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CARLTON BERNARD CYRUS

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AN ACTIVE UPONVERTER WITH GAIN

UTILIZING SILICON AVALANCHE DIODES

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NAME OF SUPERVISOR/NOM DU DIRECTEUR DE THÈSE

L. ~~LEE~~ A. STARK

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AN ACTIVE UPCONVERTER
WITH GAIN UTILIZING SILICON
AVALANCHE DIODES

by

Carlton Bernard Cyrus, B.Sc. (Eng.) (Hons.)

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

Master of Engineering

Carleton University
Ottawa, Canada
August 1979

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

"An Active Upconverter with Gain Utilizing Silicon
Avalanche Diodes"

submitted by Carlton B. Cyrus, B.Sc.(Eng.), in partial fulfillment of
the requirements for the degree of Master of Engineering.

Thesis Supervisor

Chairman, Department of
Electronics, Engineering

ABSTRACT

A solid state microwave upconverter with gain has been built and tested. The active device used in this upconverter was a silicon double-drift IMPATT diode.

The upconverter was designed to minimize instabilities and non-linearities. The output power of the upconverter was optimized and then its performance was observed under laboratory conditions.

Experimental results of gain and frequency response for the upconverter were obtained. They indicated that the building of an IMPATT-amplifier-upconverter with gain, and with a stable fairly linear performance, is possible.

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LIST OF SYMBOLS

C_p	Capacitance of IMPATT diode package (F)
CF	Centre-frequency component of upconverter frequency spectrum
$f_a(t)$	Amplitude modulation term in upconverter current (A)
f_{LO}	Local oscillator (LO) frequency (Hz)
f_0	Frequency of IMPATT oscillator (Hz)
G_0	Midband power gain of IMPATT amplifier
$g_p(t)$	Angle modulating term in upconverter current (rad.)
I_b	IMPATT diode bias current (A)
I_{CF}	Centre-frequency component of upconverter current (A)
I_{IF}	Amplitude of diode bias modulating current (A)
I_{LSB}	Lower-sideband component of upconverter current (A)
I_{LSB2}	Second-order lower-sideband component of upconverter current (A)
I_m	Maximum safe d.c. bias current of IMPATT diode (A)
I_n	Coefficient of nth harmonic in Fourier series of $f_a(t)$
I_{op}	DC operating current of IMPATT diode (A)
I_{th}	Threshold current. Bias current at onset of IMPATT oscillations (A)
I_{USB}	Upper-sideband component of upconverter current (A)
I_{USB2}	Second-order lower-sideband component of upconverter current (A)

$\bar{J}$	Current density (A/m^2)
$J_n(X)$	Bessel function of the first kind, order n and argument X .
L_p	Inductance of IMPATT diode package (H)
LSB	Lower-sideband component of upconverter frequency spectrum
LSB2	Second-order lower-sideband component of upconverter frequency spectrum
P_i	Signal power injected into injection-locked oscillator (W)
P_p	RF pump signal power level (W)
P_{uc}	Power level at circulator output of IMPATT upconverter (W)
Q_{ext}	External quality factor of IMPATT oscillator
R_D	Equivalent circuit resistance of IMPATT diode chip (Ω)
R_d	Resistance of packaged IMPATT diode (Ω)
R_L	Load resistance presented to IMPATT diode by circuit (Ω)
T_j	Junction temperature of IMPATT diode ($^{\circ}C$)
USB	Upper-sideband component of upconverter frequency spectrum
USB2	Second-order upper-sideband component of upconverter frequency spectrum
V_{IF}	Amplitude of diode bias modulating voltage (V)
V_{RF}	RF voltage amplitude (V)
X_D	Equivalent circuit reactance of IMPATT diode chip (Ω)
X_d	Reactance of packaged IMPATT diode (Ω)
Y_0	Characteristic admittance (U)

$Z_C(\omega)$	Circuit impedance seen by IMPATT diode (Ω)
Z_{in}	IF input impedance of upconverter (Ω)
Z_L	Load impedance presented to IMPATT diode by circuit (Ω)
Z_0	Characteristic impedance (Ω)
ΔR	Difference in magnitudes between load and diode resistances (Ω)
$\Delta\omega_m$	Locking range of injection-locked oscillator (Hz)
ϵ	Permittivity (F/m)
θ_n	Coefficient of nth harmonic in Fourier series of $g_p(t)$
θ_0	RF current phase angle (rad.)
ρ	Charge density (C/m^3)
ϕ	Electric potential (V)
ϕ	Phase difference between $f_a(t)$ and $g_p(t)$ (rad.)
ω_1	Angular frequency of IF modulating signal (rad./s)
ω_0	Angular frequency of IMPATT oscillator (rad./s)

CHAPTER 1

INTRODUCTION1.1 Background

The enormous increase in communications traffic over the past few years, and the subsequent crowding of the available lower frequency spectrum, has necessitated the use of higher and higher frequencies in the microwave region of the radio frequency spectrum. These higher frequencies also have the added advantage of increased information carrying capacity because of their larger available bandwidth. All of this has produced an increased demand for microwave power sources.

Recently, microwave solid state sources have demonstrated smaller size, lower weight, lower voltage operation and the potential of increased reliability over microwave tubes, which were formerly the sole source of microwave power. The present most effective solid state source of microwave power is the IMPATT diode [1,2].

The birth of the IMPATT diode occurred in 1958 when W.T. Read of Bell Telephone Laboratories first suggested that the generation of microwave power in a reverse-biased p-n junction was possible [3]. The name, IMPATT, given to the diode is an acronym for IMPact ionization Avalanche Transit Time, and summarizes the process which produces the diode's negative resistance characteristic.

The present state of the diode's development is the double-drift structure, in which the transit time of both holes and electrons

is utilized, unlike the original single-drift IMPATT which only used one type of carrier, generally electrons. The superiority of the double-drift structure results in increased power output over a wider frequency range, less noise and higher efficiency. Silicon (Si) and Gallium-Arsenide (GaAs) are presently the two semiconductor materials being used in commercially available IMPATTs. Si IMPATTs provide a larger percentage bandwidth over which useful output power can be generated, though GaAs diodes are less noisy [4]. Indium-Phosphide (InP) IMPATT diodes which seem potentially capable of offering better efficiency, noise characteristics and power output than either Silicon or Gallium-Arsenide diodes are also under development [2].

The two main commercial applications of IMPATTs are CW microwave sources and reflection type microwave amplifiers. Experimental research has also been done to investigate the possibility of using IMPATTs as the active elements in microwave upconverters. The most successful results so far have been obtained by employing an IMPATT-reflection-amplifier as the core of the upconverter.

1.2 Purpose of the Thesis

Microwave upconverters normally use varactor or Schottky-barrier diodes as their mixing elements. Of the two, varactor diodes are theoretically capable of providing gain. This is not so in practice because of the varactor's parasitic series resistance which becomes very pronounced at microwave frequencies.

The purpose of this thesis is to study the feasibility of building a solid state microwave upconverter with gain by utilizing the negative resistance available from an IMPATT diode. The intermediate frequency (IF), baseband bandwidth and upconverted frequency used are similar to those employed on a practical, operational microwave system in Canada. That system is Bell Canada's 8 GHz Digital Radio System (DRS-8) which has been recently developed to supplement its older 4 GHz microwave communication link [5].

The approach taken to obtain a practical, operational IMPATT-amplifier microwave upconverter is presented in the thesis, the outline of which follows.

1.3 Thesis Outline

In Chapter 2, the basic theory of operation of the IMPATT diode is explained, after which a theoretical analysis of the modulation process occurring in the upconverter is presented. The development of the equations used in the theoretical analysis is done in Appendix A. Chapter 3 describes the IMPATT circuit used in the upconverter and outlines the steps taken to optimize the upconverter's performance characteristics. Chapter 4 contains the laboratory results of the measurements made on the optimized upconverter, together with a sample theoretical calculation using the equations from Chapter 2; thus allowing relative comparisons to be made between the values obtained from the theoretical expressions and those experimentally observed. Finally, conclusions and recommendations are given in Chapter 5.

CHAPTER 2

THEORY OF THE IMPATT-AMPLIFIER-UPCONVERTER

2.1 Introduction

When properly biased, the IMPATT diode can produce a 90° phase lag between the RF current flowing in, and the RF voltage impressed upon the avalanche zone of the diode. Additional phase lag arises in the drift regions resulting in an RF negative resistance. The magnitude of the negative resistance produced is strongly dependent on the diode's bias current, which is exploited to operate the diode as an IMPATT-amplifier-upconverter in this research.

The basic theory of operation of an IMPATT diode is first presented, after which there is a theoretical analysis of the modulation process occurring within the IMPATT-amplifier-upconverter.

2.2 The IMPATT Mode

The IMPATT diode's negative resistance effect is produced by a time delay within the diode, the result of an avalanching process and a transit time delay of its charge carriers. The dynamic operation of the diode will be explained using the Read [6] structure as this is considered the most convenient for explaining the IMPATT's operation.

During operation, the diode is reverse-biased slightly above the breakdown voltage. With a suitable bias, if a small RF voltage is then impressed across the diode, the sum of the RF and d.c. voltages exceeds that required to give a breakdown field in the

A diode (Fig. 2.1). Avalanching then occurs in the thin p-region of the Read structure. The carriers are multiplied by impact ionization in the 'avalanche' zone, and the density of the carriers increases exponentially until the RF voltage goes negative, bringing the total voltage across the diode, and the electric field, below breakdown. The result is that the maximum avalanche current occurs when the RF voltage goes through zero, producing a 90° inductive phase lag (Fig. 2.1-c,d).

The carriers are then injected into the 'drift region' ('i' or intrinsic region) and traverse this portion of the diode at a constant scattering limited velocity. The transit time of the carriers through the drift region produces an additional phase delay to the one caused by the avalanching process. The resulting external current then lags the applied RF voltage by more than 90° (Fig. 2.1-e), producing a voltage-current characteristic which is equivalent to that of a negative resistance.

The phase delay due to the transit time depends on the frequency of the RF voltage. Thus, the total phase delay, and therefore the diode's efficiency, is frequency dependent. Also, below a certain 'resonant' frequency, the angular difference between current and voltage is reduced to less than 90° and the diode no longer shows a negative resistance. A particular diode is therefore usually manufactured to operate within a specified frequency range.

2.3 Practical Diodes

Because the Read diode is difficult to fabricate, the structure of the normal IMPATT diode is slightly different, as shown in Fig. 2.2.

Fig. 2.1 Waveforms in Read IMPATT diode during operation.

Fig. 2.2 Single and double-drift IMPATT diode structures and equivalent circuit of IMPATT.

The single-drift structure normally has electrons as the majority charge carriers, whereas the double-drift diode employs both electrons and holes as charge carriers. The double-drift IMPATT has a higher d.c. to RF conversion efficiency than the single-drift IMPATT when used as a CW source, viz. approximately 11% for the double-drift diode compared to approximately 7.5% for the single-drift diode. The double-drift diode also shows a negative resistance over a wider frequency range and has a higher power-handling capability than the single drift IMPATT [4,7]. However, the double-drift structure requires higher operating voltages than those for the single-drift structure due to the voltage across the extra drift region.

The diodes used in the experiments of this thesis were all double-drift IMPATT diodes. An equivalent circuit of the diode chip and package is also shown in Fig. 2.2. The negative resistance, R_D , and the reactance, X_D , represent the active diode chip. An inductor, L_p , and capacitor, C_p , are used to represent the package in which the chip is mounted.

2.4. Upconversion Theory

2.4.1 Preliminaries

Avalanching, together with transit time effects are responsible for an IMPATT diode's operating characteristic. Avalanching is a highly nonlinear process. Consequently, the diode is a very nonlinear device, its impedance being a highly nonlinear function of bias current, RF voltage amplitude, frequency and junction temperature. The equations which govern the diode operation are Poisson's equation,

$$\nabla^2 \phi = -\rho/\epsilon \quad (2.1)$$

and the continuity equation [6],

$$\nabla \cdot \bar{J} + \partial \rho / \partial t = 0 \quad (2.2)$$

where ϕ is the electric potential (V), ρ represents the charge density of the medium (C/m^3), ϵ is the medium's permittivity (F/m), and $\bar{J}$ is the current density (A/m^2). The nonlinear differential equation of the current flowing in the diode can be derived from the above equations. An involved analysis of the nonlinear behaviour of the IMPATT diode is not normally necessary from a practical engineering point of view. Only the characteristics of the diode which have significant effects on the experimental results will therefore be considered.

The diode impedance (Z_d) is dependent on bias current (I_b), frequency (ω), RF voltage level (V_{RF}) and junction temperature (T_j) [4]. By slowly varying (compared to the local oscillator frequency, ω_0) the bias current of the diode, the diode impedance will vary accordingly. This is the basic idea of an IMPATT-amplifier-upconverter as shown in Fig. 2.3-a.

Assuming a constant local oscillator (LO) power and frequency, the diode impedance is now mainly dependent on the bias current (I_b). If I_b is varied at an intermediate frequency (ω_1) high enough so that its period is small compared to the thermal time constant of the diode, then the diode junction temperature (T_j) may be assumed fairly constant, since it will not be able to respond to the instantaneous variations of the bias current. Now, if ω_1 is also small compared to the RF frequency

(a) Basic IMPATT-amplifier-upconverter network

(b) Equivalent circuit of upconverter network

Fig. 2.3 Basic network of IMPATT-amplifier-upconverter and its equivalent circuit.

(ω_0) , and the magnitude of the impedance variation ($|Z_d|$) produced by the varying bias current (I_b at ω_1) is small, then the magnitude of V_{RF} may be assumed constant and the frequency of the RF current through the diode is also fairly constant. The diode impedance may then be considered to be dependent only on the bias current variations (I_{IF}).

As shown in the upconverter network of Fig. 2.3 (a), an RF signal (ω_0) is being reflected off a varying (at ω_1) IMPATT impedance and is therefore modulated. Bias current variations mainly change the diode negative resistance (R_d), the corresponding change in reactance (X_d) being much smaller [3,8,9]. The resulting modulation produced is therefore mainly amplitude modulation, and the frequency spectrum of the modulated signal is expected to consist of a centre-frequency (CF) at ω_0 , and upper and lower-sidebands (USB and LSB respectively) at $(\omega_0 + \omega_1)$ and $(\omega_0 - \omega_1)$. A similar spectrum was experimentally observed. The analysis also considers the angle modulating effect of the varying reactance. This is of importance when using high indices of modulation as employed in the experiments in an effort to maximize the power output of the upconverter.

The upconverter network may be modelled by a series circuit consisting of the RF source (LO), the impedance of the IMPATT diode $\{Z_d(I_b)\}$ and the impedance of the circuit as seen by the active diode $\{Z_c(\omega)\}$, which is a function of frequency only (Fig. 2.3-b). The circuit impedance $\{Z_c(\omega)\}$ contains the circulator and source impedance, the waveguide impedance and the impedance of the diode network except the IMPATT diode itself.

The current $\{i(t)\}$ flowing in the circuit can be broken up into a sum of sinusoidal components. The total power supplied by the diode is the sum of the powers in the individual current components. The output power of the upconverter will be increased with respect to that of the RF source because of the amplification of the RF waveform provided by the diode negative resistance.

An analysis of the upconverter output will be done by analyzing the current waveform in the circuit, since the current is a representation of the upconverter's power output.

2.4.2 Analysis

In the absence of bias current modulation ($I_{IF} = 0$), the circuit acts like a basic negative resistance amplifier, and the circuit current may be represented as

$$i(t) = I_0 \cos \omega_0 t \quad (2.3)$$

where I_0 is the peak amplitude of the circuit current at the RF frequency of the local oscillator (ω_0).

I_0 can be calculated from the output power (P_{uc}), the power gain of the amplifier (G_0), and the magnitude of the diode negative resistance ($|R_d|$) using the following equation [10]:

$$I_0 = \sqrt{2 P_{uc} / |R_d| \left[1 + \frac{1}{\sqrt{G_0}} \right]} \quad (2.4)$$

On modulating the bias current of the diode, the current waveform in the circuit is described by

$$i(t) = (I_0 + f_a(t)) \cos(\omega_0 t + \theta_0 + g_p(t)) \quad (2.5)$$

where $f_a(t)$ is the amplitude modulation of I_0 produced, and $g_p(t)$ is the corresponding angle modulation of the RF signal produced by the varying impedance of the diode. θ_0 is the RF signal phase angle.

Both $f_a(t)$ and $g_p(t)$ are functions of the modulating current (I_{IF}), and are related to it through the impedance versus bias current characteristic of the diode, i.e. providing I_{IF} changes slowly enough (ω_1 small enough) to allow the amplitude and phase angle of the diode current to follow the instantaneous bias current variations. In general, $f_a(t)$ and $g_p(t)$ will both contain many harmonics of the bias modulating frequency (ω_1). A Fourier series will therefore be an appropriate form of representation for these two signals.

$$f_a(t) = \sum_{n=0}^{\infty} I_n \cos(n\omega_1 t) \quad (2.6)$$

$$g_p(t) = \sum_{n=0}^{\infty} \theta_n \cos(n\omega_1 t + \phi_n) \quad (2.7)$$

where I_n and θ_n are the coefficients of the n th harmonics of the series.

In the theory which follows, only the fundamental components of $f_a(t)$ and $g_p(t)$ will be considered. Such an assumption is valid when a small sinusoidal perturbation of the diode bias current is assumed.

While this linear approximation is lacking in generality, it can explain the main features of the IMPATT upconverter. Inclusion of higher order terms of $f_a(t)$ and $g_p(t)$ is straight forward but complicates the algebraic expressions rather unnecessarily. Thus,

$$f_a(t) = I_1 \cos \omega_1 t \quad (2.8)$$

$$g_p(t) = \theta_1 \cos(\omega_1 t + \phi) \quad (2.9)$$

Substitution of (2.8) and (2.9) into (2.5) results in

$$i(t) = \{I_0 + I_1 \cos \omega_1 t\} \cos\{\omega_0 t + \theta_0 + \theta_1 \cos(\omega_1 t + \phi)\} \quad (2.10)$$

The phase difference between the angle and amplitude modulating terms (ϕ) is due to extra delay in the amplitude response of the circuit compared to the delay in angle response, a result of the IMPATT impedance versus current characteristics. The angle variations of the current waveform lead the amplitude variations, and introduce some frequency modulation to the output waveform of the upconverter. This effect was observed experimentally and has also been reported in the literature [11,12]. Other contributions to the frequency (phase) and amplitude modulation characteristics that are due to the diode nonlinearities are assumed small and will not be considered in this analysis. As a rule, other nonlinearities became prominent when the diode was driven hard by the modulating signal or when it was in the region of saturation. Then, more complicated and extensive analysis than that presented here would be required.

Using Bessel function expansions on (2.10), the centre frequency (I_{CF}) and first and second-order upper and lower-sideband current components (I_{USB} , I_{USB2} ; I_{LSB} , I_{LSB2} , respectively) of the upconverter output may be approximated (see Appendix A) as

$$|I_{CF}|^2 = \{I_0 J_0(\theta_1)\}^2 + \{I_1 J_1(\theta_1) \cos \phi\}^2 \quad (2.11)$$

$$|I_{USB}|^2 = \left\{ \frac{I_1 J_0(\theta_1)}{2} - I_0 J_1(\theta_1) \sin \phi - \frac{I_1 J_2(\theta_1)}{2} \cos 2\phi \right\}^2 + \left\{ I_0 J_1(\theta_1) \cos \phi - \frac{I_1 J_2(\theta_1)}{2} \sin 2\phi \right\}^2 \quad (2.12)$$

$$|I_{LSB}|^2 = \left\{ \frac{I_1 J_0(\theta_1)}{2} + I_0 J_1(\theta_1) \sin \phi - \frac{I_1 J_2(\theta_1)}{2} \cos 2\phi \right\}^2 + \left\{ I_0 J_1(\theta_1) \cos \phi + \frac{I_1 J_2(\theta_1)}{2} \sin 2\phi \right\}^2 \quad (2.13)$$

$$|I_{USB2}|^2 = \left\{ I_0 J_2(\theta_1) \cos 2\phi + \frac{I_1 J_1(\theta_1)}{2} \sin \phi \right\}^2 + \left\{ \frac{I_1 J_1(\theta_1)}{2} \cos \phi - I_0 J_2(\theta_1) \sin 2\phi \right\}^2 \quad (2.14)$$

$$|I_{LSB2}|^2 = \left\{ I_0 J_2(\theta_1) \cos 2\phi - \frac{I_1 J_1(\theta_1)}{2} \sin \phi \right\}^2 + \left\{ \frac{I_1 J_1(\theta_1)}{2} \cos \phi + I_0 J_2(\theta_1) \sin 2\phi \right\}^2 \quad (2.15)$$

where $J_n(X)$ is the Bessel function of the first kind, n th order and argument X . Terms with Bessel functions of higher order than $n=2$ are assumed negligible and ignored. For values of $X \leq 0.5$, $J_3(X)$ is less than 9% of $J_2(X)$. Thus, if X is assumed to be in the above range, higher orders than 2 need not be considered. Calculations show (see Chapter 4) that terms with Bessel functions of order $n=2$ are also negligible compared to the other terms. Thus, expressions (2.11-2.15) can be further simplified if necessary.

The power in the individual spectral components of the diode current (I_k) is given by

$$P_k = \frac{|I_k|^2}{2} \times |R_d| \quad \text{watts} \quad (2.16)$$

where $|R_d|$ is the magnitude of the diode resistance at the frequency ω_k .

A comparison between (2.12) and (2.13) shows differences in signs between their second terms and fifth terms, a direct result of the angle modulating process. The result is a larger lower-sideband than an upper-sideband. This has been observed to be the general case in practice.

Typical values of the variables in (2.11-2.15) for low modulation indices are calculated in the results of Chapter 4.

2.5 Summary

The mechanism of operation of the IMPATT diode has been presented. When the impedance of the diode is varied by a modulating current (I_{IF}), the RF input signal to the upconverter network is also modulated.

The form of the frequency spectrum of the upconverter output at the circulator (Fig. 2.3-a) has been obtained by analyzing the waveform of the current $\{i(t)\}$ flowing in the circuit, giving expressions for the power in the spectral components of the upconverter output.

CHAPTER 3

CIRCUIT OPERATION AND OPTIMIZATION3.1 Introduction

A properly designed upconverter circuit is necessary to obtain optimum performance from the IMPATT diode. As explained in the previous chapter, an IMPATT-amplifier was utilized to obtain an upconverted signal by modulating the amplifier bias current, and correspondingly, the gain. So it is reasonable to assume that good upconverter performance should be obtained if the amplifier's performance is optimized.

The IMPATT network operation is explained. Following this, the stabilization of what is normally a major drawback of IMPATT diode circuits, namely bias circuit instabilities, is performed. The following section of this chapter then describes the injection-locked mode of operation used here in the IMPATT-amplifier. Power output optimization of the amplifier is then performed. Finally, the procedure followed to obtain the maximum upconverted signal (sideband) output from the upconverter is given.

3.2 The IMPATT Network

A detailed drawing of the IMPATT network which houses the IMPATT diode (Fig. 2.3) is shown in Fig. 3.1. The diode cavity was adapted from one presented in a paper by P. Ruella [13]. The network is composed of a TeCu alloy WR-112 waveguide cavity with the IMPATT

Fig. 3.1 The IMPATT network
(parts description on next page)

Figure 3.1

IMPATT NETWORK PARTS DESCRIPTION

Part No.	Description
1	Centre-conductor (brass). Bias port.
2	Biasing body (brass)
3	Quarter wave ($\lambda_0/4$) coaxial transformer.
4	Variable short-circuit
5	Plunger
6	Waveguide body. WR-112 (TeCu alloy)
7	Insulator
8	IMPATT diode
9	Diode mount (HP)
10	Tuning screw (brass)

diode kept in place with a diode mount (9), and biased by means of a centre-conductor (1) which is part of a quarter wave coaxial transformer (3). The purpose of the transformer is to transform the high characteristic wave impedance of the waveguide ($\approx 500\Omega$ at 8 GHz) down to the low impedance of the few ohms presented by the diode. The upper portion of the transformer-centre-conductor was designed with a taper to reduce coupling losses between the TE_{10} propagation mode in the waveguide and the TEM mode in the coaxial transformer, by providing a smooth transition from the horizontal electric field of the TEM mode to the vertical electric field of the TE_{10} mode. The variable short-circuit (4) is placed quarter wavelength away from the plane of the coaxial transformer and therefore simulates an open circuit at the transformer plane. A tuning screw (10) is used to make fine adjustments to the load impedance (Z_L) seen by the diode. Increasing the screw's depth raises the capacitive susceptance of the guide (Fig. 3.2), from 'A' to 'B' say. The result is a reduction in the admittance at the transformer plane (C) which is $1/3$ wavelength away from the screw. Thus the load impedance seen by the diode at the end of the quarter wave transformer is lowered.

A more detailed diagram of the diode mount and transformer circuit is shown in Fig. 3.3. The spacer (15), which is the fourth and final adjustable element in the IMPATT network, is shown in the diagram. The centre-conductor diameter, the variable short-circuit, and the tuning screw are the other variables, which were all described above. The spacer is used to vary the circuit inductance close to the

Fig. 3.2 Effect of screw tuner on waveguide admittance.

Fig. 3.3 Detailed diagram of IMPATT waveguide circuit and diode mount.
(parts description on next page)

Figure 3.3

PARTS DESCRIPTION OF DETAILED IMPATT WAVEGUIDE
CIRCUIT AND DIODE MOUNT

Part No.	Description
1	2-56 nut
2	Washer (rexolite)
3	Centre-conductor (brass)
4	Biasing body (brass)
5	Bucket choke (brass)
6	Waveguide body (WR-112)
7	Coaxial sleeve (TeCu alloy)
8	Coaxial body (TeCu alloy)
9	Collet guide
10	Clamp screw
11	Sleeve
12	Collet
13	Clamp washer
14	Clamp
15	Spacer
16	Bucket support (rexolite)

diode. The bucket-choke/bucket support assembly (5,16) simulates an RF short-circuit between itself and the waveguide at the plane where the centre-conductor enters the waveguide, but maintains a d.c. open circuit between both.

3.3 Bias Circuit Stabilization

The nonlinear microwave avalanche process of the IMPATT diode displays large signal rectification properties. These cause a low frequency RF voltage-induced negative resistance which may extend from d.c. to several tens, and perhaps hundreds, of MHz depending on the loaded circuit Q [14]. This spurious negative resistance results in excessive upconversion of bias circuit noise, bias circuit oscillations leading to diode burnout, and burnout that may accompany microwave tuning of the circuit. Instabilities are particularly prominent in the higher power, higher efficiency diodes as used here. Research in this area [14] has shown that a high impedance bias circuit reduces these instabilities. To produce a high RF impedance, yet provide a short circuit to the d.c. bias, a lossy inductor or 'resistive choke stabilizer' as it is called in [14] was placed in series with the constant current bias supply of the diode. The bias supply designed for the diode and the choke are shown in Fig. 3.4. The low frequency impedance of the choke is shown in Fig. 3.5.

3.4 Injection Locking

An oscillator may be locked or synchronized to an injected signal whose frequency is close to that of the oscillator. Provided the injected signal frequency remains within the locking range of the

Fig. 3.4 Current regulator circuit and the 'resistive choke stabilizer'

Fig. 3.5 Low frequency impedance of 'resistive choke stabilizer'

IMPATT oscillator, it behaves similarly to an amplifier of the input signal. The main advantage of this mode of operation is that much higher gain can be realized than in the stable amplifying mode, as will be explained later. An X-Y recorder plot of the oscillator spectrum of the highly nonlinear locking on process is shown in Figs. 3.6 and 3.7. The observed locking-on spectrum with its geometrical decrease in the amplitudes of the spectral lines (linear decrease on a logarithmic scale) on the ω_0 side of the LO can be explained mathematically with the use of Adler's equation [15]. Adler's equation is a mathematical representation of the dynamic behaviour of locking is as follows:

$$\Delta\omega_m = \frac{2\omega_0}{Q_{\text{ext}}} \sqrt{\frac{P_1}{P_0}} \quad (3.1)$$

where $\Delta\omega_m$ is the total locking range, ω_0 is the free running oscillation frequency, Q_{ext} is the external quality factor of the circuit, P_1 is the injected signal power, and P_0 is the free-running oscillator power.

If R_L is the load resistance seen by the diode and R_d is the value of the diode negative resistance, the midband power gain (G_0) provided by the IMPATT-amplifier is given by the well known equation,

$$G_0 = \left[\frac{R_L - R_d}{R_L + R_d} \right]^2 \quad (3.2)$$

Fig. 3.6 Locking-on spectrum of IMPATT oscillator with injected local oscillator (LO) signal (14 dBm) approaching from the left.

Fig. 3.7 Locking-on spectrum of IMPATT oscillator with injected local oscillator (LO) signal (14 dBm) approaching from the right.

Since R_d is negative, the denominator of (3.2) is the difference between the magnitudes of the load resistance and the diode resistance. The difference will be referred to as ΔR in the future. Thus, the power gain varies as the inverse square of ΔR .

An example of the IMPATT resistance variation with diode RF current, and as a function of d.c. bias current, is shown in Fig. 3.8 [10]. If the diode's optimum bias corresponded to the uppermost curve, then R_{L1} would be a typical value of load resistance chosen for operation as a stable negative-resistance amplifier. ΔR in this case is ΔR_1 . Any value of R_L that intersects the curve in question will result in oscillation. Thus a load resistance of 4.5Ω results in stable amplification. Now, if a lower value of R_L is chosen, say $R_{L2} = 3\Omega$, the circuit will oscillate at the stable operating point (1). At this resonant frequency of oscillation, the circuit's series impedance is equal to zero. A signal whose frequency is equal to or close to that of oscillation will therefore produce a current increase and move the circuit's operating point to (2) on the diagram. The circuit is now no longer operating as an oscillator but as an amplifier of the injected signal with a ΔR equal to ΔR_2 . The value of ΔR_2 is much smaller than ΔR_1 , thus giving the injection-locked-amplifier the capability of providing a larger power gain than the stable amplifier for a given operating condition. As the frequency difference between the natural oscillating frequency and the frequency of the injected signal increases, there comes a point when the conditions necessary for locking can no longer be satisfied and the signal unlocks [15]. The range over which the system remains in lock is dependent on the magnitude of the injected signal.

Fig. 3.8 Diode resistance (R_d) versus diode RF current for an X-band IMPATT diode [10].

Fig. 3.9 IMPATT oscillator power output versus diode bias current for different values of threshold current (I_{th})

The injection-locked mode of operation was used for the amplifier in the experiments since this mode offered larger conversion gains than that available from the stable amplifier.

3.5 Output Power Optimization

If conditions are ideal, the maximum power that can be added to the input RF signal of the IMPATT-amplifier is equal to the maximum possible power output of the free-running IMPATT diode oscillator, [7]. A good oscillator should therefore be the prerequisite to optimum power output from an injection-locked-amplifier, which in turn determines the upconverter's performance. To get the best performance from the upconverter, therefore, the first step taken was to optimize the power output of the free running oscillator.

Figure 3.9 displays the typical variations to be expected for the oscillator power output versus diode bias current for different values of threshold current, I_{th} . The threshold current is the current at the onset of oscillations and can be varied by changing the load resistance (R_L) presented to the diode by the circuit. The resistance was adjusted by changing the depth of the tuning screw and the diameter of the centre-conductor in the quarter wave transformer. The current, I_m , is the maximum safe operating current of the diode if thermal overload is to be avoided.

Consider the case where the load resistance is adjusted to give a threshold current I_{th1} . The maximum power obtained, curve 1, is less than the maximum the diode is capable of delivering if it were operating at its optimum bias current (I_m). Further increase in bias

current, however, results in a sharp decrease in the output power due to the saturation of the RF voltage. Diode burnout is imminent when operating on this dashed portion of the curve. B. van Iperen has suggested [16] that burnout in this operating region is due to the development of a transversely non-uniform current density within the diode, caused by the RF voltage-induced low frequency negative resistance.

On the other hand, take the case where the load resistance is increased to give a threshold current I_{th3} . Obviously, the thermal limit of the diode will be reached before the diode's maximum power output can be attained (curve 3).

There is an optimum load resistance, and corresponding threshold current (I_{th2}) which will give the maximum obtainable microwave power output from the diode when operating at the maximum permissible bias current. As a rule of thumb, this threshold current is normally in the region of one-third ($1/3$) the maximum bias current. The optimum threshold current was 60 mA in the case of the double-drift IMPATT diode used in the experiments. The maximum operating current was approximately 186 mA. The maximum free-running oscillator power extracted from this diode was 3.02 W with a free-running frequency of 7.878 GHz, and a d.c. to microwave conversion efficiency of 9.25%. Adjustments were made to the oscillating frequency by changing the inner diameter and/or thickness of the spacer ((15) in Fig. 3.3) which varied the inductance and therefore the net reactance of the load impedance provided by the circuit. Thus, the appropriate oscillation frequency, and injection-locked centre frequency, were obtained.

3.6 Sideband Magnitude Optimization

A plot of the sideband magnitude versus diode bias current for a fixed modulating current, using the optimized injection-locked-amplifier, is given in Fig. 3.10. The lower-sideband gave the maximum obtainable upconverted-signal power output and so was selected as the filtered upconverted signal. The general decrease in the sideband magnitudes at the higher d.c. bias currents could be due mainly to the saturation of the amplifier as can be seen from the shape of the centre-frequency power curve. The crossover in the sideband magnitudes may be the result of a combination of the nonlinear angle modulation which is taking place in the diode and other diode nonlinearities that are being brought into prominence by the saturation of the amplifier.

The d.c. operating current (I_{op}) was chosen to give the maximum lower-sideband output. The LSB component was selected by means of a bandpass filter as the upconverted signal output of the optimized upconverter.

3.7 Summary

With the intelligent use of the IMPATT diode's operational characteristics, the negative resistance IMPATT-amplifier-upconverter's power output has been optimized. The results obtained with the optimized circuit will be presented in the next chapter.

Fig. 3.10 Sideband magnitude and centre-frequency magnitude versus diode bias current for a fixed IF current amplitude (Nippon diode).

CHAPTER 4

RESULTS OF UPCONVERTER MEASUREMENTS4.1 Introduction

The results obtained using the optimized injection-locked IMPATT-amplifier-upconverter described in Chapter 3 are presented here. The measurement circuit used is described. A sample calculation using the theoretical results of Chapter 2 is then performed, giving typical values for the variables in the expressions (2.11-2.15). The diode's operating point was chosen in the median range between its breakdown voltage and the maximum allowable voltage, in the hope of obtaining as typical values as possible. The choice of its value of 110 mA (159 V) [see Sect. 4.3] was also made to facilitate the ease of calculations by not having too great a difference in magnitudes between the upper and lower sidebands. These values are then compared to the experimental values to illustrate the degree of accuracy to which the calculations were performed. Finally, single and swept frequency performance curves of the upconverter's operational characteristics are presented, explained and analyzed. From these, the best mode of operating the upconverter for a fairly linear, stable output is inferred and given in the summary.

4.2 Upconverter Measurement Circuit4.2.1 Circuit

The circuit used to obtain the results given later in this chapter is illustrated in Fig. 4.1. An HP 8620C sweep oscillator (3.6 - 8.6 GHz) was used as the local oscillator (LO) to supply the RF signal power (P_p at ω_0) to the system. The output power of the

Fig. 4.1 Measurement circuit for the injection-locked IMPATT-amplifier-upconverter

upconverter was then measured using an HP 432A power meter (0.01 - 10 GHz), which was in turn calibrated with the extremely accurate HP 436A digital power meter. The upconverter's frequency spectrum was displayed on an Ailtech 707 spectrum analyzer (0.001 - 12.4 GHz). A Wavetek model 2001 sweep-signal generator was used as the VHF source (I_{IF} at ω_1), and its output was amplified with a Motorola MHW580 40-300 MHz hybrid wide-band amplifier. The current transformer that monitored the IF current modulating the IMPATT bias current (I_b) was a Tektronix CT-1 ns current transformer. Both the current and voltage waveforms of the IF signal were displayed on a dual trace Tektronix 475 oscilloscope (0-200 MHz BW). To measure the d.c. voltage across the IMPATT (V_{dc}), a Fluke 8000A digital multimeter together with a 15 k Ω isolating resistor were employed.

4.2.2 Diodes

Double-drift Si IMPATTs were the diodes used in the system to obtain the final experimental results. They were the HP 5082-0608 and Nippon 468A 3W diodes. Overall better performance was obtained from the latter, so results from the Nippon diode will predominate in this chapter.

4.2.3 The Filter

Selection of the lower sideband (LSB) as the upconverted signal from the remainder of the spectrum at the circulator output was accomplished with the use of a bandpass filter. The filter was an iris-coupled 5-pole microwave cavity bandpass filter with a 92 MHz

bandwidth (7699.37 - 7791.37 MHz), and its measured attenuation characteristic is shown in Fig. 4.2. This is the actual type of filter being used on the 8 GHz Digital Radio System (DRS-8), the Bell Canada system on which the upconverter was designed to operate. The filter was employed only where both the upconverted (sideband) frequency was required, and the frequency range needed for those particular experimental results was within its bandwidth. Therefore, the filter's presence in the final upconverter measurements will be indicated on the figures containing the approximate curves.

4.3 Theoretical Calculations

The expressions for the current $\{i(t)\}$ flowing in the upconverter circuit as given in Chapter 2 (2.10), and those obtained in Appendix A (A.4-A.8) for the current components at the different frequencies are repeated here for convenience,

$$i(t) = \{I_0 + I_1 \cos \omega_1 t\} \cos\{\omega_0 t + \theta_0 + \theta_1 \cos(\omega_1 t + \phi)\} \quad (4.1)$$

where I_0 is the peak amplitude of the RF current in the circuit at frequency ω_0 . I_1 is the peak amplitude of the amplitude modulating signal, and θ_1 is the peak amplitude of the angle modulating signal due to the IF current (I_{IF}) at a frequency ω_1 , and shifted in phase from the amplitude modulating signal by an angle ϕ . Finally, θ_0 is the phase angle of the RF signal.

The main frequency components of $i(t)$ are as follows:

Fig. 4.2 Attenuation characteristic of the iris-coupled 5-pole microwave cavity filter (7699.37 - 7991.37 MHz)

i) Centre-frequency (CF) at ω_0 ,

$$|I_{CF}|^2 = \{I_0 J_0(\theta_1)\}^2 + \{I_1 J_1(\theta_1) \cos \phi\}^2 \quad (4.2)$$

ii) Upper-sideband (USB) at $(\omega_0 + \omega_1)$,

$$|I_{USB}|^2 = \left\{ \frac{I_1 J_0(\theta_1)}{2} - I_0 J_1(\theta_1) \sin \phi - \frac{I_1 J_2(\theta_1)}{2} \cos 2\phi \right\}^2 \\ + \left\{ I_0 J_1(\theta_1) \cos \phi - \frac{I_1 J_2(\theta_1)}{2} \sin 2\phi \right\}^2 \quad (4.3)$$

iii) Lower-sideband (LSB) at $(\omega_0 - \omega_1)$,

$$|I_{LSB}|^2 = \left\{ \frac{I_1 J_0(\theta_1)}{2} + I_0 J_1(\theta_1) \sin \phi - \frac{I_1 J_2(\theta_1)}{2} \cos 2\phi \right\}^2 \\ + \left\{ I_0 J_1(\theta_1) \cos \phi + \frac{I_1 J_2(\theta_1)}{2} \sin 2\phi \right\}^2 \quad (4.4)$$

iv) Second-order upper-sideband (USB2) at $(\omega_0 + 2\omega_1)$,

$$|I_{USB2}|^2 = \left\{ I_0 J_2(\theta_1) \cos 2\phi + \frac{I_1 J_1(\theta_1)}{2} \sin \phi \right\}^2 \\ + \left\{ \frac{I_1 J_1(\theta_1)}{2} \cos \phi - I_0 J_2(\theta_1) \sin 2\phi \right\}^2 \quad (4.5)$$

v) Second-order lower-sideband (LSB2) at $(\omega_0 - 2\omega_1)$,

$$|I_{\text{LSB2}}|^2 = \left\{ I_0 J_2(\theta_1) \cos 2\phi - \frac{I_1 J_1(\theta_1)}{2} \sin \phi \right\}^2 + \left\{ \frac{I_1 J_1(\theta_1)}{2} \cos \phi + I_0 J_2(\theta_1) \sin 2\phi \right\}^2 \quad (4.6)$$

In the above equations, $J_n(X)$ is the Bessel function of the first kind, nth order and argument X.

Also, from (2.4)

$$I_0 = \sqrt{2 P_{\text{uc}} / |R_d|} \left[1 + \frac{1}{\sqrt{G_0}} \right] \quad (4.7)$$

where P_{uc} is the output of the upconverter (with $I_{\text{IF}} = 0$) at the RF frequency ω_0 , and G_0 is the midband power gain provided by the amplifier.

Experimental results obtained with the Nippon diode biased at 110 mA (159 V), $f_0 = 7.885$ GHz, $P_p(\text{LO}) = 14$ dBm and $f_1 = 140$ MHz are given in Table 1. The relative magnitudes of the spectral components are displayed in Fig. 4.3.

The values of the variables in (4.2-4.7) were calculated using the following procedure. The angle, ϕ , was assumed small (as verified by calculations), so it may be assumed that $\sin \phi \approx 0$ and $\cos \phi \approx 1$. This approximation gave $|I_{\text{LSB}}|^2 = |I_{\text{USB}}|^2$. The argument, θ_1 , was then calculated using the difference between the sidebands and centre-frequency, and the value of θ_1 was used to calculate I_1 . Then using the more accurate expressions for $|I_{\text{LSB}}|^2$ and $|I_{\text{USB}}|^2$, and the measured difference between the upper and lower-sidebands, an approximate value of ϕ was found. The values thus obtained could be

TABLE 1

POWER IN SPECTRAL COMPONENTS OF
UPCONVERTER OUTPUT $(I_{IF} = 90/\sqrt{2} \text{ mArms})$

Spectral component	Power (dBm)
CF	30.8
USB	18.8
LSB	19.3
USB2	3.3
LSB2	4.3

$$P_{CF}(I_{IF} = 0) = 31.3 \text{ dBm}$$

Fig. 4.3 Spectral components of IMPATT-amplifier-upconverter with $I_{ac} = 110 \text{ mA}$, $P_p(L0) = 14 \text{ dBm}$, $I_{IF} = 90/\sqrt{2} \text{ mArms}$, $f_0 = 7.885 \text{ GHz}$ and $f_1 = 140 \text{ MHz}$.

used to select and compute more accurate values of θ_1 , I_1 and ϕ . The process is a numerical one.

First approximation results for the above example were

$$\theta_1 = 0.4 \text{ rad. (22.9}^\circ\text{)}$$

$$I_1 = 0.8 \text{ A}$$

and

$$\phi = 3.5^\circ$$

The value of I_0 used was 2.0 A and was obtained with an $|R_d| \approx 0.9 \Omega$, which is within the narrow range of operating values of R_d specified by the manufacturer. The value of R_d was assumed constant with frequency over the relatively small percentage frequency variation from $(\omega_0 - 2\omega_1)$ to $(\omega_0 + 2\omega_1)$ with respect to ω_0 ($\omega_1 = 140 \text{ MHz}$ is $\approx 2\%$ of ω_0 at 7.885 GHz). So only one value of R_d is used in the calculations. The measured and calculated results using this first approximation are given in Table 2 to show the degree of accuracy of the first approximation.

An 11% change in the value of R_d causes 6% variations in both I_0 and I_1 , and no noticeable change in ϕ . The first approximation of θ_1 does not require the use of R_d and so is not affected by a change in R_d .

4.4 Results

4.4.1 Preliminaries

The lower-sideband provided the maximum upconverted power output obtainable from the optimized upconverter, as shown in Chapter 3 which describes the circuit's optimization. This sideband was consequently used as the upconverted signal. However, the general behaviour of the upper-sideband as a function of the IF modulating signal amplitude

TABLE 2

CALCULATED AND MEASURED VALUES OF
THE RELATIVE MAGNITUDES BETWEEN THE
SPECTRAL COMPONENTS OF THE UPCONVERTER

	Calculated value (dB)	Measured value (dB)
$10 \log \frac{ I_{CF} ^2}{I_0^2}$	-0.3	-0.5
$10 \log \frac{ I_{LSB} ^2}{I_0^2}$	-11.0	-12.0
$10 \log \frac{ I_{LSB} ^2}{ I_{USB} ^2}$	+0.5	+0.5
$10 \log \frac{ I_{LSB2} ^2}{ I_{LSB} ^2}$	-16.0	-15.0

and frequency was similar to that of the lower-sideband. Thus, the shape of the curves given in the figures for the lower-sideband are of similar form to those for the upper-sideband.

A larger lower-sideband can be ascribed to the nonlinear frequency and signal-amplitude dependent behaviour of the IMPATT diode. Hines [11] has presented numerical calculations of the frequency conversion properties of IMPATT diodes which are in excellent qualitative agreement with the results in this thesis. Two sets of computed data from his paper displaying the IMPATT amplifier and upconverter gain-frequency characteristics as a function of the LO pump signal level (P_p at ω_0) are shown in Fig. 4.4. As can be seen, the lower frequency signals (lower-sidebands) are favoured in both cases for large pump signals. Such an unequal magnitude sideband response implies a mixture of amplitude and frequency modulation of the signal in the diode [11,12,17], an analysis of which has been given in Chapter 2.

The LO pump frequency (ω_0) of 7.885 GHz was chosen to allow the lower-sideband signal to fall at the centre of the passband of the filter while using the IF frequency band used in the DRS-8 system ($\omega_1 \approx 140 \pm 20$ MHz).

The level of the pump signal (≈ 14 dBm) was also chosen to be close in magnitude to that actually used in the DRS-8 system (≈ 16 dBm).

(a)

Computed data of IMPATT-amplifier gain versus frequency as a function of increasing pump signal (P_p) levels (1,2,3...; $P_{p3} > P_{p2} > P_{p1}$ etc.) after Hines [11]

(b)

Computed data of IMPATT-amplifier-upconverter gain versus frequency as a function of increasing pump signal (P_p) levels (1,2,3...; $P_{p3} > P_{p2} > P_{p1}$ etc.) after Hines [11]

Fig. 4.4

4.4.2 Data

The magnitude of the lower-sideband as a function of IF signal frequency for various values of IF voltage amplitudes (V_{IF}) is shown in Fig. 4.5. The figure displays an undulating characteristic of a form somewhat similar to the variations shown in the calculated gain of an IMPATT-upconverter given in Fig. 4.4, which is a further demonstration of the form of the gain-frequency characteristics of an IMPATT diode. The variation with frequency of the magnitude of the IF input impedance ($|Z_{in}|$) at the bias port of the diode is displayed in Fig. 4.6 along with the corresponding curve from Fig. 4.5. The IF input impedance was calculated with the monitored values of the IF current (I_{IF}) and voltage (V_{IF}). As is expected, the variation is of a similar form, though opposite in slope, to that of the sideband magnitude. The IMPATT is a current controlled/current controlling device, and an increase, say, of the output current (and power) for a fixed modulating voltage level will correspond to an increase in controlling current and therefore a drop in its input impedance as shown.

This suggests the use of a high impedance, current-driving source as the upconverter modulator in order to obtain fairly constant sideband magnitudes as the IF modulating signal frequency varies. By making point by point measurements, and readjusting the level of the IF drive to maintain a constant value of IF current, a high impedance (constant current) IF source was simulated. The lower-sideband level versus IF frequency for this case is plotted in Fig. 4.7. As expected, the variations have been reduced considerably. When the upconverter filter was added to the circuit, the power level was almost

Fig. 4.5 Lower-sideband magnitude versus IF frequency as a function of IF voltage amplitude. (Nippon diode)

Fig. 4.6 Lower-sideband magnitude and upconverter input impedance versus IF frequency for a fixed IF voltage amplitude. (Nippon diode)

Fig. 4.7 Lower-sideband magnitude versus IF frequency as a function of IF current amplitude. (Nippon diode)

constant over the smaller IF frequency range required for the DRS-8 system as shown in Fig. 4.8.

The upconverted signals in Figs. 4.7-4.8 are large enough to show gain above the fairly large pump signal power of 14 dBm for the higher modulating signals. Conversion gain, i.e. the ratio of upconverted to microwave pump power, as high as 9 and 10 dB were obtained with larger modulating signals. An attractive feature of the injection-locked mode is the ability to reduce the level of the pump signal while keeping the amplifier in lock but without causing a large reduction in sideband power level. By using a smaller pump signal, therefore, higher conversion gains than those in the experiments can be obtained.

A comparison of Figures 4.7 and 4.8 shows that the insertion of the filter in the circuit while taking the measurements shown in Fig. 4.8, resulted in a general decrease in the magnitudes of the lower sideband for different values of I_{IF} as compared to those in Fig. 4.7. The cause of the decrease could be twofold.

The amplitudes and phases of the harmonics of the centre-frequency of the upconverter being reflected back into the IMPATT network would change when the filter is employed, due to the ensuing change in harmonic termination caused by the filter. The reflected harmonics, especially the larger second harmonic, could affect the RF voltage and therefore the operating conditions of the network.

Secondly, any slight change in the reactance of the upconverter circuit as a result of the filter could cause a change in the natural oscillating frequency of the circuit and thus change the centre frequency of the injection-locked bandwidth. These two effects could change the former operating conditions of the circuit when the filter was absent and thus be possible causes of the different results.

Fig. 4.8 Filtered lower-sideband magnitude versus IF frequency as a function of IF current amplitude. (Nippon diode)

Fig. 4.9 Sideband and centre-frequency magnitudes versus IF current amplitude (log scale). (Nippon diode)

Fig. 4.10 Filtered lower-sideband magnitude versus IF frequency as a function of IF current amplitude. (HP diode)

Figure 4.9 displays the amplitude saturation characteristics of the upconverter as a function of the IF modulating current amplitude. The response is fairly linear, particularly for the LSB. However, in the region of 100 mAp, spurious responses begin to appear at the upconverter's output because of large-signal diode nonlinearities being forced into prominence by overdriving the diode.

The decrease in the centre-frequency magnitude as the IF current amplitude, and therefore the modulation index, was increased is consistent with the angle modulation which occurs within the diode. As the modulation index was increased, some of the power in the centre-frequency was spread into the sidebands.

The lower-sideband magnitude versus IF-frequency performance curves using the HP diode is shown in Fig. 4.10. There was a slightly larger spread in the data but the average gain was similar to that of the Nippon diode upconverter for this particular case.

4.5 Swept Frequency Response of Amplifier and Upconverter

4.5.1 Measurement Circuit

Figure 4.11 shows the measurement circuit used to make the swept frequency measurements of the IMPATT-amplifier and the IMPATT-amplifier-upconverter. The network analyzer was an Alfred model 7051 sweep network analyzer system with a 60 dB dynamic range. The detector used with the system was an Alfred model 1001 crystal detector with a frequency range from 30 MHz to 8 GHz. The Wavetek VHF signal generator was levelled to within 0.25 dB at a level of 5.2 dBm using an Anzac H-1-4 hybrid junction (2-1000 MHz), an HP 423A crystal detector (0.01-12.4 GHz) and the Motorola VHF amplifier. The HP sweep generator

Fig. 4.11 Swept frequency measurement circuit for the IMPATT-amplifier and IMPATT-amplifier-upconverter.

17dBm

- (a) Swept frequency response of IMPATT-amplifier. Input signal level is 14 dBm (P_p) and frequency is swept from 7.7 to 7.8 GHz.

Horiz.: 10 MHz/div.
Vert.: 2 dB/div.

7dBm

- (b) Swept frequency response of IMPATT-amplifier-upconverter. Filtered LSB magnitude is shown. Modulating signal (P_{IF}) level is 5.2 dBm and frequency is swept from 120 to 160 MHz.

Horiz.: 4 MHz/div.
Vert.: 0.5 dB/div.
 $P_p = 14$ dBm at 7.911 GHz

-3dBm

- (c) Spectrum analyzer display of filtered lower-sideband (left) and centre-frequency (right) attenuated by filter.

Vert.: 5 dB/div.
Horiz.: 100 MHz/div.

Fig. 4.12 Swept frequency response of IMPATT-amplifier and upconverter (a), (b). Spectrum of upconverter output (c).

(LO) supplying the RF signal was levelled with an HP 788C directional detector to within 0.25 dB also. The filter (dashed) was employed when taking the upconverter measurements.

4.5.2 Swept Frequency Results

The IMPATT-amplifier swept frequency response is displayed in Fig. 4.12(a). The input signal to the amplifier was swept from 7.7 to 7.8 GHz and was at a level of 14 dBm. That frequency span was chosen to cover the range of interest for the upconverted (LSB) signal.

The response of the IMPATT-amplifier-upconverter is shown in Fig. 4.12(b) for a 5.2 dBm IF modulating signal that was swept from 120 to 160 MHz. The frequency of the 14 dBm RF signal was 7.911 GHz. That particular frequency was chosen so that the highest frequency of the filtered lower-sideband was just within the bandwidth of the filter, the purpose of which was to attenuate the unwanted upper-sideband (USB) and centre-frequency (CF) as much as possible by having them as far outside the filter's bandwidth as practicable without affecting the magnitude of the wanted lower-sideband (LSB). The output spectrum of the upconverter is shown in Fig. 4.12(c). Besides the wanted lower-sideband, the only other visible frequency component is the centre-frequency which is 18 dB below the upconverted signal. With higher modulating baseband signal levels, this difference can be increased because of the corresponding increase in the LSB level.

The operating point of the IMPATT diode for the results shown was chosen to give the largest upconverted-frequency level for the limited IF drive available from the levelling system. It was 143 V at 48 mA. The slight difference between this operating point and

the one in the previous results (150 V) was due to the loading effect the filter had on the circuit. Optimization was done without the filter in the previous case, but with the filter in the second instance. Isolating the filter from the remainder of the circuit will be advantageous in a final commercial system in order to eliminate any detrimental effects the loading effect of the filter may have on the circuit.

The shape of the upconverter response in Fig. 4.12(b) is similar to that of the amplifier as shown in Fig. 4.12(a) as is to be expected. The IMPATT-amplifier-upconverter's characteristics and performance will be determined by the IMPATT-amplifier from which is constructed.

4.6 Summary

The results indicate that it is possible to obtain gain and a fairly linear response with an IMPATT-amplifier-upconverter. To help linearize the upconverted-signal level versus IF-frequency variation, it will be necessary to use a high output-impedance current-driver as the signal source. Also, it will be necessary to keep the modulating current under a certain maximum value if a linear output with respect to the modulating signal amplitude is to be obtained.

CHAPTER 5

CONCLUSIONS AND RECOMMENDATIONS

Upconversion with gain has been shown as possible with the use of an IMPATT diode at practical baseband and upconverted frequencies. In the experiments, the actual frequencies used were similar to those of an operational microwave digital radio system.

In a completely solid state system, upconversion with gain is particularly advantageous. Multistage solid state amplifiers are at present necessary to compete with the power gain a travelling wave tube (TWT) can deliver. Solid state systems are therefore not capable of easily handling the large conversion losses that can be found in a varactor upconverter used in a TWT system.

Although the present IMPATT-upconverter was operated at conventional microwave frequencies, the mm-wave region of the electromagnetic spectrum holds enormous scope for such an upconverter. IMPATTs are capable of operating at very high frequencies, much higher than a TWT. The IMPATT diode can work at frequencies of up to 300 GHz at present, and will therefore be an invaluable source of upconverted microwave power in the high frequency mm-wave region.

There were two main difficulties encountered in operating the diode in the IMPATT-amplifier-upconverter mode. These were bias circuit oscillations due to a low frequency negative resistance displayed by the diode, and the upconverter's nonlinear gain-frequency

characteristic. The former instability was eliminated with the use of a stabilizing design by Brackett [14]. Then, it was shown that the latter nonlinearity can be minimized by using a high impedance current-driving baseband signal source rather than a voltage source.

The theoretical analysis gave an insight into the modulating process occurring within the diode upconverter circuit, thus giving a better understanding of the upconverter's operational process.

The cost of an IMPATT upconverter is higher than its varactor counterpart. However, the gain provided by the IMPATT system will allow the use of a cheaper less powerful amplifier in the final stage than that of the varactor-TWT system. Therefore, the overall cost of the two systems will possibly be comparable.

To augment this feasibility study, operating such an upconverter on an actual microwave system would be very informative. This would give direct comparison performances between the IMPATT-upconverter and the commonly used upconverters, like the varactor-diode-upconverter for example. Such testing could be accompanied by final optimization adjustments, providing the opportunity to evaluate whether the gain provided by the IMPATT-amplifier-upconverter outweighs any performance or lifetime disadvantages it may have when compared to presently used varactor upconverters.

APPENDIX A

FREQUENCY ANALYSIS OF UPCONVERTER OUTPUT

As shown in Chapter 2, the current $\{i(t)\}$ flowing in the upconverter circuit may be expressed as {from (2.10)}

$$i(t) = \{I_0 + I_1 \cos \omega_1 t\} \cos\{\omega_0 t + \theta_0 + \theta_1 \cos(\omega_1 t + \phi)\} \quad (\text{A.1})$$

Using the Bessel function identity [18]

$$\cos(\alpha + X \cos \beta) = \sum_{n=-\infty}^{\infty} J_n(X) \cos(\alpha + n\beta + n\pi/2) \quad (\text{A.2})$$

together with

$$J_{-n}(X) = (-1)^n J_n(X) \quad (\text{A.3})$$

where $J_n(X)$ is the Bessel function of the first kind, n th order and argument X . And assuming the index of modulation is low enough so that $\theta_1 \ll 1$, then the Bessel functions of higher order than $n=2$ are negligible. From (A.1),

$$\begin{aligned} i(t) = (I_0 + I_1 \cos \omega_1 t) & \left[J_0(\theta_1) \cos(\omega_0 t + \theta_0) \right. \\ & + J_1(\theta_1) \cos(\omega_0 t + \theta_0 + \omega_1 t + \phi + \pi/2) \\ & \left. + J_{-1}(\theta_1) \cos(\omega_0 t + \theta_0 - \omega_1 t - \phi - \pi/2) \right] \end{aligned}$$

$$\begin{aligned}
& + J_2(\theta_1) \cos(\omega_0 t + \theta_0 + 2\omega_1 t + 2\phi + \pi) \\
& + J_{-2}(\theta_1) \cos(\omega_0 t + \theta_0 - 2\omega_1 t - 2\phi - \pi) \\
& + \dots]
\end{aligned}$$

where $(\omega_0 t + \theta_0) \equiv \alpha$, $(\omega_1 t + \phi) \equiv \beta$ and $\theta_1 \equiv X$ in (A.2)

Using (A.3) and the identity $\cos(A + \pi) = -\cos A$ in the above equation

$$\begin{aligned}
i(t) = (I_0 + I_1 \cos \omega_1 t) & \left[J_0(\theta_1) \cos(\omega_0 t + \theta_0) \right. \\
& + J_1(\theta_1) \cos\{(\omega_0 + \omega_1)t + \theta_0 + \phi + \pi/2\} \\
& + J_1(\theta_1) \cos\{(\omega_0 - \omega_1)t + \theta_0 - \phi + \pi/2\} \\
& + J_2(\theta_1) \cos\{(\omega_0 + 2\omega_1)t + \theta_0 + 2\phi + \pi\} \\
& \left. + J_2(\theta_1) \cos\{(\omega_0 - 2\omega_1)t + \theta_0 - 2\phi - \pi\} \right]
\end{aligned}$$

Now, using the identity

$$\cos A \cdot \cos B = 1/2 [\cos(A-B) + \cos(A+B)],$$

$$\begin{aligned}
i(t) = I_0 J_0(\theta_1) \cos(\omega_0 t + \theta_0) \\
+ I_0 J_1(\theta_1) \cos\{(\omega_0 + \omega_1)t + \theta_0 + \phi + \pi/2\} \\
+ I_0 J_1(\theta_1) \cos\{(\omega_0 - \omega_1)t + \theta_0 - \phi + \pi/2\}
\end{aligned}$$

$$\begin{aligned}
& + I_0 J_2(\theta_1) \cos\{(\omega_0 + 2\omega_1)t + \theta_0 + 2\phi + \pi\} \\
& + I_0 J_2(\theta_1) \cos\{(\omega_0 - 2\omega_1)t + \theta_0 - 2\phi - \pi\} \\
& + \frac{I_1 J_0(\theta_1)}{2} \left[\cos\{(\omega_0 - \omega_1)t + \theta_0\} + \cos\{(\omega_0 + \omega_1)t + \theta_0\} \right] \\
& + \frac{I_1 J_1(\theta_1)}{2} \left[\cos\{(\omega_0 + 2\omega_1)t + \theta_0 + \phi + \pi/2\} + \cos\{\omega_0 t + \theta_0 + \phi + \pi/2\} \right. \\
& \quad \left. + \cos\{\omega_0 t + \theta_0 - \phi + \pi/2\} + \cos\{(\omega_0 - 2\omega_1)t + \theta_0 - \phi + \pi/2\} \right] \\
& + \frac{I_1 J_2(\theta_1)}{2} \left[\cos\{(\omega_0 + 3\omega_1)t + \theta_0 + 2\phi + \pi\} + \cos\{(\omega_0 + \omega_1)t + \theta_0 + 2\phi + \pi\} \right. \\
& \quad + \cos\{(\omega_0 - \omega_1)t + \theta_0 - 2\phi - \pi\} \\
& \quad \left. + \cos\{(\omega_0 - 3\omega_1)t + \theta_0 - 2\phi - \pi\} \right]
\end{aligned}$$

Collecting terms of like frequency, the spectral components of $i(t)$ are

Centre-frequency, CF : $I_0 J_0(\theta_1) \cos(\omega_0 t + \theta_0)$

$$\begin{aligned}
& + \frac{I_1 J_1(\theta_1)}{2} \left[\cos\{\omega_0 t + \theta_0 + \phi + \pi/2\} \right. \\
& \quad \left. + \cos\{\omega_0 t + \theta_0 - \phi + \pi/2\} \right]
\end{aligned}$$

Upper-sideband, USB : $I_0 J_1(\theta_1) \cos\{(\omega_0 + \omega_1)t + \theta_0 + \phi + \pi/2\}$

$$+ \frac{I_1 J_0(\theta_1)}{2} \cos\{(\omega_0 + \omega_1)t + \theta_0\} + \frac{I_1 J_2(\theta_1)}{2} \cos\{(\omega_0 + \omega_1)t + \theta_0 + 2\phi + \pi\}$$

Lower-sideband, LSB : $I_0 J_1(\theta_1) \cos\{(\omega_0 - \omega_1)t + \theta_0 - \phi + \pi/2\}$

$$+ \frac{I_1 J_0(\theta_1)}{2} \cos\{(\omega_0 - \omega_1)t + \theta_0\} + \frac{I_1 J_2(\theta_1)}{2} \cos\{(\omega_0 - \omega_1)t + \theta_0 - 2\phi - \pi\}$$

Second-order upper-sideband, USB2 : $I_0 J_2(\theta_1) \cos\{(\omega_0 + 2\omega_1)t + \theta_0 + 2\phi + \pi\}$

$$+ \frac{I_1 J_1(\theta_1)}{2} \cos\{(\omega_0 + 2\omega_1)t + \theta_0 + \phi + \pi/2\}$$

Second-order lower-sideband, LSB2 : $I_0 J_2(\theta_1) \cos\{(\omega_0 - 2\omega_1)t + \theta_0 - 2\phi - \pi\}$

$$+ \frac{I_1 J_1(\theta_1)}{2} \cos\{(\omega_0 - 2\omega_1)t + \theta_0 - \phi + \pi/2\}$$

The components at each frequency are added vectorially as illustrated in Fig. A.1 to give

$$|I_{CF}|^2 = \{I_0 J_0(\theta_1)\}^2 + \{I_1 J_1(\theta_1) \cos \phi\}^2 \quad (\text{A.4})$$

$$|I_{USB}|^2 = \left\{ \frac{I_1 J_0(\theta_1)}{2} - I_0 J_1(\theta_1) \sin \phi - \frac{I_1 J_2(\theta_1)}{2} \cos 2\phi \right\}^2$$

$$+ \left\{ I_0 J_1(\theta_1) \cos \phi - \frac{I_1 J_2(\theta_1)}{2} \sin 2\phi \right\}^2 \quad (\text{A.5})$$

Fig. A.1 Vectorial components of upconverter frequency spectra as seen at circulator output

$$|I_{LSB}|^2 = \left\{ \frac{I_1 J_0(\theta_1)}{2} + I_0 J_1(\theta_1) \sin \phi - \frac{I_1 J_2(\theta_1)}{2} \cos 2\phi \right\}^2 + \left\{ I_0 J_1(\theta_1) \cos \phi + \frac{I_1 J_2(\theta_1)}{2} \sin 2\phi \right\}^2 \quad (A.6)$$

$$|I_{USB2}|^2 = \left\{ I_0 J_2(\theta_1) \cos 2\phi + \frac{I_1 J_1(\theta_1)}{2} \sin \phi \right\}^2 + \left\{ \frac{I_1 J_1(\theta_1)}{2} \cos \phi - I_0 J_2(\theta_1) \sin 2\phi \right\}^2 \quad (A.7)$$

$$|I_{LSB2}|^2 = \left\{ I_0 J_2(\theta_1) \cos 2\phi - \frac{I_1 J_1(\theta_1)}{2} \sin \phi \right\}^2 + \left\{ \frac{I_1 J_1(\theta_1)}{2} \cos \phi + I_0 J_2(\theta_1) \sin 2\phi \right\}^2 \quad (A.8)$$

These, (A.4-A.8), are the expressions given in (2.11-2.15).

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