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DIRECT BASEBAND TO MICROWAVE
MSK GENERATION BY USING
INJECTION LOCKED OSCILLATOR

by

C

Stamatis A. Myrillas, Diploma Engineering

A thesis submitted to the Faculty of Graduate Studies
in partial fulfillment of the requirements for the degree of
Master of Engineering

Department of Electronics
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The undersigned recommend to the Faculty of Graduate Studies and Research, the acceptance of the thesis:

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ABSTRACT

The feasibility of direct microwave Minimum Shift Keying generation by using injection locking for stability improvement is investigated in this thesis, theoretically as well as experimentally. The phenomenon of injection locking and the time delays introduced when this technique is used for stabilization purposes is discussed. In specific a theoretical model for the breaking-lock time is developed. This breaking lock time is numerically predicted and a method for its experimental measurement is presented. Experimental investigation of the overall proposed scheme of direct microwave MSK generation is made through the development of a system transmitting at a rate of 4Mbits/sec. Performance evaluation is carried out through noncoherent demodulation and regeneration of the transmitted data. The effect of the injected signal power level on the bit-error rate of the system is concluded.

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LIST OF PRINCIPAL SYMBOLS

A	current for the active device
A_o	free-running current
B	bandwidth
BER	bit error rate
BPF	bandpass filter
BPSK	biphase shift keying
c	velocity of light
C	capacitance
C_j	junction capacitance
C_p	package capacitance
C_{rv}	capacitance during reverse bias
C_{fw}	capacitance during forward bias
dB	decibel
DSB-SC	double sideband suppressed carrier
Δf_m	locking range
e	efficiency
fb	bit rate
f_c	carrier frequency
f_{in}	input frequency
f_{out}	output frequency
$G(f)$	normalized spectral density
h	modulation index
H	substrate thickness

IF	intermediate frequency
ILO	injection locked oscillator
λ_g	wavelength in microstrip line
L	inductance
L_p	package inductance
LPF	low pass filter
MSK	minimum shift keying
OQPSK	offset quadrature phase shift keying
P_d	dissipated power
P_i	injected power
P_o	output power
QAM	quadrature amplitude modulation
Q_{ext}	external quality factor
QPSK	quadrature phase shift keying
r	saturation factor for imaginary part
R	real part of impedance
R_{in}	input resistance
R_l	load
R_s	series resistance
s	saturation factor for real part
S	spacing between two microstrip coupled lines
S_{11}	input reflection coefficient
SRD	step recovery diode
SRM	switching reactance multiplier

Tbl	breaking lock time
Tl	locking time
Tp	impulse width
Tr	thermal resistance
Ts	transition time
Ve	even mode velocity
Vo	odd mode velocity
Vbr	breakdown voltage
Vp	phase velocity in microstrip
Vt	tuning voltage
vd	varactor diode
VCO	voltage controlled oscillator
VAM	variable reactance multiplier
W	strip conductor width
X	imaginary part of impedance
Z(A)	impedance of active device
Z(ω)	output impedance of oscillator
Zo	characteristic impedance
Zoe	even mode characteristic impedance
Zoo	odd mode characteristic impedance
Zl	load impedance
'o	free-space permeability
'eff	effective dielectric constant
'r	relative dielectric constant

φ	phase angle
φ_0	phase angle after locking
ω_0	free-running oscillating frequency
ω_c	carrier frequency
ω_i	frequency of injected signal

1. CHAPTER ONE

INTRODUCTION

1.1 Introduction

During the last several years, there has been an ever-increasing interest by planners, and manufacturers toward the transmission of information in digital form. The main advantages of digital transmission is high reliability and great flexibility.

A very important parameter in digital transmission is the choice of the modulation scheme, in relation to its efficient use of the frequency spectrum, its simplicity of practical equipment design, and its resulting overall economy. Minimum-Shift Keying (MSK) is one modulation technique that possesses a number of favourable properties concerning spectrum efficiency, circuit

and power conservation.

This modulation method has the unique feature of being equivalent to either a continuous phase Frequency-shift keying (FSK) with modulation index of 0.5 or Offset-Keyed, sinusoidally-weighted Quadrature Phase-Shift Keying (QK-QPSK), and has been a subject of much recent research. Because of its equivalence to FSK, it is feasible to generate directly MSK at microwave frequencies by feeding the baseband digital data directly to the tuning port of a Voltage-Controlled-Oscillator (VCO).

Direct generation of a modulated signal at microwave frequencies offers considerable simplicity and economy at the transmitter of the microwave systems. In conventional microwave communication systems, the modulation is carried out at intermediate frequency and the modulated signal is then upconverted at microwave frequencies, Fig. 1.1. As compared to this, direct modulation at microwave frequencies offers an attractive alternative from reliability and cost considerations point of view, Fig. 1.2. However, the frequency stability of the high power VCO may be of the order of 0.05 percent. This poor stability is unacceptable as a stability of 1×10^{-6} or better is generally specified for a communication system. One way of improving

Fig. 1.1 Conventional Microwave System

Fig. 1.2 Direct Generation Microwave System

significantly the stability of the system is by using injection locking techniques.

However, the delay that would be produced by the locked carrier during the process of breaking lock and switching from one frequency to the other, could create implications in the system and put a restriction on how fast the VCO could switch in frequency. In other words it could put a restriction on the bit-rate that such a system could transmit.

1.2 Thesis Objectives

The objective of this thesis is two-fold. First to investigate the process of breaking lock between the VCO carrier and the injected signal. For this task a theoretical model will be developed and an analysis of the situation will be presented. Since the VCO and power level of the injection signal (P_i) will be specified, the breaking lock time (T_{bl}) can be determined and its value depends on the VCO parameters and on P_i . Therefore by evaluating experimentally most of the parameters of the VCO and

by using the results of the above mentioned analysis, estimates of the breaking lock time (T_{bl}), will be derived by using computational techniques. Experimental evaluation of T_{bl} will also be achieved in order to verify our theoretical estimates.

The second objective is to study the feasibility of direct generation of MSK by using an injection locked oscillator. This will be done by implementing an experimental system capable of transmitting at a rate of 4Mbits/sec in MSK form, directly generated at 1.4 GHz. For this system a microstrip frequency multiplier will be developed by using Step Recovery Diodes (SRD). Finally demodulation of the transmitted signal will be carried out in order to evaluate the performance of the system by means of eye patterns and Bit-Error-Rate (BER) degradation for different power levels of the injected reference signals.

1.3 Thesis Organization

The subsequent chapters of this thesis are arranged as follows.

In chapter two the theoretical background for the system implementation will be given. In the first section the highlights of the MSK modulation scheme will be presented in order to obtain a better insight of the system functioning. The next two sections will deal with microstrip transmission lines and frequency multiplication by using Step Recovery Diodes and contain all the necessary information for the design of the frequency multiplier that will be built and used in our system.

The theory of Injection Locking is given in the first section of chapter three. Based on this theory, a theoretical analysis concerning the Breaking Lock situation is derived and some numerical results for the breaking lock time (T_{bl}) are evaluated in section two. In the third section the experimental

measurement of Tbl is presented.

In chapter four the implementation of the 4 Mbit/sec MSK direct generation will be carried on. Detailed presentation of the system, and especially of the design of the X7 frequency multiplier that was developed and employed in the system will be given. The rest of this chapter will be concerned with the demodulation and performance evaluation of the proposed system.

Finally the conclusions and recommendations for further work can be found in chapter five of this work.

2. CHAPTER TWO

BACKGROUND INFORMATION

2.1 Minimum Shift Keying (MSK) Modulation Scheme

2.1.1 Introduction

For a communication system, the two primary communication resources are the transmitted power and channel bandwidth. Depending on which of these two resources is more precious, most of the channels can be classified primarily as power-limited or band-limited. In power-limited channels, coding schemes would be generally used to save power at the expense of bandwidth, whereas

in band-limited channels spectrally efficient modulation techniques would be used to save bandwidth. Beyond that, certain types of nonlinear channels call for an additional feature on the modulation technique, namely, a constant amplitude. A typical example of such channels is satellite links where high power amplifiers are operated near power saturation for high efficiency.

The Minimum Shift Keying modulation (MSK), which is a simple modification of the generally accepted QPSK, satisfies all the above constraints by possessing the properties listed below:

(i) Theoretical spectral efficiency of 2 bits/sec/Hz, (1)

(ii) Constant envelope

(iii) Continuous phase

(iv) Instantaneous frequency is always one of two values (f_1, f_2) and the modulation index $h = (f_2 - f_1)/f_b$ where f_b is the bit rate of the modulating data stream

(v) Simplicity and low cost of implementation of the

weight

- Large bandwidth
- Direct bonding of active and passive devices on the dielectric substrate
- Devices and components are accessible for probing and measurement.

Assuming the quasi-TEM mode of propagation, the phase velocity in microstrip is given by

$$V_p = c / \epsilon_{\text{eff}} \quad (2.5)$$

where

c is the velocity of light, and

ϵ_{eff} is the effective dielectric constant of the substrate material given by

$$\epsilon_{\text{eff}} = \left(\frac{\epsilon_r + 1}{2} \right) + \left(\frac{\epsilon_r - 1}{2} \right) \left(1 + 10 \frac{H}{W} \right)^{-1/2} \quad (2.6)$$

where

ϵ_r is the relative dielectric constant of the substrate material

Also, the wavelength λ_g in microstrip line is given by

Fig. 2.1.1 The Phasor Diagram
 (a) BPSK
 (b) QPSK
 (c) MSK

binary component changes state, then the carrier has transitions of $\pm\pi/2$ radius. The change of state of both the binary components results in a phase change of π radius.

Skewing I and Q channel sequences in time by T, the change of state does not occur simultaneously thus avoiding phase changes of π radius. Even though any abrupt phase shift of π radius is avoided in QPSK, abrupt $\pi/2$ phase transients are still present. With the sinusoidal pulse weighting of OK-QPSK, phase transitions become continuous and this modified signal can be defined as MSK and equals

$$S(t) = a_i(t) \cos(\pi t/2T) \cos(2\pi fct) + a_q(t) \sin(\pi t/2T) \sin(2\pi fct) \quad (2.1)$$

Fig. 2.1.2 shows the various components of MSK as defined by Eq. 2.1.

b. Representing MSK as a special case of CPPSK

When viewed as a CPPSK, the modulated signal waveform can be expressed as

$$S(t) = \cos(\omega_c t + \pi U_k t/2T + I_k) \quad (2.2)$$

$$(k-1)T \leq t \leq kT$$

2.1.4 BER Performance

Bit-error-rate results from simulations and actual systems show a difference of about 1 to 2 dB's between QPSK and MSK, depending on the filtering and kind of reception.

2.1.5 Direct Microwave MSK Generation Using an Injection Locked Oscillator (ILO)

Recently special interest has been given to direct generation of MSK modulated signals at microwave frequencies (4). This technique offers considerable simplicity and economy in the transmitter of a microwave communication system.

ω_c = carrier or centre frequency

U_k = kth data bit = ± 1

X_k = recursive variable required to
maintain the phase continuity at the
switching time

Obviously for a mark ($U_k=+1$) the frequency $\omega_c+\pi/2T$ is transmitted and for a space ($U_k=-1$) the frequency $\omega_c-\pi/2T$ is transmitted. By equating the phase at the (K-1)th to the Kth interbit switching instant ($t = (K-1)T$)

$$X_k = X_{k-1} + \frac{\pi(k-1)}{2} \times (U_{k-1} - U_k) \quad (2.3)$$

Equations (2.3) and (2.2) together define the MSK waveform.

2.1.3 Spectrum of MSK

The normalized spectral density of MSK, shifted to baseband, can be expressed as, (2),

$$G(f)/T = 16/\pi^2(\cos 2\pi fT/1-16f^2T^2)^2 \quad (2.4)$$

This is shown in Fig. 2.1.3a, together with the spectrums for QPSK and OK-QPSK.

A measure of the compactness of a modulation spectrum is the bandwidth B which contains 99 percent of the total power. For MSK, $B = 1.2/T$ while for QPSK, $B = 8/T$. Thus for wide-band links where filtering is not used after the nonlinearities, MSK may be spectrally more efficient than QPSK. However MSK spectrum has a wider main lobe than QPSK and for narrowband links may not be the preferred method. In Fig. 2.1.3b the relative out-of-band power is shown, (3).

(a)

(b)

Fig. 2.1.3 (a) Spectral Density, and
(b) Relative Out-of-Band Power
For MSK, BPSK, QPSK

2.1.4 BER Performance

Bit-error-rate results from simulations and actual systems show a difference of about 1 to 2 dB's between QPSK and MSK, depending on the filtering and kind of reception.

2.1.5 Direct Microwave MSK Generation Using an Injection Locked Oscillator (ILO)

Recently special interest has been given to direct generation of MSK modulated signals at microwave frequencies (4). This technique offers considerable simplicity and economy in the transmitter of a microwave communication system.

In the conventional transmitter, an IF signal is modulated and the modulated signal is upconverted using a high stability local oscillator. Further because of the power handling limitations of the upconverter a microwave amplifier is generally required to feed the requisite power to the antenna. Filtering at microwave frequencies in such a transmitter is limited to suppressing the harmonics of the transmitter frequency; the filtering to limit the spectrum being done at an IF.

In the directly modulated system, the modulation is carried out directly at microwave frequencies. The filtering is also done at microwave frequencies. If a microwave source of sufficiently high power is used, no further amplification of the modulated signal may be required.

The generation of MSK is accomplished by feeding the binary baseband data to the tuning terminal of a microwave voltage - controlled - oscillator (VCO). However, the frequency stability of the high power microwave oscillator is of the order of 0.05 percent. This poor stability is unacceptable as a stability of 1×10^{-6} or better is generally specified for the communication systems. A higher stability is achieved by injection-locking the oscillator to a frequency reference derived from a high stability crystal oscillator. A possible scheme for MSK generation using this technique is given in Fig. 2.1.4.

Fig. 2.1.4 Possible Scheme For Direct MSK Generation

2.2 Microstrip Transmission Lines

2.2.1 Description, Wave Propagation

Microstrips, transmission lines of metallic layers deposited on a dielectric substrate Fig. 2.2.1, are very useful for the implementation of the microwave and millimeter wave integrated circuits. For their fabrication either thin-film or thick-film technology (5) is used. Here the processing steps are substantially different compared to conventional coaxial and waveguide circuit technology. Circuits built with microstrip components have the following advantages (6), (7)

- Low cost of fabrication
- Excellent miniaturization, small volume and

Fig. 2.2.1 Microstrip Transmission Line Geometry

Fig. 2.2.2 A Pair of Coupled Lines

- weight
- Large bandwidth
- Direct bonding of active and passive devices on the dielectric substrate
- Devices and components are accessible for probing and measurement.

Assuming the quasi-TEM mode of propagation, the phase velocity in microstrip is given by

$$V_p = c / \epsilon_{eff} \quad (2.5)$$

where

c is the velocity of light, and
 ϵ_{eff} is the effective dielectric constant of the substrate material given by

$$\epsilon_{eff} = \left(\frac{\epsilon_r + 1}{2} \right) + \left(\frac{\epsilon_r - 1}{2} \right) \left(1 + 10 \frac{H}{W} \right)^{-1/2} \quad (2.6)$$

where

ϵ_r is the relative dielectric constant of the substrate material

Also, the wavelength λ_g in microstrip line is given by

$$\lambda_g = V_p/f \quad (2.7)$$

where f is the frequency of interest.

The characteristic impedance Z_o is given by

$$Z_o = 1/V_{pc} \quad (2.8)$$

In terms of microstrip geometry different expressions have been reported (7), (8), (9), (10), (11). According to Wheeler (8), whose analysis is and the most accurate

$$Z_o = \frac{377}{\epsilon_r} \left(\frac{H}{W'} \right) \frac{1}{1 + 1.735 \epsilon_r^{-0.0724} \left(\frac{W'}{H} \right)^{-0.836}} \text{ ohms} \quad (2.9)$$

where

$$\begin{aligned} W' &= \text{effective conductor width} \\ &= W + t / (1 + \ln 2H/t) \end{aligned}$$

The design in this thesis will be based on curves of the characteristic impedance Z_o vs W/H obtained from the work of Wheeler (12), together with curves of the effective dielectric

constant. These curves are given in Appendix A.

2.2.2 Implementation of L, C in Microstrip Circuits

At the lower frequencies the distributed elements consume a great deal of space. Consequently lumped element inductors and capacitors are formed from series transmission line elements by the following approximate expressions, (13)

$$L = Z_0/c \quad \text{Henry/inch} \quad (2.10)$$

$$C = 1/cZ_0 \quad \text{Farad/inch} \quad (2.11)$$

where

c is the propagation velocity in the transmission medium

Z_0 is the characteristic impedance of the line

By substituting Z_0 according to (7), L and C can be expressed in terms of W/H , ϵ_r and ϵ_0 (14).

$$L = 60 \ln(8H/W + W/4H) / c \quad (2.12)$$

Also for the capacitor C an effective area A is defined, which has no fringing field and is given by

$$A = H^2/4 \left((4C/\epsilon_0 \epsilon_r H + B^2)^{1/2} - B \right)^2 \quad (2.13)$$

where

ϵ_r is the relative dielectric constant

ϵ_0 is the free-space permeability

$B = r + 1/r$ where

r is the aspect ratio of the capacitor plate b/a , and

b, a are the width and length of the capacitor plate.

2.2.3 Coupled Lines

The physical construction of a coupled pair of microstrip lines is shown in Fig. 2.2.2. The configuration is conventionally specified by the parameters W/H and S/H , together with ϵ_r . An analysis of this configuration, based on the assumption

of approximate TEM propagation and using Green's functions, is given in (15). This results in the calculation of even and odd mode impedances for various values of ϵ_r as a function of W/H and S/H . Because of the large variety of substrates used for the implementation of microstrip circuits involving coupled lines, a simple method is needed for the calculation of even and odd mode impedances for any value of ϵ_r .

An accurate and direct method for designing coupled microstrip lines for any substrate is given in (16), once data for any substrate are known. It turns out that the even and odd mode impedances for two substrate materials are approximately related by

$$Z_{oi1}/Z_{oi2} = (\epsilon_{\text{reff}2}/\epsilon_{\text{reff}1})^{1/2} \quad (2.14)$$

where $i = \begin{cases} o & \text{for odd mode impedance} \\ e & \text{for even mode impedance} \end{cases}$

ϵ_{reff} is the effective dielectric constant of the substrate given by Eq. 3.2.

According to this method the curves of Fig. 2.2.3 have been derived for $\epsilon_r=4.9$, from the curves for $\epsilon_r=9.0$, in order to

Fig. 2.2.3 Even and Odd Mode Impedance for $\epsilon_r = 4.9$

design the band-pass filter of the multiplier.

Another factor which is involved in the design of structures with coupled lines is the velocity of the even (V_e) and odd (V_o) mode propagation which differ significantly. This difference complicates the calculation of the length of the coupled lines. In this case, a medium value (17) of wave propagation is given by

$$V_m = V_e + 0.25(V_o - V_e) \quad (2.15)$$

2.2.4 Quarter Wave Transformer

The quarterwave transformer is used to provide a match between real loads, especially when broadband performance is needed. To match a real load Z_l to an input resistance Z_{in} the characteristic impedance Z_o of a line of length $l = \lambda/4$ is given by

$$Z_o = \sqrt{Z_{in} Z_l} \quad (2.16)$$

2.2.5 Microstrip Losses

Whereas microstrip lines have conductor losses of the same order as other lines of the same size, a practical line has rather high losses because of the dielectric (18). The topic of microstrip losses is well covered in the literature.

In this thesis, it is assumed that the microstrip transmission line is lossless. While this is not absolutely true, it is justified because at the frequency of interest, 1.4GHz, losses are negligible.

2.3 Varactor Diodes and Frequency Multiplication

2.3.1 Introduction

The varactor diode (v.d.) is a semiconductor device that is used as a variable reactance element. This reactance is provided by the junction capacitance which varies with the applied bias voltage. Therefore the v.d. is a nonlinear element that can produce three substantially different effects, (19).

- (i) Generation of harmonics
- (ii) Switching or modulation of microwave power
- (iii) Parametric amplification of two microwave signals,

and/or up-conversion

A special type of v.d., the step recovery diode (SRD), is usually preferred and employed in the area of harmonic generation especially for cases where a high order of frequency multiplication and circuit simplicity are desired, and where maximum power is not required. SRD frequency multipliers differ considerably from the common v.d. multipliers on the principal mechanism for generating harmonics.

2.3.2 Operation of the Step Recovery Diode (SRD)

Frequency multipliers that employ the common varactor diode can be classed as variable reactance multipliers (VRM), since it is the nonlinear reverse bias capacitance variation of the varactor which is employed as the principal mechanism for generating the desired harmonics. By contrast, the SRD multiplier can be viewed as a switching reactance multiplier (SRM) since its operation is based on the SRD having two discrete states of reactance and on its ability to switch rapidly between these two states. The two states of reactance correspond to high capacitance when the diode

is forward biased and low capacitance when it is reverse biased. Equivalent circuits in both forward and reverse bias are shown in Fig. 2.3.1, together with their simplified form.

During the forward-bias a capacitance is produced by the injection of minority carriers. When the applied voltage switches from forward to reverse, large reverse current flows from the junction for a short period, which is supplied by the carriers in the junction. After the stored injected charge is exhausted, the current abruptly decreases (transition period). The abrupt collapse of the current, in the presence of a drive inductor before the diode will result to a large voltage spike of width t_p rich in harmonic content.

2.3.3 Diode - Multiplier Interaction

A good understanding of how the multiplier performance is influenced by the device parameters is needed in order to be able to select the proper diode. Selection is mainly based on the following parameters, (20)

(a) $C_j(v) \rightarrow \infty$

(b)

Fig. 2.3.1 Equivalent Circuit of Packaged Varactor Diode
 (a) During Forward Bias
 (b) During Reverse Bias

Transition time - T_s

Known also as snap time, the transition time is the time that the diode takes to switch from a conduction to a nonconduction state. This should be less than the reciprocal of the output frequency

$$T_s < 1/f_{out} \quad (2.17)$$

Minority carrier lifetime - T_l

This is the time required for the stored charge to be recovered. It must be long enough so the RF current through the diode reaches a negative peak before it snaps. For optimum performance

$$T_l = 25/f_{in} \quad (2.18)$$

Package Parasitics - L_p, C_p

The diode package parasitics should be small enough so that the series and parallel resonances will be well above the maximum operating frequency.

Thermal resistance - T_r

This must be small enough to allow the diode to remain within the maximum allowable operating temperature. It must be commensurate with the power to be dissipated, i.e.,

$$T_r = (T_{dm} - T_c) / P_d \quad (2.19)$$

T_{dm} : maximum allowable diode temperature ($^{\circ}\text{C}$)

T_c : heat sink temperature

P_d : the RF power dissipated in the diode.

2.3.4 Frequency Multiplier Design

A general analysis of v.d. frequency multipliers is given in (21). The main assumptions made in this analysis are that the diode is lossy and that currents in the diode flow only at the input, output, and idler frequencies, since a suitable external

circuit (idler circuit) prevents other currents from flowing. Special frequency multipliers employing SRD v.d. do not require idler circuits.

Also, a computer program is described in (21) which evaluates the basic design parameters of the multiplier for different drives. Here

$$\text{Drive} = (Q_{\max} - Q_{\min}) / (Q_B - q_0) \quad (2.20)$$

where $Q_{\max}$, $Q_{\min}$ are the max. and min. of the charge on the capacitance

Q_B is the charge at breakdown

q_0 is the charge at a contract potential

Results given for multipliers up to x10 show that the highest efficiency can be obtained for a drive equal to 2.0. Also from the tables given in (21) with different values of a , A , B etc. several parameters can be calculated for the multiplier design

efficiency $e = \exp(-a \omega_{\text{out}} / \omega_c)$

input res. $R_{\text{in}} = A / \omega_{\text{in}} C j(-\frac{1}{2}v)$

output res. $R_1 = E / \omega_{in} C_j (-6v)$

where ω_{in} and ω_{out} are the input and output frequencies at the multiplier and ω_c the cutoff frequency of the diode. Although this analysis is general, the analysis presented in (22) may provide better insight to the frequency multiplication process.

The basic sections of a frequency multiplier are shown in Fig. 2.3.2 In Burckhard's analysis the input section of the multiplier is a transforming network that assures the maximum transfer of energy to the diode. Following the analysis given in (22) this part is broken into two parts, the impulse generator and the matching section. Actually the final result is almost the same. The output section consists of a quarter-wave transformer in order to match the output resistance of the v.d. to 50 ohms, and a band-pass filter for the selection of the desired output frequency. The first section of the band-pass filter also plays the role of the resonator. In addition the multiplier includes a bias section consisting of a high impedance line and a filter capacitor.

Impulse Section

The most crucial part of the multiplier is the impulse

Fig. 2.3.2 Basic Parts of a Frequency Multiplier

section. Its primary section converts the input sinusoid into a narrow impulse with harmonic content. In this section, the diode shunts the line and switches the current through a series inductor. The voltage transient across the inductor then forms the impulse shown in Fig. 2.3.3d.

The charge-controlled switching action of the v.d. in impulse formation can be explained with reference to Fig. 2.3.3a. During the positive half-cycle of the input drive, the diode is driven heavily into forward conduction ($C_{fwd} \rightarrow \infty$). This results in the storage of charge in the depletion region by minority carriers. The stored charge, Q_s , is proportional to the area A under the diode current waveform Fig. 2.3.3c. Also when the minority carrier lifetime, T_l , is much greater than the input period ($T_l \gg 10/f_{in}$), very little stored charge is lost by recombination during the half cycle. This charge is extracted during the reverse half-cycle of the driving voltage ($C_{rv} = C(-6v)$) resulting in large current. When area B equals area A in Fig. 2.3.3c, the stored charge is depleted and diode conduction ceases very abruptly. Then the inductor L, will try to maintain the collapsing current, giving rise to a large voltage spike of width T_p which will be repeated every $T = 1/f_{in}$. This can be seen as a half-cycle resonance of L with C_{rv} at four while the diode is open. Then the impulse width T_p can be

Fig. 2.3.3 Frequency Multiplication Process

approximated by

$$T_p = \pi \sqrt{LCr_v} \quad (2.21)$$

This "train" of pulses in the time domain is responsible for the frequency multiplication resulting from a "train" of pulses in the frequency domain. By adjusting the width of the pulse T_p the frequency content of the signal is adjusted to the desired harmonic content.

It has been found that the optimum impulse width lies in the range

$$1/2f_{out} < T_p(\text{optimum}) < 1/f_{out} \quad (2.22)$$

With the impulse width taken as the mean of these extreme values the other two component of the impulse section can be calculated.

$$L = (T_p/\pi)^2 \times 1/Cr_v \quad (2.23)$$

$$C_t = Cr_v / (2f_{in}T_p)^2 \quad (2.24)$$

Matching Section

This section must perform three functions:

- (i) Good impedance match between the generator (200 MHz) and the impulse section for maximum power transfer
- (ii) D.C. isolation between the source and diode
- (iii) High frequency isolation

These functions are accomplished by an impedance transforming filter. Since bandwidth requirements are modest and loss is important, a simple L-section will meet most requirements while keeping insertion loss low. Element values are given by

$$L_m = 1/\omega_{in} \times \sqrt{R_i (R_g - R_i)} \quad (2.25)$$

$$C_m = 1/\omega_{in} R_g \times \sqrt{(R_g - R_i) / R_i} \quad (2.26)$$

where

$$R_i = \omega_i L = 2\pi f_{in} L \quad (2.27)$$

3. CHAPTER THREE

INJECTION LOCKING PHENOMENA

3.1 Introduction

An oscillator is usually considered to be a device which converts DC input power to AC currents and voltages and which is capable of delivering AC power to some load (23). The objective of this chapter is, to present a better insight to the injection locking aspects of microwave oscillators. This will be a very useful tool in our attempt to derive an analysis of the breaking-lock process and therefore an estimation of the breaking lock time.

Injection locking refers to a phenomenon in which a relatively powerful oscillator may be frequency synchronized and phase locked to a relatively weak input signal. This input may

gain access to the oscillator through the oscillator output port using a circulator. In general, upon application of the weak input signal, the oscillator frequency is pulled towards the signal frequency by an amount dependent upon the original frequency separation. A beat frequency is also observed in this instance. If the original frequency separation is made small enough the oscillator will become frequency locked to the input signal and will therefore oscillate at the input signal frequency. Taking advantage of this phenomenon, injection locking has been widely used for frequency stabilization, amplification (24), amplitude limiting (25) and detection of FM and PM signals (26).

Also, it is well known that injection locking is a phenomenon associated with nonlinearities in the I-V characteristics of oscillators and is not observed in an ideal linear oscillator which, for example exhibits no frequency pulling. Historically the phenomenon of injection locking has been well known for many years, being described by terms such as "synchronization of oscillators" and "entrainment of frequency". The synchronization of two clocks placed close to each other, is an early example of this, and was observed by Huygens (1629-1695), (27). But the pioneering work was performed by Van der Pol, in the 1920s, describing many of the important characteristics of synchroniza-

tion through a qualitative analysis for the triode oscillator, (28), (29). A simple theory applicable to many types of oscillators, in contrast to the special case treated by Van der Pol, was developed by Adler in 1946, (30). Almost immediately after the publication of Adler's now classic paper, Hunton and Weiss (31) demonstrated that any oscillator, whose frequency changes with its load impedance, would obey Adler's theory. Since we are interested in microwave frequencies, the analysis of Slater (1950) is particularly interesting, (23), as also that of Kurokawa, (32). Their difference lies in the selection of the reference plane.

3.2 Analysis of Locking Behaviour

In the absence of locking and for the simplified equivalent circuit of the oscillator, Fig.3.2.1a, the condition of free-running is given by

$$Z(\omega_0) + Z_{\text{act.dev.}}(\Lambda_0, \omega_0) = 0 \quad (3.1)$$

(a)

(b)

(c)

Fig. 3.2.1 (a) Oscillator Model
(b) Oscillating Condition
(c) Saturation Factors' Representation

where A_0 , ω_0 are the free-running current flowing through the active device and the oscillating frequency. This leads to

$$Z(\omega_0) = Z(A_0) \quad (3.2)$$

because the variation of the active device impedance is a slow varying function of frequency. The real meaning of the above equations is appreciated by drawing on the complex plane $Z(\omega)$, the impedance locus, and $Z(A)$, the active device line, Fig. 3.2.1b. Eq.(3.2) can be written in the following form of a system of two equations governing the steady-state free-running oscillation.

$$R(\omega_0) - \bar{R}(A_0) = 0 \quad (3.3)$$

$$X(\omega_0) + \bar{X}(A_0) = 0 \quad (3.4)$$

When the magnitude of the RF current $i(t)$ flowing through the active device

$$i(t) = A \cos(\omega t + \phi) \quad (3.5)$$

deviates somehow from the steady-state value A_0 , by a small amount δA , Eqs. (3.3) and (3.4) become

$$R(\omega_0) - \bar{R} = (\delta A/A_0) s R_0 \quad (3.6)$$

$$X(\omega_0) + \bar{X} = (\delta A/A_0) r R_0 \quad (3.7)$$

where s, r are called the saturation factors of the device resistance and reactance and their meaning is explained in Fig. 3.2.1c where R_0 and X_0 are the free running values for the active device.

A plot of constant power and constant frequency contours on a Smith chart of the load impedance is called a Rieke diagram (33) and shows how the oscillator output frequency and power is forced to change when the load impedance changes. The frequency and power pulling of the oscillator depending on the load variation is also very useful in an attempt to gain some natural insight of the injection locking. Also, from the Rieke diagram the frequency and amplitude sensitivity of the oscillator to small changes of load can be obtained by the gradients of the constant frequency and amplitude contours.

In the presence of an injection-locking signal, Fig. 3.2.2a

$$e(t) = A_0 \cos \omega_1 t \quad (3.8)$$

Eq. (3.1) has to be replaced by

(a)

(b)

Fig. 3.2.2 (a) Injection Locking Experiment Through a Circulator
(b) Power Variation for $S < 2$ and $S > 2$

$$Z(\omega_1) + Z_{\text{act.dev.}}(A, \omega_1) = (a_0/A) e^{-j\varphi} \quad (3.9)$$

or

$$Z(A) = Z(\omega_1) + (a_0/A) e^{-j\varphi} \quad (3.10)$$

where φ is the phase between $i(t)$ and $e(t)$. Eq. (3.10) is depicted in Fig. 3.2.3. Therefore the injected signal coming into the oscillator can be interpreted from the active device as a reflection from its output or as a load variation. This forces it to change frequency and power in order to again fulfill the oscillation conditions, provided of course that the external frequency is close enough to that of self operation (23).

The range of frequencies around the free-running frequency for which locking can occur is called the locking range. Also, the phase difference is a function of time and finally settles down to a constant value which depends on the initial difference between free-running and injection frequency, Eq. (3.22).

A mathematical expression describing the process of locking can be attained with a quasi-static analysis where A and φ are assumed constant in time and then from Fig. 3.2.3, (23), the locking range $D\omega_m$ can be evaluated.

$$D\omega_m = (\omega_0/Q_{\text{ext}}) \sqrt{P_i/P_o} \cdot 1/\cos\theta \quad (3.11)$$

Fig. 3.2.3 Injection Locking Impedance Diagram

where Q is not the loaded but the external Q of the resonant cavity

$$Q_{\text{ext}} = \omega_0 L / R \quad (3.12)$$

However, clear insight into the locking transient can be achieved only with a dynamic analysis where the differential equation for the RF current amplitude A and phase ϕ will be derived. The most general and complete derivation of these equations is given in (32). Since ω_i may differ from ω_0 , and A from A_0 , we in place of Eqs. (3.6) and (3.7)

$$R(\omega_i) - \bar{R} = (DA/A_0) sR_0 + DR \quad (3.13)$$

$$X(\omega_i) - \bar{X} = (DA/A_0) rR_0 + DX \quad (3.14)$$

where

$$DR = R(\omega_i) - R(\omega_0) \quad (3.15)$$

$$DX = X(\omega_i) - X(\omega_0) \quad (3.16)$$

Finally substituting in (3.9), (3.10), and as outlined in Appendix B1

$$\begin{aligned} & \left[(DA/A_0) sR_0 + DR \right] X'(\omega_i) - \left[(DA/A_0) rR_0 + DX \right] R'(\omega_i) + \left| \bar{Z}'(\omega_i) \right|^2 \frac{1}{A_0} \frac{dDA}{dt} \\ & = \frac{a_0}{A_0} \left\{ X'(\omega_i) \cos \phi + R'(\omega_i) \sin \phi \right\} \end{aligned} \quad (3.17)$$

$$\begin{aligned} & \left[(DA/A_0) S R_0 + DR \right] R'(\omega_i) + \left[(DA/A_0) r R_0 + DX \right] X'(\omega_i) + \left| \dot{\epsilon}'(\omega_i) \right|^2 \frac{d\varphi}{dt} \\ & = \frac{QC}{A_0} \left\{ R'(\omega_i) \cos \varphi - X'(\omega_i) \sin \varphi \right\} \end{aligned} \quad (3.18)$$

When the oscillator is locked $dDA/dt = 0$, $d\varphi/dt = 0$ and from Eqs. (3.17), (3.18) we have

$$a_0/A_0 (s^2 + r^2)^{1/2} \sin(\varphi_0 + \theta) = rDR - sDX \quad (3.19)$$

where φ is the steady-state phase difference and θ is given by

$$\tan \theta = -r/s \quad (3.20)$$

A more simplified version of the above general analysis had been derived by Adler, where the assumptions that A is expected to stay near A_0 , and small signal injection were made. This leads to

$$d\varphi/dt = -(P_i/P_0)^{1/2} (P_0/Q_{ext}) \sin \varphi + D\omega \quad (3.21)$$

Direct integration of Adler's equation and closed form solution for the angle results in the conclusion that regardless of the initial phase difference, φ settles down to φ_0 when $|D\omega/D\omega_m| < 1$, where

$$\sin \varphi_0 = -D\omega/D\omega_m \quad (3.22)$$

Otherwise, φ undergoes a periodic variation and does not converge

to a constant value.

It should be pointed out that in the locking process, the locking in frequency occurs immediately when an external signal is suddenly supplied, but the phase gradually adjusts itself (34). In the case where a continuous external signal approaches the frequency for which locking is possible, the oscillator's frequency will pull toward that of the external signal, and at the same time sidebands will build up, separated from the oscillator frequency by integral multiples of the frequency separation between the oscillator and injected signal frequencies. When final locking occurs, the sidebands will disappear and the oscillator will operate at the same frequency as that of the external signal. Varying the injected frequency across the locking range, the phase between current flowing through the active device and external signal will change but the frequencies will remain synchronized. A discussion on the spectra approaching lock-in is given in (35), (36), and some photographs will be given in section 3.3.

Since power is a more meaningful quantity at microwave frequencies, than signal amplitude, usually the above discussion is made in terms of power, (23). The injection vector is equal to $2R_o(P_i/P_o)^{1/2}$, Fig. 3.2.3 and consequently Eq. (3.19) can be

rewritten in the form

$$DR\sin\theta - DX\cos\theta = 2R_0(P_i/P_0)^{1/2}\sin(\varphi_0 + \theta) \quad (3.23)$$

The output power variation during the injection locking experiment, is also very interesting Fig. 3.2.2b, and is given by, (37)

$$DP/P_0 = 2\sqrt{P_i/P_0(2/S-1)}\cos\varphi_0 \quad (3.24)$$

Because of the terms in the brackets the power variation has an elliptical form, convex for $S < 2$ and concave for $S > 2$. Therefore the observation of the output power during the injection locking experiment gives a good indication of the magnitudes of s and θ and hence r from Eq. (3.20). By examining the power variation of an actual oscillator an asymmetry in the locking range is usually observed, which is explained by the non-linearity of the oscillator tuned circuit phase response, (38).

3.3 Characterization of the VCO

3.3.1 Voltage Tuning and Power Characteristics

The voltage controlled oscillator that was employed in the experiments throughout this thesis was the transistorized oscillator model 8090 from AVANTEK. The frequency tuning and output power characteristics of the oscillator with details around 1.4 GHz , where the MSK generation through locking will be carried out, are given in Fig. 3.3.1. The measurements have been taken with the HP 5342A Microwave Frequency Counter. From these results, the near linear tuning characteristic of the oscillator around the frequency of 1400 MHz can be observed. Also from the range of tuning voltages of the varactor employed on the tank circuit of the VCO, it can be concluded that the varactor is of

Fig. 3.3.1 Tuning Voltage Characteristic of the VCO

Fig. 3.3.2 Output Power Characteristic of the VCO

an abrupt junction and therefore $n=0.5$. The output power is not very constant as would be desirable, Fig. 3.3.2, but for analysis convenience it will be assumed constant.

3.3.2 Injection Locking Experiments - Q Evaluation

The external Q of the oscillator was calculated by performing injection locking experiments for different ratios of P_i/P_o , and for a carrier frequency of the VCO of 1400 MHz. By calculating after the term $2Df/f_o$, where Df is the locking range of the oscillator, the following Table 3.1 is derived

Table 3.1

Injection Locking Experiments Results

P_o/P_i (dB)	f (MHz)	$2Df/f_o$
7	47	0.067
10	28.5	0.040
15	15.5	0.022
20	8	0.011

By plotting the above results on logarithmic paper, Fig. 3.3.3, the value of the average Q_{ext} can be obtained as the interception of the y-axis. This is obvious from Eq.(3.11). Actually $1/Q_{ext}=1.8 \times 10^{-1}$, and $Q_{ext}=5.5$.

The output variation of the locked oscillator is given in Fig. 3.3.4 for $P_i/P_o=-10\text{dB}$. From the concave variation of output power, the saturation factor s should be expected to be larger than 2, (37). For the same experiments with increased P_i , the equivalent locking range increased and the spacing between adjacent spectral lines at the output spectrum decreased. This output spectra of injection locking, for different P_i/P_o ratios, is presented in Pict. 3.3.1a-d.

3.3.3 Rieke Diagram - Evaluation of Angle θ

The angle θ can be found from the Rieke diagram, Fig. 3.3.5, since the angle between the device line and the impedance locus is equal to the angle between the constant frequency and constant power contours of the oscillator, (33, pp.224). The set-up that was

Fig. 3.3.3. Graphical Evaluation of Q_{ext}

Fig. 3.3.4 Output Power Variation of the Injection Locked Oscillator

(a)

(b)

Pict. 3.3.1 Injection Locked Oscillator

(a) $P_o/P_i = 20$ dB

(b) $P_o/P_i = 10$ dB

Hor. 30 MHz/div., vert. 10 dB/div.

used to derive this diagram is given in Fig. 3.3.6. With the line stretcher and the phase shifter several different impedances were created at the output of the oscillator. For each of these the reciprocal output power and frequency of the oscillator were also measured. The intersection of the 1400 MHz contour and the 12 dBm contour give an approximate value of θ of 15° .

3.3.4 Evaluation of Saturation Factors s and r

The set-up of Fig. 3.3.7 was used to determine the saturation factor s. With the 50 ohm termination, the oscillating VCO circuit looks like Fig. 3.3.8a, and

$$I_{rms} = 2I_{bol} = 2 P/50 = 2 (11.2mw)/50 = 29.95mA$$

or

$$I_{opeak} = 42.35mA$$

For the second case with the 50 ohm termination and T junction removed, Fig. 3.3.8b, R_o is equal to 50 ohms and

$$I_{rms} = P/50 = 38.01mw/50 = 27.5mA$$

or

$$I_{opeak} = 38.89mA$$

Fig. 3.3.6 Set-Up for the Derivation of the Rieke Diagram

Fig. 3.3.7 Set-Up for Evaluation of the s -factor

Fig. 3.3.8 Equivalent Circuits for Different Loads

From these two values the plot of Fig.3.3.9 is made, and from the slope of the curve s has been calculated and found to be $s = 5.6$.

From the values of θ and s , r can be calculated from Eq.(3.20) and is found to be $r=1.5$.

3.4 Time Delays Involved with the Injection Locking Process

The different delay times involved with the injection locking technique is of primary concern when this technique is employed for the stabilization of an MSK signal, as is the case in this thesis. Actually in such a situation, the VCO carrier switches in frequency between two reference signals, being locked alternatively to each one of them.

Assuming that the time required for the VCO to switch between the two frequencies is almost negligible, there are still two delays introduced by the nature of the injection locking

Fig. 3.3.9 Graphical Evaluation of s

process. The one is the time required for the VCO carrier to lock to the injected signal, locking time (T_l), and the other is the time needed for the VCO carrier to break the lock. These locking and lock breaking times occur when a change of state has to take place and the VCO must switch and lock to the other reference signal.

3.4.1 Locking Time (T_l)

During the locking process, the locking in frequency occurs immediately and the phase gradually adjusts itself (32) according to Eq. (3.21), (23). The time required for this adjustment can be found by integration of Eq. (3.21) from the initial value of phase ϕ_0 to the final value given by Eq. (3.22).

For small signal injection, $P_i/P_o \ll 1$, and for an initial frequency difference of $D\omega_0 = \omega_i - \omega_0$ between the two signals close to zero, the transient response of the phase angle is given by

$$\phi = 2 \tan^{-1}(e^{-\eta t} \tan \phi_n / 2) \quad (3.25)$$

where

$$D_o = \omega_o / Q_{ext} P_o / P_i \quad (3.26)$$

The phase angle variation is plotted in Fig. 3.4.1 and it is seen that, for low values of initial phase φ_{in} , the transient phase angle approaches zero exponentially, or

$$\varphi(t) = \varphi_{in} e^{-D_o t} \quad (3.27)$$

To be consistent with the conventional definition of "time constant" for transient circuits, locking time

$$T_l = 1/D_o \quad (3.28)$$

is defined as the time required for the phase angle to reduce to $1/e$ of the initial phase difference. At this moment the two signals are said to be phase locked. In Table 3.2, the locking time T_l is given for $Q_{ext}=5.5$, $f_o=1400$ MHz and different ratios of P_o/P_i .

Table 3.2

Locking Time

P_o/P_i (dB)

T_l (sec)

20

6.25×10^{-11}

30	1.93×10^{-11}
40	6.25×10^{-12}
50	1.97×10^{-12}

3.4.2 Breaking-Lock Time (T_{bl})

An analysis for the calculation of the breaking-lock time will be outlined in this next section for the special case of our oscillator, but the procedure can also be applied to any microwave oscillator. By investigating Fig. 3.2.3, we can see that after the transients, locking has been established with a constant ϕ_0 and A₁ new operating point of the active device. When an external cause, which in our case is the tuning voltage of the oscillator, forces the VCO to change oscillating condition the locking signal tries to keep track until the moment that the angle ϕ takes the value of ϕ_{cr} and the lock breaks, Fig. 3.2.3

$$\phi_{cr} = 90^\circ + \theta \quad (3.29)$$

Therefore, an expression for the change $\delta\phi$ of the angle ϕ as a

Fig. 3.4.1 Variation of $\phi(t)$ during the Locking Process

function of time integrated from φ_0 until $90+\theta$ could give us the time required for the locking to break. An expression for $\delta\varphi$, can be derived from Eqs. (3.17), (3.18) after introducing the quantities

$$DA = DA_0 + \delta A \quad (3.30)$$

$$\varphi = \varphi_0 + \delta\varphi \quad (3.31)$$

where δA is a small amount deviation from the steady-state value during locking. Elimination of δA between the two equations yields Eq. (3.32).

$$\left| Z'(\omega_i) \right|^2 \frac{d^2 \delta\varphi}{dt^2} + \left\{ sR_0 X'(\omega_i) - rR_0 R'(\omega_i) \right\} \times \frac{\alpha_0}{A_0} \left[R'(\omega_i) \sin \varphi_0 + X'(\omega_i) \cos \varphi_0 \right] \frac{d\delta\varphi}{dt} + \frac{\alpha_0}{A_0} R_0 (S^2 + r^2)^{1/2} \cos(\varphi_0 + \theta) \delta\varphi = 0 \quad (3.32)$$

where $Z(\omega)$ is the impedance seen from the active device terminals and the primes indicate differentiation with respect to ω .

In order to take into account the tuning voltage V_t of the VCO, the model shown in Fig. 3.4.2 of a practical single-tuned oscillator is employed. The change in V_t will produce a change in the varactor diode parameters, shown in Fig. 3.4.3a. The simplification of the varactor diode model is justified because R_j , the junction leakage resistance, is in the range Megohms, the package capacitance C_p has minimal effects and the package indus-

Fig. 3.4.2 VCO Model

Fig. 3.4.3 (a) Varactor Diode Model
(b) Simplified Varactor Diode Model

tance L_p can be considered as part of the tank circuit inductance. The simplified varactor diode model is shown in Fig. 3.4.3b, where R_s accounts for the total losses and depends on the tuning voltage V_t .

$$C_j = C(0) / (1 + V_t / \phi)^n \quad (3.33)$$

where

$C(0)$ is the junction capacitance in the free-running operation of the VCO

$$\phi \approx 0.7 \text{ V}$$

$n = 1/2$ for abrupt junction

$= 1/3$ for hyper-abrupt junction

Thus, after incorporating the previous modeling of oscillator and varactor diode Eq. (3.32) takes the final form of a second order nonlinear differential equation, Eq. (3.34).

$$\begin{aligned} \frac{d^2 \delta \psi}{dt^2} - \left[\frac{(\omega_i / \omega)^2 + 1}{\omega_i C_j(V_t)} \right] \{ S R_o + G \cos \psi_0 \} \frac{d \delta \psi}{dt} \\ - \left[\frac{(\omega_i / \omega)^2 + 1}{\omega_i C_j(V_t)} \right] G \sqrt{(S R_o)^2 + (r R_o)^2} \cos(\psi_0 + \theta) \delta \psi = 0 \end{aligned} \quad (3.34)$$

where

ω_i = frequency of injected signal

$C_j(V)$ = junction capacitance of the varactor

G = current gain = $a_o/A_o \sim P_i/P_o$

s, r = saturation factors for real and imaginary part of the active device

ϕ_o = angle after locking transients have settle down

θ = angle between active device line and horizontal

* The step by step derivation of Eq. (3.34) is given in Appendix B2. A numerical solution of Eq. (3.34) will be outlined in the next section.

3.4.3 Numerical Evaluation of Breaking Lock Time

By solving numerically the differential Eq. (3.34) a fairly good idea about the time required for the angle to reach the critical value of $90^\circ + \theta$ can be obtained. A computer program has been developed, Appendix B3 and used for the evaluation of $\delta\phi$ as a function of time. For a change of the tuning voltage V_i in the form of the step function, the junction capacitance of the varac-

tor as well as the other parameters of Eq. (3.34) are computed and then the differential equation is solved using the Runge-Kutta procedure for successive time instants.

The values of the different parameters of Eq. (3.34) has been found experimentally in the previous section 3.3. The only parameter of the VCO that is unknown, is the junction capacitance at free-running $C(0)$ of the varactor diode employed in the tank circuit of the VCO model, Fig. 3.4.2. However, for this type of VCO, a $C(0)$ in the range of few picofarads should be expected.

Assuming an initial value for ϕ_0 of 20° and $P_i/P_c = -20\text{dB}$, the results shown in Fig. 3.4.4 have been obtained for values of $C(0)$ of 0.8, 1.6 and 3.2 pF. These results are summarized in Table. 3.3.

Table 3.3

Breaking Lock Time

$C(0)$ pf	T_{bl} nsec	slope $^\circ/\text{nsec}$
0.8	15.75	6.6
1.6	8	11.4
3.2	4	26.6

Fig. 3.4.4 $\delta\phi(t)$ Variation for Different Values of $C(O)$ During the Breaking Lock Process

From this table it can be concluded that the breaking lock time T_{bl} is inversally proportional to $C(0)$. Evaluation of T_{bl} for different ratios of P_i/P_o close to -20dB did not show any significant change, remaining in the same range of 10 nsec.

3.4.4 Experimental Evaluation of Breaking Lock Time

The breaking lock time was obtained by measuring the differential settling times from an initial frequency flow to a frequency f_{HIGH} , once with an injected signal of frequency f_{LOW} present and once without. The difference between the two settling times is due to the supplementary time needed for the VCO to break the lock from the injected signal.

The set-up used for the measurement of both settling times is shown in Fig. 3.4.5, and is based either on observation of the output spectrum on the spectrum analyzer or either of the demodulated output pulse on the oscilloscope. The method of the spectrum analyzer is suggested in (39) and requires first calibration

Fig. 3.4.5 Set-Up for Measurement of Breaking-Lock Time

of the spectrum analyzer with the VCO tuned from a pulse generator of a variable risetime. The relative amplitude of the displayed output on the CRT is a direct indication of the speed with which the VCO signal is slewed through the IF filter of the spectrum analyzer. The output spectrum of our VCO is shown in Pict. 3.4.1a, without an injected signal at f_{LOW} and in Pict. 3.4.1b with an injected signal of frequency f_{LOW} , and for $P_i/P_o = -20$ dB. Even though there is an obvious difference in the output spectra, the spectrum analyzer was not sensitive enough to obtain the difference in settling times.

By using the second technique of down-conversion and demodulation of the output signal, Pict. 3.4.2a-b, the difference between the two settling times was estimated to be about 10% by comparing the equivalent demodulated pulses. Also from Pict. 3.4.2a the settling time is about 90nsec and therefore the breaking-lock time is about 9nsec. The same technique has been tried for different ratios of P_i/P_o , but the difference was very small and it could not be estimated.

Pict. 3.4.1 a. VCO Switching in Frequency
Between F_{Low} and F_{High}
b. VCO Switching in the Presence
of an Injected Signal of Frequency
 F_{Low}

(a)

(b)

Pict. 3.4.2 a. Demodulated Pulse
b. Demodulated Pulse in the Presence of the Injected Signal

4. CHAPTER FOUR

SYSTEM IMPLEMENTATION

4.1 Introduction

The set-up used to implement MSK at microwave frequencies using a locked oscillator is shown in Fig.4.1.1. The data to be transmitted is fed to the tuning port of the VCO (V_t) through a matching circuit in order to retain the best possible pulse shape of the data stream on its transition from the 50 ohms source, HP Data Generator, to the high impedance input port of the VCO. The two voltage levels of the data stream V_h and V_l correspond to frequencies $f_1 = (1400 - 1) \text{ MHz} = 1399 \text{ MHz}$ and $f_2 = (1400 + 1) \text{ MHz} = 1401 \text{ MHz}$. By keeping also the ratio of $(f_2 - f_1)/f_b = 0.5$, where f_b is the bit rate of the data, the output signal of the VCO which switches between f_1 and f_2 is already MSK modulated. By injecting into the

Fig. 4.1.1 4Mbits/sec Microwave System

VCO, two signals of crystal stability, one at f_1 and the other at f_2 , locking of the instantaneous frequency of the oscillator is achieved. Therefore the stability of the output signal is improved, reducing the inherent "jitter" noise of the transmitter.

These two reference signals, 2 MHz apart and surrounding 1.4 GHz have been derived from a 100 MHz crystal oscillator. As Fig. 4.1.1 shows the signal from the crystal controlled oscillator is split in two paths, one is divided by a hundred giving a signal of 1 MHz, and the other is frequency multiplied to 1.4 GHz. Finally by using a double sideband suppressed carrier mixture of the two signals the two spectral lines that will be used as locking signals has been derived.

4.2 MSK Modulator

4.2.1 Crystal Oscillator

The reference injected signals at 1399 MHz and 1401 MHz were derived from a 100 MHz crystal controlled source made by ZEIA laboratories. The total frequency deviation of the oscillator was less than 0.003% and its output power was 12.6 dBm.

4.2.2 Frequency Divider

The SP8629 integrated circuit of Plessey Semiconductors was used to divide the frequency of the crystal oscillator ($f_{in}=100$ MHz). This chip is a fixed ratio ECL divider with a minimum guaranteed toggle frequency of 150 MHz over a $+30^{\circ}\text{C}$ to $+70^{\circ}\text{C}$ temperature range. The device was operated in the single-ended mode, capacitively coupled to the signal source as is shown in Fig. 4.2.1 and produced at its output a square wave of frequency $f_{out}=f_{in}/100$.

A low-pass three-pole Butterworth filter, with a 3-dB cutoff frequency at 3 MHz, was used to smooth the output waveform and finally obtain a 1 MHz sinewave.

4.2.3 Double Sideband-Suppressed Carrier Modulator

The set-up of Fig. 4.2.2 was used in order to derive the two signals at 1399 MHz and 1401 MHz which are 2 MHz apart and surround the center frequency of 1.4 GHz.

The technique of double-sideband suppressed carrier (DSB-SC)

Fig. 4.2.1 $\div 100$ Divider

Fig. 4.2.2 DSB- SC Technique

Pict. 4.2.1 Generation of Reference Signals at $f = 1.4 \text{ GHz} \pm 1 \text{ MHz}$
(Hor. 1 MHz/div., ver. 5 dB6div.)

was implemented in order to suppress the carrier, 1.4 GHz signal, and leave only the lower and upper sidebands 1.4 GHz $\pm$ 1 MHz. Suppression was achieved by adding to the product of the two initial signals the carrier after some phase shifting required to cancel the carrier feedthrough from the DSB-SC modulator. A double-stub was properly tuned in order to bring the two signals into antiphase.

The output of this set-up in the frequency domain is shown in Pict. 4.2.1 where the suppression of the carrier by more than 35 dB is shown. Some undesirable sidebands are also shown created at the first mixing of the 1 MHz and 1.4 GHz signals, but since they are 20 dB down the overall result is still quite good.

4.2.4 Calculation of Power Level, P_o/P_i

From Section 3.3, Fig. 3.3.3 or directly from the equation

$$Q\omega = \omega_o/2Q\sqrt{P_i/P_o} \quad (4.1)$$

and for $Q_{ext} = 5.5$ and $D\omega = 2\pi \times 1 \text{ MHz}$ the ratio P_o/P_i between the power of the VCO carrier and the locking signals must be adjusted to 42 dB.

4.3 Frequency Multiplier Design and Implementation

4.3.1 Introduction

The frequency multiplier needed for the derivation of a 1.4 GHz signal starting with 200 MHz (coming from the 100 MHz crystal oscillator after passing through an ANZAC doubler) is of seventh order. Actually the design of the input section has been carried out for a x8 multiplier and after providing a 1.4 GHz band-pass filter structure at the output, the seventh harmonic was selected. The power handling and bandwidth of the multiplier was not of primary interest but the efficiency of the multiplier was

of concern and it has been optimized by using the COMPACT.

4.3.2 Description of the Device - Microstrip Data

Since high efficiency is desirable a diode with a high cutoff frequency was chosen from Microwaves Associates, model MA 44130. The diode characteristics are as follows:

Breakdown voltage = V_{br} = 47.50 volts

Thermal resistance = R_t = 13.3 C/Watt

Cutoff frequency = f_c = 188.2 GHz

Lifetime = T_l = 61 sec (=12.2 /fin)

Transition (snap) time = T_s = .10 psec ($< 1/f_{out}$ = 625 sec)

Junction capacitance = C_j = 1.934 pf at -6 volts

Package capacitance = C_p = 18 pf

Package inductance = L_p = .4nH

Another useful diode parameter is the diode series resistance

$$R_s = 1/2\pi f_c r_c = 0.40 \text{ Ohms}$$

where

$$C = C_j + C_p = 2.113 \text{ pf}$$

The substrate used for the implementation of the microstrip circuits was of the epoxy-glass type. Its relative dielectric constant ϵ_r has been measured and found to be about 4.9. The two conducting sides are copper, ideally with zero thickness, and the substrate thickness H is 60 mils.

4.3.3 Multiplier Design

a. Input Section

The input section was designed and implemented by following both the techniques described in Section 2.3. According to Burckhardt's analysis (21) for $\gamma \approx 0$ (abrupt-function) and drive $= 2.0$ (diode overdriven) the basic circuit parameters can be evaluated, from Table 4.1. The different symbols have been already explained in section 2.3.

Table 4.1

1-8 Multiplier Without Idlers

(γ = 0, Drive = 2.0)

a	21.7
β	0.0034
P _{max} /P _{norm}	1.3 X 10 ⁻⁵
W _{omax} /W _c	3.0 X 10 ⁻²
A	0.0098
E	0.0106
S _{o1} /S _{max}	0.50
S _{oout} /S _{max}	0.50
V _{onorm}	0.29

from the above Table 4.1 and for the varactor previously described, the efficiency η , input impedance R_{in} and output impedance R_{out} may be calculated. Also it is estimated that the effect of the series resistance variation is to multiply a by a factor of 1.4 (40), giving an effective $a = 30.38$ and

$$\eta = 0.77$$

$$R_{in} = 4.03 \text{ ohms}$$

$$R_{out} = 4.36 \text{ ohms}$$

Using tables given by (41), the element values for the low-pass filter form impedance transforming network were determined for a transformation ratio r of 12. The normalized parameters for a Tchebyscheff filter with a 50 ohms source, 0.2 ripple and a transformation ratio of 10, which is the closest tabulation to 12 are

$$\begin{aligned}g_0 &= 1 \\g_1 &= 1.81212 \\g_2 &= 0.521413 \\g_3 &= 5.08377 = g_2 r \\g_4 &= 0.185858 = g_1 / r \\g_5 &= 9.75\end{aligned}$$

The element values were derived for the configuration of Fig.4.3.1, (41) and

$$\begin{aligned}C1 &= 26.94 \text{ pf} \\L2 &= 22.27 \text{ nH} \\C3 &= 89.10 \text{ pf} \\L4 &= 6.73 \text{ nH}\end{aligned}$$

According to application note HP920, (as outlined in Section 2.3.4) for impulse width t_p of 500 psec, the elements of the

Fig. 4.3.1 Lumped Elements Equivalent for Input Section

Fig. 4.3.2 Distributed Form of the Optimized Input Section

impulse generator were found to be

$$L = 13.09 \text{ nH}$$

$$C_t = 48.35 \text{ pf}$$

consequently

$$R_i = \omega_m \times L = 16.33 \text{ Ohms}$$

and the values of the transforming elements were evaluated and found to be

$$L_m = 18.6 \text{ nH}$$

$$C_m = 22.85 \text{ pf}$$

Because of the similarity of the results of both design philosophies, no significant difference of the actual multiplier should be expected. Therefore we proceeded on the implementation based on the first set of results. Actually computer optimization for best matching (S_{11} close to zero) has been carried out, Appendix C, and the new optimized component values were found to be

$$C_1 = 25.6 \text{ pf}$$

$$L_2 = 18.4 \text{ nH}$$

$$C_3 = 71.9 \text{ pf}$$

$$L_4 = 6.57 \text{ nH}$$

The microstrip implementation was made according to Section 2.2 and Appendix A and the result is shown in Fig.4.3.2. It should be mentioned here that even the capacitances are usually implemented by low characteristic impedance sections. An effort was made to present these sections as pairs of open-circuited stubs in parallel in order to overcome problems that would appear otherwise in the computer optimization task.

Measurements of the input section with the diode mounted show very good matching for the frequencies of interest, Fig.4.3.3. Also in Pict.4.3.1 the comb spectrum is shown following the diode.

b. Output Section

The output section consists of a quarter-wavelength transformer followed by a three-section band-pass filter. The transformer having a characteristic impedance of 15.8 ohms transforms the real part of the varactor output impedance, 5 ohms, to 50 ohms. The band-pass filter selects the 1.4 GHz for the output of the overall circuit, its first section functioning as a resonator. The design of the band-pass filter is based on (42) and

Fig. 4.3.3 S11 for the Input Section of the Multiplier

Pict. 4.3.1 Comb Spectrum of the
Frequency Multiplier
(Hor, 200 MHz/div., ver. 5 dB/div.)

consists of half wave lines, edge coupled. The design parameters for a 0.1 dB ripple Tchebyscheff filter are given by

$$\begin{aligned} g_0 &= 1.0 & \text{fractional bandwidth} &= \omega = 0.1 \text{ (10\%)} \\ g_1 &= 1.0315 & \omega_1 &= 1.33 \text{ GHz} \\ g_2 &= 1.1474 & \omega_2 &= 1.47 \text{ GHz} \\ g_3 &= 1.0315 & \omega_3 &= 1.0 \\ g_4 &= 1.0 \end{aligned}$$

and the even and odd mode impedances are calculated:

$$\begin{aligned} \text{Section 1: } (Z_{oe}) &= 77.1 \text{ ohms} \\ (Z_{oo}) &= 38.1 \text{ ohms} \end{aligned}$$

$$\begin{aligned} \text{Section 2: } (Z_{oe}) &= 57.9 \text{ ohms} \\ (Z_{oo}) &= 44.5 \text{ ohms} \end{aligned}$$

$$\begin{aligned} \text{Section 3: } (Z_{oe}) &= 77.1 \text{ ohms} \\ (Z_{oo}) &= 38.1 \text{ ohms} \end{aligned}$$

The width and spacing of the lines can be found from Fig.2.2.3 which gives W/H and S/H for different odd and even impedances for microstrip coupled lines on a substrate of $\epsilon_r = 4.9$.

The filter is shown on the final circuit of the multiplier,

Fig. 4.3.4 X7 Frequency Multiplier

Fig. 4.3.4, and its output frequency response is given in Fig. 4.3.5. Measurements of the final circuit of the multiplier resulted in an efficiency of 45% for input power level around 0 dBm. However, it is believed that for higher power levels and for better input and output matching the multiplier efficiency could approach the theoretical predicted value of 77%.

Fig. 4.3.5 Frequency Response of the 1.4 GHz Band-pass Filter

4.4 Spectrum of Modulated Waveform

By adjusting the tuning voltage of the VCO with the two locking signals present, the carrier is locked to f_1 or to f_2 . Therefore for a "zero" a signal of frequency $f_1 = 1399,001.80$ MHz is transmitted and for an "one" a signal of frequency $f_2 = 1401,001.77$ MHz. The equivalent spectra for the locked carrier are shown in Pict. 4.4.1 a-b.

When the tuning port of the VCO is fed with a pseudorandom sequence of 4 Mbits/sec and having an amplitude such that the modulation index is kept to 0.5, the output of the VCO is MSK modulated and appears as shown in Pict. 4.4.2a-b. The same power spectra in the presence of the two locking signals is shown in Pict. 4.4.3. By increasing the environmental temperature of the VCO no drift has been observed at the locked output power spectra.

It should be noted that the above results have been obtained

(a)

(b)

Pict. 4.4.1 (a) VCO Carrier Locked at $f_1 = 1399.00203$ MHz
(b) VCO Carrier Locked at $f_2 = 1401.0028$ MHz
 $P_{out} = 14.5$ dBm (Hor. 1 MHz/div., ver. 5 dB/div.)

(a)

(b)

Pict. 4.4.2 4 Mbits/sec MSK Spectra at 1.4 GHz
 (a) Hor. 3 MHz/div., ver. 5 dB/div.
 (b) Hor. 1 MHz/div., ver. 5 dB/div.

Pict. 4.4.3 4 Mbps/sec MSK at 1.4 GHz With Injection Locking
(Hor. 1 MHz/div., ver. 5 dB/div.)

with a power ratio of $P_o/P_i=49\text{dB}$, and not with 42dB previously calculated. Actually what happens with a power ratio more than 49dB is that the signal at one reference frequency has enough power to unlock the locked carrier at the other reference frequency.

4.5 Detection of the MSK Waveform

In order to get some idea of the performance of the system, the set-up of Fig. 4.5.1 was used for the demodulation and regeneration of the digital data. From its nature the technique used was that of a noncoherent detection system, since the signal used for the down-conversion to IF came from an extra source and was not extracted from the received signal itself.

The main objectives of this section are to obtain the eye pattern for the baseband demodulated signal and also obtain measurements of probability of error (BER) after the regeneration of the data. This was repeated for different power levels of injected signals.

Fig. 4.5.1 Set-up For the Detection of 4 Mbits/sec MSK Transmitted Data

4.5.1 Description of the Set-Up

a. Demodulator (Limiter-Discriminator)

The system used for the demodulation of the MSK signal is shown in Fig.4.5.2 in detail. Because the baseband modulating signal varies only the frequency of the carrier, any variation of the amplitude would be due to noise alone. Therefore a limiter is used to suppress such amplitude-variation noise, resulting in a square-wave in its output. The bandpass filter following the limiter, selects the fundamental frequency component of the square-wave, converting the output back to a sinusoid but of constant amplitude.

The discriminator consists of two component parts. The first converts frequency to amplitude variations (differentiator), and the second the envelope detector, recovers the baseband signal from the amplitude variations.

Fig. 4.5.2 Demodulator (Limiter – Discriminator)

Fig. 4.5.3 Noise Spectral Density for FM System

The noise power N_o at the output of this system is equivalent to f_m , where f_m is the baseband bandwidth of the system (43). Fig. 4.5.3. Therefore filtering of the baseband output is an asset reducing the noise. A simple $n=3$ Butterworth low-pass filter with a 6 MHz 3-dB cutoff point was used at the output of the demodulator.

b. Regenerator

For the regeneration of the received data stream the AM685 fast voltage comparator was used. A latch enable input is provided on this chip in order to allow its use in a sample and hold mode. Whenever a clock pulse appears on this input the voltage comparator functions normally and locks to a logical state until the next clock pulse arrives.

4.5.2 Eye-Pattern

The eye-pattern was taken after the filtering of the base-

band signal and it is shown on Pict. 4.5.1. The opening of the eye is about 90% and the time-jittering at the crossing points almost insignificant. Some noise on the top and bottom of the eye remains even after the filtering. This can be due to intersymbol interference from the IF filter, which was not narrow enough, for this specific purpose of receiving a 4 Mbits/sec MSK waveform.

Pict. 4.5.1 Eye Pattern
(Hor. 0.1 μ sec/div., ver. 0.1 volt/div.)

4.5.3 Bit Error Rate

Bit error rate measurements taken for different power ratios P_o/P_i , Fig. 4.5.4, show a significant degradation on the probability of error for increasing the power of the injected signal. This is supported by the fact that by increasing P_i , the MSK spectrum is disturbed more and more by the presence of discrete spectral lines, Pict. 4.4.3. In other words intersymbol interference (ISI) is introduced from the injection locking process degrading the BER performance, (44). On the other hand by decreasing P_i , the locking range of the injected signals decreases and also the impact of this technique for frequency stability becomes less efficient. Therefore an optimization between frequency stability and BER performance should be taken into account whenever the technique of injection locking is applied for direct microwave generation.

Fig. 4.5.4 Bit - Error Rate Degradation

5. CHAPTER FIVE

CONCLUSIONS AND RECOMMENDATIONS

5.1 Summary and Conclusions

In this thesis we have proposed and studied, theoretically as well as experimentally, a possible scheme for direct microwave MSK generation of crystal stability, by using injection locking techniques. A summary of the results is described below.

Theoretical analysis concerning the time delays introduced from the injection locking process showed that the breaking lock time and locking time are small and are not expected to create any problems during the MSK generation for moderate data rates. The theoretical and experimental results were derived for a specific VCO which was employed in the experiments throughout this thesis. However the same main steps could be applied for any

VCO which could be employed in a similar microwave system. The results showed that the time needed for the VCO carrier to lock is in the range of few picoseconds. Especially for the breaking lock time, it can be concluded that this is strongly correlated to the VCO structure. For the VCO model employed in our analysis, it seems that the breaking lock time is inversely proportional to the capacitance of the tuning varactor, and was found to be within a few nanoseconds. Therefore the sum of the two time delays is much less than the bit period of 250 usec, which was used by the experimental system.

The feasibility of the proposed scheme was experimentally investigated through the development of such a system operating at a rate of 4 Mbits/sec. The spectrum of the generated waveform was found to be slightly different from the ideal MSK spectrum. The difference was due to the appearance of discrete spectral lines caused by the injection locking process, and appearing at ± 3 MHz points from the center frequency. However, the main lobe and out-of-band energy distribution of the MSK modulation scheme was not affected by the appearance of these discrete spectral lines. When the proposed scheme is used in a frequency arrangement, filtering of the generated waveform is required in order to avoid adjacent channel interference that will result from the presence of these discrete lines (45). By

observing also the output spectrum it was concluded that crystal stability was achieved by using injection locking of the VCO carrier. Finally, demodulation of the transmitted data and bit error rate measurements showed the amount of degradation experienced in the bit error rate by increasing the power level P_i of the injected signal. However for proper adjustment of P_o/P_i at 49dB the bit error rate of the system was constant at 3.6×10^{-6} .

5.2 Recommendations for Further Work

Further work should be carried out on the theoretical analysis aspect of the breaking lock time of a high power VCO. The locking time has been studied sufficiently (46) so that no further analysis of this aspect is required.

The VCO used in our experiments was a prefabricated one. Having in mind the characteristics of a VCO specifically designed for applications, such as direct microwave generation, further work should be done on the design of a faster VCO. This would allow MSK generation of bit rates higher than 4 Mbits/sec.

However, it should be noted that higher bit rates will require greater spacing between the two reference spectral lines and therefore a higher level of injected power.

Finally integration of the overall system on a single microstrip is recommended for better matching between different components.

Some thought should be given also to power level stabilization techniques that would reduce the sensitivity of the system and keep the P_o/P_i ratio constant to the optimum value.

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APPENDIX A

MICROSTRIP CHARACTERISTIC IMPEDANCE CALCULATED FROM WORK OF WHEELER

WIDE STRIP APPROXIMATION ($W/H > 1$)

NARROW STRIP APPROXIMATION ($W/H < 1.0$)

RATIO OF FREE SPACE WAVELENGTH (λ_0) TO MICROSTRIP WAVELENGTH (λ_m)
CALCULATED FROM WORK OF WHEELER

WIDE STRIP APPROXIMATION ($W/H > 1$)

NARROW STRIP APPROXIMATION ($W/H < 1.0$)

APPENDIX B1

The oscillating condition of the VCO in the case of an injected signal, Eq. (3.10), can be written in voltage terms as

$$V + \operatorname{Re}\{ZI\} = e(t) \quad (\text{B.1.1})$$

where

V is the voltage drop across the active device
 $\operatorname{Re}\{ZI\}$ is the voltage drop across $Z(\omega)$, which is in series with the active device, and
 $e(t)$ is the injected signal

If the current flowing through the active device is represented by

$$i(t) = A \cos(\omega t + \phi)$$

or

$$I = Ae^{j(\omega t + \phi)}$$

then

$$\frac{di}{dt} = -A\left(\omega + \frac{d\phi}{dt}\right) \sin(\omega t + \phi) + \frac{dA}{dt} \cos(\omega t + \phi)$$

$$= \operatorname{Re} \left\{ \left[j\left(\omega + \frac{d\phi}{dt}\right) + \frac{1}{A} \frac{dA}{dt} \right] Ae^{j(\omega t + \phi)} \right\}$$

and

$$\frac{d^n i}{dt^n} = \operatorname{Re} \left\{ \left[j\left(\omega + \frac{d\phi}{dt}\right) + \frac{1}{A} \frac{dA}{dt} \right]^n Ae^{j(\omega t + \phi)} \right\}$$

Therefore the voltage drop across the active device

$$V = -RA \cos(\omega t + \phi) - \overline{XA} \sin(\omega t + \phi) \quad (\text{B.1.2})$$

For the calculation of the term $\text{Re}\{ZI\}$, we replace in $Z(\omega)$

ω by $\omega + d\phi/dt - j(1/A)(dA/dt)$. Assuming $d\phi/dt \ll \omega$

and $(1/A)(dA/dt) \ll \omega$, we have

$$\begin{aligned} Z(\omega) &= Z\left(\omega + \frac{d\phi}{dt} - j \frac{1}{A} \frac{dA}{dt}\right) \\ &= R(\omega) + jX(\omega) + \left[R'(\omega) + jX'(\omega)\right] \left[\frac{d\phi}{dt} - j \frac{1}{A} \frac{dA}{dt}\right] \end{aligned}$$

where $R(\omega) = \text{Re}\{Z(\omega)\}$, $X(\omega) = \text{Im}\{Z(\omega)\}$

and the primes indicate differentiation with respect to ω .

Finally

$$\begin{aligned} \text{Re}\{ZI\} &= \left[R(\omega) + R'(\omega)\frac{d\phi}{dt} + X'(\omega)\frac{1}{A}\frac{dA}{dt}\right] A \cos(\omega t + \phi) \\ &\quad - \left[X(\omega) + X'(\omega)\frac{d\phi}{dt} - R'(\omega)\frac{1}{A}\frac{dA}{dt}\right] A \sin(\omega t + \phi) \end{aligned} \quad (\text{B.1.3})$$

and

$$e(t) = a_0 \cos \omega_0 t \quad (\text{B.1.4})$$

After introducing Eq. (B.1.2) to Eq. (B.1.4). to the initial Eq. (B.1.1) a final equation is obtained from where the basic equations for the amplitude and phase of the oscillating current are derived

$$\begin{aligned} \left[R(\omega) - \bar{R} \right] X'(\omega) - \left[X(\omega) + \bar{X} \right] R'(\omega) + \left[Z'(\omega) \right]^2 \frac{1}{A} \frac{dA}{dt} \\ = \frac{1}{A} \left[X'(\omega) e_c(t) + R'(\omega) e_s(t) \right] \end{aligned} \quad (B.1.5)$$

$$\begin{aligned} \left[R(\omega) - \bar{R} \right] \dot{R}'(\omega) + \left[X(\omega) + \bar{X} \right] X'(\omega) + \left[Z'(\omega) \right]^2 \frac{d\phi}{dt} \\ = \frac{1}{A} \left[R'(\omega) e_c(t) - X'(\omega) e_s(t) \right] \end{aligned} \quad (B.1.6)$$

where

$$e_c(t) = a_0 \cos \phi$$

$$e_s(t) = a_0 \sin \phi$$

By introducing into Eq. (B.1.5) and (B.1.6) the quantities given from Eq. (3.13) to (3.16) the final equations for the current and phase, Eq. (3.17) and Eq. (3.18), can be obtained.

B2. Derivation of Eq. (3.34)

With respect to Fig. 3.4.3b (simplified varactor diode model) and Fig. 3.4.2. (VCO model)

$$Z(\omega) = R_S(V_i) + R_L + j\left(\omega L - \frac{1}{\omega C_j(V_i)}\right) + X_L \quad (\text{B.2.1})$$

and with the assumption of small variation of $Z_L = R_L + jX_L$ with ω

$$Z'(\omega) = \frac{dZ(\omega)}{d\omega} = j \left[L + \frac{1}{\omega^2 C_j(V_i)} \right] \quad (\text{B.2.2})$$

or

$$|Z'(\omega_i)|^2 = - \left[L + \frac{1}{\omega_i^2 C_j(V_i)} \right]^2 = - \left[\frac{(\omega_i/\omega)^2 + 1}{\omega_i C_j(V_i)} \right]^2 \quad (\text{B.2.3})$$

because ω , the output frequency of the VCO, depends on the values of L and $C_j(V_i)$ or $LC_j(V_i) = 1/\omega^2$, where ω can be empirically expressed by

$$\omega = 2\pi \times \left[1400 \times 10^6 + 9.09 \times 10^6 (V_i - 22.8) \right] \quad (\text{B.2.4})$$

from the voltage tuning characteristic of the VCO, Fig. 3.3.1.

Replacing Eq. (B.2.3) into Eq. (3.32) yields the final form of Eq. (3.34).

B3. Computer Program for the Numerical Solution of Eq. (3.34)

```

1 - 1.000 COMMON V1,V2,T1,C0,G,PI,F0
2 - 2.000 DIMENSION X(2),Y(2)
3 - 3.000 V1=25
4 - 4.000 V2=50
5 - 5.000 T1=1E-09
6 - 6.000 C0=0.8E-12
7 - 7.000 G=0.316
8 - 8.000 PI=3.1415
9 - 9.000 F0=20.0/180.0*PI
10 - 10.000 TMAX=2E-4
11 - 11.000 DT=1E-10
12 - 12.000 X(1)=0.0
13 - 13.000 X(2)=0.000001
14 - 14.000 T=0
15 - 15.000 ICOUNT=0
16 - 16.000 10 CALL RUNGE(T,DT,X,Y)
17 - 17.000 ICOUNT=ICOUNT+1
18 - 18.000 IF(ICOUNT.LT.1)GO TO 199
19 - 19.000 ICOUNT=0
20 - 20.000 XXXX=X(1)*180.0/PI
21 - 21.000 WRITE(6,150) T,XXXX
22 - 22.000 150 FORMAT(2E20.6)
23 - 23.000 199 CONTINUE
24 - 24.000 IF(T.LE.TMAX)GO TO 10
25 - 25.000 STOP
26 - 26.000 END
27 - 27.000 SUBROUTINE RUNGE(T,DT,Y,NY)
28 - 28.000 DIMENSION DY(2),Y(2),TMP(8)
29 - 29.000 K=1
30 - 30.000 CALL DERCAL(Y,DY,T)
31 - 31.000 TMP(1)=Y(1)
32 - 32.000 TMP(2)=Y(2)
33 - 33.000 510 DO 52 I=1,2
34 - 34.000 I1=I+2
35 - 35.000 I2=I1+2
36 - 36.000 I3=I2+2
37 - 37.000 GO TO (520,530,540,550),K
38 - 38.000 520 TMP(I1)=DY(I)*DT
39 - 39.000 Y(I)=TMP(I)+0.5*TMP(I1)
40 - 40.000 GO TO 52
41 - 41.000 530 TMP(I2)=DY(I)*DT
42 - 42.000 Y(I)=TMP(I)+0.5*TMP(I2)
43 - 43.000 GO TO 52
44 - 44.000 540 TMP(I3)=DY(I)*DT
45 - 45.000 Y(I)=TMP(I)+TMP(I3)
46 - 46.000 GO TO 52
47 - 47.000 550 Y(I)=TMP(I)+(TMP(I1)+2.0*(TMP(I2)+TMP(I3))+DY(I)*DT)/6.0
48 - 48.000 52 CONTINUE
49 - 49.000 GO TO(560,570,560,570),K
50 - 50.000 560 T=T+0.5*DT
51 - 51.000 570 K=K+1
52 - 52.000 IF(K.GT.4) RETURN
53 - 53.000 CALL DERCAL(Y,DY,T)
54 - 54.000 GO TO 510
55 - 55.000 END
56 - 56.000 SUBROUTINE DERCAL(X,Y,T)
57 - 57.000 COMMON V1,V2,T1,C0,G,PI,F0
58 - 58.000 DIMENSION X(2),Y(2)

```

```

59.000  VT=(V2-V1)*T0+V1
60.000  IF (VT.GT.V2) VT=V2
61.000  W=2*PI*(1.4E+9 + 9.09E+8 *(VT-22.8))
62.000  WI =2*PI*1.4E+9
63.000  CJ =C0/SQRT(1+VT/6.5)
64.000  COM=((WI/W)**2.0+1)/WI/CJ
65.000  A= COM *(280.0 +6*COS(F0))
66.000  B=COM*6*SQRT(280.0*280.0+75.0*75.0)*COS(F0+PI/6.0)
67.000  Y(1)=X(2)
68.000  Y(2)=A*X(2)+B*X(1)
69.000  RETURN
70.000  END

```

APPENDIX C

The following initial parameters are given and optimization with respect to S11 is requested.

CURRENT LINE:

```
?   TRL  AG  SE  50  500  4.9
?   DST  BD  PA  -20  -1247  4.18
?   CST  CC  PA  -20  -1247  4.18
?   TRL  DD  SE  -84  -1304  3.37
?   DST  EE  PA  -18  -3045  4.26
?   DST  FF  PA  -18  -3045  4.26
?   TRL  GG  SE  -84  -463  3.37
?   CAP  HH  PA  .18
?   SRX  II  PA  5.12  .4  1.52
?   CAX  AA  II
?   PRI  AA  IR  50  50000
?   END
?   150  300  10
?   END
?   .15
?   0.01
?   5 * 0  0  0
?   END
?   END-OF-FILE - RESET
```

CIRCUIT OPTIMIZATION WITH 12 VARIABLES

INITIAL CIRCUIT ANALYSIS

INPUT REFL. COEF. AND VSWR IN 50. OHM SYSTEM WITH 50.0 OHM LOAD

F(MHZ)	RHO(MAGN. ANGLE)	VSWR	RET L/C (DB)	Z(R+JX)OHM		
150.000	.819	167.7	10.05:1	-1.73	8.03	8.31
160.000	.826	161.9	10.49:1	-1.36	4.88	7.87
170.000	.831	155.8	10.80:1	-1.61	4.84	10.83
180.000	.833	149.1	10.96:1	-1.59	4.91	13.70
190.000	.833	141.7	10.98:1	-1.59	5.10	17.21
200.000	.831	133.4	10.83:1	-1.61	5.46	21.34
210.000	.827	123.8	10.58:1	-1.65	6.07	26.41
220.000	.821	112.5	10.17:1	-1.71	7.08	32.94
230.000	.814	99.1	9.74:1	-1.77	8.80	41.86
240.000	.807	83.0	9.34:1	-1.87	12.02	55.10
250.000	.803	63.6	9.13:1	-1.91	19.11	77.26
260.000	.806	40.9	9.32:1	-1.87	10.52	122.29
270.000	.821	15.7	10.20:1	-1.71	173.97	238.30
280.000	.849	-10.2	12.20:1	-1.43	279.90	-301.82
290.000	.882	-34.9	15.95:1	-1.09	33.49	-182.30
300.000	.914	-56.8	22.33:1	-.78	9.84	-91.73

OPTIMIZATION BEGINS WITH FOLLOWING VARIABLES AND GRADIENTS

VARIABLES	GRADIENTS
(1): 20.0000	(1): .470281
(2): 1247.00	(2): -.500083
(3): 20.0000	(3): .470281
(4): 1247.00	(4): -.500083
(5): 84.0000	(5): -.445843
(6): 1304.00	(6): -.492334
(7): 18.0000	(7): -1.02520
(8): 3045.00	(8): 1.64151
(9): 18.0000	(9): -1.02520
(10): 3045.00	(10): 1.64151
(11): 84.0000	(11): -.715256E-01
(12): 465.000	(12): -.518560E-01
ERR. F. = 3.467	

```

( 1): 34.5077
( 2): 676.412
( 3): 34.8063
( 4): 676.412
( 5): 116.511
( 6): 1639.04
( 7): 42.4626
( 8): 845.996
( 9): 42.4626
(10): 845.996
(11): 99.7988
(12): 514.719
ERR. F.= .013

```

```

-----***-----
( 1): 45.2589
( 2): 517.345
( 3): 45.2792
( 4): 517.436
( 5): 101.271
( 6): 1363.23
( 7): 50.2840
( 8): 688.567
( 9): 50.2840
(10): 688.567
(11): 112.523
(12): 545.555
ERR. F.= .002

```

-----***-----
FUNCTION TERMINATION WITH ABOVE VALUES. FINAL ANALYSIS FOLLOWS

150.000	.012	-105.9	1.02:1	-38.24	49.65	-1.17
160.000	.012	-104.5	1.02:1	-38.46	49.69	-1.15
170.000	.011	-102.1	1.02:1	-38.81	49.75	-1.12
180.000	.011	-98.3	1.02:1	-39.25	49.83	-1.08
190.000	.010	-92.7	1.02:1	-39.74	49.94	-1.03
200.000	.010	-84.7	1.02:1	-40.19	50.08	-.98
210.000	.010	-74.2	1.02:1	-40.40	50.25	-.92
220.000	.010	-61.7	1.02:1	-40.16	50.46	-.87
230.000	.011	-49.2	1.02:1	-39.35	50.70	-.83
240.000	.012	-38.3	1.03:1	-38.06	50.98	-.79
250.000	.015	-30.0	1.03:1	-36.50	51.31	-.77
260.000	.018	-24.1	1.04:1	-34.86	51.67	-.76
270.000	.022	-20.2	1.04:1	-33.25	52.08	-.76
280.000	.026	-17.7	1.05:1	-31.72	52.53	-.83
290.000	.031	-16.3	1.06:1	-30.27	53.02	-.92
300.000	.036	-15.7	1.07:1	-28.91	53.56	-1.04

END

1	6	0	6	8	3
---	---	---	---	---	---

FIN