Design and Evaluation of a Low-Cost X-Band Synthesizer for LMDS Applications

By

Srdjan Suvakov

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Faculty of Engineering
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Acceptance of the thesis

Design and Evaluation of a Low-Cost X-Band Synthesizer for LMDS Applications

Submitted by

Srdjan Suvakov, B.Eng.
In partial fulfillment of
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Master of Applied Science

Dr. James Stuart Wight
Thesis Supervisor

Dr. M. S. Nakhla
Chair, Department of Electronics

Carleton University
Ottawa, Canada
October, 2002

Abstract

Inspired by the availability of the low-cost, high performance L-Band synthesizers and X-Band dividers, the design and performance of a low-cost X-Band synthesizer is evaluated. The X-Band synthesizer is designed for LMDS base station applications below 25GHz with high stability requirements, lower level modulation schemes and higher modulation rates. The X-Band VCO is an MHMIC design using the NEC321000 chip as its active element, with a low-cost printed spiral inductor to form the resonator and resistive buffering at its output. The L-Band synthesizer can be any off-the-shelf low-cost device. The reference frequency is obtained from a low-cost 10MHz TCVCXO.

The phase noise performance of the built synthesizer is evaluated in terms of the residual signal to noise ratio at the demodulator input. Link margin degradation due to synthesizer phase noise is then calculated for lower modulation schemes and higher symbol rates.

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Contents

ABSTRACT	III
ACKNOWLEDGMENTS	IV
LIST OF FIGURES	VIII
LIST OF TABLES	XI
LIST OF SYMBOLS	XII
1 INTRODUCTION.....	1
1.1 LOCAL OSCILLATORS IN WIRELESS BROADBAND RADIOS.....	1
1.2 LOCAL MULTIPOINT DISTRIBUTION SERVICE (LMDS)	2
1.2.1 <i>LMDS Band Allocation</i>	2
1.2.2 <i>LMDS Network Architecture</i>	3
1.2.3 <i>Modulation Methods for LMDS</i>	5
1.2.4 <i>LMDS Standards</i>	6
1.3 LOCAL OSCILLATORS IN LMDS.....	7
1.3.1 <i>CPE Local Oscillators</i>	7
1.3.2 <i>BTS Local Oscillators</i>	8
1.4 LOW-COST LOCAL OSCILLATOR FOR LMDS.....	9
1.5 SUMMARY.....	10
2 BACKGROUND INFORMATION	11
2.1 THE OSCILLATION CRITERION	11
2.2 FREQUENCY STABILITY	13
2.3 RESONANT VCOS	14
2.3.1 <i>Noise Through a Bandpass Filter</i>	16
2.3.2 <i>Phase Noise in Resonant Oscillators</i>	21
2.3.3 <i>L-Band Oscillator Phase Noise Example</i>	26
2.3.4 <i>Measuring Φ_{osc} Using a Spectrum Analyzer</i>	28

2.3.5	<i>Amplitude Modulation in Resonant Oscillators</i>	30
2.3.6	<i>Amplitude and Phase Noise Comparison</i>	30
2.4	THEORY OF NEGATIVE-RESISTANCE OSCILLATORS	31
2.4.1	<i>One-Port Negative-Resistance Oscillators</i>	32
2.4.2	<i>Example for the Stable Oscillation Conditions</i>	34
2.4.3	<i>Two-Port Negative-Resistance Oscillators</i>	38
2.4.4	<i>The Design Procedure for Two-Port Oscillator</i>	41
2.5	FREQUENCY SYNTHESIZERS.....	41
2.5.1	<i>Simple Frequency Synthesizer</i>	42
2.5.2	<i>L-Band Synthesizer Block Diagram</i>	42
2.5.3	<i>PLL Forward Loop Gain</i>	43
2.5.4	<i>Phase Noise in the PLL Loop</i>	44
2.6	QAM PHASE NOISE REQUIREMENTS	46
2.6.1	<i>QAM Demodulator</i>	47
2.6.2	<i>BER vs. SNR (Signal to Noise Ratio)</i>	49
2.6.3	<i>Residual $(S/N)_{res}$ and Fade Margin Degradation</i>	50
2.6.4	<i>Carrier Recovery PLL</i>	52
3	DESIGN AND SIMULATION	57
3.1	X-BAND VCO DESIGN AND SIMULATION	57
3.1.1	<i>X-Band DRO</i>	57
3.1.2	<i>X-Band VCO Design Specification</i>	59
3.1.3	<i>X-Band VCO Design Topology</i>	60
3.1.4	<i>Active Device Choice</i>	61
3.1.5	<i>X-Band VCO Linear Circuit Simulation</i>	63
3.1.6	<i>Device Nonlinear Model Development</i>	64
3.1.7	<i>X-Band VCO Nonlinear Circuit Simulation</i>	67
3.1.8	<i>X-Band VCO Simulation Results</i>	71
3.1.9	<i>X-Band VCO Layout</i>	74
3.2	SYNTHESIZER DESIGN AND SIMULATION	76
3.2.1	<i>X-Band Synthesizer Block Diagram</i>	76
3.2.2	<i>Synthesizer Design Parameters</i>	77
3.2.3	<i>PLL Loop Filter Design</i>	78
4	MEASUREMENTS AND PERFORMANCE	81
4.1	X-BAND VCO TEST RESULTS.....	81
4.1.1	<i>VCO Tuning Range</i>	82
4.1.2	<i>VCO Output Power</i>	84
4.1.3	<i>VCO Phase Noise</i>	84
4.2	X-BAND SYNTHESIZER TEST RESULTS	85
4.2.1	<i>L-Band Output Test Results</i>	87
4.2.2	<i>VCO Free Running Phase Noise</i>	91
4.2.3	<i>X-Band Output Test Results</i>	92
4.2.4	<i>Synthesizer Overall Performance</i>	95
4.3	X-BAND SYNTHESIZER $(S/N)_{RES}$	97

4.3.1	<i>Weighted Phase Noise</i>	97
4.3.2	<i>Link Margin Degradation</i>	98
4.4	COST ANALYSIS.....	99
5	CONCLUSIONS	100
5.1	ACCOMPLISHMENTS.....	100
5.2	FUTURE WORK	101
6	BIBLIOGRAPHY	103

List of Figures

<i>Figure 1–1: FCC LMDS band allocation.....</i>	<i>3</i>
<i>Figure 1–2: LMDS network architecture</i>	<i>4</i>
<i>Figure 1–3: CPE local oscillator chain.....</i>	<i>8</i>
<i>Figure 1–4: BTS local oscillator chain with PLDRO.....</i>	<i>9</i>
<i>Figure 1–5: BTS LO chain with X-Band synthesizers</i>	<i>10</i>
<i>Figure 2–1: The basic structure of a sinusoidal oscillator.....</i>	<i>11</i>
<i>Figure 2–2: Oscillator frequency stability vs. slope of the phase response</i>	<i>13</i>
<i>Figure 2–3: Principle of the resonant oscillator</i>	<i>14</i>
<i>Figure 2–4: Resonant VCO</i>	<i>15</i>
<i>Figure 2–5: Signal and noise through a bandpass filter</i>	<i>16</i>
<i>Figure 2–6: Power spectral density (psd) of the filtered noise $n_f(t)$</i>	<i>17</i>
<i>Figure 2–7: Orthogonal noise components, psd.....</i>	<i>18</i>
<i>Figure 2–8: Noise induced phase</i>	<i>19</i>
<i>Figure 2–9: Spectral density of noise induced phase</i>	<i>20</i>
<i>Figure 2–10: Resonator oscillator with noise voltage $n_s(t)$.....</i>	<i>21</i>
<i>Figure 2–11: Psd of the oscillator noise $n_s(t)$.....</i>	<i>22</i>
<i>Figure 2–12: Phasor diagram within the oscillator bandwidth</i>	<i>23</i>
<i>Figure 2–13: L-Band oscillator phase noise example, $\Phi_{\theta_{osc}}$ in rad^2/Hz</i>	<i>27</i>
<i>Figure 2–14: L-Band oscillator phase noise example, $\Phi_{\theta_{osc}}$ in dBc/Hz</i>	<i>29</i>
<i>Figure 2–15: Amplitude and phase noise comparison</i>	<i>31</i>

Figure 2–16: One-port negative-resistance oscillator	32
Figure 2–17: Negative-resistance device	35
Figure 2–18: Linear amplitude variation of $G(V)$	35
Figure 2–19: Two-port oscillator model	39
Figure 2–20: Simple frequency synthesizer	42
Figure 2–21: LMX-2326 synthesizer with 3 rd order loop filter	43
Figure 2–22: PLL equivalent AC model	44
Figure 2–23: Output phase noise from the free running VCO	45
Figure 2–24: Typical QAM demodulator	48
Figure 2–25: QAM constellation diagram, $M=64$	48
Figure 2–26: BER vs. S/N for M-QAM	50
Figure 2–27: Phase margin degradation for QPSK and 16-QAM	52
Figure 2–28: Model for carrier recovery PLL	54
Figure 2–29: PLL suppression function for $\zeta = 0.7$	56
Figure 3–1: Reflection type DRO	58
Figure 3–2: X-Band VCO design topology	60
Figure 3–3: NE321000 chip, dimensions and layout	62
Figure 3–4: NE321000 chip parasitics	67
Figure 3–5: X-Band VCO simulation file	68
Figure 3–6: VCO output attenuator simulation file	69
Figure 3–7: Test-bench for oscillator analysis	70
Figure 3–8: Oscillation frequency vs. varactor capacitance	71
Figure 3–9: Output power vs. varactor capacitance	72
Figure 3–10: Low-cost X-Band VCO layout	75
Figure 3–11: X-Band synthesizer, block diagram	76
Figure 3–12: Φ_{pd} , 40KHz and 100KHz loop bandwidth, L-Band	79
Figure 3–13: High pass filtering of the free running VCO phase noise	80
Figure 4–1: Two X-Band VCOs	82
Figure 4–2: Measured VCO tuning range	83
Figure 4–3: Measured L-Band output vs. control voltage	83

<i>Figure 4–4: VCO output power vs. control voltage</i>	<i>84</i>
<i>Figure 4–5: X-Band divide by 8 and L-Band synthesizer</i>	<i>86</i>
<i>Figure 4–6: Synthesizer at L-band</i>	<i>87</i>
<i>Figure 4–7: Phase noise @ 1KHz offset, L-Band.....</i>	<i>88</i>
<i>Figure 4–8: Phase noise @ 10KHz offset, L-Band.....</i>	<i>88</i>
<i>Figure 4–9: Phase noise @ 100KHz offset, L-Band.....</i>	<i>89</i>
<i>Figure 4–10: Phase noise in $\pm 500\text{KHz}$ offset, L-Band.....</i>	<i>89</i>
<i>Figure 4–11: Reference suppression at L-Band</i>	<i>90</i>
<i>Figure 4–12: Synthesizer phase noise at L-Band</i>	<i>91</i>
<i>Figure 4–13: Phase noise @ 1KHz offset, X-Band</i>	<i>92</i>
<i>Figure 4–14: Phase noise @ 10KHz offset, X-Band</i>	<i>93</i>
<i>Figure 4–15: Phase noise @ 100KHz offset, X-Band</i>	<i>93</i>
<i>Figure 4–16: Phase noise in $\pm 500\text{KHz}$ offset, X-Band.....</i>	<i>94</i>
<i>Figure 4–17: Calculated and measured phase noise at X-band.....</i>	<i>95</i>
<i>Figure 4–18: Weighted phase noise, $\zeta=0.7$, $T_d=0$</i>	<i>98</i>

List of Tables

<i>Table 1-1: LMDS modulation methods.....</i>	<i>6</i>
<i>Table 2-1: Selected parameters for some M-QAM systems.....</i>	<i>47</i>
<i>Table 3-1: DDBS DRO specification.....</i>	<i>58</i>
<i>Table 3-2: X-Band VCO design specification.....</i>	<i>59</i>
<i>Table 3-3: NEC321000 specification.....</i>	<i>62</i>
<i>Table 3-4: LIBRA parameter definition for the TOM model.....</i>	<i>66</i>
<i>Table 3-5: Simulated VCO phase noise.....</i>	<i>73</i>
<i>Table 3-6: Synthesizer design parameters, L-Band.....</i>	<i>78</i>
<i>Table 3-7: Loop filter components values</i>	<i>78</i>
<i>Table 4-1: Loop components for 75KHz bandwidth.....</i>	<i>85</i>
<i>Table 4-2: X-Band synthesizer overall performance.....</i>	<i>96</i>
<i>Table 4-3: Synthesizer residual noise contribution</i>	<i>98</i>
<i>Table 4-4: Cost analysis for active parts.....</i>	<i>99</i>

List of Symbols

$A(s)$	Forward gain response
$A_f(s)$	Gain with feedback
B	Receiver bandwidth
B_w	Noise bandwidth of the input bandpass filter
B_o	Open loop gain, $G_L(f)$ zero dB crossing
BER	Bit error rate
C	Capacitance of the negative resistance device
C_v	Varactor diode junction capacitance
C/N	Carrier to noise level
D	Fade margin
NF, F	Receiver noise figure
$F(s)$	Loop filter transfer function
F_s	Nyquist bandwidth
f_o	Frequency of oscillation, resonant frequency of the tank circuit;
f_a	Flicker noise corner frequency
f_b	Half-bandwidth of the resonant oscillator tank
f_c	Half-bandwidth of the filter following oscillator
f_i	Center frequency of the input bandpass filter
f_m	Offset frequency
f_{STEP}	Synthesizer step frequency
f_R	Reference crystal frequency
$G(s)$	PLL forward loop gain
$G(V)$	Negative conductance linear amplitude variation
$G_{amp}(f)$	Loop amplifier gain
$G_L(f)$	Carrier recovery open loop gain
$G_{LP}(f)$	Demodulator low pass filter response
$ G_{Nq} $	Nyquist filter amplitude response
g_m	Amplifier transconductance
$H_{pd}(s)$	PLL phase transfer function from θ_{pd} to θ_{opd}
$H_{vco}(s)$	Transfer function from θ_{vco} to θ_{ovco}
I_{cp}	Charge pump current gain
$I(s)$	Phase detector output current
K	PLL loop gain
k	Boltzmann constant 1.38×10^{-23} [JK ⁻¹]
K_{vco}	VCO gain
K_{cp}	Phase detector gain
$L(s)$	Loop gain
$n_{uf}(t)$	Noise before input bandpass filter
$n_f(t)$	Noise after input bandpass filter
$n_x(t), n_y(t)$	Orthogonal noise components of noise $n_f(t)$

$n_s(t)$	Noise generated in the resonant oscillator before oscillations start
N_0	Power spectral density of white noise
N	Frequency divider ratio in the feedback path
P_{osc}	Signal carrier power in dBm
P_e	Bit error probability
Q	RLC tank circuit quality factor
$Q(x)$	Marcum Q function
R	Reference divider ratio
R_{IN}	Real part of two-port network input impedance Z_{IN}
R_L	Real part of the load impedance Z_L
RSL	Received signal level
r_p	Tank resistance
S/N	Average signal power over noise in Nyquist bandwidth
$(S/N)_{res}$	Residual signal to noise ratio
$(S/N)_\delta$	Signal to noise ratio required for $BER = 10^{-6}$
T	Temperature in K
$T(s)$	Transfer function of bandpass RLC tank
T_d	Time delay of Nyquist filter
$v_i(t)$	Input voltage to bandpass filter
$v_o(t)$	Output voltage from bandpass filter
V_s	Peak value of $v_i(t)$ and $v_o(t)$
$v_{osc}(t)$	Resonator oscillator output voltage
V_{osc}	Peak value of $v_{osc}(t)$
V_R	Reverse bias for varactor diode
$x(t)$	Random amplitude modulation
$x_{osc}(t)$	Random amplitude modulation of the resonant oscillator
X_{IN}	Imaginary part of two-port network input impedance Z_{IN}
X_L	Imaginary part of the load impedance Z_L
$y(t)$	Random baseband noise
$W(s)$	Transfer function from local oscillator phase to demodulator output phase
$ W(f) $	Carrier recovery loop suppression function
Z_{IN}	Two-port network input impedance
Z_L	Two-port network load impedance
Z_T	Two-port network terminating impedance
Z_{OUT}	Two-port network output impedance
Z_0	Characteristic impedance of the port
$\beta(s)$	Feedback response
Δ	Determinant of S-parameter matrix
$\Phi_{\theta_i}(f)$	Single sided spectral density of random phase modulation $\theta_i(t)$
$\Phi_\phi(f)$	Single sided spectral density of compensating phase shift $\phi(t)$
$\Phi_{\theta_{osc}}(f)$	Single sided spectral density of oscillator phase modulation $\theta_{osc}(t)$
$\Phi_x(f)$	Single sided power spectral density of random amplitude modulation $x(t)$
$\Phi_y(f)$	Single sided power spectral density of random signal $y(t)$
$\Phi_{n_{uf}}(f)$	Single sided power spectral density of noise $n_{uf}(t)$
$\Phi_{n_f}(f)$	Single sided power spectral density of noise $n_f(t)$
$\Phi_{n_s}(f)$	Single sided power spectral density of noise $n_s(t)$
$\Phi_{\omega_i}(f)$	Single sided spectral density of random frequency modulation $\omega_i(t)$
$\Phi_{x_{osc}}(f)$	Single sided power spectral density of random amplitude modulation $x_{osc}(t)$
$\Phi_{pdfloor}(f)$	Single sided phase detector phase noise floor
$\Phi_{\theta_{vco}}(f)$	VCO free running, single sided phase noise spectral density
$\Phi_{pd}(f)$	Single sided phase noise at PLL output from phase detector noise
$\Phi_{vco}(f)$	Single sided phase noise at PLL output from VCO noise

$\Phi_n(f)$	Single sided phase noise at PLL output from VCO and phase detector noise, measured synthesizer phase noise
$\phi(t)$	Random phase, compensating phase shift in the tank circuit, signal phase
Γ_T	Reflection coefficient for terminating network
Γ_{OUT}	Two-port output reflection coefficient
Γ_{IN}	Two-port input reflection coefficient
Γ_L	Two-port reflection coefficient
$\lambda_{\theta LO}$	Local oscillator single sided phase noise spectral density
Θ_o	Power spectral density of white phase noise
$\theta(t)$	Random phase
$\theta_i(t)$	Random phase modulation of $v_o(t)$, phase of PLL input signal
$\theta_{osc}(t)$	Phase modulation of the resonant oscillator output $v_{osc}(t)$
$\theta_o(t)$	Phase of PLL output signal
$\theta_{pd}(t)$	Phase error from the phase detector
$\theta_{opd}(t)$	Phase at the PLL output due to $\theta_{pd}(t)$
$\theta_D(t)$	Phase error at demodulator output
θ_{Drms}	Rms phase noise at demodulator output
$\theta_R(t)$	Phase error of received IF signal, demodulator input phase error
$\theta_{vco}(t)$	VCO phase noise
$\theta_{ovco}(t)$	Phase at the PLL output due to $\theta_{vco}(t)$
ω_o	Resonator oscillator natural frequency, tank natural frequency
$\omega_{osc}(t)$	Frequency modulation of the resonant oscillator output $v_{osc}(t)$
ω_i	Center frequency of the input bandpass filter
ζ	PLL damping factor

Chapter One

1 Introduction

1.1 Local Oscillators in Wireless Broadband Radios

In wireless broadband radios, design of the local oscillators in the microwave range has to be cost-effective with excellent performance in terms of frequency stability and phase noise. One of the broadband wireless applications is *LMDS* (local multipoint distribution service). Current design of the local oscillators for base stations in the point-to-multipoint radios uses PLDROs (phase locked dielectric resonator oscillators) as the microwave source. They have known advantages in stability and in phase noise performance. Their disadvantage is their cost, sometimes up to 20% of the overall cost for the base station radio.

With the availability of inexpensive components for the design of X-Band synthesizers, it might be possible to use them instead of PLDROs for broadband wireless applications below 25GHz. In this work we would investigate the design of an X-Band low cost synthesizer and the possibility and limitations for its implementation as the microwave source in LMDS radios. Of course, this can be applied to other broadband wireless applications.

1.2 Local Multipoint Distribution Service (LMDS)

LMDS or *LMCS* (local multipoint communication systems), as the technology is known in Canada, is a wireless, two-way broadband technology designed to allow network integrators and communication service providers to quickly and inexpensively bring a wide range of quality services to homes and businesses. Services using LMDS technology include high-speed Internet access, real-time multimedia file transfer, remote access to corporate local area networks, interactive video, video-on-demand, video conferencing, and telephony among other potential applications. Many developing countries see this technology as a way to bypass the expensive implementation of cable or fiber optics, and leapfrog into the twenty-first century.

1.2.1 LMDS Band Allocation

In 1998, the U.S. *FCC* (Federal Communications Commission) carried out an auction of spectrum in the 28-31 GHz range for a service known as LMDS. In each geographical area, the FCC auctioned an "A block" (with bandwidth of 1150 MHz) and a "B block" (with bandwidth of 150 MHz), [1], Figure 1–1.

The LMDS bandwidth is by far the largest ever auctioned. For instance, using just the 850 MHz "downstream" band, one system developer expects to be able to offer 76 digital broadcast video channels to all users in a given cell while setting aside 1555 Mbit/s (equivalent to 1080 T-1 lines) for interactive data channels. This could be allocated into many channels (for example, 370 residential lines at 4 Mb/s each) or only a few channels (for commercial users with large scale data requirements). Similar services can be provided to the users in adjacent cells; the frequencies can be "reused" due to the short

propagation path over 20GHz radio waves and, in some schemes, by controlling the wave polarization.

Source: Federal Communications Commission

Figure 1–1: FCC LMDS band allocation

LMDS is not confined to the United States. LMDS operating frequencies around the world are in 20-40GHz range, K-Band and Ka-Band, [2].

1.2.2 LMDS Network Architecture

Various network architectures are possible within LMDS system design. The majority of system operators will be using point-to-multipoint wireless access designs, although point-to-point systems and TV distribution systems can be

provided within the LMDS system. The LMDS network architecture consists of primarily four parts, Figure 1–2.

- network operations center – *NOC*
- fiber-based infrastructure
- base station – *BTS*
- customer premises equipment – *CPE*.

Figure 1–2: LMDS network architecture

The NOC (network operations center) contains the *NMS* (network management system equipment) that manages large regions of the customer network. Multiple NOCs can be interconnected.

The fiber-based infrastructure typically consists of *SONET* (synchronous optical network) optical carrier OC-12, OC-3, and DS-3 links; *CO* (central-office) equipment; *ATM* and *IP* switching systems; and interconnections with the Internet and *PSTNs* (public switched telephone networks).

BTS (base station) equipment includes the network interface for fiber termination, modulation and demodulation functions and microwave transmission and reception equipment. If local switching is present, customers connected to the

base station can communicate with one another without entering the fiber infrastructure. This function implies that billing, channel access management, registration, and authentication occur locally within the base station. The alternative base-station architecture simply provides connection to the fiber infrastructure. This forces all traffic to terminate in ATM switches or CO equipment somewhere in the fiber infrastructure. In this scenario, if two customers connected to the same base station wish to communicate with each other, they do so at a centralized location. Billing, authentication, registration, and traffic-management functions are performed centrally.

The customer-premises configurations vary widely from vendor to vendor. Primarily, all configurations will include outdoor-mounted microwave equipment and indoor digital equipment providing modulation, demodulation, control, and customer-premises interface functionality. The CPE (customer premises equipment) may attach to the network using *FDMA* (frequency-division multiple access), *TDMA* (time-division multiple access) or *CDMA* (code-division multiple access) methodologies. The customer premises interfaces will run the full range from digital signal, level 0 (DS-0), *POTS* (plain old telephone service), 10BaseT, unstructured DS-1, structured DS-1, frame relay, ATM25, serial ATM over T1, DS-3, OC-3, and OC-1. The customer premises locations will range from large enterprises (e.g., office buildings, hospitals, campuses), in which the microwave equipment is shared between many users, to mall locations and residences, in which single offices requiring 10BaseT and/or two POTS lines will be connected. Obviously, different customer-premises locations require different equipment configurations and different price points.

1.2.3 Modulation Methods for LMDS

Modulation methods for broadband wireless LMDS systems are generally separated into *PSK* (phase shift keying) and *QAM* (quadrature amplitude

modulation) approaches. The modulation options for TDMA and FDMA access methods are almost the same.

The access modulation methods are listed in Table 1-1 and are rated to the amount of bandwidth they require for a 2-Mbps *CBR* (constant bit rate) connection, [1].

<i>Name</i>	<i>2 Mbps CBR</i>
BPSK	2.8 MHz
DQPSK	1.4 MHz
QPSK	1.4 MHz
8-PSK	0.8 MHz
4-QAM	1.4 MHz
16-QAM	0.6 MHz
64-QAM	0.4 MHz

Table 1-1: LMDS modulation methods

The TDMA link modulation methods typically do not include the 64-QAM, although this might become available in the future.

1.2.4 LMDS Standards

LMDS standards activities currently underway include activities by the *ATM* Forum, the *DAVIC* (Digital Audio Video Council), the *ETSI* (European Telecommunications Standards Institute) and the *ITU* (International Telecommunications Union), [1]. The majority of these methods use ATM cells as the primary transport mechanism.

1.3 Local Oscillators in LMDS

The BTS and CPE are where the conversion from fibered infrastructure to wireless infrastructure occurs. As LMDS operating frequencies are in 20-40GHz range, both base station and customer premises equipment would require *LOs* (local oscillators) in the microwave frequency range.

LO phase noise requirements are the same for both BTS and CPE and depend on the modulation method applied, Table 1-1.

In point-to-multipoint applications BTS serves as the stability reference for several CPEs. Different LMDS standards specify different radio frequency tolerances for the BTS. For example, ETSI specification for point-to-multipoint digital radio systems in frequency bands in the range 24.25GHz to 29.5GHz, [3], specifies that radio frequency tolerance should not exceed $\pm 15\text{ppm}$. Stability requirement would drive the price difference between the BTS and CPE.

Depending on the frequency plan, both the BTS and CPE would require a minimum one or two microwave sources for the *Tx* (transmitter) and *Rx* (receiver). Because one BTS can support up to 50 CPEs it is important to keep the cost of the CPE down.

1.3.1 CPE Local Oscillators

A CPE local oscillator chain provides LO signals to the transmit and receive sub-harmonic mixers. The oscillator is a free-running DRO but it is temperature compensated to control the output frequency. The LO signal is split into two arms to drive the two mixers, Figure 1–3.

Figure 1–3: CPE local oscillator chain

1.3.2 BTS Local Oscillators

In order to meet stability requirements, the BTS LO chain is synthesized. One possible realization for the LMDS BTS local oscillators is as on Figure 1–4. The BTS LO chain uses a *PLDRO* (phase locked DRO) to drive the LO of two sub-harmonic mixers. The IF ports of the sub-harmonic mixers are driven by L-Band synthesizers. Filters are required to remove unwanted harmonics and spurious signals. Some amplification of the up-converted signals is required to achieve the required output power per arm. Both the PLDRO and the synthesizers are locked to a common stable crystal reference.

Figure 1–4: BTS local oscillator chain with PLDRO

1.4 Low-cost Local Oscillator for LMDS

If high stability is not the requirement then a free-running DRO is the obvious choice due to inherently low-phase noise performance and low-cost design. A problem with the free-running DROs is that, even for the stability of $\pm 100\text{ppm}$ we need an algorithm for temperature compensation as the DRO itself has stability over industrial temperature range (-40°C to $+80^\circ\text{C}$) of $\pm 1\text{-}2\text{MHz}$.

For stability requirements of less than $\pm 15\text{ppm}$, the PLDRO configuration from Figure 1–4 is a possible solution. Complexity of design and cost for the PLDRO of more than \$1000 US are the prices we have to pay for both excellent phase noise and stability.

For LMDS frequencies lower than 25GHz with higher stability requirements, lower modulation schemes and higher symbol rates, we might be able to use a phase locked VCO at X-Band as shown in Figure 1–5. Here two synthesized X-band oscillators provide LO signals to the transmit and receive sub-harmonic

mixers. Both synthesizers are locked to a common temperature compensated reference.

Cost estimate for both Tx and Rx LOs is around \$100 US which is one tenth of the price for the LO chain on Figure 1–4. As only two sub-harmonic mixers are used, filtering of unwanted emissions is easier than on Figure 1–4.

Figure 1–5: BTS LO chain with X-Band synthesizers

1.5 Summary

If the phase noise performance of the LOs on Figure 1–5 is sufficient for certain modulation schemes and symbol rates, X-Band synthesizers might be attractive for LMDS applications due to their low-cost, high stability and low level unwanted emissions.

In the following chapters we will investigate design considerations, performance and possible applications of the low-cost X-Band synthesizers.

Chapter Two

2 Background Information

2.1 The Oscillation Criterion

An oscillator produces an alternating current (AC) output signal when direct current (DC) bias is applied to the oscillator supply terminals. The basic structure of the sinusoidal oscillator consists of an amplifier and a frequency selective network connected in a positive-feedback loop, Figure 2–1, [4].

Figure 2–1: The basic structure of a sinusoidal oscillator

The feedback signal is summed with the positive sign, thus the gain with feedback is given by

$$A_f(s) = \frac{x_o(s)}{x_s(s)} = \frac{A(s)}{1 - A(s) \cdot \beta(s)} \quad (2-1)$$

We define the loop gain $L(s)$ as

$$L(s) = A(s) \cdot \beta(s) \quad (2-2)$$

If at a specific frequency f_0 , $L(f_0)$ is equal to one, from (2-1) A_f will be infinite. In other words, at f_0 the circuit from Figure 2-1 will have a finite output for zero input signal. Such a circuit is by definition an oscillator and the input signal x_s can be removed. Any noise applied at the input would cause sinusoidal oscillations of frequency f_0 .

A_f is infinite at frequency f_0 when loop gain amplitude and phase are

$$\begin{aligned} |A(j \cdot 2 \cdot \pi \cdot f_0) \cdot \beta(j \cdot 2 \cdot \pi \cdot f_0)| &= 1 \\ \phi(f_0) = \angle A(j \cdot 2 \cdot \pi \cdot f_0) \cdot \beta(j \cdot 2 \cdot \pi \cdot f_0) &= 0 + k \cdot \pi, \quad k = 0, 1, 2 \dots \end{aligned} \quad (2-3)$$

This is known as the *Barkhausen* criterion. To guarantee start-up of oscillation we typically over compensate, so

$$|A(j \cdot 2 \cdot \pi \cdot f_0) \cdot \beta(j \cdot 2 \cdot \pi \cdot f_0)| > 1 \quad (2-4)$$

As the oscillation amplitude increases, nonlinearity in the active device will limit the amplitude and maintain (2-3).

2.2 Frequency Stability

It should be noted that the frequency of oscillation f_0 is determined solely by the phase characteristics of the feedback loop. The loop oscillates at the frequency for which the phase is zero. It follows that the stability of the frequency of oscillation f_0 will be determined by the manner in which the phase of the feedback loop $\phi(f)$ varies with frequency. If $d\phi/df$ is large, the resulting change in f_0 is small, Figure 2-2.

Figure 2-2: Oscillator frequency stability vs. slope of the phase response

In oscillator design it is very important to maintain high stability of frequency over time and temperature.

2.3 Resonant VCOs

The principle of the resonant oscillator is shown in Figure 2–3, [5].

Figure 2–3: Principle of the resonant oscillator

A resonant circuit, in this case a parallel L-C tank, converts the current i_1 from a current source to a voltage v_1 . At resonance,

$$f_0 = \frac{\omega_0}{2 \cdot \pi} = \frac{1}{2 \cdot \pi \cdot \sqrt{L \cdot C}} \quad (2-5)$$

the admittances of L and C cancel, and the tank has the impedance r_p . This resistance is usually a model of effective resistance due to losses in the current source and the inductor. The quality factor Q of the tank is

$$Q = \frac{r_p}{2 \cdot \pi \cdot f \cdot L} \quad (2-6)$$

At resonance, we have $v_1 = r_p \cdot i_1$. It follows that for positive feedback with unity loop gain we must have $gm = 1/r_p$. The result is oscillation at the frequency f_0 , (2–5). Oscillation frequency f_0 may be controlled by varying C electronically,

converting the oscillator to a VCO (Voltage Control Oscillator). The oscillator in Figure 2–4 includes a varactor diode as a part of the tank capacitance.

Figure 2–4: Resonant VCO

A varactor diode is the reverse-biased diode whose junction capacitance C_V is a function of the reverse bias v_R . v_R is usually applied by the control voltage v_c through a buffer resistor R_D , which keeps the source of v_c from loading down the tank. A series capacitor C_D blocks DC current that would otherwise flow through R_D and L . Therefore, $v_R = v_c$ for slow variation in v_c .

VCO voltage gain, by definition, is

$$K_{vco}[\text{rad} / \text{s} / \text{V}] = \frac{d\omega_0}{dv_c} = \frac{d\Delta\omega_0}{dv_c}$$

(2–7)

2.3.1 Noise Through a Bandpass Filter

In order to achieve optimum synthesizer phase noise performance, emphasis is placed on the resonator oscillator (VCO) phase noise theory. First, band-limited white noise will be discussed.

In this section, the relationship between band limited white noise power spectral density and equivalent phase noise spectral density will be derived.

Consider the situation on Figure 2–5, where the unfiltered noise $n_{uf}(t)$ at the input of the bandpass filter is white (either thermal or shot noise or both combined) with uniform single sided power spectral density (psd), $\Phi_{n_{uf}}(f) = N_0[W/Hz]$. The bandpass filter center frequency f_i is much higher than the filter bandwidth B_w , $f_i \gg B_w$.

Figure 2–5: Signal and noise through a bandpass filter

The noise $n_f(t)$ after the bandpass filter has the Φ_{n_f} as shown in Figure 2–6. For $f_i \gg B_w$, [6], random signal $n_f(t)$ can be divided into two orthogonal components $n_x(t)$ and $n_y(t)$

$$n_f(t) = n_x(t) + n_y(t) = x(t) \cdot \sin(\omega_i t) + y(t) \cdot \cos(\omega_i t)$$

(2–8)

Figure 2-6: Power spectral density (psd) of the filtered noise $n_f(t)$

It will be shown that $n_f(t)$ at the output of the bandpass filter induces both amplitude modulation $x(t)$, and phase modulation $\theta_i(t)$

$$v_o(t) = [V_s + x(t)] \cdot \sin[\omega_i \cdot t + \theta_i(t)] \quad (2-9)$$

Next, the psd Φ_x of the random signal x , and psd Φ_{θ_i} of the random phase θ_i have to be defined. From (2-8), since n_x and n_y are orthogonal,

$$\overline{n_f^2} = \overline{n_x^2} + 2 \cdot \overline{n_x \cdot n_y} + \overline{n_y^2} = \overline{n_x^2} + \overline{n_y^2} \quad (2-10)$$

The phase reference that established the *sin* and *cos* functions in (2-8) was arbitrary, so it must be that

$$\overline{n_x^2} = \overline{n_y^2} = \frac{\overline{n_f^2}}{2} = \frac{N_0 \cdot B_w}{2} \quad (2-11)$$

It will be shown that $n_x(t)$, comprising half the noise power induces amplitude modulation of $v_o(t)$, and $n_y(t)$, comprising the other half of the noise power, induces phase modulation on $v_o(t)$.

Since $x(t)$ and $\sin(\omega_i t)$ are independent random variables

$$\overline{n_x^2} = \overline{(x \cdot \sin \omega_i t)^2} = \overline{x^2} \cdot \overline{\sin^2 \omega_i t} = \frac{\overline{x^2}}{2} \quad (2-12)$$

The expression for $n_x(t)$ in (2-8) has the form of suppressed-carrier amplitude modulation, where $x(t)$ is the baseband signal occupying a bandwidth of $B_w/2$. Using this with (2-11) and (2-12), the equations for the psd Φ_x and psd Φ_y are obtained and graphically shown in Figure 2-7.

$$\Phi_x(f) = \Phi_y(f) = 2 \cdot \Phi_{n_f}(f + f_i) = 2 \cdot N_0 ; \quad f \geq 0 \quad (2-13)$$

Figure 2-7: Orthogonal noise components, psd

For calculating phase noise spectral density Φ_{θ_i} we have to rearrange the expression for the output signal $v_o(t)$ in the form of

$$v_o(t) = [V_s + x(t)] \cdot \sin \omega_i t + y(t) \cdot \cos(\omega_i \cdot t) \quad (2-14)$$

A phasor diagram of (2-14) is shown on Figure 2-8.

Figure 2-8: Noise induced phase

From Figure 2-8 for small values of phase noise $\theta_i(t)$, i.e., for $V_s \gg x(t)$, and $V_s \gg y(t)$, the output signal is

$$v_o(t) \approx [V_s + x(t)] \cdot \sin[\omega_i \cdot t + \theta_i(t)] \quad (2-15)$$

Equation (2-15) supports what is stated in (2-9). The random signal $x(t)$ induces amplitude modulation, and the random phase $\theta_i(t)$ induces phase modulation of the input signal $v_i(t)$.

The phasor phase on Figure 2-8 is

$$\theta_i(t) \approx \frac{y(t)}{V_s} \quad (2-16)$$

Mean square phase is

$$\overline{\theta_i(t)^2} = \frac{\overline{y(t)^2}}{V_s^2} = \frac{N_o \cdot B_w}{V_s^2} = \Theta_o \frac{B_w}{2} \quad (2-17)$$

where

$$\Theta_o = 2 \cdot N_o / V_s^2 \quad (2-18)$$

is the phase spectral density in rad^2/Hz , as shown in Figure 2–9.

Figure 2–9: Spectral density of noise induced phase

By comparing Figure 2–9 with Figure 2–6, the phase noise spectral density Φ_{θ_i} of the random phase $\theta_i(t)$ is

$$\Phi_{\theta_i}(f) = \left(\frac{2}{V_s^2} \right) \cdot \Phi_{n_f}(f_i + f) = \Theta_o; \quad 0 < f < \frac{B_w}{2} \quad (2-19)$$

In conclusion, band limited white noise induces both amplitude and phase modulation of the sinusoidal carrier. The psd of the induced amplitude modulation Φ_x is given in (2–13). The spectral density of the phase modulation Φ_{θ_i} is given by (2–18) and (2–19) and is directly proportional to single sided psd N_o of the unfiltered white noise, and is inversely proportional to the signal amplitude V_s . Both Φ_x and Φ_{θ_i} are baseband signals occupying *SSB* (single sided bandwidth) of $B_w/2$.

2.3.2 Phase Noise in Resonant Oscillators

Figure 2–10 gives a model for a resonant oscillator with a noise voltage $n_s(t)$. The noise $n_s(t)$ generated within resonant oscillator is added to the oscillation voltage $v_1(t)$ and induces both amplitude $x_{osc}(t)$ and phase modulation $\theta_{osc}(t)$ of the output signal of the output signal

$$v_{osc}(t) = [V_{osc} + x_{osc}(t)] \cdot \sin[\omega_0 \cdot t + \theta_{osc}(t)] \quad (2-20)$$

The amplitude of $v_1(t)$ is V_1 , oscillation frequency is $\omega_0 = 2 \cdot \pi \cdot f_0$, and the half bandwidth of the oscillator tank is $f_b = f_0 / 2 \cdot Q$.

Figure 2–10: Resonator oscillator with noise voltage $n_s(t)$

For large signal to noise ratios it will be shown at the end of this section that most of the resonant oscillator noise is coming from the phase modulation of the carrier $\theta_{osc}(t)$. In that case, $x_{osc}(t)$ in (2–20) can be neglected, leaving

$$v_{osc}(t) = V_{osc} \cdot \sin[\omega_0 \cdot t + \theta_{osc}(t)] \quad (2-21)$$

The phase spectral density $\Phi_{\theta_{osc}}$ of the random phase $\theta_{osc}(t)$ will now be calculated.

Noise, $n_s(t)$ is the noise in the tank before oscillations start. It can include thermal noise, shot noise, and flicker noise generated within the oscillator, and all are presented here as originating at the amplifier input. The psd Φ_{n_s} is shown on Figure 2–11, where $f_m \equiv f - f_0$ is the offset frequency.

Figure 2–11: Psd of the oscillator noise $n_s(t)$

The thermal and shot noise contribute a flat psd

$$\Phi_{n_s}(f_o + f_m) = N_o; \quad \text{for } f_m > f_a \quad (2-22)$$

At frequencies near f_0 , the flicker noise dominates with a psd proportional to $1/f_m$

$$\Phi_{n_s}(f_o + f_m) = \frac{N_o \cdot f_a}{f_m}; \quad \text{for } f_m < f_a \quad (2-23)$$

The frequency f_a below which flicker noise dominates is an empirical quantity, typically around $10^{-5} \cdot f_o$.

The addition of $n_s(t)$ within the bandwidth of the oscillator tank is represented as a phasor sum on Figure 2–12.

Figure 2–12: Phasor diagram within the oscillator bandwidth

The effect of the noise is to produce $v_2(t)=v_1(t)+n_s(t)$ shifted in phase from $v_1(t)$ by $\phi(t)$, [7]. For oscillation to be sustained, the tank circuit must produce a compensating phase shift $-\phi(t)$. By development similar to that of Φ_{θ_i} in (2–19), the phase spectral density Φ_ϕ of random phase $\phi(t)$ is calculated as follows:

$$\Phi_\phi(f_m) = \frac{2}{V_1^2} \cdot \Phi_{n_s}(f_o + f_m); \quad \text{for } f_m < f_b \quad (2-24)$$

Using Equations (2–18) and (2–19) gives

$$\begin{aligned} \Phi_\phi(f_m) &= \frac{\Theta_o \cdot f_a}{f_m}; & \text{for } f_m < f_a \\ \Phi_\phi(f_m) &= \Theta_o; & \text{for } f_m > f_a \end{aligned} \quad (2-25)$$

where

$$\Theta_o = \frac{2 \cdot N_o}{V_1^2} \quad (2-26)$$

Equations (2–25) and (2–26) give phase noise spectral density of the compensating phase shift $-\phi(t)$ produced by the resonator tank. This compensating phase shift, when changing in time, causes frequency modulation $\Delta\omega_0$ of the oscillator. In order to define phase noise at the output of the resonator, a relationship between frequency modulation $\Delta\omega_0$ and compensating phase shift $-\phi(t)$ has to be derived. The first step is to investigate the phase response vs. frequency of the tank RLC circuit.

The transfer function of the bandpass RLC tank, [4], is

$$T(s) = \frac{V(s)}{I(s)} = \frac{R \cdot \frac{\omega_0}{Q} \cdot s}{s^2 + s \cdot \frac{\omega_0}{Q} + \omega_0^2} \quad (2-27)$$

Let the frequency deviation $\Delta\omega_0$ be represented by ω_{osc} . The phase response $\phi(\omega) = \angle T(\omega)$ in the vicinity of f_0 is

$$\phi(\omega_{osc}) = \tan^{-1} \left(\frac{2 \cdot Q \cdot \omega_{vco}}{\omega_0} \right) \approx \frac{2 \cdot Q \cdot \omega_{osc}}{2 \cdot \pi \cdot f_0} = \frac{\omega_{osc}}{2 \cdot \pi \cdot f_b} \quad (2-28)$$

Solving (2–28) for ω_{osc} gives

$$\omega_{osc}(t) = 2 \cdot \pi \cdot f_b \cdot \phi(t) \quad (2-29)$$

The corresponding spectral densities $\Phi_{\omega_{osc}}$ and Φ_ϕ are related as

$$\Phi_{\omega_{osc}}(f_m) = (2 \cdot \pi \cdot f_b)^2 \cdot \Phi_\phi(f_m) ; \quad \text{for } f_m < f_b \quad (2-30)$$

Equation (2–30) establishes a statistical relationship between frequency modulation $\omega_{osc}(t)$ and compensating phase shift $-\phi(t)$.

With (2–25) and (2–26)

$$\begin{aligned}\Phi_{\omega_{osc}}(f_m) &= \Theta_0 \cdot (2 \cdot \pi \cdot f_b)^2 \cdot \frac{f_a}{f_m}; & \text{for } f_m < f_a \\ \Phi_{\omega_{osc}}(f_m) &= \Theta_0 \cdot (2 \cdot \pi \cdot f_b)^2; & \text{for } f_a < f_m < f_b\end{aligned}\quad (2-31)$$

We need to calculate phase noise spectral density of the phase modulation at the resonator output $\Phi_{\theta_{osc}}$ in terms of the spectral density of the frequency deviation $\Phi_{\omega_{osc}}$. Phase modulation $\theta_{osc}(t)$ is the integral of frequency modulation $\omega_{osc}(t)$, which means that

$$\begin{aligned}\theta_{osc}(s) &= \frac{1}{s} \cdot \omega_{osc}(s) \\ \theta_{osc}(j \cdot 2 \cdot \pi \cdot f_m) &= \frac{1}{j \cdot 2 \cdot \pi \cdot f_m} \cdot \omega_{osc}(j \cdot 2 \cdot \pi \cdot f_m)\end{aligned}\quad (2-32)$$

In terms of spectral densities

$$\Phi_{\theta_{osc}}(f_m) = \left(\frac{1}{2 \cdot \pi \cdot f_m} \right)^2 \cdot \Phi_{\omega_{osc}}; \quad \text{for } f_m < f_b \quad (2-33)$$

Equation (2–33) gives the relationship between the phase noise spectral density $\Phi_{\theta_{osc}}$ and spectral density of the frequency deviation $\Phi_{\omega_{osc}}$. From (2–33), within the tank circuit bandwidth

$$\begin{aligned}\Phi_{\theta_{osc}}(f_m) &= \Theta_o \cdot \frac{f_a \cdot f_b^2}{f_m^3} = \frac{N_o \cdot f_a \cdot f_o^2}{2 \cdot V_1^2 \cdot Q^2} \cdot \frac{1}{f_m^3}; & \text{for } f_m < f_a \\ \Phi_{\theta_{osc}}(f_m) &= \Theta_o \cdot \frac{f_b^2}{f_m^2} = \frac{N_o \cdot f_o^2}{2 \cdot V_1^2 \cdot Q^2} \cdot \frac{1}{f_m^2}; & \text{for } f_a < f_m < f_b\end{aligned}\quad (2-34)$$

Outside the tank circuit bandwidth, feedback in the oscillator is effectively broken and $\theta_{osc}(t)=\phi(t)$. Therefore the phase noise spectral density is

$$\Phi_{\theta_{osc}}(f_m) = \Theta_o = \frac{2 \cdot N_o}{V_1^2}; \quad \text{for} \quad f_b < f_m \quad (2-35)$$

This flat portion does not extend forever, otherwise the phase noise would have infinite power. In practice the curve breaks at some cutoff frequency f_c of the output bandpass filter.

Leeson, [8], has confirmed experimentally the phase noise expressions given in (2-34) and (2-35). Equations show what was expected: $\theta_{vco}(t)$ is minimized by maximizing tank Q factor, maximizing the oscillation amplitude V_1 , and minimizing the noise figure of the components N_o .

2.3.3 L-Band Oscillator Phase Noise Example

To put the theory from the previous section into practice, piecewise approximation of the phase noise spectral density, (2-34) and (2-35), is shown in the following L-Band oscillator example.

An L-Band oscillator at $f_o=1600\text{MHz}$ has a tank with $Q=30$, ($f_b=27.5\text{MHz}$). The signal amplitude at the tank is $V_1=200\text{mV}$ and the oscillator has a noise figure $NF=14\text{dB}$. This gives a noise spectral density of $N_o=-160\text{dBm/Hz}=5 \times 10^{-18}\text{V}^2/\text{Hz}$. Below the frequency of $f_a = 10^{-5} \cdot f_o = 16\text{KHz}$, flicker noise dominates. The output bandpass filter has a half-bandwidth of $f_c=50\text{MHz}$.

Piecewise approximation for $\Phi_{\theta_{osc}}$ is as follows

$$\begin{aligned}\Phi_{\theta_{osc}}(f_m) &= \Theta_o \cdot \frac{f_a \cdot f_b^2}{f_m^3} = \frac{3025}{f_m^3} \text{ rad}^2/\text{Hz} \quad \text{for } f_m < f_a \\ \Phi_{\theta_{osc}}(f_m) &= \Theta_o \cdot \frac{f_b^2}{f_m^2} = \frac{0.19}{f_m^2} \text{ rad}^2/\text{Hz} \quad \text{for } f_a < f_m < f_b \\ \Phi_{\theta_{osc}}(f_m) &= \Theta_o = \frac{2 \cdot N_o}{V_1^2} = 2.5 \times 10^{-16} \text{ rad}^2/\text{Hz} \quad \text{for } f_b < f_m < f_c\end{aligned}\tag{2-36}$$

Phase noise spectral density $\Phi_{\theta_{osc}}$ vs. carrier offset frequency f_m , (2-36), is plotted in Figure 2-13.

Figure 2-13: L-Band oscillator phase noise example, $\Phi_{\theta_{osc}}$ in rad^2/Hz

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2.3.4 Measuring $\Phi_{\theta_{osc}}$ Using a Spectrum Analyzer

Since it is impractical to fully characterize the noise source $n_s(t)$, $\Phi_{\theta_{osc}}$ must be determined by measurement. A simplest method is to use a spectrum analyzer to measure *power* spectral density $\Phi_{V_{osc}}$ of the resonant oscillator output $v_{osc}(t)$ from Figure 2–10.

Comparison of (2–21) where

$$v_{osc}(t) = V_{osc} \cdot \sin[\omega_0 \cdot t + \theta_{osc}(t)],$$

with (2–9) where

$$v_o(t) = [V_s + x(t)] \cdot \sin[\omega_i \cdot t + \theta_i(t)]$$

shows that they differ in the presence of an amplitude modulation part, $x(t)$. By neglecting the amplitude modulation part, and after replacing Φ_{n_f} with $\Phi_{V_{osc}}$ in (2–19), the result for the phase noise spectral density is simply

$$\Phi_{\theta_{osc}}(f_m) = \frac{2}{V_{osc}^2} \cdot \Phi_{V_{osc}}(f_o + f_m); \quad \text{for } f_m \geq 0 \quad (2-37)$$

Equation (2–37) gives the relationship between *phase noise* spectral density $\Phi_{\theta_{osc}}$ and measured *power* spectrum density $\Phi_{V_{osc}}$ of the resonant oscillator output $v_{osc}(t)$. The dimension for $\Phi_{V_{osc}}$ is V^2/Hz , however the spectrum analyzer reads $\Phi_{V_{osc}}$ in dBm/Hz. The spectrum is normalized to dBc/Hz simply by using

$$\frac{\text{dBc}}{\text{Hz}} = \frac{\text{dBm}}{\text{Hz}} - \text{dBm}(\text{carrier}) \quad (2-38)$$

If P_{osc} is the signal carrier power in dBm

$$P_{osc} [dBm] = 10 \cdot \log_{10} \frac{V_{osc}^2 / 2}{50\Omega \cdot 1mW} \quad (2-39)$$

then

$$\Phi_{\theta_{osc}} [dBc/Hz] = \Phi_{V_{osc}} [dBm/Hz] - P_{osc} [dBm] \quad (2-40)$$

and finally,

$$\Phi_{\theta_{osc}} [dBc/Hz] = 10 \cdot \log_{10} (\Phi_{\theta_{osc}} [rad^2/Hz]) \quad (2-41)$$

Equation (2-36) is plotted again in Figure 2-14, now with the dimension for $\Phi_{\theta_{osc}}$ in dBc/Hz.

Figure 2-14: L-Band oscillator phase noise example, $\Phi_{\theta_{osc}}$ in dBc/Hz

2.3.5 Amplitude Modulation in Resonant Oscillators

In this section a novel approach for calculating amplitude modulation of the output signal $v_{osc}(t)$ of the resonator oscillator is presented.

Noise $n_s(t)$ in Figure 2–10 also causes amplitude modulation of the output signal $v_{osc}(t)$ in the form of $x_{osc}(t)$, (2–20).

From (2–13), (2–22) and (2–23) the amplitude modulation psd $\Phi_{x_{osc}}$ in (V^2/Hz) is calculated as

$$\begin{aligned}\Phi_{x_{osc}}(f_m) &= 2 \cdot \Phi_{n_s}(f_o + f_m) = 2 \cdot \frac{N_o \cdot f_a}{f_m}; & \text{for } f_m < f_a \\ \Phi_{x_{osc}}(f_m) &= 2 \cdot \Phi_{n_s}(f_o + f_m) = 2 \cdot N_o; & \text{for } f_a < f_m < f_c\end{aligned}\tag{2-42}$$

$\Phi_{x_{osc}}$ will be in dBc/Hz after being normalized to carrier power

$$\Phi_{x_{osc}}[dBc/Hz] = \Phi_{x_{osc}}[dBm/Hz] - P_{osc}[dBm]\tag{2-43}$$

2.3.6 Amplitude and Phase Noise Comparison

For comparison, amplitude modulation noise psd $\Phi_{x_{osc}}$ and phase noise spectral density $\Phi_{\theta_{osc}}$ are plotted on the same graph, Figure 2–15. Dimensions are normalized in dBc/Hz.

From Figure 2–15 it is clear that in the case of large signal-to-noise ratios, most of the oscillator noise is coming from the phase modulation of the carrier, (i.e. the dominant noise in the resonant oscillators is the phase noise).

Also, beyond frequency f_c , the total noise spectral density at the resonator oscillators output is N_0 .

Figure 2-15: Amplitude and phase noise comparison

2.4 Theory of Negative-Resistance Oscillators

In this section the theory for designing microwave transistor oscillators using the concept of negative resistance [9] is presented. The transistors small and large signal S parameters provide all the information needed to design negative-resistance oscillators.

2.4.1 One-Port Negative-Resistance Oscillators

A general schematic diagram for the one-port negative-resistance oscillator is given in Figure 2–16.

Figure 2–16: One-port negative-resistance oscillator

The negative-resistance device is presented by the amplitude and frequency dependent impedance

$$Z_{IN}(V, \omega) = R_{IN}(V, \omega) + j \cdot X_{IN}(V, \omega) \quad (2-44)$$

where,

$$R_{IN}(V, \omega) < 0 \quad (2-45)$$

The oscillator is constructed by connecting the device to a passive load impedance

$$Z_L(\omega) = R_L(\omega) + j \cdot X_L(\omega) \quad (2-46)$$

For characteristic impedance $Z_0 = 50\Omega$, the load reflection coefficient is

$$\Gamma_L(\omega) = \frac{Z_L(\omega) - 50}{Z_L(\omega) + 50} \quad (2-47)$$

and the reflection coefficient of an active device is

$$\Gamma_{IN}(V, \omega) = \frac{Z_{IN}(V, \omega) - 50}{Z_{IN}(V, \omega) + 50} \quad (2-48)$$

The network in Figure 2-16 will oscillate at the amplitude $V = V_0$ and frequency $\omega = \omega_0$, when

$$\Gamma_{IN}(V_0, \omega_0) \cdot \Gamma_L(\omega_0) = 1 \quad (2-49)$$

After substituting (2-47) and (2-48) into (2-49), and equating the real and imaginary parts, the oscillation condition can be written as

$$\begin{aligned} R_{IN}(V_0, \omega_0) + R_L(\omega_0) &= 0 \\ X_{IN}(V_0, \omega_0) + X_L(\omega_0) &= 0 \end{aligned} \quad (2-50)$$

To be specific, the one port network is unstable in the range $\omega_1 < \omega < \omega_2$, if the net resistance of the network R_{IN} is negative and

$$|R_{IN}(V, \omega)| > R_L(\omega) ; \quad \text{for} \quad \omega_1 < \omega < \omega_2 \quad (2-51)$$

Under proper conditions, a growing sinusoidal current will flow through the circuit. That is, at the start of the oscillation when the amplitude V is small, (2-51) must be satisfied. This is expressed in the form

$$|R_{IN}(0, \omega)| > R_L(\omega) ; \quad \text{for} \quad \omega_1 < \omega < \omega_2 \quad (2-52)$$

The oscillations will continue to build up as long as the loop resistance is negative. The amplitude of the voltage must eventually reach a steady-state value $V=V_0$ and $\omega=\omega_0$, which occurs when the loop resistance is zero. To satisfy the start of oscillation condition, the build-up of oscillation and the oscillation conditions, the impedance $Z_{IN}(V,\omega)$ must be amplitude and frequency dependent. The frequency of oscillation ω_0 determined by (2-50) might not be stable since $X_{IN}(V_1,\omega_0) \neq X_{IN}(V_0,\omega_0)$. Kurokawa had shown that for the small variations of Z_{IN} around ω_0 , and for $R_L(\omega) = R_L$, the additional condition to (2-50) for stable oscillation is

$$\left. \frac{\partial R_{IN}(V,\omega)}{\partial V} \right|_{V=V_0} \cdot \left. \frac{dX_L(\omega)}{d\omega} \right|_{\omega=\omega_0} > 0 \quad (2-53)$$

The interested reader is directed to reference [10] for the derivation of (2-53).

2.4.2 Example for the Stable Oscillation Conditions

In order to better understand the concept explained in the previous section, a simple example of the negative-resistance oscillator with the linear amplitude dependence of the negative conductance will be presented as follows.

A negative resistance can be modeled by the parallel combination of a capacitor and a negative conductance, as shown on Figure 2-17

The linear amplitude dependence of the negative conductance is shown in Figure 2-18, and is given by

$$G(V) = G_M \left(1 - \frac{V}{V_M} \right) \quad (2-54)$$

Figure 2–17: Negative-resistance device

The design goal is to calculate a load circuit Z_L to provide a stable oscillation at ω_0 , and to calculate the output power. The device impedance is

$$Z_{IN} = R_{IN}(V, \omega) + jX_{IN}(V, \omega) = \frac{-G(V)}{G^2(V) + \omega^2 \cdot C^2} + j \frac{-\omega \cdot C}{G^2(V) + \omega^2 \cdot C^2} \quad (2-55)$$

Figure 2–18: Linear amplitude variation of $G(V)$

After replacing (2–54) in R_{IN} , and differentiating with respect to V

$$\frac{\partial R_{IN}(V, \omega)}{\partial V} = \frac{-1 + 2 \cdot \frac{V}{V_M} - \frac{V^2}{V_M^2} + \omega^2 \cdot \frac{C^2}{G_M^2}}{G_M \cdot V_M \cdot \left[\left(1 - \frac{V}{V_M} \right)^2 + \omega^2 \cdot \frac{C^2}{G_M^2} \right]^2} \quad (2-56)$$

From (2-50), (2-53) and (2-56), stable oscillation occurs at frequency ω_0 and oscillation voltage level V_0 , when

$$R_L = \frac{G(V_0)}{G^2(V_0) + \omega_0^2 \cdot C^2} \quad (2-57)$$

$$X_L = \frac{\omega_0 \cdot C}{G^2(V_0) + \omega_0^2 \cdot C^2} \quad (2-58)$$

and

$$\left. \frac{dX_L(\omega)}{d\omega} \right|_{\omega=\omega_0} \cdot \left(-1 + 2 \cdot \frac{V_0}{V_M} - \frac{V_0^2}{V_M^2} + \omega_0^2 \cdot \frac{C^2}{G_M^2} \right) > 0 \quad (2-59)$$

There is no direct way to solve for R_L in (2-57) and X_L in (2-58). Another design consideration such as maximizing the power delivered to R_L must be introduced. The output power at the circuit is given by

$$P = \frac{1}{2} \cdot |V|^2 R_L \cdot [G^2(V) + \omega^2 C^2] \quad (2-60)$$

Rewriting by substituting R_L from (2-57) and V from (2-54), gives

$$P = \frac{1}{2} \cdot V_M^2 \cdot \left(1 - \frac{G(V)}{G_M} \right)^2 \cdot G(V) \quad (2-61)$$

Maximizing power by differentiating (2-59) over $G(V)$ provides

$$\frac{\partial P}{\partial G(V)} = \frac{1}{2} \cdot V_M^2 \cdot \left(1 - 4 \cdot \frac{G(V)}{G_M} + 3 \frac{G^2(V)}{G_M^2} \right) = 0 \quad (2-62)$$

Solving (2-62) for $G(V)/G_M$ gives

$$\frac{G(V)}{G_M} = \frac{1}{3} \quad (2-63)$$

Substituting for the output voltage when maximum power is delivered to R_L at $V=V_0$

$$\frac{V_0}{V_M} = \frac{2}{3} \quad (2-64)$$

Values for R_L and X_L that maximize the power are

$$R_L = \frac{\frac{G_M}{3}}{\left(\frac{G_M}{3} \right)^2 + \omega_0^2 \cdot C^2}$$

$$X_L(\omega_0) = \frac{\omega_0 \cdot C}{\left(\frac{G_M}{3} \right)^2 + \omega_0^2 \cdot C^2} \quad (2-65)$$

and Equation (2-59) becomes

$$\frac{dX_L(\omega)}{d\omega} \Big|_{\omega=\omega_0} \cdot \left(\frac{\omega_0^2 \cdot C^2}{G_M^2} - \frac{1}{9} \right) > 0 \quad (2-66)$$

From (2–66), $\left. \frac{dX_L(\omega)}{d\omega} \right|_{\omega=\omega_0} > 0$ when

$$\frac{\omega_0 \cdot C}{G_M} > \frac{1}{3}$$

(2–67)

For $\frac{\omega_0 \cdot C}{G_M} \gg 1$

$$X_L(\omega_0) \approx \frac{1}{\omega_0 \cdot C}$$

(2–68)

Obviously, an inductor $X_L = \omega \cdot L$ satisfies (2–68), and the frequency of oscillation ω_0 is given by

$$\omega_0 \approx \frac{1}{\sqrt{L \cdot C}}$$

(2–69)

At this point it is necessary to check if R_L satisfies (2–52). For $\frac{\omega_0 \cdot C}{G_M}$ large, this is true because

$$\frac{R_L}{|R_{IN}(0, \omega_0)|} \approx \frac{1}{3}$$

(2–70)

Equation (2–70) provides a good design guideline for selecting R_L , which maximizes the oscillator power. We should observe that the above equation is valid when the negative input resistance varies linearly with amplitude.

2.4.3 Two-Port Negative-Resistance Oscillators

In order to achieve the negative resistance component, a transistor is used as an active device, with the two-port S parameters defined at microwave frequencies.

By choosing a proper terminating network on one of the transistor ports, the active device is made unstable at a specific frequency of oscillation. The load network on the other port is then designed in a manner described in the section for one-port negative resistance oscillators.

The general block diagram for the two-port negative resistance oscillator is shown in Figure 2–19. The transistor network is characterized by its S parameters, Z_T is the terminating network impedance, and Z_L is the load impedance.

Figure 2–19: Two-port oscillator model

When the two-port is potentially unstable, an appropriate Z_T permits the two-port to be presented as a one-port negative-resistance device with input impedance Z_{IN} . The conditions for a stable oscillation are given by (2–50) and (2–53). When the input port is made to oscillate, the terminating port also oscillates. The fact that both ports are oscillating can be proved as follows.

The input port is oscillating when

$$\Gamma_{IN} \cdot \Gamma_L = 1 \quad (2-71)$$

For a two-port network, from Figure 2–19,

$$\Gamma_{IN} = S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_T}{1 - S_{22} \cdot \Gamma_T} \quad (2-72)$$

and

$$\Gamma_{OUT} = S_{22} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_L}{1 - S_{22} \cdot \Gamma_L} \quad (2-73)$$

If we define

$$\Delta = S_{11} \cdot S_{22} - S_{12} \cdot S_{21} \quad (2-74)$$

(2-71) becomes

$$\Gamma_L = \frac{1}{\Gamma_{IN}} = \frac{1 - S_{22} \cdot \Gamma_T}{S_{11} - \Delta \cdot \Gamma_T} \quad (2-75)$$

Solving (2-75) for Γ_T gives

$$\Gamma_T = \frac{1 - S_{11} \cdot \Gamma_L}{S_{22} - \Delta \cdot \Gamma_L} \quad (2-76)$$

From (2-73)

$$\frac{1}{\Gamma_{OUT}} = \frac{1 - S_{11} \cdot \Gamma_L}{S_{22} - \Delta \cdot \Gamma_L} \quad (2-77)$$

so the terminating port is also oscillating as

$$\Gamma_{OUT} \cdot \Gamma_T = 1 \quad (2-78)$$

2.4.4 The Design Procedure for Two-Port Oscillator

A design procedure for a two-port oscillator (Figure 2–19) is as follows:

- Use a potentially unstable transistor at the frequency of oscillation ω_0 .
- Design the terminating network Z_T to make $|\Gamma_{IN}| > 1$.
- Design the load network to resonate Z_{IN} . That is, let

$$X_L(\omega_0) = -X_{IN}(\omega_0) \quad (2-79)$$

and from (2–70)

$$R_L = \frac{|R_{IN}(0, \omega_0)|}{3} \quad (2-80)$$

The frequency of oscillation will shift somewhat from its design value because the oscillation power increases until the negative resistance is equal to the load resistance, and X_{IN} varies as a function of V . But this simple, small signal analysis is of great help for getting a starting point for further nonlinear harmonic balance analysis.

2.5 Frequency Synthesizers

One way to improve stability and phase noise of a free running VCO is to lock the VCO in a phase locked loop (PLL) referenced to a stable crystal. A frequency synthesizer is built by adding a frequency divider in the feedback path of the PLL loop.

2.5.1 Simple Frequency Synthesizer

A simple frequency synthesizer block diagram is given in Figure 2–20 [5] where f_R is the reference crystal frequency, R is the reference divider counter, f_{step} is the synthesizer step frequency, f_0 is the synthesizer output frequency and N is the frequency divider in the feedback path.

Figure 2–20: Simple frequency synthesizer

When the PLL is in lock, the feed-back frequency f_0 / N equals f_R / R . Therefore, the output frequency for the fixed step size f_{step} is generated by selecting the proper integer N

$$f_0 = N \cdot f_{step} \quad (2-81)$$

2.5.2 L-Band Synthesizer Block Diagram

Most common synthesizers available on the market have charge pump at the output of the phase detector. Figure 2–21 shows a block diagram of the National

Semiconductor's LMX-2326 synthesizer using charge pump with an external 3rd order loop filter.

The output of the charge pump is a current source of high output impedance. This enables simplified loop filter design without the use of an operational amplifier. Average output current $I(s)$ is proportional to the phase error.

Figure 2-21: LMX-2326 synthesizer with 3rd order loop filter

2.5.3 PLL Forward Loop Gain

Phase detector gain is defined by

$$K_{cp} = \frac{I_{cp} [A]}{[2 \cdot \pi \cdot rad]} \quad (2-82)$$

where I_{cp} is the charge pump current gain.

VCO gain has been defined in Section 2.3 and N is the synthesizer divide ratio.

Loop gain, K is defined as

$$K = K_{cp} \left[\frac{A}{2 \cdot \pi \cdot rad} \right] \cdot K_{vco} \left[\frac{rad}{s} / V \right] = I_{cp} [\mu A] \cdot K_{vco} [MHz/V] \quad (2-83)$$

The loop filter transfer function is given by

$$F(s) = \frac{V(s)}{I(s)} \quad (2-84)$$

The PLL equivalent AC model, for Figure 2-21 is given in Figure 2-22, with the VCO replaced with linear block K_{vco} and the integrator $1/s$, [2].

Figure 2-22: PLL equivalent AC model

From Figure 2-22, by inspection, the forward loop gain is

$$G(s) = \frac{K \cdot F(s)}{s} \quad (2-85)$$

2.5.4 Phase Noise in the PLL Loop

The principal sources of phase noise in a synthesizer are the VCOs, phase detectors, resistors in the loop filter and crystal oscillators. To find the

contribution from each of these sources we need to define different phase transfer functions within the PLL.

If in Figure 2–22 $\theta_i(t)$ is replaced with the phase error from the phase detector $\theta_{pd}(t)$, equivalent circuit can be used to calculate the contribution from the phase detector, $\theta_{o_{pd}}(t)$.

By inspection of the equivalent circuit on Figure 2–22, the transfer function is

$$H_{pd}(s) = \frac{\theta_{o_{pd}}(s)}{\theta_{pd}(s)} = \frac{G(s)}{1 + \frac{G(s)}{N}} \quad (2-86)$$

Let the phase noise spectral density of the phase detector be $\Phi_{pd_{floor}}$ in rad^2/Hz .

The phase detector noise would cause phase noise spectral density at the PLL output Φ_{pd} , given by

$$\Phi_{pd} = \Phi_{pd_{floor}} \cdot |H_{pd}(s)|^2 \quad (2-87)$$

The equivalent AC loop model for calculating output phase noise component from the free running VCO is given in Figure 2–23

Figure 2–23: Output phase noise from the free running VCO

By inspection of the loop transfer function on Figure 2–23 we have

$$H_{vco}(s) = \frac{\theta_{o_{vco}}(s)}{\theta_{vco}(s)} = \frac{1}{1 + \frac{G(s)}{N}} \quad (2-88)$$

Let the phase noise spectral density of the free running VCO be $\Phi_{\theta_{vco}}$ in rad^2/Hz .

The phase noise spectral density at the output of Figure 2–23 is

$$\Phi_{vco} = \Phi_{\theta_{vco}} \cdot |H_{vco}(s)|^2 \quad (2-89)$$

For high divide ratios N , the phase noise component from the crystal oscillator and phase noise generated by the thermal noise from loop filter resistors is much smaller than Φ_{pd} and Φ_{vco} , so the total phase noise spectral density Φ_n is

$$\Phi_n[\text{rad}^2/\text{Hz}] = \Phi_{pd} + \Phi_{vco} \quad (2-90)$$

Results for Φ_n can be easily converted to dBc/Hz , (2–41).

2.6 QAM Phase Noise Requirements

QAM modulation is widely used in LMDS for achieving high data transmission rates (up to 155 Mbits/sec) over relatively narrow bandwidths as shown in Table 2-1, [11]. Digital radio systems designed using such modulation schemes must balance the effects of phase noise from local oscillators with the demodulator parameters in determining overall performance. The *rms* (root mean square) phase noise of the LO, after being filtered by the demodulator, must be low

enough to not cause bit errors and not significantly degrade the fade margin of the link.

<i>Parameter</i>	<i>4-QAM</i>	<i>16-QAM</i>	<i>64-QAM</i>	<i>256-QAM</i>
Theoretical spectral efficiency, bits/sec/Hz	2	4	6	8
Practical spectral efficiency, bits/sec/Hz	1.66	3.33	5	6.66
Min. Symbol Rate (no FEC) in Msps to transmit				
1.544 Mbits/sec (DS1)	0.77	0.39	0.26	0.19
2.048 Mbits/sec (E1)	1.02	0.51	0.34	0.26
6.312 Mbits/sec (DS2)	3.16	1.58	1.05	0.79
8.448 Mbits/sec (E2)	4.22	2.11	1.41	1.06
12.352 Mbits/sec (8 DS1)	6.18	3.09	2.06	1.54
24.704 Mbits/sec (16 DS1)	12.4	6.18	4.12	3.09
34.368 Mbits/sec (E3)	17.2	8.59	5.73	4.30
44.736 Mbits/sec (DS-3)	22.4	11.2	7.48	5.59
155.52 Mbits/sec (STS-3)	77.8	38.9	25.9	19.4
SNR for 10^{-3} BER (dB)	9.8	16.5	22.6	28.4
for 10^{-6} BER	13.6	20.4	26.6	32.6
for 10^{-7} BER	14.3	21.2	27.4	33.4
for 10^{-8} BER	15.0	21.9	28.1	34.1
for 10^{-9} BER	15.6	22.5	28.7	34.7
for 10^{-10} BER	16.1	23.0	29.2	35.2

Table 2-1: Selected parameters for some M-QAM systems

2.6.1 QAM Demodulator

A block diagram of a typical QAM demodulator is shown in Figure 2–24. To obtain coherent demodulation, a carrier recovery PLL is used to phase lock the demodulator VCO to the IF frequency. The quadrature outputs of this VCO separate the *I* and *Q* components of the signal in the conversion to baseband.

Each component then goes through a matched Nyquist lowpass filter, which has a bandwidth of half symbol rate F_s . The demodulator usually includes an adaptive equalizer which contains tapped delay lines with tap coefficients that are dynamically adjusted to minimize ISI (intersymbol interference).

Figure 2-24: Typical QAM demodulator

The decision slicer determines which of the M states is nearest to the received vector and computes the difference in phase $\Delta\theta$ between the received vector and that state, as shown in Figure 2-25.

Figure 2-25: QAM constellation diagram, $M=64$

The phase error signal is fed to a loop amplifier which outputs a tuning signal to the VCO, thus closing the PLL.

FEC (forward error correction) algorithms are used to correct infrequent bit errors at the cost of a fractional increase in the number of redundant bits that need to be transmitted.

Interleaving spreads out the bits making it easier to correct a burst of errors.

2.6.2 *BER* vs. *SNR* (Signal to Noise Ratio)

The theoretical *BER* for a matched Nyquist filter QAM receiver for $M=4, 16, 64, 256, \dots$, and assuming equiprobable symbols and no FEC is, [11],

$$P_e = \frac{1 - \left[1 - 2 \cdot \left(1 - \frac{1}{\sqrt{M}} \right) \cdot Q \left(\sqrt{\frac{3 \cdot (S/N)}{(M-1)}} \right) \right]^2}{\log_2 M} \quad (2-91)$$

where

$$Q(x) = \frac{\int_0^\infty e^{-\frac{t^2}{2}} \cdot dt}{\sqrt{2 \cdot \pi}} \quad (2-92)$$

and S/N equals average signal power over noise in the Nyquist bandwidth F_s .

These *BER* curves are shown in Figure 2-26, and are evaluated for some *BER*s in Table 2-1.

10^{-6} is a typical *BER* threshold for data transmitted on microwave links, so $(S/N)_6$ is defined as

$$P_e \left[\left(\frac{S}{N} \right)_6 \right] = 10^{-6} \quad (2-93)$$

Figure 2-26: *BER* vs. *S/N* for M-QAM

Note that there is 2.5dB difference in *S/N* between *BER* of 10^{-6} and *BER* of 10^{-10} . FEC (forward error correction) typically lowers the $(S/N)_6$ by 2-4dB relative to unFEC'd curves.

2.6.3 Residual $(S/N)_{res}$ and Fade Margin Degradation

Under normal (unfaded) conditions of an *LOS* (line of sight), the ratio of the *RSL* (received signal level) to receiver thermal noise

$$\frac{C}{N} = \frac{RSL}{K \cdot T \cdot B \cdot F} \quad (2-94)$$

is 30-65dB higher than the $(S/N)_6$. A high nominal *C/N* is used to provide margin for the signal fading. However, a low residual *BER* remains (usually $< 10^{-10}$) at

high C/N due to residual noise from equipment imperfections and external effects. Sources of the residual noise are:

- local oscillators phase noise
- amplitude and group delay distortion
- power amplifier intermodulation products
- quantization errors
- carrier recovery PLL noise
- clock recovery loop jitter
- residual ISI (inter symbol interference) from Nyquist filters

An adaptive equalizer will reduce many of the noise contributions.

It is reasonable to assume that the $(S/N)_{res}$ is independent of RSL , except at very low input levels.

Fade margin is the difference between the unfaded RSL and the RSL at which the $BER=10^{-6}$. On the occurrence of a deep fade due to heavy rain or multipath, the residual noise adding to receiver thermal noise affects performance as degradation of the fade margin.

If it is assumed that the receiver front-end thermal noise and the residual noise are additive Gaussian, then the degradation of the fade margin D is

$$D = \frac{1}{1 - \frac{\left(\frac{S}{N}\right)_6}{\left(\frac{S}{N}\right)_{res}}}$$

(2-95)

Phase margin degradation for 4-QAM (QPSK) and for 16-QAM vs. $(S/N)_{res}$ is given in Figure 2-27.

Figure 2–27: Phase margin degradation for QPSK and 16-QAM

If a simplifying assumption is made that the components making $(S/N)_{res}$ are independent additive Gaussian processes then the inverse $(S/N)s$ add powerwise and a degradation in the fade margin could be defined for each residual noise source individually.

2.6.4 Carrier Recovery PLL

Local oscillator phase noise is usually a major part of residual noise. Its effect on performance depends on the noise it produces at the output of the demodulator. In general, the mean square phase noise at the demodulator output due to noise of the LO is given by, [11]

$$\overline{\theta_D^2} [\text{rad}^2] = \int_0^\infty 2 \cdot \lambda_{\theta_{LO}}(f) \cdot |W(f)|^2 \cdot df \quad (2-96)$$

where $\lambda_{\theta_{LO}}$ is SSB phase noise spectral density in $[\text{rad}^2/\text{Hz}]$ vs. offset frequency and W is the transfer function from local oscillator phase to demodulator output phase.

The QAM demodulator is effectively a multioctave bandpass filter to phase variations from the LO since

- the Nyquist LP filter rejects frequency components above $F_s/2$
- the carrier recovery PLL acts as a HP filter by rejecting low frequency components

The purpose of the carrier recovery PLL is to track out low frequency variations in the carrier frequency and to maintain phase coherence. The IQ mixers, low pass filters, symbol decision slicer and phase error estimator form a type of phase detector with a low pass filter. The phase error is fed to a loop amplifier to tune the VCO. The model in Figure 2–28 can be used to estimate the behavior of the carrier recovery PLL with respect to phase deviations. θ_R represents the phase error of the received IF signal and θ_D represents the phase error at the demodulator output, (2–96).

The dashed box in Figure 2–28 approximates the effects of the IQ mixers, $F_s/2$ low pass filters, adaptive equalizers, decision slicer and phase error calculator. The LP function is modeled with

$$G_{LP}(f) \approx |G_{Nq}(f)| \cdot e^{-j \cdot 2\pi \cdot f \cdot T_d} \quad (2-97)$$

where $|G_{Nq}|$ is the Nyquist filter amplitude response and the phase response is modeled for simplicity by a time delay T_d .

Figure 2–28: Model for carrier recovery PLL

The open loop G_L in Figure 2–28 is

$$G_L(f) = G_{amp}(f) \cdot G_{LP}(f) \cdot \frac{K_{vco}}{j \cdot f} \quad (2-98)$$

The transfer function of most interest is the magnitude response

$$|W(f)| = \left| \frac{\theta_D(f)}{\theta_R(f)} \right| = \left| \frac{1}{1 + G_L(f)} \right| \cdot |G_{LP}(f)| \quad (2-99)$$

For a type II 2nd order PLL, the loop amplifier response G_{amp} should be in the form

$$G_{amp}(f) = K_{amp} \cdot \frac{(j \cdot f + F_z)}{j \cdot f} \quad (2-100)$$

B_0 is the zero dB crossing of the Bode plot of open loop gain, and is equal to

$$B_0 = K_{amp} \cdot K_{vco} \quad (2-101)$$

With K_{amp} in V/rad and K_{vco} in Hz/V the damping factor ζ is

$$\zeta = 0.5 \cdot \sqrt{\frac{B_0}{F_z}} \quad (2-102)$$

The magnitude response $|W(f)|$, of (2-99), is plotted in Figure 2-29 for a damping factor of 0.7 ($F_z = B_0/2$) commonly used in carrier recovery PLLs.

Transfer function $|W(f)|$, denoted the suppression function, provides rejection of 40dB per decade as the offset frequency decreases below F_z . Frequencies above $F_s/2$ are rolled off by $|G_{Nq}|$.

By making B_0 higher, the rejection of the LO phase noise and microphonics is increased and acquisition is faster. However as B_0 is increased the delay in Nyquist filter T_d may cause peaking. For $B_0 > 0.05/T_d$ the PLL significantly elevates the phase noise out to $4 \times B_0$. Note that all the curves have -6 dB point at about F_z , and are essentially identical below F_z .

Higher symbol rate demodulators have less time delay since the symbol interval is shorter and the delay is a multiple of the symbol interval. Therefore they permit higher PLL bandwidths, making high symbol rate radios usually more tolerant of close in phase noise than low symbol rate radios. Typical carrier recovery PLL bandwidths for symbol rates from 2-35 Msps QAM modems are 10KHz to 100KHz.

Figure 2–29: PLL suppression function for $\zeta = 0.7$

Chapter 3

3 Design and Simulation

3.1 X-Band VCO Design and Simulation

A low-cost X-Band VCO design is derived from existing X-Band *DROs* used in *DDBS* (digital direct broadcast system) applications. Low-cost performance is achieved by replacing the DRO's high Q dielectric resonator with a printed spiral inductor, and a varactor chip for tuning. Once the VCO topology is selected, a linear circuit simulation for the two-port negative-resistance oscillator (described in Section 2.4.4) is performed. Starting values for circuit components obtained from linear simulation are refined by applying more precise nonlinear oscillator simulations using features from *Series IV LIBRA*.

At the start of this chapter a summary of an X-Band DROs is given.

3.1.1 X-Band DRO

DROs are attractive microwave sources because of their high Q , low phase noise and good output power, [12]. Although their temperature stability is fine for the

DDBS application it is not sufficient for LMDS base-stations. Another drawback of the DRO is that it has to be mechanically tuned.

The DRO specification for the local oscillator of an X-Band *LNB* (low noise block) in a DDBS application is given in Table 3-1, [12].

Parameter	Value
Supply voltage	Up to 8V
Supply current	Up to 20mA
Operating frequency	11.25GHz
Output power	+7dBm
Phase noise @ 1 KHz	-58dBc/Hz
Phase noise @ 10 KHz	-80dBc/Hz
Phase noise @ 30 KHz	-90dBc/Hz
Phase noise @ 100 KHz	-100dBc/Hz
Phase noise @ 1MHz	-120dBc
Harmonics	-40dBc
Spurious	-80dBc
Operating temperature	-40°C to +80°C
Temperature stability	±2MHz
Output impedance	50Ω
Freq. pulling (VSWR = 2 for all phases)	±2MHz

Table 3-1: DDBS DRO specification

Typical DROs at X-band are of the reflection type, [12], as shown in Figure 3–1.

Figure 3–1: Reflection type DRO

The reflection type DRO uses the concept of a two-port negative resistance oscillator in which the resonator is placed near a terminated microstrip line connected to the input port of an unstable amplifier. Near its resonant frequency, the dielectric resonator reflects power back to the amplifier, causing an oscillation build-up.

3.1.2 X-Band VCO Design Specification

Design specification for a low-cost X-band VCO is given in Table 3-2.

<i>Parameter</i>	<i>Value</i>
Supply voltage	5V
Supply current	10mA
Operating frequency	12GHz
Tuning voltage	0V to 5V
Tuning range	More than 300MHz
Output power	-5dBm
Phase noise @ 100 KHz	-100dBc/Hz
Phase noise @ 1MHz	-120dBc
Harmonics	-40dBc
Spurious	-80dBc
Operating temperature	-40°C to +80°C
Output impedance	50Ω

Table 3-2: X-Band VCO design specification

The supply voltage and current are specified for the active device specification. The VCO phase noise is going to be improved in the loop bandwidth after locking. At this point we estimate the loop bandwidth to be around 50KHz. For the frequencies outside this loop bandwidth we would keep the DRO phase noise specifications. Output power and output impedance are specified for the resistive output buffer.

3.1.3 X-Band VCO Design Topology

The VCO topology is derived from the DRO design in Figure 3–1, and is shown in Figure 3–2. Instead of a dielectric resonator we are using an on-board printed inductor L_G as the resonating element. To convert from a DRO to a VCO, the drain stub S_D from DRO configuration is replaced with the varactor diode C_V . The resistor R_G is required for stable oscillation to occur. Two quarter wave length chokes isolate RF signal paths from the DC supply and tuning voltage input. The varactor diode is AC coupled to ground through the capacitor C_2 . The resistor R_D is for drain voltage adjustment. Using resistance R_S in the source utilizes single supply operation. Output is AC coupled with the capacitor C_S . Resistors R_{ATT1} and R_{ATT2} will provide the required load and output buffering. More detailed description of the VCO circuit and its components is presented in Section 3.1.5.

Figure 3–2: X-Band VCO design topology

3.1.4 Active Device Choice

When choosing an active device, there are many options: silicon bipolars, Si MOSFETs, GaAs FETs or Gunn/IMPATT diodes [13]. In all cases, to achieve a clean oscillation and good phase noise performance, the criteria should include a low noise figure and enough gain. The silicon bipolar is a natural choice for low noise oscillators due to its intrinsic excellent flicker noise performance. However, for any good steady oscillating operation, a good rule of thumb is to use a transistor with an f_T at least two to three times the operating frequency. These conditions would require silicon transistors with an f_T up to 35GHz. Such devices, while currently under development, are not yet readily available for high volume manufacturing. Gunn and IMPATT diodes make excellent high frequency devices (50GHz and above), but their high phase noise, need for careful mechanical design and very low power efficiency make them an unsuitable choice for high volume applications. This elimination process leaves GaAs FETs as the most suitable device. GaAs FETs naturally exhibit a very high f_T , a good loop gain and enough output power in X-Band.

The NEC NE321000, [14], pseudomorphic, hetero-junction FET chip, utilizes the junction between Si-doped AlGaAs and undoped InGaAs to create high electron mobility. Device ion implantation technology minimizes the device's flicker noise and provides low $1/f$ noise performance.

Chip dimension and layout are given in Figure 3–3.

CHIP DIMENSIONS (Units in μm)

Note: All dimensions are typical unless otherwise specified

Figure 3-3: NE321000 chip, dimensions and layout

Device specification is below in Table 3-3, [14].

Parameter	Conditions	Typical Value
Noise Figure, NF	$V_{DS}=2V$ $I_D=10mA$ $f=12GHz$	0.35dB
Associated gain, G_A	$V_{DS}=2V$ $I_D=10mA$ $f=12GHz$	13.5dB
Saturated Drain Current, I_{DSS}	$V_{DS}=2V$	40mA
Pinch of Voltage, V_P	$V_{DS}=2V$ $I_D=100\mu A$	-0.7V
Transconductance, g_m	$V_{GS}=-3V$ $I_D=10mA$	55mS
Gate to Source Leakage Current, I_{GSO}	$V_{GS}=-3V$	0.5 μA

Table 3-3: NEC321000 specification

3.1.5 X-Band VCO Linear Circuit Simulation

Although it is understood that an oscillator is a circuit that behaves non-linearly under steady state operation, a linear simulation will provide a good initial circuit layout before fine tuning the design in the nonlinear simulator. The linear simulation is used to develop the matching network, determine the appropriate resonator model, and find the appropriate reactive elements that will affect the circuit's performance.

Without feedback elements, the common source FET transistor does not make a very good oscillator because of its small feedback capacitance from input to output $C_{GD0} = 0.025\text{pF}$. To generate the required output to input feedback, the design will use a common drain configuration. This structure is very unstable and makes good oscillators by using the internal capacitance feedback of the transistor, $C_{GS0} = 0.21\text{pF}$, instead of external feedback. Having selected the topology, the next step is to follow the design procedure outlined in Section 2.4.4. By varying C_V of the varactor diode we will determine the frequency at which the series negative resistance will be generated at the gate's reflection port. Selecting the correct Z_T at the source, using a transmission line in the source TL_S , C_S and a load impedance determined by the impedance transformer, maximizes the magnitude of the reflection coefficient at the gate terminal, $|\Gamma_G| > 1$. Adjusting these parameters will provide the required amount of negative resistance at the desired frequency. We can now adjust R_G and L_G so that (2-79) and (2-80) are satisfied.

Adjusting the output matching network and the amount of output coupling C_S , Figure 3-2, will affect the output power and loaded Q_L (and therefore the phase noise) of the oscillator. A higher coupling provides more output power and robustness of oscillation build-up. However, it reduces the loaded Q and therefore the phase noise performance. A lower coupling will improve phase noise but reduces the output power, and under certain circumstances such as

high temperature or over component variance, the oscillator could fail to start oscillating.

3.1.6 Device Nonlinear Model Development

In order to use the harmonic balance simulator *Series IV LIBRA*, we need a nonlinear transistor model. LIBRA supports the *TOM* (Triquint's own model) nonlinear model supplied with NEC's NE321000 chip. The choice of a nonlinear model for a FET is determined by evaluating the DC characteristics of the device and comparing these measured characteristics to characteristics of available nonlinear models. Different models implement the DC I - V curve equations differently [15]. For the device under consideration, NEC's NE321000, it was determined that the TOM model would best represent the I - V curves because the MESFET showed an almost linear increase in drain current with increasing drain voltage at lower gate voltages, and an approximately constant drain current with respect to increasing drain voltage at higher gate voltages. The first step in the extraction process is to extract the DC model parameters so the model reflects the measured I - V curves. From Table 3-4, the main DC parameters affecting the I - V curves are V_{TO} , $ALPHA$, $BETA$, $GAMMADC$, Q , $DELTA$, R_G and R_S . A good fit to the AC data cannot be achieved until a good DC fit is obtained. The TOM model parameters that most affect the AC prediction of the model are $GAMMA$, TAU , C_{DS} , C_{GS0} , C_{DSO} , R_G , R_S and the package parasitics. Once the DC and AC performance of the model is satisfactory, the model can be optimized to fit measured power and noise data (including $1/f$ noise), where applicable. Model parameters typically affect more than one type of simulation response. The value of a parameter that results in the model providing the best S parameter fit may not provide the best fit to measured noise data across a wide range of biases and frequencies. There is usually a trade-off in device model performance when developing this type of model. In general, the DC and AC parameter prediction is

approximately equivalent. Then, depending on the target application of the device, either the power or the noise performance of the device model is optimized. Sometimes the AC performance of the model is slightly degraded to improve the power or noise prediction of the model. However, the parameters A_F and K_F are the only model parameters which affect $1/f$ noise prediction and no compromises to the AC performance need be made.

Values for model parameters for NE321000 are added in Table 3-4, [14].

<i>Name</i>	<i>Value</i>	<i>Units</i>	<i>Parameter</i>
VTO	-0.774	Volts	Nonscaleable portion of the threshold voltage
VTOSC	0	Volts	Scaleable portion of the threshold voltage
ALPHA	8	1/Volts	Current saturation parameter
BETA	0.102	A/V ²	Transconductance parameter or coefficient
GAMMA	0.085		AC drain pull coefficient
GAMMADC	0.08		DC drain pull coefficient
Q	2.5		Power law exponent
DELTA	0.8	1/W	Output feedback coefficient
VBI	0.6	Volts	Built-in gate potential
IS	1e-14	Amps	Gate junction reverse saturation current
N	1		Gate junction ideality factor
RIS	0	Ω	Source end channel resistance
RID	0	Ω	Drain end channel resistance
TAU	2e-12	seconds	Transit time under gate
CDS	0.08e-12	Farads	Drain-source capacitance
RDB	5000	Ω	Dispersion source output impedance. Infinity = 0
CBS	1e-9	Farads	Dispersion source capacitance
CGSO	0.21e-12	Farads	Zero bias gate-source junction capacitance
CGDO	0.025e-12	Farads	Zero bias gate-drain junction capacitance
DELTA1	0.3	V	Capacitance saturation transition voltage parameter
DELTA2	0.2	V	Capacitance threshold transition voltage parameter
FC	0.5		Coefficient for forward bias depletion capacitance
VBR	0	Volts	Gate-drain junction reverse bias breakdown voltage. Infinity = 0V
RD	3	Ω	Drain ohmic resistance
RG	3	Ω	Gate ohmic resistance
RS	3	Ω	Source ohmic resistance
RGMET	0	Ω	Gate metal resistance. Infinity = 0
KF	2e-12		Flicker noise coefficient
AF	1.5		Flicker noise exponent
TNOM	27	°C	Temperature
XTI	3		Temperature exponent for saturation current
EG	1.43		Energy gap or band gap voltage
VTOTC	0	V/°C	VTO temperature coefficient
BETATCE	0	%/°C	BETA exponential temperature coefficient
FFE	1		Flicker noise frequency exponent

Table 3-4: LIBRA parameter definition for the TOM model

NE321000 nonlinear model schematic with values for the parasitics is given in Figure 3–4, [14],

Figure 3–4: NE321000 chip parasitics

3.1.7 X-Band VCO Nonlinear Circuit Simulation

Final values for the circuit components are achieved after nonlinear analysis and fine tuning of the circuit previously obtained by linear analysis. As mentioned before, the nonlinear harmonic balance simulator, *Series IV LIBRA* was used. LIBRA simulates the performance of an oscillator in three steps:

- The simulator looks for the frequency of oscillation
- The power output and harmonics of the oscillator are computed
- The phase noise is calculated

Difficulties in successfully simulating the oscillator circuit are typically encountered in the first two steps. However, if the linear circuit has been properly optimized for negative resistance, the first step should result in an oscillation frequency close to that for which the circuit was designed and only small adjustments to the resonator model should be needed.

The LIBRA simulation schematic files are given in Figure 3–5 and Figure 3–6.

The LIBRA test-bench simulation file is given in Figure 3–7.

Figure 3–5: X-Band VCO simulation file

Figure 3-7: Test-bench for oscillator analysis

3.1.8 X-Band VCO Simulation Results

Simulation results are given in Figure 3–8, Figure 3–9 and in Table 3-5.

Figure 3–8 shows simulated X-Band VCO tuning range as a function of varactor capacitance C_V .

Figure 3–8: Oscillation frequency vs. varactor capacitance

The Simulated X-Band VCO tuning range is around 1GHz for varactor capacitances between 0.5pF to 2.0pF. This capacitance range is taken into account when choosing the off-the-shelf varactor chip C_V .

Figure 3–9 shows simulated X-Band VCO output power as a function of varactor capacitance C_V .

Figure 3–9: Output power vs. varactor capacitance

Output power is calculated after 7dB of resistive buffering. There is less than 1dB of output power variation within 1GHz tuning range. Results meet the specification from Table 3-2, but there is less than 2dB of predicted margin.

Table 3-5 shows simulated X-Band VCO phase noise at three different frequency offsets.

Offset frequency f_m	Simulated phase noise $\Phi_{\theta_{VCO}}$
10KHz	-73dBc/Hz
100KHz	-92dBc/Hz
1MHz	-112dBc/Hz

Table 3-5: Simulated VCO phase noise

The simulated phase noise results are well below specification for the DDBS DRO. This result was expected due to the low Q of the resonator in the X-Band VCO design.

Resonator circuit consists of the printed inductor L_G and resistor R_G . A simulated value of 25Ω for R_G was required for stable oscillation to occur. The simulated inductance value for the L_G in Figure 3–5 is around 1nH. If we neglect the losses and parasitic capacitance of the inductor, quality factor Q_R of the resonator is

$$Q_R = \frac{\omega_0 \cdot L_G}{R_G} = \frac{2 \cdot \pi \cdot 12GHz \cdot 1nH}{25\Omega} = 1.7 \quad (3-1)$$

This rather low value for the Q_R is due to high value of R_G and is 3-5 times lower than the Q factor of the spiral printed inductor [16]. Therefore the simulated phase noise for the offsets greater than 50KHz is far out from the specified values in Table 3-2.

3.1.9 X-Band VCO Layout

The VCO was built with the existing *C.R.C* (Communications Research Centre) *MHMIC* (Miniature Hybrid Microwave Integrated Circuit) fabrication technology process [17]. This process features passive elements fabricated using thin-film techniques on alumina or quartz substrates with active devices that are added after processing.

Due to wide tolerances in the fabrication of MIM (metal-insulator-metal) capacitors, discrete single layer capacitors were preferred. Discrete chip ICs were used as they were most compatible with the process. Interconnections were made using either first level metal or air bridges (second level metal), with metallized via holes providing contact to the ground plane on the reverse side of substrate. Processing is done on 0.010" alumina with $\epsilon_r = 9.8$. The process sequence developed at C.R.C. generally requires five to six masks for circuits containing transmission lines, resistors and capacitors.

The following is a summary of the passive circuitry used in this design:

- Resistors are thin-film, fabricated using a Ti:W seed layer. Sheet resistance is specified at $50\Omega/[\text{per unit square}]$ nominal. The thin-film resistor is modeled using the 'TFR' element in LIBRA.
- Spiral inductor is rectangular, modeled with LIBRA's 'MRIND' element. 'MRIND' model should be considered reliable, as for our inductance value of 1nH self-resonant frequency is higher than 12GHz.
- Capacitors are ceramic, single layer chips.
- The varactor diode, Alpha Industries GMV7821, [18], has been chosen to meet the simulation result from Figure 3–8 within an operating voltage tuning range of 5V. It is in a chip package.
- Decoupling chip capacitor C_1 , (on the VCO control voltage input) of 1000pF is used only for the evaluation of the free running VCO.

The VCO layout is given on Figure 3–10.

The resistor marked with 50 OHM is used for processing purposes.

Figure 3–10: Low-cost X-Band VCO layout

3.2 Synthesizer Design and Simulation

In this section the X-Band synthesizer design is presented. The X-Band VCO signal is divided by 8 and translated to the L-Band frequency range. The National LMX-2326 frequency synthesizer is used to lock the L-Band signal to a 10MHz reference. The design of the loop filter will be optimized for optimal phase noise performance.

3.2.1 X-Band Synthesizer Block Diagram

The synthesizer block diagram is given in Figure 3–11, and incorporates:

- An L-Band synthesizer – National Semiconductor LMX-2326, [19].
- A 10MHz TCVCXO, [20].
- A divide by 8 – Fujitsu FMM1103VJ, GaAs frequency divider, [21].
- Output buffering, done with an attenuator at the VCO output.

Figure 3–11: X-Band synthesizer, block diagram

3.2.2 Synthesizer Design Parameters

In order to simulate the synthesizer phase noise performance, (Section 2.5.4), we need to know the PLL loop parameters: K , $\Phi_{pd\ floor}$, $\Phi_{\theta_{vco}}$, and N . We could define those parameters either for X-Band or for L-Band. Here, we will define them for L-Band, and then adjust the results for X-Band frequencies.

From (2–83), the loop gain K is a function of the phase detector charge pump gain I_{cp} , and VCO voltage gain at L-Band, K_{vco} . Within a control voltage from 1V to 5V, (Figure 4–3), $K_{vco} = 5.5\text{MHz/V}$. The charge pump current gain is set to $I_{cp}=1\text{mA}$. The synthesizer step size is $f_{step} = 1\text{MHz}$, (i.e. the 10MHz crystal is divided by $R = 10$ in Figure 2–21).

The LMX-2326 phase noise floor [19], [22] is at

$$\Phi_{pd\ floor} = -210 + 10 \cdot \log_{10}(f_{step}) = -150 \text{ [dBc/Hz]} \quad (3-2)$$

Due to the low Q of the VCO resonator, $\Phi_{\theta_{vco}}$, is inherently poor. Measuring the phase noise below offset frequencies of 100KHz with the spectrum analyzer is not possible due to signal instability. To avoid needing to perform free running VCO phase noise measurements, we presume that the overall synthesizer phase noise is coming from the phase detector within the PLL loop bandwidth, and that the free running VCO contributes to the phase noise outside the loop bandwidth. As the synthesizer step-size is 1MHz for L-Band output frequency of 1340MHz, the divide ratio is $N = 1340$.

The L-Band synthesizer design parameters are summarized in Table 3-6.

<i>Parameter</i>	<i>Value</i>
Phase detector charge pump gain	$I_{cp} = 1\text{mA}$
L-Band VCO voltage gain	$K_{vco} = 5.5\text{MHz/V}$
Phase margin	50 degrees
L-Band divide ratio N	1340
Crystal frequency	10MHz
Reference frequency	1MHz
Reference divide ratio R	10

Table 3-6: Synthesizer design parameters, L-Band

3.2.3 PLL Loop Filter Design

Numerous programs for the PLL filter synthesis exist. National Semiconductor's application software "CodeLoader 2" was used to calculate the 3rd order loop filter from Figure 2-21. For PLL loop bandwidths of 40KHz and 100KHz calculated component values are presented in Table 3-7.

<i>Loop filter components</i>	
40KHz bandwidth	100KHz bandwidth
C1=33pF	C1=8pF
C2=270pF	C2=82pF
R2=51K	R2=91K
R3=39K	R3=39K
C3=10pF	C3=10pF

Table 3-7: Loop filter components values

The phase noise component, Φ_{pd} , originating from the phase detector, can now be calculated using (2-87). Simulated Φ_{pd} for both loop bandwidths is shown in Figure 3-12.

Figure 3–12: Φ_{pd} , 40KHz and 100KHz loop bandwidth, L-Band

According to (2–88) and (2–89), the PLL is performing high-pass filtering on the VCO free running phase noise, $\Phi_{\theta_{vco}}$. Simulation results in Figure 3–13 give the attenuation level of output filtering for two different loop bandwidths.

By inspection of Figure 3–12 the phase noise coming out from the phase detector Φ_{pd} is worse by 6dB at 100KHz offset for the 100KHz loop bandwidth then for the 40KHz bandwidth. But the phase noise from the VCO $\Phi_{\theta_{vco}}$ would be 10dB better at 10KHz offset for the same loop bandwidth, (Figure 3–13). It is obvious that a compromise in the loop bandwidth is necessary, depending on the $\Phi_{\theta_{vco}}$.

Figure 3–13: High pass filtering of the free running VCO phase noise

Chapter 4

4 Measurements and Performance

An X-Band synthesizer has been built according to simulation results and design procedure highlighted in Section 3. Measurement results and performance for both the X-Band VCO and the synthesizer are presented as follows.

4.1 X-Band VCO Test Results

Two X-Band VCOs, shown on Figure 4–1, were built in accordance with the layout in Figure 3–10. Two different values for the output chip capacitance C_s have been used: $C_s=1\text{pF}$ for the VCO on the left side and $C_s=0.3\text{pF}$ for the VCO on the right side.

Figure 4–1: Two X-Band VCOs

4.1.1 VCO Tuning Range

Results for the tuning range for both X-Band and L-Band are given in Figure 4–2. and Figure 4–3, respectively. Between 1 and 5V of control voltage, the tuning range at X-Band is around 300 MHz for both values for C_s .

Figure 4–2: Measured VCO tuning range

Figure 4–3: Measured L-Band output vs. control voltage

4.1.2 VCO Output Power

Results for the output power vs. control voltage are given in Figure 4–4. In the case where $C_s = 0.3\text{pF}$, there is around 3dB of difference from the simulation results from Table 3-2.

Figure 4–4: VCO output power vs. control voltage

4.1.3 VCO Phase Noise

Phase noise measurements of the free running X-Band VCO are not included in this section. Due to the low Q of the resonator in the VCO design, phase noise is inherently degraded. Within phase offsets of less than 100KHz, phase noise measurements obtained with the spectrum analyzer would not give meaningful

results. In the next section, a design procedure for approximating the X-Band VCO free running phase noise performance is described.

4.2 X-Band Synthesizer Test Results

The X-Band synthesizer from Figure 3–11 has been built to meet the design parameters from Table 3-6. Loop components values were calculated for the closed loop bandwidth of 75KHz, and are presented in Table 4-1.

75KHz bandwidth
C1=15pF
C2=130pF
R2=72K
R3=39K
C3=10pF

Table 4-1: Loop components for 75KHz bandwidth

The X-Band divide by 8, L-Band synthesizer and 10MHz reference, were built on 3 layer board, using conventional FR-4 material. Associated schematic files are presented in Appendix I and the synthesizer board is shown in Figure 4–5.

Figure 4–5: X-Band divide by 8 and L-Band synthesizer

4.2.1 L-Band Output Test Results

Figure 4–6, shows synthesizer locked at two different frequencies.

Figure 4–6: Synthesizer at L-band

Phase noise measurements for three different frequency offsets are given in Figure 4–7, Figure 4–8 and Figure 4–9.

Figure 4–7: Phase noise @ 1KHz offset, L-Band

Figure 4–8: Phase noise @ 10KHz offset, L-Band

Figure 4–9: Phase noise @ 100KHz offset, L-Band

Out of loop bandwidth phase noise performance is given in Figure 4–10

Figure 4–10: Phase noise in ± 500 KHz offset, L-Band

To check for the reference suppression, phase noise in 2.5MHz span is shown in Figure 4–11. There is more than 75dB of reference suppression at 1MHz.

Figure 4–11: Reference suppression at L-Band

The blue line in Figure 4–12 is calculated phase noise from the phase detector Φ_{pd} . The red line on the same figure is the overall synthesizer phase noise measured at L-Band.

Measured Φ_n differs from Φ_{pd} within the loop bandwidth because the high pass filtering of the $\Phi_{\theta_{vco}}$ is not enough to remove the free running VCO phase noise (Section 2.5.4).

Figure 4-12: Synthesizer phase noise at L-Band

4.2.2 VCO Free Running Phase Noise

Once the VCO is locked we can calculate the free-running VCO phase noise spectral density in the following way.

According to (2-90), the phase noise component at the synthesizer output originating from the free running VCO, Φ_{vco} , is given as a difference between Φ_n and Φ_{pd}

$$\Phi_{vco} \left[\text{rad}^2/\text{Hz} \right] = \Phi_n - \Phi_{pd} \quad (4-1)$$

Rewriting (4–1) and using (2–89), gives the free running phase noise spectral density, $\Phi_{\theta_{vco}}$

$$\Phi_{\theta_{vco}} = \frac{\Phi_n - \Phi_{pd}}{|H_{vco}(s)|^2} \quad (4-2)$$

The green line on Figure 4–12 gives calculated and interpolated $\Phi_{\theta_{vco}}$ at L-Band.

4.2.3 X-Band Output Test Results

Theoretically for a divider by 8, the phase noise at X-band should be higher by $20 \cdot \log_{10} 8 = 18\text{dB}$, than at L-band.

The synthesizer phase noise measurements at three different frequency offsets are given in Figure 4–15, Figure 4–14 and Figure 4–15.

Figure 4–13: Phase noise @ 1KHz offset, X-Band

Figure 4–14: Phase noise @ 10KHz offset, X-Band

Figure 4–15: Phase noise @ 100KHz offset, X-Band

The out of PLL loop bandwidth phase noise is given in Figure 4–16

Figure 4–16: Phase noise in ± 500 KHz offset, X-Band

The red line in Figure 4–17 shows measured overall synthesizer phase noise Φ_n at X-Band. The green line on the same graph is the approximation for the free-running X-Band VCO phase noise spectral density, $\Phi_{\theta_{VCO}}$, and the blue line is the calculated phase detector phase noise Φ_{pd} .

Figure 4–17: Calculated and measured phase noise at X-band

4.2.4 Synthesizer Overall Performance

The X-Band synthesizer test results are summarized in the Table 4-2.

The synthesizer current consumption is dominated by the divide by 8 current consumption of 90mA.

The temperature stability is the stability of the TCVCXO.

The X-Band VCO phase noise is measured as follows:

- In Figure 4–13 marker reading is -57.5dBc and spectrum analyzer resolution bandwidth is 30Hz. Therefore, phase noise at 1KHz offset is $-57.5 - 10 \cdot \log_{10} 30 = -72\text{dBc}$.

- In Figure 4–14 phase noise at 10KHz offset is –71dBc/Hz.
- In Figure 4–15 phase noise at 100KHz offset is –74dBc/Hz.
- In Figure 4–16 phase noise at 500KHz offset is –98dBc/Hz.

<i>Parameter</i>	<i>Value</i>
Center frequency	10.7GHz
Bandwidth	±100MHz
Supply voltage	5V
Tuning voltage	5V
Supply current	100mA
Phase noise @ 1 KHz	-71dBc/Hz
Phase noise @ 10 KHz	-71.5dBc/Hz
Phase noise @ 100 KHz	-75dBc/Hz
Phase noise @ 100 KHz	-98dBc/Hz
Phase noise @ 1MHz	-108dBc/Hz
Spurious	<-75dBc
Operating temperature	-30°C to +80°C
Temperature stability	±5ppm
Tuning	8MHz step

Table 4-2: X-Band synthesizer overall performance

Further improvement of the X-Band synthesizer is discussed in Chapter 5.

4.3 X-Band Synthesizer $(S/N)_{res}$

For LMDS applications (Figure 1–5), X-Band synthesizers are used in the base stations as local oscillators for the transmit and receive sub-harmonic mixers. Both transmit and receive LO phase noise are sources of residual noise that will degrade the fade margin D (defined in the link budget). We will perform the link margin degradation analysis for the receive path; a similar analysis applies for the transmit path.

4.3.1 Weighted Phase Noise

Because the X-Band synthesizers are driving sub-harmonic mixers, the output phase noise spectral density, Φ_n in Figure 4–17, is degraded by $20 \cdot \log_{10} 2 = 6\text{dB}$. In the equation for calculating phase noise at the demodulator output (2–96) we thus have to change $\lambda_{\theta_{LO}}$ by $2^2 \cdot \Phi_n$

$$\overline{\theta_D^2} [rad^2] = \int_0^{\infty} 2 \cdot 4 \cdot \Phi_n(f) \cdot |W(f)|^2 \cdot df \quad (4-3)$$

In (4–3), the X-Band synthesizer phase noise spectral density is weighted by the PLL suppression function $|W(f)|^2$ from Section 2.6.4. The weighted $4 \cdot \Phi_n$ is shown in Figure 4–18 for three different carrier recovery loop bandwidths B_0 .

Figure 4–18: Weighted phase noise, $\zeta=0.7$, $T_d=0$

4.3.2 Link Margin Degradation

After integration of (4–3) the rms phase noise, $\theta_{D_{rms}}$, for three different loop bandwidths B_0 is given in Table 4-3.

B_0 [KHz]	$\theta_{D_{rms}}$ [rad]	$(S/N)_{res}$ [dB]	D [dB] QPSK
10	0.217	13.25	
50	0.157	16.07	3.62
100	0.096	20.30	1.05

Table 4-3: Synthesizer residual noise contribution

Defining the residual signal to noise ratio as in Kucar and Feher, [23]

$$\left(\frac{S}{N}\right)_{res} = 20 \cdot \log_{10} \left(\frac{1}{\theta_{D rms}} \right) \quad (4-4)$$

The residual SNR 's are also given in Table 4-3.

The link margin degradation D was calculated, (2-95), for a QPSK demodulator using carrier recovery loop bandwidths of $B_o=50\text{KHz}$ and $B_o=100\text{KHz}$. Results are in Table 4-3.

From Figure 2-27 it is obvious that the $(S/N)_{res}$ is not good enough for the 16-QAM LMDS applications Also, for $B_o=10\text{KHz}$, the residual SNR is too low for QPSK modulation.

4.4 Cost Analysis

The X-Band synthesizer cost analysis for the active parts used in the design is presented in Table 4-4. Prices are in US currency. The synthesizer cost estimate was done for quantities less than 1K. It does not include passive components, printed circuits boards or packaging.

<i>Part description</i>	<i>Unit price/Qty</i>
Microwave Transistor, NEC 321000	\$35/100
Divide by 8, Fujitsu FMM1103V	\$32\$/10
L-Band Synthesizer, National LMX-2326	\$3/1000
10 MHz TCXO, off-the-shelf	<\$4/1000
Varactor diode, Alpha Industries GMV7821	\$1/1000

Table 4-4: Cost analysis for active parts

The estimated cost for the active parts for quantity of 1000 is in the range of \$50US.

Chapter 5

5 Conclusions

5.1 Accomplishments

In LMDS applications microwave LO sources usually drive sub-harmonic mixers to achieve output frequencies in the 20-40GHz range, so the LO phase noise performance is degraded by 6dB. That is why the LO phase noise performance is usually achieved with a DRO. To obtain the required stability for the base stations, the DRO is also phase locked to a stable reference. For frequency stepping, additional synthesizers are required together with mixers and filters to suppress unwanted spurious products. This multiplies the cost and the complexity of the LO source significantly.

This thesis has explored the design and development of a low-cost X-band synthesizer. The high stability and simple design makes it an attractive candidate for microwave oscillators in broadband wireless communications. As the phase noise performance is degraded in comparison with the PLDRO, fade link margin would be degraded. In the case of lower modulation schemes and higher

modulation rates, this degradation might not be significant, or could be compensated with other radio parameters. The X-Band synthesizer developed and designed in this thesis can support a QPSK modulation where carrier recovery loop bandwidth is close to 100KHz.

In Section 2.3.5 a novel approach has been introduced to evaluate amplitude modulation in resonant oscillators. Calculated power spectral density $\Phi_{x_{osc}}$ is then normalized in dBc/Hz and compared with the phase noise spectral density $\Phi_{\theta_{vco}}$. As expected, in the case of large signal-to-noise ratios, most of the oscillator noise is coming from the orthogonal noise component, which induces phase modulation of the carrier.

5.2 Future Work

For higher modulation schemes significant improvements are required:

Firstly and the most important requirement is to improve the phase noise of the free running VCO. By improving the phase noise $\Phi_{\theta_{vco}}$, (Figure 4–17), by 10dB, higher modulation schemes such as 16-QAM might be easily supported. The low Q of the resonator circuit in the gate of the X-Band VCO should be improved by capacitive matching at the frequency of oscillation, and by removing the resistor R_G . A higher resonator Q would directly reduce $\Phi_{\theta_{vco}}$ within the resonator tank.

Secondly an improved $\Phi_{\theta_{vco}}$ would allow a narrower PLL design. Resistors R2 and R3 in the loop filter in Figure 2–21, have large values for a loop bandwidth of 75KHz, (Table 4-1). The rms value of the resistor thermal noise is applied to the varactor diode and modulates the VCO, creating frequency noise. Integrated frequency noise adds as the phase noise to the free running VCO phase noise

$\Phi_{\theta_{vco}}$. By reducing resistor values using narrower loop filters, the phase noise component from the resistor thermal noise could be significantly reduced.

Third, there is a 7dB of loss in the resistive buffer at the VCO output in Figure 3–10. Synthesizer output power is thus too low to drive any type of millimeter wave mixer. Whereas a properly matched VCO output over the operating frequency range could have increased the output power level significantly.

Lastly, building an X-Band VCO in SMD technology and new thin teflon circuit board materials, should significantly reduce the synthesizer costs, size and improve overall performance up to 20GHz.

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APPENDIX I