INTEGRATED SILICON BIPOLAR
WIDEBAND FREQUENCY MODULATION CIRCUITS
FOR HIGH-PERFORMANCE
ANALOG LIGHTWAVE TRANSMISSION

by

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We accept this thesis as conforming
to the required standard

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April 2003
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Date April 30, 2003
This thesis describes the research done to achieve wideband frequency modulator and demodulator circuits for transmission of analog multi-channel cable television (CATV). These circuits were fabricated in an advanced silicon bipolar technology and resulted in the best-reported fully electronic implementation of a modulator for wideband frequency modulation (WFM) of CATV signals. Design of the modulator is achieved by using an emitter-coupled multivibrator (ECM) based oscillator. The viability of this approach is based on the ECM wideband tuning linearity at high-speed operation, particular noise requirements of CATV transmission, and a novel phase noise reduction technique.

Results for an ECM-based current-controlled oscillator (ICO) show linear operation in the range of 1GHz to 2.5GHz. Results for a linearized ECM-based voltage-controlled oscillator (VCO) show linear operation in the range of 1.5GHz to 2GHz. Linearity, from static tuning measurements, show maximum deviations from an ideal linear fit of 15MHz, and 1.5MHz for the ICO and VCO respectively. Results for a wideband delay-line demodulator show linear operation in a range of 2.5GHz, with back-to-back modulator/demodulator modulation bandwidth of 1GHz.

A state-of-the-art phase noise measurement system, based on automatic data gathering with a spectrum analyzer followed by mathematical post-processing, is presented. This system measures the phase noise of a variable oscillator versus tuning input, at a fixed offset from the carrier, with better accuracy than the best dedicated commercial measurement instrument available at the time.
The first study of phase noise in high-speed ECMs is presented from experimental measurements and circuit noise simulations. This study revealed new findings on the fundamental noise limits of ECMs. It was found that, at high frequencies, shot noise of the ECM core transistors dominates oscillator phase noise. Furthermore, phase noise was found to be directly proportional to the ECM tail current and inversely proportional to the square of ECM timing capacitor. This lead to a simple and novel optimization design approach to reduce phase noise by scaling up, by the same factor, the tail current and timing capacitor with minimal effects on tuning linearity. In addition, the modulator perspectives in a narrowcast WFM system are explored through system calculations.
Table of Contents

Abstract ....................................................................................................................... ii
List of Tables ........................................................................................................... vii
List of Figures ......................................................................................................... viii
Publications ............................................................................................................ x
Acknowledgements ................................................................................................. xi
1 Introduction ............................................................................................................ 1
   1.1 Introduction to Thesis ....................................................................................... 1
      1.1.1 Overview ................................................................................................. 1
      1.1.2 Outline of Chapter ................................................................................. 2
   1.2 Hybrid Fiber-Coax CATV Systems ................................................................. 3
   1.3 Wideband Frequency Modulation ................................................................ 7
   1.4 High-Speed Monolithic Voltage-Controlled Oscillators ......................... 9
   1.5 Phase Noise .................................................................................................... 15
      1.5.1 Fundamentals ....................................................................................... 15
      1.5.2 Phase Noise in Multivibrator Oscillators ......................................... 19
   1.6 Outline of the Thesis ...................................................................................... 22
2 Ultralinear WFM Circuits .................................................................................... 23
   2.1 Introduction to Chapter .................................................................................. 23
   2.2 Design Goals and Considerations ............................................................... 24
<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>System Perspectives</td>
<td>83</td>
</tr>
<tr>
<td>5.1 Introduction to Chapter</td>
<td>83</td>
</tr>
<tr>
<td>5.2 Composite-Second-Order (CSO) and Composite-Triple-Beat (CTB)</td>
<td>84</td>
</tr>
<tr>
<td>Performance</td>
<td>84</td>
</tr>
<tr>
<td>5.3 Carrier-to-Noise Ratio (CNR) Performance</td>
<td>91</td>
</tr>
<tr>
<td>5.4 Summary and Conclusions</td>
<td>99</td>
</tr>
<tr>
<td>Summary and Conclusions</td>
<td>101</td>
</tr>
<tr>
<td>Bibliography</td>
<td>109</td>
</tr>
</tbody>
</table>
List of Tables

1.1 Comparison of reported high-speed LC, ring, and relaxation monolithic VCOs ........................................ 11
4.1 Measured tuning and noise performance for three variants of the ICO with timing capacitors of 300, 150, and 100fF ......................................................... 68
4.2 Summary of principal noise contributors at one tuning point, 150fF ICO .............. 76
4.3 Simulated and predicted phase noise improvement from scaling C and \( I_{cirt} \) ............ 79
6.1 Summary of tuning performance for the simple ICO and the linearized VCO .... 102
6.2 Comparison of significant high-speed oscillator designs and WFM modulators reported to date ........................................................................................................... 103
# List of Figures

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.1</td>
<td>Typical HFC system for one-way CATV transmission</td>
<td>3</td>
</tr>
<tr>
<td>1.2</td>
<td>Conventional CATV transmission</td>
<td>5</td>
</tr>
<tr>
<td>1.3</td>
<td>WFM transmission system</td>
<td>8</td>
</tr>
<tr>
<td>1.4</td>
<td>Power spectrum of a stable frequency source</td>
<td>16</td>
</tr>
<tr>
<td>1.5</td>
<td>Typical SSB phase noise of an oscillator versus offset frequency</td>
<td>18</td>
</tr>
<tr>
<td>1.6</td>
<td>Noise mechanisms leading to phase noise in ECMs</td>
<td>20</td>
</tr>
<tr>
<td>2.1</td>
<td>Schematic of Emitter-Coupled Multivibrator</td>
<td>27</td>
</tr>
<tr>
<td>2.2</td>
<td>Implemented ICO</td>
<td>30</td>
</tr>
<tr>
<td>2.3</td>
<td>Schematic of linear transadmittance amplifier</td>
<td>33</td>
</tr>
<tr>
<td>2.4</td>
<td>Delay-line demodulator</td>
<td>35</td>
</tr>
<tr>
<td>2.5</td>
<td>Static response of simple ICO</td>
<td>37</td>
</tr>
<tr>
<td>2.6</td>
<td>Static response of the linearized VCO</td>
<td>39</td>
</tr>
<tr>
<td>2.7</td>
<td>Effect of switching delay on linearity of simple ICO</td>
<td>41</td>
</tr>
<tr>
<td>2.8</td>
<td>Effect of transadmittance amplifier response on deviation from linearity</td>
<td>43</td>
</tr>
<tr>
<td>2.9</td>
<td>Linearized VCO, time domain single-ended output at 1.8GHz</td>
<td>44</td>
</tr>
<tr>
<td>2.10</td>
<td>ECM-VCO dynamic response</td>
<td>46</td>
</tr>
<tr>
<td>2.11</td>
<td>Demodulator differential static-response</td>
<td>48</td>
</tr>
<tr>
<td>2.12</td>
<td>Microphotograph of back-to-back modulator/demodulator</td>
<td>49</td>
</tr>
<tr>
<td>2.13</td>
<td>Measured back-to-back modulator/demodulator response</td>
<td>50</td>
</tr>
<tr>
<td>3.1</td>
<td>Custom phase noise measurement setup</td>
<td>56</td>
</tr>
</tbody>
</table>
3.2 Mathematical fitting during the post-processing stage .................................. 60
3.3 Post-processed results versus input control current ........................................ 62
3.4 Measured phase noise at 5MHz offset for two ICOs with 150 and 300fF .......... 64
4.1 Measured phase noise versus tuning current .................................................. 68
4.2 Predicted and measured phase noise versus tuning current ............................. 72
4.3 Graphical demonstration of phase noise dependence with tuning current $I_{ctrl}$
   and timing capacitor $C$ ..................................................................................... 74
4.4 Circuit diagram of the simulated ICO ............................................................. 76
5.1 System transfer function used to calculate CSO and CTB ................................. 86
5.2 CSO versus channel module number and CTB for highest channel ................. 88
5.3 Worst case CSO versus per channel frequency deviation for 80-channels ...... 90
5.4 CNR of the jth AM carrier versus carrier frequency for 80-channels ............ 93
5.5 CNR of channel 80 versus frequency deviation and modulator phase noise ..... 95
5.6 Performance variables in 80-channel transmission ................................--------- 97
5.7 Performance variables in 40-channel transmission ......................................... 98
Publications

I hereby declare that I am the sole author of this thesis.

In accordance with University of British Columbia guidelines, I hereby declare that parts of this thesis have been or will be published under the following titles:


Please note that this thesis is the result of collaborative work with Thomas & Betts Photon Systems, now part of Scientific Atlanta Canada. Due to this partnership, the publication of results from this thesis has been in part delayed until recently.
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Chapter 1
Introduction

1.1 Introduction to Thesis

1.1.1 Overview

In this thesis we describe a study of high-speed silicon bipolar integrated circuits for optical transmission of analog multi-channel cable television (CATV). To this aim we performed the design, experimental characterization and performance analysis of ultra-linear wideband frequency modulator and demodulator circuits. This research is motivated by the needs of wideband frequency modulation (WFM) transmission and is framed in the context of the North American cable television standard NTSC. As WFM transmission is a prospective scheme to improve on the performance of existing hybrid fiber-coax (HFC) CATV systems, our research has drawn the industrial support and collaboration of Photon Systems Corp. (now a division of Scientific Atlanta). The circuits that resulted from this work constitute, to our knowledge, the best reported implementations of modulator and demodulator for WFM transmission. In addition, the study of noise in the modulator led to the implementation of a state-of-the-art phase noise measurement setup and to new findings on the fundamental noise limits of emitter-coupled multivibrator (ECM) circuits.

Three distinct topics are addressed in this thesis. The first topic is the design and performance testing of ultra-linear WFM modulator and demodulator circuits. The second topic is a systematic study of phase noise origins in the modulator. This includes the design of the aforementioned measurement setup, followed by experimental and numerical analysis. The
third and final topic examines numerically the perspectives of using the designed modulator in a WFM transmission system.

1.1.2 Outline of Chapter

In this chapter we provide background information relevant to this thesis. The application context of our work is introduced in Section 1.2 by a general description of hybrid fiber-coax CATV systems. In Section 1.3 we explain how system performance of existing CATV systems can be improved by the use of WFM transmission, and the need this creates for fully electronic modulator and demodulator circuits with extremely demanding requirements on linearity and tuning bandwidth. In Section 1.4 we review previous work done in the field of high-speed monolithic voltage-controlled oscillators (VCOs). Emphasis is given to the multivibrator VCO configuration, on which we based our designs. As phase noise performance is critical for this configuration, we provide in Section 1.5 a review of phase noise fundamentals, followed by a summary of previous work on noise in multivibrator oscillators. Finally in Section 1.6 we provide an outline for the remaining chapters of this thesis.
1.2 Hybrid Fiber-Coax CATV Systems

Current cable television systems are based on a hybrid fiber-coax (HFC) architecture in which optical links are used for long distance transmission and local distribution is done electrically through coaxial cables. Fig. 1.1 represents a typical HFC system for the simple case of one-way transmission of multi-channel television. The signals are transmitted from a central location via independent trunk connections to several neighborhood local feeders and from there distributed to the subscriber’s homes. A fiber link is used for every trunk portion of the network creating a set of independent local coax-cable systems [1]. This architecture takes advantage of low fiber losses, and high transmitter optical output powers to affordably extend system reach. When introduced approximately in 1989, it replaced a previous tree-and-branch topology, in which a single high-power coaxial trunk was branched continuously to extend system coverage.

![Diagram of typical HFC system for one-way CATV transmission](image)

Figure 1.1: Typical HFC system for one-way CATV transmission [1].
Chapter 1. Introduction

The transmission of the multi-channel CATV signal is based on a frequency division multiplexing approach (FDM) where many television channels, each of amplitude-modulated video with vestigial sideband (AM-VSB), are combined into a single broadband electrical signal. In the NTSC system (standard of the National Television System Committee) a typical transmission involves 79 channels, of 6MHz bandwidth each, distributed in a carrier allocation plan that ranges from 55.25 to 559.25MHz for a total bandwidth of 510MHz. This is commonly referred to as an 80-channel system. Details are illustrated in Fig. 1.2, where (a) shows the frequency spectrum of a single channel, and (b) the multi-channel frequency allocation. In Fig. 1.2 (a) the video carrier is 1.25MHz above the vestigial lower sideband (AM LSB), and followed by a 4MHz wide upper sideband (AM USB), and sound. In Fig. 1.2 (b) 79 such channels are distributed in frequency and identified with a channel module number. As shown, the spacing of the lowest carriers is not uniform, and the gaps exist for historical reasons, specifically for compatibility with commercial FM and other radio bands. For the same historical reasons the channel numbers, as known by the end-users, are not distributed sequentially, and consequently module numbers are used to label the channels with increasing carrier frequency\(^1\). The resulting multi-channel CATV signal is an extremely wideband amplitude modulated signal that must be transmitted with a minimum of degradation. Consequently, amplitude noise is an important limiting quality factor, as well as distortion produced by intermodulation products, which originate when the multi-channel signal passes through nonlinear components. To emphasize the latter Fig. 1.2 (c) shows the number of dominant second (\(\bullet\)) and third (\(\bigcirc\)) order intermodulation products that beat on each carrier in an 80-channel system.

\(^1\) This notation of showing results versus channel module number will be used throughout the thesis.
Figure 1.2: Conventional CATV transmission. (a) AM-VSB Spectrum of a single video channel. (b) NTSC frequency allocation plan for 80-channel transmission. (c) Number of dominant second (●), and third (○) order intermodulation products that beat on each channel.
Dominant second order terms are caused by products with frequencies of the type $f_1 + f_2$, where $f_1$ and $f_2$ are the frequencies of any two video carriers. These terms fall at 1.25MHz above the video carrier and cause visual diagonal lines interfering with the image. Dominant third order terms are caused by products with frequencies of the type $f_1 \pm f_2 \pm f_3$. These terms fall on the carriers and can cause streakiness in the picture [2]. The precise impact of these and other intermodulation products depends on the transfer function describing the nonlinearities and the amplitudes of the video carriers. For these reasons, metrics of carrier-to-noise ratio (CNR), composite-second-order (CSO) distortion, and composite-triple-beat (CTB) distortion are used to quantify signal quality. The former two metrics are typically expressed in dB with respect to the carrier, i.e., in dBc. Commonly adopted quality targets, at the television receiver end, are CNR better than 46dB, and CSO and CTB better than $-53$dBc for signal levels of 0 to +3dBm [2]. With these aims, the design of the overall HFC system is carefully balanced, in order to distribute the available noise and distortion budget throughout its different components. From this budget allocation, typical requirements for the trunk fiber equipment are CNR better than 52dB, and composite distortion CSO, CTB better than $-65$dBc, for a minimum transmission length of 20km [1].

At the fiber backbone optical transmission is performed by intensity modulating the optical carrier with the multi-channel CATV signal; the resulting optical signal is carried several tens of kilometers by a fiber-optic link and demodulated at the receiver end by a $pin$ diode photo-detector. In this simple analog optical transmission, AM-VSB modulation is maintained throughout the system and equipment costs are low. Single-mode standard fiber is generally used, exhibiting zero chromatic dispersion at a wavelength near 1310nm and minimum attenuation, typically 0.25dB/km, near 1550nm. The simplest systems operate at a
wavelength of 1310nm, and involve direct modulation of a semiconductor laser at the
transmitter. However, to reduce attenuation in the fiber link as well as to allow for the use of
erbium doped fiber amplifiers (EDFAs) the transmission is most commonly done at 1550nm.
This results in fiber dispersion, which induces CSO and CTB distortion. Thus, to reduce
dispersion cooled distributed feedback (DFB) lasers are used. System noise is normally
dominated by shot noise from the photo detector, but also includes optical amplifier noise,
laser relative intensity noise (RIN), and photo detector thermal noise. The demanding CNR
requirements necessitate high received optical powers; this sets a tight budget for transmission
reach and forces the use of high launched powers, which accentuate the effects of fiber
nonlinearities at the transmitter end.

1.3 Wideband Frequency Modulation

An alternative transmission approach that has been proposed to improve the CNR performance
of CATV optical links is that of wideband FM transmission (WFM) [3,4]. Fig. 1.3 shows a
block diagram of the WFM scheme. The entire multi-channel signal \((RF_{in})\) is used as input to a
wideband frequency modulator (FM MOD), and modulated onto a single carrier with a
frequency, \(f_c\), in the range of several GHz; the resulting wideband FM signal is used to
intensity modulate a laser source in the transmitter \((T_x)\), and is then launched into an optical
fiber link. At the receiver \((R_x)\), the incoming optical power is first converted into an electrical
signal, and later demodulated in frequency by a wideband FM demodulator (FM DEMOD);
thus, the signal is restored to its original composite AM-VSB form.
This approach exploits a general property of FM modulation, that is, the ability to increase system noise immunity by using a larger transmission bandwidth. This increased bandwidth is easily supported by optical fibers. By improving the CNR it allows either reduction of the transmitted power or extension of the transmission distance. It also provides a transmitted signal of constant peak intensity, which relaxes the demands on linearity of the optical components. For this reason, WFM transmission can also be used to take advantage of FM insensitivity to nonlinearities of the transmitter diode. This would allow the use of cost-effective alternatives to cooled DFB lasers such as uncooled, medium power, DFB lasers making economically feasible the distribution of specialized CATV programming to small communities. In this application direct modulation of a laser diode can be used in a 1310nm system for “narrowcast” transmission, aiming for short distances such as to provide high-received powers of about 0dBm.

Previous work to implement viable WFM systems has predominantly been carried on by Japanese groups; namely, at Matsushita Electric Industrial Co., where the concept was originally proposed [3], and at NEC Corporation. Their implementations target the Japanese CATV frequency allocation standard, which differs slightly from the NTSC standard. Their work has resulted in systems with limited performance for low channel counts, 20 to 40 channels, with attempts to reduce distortion by using dispersion-shifted fibers and pre-
distortion circuitry [3-6]. Their results have been limited by the performance of the WFM modulators. In the first work presented on wideband FM, an elaborate optoelectronic approach is used for the modulator based on interferometric interaction between two lasers [3]. In this scheme two optical waves are coherently combined and detected with a photodiode in the modulator; one of the waves is produced by modulating a laser with the multichannel AM signal to induce frequency shifts in the laser; and the other is unmodulated. The result is an electrical wideband FM signal, that is used to intensity modulate a third DFB laser diode in the transmitter [3]. In addition to having a limited performance, this system is complex, costly, and bulky. Therefore, a fully electronic modulator design is an attractive alternative. More recently, an electronic modulator based on silicon bipolar integrated circuits has been reported [5]. Unfortunately, modulator linearity was poor for a system of only 20 channels, despite use of pre-distortion. The electronic approach requires the design of a high-speed voltage-controlled oscillator (VCO) circuit with extremely demanding requirements on linearity and tuning bandwidth. Although VCOs have been extensively studied and are still the focus of active research, no design fulfilling the requirements of this application has ever been reported. As it is the goal of this work to implement such a circuit, we next proceed with a comparative review of high-speed monolithic VCO implementations.

1.4 High-Speed Monolithic Voltage-Controlled Oscillators

Strictly speaking, VCOs are oscillators in which the frequency of oscillation can be tuned by varying an applied voltage. By extension however, the term VCO is commonly used when referring to any tunable oscillator regardless of the nature of the tuning variable. In that sense, the development of VCO configurations has, by and large, been a byproduct of the evolution of fixed oscillators. Tunable and fixed oscillators have nonetheless fundamental conflicting
requirements. A stable frequency reference, thus low phase-noise, implies a natural preference for a certain frequency; good tunability, on the other hand, implies no particular frequency preference at all [7]. Consequently, as will be shown throughout this section, the design of VCOs unavoidably involves a trade off between tuning and phase noise performance.

For decades, the most common types of VCOs used in high-speed/RF circuits belong to the category of so-called "linear oscillators"; these oscillators consist of an amplifier with positive feedback and a frequency selective circuit. The term "linear" is used despite the fact that all oscillators depend upon circuit nonlinearities for operation, to indicate that the oscillations are due to weakly nonlinear operation of the amplifier. This is in opposition to the operation of nonlinear oscillators, that involve highly nonlinear switching of active devices, like in ring oscillators and relaxation-oscillators/multivibrators [8] discussed below. Some commonly known linear oscillators include the Colpitts and Clapp oscillators, but in general different configurations are possible, all involving a LC resonator circuit as the frequency selective network. For this reason, these circuits are most commonly referred to as LC oscillators. LC oscillators benefit from low noise output spectra due to the fact that the active devices are operated in their linear region and also to the filtering action of the tuning circuit. Unfortunately, this results in poor tuning linearity and narrow tuning range. Therefore, LC oscillators are ideal for applications dominated by low phase noise requirements. In practice, however, monolithic implementation of LC-oscillators with a high enough quality factor ($Q$) is difficult due to the resulting large physical sizes of the resonant circuit elements, most particularly the inductors. Furthermore, parasitic substrate losses degrade $Q$ in monolithic silicon circuits.

Table 1.1 summarizes the performance achieved, up to this date, by several reported high-speed monolithic VCOs. The table covers results for the three types of VCOs discussed in
this section; namely, LC, ring and relaxation oscillators. Notice that although LC oscillators working well into the microwave range have been reported, only LC designs operating in our frequency range of interest are included in Table 1.1. The central oscillation frequency is referred to as “$f_c$”, and the tuning ranges are, as reported in the references, defined in terms of the attainable output frequency ranges with no consideration of tuning linearity. Exceptions to this are explained further below. For comparison, tuning range is also shown as a percent of the central frequency.

Table 1.1: Comparison of reported high-speed LC, ring, and relaxation monolithic VCOs. Tuning performance is as reported in the references for full dynamic range regardless of linearity considerations. Exceptions to this are indicated by *. In addition designs marked with # show a narrower, but distinguishable, linear region.

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Year</th>
<th>$f_c$ (GHz)</th>
<th>Tuning range (Hz)</th>
<th>Phase noise (dBc/Hz) @ offset (Hz)</th>
<th>Oscillator Type</th>
</tr>
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<tbody>
<tr>
<td>[9]</td>
<td>1992</td>
<td>1.77</td>
<td>180M ± 5.0%</td>
<td>-88.0 @ 100k</td>
<td>LC</td>
</tr>
<tr>
<td>[10]</td>
<td>1995</td>
<td>1.60</td>
<td>245M ± 7.6%</td>
<td>-88.0 @ 1k</td>
<td>LC</td>
</tr>
<tr>
<td>[11]</td>
<td>1996</td>
<td>1.70</td>
<td>200M ± 5.9%</td>
<td>-85.0 @ 100k</td>
<td>LC</td>
</tr>
<tr>
<td>[12]</td>
<td>1997</td>
<td>1.71</td>
<td>120M ± 3.5%</td>
<td>-100.0 @ 500k</td>
<td>LC</td>
</tr>
<tr>
<td>[13]</td>
<td>1997</td>
<td>1.48</td>
<td>150M ± 5.0%</td>
<td>-105.0 @ 100k</td>
<td>LC</td>
</tr>
<tr>
<td>[14]</td>
<td>1999</td>
<td>6.50</td>
<td>1.02G ± 7.8%</td>
<td>-98.4 @ 1M</td>
<td>LC</td>
</tr>
<tr>
<td>[15]</td>
<td>2000</td>
<td>4.76</td>
<td>580M ± 6.1%</td>
<td>-98.0 @ 100M</td>
<td>LC</td>
</tr>
<tr>
<td>[16]</td>
<td>2001</td>
<td>2.33</td>
<td>605M ± 13.0%</td>
<td>-117.0 @ 600k</td>
<td>LC</td>
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<tr>
<td>[17]</td>
<td>2001</td>
<td>5.00</td>
<td>850M ± 8.5%</td>
<td>-114.0 @ 1M</td>
<td>LC</td>
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<td>[18]</td>
<td>2002</td>
<td>2.10</td>
<td>650M ± 15.5%</td>
<td>-139.0 @ 3M</td>
<td>LC</td>
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<td>[19]</td>
<td>2002</td>
<td>1.57</td>
<td>330M -14, +7 %</td>
<td>-133.5 @ 600k</td>
<td>LC</td>
</tr>
<tr>
<td>[20]</td>
<td>1997</td>
<td>0.62</td>
<td>606M ± 49.0%</td>
<td>-83.0 @ 100k</td>
<td>Ring</td>
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<tr>
<td>[21]</td>
<td>1997</td>
<td>0.97</td>
<td>1.44M ± 74.0%</td>
<td>Not reported</td>
<td>Ring</td>
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<tr>
<td>[22]</td>
<td>1999</td>
<td>0.97</td>
<td>450M ± 23.0%</td>
<td>-117.0 @ 600k</td>
<td>Ring</td>
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<tr>
<td>[23]</td>
<td>2000</td>
<td>4.50</td>
<td>1.00G ± 11.1%</td>
<td>Not reported</td>
<td>Ring</td>
</tr>
<tr>
<td>[24]</td>
<td>2001</td>
<td>0.96</td>
<td>609M ± 31.7%</td>
<td>-107.5 @ 600k</td>
<td>Ring</td>
</tr>
<tr>
<td>&quot;</td>
<td>&quot;</td>
<td>1.18</td>
<td>781M ± 33.0%</td>
<td>-106.9 @ 600k</td>
<td>Ring</td>
</tr>
<tr>
<td>[25]</td>
<td>2001</td>
<td>5.75</td>
<td>6.50G ± 56.5%</td>
<td>-82.0 @ 1M</td>
<td>Ring</td>
</tr>
<tr>
<td>[26]</td>
<td>1990</td>
<td>0.50</td>
<td>999M ± 99.9%</td>
<td>Not reported</td>
<td>Relaxation</td>
</tr>
<tr>
<td>[27]</td>
<td>1992</td>
<td>1.47</td>
<td>1.43G -41, +56%</td>
<td>Not reported</td>
<td>Relaxation</td>
</tr>
<tr>
<td>[5]</td>
<td>1997</td>
<td>3.95</td>
<td>2.90G ± 36.7%</td>
<td>Not reported</td>
<td>Relaxation</td>
</tr>
<tr>
<td>[28]</td>
<td>1998</td>
<td>0.41</td>
<td>780M ± 95.0%</td>
<td>-99.0 @ 2M</td>
<td>Relaxation</td>
</tr>
<tr>
<td>[29]</td>
<td>2001</td>
<td>4.30</td>
<td>2.00G ± 23.3%</td>
<td>-113.0 @ 600k</td>
<td>Relaxation</td>
</tr>
</tbody>
</table>
Measurements of phase noise are also given in Table 1.1 but cannot be compared directly as they are reported for different $f_c$'s and offsets. Nonetheless they are still useful for the purpose of a qualitative comparison. While the topic of phase noise will be reviewed in Section 1.5, it is worth recalling at this stage that phase noise is commonly measured as a spectral power ratio of sideband noise to the carrier, with the noise obtained from a single sideband, on a one Hertz bandwidth, and at a particular offset from the carrier. Phase noise is expected to increase with $f_c$ and decrease for increasing offsets. In principle, simple extrapolations can be performed from a known $f_c$ and an offset reading to obtain the expected phase noise at a different central frequency or offset. For example, phase noise is generally expected to increase directly with the square of the oscillator output frequency. In practice, however, a meaningful comparison requires knowledge of phase noise for a range of outputs and offsets.

As can be seen from Table 1.1, despite constant progress during the last decade, LC oscillators offer a limited tuning range. Among all LC designs listed, Refs. [11,18] are the only to provide a qualitatively linear region. Ref. [11] reports directly their linear tuning range of $f_c \pm 5.9\%$, while in Ref. [18] a linear range of $f_c \pm 9.5\%$ can be observed within their reported total tuning range of $f_c \pm 15.5\%$. In contrast, all other LC oscillator designs in Table 1.1 lack a significant linear region throughout their full tuning range. In fact, none of these designs is intended for wideband linear tuning; while low phase noise performance, on the other hand, is consistently targeted reaching an impressive $-133.5$dBc/Hz @ 600kHz in Ref. [19]. To avoid the restrictions of monolithic LC-oscillators, another approach consists of using ring-oscillator-based VCOs. These consist of a closed chain of identical inverters, with a frequency of oscillation determined by the total number of gates. These circuits do not use a resonant
network, and consequently result in higher levels of phase noise. Additionally they have lower attainable frequencies than LC-oscillators, in similar technologies, since their maximum frequency of oscillation is limited by the minimum number of gates, generally three, and by the minimum gate delay [8]. Performance values of typical monolithic ring oscillator designs are listed in Table 1.1. Compared to the LC type, the ring oscillators designs offer larger tuning ranges, but with higher phase noise; and like LC oscillators they exhibit poor to non-existing linear tuning. Designs reported in Refs. [20-22] show typical results with moderately low $f_c$'s and phase noise. More recent publications describe higher $f_c$'s and wider tuning ranges by means of various circuit optimizations [23-25]. However, regarding linear tuning, results for two oscillators in [24] are the only applicable attaining a qualitatively linear tuning of ±11.5% out of a total range of ±31.7 % and ±15.3% out of ±33.0%.

A third approach consists in the use of traditional relaxation oscillators, also called multivibrator, VCOs. These circuits employ regenerative switching between two internal voltage levels to produce a continuous charge and discharge cycle on a timing reactance, most commonly a capacitor. Due to their simple monolithic implementation and inherent linear/exponential behavior, multivibrators have been used extensively for frequencies up to hundreds of mega-Hertz; however, little effort has been devoted to exploiting their capabilities in the range of 1 to 10GHz [26, 30]. The lower section of Table 1.1 summarizes results for the most significant high-speed multivibrator VCOs reported to date.

Early work in Ref. [26] shows that wideband tuning and high frequency operation, ±99.9% up to 1GHz, are attainable employing a relaxation oscillator. Wideband linearity nonetheless was limited to the lower frequencies in a range of ±98MHz around 102MHz, and noise measurements were not reported. Soon after, work in Ref. [27] revealed that, for an
Chapter 1. Introduction

advanced silicon bipolar process of cutoff frequency $f_T$ of 30GHz, it is possible to achieve oscillations up to 7.4GHz with an emitter-coupled multivibrator oscillator. In particular, Ref. [27] presents results for VCOs with center frequencies of 1GHz to 5GHz with wide tuning ranges; for example for $f_c = 1.47$GHz the tuning range is 1.4GHz ($-41\%$ to $+56\%$ of $f_c$). Measurements of linearity and noise performance are not presented in [27], but an observation of increasing jitter with increasing $f_c$ is reported. The fully electronic WFM modulator [5] mentioned earlier was based on the emitter-coupled-multivibrator configuration. It was implemented in an advanced silicon bipolar process of cutoff frequency $f_T$ of 40GHz, and achieved a variable linear frequency range from 2.5GHz to 5.4GHz. Details on linearity and noise performance of the modulator are not discussed in [5], but as mentioned earlier in Section 1.3, pre-distortion was necessary to compensate for modulator second and third order nonlinearities, with overall poor results for 20-channel transmission. In terms of linearity, results in [28] are significant as they show wideband tuning of $\pm 95\%$ around 390MHz, with a linear range of $\pm 242$MHz around 262MHz. More recently efforts in Ref. [29] are characteristic of attempts to improve on the phase noise of multivibrators, but at the expense of tuning linearity. In summary, previous works show promise for wide linear tuning but poor phase noise performance. As mentioned at the beginning of this section, broadband tuning and low phase noise are conflicting requirements and for this reason multivibrators have received less attention than other types of oscillators. Moreover, the study of multivibrator phase noise is complicated by their highly nonlinear nature, requiring a different treatment from the well-known theory developed for LC oscillators. Nonetheless, meticulous studies of phase noise and jitter in multivibrators have been reported. A review of these studies and their conclusions is provided in Section 1.5.2.
1.5 Phase Noise

1.5.1 Fundamentals

To better understand the issues of oscillator phase noise prediction and measurement it is important to review some principles of phase noise theory. Accordingly, this section begins with a review of phase noise fundamentals. For a complete coverage of phase noise and frequency modulation theory the reader is referred to references [7,31-33].

In all oscillators, the random noise sources of the internal components cause small amplitude and frequency perturbations of the oscillator output. In the time domain, this results in small fluctuations of the signal period, referred to as timing jitter, and of the signal amplitude, referred to as amplitude noise. In the frequency domain, the perturbations result in combined amplitude and phase modulation sidebands around the oscillator carrier and its harmonics. In general, the amplitude fluctuations are assumed small enough to be neglected and the sidebands are exclusively considered as phase noise sidebands. In practice, however, this might not be the case and thus amplitude limiting is commonly used to remove unwanted AM noise components. Fig. 1.4 shows a representation of the power spectrum of a stable frequency reference [31]. The narrow peak section in Fig. 1.4 is usually referred to as the carrier portion of the signal spectrum and corresponds to the theoretically expected impulse spectrum of a noise free sinusoidal signal. The presence of noise is responsible for widening the base of the impulse creating sidebands on both sides of the carrier. Phase noise causes symmetrical sidebands of opposite phase, while amplitude noise causes symmetrical sidebands of equal phase; combined they result in asymmetrical sidebands on a spectrum analyzer. In Fig. 1.4 AM noise is negligible and only phase noise sidebands are shown. From this figure, the negative effect of phase noise in communication systems can easily be appreciated by considering the interference the sidebands could cause if overlapping on an adjacent carrier.
Phase noise sidebands as depicted in Fig. 1.4, are generally treated as caused by FM modulation of the oscillator by an equivalent white noise source, which represents the total noise at the oscillator output nodes. For calculation purposes, white noise can be modeled as composed of an infinite number of sinusoids, distributed in frequency, with equal amplitude, and random phases [34]. Thus, for linear systems, superposition of individual sinusoidal component effects is commonly used to obtain the total impact of white noise. Although frequency modulation is a nonlinear process, for very small modulation indexes, a linear approximation is valid. In such cases, the spectrum of single tone modulation is simply composed by the carrier, \( f_c \), and two sidebands at \( \pm f_m \) offsets, where \( f_m \) is the frequency of the modulating signal. From the Bessel coefficients describing the FM spectrum, it follows that the amplitude of the sidebands is directly proportional to the modulation index; thus, proportional to the noise amplitude and inversely proportional to \( f_m \). This is the case for FM noise.
modulation, where the total modulation index, accounting for all sinusoidal components, is much smaller than one. Therefore, by superposition, the phase noise sidebands of Fig. 1.4, are considered as composed of a large number of FM sidebands pairs, such as the power addition of their corresponding sinusoidal components equals the noise mean power of the equivalent white noise source modulating the carrier [31].

The previous considerations are the basis for the definition of the most commonly used oscillator noise measure. Single-sideband (SSB) phase noise, $L(f_m)$, is defined as the ratio of noise power in a one Hz bandwidth SSB component, to the total signal power; though in practice, the ratio of noise to carrier peak power is generally used for convenience. Fig. 1.4 illustrates a typical measurement of SSB phase noise. First, a SSB noise component is located at an offset $f_m$ from the carrier and its power is measured in a limited bandwidth $\Delta B$. Ideally $\Delta B$ is one Hz wide; however, in experimental measurements this might not be possible. Then the noise power in the region $\Delta B$ is assumed flat, and its total measured power is normalized to a per Hz basis. Finally, the SSB phase noise is obtained from the ratio of the noise power, per Hz, to the carrier peak power and is given in units of dBc/Hz at a specific offset, $f_m$, from a specific carrier, $f_c$.

Fig. 1.5 shows a typical plot of an oscillator's SSB phase noise, versus offset frequency. Three distinct regions are visible. At large offsets from the carrier $L(f_m)$ is dominated by white noise sources present at the oscillator output, or, for experimental data, by the noise floor of the test setup. The middle section with slope of $-20$ dB per decade is caused by FM modulation of the oscillator by white noise. This follows from the previous discussion, as sideband noise power is inversely proportional to the square of $f_m$. A third section with $-30$ dB per decade
Chapter 1. Introduction

slope is commonly observed, and is due to flicker noise sources in the oscillator, which, being inversely dependent of frequency contribute to the slope with an additional $-10\text{dB per decade.}$

![Graph showing phase noise](image)

Figure 1.5: Typical SSB phase noise, $L(f_m)$, of an oscillator versus offset frequency $f_m$.

There are different techniques to measure SSB phase noise. Simply, it can be obtained from the oscillator’s power spectrum, as in Fig. 1.4, with a spectrum analyzer. However, the sidebands measured this way may include unwanted AM noise components. In addition, low offset measurements are limited in this approach by the phase noise of the analyzer’s front-end local oscillator; furthermore, several measurement inaccuracies must be accounted for, and the practical offset span is limited by the analyzer noise floor. Therefore, alternative measurement approaches are preferred either in the presence of AM noise, low phase noise, or for close-in offsets. These approaches are based on the removal of the carrier component and of the AM fluctuations, and provide a direct reading of SSB phase noise spectrum such as shown in Fig. 1.5. Thus, the choice of the measurement technique depends heavily on practical requirements. But in general, regardless of the technique used, accurate phase noise measurement remains a
Chapter 1. Introduction

difficult task. This is reflected in the small selection range and high cost of commercially available solutions. An alternative, low cost, state-of-the-art, measurement technique based on a spectrum analyzer was developed for the requirements of this thesis. The use of this technique is made possible due to the moderate demands of multivibrator phase noise levels, and by the phase noise offsets of interest in WFM transmission. Nonetheless, the custom measurement technique increases considerably the accuracy and repeatability of the SSB phase noise measurements. This subject is further explained in Chapter 3.

1.5.2 Phase Noise in Multivibrator Oscillators

Several analytical models of SSB phase noise [33] have been developed to account for measured behaviors such as shown in Fig. 1.5. Notably, a large body of work has been reported for LC oscillators. Nonetheless, the issue remains an active area of research for all oscillator configurations. As this thesis deals with the topic of multivibrator oscillator phase noise, we proceed in this section with a short review of reported studies in that area.

An early attempt to predict timing perturbations in relaxation oscillators was presented in Ref. [30] and was successful in identifying internal circuit contributions to the total noise at the output nodes as well as provided a general expression for the calculation of jitter. However, from the results in [30] it is not easy to predict performance in terms of phase noise because time correlations are not included in the calculations. Further work reported in Refs. [35, 36] studies the origins of multivibrator phase noise and provides analytical tools to predict noise sidebands as measured with a spectrum analyzer. These analyses are based on defining equivalent voltage and current noise sources in the vicinity of the multivibrator timing capacitor.
Figure 1.6: Noise mechanisms leading to phase noise in ECMs. (a) Block diagram of a multivibrator oscillator with equivalent current and voltage noise sources. (b) Effects of the equivalent noise sources on the capacitor voltage.

Fig. 1.6 illustrates the noise mechanisms leading to phase noise in a multivibrator oscillator as described in Refs. [30,35-36]. A triangular periodic voltage is created across the timing capacitor $C$, by continuously charging and discharging the capacitor between two threshold levels, $V_{\text{trigger1}}$ and $V_{\text{trigger2}}$, with a constant current source. The charge and discharge are achieved by inverting the polarity of the current source $I$ with a Schmitt trigger circuit. The equivalent noise sources are indicated in Fig. 1.6 (a) by $V_n$ and $I_n$. These sources affect the output timing in two different manners. As shown in Fig. 1.6 (b), $I_n$ modifies the slope of the capacitor ramp, thus changing the timing of the signal; the magnitude and rate of the slope change is set by the magnitude and frequency of $I_n$. Since $I_n$ is composed of a large number of different frequency components many scenarios are possible depending on the relative magnitudes of the noise and oscillator frequencies. $V_n$, on the other hand, affects the threshold levels thus delaying or advancing the switching of the current source, which in turn affects the timing of the signal. Again, several scenarios are possible for different frequency components of $V_n$. From the analysis of all possible scenarios, and their impact on phase noise, it is
concluded in Refs. [35,36] that noise at low and high frequencies contribute to phase noise in different ways. The contribution from low frequency noise is dominated by $I_n$ components, and in practical ECM circuits is attributed directly to thermal noise from resistors used in the implementation of the charging current sources. High frequency noise contributions on the other hand are dominated by $V_n$ components, which, through a noise folding mechanism caused by the sampling action of the oscillator switching, are down-converted to lower frequencies and dominate the overall phase noise. These conclusions are experimentally confirmed in [36] for a relatively low frequency, $f_c = 110$MHz, design. More recently, Ref. [37] presented a similar theoretical analysis for a very low frequency, $f_c = 1.1$MHz, relaxation oscillator, and again, confirmed dominance of the noise folding effect from fluctuations of the threshold switching levels. In addition, a systematic experimental proof of these results is provided in Ref. [37]. The work presented in [35, 36] was further advanced in [7] by devising a model for the oscillator sampling transfer function such as to compute directly the effects of the noise equivalent sources in different oscillation parameters, such as phase noise, signal period, and signal duty cycle. In addition, the number of equivalent noise sources is expanded in Ref. [7] from two to four by including correlated and uncorrelated voltage and current sources. These sources reflect correlation with the switching states of the oscillator. Accounting for all four cases, again, a dominance of voltage noise folding is theoretically found.

Overall, previous works have been successful explaining the excessive noise observed in relatively low frequency, kilo-Hertz to hundreds of mega-Hertz, designs. Based on these conclusions several improved circuits were designed in [22,29,30,36] to reduce phase noise. Most of these, however, operate at sub GHz frequencies and all involve architectural modifications to the multivibrators at the expense of tuning linearity.
Chapter 1. Introduction

1.6 Outline of the Thesis

This thesis has six chapters. In Chapter 1, “Introduction”, we provided background information for the thesis. In Chapter 2, “Ultralinear WFM Circuits”, we describe the design and experimental testing of ultra-linear high-speed wideband modulator and demodulator circuits. In particular, we present designs for a current-controlled ECM, a linearized voltage-controlled ECM, a delay line FM demodulator, and a back-to-back modulator/demodulator. Experimental and expected results are compared showing record performance for WFM systems.

Optimizations to the modulator required a systematic study of phase noise in ECMs. For this reason a custom phase noise measurement setup was developed as described in Chapter 3, “Phase Noise Measurement Technique”.

In Chapter 4 “Noise Limits in High-Speed ECMs”, phase noise measurements are presented for several versions of the current-controlled ECM running at different frequencies. Results are then compared with state-of-the-art noise simulations. The challenges confronted to obtain accurate simulation results are explained with a brief discussion on simulation requirements of time-variant circuits. From results, it is shown that, through frequency folding, phase noise of high-speed ECMs is dominated by internal shot noise sources. From this, design optimizations are proposed to improve oscillator noise performance without affecting tuning linearity.

In Chapter 5, “System Perspectives”, application perspectives in WFM transmission for the linearized modulator are explored. This is done numerically to evaluate system CNR, CSO and CTB from the oscillator static tuning response. System parameters are chosen for narrowcast WFM transmission. Finally, we present conclusions in Chapter 6.
Chapter 2
Ultralinear WFM Circuits

2.1 Introduction to Chapter
As explained in Chapter 1 the primary goal of this thesis is the design, implementation, and study of fully electronic circuits for wideband FM transmission. The focus is limited exclusively to circuits, and the impact of transmission over real fiber with practical optoelectronic transmitters and receivers is beyond the scope of this work. To this purpose, we implemented proof of principle designs to determine practical performance limitations and the feasibility of a design approach. The successful implementation of these circuits validated the electronic implementation approach, provided modulator and demodulator devices for future experimental testing of the WFM system, and holds promise for a commercial implementation of electronic-based WFM systems.

In this chapter, we present design considerations and experimental results for high-speed, ultralinear wideband modulator and demodulator circuits. All designs are fully electronic, monolithic, and implemented in an advanced silicon bipolar process with cutoff frequency of 25GHz. All circuits were tested on bare dice using multi-contact wedge probes. Details are presented as follows. In Section 2.2, we describe design goals and WFM system considerations leading to the selection of ECM-based modulator and delay-line based demodulator circuits. Section 2.3 covers design specifics for an ECM-based current-controlled oscillator (ICO), a linearized ECM-based VCO, a delay-line demodulator, and a back-to-back modulator/demodulator. The ICO was implemented for studies of the intrinsic limitations of the ECM core, while the VCO is intended for WFM transmission. The demodulator was
implemented for testing of the VCO and based on previously reported hybrid designs presented in [3, 4]; with it, the back-to-back linearized VCO and demodulator was built to measure end-to-end response. For each of these circuits simulated and experimental results are presented throughout Section 2.4. The chapter is closed in Section 2.5 with a summary and conclusions.

2.2 Design Goals and Considerations

Design goals were set for a narrowcast WFM transmission system of 80 CATV channels and 0dBm received power levels. As per HFC standards, system performance targets are CNR of at least 52dB, and CSO and CTB better than -65dBc. These system specifics have several implications on the performance of the modulator. First, the selected level of received power, 0dBm, reduces the relative importance of receiver shot noise, making the system noise performance largely determined by the modulator phase noise. Second, the target distortion levels require, from the modulator, a highly linear frequency versus voltage response. Considering the poor linearity of previously reported designs, this is clearly a serious design challenge. Third, practical system considerations for 80-channel WFM transmission set particular bandwidth requirements that define the oscillator central frequency and tuning bandwidth. Increasing the system FM bandwidth improves the CNR but also increases CSO and CTB levels [38]; therefore, a compromise must be made. A total FM modulation index near 1 and a carrier frequency, $f_c$, close to 2GHz were chosen based on the observations presented in Ref. [3] and WFM system calculations [38]. Applying Carson's rule, the estimated FM bandwidth for an 80-channel modulating signal on a 2GHz carrier and a modulation index of 1 results in a total bandwidth of 2.2GHz, with a maximum frequency deviation, $\Delta f$, of 565MHz. As will be described in Section 2.3 a practical input signal swing is 0.2Vp. Thus, for
an input modulating signal of 0.2Vp, the voltage to frequency conversion sensitivity is 2.8MHz/mV. Finally, the dynamic range of the modulator must be large enough to accommodate the tuning band defined by maximum deviations of ±565MHz around a carrier frequency of 2GHz; this represents a variation of ±30% of $f_c$.

From the previous discussion it follows that the wideband FM modulator is essentially a VCO with requirements for (a) very wide tuning range, (b) modulation bandwidth exceeding 565MHz, (c) highly linear tuning characteristic (d) low phase noise. As seen in Chapter 1 the large body of work on VCOs for wireless communication and clock synthesis/recovery describes circuits poorly adapted to this application. In particular, the first three requirements cannot be met by conventional LC-tuned VCOs. An emitter-coupled multivibrator (ECM) is a natural choice to meet the linearity requirement because of its current-based tuning characteristic, involving only a capacitive timing component. Furthermore, the broadband nature of the ECM also provides the wide tuning range required. However, the ECM relatively poor phase noise represents a drawback. Nonetheless, in wideband FM transmission, the lower offsets phase noise is much less of an issue. The reason for this is that the CATV band starts at approximately 55MHz; this means that low frequency, close-in offset, phase noise present in the VCO, and appearing below 55MHz after demodulation at the receiver, is out of band and therefore of no concern. In addition 55MHz is well above the 30dB per decade slope portion of the phase noise, as shown in Ref. [36] for phase noise measurements of a multivibrator, where the transition between the 30 and 20dB per decade slopes is at 1kHz. The higher frequency phase noise occurring between 55 and 565MHz, although still of concern, can be dealt with through careful circuit design; this is discussed in detail in Chapter 4. The application considered here is in sharp contrast with the low, close-in phase noise requirements of wireless and narrow-band clock recovery circuits, for which the ECM is not suitable.
Fig. 2.1 shows the simplified schematic of an ECM. The functioning of the circuit is based on the same principles described in Section 1.5.2. while discussing the block diagram of a multivibrator oscillator presented in Fig. 1.6. In Fig. 2.1, the top cross coupled transistors ensure that the transistors in the differential pair alternate between on and off states, in this manner the capacitor is periodically charged and discharged with an alternating flow of current, $I$, from the two tail current sources. The state of the differential pair is switched when the voltage of the capacitor reaches a fixed threshold, $\pm V_s$, which is set by the clamping action of the load diodes. To describe circuit operation, the classic simple approach, assumes instantaneous transistor switching and constant tail currents $I$, leading to the following expression for the oscillation frequency $f_c$:

$$f_c = \frac{I}{4 \cdot C \cdot V_s}$$

Thus the oscillation frequency of an ECM is directly proportional to $I$, and inversely proportional to the timing capacitor $C$ and the forward voltage drop $V_s$ of diodes $D_1$ and $D_2$. Therefore, control of the DC tail current provides control of the oscillator frequency. A more precise analysis presented in [39] reveals that the effect of transistor parameters on the output frequency is dependent on the value of the load resistors $R_l$. Furthermore, it shows that, if the currents through the load resistors and the diodes are made comparable, during the conducting stage of the load diodes, then the output frequency is predicted by the results of the classic simple analysis. This outlines the intrinsic linear nature of the ECM and indicates the need to design a linear tail current control circuit.
For the implementation of a monolithic demodulator we followed a previous successful design reported in Refs. [3,4] for WFM transmission. This demodulator is based on a delay-line detection scheme, which functioning is described in Section 2.3.3, and was previously implemented in Ref. [3,4] using discrete components and distributed transmission lines. The delay line demodulator achieved a qualitatively linear range from DC to 3.5GHz, which is ideal to match the linear range of our modulator design. This type of circuit is, however, convenient for monolithic implementation since it is composed of digital cells and a delay line that can be implemented in a small area using a cascade of buffer gates. For these reasons a monolithic silicon bipolar version, $f_T = 40$GHz, of the same circuit was reported in Ref. [5] for application in WFM transmission. The range of this circuit is reported to be 5.5GHz, but its linearity is not described.
Chapter 2. Ultralinear WFM Circuits

As for the implementation technology, both ECM and the delay-line modulator can be realized with high-speed bipolar designs. Constant improvements in silicon bipolar technologies have made possible silicon bipolar circuits for application speeds well into the giga-Hertz range. Careful circuit design considering transistor sizing, differential operation, low voltage swings and circuit optimization has allowed the demonstration of digital bipolar circuits with clock frequencies up to the transistor cutoff frequency, $f_T$, or broadband amplifiers with cutoff frequencies of $0.5f_T$ [40]. Based on these capabilities of Si bipolar integration and on previously reported high-speed VCO designs, we selected Nortel's NT25 advanced silicon bipolar process, with an $f_T$ of 25GHz, as the target process. This process is a self-aligned, double-poly, implanted base, silicon bipolar technology, developed in the years of 1996-97 for RF applications operating in the range of 1GHz to 3GHz [41]. This technology is particularly attractive for high-speed design as it targets circuit optimization through transistor sizing. To this purpose NT25 incorporated novel transistor modeling features for circuit simulation. It provides scalable transistor models based on a geometry and process scalable variant of the SPICE Gummel-Poon (SGP) bipolar transistor model. The scalability is based on equations previously developed for the HICUM transistor model, and is fully described in Refs. [42-43]. In addition, the models also include novel scalable high-frequency noise parameter equations for bipolar transistors, that were originally presented, in Ref. [43], by developers of the NT25 process. These features provided a powerful simulation tool that proved invaluable in their accurate prediction of our circuit designs. A detailed discussion on the final circuit implementations follows.
2.3 Circuit Design

2.3.1 Simple Current-Controlled Oscillator

For studies of tuning and linearity, a simple ECM driven by a current mirror was implemented. Fig. 2.2 (a) shows the block diagram of the simple ECM-based ICO design. All transistors used are of the same emitter width and length, W=0.5μm, L=5.0μm. Transistor sizing was selected such as to optimize cutoff frequency versus collector current density [40], for collector currents of 1mA. For convenience this same transistor is used throughout all other blocks designed as no smaller sizes were required, and optimizing power consumption was not a goal. The timing capacitor C is 150fF for a targeted central frequency of 1.7GHz at a 1mA tail current. Particular attention was paid to the ECM core collector loads with best linearity achieved when the diode and resistor currents are comparable. As shown in Fig. 2.2 (a) the load diodes, D1-D2, are implemented with diode-connected transistors. Emitter resistors R2 were selected to reduce the collector current density of transistors Q3 and Q4 from the optimum value. This was done to reduce ringing at the output of the emitter followers, caused by their low inductive output impedance and the input capacitance of the following stages [40]. Current mirror transistors are identical and the input Iin is for an external, off-chip, current source.

The differential square output of the ECM, q+ q−, is connected to a limiting amplifier (LIM) followed by an output driver (DRV); these are shown in detail in Figs.2.2 (b) and (c). The limiter consists of a cross-coupled pair of differential amplifiers used to cancel capacitive feedthrough of the ECMs high-speed transitions. This block plays an important role in obtaining a constant-amplitude output while the ECM core is tuned. In principle, the magnitude of the ECM output is constant and twice the forward voltage of the load diodes. In practice variations in the output magnitude occur with changing output frequency due to the finite
Chapter 2. Ultralinear WFM Circuits

Figure 2.2: Implemented ICO. (a) Diagram of simple current-controlled oscillator, showing ECM core details and output stage blocks. (b) Schematic of Limiting Amplifier "LIM". (c) Schematic of Output Driver, "DRV". Outputs are open collector and drive external 50Ω loads.
transistor switching times [28]. These amplitude variations are effectively removed by the limiting stage. On the other hand, the use of limiting does not help remove the phase noise component of the VCO output signal, and under modulation may introduce AM-to-PM conversion. This conversion effect is minimized by the use of fully differential bipolar circuits, as demonstrated in Ref. [44] for a 30GHz limiting amplifier. As explained in Ref. [44], in bipolar limiting amplifiers, the nonlinearity of the base-emitter input capacitance is responsible for amplitude dependent phase variations of the output; under differential operation, the phase variations at the two outputs are of opposite sign and therefore cancel.

The limiting circuit also performs the task of converting the large output swing of the ECM to a lower value. The use of transistor base-emitter junctions as clamping diodes provides a large voltage swing across the timing capacitor, approximately 1.8Vp-p, which in principle helps reduce phase noise [30]. However, for Si-bipolar high-speed ICs the use of current mode logic gates (CML) is convenient and optimal values for differential peak-to-peak voltages are 400mVp-p [40]. For this reason, the limiting amplifier is also used to reduce the output of the ECM to 400mVp-p differential. Additionally, the limiting amplifier acts as an intermediate buffer for the higher current output driver stage.

The output driver is a simple differential pair with open collectors to drive matched 50Ω impedance lines; it was designed to handle 4mA on each collector such as to produce 200mV swings on 50Ω loads; 400mVp-p differential. Arrays of four parallel transistors are used to handle the corresponding currents and the layout is optimized to reduce parasitic capacitances. An external current source load resistor $R_{drv}$ provides control of the output amplitude. The voltage power supply $V_{ee}$ is $-4.5\text{V}$. 
A compact and fully symmetrical physical layout was used to implement all blocks of the ICO. This was done for two reasons; first, to improve immunity to common mode noise and signal pickup in the differential circuit blocks, such as the limiter and output driver; second to minimize fabrication mismatches between circuit components. Notice that despite the symmetrical appearance of the ECM its operation is essentially asymmetrical due to the switching action of the transistors. The ICOs tuning and linearity results are discussed in Section 2.4.1.

2.3.2 Linearized Voltage-Controlled Oscillator

For WFM system tests we designed a linearized voltage-controlled version of the ECM modulator. This VCO contains the same ECM core, limiting amplifier and output driver as the ICO previously described, but replaces the current mirror circuitry with a linear transadmittance amplifier ($Y$) shown in Fig. 2.3. The $Y$ block provides 50Ω input impedance matching for the external RF, 55-565MHz, modulating signal ($R_{Fin}$) and linear voltage to current conversion to drive the two tails of the ECM core with identical currents ($I_1$ and $I_2$). The circuit consists of a single differential pair stabilized with negative feedback and single pole compensation, for a total 3dB bandwidth of 1GHz. The feedback is of the type “series-series” in the sense that it samples the output current and feeds back a series input voltage; this type of circuit provides a very stable transadmittance as well as high input and output impedance. The transadmittance transfer function provides a linear output in the range of 0.75mA to 1.25mA for a single-ended input range of ±200mV. The resulting central frequency of the VCO is 1.8GHz; $V_{ee}$ is −4.5V. Again, a fully symmetrical layout was employed for the same reasons described in Section 2.3.1. For the asymmetrical transadmittance amplifier,
dummy loads were used to maintain layout symmetry when necessary. Complete static tuning and dynamic tests were carried on the linearized VCO, these are described in detail in Section 2.4.2.

Figure 2.3: Schematic of linear transadmittance amplifier “Y”. \textit{RFin}, \textit{I1} and \textit{I2} are the RF voltage input and the two output currents respectively.
2.3.3 Delay-Line Demodulator and Back-to-Back Modulator/Demodulator

A wideband demodulator was also implemented for WFM system tests. A block diagram of the demodulator is shown in Fig. 2.4 (a). The demodulator consists of a delay-line XOR-based discriminator [3,4]. In this circuit, the incoming signal is first limited and buffered by two differential input buffers, then combined with a delayed version of itself by an XOR gate, and finally externally low-pass-filtered. The resulting output is linearly proportional to the input frequency from DC to a frequency equal to $(2\tau)^{-1}$, where $\tau$ is the differential delay between the two XORed paths. The circuit was implemented with fully differential current-mode-logic buffers and gates. Fig. 2.4 (b) is the schematic of the buffer cell used, consisting of a simple differential pair with inputs $AT, AF$, and outputs $QT, QF$. The differential delay $\tau$ was chosen to be approximately 80ps, achieved by a cascade of five buffer elements. This provides a linear demodulation range centered at approximately 2GHz, to match the modulator center frequency.

The schematic of the XOR gate is shown in Fig. 2.4 (c), $AT, AF$, and $BT, BF$ are the differential gate inputs. The outputs $QT, QF$ are open collector to drive external 50Ω loads for 200mVp-p differential swings. Again, transistor sizing was used to optimize cutoff frequency with collector current. All transistors are of the same dimension, $W=0.5\mu m, L=5.0\mu m$, to optimize cutoff frequency for collector currents of 1mA, except for the transistors in the bottom differential pair of the XOR gate which have dimensions of $W=0.5\mu m, L=10.0\mu m$, to drive 2mA currents.

A back-to-back modulator demodulator was also implemented for end-to-end tests. It consists of the linearized VCO described in Section 2.3.2, without the output driver block, followed by the delay-line demodulator described here. Again, layout is symmetric and $V_{ce}$ is -4.5V. Experimental results for these circuits are described in Section 2.4.3.
Figure 2.4: Delay-line demodulator. (a) Diagram of the delay-line frequency modulator. (b) Buffer block used to implement the delay line. (c) XOR gate, outputs are open collector to drive external 50Ω loads.
2.4 Predicted and Experimental Results

2.4.1 Simple ICO

The static tuning response for the simple ICO is shown in Fig. 2.5 (a). Open circles are predicted behavior from circuit simulation and show linear operation over a wide range from 1 to 2.5GHz. Closed circles are measurement results collected with a spectrum analyzer for static input currents and are in excellent agreement with predictions. Current starving of the ECM load diodes sets the lowest limit of the input sweep, while the highest limit is set by the maximum power dissipation chosen for on-chip resistors. The central oscillation frequency, defined at 1mA input, is 1.7GHz. To quantify nonlinearity in the ECM response we performed a linear fit of the tuning data constrained to the point of central frequency; this is shown for measured data by the solid line in Fig. 2.5 (a). We then calculated the difference of each point to the linear fit as shown in Fig. 2.5 (b). Open circles are from predicted response and closed circles from measurements; the maximum absolute deviation from linearity is 15MHz and the qualitative shape of the experimental curve is very well reproduced by the design calculations. Additional simulations, replacing the tail current mirror by ideal current sources, show that this effect is intrinsic to the ECM core. The source of this behavior is analyzed in more detail in Section 2.4.2, when comparing the deviations from linearity of the ICO and VCO designs.
Figure 2.5: Static response of simple ICO. (a) Tuning. (b) Deviation from linearity. Closed (open) symbols are measurements (simulations).
2.4.2 Linearized VCO

The total circuit size of the linearized VCO is approximately $150 \times 400 \mu m^2$ and the power consumption is $150 mW$ from a $-4.5 V$ supply. Measured $f_c$ at a $0 V$ input is $1.8 GHz$. VCO static tuning response and deviation from linearity are presented in Fig. 2.6. Open circles are predictions and closed symbols are measurements. Results show linear operation in the range of 1.5 to $2.5 GHz$. As mentioned in Section 1.4 the widest linear range previously reported is 484MHz centered at 262MHz in Ref. [28]. In comparison, the linearized VCO achieves a 1GHz linear tuning range centered at 1.8GHz, this represents an improvement of a factor of 2 in range and 6.8 in center frequency.

The tuning range of the linearized VCO is smaller than that measured for the simple ICO since the inputs of the input transadmittance amplifier are limited to $\pm 200 mV$, covering only a fraction of the current tail sweep applied to the simple ICO in Fig2.5. Measured performance is in good agreement with simulation, and over the 1GHz tuning range the largest deviation is less than $1.5 MHz$; thus, proving the suitability of the linearization feedback, which provides an improvement in deviation of a factor of 10 over that of the simple ICO. An analysis of these linearity results follows.

From circuit simulations of the ICO, the multivibrator switching delay is approximately 16ps. At GHz operation this delay represents a non-negligible portion of the ECM oscillation period. Assuming a constant switching delay, $t_{sw}$, along the total tuning range, the effect of this delay on the output frequency can be expressed as:

$$f_c' = \frac{1}{T + 2 \cdot t_{sw}} = \frac{1}{I_c \cdot T_c + 2 \cdot t_{sw}}$$

(2.2)
Figure 2.6: Static response of the linearized VCO. (a) Tuning. (b) Deviation from linearity. Closed (open) symbols are measurements (simulations).
where $T$ represents the ECM oscillation period under instant switching conditions as expressed in equation 2.1, $I_C$ and $T_C$ are the input current and output period when the ICO is operating at the central frequency, and $I$ is the input tuning current. From this equation it is possible to generate plots of tuning response and deviation from linearity similar to those presented earlier for the ICO and VCO. In Fig. 2.7 we compare the results of such calculations with the experimental results of the ICO. In Fig 2.7 (a) we reproduce the ICOs measured tuning response and linear fit constrained to the point of central frequency. In Fig. 2.7 (b) the closed symbols are the ICOs deviation from linearity, and the solid line is the deviation from linearity as predicted by equation (2.2) for a linear fit constrained to the point of central frequency. In these calculations $t_{sw}=16\text{ps}$, $I_C=1\text{mA}$, $T_C=588\text{ps}$.

The results show a good match between measurements and calculations in the range of approximately 0.7mA to 1.4mA, suggesting that the ICO linearity is near the fundamental limit set by the multivibrator switching delay. The deviations from the predicted behavior at the lower currents and above 1.4mA can be explained as follows. At low input currents the two ECM load diodes are current starved and thus fail in their clamping action. These diodes are necessary to turn what would otherwise be a fixed frequency multivibrator oscillator into a tunable oscillator. The convergence to a fixed frequency at low currents is responsible for the upward curvature in the deviation from linearity. This behavior is consistently predicted by the ICO simulations shown in Fig. 2.5. On the other hand, above 1.4mA inputs, the ICO measured deviations diverge from both, predictions based on a finite switching delay (Fig. 2.7), as well as circuit simulations (Fig. 2.5). For this reason we suspect that transistor thermal effects, which are not accounted for in the simulation, may be playing a role. As an example of this,
Figure 2.7: Effect of switching delay on linearity of simple ICO. (a) Measured tuning response of ICO, and linear fit of the experimental data. (b) Deviation from linearity. Closed symbols are from measurements; solid line is from theoretical prediction accounting for a constant switching delay in the multivibrator. The tuning inputs are normalized with respect to 1mA.
consider the reduction of \(-2\text{mV/°C}\) of the load diode voltages, which in turn reduces the capacitor voltage swing and increases the ECM oscillation frequency. Experimentally, it was observed that as the tuning current is increased above 1.5mA the ICO exhibits an unstable behavior with an increased tendency to drift towards higher frequencies, thus suggesting that transistor self-heating effects may be playing a role.

Considering the fundamental limit on ECM linearity discussed above, it is surprising that the VCO would exhibit lesser deviations than the ICO. This fact suggests that non-linearity of the transadmittance amplifier tuning response may be compensating the non-linearity of the ECM core. To explore this further, we show on Fig. 2.8 (a) the simulated tuning response of the transadmittance amplifier (Y block) with its linear fit constrained to the tuning point of central frequency. Fig. 2.8 (b) is the resulting deviation from linearity for the Y block, and (c) is the frequency deviations that would result at the ECM output, simply considering the slope of the ICO measured tuning response. These results show a positive contribution to the linearity of the ECM, but their magnitude is too small to account for the measured behavior of the VCO. From these observations it can be concluded that although the static linearity analysis provides useful insight on the intrinsic ECM core performance it is far too simple to explain the improved linearity of the VCO, which may be caused by a more complex, dynamic interaction between the tail circuit and the ECM core.

In Fig. 2.9 we show the time domain single-ended output of the VCO at 1.8GHz. The measurement was obtained using a 20GHz sampling oscilloscope with an input attenuation of 6dB. Scales are 200ps/div and 20mV/div; accounting for the input attenuation, the corresponding differential amplitude is approximately 360mVp-p at the circuit output. The signal is fairly square and exhibits low ringing, confirming the proper functioning of the circuit output driver.
Figure 2.8: Effect of transadmittance amplifier (Y block) response on deviation from linearity. (a) Simulated tuning response of Y block, and linear fit of simulation results. (b) Deviation from linearity of simulated Y block response. (c) Calculated resulting frequency deviations at the output of the ECM core.
Figure 2.9: Linearized VCO, time domain single-ended output at 1.8GHz with 6dB attenuation. Scale: 200ps/div, 20mV/div. From this, total differential amplitude is 360mVp-p.
Performance under modulation was measured from carrier-null tests as shown in Fig. 2.10. This is a common technique to measure the frequency deviation of a FM modulator and is based on mathematical principles of FM modulation. The amplitudes of all FM spectral sidebands, and the carrier, are Bessel functions of the modulation index. Each modulation index results in a particular output spectral pattern, among which some have a carrier spectral component of zero amplitude and are called carrier nulls. Since the modulation index is the ratio of the maximum frequency deviation to the modulating frequency it is possible to search for carrier null patterns by sweeping the modulating frequency. For example the first carrier occurs at a modulation index of 2.4048 and the second at 5.5201. To perform this test a single-tone of 400mVp-p was used. Fig. 2.10 (a) and (b) show the spectra of the first and second carrier nulls which occur at input frequencies of 170MHz and 76MHz respectively. From these results the VCO frequency deviation is 414MHz while the expected value is 423MHz, thus confirming FM performance and the quality of the input signal matching. In Fig. 2.10 (a) and (b) the input modulating frequencies are reflected in the spectra by the frequency offset between any two peaks; this is indicated in Fig. 2.10 (a) by the location of the markers. The additional spikes seen at the higher end of both plots are caused by harmonics of the VCOs non-sinusoidal output. Both plots in Fig. 2.10 also show spectral asymmetry, which indicates the presence of combined AM and FM modulation and which we attribute to frequency dependence of the limiter and output driver.

\[\text{Notice that location of the null requires judgment from the operator, for this reason the disparity between our expected and measured deviations can simply be attributed to an error of } 4\% \text{ in the selection of the input frequency. Nonetheless the measurements are particularly useful in the sense that they provide insight on the product of the effective input voltage and the dynamic voltage-to-frequency gain. As an example of the practical value of this information, it was observed that a mismatch of the input impedance, caused by wear-off of the input pads, can significantly affect the effective input voltage, to the point of reducing the frequency deviations by as much as half of the expected value.}\]
Figure 2.10: ECM-VCO dynamic response. (a) first and (b) second carrier null tests, single ended output. Nulls occur for input modulating frequencies of 170MHz (a) and 76MHz (b).
2.4.3 Demodulator and Back-to-Back Modulator/Demodulator

Total circuit size of the demodulator is approximately $150 \times 450 \mu m^2$, and it dissipates 135mW at a supply voltage of $-4.5V$. The static tuning characteristic of the demodulator was measured using a single-tone source and an oscilloscope. A differential splitter and appropriate DC level shifting were used to create the appropriate differential inputs. The average differential output is shown in Fig. 2.11 as a function of the input frequency; open circles are from predicted results and closed symbols are from measurements. The response is in good agreement with that expected from the design calculations, although the effective delay $\tau$ of 110ps is slightly greater than the design value, resulting in a steeper slope. Measured linearity is excellent in the range from 1 to approximately 3.5GHz, achieving a wider linear range than previously reported hybrid wideband FM demodulators [3,4]. This is 2GHz less than the linear range of the best silicon wideband FM demodulator reported in [5], which was, however, implemented in a process with an $f_T$ of 40GHz. The deviation from linearity in the present design occurring roughly above 3.5GHz is consistent with the triangular response of output versus input frequency expected for a delay-line discriminator, when considering the finite switching times of the CML gates. Note that for this type of circuit, additional distortion may occur if the end-to-end demodulator group delay is not flat. This could occur if the switching delays of the XOR and buffer elements vary with dynamically changing input frequency. These frequency dependent effects are not considered by the measurements presented in this section and remain a topic for further study.
Figure 2.11: Demodulator differential static-response. Closed (open) symbols are measurements (simulations).
Fig. 2.12 is a microphotograph of the back-to-back linearized VCO, and delay-line demodulator. Text labels on the image indicate the location of different circuit blocks and I/O nodes.

Figure 2.12: Microphotograph of back-to-back modulator/demodulator. Labels show modulator’s transadmittance amplifier "Y", "ECM" core, limiting amplifier "LIM", and demodulator’s "delay line" and "XOR" gate. RF input is "Vin"; “Q+,Q-” are the differential open-collector outputs.

This circuit was implemented to verify end-to-end modulation bandwidth performance. For this purpose an input tone of 200mVp was applied and the magnitude of the single-ended output Q+ was measured versus input frequency. The result is shown in Fig. 2.13, and was obtained using a spectrum analyzer to achieve accurate results in the presence of out-of-band carrier feedthrough occurring in the demodulator. The response is monotonic with a 3dB bandwidth of approximately 1GHz, which equals the design value. These results validate the high-frequency operation and input matching of the linearized transadmittance amplifier, the high-bandwidth tunability of the ECM core, and the dynamic operation of the demodulator.
Figure 2.13: Measured back-to-back modulator/demodulator response, single-ended output.
Chapter 2. Ultralinear WFM Circuits

2.5 Summary and Conclusions

In this chapter, Silicon bipolar designs and experimental results for wideband ultralinear modulator and demodulator circuits were presented. Results confirm the viability of the electronic implementation approach and revealed the suitability of the ECM-based modulator design, surpassing the performance of previously reported WFM modulators. Static experimental results for a simple ECM-based ICO proved possible linear operation in a range of 1.5GHz with an $f_c$ of 1.7GHz and a maximum deviation from linearity of 15MHz. Also, a WFM ECM-based VCO modulator was achieved with a 1.8GHz center frequency and a 1GHz tuning range, while maintaining record linearity. To our knowledge, this modulator is much more linear over a wider operating range than any other design reported to date. Feedback linearization of the VCO proved to be effective, reducing deviation from linearity by a factor of 10 with respect to the simple ICO. A static linearity analysis of the results indicates that the ICO linearity is near the fundamental limit set by the finite multivibrator switching delay. A similar analysis proved inadequate to explain the improved linearity of the VCO, which may be caused by a more complex dynamic interaction between the transadmittance amplifier and the ECM core. Modulation tests confirmed the dynamic operation of the modulator.

A WFM demodulator was also implemented. Experimental operation exhibits excellent linearity in a range of approximately 2.5GHz. Back-to-back testing of the linearized VCO and demodulator show end-to-end bandwidth of 1GHz, as designed. These constitute the first back-to-back measurements of an integrated modulator and delay-line demodulator reported to date.
Chapter 3
Phase Noise Measurement Technique

3.1 Introduction to Chapter

Preliminary phase noise measurements of the simple ICO and linearized VCO resulted in
-128dBc/Hz at 50MHz from a 1.7GHz carrier, and -122.5dBc/Hz at 50MHz from a 1.8GHz
carrier respectively. Both of these results surpass the performance of the best-reported WFM
optoelectronic modulator [3], which is -118.5dBc/Hz at 50MHz from a 1.7GHz carrier.
However, WFM system simulations for 80 and 40-channel transmission, described in detail in
Chapter 5, revealed that in order to achieve the CNR required by CATV standards, the
linearized VCOs phase noise needs to be, at least, 10dB lower. Unlike previous works
described in Chapter 1, it is imperative to improve phase noise performance without sacrificing
linearity. This situation led us to a systematic analysis of the origins of phase noise in high­
speed ECMs, for which we developed the custom measurement technique that is the object of
this chapter.

Measuring phase noise accurately is not an easy task. Therefore, a substantial effort was
invested in the development and validation of the setup described here. The result is a low cost,
automated, state-of-the art solution based on a portable spectrum analyzer, a high-precision
controlled current source, and mathematical post-processing. In Section 3.2 we explain the
practical limitations of spectrum analyzer-based phase noise measurements and their impact in
our approach. A complete description of the developed custom setup is given in Section 3.3,
followed by results in Section 3.4. The chapter closes with a summary and conclusions in
Section 3.5.
3.2 Measuring Phase Noise with a Spectrum Analyzer

In Section 1.5.1 we explained how phase noise can be measured directly using a spectrum analyzer. This approach offers a simple, flexible, and economical solution based on an instrument of common use and availability. However, this approach suffers from serious limitations that motivate the existence of alternate measurement techniques. Nonetheless, depending on the application, these limitations can either be neglected or properly dealt with. This course of action is extremely attractive when considering the cost of commercial measurement solutions. For example, a specialized commercial phase noise dedicated instrument operating up to 6GHz, Agilent model E5502B, has a cost of six figures. The sophistication of such instrument resides in its ability to perform measurements from extremely close-in offsets, 0.01Hz, to a maximum of 100MHz, and accuracy is limited to ±4dB above 4MHz offsets. In contrast our experimental requirements are to achieve low-cost phase noise measurements, at fixed 5MHz offsets, and with the best possible accuracy. In addition, automation of the measurement procedure is required to collect phase noise data versus varying tuning input. The frequency range of the tests is from 0.5 to 3.0GHz. This is to accommodate the tuning bandwidth of different ICOs implemented for this study. These considerations motivated us to implement a custom test technique based on data gathered with a spectrum analyzer.

The two most serious draw-backs to direct phase noise measurements with a spectrum analyzer are, first, the limitation of discernable close-in offsets due to the noise sidebands of the instrument's local oscillator; and second the instrument's inability to distinguish between AM and FM noise. The first constraint is not an issue at 5MHz offsets, while for the second we verified that the AM noise contribution is negligible from symmetry of the static output spectra.
of the oscillators. Another drawback, mentioned earlier, is set by the noise floor of the spectrum analyzer. For the instrument used in this work, Agilent model 8593E, the noise floor limits the observable sidebands of our ICO measurements to offsets well below 10MHz, while we are interested in noise at a 50MHz offset. Nonetheless, from our observations and in accordance with [36] the non-close in ECM phase noise has a conventional slope of –20dB per decade. Thus measurements at 5MHz can easily be used to estimate phase noise at a 50MHz offset. Finally, three measurement adjustments must be taken into account when calculating the phase noise from measured peak and sideband powers [45-47]. First the sideband noise must be normalized to 1Hz. The actual bandwidth of the noise measurement is set by the resolution bandwidth (RESPBW) of the spectrum analyzer's “IF filter”. This is the 3dB bandwidth of the front-end bandpass filter against which the input signal is swept. The normalization is simply performed by subtracting 10×log₁₀(RESBW) from the sideband readout. This calculation assumes flat noise spectra and an ideal filter with a flat passband; in practice however the filter's spectral power gain is closer to a bell shape. To compensate for this an equivalent noise bandwidth is defined for the filter. This is the 3dB width of an ideal rectangular bandpass filter, with the same peak gain and equivalent area as the IF filter. For the type of IF filter used in our measurements the equivalent bandwidth has the effect of increasing the RESBW by a factor of 1.128. Consequently, an additional –0.52dB must be included when normalizing noise sidebands. The last adjustment to consider is caused by an under-response of the spectrum analyzer to noise. This results from the instrument averaging logarithmic measures of bandpass noise envelopes, rather than computing the logarithm of average measurements. To compensate for this effect a +2.5dB correction factor must be added to the noise readout. A further description of these calibration issues can be found in Refs. [45-47].
3.3 Custom Measurement Approach

To accurately measure phase noise versus input tuning for several versions of the ICO we designed and implemented the custom measurement setup described in this section. The measurement procedure involves automatic data gathering and post-processing, this is illustrated by the top and bottom sections of Fig. 3.1. During the data gathering stage the design under test (DUT) is electrically probed, on-wafer, with two microwave probes, GGB model MCW40A. These probes are coplanar multi-contact wedges with a probe spacing of 150µm, and a custom footprint of eight signals each, from which three are independent signals, two are independent DC powers, and three are grounds. A semiconductor parameter analyzer, Agilent model 4155A, is used as high-precision controllable DC current source. This instrument is connected to the "Iin" input of the ICO, which sets the value of the tail currents to the ECM. The resulting single-ended output is measured with a spectrum analyzer, Agilent model 8593E, connected to one of the differential ICO outputs, while the remaining output is terminated on probe to 50Ω. Because the spectrum analyzer's input is AC coupled, a RF bias tee is used, on probe, to provide different current paths for the DC and AC output components.

Several precautions are necessary to reduce noise pickup, interference and measurement disturbances. Among these, noise filtering at the input of the ICO, and powering the DUT with batteries, are indicated in Fig. 3.1. The filtering consists of a sequence of low pass filters and de-coupling inductors. Other precautions not shown in the figure include built-in decoupling capacitors on the microwave probe power pins, placing the test setup on a vibration-isolated table, running the measurement sequence in the dark, shielding of sensitive connections, and careful distribution of ground paths.
Figure 3.1: Custom phase noise measurement setup. The top section shows a block diagram of the test setup and its sequence of operation during the data gathering stage. The bottom section shows the sequence followed during the post-processing stage.
The data gathering process is controlled and automated with an external computer that communicates with the input current source and the spectrum analyzer via GPIB interfaces. To this effect a custom software control routine was created in National Instruments LabView instrumentation control package.

The sequence for data gathering is indicated on Fig. 3.1 by steps labeled 1 to 6. During the gathering process the input current is swept in fixed increments to cover a certain input range of the ICO. The increment, start and stop values of the sweep are initial parameters selected by the user. For each fixed input the fundamental of the output spectrum is located and centered on the spectrum analyzer. Then, the span view is set in accordance to initial user settings. To increase dynamic range, the spectrum is averaged 100 times; this has the additional advantage of smoothing spurious picked-up noise on the sidebands. The resulting averaged spectrum is then downloaded from the spectrum analyzer into the computer and saved in ASCII format in a new data file of its own. This process is repeated for each value of the input current sweep, and takes approximately 20 seconds per input. As a result, one sweep of the data gathering process produces a collection of output spectrum data files that must be post-processed to extract the exact carrier power, sideband power, and compute phase noise. In addition, as will be explained soon, because the ICO output suffers from low frequency drifting the gathering sequence is repeated twice at a span of 10MHz and 100MHz.

The post-processing sequence is illustrated at the bottom of Fig. 3.1. The peak and sideband data required to compute phase noise are extracted separately, one from the data collected at 100MHz span, and the other from the data collected at 10MHz span. The reasons for this are as follows. During the gathering process, as the video averaging of 100 traces is performed, any drifting of the narrow-shaped ICO peak considerably affects the final peak value saved into file. As the spectrum analyzer's span is reduced the frequency resolution is
also reduced and therefore the drifting effect is more noticeable; this drifting still occurs
despite the use of built-in tracking features in the spectrum analyzer. At a 10MHz span, the
drifting effect was observed to shift the peak no more than a single data frequency point during
the complete averaging of 100 traces. For each tuning point the averaging time in our setup is
30 seconds long, over which the maximum drifting was observed to be approximately 500kHz.
Using the built-in tracking features of the Spectrum Analyzer this drift is not eliminated, but it
is reduced to approximately 25kHz over 30 seconds. The resulting effect is an error of close to
5dB in the averaged magnitude, and thus in the final phase noise result. 25kHz is also the
frequency resolution of the analyzer at 10MHz span, and for that reason the drifting effect is
occasionally observed. On the other hand at a 100MHz span the resolution is limited to a range
of 250kHz, over which the analyzer detects, and averages, the peak amplitude of the signal.
Thus the effect of drifting can easily be avoided using a larger frequency span. In contrast the
averaging of sideband data is not significantly affected by drifting, and is more accurate at
lower spans. For example at 5MHz offset from the carrier and with a −20dB per decade slope,
a frequency drift of 25kHz results, not considering averaging, in a change of 0.04dB in the
sideband magnitude. For these reasons, two different spans are used to extract the peak and
sideband information. There are other well-known phase noise measurement techniques that by
employing phase or frequency demodulation are insensitive to drifting of the oscillator under
test. These setups however require the use of mixers, filters, reference oscillators or delay lines,
which add uncertainty to the final result. In contrast the two span approach described above
provides a practical solution to maintain a setup that by its simplicity minimizes measurement
uncertainties.
The data is also mathematically fitted to reduce discretization errors originating from digital handling and discrete displaying of data by the spectrum analyzer. In particular, for the instrument used, data is limited to 401 frequency points regardless of the span selected. Peak data collected at 100MHz is fitted, in dBm, with a Gaussian function, and two-sided sideband data collected at 10MHz span is fitted, in mW, with a \( \alpha/(f-\beta)^2+c \) function. At the higher span the IF filter is set to a large resolution bandwidth (RESBW) of 1MHz. The spectral power function of this filter is Gaussian in shape, and when swept against a signal with a narrower line-width its full shape is depicted on screen; thus the Gaussian fitting performed to find the location and magnitude of the peak. The fitting equation is of the type \( \alpha \times 2.72^\left(- (f-\beta)^2/\gamma\right) + c \).

Fig. 3.2 (a) illustrates the effect of peak Gaussian fitting for one set of data collected from an ICO with a 150fF timing capacitor. Only a limited set of data points is used around the peak. The sidebands on the other hand follow a slope of \(-20\)dB per decade until limited by the noise floor of the spectrum analyzer. Since the power of non-coherent signals add linearly, it follows that the combined sideband plus noise floor can be fitted, in milliwatts, with a \( \alpha/(f-\beta)^2+c \) equation. From this calculation an accurate value of sideband noise at 5MHz offset is obtained.

Fig. 3.2 (b) shows sideband fitting for a 150fF ICO, in particular, under interference pickup conditions. The left and right sidebands are shown in separate plots only for convenience, as the fitting is performed considering simultaneously data from both sidebands with no constrains to the center point. The interference in the data is visible from the uneven distribution of data points, and appears sporadically, from external radiation, despite the considerable precautions to reduce noise pickup in the experimental setup. The interference is responsible for the fitting mismatch below approximately 2.5MHz offset, nonetheless at 5MHz
Figure 3.2: Mathematical fitting during the post-processing stage. (a) Gaussian fitting of peak data. (b) Sideband fitting. Open circles are data collected from a 150fF ICO, solid lines are fitting curves.
offset the fitting is in good agreement with the data. In the nonappearance of external pickup the fitting is excellent in the full range of 2MHz to 5MHz offsets. As seen from Fig. 3.2 this step proved to be extremely effective in filtering spurious interference from the sidebands and consequently plays a major role in the repeatability of measurement results.

The mathematical fittings provide accurate values of peak location, peak magnitude, and sideband levels that are used to calculate phase noise. Fig. 3.3 illustrates the obtainment of phase noise from the magnitude and sideband data, and the effectiveness of the two-span data gathering approach with data collected from a 150fF ICO. Fig.3.3 (a) shows fitting results of peak magnitude for data collected at both 10MHz and 100MHz spans. Fig. 3.3 (b) shows fitting results for sideband noise at 5MHz offset from the carrier collected at 10MHz span. Open symbols are for 10MHz and closed symbols for 100MHz span, and measurements are plotted for the complete tuning range of the ICO from 0.5mA to 1.5mA. On Fig. 3.3 (a) low frequency drifting is responsible for the scattered data points at 10MHz span, while the data obtained at 100MHz shows no drifting effects. This shows how the effectiveness of employing a larger span to reduce drifting effects on the averaged data. On the other hand, on Fig. 3.3 (b), the 5MHz-sideband noise obtained at 10MHz shows no significant distortion from drifting. Fig. 3.3 (c) shows the calculated phase noise versus control current for two cases, first combining 10MHz peak and sideband data, second combining 100MHz peak and 10MHz sideband data. Open symbols are from employing peak and sideband data collected at 10MHz, while closed symbols are from combining 100MHz peak and 10MHz sidebands. From this the advantages of the two-span approach are obvious. In the calculation of phase noise the RESBW of the spectrum analyzer, as well as the filter equivalent noise bandwidth, and the +2.5dB correction factor are taken into account.
Figure 3.3: Post-processed results versus input control current. (a) Peak magnitude for data collected at spans of 10MHz, open symbols, and 100MHz, closed symbols. (b) Sideband data from 10MHz span. (c) Phase noise computed from 10MHz peak and sideband data, open symbols, and from combining 100MHz peak and 10MHz sideband data, closed symbols.
3.4 Results

Phase noise measurements for two ICOs fabricated on the same chip but with different timing capacitors are shown in Fig. 3.4. Subplots (a), (b) and (c) are results for both ICOs from three different dice. The thin lines are for a timing capacitor of 150fF and the thick lines for 300fF. The data was collected at tuning steps of 5μA across the full tuning range of both oscillators. Repeatability, not shown, is ±0.5dBc/Hz. The consistency of the results across different dice is excellent, despite the highly structured appearance of the curves. We attribute this effect to reflections from the spectrum analyzer front end, whose input return loss is 20dB. These reflections result in an added measurement uncertainty of ±1.5dB/Hz, for a total accuracy of ±2dB/Hz.

The results for the 300fF ICO are shown here as an example of the lowest phase noise values dealt with in this work, since this is the slowest oscillator implemented. In fact, the minimum measurable phase noise at 5MHz offset is approximately −118dBc/Hz. This is due to a −75dBm noise floor limit from the portable spectrum analyzer used. Nonetheless this measurement performance is adequate for our purposes.

The frequency range of the measurements is in principle limited only by the bandwidth of the spectrum analyzer, which extends from 9kHz to 22GHz. Such a large bandwidth is achieved by the instrument automatically selecting among internal local oscillators. This has the effect of segmenting the practical range of our measurements into several ranges, among which the most relevant to our tests are from a few MHz to 2.9GHz, and from 2.75GHz to 6GHz. It is also important to note that the setup can easily be adapted to sweep input voltages in steps of micro Volts and thus be used for testing of the linearized VCO.
Figure 3.4: Measured phase noise at 5MHz offset for two ICOs with 150 and 300fF capacitors fabricated on the same chip. (a), (b), and (c) show results for three different dice. Thin lines are for 150fF and thick lines for 300fF.
3.5 Summary and Conclusions

In this chapter we described a custom phase noise measurement technique developed to automatically measure noise versus input control of our ECM-based oscillators. Reliable measurements are achieved from post-processing output oscillator spectra captured with a portable spectrum analyzer. The setup successfully filters noise and interference pickup from external sources, and is designed to remove unwanted effects from oscillator thermal drifting. Measurements are performed at a fixed 5MHz offset from the carrier and can be taken for carrier frequencies over a range of a few MHz to 2.9GHz or from 2.75GHz to 6GHz. These ranges fully cover the bandwidth of our oscillators. Input sweep control can be set as low as hundreds of nano Amperes or hundreds of micro Volts. The minimum possible measurement is $-118\text{dBc/Hz}$ and the accuracy is $\pm 2\text{dB/Hz}$. Repeatability is excellent.

The measurable range of phase noise offsets is very limited compared to that of dedicated phase noise measurement instruments, but nonetheless it is more than sufficient for our purposes. In addition, the setup provides a low cost and practical solution to our experimental work with better accuracy than that of available dedicated instruments. In the next chapter, the phase noise data presented in Fig. 3.4, as well as additional results obtained with our custom phase noise measurement setup, are analyzed and compared with theoretical predictions of phase noise in ECMs.
Chapter 4
Noise Limits in High-Speed ECMs

4.1 Introduction to Chapter

In Section 1.5.2 we reviewed previous studies on the origins of noise in multivibrator oscillators. These studies have been successful in predicting the comparatively poor noise performance of several multivibrator designs, and have been the basis for several circuit optimizations. However, experimental validation has only been done for circuits operating at low speeds relative to their transistor cutoff frequency, $f_T$. For example, in Refs. [30, 35, 37] the test designs operate at 1kHz, 5kHz, and 1MHz respectively; although no details are given of the transistor performances, it is reasonable to assume $f_T$'s of no less than 200MHz. A higher speed of operation is covered in Ref [36], where theoretical predictions are successfully confirmed with a 100MHz design, for a process of 3GHz $f_T$. In addition to this, circuit optimizations proposed to lower phase noise [29, 35] have resulted in lesser tuning linearity than that of a simple ECM core. On the other hand, the WFM modulator designed in this thesis operates at a higher fraction of $f_T$, and maintaining tuning linearity is critical to reduce distortion of the multicarrier CATV signal. It is therefore important for the purposes of this work to investigate the phase noise performance of the ECM in high-speed operation, and to seek for alternate approaches to phase noise improvement.

This chapter presents the results of such investigation. In Section 4.2 we present phase noise measurements for three versions of the simple ICO with timing capacitors of 300fF, 150fF, and 100fF, which combined cover an output range from 0.5GHz to 2.8GHz.
These experimental results were obtained using the custom measurement setup described in Chapter 3. Section 4.3 reviews modeling considerations that are critical for phase noise prediction in time-varying circuits, and presents modeling results obtained from the SpectreRF circuit simulator program. The experimental and predicted phase noise results are discussed in Section 4.4. Analysis of the results reveals novel findings about the noise limits of high-speed ECMs, and provides simple circuit optimizations to improve noise performance without resorting to topological modifications. Section 4.5 closes the chapter with a summary and conclusions.

### 4.2 Phase Noise Measurement Results

For our study of noise in high-speed ECM we performed phase noise measurements of the simple ICO with the test setup described in Chapter 3. We concentrated on the performance of the simple ICO as its operation is closest to the intrinsic ECM. Furthermore, to extend the frequency range of the tests we implemented a new IC with two variants of the ICO by replacing the 150fF timing capacitor \( C \) with capacitors of 100fF and 300fF; we also included in the new IC an instance of the 150fF ICO. Except for the values of the timing capacitor the three ICO circuits are identical in all respects. Experimental results for the three circuits are shown in Fig. 4.1 and summarized in Table 4.1. In Fig. 4.1 (a) we show static tuning response and in Fig. 4.1 (b) phase noise versus tuning current at a 5MHz offset. All data was collected at tuning steps of 5\( \mu \)A. In Fig. 4.1 (a) the tuning range of the 100fF ICO is truncated at 1.25mA to avoid complications that arose at higher frequencies from switching of the spectrum analyzer's internal local oscillator. For all versions, the response is qualitatively linear over the full tuning range.
Table 4.1: Measured tuning and noise performance for three variants of the ICO with timing capacitors of 300, 150, and 100fF. Phase noise is measured at the operating point of the central frequency, $f_c$.

<table>
<thead>
<tr>
<th>C (fF)</th>
<th>$f_c$ (GHz)</th>
<th>Tuning range (GHz)</th>
<th>SSB phase noise at 1mA input (dBc/Hz) @ 5MHz offset</th>
</tr>
</thead>
<tbody>
<tr>
<td>300</td>
<td>0.86</td>
<td>0.5 – 1.2</td>
<td>-112</td>
</tr>
<tr>
<td>150</td>
<td>1.61</td>
<td>1.0 – 2.3</td>
<td>-108</td>
</tr>
<tr>
<td>100</td>
<td>2.29</td>
<td>1.3 – 2.8</td>
<td>-105</td>
</tr>
</tbody>
</table>

Figure 4.1: Measured phase noise versus tuning current for three versions of the simple ICO implemented on the same dice with timing capacitors of 100fF, 150fF, and 300fF. (a) Static tuning response. (b) Phase noise versus tuning current at 5MHz offset.
Chapter 4. Noise Limits in High-Speed ECMs

To verify the validity of these experimental results as well as to further analyze the internal mechanisms of phase noise generation we proceeded to obtain phase noise predictions with the help of a circuit simulator. The experimental and simulated results are discussed in more detail in Section 4.4.

4.3 Phase Noise Simulations

The NT25 Advanced Silicon Bipolar design kit offers device models for four different analog circuit simulation packages; namely, Hspice, Eldo, Spectre, and Libra. Among these we employed the Hspice models for all circuit simulations that have been presented in the previous chapters. This choice was motivated by familiarity with the tool, and the need for time-domain simulation during the design phase of the project. However, when pursuing phase noise predictions, the complex nature of phase noise generation in multivibrators called for different simulation techniques not available in Hspice. For this reason, we performed all phase noise calculations with the SpectreRF Circuit Simulator. In this section we will discuss the underlying principles that motivate this choice and will present simulation results for the three versions of the simple ICO.

Traditional SPICE-derived time domain circuit simulation programs do not provide for direct phase noise computation. They do, however, support noise calculations to find total and individual noise contributions to an output voltage noise port due to all noise generators present in a circuit. These calculations are performed as an option of the standard small-signal AC analysis, and as such are based on circuit linearizations around a DC operating point. For this reason, they are valid only for circuits with a fixed DC bias. Such circuits are categorized as linear time-invariant circuits, as opposed to time-varying circuits in which the operating point varies with time. Nonlinear oscillators are time-varying by
nature, with periodic variation of their operating point, and, therefore, require a different noise analysis approach. One alternative with SPICE is to study the impact of injected noise on the output spectrum of an oscillator following post-processing of a transient analysis. For this the user must generate a custom noise source, connect it to a particular node, and perform a Fast Fourier Transform (FFT) of the transient results. This approach has been attempted in [48] for several oscillators, but shown to introduce considerable errors from internal interpolations that occur during the computation of steep transitions.

The solution of time-varying circuits is further complicated by two additional noise conversion mechanisms [49]. First, the variations of the operating point are responsible for modulating bias dependent noise sources, such as shot noise in bipolar transistors. Second, variations of the operating point cause modulation of the internal circuit noise transfer functions. This effect is responsible for noise frequency conversions generally referred to as noise folding effects. The noise created by both of these mechanisms exhibits random statistical properties that vary periodically with time, and for that reason is said to be cyclostationary. In the frequency domain, cyclostationary noise can be computed from the convolution of the noise sources with the periodic modulating transfer function [49,50]. As a consequence, the noise is replicated around all harmonics of the periodic signal, creating a correlated overlap of up and down converted noise components, which add up to the total cyclostationary noise power spectrum [50]. This is the cause of the noise folding effect.

Noise folding effects can neither be predicted with SPICE-type simulators, nor can their overlapping components be separated with a spectrum analyzer. Fortunately, the SpectreRF Circuit Simulator contains solution algorithms to predict cyclostationary noise, as well as built-in capability for direct phase noise calculation. The simulation sequence requires performing a “periodic steady state analysis” (PSS) first, to find the time-varying
operating point of the circuit, followed by a "periodic noise analysis" (Pnoise) [51]. Phase
noise simulation results obtained with SpectreRF are shown in Fig. 4.2.

The results in Fig. 4.2 are for the intrinsic ECM with the current mirror tail control. The limiting amplifier and output driver are not taken into account in the simulations, as they are not expected to contribute substantially. In all simulations the current was swept in steps of 5μA from 0.5mA. Diamonds, circles and squares indicate results for the 100, 150, and 300fF ICOs respectively. The computations represent phase noise, at 5MHz offset, as seen across the timing capacitor; this is where the timing of all the oscillator nodes is defined, and is equivalent to measuring the phase noise at the output of the ECM, i.e., at the bases of the ECM core emitter coupled pair transistors as shown in Fig. 2.2. Noise folding contributions are considered in the computations for the first nine harmonics of the oscillator's large signal output. Higher harmonics are not included as their contribution was found to be negligible. Measurements are reproduced for comparison and indicated by the thick lines. Calculations are in excellent agreement with measurements, in the absence of any adjustable parameters thus confirming the validity of the measurements and the performance of the scalable bipolar models used. An in depth analysis of these results and their implications follows.
Figure 4.2: Predicted and measured phase noise versus tuning current for three versions of the simple ICO implemented on the same dice with timing capacitors of 100fF, 150fF, and 300fF. All results are for 5MHz offset from the carrier. Solid lines are measurements. Diamonds, circles, and squares are simulation results for the 100, 150, and 300fF ICOs respectively.
Chapter 4. Noise Limits in High-Speed ECMs

4.4 Discussion

Since the value of the timing capacitor and the magnitude of the tuning current play a direct role setting the output frequency of the ECM core, it is reasonable to expect that perturbations linked to these two parameters will, in some manner, result in unwanted output frequency variations leading to phase noise. In that respect, careful analysis of the measurement results reveals that as the ICO operation increases in frequency its phase noise becomes directly proportional to the tuning current and inversely proportional to the square of the timing capacitor. Consequently, for each version of the ICO the rising slope with frequency, as seen in Fig. 4.2, is simply caused by this linear dependence with the current, while the relative offset between any two curves in Fig 4.2 is proportional to the squared ratio of their timing capacitors. These relationships are demonstrated graphically in Fig. 4.3 with plots of normalized phase noise versus oscillator output frequency. Fig. 4.3 (a) illustrates the phase noise dependence with the tuning current $I_{ctrl}$, by plotting the phase noise of all three ICOs normalized by $I_{ctrl}$. Fig. 4.3 (b) illustrates the phase noise dependence with the timing capacitor $C$ by plotting phase noise of all three ICOs multiplied by $C^2$ and normalized by $I_{ctrl}$. In Fig 4.3 (a) the normalization by $I_{ctrl}$ effectively flattens the three curves but with better outcome as frequency increases. The deviation from a flat line is more pronounced for the 300fF ICO but extends to the lower frequency end of the 150fF; this behaviour suggest the presence of an additional phase noise mechanism that raises the ECM noise at the lowest frequencies. In Fig. 4.3 (b) the offsets that exist between the curves in Fig. 4.3 (a) are cancelled, thus confirming the dependence with $1/C^2$, and only the same low frequency discrepancy already observed in Fig. 4.3 (a) diverges from a constant level.

Considering previously published theories of phase noise generation in multivibrator oscillators, the observations presented above are surprising. Recalling from Chapter 1, such
Figure 4.3: Graphical demonstration of phase noise dependence with tuning current $I_{ctrl}$ and timing capacitor $C$. Curves are based on phase noise measurements at 5MHz offset for three versions of the ICO with timing capacitors of 100fF, 150fF, and 300fF. (a) Is measured phase noise versus oscillation frequency normalized by $I_{ctrl}$. (b) Is measured phase noise versus oscillation frequency multiplied by $C^2$ and normalized by $I_{ctrl}$, where $C$ is each oscillator's corresponding capacitor.
Chapter 4. Noise Limits in High-Speed ECMs

studies predict that phase noise is generated from two mechanisms. First, current noise flowing through the ECM timing capacitor directly modulates the output frequency. In particular, equations derived in [36, 37] show that the resulting contribution to phase noise is independent of the multivibrator oscillation frequency, and inversely proportional to the square of $C$ and of the noise frequency $f_m$. In Ref. [36] it was determined that this current noise is commonly generated from thermal noise at the emitter resistors of the tail current sources. Second, voltage noise, originating in the voltage loop of the timing capacitor affects the frequency of the oscillator by disturbing the multivibrator regeneration threshold levels. This noise increases with the square of the oscillation frequency and decreases with the square of the timing capacitor $C$ and of the noise frequency $f_m$ [36, 37]. In addition the noise is downconverted by the switching action of the multivibrator, with overlapped contributions from all possible harmonics within the oscillation frequency and the bandwidth limit of the multivibrator. From the two mechanisms the second is demonstrated to be dominant in Refs. [35-37]. In accordance with these studies, the results suggest that frequency folding may account for the higher noise observed at the lower end of the 300fF ICO tuning range, and is progressively reduced by the limited bandwidth of the folding mechanism. However, the simple and direct relation of phase noise with $I_{ctrl}$ and $C^2$, which dominates the results, is unexpected. To further explore these issues we must turn our attention to the results gathered from circuit simulations.

From the simulation program it is possible to obtain individual contributions from each noise source to the total noise across the capacitor. In addition, it is possible to view the noise folding contribution from each harmonic. Results obtained in this manner are indicated in Fig. 4.4 and Table 4.2. Fig. 4.4 is a schematic of the simulated circuit with
Figure 4.4: Circuit diagram of the simulated ICO. Highlighted sources are the relevant noise contributors to the total noise across the capacitor C. Thermal voltage noise sources are labelled $vn_{th}$, and collector shot noise current sources are labelled $ic_{shot}$. All noise sources are uncorrected.

Table 4.2: Summary of principal noise contributors at three tuning points, as obtained from prediction calculations for the circuit shown in Fig. 4.4 with 150fF.

<table>
<thead>
<tr>
<th>Noise origin</th>
<th>Contribution to total noise vs. tuning current</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$I_{ctr} = 0.5mA$</td>
</tr>
<tr>
<td></td>
<td>$V^2/Hz$</td>
</tr>
<tr>
<td>$Q_3, Q_4$ ic shot (emitter followers)</td>
<td>5.84e-13</td>
</tr>
<tr>
<td>$Q_1, Q_2$ ic shot (emitter-coupled pair)</td>
<td>4.38e-13</td>
</tr>
<tr>
<td>$R_{C1}, R_{C2}$ (ECM load resistors)</td>
<td>7.55e-13</td>
</tr>
<tr>
<td>$R_{EA}, R_{EB}, R_{EC}$ (tail emitter resistors)</td>
<td>6.16e-13</td>
</tr>
</tbody>
</table>
additional highlighted sources that indicate the relevant noise contributors to the total noise across the capacitor. Voltage noise sources indicated as $vn_{th}$ are from thermal noise generated by the current mirror emitter resistors $R_{E_A}$, $R_{E_B}$, and $R_{E_C}$, and the load resistors $R_{C_1}$, and $R_{C_2}$. Current noise sources indicated as $ic_{shot}$ are from the collector shot noise generated by emitter follower transistors $Q_3$, and $Q_4$, and emitter coupled pair transistors $Q_1$, and $Q_2$. All noise sources are uncorrelated. Table 4.2 summarizes the principal noise source contributors for $C=150fF$ at three different operating points. For each noise source the absolute noise contribution, and its percentage contribution to the total noise across the timing capacitor, are indicated.

At the lowest frequency of operation, $I_{ctrl}=0.5mA$, the noise is dominated by the resistors. The noise from resistors $R_{E_A}$, $R_{E_B}$, and $R_{E_C}$ is consistent with the dominant noise sources expected from a current mirror circuit [52]. As shown in Table 4.2, the noise from resistors $R_{C_1}$ and $R_{C_2}$ plays an important role and is the dominant contributor at $I_{ctrl}=0.5mA$. As the VCO frequency increases, with tuning current, the collector shot noise of the transistors becomes dominant. This occurs through accumulative noise folding contributions from higher harmonics. In contrast, a similar analysis of the 300fF ICO shows that despite a progressive increase of shot noise with frequency it never overcomes the effect of the resistor's thermal noise. While for the 100fF ICO, collector shot noise is dominant throughout the complete tuning range, and base shot noise sources begin to play a noticeable role. Notice, in Table 4.2, that the contribution from shot noise of transistors $Q_1$, $Q_2$ increases by a factor of 4.4 with a current increase of a factor of 3. For the 100fF ICO over the same range the contributions increased by a factor of 7. Thus, shot noise contributions are not simply scaling with bias current. For this reason we suspect that as the reactances of the ICO small signal circuit change with increasing frequency, the coupling
between noise sources in the ECM core and the switching capacitor nodes also increases. This will explain in Table 4.2 the raising contribution from the load resistors, which was also observed for the 100fF ICO. On the other hand, for all three versions of the ICO, the contribution from thermal noise of the tail emitter resistors remains approximately constant over the full tuning range.

The following conclusions can be drawn from these results. First, that as the oscillation frequency of the ICO is increased collector shot noise of the ECM switching transistors becomes the principal source of phase noise. This is due to increasing noise coupling to the capacitor nodes; and, becomes noticeable at a 0.06 fraction of \( f_T \), or a 0.3 fraction of the maximum operating speed of the ICO, which was observed to be of approximately 5GHz. Therefore, this effect is seen here for the first time either due to the high operating speeds of the ICO, the reduced intrinsic parasitics of the high-speed bipolar technology, or a combination of the two. Second, the increased noise level observed at the lower tuning end of the 300fF ICO is not caused by band-limited noise folding but rather by the dominance of thermal noise from the tail resistors which is then overcome by the increasing coupling of the ECM noise sources. Third, the dominance of shot noise across the capacitor is consistent with the experimental observation of phase noise proportionality with \( I_{ctr} / C^2 \). Compared to previous works, the direct proportionality with \( I_{ctr} \), and thus with \( f_0 \), can be explained by an equivalent parallel noise source (proportional to \( I_{ctr} \)), but not by the effects from an equivalent series voltage noise source. However, among these sources, the voltage source was expected to dominate the phase noise behavior at higher frequencies of operation. This reversal of the relevant noise mechanism at high speed, combined with the unexpected dominance of shot noise constitutes the principal difference between our observations and those reported in previous works [30, 35-37].
Chapter 4. Noise Limits in High-Speed ECMs

From our findings it follows that through careful selection of \( C \) and \( I_{ctrl} \), phase noise can be improved for a fixed frequency range of operation. This is because the output frequency of oscillation is directly proportional to the ratio \( I_{ctrl}/C \), while the phase noise is directly proportional to the ratio \( I_{ctrl}/C^2 \). This means that in order to reduce phase noise and maintain a particular central frequency \( f_c \), a designer must scale both the tuning current and the timing capacitor by the same factor. The predicted improvement from this technique is \( 10 \log_{10}(\text{scale factor}) \) dB. This was confirmed through simulations of an ECM core with ideal current sources. Results at one tuning point are presented in Table 4.3, and show excellent agreement with expected phase noise improvements.

Table 4.3: Simulated and predicted phase noise improvement from scaling \( C \) and \( I_{ctrl} \). Results are for simulations of an ECM core with ideal tail current sources. Improvements are referred to performance at 1mA and 150fF.

<table>
<thead>
<tr>
<th>( C ) (fF)</th>
<th>( I_{ctrl} ) (mA)</th>
<th>Scaling factor</th>
<th>( \mathcal{L}(5\text{MHz}) ) (dBc/Hz)</th>
<th>Improvement from simulations</th>
<th>Predicted improvement</th>
</tr>
</thead>
<tbody>
<tr>
<td>150</td>
<td>1</td>
<td>1</td>
<td>-107.8</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>300</td>
<td>2</td>
<td>2</td>
<td>-111.1</td>
<td>3.3dB</td>
<td>3.0dB</td>
</tr>
<tr>
<td>750</td>
<td>5</td>
<td>5</td>
<td>-115.0</td>
<td>7.2dB</td>
<td>7.0dB</td>
</tr>
</tbody>
</table>

The practical limits of this scaling approach are, at first sight, set by the capabilities of the monolithic technology. For example, to implement a 1.5pF, in the NT25 technology, requires a physical area of approximately 40\times40\mu m^2, while a 1nF capacitor would require an area of 1000\times1000\mu m^2, which is impractical. To accommodate for larger currents, NPN transistors with collector currents of up to 12mA are standard library components in NT25 and higher currents can be achieved by connecting several transistors in parallel. In practice, however, the limits of the scaling design approach are more likely to be set by the effects of circuit sizing on the tuning linearity of the ECM, which still remains a topic for further study. However, from preliminary simulations of an ECM core, with ideal tail current
sources, a scaling factor of two, and a tuning range of 0.5mA to 1.5mA, it was observed that
the deviation from linearity worsen approximately 10MHz at the extremes of the tuning
range but remained approximately the same in the range of 0.75mA to 1.25mA, which is the
ECO tail tuning range employed in the linearized VCO. These observations must still be
confronted with various issues such as the use of non-ideal current sources, dependence
with scaling factor, transistor scaling to accommodate for larger currents, feedback
linearization, and prospect performance in a WFM system.

There is an additional finding worth mentioning, but which in lack of a solid
explanation is presented here merely as an observation. It was found that phase noise
measurements of the ICO as shown in Fig. 4.1 can closely be predicted by assuming direct
frequency modulation of the ICO output from an equivalent shot noise source, \( i_{\text{noise}} \), of
value \( 2qI_{\text{ctrl}} \). In such a situation, an expression for phase noise can easily be derived from
simple frequency modulation theory to be:

\[
L = \frac{i_{\text{noise}}^2}{2 \cdot I_{\text{ctrl}}^2} \left( \frac{f_c}{f_m} \right)^2
\tag{4.1}
\]

Where, \( f_c \) is the ICO oscillation frequency and \( f_m \) the phase noise offset from the carrier.

From the ICO tuning expression:

\[
f_c = \frac{I_{\text{ctrl}}}{4 \cdot C \cdot V_s}
\tag{4.2}
\]

where \( V_s \) is the forward voltage drop of the ECM load diodes. Equation (4.2) is equivalent
to equation (2.1) but is repeated here for the convenience of the reader.

Therefore, from substituting (4.2) into (4.1), it follows that:

\[
L = \frac{q}{16 \cdot V_s^2} \left( \frac{I_{\text{ctrl}}}{f_m^2 \cdot C^2} \right)
\tag{4.3}
\]
Chapter 4. Noise Limits in High-Speed ECMs

Equation (4.3), accounts for the dependence with $I_{ctrl}$ and $C$ as discussed before, for the observed -20dB per decade slope with offset, and for the absolute value of the phase noise as obtained from both measurements, and simulations. As in the graphical analysis shown in Fig. 4.3, the match of equation (4.3) to experimental results improves with increasing frequency and is less effective at the lower end of the 300fF ICO. In addition, Equation (4.3) is consistent with the dominance of shot noise observed from simulations. However, this dominance of shot noise is caused by the contribution of several transistors, each exhibiting noise folding effects. Despite the agreement with experimental data, no explanation could be found to account for the addition of all shot noise sources into a single equivalent current noise source of value $2qI_{ctrl}$. It is therefore our opinion that it is only by chance that the simplistic analysis leading to equation (4.3) leads to such accurate prediction of phase noise, both in the scaling behaviour with $I_{ctrl}$ and $C$, and also the absolute value.

4.5 Summary and Conclusions

In this chapter, we presented experimental and predicted phase noise results for three versions of the ICO with different timing capacitors, which combined cover a range from 0.5GHz to 2.8GHz. It was explained that due to the time-varying nature of the ECM, circuit predictions must account for cyclostationary noise effects. Agreement of measured and predicted phase noise is excellent and confirms the validity of our custom measurement setup as well as the performance of the scalable bipolar models used. Results show that at high-speed operation phase noise is directly proportional to the tuning current and inversely proportional to the square of the timing capacitor. This provides a design approach to reduce phase noise while maintaining the frequency tuning range intact, and the basic topology of the ECM. The results obtained are novel in view of previous theoretical and experimental
work [30, 35-37], which had only been confirmed until now at lower fractions from the
maximum multivibrator speed. In particular, it was found that with increasing oscillator
frequency, folding of shot noise from the ECM switching transistors is responsible for the
dominance of current on the phase noise.

The results obtained here can be used to improve the noise performance of the
linearized VCO. As mentioned earlier, phase noise of the VCO is 5.5dB higher than that of
the ICO. This is due to additional noise sources present in the transadmittance amplifier
stage, and can be solved through careful redesign of this block. But as shown in this chapter,
the phase noise of the VCOs ECM core can also be improved by scaling the tuning current
and timing capacitor. For example, for an improvement of 10dB in noise the scaling factor
is 10. This is easily achievable by implementing the ECM with transistors optimized for
10mA collector currents and a 1.5pF capacitor. Practical scaling factors of a few tens are
feasible with current monolithic technology. In addition, load diodes, in the ECM core, with
lower forward voltage can also be used to help reduce the needed tuning currents. However
it is likely that the limits of this scaling approach are set by the effects of circuit scaling on
the tuning linearity of the ECM, which still remains a topic for further study.
Chapter 5
System Perspectives

5.1 Introduction to Chapter

From an early stage in our work, it became possible to numerically obtain some estimates of modulator performance in the context of the wideband FM system. This was necessary to validate the design approach and to determine any required improvements. In this Chapter we present the results of such investigation. The motivation to measure and improve phase noise performance, as described in Chapters 3 and 4, is a direct consequence of the conclusions presented in this chapter. The figures of merit we used for such an assessment are the composite-second-order (CSO), composite-triple-beat (CTB), and the carrier-to-noise ratio (CNR) of a WFM system with the linearized VCO as the FM modulator. As mentioned earlier the performance targets we are aiming for in the WFM system are CSO and CTB below -65dBc, and CNR above 52dB.

This chapter is organized as follows. In Section 5.2 we explore the CSO and CTB performance of the linearized VCO in a system with ideal optical link, ideal demodulator, and 80-channel transmission. For this purpose distortion calculations are performed in MATLAB based on the VCO transfer function obtained from static tuning predictions. In Section 5.3 we explore the CNR performance of the VCO in a typical narrowcast system with an ideal demodulator. We present simulation results that prove unnecessary the use of pre-emphasis, and provide combined performance estimates of distortion and noise are presented for 80 and 40-channel transmission. Finally, Section 5.4 closes the chapter with a summary and conclusions.
5.2 Composite-Second-Order (CSO) and Composite-Triple-Beat (CTB) Performance

In Chapter 2 we appraised the linearity of the VCO static tuning response by calculating its deviation from a linear fit. This approach was useful to qualitatively illustrate the record performance of the VCO, and to quantify the improvement obtained from the linearization of the VCO over the simple ICO. Still, to accurately quantify the effects of VCO nonlinearities in a working multichannel WFM system, it is necessary to perform measurements of CSO and CTB distortion. Such measurements would contain a number of contributors. First, the nonlinearities of the modulator quasi-static response. In addition dynamic effects, such as frequency dependent phase shifts that may occur in the limiting and output driver block, could play a role. Next, nonlinearities in the optical link components. And finally, further distortion introduced by nonlinearities in the response of the wideband demodulator.

In this complex situation we have made quasi-static estimates of distortion from the predicted linearity performance of the VCO. These provide an incomplete picture, but also useful guidance for future measurements, and allow us to place some bounds on attainable performance. It is important to keep in mind that the results we present in this section are exclusively focused on the distortion induced by the modulator. In that sense the experimental characterization of a full system setup is essential, and the present estimates must be compared with those measurements, once available.

These quasi-static calculations were performed using a MATLAB routine where the VCO transfer function was modeled with a polynomial fitting of the predicted static transfer function obtained by circuit simulation. A transfer function from static measurements of the VCO was not used, since the measurements obtained at this stage of our work suffered from what was thought to be short-term thermal drifting. This problem was corrected later by improving the accuracy of our static tests, and as seen in Fig. 2.6 (b) the match between
predicted and measured results is now excellent. For this reason the results presented in this section are also considered valid for the implemented VCO. For the simulations the optical link and demodulator were assumed ideal, the 80 channels were generated according to the NTSC frequency plan standard and with independent random phases, and the resulting CSO and CTB were averaged over a hundred iterations.

In general FM transmission, white noise in the link produces in the demodulated output a decreasing CNR with frequency. To compensate for this effect pre-emphasis of the highest frequency components, prior to transmission, is commonly used. However the use of pre-emphasis modifies the relative contribution of different carriers to the intermodulation beats, and consequently affects the CSO and CTB levels. For this reason the presence and absence of pre-emphasis was also included in the simulations.\(^3\)

In Fig 5.1 we show a plot of the transfer function used to represent a system composed of the VCO followed by an ideal demodulator. This transfer function was obtained from a polynomial fit of the static response of the VCO to a set of discrete inputs in the range of ±520mV. The VCO response was obtained from Hspice simulations and the polynomial fit is of order 20. The fit over the linear range of the modulator response is excellent; while the large deviations seen at the limits of the ±520mV range are typical of polynomial fits, and not expected to play a large role in the results.

\(^3\) Notice that the use, or not, of pre-emphasis is a system level consideration for a full WFM link. The pre-emphasis, if necessary, is performed to the multi-channel CATV signal prior to FM modulation.
Figure 5.1: System transfer function used to calculate CSO and CTB, for back-to back VCO and ideal demodulator. The transfer function is valid in the input range of ±520mV.
Fig. 5.2 is a plot of the resulting average CSO versus channel module number and CTB for the highest channel for cases with and without use of 6dB per octave pre-emphasis. Carrier amplitudes were set to 6.5mV for no pre-emphasis and to 18mV at the highest channel for pre-emphasis; these values were obtained through successive simulations to approximately reach the target limit of -65dBc. The dashed line at -65dBc indicates target CSO and CTB, thus increasing the carrier amplitudes would result in CSO levels above the -65dBc line. In the WFM system the carrier amplitudes used produce maximum per channel deviations, $\Delta f_{\text{max,CH}}$, of 14MHz and 40MHz for the signals without and with pre-emphasis respectively. On Fig. 5.2 CTB levels are shown with “0” symbols and are only for their worse case, i.e., highest channel. The resulting worse case CTB levels are -79dBc for no pre-emphasis and -65.5dBc for 6dB per octave pre-emphasis.

The plot in Fig. 5.2 allows comparing CSO performances with and without pre-emphasis, and provides an estimate of the maximum acceptable input amplitudes. These results show that the maximum allowed carrier amplitude, and thus $\Delta f_{\text{max,CH}}$, is higher if 6dB per octave pre-emphasis is used. In principle, carrier-to-noise performance improves with increasing frequency deviation. Therefore it will seem that the use of pre-emphasis is more advantageous to overall system performance. Despite this, as will be discussed in Section 5.3, CNR simulations revealed that for a narrowcast transmission system pre-emphasis is unnecessary. For this reason we will concentrate from this point forward exclusively on transmission without pre-emphasis; consequently, all carrier amplitudes are equal. Also, from Fig. 5.2 it can be seen that the highest channel presents the worst CSO level, thus CSO results are presented hereafter only for the highest channel.
Chapter 5. System Perspectives

Figure 5.2: CSO versus channel module number and CTB for highest channel. Dotted and crossed lines are for cases with no pre-emphasis and 6dB per octave pre-emphasis respectively. In each case carrier amplitudes were chosen to reach target CSO and CTB levels. ◊ Indicate CTB levels for channel 80, with values of -79dBc for no pre-emphasis and -65.5dBc for 6dB per octave pre-emphasis. Dashed line at -65dBc indicates target CSO and CTB.
Fig. 5.3 shows a plot of worst case CSO versus $\Delta f_{\text{max}_{\text{CH}}}$. The plot was built from successive simulations of CSO for different carrier amplitudes. The frequency deviations on the x-axis of Fig. 5.3 are calculated from the VCO voltage to frequency response. From the plot, the projected intersection point with the target line of $-65\text{dBc}$ corresponds to a deviation of 12.5MHz, which corresponds to a 5.8mV input amplitude; this deviation is close to the estimate of 14MHz used to generate Fig. 5.2. The fast increase of CSO distortion with increasing deviation shows that, despite the record linearity achieved with the linearized VCO, the transmission of 80 channels is demanding enough to limit the practical maximum frequency deviation, and thus channel amplitudes, to very low values. This in turn sets a limit to carrier powers and thus is expected to restrict the CNR performance. The corresponding analysis of CNR performance versus frequency deviation and its combined effect with distortion is discussed in the next section.
Figure 5.3: Worst case CSO versus per channel frequency deviation, $\Delta f_{\text{max,CH}}$, for 80-channel transmission and no pre-emphasis. The dotted line at -65dBc indicates target CSO.
Chapter 5. System Perspectives

5.3 Carrier-to-Noise Ratio (CNR) Performance

The CNR performance of the complete WFM system depends on the power of the transmitted carriers, the phase noise of the modulator, the shot noise of the photodetector, the laser source relative intensity noise (RIN), photodetector thermal noise, and demodulator output noise. For our CNR calculations we assumed an ideal demodulator, and used for the optical link components similar system parameters as in references [3,4], but adapted to narrowcast transmission.

The CNR of the jth AM carrier in the multichannel WFM system can be calculated using the following equations derived from work presented in [3,4]:

$$\text{CNR}_{j\text{AM}} = \frac{1}{2 \cdot B_{AM}} \frac{\Delta F_j^2}{f_j^2} \cdot \text{CNR}_{FM}$$  \hspace{1cm} (5.1)

With,

$$\text{CNR}_{FM} = \left(\frac{C_{FM}}{n_{FM}}(f)\right)^{-1} = \left(\frac{C_{MOD}}{n_{MOD}}(f)\right)^{-1} + \left(\frac{C_{ONU}}{n_{ONU}}(f)\right)^{-1}$$  \hspace{1cm} (5.2)

$$\frac{C_{MOD}}{n_{MOD}}(f) = \frac{\frac{1}{2}Is^2}{\sigma_{PN}(f)^2}$$  \hspace{1cm} (5.3)

$$\frac{C_{ONU}}{n_{ONU}}(f) = \frac{\frac{1}{2}m^2\eta^2P_r^2}{\sigma_{TH}^2 + \sigma_{PIN-SHOT}(f)^2 + \sigma_{PIN-RIN}(f)^2}$$  \hspace{1cm} (5.4)

Equation (5.2) describes the total CNR of the FM modulated signal after transmission, with $C_{FM}$ and $n_{FM}$ as the total FM carrier and noise powers. Parameters $C_{MOD}$ and $n_{MOD}$ are, respectively, the carrier and noise power at the transmitter, and parameters $C_{ONU}$ and $n_{ONU}$ are, respectively, the carrier and noise power at the receiver end. Equation (5.3) provides the CNR of the modulator and has been modified from the expression proposed in [3,4] to fit an
Chapter 5. System Perspectives

An electronic modulator rather than a heterodyne detection system. Equation (5.4) provides the CNR at the receiver end. In the expressions above, \( f_j \) and \( \Delta F_j \) are the AM carrier frequency and deviation of the jth carrier, \( B_{AM} \) is the AM bandwidth of one channel, \( \frac{1}{2}Is^2 \) is the transmitted power, \( \sigma_{PN}(f)^2 \) is the variance per unit frequency of the modulator phase noise, \( m \) is optical modulation depth, \( \eta \) is the responsivity for the receiver PIN photo-diode (PD), \( P_r \) is the power received, and \( \sigma_{TH}(f)^2, \sigma_{PIN-SHOT}(f)^2, \sigma_{PIN-RIN}(f)^2 \) are the variances per unit frequency of the PD thermal, shot and RIN noise. The performance of the modulator is reflected in these calculations through the phase noise variance; and also by the practical range of frequency deviations, which relate to the VCO central oscillation frequency, its voltage-to-frequency sensitivity, and its linear range. System values used for the calculations are: \( B_{AM} = 6 \text{MHz} \), transmitted power of 5dBm, worst case modulator phase noise of -113.5dBc/Hz at 50MHz from the carrier and decreasing at 20dB per decade, optical modulation depth of 0.9, PD responsivity of 0.7A/W, received power of 0dBm, thermal noise of 22 pA/\sqrt{Hz} , and laser relative intensity noise (RIN) of -159dB/Hz.

Calculations employing the system values listed above show that, due to the high received power and low DFB laser RIN, the CNR of the modulator dominates the overall CNR figure. Consequently, the CNR of the FM modulated signal, \( CNR_{FM} \), follows the pattern of the modulator CNR, increasing with frequency at a rate of 20dB per decade, i.e., 6dB per octave, due to the correspondingly decreasing phase noise. The effect this phase noise dominance has on the overall CNR is shown in Fig. 5.4 for cases with and without use of 6dB per octave pre-emphasis. Carrier amplitudes used are the same as in Fig. 5.2 to achieve minimum acceptable CSO performance.
Figure 5.4: CNR of the jth AM carrier versus carrier frequency for 80-channel transmission with and without use of 6dB per octave pre-emphasis. Dots represent channels. For 6dB per octave pre-emphasis: amplitude of top carrier is 18mV, maximum frequency deviation 40MHz, slope of response is 20dB per decade, i.e., 6dB per octave. For no pre-emphasis: amplitude of carriers is 6.5mV, maximum frequency deviation is 14MHz, and response is flat.
The results shown in Fig. 5.4 contrast with general systems where pre-emphasis is required to level the CNR off all carriers; in such cases the dominant source of noise is white which results in increasing FM demodulated noise with increasing frequency, thus the need to boost the amplitudes of the higher carriers. In the narrowcast system we are considering, the modulator phase noise dominates all sources of white noise in the frequency range of interest, and by decreasing at a rate of 20dB per decade counteracts the demodulation-induced increase of noise with frequency, producing a constant CNR response for all carriers. For no pre-emphasis, the plot also indicates that the CNR performance is below the minimum target of 52dB and is slightly worse at the highest channel. These results however are dependent on the chosen modulator phase noise and per channel deviation.

We performed further calculations to analyze CNR versus phase noise and deviation. In Fig. 5.5 we show a plot of CNR of the 80th channel, versus per channel frequency deviation for different values of phase noise. The phase noise is indicated as if measured at an offset of 50MHz from the FM carrier, and is increased in steps of 5dB from the minimum value of -113.5dBc/Hz. The choice of this minimum value was made based on preliminary crude phase noise measurements. The results on Fig. 5.5 show that in order to reach the target CNR of at least 52dB, the modulator phase noise and maximum frequency deviation must be at least -133.5dBc/Hz at 50MHz and 35MHz. For the linearized VCO implemented in this thesis the phase noise performance is -122.5dBc/Hz at 50MHz offset and the maximum frequency deviation is approximately 14MHz. It is evident from Fig. 5.5 that improving the phase noise alone will not help reach the target CNR. On the other hand increasing the maximum frequency deviation beyond the limit of 45MHz shown in the figure would eventually help reach the target CNR. The required deviation was calculated to be 54MHz, which represents an
Figure 5.5: CNR of channel 80 versus maximum per channel frequency deviation and modulator phase noise. Phase noise is indicated as measured at an offset of 50MHz from the FM carrier. Dotted line at 52dB indicates target CNR.
increase in the carrier amplitudes of a factor of 4. The CSO distortions are proportional to the
carrier magnitudes, the polynomial coefficient describing the second order nonlinearity, and
the number of distortion beats. All other variables constant it can be estimated that the second
order polynomial coefficient, thus roughly the deviations from linearity, would have to be
reduced by a factor of 4. This would represent an improvement of a factor of 40 over the
performance observed for the ICO. On the other hand, as discussed in Chapter 4, an
improvement of 10dB in phase noise seems feasible and thus a frequency deviation of 35MHz
would be enough to reach the target CNR. Thus, it can be concluded that to achieve the target
CNR it is more practical to improve both VCO phase noise performance, and extend the VCO
maximum frequency deviation.

The compromise between noise and distortion performance is illustrated in Fig. 5.6
where we show CNR versus CSO and phase noise for the highest carrier in an 80-channel
system. Circles correspond to frequency deviations spaced by 5MHz in the range of 5MHz to
40MHz. This plot summarizes the results presented so far in this section.

In conclusion for 80-channel transmission, the target CNR, CSO and CTB cannot be
achieved with the current version of the wideband modulator. Improving the performance of
the modulator will require both reducing phase noise by 10dB and increasing the maximum
possible frequency deviation from its current value of 14MHz to approximately 35MHz.
Another alternative to help reach the target CSO and CNR is to decrease the number of
transmitted channels, this would improve distortion performance by reducing the number of
intermodulation and harmonic beats, but should not affect the CNR if no pre-emphasis is used.
Fig. 5.7 shows the results for a system consisting of 40 channels. The plots on Fig. 5.7 are
equivalent to those discussed for an 80-channel system but the range of frequency deviations
Figure 5.6: Performance variables in 80-channel transmission: CNR versus CSO and phase noise. CNR and CSO are for highest channel. Phase noise is indicated as measured at an offset of 50MHz from the FM carrier. Circles are for points with equal frequency deviation. Dotted vertical line at -65dBc indicates target CSO. Dotted horizontal line at 52dB indicates target CNR.
Figure 5.7: Performance variables in 40-channel transmission: (a) CSO versus maximum per-channel frequency deviation. (b) CNR versus maximum per channel frequency deviation and modulator phase noise. (c) CNR versus CSO and phase noise. Circles are for points with frequency deviations spaced by 5MHz in the range of 5 to 60MHz. In all plots: CNR and CSO are for the highest channel, and phase noise is indicated as measured at a 50MHz offset from the FM carrier. Dotted lines at 52dB and -65dBc indicate target CNR and CSO.
has been increased from 40MHz to 60MHz to account for the larger per-channel amplitudes possible in a lower channel count. Fig. 5.7 (a) shows worst case CSO versus maximum frequency deviation, $\Delta f_{\text{max},CH}$. The intersection point with the –65dBc line is approximately 22MHz. Fig. 5.7 (b) is a plot of CNR for the 40th channel, versus frequency deviation and modulator phase noise at 50MHz offset. The results from Fig. 5.7 (a) and (b) are combined on Fig. 5.7 (c) in a plot of CNR versus CSO for the highest channel in a 40-channel system. The plots show a consistent CNR performance with a small improvement over to the slightly worse CNR at channel 80. The plots also show an improvement in maximum frequency deviation that would make possible a CNR of 52dB, with a CSO of –65dBc, by exclusively improving the phase noise by 10dB.

5.4 Summary and Conclusions

In this chapter we explored the perspectives for application of the linearized VCO in a WFM transmission system. For this purpose, quasi-static estimates of CSO and CTB performance were obtained from the predicted tuning response of the VCO assuming an ideal optical link, and ideal demodulator. Also CNR performance was calculated from the VCOs phase noise and frequency response for a typical narrowcast system with received powers of 0dBm, and an ideal demodulator. Results show that phase noise of the modulator dominates the CNR of the FM signal above photodetector shot noise and laser RIN noise. As a consequence, the drop of VCO phase noise with frequency offset makes unnecessary the use of pre-emphasis.

For 80-channel transmission the present performance of the VCO is short from reaching the target CSO and CTB of less than –65dBc, and the target CNR of at least
Chapter 5. System Perspectives

52dB. Measured performance metrics of the VCO are maximum acceptable frequency deviation of 14MHz and phase noise of $-122\text{dBc/Hz}$ at 5MHz offset. Calculations show that modulator linearity must be further improved to achieve a maximum frequency deviation of 54MHz. Alternatively, if the phase noise is improved by 10dB then the required frequency deviation is 35MHz. In contrast, for 40-channel transmission, linearity is adequate and only the phase noise improvement of 10dB is required. If such improvements were made possible without affecting tuning linearity, it would provide a solution for 40-channel transmission with record performance if compared to previous results reported in [2,3,4,5].
The work presented in this thesis resulted in three main contributions. The first contribution is the implementation and testing of record monolithic designs for WFM transmission of analog CATV. These designs proved attainable, for the first time, the transmission of 40 channels without use of pre-distortion, and hold promise for the first implementation of an 80-channel WFM system. In particular, the implementation of a fully electronic, monolithic, WFM modulator is proved feasible by using an ECM-based design. The viability of this approach is based on the ECMs wideband tuning linearity at high-speed operation, and a combination of the undemanding close-in phase noise requirements of CATV transmission with a simple and novel phase noise reduction technique devised in this thesis for high-speed ECMs.

The second contribution is the design and implementation of a high-accuracy, economic, custom measurement technique to obtain oscillator phase noise versus tuning input. The resulting measurement setup was used in our studies of ECM phase noise, and the phase noise results obtained with it were validated by circuit noise simulations. The third contribution is fundamental findings on the noise limits of high-speed ECM oscillators that lead to the aforementioned simple design optimization technique to improve phase noise performance, therefore increasing the application prospects of this topology in high-speed circuit design. In this chapter we will summarize the results and observations presented throughout this thesis and comment on their implications to the present and possible future work.
Based on the need for a fully electronic integrated WFM modulator, we described in Chapter 2 the design considerations that led to the selection of an ECM-based topology. Two circuits were presented based on this topology. The first is a simple current-controlled oscillator, ICO, which was implemented for the purpose of linearity and noise studies of the intrinsic ECM core. The second is based on a linearized voltage-controlled oscillator, VCO, to be used as the actual WFM modulator. The predicted and experimental results confirmed the viability of the electronic approach and the advantages of employing ECM-based designs, which achieved record performance compared to all previously reported optoelectronic [3,4,6] and monolithic [5] WFM modulators. Furthermore the wideband linearity of both the ICO and the VCO is noteworthy. To our knowledge both of these circuits are by far much more linear over a wider range than any previously reported high-speed controlled oscillators. The tuning performances of the ICO and VCO designs are summarized in Table 6.1. For comparison, Table 6.2 summarizes the performance of the most significant linear high-speed VCOs and WFM modulators reported to date.

<table>
<thead>
<tr>
<th>Performance parameters</th>
<th>Simple ICO</th>
<th>Linearized VCO</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_c$ Tuning range</td>
<td>1.7GHz (1.0 to 2.5) GHz</td>
<td>1.8GHz (1.5 to 2.5) GHz</td>
</tr>
<tr>
<td>Tuning as % of $f_c$</td>
<td>(-41, +47)%</td>
<td>(-17, +39)%</td>
</tr>
<tr>
<td>Max. deviation from linear fit</td>
<td>15MHz</td>
<td>1.5MHz</td>
</tr>
<tr>
<td>External input control range</td>
<td>(0.5 to 1.5) mA</td>
<td>(-200 to +200) mV</td>
</tr>
<tr>
<td>ECM core tail current range</td>
<td>(0.5 to 1.5) mA</td>
<td>(0.75 to 1.25) mA</td>
</tr>
</tbody>
</table>
Table 6.2: Comparison of significant high-speed oscillator designs and WFM modulators reported to date. Designs marked with # are originally not reported in terms of linear tuning but show a qualitatively linear region that is indicated in this table.

<table>
<thead>
<tr>
<th>Description</th>
<th>Ref.</th>
<th>$f_c$ (GHz)</th>
<th>Linear range (MHz)</th>
<th>Linear range as % of $f_c$</th>
<th>Phase noise (dBc/Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>LC oscillator #</td>
<td>[18]</td>
<td>2.10</td>
<td>400</td>
<td>± 9.5 %</td>
<td>−133.5 @ 600kHz</td>
</tr>
<tr>
<td>Ring oscillator #</td>
<td>[24]</td>
<td>1.18</td>
<td>361</td>
<td>± 15.3 %</td>
<td>−106.9 @ 600kHz</td>
</tr>
<tr>
<td>Electronic WFM modulator</td>
<td>[5]</td>
<td>3.95</td>
<td>2900</td>
<td>± 36.7 %</td>
<td>Not reported</td>
</tr>
<tr>
<td>Optoelectronic WFM modulator</td>
<td>[3,4]</td>
<td>1.7</td>
<td>Not reported</td>
<td>−</td>
<td>−113.7 @ 91MHz</td>
</tr>
<tr>
<td>Relaxation oscillator #</td>
<td>[28]</td>
<td>0.39</td>
<td>485</td>
<td>± 62.2 %</td>
<td>−99 @ 2MHz</td>
</tr>
<tr>
<td>Linearized VCO</td>
<td>This work</td>
<td>1.8</td>
<td>1000</td>
<td>−17, +39 %</td>
<td>−102.5 @ 5MHz</td>
</tr>
</tbody>
</table>

The LC, ring, and relaxation oscillator included in Table 6.2 were selected from those described in Table 1.1, Section 1.4, and represent the designs with widest tuning linearity of their kind. These three oscillators are originally not reported in terms of linear tuning performance but exhibit a qualitatively region that is indicated in Table 6.2, none of which, however, is qualitatively as good as that of the linearized VCO. As for the electronic and optoelectronic WFM modulators in Table 6.2, their poor linearity has restricted these designs to no more than 20-channel systems with the use of external pre-distortion. From the previous comparison of tuning range and linearity the results obtained in this thesis are clearly significant. In addition it was found for the ICO that the qualitative shape of its deviation from linearity is intrinsic to the ECM core, and is near the fundamental limit set by the finite multivibrator switching delay. However it was also shown for the VCO that linearity of the ECM is effectively improved by linearizing the tail control circuit with negative feedback. This technique resulted in an improvement of a factor of 10 in the deviation from linearity. A simple static analysis of the interaction between the transadmittance amplifier and ECM nonlinearities failed to explain the
improved linearity of the VCO, which may be caused by a more complex dynamic interaction between these two blocks. Beyond the achievement of implementing the best WFM modulator to date, these results also represent a first time demonstration of ECM suitability in a high-speed application. In that sense, the ECM wideband linearity demonstrated may prove valuable for other high frequency applications.

A wideband demodulator was also presented for experimental testing of the VCO. Its design followed the architecture of a previous hybrid design presented in [3,4] and a monolithic design presented in [5], and as such constitutes no novelty. As the oscillators the demodulator was also implemented in Nortel's NT25 Advanced Silicon Bipolar process, with an $f_T$ of 20GHz. The resulting demodulator achieved a linear range from 1GHz to 3.5GHz to match that of the modulator. This performance is better than that of the hybrid demodulator in [3,4] but 2GHz short of the monolithic demodulator in [5], which however was implemented in a process with an $f_T$ of 40GHz. Also, the first reported back-to-back modulator/demodulator measurements were presented for the VCO and the delay-line demodulator, confirming a combined 3dB modulation bandwidth of 1GHz.

In general, the match between predicted and measured results is excellent. This is noteworthy, considering the highly nonlinear nature of the circuits, the frequency ranges involved, and the difficult computations performed, such as prediction of cyclostationary noise in the ECM. In that sense, the quality of the NT25 device models and the suitability of the circuit design techniques described in Chapter 2 proved to be invaluable. This stresses the quality of state-of-the-art high-speed design tools, and the accuracy that can be expected from careful circuit design prior to fabrication.
Chapter 6. Summary and Conclusions

As expected, the record tuning linearity of the ECM-oscillators resulted in low phase noise levels, which are $-128\text{dBc/Hz}$ and $-122.5\text{dBc/Hz}$ at 5MHz offset, for the ICO and VCO respectively. The 5.5dB lower phase noise of the VCO is due to additional noise contributions from the transadmittance amplifier block, Y. However, circuit analysis and simulations showed that noise could be lowered through careful redesign of the input stage of the Y block and reselection of the feedback resistors.

The best optoelectronic modulator reported to date has a phase noise of $-113.7$ at 91MHz offset from a carrier of 1.7GHz [3,4], where 91MHz is the location of the first CATV channel in the Japanese frequency allocation system. Extrapolating the phase noise of the linearized VCO to 91MHz offset, assuming a $-20\text{dB}$ per decade slope, results in $-127.7\text{dBc/Hz}$, which represents an improvement of 14dB over that of the best optoelectronic modulator. From Table 6.2, it can be seen that the phase noise performance of the linearized VCO is lower than that of the best LC and ring oscillators, which in contrast do not provide for wideband linearity. Still, as explained in Chapter 5, the actual phase noise performance requires 10dB improvement for 40 and 80-transmission, and motivated us to pursue a study on noise origins in high-speed ECMs.

In Chapter 3 we described a custom technique to measure phase noise versus tuning input at a fixed offset from the carrier. Although specifically engineered for our experimental needs, this technique is significant in its economy and accuracy, surpassing in both of these aspects the best dedicated phase noise instrument available at the time. The custom measurement approach is based on automatic capture of the oscillator output with a spectrum analyzer, followed by mathematical post processing. Particular effort was invested to reduce interference pickup, compensate for oscillator thermal drifting, and account for several inaccuracies originating from the spectrum analyzer. The
resulting setup is capable of reliable measurements with an accuracy of $\pm 2\text{dB/Hz}$, and with a minimum floor of $-118\text{dBc/Hz}$ at 5MHz offsets. The sweeping tuning variable can either be a current in minimum steps of micro Amperes, or a voltage in minimum steps of micro Volts. The frequency range of the measurements is from a few MHz to 2.9GHz and from 2.75GHz to 6GHz; where the partitioning is caused by the analyzer automatically switching between internal local oscillators. Nonetheless, the measuring range can easily be extended to match that of the spectrum analyzer, which is 26.5GHz. The main limitations of this measurement approach are its relatively high phase noise floor, and the impossibility to make accurate measurements at close-in offsets. In general, however, the technique presented here still represents a simple, attractive, and flexible solution, and its performance was more than adequate for our experimental purposes.

In Chapter 4 we presented the results from our study of the origins of noise in high-speed ECMs. Experimental and predicted results showed new findings in ECM phase noise behavior. Previous theories predict dominance in phase noise generation from thermal fluctuations of the multivibrator regenerative thresholds. This dominance results from the accumulative contribution from a frequency folding mechanism, and has been experimentally confirmed for designs operating at low speeds relative to their transistor cutoff frequency [30, 35–37]. The experimental study presented in Chapter 4, constitutes the first investigation of phase noise generation in ECMs at GHz operation, and at a higher ratio of transistor cutoff frequency. Our results show that as the frequency of the ECM is increased, the shot noise of the ECM core transistors becomes the dominant source of phase noise. Furthermore it is shown that phase noise is proportional to the ratio $I_{\text{ctrl}}/C^2$; where $I_{\text{ctrl}}$ is the tuning tail current and $C$ the value of the timing capacitor. Since the ECM output frequency is proportional to the ratio $I_{\text{ctrl}}/C$, it follows
that scaling the current and capacitor by the same factor maintains the output frequency while reducing phase noise. This was confirmed with simulations, for the ICO at one tuning point, for scaling factors of 2 and 5, with resulting improvements of 3 and 7dB as expected. The effects of this scaling approach on tuning linearity remain an area for further study; nonetheless, preliminary simulations indicate that the effect is minimal particularly when compared with that of other design approaches reported in literature to improve ECM phase noise. As a consequence of the scaling design approach proposed here, the phase noise of the VCO can be improved without affecting the ECM topology, thus maintaining wideband tuning, and minimizing the loss of linearity.

These findings have several implications. First, they make feasible improving the ECM for 40 and 80-channel transmission. Previous WFM modulator designs have been limited both in linearity and phase noise performance, and for that reason have been confined to low channels count proof-of-principle systems. Over the last 10 years, other groups have not been able to solve these limitations despite considerable efforts. As a consequence the linearity provided by the ECM designs with now the possibility of improving the phase noise constitutes an important achievement in this area. Nonetheless, the implementation of an improved modulator and its testing in a full system is still a pending task. A second implication is the need for further study of ECM phase noise generation.

In Chapter 5, we presented system calculations to explore the performance of the VCO in a narrowcast WFM system. For CSO and CTB calculations we assumed ideal fiber link, and demodulator. For CNR calculations we assumed an ideal demodulator. The results are therefore estimates of VCO performance and for the purposes of a practical system implementation a full experimental characterization is still required. In
addition, the calculations are based on a quasi-static approach, and as such do not account for any possible frequency dependent distortion effects. The results however were useful in several respects. First they showed that in a narrowcast transmission the FM CNR is dominated by the phase noise of the modulator. As a consequence of this it was found that pre-emphasis is not needed. Second the calculations showed that based on the actual performance of the VCO, 40-channel transmission is only possible if the phase noise is improved in 10dB, while with the same noise reduction, 80-channel transmission still requires improving the modulator linearity.

In conclusion, we have successfully fulfilled our initial goal of developing fully electronic circuits for WFM transmission of analog CATV. In addition this work has contributed to enlarge the understanding and the application range of ECM circuits, and has produced a low cost but accurate approach to phase noise measurement. However, future work is still required to apply the results of this thesis for the first implementation of a viable WFM system for 40 and 80-channel transmission.
Bibliography


