PM-1 3½"x4" PHOTOGRAPHIC MICROCOPY TARGET
NBS 1010a ANSI/ISO #2 EQUIVALENT

PRECISIONSM RESOLUTION TARGETS

PIONEERS IN METHYLENE BLUE TESTING SINCE 1870

 

National Library
of Canada

Bibliothèque nationale
du Canada

Canadian Theses Service

Service des thèses canadiennes

Ottawa, Canada
K1A 0N4

NOTICE

The quality of this microform is heavily dependent upon the quality of the original thesis submitted for microfilming. Every effort has been made to ensure the highest quality of reproduction possible.

If pages are missing, contact the university which granted the degree.

Some pages may have indistinct print especially if the original pages were typed with a poor typewriter ribbon or if the university sent us an inferior photocopy.

Reproduction in full or in part of this microform is governed by the Canadian Copyright Act, R.S.C. 1970, c. C-30, and subsequent amendments.

AVIS

La qualité de cette microforme dépend grandement de la qualité de la thèse soumise au microfilmage. Nous avons tout fait pour assurer une qualité supérieure de reproduction.

S'il manque des pages, veuillez communiquer avec l'université qui a conféré le grade.

La qualité d'impression de certaines pages peut laisser à désirer, surtout si les pages originales ont été dactylographiées à l'aide d'un ruban usé ou si l'université nous a fait parvenir une photocopie de qualité inférieure.

La reproduction, même partielle, de cette microforme est soumise à la Loi canadienne sur le droit d'auteur, SRC 1970, c. C-30, et ses amendements subséquents.

**The Use of GaAs Astable Multivibrators for
High Frequency Oscillators**

by

Dennis Rosenauer, B.A.Sc.

A thesis submitted to
the Faculty of Graduate Studies and Research
in partial fulfilment of
the requirements for the degree of

Master of Engineering

Ottawa-Carleton Institute for Electrical Engineering
Faculty of Engineering
Department of Electronics
Carleton University

© copyright 1991

Dennis Rosenauer

National Library
of Canada

Bibliothèque nationale
du Canada

Canadian Theses Service Service des thèses canadiennes

Ottawa, Canada
K1A 0N4

The author has granted an irrevocable non-exclusive licence allowing the National Library of Canada to reproduce, loan, distribute or sell copies of his/her thesis by any means and in any form or format, making this thesis available to interested persons.

The author retains ownership of the copyright in his/her thesis. Neither the thesis nor substantial extracts from it may be printed or otherwise reproduced without his/her permission.

L'auteur a accordé une licence irrévocable et non exclusive permettant à la Bibliothèque nationale du Canada de reproduire, prêter, distribuer ou vendre des copies de sa thèse de quelque manière et sous quelque forme que ce soit pour mettre des exemplaires de cette thèse à la disposition des personnes intéressées.

L'auteur conserve la propriété du droit d'auteur qui protège sa thèse. Ni la thèse ni des extraits substantiels de celle-ci ne doivent être imprimés ou autrement reproduits sans son autorisation.

ISBN 0-315-75973-9

Canada

The undersigned recommend to the Faculty of Graduate Studies
and Research acceptance of the thesis

**The Use of GaAs Astable Multivibrators for
High Frequency Oscillators**

submitted by Dennis Rosenauer, B.A.Sc.
in partial fulfilment of the requirements
for the degree of
Master of Engineering

A handwritten signature in dark ink, appearing to read "J. Wright", is written over a horizontal line.

Thesis Supervisor

A handwritten signature in dark ink, appearing to read "J. Wright", is written over a horizontal line.

Chairman, Department of Electronics

Carleton University

September 13, 1991

Abstract

High frequency and microwave oscillators are used in many communications systems. Their design is most often based on feedback amplifier techniques which does not lend itself to efficient integrated circuit fabrication. This thesis explores the use of astable multivibrators and other relaxation oscillators, built as GaAs Monolithic Microwave Integrated Circuit components, as microwave signal sources. Several oscillator topologies are explored and compared against conventional oscillators for both fixed frequency and voltage controlled oscillator applications.

Acknowledgments

I would like to take a moment to express my sincere appreciation to those who helped make this work possible. First, I would like to thank Professor J. S. Wight for his guidance and supervision as well as providing relevant research topics for study. I would also like to express my gratitude to Dr. M. Stubbs at the Communications Research Centre for making facilities and resources available to fabricate and measure the circuits. Finally, I would like to thank Carole Glaser for the time spent in probing wafers and laying out the MMIC oscillators.

Table of Contents

Abstract	iii
Acknowledgments	iv
Table of Contents	v
List of Figures	viii
List of Symbols and Terms	x
 Chapter 1 Introduction	 1
1.0 Introduction	1
1.1 Thesis Objective	2
1.2 Thesis Outline	3
 Chapter 2 Analog Oscillators	 4
2.0 Introduction	4
2.1 Basic Oscillator Design	4
2.2 Oscillation Conditions	5
2.3 Maximum Oscillator Power	10
2.4 Phase Noise	13
 Chapter 3 Astable Multivibrators	 18
3.0 Introduction	18
3.1 Topologies	18
3.2 Cross Coupled AMV	24

3.3	Opamp AMV	29
3.4	Phase Noise	34
3.5	Opamp AMV Phase Noise	34
3.6	Time Domain Oscillator Phase Noise Analysis	37
Chapter 4 Oscillator Design		44
4.0	Introduction	44
4.1	Device Models	44
4.2	Cross Coupled AMV	48
4.3	Opamp Oscillator	66
4.4	Phase Noise	72
Chapter 5 Experimental Results and Analysis		73
5.0	Introduction	73
5.1	Test Results	73
5.2	Analog Oscillator Comparison	78
5.3	Phase Noise Considerations	82
Chapter 6 Conclusion		87
6.0	Oscillator Comparison	87
6.1	Areas for Further Research	89
Bibliography		90
Appendix 1: NEC71000 Cross Coupled AMV Test CKT (embedded)		92
Appendix 2: COMSAT GaAs FET Cross Coupled AMV		94

Appendix 3: COMSAT GaAs FET Cross Coupled AMV (with parasitics)	95
Appendix 4: Giga-Bit Logic Comparator Transfer Function Estimate	96
Appendix 5: Opamp Oscillator	97

List of Figures

Fig 1	Amplifier With Passive Embedding Network	7
Fig 2	Oscillator Topologies	9
Fig 3	Oscillator Equivalent Noise Model	14
Fig 4	Typical Oscillator Noise Spectrum	15
Fig 5	Three Inverter Ring Oscillator	19
Fig 6	Phase Shift Oscillator	20
Fig 7	Cross Coupled Astable Multivibrator	21
Fig 8	A Typical Clock Generator Oscillator	22
Fig 9	Opamp Comparator Oscillator	23
Fig 10	Cross Coupled AMV With Gate Diodes	25
Fig 11	Cross Coupled AMV Timing Diagram	26
Fig 12	Opamp Astable Multivibrator	29
Fig 13	Opamp AMV Timing Diagram	31
Fig 14	Opamp Oscillator	33
Fig 15	Nonlinear Voltage - Current Transfer Function	35
Fig 16	Relaxation Oscillator Waveforms Showing the Effect of Noise	39
Fig 17	Representation of Noise in the Relaxation Oscillator	41
Fig 18	Capacitor Voltage Waveforms	42
Fig 19	Probe Measurement Setup	47
Fig 20	NE71084 With Additional Transmission Lines	48
Fig 21	NE71084 Cross Coupled AMV	50
Fig 22	NE71084 AMV Including Parasitics and Test Jig Loading	51
Fig 23	Calculated AMV Gate and Drain Waveforms	52
Fig 24	Calculated AMV Output Waveform	53
Fig 25	Calculated AMV Output Spectrum	54

Fig 26	Cross Coupled AMV Using COMSAT MMIC FETS	55
Fig 27	Calculated AMV Drain and Gate Waveforms (1.33 GHz)	57
Fig 28	Calculated AMV Output Waveform (1.33 GHz)	58
Fig 29	Calculated AMV Drain and Gate Waveforms (360 MHz)	59
Fig 30	Cross Coupled AMV VCO Using COMSAT MMIC FETS	61
Fig 31	Calculated VCO Transfer Function (minimum parasitics)	62
Fig 32	Calculated VCO Transfer Function (including parasitics)	63
Fig 33	Cross Coupled AMV (COMSAT layout, fixed frequency)	64
Fig 34	Cross Coupled AMV (COMSAT layout, VCO)	65
Fig 35	Limiting Opamp Transfer Function	67
Fig 36	Limiting Opamp Model	68
Fig 37	Opamp Oscillator	70
Fig 38	Opamp Oscillator Timing Capacitor and Output Waveforms	71
Fig 39	Output Waveform	74
Fig 40	Gate Waveform	75
Fig 41	Drain Waveform	76
Fig 42	Spectrum from 0 to 1.5 GHz	77
Fig 43	Spectrum from 0 to 200 MHz	78
Fig 44	Close in Spectrum	79
Fig 45	Measured Phase Noise Slope Compared to a -9 dB Slope	83

List of Symbols and Terms

A	Embedding network gain
B	Bandwith
C_o	Carrier power
F	Noise factor at a given power level
G	Small signal gain
$G(\omega)$	Frequency dependent transconductance
I_s	Saturation current
K_{vco}	Vco transfer function coefficient
N_o	Noise power
$\tilde{P}_{in}$	Input power
P_o	Output power
$P_{osc(max)}$	Maximum oscillator output power
P_{sat}	Saturated output power
R_{eq}	Equivalent noise source resistance
V_o	Saturation voltage
V_t	Threshold voltage
V_{dd}	Drain supply voltage
$\Delta\omega$	Frequency deviation in radians
f_n	Noise bandwidth cutoff frequency
f_m	Frequency deviation from the carrier frequency
f_o	Oscillator carrier frequency
f	Frequency (in Hz)
kT	Thermal noise power
m_i	Modulation index
μ_i	Mean oscillator period

n	harmonic number
ω	Radian frequency
ω_o	Oscillator carrier frequency in radians
$\sigma(\delta f)$	Standard deviation of the oscillator jitter
σ_t	Standard deviation of the oscillator period
t	Time
τ	RC time constant
θ_d	Peak phase deviation
AMV	Astable multivibrator
FET	Field Effect Transistor
FM	Frequency modulation
GaAs	Gallium Arsenide
MMIC	Monolithic microwave integrated Circuit
Q	Resonator quality factor
RF	Radio Frequency
T	Oscillator period
VCO	Voltage controlled oscillator

Chapter 1

Introduction

1. Introduction

The primary function of an oscillator in radio frequency (RF) circuits, is to generate a given frequency and to maintain that frequency within certain limits. Oscillators are essential in communication electronics as they provide the ultimate signal source which may be modulated. They are also used to provide an auxiliary signal source which is needed for frequency conversion.

Conventional RF and microwave oscillator designs are able to meet the electrical requirements of almost any application, however, they are also large and tend to be expensive to manufacture and as a result they are only used in commercial and military communications equipment. With the current interest in personal communications there is a need for small low power communications circuitry. With silicon based integrated circuits it is possible to manufacture compact, low power and complex electronic circuitry inexpensively, however, silicon devices are limited in their high frequency ability. The inability to manufacture inexpensive high frequency integrated circuits has limited the use of the microwave frequency allocations. Since radio spectrum is a finite resource and all the low frequency allocations are already overcrowded, any new communications applications must use the high frequency and microwave allocations.

To obtain higher operating frequencies, either the circuit topology must be optimized to work with limited device parameters or a new active device technology must be used. Gallium arsenide (GaAs) devices are now becoming available in integrated form and have a much higher frequency of operation than silicon devices. As a result simple designs which are more suitable for integrated manufacture are available for use

at higher frequencies.

This thesis studies the problem of designing high frequency oscillators and voltage controlled oscillators in integrated form which differ from conventional microwave design. Integrated designs often use fewer passive components and more active devices. It is preferable to use devices which are fabricated with a single semiconductor process, and for microwave circuits, to use circuit designs that do not make use of the mechanical housing or similar structure as a resonator.

1.1. Thesis Objective

Conventional high frequency and microwave oscillator designs are based on either a resonant network feedback approach or a negative resistance approach. These design techniques will yield an oscillator which produces a nearly sinusoidal output waveform and will operate near the active device cutoff frequency. The drawback with this approach for integrated design is fabricating a resonator on a GaAs substrate. If a voltage controlled oscillator is desired then the additional problem of fabricating an integrated varactor diode becomes significant.

This thesis explores the limits of conventional oscillator design on a GaAs substrate and also explores the use of integrated GaAs astable multivibrator oscillators. It is hypothesized that although a conventional oscillator will operate near the cutoff frequency of the active device, the difficulty in fabricating a resonant structure and the large substrate area required will favor the use of an astable multivibrator which does not need a resonator and can be fabricated in a small chip area. Also it is expected that the phase noise performance of the astable multivibrators will be similar to the conventional oscillator since the on chip resonator is of low quality.

1.2. Thesis Outline

Chapter 2 outlines the basic performance expected from conventional analog oscillator designs. The conditions for oscillation at a given frequency is examined as well as estimates for the oscillator output power, spectrum and phase noise. These results are used to define a performance baseline to which astable multivibrator oscillators can be compared.

Chapter 3 outlines the performance expected from an astable multivibrator oscillator. Various topologies are examined for their suitability for fabrication using GaAs FET devices. Also the frequency of oscillation, output power, spectrum and phase noise are examined.

Chapter 4 presents the calculated performance expected from an astable multivibrator oscillator. Several topologies are discussed and particular attention is paid to MMIC designs.

Chapter 5 presents experimental results of the astable multivibrator and compares the results to previously simulated performance. A comparison is also made between astable multivibrators and analog oscillators using the same active device.

Chapter 6 summarizes the research conclusions and gives possible areas for future research.

Chapter 2

Analog Oscillators

2. Introduction

In this chapter we will consider the design of conventional analog oscillators and VCOs. By analog oscillators we mean oscillators which consist of an amplifier with a feedback network arranged such that positive feedback occurs at a given frequency. In contrast Astable Multivibrators (AMVs) consist of a device configured as a switch or switches coupled by a delay line. Analog oscillators must be considered in this thesis since they are the performance standard to which the AMV type oscillators will be judged.

When an oscillator is required for a given application the designer must know the desired frequency or frequencies of operation, the output power, the spectral purity, the long term frequency stability and close in carrier phase noise. When an analog oscillator is designed for fabrication as a GaAs MMIC the limitations of the MMIC fabrication process puts performance limits on the oscillator. These performance limits can be quite severe and will be examined in the following chapter.

2.1. Basic Oscillator Design

A linear amplifier, or any linear circuit for that matter, can be described in terms of small signal parameters. For microwave circuits scattering or S parameters are frequently used. S parameters relate the reflection and transmission coefficients of any given port to all other ports of the network or amplifier. At lower frequencies Z, H or Y parameters are used since they can be easily related to voltages or currents. S parameters are directly related to incident and reflected waves and while easier to use with transmission line type problems do not lead to a clear understanding of circuits using lumped elements. Since we are primarily concerned with lumped element

representations our analysis will be done using Y and Z parameters. If the active device is described in S parameters a conversion can be easily made to Y and Z parameters [1].

An oscillator is an inherently nonlinear device, if it were a linear device the amplitude of oscillation would build up to infinity, and would require an infinite amount of DC power to do so. There must be a mechanism in every oscillator which limits the amplitude of the oscillation. Usually a limiting action occurs due to cutoff or saturation of the active device and limits the oscillators output signal level. This would imply that linear device models would be unusable for oscillator design. This is true for determining the oscillator output power and spectrum however, the initial oscillation conditions and a very good approximation to the oscillator operating frequency can be made using linear circuit analysis.

2.2. Oscillation Conditions

In high frequency amplifier design we are concerned with the stability of the amplifier. In this case we wish to amplify a given signal and do not want the amplifier to generate a signal of it's own. In oscillator design we want to establish exactly those conditions which make an amplifier unstable. In order to make an oscillator operate at a given frequency we must meet several conditions. First, a portion of the output signal must be fed back into the input of the amplifier. This may be done by an external network, or in some cases by the parasitic elements of the active device itself. The second condition is that the gain of the amplifier in cascade with the feedback network including the loading effects of the feedback network must be greater than unity. If the gain were less than 1.0 then insufficient energy would be fed back to the input to sustain oscillation. The final condition is that the total phase shift of the amplifier, feedback network combination is 0 or a multiple of 360 degrees at the frequency of operation.

These three conditions must be met simultaneously in order to start oscillations at a given frequency. When designing an oscillator we must contrive an electrical network which will provide feedback around an amplifier to meet the above conditions. There are many networks available which can meet these requirements. We will consider only a pi network. It can be easily shown that a T network will give an equally useful oscillator and would be chosen to meet physical constraints rather than electrical ones.

Consider the amplifier with passive feedback network shown in Fig 1. The amplifier is described by a set of complex Y parameters at any given frequency and the feedback or embedding network consists of three complex impedances. We now allow the amplifier input and output voltages and currents to be V_1, I_1, V_2, I_2 respectively. If we apply the analysis shown in [2] and define a complex voltage ratio

$$A = \frac{V_2}{V_1} = A_r + jA_i \quad [2.1]$$

and set the currents

$$I_1 + I_H = 0 \quad [2.2]$$

$$I_2 + I_L = 0 \quad [2.3]$$

we can meet the conditions for oscillation at any given frequency up to the point where the active device can no longer provide sufficient gain to make the total loaded loop gain greater than one. If we separate the embedding network load into real and imaginary conductances, and substitute AV_1 for V_2 and solve the node equations we get a set of four simultaneous equations.

Fig 1 Amplifier With Passive Embedding Network

$$\begin{bmatrix} 1 & 0 & 1-A_r & 0 & 0 & A_i \\ 0 & 0 & -A_i & 1 & 0 & 1-A_r \\ 0 & A_r & A_r-1 & 0 & -A_i & -A_i \\ 0 & A_i & A_i & 0 & A_r & A_r-1 \end{bmatrix} \begin{bmatrix} G_1 \\ G_2 \\ G_3 \\ B_1 \\ B_2 \\ B_3 \end{bmatrix} = \begin{bmatrix} -g_{11} - \text{Re } Ay_{12} \\ -b_{11} - \text{Im } Ay_{12} \\ -g_{21} - \text{Re } Ay_{22} \\ -b_{21} - \text{Im } Ay_{22} \end{bmatrix} \quad [2.4]$$

If two of the six unknowns ($G_1, G_2, G_3, B_1, B_2, B_3$) are chosen then the remaining elements can be calculated provided that A is chosen such that the device generates power and that the conductances G_1, G_2, G_3 are chosen to be positive and such that the total power dissipated is less than the power generated by the active device.

Vehovec [2] states that an optimum value for A exists which delivers the maximum power to the embedding network. Since we must choose two of the unknowns in the embedding network, we choose two of the conductances equal to zero. We do this since we want all of the oscillator power delivered to one load. If we now optimize A in both magnitude and phase until the real part of the power dissipated in the active device is maximally negative (it generates a signal) we find in [2] that the optimum value for A is

$$A_{opt} = \frac{-(y_{21} + y_{12}^*)}{2g_{22}} \quad [2.5]$$

If we set G_2 and G_3 to zero and substitute [2.5] into [2.4] we can get

$$G_1 = \frac{|y_{21} + y_{12}^*|^2 - 4g_{11}g_{22}}{4g_{22}} \quad [2.6]$$

This now gives us the maximum load conditions for port 1. We can also solve for G_2 or G_3 with the other conductances set to zero. This then gives us three possible oscillator topologies. If we were to repeat this analysis using a T embedding network and worked in Z parameters, following the analysis given by Kotzebue [3] we would also get three more oscillator topologies.

Fig 2 Oscillator Topologies Obtained by Setting $G_2 G_3$,
 $G_1 G_3$ or $G_1 G_2$ to Zero

The embedding networks determined by the above procedure are optimal from the point of view of getting the maximum power from a given device. If the oscillator is loaded any more heavily than what is calculated by [2.6] above, there will not be enough energy fed back into the amplifier input to sustain oscillation. Since the

analysis done so far is based only on linear device parameters we cannot yet determine the oscillator output power or spectrum. The embedding network calculated above consists only of the admittances required to start oscillation. To determine what frequency we oscillate at we must convert the admittances to inductors and capacitors at the given oscillation frequency. Also the Y parameters for the active device must be those for the oscillation frequency. If we must design an oscillator which is to be adjustable over a frequency range, we will have to obtain device Y parameters over the adjustment range and calculate the embedding required for a number of points within the required frequency range. It should be noted that usually one of the embedding elements (usually a capacitor) is adjusted for a variable frequency oscillator and the others remain constant. The oscillator will not be *optimal* at all frequencies, however we must in any practical oscillator reduce the load by a sufficient amount to maintain reliable oscillation over a normal variation of device parameters.

2.3. Maximum Oscillator Power

To determine the oscillator output power we can no longer use true linear analysis techniques. A nonlinear device model and some form of nonlinear analysis must be used. The Y parameter model used to describe the active device is based on the assumption that the device is linear. This assumption is valid during the start of oscillation and thus can be used to determine the starting conditions for oscillation. As the signals become large we can no longer use the small signal representations. To determine the general solution for a nonlinear active device for delivering maximum power to a given load is far too complex to afford a general treatment. However, if some simplifying assumptions can be made a solution is possible for a large class of quasiharmonic oscillators containing nonlinear active devices.

In a paper by Pucel [4] it was shown that the power gain saturation characteristic of a FET amplifier could be approximated by

$$P_{out} = \frac{G P_{in}}{1 + \frac{G P_{in}}{P_{sat}}} \quad [2.7]$$

Where: G = small signal gain

P_{sat} = saturated output power of the device as an amplifier

Johnson [5] later showed that a better approximation to the saturation characteristic shown in [2.7] was

$$P_{out} = P_{sat} \left[1 - \exp \frac{-G P_{in}}{P_{sat}} \right] \quad [2.8]$$

We can use [2.8] to determine the maximum output power that can be obtained for an oscillator. This will be an approximation but is quite useful for getting a good oscillator which may then be simulated more accurately. If we want to maximize the two port added power by the oscillator we can use the argument given in [5] and maximize

$$\frac{P_{out} - P_{in}}{P_{in}} = \text{maximum} \quad [2.9]$$

which occurs when $P_{out} - P_{in}$ is maximized or

$$\frac{\partial (P_{out} - P_{in})}{\partial P_{sat}} = 0 \quad [2.10]$$

which gives

$$\frac{1}{G} = \exp \frac{-G P_{in}}{P_{sat}} \quad [2.11]$$

and substituting [2.11] into [2.9] we get

$$P_{out(max)} = P_{sat} \left[1 - \frac{1}{G} - \frac{\ln G}{G} \right] \quad [2.12]$$

We now have a simple expression for the maximum oscillator power output.

Equation [2.12] gives us an approximation of the oscillator output power. It does not however, give us any idea of the oscillator output waveform. Most quasiharmonic oscillators more or less provide an output which approximates a sinewave. In order to determine the harmonic content we must do a full nonlinear analysis. This cannot be done in the general case. Once an oscillator is designed a numerical analysis can be done for a given oscillator. Computer analysis tools like SPICE or Libra can simulate a complete oscillator if the nonlinear device model is accurate. They are also capable of either displaying the time domain output waveform or spectrum as the user desires.

A few general statements can be made about analog oscillator output waveforms. First, since the embedding network will likely look like a bandpass filter all harmonics will be attenuated to some extent. Therefore the output signal will appear to be more or less sinusoidal with second and third harmonic levels about 20 to 30 dB below the carrier for well designed oscillators. Second, if the output waveform is taken between the embedding network output and the amplifier input the signal will be *cleanest* since the embedding network will filter the limited waveform coming from the amplifier. It should be noted that any reactive loading at this point will probably cause a greater frequency shift than loading at the output of the amplifier since the amplifier acts as a buffer. Finally, care should be taken in the design of the amplifier bias networks. If there is a negative resistance region existing at other frequencies than the desired output there may be an additional oscillation due to feedback in the bias network. This is known as *squegging*. The output waveform will appear as an overmodulated AM signal. The oscillator design frequency will be the carrier frequency and the low frequency bias network oscillation will look like a pair of sidebands.

2.4. Phase Noise

In many applications it is necessary to have a low phase noise signal source. Phase noise is a random short term frequency instability of an oscillator and it can only be described in terms of a random variable. The phase noise performance of an analog oscillator can be modeled if we consider the model shown in Fig 3.

In this model the embedding network as with most oscillators is a single tuned circuit with some impedance matching. There are however, a number of noise sources fed into summing junctions and across nodes. It is the noise from these sources which can be manipulated by circuit design to minimize the phase noise of the oscillator. Also this model shows the active device noise source as a single composite noise source at the input of the device. For FETs this noise source actually consists of two noise sources, one fully correlated with the forward transconductance and the other not correlated [6]. For ease of calculation however, most phase noise analysis combine the FET noise sources into a single composite source.

A typical oscillator phase noise spectrum is shown in Fig 4. It consists of three components. The first is a background noise level due to the intrinsic noise figure of the active device and resistances in the circuit. In general the noise figure of the device predominates in this region and is given by [7] as:

$$\frac{N_o}{C_o} = \frac{F_{osc} kT}{2P_s} \quad [2.13]$$

As we move closer to the oscillator's frequency of operation we will enter a region which is increasing in noise but is shaped by the frequency characteristics of the embedding network's filter. Since the embedding network is usually a two pole resonant structure then we can speak in terms of Q and bandwidth. Within one half of the loaded resonator's bandwidth from the oscillator frequency two other effects predominate. Close in to the carrier the $1/f$ noise of the active device is predominant.

Fig 3 Oscillator Equivalent Noise Model

g'' is a composite device dependent noise parameter

SSB Phase Noise / Carrier (dB / Hz)

Fig 4 Typical Oscillator Noise Spectrum

It is due to junction surface states and the external circuit topology has little influence on it. The $1/f$ noise has a slope of 9dB/octave and intersects a 6dB/octave section of the curve at f_c . The intersection point must be measured. It cannot be calculated since at present no expression exists that describes the dependence of $1/f$ noise on oscillator configuration or on the parameters of the circuit components.

In the 6dB slope region the phase noise is controlled by the loaded Q of the embedding network's resonator. Here the noise spectrum can be shaped by the embedding network. If you have a high Q resonator you will have a narrower phase noise bandwidth. An expression can be derived for the phase noise by considering the oscillator as a feedback network. A noise source is then added to the input of the amplifier and if the total steady state loop gain is 1.0 then the noise in the 6dB slope region is given in [8] as:

$$\frac{N_o}{C_o} = \frac{FkTf_o^2}{8CQ^2f_m^2} \quad [2.14]$$

At this point we have considered only a fixed frequency oscillator. In many designs it is necessary to devise a low noise voltage controlled oscillator (VCO). In this case an additional effect must be considered. The control voltage shifts the oscillator frequency. If a noise voltage is present on the control line then we have an oscillator frequency modulated by noise. Consider a VCO with a transfer coefficient of K_{vco} and an equivalent noise resistance of R_{eq} ohms placed across the control voltage input. The RMS frequency deviation due to this noise component in the bandwidth of the oscillator resonator B and neglecting $1/f$ noise is given in [8] as:

$$f_{rms} = K_{vco} \sqrt{2kTR_{eq}B} \quad [2.15]$$

If we convert this to peak phase deviation we get:

$$\theta_d = \frac{\sqrt{2}K_{vco} \sqrt{2kTR_{eq}B}}{f_m} \quad [2.16]$$

Finally if we compute the SSB carrier to phase noise ratio we get:

$$\frac{C_o}{N_o} = 20 \log \left[\frac{\sqrt{2} K_{vco} \sqrt{2kTR_{eq}B}}{2f_m} \right] \quad [2.17]$$

This can be a very significant factor in wide band VCOs. In some cases this effect will be 20 to 40 dB higher than the noise of the same oscillator used as a fixed frequency oscillator.

In this chapter we have considered analog oscillators. This was done in order to find a baseline for the comparison of astable multivibrator type oscillators. The conditions for establishing an oscillation at a given frequency and power have been examined. Oscillator phase noise was also considered. In the next chapter an attempt to calculate the same parameters for high frequency astable multivibrators will be made.

Chapter 3

Astable Multivibrators

3. Introduction

In this chapter we will consider the design of astable multivibrators (AMVs) and their use as RF oscillators. AMVs have not been commonly used for RF oscillators in the past. With the advent of bipolar integrated circuits some single chip phase locked loops have used a form of emitter coupled AMV for the voltage controlled oscillator. In this chapter we will examine what type of AMV topologies are suitable for use with GaAs field effect devices and show some basic design equations. As in the case of analog oscillators considered previously we will consider the AMV's operating frequency, output power, output spectrum, frequency stability and phase noise.

3.1. Topologies

AMVs generally consist of one or more pulse amplifiers interconnected by delay lines. The simplest form is a ring oscillator (Fig 5.) It consists of nothing more than an odd number of inverting pulse amplifiers. The total propagation delay through all the amplifiers is one half the oscillator period. Ring oscillators are commonly used as process control devices in digital integrated circuits as the phase delay through an inverter and hence the oscillation frequency is a strong function of the gate capacitance.

If we modify the ring oscillator by removing some of the pulse amplifiers and substitute a delay line, a similar type of oscillator can be built. This is known as a phase shift oscillator. It can be analyzed in two ways, first by considering the delay through the amplifier and then the delay through the phase shift network. It can also be analyzed by considering the phase shifts only. Normally design equations are developed by the use of the phase shift mode¹ since the loading effects of the phase

$t_d = \text{propagation delay}$

$$\frac{1}{1/2 \text{ oscillator period}} = 3 t_d$$

oscillator period $T = 6 t_d$

Fig 5 Three Inverter Ring Oscillator

shift network must be considered otherwise the amplifier will have insufficient gain to sustain oscillations.

There are a large number of AMV topologies that have been very successfully used with bipolar transistors. Many of these would appear at first to be useful with GaAs field effect devices however, this is often not the case. A bipolar device intrinsically has a sharp switching threshold when the base goes from reverse to forward bias. A FET going from cutoff to conduction does so very smoothly. This effect does not

Fig 6 Phase Shift Oscillator

allow a sharp switching edge to be generated and as a result the step voltage which is usually applied to a capacitor is a smoothed *bump* which reduces the switching level during the next cycle until the oscillator stops. Another problem with FETs is that there cannot really be any gate current during normal operation. As a result care must be taken in the design to control the gate voltage excursions. Some gate current is not detrimental to the device but it must be fairly small as the gate conductor can be easily destroyed. Conventional GaAs FETs are also only available as N channel depletion mode devices and as a result complimentary metal oxide semiconductor (CMOS) designs do not readily work. Recently pseudomorphic high electron mobility devices (PHEMTs) have become available and these are enhancement mode devices. They

have not as yet been tried in this application. As a result of the above, the correct choice of FET characteristics and topologies is required to produce a workable AMV.

In an effort to find suitable AMV topologies some unlikely sources were exploited. A FET has a I/V characteristic similar to a tetrode or pentode vacuum tube. Some older papers and texts were explored to find circuit topologies. A cross coupled grid plate AMV and a cathode coupled AMV topology were found. Of these the cross coupled grid plate AMV was found to be the most suitable since the amount of gate current during the cycle was limited.

Fig 7 Cross Coupled Astable Multivibrator

Other sources of AMV topologies are those used for *on chip* clock generators found in silicon NMOS or PMOS integrated circuits. These designs which appear complex are

Fig 8 A Typical Clock Generator Oscillator

very robust. They do not rely on capacitive coupling and therefore do not suffer from the capacitive coupling problems of simpler topologies. These designs merely use FETs as simple switches. A FET switch is used to cycle the charging and discharging of a capacitor. There is some associated logic used to sequence the switches. Due to

the large number of devices involved in the switching loop these designs tend to be somewhat slow. The biggest difficulty with these designs is getting the output voltage swings to match the input switching threshold of the next stage.

Another AMV design follows from an operational amplifier. In this case the opamp is configured as a comparator with hysteresis and a feedback resistor from output to inverting input provides a source for charging and discharging the timing capacitor.

Fig 9 Opamp Comparator Oscillator

This topology is very simple and does not rely on a sharp switching transition. Fast switching is accomplished by positive feedback and therefore is suitable for fabrication using GaAs FETs. An additional advantage is that an analytical expression for the

phase noise of this topology exists. [9][10]

There are a few other AMV topologies which were investigated and found to be unsuitable. Most designs which followed from an emitter or cathode coupled topology forward biased the gate excessively in a source coupled FET version. The devices simply would not tolerate the gate current. The gate current may be limited by the use of external diodes which carry the excess gate current. This works fine, however, the additional capacitance of the diodes will lower the operating frequency of the AMV.

3.2. Cross Coupled AMV

The first AMV topology to be examined is the cross coupled AMV. It is a common design which works well with both bipolar and field effect devices. Since it is expected that AMVs will not be sufficiently stable to act as a frequency reference most designs will be used as a voltage controlled oscillator. By replacing the gate timing resistors with FETs a voltage controlled oscillator can be easily made. The design is very simple and takes very little chip area.

To determine the frequency of oscillation we can make a number of simplifying assumptions. First we will use a low frequency approximation. This neglects all parasitic capacitances and inductances of the device as well as any transmission lines used for interconnection. At low frequencies this is valid and makes a simple closed form solution possible. If an accurate high frequency solution is desired a full circuit simulator like SPICE must be employed. The second assumption is that the device will have a fairly sharp switching action occurring at or near the device threshold voltage. GaAs FETs do not really have a sharp threshold, however, this is an approximation. Finally it will be assumed that the device load resistors are much smaller than the timing resistors on the gates. If we assume that Q1 has just turned on and Q2 has just turned off then the remaining charge on capacitor C1 has been dumped into the gate of Q1 putting it into full saturation. C2 which has been charged up to nearly Vdd

Fig 10 Cross Coupled AMV With Gate Diodes

Fig 11 Cross Coupled AMV Timing Diagram

has caused the gate of Q2 to approach $-V_{dd}$ turning Q2 off. Q2 will begin to conduct when C2 discharges through R2 as the gate voltage on Q2 approaches its threshold voltage. This can be expressed as:

$$V_t = -V_{dd} \exp \frac{t}{\tau} \quad [3.1]$$

Where: $\tau = RC$; the RC time constant

at V_t we get one half of the oscillator period which gives:

$$V_t = -V_{dd} \exp \frac{t}{\tau} \quad [3.2]$$

rearranging terms we get for the oscillator period:

$$T = 2\tau \ln \left(\frac{-V_{dd}}{V_t} \right) \quad [3.3]$$

The same action as before will repeat to turn Q1 off and Q2 on.

The diodes connected from the gate to ground are used to limit the gate current when the device turns on. Some charge is left on the timing capacitors after switching and this could damage the devices in some cases. When the gate is forward biased the FET is simply a diode, and the gate voltage will be limited to the forward turn on voltage of the FET gate-channel junction. As long as the gate diode can withstand the current from the discharge of the capacitor there will be no device damage.

As we attempt to use this approximation at higher frequencies we can include some effects to help increase accuracy. If we estimate the transmission delay time of the interconnecting lines, especially between the timing capacitors and the gate resistors, and add this time delay to the oscillator period we will account for some transmission line effects. To improve accuracy to the point where we can use this

oscillator near the high frequency limit we must make use of a full nonlinear device model and include all of the parasitic elements of the device and the interconnects. This cannot be solved analytically, however, this task is not beyond a good time domain circuit simulator. It should be noted that the analysis for AMVs takes place in the time domain, whereas for analog oscillators the circuit analysis took place in the frequency domain. Time domain analysis is somewhat easier for nonlinear devices since virtually all models are time domain models.

To make a useful RF oscillator we must be able to operate the AMV at a fairly high frequency. In this case the timing resistors, capacitors, and load resistors will be as small as possible. The load resistors are limited by the amount of current the device can handle. We can design the AMV with *power* devices, however, the turn on threshold becomes very extended thus degrading performance. As a lower limit on the load resistors we must be able to guarantee that the devices will be fully turned *on* at I_{dss} . As the timing resistor becomes smaller, the approximation made earlier that the timing resistor was large compared to the load resistor is no longer valid. In this case when a device just turns on there is an appreciable voltage drop across the load resistor due to the capacitor charging current. All these effects make simple calculation of the oscillator period difficult. To design this type of oscillator for high frequency use one initially uses a crude approximation neglecting virtually all parasitic effects and then uses a circuit simulator with an optimizer to adjust the component values for the desired operating frequency.

In the first approximation the output spectrum from the cross coupled AMV is essentially that of a square wave. If the oscillator's load is the load resistor then the output power is the energy contained in the square wave swinging from V_{dd} to ground across the load resistor. In the cross coupled AMV there are two outputs available each operating 180 degrees out of phase with respect to each other. This can be useful

for some designs. Again at high frequencies approaching the limit of the active devices the simple model fails. For more accurate output power and spectrum a detailed simulation is required.

3.3. Opamp AMV

In an effort to overcome the problem of the lack of sharp switching transitions in GaAs FET devices another topology was explored which used positive feedback to speed up switching.

Fig 12 Opamp Astable Multivibrator

If we make a simple operational amplifier using GaAs FET's as a differential amplifier coupled to a source follower we can make another type of AMV. Using an ideal opamp the circuit in Fig 12 works as follows: Assume initially the capacitor is discharged and at switch-on the voltage on the inverting input was slightly lower than that on the noninverting input. The output then will be driven to the positive rail. Positive feedback occurs by resistors R1 and R2 making the noninverting input even more positive until, the output reaching the positive rail, the noninverting input is at a voltage set by the R1, R2 voltage divider. The output being positive now sources a current through Rx charging the timing capacitor C. The voltage on the inverting input increases until it reaches the voltage set by the R1, R2 divider. At this point the output goes to the negative rail. Again positive feedback by R1, R2 moves the noninverting input further negative. This positive feedback action creates a very sharp switching transition independent of the device characteristics. The timing capacitor now charges from the positive threshold voltage to the negative rail voltage. When the voltage at the inverting input passes the negative threshold the opamp will change state again and the cycle will repeat. An expression for the frequency of oscillation can be developed as follows: Assuming a symmetric saturation voltage of the opamp of $\pm V$ then the threshold voltage V_x is:

$$V_x = V \left[\frac{R_2}{R_1 + R_2} \right] \quad [3.4]$$

The timing capacitor charges in 1/2 period from $-V_x$ to $+V_x$ but is attempting to charge all the way to $+V$. This gives

$$V_{cap} = (V_x + V) \left[1 - \exp \frac{-t}{\tau} \right] \quad [3.5]$$

Fig 13 Opamp AMV Timing Diagram

Where: $\tau = R_x C$

Switchover occurs at $V_{cap} = V_x$

$$V_x = (V_x + V) \left(1 - \exp \frac{-t}{\tau} \right) \quad [3.6]$$

Rearranging terms we get for 1/2 period:

$$t = -\tau \ln \left(\frac{V}{V_x + V} \right) \quad [3.7]$$

If the oscillator operates with a nonsymmetrical output the the switching voltage threshold is different for the positive and negative swings. In this case V_x is modified for the positive and negative swings.

This oscillator can be easily made into a voltage controlled oscillator by using a control voltage on the ground end of the R1, R2 voltage divider. This changes the switching threshold V_x and thus the oscillator period as described by [3.7].

The output signal of this oscillator is again a square wave with a peak to peak value of $2V$. This also generates a harmonic-rich spectrum of fairly high power. The oscillator can drive a fairly large load, the amount of load is dependent upon the opamp design. The simplest differential opamp is a differential amplifier. Since the voltage gain should be very high the drain loads have to be very high as well. This reduces the frequency response of the opamp since the parasitic capacitances will decrease the drain load. Also if we want the opamp to be a low impedance source we will need to follow the differential amplifier with some additional gain. If we use many stages the DC gain will be quite high however, the ac gain will fall off drastically at high frequencies. A usable compromise is to follow the differential amplifier with a common drain amplifier.

Fig 14 Opamp Oscillator

3.4. Phase Noise

Analytical expressions for the phase noise of AMVs are difficult due to the extreme nonlinear nature of the oscillator. A lengthy search through the open literature yielded only two relevant papers on the calculation of the phase noise of relaxation oscillators. The paper by Boon [9] only considered the opamp type of AMV operating at 5 KHz. Although experimental results in this paper were at low frequencies there is no reason to believe that they would not be usable at microwave frequencies, especially since the transition time for the oscillator to change states is considered.

Abide [10] initially considers several oscillator designs; unfortunately only time domain statistics on the variation of oscillator period result. In order to perform a noise spectra calculation some statistics on the rate of change of oscillator period are required as well.

With numerical methods it is also difficult to model the phase noise of an oscillator since the oscillator is a nonlinear device and is modeled in the time domain. Noise however, is best described in the frequency domain. This requires a mixed mode analysis tool similar to *harmonic balance*. Also we cannot use superposition to add the noise sources to a previously modeled circuit. Since the oscillator is quite nonlinear we must include the noise sources in the nonlinear device model, otherwise the autocorrelation effects of the nonlinearities of the device would be neglected. Also numerical techniques do not provide as much insight to the noise spectrum shaping mechanisms as analytical techniques, however, they do allow analysis of a given design.

3.5. Opamp AMV Phase Noise

Boon [9] modeled the noise spectrum of an opamp AMV and found an analytic

expression describing the shape of the phase noise spectrum and the noise bandwidth. The oscillator noise can be modeled by assuming that the noise is a small signal. In this case the large signal conditions of oscillation set the operating point as a function of time and the noise signal is a small signal injected which perturbs the voltages and currents throughout the oscillator. The passive elements of the oscillator are considered linear over all signal levels, however, the opamp with its feedback networks is considered as a nonlinear voltage to current converter.

Fig 15 Nonlinear Voltage /Current Transfer Function

There are a number of noise sources in the oscillator. Each resistor, since it is above absolute zero, is generating thermal noise. The opamp has a number of noise sources as well. Since the noise is a small signal at each of the oscillator's states we can consider the opamp to be linear between state transitions. Therefore between state

transitions we can transform all the noise sources to a single equivalent noise source at the input to the opamp. The state transitions of the oscillator cause a sampling of the various noise spectra and allow the AMV noise to be modeled as a sampled system [9].

Since the noise is a small signal the transfer of noise to oscillator sidebands is linear. This allows the small modulation index for frequency modulation to be used and superposition holds for the sideband magnitudes. By modeling the equivalent noise source as a Fourier series with random phase one can follow the derivation in [9] to obtain the noise sideband magnitude relative to the total oscillator signal level equal to one half the modulation index:

$$\frac{m}{2} = \frac{a}{\Delta V_o} \left| \cos \frac{\Delta \omega T_o}{2} \right| \left| \frac{\omega_o}{\Delta \omega} \right| \quad [3.8]$$

This equation assumes that the transitions from oscillator state to state are instantaneous. This would be ideal sampling. AMV's operating at high frequencies have a finite switching time. This then causes the noise conversion to be bandlimited.

In this case the switching time of the opamp must be calculated. For a step voltage input the output current fed back to the timing capacitor must be known. This can be easily accomplished by a nonlinear circuit simulator. If the switching waveform is approximated by a straight line during the transition of slope S , and if the small signal transconductance of the opamp during the switching transition is computed and transformed to the frequency domain, we get:

$$\begin{aligned} \left[\frac{m}{2} \right]_{total}^2 &= \sum_{n=1,3}^{\infty} \left| \frac{G(n\omega_o - \Delta\omega)}{G_o} \right|^2 \cos^2 \left[\frac{\Delta\omega T_o}{2} \right] \frac{2S_{un}(n\omega_o - \Delta\omega)}{\Delta V_o^2} \left[\frac{\omega_o}{\Delta\omega} \right]^2 \\ &+ \sum_{n=1,3}^{\infty} \left| \frac{G(n\omega_o + \Delta\omega)}{G_o} \right|^2 \cos^2 \left[\frac{\Delta\omega T_o}{2} \right] \frac{2S_{un}(n\omega_o + \Delta\omega)}{\Delta V_o^2} \left[\frac{\omega_o}{\Delta\omega} \right]^2 \quad [3.9] \end{aligned}$$

This expression takes into account the bandwidth limitation of sampling. If we let the sideband noise sampling frequency $\Delta\omega$ approach the carrier we can simplify [3.9] once more and get:

$$\left(\frac{m}{2}\right)_{total}^2 = \sum_{n=1,3}^{\infty} \left| \frac{G(n\omega_o)}{G_o} \right|^2 \cos^2 \left[\frac{\Delta\omega T_o}{2} \right] \frac{4S_{\mu}(n\omega_o)}{\Delta V_o^2} \left(\frac{\omega_o}{\Delta\omega} \right)^2 \quad [3.10]$$

We now have an expression for the phase noise spectra of the opamp AMV.

Since the slope S and the saturation current I_o of the voltage to current converter completely determine the function $G(\omega)$, the effective noise bandwidth B_{eff} is:

$$B_{eff} = \frac{\int_0^{\infty} |G(\omega)|^2 d\omega}{|G(\omega)|^2} = \frac{2\pi}{3} \frac{S}{2I_o} \quad [3.11]$$

This then gives us a simple expression the bandwidth of the noise spectrum for an opamp AMV.

3.6. Time Domain Oscillator Noise Analysis

The noise analysis considered in the previous section is essentially a frequency domain analysis. It is based on time domain sampling of the AMV waveforms. Astable multivibrators, in general, are best described in the time domain and as a result a more general noise analysis can be made in the time domain. The problem with time domain analysis is one of interpreting results. Oscillator phase noise is described in the frequency domain and the conversion from the time domain solution can be difficult.

In the time domain, oscillator frequency uncertainty is called *jitter* and can be statistically defined in terms of a mean oscillator period μ , and the standard deviation σ , from the mean of the period.

$$jitter = \frac{\sigma_t}{\mu_t} \quad [3.12]$$

Abide [10] has performed a fairly general time domain analysis of relaxation oscillator noise. The analysis assumes that an oscillator consists of a charging capacitor to which a voltage sensor is attached. When the capacitor voltage crosses one of two trigger points, the sensor changes state which causes the capacitor to be discharged. The voltage sensor has a vanishingly small gain while the capacitor voltage is between trigger points. However when the trigger point is approached, the operating points of the active devices in the voltage sensor change so that the voltage sensor becomes an amplifier of varying gain. The small signal gain of the circuit is determined by an internal positive feedback loop and becomes infinitely large at the trigger point which causes the voltage sensor to regeneratively change state and reverse the capacitor charging current.

Like all other noise analysis encountered so far all the noise sources of the circuit are lumped together in a single equivalent noise source. This noise source acts to perturb the capacitor as seen by the voltage sensor. The voltage sensor's transfer function slope, being quite nonlinear because of positive feedback, approaches infinity as the trigger point is approached. The noise in the circuit is then amplified and randomly modulates the time at which the circuit switches state.

Analysis of most oscillator topologies result in an integral equation which includes voltage contributions due to the noise source. The introduction of the random variable representing the noise source makes a closed form solution virtually impossible, however some properties of the oscillator noise can be determined. Abidi [10] has observed:

- 1) The jitter appears to be directly proportional to the rms noise current for a fixed slope of the timing waveform.

Fig 16 Relaxation Oscillator Waveforms Showing the Effect of Noise

2) If the dominant noise power in the oscillator lies at high frequencies, it acts to reduce the mean period of oscillation.

3) If low frequency noise is dominant, the resulting jitter will be greater than it would be if the same noise power were concentrated at higher frequencies.

The primary result of this analysis is that the oscillator can be simplified into an equivalent noise source in series with the voltage sensor. This has the effect of adding noise to the waveform on the timing capacitor to produce an uncertainty in the time of regeneration. However it should be noted that this model only represents the oscillator for an incrementally small time before the onset of regeneration. From (1) above, if we can find the RMS voltage of the equivalent noise source we can then compute the jitter by:

$$jitter = \frac{\text{rms noise voltage}}{\text{slope of waveform at triggering point}} \times \text{a constant} \quad [3.13]$$

This is shown more clearly in Fig 18.

If we now consider the uncertainty at each transition to be statistically independent the variation of the oscillator period becomes:

$$\delta T = \frac{V_n(t_1)}{S_1} + \frac{V_n(t_2)}{S_1} + \frac{V_n(t_2)}{S_2} + \frac{V_n(t_3)}{S_2} \quad [3.14]$$

If the waveform is symmetrical in which case $S_1 = S_2 = S$ then the jitter $\sigma(\delta T)$ is:

$$\sigma(\delta T) = \alpha\sqrt{6} \frac{V_{n\text{ rms}}}{S} \quad [3.15]$$

Fig 17 Representation of Noise in the Relaxation Oscillator

Where: α is a constant depending upon whether the noise power is contained at high or low frequencies

Abide [10] shows that the distribution of jitter is Gaussian. It is however, only a statistic about the change in oscillator period. It is not in itself sufficient to determine the noise spectrum of the oscillator. In addition to statistics on the oscillator period we need statistics on the rate of change of period. Oscillator noise spectra is computed by taking a small modulation index approximation to an FM modulated signal. In FM modulation both the deviation as well as the modulating frequency is needed for the

Fig 18 Capacitor Voltage Waveforms

analysis. Many authors make the assumption that the modulation index is small and then can ignore deviation. Unfortunately we have no statistics on the rate of change of period which is the modulating frequency and cannot make a similar assumption.

If we make the possibly false assumption that the modulating frequency or the rate of change of period is small compared to the deviation or jitter, then a simple analysis can be done. We simply compute the probability of the oscillator's frequency being in any given 1 Hz interval at a given frequency away from the carrier. This

would then be proportional to the energy in the 1 Hz interval. If this is then done for all 1 Hz intervals away from the carrier, the resultant would be a graph of noise spectra.

In this chapter we have examined some of the attributes of astable multivibrators which may make them useful as RF oscillators. Operating frequency, power, output spectrum and phase noise have all been examined. Unlike the analog oscillators of the previous chapter there are not as many analytic expressions which may be used for the analysis and design of astable multivibrators, especially at high frequency. Most work needs to be done using simple time domain equations and then the result optimized using a circuit simulator. Oscillator phase noise can also be determined in some cases using analytic expressions by considering the oscillator as a sampled amplifier which cyclically changes states or by considering the oscillator in the time domain and calculating jitter statistics.

Chapter 4

Oscillator Design

4. Introduction

In this chapter we will examine the detailed design and estimated performance of AMVs as RF oscillators. Particular attention is paid to the fabrication of GaAs FET MMIC oscillators.

An initial design is done using closed form estimates and the result used as a starting point for a more detailed design employing the circuit simulator SPICE as an analysis tool.

4.1. Device Models

In order to design an oscillator of any kind we must know the characteristics of the active device. For an initial estimate very little other than the D. C. characteristics need to be known since the parasitic elements are usually ignored. In the case of an analog oscillator, which is used as a design performance reference, we make use of published S parameters for the initial design.

Obtaining a valid nonlinear device model can be very challenging indeed. This thesis explored analog oscillator and AMV designs using a millimetre wave GaAs FET available in an MMIC process from COMSAT Laboratories and a low noise packaged GaAs FET from NEC. An OpAmp type AMV was also explored using a Giga-Bit Logic comparator as a device prototype.

There are several GaAs FET models in use today. The most common, and the one used for this analysis is the *Curtice Ettenberg* in [11]. The *Curtice Ettenberg* model consists of a time domain segment which models the nonlinear portion of the device and a frequency domain segment which models the linear input and output

parasitics. The use of this model to determine the oscillator characteristics requires the use of a suitable nonlinear circuit simulator. All the analysis in this thesis was done using a variant of Berkeley SPICE marketed by EEsof which has a *Curtice Ettenberg* GaAs FET model included.

In order to obtain accurate simulations the device model coefficients must also be known. Initially the GaAs FET manufacturers were contacted to determine if any nonlinear device models were available. In many cases nonlinear device modeling coefficients are not available or in error. This was particularly the case with one MMIC vendor's GaAs FETs. It is thus necessary to determine the modeling coefficients directly from measured data. This is done by first measuring the device IV characteristics and fitting the DC component of the model to the measured values. Next the device is biased as an amplifier and the S parameters are measured over the frequency range of interest. The amplifier bias point is varied to obtain a set of S parameters at several operating points. The accuracy of the model will be improved if S parameters can be measured near cutoff and saturation as well. The device parasitic elements are now added to the model and the DC component of the device is fixed to the bias point where a given set of S parameters have been measured. At this point an optimizing algorithm is used to adjust the parasitic model elements until the best fit is obtained against the measured S parameters. This is repeated for all sets of S parameters at their respective bias points. A weighting can be added to the optimizing algorithm to enhance or reduce the effect of any given set of measured parameters. This is done to get the best overall fit of the model coefficients to the measured data.

To obtain nonlinear models for the devices used to build a test AMV both an MMIC GaAs FET from COMSAT Laboratories and a low noise packaged GaAs FET were measured. The COMSAT GaAs FET is a 0.25 μm by 0.4 mm millimetre wave power device and was measured using wafer probing techniques. The DC characteris-

tics were measured by using a programmable power supply attached to bias tees and the S parameters were measured from 1.0 to 26.0 GHz using an HP 8510 vector network analyzer.

The measured data was transferred to a larger computer and a commercial optimizing program called *Harpe* from Optimization Associates was used to fit the model coefficients to the measured data. The result was a device model which described the MMIC GaAs FET. This was done for several more devices on the same wafer to obtain a qualitative estimate of the device parameter spread.

The NE71084 is a packaged low noise GaAs FET. A set of nonlinear model coefficients were available from the vendor for an unpackaged device. Since the ceramic package adds significant parasitic elements to the overall device model additional S parameter measurements were made of the packaged device. A full optimization of the packaged device was not needed since the nonlinear model coefficients available from the manufacturer were retained. This caused an error in the S parameters obtained from the model since the parasitic elements of the package are ignored. An assumption was made that the device package was a short section of 50 ohm transmission line and that the measured S parameters of the packaged device would be a phase shifted version of the unpackaged device parameters. A new device model consisting of the unpackaged device with short transmission lines on the base and drain was then simulated to generate a new set of S parameters. When the transmission line segments on both the input and output were chosen to be 53 degrees at 7 GHz the calculated S parameters agreed fairly well with the measured S parameters of the packaged device up to about 5 GHz, which was well beyond the range of interest.

Armed with usable device models we can now begin the design and detailed simulation of the oscillators.

Fig 19 Probe Measurement Setup

Fig 20 NE71084 With Additional Transmission Lines

4.2. Cross Coupled AMV

As shown in Chapter 3 the cross AMV can be used as a simple oscillator. Fabricating this oscillator with GaAs FETs is not without some difficulties. GaAs FETs do not have a large breakdown voltage and power devices have a large threshold voltage which makes quick switching transitions difficult. A cross coupled AMV however, can be designed using GaAs FETs if these limitations are taken into account.

To determine if the AMV design would work with GaAs FETs, an initial design

was done using NEC71084 GaAs FETs. The design frequency was chosen to be about 50 MHz so that parasitic effects could be ignored. Using the approximate design equations developed in Chapter 3 the initial design is shown in Fig 21.

The component values in this design were then used for an accurate simulation using SPICE. The SPICE model included a detailed GaAs FET model which accounted for parasitic elements. Also included in the SPICE model was some stray capacitance on the gate and loading effects of the output circuit. The output circuit consisted of a coupling capacitor and resistor as well as the capacitance of the coaxial line going to the spectrum analyzer and oscilloscope. When the design was modeled without the loading effects the frequency of oscillation was about 80 MHz and when the loading effects shown in Fig 22 were included the frequency dropped to 40 MHz. As a result it should be noted that the cross coupled AMV is affected by output loading.

The SPICE simulation was used to calculate the gate and drain waveforms as well as the output spectrum. The output spectrum was computed by taking a single cycle of the drain voltage transient response and applying a Fourier Transform to the time domain waveform. These results are shown in Figs 23, 24 and 25. It should be noted that the calculated output spectrum amplitude uses a linear scale, not logarithmic as used on most spectrum analyzer plots.

This low frequency test showed that it would be worthwhile to fabricate a cross coupled AMV in MMIC form using devices with the highest f_t available. A millimetre wave power GaAs FET process was available from COMSAT through CRC and a design was completed and subitted for fabrication. This design was chosen to use the smallest practical timing components in order to achieve the highest possible frequency of oscillation. The COMSAT GaAs FET model which was based on measured devices as explained above was used and a simulation was again done using SPICE. The

Fig 21 NEC71084 Cross Coupled AMV

Fig 22 NEC71084 AMV Including Parasitics and Test Jig Loading

EE50f -- mwsPICe - 7/21/91 - 13:39:9 - DTA
 NEC71000 CROSS COUPLED AMV TEST CKT (WITH EMBEDDING)

Fig 23 Calculated AMV Gate and Drain Waveforms

EN506 - RWSPIRE - 7/2/91 - 13:50:30 - 1.7A
 NEG 1000 CROSS COUPLED AMV TEST ON WITH OVERLOAD

Fig 24 Calculated AMV Output Waveform (including loading)

EEsof - mwsPICE - 7/21/91 - 15:13:53 - JTA
 NEC71000 CROSS COUPLED AMV TEST CKT (WITH EMBEDDING)

Fig 25 Calculated AMV Output Spectrum (linear magnitude)

Fig 36 Limiting Opamp Model

interconnections between various circuit components was modeled using sections of transmission lines and this gave the design shown in Fig 26.

The SPICE simulation showed it was possible to achieve oscillation up to 1.7 GHz with zero line lengths interconnecting the AMV halves. The oscillation frequency dropped to 400 MHz with 0.5 mm interconnect lines. It is therefore necessary to minimize the interconnection line length in order to achieve the maximum frequency of oscillation. Also in this simulation it was assumed that the drain load resistors would in reality be the load of next stage driven by the oscillator so no external output load was included in the simulation.

Again the the gate and drain waveforms were calculated as well as the output spectrum. It should be noted that unlike an analog oscillator whose output spectrum is fairly clean of harmonics the astable multivibrators do have a large number of harmonics. While this is a disadvantage in designs which require spectral purity, it can be used to an advantage in cases where a frequency higher than what can be produced by the AMV is desired. In these cases the desired harmonic can be simply extracted with a bandpass filter.

During the design of the MMIC AMV an undesirable characteristic of high frequency design was noted. The COMSAT FETs are not unconditionally stable at all frequencies and there are some complex loads and sources which will cause the FET to become an *analog* oscillator. This attribute is put to good use in the design of analog oscillators but makes for difficult AMV design. As the devices cycle between saturation to cutoff the effective small signal input and output impedances vary a great deal. If the impedance presented to a device falls into the unstable region the AMV may either stop oscillating at the AMV design frequency and switch modes to an analog oscillator mode or break into an analog oscillator mode during part of the AMV cycle. To solve this problem, apart from obtaining devices which are unconditionally

Fig 27 Calculated AMV Drain and Gate Waveforms (1.33 GHz operation)

FES01 - mWSPICE - 7/28/91 - 17:21:7 - DTA
 COMSAT GAAS FET 1 CROSS COUPLED AMV

Fig 28 Calculated AMV Output Spectrum (1.33 GHz operation)

Fig 29 Calculated AMV Drain and Gate Waveforms (360 MHz operation)

stable (which is virtually impossible for microwave FETs), is to either load the input or output of the device with a fixed resistor as is done in broadband pulse amplifiers or to design a broadband embedding network to stabilize the device. In some cases it is possible to configure the design so that the devices will not see an input or output impedance which will place them into an unstable operating region.

The COMSAT AMV design was found to be stable as long as the interconnecting lines between the devices were short. As the length was increased the AMV exhibited an increased desire to oscillate at about 6 times the AMV oscillation frequency. To solve this problem in the COMSAT AMV oscillator the interconnect lengths were minimized and the drain load and gate timing resistors were chosen to provide a fixed load thus preventing excursions into the unstable region.

In most designs the inherent long term frequency stability of an AMV is inadequate and the AMV would be used as an on chip voltage controlled oscillator. In these cases it is necessary to control the oscillation frequency electrically. By replacing the gate timing resistors with FETs a voltage controlled oscillator can be built. These FETs act as voltage controlled resistors and change the RC time constant thus changing the frequency of oscillation. This design was also simulated using SPICE and a voltage to frequency transfer function was plotted in Figs 31 and 32.

By using two control FETs in the AMV design a symmetric waveform is generated. If however an asymmetric waveform is needed only one timing resistor is replaced by a control FET.

This design was laid out for fabrication by COMSAT. At the time of writing COMSAT has been unable to deliver the completed MMIC as initially scheduled. As a result no measured data are available.

Fig 30 Cross Coupled AMV VCO Using COMSAT MMIC FETS

Fig 31 Calculated VCO Transfer Function (minimum parasitics)

Fig 32 Calculated VCO Transfer Function (including interconnect parasitics)

Fig 33 Cross Coupled AMV (COMSAT layout, fixed frequency)

Fig 34 Cross Coupled AMV (COMSAT layout, VCO)

4.3. Opamp Oscillator

Another topology which can be built with minimal components is the Opamp oscillator described in Chapter 3. In this case a GaAs FET operational amplifier or comparator needs to be made. Cells for these functions are starting to become available from GaAs FET foundries as well as packaged devices from some vendors. Giga-Bit Logic for example has a comparator available as a packaged device which is claimed to work to 1 GHz.

An initial analysis was done to determine how much differential gain is required to make a useful comparator for oscillator applications. A model for a saturating opamp was made consisting of a voltage controlled voltage source of a fixed gain followed by a diode limiter and the output of the limiter is fed into a voltage follower again consisting of a voltage controlled voltage source with feedback.

If the gain is made too low there will not be sufficient output voltage swing to shift the comparison threshold at the switching points. Calculation would also be difficult since the comparison threshold would not be fixed because the opamp output would not be saturated. An initial calculation indicated that a usable gain which would put the opamp into saturation would be on the order of 100 to make a useful oscillator. With this in mind an attempt was made to fabricate an opamp using NEC71084 devices. The differential amplifier itself worked, however obtaining sufficient gain at both DC and high frequencies proved to be difficult. In order to obtain high gain a large drain load resistor or current source is needed. Since we are dealing with high frequency devices it is impossible to produce a high impedance without a large stray capacitance. As a result the differential amplifier gain was fairly high at DC and fell off rapidly as the stray capacitance effectively lowered the load impedance as frequency increased. This severely limited the frequency response. As a compromise a design was tried using a number of low gain stages to make an opamp.

Fig 35 Limiting Opamp Transfer Function

Fig 36 Limiting Opamp Model

A low gain differential amplifier was followed by a number of low gain stages in order to bring the differential voltage gain up to about 100. This could be done fairly easily for ac coupled stages but when DC coupling was tried the bias voltage levels required a level shifter at each stage interconnect. This then required a large number of devices, far too many for a test fabrication using packaged devices. Ac coupling was ruled out since the waveforms may not necessarily be symmetric about zero volts and thus when coupled would need to be clamped to a bias level to make the following stage work properly. The clamping circuits would again make the circuit too complicated for test fabrication. As a result a test design using packaged devices was abandoned.

In order to estimate the performance of an opamp AMV a commercially available comparator was used as a performance measuring stick. The highest speed GaAs comparator readily available is the Giga-Bit Logic 10G012B. The data sheets shows its differential gain starts to drop off at about 1 GHz. Since no model exists for the comparator which could be used by SPICE a model was made up from the limiting opamp model in Fig 36 above. This model consisted of a limiting opamp and some time delay elements to account for the propagation delay of the comparator. Also included in the model were resistors to account for the input leakage current and the output was also level shifted and set to 50 ohms. The sub circuit shown in Fig 36 was then used to model the operation of the high speed opamp oscillator shown in Fig 37.

As in the case of the COMSAT MMIC AMV above an attempt was made to run the oscillator at the highest possible frequency. In this case an estimate of the smallest resistors and capacitors that would be practical was made and a design completed. The result was simulated again using SPICE.

Simulation of the oscillator showed it would operate at 500 MHz. The capacitor voltage and output waveforms show a crude approximation to an actual oscillator since

Fig 37 Opamp Oscillator

stray effects cannot really be simulated as there was no component level model of the comparator. An actual MMIC fabrication of the oscillator would be more suitable for simulation since the comparator internals would be known. No MMIC fabrication was attempted since the COMSAT foundry was as yet unable to return the simple cross coupled AMV and the comparator would have had to be on a second foundry run.

No other astable multivibrator topologies were attempted.

Fig 38 Opamp Oscillator Timing Capacitor and Output Waveform

4.4. Phase Noise

As noted in Chapter 3 there is little information regarding the modeling of phase noise of relaxation oscillators. The Abide paper [10] has the most general noise analysis unfortunately it does not cover the cross coupled AMV. The analysis presented however, could serve as a prototype to assist in the development of a noise model for a cross coupled AMV. The NE71084 cross coupled AMV shown previously has been fabricated and the phase noise was measured with results presented in Chapter 5.

The opamp oscillator design was not fabricated since a second MMIC fabrication run was unavailable. The phase noise of the Giga-Bit Logic design was not computed since the equivalent noise source values are not available from the manufacturer's data sheets. A special test apparatus and measurements would have been required to determine the necessary noise parameters. Since the Giga-Bit Logic design was used only to conduct a feasibility analysis in using a high frequency comparator as an oscillator no oscillator was constructed. As a result no noise measurements were made.

In this chapter we have examined the detailed design and performance of several AMVs. They provide a high power, harmonic rich output and can be used to design VCOs with wide tuning ranges. Currently available design tools allow simulation of oscillator performance with the exception of phase noise. AMVs can be designed using several circuit topologies each with distinct advantages. Simple designs like the cross coupled AMV operate at fairly high frequencies but are more susceptible to frequency change from parameter variations such as threshold voltage. Opamp AMVs cannot operate at as high a frequency as the cross coupled AMV, however they are much less sensitive to device variation. In all cases, however, the AMVs are designed using only FETs resistors and capacitors; inductors and varactors are not needed.

Chapter 5

Experimental Results and Analysis

5. Introduction

In this chapter measurement results are presented and a comparison between these results, simulation and analog oscillators is made. Due to the unfortunate scheduling of COMSAT MMIC foundry no MMICs were available for measurement. Initially two foundry runs were to have been available which would have allowed the fabrication of both the cross coupled AMV and the opamp AMV. There is currently in progress a wafer which holds a fixed frequency cross coupled AMV and a VCO version of the cross coupled AMV. The second run was to have held an opamp AMV. The opamp AMV was put on the second run due to space limitations on the first run and since it required a fairly complex design the first MMIC run was to have provided additional modeling data on the COMSAT GaAs FETs.

In an effort to prove the concept of using an AMV in place of a conventional oscillator a cross coupled AMV was fabricated using NE71084 low noise packaged devices. The design frequency was deliberately low so that parasitics could be safely ignored. An AMV was fabricated and tested on a piece of *Duroid* board. The oscillator worked as designed.

5.1. Test Results

The oscillator drain and gate waveforms were measured using a Tektronix model 2445 oscilloscope using a X 10 probe. The output waveforms were measured using the same oscilloscope with a 50 ohm termination at the scope input. This was done in order to have the same loading as the spectrum analyzer which was used to measure the output spectrum. The oscillator output spectrum and phase noise was measured using a Hewlett Packard 8568B Spectrum Analyzer.

Fig 39 Output Waveform

Fig 39 shows the measured output waveform at the end of the isolation section used to limit the effects of the 50 ohm input impedance of the test instrument. It shows a nominal square wave of 0.2V pp into a 50 ohms load. This is in good agreement to the calculated waveform in Fig 24.

Fig 40 shows the gate waveform measured with a X 10 probe on the actual circuit. This waveform shows some *ringing* due to the somewhat long ground lead on the oscilloscope probe. If you ignore the ringing the waveform is in fairly good agreement with the calculated time domain waveform of Fig 23. It should be noted that the gate makes a definite positive excursion limited only by the gate channel junction diode going into conduction. It can also be seen that the gate voltage which should go

Fig 40 Gate Waveform

as low as $-V_{dd}$ as expected from the simple design equations, has a limiting value. However, the detailed simulation from SPICE does show a limiting value close to the measured value.

Fig 41 shows the drain waveform measured with a X 10 probe. The output with only the drain resistor loading the circuit is 4.90 V (p-p). If the drain resistor was in fact the actual load of the next stage the oscillator would be delivering about 15 milliwatts of power to each drain resistor.

Fig 42 shows the oscillator output spectrum measured from 0 to 1.5 GHz. This is measured as the output of the isolation section. The oscillator output is very rich in harmonics, which may be desirable in some applications since a frequency multiplier

Fig 41 Drain Waveform

can be avoided.

Fig 43 shows the oscillator output spectrum limited to the first 5 harmonics. At the output of the isolation section the harmonic levels are:

Harmonic No.	Measured Level (dBm)	Calculated Level (dBm)
1	-8.70 dBm	-9.15 dBm
2	-29.0 dBm	-22.4 dBm
3	-20.0 dBm	-22.4 dBm
4	-30.0 dBm	-26.9 dBm
5	-25.0 dBm	-30.1 dBm

Fig 42 Spectrum From 0 to 1.5 GHz

The isolation section reduces the measured level about 20 dB from what is generated at either drain of the oscillator. There is also some high frequency rolloff which reduces the higher harmonic amplitudes. The levels at the drain of the devices would be about 20 dB higher in an actual circuit. The calculated value shown is computed from Fig 25 which includes the effects of the isolation section.

Fig 44 shows the close in spectrum near the oscillator's fundamental frequency. The spectrum analyzer with no input has a noise floor of about the -95 dBm with a 10 KHz bandwidth. This corresponds to a -135 dBm noise floor at 1 Hz bandwidth. When the oscillator's SSB phase noise is plotted the noise spectra exhibits a measured slope -10 dB per octave. Since the oscillator is operating at a low frequency we can-

Fig 43 Spectrum From 0 to 200 MHz

not determine if a break point exists to a 6 dB per octave slope as does in an analog oscillator or if the $1/f^3$ characteristic extends all the way down to the intrinsic oscillator noise floor. To make these measurements we would have to use a varactor or similar low noise multiplier to bring up the oscillator frequency to a point where the spectrum analyzer noise floor would no longer be a problem or we would need to acquire a phase noise test system.

5.2. Analog Oscillator Comparison

The reason for examining AMVs as an alternative to conventional analog oscillators was to determine if there were any advantages in using AMVs in MMIC designs. There is little possibility of using an AMV in a hybrid design since AMVs are more

Fig 44 Close in Spectrum

complex and are likely to use a large number of FETs. In MMIC fabrication a large number of FETs is not necessarily a disadvantage since they are fabricated all at once.

The results of a test cross coupled AMV using NE71084 packaged FETs was described in the previous section. The NE71084 is a packaged low noise GaAs device often used for low noise amplifier applications up to about 12 GHz. It has an f_t of 90 GHz which would technically allow an analog oscillator to be built at that frequency. Its practical limit, however, would be about 15 GHz since the embedding network would become difficult to fabricate for a packaged device. Its use as an AMV would only be possible at a much lower frequency and would be limited by the external parasitic capacitances to about 500 MHz. The cross coupled AMV requires that the

drain resistors are small enough so that the timing capacitors can charge to near V_{dd} when the appropriate FET is *on*. The drain resistors minimum value is limited by the maximum drain current the FET can withstand. The other limiting component is the smallest size of timing capacitor we can use. If the timing capacitor is too small there will not be enough charge transferred to the gate capacitance to drop the gate voltage low enough to turn the FET off. Some of the charge from the timing capacitor will be wasted charging the parasitic capacitances shunting the gate lead. A SPICE simulation was done for the packaged device minimizing the timing components and the maximum frequency of oscillation as a cross coupled AMV was about 500 MHz.

The output power of the NE71084 as an AMV was computed and measured to be about 15 milliwatts supplied to each drain load when the device was used as a 40 MHz cross coupled AMV. This is only the power contained in the fundamental. There is significant energy in harmonics since the output waveform is approximating a square wave. An analog oscillator fabricated from the same device would be capable of:

$$P_{osc} = P_{sat} \left[1 - \frac{1}{G} - \frac{\ln G}{G} \right] \quad \text{from [2.12]}$$

From the manufacturer's data book we find:

$$\begin{aligned} P_{sat} &= 14.5 \text{ dBm} = 28.18 \text{ mW} \\ G_{max} &= 17 \text{ dB} = 50.1 \end{aligned}$$

Which gives:

$$P_{osc} = 30.94 \text{ mW} = 14.9 \text{ dBm}$$

The cross coupled AMV gives about 3 dB more power since it has two outputs rather than the single output of an analog oscillator. Any single output on the AMV pro-

duces about the same output power as the analog oscillator. This stands to reason since any AMV FET is going from saturation to cutoff using the minimum possible load resistance and this limits the maximum power available. In the analog oscillator a limiting mechanism must be in force which is usually the device running from saturation to cutoff as well. It is then expected that the power from an AMV oscillator is about the same as from an analog oscillator using the same device.

This picture changes somewhat when we consider an MMIC implementation. In this case we find that the COMSAT cross coupled AMV operates at a much higher frequency. Simulation shows that the fixed frequency oscillator will operate at a frequency greater than 1 GHz. The VCO shown in Fig 30 operates from 800 to 1750 MHz. There are two primary reasons for this increase in frequency, first, the COMSAT device is a millimetre wave FET which is optimized for operation in the 40 GHz range as opposed to the NE71084 which is optimized for low noise 12 GHz amplifier service. Second, the MMIC does not suffer from the parasitic capacitances and inductances introduced by the NE71084 package. This reduction in packaging parasitics can allow a considerable increase in operating frequency without changing the f_t of the active device. Like the NE71084 when the COMSAT device is used as an analog oscillator it can technically oscillate up to its f_t however, its usable oscillation frequency of about 60 GHz is much lower.

The fundamental output power of the COMSAT cross coupled AMV is about +19 dBm delivered to each load resistor. The COMSAT FETs are a millimetre wave power device. They are capable of in excess of 100 milliwatts at 44 GHz. They are not as well suited to AMV operation as the low noise NEC devices since their turn on threshold is very large.

In MMIC form an analog oscillator is about the same size as the cross coupled MMIC oscillator. Most of the area in an analog oscillator is taken up by the spiral

inductors. As the frequency decreases the inductors get larger but the AMV design does not change in size appreciably since the timing resistors and capacitors do not change in area as rapidly as the inductor area. An area crossover point is reached in the 1 GHz region based on observations of some of CRC's analog oscillator designs. Below 1 GHz the cross coupled AMV design is smaller. Other AMV designs will be larger than the cross coupled design since they contain more active components, however, they allow the practical design of oscillators at VHF as part of a more complex MMIC.

An area where AMV design surpass analog designs is in the insensitivity to special processing and process variations. An analog VCO requires a varactor to tune the oscillator. The process for making varactors is quite different from the process needed to make good FETs. As a result you cannot get both a good varactor and a good FET on the same substrate. The AMV on the other hand only require one type of FET for all it's circuitry. Also since the AMV's FETs are operating as switches they are more insensitive to process variations compared to an analog design. In FETs I_{dss} varies a great deal and can play havoc with the device bias. In AMVs the device merely has to switch on and off. If it is a little further into saturation than initially designed it makes no difference to the oscillator performance.

5.3. Phase Noise Considerations

The inability to compute a phase noise spectrum for the cross coupled AMV has necessitated the use of measured phase noise results. The measured phase noise exhibited a -10 dB per octave slope down to the spectrum analyzer noise floor. This agrees quite well with the -9 dB per octave slope exhibited by an analog oscillator near the carrier. An analog oscillator has a $1/f^3$ characteristic in the $1/f$ noise region. If we view a feedback oscillator as a phase servo having positive feedback we can obtain a transfer function of the phase servo [12]. The result of this analysis is that

Fig 45 Measured Phase Noise Slope Compared to a -9 dB Slope

the resonator contributes a $1/f^2$ response within the bandwidth of the resonant embedding network. The $1/f$ noise near the carrier is then superimposed on the $1/f^2$ noise creating a $1/f^3$ characteristic. In an AMV it appears that a similar mechanism is taking place, however we do not know the extent of the $1/f^2$ characteristic. If we can extend the work of Boon[9] to cover cross coupled AMVs then an effective bandwidth B_{eff} should exist describing the width of the $1/f^2$ region.

To compare the AMV phase noise to the phase noise of an analog oscillator we need to determine the phase noise power of the AMV at the crossover frequency when the noise slope changes from a $1/f^3$ characteristic to a $1/f^2$ characteristic. For the AMV using NE71084s it was found that the phase noise was decreasing at 9 dB per octave down to the noise floor of the spectrum analyzer used to measure the noise. The noise floor was about -135 dBm in a 1 Hz bandwidth. No breakpoint changing slope was noted, however one was expected to occur near the noise floor. If we make this assumption to get an estimate of a worst case condition and let the crossover frequency from the $1/f^3$ to $1/f^2$ region occur at the noise floor of -135 dBm we can use equation [2.14] to compute the equivalent Q of an analog oscillator which has the same noise characteristics. This gives:

$$Q = \left[\frac{FKTf_o^2}{8Cf_m^2 \left(\frac{N_o}{C_o} \right)} \right]^{\frac{1}{2}} \quad [5.1]$$

using:

$$f_c = 400\text{KHz at an } \frac{N_o}{C_o} \text{ of } -135 \text{ dBm}$$

$$C = \text{noise figure of device as an oscillator} = 10 \text{ dBm}$$

$$Q = 0.326$$

2

of/de

2

PM-1 3½"x4" PHOTOGRAPHIC MICROCOPY TARGET
NBS 1010a ANSI/ISO #2 EQUIVALENT

PRECISIONSM RESOLUTION TARGETS

PIONEERS IN METHYLENE BLUE TESTING SINCE 1974

1980 COUNTY ROAD 5 BURNINGWELL, WV 26037 USA
TEL: 812/426-7887 FAX: 812/426-7887 T.LX: 8100000400

which is pathetically low but gives us a consistent comparison to an analog oscillator. The loaded Q of an analog oscillator operating at 40 MHz could easily be 100 with the loaded Q increasing to 4000 for a crystal oscillator. It is therefore evident that an AMV is not a good choice for a low noise reference oscillator.

This picture changes somewhat when we consider a high frequency VCO application. In this case the AMV will be the COMSAT cross coupled AMV VCO which operates from 800 to 1750 MHz. If we extrapolate the noise performance of the NEC71084 cross coupled AMV to 1.0 GHz the Q does not change. The ratio of f_o/f_m is constant and if we assume the noise figure of the devices as an oscillator does not change appreciably then the COMSAT cross coupled AMV intersects the -135 dBm noise level at 10 MHz.

The design of an analog VCO with an octave tuning range at 1.0 GHz requires a compromise in the embedding network's resonant circuit Q. Solving for a varactor tuned resonant circuit which will resonate between 800 and 1750 MHz gives:

$$C = \frac{1}{2\pi f^2 L} \quad [5.2]$$

$$C = \frac{1}{(2\pi 800MHz)^2 2.64nH} = 1.5pF \quad [5.3]$$

$$C = \frac{1}{(2\pi 1750MHz)^2 2.64nH} = 3.1pF \quad [5.4]$$

If we attempt to raise the inductance value in an effort to increase the inductor Q we will find the the minimum capacitance required is so small that the circuit stray capacitances will be greater than the minimum. Examining the design manual for several MMIC manufacturers we find for a 2.6 nH spiral inductor at 2 GHz we get a Q of 40. This is the unloaded Q, if we assume the varactor and other capacitor Q's are high enough to be neglected, then an estimate given by Robbins [8] states that the loaded Q

is approximately the unloaded Q divided by 4. This then gives us a loaded Q of 10.

This would give us far better performance than the AMV which still has a Q of 0.326 from our previous estimate. This estimate is based on pessimistic performance on the part of the AMV, in fact we do not yet know where the cross over frequency from $1/f^3$ to $1/f^2$ region exists. Further work in measuring the phase noise to lower signal levels is required. As the VCO range is increased the analog oscillator Q will have to be decreased in order to maintain the tuning range or a range switch mechanism used. The AMV however could be extended until the leakage currents of the control FETs overwhelm the timing capacitor charging currents.

In this section we have presented our measured results on a cross coupled AMV using packaged devices. The output is of reasonably high power and is harmonic rich. The circuit simulations using SPICE agree quite well to the measured results. We have also compared the AMV to an equivalent analog oscillator to get a perception of the advantage of either design. AMVs provide a high power, harmonic rich output and can be used to design VCOs with wide tuning ranges. Also the AMV is less sensitive to device fabrication variations and does not need any special device fabrication steps. The analog oscillator on the other hand provides a *cleaner* output of about the same power. Its primary advantages are a lower phase noise and higher operating frequency.

Chapter 6

Conclusions

6. Introduction

In this thesis we have explored the use of astable multivibrators as an alternative to conventional analog oscillators for use in RF circuits. Since a large number of RF circuits are becoming integrated by the use of GaAs FET MMIC technology it is expected that circuit designs that are robust and do not require elaborate processing will be needed.

6.1. Oscillator Comparison

Traditional designs use an analog oscillator which is derived from a feedback amplifier. This topology, although simple, does not always lend itself to MMIC fabrication. In MMICs the circuits must be planar and when the resonant structure of the embedding network is converted to a planar layout it loses many desirable properties. The inductors needed for the embedding network must be of a planar spiral type. The Q of these inductors are quite low due to the high resistivity of the metallization and they also have a large number of parasitic capacitances associated with them since they are fabricated over a ground plane. Other resonant structures such as cavities and transmission lines are not possible due to their 3 dimensional nature. In some millimetre wave applications microstrip transmission lines may be used but again they are lossy due to conductor resistivity and dielectric losses. Another problem with analog oscillators is when a voltage controlled oscillator is needed. In this case a varactor is required to tune the oscillator with a control voltage. The varactor needs a different process than the GaAs FETs. This does not allow one to fabricate a good varactor and a good GaAs FET on the same substrate. In hybrid designs this is not a problem since components can come from different processes. Also since analog oscillator designs

require the embedding network to include the parasitic elements of the active device any change in device parasitics will result in a change in operating frequency. The device parasitics are also sensitive in some cases to the operating bias of the device and if the DC characteristics shift sufficiently the operating frequency can also be changed. This requires that the process be consistent otherwise the oscillator will shift from the design frequency.

Astable multivibrators use FETs as simple switches and there is no other active device required. As a result the AMV is less sensitive to process variations than the analog design. A shift in I_{dss} for example will only cause the device to change how heavily it saturates and does not affect the operating point. Also no inductors are required. The AMV derives its operating frequency from an RC time constant. When some of the timing resistors are replaced with FETs acting as voltage variable resistors a simple VCO can be fabricated. Simulation has shown it is quite easy to design a VCO with an octave tuning range and this is limited only by the leakage current of the devices. Since only FETs, resistors and capacitors are needed to make an oscillator the design lends itself quite well to planar layout.

The AMV and analog oscillator differ in the nature of their output signal. They are both of about the same level for a given active device however, the AMV contains an output that has many harmonics. This is undesirable in some applications, however it can be used to an advantage in order to obtain an output of higher frequency than is normally possible with an AMV. The desired harmonic is simply extracted with a bandpass filter.

Analog oscillators are at an advantage when a low phase noise source is desired. For fixed frequency and narrow band VCO applications the analog oscillator is far superior to the AMV with respect to phase noise. For wide band VCO operation however, the gap narrows. The loaded Q of the embedded resonant circuit is quite low

since a wide tuning range is needed. It then approaches the effective Q of an AMV. It is expected as better devices are available in MMIC form that the gap will narrow further and be in favor of the AMV approach for very wide band VCOs.

Different AMV topologies were examined for their suitability for use as RF oscillators and when GaAs FETs are used only the cross coupled AMV and the opamp AMV are usable. Other topologies such as a source coupled oscillator would have the FET drawing too much gate current which would likely damage the device.

6.2. Areas For Further Research

The major difficulties encountered in this work were the lack of an algorithm to compute the phase noise of any AMV topology, and the lack of measurement equipment to measure phase noise at low frequencies. Of these the phase noise calculation techniques is the most pressing. At this point only two paper were found which examined the phase noise of relaxation oscillators and they were primarily for the opamp and source coupled topologies. A more general work is needed. An alternate approach would be to find the phase noise spectrum by numerical methods. Discussions with others has indicated that a possible approach would be to modify a harmonic balance nonlinear circuit simulator to include noise analysis. This has already been done in the case of microwave mixer analysis by harmonic balance techniques.

Phase noise measurement was hampered by the lack of fabricated MMIC oscillators which would have allowed the measurement at a higher frequency where the equipment at CRC would have been adequate. Since only a low frequency AMV was available for test it was not possible to see the $1/f^3$ to $1/f^2$ crossover.

- [9] Boon, C. A. M., Ruten, I. W. J. M., Nordholt, E. H., *Modeling the Phase Noise of RC Multivibrators*, Proceedings of the Twenty Seventh Midwest Symposium on Circuits and Systems, Morgantown W.Va, 11-12 June 1984 pp 421-424 vol 2
- [10] Abide, Asad A., Meyer, Robert G., *Noise in Relaxation Oscillators*, IEEE Journal of Solid-State Circuits, Vol SC-18, No. 6 pp 794-802 Dec 1983
- [11] Curtice, Walter R., Ettenberg M. *A Nonlinear GaAs FET Model for Use in the Design of Output Circuits for Power Amplifiers*, IEEE Trans. on Microwave Theory and Techniques, Vol MTT-33, No. 12 pp 1383-1393 Dec 1985
- [12] Leeson, D. B. *Short Term Stable Microwave Sources*, The Microwave Journal, pp 59-69 June 1970

Bibliography

- [1] Gonzalez, Guillermo, *Microwave Transistor Amplifiers*, Prentice Hall 1984, pp 24-25
- [2] Vehovec M., Houselander L., Spence R. *On Oscillator Design for Maximum Power*, IEEE Trans. on Circuit Theory, Vol CT-15 pp 281-283 Sept 1968
- [3] Kotzebue K. L., Parrish W. J., *The use of Large Signal S-parameters in Microwave Oscillator Design*, Proc 1975 IEEE Int. Microwave Symposium on Circuits and Systems
- [4] Pucel R.A., Bera R., Masse D., *Experiments on Integrated Gallium Arsenide FET Oscillators at X-Band*, Electron Lett. Vol 11 No. 10 pp 219-220 May 15, 1975
- [5] Johnson K. M., *Large Signal GaAs MESFET Oscillator Design*, IEEE Transactions on Microwave Theory and Techniques MTT-27 No 3 March 1979 pp 217-227
- [6] Van der Ziel, A., *Noise in Solid State Device and Circuits*, John Wiley and Sons 1986 pp 183
- [7] Manassewitsch V., *Frequency Synthesizers Theory and Design*, John Wiley and Sons 1980 pp 114
- [8] Robbins W. P., *Phase Noise in Signal Sources*, Peter Peregrinus Ltd 1984 on behalf of the Institution of Electrical Engineers (UK)

Appendix 1

Spice input file for simulating the cross coupled AMV using a NEC71084 GaAs FET at low frequency. This file also generates the Fourier output spectrum by sampling the last period in the transient time domain waveform.

```
nec71000 cross coupled amv test ckt (with embedding)
r1 1 5 470
x1 1 2 0 necfet
r2 2 6 200
cs1 1 0 1pf
c1 1 3 15pf
r3 4 0 470
r4 3 6 200
cs2 4 0 1pf
c2 2 4 15pf
x2 4 3 0 necfet
vs 5 0 tran=pulse(-3 0 1ns)
vcc 6 0 dc=5
*
* equiv test ckt
*
cx 3 7 470pf
cy 7 0 30pf
rx 7 8 1000
ry 8 0 50
*
* nec71000 with extrinsic components to simulate packaged device
*
subckt necfet 100 101 102
t1 100 0 0 103 z=50 e=53deg f=7ghz
a1 104 103 102 nec4u
t2 104 0 0 101 z=50 e=53deg f=7ghz
ends necfet

model nec4u GAS MODEL=2
+A0=0.047330 A1=0.129082 A2=0.117348 A3=0.03556
+GAMMA=2.0 BETA=0.0 VDSO=3.0 vdsdc=3.0 rdso=132
+cgd=36.88ff cgs=367.5ff r1=10 r2=0 rf=5 rin=11.5
+rs=1.5 rd=1.5 rg=6.54
+vbr=6.0 vbi=0.275 cds=48.45ff

tran .1ns 100ns
file tran v(1) v(2) v(8)
end

func
tran
f1=fft(.75e-7 1e-7 10 v[8])
```

```
out
tran
  re[v[1]] gr1
  re[v[2]] gr1
  re[v[8]] gr2
  mag[f1] gr3

grid
tran
  range 0 100e-9 10e-9
  gr1 -5 5 1
  gr2 -.5 .5 .1

  range 0 10 10
  gr3 0 .1 .01
```

Appendix 2

Spice input file for simulating the cross coupled AMV using COMSAT GaAs FETs without circuit parasitics.

```

comsat gaas fet 1 cross coupled amv
c1 1 2 5pf
r1 1 0 100
x1 1 3 0 comfet
r2 3 5 25
c2 3 4 5pf
r3 4 0 100
x2 4 2 0 comfet
rm 2 6 25
vdd 5 0 dc=5
vs 6 0 pulse(0 5 .1ns)
*
* comsat fet 1 on wafer prober
* we fiddle node numbers in the subckt
*
subckt comfet 100 101 102
a1 101 100 102 com4u
ends comfet
*
model com4u GAS MODEL=2
+A0=0.12067 A1=0.108896 A2=0.000378198 A3=-0.013507
+GAMMA=2.29973 BETA=0.0635824 VDSO=0.325581 TAU=4.71682e-21
+IS=2.39527e-11 N=1.34611 CGSO=6.83556e-13 CGDO=7.13898e-14
+FC=0.5 VBI=0.4177 VBR=20
*
tran .01ns 5ns
file tran v(1) v(6) v(2)
end

out
tran
re[v(1)] gr1
re[v(6)] gr1
re[v(2)] gr1

grid
tran
range 0 5e-9 .5e-9
gr1 -10 10 2

```

Appendix 3

Spice input file for simulating the cross coupled AMV using COMSAT GaAs FETs with circuit parasitics.

```
comsat gaas fet 1 cross coupled amv (including line prop delay)
c1 1 11 5pf
t1 11 0 0 2 z=110 e=1.5deg f=1ghz
r1 1 0 100
x1 1 3 0 comfet
r2 3 5 25
c2 3 10 5pf
t2 10 0 0 4 z=110 e=1.5deg f=1ghz
r3 4 0 100
x2 4 2 0 comfet
rm 2 6 25
vdd 5 0 dc=5
vs 6 0 pulse(0 5 .1ns)
*
* comsat fet 1 on wafer prober
* we fiddle node numbers in the subckt
*
subckt comfet 100 101 102
a1 101 100 102 com4u
ends comfet
*
model com4u GAS MODEL=2
+A0=0.12067 A1=0.108896 A2=0.000378198 A3=-0.013507
+GAMMA=2.29973 BETA=0.0635824 VDSO=0.325581 TAU=4.71682e-21
+IS=2.39527e-11 N=1.34611 CGSO=6.83556e-13 CGDO=7.13898e-14
+FC=0.5 VBI=0.4177 VBR=20
*
tran .02ns 20ns
file tran v(1) v(6) v(2)
end

out
tran
re[v{1}] gr1
re[v{6}] gr1
re[v{2}] gr1

grid
tran
range 0 20e-9 1e-9
gr1 -10 10 2
```

Appendix 4

Spice input file for calculating the comparator transfer function.

```
* gigabit logic comparator transfer function estimate
vin 101 102
e1 100 0 101 102 100
r110 100 104 1000
d1 104 105 diode
d2 105 104 diode
vc1 105 0 dc=-1.2
e2 106 0 104 106 100
r101 101 0 20000
r102 102 0 2(KXX)
r103 107 106 50
r104 107 0 1meg
*
model diode d rs=10
dc vin -.5 .5 .01
file dc v(107)
end

out
dc
  re[v(107)] gr1

grid
dc
  range -.5 .5 .1
  gr1 -2.5 0 .5
```

Appendix 5

Spice input file for simulating the behavior of an opamp AMV based on the Giga-Bit Logic comparator.

```
* opamp oscillator based on gigabit logic comparator
vin 5 4 tran=pulse(2 0 1ns 1ns)
x1 1 2 3 comp
r1 3 2 3000
r2 2 4 1000
rx 3 1 100
ct 1 5 5pf
vb 4 0 dc=-1.2
*
subckt comp 102 101 107
e1 100 0 101 102 100
r110 100 104 1000
cd 104 0 10pf
d1 104 105 diode
d2 105 104 diode
vc1 105 0 dc=-1.2
e2 106 0 104 106 100
r101 101 0 20000
r102 102 0 20000
r103 107 106 50
r104 106 0 1meg
ends comp
*
model diode d rs=10
tran .01ns 20ns 0ns .01ns
file tran v(1) v(2) v(3)
end

out
tran
re[v[1]] gr1
re[v[2]] gr1
re[v[3]] gr1

grid
tran
range 0 20e-9 2e-9
gr1 -2.5 0 .5
```


END

231292

FIN