

THE DESIGN OF A 94 GHz HIGH RESOLUTION COHERENT RADAR

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at the University of Cape Town

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I declare that this thesis is my own work and is based on results of experiments and research work done by me. The material has not been submitted for any other degree.

Signed by candidate

Abraham Francois Celliers

ABSTRACT

This thesis outlines the design procedure for a 94 GHz high resolution coherent radar. The technique used to obtain coherency is unique in the sense that the Pulsed Locked Oscillator (PIL0) and the first and second coherent oscillators (Coho) are phased locked to a low frequency (300 MHz) crystal oscillator. This technique assures that the phase of the transmit signal is preserved in the RF and IF Coho sources respectively.

A unique method is derived whereby the coherency of an injection locked system can be measured. The mathematical derivation gives the coherency of the system as the phase difference between the reference signal and the transmit signal.

A typical application of this high resolution radar is to use it as an early warning radar against high voltage transmission lines.

The thesis consists of four main sections, i.e. system specifications, measurement techniques, measured results and component design.

In the section on system specifications, basic parameters and specifications are derived for the carrier vehicle and the sensor. It is shown how certain manoeuvrability properties of the carrier vehicle influence the electrical properties and abilities of the sensor.

In the following section measuring techniques for the coherency of pulsed injection locked sources and phase noise of CW sources are described. Once again, the author wants to emphasize the uniqueness of the coherency measurement technique. Measured results obtained with this method show good comparison with published data. The attractiveness of this method is that it gives the coherency of a system as the phase difference between the transmit and reference signal in degrees, and not as the peak to valley ratio in dBc.

In Section 3, measured results are shown and compared with theoretically predicted data. The achieved results are as good and in some cases even better than the predicted and published data.

In the last section the design procedure for the different components, i.e. CW second harmonic Gunn oscillators, CW and pulsed Impatt oscillators, antenna, modulator, etc. is described and the characterization of the components are done. The results achieved compared favourably with published data of commercially available components.

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NOMENCLATURE

θ	antenna beamwidth in degrees.
d	distance vehicle can move in metres.
a	acceleration (m/s^2).
t	time constant in seconds.
v	velocity m/s .
R	distance in metres.
σ_c	clutter cross section in m^2 .
f_d	doppler shift in Hz.
G	antenna gain in dBi.
P_r	received power in dBm.
P_t	transmit power in dBm.
σ	radar cross section of target in m^2 .
L	plumbing and atmospheric losses.
N	thermal noise in dBm.
k	Boltzman's constant.
T	equivalent noise temperature.
B	bandwidth in Hz.
F	noise figure in dB.
$Z(\omega)$	circuit impedance.
$-Z(A)$	device impedance.
$-\bar{Z}(A)$	complex conjugate of $-Z(A)$.
ω	instantaneous oscillating frequency ($2\pi f$).
ϕ	phase difference between injected current and output voltage.
α	pulse envelope $\alpha > 0$ on rise period $\alpha < 0$ on fall period. $\alpha = 0$ on pulse pedestal.

Q_{ext}	external Q of locked source.
A_0	signal amplitude of sine wave.
$\phi_r(t)$	randomly fluctuating phase term.
$\Delta\phi$	absolute phase difference between two signals.
$V'_{\text{ref}}(t)$	signal via reference line.
$B_{\text{op}}(t)$	pulsed signal amplitude.
$\rho(t)$	1 for $nT \leq t < nT + d$. ($-\infty < n < +\infty$). 0 elsewhere.
d	pulse width.
ω_p	pulsed signal frequency.
$\phi_p(t)$	pulsed oscillator phase at start of pulse.
T	period of pulse train.
ω_r	reference signal frequency.
ϕ_r	phase of reference signal.
ϕ_d	instantaneous phase difference of the pulsed and reference oscillator.
$\phi_d(t)$	is a random variable.
$\overline{\phi_d(t)}$	mean or average value of $\phi_d(t)$.
$R_\phi(\tau)$	autocorrelation and is defined by: $\lim_{T \rightarrow \infty} \frac{1}{T} \int_{-T/2}^{T/2} \phi(t) \phi(t + \tau) dt$
$\overline{\phi^2(t)}$	mean square value of $\phi(t)$.
$\overline{\phi(t)^2}$	mean squared value of $\phi(t)$.
$\langle \phi(t) \rangle$	average or mean value of $\phi(t)$.
$E[\phi(t)]$	ensemble or statistical averages of $\phi(t)$.
$\overline{\phi(t)}$	mean value or DC component of signal.
$\overline{\phi(t)^2}$	mean square value or DC power.

$\overline{\phi^2(t)}$	mean square value or total average power.
$\overline{\sigma_\phi^2}$	$\overline{\phi^2(t)} - \overline{\phi(t)}^2$ is the variance or the AC power.
σ_ϕ	standard deviation or the rms value or the AC power.
$\epsilon(t)$	amplitude fluctuations of a sine wave.
$\Delta\phi(t)$	randomly fluctuating phase term.
$V_n(\omega)$	noise spectrum of a noisy sine wave.
$\Delta\phi(\omega)$	spectrum of $\Delta\phi(t)$.
$S_{\Delta\phi}(f)$	one-sided spectral distribution of the phase fluctuations on a per hertz basis.
$L(f)$	power of the phase fluctuations in a sideband in dBc/Hz.
ℓ_2	loss of delay line.
$\overline{Z}_T$	wave input impedance.
$\overline{Z}_i$	input wave impedance.
Z_i	load wave impedance.
$J_n(kr)$	Bessel's function of the first kind of order n .
$N_n(kr)$	Newman's function of the second kind of order n .
ϵ_r	dielectric constant.
λ_g	guided wavelength.
λ_c	cut-off wavelength.
β	Snell's law.

CHAPTER 1

INTRODUCTION

This thesis describes the design and performance of a 94 GHz short-pulse, low duty cycle, high resolution coherent injection locked radar system for sensor applications, with the specific use as an early warning radar against high voltage transmission lines.

The recent development of solid-state components for frequencies around 94 GHz has made it possible to design coherent millimeter-wave radar systems. Key components of such systems are high power pulsed silicon Impatt oscillators, CW Impatt oscillators, second harmonic Gunn oscillators, filters, circulators, isolators, couplers and antennas.

Elementary system specifications are derived for the delivery vehicle and the millimeter-wave sensor. Each of the separate components of the system were designed, built and tested. Measurements were taken with the sensor and are presented. Specific attention is given to the measurement of intra-pulse phase ripple, additive noise, injection locking and coherency. The above-mentioned parameters are critical in the design of a coherent sensor and special care should be given to the various components in the design stage.

The sensor is not built into a carrier system yet, but was built up as a laboratory model and measurements were taken from the laboratory to various objects and distances outside the laboratory.

There is only one article in literature [1] which described a 66 GHz collision warning sensor for helicopters. The system is a non-coherent pulsed radar with the following specifications:

Operating frequency	66 GHz
Output power	1 W
Pulse length	50 ns
PRF	125 kHz
Antenna gain	37 dB
-3 dB beamwidth	2°
Receiver noise figure	13 dB
Receiver bandwidth	100 MHz
Field of view (azimuth and elevation)	180° x 30°
Detection range: 3 mm ϕ wires	250 m maximum
30 mm ϕ wires	600 m maximum

CHAPTER 2

OPERATIONAL AND DYNAMIC SPECIFICATIONS
OF A HIGH RESOLUTION MILLIMETER-WAVE COHERENT RADAR

2.1 INTRODUCTION

Avoiding high voltage transmission lines seems to be one of the major tasks of a helicopter pilot [1-3], and therefore the aim of this work is to design a high resolution radar which will be used as an early warning radar to avoid high voltage transmission lines in a situation where the system will not be confused by natural clutter. The following assumptions will be made:

- 1) The measuring area is subject to all environmental deterrents.
- 2) The accuracy of the system must allow for a 100% detection.
- 3) The obstacles are stationary.
- 4) There will not be countermeasures from the obstacle.

The question is often asked, what is the reason for working at millimeter-waves, especially at 94 GHz. The answer to this is probably that there is a commonality between millimeter-waves and electro-optical (EO) sensors (e.g. spatial angle resolution), and a commonality between millimeter-waves and cm-wave sensors (e.g. range and doppler resolution), while at the same time the weather penetration capability resides between that of EO and a CM-wave sensor.

It is this quasi-optical (or quasi-microwave) behaviour of devices, components and propagation which makes mm-wave sensing simultaneously attractive and problematic.

It is the unique atmospheric penetration characteristics, see Figure 2.1, and the substantial progress in the development of millimeter-wave components and solid state devices which promise high reliability and low cost which have urged the interest in this field.

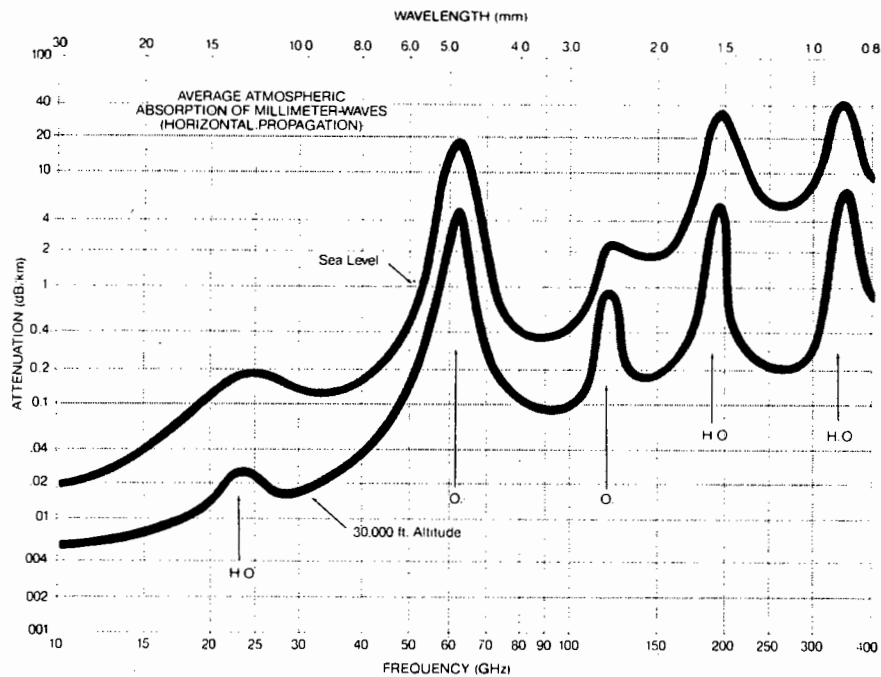


FIGURE 2.1 : Spectrum of atmospheric attenuation versus frequency.

From Figure 2.1 it can be seen that there are millimeter-wave windows at 35, 94, 140 and 220 GHz.

The requirements of a sensor includes adverse weather operation, small size, low cost and high effectiveness.

Radar sensors operating in the millimeter-wave band (30-300 GHz) provide the promise of penetrating bad weather such as fog, rain, smoke and dust. In some cases the mm-wave sensor appears to provide the practical solution to the scenarios outlined above. Advances in solid-state components, coupled with a civilian need, is enough driving force for the design of a high resolution mm-wave radar.

2.2 THE PROS AND CONS OF MILLIMETER-WAVES

There are various advantages as well as disadvantages of millimeter-waves [4-6]. It is of great importance that the system designer takes note of it and bears it in mind when designing a system.

2.2.1 Advantages of Millimeter-Waves

- 1) Relatively low propagation losses at 94 GHz (0,4 dB/km, Figure 2.1).
- 2) Reasonable resolution can be obtained with small diameter antenna apertures (e.g. a 150 mm diameter antenna has a 1,46° beamwidth compared to 14° at 9 GHz; $\theta = \lambda/D$).
- 3) Obstacles have significant radar cross section at mm-wave frequencies.
- 4) Solid-state mm-wave component technology provides a mature base for advanced developments up to 94 GHz.
- 5) The shorter wavelengths make it possible to reduce the size of components and systems considerably.
- 6) At each atmospheric window in the mm-wave band, the bandwidths are extremely large, e.g. at 94 GHz the bandwidth is 35 GHz. This means that there are more frequencies that can be used which make countermeasures more difficult and also increase the immunity to interference.

2.2.2 Disadvantages of Millimeter-Waves

- 1) Relatively short target detection range (2 km).
- 2) Poor target resolution as compared with electro-optical seekers, but much better than cm-wave seekers.
- 3) Relatively scarce clutter/target data base.
- 4) Major investment in manufacturing techniques to bring production cost to an acceptable level.
- 5) The narrow beamwidths make it difficult for millimeter-wave systems to search and acquire targets.

Table 2.1 gives a comparison [6] of the various advantages and disadvantages of different techniques.

TABLE 2.1 : Comparison of mm-waves, cm-waves and electro-optical techniques.

MM-WAVE CHARACTERISTIC	FREQUENCY		
	MICROWAVE	MM-WAVE	ELECTRO-OPTIC
Range in fog, clouds and smoke	Excellent	Good	Poor
Components	Large, rugged	Small, rugged	Smaller, fragile
Aperture	Large antenna	Small antenna	Very small aperture
Mode	Active only (radar)	Active or passive	Mostly passive
Imaging	Impractical	Possible	Excellent
Atmosphere	Atmospheric transparent	Good spectral responses	Fog and clouds opaque

Figure 2.2 shows the effect of rain and fog on mm-waves [6].

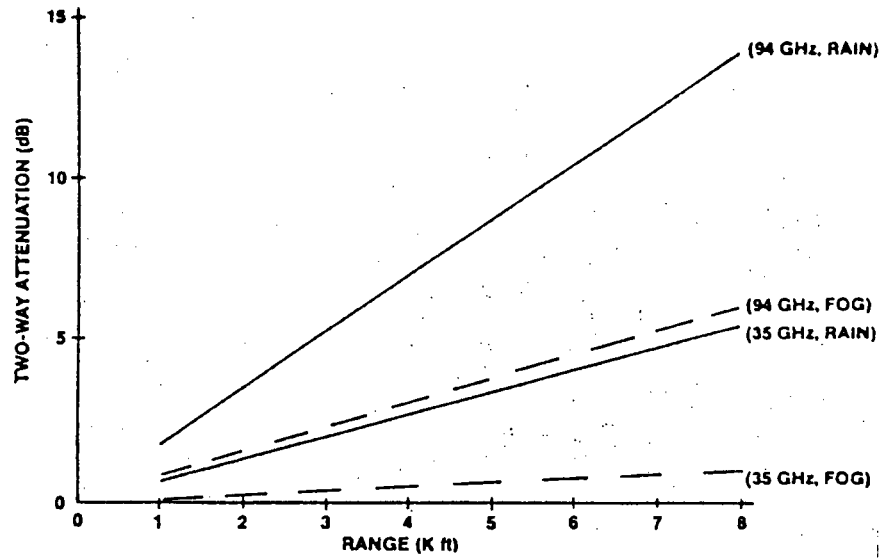


FIGURE 2.2 : Rain and fog effects on mm-waves.

2.3 TECHNICAL SPECIFICATIONS

The technical specification of a sensor can be divided into two categories. The specifications of the carrier system and the specifications of the sensor head itself.

2.3.1 Carrier System Specifications

In order to consider the optimum acquisition area, some information of the carrier system must be known.

There is a certain minimum range that the carrier system can be from the obstacle in which the sensor can tell it to pull a maximum manoeuvre, this is called the end game manoeuvrability [4]. The vehicle time constant (t) limits the response to such a command. The following formula is used:

$$d = \frac{1}{2} at^2 \quad (2.1)$$

where d = distance vehicle can move
 a = acceleration (e.g. 0,2 g)
 t = time constant (e.g. 1 s)

By substituting these values into (2.1), the distance is obtained as:

$$d = 0,98 \text{ m}$$

Therefore, if a sensor is detecting a transmission line, the maximum end manoeuvre cannot prevent a collision if applied in the last second.

The distance of the vehicle one second before it will collide with a transmission line, with a velocity of 80 m/s, is given by:

$$d = vt \quad (2.2)$$

$$= 80 \text{ m}$$

Thus, if the receiver burns out 80 m or less from the transmission line, the vehicle will still collide with the obstacle.

The mm-wave sensor is usually mounted in a helicopter. The illustration [4] in Figure 2.3 shows the vehicle is on a shallow 20° depression angle trajectory towards the obstacle, and use a search algorithm that is referred to as a "push-broom search".

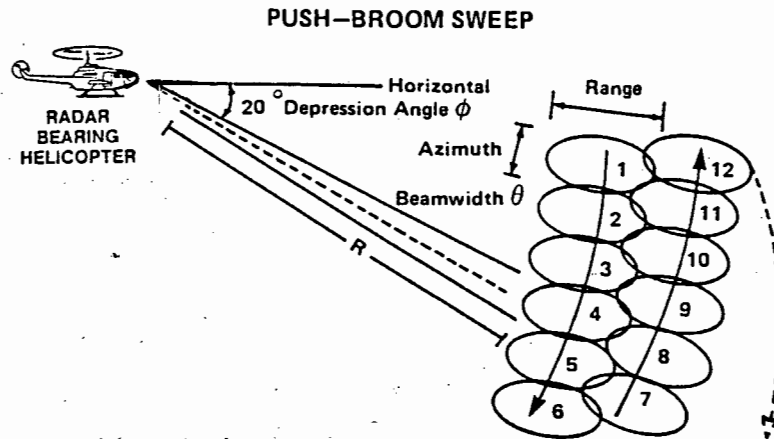


FIGURE 2.3 : Search algorithm of sensor.

This algorithm is very practical, since only the azimuth pointing errors are critical in the search mode. Elevation errors will not have a significant effect on the search because the vehicle will sweep the ground in front of it as it flies towards the obstacle.

The sensor pattern results from a pencil beam that scans from side to side (azimuth) as the vehicle flies towards the obstacle. The dimensions of a single beam on the ground is defined by:

$$R\theta \quad \text{in azimuth} \quad (2.3)$$

and $\frac{R\theta}{\sin\phi}$ in range (2.4)

and the clutter is illuminated elliptically [4] as shown in Figure 2.4.

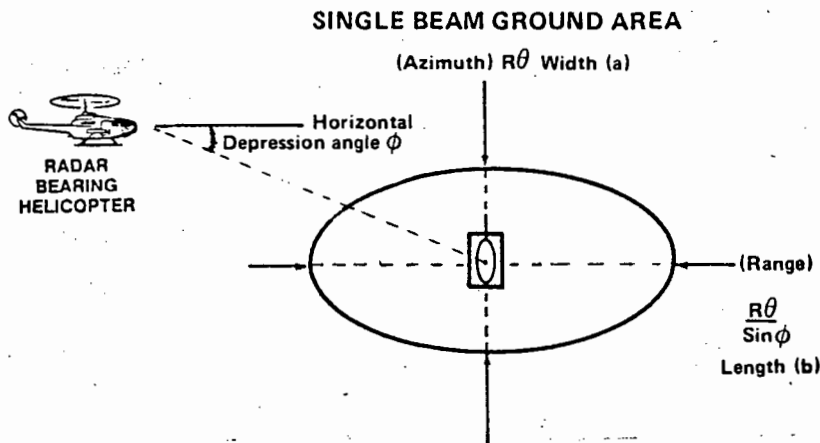


FIGURE 2.4 : Single beam area on ground.

Assume that the vehicle is 2 km from the obstacle, and the beamwidth is 1.46° then the axis of the ellipse are, using equations (2.3) and (2.4)

$$a = 51 \text{ m}$$

$$b = 149 \text{ m}$$

and the total area illuminated is

$$\begin{aligned} A &= \pi \frac{a}{2} \cdot \frac{b}{2} \text{ m}^2 \\ &= 5\,968 \text{ m}^2 \end{aligned} \quad (2.5)$$

It is obvious that the wider the beamwidth, the larger the area that can be illuminated per search interval, and fewer beams are necessary to scan the azimuth width on the ground.

Assume that the obstacle is a high voltage transmission line with a diameter of 10 mm, then the reflected area is 30 m^2 , which is a reasonable value [7] to use for the calculations. It might happen that smaller and larger reflections are measured, but the output of the receiver will be integrated over a number of pulses, so therefore the average value is used.

Because ground clutter is not as good a reflector as metal, assume that the cross section is -20 dB/m^2 . Therefore, the relative cross section area is $59,7 \text{ m}^2$. This is much larger than the cross section of the obstacle.

The signal to clutter ratio is:

$$\begin{aligned} \frac{S}{C} &= \frac{30}{59,7} \\ &= -3 \text{ dB} \end{aligned}$$

To achieve a more desirable, positive S/C, three options are possible:

- 1) Make the beamwidth narrower.
- 2) Make the resolution cell smaller.
- 3) Make the depression angle larger.

It must be remembered that the sensor is a wide open receiver and will gather energy from the clutter reflections as well as obstacle reflections, and the output would be the sum of the two per resolution cell.

The simplest solution is to make the beamwidth narrower, but since the diameter of the antenna is fixed, as well as the frequency, this option is out.

The next option is to make the depression angle larger, but even if the angle $\phi = 90^\circ$, the S/C = +1,4 dB which is only slightly better.

The last solution is to reduce the length of the resolution cell, this can be done by cutting the beam into narrow strips, using a radar range gate, i.e. to sample the return signal at finite time intervals. The narrow strips on the ground resulting from the range-gating technique [4] are illustrated in Figure 2.5.

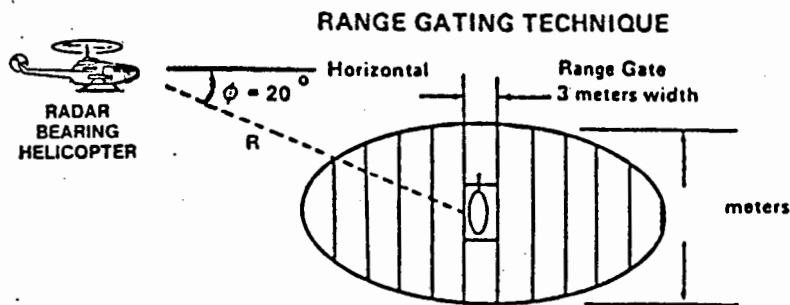


FIGURE 2.5 : Range gating technique.

The minimum range gate is taken as 3 m which is still much larger than the diameter of the high voltage transmission line.

Therefore, the clutter cross section is

$$\begin{aligned}\sigma_c &= [(3 \times 51) - 30 \text{ m}^2]/100 \\ &= 1,23 \text{ m}^2\end{aligned}$$

which gives

$$\frac{S}{C} = 13,9 \text{ dB}$$

These calculations show that the S/C is now acceptable, which means that the obstacle will appear much larger than the surrounding natural clutter. Additional signal processing is needed to raise the S/C even higher.

The next question to be asked is how many independent beamwidths can be tolerated as the beam swept from side to side to ensure complete ground coverage? Overlap of the beams are allowed, as long as there are no "holes" in the coverage, see Figure 2.3.

Using the previous assumed vehicle velocity of 80 m/s, and equation (2.2), it will take 1,86 s for the beam to move the length b of the ellipse as the vehicle comes closer to the target. Let one azimuth sweep be completed in half this time, and assume a practical mechanical scan rate of 50°/s. Thus the beam can scan 46,56 in azimuth mechanically, which means $\frac{46,56^\circ}{1,46^\circ} \approx 32$ beamwidth, and the width of one scan is 1 632 m.

This width will shrink slightly as the range decreases, but this calculation is a fairly good estimate of the azimuth range that the radar can sweep and still be able to detect an obstacle.

The system's ability to measure doppler (moving targets) is also very important, and is greatly enhanced by mm-waves. For example, let the transmission line swing forward and backwards at 1 km/hr, then the doppler shift from [8] is

$$f_d = \frac{2v_r}{\lambda} \quad (2.6)$$

$$= 87 \text{ Hz}$$

while at 9,4 GHz, the doppler shift will be 8,7 Hz. The higher the doppler frequency, the easier the filtering in the signal processing.

2.3.2 Circuit Description of Coherent Sensor

Figure 2.6 shows a block diagram of the sensor receiver and transmitter. The operation of the circuit can be classified as a superhetrodyne principle with a double coherent conversion. This double coherent conversion is a unique idea and no reference could be found in literature on it.

The elements of the transceiver, as shown in Figure 2.6, are:

- 1) Crystal Synthesized Reference Oscillator.
- 2) Bias Injection Locked Stable Master Oscillator (STAMO).
- 3) Impatt Pulsed Injection Locked Oscillator chain (PILO).
- 4) RF Coherent Local Oscillator (RF, Coho).
- 5) IF Coherent Local Oscillator (IF, Coho).

All the RF and IF signals are phase locked to the crystal oscillator. The return signal from a target is mixed down to a 300 MHz IF signal before it is detected. As can be seen from Figure 2.6, the RF and IF Coho signals are coherent signals obtained from the crystal reference source.

 f_1

✓

The reference signal which is required to generate the RF and IF Coho signals, is generated by a crystal oscillator operating at a nominal frequency of 300 MHz. Two output signals are available from the synthesizer; one is fed into the DC bias port of the STAMO, while the other is fed to the coherent detector.

2.3.2.2 Bias injection locked stable master oscillator (STAMO)

The 300 MHz nominal frequency from the crystal synthesizer is fed into the DC bias port of a low noise, low power (+7 dBm) 93,5 GHz Gunn source. The output of this source consists of three frequency components, i.e. f_0 and $f_0 \pm 0,300$ GHz, which is phase locked to the crystal source [9,10]. This signal is split into two signals and are fed to the PILO and RF Coho which are tuned to 93,5 and 93,80 GHz respectively.

The signal that is fed to the PILO is switched by a PIN-diode switch. This switch is turned on about 200 ns before the high power source and turned off about 20 ns after the high power source is switched off. The reason for doing this is to prevent the leakage signal from saturating the logarithmic amplifier.

2.3.2.3 Pulsed impatt injection locked oscillator chain (PILO)

The PILO amplifier chain consists of a medium power (+18 dBm) CW Impatt oscillator, that is used in the pulsed mode, and a high power (5 W) Pulsed Impatt source.

The CW source is biased to a point just before switch-on. This ensures that the diode is already heated to a thermal point which results in a virtually chirpless output signal when it is switched on by the pulse modulator. This source is switched on about 20 ns before the high power source, and is switched off about 20 ns after the high power source is switched off. This ensures that the medium power source has stabilized and no transients exist.

The output of the PILO is fed via the circulator to an aplanatic lens antenna.

The PILO is injection-locked with the 94 GHz signal from the STAMO.

2.3.2.4 RF coherent local oscillator (RF Coho)

The 93,80 GHz signal from the STAMO is used to injection-lock the RF Coho. This ensures that the phase of the transmit signal is preserved in the RF Coho signal. This Coho is a low noise, low power (+7 dBm) Gunn oscillator tuned to the higher sideband of the injected frequency. This signal is used together with the return signal from a target to mix down to an IF signal of 300 MHz.

2.3.2.5 Coherent local oscillator source at VHF (IF Coho)

The 300 MHz IF signal is amplified and fed to the coherent detector. The phase of the transmit signal is also preserved in the IF Coho which is used as the reference signal for the coherent detector.

The log amplifier also has a video output which is used to determine the target range.

The output of the coherent detector is fed to the signal processing circuitry which is beyond the scope of this thesis.

2.3.3 Sensor Specifications

In Section 2.2.1 the antenna diameter was taken as 150 mm. The gain and beamwidth is a function of frequency and is given nominally as:

$$\theta = \lambda/D \quad (2.7)$$

$$G = \frac{4\pi A_e}{\lambda^2} \quad (2.8)$$

where A_e = effective antenna area.

Thus the higher the frequency, the narrower the beamwidth and the higher the gain. At 93,5 GHz the gain is calculated as:

$$G = 42 \text{ dBi}$$

Assume that the following sensor specification can be met:

Beamwidth	:	1,46°
Antenna diameter	:	150 mm
Antenna gain	:	42 dB
Transmitter power (peak)	:	5 W
Signal format	:	Pulsed
Pulse repetition rate	:	50 kHz
Signal processing	:	Range gated - Log video Square law detection Video integration and MTI
Pulse length	:	40 ns

2.3.3.1 Minimum detectable signal and maximum acquisition range

The standard radar equation from [4] is:

$$P_r = P_t + 2G_t + \sigma - L + 10 \log_{10} \left[\frac{\lambda^2}{(4\pi)^3 R^4} \right] \quad (2.9)$$

P_r = received power in dBm

P_t = transmit power in dBm

σ = radar cross section of target [i.e. $30 \text{ m}^2 = 14,7 \text{ dB}$].

L = plumbing and atmospheric losses (2,5 + 0,4 dB/km)

f = 93,5 GHz

The thermal noise of an ideal receiver (i.e. $F = 0$ dB) is given by [4] as:

$$N = kTB$$

where $k = 1,38 \times 10^{-23}$ J/K

$T = 300$ K

$B =$ bandwidth (16 MHz for 60 ns pulsewidth)

therefore $N = -101,6$ dBm

Assume that the receiver has a noise figure of

$$F = 10 \text{ dB}$$

then the received signal should exceed

$$P_r \geq -91,6 \text{ dBm}$$

Substituting this into (2.9) gives a maximum acquisition range of

$$R = 2\,723 \text{ m}$$

2.4 CONCLUSION

With the previous calculations it seems possible for a helicopter equipped with a mm-wave sensor to detect an obstacle and avoid an accident.

CHAPTER 3

THEORETICAL AND PRACTICAL DETERMINATION OF THE COHERENCY OF MILLIMETER-WAVE PULSED SOURCES

3.1 INTRODUCTION

This section describes the technique of injection locking pulsed sources, and a mathematical analysis is performed where a unique method is derived with which the coherency of injection locked pulsed sources can be determined [11]. Nowhere in the literature could any reference to such a method be found.

Many modern radars work on the principle that they transmit and receive RF pulse energy instead of continuous waves (cw). A large portion of these radars are coherent type radars. That means that the phase of the transmit signal is preserved in the reference signal. The coherency is measured as the phase error in degrees between the transmit signal and the reference signal. Normally a pulsed oscillator "chirps" which means a variation in frequency and phase between the two sources.

3.2 INJECTION LOCKING

The concept of injection locking a noisy RF source to a "clean" isolated source to improve the FM stability is not new. Van der Pol [12,13] derived mathematical expressions to describe under what conditions injection locking could be expected. The theory was refined and extended by [14,15] and represented in a more usable manner with reference to microwave oscillators in particular. Similar work has been reported with respect to solid state microwave oscillators [16,17].

The principle behind injection locking is to force a source with high phase noise to synchronise with a lower power level stable source by injecting the output from the low phase noise signal into the input port of the noisy oscillator. Provided sufficient isolation exists [18] between the two sources, so that the clean source is not affected by the noisy source RF output, the latter is forced to synchronise its output signal with that of the injector.

The circuits to be used in the argument are presented in Section 3.3.

3.3 THE FREE RUNNING OSCILLATOR

Figure 3.1 shows a general solid state oscillator circuit with its equivalent circuit [12,13]. $Z(\omega)$ is the circuit impedance seen from the device and $-Z(A)$ is the device impedance. The former is highly frequency dependent only and is a passive network and the latter dependent principally on the RF current amplitude flowing in the external circuit.

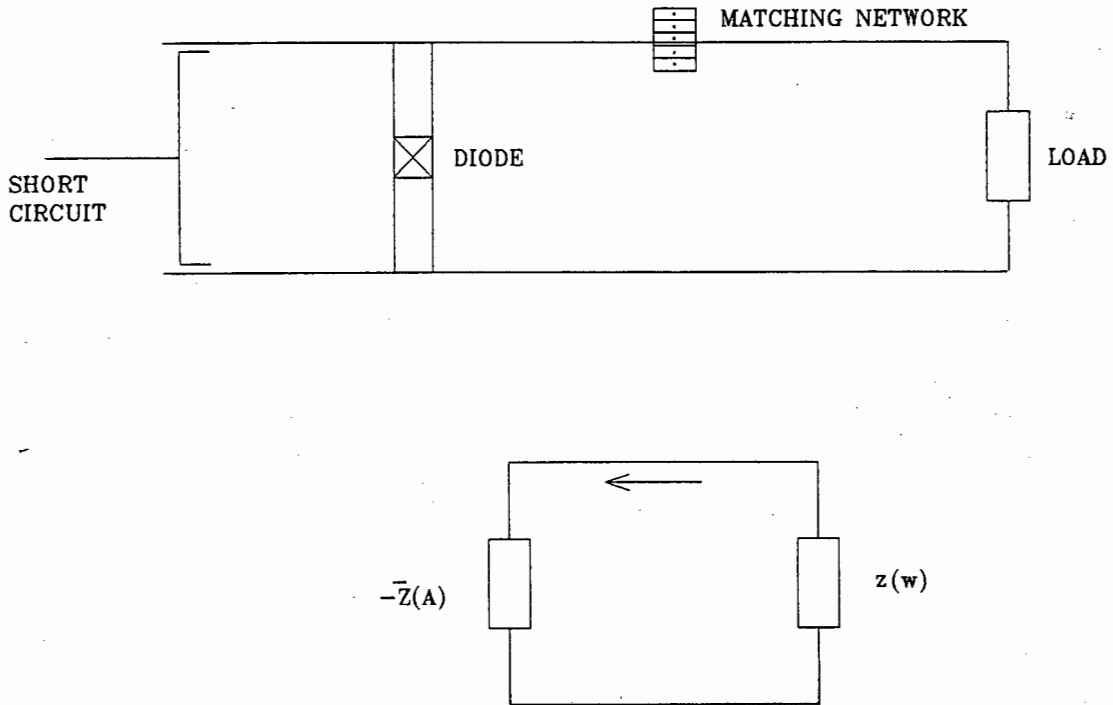


FIGURE 3.1 : Schematic diagram of a noise free solid state source with equivalent circuit.

For free running operation the following equation must be satisfied:

$$[Z(\omega) - \bar{Z}(A)] I = 0 \quad (3.1)$$

and for non-zero I the operating point is fixed at the intersections of the $Z(\omega)$ and $\bar{Z}(A)$ loci [12,13] see Figure 3.2.

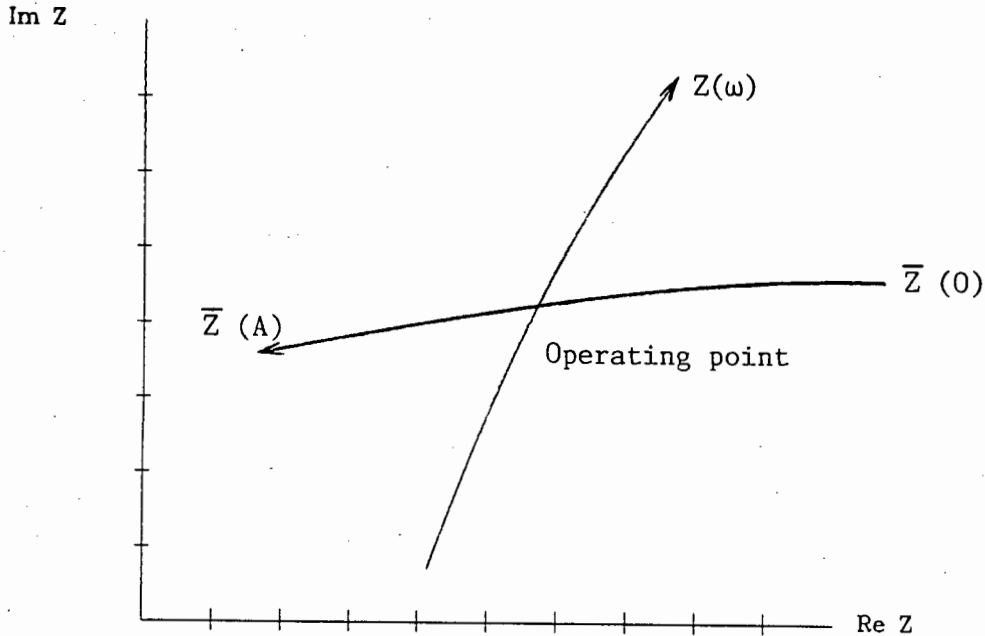


FIGURE 3.2 : Complex plane plot of impedance ($Z(\omega)$), and device ($Z(A)$), loci.

$\bar{Z}(A)$ is the complex conjugate of $Z(A)$.

If noise is present equation (3.1) is modified to

$$[Z(\omega) - \bar{Z}(A)] I = e \quad (3.2)$$

where e is random in both magnitude and phase. The complex plane plot in Figure 3.2 is modified to that in Figure 3.3, where the operating point is now no longer defined by a single fixed intersection point between the two loci, but is instead represented by the random, time varying intersection produced by a horizontal and vertical translation of the $\bar{Z}(A)$ locus with the $Z(\omega)$ locus.

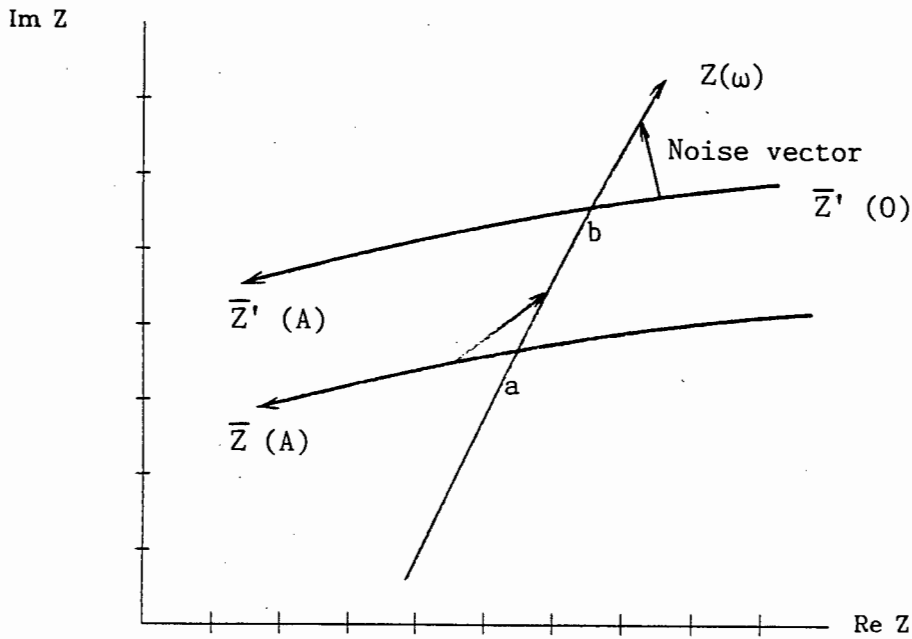


FIGURE 3.3 : Effect of noise vector on operating point. Latter can lie between "a" and "b".

$\bar{Z}(0)$, $\bar{Z}(A)$, $\bar{Z}'(A)$ and $\bar{Z}'(0)$ are the impedance loci as a function of operating point, and is plotted on a complex impedance scale.

3.4 THE LOCKED CW OSCILLATOR

If a highly stable RF voltage "E" is now added to equation (3.2) with amplitude very much greater than the noise generator "e", equation (3.2) now becomes

$$[Z(\omega) - \bar{Z}(A)] I = e + E \quad (3.3)$$

The RF current I may now be written as

$$I = A_0 e^{j\phi_0} \quad (3.4)$$

Where A_0 and ϕ_0 are near constants.

The complex plane plot in Figure 3.4 now shows the locking vector overriding the noise vector and forcing the operating point away from the intersection of the two loci to point A. The frequency of the locking signal defines the point on $Z(\omega_1)$ locus fixed by the arrow head and as ω_1 is varied within the locking range ω_1 to ω_2 shown in Figure 3.4 the operating point moves from the intersection 0 out to A' and back to 0. If ω is taken outside the limits ω_1 and ω_2 , in order to preserve lock, the injected vector amplitude must be increased to satisfy equation (3.3).

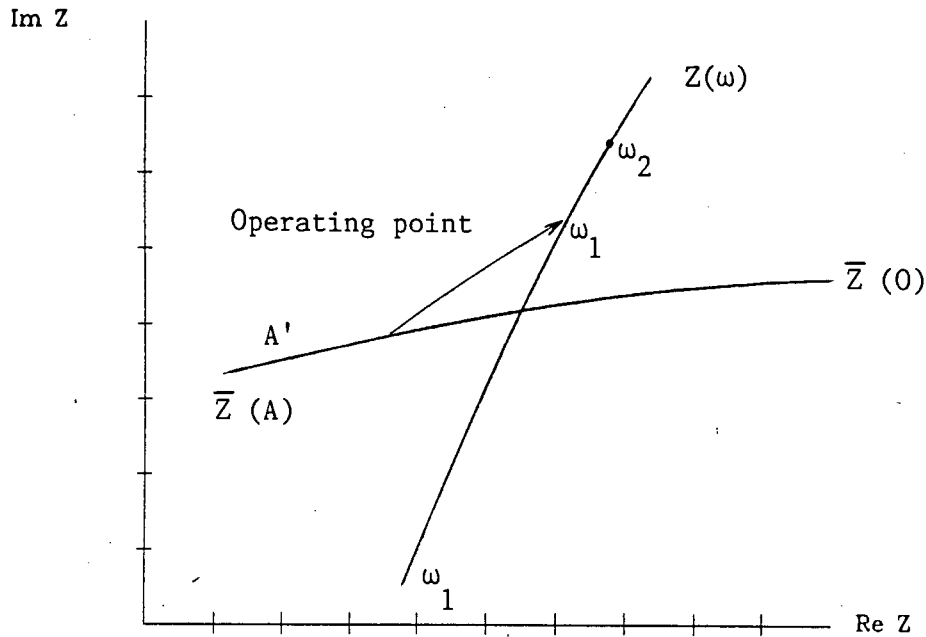


FIGURE 3.4: Effect of locking vector on operating point.

Although the oscillator noise is still present, the restraint on movement of $\bar{Z}(A)$ by the locking vector limits the vibration of the device line $\bar{Z}(A)$. Hence the position of point A with respect to ω_1 produces a change in orientation of the locking vector about the point ω_1 . This change in phase is still random in time and represents a random rate of change of phase of the RF current in the circuit. As the rate of change of phase is frequency, residual FM is still present but drastically reduced from the free running case.

For a fixed displacement of the $\bar{Z}(A)$ locus, the angular orientation and hence the residual FM is minimum for ω_1 at or near the locking range centre.

3.5 THE LOCKED PULSED OSCILLATOR

So far the discussion has been limited to CW steady state locked sources in describing the synchronising mechanism within the locked source, but no mention has been made of synchronisation conditions at, and following the point of switch-on from the zero RF amplitude state.

In the case of the pulsed source the RF voltage envelope across the active device may be represented by

$$V(t) = V_0 e^{\alpha t} e^{j(\omega t + \phi)} = V e^{j(\omega t + \phi)} \quad (3.5)$$

ϕ is phase difference between injected current and output voltage.

ω is instantaneous oscillation frequency

α describes pulse envelope $\alpha > 0$ on rise period

$\alpha < 0$ on fall period

$\alpha = 0$ on pulse pedestal.

The time derivatives of the real part of $V(t)$ is

$$\frac{d}{dt} \text{Re}(V(t)) = \text{Re}\left(j\left(\omega + \frac{d\phi}{dt}\right) + \frac{1}{V} \frac{dV}{dt}\right)V(t) \quad (3.6)$$

$$\equiv \text{Re}(j\omega V(t)) \quad (3.7)$$

Kurokawa [16,17] and Pavlidis [18] argue that if $\frac{d\phi}{dt} + \frac{1}{V} \frac{dV}{dt}$ are much smaller than ω then equation (3.6) should replace equation (3.7) in the general circuit equation. After manipulation and expansion by Taylor Series, it is possible to obtain information on the phase transients at and during switch on [18], and also to show that after some time on the pulse pedestal, the transient settles to a constant

value and the pulse source is again synchronised to the injector, Figure 3.5.

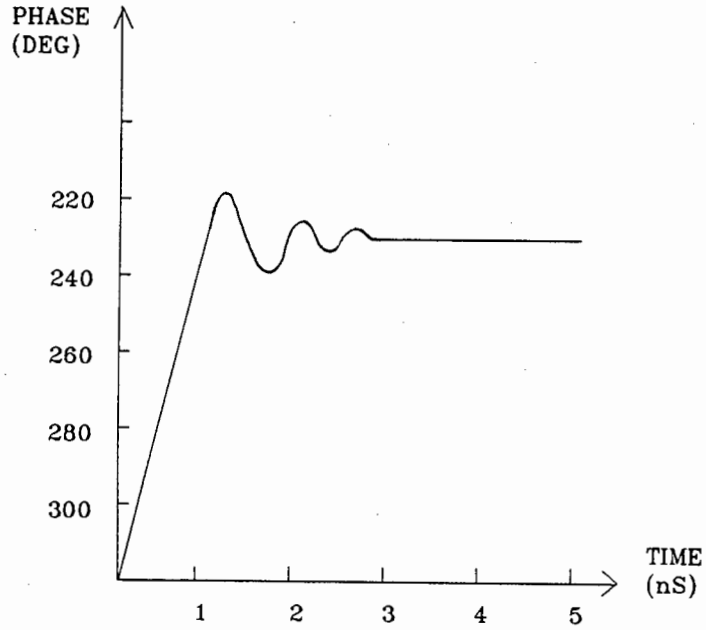


FIGURE 3.5: Phase transient on pulsed source injection locked to high stability RF injector.

3.6 THE INJECTION PRIMED SOURCE

A useful general equation derived by Adler [14] for predicting the locking range ($\omega_1 - \omega_2$) of an injection locked source is

$$\omega_1 - \omega_2 = \frac{\omega}{Q_{\text{ext}}} \sqrt{\frac{P_i}{P_o}} \quad (3.8)$$

ω is free running instantaneous frequency of the locked source

Q_{ext} is the external Q of the locked source

P_i is the injected source power

P_o is the locked source power

From equation (3.8) it can be seen that if

$$Q_{\text{ext}} \gg \sqrt{\frac{P_i}{P_o}}$$

the locking range can be a very small fraction of ω which can cause system problems with frequency tracking when temperature and frequency pulling effects on the injected and injecting sources are considered. In a condition of "no lock", i.e. ω_i is outside the locking range $\omega_1 - \omega_2$ the injected source is no longer in continuous synchronism with the injecting stable source.

It should be noted that Q_{ext} of an oscillator has two conflicting effects:

- 1) If Q_{ext} is large, $\omega_1 - \omega_2$ is small, and therefore it would appear that a small Q_{ext} is required.
- 2) If Q_{ext} is small, the oscillator stability, noise performance and frequency drift within the pulse are all very poor.

Therefore, from these arguments a high Q_{ext} is required.

In the CW case, the injected source tends to "cycle" in and out of synchronism with the injector with a pseudo random period.

In the case of the pulse source however, the amplitude of the injecting signal at the start of the pulse is large enough to stimulate synchronism of the pulsed oscillator where RF current amplitudes are small, instead of the RF pulse starting up from the instantaneous noise fluctuation. This condition is injection priming, where the start up of the RF pulse is synchronised to the injector. As RF amplitude builds up past the "lock condition", defined by equation (3.8), the oscillator "free wheels" and synchronism is lost over the remainder of the pulse period.

3.7 MEASUREMENT TECHNIQUE TO DETERMINE PHASE COHERENCY

An indication of how well a pulsed oscillator is primed or locked is given by the depth of the noise floor from the interspectral line, Figure 3.6. This is usually expressed in terms of dBc, as shown in Figure 3.6, the ratio of line to noise power and is termed peak to valley ratio (P/V ratio). Although this qualitative measurement can give a good general guide to expected sub-clutter visibility [19,20], the experimental observations and measurements on the interspectral line may only be made when the following conditions are fulfilled:

- (a) The PRF is high enough to separate the interspectral lines with their LO noise sidebands.
- (b) The duty ratio is high enough to overcome the sensitivity limitations of many spectrum analyser units from the "desensitisation factor". This is an effect resulting in the need to drastically increase IF gain and also the level of the instrument noise floor in order to display the spectrum outline.

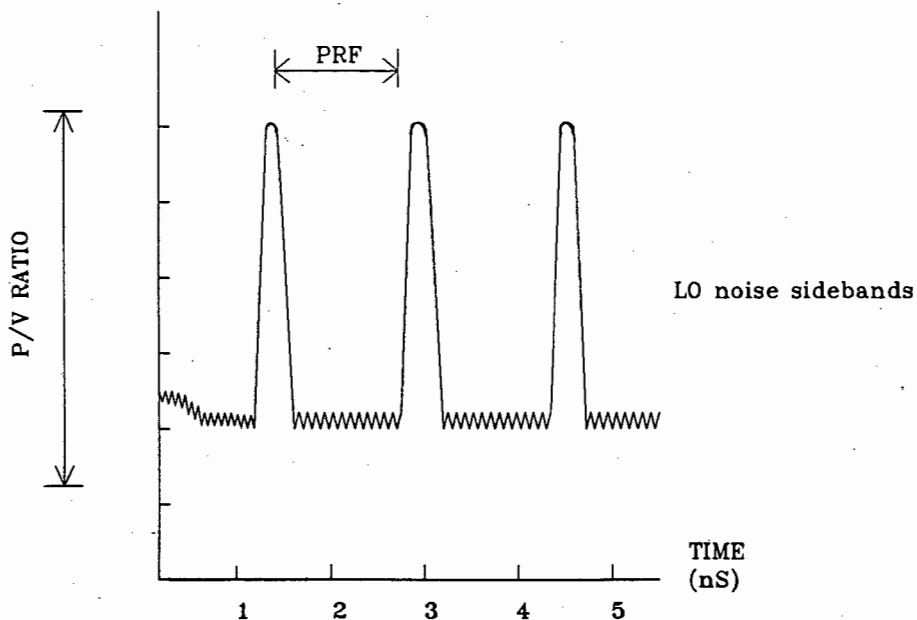


FIGURE 3.6 : Display of interspectral lines for high PRF source with injection priming.

In order to obtain quantitative measurements on transmitter stability with low duty ratios, a measurement system based on a radar's signal processing approach was used. The schematic of Figure 3.7 shows a general measurement system. The CW source acts as priming source and STALO and mixes with the pulsed output at RF in the balanced mixer. The video output is sampled in the sample/hold unit whose output during the PRF interval is constant and is a measure of the video phase at each sampling interval.

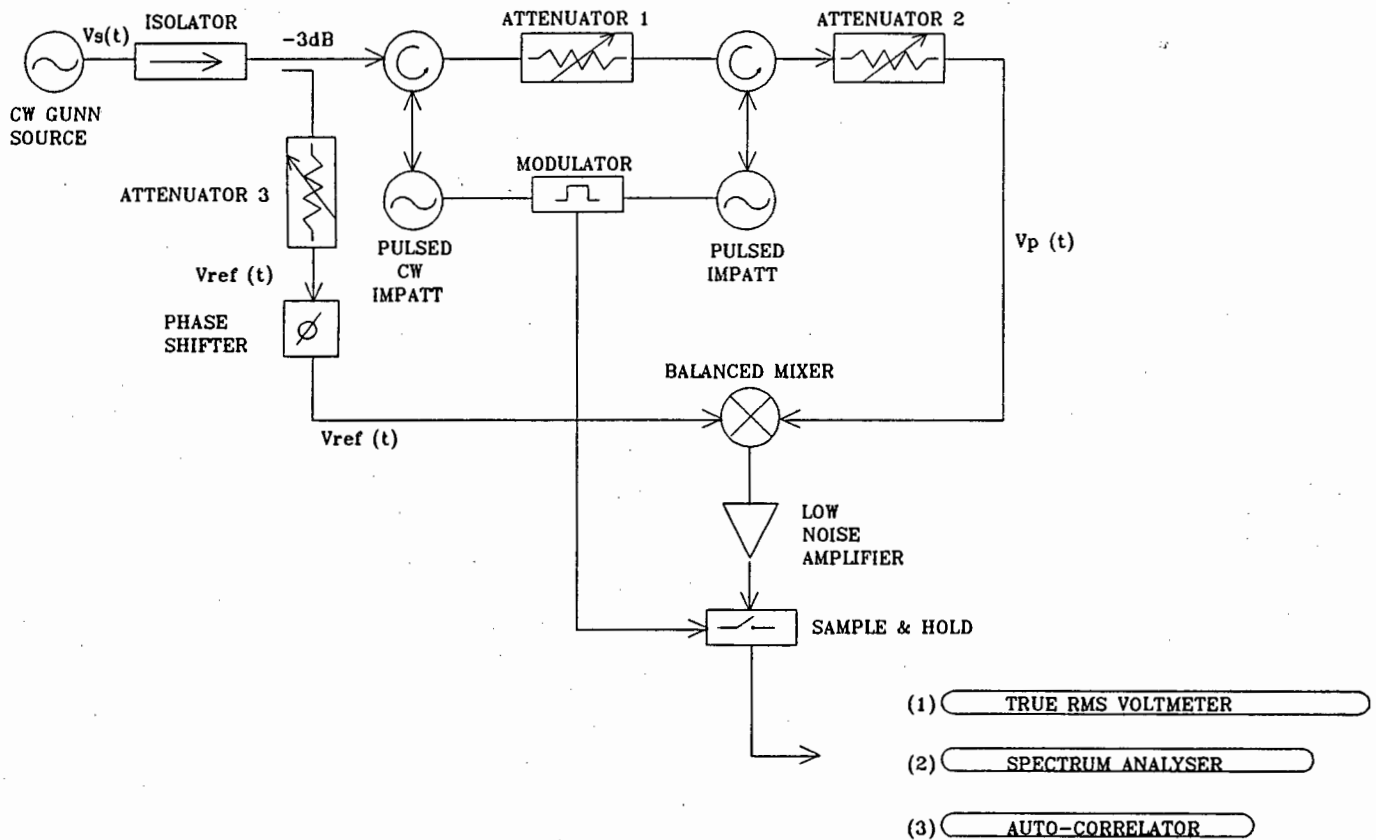


FIGURE 3.7 : Injection priming circuit for phase noise measurements.

If there is no amplitude jitter (caused by starting phase jitter in the pulsed RF) on the video pulse, Figure 3.8(a), then the sample/hold output is constant, Figure 3.8(b). If phase jitter exists due to inadequate priming/locking of the pulsed source, Figure 3.8(c), the sample/hold output varies randomly from one PRI to the next, Figure 3.8(d).

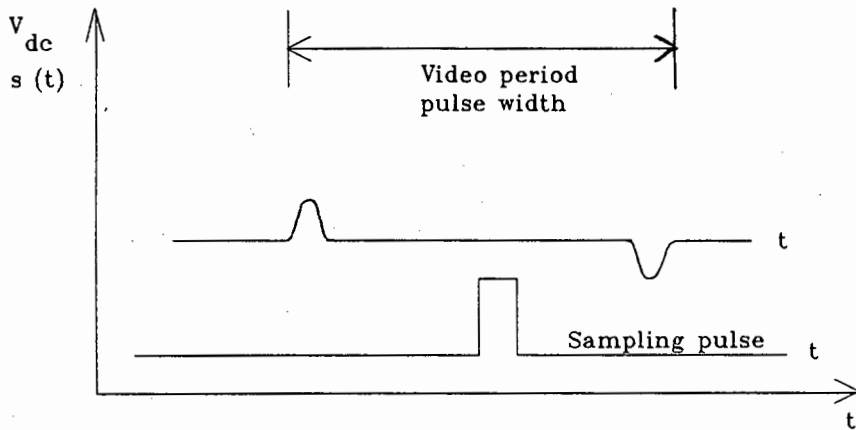


FIGURE 3.8(a) : Balanced mixer video output with locked pulsed RF source and sampling pulse.

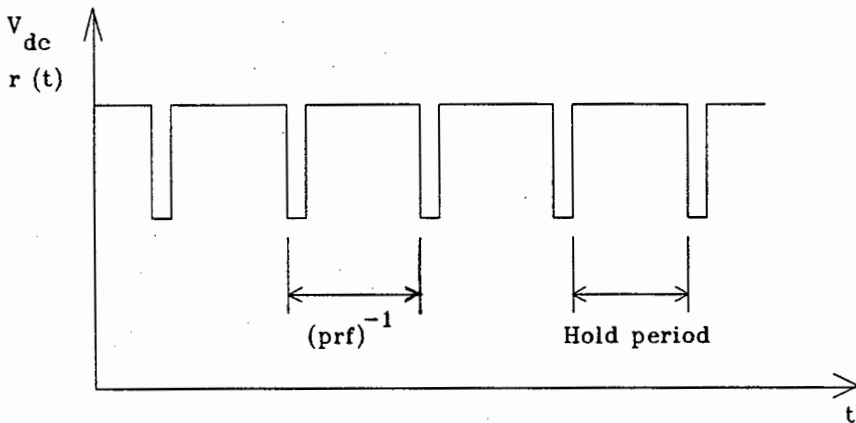


FIGURE 3.8(b) : Sample and hold output for injection locked pulsed RF source.

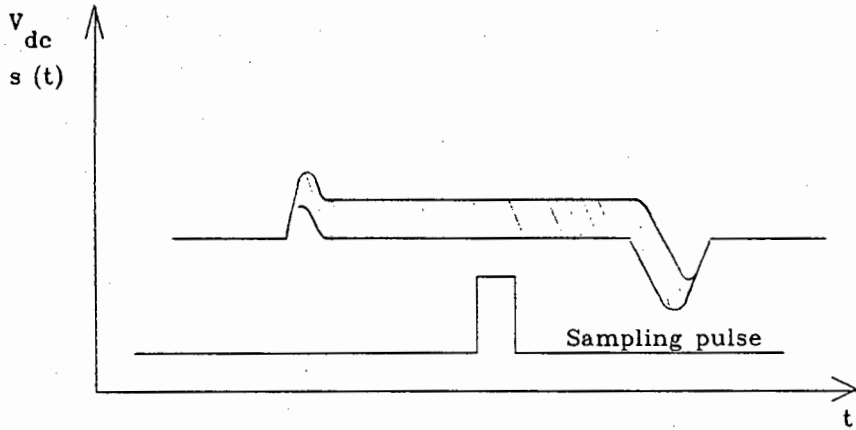


FIGURE 3.8(c) : Balanced mixer video output with primed pulsed RF source and sampling pulse. The broadening is caused by non-ideal priming.

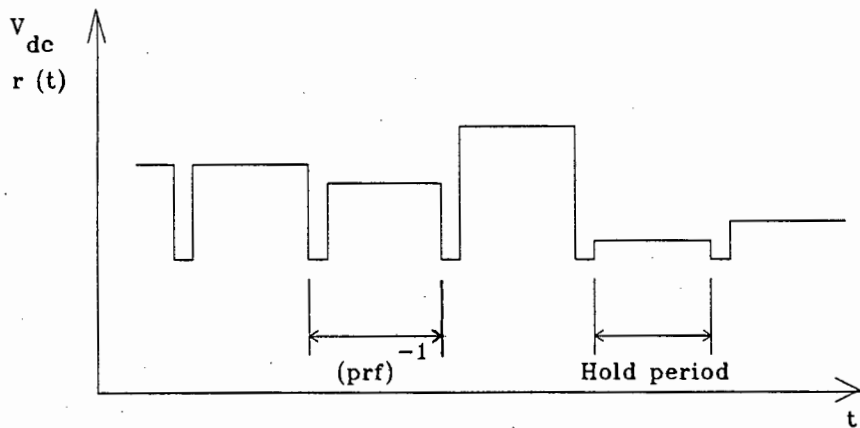


FIGURE 3.8(d) : Sample and hold output for non-ideal injection priming.

The sample/hold output may be fed into a dc to 10 kHz spectrum analyser for a qualitative measurement of the noise power spectral density or for more accurate results into an auto correlator to determine the variance by measuring the auto correlation function. This is then directly proportional to RMS phase jitter.

3.8 MEASUREMENT SET-UP FOR COHERENCY DETERMINATION

The signal of the CW Gunn source shown in Figure 3.7 is given by equation (5.1), Chapter 5 as:

$$V_s(t) = A_o \cos(\omega_r t + \phi_r(t)) \quad (3.9)$$

where A_o = signal amplitude.

ω_r = reference frequency.

$\phi_r(t)$ = randomly fluctuating phase term.

As can be seen from Figure 3.7, the signal from the Gunn source is split into two signals with a coupler (ideally -3 dB). The one signal (V_{ref}) is fed via attenuator 3 and the phase shifter to the balanced mixer. The other signal is used to injection prime the pulsed Impatt sources. The amplitude of the injection signal is controlled by attenuator 1, and the amplitude of the pulsed signal by attenuator 2. The nett effect of equal arm lengths of the phase bridge is the cancellation of the phase noise of the reference oscillator at the phase detector. The effect of the additive phase noise from the Impatt injection-locked oscillator can be observed as noise on the waveform of the phase detector output when the phase shifter is adjusted to the position at which the balanced mixer phase detector is most phase sensitive, see Figure 3.8.(c).

The CW RF bias to the mixer produced a dc offset voltage which has to be taken into account for the phase noise measurements [21] although not to such an accuracy as might be required for absolute phase measurements. Figure 3.9 shows the effect of dc offset on the video output from the mixer where two identical frequency constant amplitude signals are applied to the R & L ports of the mixer. The resulting dc output is proportional to the cosine of the phase difference and appears at the mixer I port.

$$V_{IF} = V_m \cos(\Delta\phi + \pi) \quad (3.10)$$

where $\Delta\phi$ is the absolute phase difference between the two signals [22].

Figure 3.9 indicates the error in absolute phase measurements if V_{DC} , the offset voltage is not taken into account.

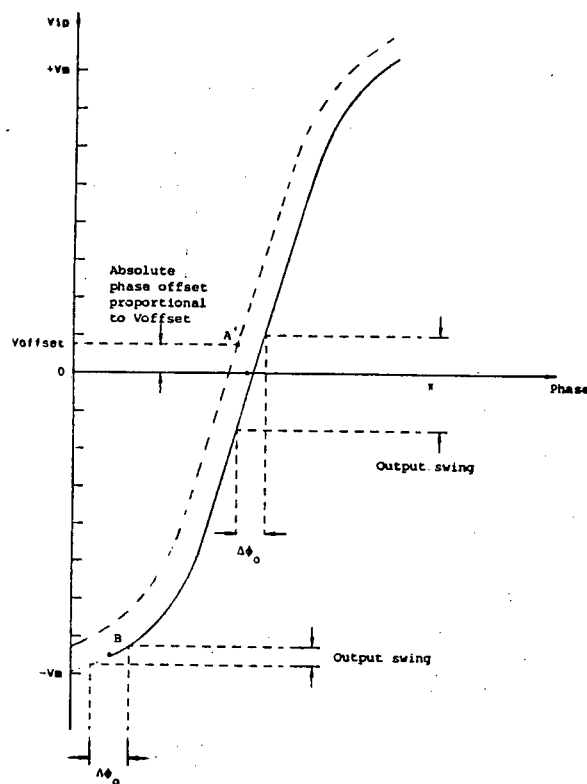


FIGURE 3.9 : Response curve for ideal balanced mixer used as a phase detector.

However, when measuring phase jitter it is sufficient to ensure operation is confined to the almost linear portion of the cosine curve, to ensure maximum sensitivity at the I port to the phase swing about the centre point. A false low reading of phase jitter would be obtained if the mixer were biased at B instead of A, the 90° point.

3.9 MATHEMATICAL THEORY BEHIND PHASE COHERENT MEASUREMENTS

Let the signal via the reference line in Figure 3.7 to the balanced mixer be:

$$V_{\text{ref}}'(t) = l_0 l_3 A_o \cos(\omega_r t + \Delta\theta + \phi_r) \quad (3.11)$$

where l_0 = loss of the coupler.

l_3 = attenuation setting on attenuator 3.

$\Delta\theta$ = phase shift introduced by phase shifter.

and the pulsed signal is given by:

$$V_p(t) = l_0 l_1 l_2 B_o \rho(t) \cos[\omega_p t + \phi_p(t)] \quad (3.12)$$

where $B_o \rho(t)$ = pulsed signal amplitude.

$\rho(t) = 1$ for $nT \leq t < nT + d$ ($-\infty < n < +\infty$).

$= 0$ elsewhere.

d = pulse width.

ω_p = pulsed signal frequency.

$\phi_p(t)$ = pulsed oscillator phase at the start of pulse n (i.e. at $t = nT$).

T = period of the pulse train.

l_2 = attenuation setting on attenuator 2.

Initially it is assumed that no locking occurs, and that there is no coherency between the reference signal and the pulsed signal.

The output of the phase detector is expressed as

$$\begin{aligned}
 V_m(t) = & \frac{1}{2} l_0^2 l_1 l_2 l_3 l_4 A_0 B_0 \rho(t) [\cos[(\omega_r - \omega_p)t + \Delta\theta + \phi_r - \phi_p(t)] \\
 & + \cos[(\omega_r + \omega_p)t + \Delta\theta + \phi_r + \phi_p(t)]] \quad (3.13)
 \end{aligned}$$

where l_4 = conversion loss of the mixer.

After low pass filtering and amplification, the input to the sample and hold circuit may be written as:

$$V_o(t) = k \cos[(\omega_r - \omega_p)(t - nT) + \Delta\theta + \phi_d(t)] \quad (3.14)$$

$$\text{where } k = \frac{1}{2} g_0 l_0^2 l_1 l_2 l_3 l_4 A_0 B_0 \rho(t)$$

g_0 = gain of low noise amplifier.

$\phi_d = \phi_r - \phi_p$ is the instantaneous phase difference of the pulsed oscillator relative to the reference oscillator at the start of pulse n (i.e. at $t = nT$).

When operating the mixer as a phase detector, the signals must be at the same frequency and 90° out of phase, therefore

$$\omega_r = \omega_p$$

If it is assumed that a degree of locking primary exists, then $\phi_d(t)$ is a random variable with a small variation around a mean value $\overline{\phi_d(t)}$. To measure the variance of $\phi_d(t)$ the phase shifter is adjusted until: $\Delta\theta - \overline{\phi_d(t)} = -\frac{\pi}{2}$, and sample at $t = nT$, (in practice just after that to avoid transients).

Substituting this into equation (3.14), the output of the sample and hold circuit is

$$V_n(t) = k \cos\left[-\frac{\pi}{2} + \overline{\phi_d(t)} - \phi_d(t)\right] \quad (3.15)$$

$$V_n(t) = k \sin[\phi_d(t) - \overline{\phi_d(t)}] \quad (3.16)$$

$$V_n(t) = k [\phi_d(t) - \overline{\phi_d(t)}] \text{ for small } \phi_d(t) - \overline{\phi_d(t)} \quad (3.17)$$

The waveform of $V_n(t)$ is shown in Figure 3.8 for the locked and loosely primed conditions.

The coherency of two signals i.e. a reference - and an injection locked signal can be specified in terms of the autocorrelation function. The autocorrelation function is a mathematical operation whereby a signal is compared with itself, but displaced in time. By definition the autocorrelation process will tell us about the time variation of the phase $\phi(t)$ in an integrated or averaged sense. The autocorrelation of the phase function is defined from [23,24] as:

$$R_\phi(\tau) = \lim_{T \rightarrow \infty} \frac{1}{T} \int_{-T/2}^{T/2} \phi(t) \phi(t + \tau) dt \quad (3.18)$$

for an ergodic process.

If $|R_\phi(\tau)|$ is large, then $\phi(t - \tau)$ is very similar to $\phi(t)$ for that particular value of τ , or conversely if $R_\phi(\tau) = 0$ for some value of τ , then $\phi(t)$ and $\phi(t + \tau)$ are orthogonal. A simple way to see how the autocorrelation process works is explained intuitively. Take the function $\phi(t)$ and let it be fixed at some specific time t , then take the same function and shift it in time with time τ , time τ must be large enough so that there is no overlapping of the two functions $\phi(t)$ and $\phi(t + \tau)$. Now start making τ smaller, each time multiply $\phi(t)$ and $\phi(t + \tau)$ and compute the area under the overlapping part. Repeat this up to $\tau = 0$, and for negative τ , until $\phi(t + \tau)$ and $\phi(t)$ does not overlap anymore. The maximum value of $R_\phi(\tau)$ should be obtained at $\tau = 0$ if the two signals are very similar.

To determine the coherency, it is necessary to define the mean value $\overline{\phi(t)}$, the mean square value $\overline{\phi^2(t)}$ and the mean squared value $\overline{\phi(t)^2}$.

The mean value $\overline{\phi(t)}$ is also called the average value and is expressed as $\langle \phi(t) \rangle$ from [23]. There are two different kinds of averages involved:

- 1) the first one is to take successive measurements of one sample function and find the time averages $\langle \phi_1(t) \rangle$;
- 2) and the second is to take a lot of sample functions at a particular time t , and find the ensemble averages $E[\phi(t)]$.

Clearly the ensemble averages are the same as the statistical average at $t = t_1$.

Due to the fact that the random signal in communication systems come from ergodic processes, meaning that the time and ensemble averages are identical, i.e.

$$E[\phi(t)] = \langle \phi(t) \rangle \triangleq \lim_{T \rightarrow \infty} \frac{1}{T} \int_{-T/2}^{T/2} \phi(t) dt \quad (3.19)$$

an ergodic process is also stationary, since ensemble averages are independent of the time of observation.

In the remainder of this section we shall deal exclusively with random signals that are sample functions of ergodic processes. Therefore, these signals are power signals, since $E[\phi^2(t)]$ does not depend on time t . The following statements can now be made:

- 1) The mean value $\overline{\phi(t)}$ is the D.C. component of the signal.
- 2) The mean squared value $\overline{\phi(t)^2}$ is the power in the D.C. component, D.C. power.
- 3) The mean square value $\overline{\phi^2(t)}$ is the total average power.
- 4) The variance $\sigma_\phi^2 = \overline{\phi^2(t)} - \overline{\phi(t)}^2$ is the power in the time varying component, A.C. power.
- 5) The standard deviation σ_ϕ is the rms value of the A.C. power.

The mean average or expected value of the phase is defined by [24], see Figure 3.10 and equation (3.19).

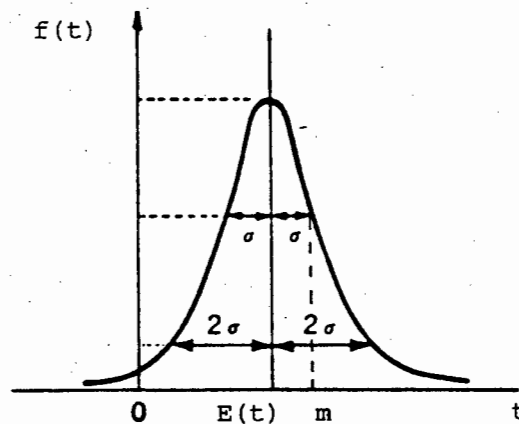


FIGURE 3.10 : The Gaussian density function.

$$E[\phi(t)] \triangleq \overline{\phi(t)} = \lim_{T \rightarrow \infty} \frac{1}{T} \int_{-T/2}^{T/2} \phi(t) dt \quad (3.20)$$

the mean square value as:

$$E[\phi^2(t)] \triangleq \overline{\phi^2(t)} = \lim_{T \rightarrow \infty} \frac{1}{T} \int_{-T/2}^{T/2} \phi^2(t) dt \quad (3.21)$$

the mean squared value as

$$\overline{\phi(t)^2} \triangleq E[\phi(t) \overline{\phi(t)}] \quad (3.22)$$

and the variance as

$$\begin{aligned} \sigma^2 &\triangleq E[\{\phi(t) - \overline{\phi(t)}\}^2] \\ &= \overline{\phi^2(t)} - \overline{\phi(t)}^2 \end{aligned} \quad (3.23)$$

If the variable phase shifter in Figure 3.7 is adjusted to the point where the mean value of the phase is zero, i.e.

$$\overline{\phi(t)} = 0$$

then the mean squared value $\overline{\phi(t)}^2 = 0$.

Let the time τ in equation (3.18) be equal to zero, that is where the autocorrelation function should be a maximum, then

$$R_{\phi}(0) = \lim_{T \rightarrow \infty} \frac{1}{T} \int_{-T/2}^{T/2} \phi^2(t) dt \quad (3.24)$$

with the assumption, the autocorrelation at $\tau = 0$ is equal to the mean square value, therefore

$$R_{\phi}(0) = \overline{\phi^2(t)} = E[\phi^2(t)] \quad (3.25)$$

and consequently the variance of the phase is

$$\sigma_{\phi}^2 = E[\phi^2(t)] \quad (3.26)$$

Taking equation (3.17) and substitute that into equation (3.25) resulting in:

$$R_{\phi}(0) = E[k^2 \phi_d^2(t)] \quad (3.27)$$

and substitute equation (3.26) into equation (3.27) resulting in

$$\begin{aligned} R_{\phi}(0) &= k^2 \sigma_{\phi}^2 \\ &= k^2 \overline{\phi^2(t)} \end{aligned} \quad (3.28)$$

taking the square root on both sides of equation (3.28)

$$\sqrt{R_{\phi}(0)} = k\sigma_{\phi} = k\sqrt{\overline{\phi^2(t)}} \quad (3.29)$$

where σ_{ϕ} is called the standard deviation and is the root mean square (rms) value of the phase $\phi(t)$ and is a measure of the spread of observed values.

To determine σ_{ϕ} , it is necessary to know what the autocorrelation of $V_n(t)$ (the output signal of the sample and hold) is.

The sample and hold circuit acts as a low pass filter with a transfer function of $\frac{\sin x}{x}$ which is necessary to measure the variance accurately [25].

If it is assumed that ϕ_d is statistically independent of ϕ_{d+1} , the autocorrelation function will appear as shown in Figure 3.11.

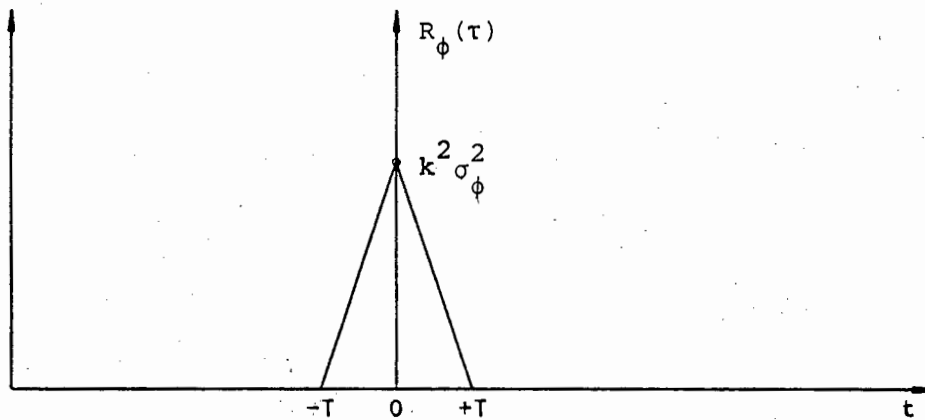


FIGURE 3.11 : Auto-correlation function.

The value of $R_\phi(0)$ which is equal to $\sigma^2[\phi(t)]$ can be obtained from an autocorrelator unit (HP 3721A) or by using a true RMS meter which measures $\sigma[\phi(t)]$.

Finally, it is necessary to know the value of k . This is achieved by changing the reference frequency ω_r so that $\omega_r - \omega_p \neq 0$. From equation (3.14) it can be seen that $V_o(t)$ will have the form as shown in Figure 3.12.

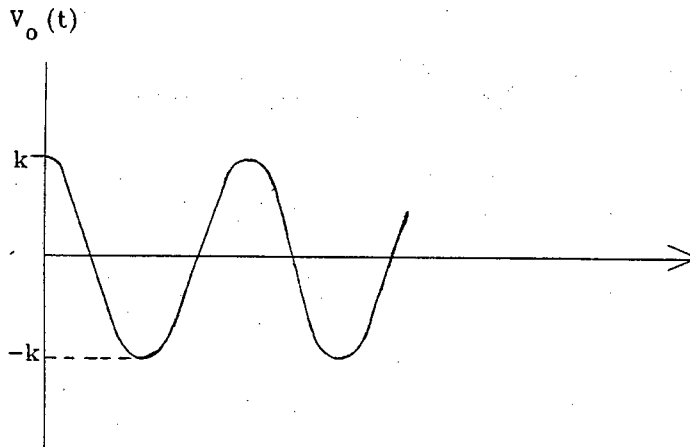


FIGURE 3.12 : Output if $\omega_r - \omega_p \neq 0$.

This can be observed on an oscilloscope and the peak voltage read off directly to obtain the value of k .

The zero crossing corresponds to 90° phase difference between the two inputs, the peak positive point on the curve to 180° and the peak negative point to 0° .

Therefore, the phase deviation in degrees $\phi_d(t)_{\text{rms}}$ can now be calculated from equation (3.29) as

$$\phi_d(t)_{\text{rms}} = \frac{\sqrt{R_r(0)}}{k} \cdot \frac{180}{\pi} \quad (3.30)$$

This measurement gives the total rms phase variations in an infinite bandwidth and the assumption is made that the variance of $\phi_d(t)$ is small, which is certainly valid for locked and primed conditions.

The method applied above does not assume any specific probability density function or spectrum of $\phi_d(t)$. As long as the tails of the PDF are not so long that the small angle assumption above is not abused, the approach may be considered valid.

Measurements were done on a system and are given in Section 5.

CHAPTER 4

SYSTEM PERFORMANCE MEASUREMENTS

4.1 INTRODUCTION

A laboratory bench model of the high resolution radar, of which the block diagram is shown in Figure 2.6, was designed and tested. The various components of the radar were first tested before it was integrated into the system. The design procedure and measured results of the various components are outlined in Chapters 6-9.

The tuning-in and setting-up procedure is relatively simple and can be done easily. It should, however, be noted that expensive capital equipment is required to do this.

4.2 SYSTEM CONFIGURATION

The system can basically be divided into two sub-systems, i.e.

- 1) injection locking of the amplifier chain;
- 2) amplification, mixing and detection of the IF signal.

The measurements on the sub-systems will be described first before the total performance of the system will be described.

4.3 SUB-SYSTEM MEASUREMENTS

4.3.1 Injection Locking of Amplifier Chain

The amplifier chain consists of a low power CW Gunn source (+10 dBm), a medium power CW Impatt source (+18 dBm) which is pulsed (from now on referred to as a CW source) and a high power pulsed Impatt source (+37 dBm).

To measure the locking bandwidth and locking gain of the various stages, a measuring set-up as shown in Figure 4.1 was assembled.

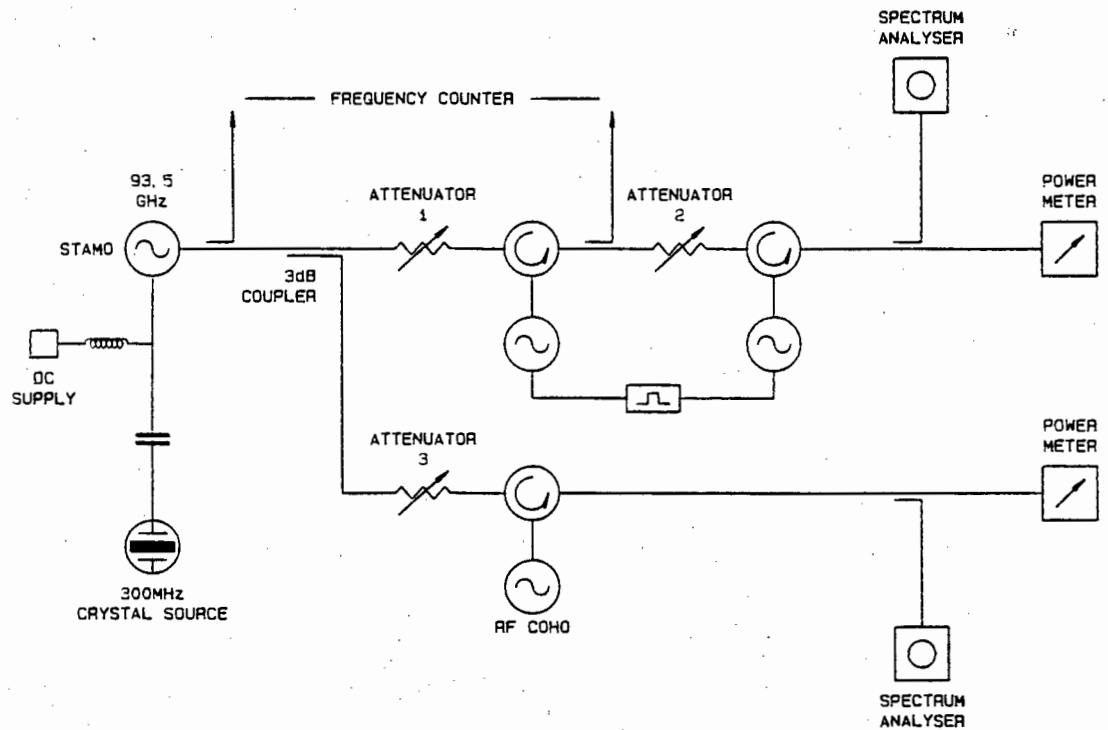


FIGURE 4.1 : Block diagram for measuring locking-gain and bandwidth.

The power levels at the various stages are:

- 1) Input power to attenuator 1 = 7 dBm.
- 2) Input power to attenuator 2 = 18 dBm.
- 3) Input power to attenuator 3 = 7 dBm.

The output spectrum of the Gunn oscillator consists of three frequency components, i.e. at 93,5 GHz and $93,5 \pm 0,300$ GHz. Initially it was decided to use bandpass filters to filter out the 93,5 GHz and 93,80 GHz frequency components which are needed to injection lock the PILO and RF Coho sources respectively. Through experimental investigations it was found that the injection locking could be done without using bandpass filters. The main advantage of using filters is that it improves the immunity of the radar against spurious signals as well as jamming.

The locking gain versus locking bandwidth of the RF Coho, CW Impatt and Pulsed Impatt Sources were measured as a function of injection power. See Figure 4.2. To measure the locking bandwidth of the Pulsed Impatt Source, the Stamo and CW Impatt oscillators were regarded as one source with attenuator 1 set to a minimum.

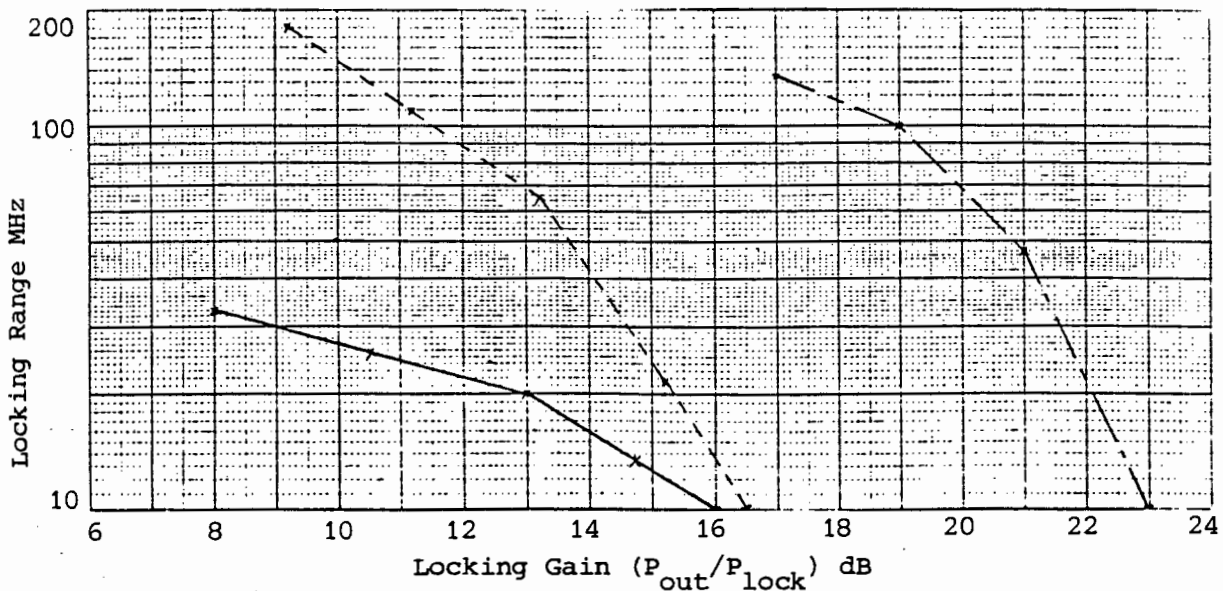


FIGURE 4.2 : Locking range versus locking gain for (a) COLO (—), (b) CW Impatt Source (-----) and (c) Pulsed Impatt Source (— - —).

4.3.2 Coherency of PILO

The coherency of the PILO was measured using the procedure outlined in Section 3.9, see Figure 4.3.

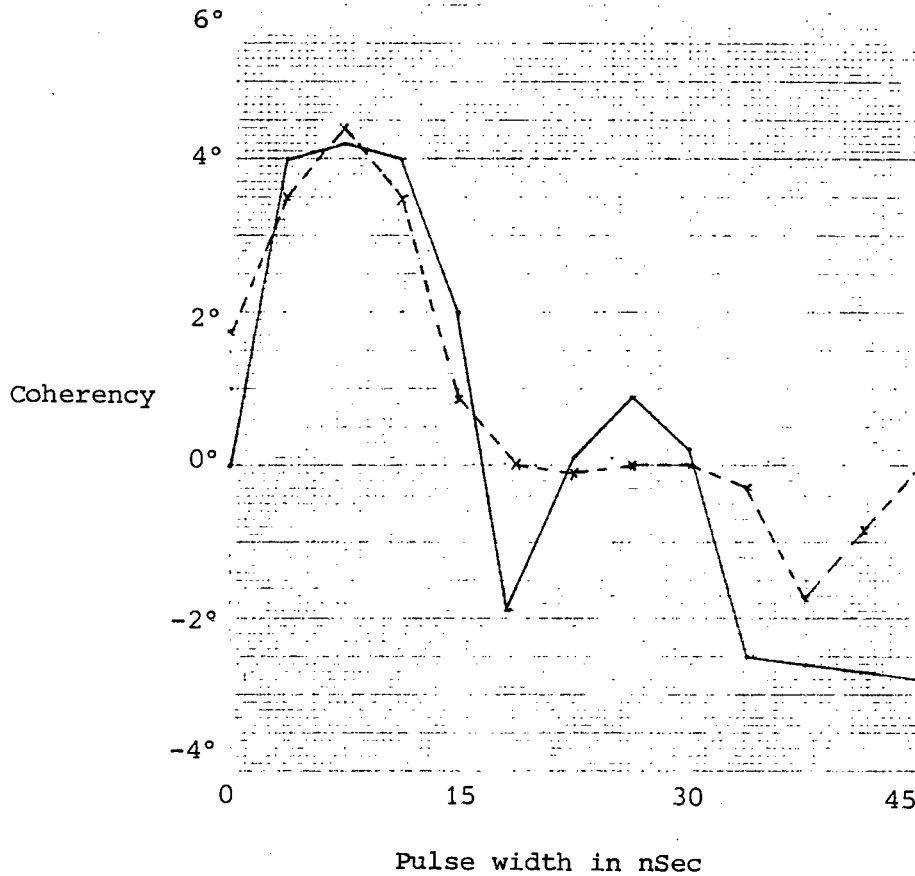


FIGURE 4.3 : Coherency response of PILO. (a) Results obtained by [26] (—) and (b) measured results obtained from the system described in this thesis (----).

Figure 4.4 shows the output of the phase detector for various settings of the phase shifter.

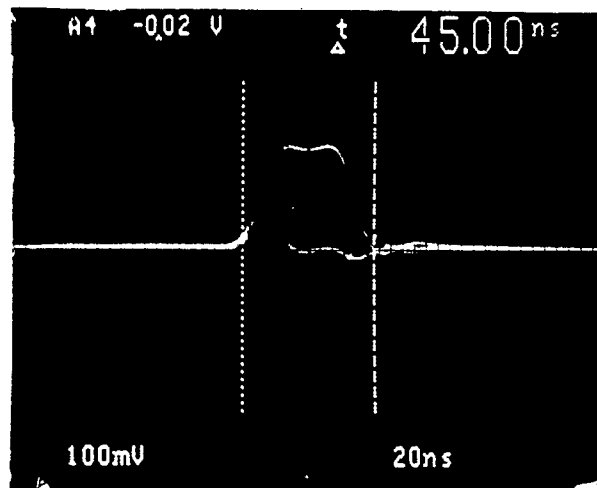


FIGURE 4.4 : Output of phase detector for various phase shifter settings. (a) Top trace is 0° and (b) middle trace is 90° .

4.3.3 Transfer Characteristic of IF Amplifier, Mixer, Bandpass Filter and IF Log Amplifier/Detector

The 300 MHz output signal from the mixer is amplified by a Miteq amplifier with a gain of 30 dB and a bandwidth of 0-500 MHz. The output signal of the amplifier is fed through a bandpass filter with a 80 MHz bandwidth and centre frequency 300 MHz. The output signal of the filter is fed into a log IF amplifier/detector. The detected output is used for range calculations and the IF output is fed into a coherent detector.

The transfer characteristics of the chain, i.e. from IF amplifier to detector is shown in Figure 4.5. The input signal was an RF pulse with a pulse width of 50 ns and PRF of 50 kHz.

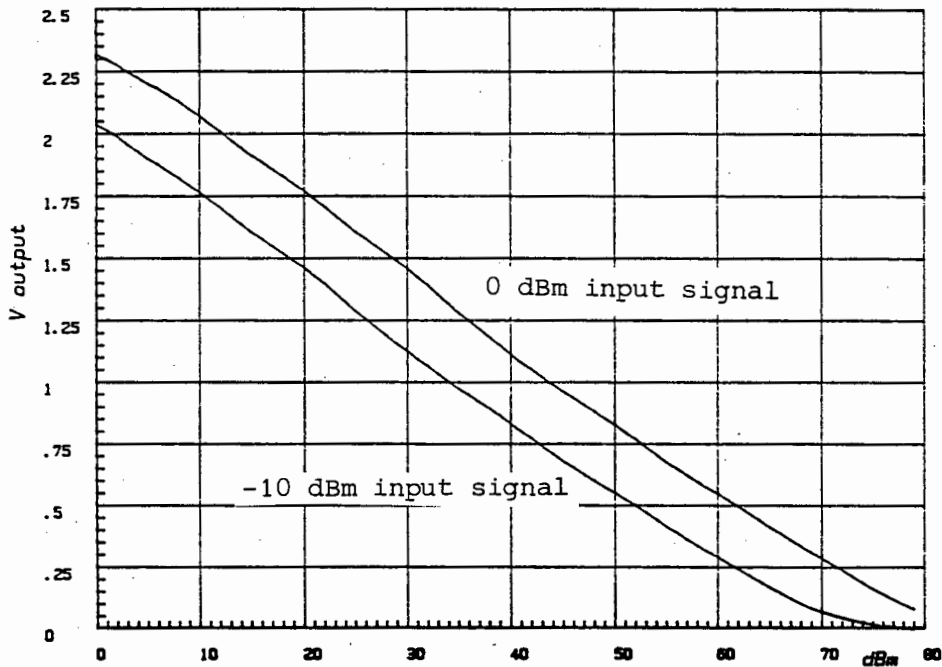


FIGURE 4.5 : Transfer characteristic of IF amplifier, mixer, filter and log IF amplifier/detector chain.

4.3.4 Coherency of Quadrature Detector

The coherency of the IF section was measured and is shown in Figure 4.6. From Figure 4.6 it is clear that the coherency of both signals are good at the centre of the pulse, and start deteriorating towards the edges. This effect is caused by the sharp phase transients at the edge of a pulse.

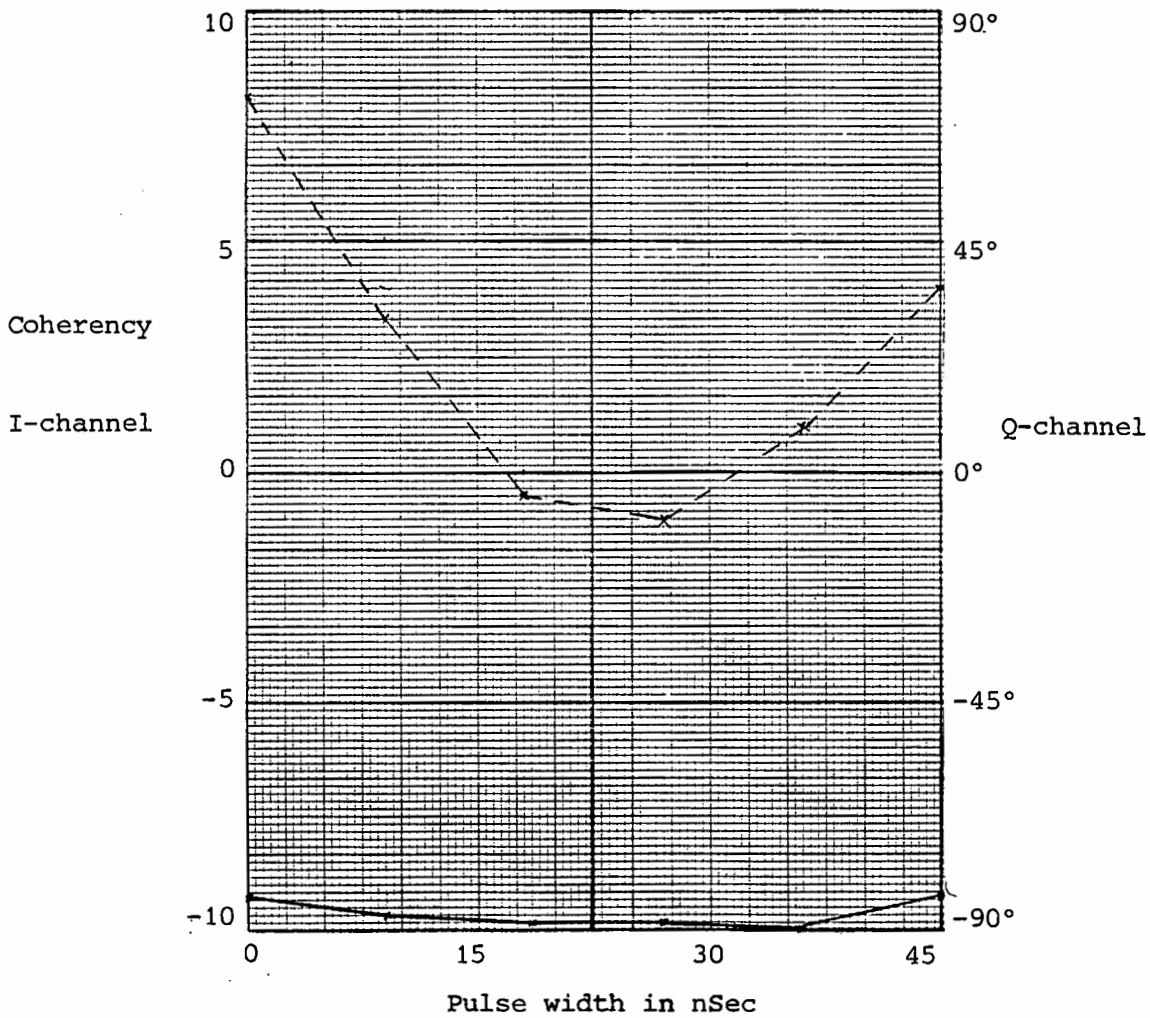


FIGURE 4.6 : Coherency of IF section. (a) I-channel (----) and (b) Q-channel.

Figure 4.7 shows the output signals of the quadrature detector.

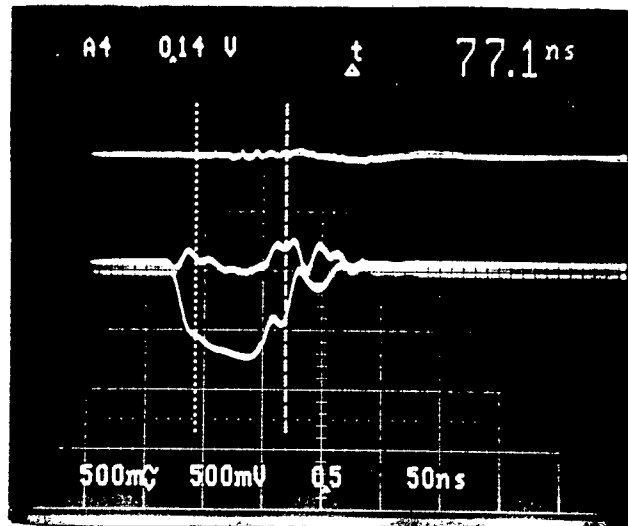


FIGURE 4.7 : Output signals of quadrature detector. (a) Bottom trace, I-channel, (b) middle trace, Q-channel and (c) top trace is output of sample and hold circuit.

4.4 SYSTEM MEASUREMENTS

The measurements made with the radar are divided into five categories, i.e.

- 1) System coherency.
- 2) Air targets without clutter.
- 3) Ground targets with ground clutter.
- 4) Distance measurements.

Each of the subjects will be discussed separately and results obtained will be compared with theoretically predicted data.

4.4.1 System Coherency

The coherency of the system as a whole was measured and is shown in Figure 4.8.

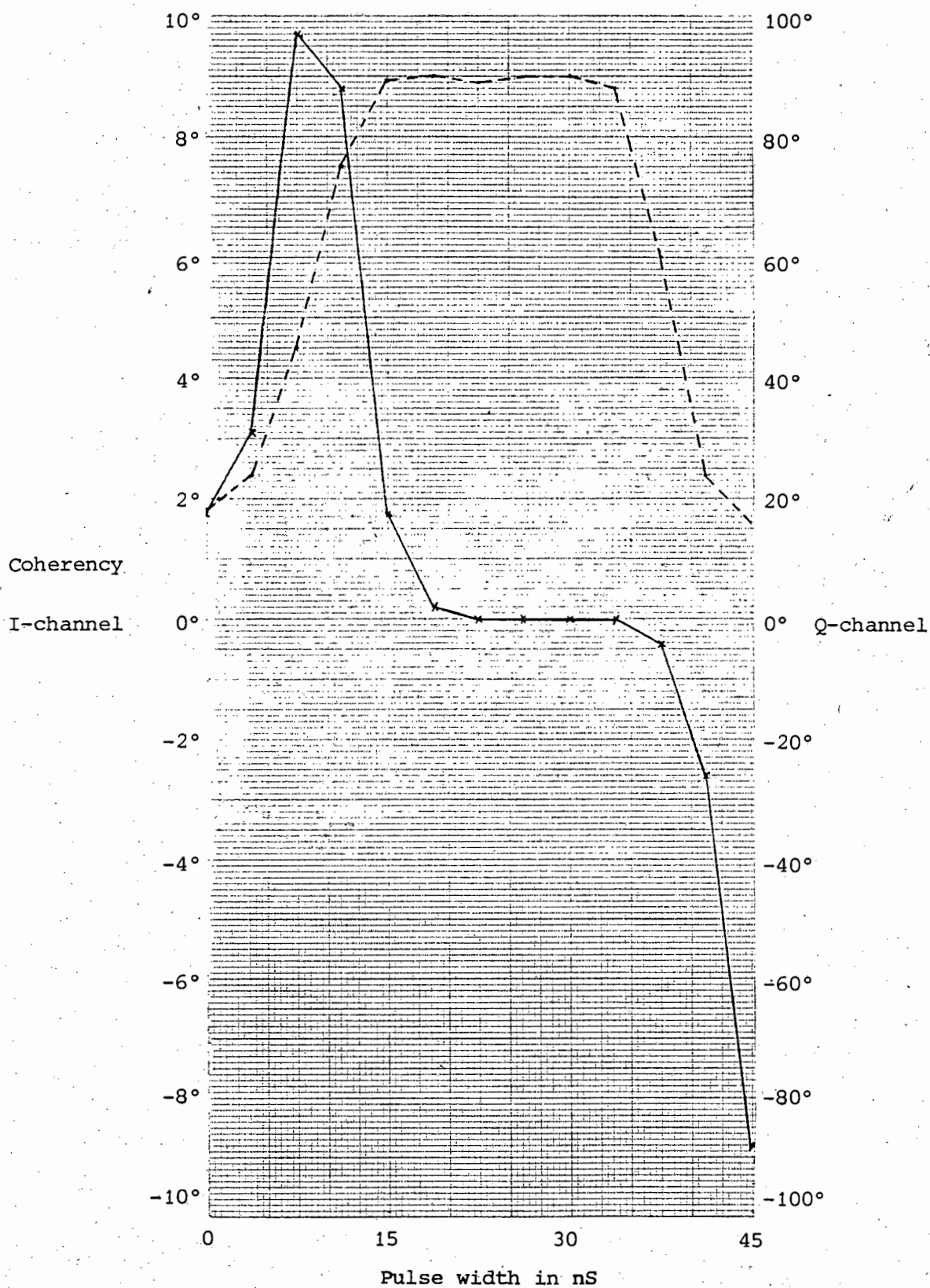


FIGURE 4.8 : System coherency. (a) I-channel (----) and (b) Q-channel (—).

Figure 4.9 shows the output signal of the I and Q-channels respectively.

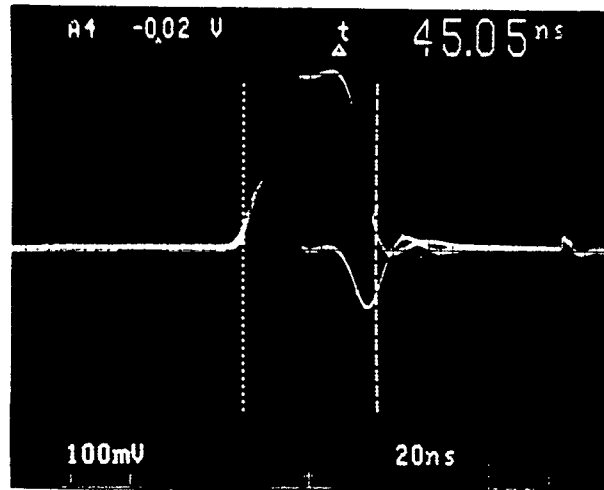


FIGURE 4.9 : Output signals of integrated system. (a) I-channel top trace and (b) Q-channel middle trace.

4.4.2 Air Targets Without Clutter

A corner reflector with a calculated radar cross section of 721 m^2 was mounted on a 40 m sectional pole mast at a distance of 300 m. By using equation (2.9), the signal level out of the first mixer should be:

$$P_{m1} = -50,5 \text{ dBm}$$

if the conversion loss of -9 dB of the mixer is taken into account.

By using this signal level, it is found from Figure 4.5 that the detected output signal should be 0,83 V.

The actual measured signal is shown in Figure 4.10.

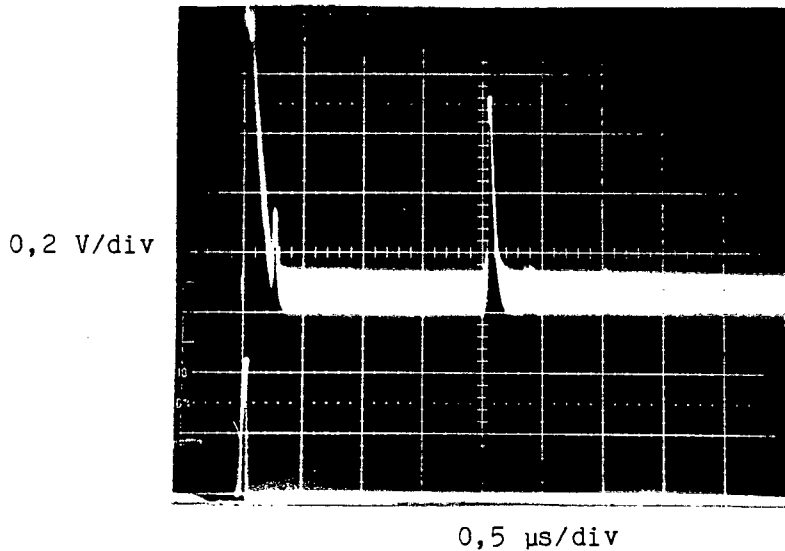


FIGURE 4.10 : Measured signal level from corner reflector.

The measured and predicted signals compared favourably and the difference of 0,12 V could be due to losses not accounted for in the calculations.

4.4.3 Ground Targets With Ground Clutter

Measurements were taken on a vehicle with dimensions 3 x 1,8 x 1,8 m at a 300 m distance. Assume that the average radar cross section is 6 m².

By using equation (2.9), the received signal from the vehicle out of the first mixer should be:

$$P_{rv} = -54 \text{ dBm}$$

The measured return signal from the vehicle is shown in Figure 4.11.

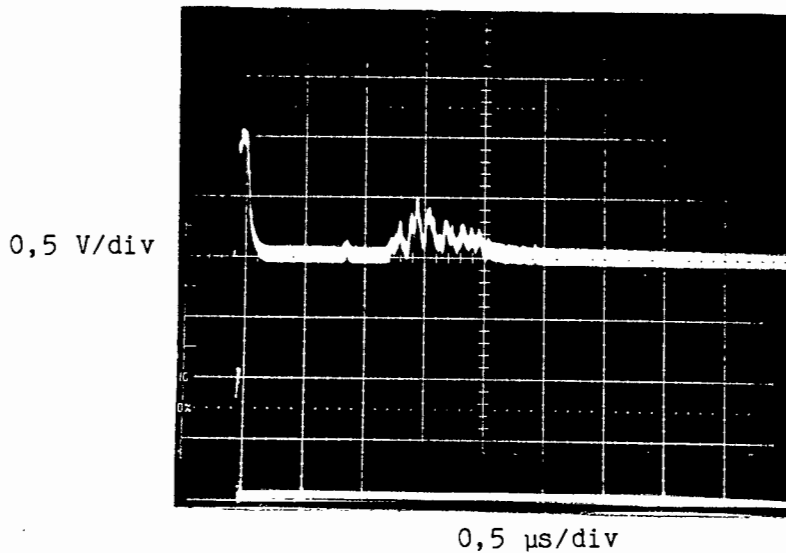


FIGURE 4.11 : Measured signal level from vehicle.

The reason why the target could still be detected without the use of range gating techniques is that the ground is a much worse reflector of electromagnetic waves than the -10 dB/m^2 as was initially estimated.

By converting the received signal back to determine what the relative cross section of the ground is, it is found that it is in the order of -30 dB/m^2 .

4.4.4 Distance Measurement

The maximum distance that the vehicle could still be detected was 1,5 km. The measured signal level is shown in Figure 4.12.

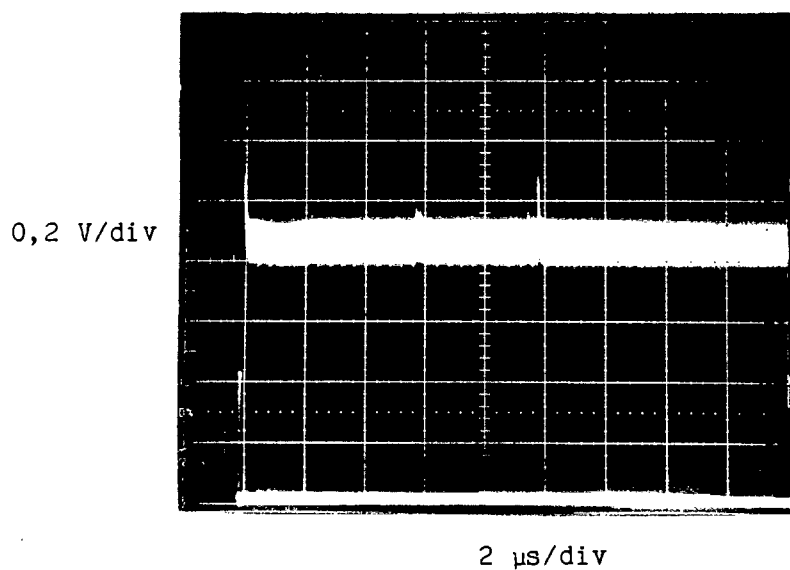


FIGURE 4.12 : Measured signal level of vehicle at 1,5 km.

This distance was measured just before the signal disappeared into the noise. The predicted signal level of 0,3 V is close to the measured signal of 0,28 V.

4.5 CONCLUSIONS

The measured results of the various aspects of the system is close to the theoretically predicted values.

An advantage of the system is that it is relatively simple and uses a minimum of components. Apart from its usage as a radar, it can also be used as an instrument to measure radar cross sections.

CHAPTER 5

PHASE NOISE MEASUREMENTS OF CW SOURCES

5.1 INTRODUCTION

This chapter describes the phase noise phenomena and the technique to measure it.

A very important parameter of a radar system is the phase noise of the oscillators. If the phase noise is large compared to the returning echo from a target, the echo will be buried in the phase noise and no detection would be possible.

The other important point is that if a noisy source is used to injection lock an amplifier chain, the phase of the transmit signal will vary according to the reference signal and it would be difficult to obtain coherency.

5.2 WHAT IS NOISE?

Frequency stability can be defined as the degree to which an oscillator source produces the same frequency throughout a specified period of time. Every microwave source exhibits some degree of frequency instability. This stability can be broken down into two components, long term and short term stability.

Long term stability describes the frequency variations that occur over long time periods, expressed in parts per million per hour, day, month or year. Short term frequency stability contains all elements causing frequency changes about the nominal frequency of less than a few seconds duration.

A modified version of the mathematical description of a sine wave [27] with amplitude and phase noise on it is described as:

$$V_S(t) = |A_0 + \epsilon(t)| \cos(\omega_c t + \Delta\phi(t)) \quad (5.1)$$

A_0 = signal amplitude

$\epsilon(t)$ = amplitude fluctuations

ω_c = carrier frequency

and $\Delta\phi(t)$ = randomly fluctuating phase term

The randomly fluctuating phase term could be observed on an ideal spectrum analyzer [31] (one which has no sideband noise of its own) as in Figure 5.1.

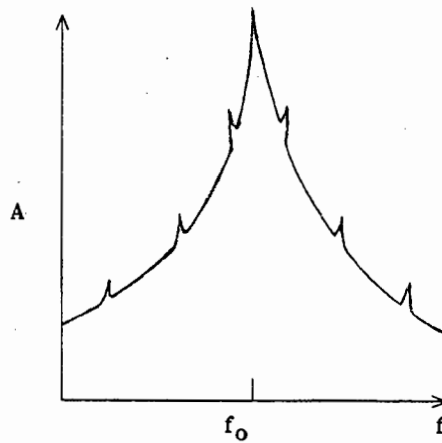


FIGURE 5.1 : RF sideband spectrum.

There are two types of fluctuating phase terms. The first, deterministic, are discrete signals appearing as distinct components in the spectral density plot. These signals, commonly called spurious, can be related to known phenomena in the signal source, such as power line frequency, vibration frequencies, or mixer products.

The second type of phase instability is random in nature, and is commonly called phase noise. The source of random sideband noise in an oscillator include thermal noise, shot noise and flicker noise.

Many terms exist to quantify the characteristic randomness of phase noise. Essentially, all methods measure the frequency or phase deviations of the source under test in either the frequency or time domain. Since frequency and phase are related to each other, all of these terms are also related.

One fundamental description of phase instability or phase noise is the one-sided spectral density of phase fluctuations on a per-Hertz basis. The term spectral density describes the energy distribution as a continuous function expressed in units of energy within a specified bandwidth.

For noise considerations it can be assumed that the magnitude of $\Delta\phi(t)$ in equation (5.1) is small or the maximum phase deviation of $\Delta\phi(t)$ is much less than one radian. This implies that the phase noise can be treated as narrow band phase modulation. Expanding equation (5.1) yields:

$$V_s(t) = |A_o + \varepsilon(t)| \{ \cos \omega_c t \cos \Delta\phi(t) - \sin \omega_c t \sin \Delta\phi(t) \} \quad (5.2)$$

Since $\Delta\phi(t) \ll 1$ the following approximation can be made:

$$\cos \Delta\phi(t) \approx 1$$

$$\sin \Delta\phi(t) \approx \Delta\phi(t)$$

Equation (5.2) can then be rewritten as

$$V_s(t) = |A_o + \varepsilon(t)| \{ \cos \omega_c t - \Delta\phi(t) \sin \omega_c t \} \quad (5.3)$$

Equation (5.3) shows that the generation of phase noise is a linear modulation process in which the noise spectrum of $V_s(t)$ is the spectrum of $\Delta\phi(t)$ translated by ω_c . This can be expressed in the frequency domain referred to the baseband as:

$$V_n(\omega) = \frac{1}{2} |V_o + \varepsilon(t)| \Delta\phi(\omega) \quad (5.4)$$

where $V_n(\omega)$ = noise spectrum of $V_s(t)$

$\Delta\phi(\omega)$ = spectrum of $\Delta\phi(t)$ [31], see Figure 5.2.

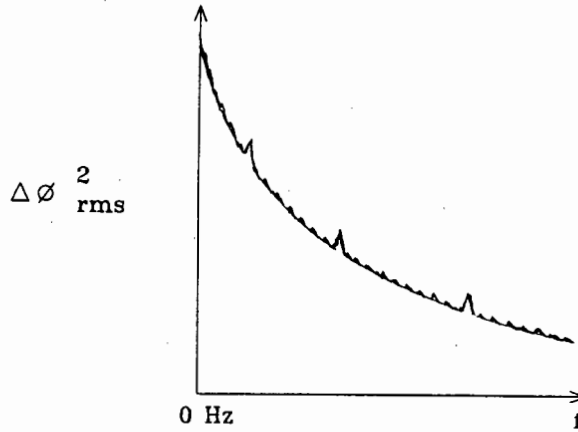


FIGURE 5.2 : Phase noise sidebands.

Thus $S_{\Delta\phi}(f)$ in rad^2/Hz is defined as the one-sided spectral distribution of the phase fluctuations on a "per hertz" basis and is calculated in [28].

$$S_{\Delta\phi}(f) = \frac{\Delta\phi_{\text{RMS}}^2}{\text{Bandwidth used to measure } \Delta\phi_{\text{RMS}}} \frac{\text{rad}^2}{\text{Hz}} \quad (5.5)$$

A useful measure of frequency stability is to relate the sideband power associated with phase fluctuations to the carrier power level. The defined measurement unit is called $L(f)$. $L(f)$ is defined as the ratio of power in one sideband referred to the input carrier frequency on a per hertz of bandwidth spectral density basis, to the total signal power at Fourier frequency f from the carrier. It is a normalized frequency domain measure of phase fluctuation sideband power [31], see Figure 5.3.

$$L(f) = \frac{\text{Power Density (one sideband)}}{\text{Carrier Power}} \text{ Hz}^{-1} \quad (5.6)$$

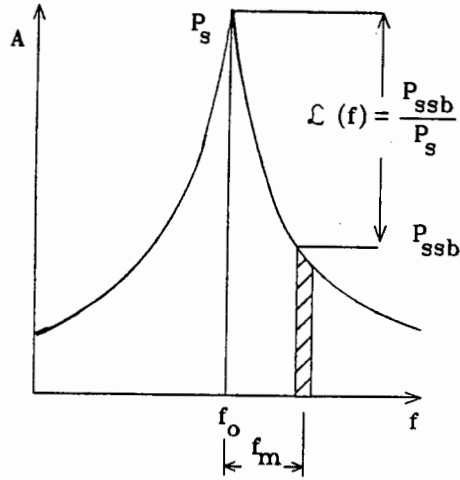


FIGURE 5.3 : Frequency domain measure of phase fluctuations.

For the condition that the phase fluctuation, occurring at rates f and faster, are small compared to one radian, a good approximation of equation (5.5) as defined by [29] is:

$$S_{\Delta\phi}(f) = 2 \text{ rad}^2 L(f) \quad \text{rad}^2/\text{Hz} \quad (5.7)$$

$L(f)$ is often expressed in decibels relative to the carrier per hertz (dBc/Hz), which is calculated as:

$$L(f) = 10 \log \frac{S_{\Delta\phi}(f)}{2 \text{ rad}^2} \quad \text{dBc/Hz} \quad (5.8)$$

5.3 NOISE MEASUREMENTS

5.3.1 Measuring Set-Up

The set-up used to measure the phase noise of an oscillator is shown in Figure 5.4, as described by [27,31].

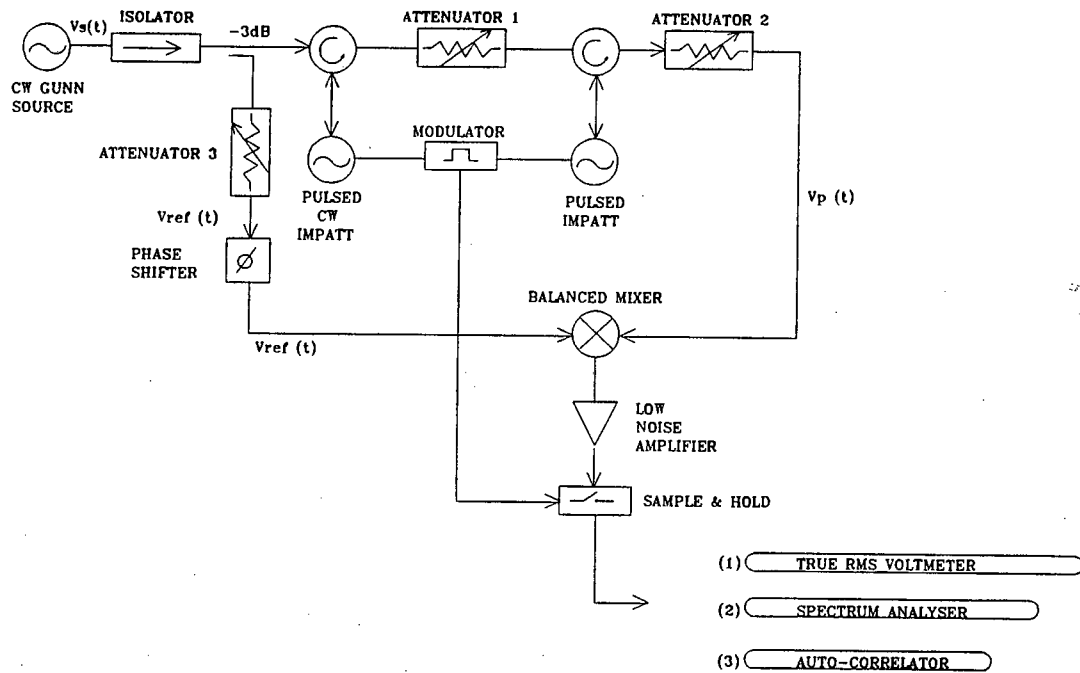


FIGURE 5.4 : Basic delay line discriminator set-up.

It consists of a power splitter, a delay line that provides the additional phase shift from which the noise information will be extracted, phase shifter, mixer and a low noise video amplifier. Measurements are made on a baseband spectrum analyzer.

5.3.2 Theory of Operation

Normally AM noise is much less than phase noise in Gunn and Impatt Oscillators [32], and therefore this discussion will describe the measuring procedure for phase noise. In fact, the delay line discriminator measuring set-up is insensitive to AM noise.

Suppose a signal in the form of equation (5.1) enters the delay line discriminator depicted in Figure 5.4. It is first separated into a reference signal and a delay line signal with the power splitter. At this point the two signals have identical phase information, ignoring the amplitude fluctuations because the delay line discriminator is insensitive to AM noise.

$$V_{\text{delay}}(t) = \ell_1 A_0 \cos(\omega_c t + \Delta\phi(t)) \quad (5.9)$$

$$V_{\text{ref}}(t) = \ell_1 A_0 \cos(\omega_c t + \Delta\phi(t)) \quad (5.10)$$

(ℓ_1 is the loss of the power splitter, ideally 3 dB).

The delay channel signal passes through the delay line with loss ℓ_2 and fixed delay time t_d .

$$V_{\text{delay}}(t) = \ell_1 \ell_2 A_0 [\cos(\omega_c(t-t_d) + \Delta\phi(t-t_d))] \quad (5.11)$$

The reference channel signal passes through the phase shifter which adds the phase $\phi\Delta$.

$$V'_{\text{ref}}(t) = \ell_1 \ell_3 A_0 [\cos(\omega_c t + \Delta\phi(t) + \phi\Delta)] \quad (5.12)$$

The processed reference and delay channel signals then enter the mixer with conversion loss ℓ_3 . The output signal is:

$$\begin{aligned} V_m(t) = \ell_1 \ell_2 \ell_3 A_0 \{ & \cos(2\omega_c t - \omega_c t_d + \Delta\phi(t-t_d) + \Delta\phi(t) + \phi\Delta) \\ & + \cos(\omega_c t_d + \phi\Delta + \Delta\phi(t) - \Delta\phi(t-t_d)) \} \end{aligned} \quad (5.13)$$

Ignoring the higher frequency components and amplifying leaves the baseband signal

$$V_o(t) = g l_1 l_2 l_3 A_o \cos(\omega_c t_d + \phi\Delta + \Delta\phi(t) - \Delta\phi(t-t_d)) \quad (5.14)$$

where g is the gain of the amplifier. It is this signal from which the phase noise information of the source can be extracted.

The first pre-requisite for accurate phase noise measurements is to put the two input signals to the mixer in phase quadrature. This effectively removes the carrier contributed DC component, rejects AM noise and provides maximum mixer phase sensitivity and linearity. Placing the inputs in phase quadrature is accomplished by adjusting $\phi\Delta$ such that the sum of $\phi\Delta + \omega_c t_d = \pi/2$. Substitute this into equation (5.14) gives

$$V_o(t) = g l_1 l_2 l_3 A_o \cos\left[\frac{\pi}{2} + \Delta\phi(t) - \Delta\phi(t-t_d)\right] \quad (5.15)$$

substitute cosine with sine.

$$V_o(t) = g l_1 l_2 l_3 A_o \sin[\Delta\phi(t) - \Delta\phi(t-t_d)] \quad (5.16)$$

Since the magnitude of $\Delta\phi(t) - \Delta\phi(t-t_d)$ can never exceed twice the magnitude of $\Delta\phi(t)$, and $\Delta\phi(t)$ is assumed to be small, the magnitude of $\Delta\phi(t) - \Delta\phi(t-t_d)$ can also be guaranteed as small. This allows use of the approximation.

$$\sin[\Delta\phi(t) - \Delta\phi(t-t_d)] \approx \Delta\phi(t) - \Delta\phi(t-t_d) \quad (5.17)$$

$$\text{then } V_o(t) = g l_1 l_2 l_3 A_o [\Delta\phi(t) - \Delta\phi(t-t_d)] \quad (5.18)$$

Measurements will be made in the frequency domain, so it is therefore appropriate to represent $\Delta\phi(t)$ as the Fourier integral

$$\Delta\phi(t) = \int_{-\infty}^{\infty} \Delta\phi(\omega) e^{j\omega t} d\omega$$

where $\Delta\phi(\omega)$ is the spectrum of $\Delta\phi(t)$. The same applies to $\Delta\phi(t-t_d)$.

The expression for $V_o(t)$, equation (5.18), can now be written as

$$V_o(t) = g l_1 l_2 l_3 A_o \int_{-\infty}^{\infty} \Delta\phi(\omega) e^{j\omega t} [1 - e^{-j\omega t_d}] d\omega \quad (5.19)$$

or in the frequency domain as

$$V_o(\omega) = g l_1 l_2 l_3 A_o \Delta\phi(\omega) (1 - e^{-j\omega t_d}) \quad (5.20)$$

If ωt_d is chosen to be much less than one, the expression

$$1 - e^{-j\omega t_d} \approx j\omega t_d \quad (5.21)$$

or
$$V_o(\omega) = g l_1 l_2 l_3 A_o \Delta\phi(\omega) j\omega t_d \quad (5.22)$$

Equation (5.22) contains the important result that for all ω such that $\omega t_d \ll 1$, the output $V_o(\omega)$ of the delay line discriminator is the spectrum of $\phi(t)$ weighted by the additional factor ω . Recalling equation (5.4),

$$V_n(\omega) = \frac{1}{2} \frac{V_o(\omega)}{l_1 l_2 l_3 g j\omega t_d} \quad (5.23)$$

Thus from equation (5.23), $V_o(\omega)$ can be measured and if the constants are known, the phase noise spectrum can be determined and $L(f)$ can be determined.

5.3.3 Design Considerations

The key element in the design of the delay line discriminator is the delay line itself. For accurate measurements it was shown that the delay t_d must be chosen such that $\omega_m t_d \ll 1$, where ω_m is the maximum frequency of the components to be measured. From the analysis it can be shown that for $\omega_m t_d \leq 0,48$, the error generated in the magnitude of the output spectrum of $V_o(\omega)$ will be less than one per cent.

Besides assuming that $\omega_m t_d \leq 0,48$, the loss ℓ_2 of the delay line will effect the overall sensitivity of the measuring set-up, ℓ_2 is given by the formula

$$\ell_2 = e^{-\alpha L} \quad (5.24)$$

where α is the attenuation factor and L is the delay line length. The delay of the line is given by

$$t_d = \frac{L}{v} \quad (5.25)$$

where v is the propagation velocity of the line.

From equations (5.23) and (5.24) it is seen that for increasing L , L_d and ℓ_2 increases. The product for which $\ell_2 t_d$ is maximized is given by

$$L = \frac{1}{\alpha} \quad (5.26)$$

Indicating that the optimum loss is one neper or 8,69 dB, and the optimum delay

$$t_d = \frac{1}{\alpha v}.$$

5.3.4 Calibration Procedure

To calibrate the system, it is necessary to modulate the oscillator with a known $V_n(\omega)$. This is accomplished by modulating the oscillator so that the sideband levels can be measured on a high frequency spectrum analyzer. Before the calibration procedure is started, the variable phase changer must be adjusted to give zero DC volts at the output of the mixer. The assumption that $|\Delta\phi(t)| \ll 1$ must not be violated, and as a rule $|\Delta\phi(t)| < 0,2$, or the sideband levels should not exceed -20 dBc. Measurements were also performed where the modulation index was 2,405, that is when the amplitude of the modulating signal is increased until the carrier of the oscillator under test is reduced to the first Bessel null. The accuracy of the system with both modulating levels were the same.

The calibration factor is defined as

$$CF = \frac{V_n(\omega)_{rms}}{V_o(\omega)_{rms}} \quad \text{Hz/V} \quad (5.27)$$

Where $V_m(\omega)$ is the rms frequency deviation of the carrier due to the intentional modulation and $V_o(\omega)$ is the spectrum analyzer voltage of the modulation sideband. The peak deviation of a f_m signal is defined as

$$V_n(\omega) = mf_m \quad \text{Hz} \quad (5.28)$$

where f_m is the modulation signal and m is the modulation index, therefore

$$V_n(\omega)_{rms} = \frac{mf_m}{\sqrt{2}} \quad \text{Hz} \quad (5.29)$$

and the calibration factor of the discriminator is

$$CF = \frac{mf_m}{\sqrt{2} V_o(\omega)_{rms}} \quad \text{Hz/V} \quad (5.30)$$

To determine $V_o(\omega)_{rms}$, the power reading in dBm on the spectrum analyzer can be converted to the corresponding voltage.

5.3.5 Correction Factors

The power reading in dBm on the spectrum analyzer needs some correction factors [30,31]:

- 1) The error introduced by peak detectors rather than true rms detector gives rise to an error of 1,05 dB.
- 2) The logarithmic display compression reduces the apparent average value and gives an error of 1,45 dB.
- 3) The noise bandwidth is approximately 0,8 of the resolution bandwidth.
- 4) A 6 dB error due to the mixer used as a phase detector.

5.3.7 Measurement of Phase Noise

By removing the modulation signal, each Fourier frequency noise power reading P_n (dBm) is converted to the corresponding rms voltage $\Delta V_{1 \text{ rms}}$ by using equation (5.33). The rms frequency fluctuations are calculated as:

$$\Delta V_{\text{rms}} = \Delta V_{c \text{ rms}} \times CF \quad \text{Hz} \quad (5.34)$$

The spectral density of frequency fluctuations is calculated as:

$$S_{\Delta V}(f) = \frac{\Delta V_{\text{rms}}^2}{B} \quad \text{Hz}^2/\text{Hz} \quad (5.35)$$

where B is the effective noise power bandwidth of the spectrum analyzer.

The spectral density of phase fluctuations is calculated as:

$$S_{\Delta \phi}(f) = \frac{S_{\Delta V}(f)}{f^2} \quad \text{rad}^2/\text{Hz} \quad (5.36)$$

where f is the measuring frequency. The normalized phase noise sideband power spectral density is calculated as:

$$L(f) = \frac{S_{\Delta \phi}(f)}{2} \quad \text{Hz}^{-1} \quad (5.37)$$

The correcting factor of 2 is left out because it is already included in the subtraction of the 6 dB in Section 5.2.5.

The above steps can be included into one equation

$$L(f) \text{dBc/Hz} = 20 \log \left[\frac{\Delta V_{c \text{ rms}} \cdot CF}{\sqrt{B} \cdot f} \right] \text{dBc/Hz} \quad (5.38)$$

5.3.8 AM Noise Rejection by Delay Line Discriminator

The assumption is made that when the two signals to the mixer are placed in phase quadrature, the delay line discriminator will be insensitive to AM noise. This analysis is somewhat simplified because it does not include the phase noise components.

A signal including AM noise can be represented as

$$V_s(t) = (A_o + \varepsilon(t)) \cos \omega_c t \quad (5.39)$$

where $\varepsilon(t)$ describes the amplitude variations with time. A signal of this form entering the delay line discriminator in Figure 5.4 will generate the following input to the mixer.

$$V'_{\text{ref}}(t) = l_1(A_o + \varepsilon(t)) \cos(\omega_c t + \Delta\phi) \quad (5.40)$$

$$V'_{\text{delay}}(t) = l_1 l_2 (A_o + \varepsilon(t)) \cos \omega_c (t - t_d) \quad (5.41)$$

mixing and audio amplification leaves the baseband output

$$V_o(t) = \frac{l_1 l_2 l_3 g A_o}{2} \left[\cos(\Delta\phi + \omega_c t_d) + \varepsilon(t - t_d) \cos(\Delta\phi + \omega_c t_d) \right. \\ \left. + \varepsilon(t) \cos(\Delta\phi + \omega_c t_d) + \varepsilon(t) \varepsilon(t - t_d) \cos(\Delta\phi + \omega_c t_d) \right] \quad (5.42)$$

For quadrature $\Delta\phi$ is adjusted so that $\Delta\phi + \omega_c t_d = \pi/2$ which means that $V_o(t) = 0$.

The inclusion of both AM and phase noise will in fact generate some AM noise related components in the output of the delay line discriminator. Typical AM rejection of the delay line is greater than 20 dB.

Because AM noise is usually much smaller than phase noise in Gunn and Impatt oscillators [32,33], no attempt was made to measure it.

5.3.9 Actual Measurements

Measurements were performed on the second Harmonic Gunn source and is shown in Chapter 6.

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CHAPTER 6

THE DESIGN OF SECOND HARMONIC MILLIMETER-WAVE GUNN OSCILLATORS

6.1 INTRODUCTION

Harmonic oscillations can be accomplished either in a separate stage (e.g. varactor-multiplier) or in the oscillator itself. At millimeter-waves, a second harmonic oscillator has a comparable efficiency and is more cost effective.

To provide low noise power for such applications as local oscillators and master oscillators in power amplifier chains, second harmonic oscillators are often used.

Gallium Arsenide (GaAs) diodes can effectively only deliver power at the fundamental frequency up to about 60 GHz. Above 60 GHz Impatt or Indium Phosphide (InP) Gunn diodes have to be used. The advantage of Impatt diodes are that they can deliver more power at higher frequencies, but the phase noise is much higher. InP Gunn diodes can also deliver more power at higher fundamental frequencies with similar phase noise than Gunn diodes. InP diodes, however, only became commercially available during the past year, and deliver less power than Impatt diodes.

Harmonic oscillators exhibit some interesting features:

- 1) They are relatively low noise.
- 2) They have much higher load pulling stability.
- 3) It is less expensive to generate the same power with harmonic operation than it is to use a fundamental mode device.
- 4) The technology of Gunn elements at lower frequencies are more mature, have greater reliability and a longer life can be expected.

6.2 PRINCIPLE OF OPERATION

The equivalent circuit [37] for a second harmonic oscillator is shown in Figure 6.1.

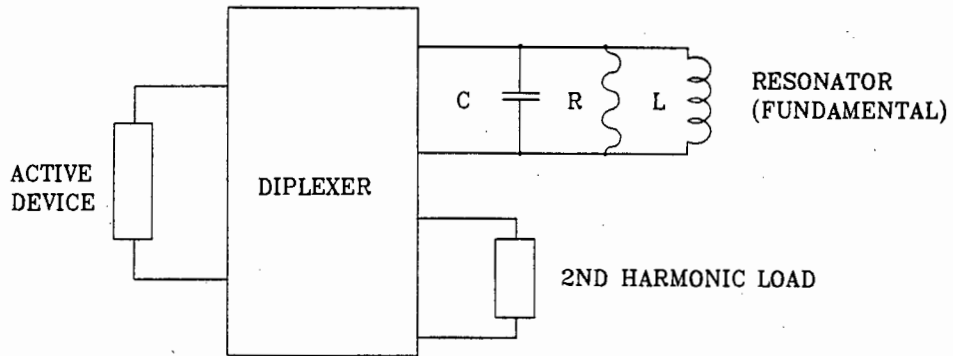


FIGURE 6.1 : Schematic representation of second harmonic oscillator.

Gunn as well as Impatt diodes are inherently non-linear devices [34-36]. If such a device is placed in a resonant circuit at the fundamental frequency, the current waveform is non-linear and consists of various harmonics, and is written as

$$\bar{I} = \bar{C}_1 V + \bar{C}_2 V^2 + \bar{C}_3 V^3 \quad (6.1)$$

The quadratic term in equation (6.1) determines the second harmonic operation. With numerous mathematical manipulation and by applying certain conditions to equation (6.1), it can be shown [37] that the maximum obtainable second harmonic power is 11 dB below the maximum obtainable fundamental power [37], see Figure 6.2.

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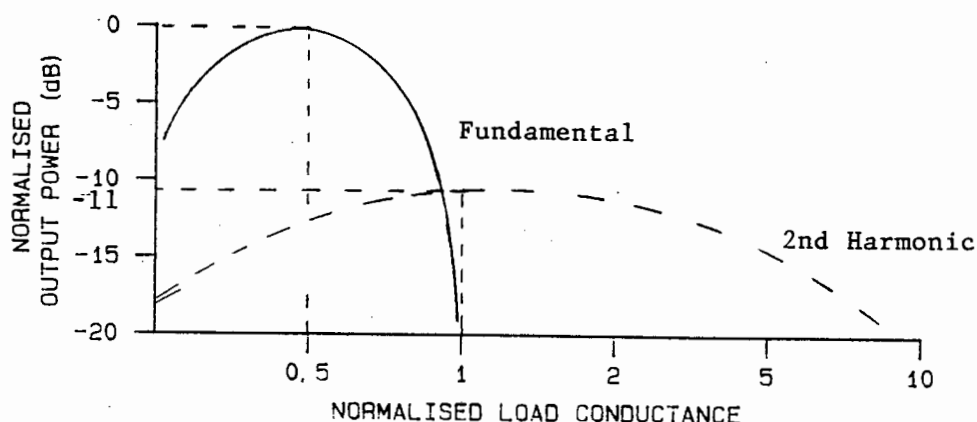


FIGURE 6.2 : Fundamental and second harmonic power versus load conductance.

It is, therefore, necessary to give special attention to the design of the resonant circuit and cavity if optimum efficiency for second harmonic oscillation is required.

Various articles [38-40] have been published on second harmonic oscillators. As can be seen in Figure 6.1 the active element operates into a resonator at the fundamental frequency and power is delivered to the load at the second harmonic. The network performing as a diplexer can be a conventional filter. If the active element and the resonator are placed in a waveguide cavity of which the cut-off frequency is higher than the fundamental frequency, the waveguide cavity acts as a filter. The high Q factor of the filter assure that the fundamental resonator is greatly decoupled from the second harmonic load. This leads to less noise and load pulling, and a decrease in output power.

From Figure 6.2 it is seen that the source impedance of the harmonic oscillator is about a factor of two lower than the fundamental source impedance. This implies that it is more difficult to match the relatively low impedance (10Ω) to the high impedance waveguide (377Ω).

6.3 COAX/WAVEGUIDE CIRCUIT

The cross section structure of a second harmonic oscillator is shown in Figure 6.3.

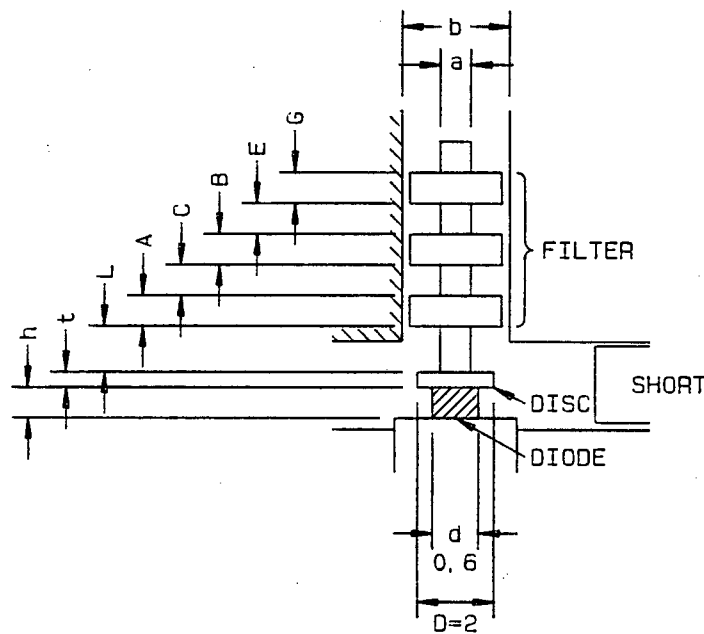


FIGURE 6.3 : Cross section of a second harmonic oscillator.

A Gunn diode is mounted in the broad wall of a WR10 waveguide and a moveable short is located on one side of it. A bias choke acting from the top incorporates a five section high-low-high filter section and a two-section impedance transformer.

The transformer section consists of a radial disc of diameter D and thickness t , and a post of length L and diameter d .

The transformer acts as a resonant circuit at the fundamental frequency, while at the second harmonic frequency it matches the characteristic impedance of the waveguide to the low source impedance of the Gunn device, with the imaginary part of the load impedance being tuned by the short.

Figure 6.4 shows an equivalent circuit [38] valid at the fundamental frequency

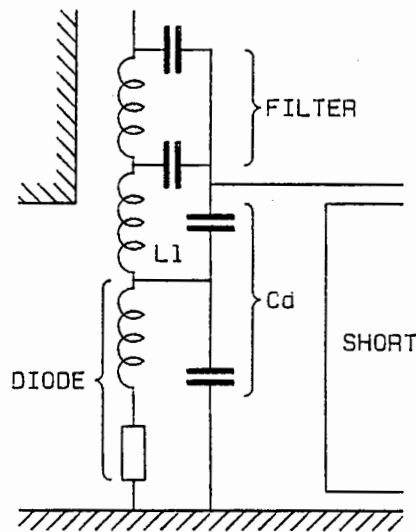


FIGURE 6.4 : Equivalent circuit of oscillator at f_o .

The post is inductive and represents L_L while the radial disc is capacitive and represents C_d . The parallel resonance frequency of the circuit with respect to the terminals of the active device is determined by L_L , C_d plus the input impedance of the choke and the parasitic elements of the Gunn device. At the second harmonic the equivalent circuit [38] is shown in Figure 6.5.

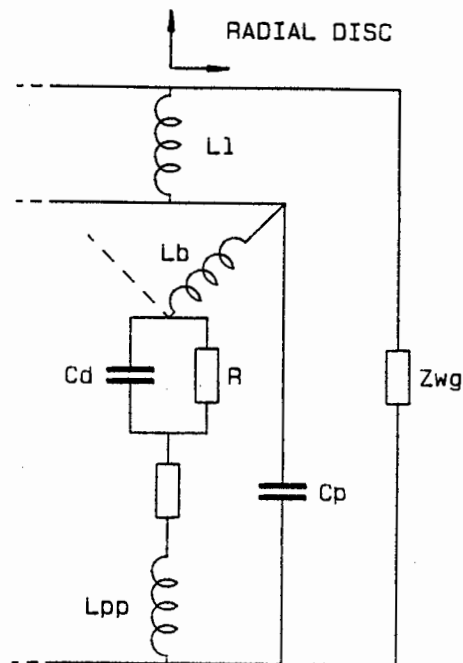


FIGURE 6.5 : Equivalent circuit of oscillator at $2f_0$.

The impedance transformation structure is that of the radial disc transformer [41-46].

6.4 EQUIVALENT CIRCUIT OF A GUNN DIODE

Figure 6.6 shows the schematic diagram of a Gunn package device with the parasitic reactance components and a dimensional diagram from a manufacturer's catalogue.

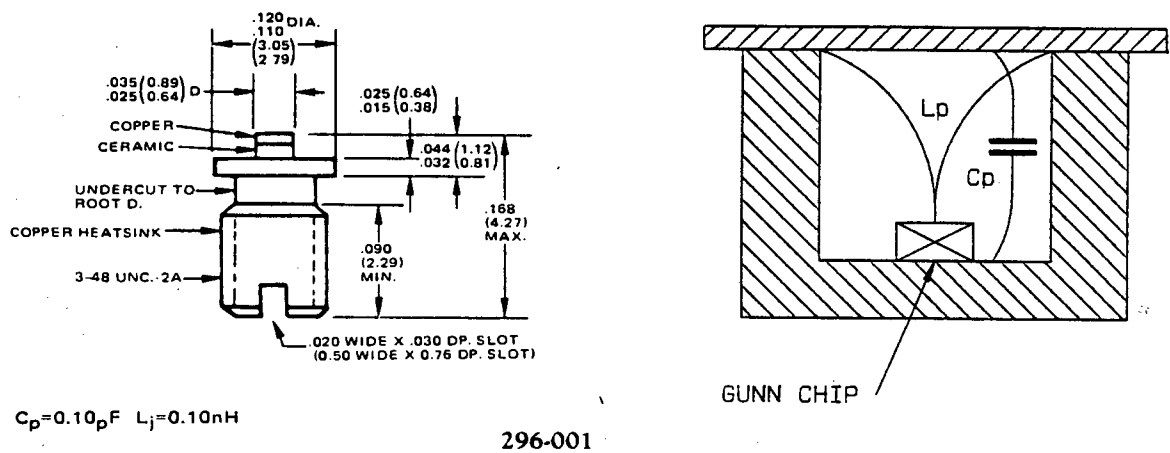


FIGURE 6.6 : Gunn device package and parasitic reactances.

Figure 6.7 shows a simple equivalent [37] circuit for the device.

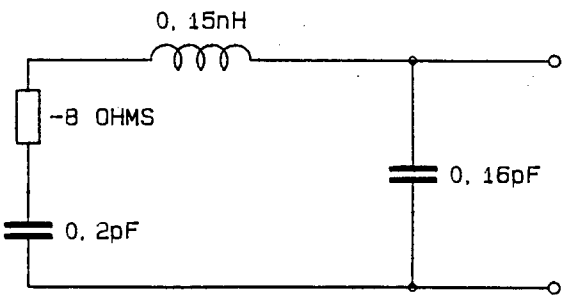


FIGURE 6.7 : Equivalent circuit of a Gunn device.

These circuit elements are approximate values, and there is no easy means of measuring these elements accurately. One of the principle problems in source design is the determination of the package circuit used in the design exercise. At lower microwave frequencies several techniques have been devised to obtain good approximation to the actual circuit model and fairly accurate values for the elements. The situation becomes more complicated at higher frequencies (mm-waves) where stray waveguide mount reactances, which are insignificant at low frequencies, further modify any package model. Package reactance studies are well documented at lower frequencies [47,48], but little information is available on the actual chip reactance and negative resistance. It is these values which are transformed through the package network to the package terminals which are of interest to the source designer. The effect of transformation considerably modifies the chip circuit values, so that accurate information on both chip and package element values are necessary for a realistic package terminal impedance. A further complication exists in the attempt to obtain package terminal impedances, and that is device chip and package variations. The manufacturer of any package is subject to mechanical and electrical tolerances, so variations in package circuit values will vary from batch to batch, as well as within a batch.

Package circuit models are, therefore, an approximation at frequencies above 40 GHz. The circuit in Figure 6.7 can be reduced to the one shown in Figure 6.8 after some mathematical manipulation.

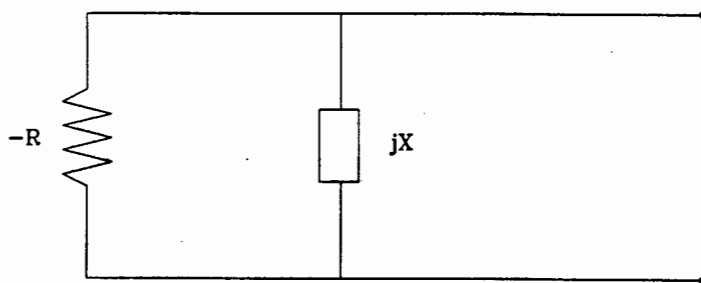


FIGURE 6.8 : Simplified equivalent circuit of a Gunn device.

The input impedance of the package device at 47 GHz is given by:

$$Z_{in} = -34,98 - j48,25 \, \Omega \quad (6.2)$$

6.5 BIAS CHOKE AND MATCHING NETWORK

The bias choke and matching network have to comply to several conditions for the oscillator to operate at the correct frequency and supply sufficient output power.

- 1) The circuit must offer positive load resistance smaller than or equal to the diode negative resistance for highest output power:

$$R_{\text{diode}}(f_o) + R_{\text{circuit}}(f_o) \leq 0 \quad (6.3)$$

- 2) The circuit must offer a reactance that is the conjugate of the device reactance at resonance:

$$X_{\text{diode}}(f_o) + X_{\text{circuit}}(f_o) = 0 \quad (6.4)$$

- 3) Bias current oscillations must not occur. Gunn devices exhibit a negative resistance at RF, but the RF voltage swing across the device induces a low frequency negative resistance which exists essentially from DC to several hundred MHz. Care in the design of the RF choke and DC power supply output must be exercised to ensure that these oscillations cannot exist.
- 4) Parasitic oscillations may exist if certain load impedance characteristics are presented to the diode at frequencies far from the operating frequency. This effect can also produce noise on the fundamental carrier.
- 5) In the design of the bias choke the diameter of the low-high-low impedance sections must be chosen such that no higher order mode can be excited or propagate, and that maximum attenuation is obtained at the oscillation frequency.

6.6 CALCULATION OF RADIAL DISC AND POST DIMENSIONS

Theoretical and experimental investigations [41-46] on the radial disc configuration indicate a susceptance of the disc B_D as shown in Figure 6.9.

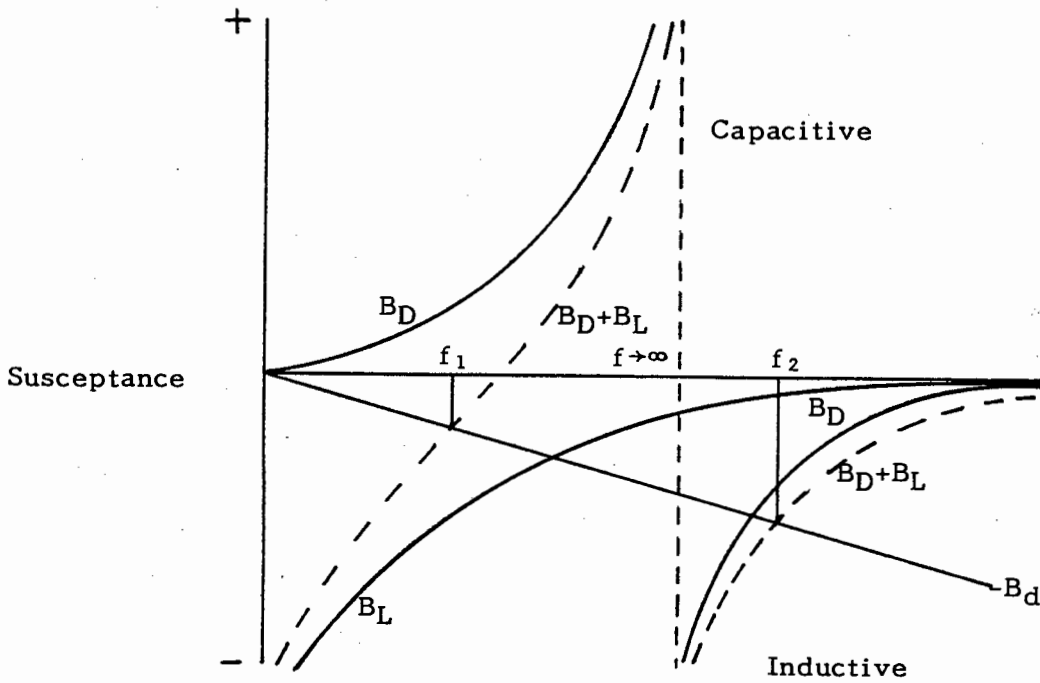


FIGURE 6.9 : Post and disc susceptance versus frequency (solution of equations 6.3 and 6.4).

The susceptance of the disc, as seen from the diode (package) terminals, is capacitive at low frequencies and inductive at high frequencies. The post that connects the disc to the bias filter has an inductance L_L , which transforms the total susceptance seen by the diode, as indicated by the line $B_D + B_L$ in Figure 6.9. This results in two oscillation conditions, one at f_1 and another at f_2 , that is at the intersection of the diode susceptance line B_D with the curve $B_D + B_L$. For very high values of L_L ($B_L \approx 0$), only one resonance exists at f_2 . To ensure that oscillations will take place at f_1 , condition (6.2) in Section 6.5 has to be satisfied.

The equivalent circuit of the disc and post is shown in Figure 6.10.

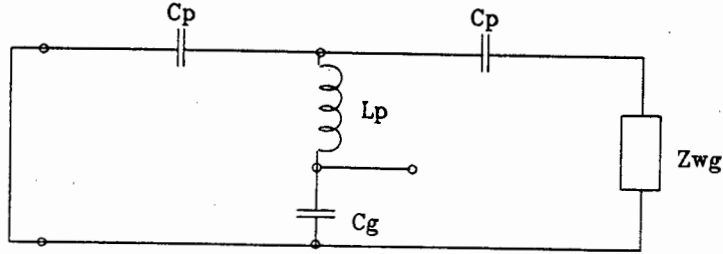


FIGURE 6.10 : Equivalent circuit of disc and post.

To determine the thickness, diameter and length of the disc and post for a specific frequency and output power is not an easy task, because there is a fair amount of interaction between the different dimensions. The length and diameter of the post, the thickness and diameter of the disc, and the height of the disc above the floor of the waveguide have an influence on the frequency and output power.

With an increase in disc thickness and diameter, the frequency and output power will increase, and for a decrease in post length and diameter the frequency and output power will increase.

To satisfy the oscillation conditions (6.3) and (6.4) in Section 6.5, the impedance of the disc and post combination should be

$$Z_{\text{disc}||\text{post}} \leq 35 + j48 \, \Omega \quad (6.5)$$

6.6.1 Impedance of a Post in a Waveguide

It was empirically found that with a post diameter of 0,6 mm the best performance was obtained, and therefore all the calculations will be based on this dimension.

The equivalent circuit of a post in a waveguide is shown in Figure 6.11, as determined by [49].

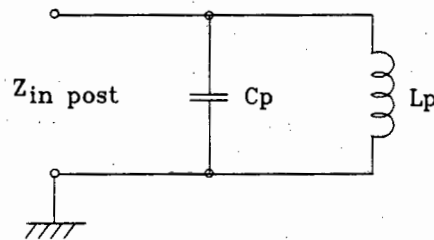


FIGURE 6.11 : Equivalent circuit of a post in a waveguide.

where L_p, C_p = post inductance and capacitance
 C_g = gap capacitance
 $Z_s = jZ_0 \tan \beta l$
 Z_L = waveguide load impedance

It can be shown that the input impedance of the post as seen from the diode, is given by:

$$Z_{in \text{ post}} = -\frac{j}{\omega C_g} \parallel \left[j\omega L_p - \frac{j}{\omega C_p} + Z_0 \right] \quad (6.6)$$

if $l = \lambda_g/4$

The imaginary part of equation (6.6) is

$$I_m Z_{in \text{ post}} = -\frac{j}{\omega C_g} \parallel \left[j\omega L_p - \frac{j}{\omega C_p} \right] \quad (6.7)$$

In [49] the element values of the equivalent circuit for a number of post diameters are given. Those values were obtained during measurements taken in WR62 (15,799 x 7,89 mm) waveguide.

The scale factor between the WR62 and WR10 (2,54 x 1,27 mm) waveguide is 6,22, thus for a post diameter of 0,6 mm in WR10 waveguide, the corresponding diameter in WR62 waveguide is 3,77 mm. The element values of the circuit in Figure 6.11 is obtained from [49] as:

$$L_p = 0,420 \text{ nH}$$

$$C_p = 0,415 \text{ pF}$$

$$C_g = C - C_1 (0,396 - 0,16 \text{ pF})$$

$$= 0,236 \text{ pF}$$

where C_1 = diode capacitance.

At 47 GHz the imaginary part of the input impedance of the post as calculated from equation (6.7) is

$$I_m Z_{in \text{ post}} = j16,5 \Omega (0,056 \text{ nH}) \quad (6.8)$$

The inductance of a piece of wire is given by [50] as:

$$L = 0,452 \ell \log_{10} \left(\frac{4\ell}{d} - 0,75 \right) \quad (6.9)$$

ℓ = length of post in mm.

d = diameter of post in mm.

L = inductance in nH.

By using equation (6.9) the length of the post is computed as

$$L = 0,405 \text{ mm}$$

It must be remembered that all the dimensions obtained are first order approximations, and have to be optimized empirically.

6.6.2 Impedance of a Radial Disc

From equation (6.5) the parallel combination of the disc and post impedance should be

$$Z_{\text{disc} \parallel \text{post}} = + j48,25 \, \Omega = - \frac{jX_{\text{disc}} \cdot jX_{\text{post}}}{jX_{\text{disc}} + jX_{\text{post}}} \quad (6.10)$$

Substituting equation (6.8) in (6.10) give

$$X_{\text{disc}} = - j25,2 \, \Omega \quad (6.11)$$

Knowing the disc impedance, all that remains is to determine the disc diameter and thickness.

The radial disc geometry is shown in Figure 6.12.

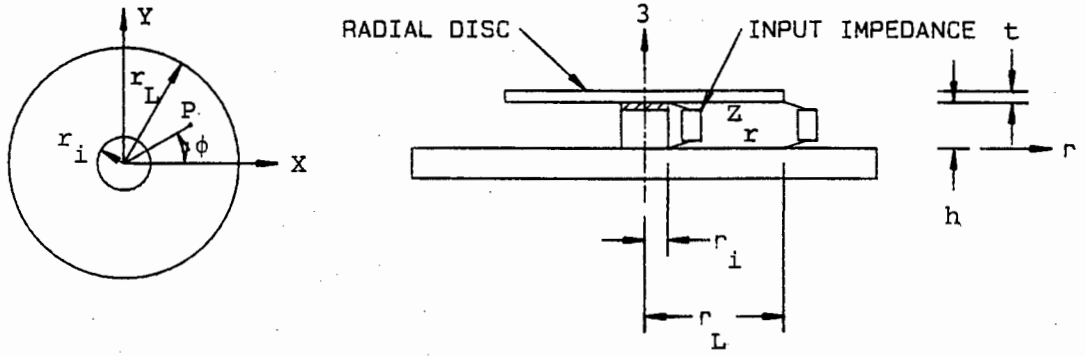


FIGURE 6.12 : Geometry of radial disc.

From [46] the wave input impedance is given by:

$$\bar{Z}_T = - \frac{h \bar{E}_z(r)}{2\pi r_i \bar{H}_\phi(r)} = - \frac{h}{2\pi r_i} \bar{Z}_i(r) \quad (6.12)$$

where the input wave impedance is

$$\begin{aligned} \bar{Z}_i &= \frac{\bar{E}_z(r_i)}{\bar{H}_\phi(r_i)} \\ &= Z_{oi} \left[\frac{Z_L \cos(\theta_i - \psi_L) + jZ_{oL} \sin(\theta_i - \theta_L)}{Z_{oL} \cos(\psi_i - \phi_L) + jZ_L \sin(\psi_i - \psi_L)} \right] \end{aligned} \quad (6.13)$$

where the load wave impedance is

$$Z_L = \frac{E_z(r_L)}{H_z(r_L)} \quad (6.14)$$

and the impedance is

$$Z_o(kr) = \sqrt{\frac{\mu_o}{\epsilon_o \epsilon_r}} \sqrt{\frac{J_o^2(kr) + N_o^2(kr)}{J_1^2(kr) + N_1^2(kr)}} \quad (6.15)$$

where $J_n(kr)$ are Bessel functions of the first kind of order n .
 $N_n(kr)$ are Newman's functions of the second kind of order n .

$$k = \omega \sqrt{\epsilon_r \epsilon_o \mu_o}$$

$$\theta(kr) = \tan^{-1} \left[\frac{N_o(kr)}{J_o(kr)} \right]$$

$$\psi(kr) = \tan^{-1} \left[-\frac{J_1(kr)}{N_1(kr)} \right]$$

for $Z_L \gg X_{oL}$ (ideally $Z_L \rightarrow \infty$).

$$\bar{Z}_T = j \frac{h}{2\pi r_i} Z_o(kr_i) \frac{\cos[\theta(kr_i) - \psi(kr_L)]}{\sin[\psi(kr_i) - \psi(kr_L)]} \quad (6.16)$$

where $\mu_o = 4\pi \times 10^{-7}$ H/m.

$\epsilon_o = 8.85 \times 10^{-12}$ F/m.

$\epsilon_r = 1$

L = height of disc above floor of waveguide.

T = thickness of radial disc.

For a first order approximation the Bessel functions can be written as:

$$J_0(x) = 1 - \frac{x^2}{2} + \frac{x^4}{64} - \frac{x^6}{2304} \quad (6.17)$$

$$J_1(x) = \frac{x}{2} - \frac{x^3}{16} + \frac{x^5}{384} - \frac{x^7}{18432} \quad (6.18)$$

and Newman's functions as:

$$N_0(x) = \frac{2}{\pi} \left(\ln \frac{x}{2} + \gamma \right) J_0(x) + \frac{2}{\pi} \left[\frac{x^2}{4} - \frac{1,5 x^4}{64} + \frac{1,833 x^6}{2304} \right] \quad (6.19)$$

$$N_1(x) = \frac{2}{\pi} \left(\ln \frac{x}{2} + \gamma \right) J_1(x) + \frac{2}{\pi x} \left[\frac{x}{2} - \frac{2,5 x^3}{16} + \frac{3,33 x^5}{284} \right] \quad (6.20)$$

and $\gamma = 0,5772$

A computer program has been written to compute the input impedance of the radial disc, see Appendix 1, with a graphic representation in Figure 6.13. For a disc impedance of $-j25,2 \Omega$ the disc diameter is calculated from Figure 6.13 as

$$D = 2,26 \text{ mm}$$

From [51] an approximation is derived which states that

$$\text{Disc diameter} = \lambda_g / 2 \text{ at } 2f_0$$

which gives $D = 2,05 \text{ mm}$.

With all the assumptions that have been made in determining the diameter of the disc, there is good agreement between the method just described and [51].

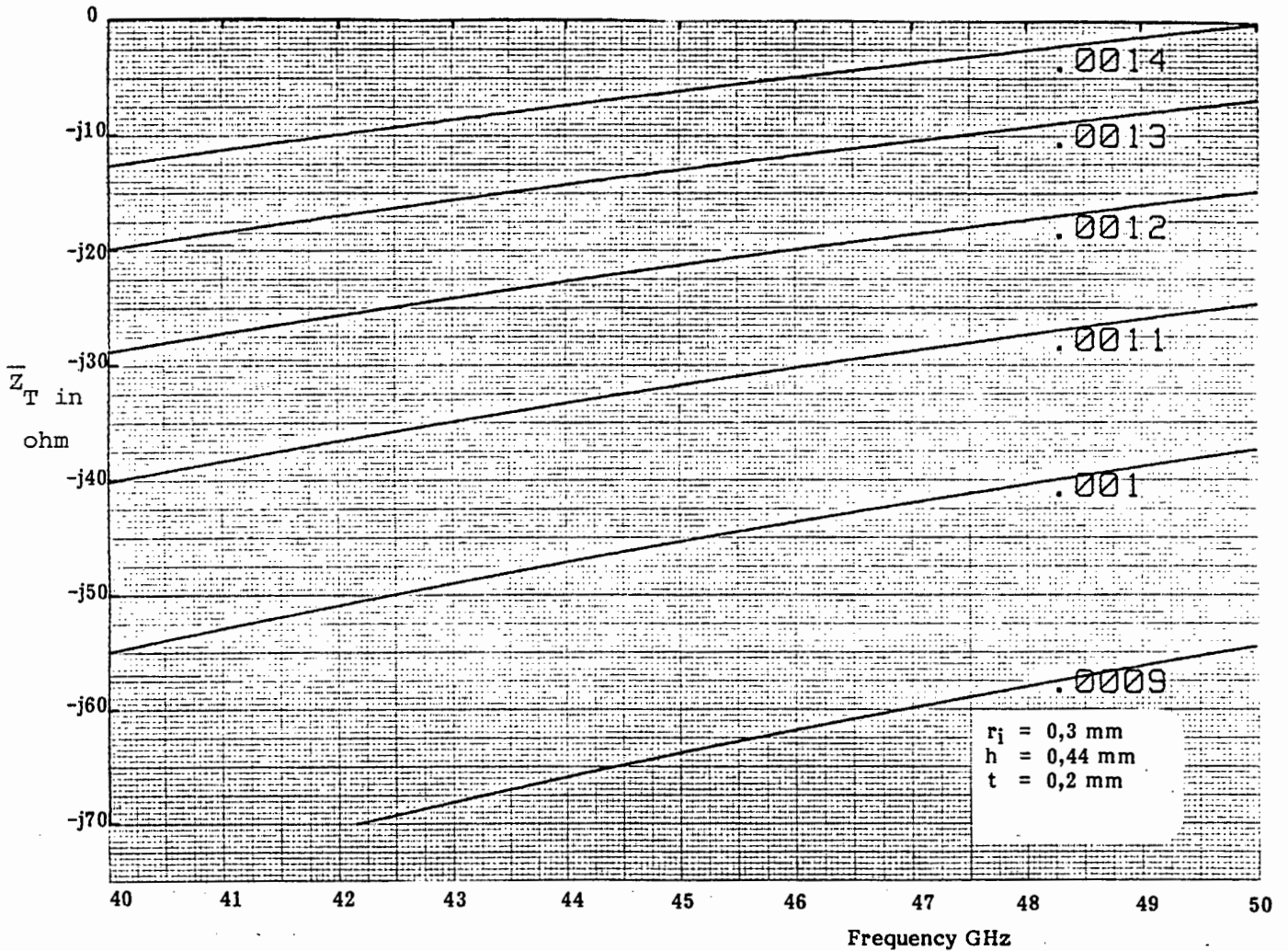


FIGURE 13 : Input impedance of Radial Disc

FIGURE 6.13 : Input impedance of radial disc.

6.7 DESIGN OF THE BIAS CHOKE

To satisfy condition (6.3) in Section 6.5, it is necessary that certain precautions have to be taken to prevent propagation of higher order modes. It is, therefore, important to obtain values for a and b , the diameters of the low and high impedance sections, that will prevent the excitation of such higher modes.

To comply with this condition it is necessary that:

- 1) The mean value of the circumference of a and b must be equal to the cut-off frequency of the shaft where the filter fits in.
- 2) The cut-off frequency must be higher than the fundamental frequency.

$$\therefore \lambda_c = \left(\frac{d}{2} + \frac{D}{2} \right) \quad (6.21)$$

choose $f_c = 56 \text{ GHz.}$

$$a = 0,6 \text{ mm.}$$

and $b = 2,8 \text{ mm.}$

With the values of a and b fixed, the lengths of the different sections of the filter (A, B, C, D and E) have to be chosen such that no signal above a certain frequency propagate to the bias connection. The impedance values of the sections are:

$$Z_L = \frac{60}{\sqrt{\epsilon_r}} \ln (F/b) \quad (6.22)$$

$$Z_H = \frac{60}{\sqrt{\epsilon_r}} \ln (F/a) \quad (6.23)$$

For isolation purposes the filter is covered with a teflon tube ($\epsilon_r = 2,2$), which gives

$$Z_L = 2,8 \, \Omega$$

$$Z_H = 96,5 \, \Omega$$

The cut-off frequency of the filter was chosen to be 9 GHz. It is not necessary to have a filter with low ripple in the passband, because the transfer characteristics will be influenced by non 50 Ω loads. From [52] for a fifth order filter with 3 dB ripple in the passband, the prototype is shown in Figure 6.14.

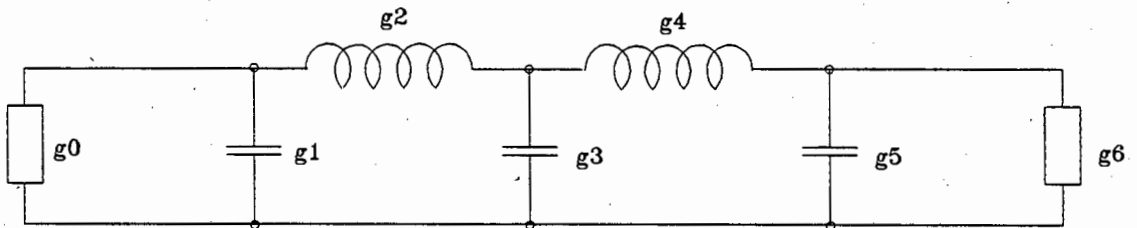


FIGURE 6.14 : Filter prototype.

After scaling in frequency and impedance, the filter network shown in Figure 6.15 is obtained.

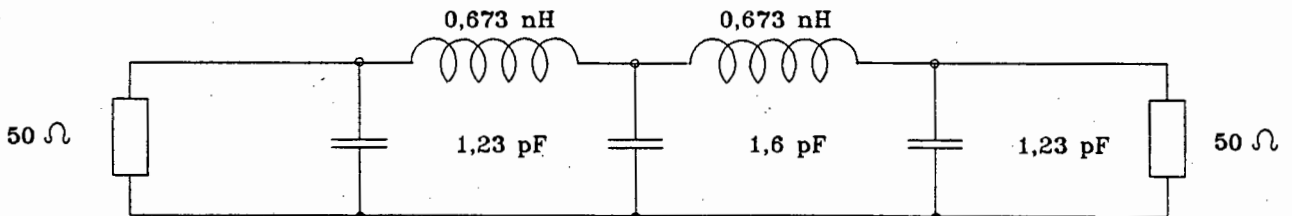


FIGURE 6.15 : Filter prototype after scaling.

From [51] the lengths of the different sections can be computed as:

$$\omega L = Z_h \sin\left(\frac{\omega l}{c}\right) \quad (6.24)$$

$$\omega C = \frac{Y_L l \omega}{c/\sqrt{\epsilon_r}}$$

With this type of filter design approximately 2 dBm more output power was obtained compared with the conventional $\lambda_g/4$ sections at 47 or 94 GHz. The same results were obtained by [53].

The dimensions shown in Figure 6.3 are as follows:

$$a = 0,6 \text{ mm}$$

$$b = 0,8 \text{ mm}$$

$$d = 0,6 \text{ mm}$$

$$D = 2,05 \text{ mm}$$

$$t = 0,2 \text{ mm}$$

$$L = 0,405 \text{ mm}$$

$$A = 0,7 \text{ mm}$$

$$B = 0,9 \text{ mm}$$

$$C = 1,8 \text{ mm}$$

$$E = 1,8 \text{ mm}$$

$$G = 0,7 \text{ mm}$$

6.8 SLIDING SHORT

The sliding short is designed according to [37] and is shown in Figure 6.16.

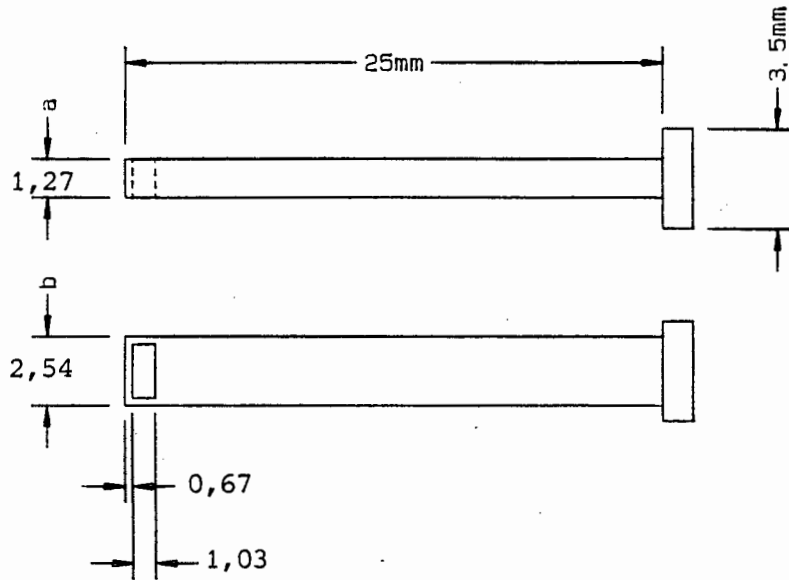


FIGURE 6.16 : Sliding back short.

A refinement was carried out by anodizing the short for isolation purposes. The short is normally placed $\lambda_g/4$ from the diode. The position of the short determines only the output power and has no influence on the frequency. The short is adjustable to maximize the output power.

6.9 MEASURED RESULTS

Figure 6.17 shows the output power versus frequency. The specifications of the diode as given by the manufacturer is as follows:

$$P_{\text{out}} = 20 \text{ dBm}$$

$$f_o = 50,56 \text{ GHz}$$

As can be seen from Figure 6.17, the maximum output power as the second harmonic is 9 dBm, which is 11 dB below the output power at the fundamental frequency as was determined in Section 6.2.

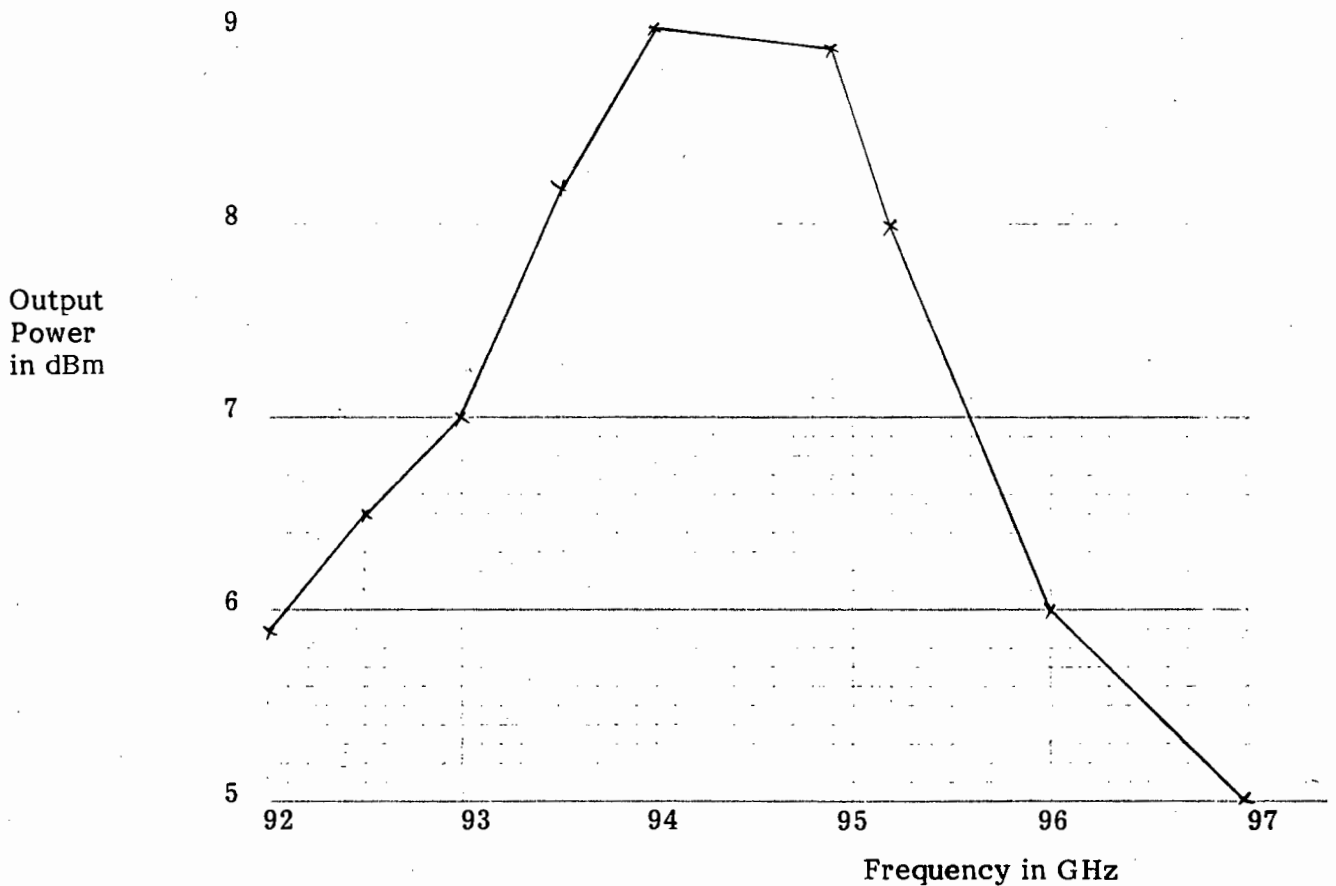


FIGURE 6.17 : Output power versus frequency of Gunn oscillator.

The frequency is adjusted by changing the height of the diode in the waveguide. The direct current source as well as the back short were adjusted, for each different height position of the diode, to maximize the output power. The spectrum is shown in Figure 6.18.

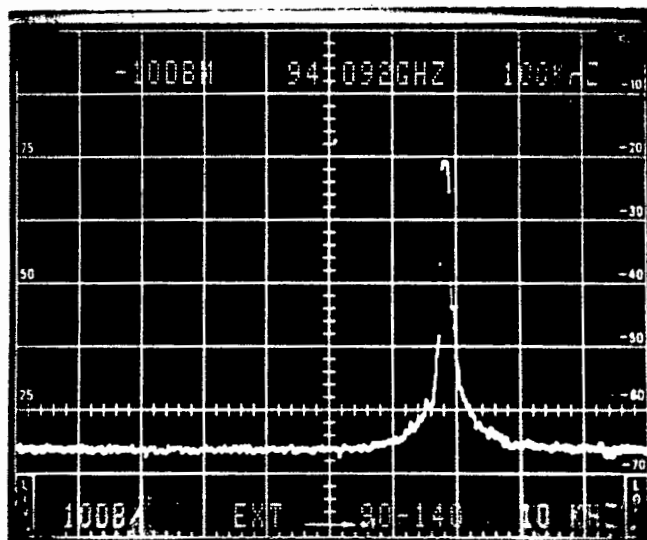


FIGURE 6.18 : Spectrum of Gunn oscillator.

6.10 PHASE NOISE MEASUREMENTS

The phase noise of the source was measured using the method described in Chapter 5. The phase noise measurements are shown in Figure 6.19.

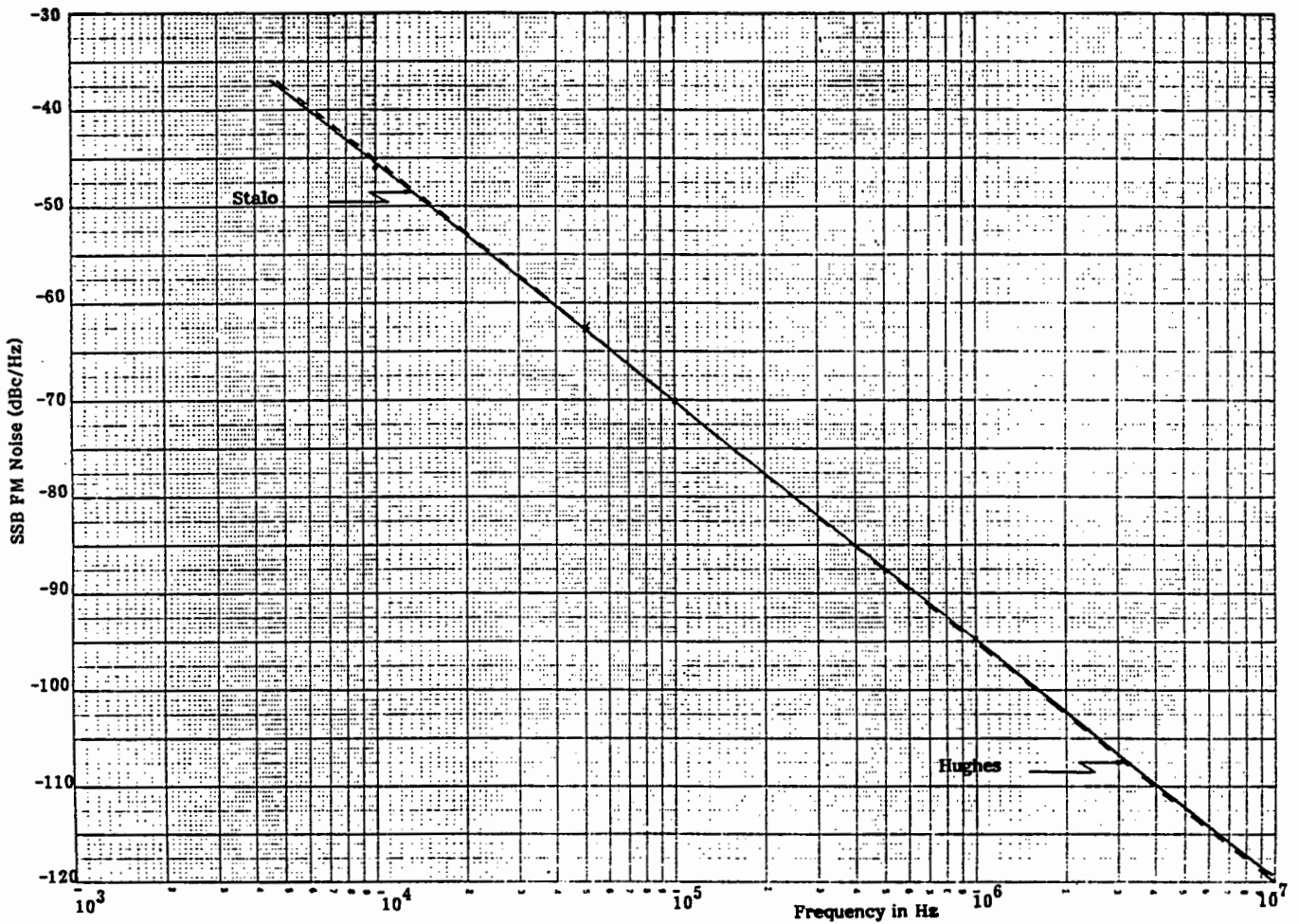


FIGURE 6.19 : Single sideband phase noise of the Gunn oscillator.

Published phase noise of a free running Hughes oscillator is also shown in Figure 6.19. There is no difference in the published data and that of the oscillator developed here and reported in this thesis. Figure 6.20 shows the spectrum of the source while it was modulated with a 500 kHz signal and the modulation index $m = 2,405$.

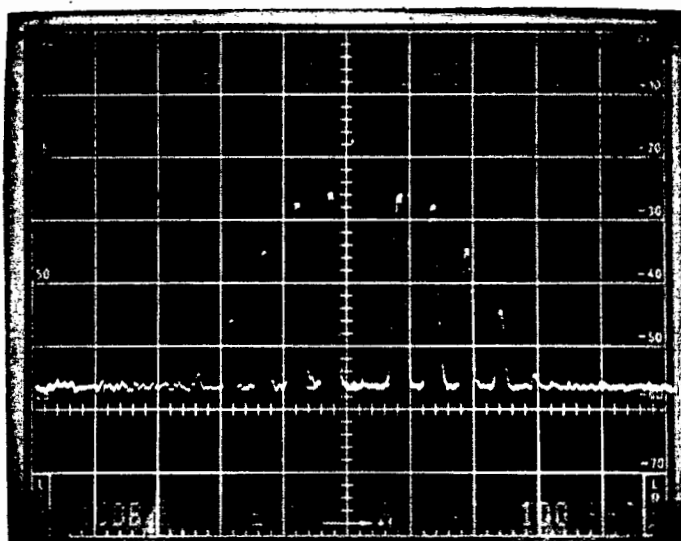


FIGURE 6.20 : Spectrum of modulated signal.

6.11 MANUFACTURING PROBLEMS

It is very difficult to manufacture the different mechanical parts, because they are very small and have tight tolerances. A special lathe and a well trained precision toolmaker is a necessity for the manufacture of such a source. Figure 6.21 shows a cross section of the assembled unit.

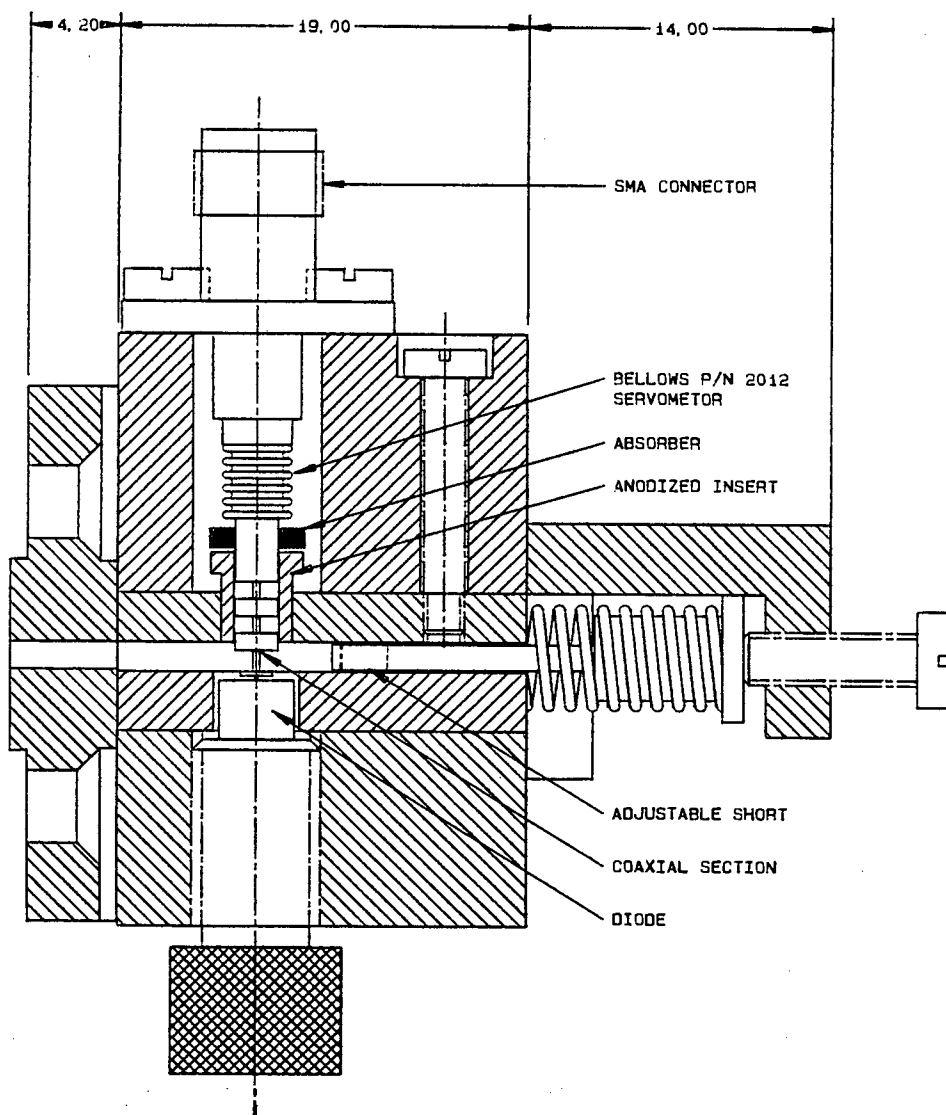


FIGURE 6.21 : Detailed cross section of Gunn source.

APPENDIX 6.1 : COMPUTER PROGRAM TO COMPUTE Z_T .

```

10      ! INPUT IMPEDANCE OF RADIAL DISC
20      DIM J0ri(1:101),J1ri(1:101),N0ri(1:101),N1ri(1:101)
30      PLUTTER IS 705,"HPGL"
40      PEN 2
50      Eps_r=1.
60      Freq_min=4.0E+10
70      Freq_max=5.0E+10
80      Freq_step=1.E+8
90      WINDOW Freq_min,Freq_max,-70,0
100     AXES 1.E+9,10,4.0E+10
110     PEN 1
120     Rad_in=3.0E-4
130     Height=4.44E-4
140     N=1
150     FOR Freq=Freq_min TO Freq_max STEP Freq_step
160         Prop_k=2*PI*Freq/3.E+8
170         CALL Bessel(Prop_k,Rad_in,J0ri(N),J1ri(N),N0ri(N),N1ri(N))
180         N=N+1
190     NEXT Freq
200     I=1
210     SELECT I
220     CASE 1
230         Rad_out=1.000E-3
240         GOSUB 410
250     CASE 2
260         Rad_out=1.100E-3
270         GOSUB 410
280     CASE 3
290         Rad_out=1.20E-3
300         GOSUB 410
310     CASE 4
320         Rad_out=1.3E-3
330         GOSUB 410
340     CASE ELSE
350         PENUP
360         PEN 0
370         STOP
380     END SELECT
390     I=I+1
400     GOTO 210
410     N=1
420     FOR Freq=Freq_min TO Freq_max STEP Freq_step
430         Prop_k=2*PI*Freq/3.E+8
440         CALL Bessel(Prop_k,Rad_out,J0ro,J1ro,N0ro,N1ro)
450         Imp=376.81944/SQR(Eps_r)*Height/2/PI/Rad_in
460         Imp=-Imp*(J1ro*N0ri(N)-N1ro*J0ri(N))/(J1ri(N)*N1ro-N1ri(N)*J1ro)
470         PLOT Freq,Imp,1
480         N=N+1
490     NEXT Freq
500     PENUP
510     IPLOT -2.E+9,-5.,-2
520     LABEL Rad_out
530     PENUP
540     RETURN
550     END
560     SUB Bessel(K,R,J0,J1,N0,N1)
570     ! SUBPROGRAM TO COMPUTE THE J0, J1, N0, N1 BESSEL FUNCTIONS WITH ARGUMENT
K*K
580     Arg=K*K
590     J0=1.
600     Kk=1.

```

```

610  FOR I=1 TO 30
620      Kk=Kk/(2*I)^2
630      Jj=(-1)^I*(Arg^(2*I))*Kk
640      J0=J0+Jj
650      IF ABS(Jj)<1.E-6*ABS(J0) THEN 680
660  NEXT I
670  PRINT "J0 DID NOT CONVERGE IN 30 ITERATIONS ",J0,Jj
680  J1=0
690  Kk=1
700  FOR I=1 TO 30
710      Ii=2*I-1
720      IF I=1 THEN
730          Kk=Kk/2
740      ELSE
750          Kk=Kk/(2*I)/(2*I-2)
760      END IF
770      Jj=(-1)^I*(-1)*Arg^Ii*Kk
780      J1=J1+Jj
790      IF ABS(Jj)<1.E-6*ABS(J1) THEN 820
800  NEXT I
810  PRINT "J1 DID NOT CONVERGE IN 30 TERMS ",J1,Jj
820  N0=2./PI*(LOG(Arg/2)+.57721566)*J0
830  N00=0
840  Kk=1.
850  Kkk=0.
860  FOR I=1 TO 30
870      Kk=Kk/((2*I)^2)
880      Kkk=Kkk+1./I
890      N0i=-(-1)^I*Arg^(2*I)*Kk*Kkk
900      N00=N00+N0i
910      IF ABS(N0i)<1.E-6*ABS(N00) THEN 940
920  NEXT I
930  PRINT "N0 DID NOT CONVERGE IN 30 ITERATIONS ",N00,N0i
940  N0=N0+2/PI*N00
950  N1=2/PI*(LOG(Arg/2)+.57721566)*J1
960  N11=Arg/2-2.5*Arg^3/16+3.3333333*Arg^5/284-3.91663*Arg^7/18432
970  N1=N1-2/(PI*Arg)-1/PI*N11
980  SUBEND

```

CHAPTER 7

THE DESIGN OF A 94 GHz PULSED IMPATT OSCILLATOR

7.1 INTRODUCTION

There is a considerable interest in using solid-state millimeter-wave pulsed sources as transmitters in active seekers for missile applications. Silicon Impatt diodes are presently the only solid-state device capable of producing high power output at millimeter-wave frequencies. Direct fundamental frequency power generation allows for simple circuit design, thus reducing size and weight of the transmitter, which are important considerations for seeker applications.

This chapter describes the design and development of a Pulsed Impatt source at 94 GHz. A detailed description of the design procedure of the source and pulse modulator will be given as well as results achieved.

7.2 GENERAL CHARACTERISTICS OF PULSE IMPATT OSCILLATORS

A short pulse (<100 ns), low duty cycle (50 kHz) Impatt oscillator has been developed which produce 5 Watts peak power at 94 GHz, using a double dirft silicon diode mounted in a hermetically sealed package on a copper heatsink. Operated in the pulse mode, the device is biased at power levels exceeding 200 Watts (8 A at 35 V) peak. Far more power is dissipated in the Silicon junction during the short pulse than the 5 to 10 Watts which can be dissipated safely on a CW basis. A junction temperature increase of about 200°C takes place during the 100 ns bias pulse.

In the case of a rectangular bias current pulse, this thermal transient causes a variation of the device impedance and of the noise level during the pulse. This causes frequency chirp, or in the case of an injection locked oscillator intra-pulse phase ripple.

To some extent the frequency chirp or phase ripple can be reduced by shaping the bias current pulse. However, experience indicates that the higher power pulsed Impatts are more difficult to control in this respect and the range of compensation is more limited.

Another significant characteristic peculiar to pulsed oscillators are the finite rise and fall time of the bias current pulse. During the leading and trailing edges of the bias current pulse of a free running oscillator the noise level is high and there are large transients in the output frequency. In the case of an injection locked oscillator the edges of the pulse are unlocked with high phase noise. With typical rise and fall times of the bias current pulse between 10 to 20 ns, the leading and trailing edge of the output pulse, nominally 60 ns, are not coherent. This leaves a pulse of between 20 to 40 ns which is coherent. For certain applications this is not acceptable in which case p-i-n switches may be used to blank the non-coherent edges of the pulse. There is a trade off that has to be considered, i.e. between the insertion loss of the p-i-n switch and the reduction of phase noise on the edges of the pulse.

7.3 SOURCE CONSTRUCTION

Figure 7.1 shows a cross section of the pulsed Impatt oscillator.

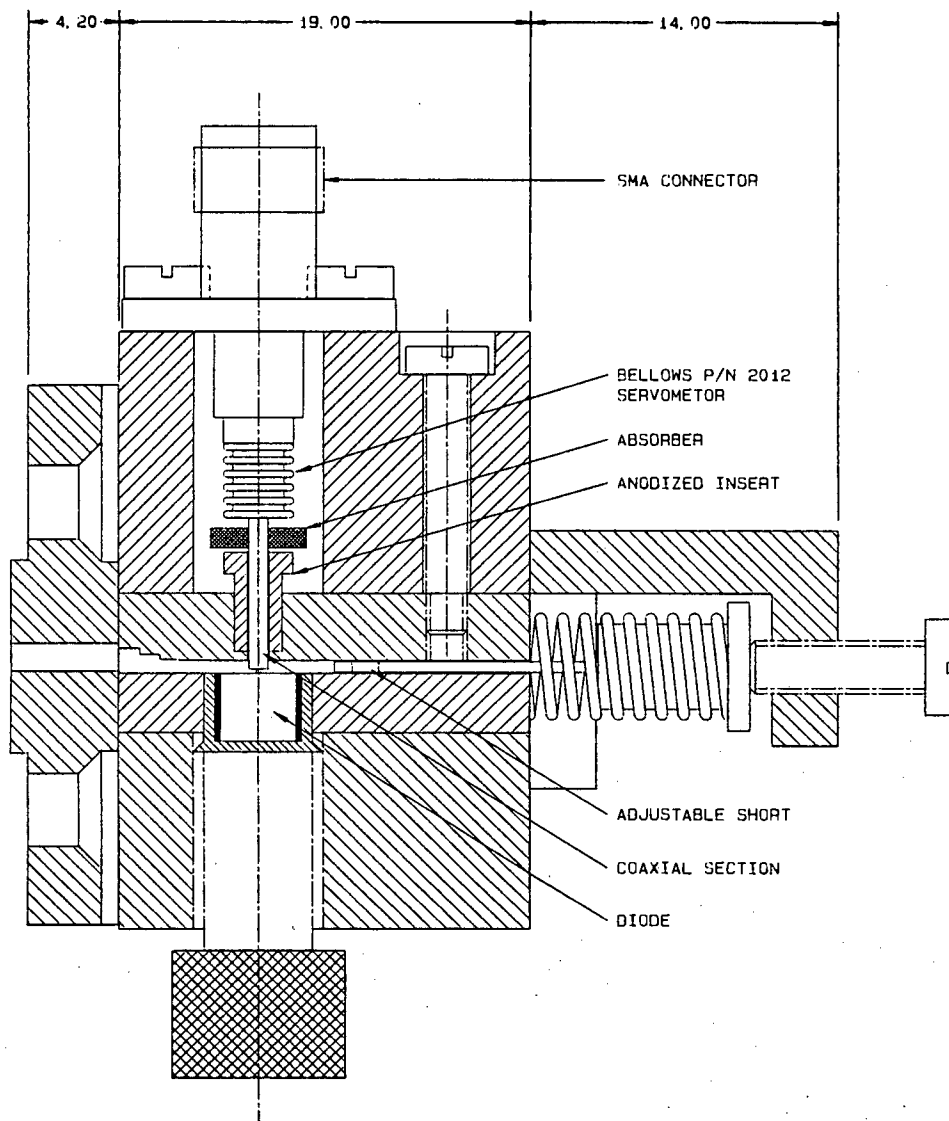


FIGURE 7.1 : Cross section of the pulsed Impatt oscillator.

The diode is mounted in the broad wall of a reduced height WR-10 waveguide, with an anodized (isolated) sliding back-short to the one side of it, which allows for frequency and power tuning.

The diode package is soldered into a copper collet, which acts as a heatsink, which is located directly under the centre conductor of a coaxial line. The centre conductor of the coaxial line also acts as the bias pin. Besides the impedance matching of the sliding short, additional impedance matching is obtained by reducing the waveguide height and by changing the post radius r_i and length l_c of the coaxial section, see Figure 7.6. An anodized aluminium insert acts as the termination for the coaxial section with a disc of absorbing material fitted on top of it, around the bias pin. The bias pin is pressed down on the diode with the bellows which are soldered onto a SMA-connector through which the bias pulse is fed. The Q of the circuit is relatively low, which causes the oscillator characteristics to be very sensitive to external load changes. The position of the anodized aluminium insert (l_c) relative to the reduced height waveguide also determines the operating frequency of the oscillator. To lower the frequency the insert has to protrude into the waveguide and vice versa for an increase in frequency.

7.4 EQUIVALENT CIRCUIT OF A PULSED IMPATT DIODE

A typical package of an Impatt diode is shown in Figure 7.2, obtained from [54].

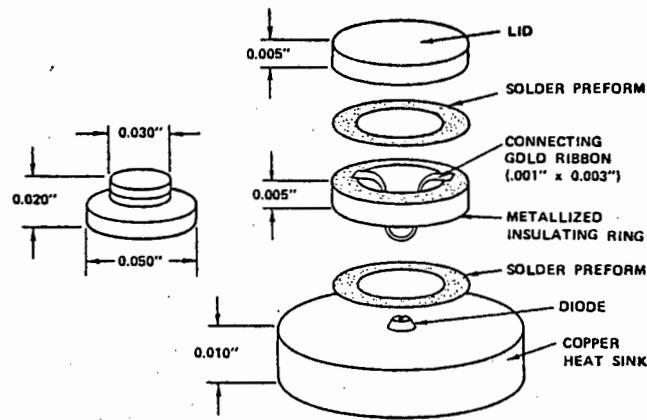


FIGURE 7.2 : Details of a millimeter-wave Impatt diode package.

The small signal admittance is shown in Figure 7.3 from [55].

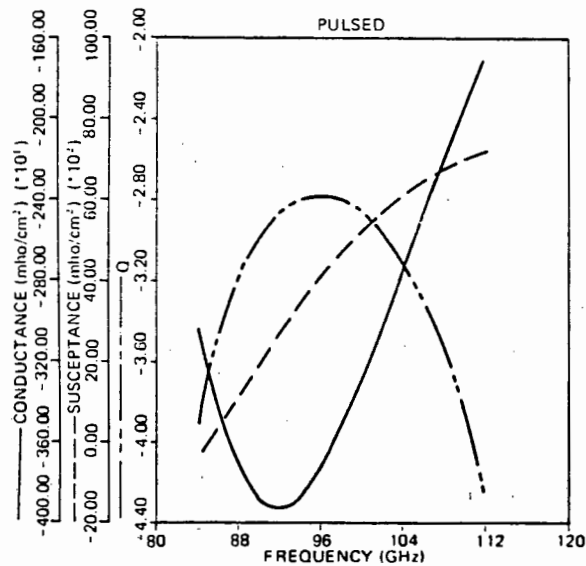


FIGURE 7.3 : Small signal admittance of a pulsed Impatt Diode.

Normally the small signal characteristics of the diode impedance are used to obtain a first order approximation of the matching network and resonator circuit. Once a first order approximation is obtained, the source can be optimized empirically.

Because the small signal impedance differs considerably from the actual large signal impedance, it is better to use the calculated large signal impedance [56], as a first order approximation. The equivalent circuit of the chip and package is shown in Figure 7.4.

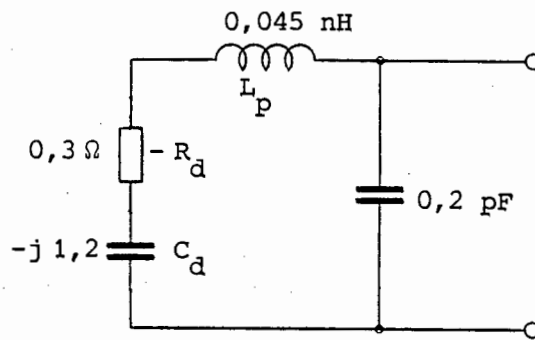


FIGURE 7.4 : Equivalent circuit of pulsed Impatt chip and package.

The circuit of Figure 7.4 can be simplified to the one shown in Figure 7.5.

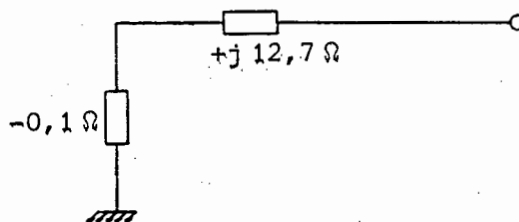


FIGURE 7.5 : Equivalent circuit as seen from the package terminals.

This low impedance of the package device has to be matched to the high impedance of the waveguide.

7.5 RESONATOR CIRCUIT

The bias pin, sliding short, length of coaxial section and the reduced height cavity have to comply with several conditions to ensure that the desired frequency and output power are obtained. The conditions for oscillation are:

- 1) The resonator circuit must offer positive load resistance smaller than or equal to the diode negative resistance for highest output power.

$$R_{\text{diode}}(f_o) + R_{\text{circuit}}(f_o) \leq 0$$

- 2) The resonator circuit must offer a reactance that is the conjugate of the device reactance at resonance:

$$X_{\text{diode}}(f_o) + X_{\text{circuit}}(f_o) = 0$$

The resonator circuit can be broken up into four different sections that have an influence on the matching:

- 1) The reduced height waveguide;
- 2) the bias pin;
- 3) the length of the coaxial section;
- 4) and the sliding back-short.

Each of these sections will be described separately to show their influence on the matching process.

7.5.1 Reduced Height Waveguide Transformer

As was calculated in paragraph 7.4, the small signal impedance of the diode is fairly low (3Ω), under large signal conditions as in the case of a pulsed oscillator, this impedance can be an order of magnitude smaller ($0,3 \Omega$). For the bias pin, sliding short and coaxial section to do the transformation in one step from 377Ω (full height waveguide) to the diode impedance ($0,3 \Omega$) is an enormous step. It was, therefore, decided to reduce the height of the waveguide to $0,51 \text{ mm}$ (160Ω), which makes it easier to do the matching.

The lengths and heights of the steps are determined according to [57]. Figure 7.6 shows a schematic diagram of the stepped transformer in the resonator circuit.

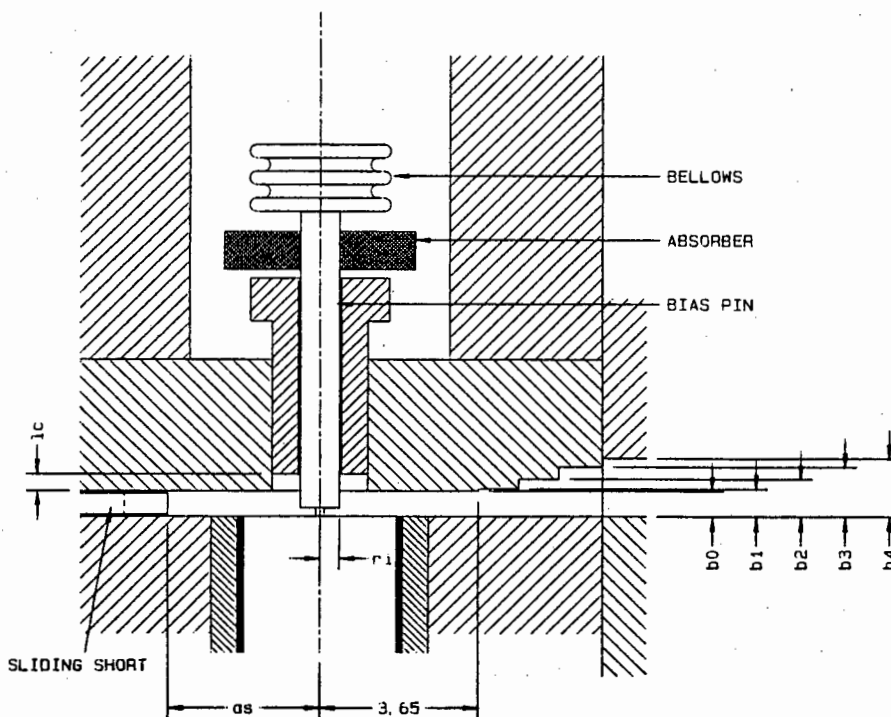


FIGURE 7.6 : Resonator circuit.

Let the frequency band over which the transformation has to take place, be 90-100 GHz. (f_1 and f_2 respectively). The fractional bandwidth is:

$$W_q = \frac{2(\lambda_{g_1} - \lambda_{g_2})}{(\lambda_{g_1} + \lambda_{g_2})} \quad (7.1)$$

where

$$\lambda_g = \frac{\lambda_o}{\sqrt{1 - \left(\frac{\lambda_o}{\lambda_c}\right)^2}} \quad (7.2)$$

$$\lambda_c = 2a \text{ for a TE}_{10} \text{ mode} \quad (7.3)$$

$a = 2,54$ mm, the width of the reduced height waveguide which is the same as the standard waveguide.

By substituting f_1 and f_2 into equations (7.1), (7.2) and (7.3) it is found that:

$$W_q \approx 0,2$$

$$\lambda_{g_1} = 4,4 \text{ mm}$$

$$\lambda_{g_2} = 3,7 \text{ mm}$$

$$\lambda_c = 5 \text{ mm}$$

The length L of each step is given by:

$$L = \frac{\lambda_{g_1} \lambda_{g_2}}{2(\lambda_{g_1} + \lambda_{g_2})} \quad (7.4)$$

$$= 1 \text{ mm}$$

The impedance ratio is:

$$R = \frac{b_4}{b_0} \quad (7.5)$$

$$= 2,5$$

From Tabel 6.04-2 [57], the normalized impedances for a three section Tchebychev quarter-wave transformer are:

$$Z_1 = 1,12422$$

$$Z_2 = 1,58114$$

$$Z_3 = 2,2237$$

The different step heights are:

$$b_0 = 0,51 \text{ mm}$$

$$b_1 = Z_1 \cdot b_0 = 0,57 \text{ mm}$$

$$b_2 = Z_2 \cdot b_0 = 0,81 \text{ mm}$$

$$b_3 = Z_3 \cdot b_0 = 1,134 \text{ mm}$$

$$b_4 = 1,27 \text{ mm}$$

7.5.2 Bias Pin and Reduced Height Cavity

A detailed theoretical description of a pin in a waveguide is given by [58] for frequencies up to 75 GHz. This information was used by [56] to compute the impedance of a pin in a reduced height waveguide as a function of pin diameter and waveguide height with the effect of the sliding short not taken into account. Figure 7.7 shows the real and imaginary parts of the impedance.

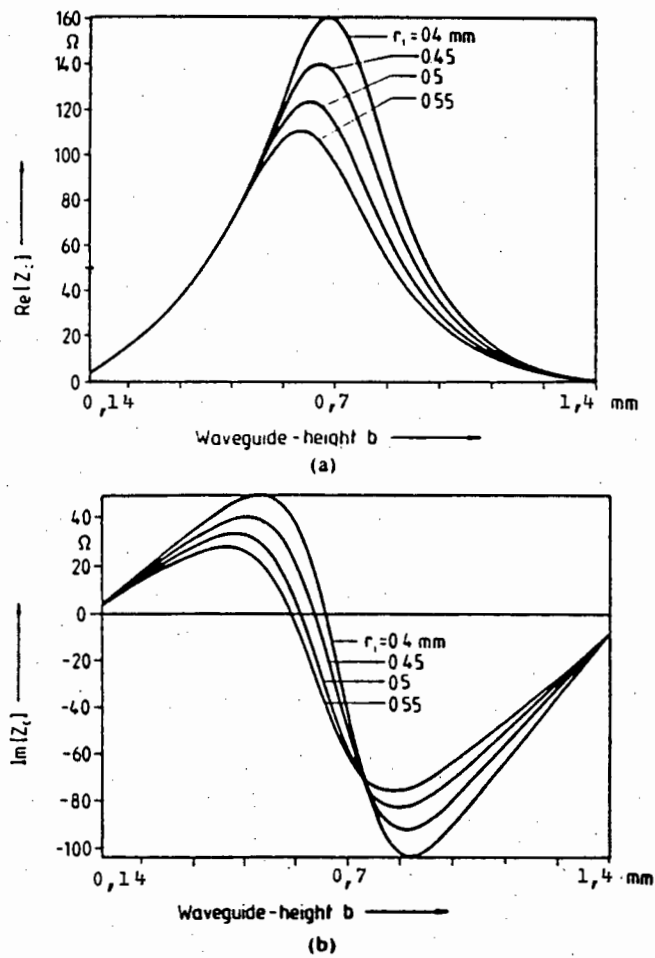


FIGURE 7.7 : (a) Real part and (b) imaginary part of circuit impedance, as a function of pin diameter and waveguide height.

Relatively small values for the real and imaginary parts of the circuit impedance Z_c are obtained if the waveguide height is either strongly reduced or nearly unreduced. For medium reduction, a resonance occurs, i.e. $\text{Re}[Z_c]$ becomes relatively large and $\text{Im}[Z_c]$ changes from inductive to capacitive. Except at resonance, the diameter of the inductive post has little influence on the circuit impedance.

With a reduced waveguide height of 0,51 mm and a bias pin diameter of 0,88 mm, the resonator circuit impedance is approximately

$$Z_c = 90 + j 39 \Omega$$

This impedance is lower than the reduced height waveguide impedance, but still far from the conjugate value of the diode impedance, and therefore some extra transformation has to be done.

7.5.3 Sliding-back-short

The sliding back short was designed according to [54] and is shown in Figure 7.8.

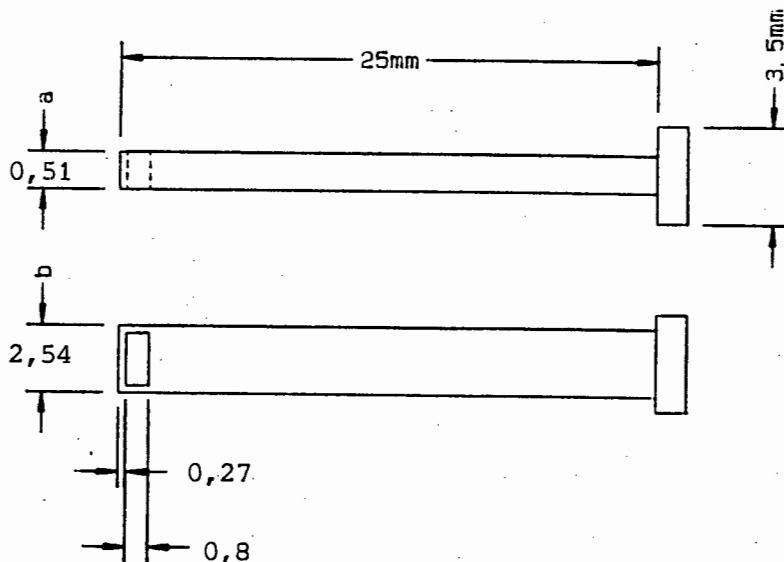


FIGURE 7.8 : Sliding back short.

Calculating the impedance of the resonator circuit becomes difficult if the location of the sliding short and the length of the coaxial section are taken into account. The Smith Chart in Figure 7.9 shows the tuning behaviour of the resonator for a reduced height waveguide (0,5 mm) and a bias pin of 1 mm [56]. The circles show typical behaviour of periodic tuning by sliding the short for different values of l_c (from 0 - 1 mm) as a parameter.

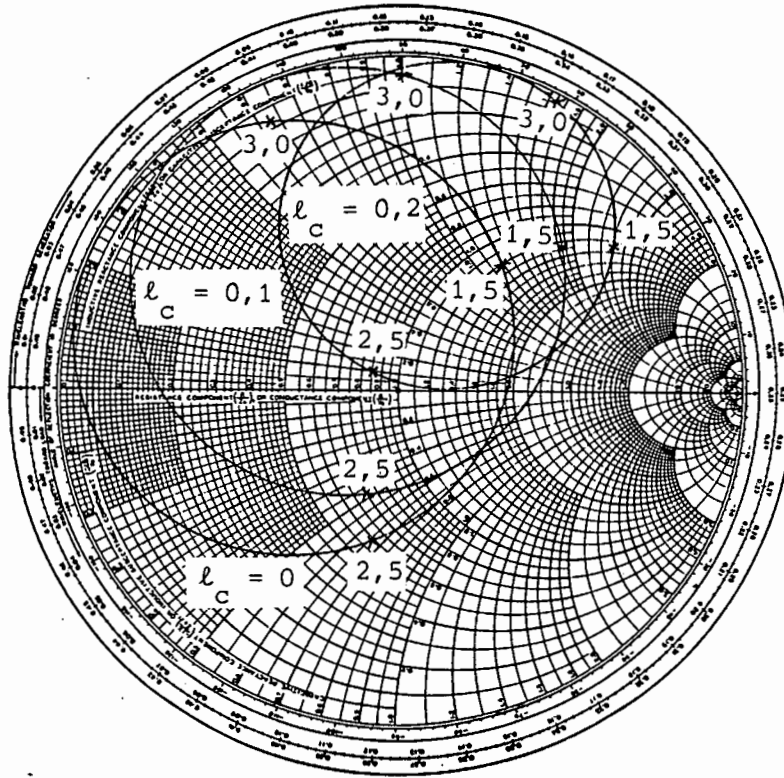


FIGURE 7.9 : Circuit impedance as a function of l_c and d_s with r_i and b_0 fixed.

The numbers on the circles denote the distance d_s in millimeters from the sliding short to the centre of the diode.

From the Smith Chart it is obvious that theoretically a large range of impedances can be covered by changing l_c and d_s . It seems possible to obtain a conjugate match with small values of l_c (0 - 0,2 mm) and with a short between 2,5 and 3 mm from the centre of the diode. The same results were obtained by [59]. In the final source construction $l_c = 0$ mm and d_s was tuned for optimum power.

7.6 PULSE MODULATOR

The schematic block diagram of the pulse modulator is shown in Figure 7.10 as obtained from [60].

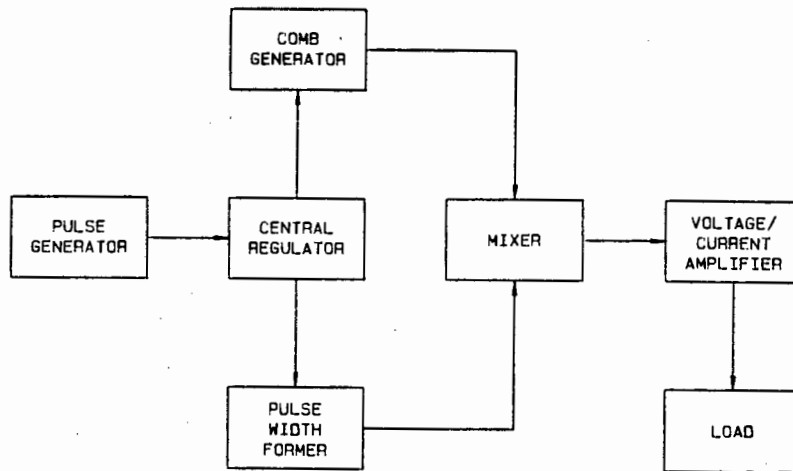


FIGURE 7.10 : Block diagram of the pulse modulator and amplifier.

It consists of an integrated timer which creates rectangular pulses which are formed to the exact pulse width by a number of logical gate. Rise and fall times of the pulse as well as duty factor are fixed in this part of the circuit. This signal is then fed to the central regulator where it is split and fed to the pulse-forming network and pulsewidth former.

The pulseforming network consists of a delay line which delays the incoming signal with a fixed time. The output consists of ten pulses of which the amplitude can be adjusted individually. By adding the single pulses, a variable composite pulse is obtained. The top of the pulse can be adjusted to obtain a linear ramp or an exponential curve.

The sloped pulse-top is then mixed with a rectangular signal from the pulse width forming network. The multiplying network ensures that the pulse width is only determined by the rectangular pulse and the pulse top is dependant upon the shaping network.

A three stage voltage amplifier follows the mixer. After the second stage the pulse amplitude can be varied continuously from zero to maximum and the fall and rise time of the pulse can be varied with the aid of a chain of switched capacitors. A direct contact between the pulse amplifier and Impatt diode produces the best results with regard to fall and rise time.

Following the voltage amplifier a two stage current amplifier completes the circuit. A current probe mounted close to the output enables the ability to display the bias pulse without any delay.

The resulting specifications of the modulator are:

Maximum Output current	$\leq 12A$
Pulse Width	Variable mechanically from 10 - 100 ns
Pulse repetition rate	50 kHz fixed or use external PRF input
Triggering	External or Internal
Rise Time	≤ 8 ns at 6A
Fall Time	≤ 15 ns at 6A

The circuit diagram of the pulse modulator is shown in Figure 7.11.

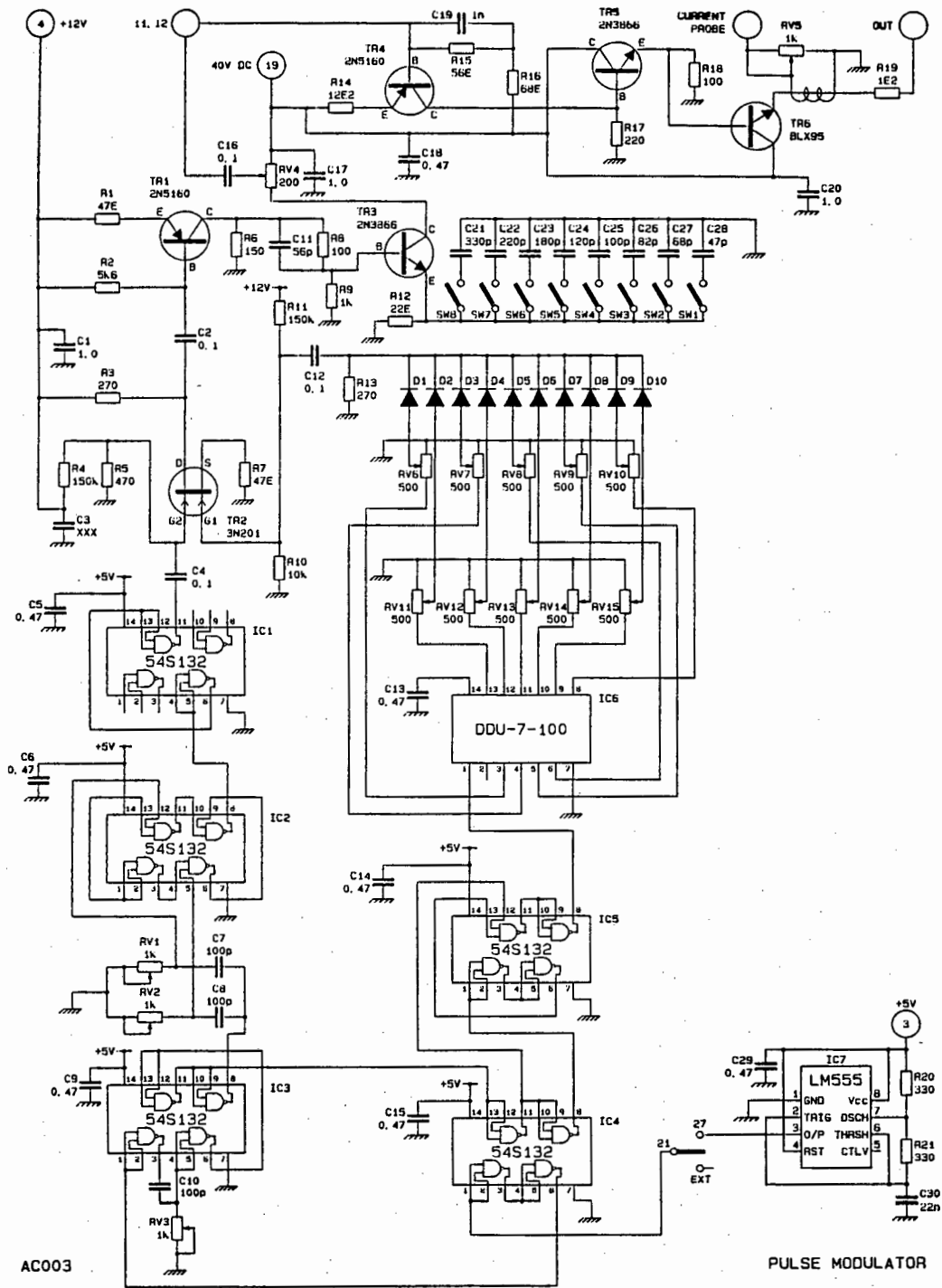


FIGURE 7.11 : Circuit diagram of pulse modulator.

7.7 MEASURED RESULTS

Figure 7.12 shows the detected RF pulse and current waveform. The dipmeter was adjusted to various positions to measure the chirp.

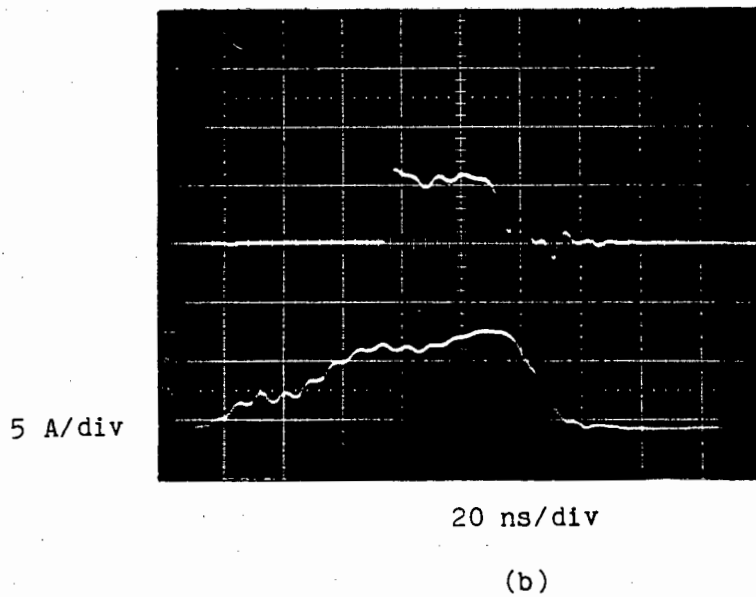
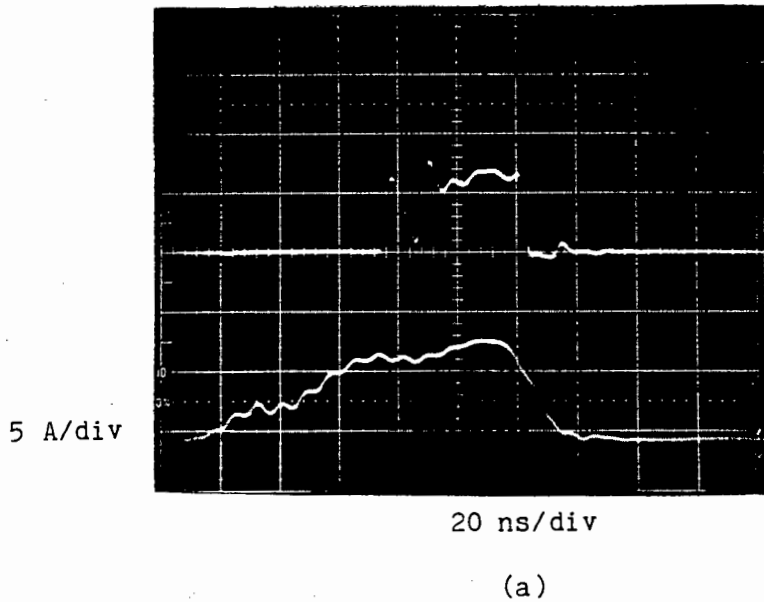
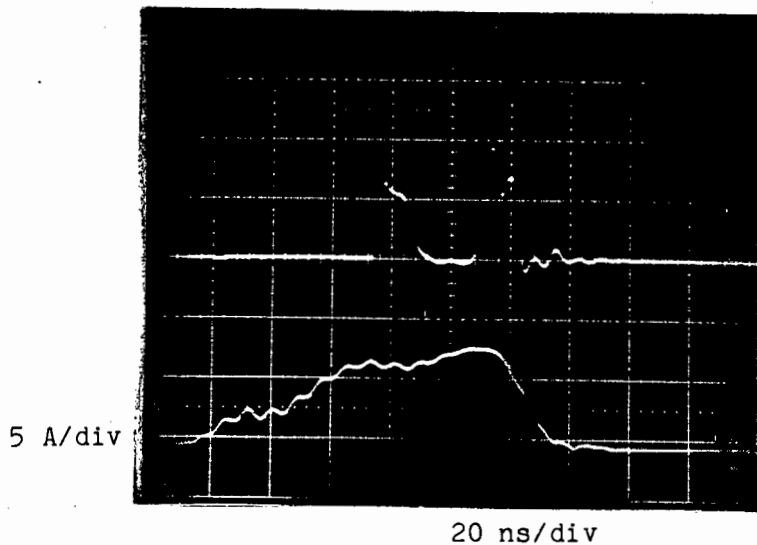


FIGURE 7.12 : Detected RF pulse. (a) Dipmeter at 93,55 GHz, (b) at 93,35 GHz and (c) at 93,5 GHz.



(c)

FIGURE 7.12 : (Continued)

The measured results are:

Output power	=	5 W (37 dBm)
Frequency chirp	=	400 MHz
Current pulse	=	6 Amp
Centre frequency	=	94 GHz

The chirp can be varied by changing the slope of the pulse-top. The setting-up procedure is as follows:

- 1) Adjust all the resistors of the comb generator to zero and switch capacitors off.
- 2) Make sure that the amplitude adjustment resistor is set to minimum.
- 3) Switch on and slowly rise amplitude of current pulse to about 4 Amps.

- 4) Switch a combination of the capacitor on to obtain a slightly up-going pulse-top.
- 5) Adjust current amplitude to the desired level.
- 6) Adjust the various resistors to obtain a slope on the pulse-top that gives the minimum chirp.

If any tuning has to be done, i.e. adjusting the diode up or down in the waveguide, the current must first be lowered, because such tuning causes instabilities which might blow the diode.

The frequency chirp characteristics of a pulsed Impatt source is shown in Figure 7.13.

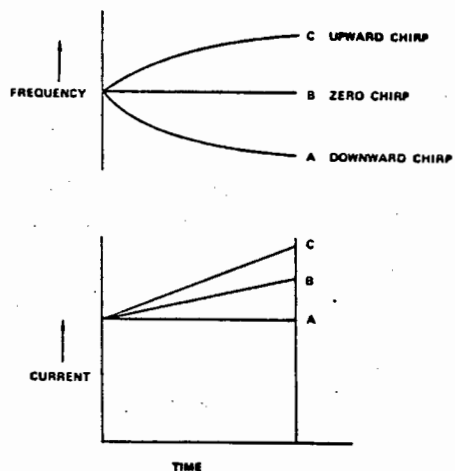


FIGURE 7.13 : Frequency chirp characteristics of a pulsed Impatt source.

The intra pulse frequency chirp is caused by the temperature rise of the device within a pulse width. As the temperature changes, the diode impedance changes which results in a decrease of oscillation frequency. Since the oscillation frequency is also dependant upon bias current, the amount of frequency chirp can be controlled by shaping the bias current pulse waveform as shown in Figure 7.13.

7.8 CONCLUSIONS

Pulsed Impatt oscillators are very sensitive for circuit adjustments, and can easily be blown. It is, therefore, necessary to turn the D.C. voltage down before any adjustments are done.

Again fine milling and turning work needs to be done to meet the tight tolerances on the very small dimensions.

The electrical performance of the oscillator meets the required specifications.

CHAPTER 8

THE DESIGN OF A 94 GHz CW IMPATT OSCILLATOR

8.1 INTRODUCTION

This chapter describes the design and development of a CW Impatt source at 94 GHz. A detailed description of the design procedure as well as measured results will be given.

It must be emphasized that the complexity in analyzing the millimeter-wave circuit and package precludes an exact analysis for quantitative predictions.

The object of this work is to understand the qualitative behaviour of the device and circuit, and more important to identify the controlling parameters to facilitate further circuit improvement. Therefore, a first order design will be done, that is to get the oscillator working, with a subsequent empirical approach to tailor the different parameters of the source to get the desired results.

8.2 GENERAL CHARACTERISTICS OF CW IMPATT SOURCE

A CW Impatt oscillator has been developed which produces 80 mW at 94 GHz, using a double drift silicon diode mounted on a copper heatsink.

This source is injection-locked with the 2nd harmonic Gunn source, and is used to injection lock the high power pulsed Impatt source.

Due to the system specifications laid out in Section 2, it is necessary to stop the injection locking process during the period between pulses. To achieve this, it was decided to pulse the CW source instead of using a pin switch. The reason for this is that a pin switch has about 3 dB insertion loss, and it means an extra component in the circuit. The advantage of pulsing the CW source is that no extra components are required, because the switching signal is obtained from the pulse modulator required for the Pulsed Impatt Source. The CW source is biased to a point just before switch on. This ensures that the active device is pre-heated and only small impedance changes will take place during the on period which means that there will virtually be no chirp.

8.3 SOURCE CONSTRUCTION

Figure 8.1 shows a cross section of the CW Impatt oscillator.

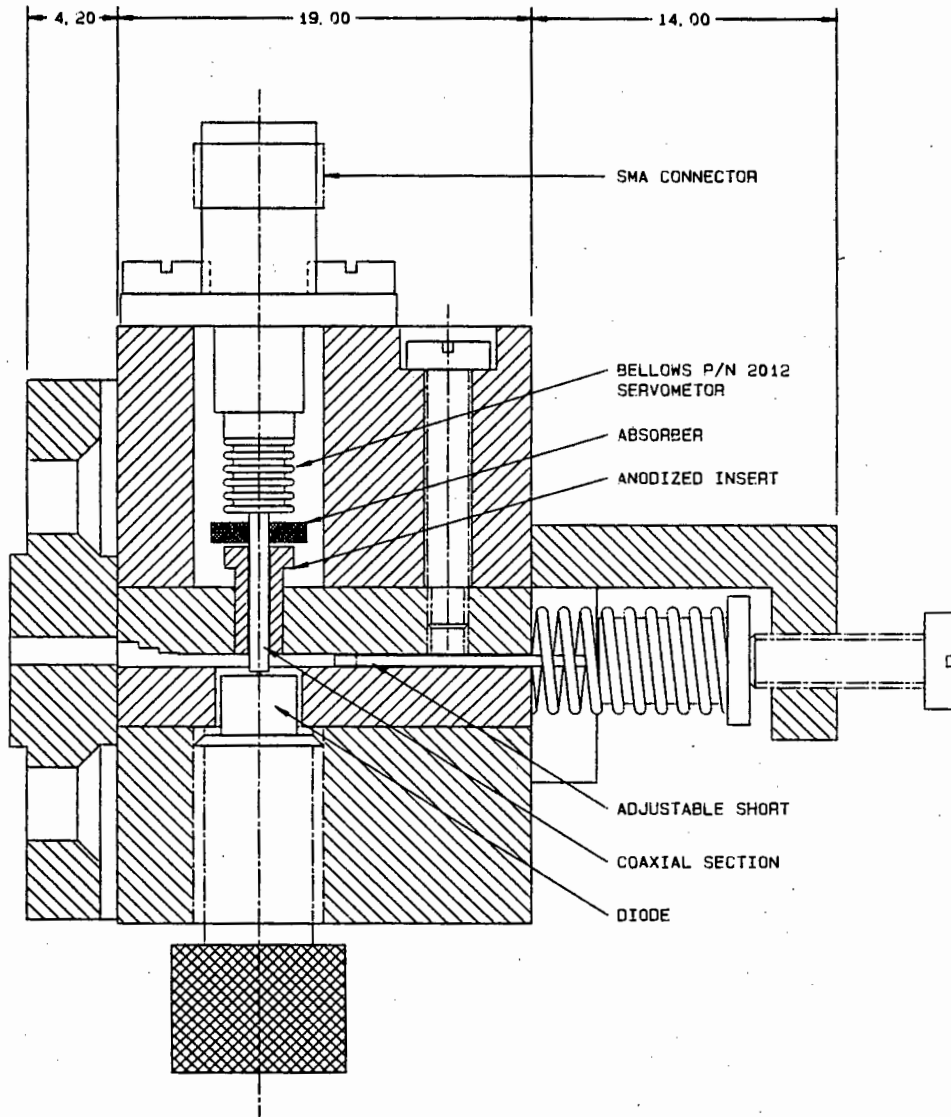


FIGURE 8.1 : Cross section of CW Impatt oscillator.

The construction of the source is exactly the same as for the pulsed source, except for two differences. That is, there is an extra spacer between the diode and the reduced height waveguide, see Figure 8.2 for a detailed drawing of the cavity.

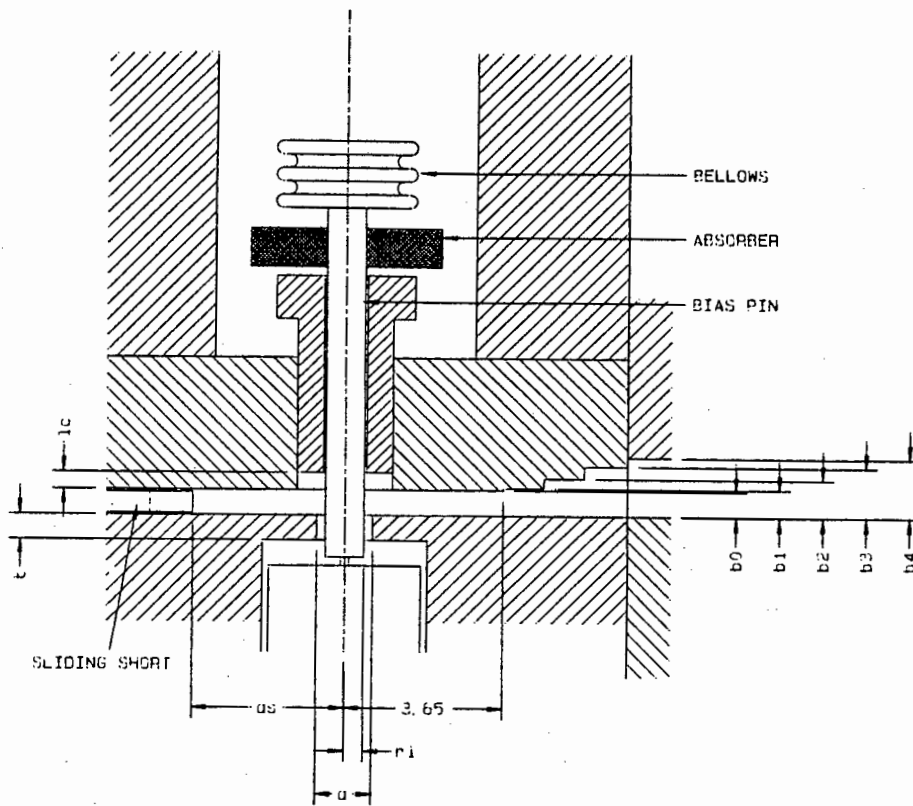


FIGURE 8.2 : Detailed drawing of CW source cavity.

The diode spacer acts as another tuning element. Basically it is a coaxial section, where the hole diameter in the spacer and the diameter of the diode tip (bias pin) forms the coaxial section. The effect that can be achieved with this spacer is:

- 1) The smaller the hole diameter (d), the higher the power. Normally the optimum hole size is twice the diameter of the bias pin;
- 2) the thicker the spacer (t), the lower the frequency.

The other difference is that the diode is not soldered into a collet, but it fits into the diode spacer. The reason for doing this is to adjust the diode up or down in the spacer which gives a certain amount of frequency tuning. The other advantage of the diode spacer is that it makes the Q of the resonant cavity higher, which gives a spectrum with lower phase noise.

All the other properties of the oscillator are the same as far as matching and tuning properties are concerned. See Section 7.3.

8.4 EQUIVALENT CIRCUIT OF A CW IMPATT DIODE

The small signal admittance of a CW Impatt diode is shown in Figure 8.3 [61].

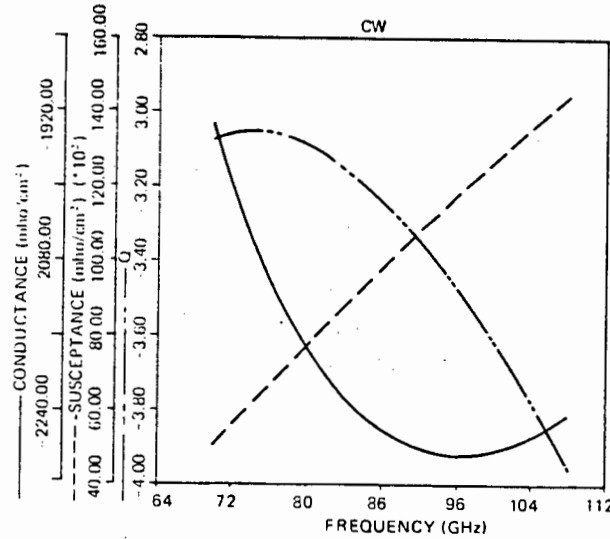


FIGURE 8.3 : Small signal admittance of a CW Impatt diode.

Comparing this with Figure 7.3, Section 7.4, it is obvious that there is a big difference between the CW and Pulsed Impatt impedances.

The equivalent circuit of the chip and package is shown in Figure 8.4 from [62].

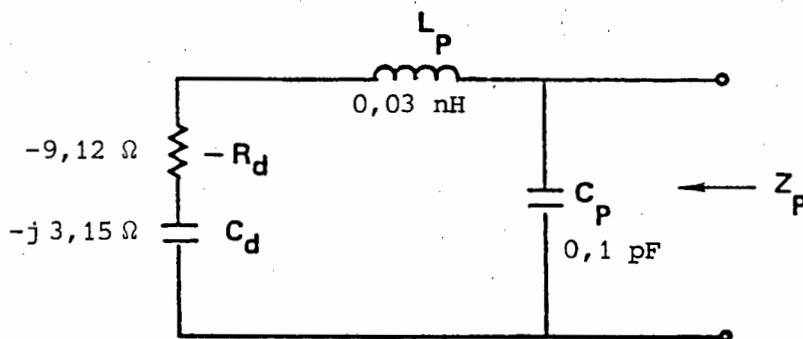


FIGURE 8.4 : Equivalent circuit of CW Impatt chip and package.

This circuit can be simplified to the one shown in Figure 8.5.

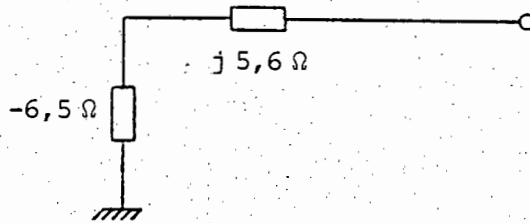


FIGURE 8.5 : Equivalent circuit as seen from package terminals.

It is this impedance that has to be matched to the high impedance of the waveguide.

8.5 RESONATOR CIRCUIT

The conditions for oscillation are the same as specified in Section 7.5. The only difference is that the resonator circuit must now reflect an impedance that is the complex conjugate of the CW Impatt impedance which is different from the Pulsed Impatt impedance.

Therefore, the design of the reduced height waveguide transformer, bias pin and sliding back short are exactly the same as for the pulsed source as well as the tuning and matching effects. The coaxial section in the lower part of the reduced height waveguide is just another matching element as described in Section 8.3.

The final dimensions of the various tuning elements are:

$$l_c = 0,2 \text{ mm}$$

$$L = 0,3 \text{ mm}$$

$$d = 1,5 \text{ mm}$$

$$d_s = \text{between } 1,2 - 1,5 \text{ mm}$$

$$r_i = 0,44 \text{ mm}$$

8.6 MEASURED RESULTS

Figure 8.6 shows the spectrum of the CW Impatt source in the free running mode, and Figure 8.7 the detected output pulse when the source is pulsed.

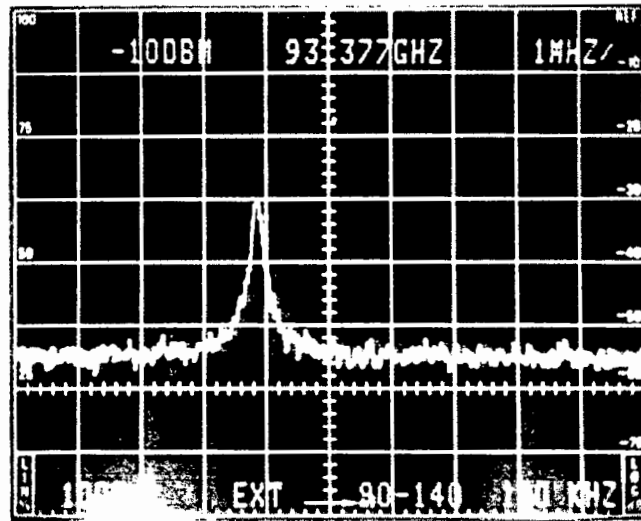


FIGURE 8.6 : Spectrum of CW Impatt source.

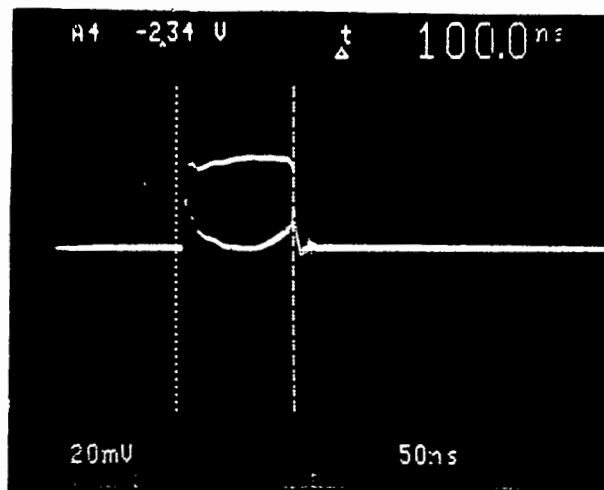


FIGURE 8.7 : (a) Detected output pulse top trace and (b) with the dip-meter set to 93,5 GHz.

CHAPTER 9

APLANATIC LENS ANTENNA

9.1 INTRODUCTION

A plano-convex crossed linked polystyrene dielectric lens is used together with a 94 GHz feedhorn to form the aplanatic mm-wave lens antenna. (Aplanatic means free from spherical aberration and coma, see [63-65]).

Due to the light weight of the dielectric material it is natural to use it at higher millimeter wave frequencies where space and weight are a problem.

Smaller lenses for mass production can easily be manufactured with casting and molding techniques. The other advantage of lenses are the low sidelobe levels, typical 25 dB and lower can easily be obtained.

9.2 DESCRIPTION OF LENS ANTENNA

The lens and feedhorn is shown in Figure 9.1.

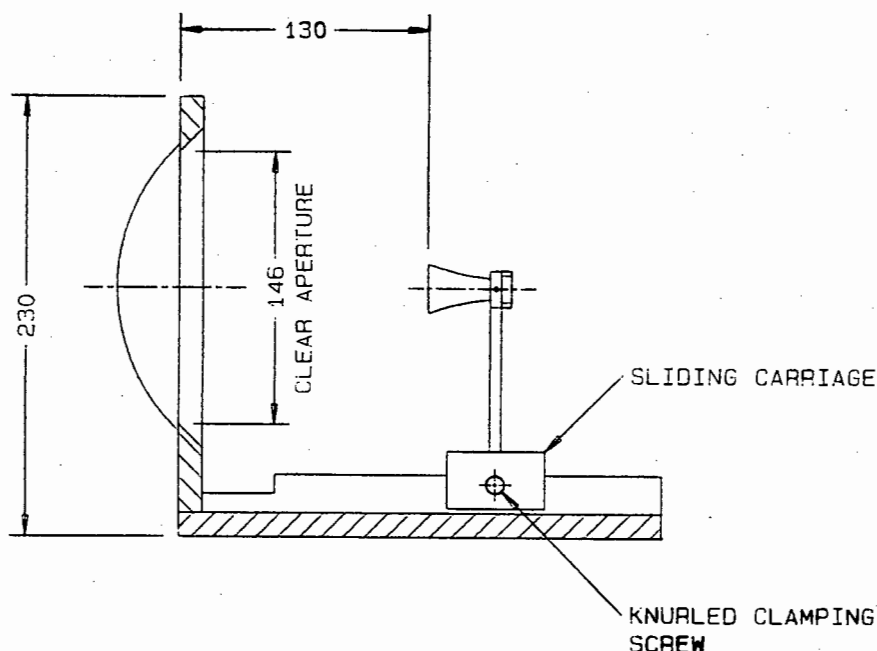


FIGURE 9.1 : Physical construction of lens and feedhorn.

The lens is fitted into a stand made out off clear bakelite. The feedhorn is clamped onto a pedestal which can be moved on axis in the x-direction as well as off axis. This enables the system designer to simulate beam scanning as well as beam broadening.

The lens is manufactured from cross linked polystyrene with a refracting index $n = 1,594$. The lens is 150 mm in diameter with a 130 mm focal length and the thickness of the lens on-axis is 33,83 mm.

The feedhorn illuminates the lens, and is manufactured by the electro-forming process. The lens is a broadband structure and can be used from about 7 GHz to well over 300 GHz. The surface tolerance of about 0,05 mm is the main limitation.

9.3 DESIGN OF THE LENS

The lens is the same as the one designed by [66], and is a double refracting lens.

The lens was fitted in the stand with the plane side towards the feedhorn and the curved convex surface away from the source. This configuration has the following advantages.

- 1) The amplitude distribution produced by the feedhorn is preserved on the exit side of the lens.
- 2) Spherical aberration for such a lens is less than when the convex side is towards the feed.
- 3) Although it is a double refracting lens, neither of the surfaces present an equi-phase contour, so that the energy reflected from the edges does not add in-phase at the feed to cause large mismatches.
- 4) A reduction in focal length increases the beamwidth.

The lens was machined from a 150 mm diameter rod of cross linked polystyrene. The design equations of the lens is found in [67] of which the geometry is shown in Figure 9.2.

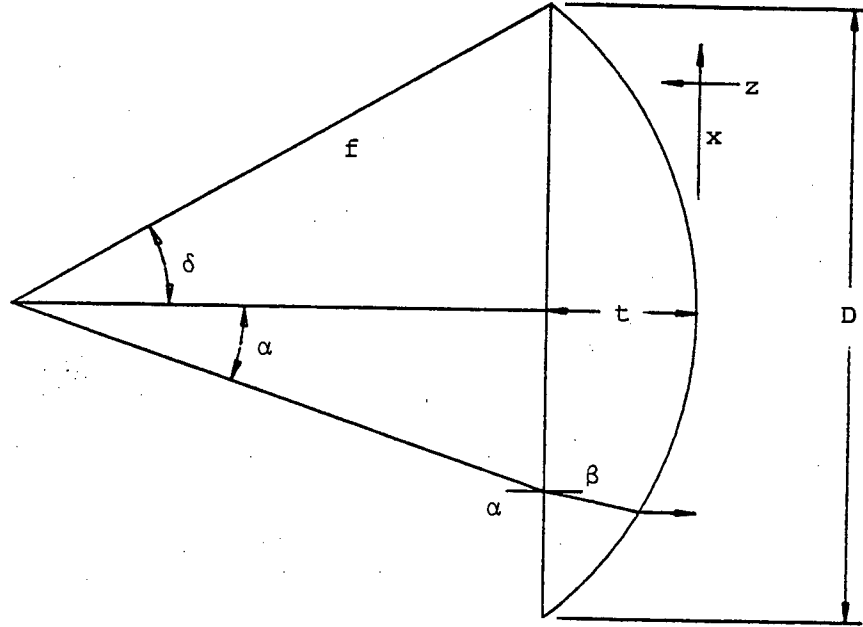


FIGURE 9.2 : Lens geometry.

The plano-convex aplanatic lens profile is defined by the equations:

$$\beta = \sin^{-1}\left[\frac{\sin\alpha}{n}\right] \quad \text{Snell's law} \quad (9.1)$$

$$t = \frac{f(1 - \cos\delta)}{n - 1} \quad (9.2)$$

$$z = t - \cos\left\{\frac{(n - 1)t - f\cos\delta(\sec\alpha - 1)}{n - \cos\beta}\right\} \quad (9.3)$$

$$x = f\sin\alpha\left\{1 + \frac{\left(1 - \frac{\cos\delta}{\cos\alpha}\right)(\cos\beta - 1)}{n(n - 1)(n - \cos\beta)} + \cos\delta(\sec\alpha - \sec\delta)\left[\frac{n^2 - n - 1}{(n - 1)}\right]\right\} \quad (9.4)$$

9.4 DESIGN OF THE PYRAMIDAL FEEDHORN

From [66] it is found that the Narda V637 horn gave good illumination and low sidelobes when it is used together with the 150 mm lens. It was, therefore, decided to scale this horn to operate at 94 GHz and have the same illumination i.e. -15 dB down at the edges of the lens.

The final dimensions of the feedhorn is shown in Figure 9.3.

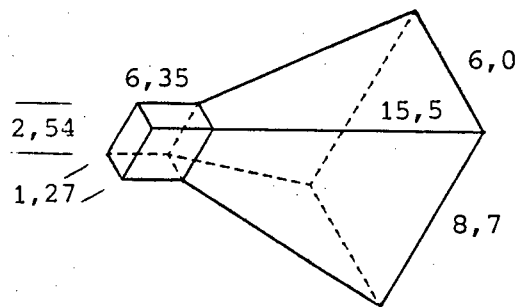


FIGURE 9.3 : Feedhorn construction.

The horn is manufactured by the electroforming process and is gold plated.

9.5 MEASURED RESULTS

The reflection coefficient of the feedhorn was first measured without the lens, and then with the lens in front of the feedhorn.

The reflection coefficient of the two situations are shown in Figure 9.4.

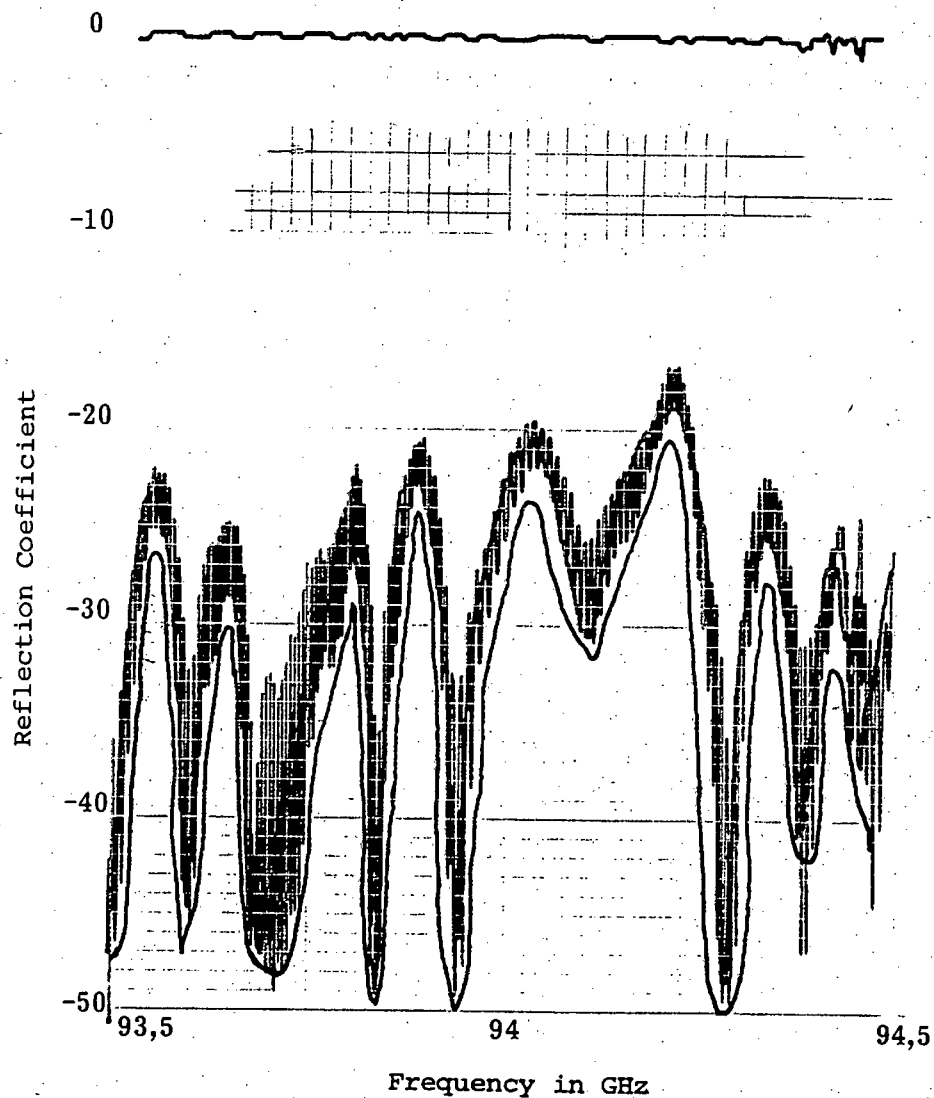


FIGURE 9.4 : Reflection coefficient of (a) feedhorn (smooth curve) and (b) feedhorn and lens (graph with high frequency ripple).

The gain of the lens antenna was measured as 41 dBi and the beamwidth as $1,5^\circ$.

From Figure 9.4 it can be seen that the two graphs are virtually the same. The high frequency ripple on the curve of the lens assembly is caused by phase summation and cancellation effects due to the surface tolerance.

The H-plane radiation pattern is shown in Figure 9.5. The horn and lens were measured on the $\pm 9^\circ$ scale, and the horns alone on the $\pm 45^\circ$ scale.

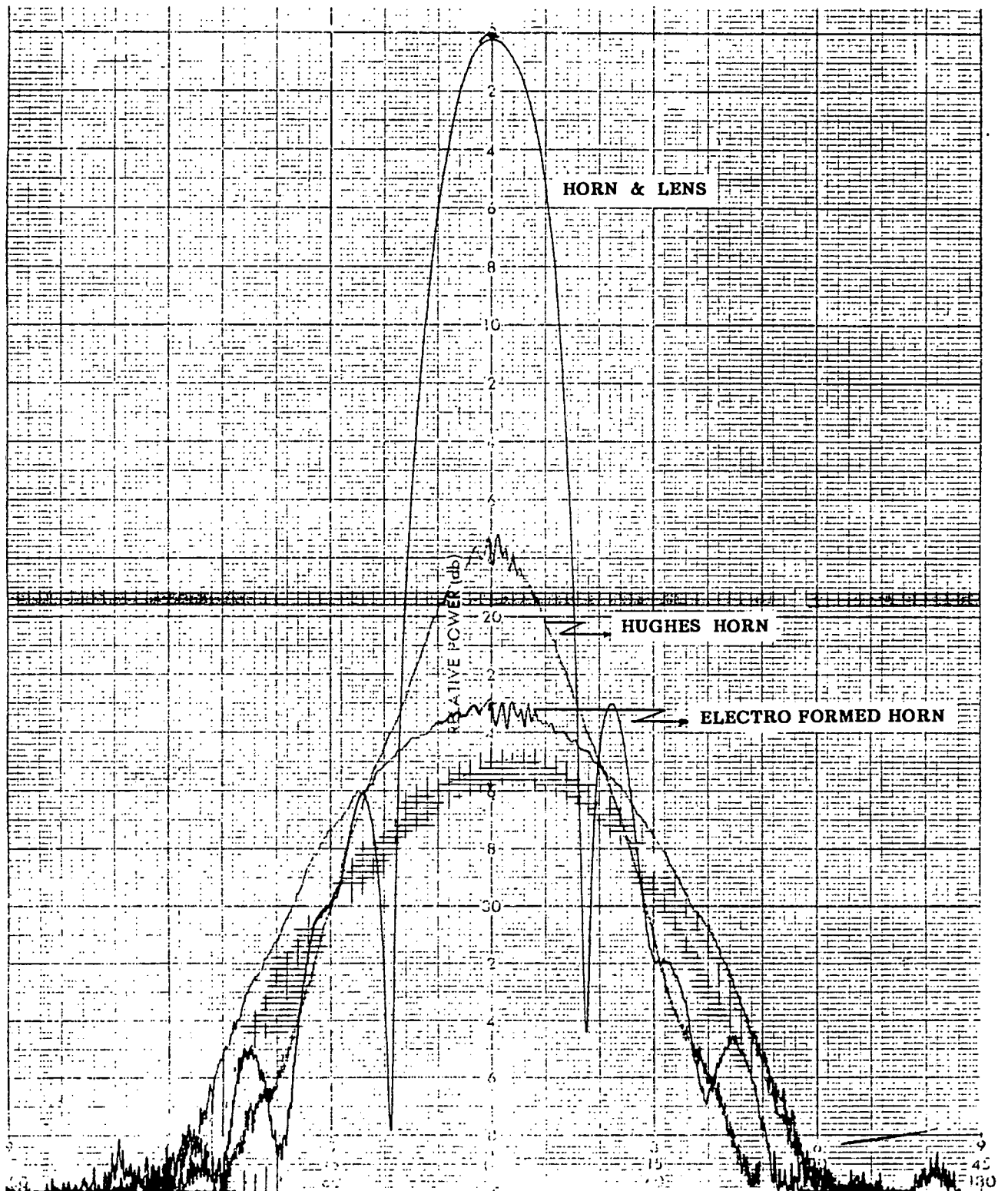


FIGURE 9.5 : H-plane radiation pattern.

9.6 CONCLUSIONS

The lens antenna gave low sidelobe levels without undue precautions. This is a result obtained with the absence of aperture blockage by the feed.

A disadvantage of the lens is that it is not as compact as a prime focus type antenna.

CHAPTER 10

GENERAL CONCLUSIONS

This work demonstrates that it is possible to build a sensor, acting as an early warning radar, applying the latest technology available at mm-wave frequencies.

The technique used to obtain coherency is unique in the sense that the Pulsed Locked Oscillator (PILO) and the first and second coherent oscillators (Coho) are phased locked to a low frequency (300 MHz) crystal oscillator. This technique assures that the phase of the transmit signal is preserved in the RF and IF Coho sources respectively.

Although no measurements were taken on high voltage overhead transmission lines, the system was calibrated with a corner reflector with known radar cross section.

The typical radar cross section of a 30 mm outside diameter overhead line is 28 m^2 at 600 m [7], and that of the corner reflector 721 m^2 .

From Figure 4.10 the detected output voltage as measured from the corner reflector is 0,71 V, which corresponds closely to the calculated value of 0,83 V using equation (2.9) and Figure 4.5.

Due to the small difference between the measured and calculated values (1,5%) it is realistic to use equation (2.9) and Figure 4.5 with good accuracy.

By using this method, the calculated detected voltage of an overhead line with outside diameter of 30 mm at a distance of 600 m is 0,15 V. Reducing the output power from 5 W to 1 W, the detected voltage should be 0,06 V, which is still detectable without any signal processing.

These calculations/measurements correspond with those mentioned in [1], namely to detect a 30 mm line at a distance of 600 m.

The main contribution of this work is, however, the method derived to measure coherency in a pulse system.

The mathematical derivation gives the coherency of the system as the phase difference between the reference signal and the transmit signal.

With this method, the coherency of any system, no matter what the frequency is, can be determined accurately.

The value of $\sqrt{R_r(0)}$ is obtained by taking the squared RMS value from the output of the sample and hold circuit, which is measured with a true RMS voltmeter, substitute that into equation (3.30), and thereby determine the phase difference between any two signals.

Good correlation is obtained with this method and published data.

The empirical design method for the various components prove to give good results. Due to a lack in measuring instruments at this frequency it is impossible to do an exact design. The results achieved compared favourably with published data of commercially available components.

CHAPTER 11

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