td>x1</td>
<td>700</td>
<td>10.17</td>
</tr>
<tr>
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<td>9.51</td>
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<tr>
<td></td>
<td>x1</td>
<td>500</td>
<td>10.32</td>
</tr>
<tr>
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<td>001</td>
<td>700</td>
<td>10.07</td>
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<tr>
<td></td>
<td>01</td>
<td>700</td>
<td>10.22</td>
</tr>
<tr>
<td></td>
<td>111</td>
<td>1123(^{(*)})</td>
<td>10.33</td>
</tr>
<tr>
<td>Three-row MEMS @2.5 mm depth.</td>
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<td>9.42</td>
</tr>
<tr>
<td></td>
<td>001</td>
<td>250</td>
<td>10.05</td>
</tr>
<tr>
<td></td>
<td>01</td>
<td>350</td>
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</tr>
<tr>
<td></td>
<td>111</td>
<td>826(^{(*)})</td>
<td>10.65</td>
</tr>
<tr>
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<td></td>
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<td>9.998</td>
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<tr>
<td>Empty PCB @2.5 mm depth.</td>
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<td>9.925</td>
</tr>
<tr>
<td>Empty cavity [^{[22]}]</td>
<td></td>
<td>3800</td>
<td>9.93</td>
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</tbody>
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\(^{(*)}\)The MEMS-switches corresponding to these states have had lower \(R_s\) than modelled. \(^{(**)}\)Too low reflection to get oscillation with the used amplifier.

From Table I we can see that increasing the intrusion depth from 1 mm to 2.5 mm increases the tuning range by almost a factor of 4 for a one-row setup, and about a factor of 2.5 for a three-row setup. These ratios are almost the same for the differential increase per state comparing the two depths. The incremental increase saturates by the number of MEMS’s put in closed state, due to the disturbed RF-field around the switches. This complicates to achieve a uniform frequency resolution.

The measured phase noise is almost scaled by the Q-factor, and worse for the intermediate states. The disturbances of the RF-field for some intermediate states are simulated in Fig. 28(a) for the state 010, and in Fig. 28(b) for the state 101. In Fig. 28(b) the single open MEMS in the middle is not enough for the RF-field to completely penetrate through the substrate, which makes the field to be similar to the case in Fig. 28(c) with all MEMS’s in closed state. This explains the compression of the tuning range when several MEMS’s are set in the closed state.

Fig. 28. HFSS-simulations for intermediate states for a three-row MEMS setup at 2.5 mm depth compared to all in closed state. (a) State 010, \(f_0=10.18\) GHz. (b) State 101, \(f_0=10.51\) GHz. (c) State 111, \(f_0=10.66\) GHz.

The impact of an inhomogeneous field around the switches for intermediate states also has a negative impact on the Q-factor. The states performing best Q-factor are those set to the extreme conditions, i.e. with all RF-MEMS’s switched to open state or to closed state.

To further verify the building practice, a measurement compared to a simulation of the reflection coefficient with an empty substrate at an intrusion depth of 1 mm and 2.5 mm are shown in Fig. 29. As reference a trace for the cavity with no PCB is added. In the latter case the cavity shrinks by the height of the PCB and will have a lower dielectric constant causing a higher resonance frequency.

Fig. 29. Measured and simulated reflection coefficient of a cavity with empty PCBs (no switches mounted) placed at 1 mm and 2.5 mm depth, respectively. The reflection coefficient of a completely empty cavity, i.e., without PCB, is also shown.
The simulated E-field of an empty substrate at 1 mm depth and at 2.5 mm depth is shown in Fig. 30(a) and (b) respectively, showing a uniform field distribution, as compared to the cavity simulation with no PCB shown in Fig. 30(c).

Fig. 30. HFSS simulation of the E-field of an empty substrate. (a) At an intrusion depth of 1.0 mm. (b) At an intrusion depth of 2.5 mm. (c) No PCB.

Table II shows a comparison to other tunable high-Q oscillators. Some designs have better phase noise near carrier, while others, for example YIGs have better tuning range. This work reports the best trade-off between phase noise and tuning range, and an excellent tuning figure of merit, FOM\(_{\text{m}}\). The noise far out at 1 MHz offset is also considerably lower for our work due to the advantage of GaN.

### Table II

Comparison to other tunable high-Q oscillators

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<th>fc (GHz)</th>
<th>Tuning ratio (TR) (%)</th>
<th>Q(_{\text{L}})</th>
<th>FOM(_{\text{m}}) @ 100 kHz</th>
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<td>500</td>
<td>196</td>
<td>-198</td>
<td>MEMS @ 2.5mm GaN</td>
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</tbody>
</table>

\(\text{FOM}_{\text{m}} = \text{FOM} - 20\log(\Delta f) - 10\log(P_{\text{in}}/1\text{mW}) - 20\log(\text{TR}/10)\) or \(\text{FOM}_{\text{m}} = \text{FOM} - 20\log(\text{TR}/10)\)

The electroacoustic resonance frequencies can be estimated as [7],

\[\omega_0 = \sqrt{k/m_{\text{modal}}}\]  

where \(k\) is the membrane spring constant, \(m_{\text{modal}}\), which can be approximated as 0.35 to 0.45 times the static mass, \(m_0\), of the membrane to average the mass in motion for a cantilever MEMS at the fundamental mode [7]. Inserting numerical values, the mechanical resonance frequency in this study can be estimated. The static mass may be estimated as \(m_0 = \rho A t\), where \(\rho = 19300 \text{ kg/m}^3\) (gold), membrane thickness \(t = 3 \mu\text{m}\), and \(A = 0.170 \times 0.111 \text{ mm}\). The mass then evaluates to \(m_0 = 1.0 \cdot 10^{-9} \text{ kg}\). Using the previously calculated spring constant, \(k = 8 \text{ N/m}\) from (6), (7) and the modal mass to 0.45 times the static mass in (8) leads to a resonance frequency of \(\omega_0 = 133 \cdot 10^3 \text{ rad/s}\), or 21 kHz. This corresponds fairly well to the experimentally detected spuriouses occurring slightly above 10 kHz.

In the measured phase noise spectra, e.g., Fig. 25(b), it is also seen that the phase noise above the electroacoustic spuriouses rolls off with approximately 60 dB as predicted by the theory of the mechanical transfer function for a membrane stimulated by an electrostatic force [12]. It is shown to have no filtering impact (0 dB) below the mechanical resonance frequency and a 40 dB/decade roll-off above. The peak at the resonance frequency depends on the damping factor or Q-factor of the mechanical system. In an oscillator, the mechanical frequency response, will further shape the phase noise response versus frequency on the up-converted noise as reported in [12].

In summary, electroacoustic resonances are indeed an issue. For MEMS-switches, as in this case, the problem may be addressed by modulating the switch so that the membrane is fixed to a metal pad also in the open state. In a MEMS-varactor, on the other hand, that would not be possible.

In the presence of electroacoustic resonances, the oscillator may still be useful if the resonances occur at frequencies low enough to be suppressed by the loop filter of the phase locked loop (PLL).

In addition to the electroacoustic resonances another issue in the measurements was that the used ohmic cantilever RF-MEMS-switches, in the closed state, could change on-resistance, or even worse, stick to the metal pad in the closed state after several cycles, which affect the lifetime [31]. An alternative solution is to use nonconductive capacitive RF-MEMS-switches [32].

### VII. CONCLUSIONS

Analysis of a low phase noise reconfigurable X-band cavity oscillator based on ohmic cantilever RF-MEMS-switches has been reported. The MEMS’s have been mounted on a PCB, intruded in the cavity. The number of MEMS’s, their horizontal position on the PCB, and the vertical position of the PCB itself inside the cavity have been investigated with respect to phase noise and tuning range.

In conclusion, it is demonstrated that good phase noise, better than -133 to -123 dBc/Hz, can be achieved with fairly wide tuning range, 12%. The tuning range can be increased, alternatively, by adding more switches or by placing the PCB deeper into the cavity. Either way it will degrade the phase noise, but to a much smaller extent compared to if varactors were used as the only tuning element.
In addition, it is found that MEMS that are free to vibrate in the open state will cause electroacoustic resonances that may result in increased phase noise. Electroacoustic-resonances seen in this work can be modeled with a simple electromechanical model; the frequencies are sufficiently low to be suppressed inside the bandwidth of a PLL.

Future work includes integration of a weakly connected varactor on the PCB to enable PLL-locking.

ACKNOWLEDGMENT

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REFERENCES


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Paper [B]
RF-MEMS Tuned GaN HEMT based Cavity Oscillator for X-band

M. Hörberg, T. Emanuelsson, P. Ligander, S. Lai, H. Zirath, D. Kuylenstierna

RF-MEMS Tuned GaN HEMT based Cavity Oscillator for X-band

Mikael Hörberg, Thomas Emanuelsson, Per Ligander, SzhaLai, Herbert Zirath, Fellow, IEEE, and Dan Kylenstierna

Abstract—This letter presents a radio frequency micro-electromechanical systems (RF-MEMS) tuned cavity oscillator for X-band. The active part of the oscillator is implemented in GaN-HEMT MMIC technology. The RF-MEMS-switches are realized on a quartz substrate that is surface mounted on a low loss PCB. The PCB is intruded in an aluminum cavity acting as an electrically moveable wall. For a three-row RF-MEMS setup, a tuning range of 5 % around an oscillation frequency of 10 GHz is demonstrated in measurements. The phase noise is as low as $-140$ dBC/Hz to $-129$ dBC/Hz at 100 kHz from the carrier, depending on the configuration of the RF-MEMS.

Index Terms—Cavity, GaN HEMT, oscillator, phase noise, radio frequency microelectromechanical systems (RF-MEMS).

I. INTRODUCTION

UNABILITY and reconfigurability in frequency are crucial in communication systems. Modern radio equipment is often generic products where software is used for sub-band selection. This sets requirements on the hardware in terms of flexibility, beside constraints in terms of low weight, volume, and low cost. For oscillators it is a particular challenge to achieve wide tuning range with maintained phase noise performance. Low phase noise oscillators are required for more advanced modulation schemes and higher data rates. The near-carrier phase noise is determined by the unloaded quality (Q) factor, and the power coupled to the resonator [1].

Examples of available techniques for electric reconfigurations are switchable cap-banks or varactors. These elements are usually degrading the unloaded Q-factor of the resonator. Solutions with varactor diodes for switching is reported in [2], but has a disadvantage of high loss and linearity problems at microwave frequencies. Substrate integrated waveguide (SIW) cavity resonators with integrated varactor diodes have been demonstrated in [3] with a tuning range of 10-15 % but with moderate unloaded Q-factor around 100-200.

Today, micro machined technology, e.g., RF-MEMS, is of extensive interest. In [4] an RF-MEMS varactor was used to tune the electric field inside a ridged cavity about 10 % around 5 GHz, with a measured unloaded Q-factor of 500-800. Similar results for a 13 GHz cavity with RF-MEMS varactors are reported in [5]. In [6], a capacitive RF-MEMS shows very good tunability around 5 GHz with unloaded Q-factor of 300 to 500.

This work demonstrates a low-phase noise X-band oscillator tuned by ohmic cantilever RF-MEMSs integrated in a metal cavity. The ohmic cantilever RF-MEMS is a good choice due to high power capability and relatively low loss. The same type of MEMSs have previously been used in high-Q tunable waveguide filters [7]. The oscillator reported in this letter is a further development of the state-of-the art cavity oscillator reported in [8]. This work demonstrates that tuning functionality can be added without severely degrading the phase noise.

II. BUILDING PRACTICE AND MEMS INTEGRATION

The ohmic cantilever RF-MEMS-switches are processed at FBK (Fondazione Bruno Kessler). Key parameters for this type of RF-MEMS are high power capability, high linearity, and relatively low loss. They are processed on a low loss quartz substrate and have electrostatic actuators to be biased at 0 V in OFF-state and 70 V in ON-state, with cantilever in up-position and in down-position, respectively.

The dies are mounted on a low loss Rogers’ 5870 substrate, intruded in an aluminum cavity. In this way an electrically controlled ground plane is created for one of the cavity walls, which will set the resonance frequency.

The cavity is fabricated of two blocks, one deep and one shallow part. The two blocks are clamped together with screws and with the PCB placed in between. The PCB is 4-layer, plated with silver and gold on top and bottom. The total height of the PCB is 800 $\mu$m, the relative permittivity $\varepsilon_r = 2.2$ and the loss tangent is $\tan \delta = 0.0017$. The RF-MEMS-switches are mounted in rows on the PCB. Every row has two MEMS-dies with three switch elements each in serial to achieve good isolation in the OFF-state. The RF-field will penetrate through the substrate in the OFF-state, and be grounded in the ON-state. There are large amounts of through-vias on the PCB to keep the ground plane at the cavity walls intact.

The cavity is placed on top of a 1.1 mm wide exciting microstrip line on a 0.38 mm thick substrate, placed in a 1 mm deep and 4 mm wide trench in a brass plate, as was presented in [8]. This plate also makes the ground plane at the bottom of the cavity. The cavity wall is thinned where it is crossing the microstrip trench for minimizing unwanted coupling to the exciting microstrip line beneath the wall. A CAD drawing, schematic and layout of the PCB are shown in Fig. 1.
The cavity is designed to resonate in the TE_{101}-mode. The dimensions, as referred in Fig. 1 are \( a = d = \frac{\lambda_g}{2} = 20.7 \text{ mm} \) while \( b = \frac{\lambda_g}{4} \) to assure a maximum margin to other resonant modes. The resonant frequency of about 10 GHz is calculated by

\[
f_{mn1} = \frac{c_0}{2\pi \sqrt{\mu_r \varepsilon_r}} \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2 + \left(\frac{l\pi}{d}\right)^2}
\]

The two cavity parts have a ground collar around the opening, to increase the pressure on the PCB to avoid leakage and make a well-defined ground connection. The parts are shown in Fig. 2.

The intrusion depth of the PCB into the cavity and the position of the MEMS-switches are co-optimized for both good tuning and phase noise performance. PCBs with different mounting alternatives for one-row and three-row configurations are shown in Fig. 3 (a,b). Fig. 4 (a) and Fig. 4 (b) show a chip photo and a schematic cross section, respectively, of the RF MEMS.

The setup has flexibility to adjust the phase condition by sliding the cavity along the microstrip line, and perpendicular to the line for optimization of the resonator coupling. The RF-MEMSs on the intruded PCB shift the resonance frequency depending on the switch states.

### III. Measurements and Characterization of Cavity

The cavity with integrated RF-MEMS switches has been characterized by two port S-parameter measurements for different MEMS-states, denoted by “0” for OFF, and “1” for ON. Fig. 5(a) and Fig. 5(b) show measured reflection coefficients compared to simulations for MEMS cavities with one and three rows, respectively. The peaks correspond to the used serial resonances.

TABLE I summarizes the unloaded Q-factor, resonance frequency and tuning range for the studied setups.

### IV. Oscillator Performance

The cavity with integrated RF-MEMSs has been connected to the MMIC reflection amplifier to form the oscillator. The phase noise of the assembled oscillator is measured by an FSUP50 signal source analyzer from Rhode & Schwarz. Fig. 6 (a-d) show the result for the three-row-setup at 100 kHz for four different MEMS-states versus different \( V_{gs}/V_{d} \) amplifier bias. The corresponding oscillation frequencies are...
9.84 GHz, 10.08 GHz, 10.22 GHz and 10.33 GHz. The minimum phase noise at 100 kHz ranges from −9.84 GHz, 10.08 GHz, 10.22 GHz and 10.33 GHz. The Vg is swept from −2 to −1.4 V and Vd from 4 to 12 V. Best and worst traces for phase noise at 100 kHz are marked and noted with bias level.

Table II

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<td>-194 microstrip, SiGe</td>
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Table II

Comparison to Other Tunable High-Q Oscillator

The minimum phase noise at 100 kHz ranges from −140 dBc/Hz to −129 dBc/Hz, depending on configuration of the MEMSs and variations in on-resistance of 1 to 3 ohm. The shift in optimum bias is due to a variation of the phase condition as the MEMS-states changed, which also affects the performance.

Fig. 7 shows phase noise versus offset frequency for the four MEMS-states with gate and drain bias swept. A plausible explanation to the phase noise peak about 15 kHz is electroacoustic resonances of the gold membranes in OFF-state of the RF-MEMSs, [9].

Table II compares this work to other state-of-the-art tunable oscillators. Some designs have better phase noise, while other (YIGs) have better tuning range. In summary, this work reports a trade-off between phase noise and tuning range, and thus an excellent tuning figure of merit.

V. Conclusion

A cavity oscillator with integrated RF-MEMS-switches acting as an electrically moveable wall inside the cavity has been demonstrated. Depending on configuration of the MEMSs, the oscillator presents a phase noise in the range −140 to −129 dBc/Hz at 100 kHz offset from the carrier frequency that is tuned from 9.84 GHz to 10.33 GHz. To the authors’ best knowledge this is the best phase noise reported for a tunable oscillator at X-band. The significantly reduced phase noise enables more advanced modulation schemes and higher data rate for future radio communication products. A future work is to integrate a weakly coupled varactor for fine analogue tuning and PLL-locking. Techniques for suppression of electroacoustic resonances are also to be investigated, for example a PLL-loop with sufficient loop bandwidth may suppress the resonance.

References

Paper [C]
Phase Noise Analysis of an X-Band Ultra-low Phase Noise GaN HEMT based Cavity Oscillator

M. Hörberg, T. Emanuelsson, S. Lai, T. N. T. Do, H. Zirath, D. Kuylensierna

Phase-Noise Analysis of an X-Band Ultra-Low Phase-Noise GaN HEMT Based Cavity Oscillator

Mikael Hörberg, Thomas Emanuelsson, Szhau Lai, Graduate Student Member, IEEE, Thi Ngoc Do Thanh, Herbert Zirath, Fellow, IEEE, and Dan Kuylenstierna, Member, IEEE

Abstract—This paper reports on an ultra-low phase-noise oscillator based on a GaN HEMT monolithic microwave integrated circuit reflection amplifier and an aluminum cavity resonator. It is experimentally investigated how the oscillator’s phase noise depends on the cavity coupling factor, phase matching, and bias condition of the reflection amplifier. For the optimum bias and cavity position phase noise of $-145$ dBc/Hz and $-160$ dBc/Hz at offsets of 100 and 400 kHz, respectively, from a 9.9-GHz carrier frequency is reached. This is, to the best of the authors’ knowledge, a record in reported performance for any oscillator based on a GaN HEMT device. The optimum performance at 400-kHz offset corresponds to a power normalized figure of merit of 227 and compensating for finite efficiency in the reflection amplifier, the achieved result is within 7 dB from the theoretical noise floor, assuming a linear theory.

Index Terms—Phase noise, oscillator, cavity, GaN HEMT.

I. INTRODUCTION

LOW PHASE noise is required in a communication system to handle advanced modulation schemes such as higher order quadrature amplitude modulation (QAM) and orthogonal frequency division multiplexing (OFDM). In modern systems with high data rates, far-carrier phase noise is of increased importance [1]. Far-carrier phase-noise performance, e.g., white phase noise, is determined primarily by the noise figure of the active device and the power swing over the resonator [2], [3]. In this perspective, GaN HEMT technology, with its high breakdown voltage, is a good choice for the active device. Beside high breakdown voltage, the GaN HEMT has better noise figure compared to bipolar technologies such as a SiGe HBT and an InGaP HBT, and thus better capability for good far-carrier phase-noise performance.

A few GaN HEMT oscillators have been reported in the open literature and some of them show promising far-carrier phase-noise performance [4]–[6]. However, near-carrier phase noise is not as good as required in many application due to significant flicker noise. To meet the near-carrier phase-noise specifications, despite the higher flicker noise, a GaN HEMT oscillator based on a high-$Q$ cavity resonator has been designed and is reported in this paper.

The designed oscillator is based on a GaN HEMT monolithic microwave integrated circuit (MMIC) reflection amplifier and a rectangular aluminum cavity. It oscillates at 9.9 GHz with an excellent phase noise of $-145$ dBc/Hz at 100 kHz and $-160$ dBc/Hz at 400 kHz. To the best of the authors’ knowledge this is by far the best performance reported for any oscillator based on a GaN HEMT device. It is also, regardless of active device technology, better than many microwave oscillators based on high-$Q$ resonators, e.g., metal cavity or dielectric resonators.

The performance of the cavity-based GaN HEMT oscillator is also benchmarked versus the theoretical noise floor according to Everard’s theory [7]. The circuit is designed with flexibility to adjust the coupling factor between the cavity and the amplifier, by shifting the offset between the cavity and a microstrip line coupling to the amplifier. For optimum coupling factor and bias condition, the measured phase noise is within 7 dB from the theoretical noise floor, after compensation for finite efficiency of the amplifier.

In the theoretical part of this paper, the theoretical noise floor according to Everard’s theory [7] is used to express a relation between power normalized phase-noise figure of merit (FOM) [8] and the unloaded quality factor ($Q_0$) of the resonator. As previously reported in [9], maximum achievable FOM depends only on $Q_c$ according to a linear theory. However, in contrast to [9], this work takes into account the finite coupling factor between the resonator and active device, which gives an about 2 dB harder bound. This affects the effective noise figure as proposed in [9].

II. THEORY AND BACKGROUND

Oscillator phase noise at offset frequency $f_m$ from an oscillation frequency $f_0$ can be quantitatively modeled by Leeson’s equation [2]

$$L(f_m) = 10\log \left[ \frac{FkT}{2P_s} \left[ 1 + \left( \frac{f_0}{2Q_L f_m} \right)^2 \right] \left[ 1 + \frac{f_{1/f_2} f_m}{f_m} \right] \right]$$

where $k$ is the Boltzmann’s constant, $T$ is the temperature in Kelvin, $P_s$ is the RF-power dissipated in the resonator, and $Q_L$ is its loaded quality factor. $F$ and $f_{1/f_2}$ are fitting parameters used for fitting the noise level depending on the active device noise as well as nonlinear noise conversion [10]. Accurate prediction of the nonlinear noise conversion requires cyclostationary noise calculations [6], [10]–[13].
From (1), it can be intuitively understood that minimization of phase noise requires maximization of the product $P_s Q_f^2$, which is about finding the optimum coupling, $\beta$, between the resonator and the reflection amplifier. If the coupling approaches zero, $Q_L$ approaches $Q_0$, but the power $P_s$ simultaneously approaches zero. Contrary if the coupling becomes too strong, $Q_L$ drops. Thus, there must be an optimal coupling somewhere between.

The purpose here is not to present another method for cyclo-stationary noise calculation. Instead, the purpose is to discuss the bound on the minimum phase noise that can be reached for an oscillator with given restrictions on unloaded quality factor $Q_0$ and $\eta$, and to relate this to the power normalized FOM, as defined by Wagemans et al. in [8]. Similar ideas were presented in [9] and [14]. However, in those works the optimal coupling factor between the resonator and the active device was not considered. It is unrealistic to maintain $Q_L = Q_0$ while maintaining 100% conversion efficiency, i.e., $P_{\text{sig}} = P_{\text{DC}}$. A better bound on $L(f_m)$ may be derived if the optimal coupling between the active device and the resonator is considered [7].

Fig. 1 shows a schematic of a negative resistance-type oscillator based on a series resonator and a reflection amplifier modeled as a negative resistance and a noise source. Fig. 1 may also be seen as a high-efficiency feedback oscillator with zero output impedance [15].

Ignoring the transposed flicker noise and the noise floor region, Everard et al. [7] has shown that the single-sideband (SSB) phase noise spectrum of a high-efficiency oscillator can be expressed in terms of the ratio between $Q_L$ and $Q_0$.

$$\mathcal{L}(f_m) = \frac{FkT}{8Q_0^2 (Q_L/Q_0)^2 (1 - Q_L/Q_0) P_{\text{RF}}} \left( \frac{f_0}{f_m} \right)^2$$

(2)

where $P_{\text{RF}}$ is the total RF power inside the oscillator, i.e., dc power $\times$ efficiency. $Q_0$ and $Q_L$ may be related through the coupling coefficient $\beta$,

$$Q_L = \frac{Q_0}{1 + \beta}.$$  

(3)

Inserting (3) into (2), the phase noise can be expressed in terms of $\beta$ as

$$\mathcal{L}(f_m) = \frac{FkT}{8Q_0^2 \beta/(1 + \beta)^3 P_{\text{RF}}} \left( \frac{f_0}{f_m} \right)^2.$$  

(4)

Equation (4) can then be differentiated to show that the minimum phase noise occurs for an optimum coupling of $\beta = 1/2$, which inserted back into (4) yields a minimum $1/f^2$ phase noise of

$$\mathcal{L}(f_m) = \frac{27FkT}{32\eta P_{\text{DC}}Q_0^2} \left( \frac{f_0}{f_m} \right)^2$$

(5)

where $\eta$ is the oscillator efficiency, $P_{\text{DC}}$ is the dc power, and $F$ is the transistor noise factor. Assuming $\eta = 100\%$ and $F = 1$, the minimum achievable SSB phase noise expressed in dBc/Hz becomes

$$\mathcal{L}_{\text{min}}(f_m) = -174.6 - 10\log \left( \frac{27FkT}{32Q_0^2 P_{\text{DC}}} \right) + 20\log \left( \frac{f_0}{f_m} \right)$$

$$- 20\log(Q_0) + 20\log \left( \frac{f_0}{f_m} \right)$$

(6)

with $P_{\text{DC}}$ expressed in mW. Equation (6) presents the minimum phase noise that may be achieved if a resonator with unloaded quality factor $Q_0$ is optimally coupled to an active device fed with dc power $P_{\text{DC}}$. In reality, the measured phase noise is generally higher due to deficiencies in design and technology, e.g., finite $\eta < 1$, noise from the active device $F > 1$, and nonoptimum coupling $\beta \neq 1/2$, as well as nonlinear conversion. In many cases it is difficult to spot exactly what is the reason why $L_{\text{min}}(f_m)$ has not been reached. However, the difference between measured SSB phase noise $L_{\text{meas}}(f_m)$ and $L_{\text{min}}(f_m)$ can be used as a FOM for the oscillator, i.e., design efficiency as proposed in [14]. In this work, we call the difference the effective noise figure ($F_{\text{eff}}$) that can be expressed as follows:

$$F_{\text{eff}} = L_{\text{meas}}(f_m) - 20\log \left( \frac{f_0}{f_m} \right) + 174.6$$

$$+ 10\log \left( \frac{f_0}{f_m} \right) - 20\log(Q_0)$$

$$= 174.6 + 20\log(Q_0) - \text{FOM}$$

(7)

where, in the last step, the FOM [8] has been used to simplify the equation.

It should be pointed out that $F_{\text{eff}}$, as defined in (7), is the effective noise figure compared to an optimally coupled resonator with 100% efficiency, i.e., according to (6). If $\eta$ and $Q_0$ are known, the effect of these parameters may be inserted into (7) to calculate an operating noise figure ($F_{\text{op}}$) including the active device noise and its nonlinear conversion in the oscillator,

$$F_{\text{op}} = 175.9 + 20\log(Q_0) - \text{FOM} + 10\log \left( \frac{8\eta \beta}{(1 + \beta)^3} \right)$$

(8)

If the oscillator operates fairly linear, $F_{\text{op}}$ will be similar to the active device microwave noise figure, while if there is strong nonlinear noise conversion, it will be higher (compare Hajimiri’s impulse sensitivity theory [10]).

Before ending this discussion about how to benchmark oscillators with FOMs and relative to absolute noise floor, it is worth pointing out that the conventionally used FOM [8] does not relate oscillator phase noise directly to the absolute noise floor, and furthermore, it does not normalize versus $Q_0$ of the oscillator.
resonator. In fact a high-Q oscillator may have comparatively good FOM without being particularly well designed, i.e., not well utilized Q factor.

Setting $P_{\text{eff}} = 0$ dB in (7), it is found that the maximum achievable FOM turns out to be bound only by $Q_0$ of the resonator, i.e.,

$$FOM_{\text{Max}} = 174.6 + 20\log(Q_0).$$

Comparing (9) to [9, eq. (6)], there is about a 2-dB difference in the numerical factor, which originates in the fact that [9] does not take into account the coupling between the active device and the resonator.

### III. DESIGN AND EXPERIMENTS

#### A. MMIC Technology and Device

The amplifier is manufactured in a 0.25-μm GaN HEMT process, TriQuint 3MI process. The model used for design is Chalmers’ own model [16] extracted from in-house measurements.

To compare measured and simulated results, the low-frequency (LF) noise model parameters of the transistor have been extracted from measurements, [17]. Fig. 2(a) shows the measured LF noise versus frequency for different drain currents at fixed $V_d$, and Fig. 2(b) shows the measured LF noise at 100-kHz offset for different $V_d/I_d$ setting. Simulated and measured $I/V$ from a transistor on the same batch as the reflection amplifier is shown in Fig. 3(a) and (b). LF noise and dc measurements are carried out on a transistor in a common source configuration. In the reflection amplifier, a 15-Ω resistance is connected between the source and ground, causing a difference in the $I/V$ characteristics (cf. Figs. 3–7).

The noise current density is modeled with the standard flicker noise model

$$<i_n^2> = K_f I_d f_{f_{\text{ref}}},$$

where the involved parameters extracted from measurements shown in Fig. 2 are $K_f = 6.4 \times 10^{-10}$, $A_{f_{\text{ref}}} = 0.87$, $f_{f_{\text{ref}}} = 1.38$. $I_d$ is the drain bias current and $f$ is the frequency.

#### B. Reflection Amplifier

To simplify the building practice and minimize the number of interconnects associated with model uncertainties, the reflection amplifier is integrated on a MMIC. Fig. 4 shows the MMIC reflection amplifier based on a GaN HEMT with 400-μm gate
periphery. To assure oscillation condition, the GaN HEMT has an inductive termination at the drain, together with a cap for dc blocking. This inductor is used to set the reflection gain. A parallel $L C$ network is connected to the source, to peak gain at about $10 - 12$ GHz. A $15-\Omega$ resistor is connected in series with the source inductance, mostly used for bias stabilization over temperature and process variations. The source termination and the drain termination are optimized together to set the desired instability and reflection gain seen at the gate.

Fig. 5 shows the gain magnitude versus frequency for the reflection amplifier shown in Fig. 4. The gain is shaped to peak in the desired resonance band, and roll off outside to suppress undesired out-of-band oscillations. Fig. 6(a) and (b) shows the gain and phase of the amplifier, respectively. At frequency 10 GHz, the measured reflection gain at $V_d = 5$ V and $I_d = 40$ mA, $|\Gamma_{\text{amp}}| = -8$ dB, corresponding to a $R_N = -21.6 \, \Omega$, when the imaginary part is tuned out.

In Fig. 6(a), higher gain forces the oscillator to run deeper in compression. There might be a risk that the phase is changed by bias, and the oscillation frequency will not appear where $Q$ is highest. Fig. 6(b) shows that is a minor effect, and can be neglected.

Bias points for the amplifier compared to simulated results are shown in Fig. 7. The small source resistance in the bias network contributes to additional slope in the $I V$ characteristics.

C. Cavity

The metal cavity is made of aluminum suitable for industrial implementation. It is designed to resonate at 9.9 GHz in the $\mathrm{TE}_{101}$ mode. The frequencies for different modes are given by

\begin{equation}
 f_{m\ell n} = \frac{c_0}{2\pi \sqrt{\mu_r \epsilon_r}} \sqrt{ \left( \frac{m\pi}{a} \right)^2 + \left( \frac{n\pi}{b} \right)^2 + \left( \frac{l\pi}{d} \right)^2 } \tag{11} \end{equation}

where $a$, $b$, and $d$ are the width, depth, and length of the cavity. For the $\mathrm{TE}_{101}$ mode this means that $a = d = \lambda_g/2 = 21.7$ mm while $b = \lambda_g/4$ to assure maximum margin to other resonant modes. For practical reasons, the corners of the cavity are rounded, which affects the resonant frequency slightly, but has no negative effect on the $Q$ factor. Fig. 8(b) shows an HFSS setup of the metal cavity coupled to a microstrip line. The coupling to the resonator is accomplished by a microstrip line recessed in a brass plate acting as a ground wall for the aluminum cavity. The end of the microstrip line is terminated in 50 $\Omega$, which effectively suppresses unwanted resonances.

The coupling factor can be controlled by varying the perpendicular offset position between the microstrip line and the center of the cavity, changing the offset from 0 to 9 mm, $R_s$ changes...
HÖRBERG et al.: PHASE-NOISE ANALYSIS OF X-BAND ULTRA-LOW PHASE-NOISE GaN HEMT BASED CAVITY OSCILLATOR

Fig. 7. Bias points for the amplifier, measured (blue dots in online version) compared to simulated (red dots in online version). (a) $I_d$ versus $V_d$. (b) $I_d$ versus $V_g$.

from about 7 to 25 $\Omega$. The highest possible $R_s$ for a weakly coupled cavity is determined by the load termination at the end of the microstrip line.

The small-signal reflection coefficient ($\Gamma'$) of the designed cavity was measured with an E8361A 67-GHz vector network analyzer (VNA) from Agilent Technologies. Fig. 9(a) and (b) shows the measured and simulated $\Gamma$ from the cavity board at different cavity positions.

The measured $\Gamma$ was converted to input impedance, and $Q_0$ of the cavity was calculated from the phase slope of the input impedance

$$Q_0 = \frac{\omega_0}{2} \frac{\partial \phi(Z_{in})}{\partial \omega} \omega = \omega_0,$$

The measurements show that $Q_0$ is marginally affected by the offset position while the impedance level is strongly affected. The measured $Q_0 \approx 3800$, and when the cavity is positioned for maximum coupling (microstrip line exciting in the middle), measured $\Gamma_{tank} = -2.6$ dB, corresponding to $R_s = 7.5$ $\Omega$.

D. Oscillator Test Bench

After characterization of the individual building blocks, i.e., the reflection amplifier and the cavity, the two parts are merged together to form a complete oscillator. The amplifier is soldered on a brass plate, with a printed circuit board (PCB) for biasing.
This plate is attached to the cavity plate. A clamp presses the cavity in place [see Fig. 10(a)]. The output power is extracted through an SMA connector at the end of the microstrip line on opposite side of the cavity. A 3-dB attenuator, connected between the microstrip line and the source analyzer, ensures a good 50-Ω termination suppressing potential spurious resonances.

Fig. 11 shows open loop gain, $G_{\text{loop}} = \Gamma_{\text{amp}} \Gamma_{\text{tank}}$, derived from measurements of the reflection coefficients of the reflection amplifier for different cavity positions. Fig. 11(a) shows the magnitude of the open loop gain in dB and Fig. 11(b) shows the open loop gain around resonant peak in polar form.

Fig. 11 shows that the maximum cavity offset position that can maintain oscillation, under the given amplifier bias condition, is 6.8 mm. The loop gain variation with bias is illustrated in Fig. 12, which presents open loop gain versus $V_g$ and $V_d$ for the cavity positioned at 5.9-mm offset. It is seen that the gain of the reflection amplifier is pretty invariant with $V_g$ while it increases about 4 dB when $V_d$ is increased from 4 to 15 V. The gain variation has implications on the phase noise as it will affect the device compression level, an effect that will be further discussed in Section IV.

The output power at fundamental frequency is measured in the range from $+2$ dBm to $+15$ dBm, and efficiency from 5% to 10% for the swept bias.

The cavity oscillator is characterized with a Rhode & Schwarz signal source analyzer (FSUP50). A direct measurement with the FSUP reveals that the oscillator is capable of a phase noise comparable to or below the noise floor of the measurement system, which is specified to a typical value of $-138$ dBc/Hz @ 100 kHz for a 10-GHz signal. Therefore, a setup based on two almost identical cavity oscillators was built and the outputs were down-converted to a much lower measurement frequency [see Fig. 13(a)], where the noise floor...
of the system is better. The IF frequency is 30 MHz due to slightly different dimensions of the otherwise identical cavities. On one of the oscillators (the reference oscillator), an isolator is placed to prevent load pulling, and on the other oscillator [the device-under-test (DUT)], a small attenuator (0–6 dB) is placed to form a broadband load. The stability from this isolation was found sufficient. The external components used in the setup are listed in Table I.

The two reflection amplifiers are identical, using the same batch of active device. In the first phase-noise measurement, both oscillators were tuned regarding cavity position and bias in the same way, until an optimum phase noise of $-141$ dBc/Hz @ 100 kHz was measured for the down-converted signal. The best value was found at $V_g = -2.5$ V. Subtracting 3 dB from this value, due to the mixing process, it is found that the minimum phase noise of a single oscillator is $-144$ dBc/Hz @ 100-kHz offset for optimum cavity position and optimum bias (see Fig. 14). In Fig. 15, the phase noise versus offset frequency for best bias is shown.

In further measurements, presented in Section IV, the reference oscillator is kept fixed while the DUT is varied to investigate variations with cavity position and bias. All phase-noise measurements are linearly corrected for the contribution from the reference oscillator, i.e., $-144$ dBc/Hz is subtracted. The output power from the reference oscillator is about $+10$ dBm, which is sufficient LO power for the mixer.

IV. RESULTS AND MEASUREMENTS OF OSCILLATOR

A. Measurements of Oscillator

The optimum phase-noise performance was presented in Figs. 14 and 15. Comparing Fig. 14 to 12, it is seen that the variation in phase noise at 100 kHz approximately follows the variation in small-signal open loop gain. The best phase-noise performance is seen for low $V_d$, where the loop gain is small. At increased $V_d$, and associated loop gain, the 100-kHz phase noise quickly increases. The degradation in phase noise with increased loop gain is expected due to nonlinear noise conversion being more important as the device is forced into harder compression [10] to keep unity loop gain in the closed loop.

The gain compression can be controlled either by adjusting the coupling factor to the cavity by changing the distance of the cavity perpendicular to the exciting microstrip line or by limiting the gain of the amplifier by choosing lower $V_d$. Table II presents optimum oscillator performance for four
TABLE II
SUMMARY OF OPTIMUM PERFORMANCE FOR DIFFERENT CAVITY PLACING AT OFFSET FREQUENCY 100 AND 400 kHz

| Cavity position offset (mm) | $R_s$/Loss (Ω)/(dB) | Optimum phase noise performance @100 kHz (dBc/Hz) @ bias level ($V_d/V_g$) | Phase noise @400 kHz (dBc/Hz) | FOM @ 400 kHz (dB) | $F_{op}$ @ 400 kHz (dB) | Gain (dB) | Reflection Amp @ opt for 100 kHz bias ($V_g/V_d$) | $|R_m|$ (Ω) | Open loop gain (dB) |
|---------------------------|---------------------|--------------------------------------------------------------------------------|-------------------------------|---------------------|---------------------|-----------|-----------------------------------------------|----------------|-----------------|
| 0.6                       | 8.1/2.8             | -139 @4/-1.9                                                                  | -152.5                        | 219                 | 15                  | 6.8       | 18.7                                         | 4              |                 |
| 3.4                       | 14/5                | -141 @4/-1.8                                                                  | -155.5                        | 223                 | 10                  | 7.0       | 19.1                                         | 2              |                 |
| 3.9                       | 19/2.7              | -141.5 @5/-1.6                                                                | -156                           | 221                 | 9                   | 8         | 21.6                                         | 1              |                 |
| 5.9                       | 20/7.4              | -144.1 @5/-1.8                                                                 | -160                           | 227                 | 7                   | 7.9       | 21.3                                         | 0.5            |                 |

Fig. 16. Measured phase noise @ 100 kHz versus bias. (a) Cavity at offset 0.6 mm from the center coupling line. (b) Cavity at offset 5.9 mm from the center coupling line.

different cavity coupling factors. In summary, the lower the loop gain, the better the phase-noise performance.

To further illustrate how the phase-noise performance varies with bias condition, Figs. 16 and 17 present phase noise versus bias condition at offset frequencies of 100 and 400 kHz, respectively, for two different cavity couplings, i.e., the strongest coupling (0.6 mm) and the weakest coupling (5.9 mm).

Fig. 16 shows that 100-kHz phase noise quickly degrades with $V_g$ while very little variation in phase noise is detected for a change in $V_d$. The phase-noise degradation with $V_d$ can be understood as an effect of the increased nonlinear noise conversion associated with the increase in open loop gain, cf. Fig. 12. Similarly, the invariant phase noise seen for increased $V_g$ can be interpreted as the balancing of three effects, which are: 1) increased flicker noise degrading phase noise; 2) increased power improving phase noise; and 3) increased efficiency improving phase noise.

Fig. 17, presenting phase noise at 400 kHz, also shows that phase noise is invariant to a change in $V_g$. However, in contrast to Fig. 16, little variation is seen for a change in $V_d$. A reason is that 400 kHz, in contrast to 100 kHz, is primarily in the $1/f^2$ region. As 400 kHz is in the $1/f^2$ region, it is in accordance to the theory in Section II a reasonable frequency for calculating FOMs. At 5.9-mm offset, an excellent FOM of 227 is reached.

It is interesting to compare the measured FOM to the fundamental limit that can be calculated from (9). Given the measured $Q_0 = 3800$, $\text{FOM}_{\text{M00}} = 246$ according to (9). Thus, a 19-dB difference compared to the theoretical noise floor, where it is assumed that the noise of the active device is 0 dB, and the efficiency is 100% ($P_a = P_{DC}$), by setting $F_{\text{eff}} = 0$ dB in (7). Out of this, 12 dB can be deduced to finite efficiency in the reflection amplifier. Calculating $F_{op}$ according to (8) shows that after compensation for finite efficiency, a minimum $F_{op}$ of 7 dB is reached for $V_d/V_g = 5 \, \text{V}/-2.1 \, \text{V}$. The optimum phase noise, FOM, and $F_{op}$ are reached for an open loop gain near unity, i.e., a coupling factor near $\beta = 1$, and not $\beta = 1/2$ as deduced in the theory. Thus, while useful for comparison versus noise floor, a linear theory is inadequate for prediction of optimum operating condition with respect to phase noise. When it comes to optimization for low phase noise, it is more important to minimize the nonlinear noise conversion compared to optimizing the power transfer and loaded quality factor.

B. Measurements Compared to Simulations

Fig. 18 compares measured phase noise versus offset frequency, with bias condition as a parameter, to simulations using ADS harmonic balance with LF noise parameters according to (10) extracted from field-effect transistor (FET) characterization. A pretty good agreement is seen for the $1/f^2$ region, while there is significant discrepancy in the $1/f^3$ region due to non-accurate representation of the nonlinear noise conversion. This is due to a stationary handling of the noise sources, and that (10) is strictly valid only at dc.

Fig. 19(a) and (b) present phase noise versus offset frequency at 100 and 400 kHz, respectively, with $V_g$ as a parameter. The figure shows simulations both with and without a flicker noise...

Fig. 16. Measured phase noise @ 100 kHz versus bias. (a) Cavity at offset 0.6 mm from the center coupling line. (b) Cavity at offset 5.9 mm from the center coupling line.
HÖRBERG et al.: PHASE-NOISE ANALYSIS OF X-BAND ULTRA-LOW PHASE-NOISE GaN HEMT BASED CAVITY OSCILLATOR

**Fig. 17.** Measured phase noise @ 400 kHz versus bias. (a) Cavity at offset 0.6 mm to the center coupling line. (b) Cavity at offset 5.9 mm to the center coupling line.

**Fig. 18.** Measurements compared to simulations. Dots are the simulated data for the same bias range, \( V_{dd} = 10 \text{ to } 13 \text{ V}, V_{gs} = -2.2 \text{ to } -1.6 \text{ V} \). Guidelines for \(-20 \text{ dB/dec}, \) respectively. \(-30 \text{ dB/dec} \) slope are added in graph.

At 100-kHz offset, the discrepancy between measurements and simulations including flicker noise is about 5–7 dB, while it is about 2–3 dB at 400-kHz offset. The larger discrepancy at 100-kHz offset is due to nonlinear conversion of flicker noise being more important near the carrier. The reason for showing the simulations without flicker noise is that it makes it easy to compare to the fundamental noise limit and FOMs derived in Section II.

**Fig. 19.** (a) Simulated phase noise at 100 kHz. (b) Simulated phase noise at 400 kHz with and without model considering LF noise (LFN), compared to measurement at \( V_{dd} = 4 \text{ V} \).

**Fig. 20.** Simulated \( F_{eq} \) compared to measurement at \( V_{dd} = 4 \text{ V} \). (a) 100 kHz. (b) 400 kHz.

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**Fig. 20.** Simulated \( F_{eq} \) compared to measurement at \( V_{dd} = 4 \text{ V} \). (a) 100 kHz. (b) 400 kHz.

**C. Results Related to Other Work**

The result of the cavity oscillator reported in this paper has been benchmarked versus performance of dielectric resonator oscillators (DROs) and metal cavity based oscillators in the open literature.

**Table III** presents a summary of the benchmark with a focus on phase noise and FOM at offset frequency of 100 kHz.

It is found that the oscillator of this work performs very well in the comparison. In fact, only one oscillator with comparable FOM has been found, it is a DRO based on an InGaP HBT [20], which has comparable FOM below 100-kHz offset. However, it should be mentioned that the oscillator in this work is based on a resonator with a lower \( Q \) factor, \( Q_{dl} = 3800 \), compared to \( Q_{dl} = 8800 \) for the dielectric resonator in [20]. Compensating...
for the Q factor, the oscillator of this work presents an effective noise figure that is 7 dB closer to the theoretical noise floor.

To compare the GaN technology to other GaN oscillators, the parameter $F_{\text{eff}}$ can be used, as that is independent of $Q_0$ and make a relevant comparison regardless of whether or not the resonator architecture is internal or external. Table IV is showing state-of-the-art oscillators in the GaN integrated in the MMIC compared to this cavity oscillator. It is seen that $F_{\text{eff}}$ for this oscillator is comparable to the best oscillators, indicating a good design.

V. CONCLUSION
A GaN HEMT oscillator based on a MMIC reflection amplifier and an aluminum cavity has been designed, manufactured, and characterized. Excellent phase noise of $-145$ dBc/Hz @ 100 kHz and $-160$ dBc/Hz @ 400 kHz from a 9.9-GHz oscillation frequency has been experimentally demonstrated. At 400-kHz offset, the power normalized FOM is 227. To the authors’ best knowledge, this is the best phase noise and FOM reported for any oscillator based on a GaN HEMT device. It is also state-of-the art for cavity based oscillators regardless of active device technology.

Beside having demonstrated excellent phase noise, this work has also investigated experimentally how the phase noise depends on the coupling factor and bias conditions. It has been found that minimum phase noise near carrier appears for relatively low drain voltage ($V_D - 3 - 4$ V), moderate current, and with open loop gain as low as possible. Compensating for finite efficiency in the reflection amplifier, the measured phase noise is within 7 dB from the theoretical noise floor.

This work has demonstrated that oscillators with excellent phase-noise performance can be designed in GaN HEMT technology, provided that coupling factor and bias levels are well controlled in order to avoid nonlinear conversion of flicker noise.

REFERENCES

TABLE III
BENCHMARK OF LOW PHASE-NOISE OSCILLATORS BASED ON EXTERNAL RESONATORS

<table>
<thead>
<tr>
<th>Part</th>
<th>$f_0$ (GHz)</th>
<th>$f_0$ (MHz)</th>
<th>$F_0$ (mW)</th>
<th>FOM (dBc/Hz)</th>
<th>Active device Technology</th>
<th>Resonator Technology</th>
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<td>N/A</td>
<td>N/A</td>
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<td>Silver coated cavity</td>
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<td>GaN HEMT</td>
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<td>N/A</td>
<td>SiGe HBT</td>
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<td>220</td>
<td>GaAs InGaP HBT</td>
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<td>GaN HEMT</td>
<td>Aluminum cavity</td>
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TABLE IV
COMPARISON TO STATE-OF-THE-ART GaN HEMT BASED OSCILLATORS

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</table>
Mikael Hörberg was born in Olofström, Sweden, in 1972. He received the M.Sc. degree in engineering physics from the Chalmers University of Technology, Göteborg, Sweden, in 1995, and is currently working toward the Ph.D. degree in microtechnology and nanoscience at the Chalmers University of Technology.

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Mr. Lai was the Transistor Modeling Competition winner in 2012 IEEE Microwave Theory and Techniques Society (IEEE MTT-S), International Microwave Symposium (IMS).
Paper [D]
A GaN HEMT X-band Cavity Oscillator with Electronic Gain Control

M. Hörberg, D. Kuylenstierna

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A GaN HEMT X-band Cavity Oscillator with Electronic Gain Control

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Abstract — This paper reports on a very low phase-noise GaN HEMT cavity oscillator at 8.5 GHz based on a reflection amplifier with electronic gain control. The gain control functionality is essential in order to control the open loop gain, which is critical for the phase noise performance. A large loop gain forces the oscillator in deep compression, resulting in increased noise conversion and degraded phase noise. On the other hand, a sufficient gain margin is mandatory to ensure satisfaction of the oscillation condition with margin that covers temperature drift and individual spread.

The electronic gain control uses varactors to change the output termination of a reflection amplifier. In this way the loop gain can be set independently of the bias point of the active device and the position of the metal cavity. A minimum phase noise of -136 dBc/Hz@ 100 kHz off-set is achieved, which is comparable to what is reached for a mechanically tuned oscillator in the same process.

Index Terms — Phase noise, oscillator, GaN HEMT, cavity, gain control

I. INTRODUCTION

Low phase noise oscillators are important in modern communication systems with high complexity modulation [1]. To design a low phase noise oscillator, the coupling between the active device and the resonator needs to be optimized so that the loaded quality factor and the power over the resonator are simultaneously high. This is partly contradictory, as the loaded quality factor will be degraded for a strong coupling [2]. A strong coupling may also result in extensive non-linear noise conversion with negative influence on near carrier phase noise [3].

It has recently been demonstrated that GaN-HEMT technology, thanks to high power and high efficiency can be used for design of low-phase noise oscillators [4], [5], [6]. In particular an ultra-low phase noise GaN HEMT oscillator based on a metal cavity oscillator was reported in [7]. Although very good performance is demonstrated in [7], the method requires extensive manual tuning of cavity position to ensure optimal loop gain. Such a method is not desired in volume production, a method that allows electronic tuning would be better.

A commonly used method to adjust the gain of the amplifier is to change the gate or drain bias of the active device. However, that method has two obvious drawbacks as the optimum power condition cannot be reached independent of the optimum loop gain and further the flicker noise changes with gate and drain bias [8], [9].

In this paper we are utilizing varactors in the output termination network of a reflection amplifier to adjust the effective feedback path and thereby control the gain. As the varactors are isolated from the resonator, they do not influence the Q factor of the resonator. Neither do they cause any modulation noise.

II. THEORY

A schematic of a reflection-type oscillator is shown in Fig. 1. The phase noise can be expressed as [2],

\[ \mathcal{L}(\Delta f) = \frac{FKT}{8(Qo)^2\beta/(1+\beta)^3 P_{RF} (\Delta f)} \]

(1)

where the resonator coupling is defined as \( \beta = R_s/R_n \). It has the optimum for \( \beta = 1/2 \).

Fig. 1 Schematic of a reflection-type oscillator
The coupling factor $\beta$ may be optimized either by changing the resistance $R_N$ of the resonator or by changing the negative resistance $R_0$. The latter can be electronically tuned with a variable-gain reflection amplifier which may be realized with a varactor in the feedback path. It is important to ensure that the gain control does not affect the oscillation frequency with modulation noise as a result. To account for modulation phase noise in a VCO, Leeson’s equation can be extended with one term [10]. The phase noise at offset frequency $\Delta f$ from the carrier frequency $f_0$ can then be expressed

$$L(\Delta f) = 10\log \left[ \frac{K_T}{2P_s} \left[ 1 + \left( \frac{f_0}{2Q_s\Delta f} \right)^2 + 1 + \frac{\Delta f_{1/f}^2}{\Delta f^2} + \frac{K_s^2V_m^2}{8\Delta f^2} \right] \right]$$

(2)

where $k$ is Boltzman’s constant, $T$ the temperature in Kelvin, $P_s$ the RF-power dissipated in the resonator, and $Q_s$ its loaded quality factor. $F$ and $\Delta f_{1/f^2}$ are fitting parameters for the noise level of the active device. The last term includes the modulation noise, where $K_s$ is the tuning sensitivity (Hz/V), and $V_m$ the applied voltage noise on the varactor.

III. DESIGN

The MMIC reflection amplifier is designed in the GH25-10 process from UMS. The gate periphery of the active device is 400 um. To enable a reflection gain at the gate side, a 2.6 nH inductor in series with two back-to-back connected varactors terminates the drain. The varactors are biased through two 5 kΩ resistors which act as RF-choke for bias supply and DC-return. On the source side, there is a small resistor of 10 Ω in series with an inductor, implemented as a transmission line of length 1000 um and width 20 um, and a capacitor of 0.5 pF in parallel. The source network tunes the peak of the gain. The capacitance of the varactors is used to set the reflection gain, without changing the operation point in terms of $V_{gS}$ and $V_{dS}$ bias of the active HEMT. A picture and a schematic of the reflection amplifier are shown in Fig. 2.

![Fig. 2](image_url)

Fig. 2 (a) Chip photo of the reflection amplifier. Size 1.0x1.0 mm. (b) Schematic.

Two back-to-back varactors are used to increase the voltage headroom and to improve the linearity of the $C(V_{varactor})$ characteristics due to the RF-voltage swing across them. The size of each varactor is 4x60 um, and the junction capacitance is formed between gate and drain-source. An external cavity tank is connected to the gate side of the active HEMT, which also provides an RF-output port, and a termination for unwanted reflections. Measured unloaded $Q$-factor of the cavity is around 3000.

IV. RESULTS

The measured tune ability of the amplifier gain versus frequency for different varactor voltage is shown in Fig. 3 (a). In Fig. 3 (b) the reflection coefficient versus the same varactor sweep is presented in a polar plot. The phase variation at the designed oscillation frequency of 8.5 GHz is measured to be less than 20 degrees.

![Fig. 3](image_url)

Fig. 3 Amplifier gain with varactor voltage as parameter, $V_{var}$= 5 to 25 V. (a) Rectangular plot. (b) Polar plot.

The loop gain of the oscillator can also be adjusted by positioning the cavity at different offset perpendicular to a microstrip that couples to the resonator. A similar test setup of this is described in [7]. The reflection coefficient and the corresponding $R_s$ at resonance of the cavity at different offset is shown in Fig. 4.

![Fig. 4](image_url)

Fig. 4 Reflection coefficient of the cavity for different cavity positions (measured as offset between cavity center line and the microstrip line).

For a cavity position corresponding to a strong coupling (0.6 mm from the center-line), the optimum varactor voltage was close to 8 V, and the corresponding minimum phase noise was obtained for an amplifier bias of $V_d/V_g=4$ V/-3 V, see Fig. 5 (a). In Fig. 5 (b), the cavity is positioned for a weaker coupling, 4.2 mm offset from the center. The optimum varactor voltage is then shifted to about 6.5-7 V which is obtained for an amplifier bias of $V_d/V_g=5$ V/-2.9 V. In both cases, the minimum achieved phase noise is -136 dBc/Hz@100 kHz off-set from 8.5 GHz carrier.
Fig. 5 Phase noise at 100 kHz offset vs varactor voltage for different $V_g$ and $V_d$. (a) Centered cavity. (b) Cavity offset of 4.2 mm from center line.

The different cavity positions according to Fig. 4 causes $R_s$ to change from 9 $\Omega$ to 18 $\Omega$, which corresponds to a reflection coefficient change from -3.2 dB to -6.5 dB. When the varactor voltage is changed from 8 V to 6.5 V, the gain of the reflection amplifier changes from 5 to 6.5 dB. The difference in gain for the optimum performance is also due to a slight phase change with the varactor voltage, as is shown in Fig. 3 (b). In Fig. 5 (b), it is shown that the lowest bias level also caused the oscillator to stop oscillate due to a loop gain below unity for a varactor voltage higher than 6.5 V.

Fig. 6(a) shows measured phase noise contours versus $V_g$ and $V_d$ with varactor voltage fixed to 8 V. Fig. 6(b) shows phase noise versus off-set frequency.

Fig. 7 The frequency shift in ppm of center frequency versus the varactor voltage. The dashed line is calculated tuning sensitivity.

As a comparison, a circuit without gain control was designed in the same process and characterized with the same method. Measured result at a cavity position manually tuned for the best phase noise is shown in Fig. 8.

Fig. 8. Phase noise measurement at 100 kHz, without gain control, at best cavity position at low loop gain and weak coupling.

Best phase noise with gain control from Fig. 6 compared to the best phase noise without gain control from Fig. 8, shows a degradation of about 2-3 dB. The gain controlled variant also shows that phase noise is more dependent of $V_d$, probably due to that higher $V_d$ and therefore higher RF-swing affects the loaded termination of the varactors.
A comparison to other oscillators based on external resonators is shown in TABLE I, [7].

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<td>200</td>
<td>212</td>
<td>GaN HEMT</td>
<td>Aluminum coated cavity</td>
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(*) The plot in [11] is used to estimate the phase noise. (**) A buffer amplifier is included, but output power is comparable with other output power in this work. (***) calculated from power density and transistor size.

Some oscillators are better during their optimized test condition. In a robust design, margins for loop gain, start up condition, temperature drift and batch variation makes it difficult to meet the optimum point at all conditions, and some degradation is therefore necessary. By using the electronic gain control proposed in this report, these effects can be adaptively compensated, to reach near optimum performance even in presence of a statistical spread and/or model uncertainties.

V. CONCLUSION

A reflection type oscillator, based on an electrically gain-controlled amplifier in GaN-HEMT, together with an external cavity, shows very good phase noise of -136 dBc/Hz@100 kHz at an oscillation frequency of 8.5 GHz. The gain control can adjust for variations due to resonator coupling, temperature drift, and create margin for start-up condition, without changing the bias condition. Compared to circuitries without gain-control and manually tuned circuits, only a minor degradation in absolute phase noise is measured.

ACKNOWLEDGMENT

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REFERENCES

Paper [E]
Low phase noise power-efficient MMIC GaN-HEMT Oscillator at 15 GHz based on a Quasi-lumped on-chip resonator

M. Hörberg, D. Kuylenstierna

Low phase noise power-efficient MMIC GaN-HEMT Oscillator at 15 GHz based on a Quasi-lumped on-chip resonator

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Abstract—This paper reports on a negative resistance 15 GHz GaN HEMT oscillator using a quasi-lumped integrated resonator. The resonator is based on a lumped element parallel LC resonator and a piece of transmission line acting as impedance transformer and phase compensation. The main advantage of this type of resonator is that it gives good flexibility in choice of impedance level so that it is easy to control the coupling factor between the active device and the resonator which is mandatory to reach good phase noise. An excellent phase noise of -106 dBc/Hz@100 kHz from a 15 GHz carrier is experimentally demonstrated. The oscillator is also very power efficient with a power normalized figure of merit (FOM) of 191 dB, extracted at 100 kHz offset. To the authors’ best knowledge this is the highest FOM reported for a GaN HEMT MMIC oscillator.

Index Terms—LF-noise, phase noise, oscillator, GaN HEMT.

I. INTRODUCTION

Low phase noise oscillators are critical components in a communication system to handle advanced modulation schemes such as higher order QAM and OFDM. In modern systems with high data rates, far carrier phase noise is of increased importance [1]. Far-carrier phase noise performance, i.e., white phase noise, is determined primarily by noise figure of the active device and the power swing over the resonator [2]. In this perspective GaN HEMT technology, with its high breakdown voltage, is a good choice for the active device. Beside high breakdown voltage, GaN HEMT has better noise figure compared to bipolar technologies such as SiGe HBT and InGaP HBT.

MMIC design provide high integration level, is very repeatable against process variations, and are cost effective in volume production. To minimize phase noise, the resonator must be well matched to the active device in order to trade off power over the resonator and loaded Q factor. This is particularly important in MMIC technology, where the unloaded Q factor is relatively low. A common approach for impedance transformation is to use tapped resonator topologies, e.g., Colpitts or Hartley [3], with reactive impedance transformation that provides flexibility in impedance level. The Colpitts and Hartley topologies have also been shown to support waveforms resulting in low up-conversion of flicker noise [4]. Excellent phase noise has been demonstrated for balanced Colpitts and Hartley MMIC oscillators [5] [6].

Recently, it was demonstrated that simple reflection type oscillators can reach good phase noise and figures of merit (FOM), provided that the waveforms are controlled [7], which requires careful design of the resonator. A challenge in MMIC technology is to design resonators yielding reasonable impedance levels. In [7] a coupled distributed resonator was used, but that is too space consuming for MMIC integration.

This paper demonstrates a low phase noise power efficient 15 GHz GaN HEMT MMIC oscillator using a quasi-lumped integrated resonator that can be made fairly compact. The resonator is based on a lumped element parallel LC resonator and a piece of transmission line acting as impedance transformer and phase compensation. It gives good flexibility in choice of impedance level so that it is easy to control the coupling factor between the active device and the resonator with component values maximizing quality factor in MMIC technology.

II. THEORY

Oscillator phase noise can be quantitatively modeled by Leeson’s equation, [8],

\[
L(\Delta f) = 10 \log \left[ \frac{F_k T}{2 P_s} \left[ 1 + \left( \frac{f_0}{2 Q_L \Delta f} \right)^2 \right] \left[ 1 + \frac{\Delta f_{1/f}}{\Delta f} \right] \right] \tag{1}
\]

where \( P_s \) is RF power swing and \( Q_L \) is the loaded quality factor of the resonator. To reach a good phase noise, \( P_s \) and \( Q_L \) must be simultaneously high which is contradictory as \( Q_L \) is reduced due to loading when more power is coupled to the resonator. The principle can be easily understood from a simple reflection-type oscillator model as in Fig. 1.
Inside the resonator bandwidth, $\Delta f < f_0/2Q_o$, $\mathcal{L}(\Delta f)$ is minimized for an optimum coupling between the resonator to the active device. If the coupling coefficient is expressed $\beta = R_s/R_N$, see Fig. 1 [2], $\mathcal{L}(\Delta f)$ in the $1/f^2$ can be expressed

$$\mathcal{L}(\Delta f) = \frac{FKT}{8(Q_0)^2 \beta/(1 + \beta)^3} \frac{f_0^2}{\Delta f^2}$$

By differentiation of (2), it can be shown that the optimum phase noise is reached for $\beta=1/2$.

Designing for a certain $\beta$ is about finding appropriate $R_s$ and $R_N$. Beside considering the ratio between $R_s$ and $R_N$, the absolute values of $R_s$ and $R_N$ are also important. A higher value of $R_s$ would make the circuit less sensitive to gain variations in the amplifier. It may seem contradictory to design for high $R_s$ in a series resonator as the unloaded quality factor $Q_0$ of the resonator decreases with $R_s$ as

$$Q_0 = \frac{\omega_0 L}{R_s} = \frac{1}{R_s \omega_0 C}$$

where $\omega_0$ is the angular center frequency.

A detailed look at (3) shows that it is possible to reach high $Q_0$ simultaneously with relatively high $R_s$ under the condition that the impedance level of the resonator is increased accordingly, i.e. increasing $L$ and reducing $C$. However, $R_s$ cannot be scaled arbitrarily for two reasons. Firstly, a too high $R_s$ would result in a loadline not maximizing the power swing. Further, it is not practically possible to scale the impedance level to maintain high $Q_0$ for very high $R_s$. In MMIC technology, a reasonable $R_s$ is about 10 $\Omega$ which results in good coupling factor to an amplifier with a gain about 5-10 dB, corresponding to $R_s$ from 14 to 26 $\Omega$.

The challenge is to find a resonator topology providing the desired impedance level with a simultaneously high $Q$-factor. The first and most obvious choice is a lumped-element series resonator and another choice is an open quarter-wave ($\lambda/4$) microstrip resonator. Since, lumped-element inductors in MMIC technology are based on microstrips, the quality factor and impedance of a microstrip resonator is considered first, (a). With some approximations the series resistance of a $\lambda/4$ resonator can be expressed

$$R_s \approx \frac{c_0}{4W} \frac{\pi \mu_0}{\sigma \epsilon_r f}$$

where $c_0$, is the free space velocity of, $\epsilon_r$ relative permittivity, $\mu_0$ permeability of free space, $\sigma$ metal conductivity, $f$ the frequency and $W$ the width of the microstrip line. According to (4) the strip width has to be reduced below 5$\mu$m in order to reach reasonable $R_s$ which would also result in poor $Q$-factor. As an open stub cannot reach $R_s$ high enough with decent $Q_0$, the same applies to a lumped series $LC$ resonator. Instead resonators for reflection-type oscillators are often implemented with some kind of impedance transformation, e.g., the coupled-line half-wave resonator shown in (b). This type of resonator can easily achieve the right impedance level with reasonable $Q$ factor and be adjusted by the coupling factor that is controlled with separation of the lines in the coupler. A drawback although is relatively large dimensions for the $\lambda/2$ long wide strips required in the on-chip coupler. Instead, the oscillator in this work is based on a lumped parallel resonator that is transformed to a series resonator through a high-impedance $\lambda/4$ line, (c). The component values of the transformed parallel resonator can be expressed

$$L = \frac{Z_c^2}{R_s \omega_0 Q}$$

$$C = \frac{R_s Q}{Z_c^2 \omega_0}$$

where $R_s$ is the desired effective series resistance and $Z_c$ is the characteristic impedance of the $\lambda/4$ transformer.
III. DESIGN

A reflection type oscillator, demonstrating the concept of the quasi-lumped resonator, is implemented in a 0.25 um GaN HEMT process from Triquint. Fig. 3(a) and Fig. 3(b), respectively show chip photo and schematic of the designed amplifier with the resonator indicated. The reflection amplifier is built with a 400 um (8x50 um) gate periphery. The source is grounded through an inductor with a small resistance (15 Ω), in parallel with a capacitor for increased gain. The source resistance is mostly used for bias stabilization. On the drain side, an inductive termination is used, creating a reflective gain on the gate side, where the resonator is placed. The resonator has $Q_0=38$ and $R_s=10\Omega$, extracted from EM simulations in ADS Momentum.

![Chip photo](image)

**Fig. 3. a) Chip photo of the oscillator, size 2.0x1.0 mm and b) Schematic.**

The measured reflection gain of the amplifier is $|S_{21}| = 5$ dB, corresponding to $R_N = -14 \Omega$ at serial resonance. Thus, the coupling factor is $\beta = R_s/R_N = 0.7$, which is slightly different from the ideal theoretical coupling $\beta = 1/2$ [9]. To suppress unwanted resonances out-of-band, a 50 Ω load is added in parallel with the resonator. The 50 Ω load has negligible influence at the fundamental oscillation frequency.

IV. RESULTS

The MMIC oscillator was measured with an FSUP-50 signal source analyzer from Rohde Schwarz. The oscillation frequency is varying from 15.01 GHz to 15.12 GHz over the measured bias range, mostly dependent on $V_d$. The output power is coupled out from the resonator by a small capacitor of 0.5 pF, and is varying from -10dBm to +2dBm for all swept bias points.

Measured phase noise is shown in Fig. 4. Fig. 4(a) shows phase noise versus offset frequency for various bias points and. Fig. 4(b) shows a contour plot of 100 kHz phase noise as a function of $V_g$ and $V_d$.

**Fig. 4 Measured phase noise [dBc/Hz] alternatively, (a) Versus offset frequency. Guide lines for -30dB/dec resp -20dB/dec are added. (b) offset 100 kHz.**

It is seen from Fig. 4 that the minimum phase noise is -106 dBc/Hz and -133dBc/Hz, respectively at 100 kHz and 1 MHz offsets, both achieved at low $V_g/V_d$. For increased $V_g/V_d$, phase noise increases due to higher amount of flicker-noise that is up-converted to phase noise. Increasing RF-power in resonator by increasing bias is not beneficial at this low offset frequency. The designed oscillator is very power efficient with a power normalized figure of merit (FOM) [10] of 191 dB and 198 dB, respectively measured at 100 kHz and 1 MHz offset frequencies. TABLE I presents a comparison versus other GaN HEMT oscillators reported in open literature. There are some oscillators with comparable phase noise, [7], [11], but the power normalized figure of merit is better for the oscillator reported in this work. In fact the FOM reached in this work is higher than for any MMIC oscillator based on a GaN HEMT device. The 1MHz FOM is also better than for any CMOS oscillator and most InGaP HBT [5] [6] oscillators published.

### TABLE I

**Comparison to State of the Art GaN HEMT Based Oscillators**

<table>
<thead>
<tr>
<th>Freq (GHz)</th>
<th>Power (mW)/ Pout (dBm)</th>
<th>Phase noise @100k (dBc/Hz)</th>
<th>FOM* @100k</th>
<th>Phase noise @1MHz (dBc/Hz)</th>
<th>FOM* @1MHz</th>
<th>Ref</th>
</tr>
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<tr>
<td>1.95</td>
<td>400/4</td>
<td>-119</td>
<td>179</td>
<td>-149</td>
<td>189</td>
<td>[7]</td>
</tr>
<tr>
<td>9.1</td>
<td>600/15</td>
<td>-101</td>
<td>172</td>
<td>-130</td>
<td>181</td>
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<tr>
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<td>-87</td>
<td>146</td>
<td>-115</td>
<td>154</td>
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<td>600/10</td>
<td>-105</td>
<td>177</td>
<td>-135*</td>
<td>187</td>
<td>[14]</td>
</tr>
<tr>
<td>15.05</td>
<td>63/3</td>
<td>-106</td>
<td>191</td>
<td>-133</td>
<td>198</td>
<td>This work</td>
</tr>
</tbody>
</table>

FOM$^*$ - $FOM^* = 10\log10(\frac{P_{DC}}{P_{amp}})-10\log10(P_{out})$, where $L(\text{phasenoise})$ is the phase noise at offset frequency, 100kHz resp. 1MHz offset are used in the calculation, $P_{amp}$ is the oscillation frequency, $P_{out}$ is the DC power measured in mW,[10]. *Phase noise scaled to 1MHz offset by 30dB/decade.
V. CONCLUSIONS

A low phase noise power-efficient GaN HEMT MMIC oscillator based on a quasi-lumped MMIC resonator is reported. The quasi-lumped resonator topology, incorporating a parallel LC resonator and a piece of microstrip, offers good flexibility in impedance level with good Q factor, which is important in order to match the resonator and the active device in reflection type oscillators. The prototype oscillator based on the proposed resonator demonstrates an excellent phase noise of -106 dBc/Hz at 100 kHz offset from a 15 GHz carrier frequency.

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Paper [F]
Phase noise analysis of a tuned-input/tuned-output oscillator based on a GaN HEMT device

M. Hörberg, S. Lai, T. N. T. Do, D. Kuylenstierna

Phase noise analysis of a tuned-input/tuned-output oscillator based on a GaN HEMT device

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Abstract—This paper reports on an experimental analysis of phase noise in a tuned-input/tuned-output oscillator based on a bare-die GaN HEMT device. To investigate phase noise dependency on resonator coupling factor ($\beta$), the circuit is designed with flexibility to modify the resonant tank, in terms of unloaded quality factor ($Q_0$) and impedance level. The reflection coefficient ($\Gamma_{\text{amp}}$) from the reflection amplifier can also be varied. With exception for very high values of $\Gamma_{\text{amp}}$, the circuit is robust to variations in $\beta$. It is more important to choose a bias point where the flicker noise is low and the power reasonably high. A minimum phase noise of -150dBc/Hz @ 1MHz off-set from a 1GHz oscillation frequency and a power normalized figure of merit (FOM) of 186 are reached.

An interesting analytical result is also presented; power normalized figure of merit, commonly used for benchmark of oscillators is bound by $Q_0$ and can be related to the theoretical noise floor limit.

Keywords—LF-noise, phase noise, oscillator, GaN HEMT.

I. INTRODUCTION

LOW phase noise oscillators are critical components in wireless communication systems. For a given resonator quality factor, the phase noise performance at room temperature is fundamentally limited by the RF power [1]. From this perspective Gallium Nitride (GaN) high electron mobility transistor (HEMT) technology is attractive. The drawback with GaN HEMT and other field-effect-transistor (FET) technologies is relatively high flicker noise limiting near-carrier phase noise performance. A couple of GaN HEMT oscillators have been published and most have a $1/f^3$ slope all the way down into noise floor, clearly indicating influence from flicker noise [2]-[5]. There is a need to study in more detail how the phase noise of GaN HEMT oscillators depend on bias conditions, in particular how the effect of signal power and flicker noise are best traded versus each other.

In this paper we present a tuned-input/tuned-output oscillator based on a GaN HEMT device. The circuit is designed with flexibility to change magnitude and phase of open-loop gain. It is investigated under what conditions the oscillator performs the best, with the least up-conversion of low-frequency noise. The unloaded-quality factor ($Q_l$) of the resonator is also measured individually and phase noise measurements compared to low-frequency noise measurements of the GaN HEMT device. The measured phase noise is compared to the theoretical phase noise limit that would be achieved if the active device acts as a noise free energy restorer, optimally coupled to the resonator. It is also analytically shown that the theoretical phase noise limit can be related to power-normalized oscillator figure of merit [6].

II. THEORY

Oscillator phase noise can be quantitatively modeled by Leeson’s equation, [1].

$$L(\Delta f) = 10\log \left[ \frac{F_k T}{2 P_r} \left( 1 + \frac{f_0}{2 Q_0 \Delta f} \right)^2 \right] \left[ 1 + \frac{\Delta f_{1/f^2}}{\Delta f} \right]$$

(1)

where $P_r$ is RF power swing and $Q_l$ is the loaded quality factor of the resonator. To reach a good phase noise, $P_r$ and $Q_l$ must be simultaneously high which is contradictory as $Q_l$ is reduced due to loading when more power is coupled to the resonator. The principle can be easily understood from a simple reflection-type oscillator model as in. Fig. 1.

Fig. 1 Schematic of negative resistance oscillator.
In Fig. 1, the reflection amplifier is modeled as a signal source with internal resistance $R_{neg}$. At the oscillation frequency $f_o$, the source injects energy into the resonator; the power coupled to the resonator can be expressed

$$P_s = \frac{|V_d|^2 \beta}{2R_{neg}(1 + \beta)^2} \leq \eta P_{DC} \frac{\beta}{(1 + \beta)^2}$$  \hspace{1cm} (2)

where $\beta = R_{neg}/R_i$ is the coupling coefficient between the source and the resonator and $\eta$ is the DC to RF conversion efficiency. Also $Q_i$ can be expressed in terms of $\beta$ as

$$Q_i = \frac{Q_0}{1 + \beta}$$  \hspace{1cm} (3)

where $Q_0$ is the unloaded quality factor of the resonator. Combining (1)-(3) and focusing on the $1/f^2$ region, where flicker noise has no influence, the single-side band phase noise in logarithmic scale can be expressed

$$\mathcal{L}(\Delta f) = 10\log \left[ \frac{FKT}{8P_{DC}Q_0^2} \left( \frac{f_0}{\Delta f} \right)^2 \right]$$  \hspace{1cm} (4)

The finite efficiency $\eta$ is merged into $F$. Based on (4) minimization of phase noise is a matter of optimizing the coupling coefficient $\beta$. Differentiation reveals an optimum $\beta = 1/3$ which inserted into (4) shows that the theoretical minimum phase noise that can be reached is

$$\mathcal{L}_{\text{min}}(\Delta f) = -173.1 - 20\log Q_0 + 10\log \left[ \frac{1}{P_{DC} \Delta f} \left( \frac{f_0}{\Delta f} \right)^2 \right].$$  \hspace{1cm} (5)

In (5), $F$ is set equal to unity which is the theoretical limit in the case where the reflection amplifier acts as a lossless energy restorer. The discrepancy between minimum phase noise in (5) and a measured phase noise $\mathcal{L}_{\text{meas}}(\Delta f)$ can be defined as a figure of merit on how well an oscillator is designed, i.e.,

$$F = \mathcal{L}_{\text{meas}}(\Delta f) - \left( -173.1 - 20\log Q_0 + 10\log \left[ \frac{1}{P_{DC} \Delta f} \left( \frac{f_0}{\Delta f} \right)^2 \right] \right) = 173.1 + 20\log Q_0 + \text{FOM}$$

where FOM is commonly used power normalized-oscillator figure of merit [6]. Equation (6) is an interesting result as it shows that FOM is linearly proportional to the theoretical minimum phase noise, and also that minimum achievable FOM depends on the resonator’s $Q_i$ as only input parameter.

III. DESIGN OF EXPERIMENT

A 1 GHz tuned-input/tuned-output oscillator is implemented on gold-coated PCB with a bare-die GaN HEMT device connected by bond wires. Schematic and photo of the circuit are shown in Fig. 2(a) and (b), respectively. The resonant tank, placed at the drain-side, is based on surface-mount components while the gate is terminated in a piece of transmission line in series with a capacitance $C_{\text{term}}$, which determines the reflection-gain ($I_{\text{amp}}$) seen at the drain terminal. The gate termination and the resonator are implemented on separate PCBs with the transistor in-between, so that $I_{\text{res}}$ and $I_{\text{amp}}$ can be measured individually to ensure satisfaction of small-signal open-loop gain.

$I_{\text{res}}$ depends on the equivalent series resistance $R_{\text{tank}}$ of the resonator. For a given $Q_{0}, R_{\text{tank}}$ can be changed by the impedance level $Z_{\text{c}} = (L/C)^{1/2}$. Note that $R_{\text{tank}}$ is not a physical element but represents the parasitic loss of the surface mounted $L$ and $C$. $|I_{\text{amp}}|$ can be increased or decreased by changing $C_{\text{term}}$. Fig. 3(a) shows $I_{\text{res}}$ for two resonators with different $Z_{\text{c}}$. Fig. 3 (b) shows how $I_{\text{amp}}$ changes with $C_{\text{term}}$. If $C_{\text{term}} = 5\text{pF}$, $I_{\text{amp}}$ is too low and the oscillator will not oscillate. In the case of $C_{\text{term}} = 100\text{pF}$, $I_{\text{amp}}$ is very high and the oscillator goes easily into compression with negative consequences for phase noise and difficulty to control oscillator with bias. Fig. 4 shows how $|I_{\text{amp}}|$ varies with $Vd/Vg$ for $C_{\text{term}} = 6.8\text{pF}$ and $C_{\text{term}} = 10\text{pF}$. The absolute gain is higher for $C_{\text{term}} = 10\text{pF}$ and the variation with bias larger. The rest of this paper will focus on $C_{\text{term}} = 6.8\text{pF}$ for which $|I_{\text{amp}}|$ has a smooth and relatively small variation with $Vg/Vd$. 

![Schematic and Photo of the oscillator.](image)

Fig. 2 a) Schematic and b) Photo of the oscillator.

The possibility to adjust $Z_{\text{c}}$ and $C_{\text{term}}$ can ensure that the magnitude of the open loop-gain is sufficiently high.
However, any effort would be in vain if the phase condition cannot be guaranteed. To cope with the phase, the layout shown in Fig. 2 (b) includes five alternative microstrip paths that can be interchanged with bond-wires. Before the resonator and reflection amplifier are connected together, the bond-wires are connected to the path that ensures $\angle (\Gamma_{res} \Gamma_{amp}) = 0$. Fig. 5 shows contour plots of phase and magnitude of the open loop-gain versus Vd/Vg for a reflection amplifier with $C_{term}=6.8\text{pF}$ connected to tank A.

![Fig. 3 Measured S parameters. (a) Resonator. (b) Reflection amplifier at Vg=−2V, Vd=10V.](image)

![Fig. 4 Reflection gain vs. bias level for different termination.](image)

![Fig. 5 Measured open-loop gain ($\Gamma_{res} \Gamma_{amp}$). (a) Magnitude. (b) Phase.](image)

![Fig. 6 Phase noise vs. offset frequency for all bias. With fluctuation line, -30dB/dec, -20dB/dec. Spurious is removed.](image)

### IV. MEASUREMENT AND ANALYSIS

Fig. 6 shows measured phase noise versus off-set frequency under different bias conditions for the case of $C_{term}=6.8\text{pF}$ connected to TankA ($Q_{res}=49$ and $R_{tank}=11\text{ohm}$, calculated from small-signal S parameter measurements of resonator). It is seen that the slope depends on bias condition. To better illustrate the bias dependency, Fig. 7 shows contour plots of $L(\Delta f)$ at 100kHz and 1MHz, respectively. At $\Delta f=1\text{MHz}$ phase noise improves with increased drain voltage and current, a minimum of $-150\text{dBc/Hz}$ is achieved at $V_d/I_d=10V/70\text{mA}$. This is good as the motivation to use GaN HEMT for oscillators is that phase noise should improve with increased power. On the other hand, phase noise at 100 kHz stays essentially invariant along the same valley of increased Vd/Vg. The reason is that the increased power is counteracted by flicker noise which increases with Id. Fig. 8 shows measured flicker noise for the 8x50um GaN HEMT device used in the oscillator, the measured data can be fit with a standard flicker noise model with $k_F=6\times10^{-10}$ and $A_F=0.87$. If flicker noise would be only current dependent and unaffected by bias the phase noise would be invariant with Id for $A_F=1$. 

![Fig. 6 Phase noise vs. offset frequency for all bias. With fluctuation line, -30dB/dec, -20dB/dec. Spurious is removed.](image)
It is interesting to compare Fig. 7 to Fig. 5, the valley of minimum phase noise agrees pretty well with fulfillment of the phase condition at the resonant frequency of the resonator. If $\angle \left( \frac{\Gamma_{\text{res}}}{\Gamma_{\text{amp}}} \right) \neq 0$, the circuit will not oscillate where $Q_0$ is the best which would have negative consequences for phase noise.

Focusing on 1MHz which is essentially in the $1/f^2$ region (6) can be used to calculate FOM and discrepancy compared to theoretical noise floor. Along the gradient of minimum phase noise $\text{FOM}=186-183$ for $F=21-24$, out-of-which 2-3dB can be explained due to a coupling that deviates slightly from optimum $\beta=1/3$ and a finite internal oscillator efficiency which is about 65%, calculated from simulated wave-forms.

Table I compares this work to GaN oscillators in open literature. To calculate $F$ for oscillators in literature, $Q_0=40$ is assumed which is a reasonable value for MMIC resonators.

V. CONCLUSIONS

A GaN HEMT based tuned-input/tuned-output oscillator with possibility to control magnitude and phase of open-loop gain has been built. The circuit is used to draw conclusions about phase noise in a GaN HEMT based oscillator. In the $1/f^2$ slope-region, phase noise scales approximately with DC power, a minimum of -150dBc/Hz@1MHz offset from 1GHz is achieved. Normalized to DC power, a figure of merit of 186dB, among the best reported for GaN HEMT based oscillators is reached.

It is found that at 100kHz off-set, the performance is limited by flicker noise and stays essentially constant if power is increased and the phase match condition is maintained. On the other hand at 1MHz off-set, the phase noise is power limited and it improves with power if the phase match condition is maintained.

Other configurations, like common gate or common drain may give other result on figure of merit, and are topics for further work.

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Paper [G]
An X-band varactor-tuned cavity oscillator

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An X-band varactor-tuned cavity oscillator

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Abstract — This paper reports on an X-band varactor-tuned cavity oscillator. The varactors are mounted on a low loss printed circuit board (PCB) that is intruded inside the cavity, which enables efficient coupling to the RF-field. The varactors’ positions are changed by adjusting the intrusion depth of the PCB as well as the horizontal positions of the varactors on the PCB. This compromises between the tuning range and the resonator Q-factor.

The active part, i.e., the reflection amplifier, is implemented in GaN-HEMT MMIC technology. A microstrip line couples the cavity to the reflection amplifier. The coupling factor can be controlled by mechanically adjusting the cavity’s position. A lateral displacement changes the amplitude coupling, while a displacement along the microstrip line changes the phase condition.

A tuning range of 1.6 % about 10 GHz is reached with the PCB placed at 1 mm depth from the cavity wall. The measured phase noise at 100 kHz and 1 MHz offset, respectively, ranges from -111 dBc/Hz to -118 dBc/Hz, and -138 dBc/Hz to -146 dBc/Hz over the tuning range. Increased tuning range can be reached if the varactors are placed deeper into the cavity, but then the phase noise degrades due to the increased tuning sensitivity that causes modulation noise and degradation of large-signal quality factor as the varactors are exposed to a stronger RF-field.

Index Terms — Phase noise, oscillator, GaN HEMT, cavity, varactor tuned, VCO, Q factor

I. INTRODUCTION

Low phase noise oscillators are important in wireless communication systems with high complexity in modulation, and for products enabling higher frequency bands [1]. Tunability and reconfigurability are key features; and the tuning elements are generally the bottleneck for maintaining a high quality factor of the resonator over a broad tuning range.

The tuning functionality may be achieved either by switches or varactors. Monolithic microwave integrated circuit (MMIC) resonators based on varactors are used in Colpitts oscillators where wide tuning range is demonstrated [2]. However, the phase noise of MMIC VCOs is not very competitive due to limitations in achievable resonator quality (Q) factor. In external high-Q resonators like DROs, on the other hand it is difficult to achieve wide tuning range due to very limited varactor coupling. The commonly used tuning method in this case is to change the load capacitance outside the resonator [3], [4], or changing the phase condition with a drawback of not fully utilizing the Q-factor of the resonator [5]. These techniques provide tuning ranges of only a few tenths percent. More attractive is to embed the varactor inside the resonator, as demonstrated in SIW-resonators [6], or using high-Q-MEMS varactors as in [7].

In this paper we are utilizing an aluminum cavity and bare die GaAs varactor diodes that are mounted on a PCB which is intruded inside the cavity. Changing the varactor capacitance will perturb the electromagnetic field inside the cavity and change the resonance frequency.

II. DESIGN & ASSEMBLY

Leeson’s extended equation predicts the phase noise as [8], [9],

\[
\mathcal{L}(\Delta f) = 10 \log \left[ \frac{kFkT}{2P_s} \right] \left[ 1 + \left( \frac{f_0}{2Q_L \Delta f} \right)^2 \right] \left[ 1 + \frac{\Delta f_{1/f^2}}{\Delta f} \right] + \frac{K_0\nu_{1/f^2}}{8\Delta f^2}
\]

(1)

where \( k \) is Boltzman’s constant, \( T \) the temperature in Kelvin, \( P_s \) the RF-power dissipated in the resonator, and \( Q_L \) its loaded quality factor. \( F \) and \( \Delta f_{1/f^2} \) are fitting parameters for the noise level of the active device. The last term describes the impact of the modulation noise, where \( K_0 \) is the tuning sensitivity (Hz/V), and \( V_{1/f} \) the voltage noise on the varactor [9]. From (1) it is seen that maximizing \( PsQ_L^2 \) will give lowest phase noise. The impact of the tuning elements will beside lowering of \( Q_L \) also add modulation noise. Minimizing \( V_{1/f} \) demands efficient coupling with low resistive loss in coupling network and low noise in supply.

The cavity resonator is built from two pieces, an open rectangular waveguide, and a shallow shortened lid with a low loss PCB as chip carrier clamped in between. Via holes around the cavity opening make the ground plane intact. Two different versions of the lid, with depth 1 mm and 2.5 mm, respectively, are fabricated to vary the intrusion depth of the PCB. Two antiparallel varactor-diodes are mounted in two symmetrical rows close to the edge, for best power handling, linearity of \( C(\phi) \) and for minimizing the effect of self-biasing due to RF-swing. Fig. 1 (a) shows a photo of the deep and shallow cavity.
The varactors are bare-die GaAs diodes with abrupt doping profile, manufactured by MA/Com, and flip-chip mounted onto the PCB, aligned with the E-field for the TE$_{10}$-mode in the cavity.

Fig. 2 shows a picture of the complete setup. The cavity is coupled to a microstrip line, connected to a MMIC-reflection amplifier designed in 0.25 µm GaN HEMT MMIC technology.

As the reverse varactor voltage is reduced, the varactor capacitance is increased, which lowers the resonance frequency. The varactor is characterized using a Deloach resonance-structure, extracting serial loss and junction capacitance to $R_s$ and $C_j$ to 2 ohm and 40 fF, respectively, for a reverse varactor voltage of -20 V. The capacitance of the varactor tunes the frequency of the shorted $\lambda_g/2$-waveguide resonator as is depicted in Fig. 3 (a). Fig. 3 (b) shows a block-schematic of the setup.

Fig. 3. (a). Equivalent narrowband circuit model of the tuned waveguide resonator. (b) Block-schematic of the coupled cavity.

III. RESULTS & DISCUSSIONS

The tunability of the cavity depends on the intrusion depth of the PCB. The measured tunability for 1 mm and 2.5 mm depth, are 1.6% and 2.0%, respectively. Fig. 4 (a) and Fig. 4 (b), show reflection coefficient and Q factor for an intrusion depth of 1 mm. The same parameters for an intrusion depth of 2.5 mm are shown in Fig. 5. The measurements are compared to simulations that are based on an HFSS-model with internal ports connecting to a circuit-based non-linear varactor model. Compared to the simulations, the measurements are somewhat shifted in frequency, likely due to tolerances in the assembly as well as mechanical tolerances.

Despite shifts in absolute frequency, the relative tuning range as well as qualitative behavior of the quality factor are well predicted both for the 1 mm and 2.5 mm cases. Note that the Q factor degrades with frequency which means that it is not limited by the varactor for which Q increases with frequency. The best measured Q-factor is lower than simulated, probably due to the substrate loss and RF-leakage which may be underestimated in the model.

Phase noise was measured for the cavity with PCB at 1 mm depth. Firstly, the gate and drain bias was swept to find optimum bias condition. Fig. 6 (a-b) show measured phase noise at 100 kHz offset versus bias level $V_d/V_g$ for varactor voltages of 0 V and -15 V, respectively. For a varactor voltage of 0 V, the optimum phase noise is -111 dBc/Hz and for a varactor voltage of -15 V it is slightly better, -116 dBc/Hz.
The phase noise reaches minimum with increasingly reversed varactor voltage, corresponding to a higher frequency. Comparing, Figs. 9(a) and 9(b) it is seen that the phase noise variation follows the tuning sensitivity of the varactor. Thus the varactor is a bottleneck for phase noise even if its losses do not limit the Quality factor of the resonator.

Finally, oscillator phase noise was simulated in ADS using a linear resonator model fitted to small signal measurements. Fig. 10 (a) presents phase noise versus offset frequency with gate and drain voltage as parameters. Fig. 10 (b) presents a contour plot of phase noise at 1 MHz offset versus gate and drain bias. Comparing to the measured result in Fig. 7 versus bias, there is a significant discrepancy at $V_{varactor}$=0 V while the agreement at $V_{varactor}$=-15 V is better and similar in shape although the measured optimum is still 9 dB worse compared to simulations. This indicates that the phase noise is not primarily limited by the unloaded Q of the cavity but rather by modulation noise proportional to the tuning sensitivity and large-signal effects. Further, the measured oscillator has higher bias current and gain, which may introduce more flicker noise.

Another reason for the deviation is the effect of power dependent Q-factor due to RF-voltage limitations across the varactors as discussed in [10], [11]. Fig. 11 (a) shows measured Q-factor of the resonator versus power with tuning voltage as parameter. Fig. 11 (b) shows simulated Q-factor versus power with associated frequency change indicated. Simulated Q-factor is higher due to underestimated loss and leakage in the model as previously commented. The power in the resonator of the oscillator is around +15 dBm, which would degrade the Q-factor to less than 300 according to Fig. 11 (b).

Table I shows a comparison to other state-of-the-art tunable oscillators. With respects to far out noise at 1 MHz offset and FOM at 1 MHz this work is comparable to others result. However, the limited tuning range is disadvantageous for FOM.
A varactor-tuned cavity oscillator has been designed, built and characterized. The cavity is tuned with varactor diodes on a PCB mounted inside the cavity. This method enables significantly better tuning range compared to DROs and better phase noise compared to MMIC VCOs. A challenge though is the high RF-field inside the cavity that contributes to varactor modulation noise limiting further increased tuning range is the high RF-field and degradation of large-signal Q factor.

Table I
Comparison to other tunable high-Q oscillator

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<th>Ref</th>
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<th>Phase noise @ 100 kHz (dBc/Hz)</th>
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FOM=\left[\frac{\Delta f}{f}\right]+20\log[20\log(\text{PDC/1 mW})]+20\log(\text{TR/10})\) or FOM=FOM+20\log(\text{TR/10})

(*) Noise floor in measurement

IV. CONCLUSION

A varactor-tuned cavity oscillator has been designed, built and characterized. The cavity is tuned with varactor diodes on a PCB mounted inside the cavity. This method enables significantly better tuning range compared to DROs and better phase noise compared to MMIC VCOs. A challenge though limiting further increased tuning range is the high RF-field inside the cavity that contributes to varactor modulation noise and degradation of large-signal Q factor.

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Paper [H]
Low Frequency Noise Measurements - A Technology Benchmark with Target on Oscillator Applications

T. N. T. Do, M. Hörberg, S. Lai, D. Kuylenstierna

Abstract—This paper presents low frequency noise (LFN) measurements of some commonly used microwave transistor technologies, e.g., GaAs-InGaP HBT, GaAs pHEMT, and GaN HEMT. It investigates how the flicker noise scales with current and voltage in the different technologies. The target application is low-phase noise oscillators. From this perspective, low-frequency noise at given frequency normalized to DC power is used as benchmark parameter. A comparison between different measurement set-ups is also included. The problem of measuring low-frequency noise at high drain voltages and currents is considered. It is found that the flicker noise of GaN HEMT technology is in about the same level as of GaAs pHEMT, but when normalized with the DC power, GaN HEMT offers a better performance. For this reason, GaN HEMT is considered to have better potential in oscillator applications. Concerning InGaP HBT, when measured at 10 kHz it provides better performance in term of both absolute noise level and normalized values. Higher frequencies are in favor for GaN HEMT technology.

Keywords—Low frequency noise, flicker noise, baseband noise, oscillators

I. INTRODUCTION

Flicker noise is a physical property of all active electronic components [1]. It has large impact on circuits like power detectors, oscillators, and baseband low-noise amplifiers. Although the exact origin of flicker noise is unclear, the level is known to differ between technologies. It is well known that the levels of flicker noise are lower in bipolar junction transistors (BJTs) compared to field-effect transistor (FETs). It is also clear that the levels depend on the semiconductor lattice quality. The level is generally higher in a material with more defects. Further, in a material with high density of defects, the low-frequency noise has less clear 1/f slope. Defects act as generation-recombination (G-R) centers resulting in a Lorentzian type of noise [2]. It has also been shown that the level and characteristics of the flicker noise is related to device reliability [3].

Hence, there are several reasons to study the flicker noise. The purpose of this paper is to benchmark different technologies for oscillator applications. In an oscillator, the flicker noise is up-converted to phase noise around the microwave signal [4]. Therefore, LFN measurement is very important for the characterization of devices. In this perspective, a device with low flicker noise should be used. On the other hand, the phase noise is directly proportional to signal-to-noise ratio [5], which favors devices that can deliver high power. To weight both flicker noise and power, this paper benchmarks devices versus low-frequency noise, measured at a certain frequency, e.g. 10 kHz and normalized with DC power. Three different technologies InGaP HBT, GaAs pHEMT, and GaN HEMT are compared in this paper.

II. MEASUREMENT SETUP

The LFN measurements of devices in this paper are based on two different setups.

A. Current-Voltage Amplifier based setup

This LFN measurement setup, called setup A and proposed by Franz Sischka, Agilent Technologies [6], uses a current to voltage preamplifier from Standford Research (SR570). An internal voltage supply of the SR570 is used for collector/drain biasing while the base current/gate voltage is biased through a parameter analyzer for accurate current /voltage control and a 1Hz low-pass filter for the noise leakage elimination. A dual channel dynamic signal analyzer (DSA) is used for FFT calculation. Due to limitations of the DSA, the LFN measurement range is limited from 1 Hz to 105.7 kHz. Thanks to the good noise floor achieved with internal bias from the SR570, this setup is perfect for LFN measurement of low noise devices, e.g. HBTs, at low current/voltage biasing.

![Fig. 1. Franz Sischka’s setup with external bias tee.](image)

A limitation is that the internal voltage supply of the SR570 cannot support a voltage larger than 4V and a current larger than 6 mA. To improve the voltage/current handling capability, an external bias tee is added, as shown in Fig.1. The additional bias tee has a cut-off frequency of 3 Hz, leading to a discontinuity in the LFN spectrum. Further, the big electrolyte capacitors used in...
the bias tee are noisy which limits the noise floor. Besides being noisy, the big capacitors also require long time for charging/discharging at every collector/drain bias point. The long time constants leads to difficulties in automatic control of the setup.

B. Voltage-Voltage Amplifier based setup

An alternative to set up A completed with external bias tee is to use a voltage to voltage preamplifier from Stanford Research (SR560). Fig. 2 shows such a set-up based on the SR560, we call this set-up B. An internal voltage supply of the SR570 is used for gate biasing. At the drain side, the SR 560 is connected to the device and the drain voltage is biased from a parameter analyzer through a constant load resistor $R_L$ of 110Ω. The DC I-V measurement is done before the LFN measurement to calculate the compensation for the voltage drop across $R_L$. In every measurement, the channel resistance $R_{dc}$ also needs to be determined from the measured I-V curve ($R_{dc} = \Delta V/\Delta I$). Finally, the measured drain noise current is calculated by normalizing the measured drain noise voltage (which is performed in DSA and multiplied to the sensitivity of the SR560) to the parallel combination of resistances of $R_{dc}$ and $R_L$. This method also allows LFN measurement at the gate side. The accuracy of this method is very good in the forward-active region of the IV curve where $R_L$ is dominating resistance over $R_{dc}$. Our experiments are carried out for a minimum measured drain voltage of 2V for GaN HEMTs and 0.5V for GaAs pHEMT. All measurements are controlled automatically by Matlab.

C. Setup Verification

The two set-ups are verified based on a 4x125μm GaAs pHEMT with 220nm gate length. Its LFN is measured with the different measurement setups (setup A, setup B and setup B complemented with battery) for both low current biasing and high current biasing. The results agree very well at high current biasing. However, at low current biasing setup B is limited by the noise floor of the parameter analyzer as seen in Fig.3. To measure LFN below this level the parameter analyzer can be replaced by a battery, but then it is difficult to control the bias automatically. Despite having a noise floor limited by the parameter analyzer, setup B is still preferred for measurements of devices having 1/f noise levels higher than the noise floor of the parameter analyzer.

III. MEASUREMENT RESULTS AND DISCUSSION

In this part, the LFN results of some commonly used MMIC transistor technologies are presented, e.g., GaAs-InGaP HBT, GaAs pHEMT, and GaN HEMT. GaAs InGaP HBTs require a low noise floor and are measured with set-up A, while GaAs pHEMT and GaN HEMTs are measured with setup B. Figs. 4-6 show chosen LFN spectra of InGaP HBTs, GaAs pHEMT and GaN HEMTs, respectively. Data from all measured devices are summarized in Table I.
collector voltage will increase LFN although less pronounced than the current.

Fig. 5 shows LFN spectrum of a 150nm GaAs pHEMT device with 80µm gate-periphery. It is found that the level of noise is higher compared to InGaP HBT, and the shape of the noise is very near the ideal 1/f which indicates the crystal quality of the material is good with low density of localized traps which would result in Lorentzian GR-type noise [2]. GaAs pHEMT devices with 100nm gate-length have also been measured, their spectrum is similar to the 150nm gate-length devices but the level of noise is higher. An assumption is that the higher noise in devices with shorter gate-length is due to more defects along the gate.

Fig.5. LFN measurements of a 4x20µm pHEMT with 150nm gate length versus the frequency at Vd =3V and different gate bias points. Right up corner is the DC I-V characteristic of the device.

Several GaN HEMT devices from different processes such as UMS, TRIONQUIT, and Chalmers have also been measured. These devices all have different gate widths and number of fingers. For GaN HEMT devices, the drain voltages are held constant at 2V, 6V and 10V while the gate bias is swept. A typical LFN spectrum of a 2x75µm GaN HEMT with 250nm gate length at 10V drain voltage is shown in Fig.6. It is observed that the slope is somewhat steeper than 1/f. Some previous works have shown that the slope of GaN HEMT, γ, varies from 1 to 1.3 [7]. Our experiment shows an even deeper slope, from 1.3 to 1.5. For Vgs>-2.2V, a Lorentzian can also be spotted which is a sign of GR-type noise that can be related to defects acting as traps. γ>1 and GR-noise is likely a sign of non-perfect crystal quality. As crystal quality improves, it is expected that the GR noise disappears and the slope of flicker noise approaches unity.

Table I and Figs. 4-6 clearly verifies the well-known fact that bipolar devices have lower flicker noise compared to field-effect-transistors. Although there are differences between the different samples it is clearly seen that GaAs pHEMT and GaN HEMT devices have noise in the same order of magnitude, while the noise of InGaP HBTs is considerably lower. However, the measurements of different devices are taken for quite different bias points. In a fair comparison, the measured noise level should be normalized versus power. With focus on oscillator applications, we may initiate from Leeson’s equation [5] in which the oscillator’s single-side-band phase noise spectrum is defined as follows [4].

\[ L(\Delta f) = 10 \log \left[ \frac{F^2 k T}{Q P_s} \left[ 1 + \left( \frac{f_0}{2 \Delta f} \right)^2 \right] \left[ 1 + \frac{\Delta f}{\Delta f_1/f_3} \right] \right] \]  

where \( f_0 \) is the output frequency, \( \Delta f \) is the offset from the output frequency, \( F \) is the noise factor, \( k \) is Boltzmann’s constant, \( Q \) is the loaded quality factor, \( P_s \) is the power swing over the resonator, and \( \Delta f_1/f_3 \) is the 1/f^3 corner frequency.

Equation (1) shows that in the 1/f^3 region, oscillator phase noise is proportional to \((\text{flicker noise})/P_s\) [5]. Assuming a constant conversion efficiency, a fair benchmark for oscillator applications is \( \text{LFN/P}_{dc} \), where \( P_{dc} = V_d \cdot I_d \). Table I shows that when LFN is normalized with dc power, GaN HEMT has a better performance compared to GaAs pHEMT. For example, at 10 kHz the best GaN HEMT device demonstrates a benchmark parameter approximately five times better than the best GaAs pHEMT one, corresponding to an improvement of 7dB in the phase noise. At 100 kHz, the best GaN HEMT device has better flicker noise normalized power than the InGaN HBTs measured in this study.

Fig.7. Benchmark parameters at 10 kHz of 250nm GaN HEMT devices with different sizes versus the gate bias points.
It should be mentioned that the measurements listed in Table I are taken in the forward active region, where the flicker noise normalized versus power has its lowest value. Fig. 7 shows $LFN(10\,kHz)/P_{dc}$ versus gate voltage for 250nm GaN HEMT devices with different gate-periphery length. The flicker noise increases with current until the device is saturated. For low $V_{gs}$, large devices have better performance, while for larger currents, there is no distinct difference in performance between devices with different gate periphery length.

IV. CONCLUSION

Low-frequency noise of GaAs-InGaP HBTs, GaAs pHEMT and GaN HEMT devices have been compared. The work is carried out with target on oscillator applications. From this aspect, LFN at given frequency scaled with DC power is a relevant benchmark parameter according to Leeson’s equation. It is found that the best performance in terms of absolute flicker noise levels is obtained for GaAs-InGaP HBTs. However, at larger frequencies where shot noise may limit HBTs, GaN HEMTs have potential for better performance. Thus for oscillator design, GaN HEMT may be an attractive technology, in particular if phase noise at large off-set frequencies is of concern.

REFERENCES


TABLE I.

<table>
<thead>
<tr>
<th>Device</th>
<th>Size (μm)</th>
<th>Lg/We</th>
<th>μ</th>
<th>Id (mA)</th>
<th>Vd (V)</th>
<th>LFN@1kHz (A^2/Hz)</th>
<th>Vg, Iq (V)</th>
<th>LFN@10kHz (A^2/Hz)</th>
<th>Vg, Iq (V)</th>
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471