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Canada

**Design and Realization of a
24 GHz Receiver Front-End
in Integrated Quasi-Planar Technique**

by

Andre Gagnon

**A Thesis presented to the
University of Ottawa in partial
fulfillment of the requirements for
the degree of Masters of Applied Science
in the
Department of Electrical Engineering**



André Gagnon, Ottawa, Canada, 1990



UNIVERSITÉ D'OTTAWA
UNIVERSITY OF OTTAWA

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ABSTRACT

This thesis describes the design, realization and testing of an integrated millimeter wave receiver front-end in E-plane configuration. In particular, the combination and integration of a balanced mixer, two filters, and a local oscillator stabilized by a dielectric resonator are described. The following is a brief description of those three major components that constitute the receiver front-end.

The Balanced Mixer. The heart of the mixer is a classical 180 degrees hybrid junction formed by the juxtaposition of unilateral fin-line and coplanar waveguide. Matching of the RF signal to the diodes is critical for good conversion loss. It is achieved by a tapered fin-line section followed by a quarter-wave transformer immediately in front of the diodes. To minimize the conversion loss, two diodes in a beam-lead T-configuration have been used. In addition to providing improved symmetry and reduced parasitics, this specialized element is also easier to mount than two separate diodes.

The Filters. The receiver contains two filters. The first is a bandpass in the RF input section. It is an optimized three-resonator bilateral filter of full waveguide height, designed to suppress unwanted signals and noise in the image band. It has a 2.5 percent bandwidth, and its passband attenuation is less than 0.3 dB. The second filter is a microstrip lowpass in the IF output port. This seventh order Chebyshev filter has a cutoff frequency of 14 GHz and an insertion loss of 0.1 dB, including the microstrip-to-SMA connector transition.

The Local Oscillator. In order to ensure compatibility with the mixer, the 24 GHz local oscillator was realized in such a way that the FET was also situated in the E-plane. The fin-line oscillator, proposed by Jacob and Ansorge, features a pair of boomerang-shaped slots which are coupled to the gate and the drain of the NE67383 FET respectively. Biasing of the transistor is straightforward. The frequency of oscillation is determined either by a half-wave strip resonator or by a dielectric disc resonator on the backside of the substrate. An efficiency of 30% has been achieved by optimizing the position of the short-circuits in three of the four fin-line branches. The maximum output power of the local oscillator was 20 mW.

The components have been realized on 10 mils thick RT-Duriod 5880 (dielectric constant $\epsilon_r = 2.22$) suspended in the E-plane of a WR-42 waveguide.

The receiver incorporates a 24.0 GHz oscillator (LO) and accepts an input radio frequency signal (RF) at 24.4 GHz with a bandwidth of 500 MHz. The intermediate frequency band (IF) is centered at 400 MHz with a bandwidth of 500 MHz as well. The conversion loss is 5.5 dB in the band center, and less than 6 dB throughout the entire bandwidth. The noise figure is 0.5 to 1 dB higher than the conversion loss. The VSWR at the signal RF port is better than 1.5:1. (Return loss less than 14 dB).

INTRODUCTION

It is well known that most of the technology applied in Canada is imported. However, there are some sectors in which Canadian technology is recognized worldwide. One of them is the communications sector.

The goal of this thesis was to design a receiver front end by integrating a balanced mixer, a local oscillator stabilized by a dielectric resonator, and two filters in E-plane configuration. Even though all of these individual components have been realized separately in the past, the author believes that it is the first time that such a combination of E-plane components has been integrated on an single substrate, particularly the combination of band-pass filters with the mixer as well as the combination of the novel local oscillator with the mixer.

The thesis is subdivided into three chapters, each describing a specific element of the receiver, followed by a chapter on the overall receiver.

Each chapter presents the theory, design, configuration, and results on its respective subject.

Appendices contain additional information which can be very general (such as other possible circuit configurations) or very specific (such as the electrical specifications of diodes).

Chapter 1 presents the theory of the single balanced mixer followed by the design, configuration, and experimental results of the balanced fin-line mixer used in this thesis. Also discussed in Chapter 1 are the different transmission lines and transitions used to integrate the mixer with its surrounding elements.

In Chapter 2 the theory, design, and the experimental results of low-pass filters and band-pass filters built during this thesis are discussed.

Chapter 3 describes the dielectric resonator, oscillator (DRO) theory, followed by the design, configuration, and experimental results of the fin-line DRO.

Finally, Chapter 4 presents the overall receiver circuit and housing configuration developed in this thesis, followed by experimental results.

Appendices A and D present the properties of the fin-line and coplanar line, respectively.

Appendix B, C, E, and F present a review of the different elements of the receiver, namely the mixer, filter, and local oscillator.

Finally, Appendix G contains the electrical specifications of the different devices, namely the beam lead Schottky diode, FET, and dielectric resonator.

CHAPTER 1

MIXER THEORY AND DESIGN

The mixer is a critical component in modern RF systems. In fact it is usually the first or second device after the RF input. The performance of the mixer is crucial to the overall operation of the system. Important mixer parameters such as conversion loss, bandwidth, noise figure, interport isolation, intermodulation, dynamic range, intercept point and VSWR (voltage standing wave ratio) must be optimised to produce the type of mixer necessary for today's sophisticated RF systems. Refer to appendix C section C-1 for a detailed description of the mixer parameters.

A mixer converts RF power at one frequency into power at another frequency to make signal processing easier and less expensive. Another fundamental reason for frequency conversion is to allow for the practical transmission of audio and other low frequency information to use free space, high frequency links. Thus in both the transmitting and receiving cases the input signal frequency must be converted to another frequency. This is done through the mixing process.

Mixing the input signal (representing the desired information) with a local oscillator signal generates upper and lower sidebands, each containing the modulation present in the input frequency. The upper sideband is the sum of the input and the local oscillator (LO) frequencies. The lower sideband is the difference between the input and the LO frequencies. The upper or lower sideband is called the intermediate frequency (IF) band. In most receiving systems, the lower sideband (LSB) is used, whereas in the transmitting systems the upper sideband (USB) is used. Since it is less expensive to amplify a signal in the MHz range than in the GHz range, the incoming microwave signal is

first downconverted in frequency and then amplified and processed. Conversely, in a transmitter it is less expensive to generate, modulate, and amplify a signal in the MHz range and then upconvert it in frequency into the GHz range.

Figure 1.0.1 shows the placement of a mixer in a receiver front end. Received RF, input LO and output IF ports are designated by R, L, and I respectively. The low noise amplifier (LNA) is used to improve the overall noise figure. The three attenuators (pads) are used to improve the matching as well as to reduce the intermodulation products.

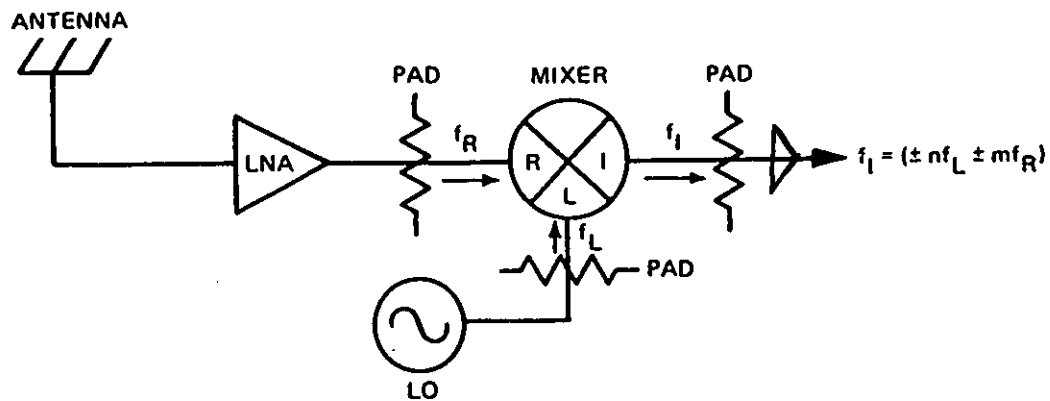


Figure 1.0.1 Mixer Placement in Receiver Front End

The LO signal (f_L) is usually stronger than the received signal (f_R). This causes the mixer to have a better intermodulation suppression than would be possible if the LO and RF power were similar. Theoretically, for infinite LO power (perfect switching process) the intermodulation products are minimized. Practically, the LO signal power should be at least 20 dB higher than the RF input power.

The intermediate frequency (f_I) is simply given by:

$$f_I = \pm mf_R \pm nf_L \quad (1.0.1)$$

where,

$$m = 0, 1, 2, 3, \dots$$

$$n = 0, 1, 2, 3, \dots$$

The desired output products are the sums (for upconverting) and differences (for downconverting) of the fundamentals of the received and LO signals. This is the case for which:

$$m = n = 1$$

giving,

$$f_I = \pm f_L \pm f_R \quad (1.0.2)$$

While this formula implies that negative frequency products occur, these can be ignored in practical mixer applications. For the case where $f_L > f_R$, which is called high-side LO, $f_I = f_L \pm f_R$. For the other case where $f_R > f_L$, which is called low-side LO, $f_I = f_R \pm f_L$. The higher-order products having $m = 1, 2, 3, \dots$ and $n = 1, 2, 3, \dots$ for which m and n are not simultaneously equal to 1 are called higher-order intermodulation products.

The RF and LO harmonics for which $m = 1, 2, 3, \dots$, $n = 0$ or $n = 1, 2, 3, \dots$, $m = 0$ appear at the terminals of the diode and leak to the IF output port due to the finite interport isolation.

1.1 Single Balanced Mixer Theory

In this section the analysis and characterization of a balanced mixer will be presented.

1.1.1 Mixer Circuit

There are basically five types of mixer circuits: single-ended (SE), single-balanced (SB), double-balanced (DB), double-double-balanced (DDB) and harmonic mixers (H). The single-balanced (SB) mixer has been chosen by the author for its good performance and ease of realization in integrated circuit technology. Refer to Appendix C, Section C-2, for a description of the SE, DB, DDB and H mixers. In the following, the single-balanced mixer (SB) (henceforth called balanced mixer) will be described.

Single-balanced mixers are made by combining two single-ended mixers as presented in Figure 1.1.1.1. Figure 1.1.1.2 shows the general configuration that the SB mixer can take. The L-port balun balances the diodes and interfaces them with the unbalanced LO input. The most important characteristic of the balun is its ability to maintain the phase angles with respect to the ground, of ± 90 degrees at B, ∓ 90 degrees at D, and 0 degrees at C (for center tap application). When these three angles are maintained, the balun is said to be well-balanced.

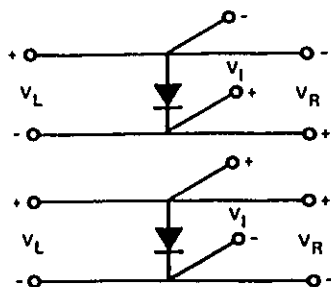


Figure 1.1.1.1
Combination of Two Single-Ended Mixers
to form a Balanced Mixer

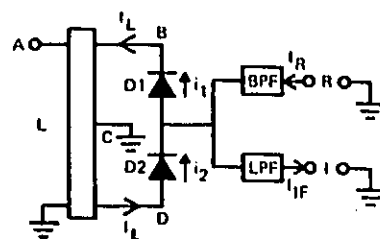


Figure 1.1.1.2
Single Balanced Mixer

1.1.2 The Balun

Balanced mixers contain one balun and two diodes. The balun balances the diodes and interfaces them with the unbalanced input line. It also serves as a matching circuit between the input and the mixer diodes, and provides interport isolation.

Balun circuits can be realized with various technologies such as: coils (dc to 4 GHz) [1], coaxial cable (dc to 8 GHz), microwave integrated circuits (0.5 to 10 GHz), waveguide (0.5 to 100 GHz) and a fin-line - coplanar configuration (0.5 to 240 GHz). The latter is described and analyzed in Section 1.2.

1.1.3 Single-balanced Mixing Process

Mixing can only be performed by devices which have current-voltage relationships that are non-linear, or that change as a function of time, or both. Switches are time-variant devices because they form either a short or an open circuit as a function of time. Diodes are non-linear; both their time-variant and non-linear properties are used to perform mixing. See Appendix C, Section C-3, for a description of different types of mixer diodes.

The non-linear characteristic of a Schottky diode is presented in Appendix C, Section C-3, by the relations C.3.1. Expanding this i-v relationship into a power series for a forward biased diode yields;

$$i = I_0 + a_1 v + a_2 v^2 + a_3 v^3 + \dots \quad (1.1.3.1)$$

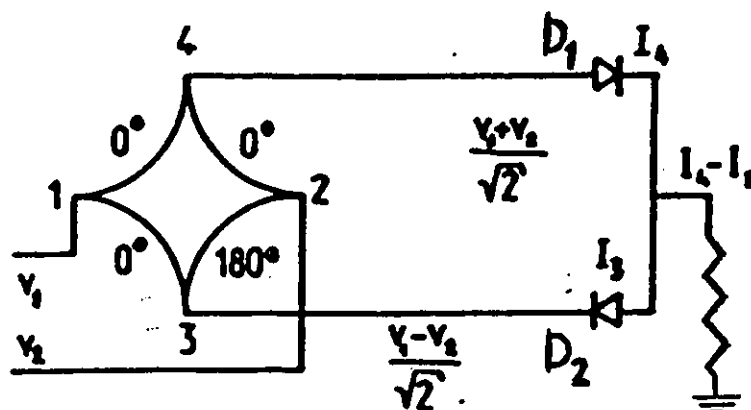


Figure 1.1.3.1 Simplified Balanced-Mixer Circuit

Referring to Figure 1.1.3.1 and assuming a perfect balun and matched diodes, the voltages that appear across diodes D1 and D2 are $(v_1 + v_2)/\sqrt{2}$ and $(v_1 - v_2)/\sqrt{2}$ respectively. Substituting these voltages into (1.1.3.1) gives the current flowing through the diode:

$$\sqrt{2} i_4 = I_0 + a_1 v_1 + a_2 v_2 + a_2 v_1^2 + a_2 v_2^2 + 2a_2 v_1 v_2 + \dots \quad (1.1.3.2a)$$

$$\sqrt{2} i_3 = I_0 + a_1 v_1 - a_1 v_2 + a_2 v_1^2 + a_2 v_2^2 - 2a_2 v_1 v_2 + \dots \quad (1.1.3.2b)$$

where I_0 is the reverse bias saturation current. The current supply to the IF load is given by subtracting (1.1.3.2a) from (1.1.3.2b).

$$\sqrt{2}(i_4 - i_3) = 2a_1 v_2 + 4a_2 v_1 v_2 + 6a_3 v_1^2 v_2 + 2a_3 v_2^3 + \dots \quad (1.1.3.3)$$

By setting $v_1 = v_1 \cos(\omega_1 t + \Theta_1)$ and $v_2 = v_2 \cos(\omega_2 t + \Theta_2)$ into (1.1.3.3), a number of distinct terms are obtained. Among them,

- the first term contains only the frequency f_2 ;
- the second term contains the frequencies $f_1 - f_2$ and $f_1 + f_2$;
- the third term contains the frequencies $f_2 \pm 2f_1$;
- the following term contains the frequencies $\pm nf_2 \pm mf_1$.

For the downconverting process, only the second term containing the frequency $(f_1 - f_2)$ is of interest. It represents the intermediate frequency, IF. Higher order terms (or modes) represent higher order harmonics, and their amplitude decreases with increasing order. The first non-desirable harmonic is therefore the third order mode represented by the third term. This harmonic is critical because it is the strongest harmonic signal and often very close to the IF. For example if $f_1 = 8$ GHz and $f_2 = 12$ GHz, the IF is 4 GHz $(12 - 8)$ and the third harmonic is also 4 GHz $[12 - (2 \times 8)]$. In that case, no filter can be used to attenuate the third harmonic frequency.

1.1.4 Linear and Non-Linear Analysis

To optimise the conversion loss, it is obvious that the diode parasitics must be included in the analysis, particularly at millimetric frequencies. Many authors [2], [3], [4] have developed linear and non-linear mixer analysis procedures. Hicks and Khan [5] have modified the harmonic balance approach such that it can be applied specifically to balanced mixers.

The numerical analysis shows that for reducing the conversion loss, it is advantageous to:

- reduce the series resistance, R_s
- reduce the junction capacitance, C_j
- reduce the stray capacitance, C_s
- reduce the ideality factor, n
- reduce the lead inductance, L_p
- use identical diodes (matched pair diodes).

1.1.5 Characteristics of the Diode used for the Mixer

The diodes used in this study are a matched pair of medium barrier beam lead Schottky diodes in T-configuration, fabricated by Metelics Corporation, part number MSS40, 241-B20.

The manufacturer's specifications are listed in Appendix C, section G-1. The most important parameters are as follows:

- series resistance, $R_s = 10$ ohms
- junction capacitance, $C_j(0) = 0.06$ pF
- stray capacitance, $C_s = 0.04$ pF
- lead inductance, $L_p = 0.1$ nH
- ideality factor, $n = 1.07$

1.1.6 Technology Considerations

Different Types of Mixer Construction

Mixers constructed in essentially four different technologies are commercially available; their selection depends on the receiver application as well as on the frequency to be down- or up-converted.

They are:

- Waveguide mixers
- Fin-line mixers
- Microstrip mixers
- Monolithic mixers.

Refer to Appendix C, Section C-4, for a brief description of each mixer type.

Comparison of Mixer Types

Table 1.1.6.1 summarizes the present trends in mixer design [6].

SPECIFICATION \ MIXER TYPE	CROSS-BAR	FIN-LINE	MICROSTRIP LINE	MONOLITHIC GaAs
Production, Engineering Design	Established	Transition from lab to production	Transition from lab to production	In development
Shocks	400g, 10ms	not shown	>400g 10ms	Not shown
Operating Temperature range (degrees C)	-55 to +70	-50 to +70	-55 to +70	Not shown
Maximum power before failure (dBm)	RF +20 LO +10	RF +20 LO +7	RF +20 LO +13	Not shown
Hermetic Package	Available	Needs to be developed	Available	Available
Frequency Range (GHz)	140	140	60	45

Table 1.1.6.1 Types of Mixers

1.2 Design of a Balanced Fin-Line Mixer

In this section, the design and realization of a balanced mixer in fin-line configuration will be described.

1.2.1 Fin-line Balanced Mixers

The basic element of a fin-line balanced mixer is a 180 degree hybrid junction formed by the juxtaposition of a fin-line and a coplanar line [9].

Figure 1.2.1.1 shows the final arrangement of the fin-line mixer designed in this thesis. The RF signal input consists of a symmetrical unilateral fin-line taper. Schottky-barrier beam lead diodes are placed at the junction between fin-line and coplanar line, as shown in the enlarged section of figure 1.2.1.1. The local oscillator power is fed to the diode via a transition from an asymmetrical fin-line to a coplanar line, working like a "probe type" transition from standard waveguide to coaxial line. For proper operation, a variable short is placed a quarter wavelength away from the transition. It allows a final adjustment of the LO-input match. A microstrip low pass filter connects the coplanar line section to the IF output port.

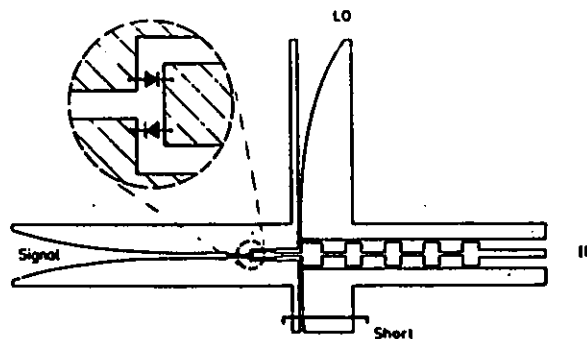


Figure 1.2.1.1 Setup of the Designed Balanced Fin-Line Mixer [9].

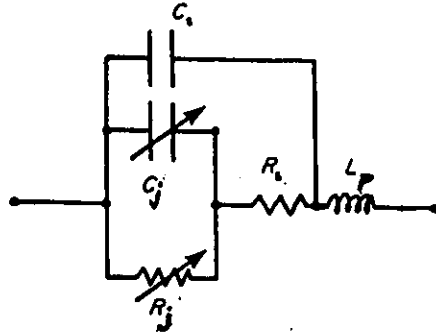
As can be seen from the layout in Figure 1.2.1.1 different types of lines such as symmetrical and unsymmetrical fin-line, coplanar line (odd and even mode of propagation), and microstrip are combined to form the mixer circuit. For the design of these circuits, accurate transmission line models are required. Refer to Appendix A and D for fin-line, and coplanar line characteristics, respectively. Furthermore, the characteristics of the diodes are of great importance. The equivalent circuit of the diodes must contain all parasitic elements. These cannot be neglected at higher frequencies, and lead to reduced mixer efficiency and increased LO power requirements. The details of the design procedure are described in the following sections.

1.2.2 Design of the Balanced Fin-line Mixer

The design of the mixer has been accomplished in three phases: the modelling of the diode, the modelling of the 180° hybrid junction and the design of the diode matching circuit.

1.2.2.1 Derivation of the Diode Impedance

The equivalent circuit for the beam-lead diodes is discussed in detail in Appendix C, Section C-3, Figure C.3.5b). For convenience, Fig. 1.2.2.1.1 repeats the equivalent circuit together with basic parameters for the beam-lead diodes used in this mixer design.



Series resistance, $R_s = 10 \Omega$

Zero Bias Junction capacitance, $C_j(0) = 0.06 \text{ pF}$

Stray capacitance, $C_s = 0.04 \text{ pF}$

Lead inductance, $L_p = 0.1 \text{ nH}$

Figure 1.2.2.1.1 Equivalent Circuit for Beam-Lead Mixer Diode

From these basic data, the values of the junction capacitance (C_j) and resistance (R_j) at the operating point can be computed using equations (C.3.4) and (C.3.3), respectively, in Appendix C.

In a private communication [14], Metelics engineering staff recommended to drive the diode with a LO power generating a short-circuit current of 3 mA for minimum conversion loss. In this condition:

$$R_j \approx 28/I(\text{mA}) = 28/3 \approx 9 \Omega$$

Since the barrier potential values were not available from the manufacturer, the general relation given in (C.3.4b) was used. Thus:

$$C_j = 1.5 C_j(0) = 1.5 \times 0.6 \text{ pF} = 0.09 \text{ pF}$$

With these parameters the single beam-lead diode impedance at 24.4. GHz is obtained as:

$$Z_{\text{diode}} = (18.4 + j 12.1)\Omega \quad (1.2.2.1.1)$$

For the realization of the mixer, a matched pair of diodes in beam-lead T-configuration was selected because it is easier to install in the circuit and is optimally balanced (see Fig. 1.2.2.1.2).

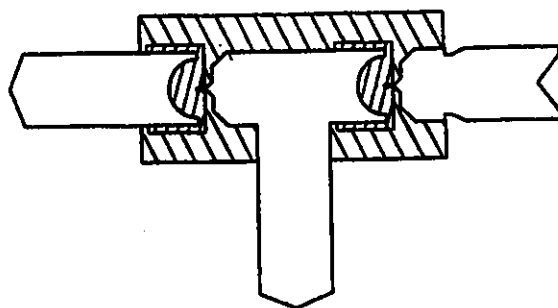


Figure 1.2.2.1.2 Beam-Lead Diodes in T-Configuration
(Metelics Corporation MSS40,241-B20)

1.2.2.2 Evaluation of the Impedance presented to the RF, LO, and IF Ports

The 180° hybrid is formed by the junction of a balanced fin-line and an unbalanced coplanar line. The diodes are mounted across the junction as shown in Figure 1.2.2.2.1. The RF signal is fed to the diodes in phase, and the LO signal is fed to the diodes 180° out of phase. Therefore, the diode pair appears in series from the RF port and in parallel from the IF and LO ports. Given the impedance values for a single diode (1.2.2.1.1) the impedances seen from the RF, LO and IF ports are:

$$Z_{RF} = 2 Z_{diode} = (36.8 + j 24.2)\Omega$$

for Z_{diode} at 24.4 GHz (1.2.2.2.1)

$$Z_{LO} = Z_{diode}/2 = (9.2 + j 6.1)\Omega$$

for Z_{diode} at 24.0 GHz (1.2.2.2.2)

$$Z_{IF} = Z_{diode}/2 = (9.5 + j 0.1)\Omega$$

for Z_{diode} at 400 MHz (1.2.2.2.3)

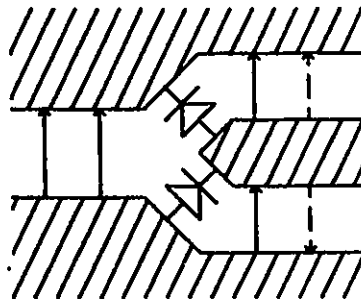


Figure 1.2.2.2.1 180-Degree Planar Hybrid Showing Location of Beam-Lead Diodes and direction of E-Field

→ RF signal
 → LO signal

1.2.2.3 Realization of the RF and LO Matching Circuitry

All three mixer ports must be matched to the diode for proper operation. Mismatch loss present at the RF and IF ports is directly added to mixer conversion loss and noise figure. Mismatch at the LO port might cause the oscillator to oscillate at a frequency different from the design value (Frequency pulling). Therefore, proper diode matching is crucial in mixer design.

1.2.2.3.1 RF Matching Circuit

To achieve good matching between the RF input waveguide and the fin-line medium, a cosine square fin-line taper is incorporated into the circuit. The taper transition is described in Section 1.2.3. In view of the relatively low diode impedance and in view of the difficulties in fabricating a very narrow slot, the gap width (d) has been chosen to fit the size of the T-configuration beam lead diode, ($d = 0.27 \text{ mm}$).

A quarter-wave transformer in front of the mixer diodes provides the remaining impedance transformation over the required 500 MHz bandwidth. Figure 1.2.2.3.1 shows the design parameters of the transformer, where the length (L_t) and the width (D_t) of the transformer plus its distance (L_p) from the diode represent the unknown variables.

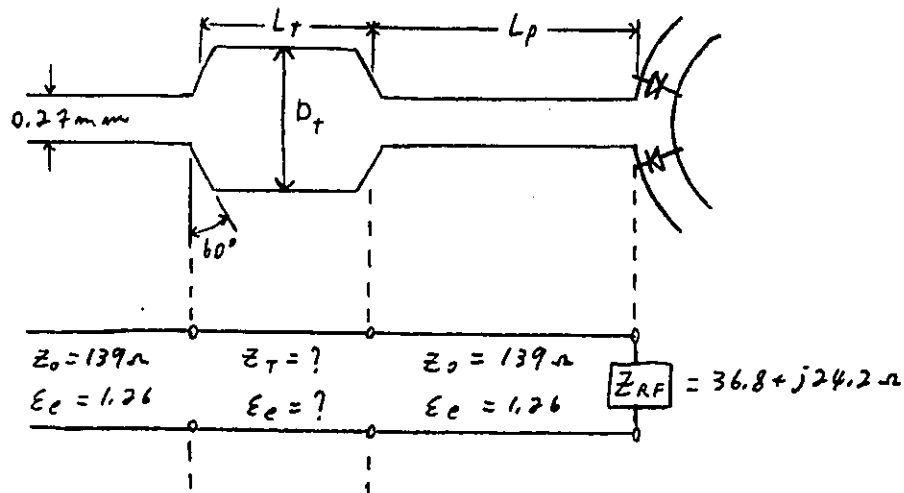


Figure 1.2.2.3.1: RF Quarter-Wave Matching Circuit
(Unilateral Fin-Line)

From Appendix A, for WR-42, $f = 24.4$ GHz, $\epsilon_r = 2.22$, substrate thickness (s) = 0.254 mm (10 mils), and $d = 0.27$ mm

$$Z_0 = 139 \, \Omega \quad (\text{Characteristic Impedance})$$

$$\epsilon_e = 1.26 \quad (\text{Effective Dielectric Constant})$$

Knowing Z_0 , ϵ_e , and the load impedance ($Z_{RF} = [36.8 + j24.2]\Omega$) from (1.2.2.2.1), the transformer distance from the load (L_p) is obtained as follows:

$$Z_{RF} \text{ normalized to } Z_0 \text{ gives } z_{rf} = 0.2644 + j 0.1745$$

$$\text{for which } z_{\text{minimum}} = .260 \text{ for } L_p = 0.718 \lambda_g$$

$$z_{\text{maximum}} = 3.85 \text{ for } L_p = 0.218 \lambda_g.$$

Since z_{maximum} is easy to realize and shorter than z_{minimum} :

$$L_p = 0.218 \lambda_g = 0.218 \times 10.95 \text{ mm} = 2.39 \text{ mm}$$

Knowing Z_0 and z_{maximum} , we obtain the transmission line characteristic of the transformer as follows:

$$Z_t = (Z_0 \times Z_0 \times z_{\text{maximum}})^{1/2} = 273 \, \Omega,$$

$$\text{Then from Appendix A, for } Z_t = 273 \, \Omega \text{ and } \epsilon_e = 1.023, \quad D_t = 1.35 \text{ mm}$$

Finally, knowing the characteristics of the transformer (ϵ_e), we obtain its length (L_t) at 24.4 GHz ($\lambda_0 = 12.25$ mm) as follows:

$$\text{Knowing that, } L_t = \lambda_g/4 \text{ which is identical to}$$

$$L_t = \lambda_0/(4\sqrt{\epsilon_e}), \text{ we obtain } L_t = 3.03 \text{ mm.}$$

1.2.2.3.2 LO Matching Circuit

As mentioned before, the LO is fed to the mixer diodes through the coplanar line (odd-mode) via a coax-to-waveguide transition. The fin-line provides the RF return for the diode through the mixer housing if there is no external bias circuitry. The coplanar line (shown in Appendix D, Section D.1, Figure D.1.1) has two modes of propagation, the odd mode and even mode, which are shown in Appendix D in Figures D.2.1 and D.3.1, respectively. Due to the mixer characteristic, both modes are present. The LO and IF propagate as odd mode while the RF propagates as even mode.

Figure 1.2.2.3.2 shows the known and unknown variables regarding the coplanar-line section circuitry. The coplanar line must satisfy the E-field distribution at both transitions, coplanar to fin-line and coplanar to microstrip. At the diode junction the ratio W_1/S_1 must be very small so that the majority of the RF E-field is concentrated in each slot, thus increasing the RF energy transfer to each diode. At the LO junction the ratio W_2/S_2 must be large to increase the LO energy coupled to the coplanar line. Note that since the IF frequency is much lower than the RF (ratio of 60:1) the step discontinuity created by the S_1W_1 to S_2W_2 transition has nearly no effect at IF.

The coplanar line must provide a way to return the percentage of RF which has not been converted to IF. Two step impedances are used to this end. The first discontinuity presents a nearly open circuit at RF which is created by the S_1W_1 to S_2W_2 transition. The second discontinuity is caused by the IF input impedance, presenting almost a short circuit at RF.

The length of each coplanar section must be chosen such that an open circuit appears in parallel with the diodes at RF. The diode impedance, and hence the RF input impedance, is thus not affected by the coplanar circuit port.

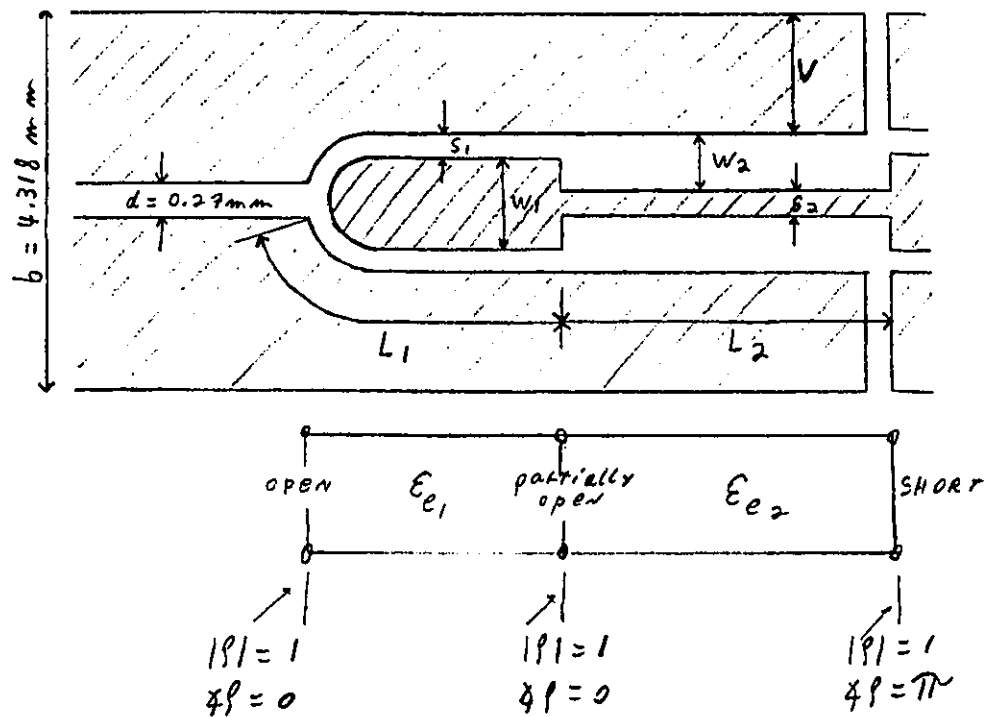


Figure 1.2.2.3.2 Coplanar-line Section Circuitry

To reduce complexity, V is chosen such that the presence of the waveguide inner wall does not have to be considered in the calculation of Z_0 and ϵ_e given by an open structure model. Hatsude [10] proved that if $V/(S/2 + W)$ is larger than 1.5 the presence of the waveguide inner wall has no effect. Refer to Figure 1.2.2.3.2 for definition of S , V and W . Therefore;

$$V/U \approx 2 \quad (\text{to allowed margin})$$

$$V + U = b/2$$

where,

$$U = S/2 + W$$

$$b = 4.32 \text{ mm (WR-42 waveguide b-dimension)}$$

then:

$$S/2 + W = 0.070 \text{ mm}$$

$$V = 1.46 \text{ mm}$$

As mentioned previously, in order to concentrate the RF field within the coplanar slot (S), the ratio of W_1/S_1 must be as small as possible. For optimum matching, W_1 is limited by the physical dimension of the beam-lead diode, hence $W_1 = 0.20 \text{ mm}$. Knowing W_1 , S_1 can be found as:

$$S_1 = 2(0.70 \text{ mm} - W_1) = 1.0 \text{ mm}$$

Similarly, as mentioned before, $W_2 \gg S_2$. The maximum ratio for W_2/S_2 is limited to 3. This limitation is due to the parameter boundary imposed by the empirical formula given in Appendix D, Section D-3 (even-mode). Therefore, for

$$W_2/S_2 \approx 3, \text{ and}$$

$$S_2/2 + W_2 = 0.70 \text{ mm}$$

we obtain:

$$W_2 = 0.60 \text{ mm}$$

$$S_2 = 0.20 \text{ mm}.$$

As mentioned previously, L_1 is chosen such as to present an open circuit at the mixer diode junction.

Then:

$$L_1 = \lambda_g/2 = \lambda_0/(2\sqrt{\epsilon_{e1}}),$$

where:

$$\lambda_0 = 12.3 \text{ mm at RF}$$

ϵ_{e1} is given by Appendix D, Section D-3 (even mode).

For $S_1 = 1.0 \text{ mm}$ and $W_1 = 0.20 \text{ mm}$:

$$Z_1 = 50 \Omega, \epsilon_{e1} = 1.6, \text{ and}$$

$$L_1 = 4.8 \text{ mm}.$$

As mentioned subsequently, L_2 is required to transform the short circuit to an open circuit, which through L_1 transforms into an open circuit at the mixer diode junction. Then:

$$L_2 = 3\lambda_g/4 = 3\lambda_0/(4/\epsilon_{e2})$$

where,

$$\lambda_0 = 12.3 \text{ mm at RF}$$

ϵ_{e2} is given by Appendix D, Section D-3 (even mode).

For $S_2 = 0.20 \text{ mm}$ and $W_2 = 0.60 \text{ mm}$;

$$Z_2 = 250 \Omega, \epsilon_{e1} = 3.4, \text{ and}$$

$$L_2 = 4.7 \text{ mm.}$$

1.2.3 Design of Fin-Line Transmission Line Transition

The realization of a balanced mixer is a good example for illustrating the necessity of integrating different transmission-lines on a single substrate. However, we must realize and characterize the transitions between them. Ideally a transition must satisfy many constraints, such as: low VSWR, wide bandwidth, small dimensions, low loss and ease of fabrication. However, it is natural that some criteria cannot be achieved simultaneously, such as wide bandwidth and small dimensions for example.

Transitions can be divided into two kinds, namely the input/output-to-mixer transitions, and transitions within the mixer. The former include the waveguide to fin-line taper, the latter include the fin-line to coplanar even mode hybrid junction, the fin-line to coplanar odd mode (LO 'probe type') coupling and the coplanar-to-microstrip (IF filter) transition which will be discussed in Chapter 2, Section 2.1.3.

1.2.3.1 Empty to Partially-Dielectric-Loaded Waveguide Transition

This transition is located at the RF-port and is used to match the empty WR-42 waveguide to a partially-dielectric-loaded WR-42 waveguide required to support the receiver front-end E-plane circuitry. The dielectric is made of RT-Duroid soft substrate ($\epsilon_r = 2.22$). This transition is used only when the RF-band pass filter is not present since the BPF is designed for empty WR-42 waveguide matching.

Design

Closed-form expressions for the design of a quarter-wave transformer in the form of a protrusion in the fin-line substrate at the interface with a waveguide have been proposed by Verver and Hoefer [11].

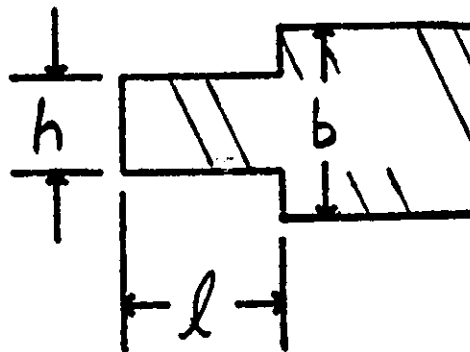


Figure 1.2.3.1.1 Empty to Partially-Dielectric-Loaded Waveguide Transition

Figure 1.2.3.1.1 shows the protrusion structure, where h and l are defined as;

$$h = \frac{(\sqrt{k_{e2}} - 1)ab}{\sqrt{k_{e2}} (\epsilon_r - 1)S} \quad (1.2.3.1.1)$$

$$l = \frac{\lambda}{4} (k_{e2} - p^2)^{-1/2} \quad (1.2.3.1.2)$$

where

$$k_{e2} = p^2 + [(1-p^2)(k_{e3}-p^2)]^{1/2}$$

$$\text{where } k_{e3} = [1 - (\epsilon_r - 1) s/a]^{-2},$$

$$p = \lambda/(2a).$$

a , b , and s are defined in Appendix A, Section A-4, Figure A.4.1..

Then, at 24.4 GHz (RF-port):

$$h = 2.16 \text{ mm}$$

$$l = 3.60 \text{ mm.}$$

1.2.3.2 Waveguide to Fin-Line Transition

Transferring electromagnetic energy from a waveguide to a dielectric supported fin-line built in the body of the same waveguide is often achieved by tapering the metallic ridges forming the fins. In this way a portion of incident energy which was initially TE_{10} mode is converted into the dominant fin-line mode. Obviously, the taper efficiency is a function of its profile.

Design

To achieve good RF matching, a double exponential fin-line taper developed by Pramanick [12] is incorporated between the RF bandpass filter and the mixer hybrid junction. It is shown in Figure 1.2.3.2.1.

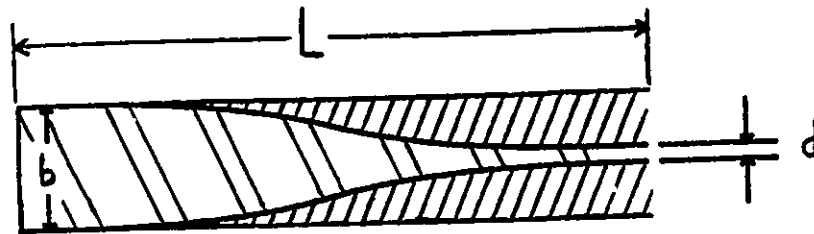


Figure 1.2.3.2.1 Double Exponential Fin-Line Taper

$D(Z)$ describes the profile of the double exponential taper:

$$D(Z) = D_1 - \left(\frac{D_1 - D_2}{2} \right) \left(\frac{2Z}{L} \right)^t \quad \text{for } 0 \leq Z \leq L/2 \quad (1.2.3.2.1a)$$

$$= D_2 + \left(\frac{D_1 - D_2}{2} \right) \left(2 - \frac{2Z}{L} \right)^t \quad \text{for } L/2 \leq Z \leq L \quad (1.2.3.2.1b)$$

where: $D_1 = b/2$, $b = 4.318$ mm, the WR-42 height
 $D_2 = d/2$, $d = 0.27$ mm, the final gap size of the fin-line for achieving the optimum impedance matching to the mixer diodes
 $L = 2\lambda_g$, $\lambda_g = 6.4$ mm, the guided wavelength corresponding to the lower limit of the RF band
 $t = 2.5$, optimum value for unilateral fin-line [12], (obtained empirically).

Result

The transition was first built and tested as an individual component. Figure 1.2.3.2.2 shows the theoretical and experimental insertion loss for a back-to-back transition. Typical insertion loss of about 0.15 dB per transition is obtained.

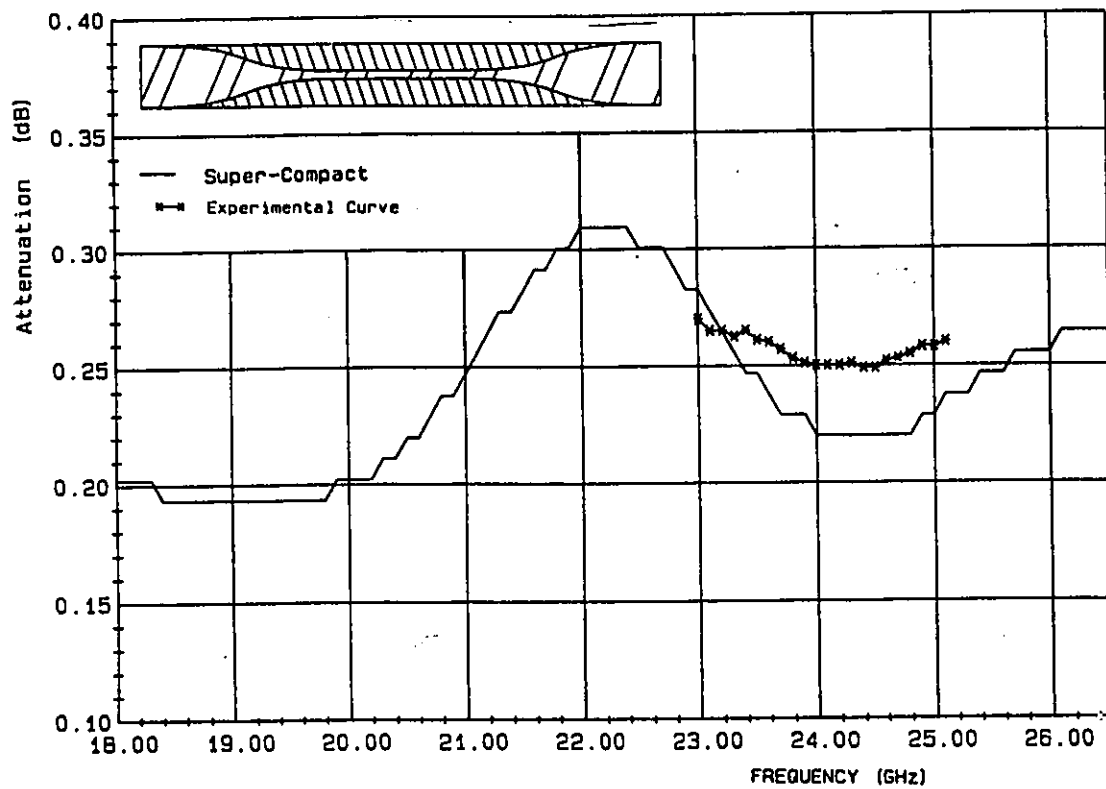


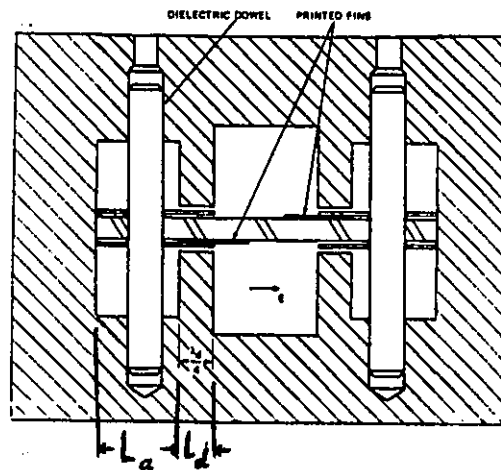
Figure 1.2.3.2.2 Back-to-Back Unilateral Fin-Line Transition Measurement

1.2.3.3 Microstrip to Coaxial Transition

A straightforward classical microstrip to coaxial transition is realized by using a female SMA microstrip launcher with a typical insertion loss of less than 0.05 dB.

1.2.3.4 Serrations in the Substrate Clamping Region

Two different styles of housing are currently used. The main differences between the two designs is in the clamping region, where the waveguide housing clamps the fin-line substrate. In one approach, the substrate is cut to fit in a narrow groove in the waveguide block. This technique requires tight tolerances for the groove height (roughly ± 10 micrometer) to achieve proper electrical contact between metalization and housing. In another approach, originally used by Meier [13], the substrate is positioned and held in place by alignment pins in the split waveguide block as shown in Figure 1.2.3.4.1. To avoid power leakage through the clamping region, the length of the clamping region L_d is chosen as $\lambda_{\epsilon_r}/4$, where λ_{ϵ_r} is the wavelength in the substrate of dielectric constant ϵ_r .



$$L_d = (\lambda_0 / 4 \sqrt{\epsilon_r}) = 2.06 \text{ mm at } 24.4 \text{ GHz}$$

$$L_a = \lambda_0 / 4 = 3.07 \text{ mm at } 24.4 \text{ GHz}$$

Figure 1.2.3.4.1 Fin-Line Housing with Open Clamping Region [13]

Design

To suppress longitudinal currents, and thus the propagation of parasitic modes in the clamping region, a serration pattern is incorporated in the circuit-board metalization as represented in Figure 1.2.3.4.2. Note that a thin insulating sheet must be placed between housing and serrations in order not to DC short-circuit them.

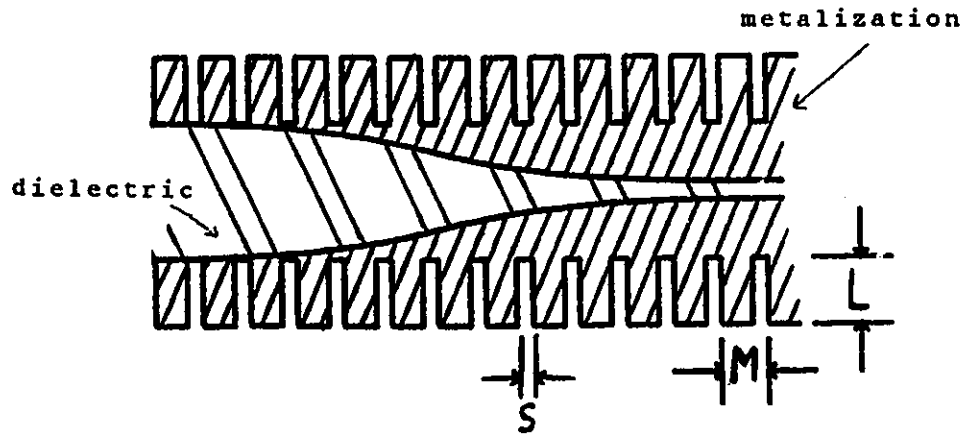


Figure 1.2.4.3.2 Serration Pattern

Here: $L = L_d = 2.06 \text{ mm}$ at 24.4 GHz
 $M = \lambda_0 / (10\sqrt{\epsilon_r}) = 0.82 \text{ mm}$ at 24.4 GHz
 S is the minimum etchable slot, $S = 0.1 \text{ mm}$.

1.3 Experimental Results for the Balanced Fin-Line Mixer

Figure 1.3.0.1 shows the equipment set-up for the mixer performance measurement.

All measurements have been performed with a nominal LO power of 10 dBm and a maximum RF power of -20 dBm.

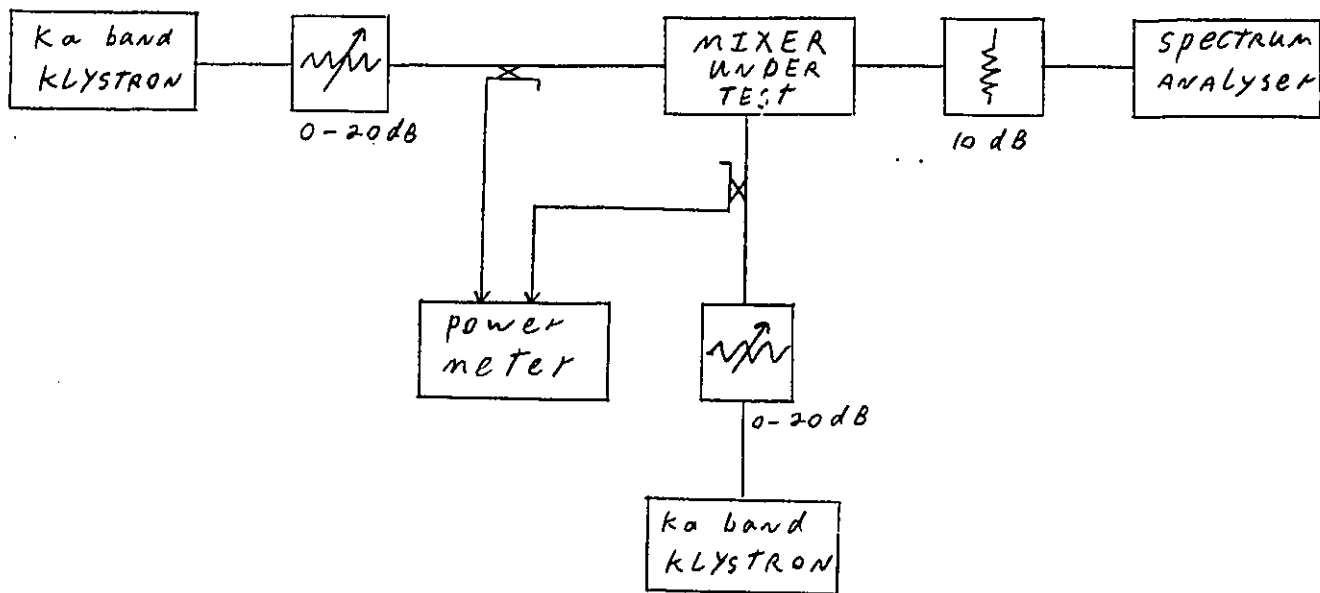


Figure 1.3.0.1 Mixer Performance Measurement Set Up

1.3.1 RF Input Return Loss

Figure 1.3.1.1 presents the theoretical and the measured return loss at the RF input port, the latter being 15 dB on the average over 500 MHz bandwidth, which corresponds to VSWR of 1:1.4. Note that the RF bandpass filter is not present.

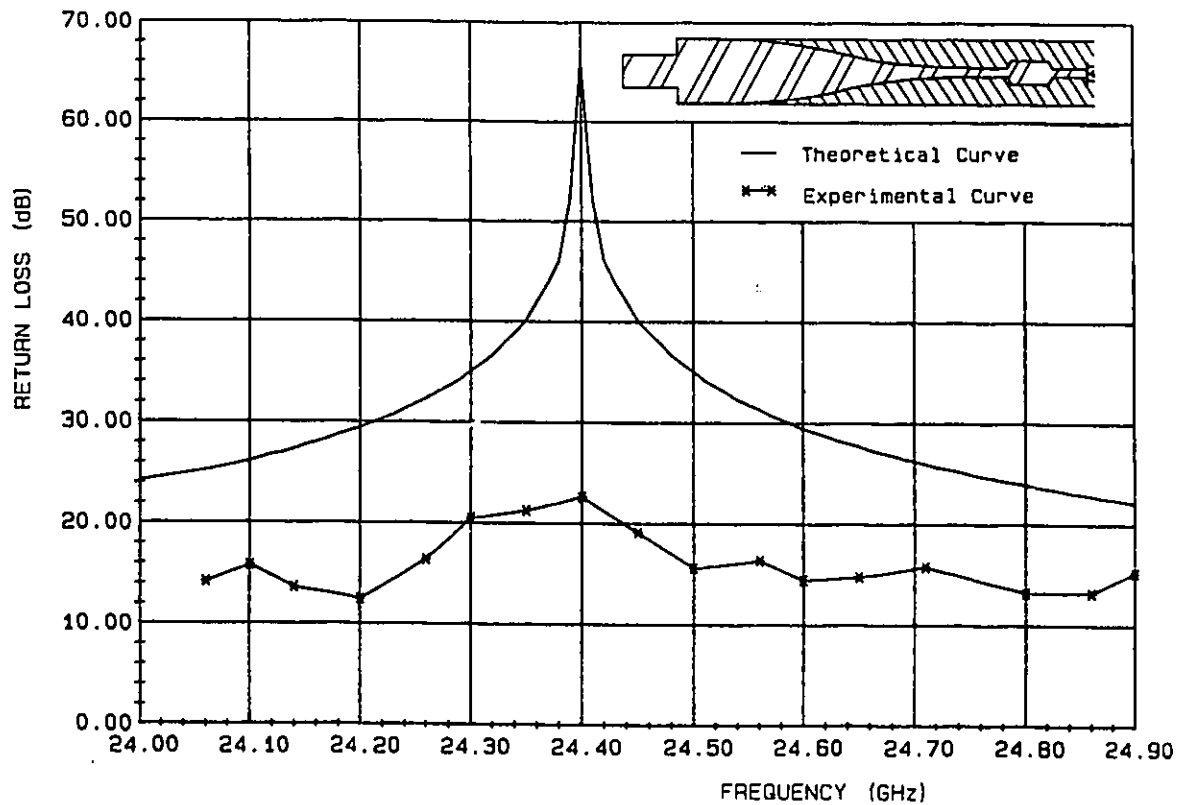


Figure 1.3.1.1 Mixer RF Input Return Loss

1.3.2 IF Output Return Loss

Figure 1.3.2.1 shows the theoretical and experimental curve for the IF output return loss of the mixer. The experimental curve resembles the theoretical curve. The lowest return loss is 8 dB at 150 MHz, whereas 18 dB is achieved at 450 MHz.

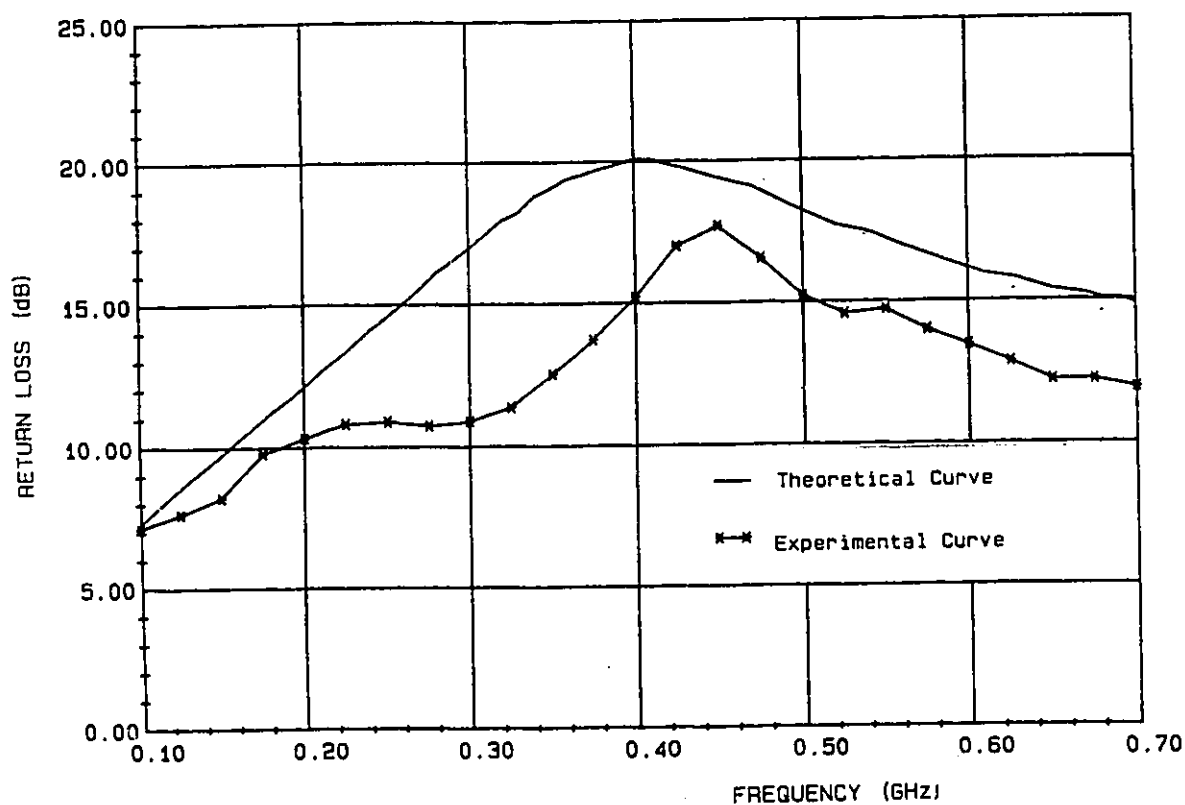


Figure 1.3.2.1 Mixer IF Output Return Loss

1.3.3 Isolation

Very good isolation between the three ports is obtained due to the properties of the 180 degrees hybrid and the filters in the RF and IF ports. The measurements have been performed only at band center (RF = 24.4 GHz) and are as follows:

Ports	Isolation (dB)
LO - RF	42
LO - IF	53
RF - IF	40

Table 1.3.3.1 Mixer Port Isolation (RF = 24.4 GHz and LO = 24 GHz)

1.3.4 Conversion Loss

The theoretical minimum conversion loss for this type of mixer approaches 3 dB, since it is not an image recovery mixer. Because of the non-ideal mixer diode characteristics, this limit goes up to about 4.5 dB. The culprits are a series resistance of 7 ohms and an ideality factor of 1.07. Other features limiting the conversion loss are the insertion loss of the bandpass filter (0.5 dB), of the waveguide to fin-line transition (0.1 dB), and the insertion loss of the IF filter, including the SMA connector (0.2 dB). Given these limitations, the expected optimum conversion loss is 5.4 dB. Figure 1.3.4.1 shows the measured value (with the bandpass filter included) around 5.5 dB in the band center, and around 5 dB at the band edges.

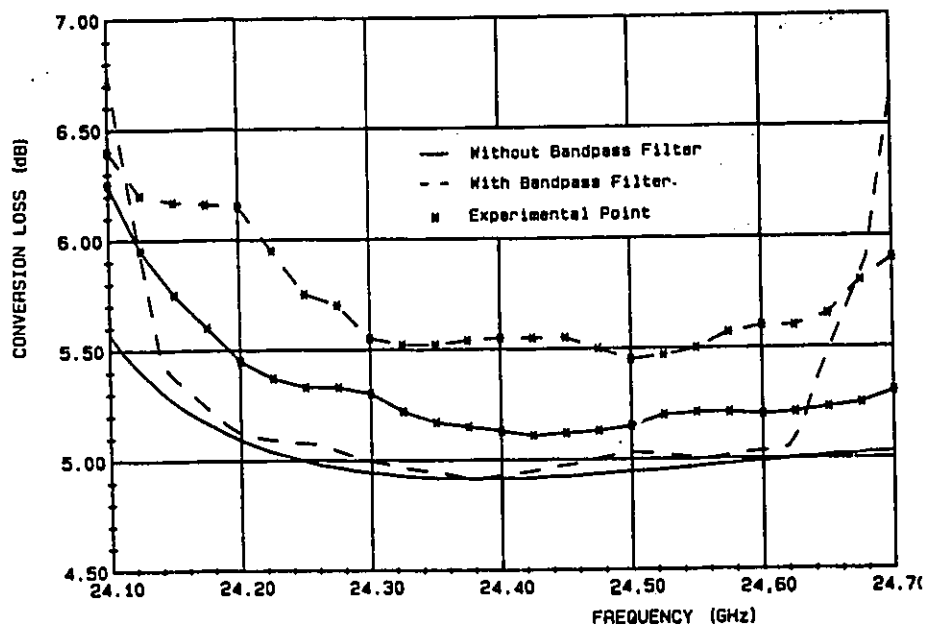


Figure 1.3.4.1 Mixer Conversion Loss

1.3.5 Single Sideband Noise Figure

Noise is generated by three different sources in this front end. The predominant contributions are the thermal noise associated with lossy components, and mismatch. This part is equivalent to the conversion loss, namely 5.5 dB. An additional contribution is the AM- FM- and PM- noise generated by the local oscillator. Thanks to the hybrid junction, the AM noise is practically eliminated. Roughly one additional dB of noise is thus generated by the LO. The flicker noise which is a third contribution, has the smallest effect. It is generated by the bias and saturation currents in the mixer diodes. Good diodes generate around 0.5 dB. In all, a single sideband noise figure of around 7 dB has been achieved.

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CHAPTER 2

FILTER THEORY AND DESIGN

In any mixer, a low pass filter is required to pass the IF frequency and reject the LO and RF signals. In the crossbar mixer (see Appendix C, Section C.4, Figure C.4.1), the RF and LO ports are isolated because the input waveguides are orthogonal to each other. In the fin-line mixer, however, a band pass filter is needed to increase the LO/RF isolation. Three solutions regarding the location of the band pass filter can be considered. First, a RF-BPF (band pass filter with center frequency at RF) and LO-BPF may be implemented in the RF and LO ports, respectively. Even though this measure will increase the LO/RF isolation the presence of both filters will enlarge considerably the size of the mixer. As a second solution, only one BPF may be implemented and located in the LO port. This configuration has the advantage of presenting a good VSWR to the oscillator, hence increasing the oscillator stability. As a last solution, a unique BPF can be realized and located in the RF port. This configuration has the great advantage of rejecting the image frequency band as well as reducing the LO-radiation at the RF-port. This last configuration has been chosen for its good image-attenuation as represented in Figure 2.0.1. Refer to Appendix E, Section E-1, for a detailed description of the parameters required to evaluate the performance of the filter.

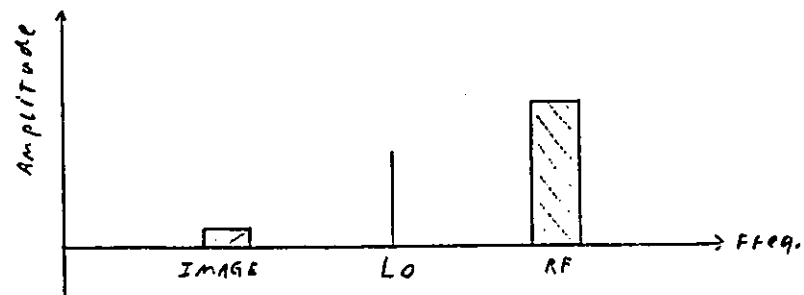
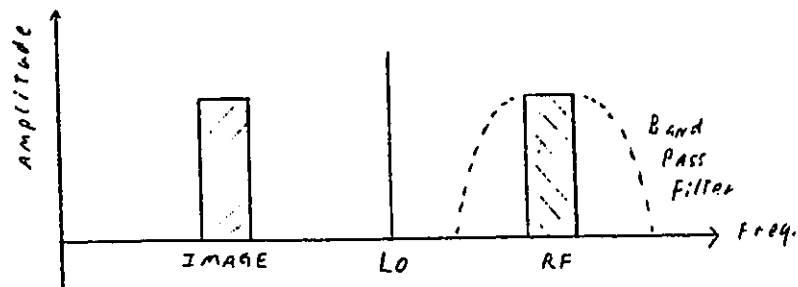


Figure 2.0.1 RF Input Frequency Spectrum Before and After Filtering

2.1 Low Pass Filter Theory and Design

2.1.1 Low Pass Filter Theory

Firstly, it should be mentioned that the design procedure as well as the relevant equations were obtained from Edwards, [1]. The design employs cascaded sections of microstrip as indicated in Table 2.1.1.1 in which the lumped filter elements are replaced by microstrip sections.

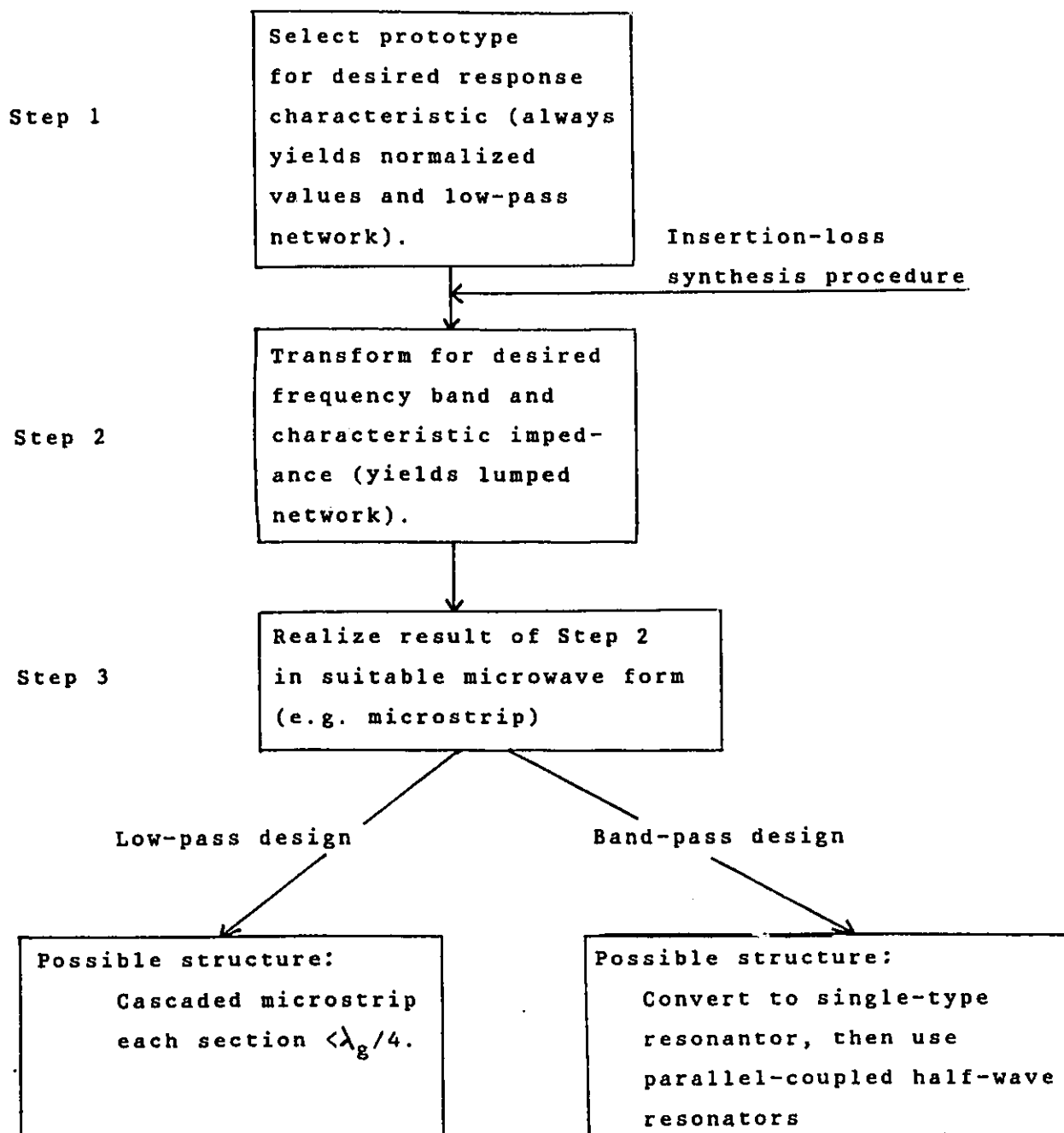
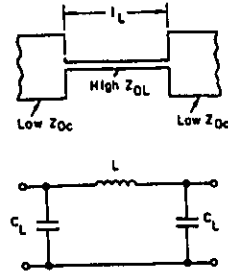


Table 2.1.1.1 Procedure for Microstrip LPF Design

Short ($< \lambda_g/4$) lengths of relatively high impedance line will behave predominantly as series inductances. Also, a very short ($\ll \lambda_g/4$) length of relatively low impedance line will act predominantly as a shunt capacitance. So a Π network of lumped elements can be realized with alternate sections of low and high impedance microstrip lines. The following Table 2.1.1.2 presents a summary chart [1] showing the results applicable to inductive and capacitive elements realized in the form of microstrip lines.

Mainly inductive element



$$X_L = Z_{0L} \sin\left(\frac{2\pi l_L}{\lambda_{gL}}\right) \quad (2.1.1.1a)$$

End-capacitances

$$C_L \approx \frac{l_L}{2f Z_{0L} \lambda_{gL}} \quad (2.1.1.1b)$$

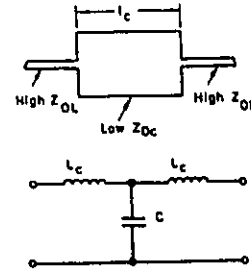
Main inductive length

$$l_L = \frac{\lambda_{gL}}{2\pi} \sin^{-1}\left(\frac{\omega L}{Z_{0L}}\right) \quad (2.1.1.1c)$$

For short-lengths only

$$l_L \approx \frac{f \lambda_{gL} L}{Z_{0L}} \quad (2.1.1.1d)$$

Mainly capacitive element



$$B_C = \frac{1}{Z_{0c}} \sin\left(\frac{2\pi l_c}{\lambda_{gc}}\right) \quad (2.1.1.2a)$$

End-inductances

$$L_c \approx \frac{l_c Z_{0c}}{2f \lambda_{gc}} \quad (2.1.1.2b)$$

Main capacitive length

$$l_c = \frac{\lambda_{gc}}{2\pi} \sin^{-1}(\omega C Z_{0c}) \quad (2.1.1.2c)$$

Since length are usually short, we can often use

$$l_c \approx f \lambda_{gc} Z_{0c} C \quad (2.1.1.2d)$$

Table 2.1.1.2 Summary Chart for Inductive and Capacitive Element of a Microstrip Lowpass Filter [1]

The step discontinuity is significant, since it results in some kind of end-capacitance and end-inductance which will interact with the adjacent microstrip sections. The end-inductance is usually negligible while the end-capacitance is not. The capacitance calculated for the capacitive length of line must in fact be less than the required total lumped value C_T because the inductive sections of line and the steps in width already contribute their capacitances, C_L and C_S , respectively.

The capacitive line is finally given by:

$$C_{\text{line}} = C_T - C_{L1} - C_{L2} - C_{S1} - C_{S2} \quad (2.1.1.3)$$

where

C_{L1} and C_{L2} are obtained from 2.1.1.1b)

C_{S1} and C_{S2} are derived by curve fitting from Garg [2].

For $\epsilon_r \leq 10$; $1.5 \leq w_2/w_1 \leq 3.5$;

$$\frac{C}{\sqrt{w_1 w_2}} = (10.1 \log \epsilon_r + 2.33) \frac{w_2}{w_1} - 12.6 \log \epsilon_r - 3.17 \text{ pF/m} \quad (2.1.1.4a)$$

For $\epsilon_r = 9.6$; $3.5 \leq w_2/w_1 \leq 10$:

$$\frac{C}{\sqrt{w_1 w_2}} = 130 \log(w_2/w_1) - 44 \text{ pF/m} \quad (2.1.1.4b)$$

where w_2/w_1 is the step ratio as requested in Figure 2.1.2.2.

2.1.2 Design of a Low Pass Microstrip Filter

The choice of the cut-off frequency of the low pass filter is governed by three considerations. Firstly, the bandwidth must be as small as possible to reduce the noise at the IF output port. Secondly, the RF-IF and LO-IF isolations must be as large as possible. Thirdly, the physical size of the filter must be as small as possible. In the design of the filter, one must also consider the fact that the pass band will be repeated for each multiple of the cut-off frequency, and may include the RF and/or the LO frequency. Based on all these factors, the cut-off frequency of the IF - low pass filter has been set at 2 GHz, with an IF bandwidth of 400 MHz.

The number of filter sections (poles and zeros) and the kind of filter are also dependent on the application. Too many filter sections will increase the losses, the passband ripple, and the size, but will present a greater stopband attenuation. A Chebyshev type bandpass filter with five sections and a 0.1 dB ripple has been chosen.

According to Matthaei [3], the low-pass filter attenuation characteristic is expressed mathematically as:

$$L_A \text{ (dB)} = 10 \log_{10} \left\{ 1 + E \cos^2 \left[n \cos^{-1} \left(\frac{f}{f_c} \right) \right] \right\} \quad \text{for } f \leq f_c \quad (2.1.2.1a)$$

and

$$L_A \text{ (dB)} = 10 \log_{10} \left\{ 1 + E \cos^2 \left[n \cos^{-1} \left(\frac{f_c}{f} \right) \right] \right\} \quad \text{for } f > f_c \quad (2.1.2.1b)$$

where; L_A represents the attenuation in dB

n is the number of sections

$$E = \left[\text{antilog}_{10} \left(\frac{L_{ar}}{10} \right) \right] - 1$$

where L_{ar} is the passband peak-to-peak ripple in dB.

For: $n = 5$, $L_{ar} = 0.1$ dB and $f_c = 400$ MHz. This theory predicts 115.6 dB and 116.3 dB for RF and LO attenuations respectively, which is much more than required.

This terminates Step 1 of Table 2.1.1.1. The next step, defined as stage 2 in table 2.1.1.1 is the calculation of the lumped low-pass network. Figure 2.1.2.1 shows the network topology, where the lumped element values given by Matthaei [3] are,

$$L_1 = L_5 = 28.67 \text{ nH}, L_3 = 49.38 \text{ nH}, C_1 = C_4 = 13.71 \text{ pF}$$

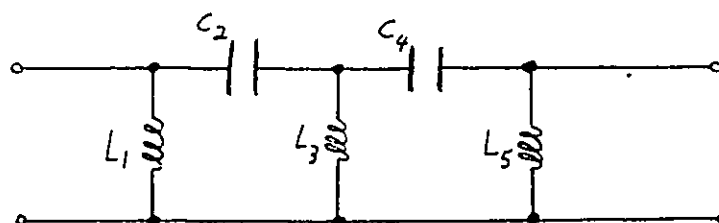


Figure 2.1.2.1 Network Topology of the LPF Prototype

At the fairly low microwave frequency of 2 GHz, dispersion can be neglected. To arrive at the general microstrip layout needed to realize this filter, the lumped components shown in Figure 2.1.2.1 are replaced by an appropriate length of microstrip. The result is illustrated in Figure 2.1.2.2.

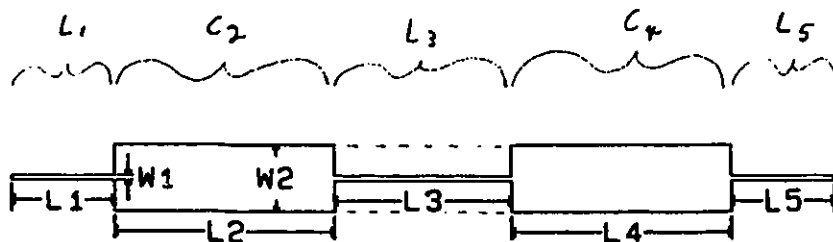


Figure 2.1.2.2 Microstrip Configuration for a Five-Section LPF

It is reasonable to initially fix the characteristic impedances. Their orders of magnitude are mainly determined by the following considerations:

- a) Z_{oc} (capacitance section); the resulting width $W_c = W_2 = W_4$, must not allow transverse resonances to occur at the highest frequencies of operation, and must also fit into a WR - 42 waveguide.
- b) Z_{ol} (inductance section); the inductance must not be so large that it becomes unrealizable or its current-carrying capability is severely limited, resulting in increased insertion loss.

Because the parameters for the various lengths of line are interrelated we must perform initial calculations and then correct progressively as follows:

- a) Determine initial values for the lengths of microstrip line (l_L) to replace inductances.
- b) Calculate the end capacitances associated with these lengths l_L .
- c) Calculate the step-discontinuity capacitance.
- d) Reduce the originally specified lumped capacitance values by the appropriate amounts as calculated in b) and c), taking care to refer to the microstrip configuration so that the correct subtractions are carried out.
- e) Calculate initial values for the lengths of microstrip lines (l_c) to replace these new capacitances.

- f) Determine the inductances of the T-sections associated with these capacitive lengths of line.
- g) Reduce the originally specified lumped inductance values by the appropriate amounts as calculated in f).
- h) Repeat the entire procedure, starting again with the new values (from step g) in step a).

When successive values converge to within about 5 per cent, the final values can be accepted with little residual error.

A fully implemented CAD routine is listed in Appendix E, section E-3. For this particular design, the following filter dimensions have been obtained. See Figure 2.1.3.1 for filter geometry.

$L_1 = L_5 = 6.720 \text{ mm}$
 $L_2 = L_4 = 14.922 \text{ mm}$
 $L_3 = 11.425 \text{ mm}$
 $W_1 = 0.2 \text{ mm}$
 $W_2 = 3.0 \text{ mm}$
 for $\epsilon_r = 2.22$ and $S = 20 \text{ mils.}$

2.1.3 Characteristics and Measurement Results of the Low Pass Microstrip Filter

Figure 2.1.3.1 shows three curves. The theoretical curve is the characteristic of the lumped element model in Figure 2.1.2.1. The dashed line has been obtained with Super-Compact, and the broken line is the measured characteristic. The experimental results have been predicted accurately by the Super-Compact analysis. The differences between the theoretical curve given by expression 2.1.2.1 and the two other curves are mainly due to intercoupling between elements which is not taken into consideration by the theoretical mathematical model.

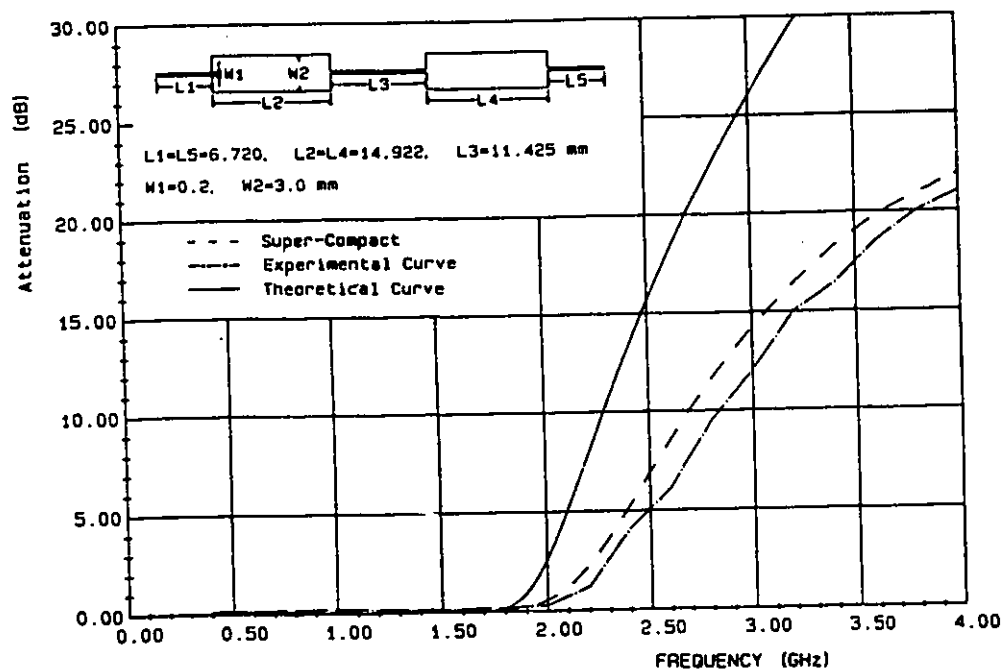


Figure 2.1.3.1 Microstrip LPF Dimensions and Theoretical, Analytical, and Experimental Results

2.2 Theory and Desing of the RF Band Pass Filter

2.2.1 Band Pass Filter Theory

The band pass filter should possess low insertion loss in the passband as its attenuation is added to the mixer noise figure. The fin-line structure has been chosen because in addition to having low insertion loss, the integration with the overall receiver is straightforward. In order to reduce the insertion loss, the height of the fin-line resonators has been made equal to the waveguide height at the RF port.

The fin-line filter has been realized in bilateral configuration to minimize insertion loss and preserve electrical symmetry in the structure in order to reduce the excitation of odd higher order modes. The attenuation is influenced by the smoothness and conductivity of the fin-line surface, mainly at the high current density edges. An important requirement is that the filter should possess a significant attenuation at the image and local oscillator frequencies.

2.2.2 Design of a Band Pass Fin-Line Filter

The filter is designed to pass the RF signal (center frequency 24.4 GHz with a bandwidth of 500 MHz) and a maximum of 0.5 dB insertion loss, while attenuating the image signal (23.6 GHz, bandwidth 500 MHz) by at least 20 dB. An optimized three-resonator bilateral fin-line filter of full waveguide height has been chosen.

The filter has been performed with the EPLANFIL software developed by R. Vahldieck based on previous work [4], and realized on 20 mils thick RT-Duriod 5880 ($\epsilon_r = 2.22$) suspended in the E-plane of WR-42 waveguide.

The physical dimensions of the filter are given in the upper right corner of the figure 2.2.3.1 where L_2 , L_4 , L_6 and L_8 are the lengths of metallic septa of height 4.318 mm. Note that stringent tolerances must be respected during circuit fabrication.

2.2.3 Characteristics and Measurement Results of the Band Pass Fin-Line Filter

Both the theoretical and experimental curves of the filter are given in Figure 2.2.3.1. The theoretical curve for $|S_{12}|$ predicts the RF passband insertion loss within less than 0.3 dB. The image rejection at 23.6 GHz is greater than 25 dB as predicted by the theoretical curve for $|S_{12}|$. The return loss characteristics $|S_{11}|$ are also in general agreement, the high peaks in the theoretical curve resulting from the absence of losses in the numerical model.

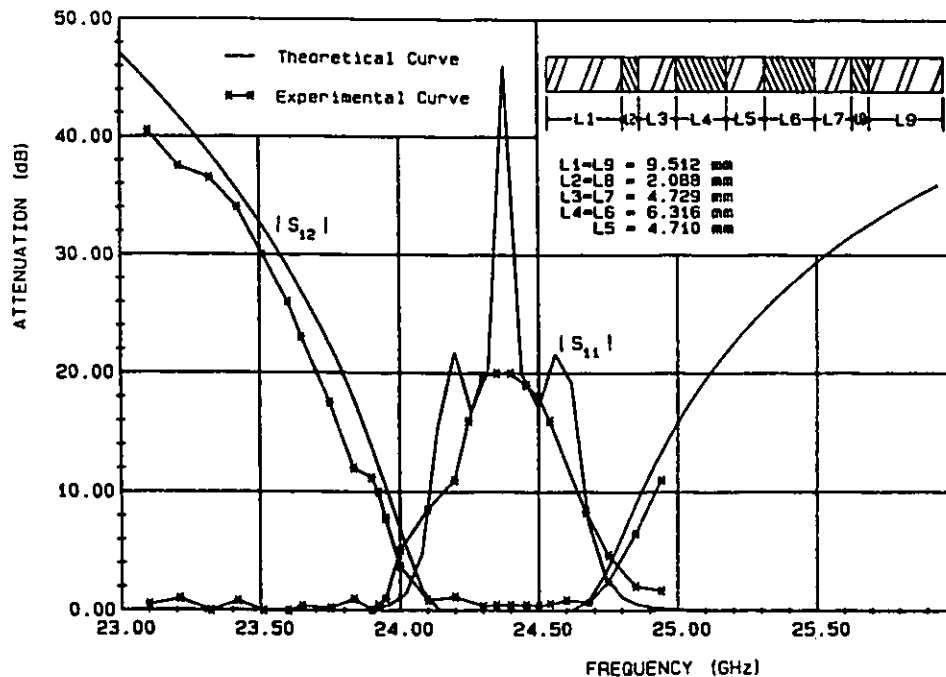


Figure 2.2.3.1 Fin-Line BPF Dimension and Analytical Experimental Result

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CHAPTER 3

THEORY AND DESIGN OF THE LOCAL OSCILLATOR

The local oscillator is a critical component in modern RF systems. Important parameters such as stability, output power, efficiency, mechanical/electrical tuning, spurious outputs and AM/FM noise must be optimized to produce the type of device necessary for today's sophisticated RF systems. Refer to Appendix F, Section F-1 for a detailed description of the oscillator parameters.

Ideally, an oscillator converts 100% of the bias energy into energy at a unique frequency required for the frequency conversion process.

The availability of highly stable, low loss dielectric material has led to the development of a new class of microwave signal sources, the dielectric resonator oscillator. The generation of stable, narrowband microwave signals has typically been performed by cavity-stabilized Gunn oscillators or crystal-multiplier chains. However, with the need for smaller, more efficient, more reliable sources, the DRO is finding increasing application, and will be used as a local oscillator required by the front-end receiver.

The dielectric resonator can be integrated with a variety of active devices to cover the microwave frequency spectrum. GaAs FET's has been selected due to their low cost.

FET oscillators utilizing a dielectric resonator as a stabilizing element are classified in three types of oscillator circuits, the DSO (dielectrically stabilized oscillator), the DRO (dielectric resonator oscillator) organized in series feedback and the DRO organized in shunt feedback. It is the shunt feedback DRO which has been selected for this thesis. Refer to Appendix F, section F-2 for a description of the DSO and the DRO in series feedback.

DRO (Dielectric Resonator Oscillator) in Shunt Feedback

As shown in Figure 3.0.1, in the shunt feedback arrangement, the dielectric resonator is placed between the output and input of the device (e.g. between the gate and source or gate and drain circuits). The conditions for oscillation are met at the resonant frequency of the dielectric resonator. However, in the shunt feedback scheme, the resonator is very strongly coupled to the drain and gate transmission lines.

While the DRO offers several advantages over other microwave signal sources, it is not perfect for all applications. Thus, its characteristics should be compared with those of other potential sources. Table 3.0.1 offers a brief summary comparison of the Gunn cavity oscillator, the crystal multiplier chain, and the DRO [1].

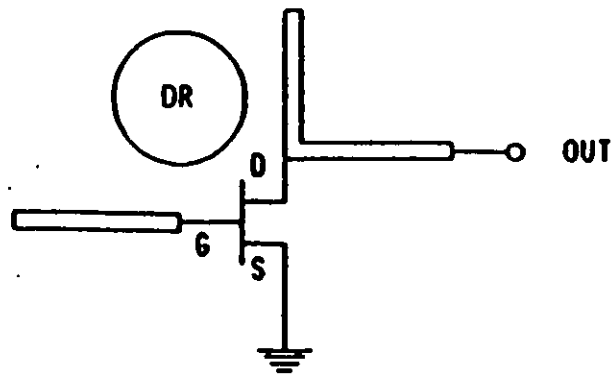


Figure 3.0.1 Basic Configuration of Shunt Feedback DRO

Characteristic	Gunn Cavity Oscillator	Crystal Multiplier	DRO*
Efficiency	Low	Low	High
Temperature Range	Poor	Good	Excellent
Size	Small	Large	Smallest
Power Variation	High	High	Low
FM Noise	Excellent	Very Good	Excellent
Frequency Stability	Good	Excellent	Very Good
Environmental Stability	Fair	Fair	Excellent
Reliability	Fair	Good	Excellent

*for both, series and shunt feedback.

Table 3.0.1 Summary of Comparison of Oscillators [1]

3.1 Dielectric Resonator Oscillator Theory

In this section, the analysis and characterization of the dielectric resonator oscillator will be described.

3.1.1 Oscillator Circuit

The oscillator circuit is based on a field effect transistor (FET) in common source configuration where the gate load consists of a dielectric resonator. The drain provides the oscillator output. The theory of the dielectric resonator is described later in Section 3.1.4.

The design consists of determining the FET (common source) S-parameters at different frequencies around the design frequency, and then tracing the output plane's instability circle. Finally, the position of the dielectric resonator is obtained. This design procedure is described step-by-step with the relevant equations in the following paragraphs.

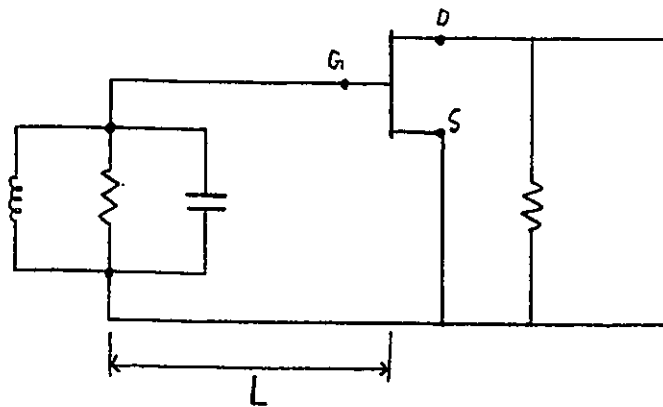
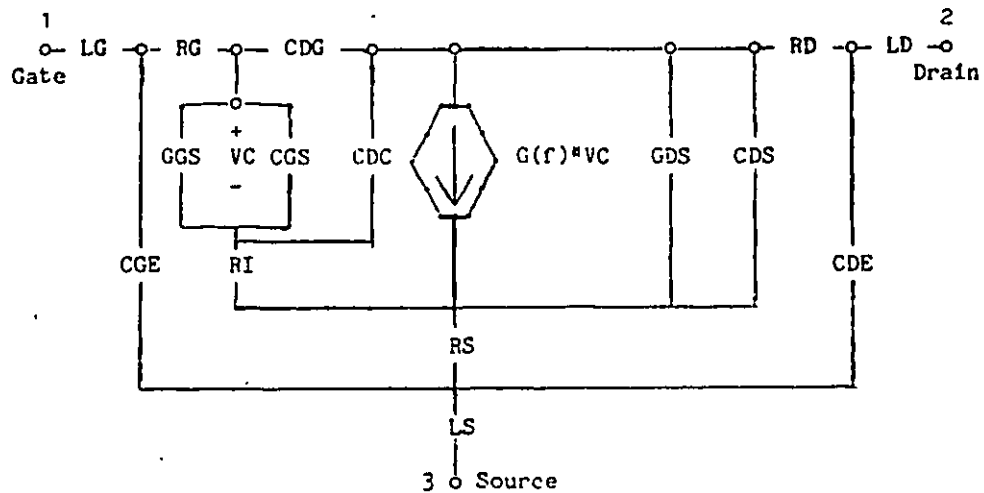


Figure 3.1.1.1 Basic Common Source FET-DRO Circuit

3.1.2 FET (Common Source) S-Parameters

Knowing the FET-model, the FET S-parameters are obtained as follows:

An accurate FET model is necessary to predict the S-parameters of the FET at the oscillator design frequency. Figure 3.1.2.1 shows the FET model used by Super-Compact.



$$G(f) = G_0 * \exp(-j * \Omega * T_d) / (1 + jf/F_{3dB})$$

$$\Omega = 2 * \pi * f, \quad f = \text{frequency}$$

In this figure:

G0	Transconductance at dc	GDS	Drain-source conductance
CGS	Gate-source capacitance	RI	Channel resistance
F3dB	3-dB roll-off frequency	RG	Gate resistance
Td	Time delay	RD	Drain resistance
GGs	Gate-source conductance	RS	Source resistance
CDG	Drain-gate capacitance	ODE	External gate capacitance
CDC	Dipole layer capacitance	LG	Gate lead inductance
CDS	Drain-source capacitance	LD	Drain lead inductance
LS	Source lead inductance		

Figure 3.1.2.1 Field-Effect Transistor Model

Knowing the S-parameters at lower frequencies using the data sheet, a FET model can be evaluated and used to predict the S-parameters at higher frequencies. Super Compact has been used to evaluate the FET NE677383 model parameters at 24 GHz from the known FET S-parameters at frequencies 2 to 18 GHz (refer to Appendix G, Section G-2).

Figure 3.1.2.2 reproduces the Super Compact optimization file listing where FET NE67383 model parameters are obtained.

```
*FET MODEL OPTIMISED FROM 2 TO 18 GHZ.
* SCOMPACT FILE TO MODEL THE FET.
*
*
LAD
  FET 1 2 G=? .03769? CGS=? .24075PF? T=?7.1733E-12? GCS=?1.1191E-3?
+ CDG=? .01PF? CDC=? .018PF?
+ CDS=? .0713PF? R1=?15.556? RD=?3.8259? RG=?5.9256?
+ RS=?4.1081?
+ GDS=?4.0144E-3? CGE=? .03903PF? CDE=? .07036PF?
+ LG=? .60504NH? LD=? .53031NH? LS=?5.9947E-3NH?
MODEL: 2POR 1 2
END
FREQ
  STEP 10GHZ 22GHZ 2GHZ
  STEP 24.0GHZ 24.8GHZ .1GHZ
  40GHZ
END
OUT
  PRI MODEL S
END
OPT
MODEL S=R W 10 10 10 10
END
DATA
Q: NE67383 FILE=FILE0.2
END
```

Figure 3.1.2.2 Super Compact Listing of FET NE67383 Model

In table 3.1.2.1, FET NE67383 S-parameters are listed for some frequencies of interest, especially at 24.0 GHz. The S-parameters for 10, 12, 14, 16 and 18 GHz match with the S-parameters listed in the data sheet shown in Appendix G, Section G-2. Therefore, it can be assumed that the S-parameter computed for higher frequency are accurate.

JPER-COMPACT Version 1.7 + 001 04/09/84 15-NOV-84 14:00:24

CIRCUIT: MODEL

MATRIX, ZS = 50.0+J 0.0 ZL = 50.0+J 0.0

Freq	S11		S21		S12		S22		S21
GHZ	Mag	Ang	Mag	Ang	Mag	Ang	Mag	Ang	dB
10.00000	0.531	-138.4	2.541	41.0	0.092	16.7	0.538	-84.6	8.10
12.00000	0.493	178.0	2.270	9.9	0.093	-3.1	0.473	-113.4	7.12
14.00000	0.523	141.4	1.919	-19.7	0.097	-22.3	0.445	-146.7	5.66
16.00000	0.577	115.8	1.556	-47.2	0.095	-40.2	0.474	-179.7	3.84
18.00000	0.630	98.4	1.223	-72.4	0.089	-56.5	0.546	152.4	1.75
20.00000	0.678	86.2	0.940	-95.2	0.080	-71.0	0.635	130.4	-0.54
22.00000	0.721	77.2	0.711	-115.7	0.071	-83.6	0.718	113.1	-2.96
24.00000	0.759	70.1	0.534	-133.7	0.061	-94.3	0.786	99.3	-5.45
24.10000	0.761	69.8	0.527	-134.6	0.061	-94.8	0.789	98.7	-5.57
24.20000	0.763	69.4	0.519	-135.4	0.060	-95.3	0.792	98.1	-5.70
24.30000	0.765	69.1	0.512	-136.2	0.060	-95.8	0.795	97.5	-5.82
24.40000	0.766	68.8	0.504	-137.1	0.060	-96.3	0.798	96.9	-5.95
24.50000	0.768	68.5	0.497	-137.9	0.059	-96.7	0.801	96.3	-6.07
24.60000	0.770	68.2	0.490	-138.7	0.059	-97.2	0.804	95.7	-6.20
24.70000	0.772	67.9	0.483	-139.5	0.058	-97.7	0.806	95.1	-6.32
24.80000	0.773	67.6	0.476	-140.3	0.058	-98.1	0.809	94.6	-6.45
40.00000	0.920	40.2	0.062	136.3	0.020	-135.6	0.968	48.7	-24.21

Table 3.1.2.1 Estimated S-Parameters of the FET NE67383

3.1.3 Analysis of Instability

The instability factor (k) is defined as:

$$K = \frac{1 + |D|^2 - |S_{11}|^2 - |S_{22}|^2}{S |S_{12} S_{21}|} \quad (3.1.3.1)$$

is: The distance of the center position of the instability circle

$$OC = \frac{N^*}{|S_{22}|^2 - |D|^2} \quad (3.1.3.2a)$$

with radius

$$R = \frac{|S_{12} S_{21}|}{||S_{22}|^2 - |D|^2|} \quad (3.1.3.2b)$$

$$\text{where } D = S_{11}S_{22} - S_{12}S_{21}$$

$$N = S_{22} - DS_{11}^*$$

*denotes the conjugate of...

Figure 3.1.3.1 represents all twelve possible cases for instability [2]. A FET is called unconditionally stable, if it is stable for any passive load located at the input and output. The input and output stability circles must not overlap with the fundamental Smith chart. For $S_{11} < 1$, this is satisfied in cases 1 and 4. For $S_{11} > 1$, it is never satisfied. A FET is called conditionally stable, if it is stable for some passive loads at the input and output. This is true for cases 2, 3, 5, 6, 8, 9, 11, and 12. Cases 7 and 10 are unconditionally unstable.

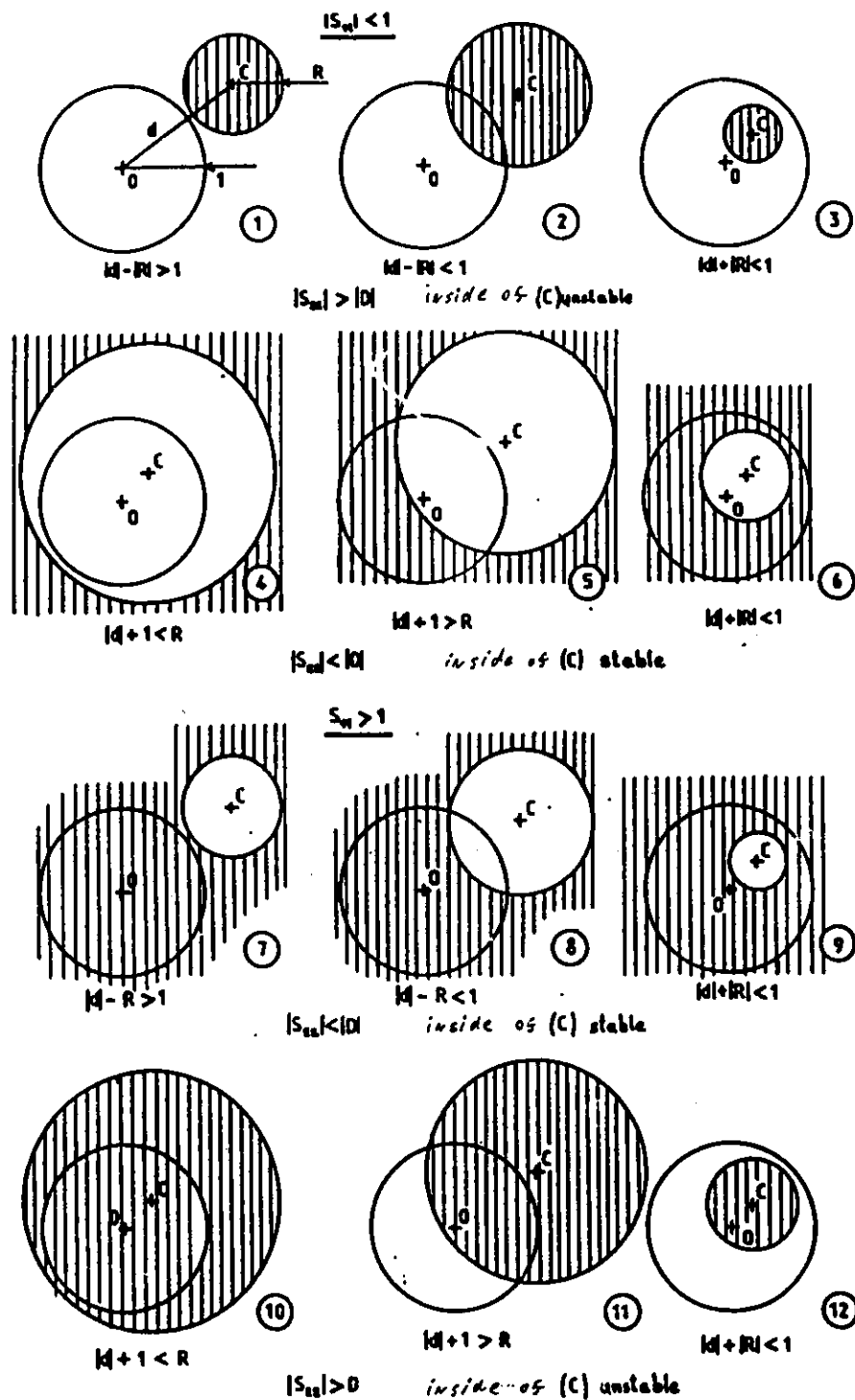


Figure 3.1.3.1 Twelve Cases of Instability [2]

3.1.4 Characteristics of the Dielectric Resonator Used for the Oscillator

In 1939, Richmyer [3] showed that unmetallized dielectric objects can function as resonators just like metallic cavities. He called them dielectric resonators (D.R.s). Practical applications of dielectric resonators to microwave circuits, however, began to appear only in the late 60's as resonating elements in waveguide filters. Recent developments in ceramic material technology have resulted in improvements including small controllable temperature coefficients of the resonant frequency over the useful operating temperature range, and very low dielectric losses at microwave frequencies. These developments have generated interest in dielectric resonator applications for a wide variety of microwave circuit configurations and subsystems [4].

Some of the advantages of the substitution of conventional resonators by dielectric resonators are:

- 1) Smaller circuit sizes.
- 2) Greater degree of circuit and subsystem integration.
- 3) Better circuit performance, when compared to MIC line resonators, with regard to both temperature and losses.
- 4) Reduction of overall circuit cost for comparable performance.

The important material properties for the dielectric resonator applications are: the temperature coefficient of the resonant frequency (τ_f), the unloaded Q factor (Q_0) and the dielectric constant of the material (ϵ_r).

Among all the theoretically possible geometries, the cylindrical shape (Figure 3.1.4.1) has been widely accepted as the most advantageous one.

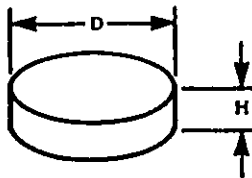


Figure 3.1.4.1
Dielectric Resonator
Cylindrical Shape [1]

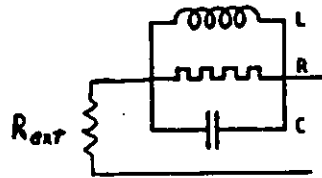


Figure 3.1.4.2
DR Equivalent Circuit

Figure 3.1.4.2 presents the equivalent electrical circuit for the dielectric resonator. The coupling factor (β), being the ratio of the energies present outside and within the resonator, is given by:

$$\beta = \frac{R}{R_{ext}} \quad (3.1.4.1)$$

where, R is shown in Figure 3.1.4.2
 R_{ext} is the characteristic impedance of the coupling transmission line.

Dielectric resonator frequencies are usually computed by assuming that the dielectric resonator is placed in unbounded space. In real conditions, however, dielectric resonators are placed in microwave structures, like waveguides, striplines, microstrip or fin-line transmission lines. Because these microwave structures are close to the resonators, they disturb the resonator's external fields, and alter their resonant frequency. A complete solution for the resonant frequency of the DR is obtained through the transverse resonance procedure [5]. From that, the physical dimension of the DR are obtained.

$$D = 2.405 \frac{C}{2\pi f} \left[\frac{1}{\sqrt{\epsilon_r}} + \frac{1}{\sqrt{\epsilon_s}} \right] \quad (3.1.4.2)$$

$$D/H \approx 3$$

where: D and H are shown in figure 3.1.4.1

ϵ_r is the relative dielectric constant of the resonator material

ϵ_s is the relative dielectric constant of the substrate material.

3.1.5 Location of the Dielectric Resonator

Two parameters are needed to determine the position of the DR (dielectric resonator), namely the coordinates of the instability circle and the DR coupling factor. The latter has been described in the previous section. The procedure is as follows: locate the output instability circle, rotate the circle by 180 degrees about the center of the Smith chart to obtain its image, trace the resonator admittance circle which corresponds to a conductance of $1/\beta$ where β is given by (3.1.4.1), and finally rotate the admittance circle until some overlap is obtained with the image of the circle. If maximum overlap is chosen, then it results in low Q and less sensitivity to oscillator output load. However, if minimum overlap is selected, it results in high Q and high sensitivity to oscillator output load. Therefore, as a rule of thumb, measure the overlap area for both cases and set the operating point such that the overlap area is the average of both extreme cases. The rotation of the admittance circle corresponds to the electrical distance of the DR from the FET, hence the position of the DR represented by 'L' on Figure 3.2.1.1. Note that if there is no overlap of both circles, the FET will never oscillate.

3.2 Design of a 24 GHz Fin-Line Dielectric Resonator Oscillator

In this section, the design and realization of a 24 GHz dielectric resonator oscillator in fin-line configuration will be described.

3.2.1 Fin-Line Dielectric Resonator Oscillator

The main problem in designing a fin-line FET dielectric resonator resonator oscillator (FET DRO) is to obtain a circuit configuration which permits simple installation and biasing of the transistor. Such a configuration has been found in [6]. The circuit consists of two fin-lines having a boomerang shape and forming a cross with four ports, as presented in Figure 3.2.1.1. Port 4 is used as the output of the oscillator, while ports 1, 2, and 3 are short-circuited in order to establish the appropriate resonance conditions.

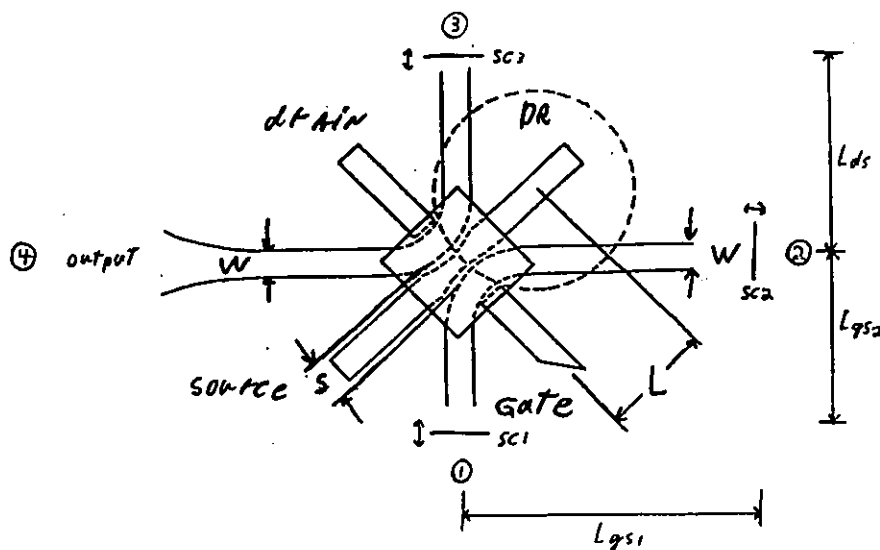


Figure 3.2.1.1 Basic Set Up of a Fin-Line FET Oscillator

The FET is placed at the center of the cross where the two slots are closest to each other. Bias voltages are applied directly to the fins via the serrations. Low pass filters and chokes are used on all bias lines to present a high impedance at RF. The transistor is mounted in a stripline package. (Low-noise Ku to K band GaAs - MESFET, NEC part number NE67383). Refer to Appendix G, Section G-2 for specifications.

The feedback is accomplished via the dielectric resonator which is placed on the back of the substrate and positioned to couple the gate to the source (DRO effect) as well as the drain to the source (DSO effect) (see Figure 3.2.1.1). The coupling between the lines and the resonator can be adjusted by moving the resonator along the direction of the source leads. The dielectric resonator is made of ZrSn Titanate ($\epsilon_r = 37.9$, Trans-Tech part number D-8513.125.050). Refer to Appendix G, Section G-3 for specifications.

To permit tuning of the frequency (SC2 at port 2 and SC1 at port 1) and of the power (SC3 at port 3), the three short-circuits are made adjustable. They consist of blocks of brass with a slot at one end to accept the substrate. Each short-circuit can be moved by a screw.

3.2.2 Design of the Fin-Line Dielectric Resonator Oscillator

The oscillator design has been divided into three phases: the localization of the instability circle, the realization of the feedback circuitry, and the realization of the transmission-line.

3.2.2.1 The Localization of the Instability Circle

Using the FET NE67383 specifications (see Appendix G, Section G-2) with the FET Model (see Section 3.1, Figure 3.1.1), the FET S-parameter at 24 GHz were obtained and are listed below:

$$\begin{aligned} S_{11} &= 0.759 \angle 70.1^\circ = 0.258 + j0.714 \\ S_{21} &= 0.534 \angle -133.7^\circ = -0.369 - j0.386 \\ S_{12} &= 0.061 \angle -94.3^\circ = -0.0046 - j0.061 \\ S_{22} &= 0.786 \angle 99.3^\circ = -0.127 + j0.776 \end{aligned}$$

If the S-parameters are known, the stability factor (K) and the instability circle coordinates (OC and R) are obtained through equations (3.1.3.1) and (3.1.3.2) respectively, then:

$$\begin{aligned} k &= 1.013, \quad (\text{instability, since } K > 1) \\ |OC| &= 1.21 \\ \angle OC &= 83.1 \text{ degrees} \\ R &= 0.311 \end{aligned}$$

Refer to Figure 3.2.2.2.2 for graphical representation.

3.2.2.2 Realization of the Feedback Circuitry

As mentioned in the introduction, the feedback is realized through the use of a dielectric resonator (DR) in order to obtain good phase noise performance. After the DR dimensions have been defined, the coupling factor is obtained experimentally. Finally, from the instability circle coordinates, the physical location of the DR is obtained.

The physical dimensions of the DR are given by the relations (3.1.4.2), where at $f = 24.0$ GHz and for

$$\epsilon_r = 2.22 \text{ and } \epsilon_s = 37.9:$$

the diameter, $D = 0.157$ mm, and

the height, $H = 0.052$ mm.

Therefore, from Appendix G, Section G-3, the DR part number D-8513.125.050 is selected.

The final parameter required to determine the location of the DR, is the coupling factor (β). The coupling factor can be obtained analytically using Equation 3.1.4.1, or experimentally. The experimental approach has been selected since the characteristic impedance of the coupling transmission line (R_{ext} , (3.1.4.1)) is unknown for the present fin-line-to-DR coupling structure.

The coupling factor (β) is obtained experimentally by measuring the coupled energy at the resonant frequency (24 GHz) and any off resonance (i.e. 22 GHz). The coupled energy is obtained by measuring the insertion loss for a signal input at port 4 and measured at port 2 without the FET installed. The short circuits SC1 and SC3 at port 1 and 3 respectively are readjusted for maximum coupling for each different value of L. The coupling factor (β) is equivalent to the ratio of S_{21} at 22 GHz to S_{21} at 24 GHz, where S_{21} is represented in Figure 3.2.2.2.1. The variable "L" represents the distance between the center of the DR and the origin as shown in Figure 3.2.1.1.

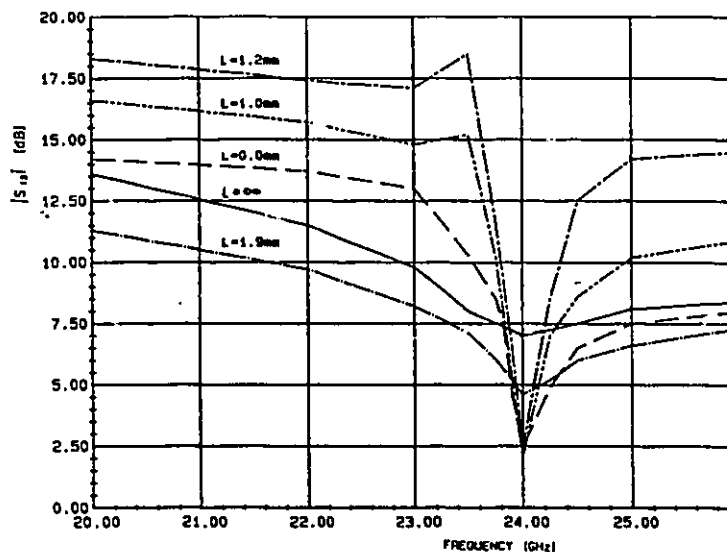


Figure 3.2.2.2.1 Coupling Measurement of the DR

Since L is a function of β and vice-versa, a range for the most likely value of L needs to be assumed first in order to evaluate β . Therefore, β is determined for $L = 0$ mm and 1.2 mm, which is:

$$\begin{aligned} \text{for: } L = 0 \text{ mm, } & \approx \text{ALOG} ((14.5 - 2.5)/10) = 15.8 \\ L = 1.2 \text{ mm, } & \approx \text{ALOG} ((16.5 - 2.5)/10) = 31.6 \end{aligned}$$

therefore, with some margin, $\beta = 20$.

It is now possible to determine the location of the DR represented by (L) in Figure 3.2.1.1.

The procedure for finding the location of the DR has been fully described in Section 3.1.5, and will not be repeated. As mentioned, two parameters are needed to determine L, namely the coordinates of the instability circle and the DR coupling factor. Figure 3.2.2.2.2 represents the graphical solution for L, where:

$$L = 0.05\lambda_g, 0.55\lambda_g, \dots,$$

and for $\lambda_g = 10.92$ mm (refer to Section 3.2.2.3)

$$L = 0.55, 6.01 \dots \text{mm}, \text{ respectively.}$$

Finally, $L = 0.55$ mm has been selected for ease of realization.

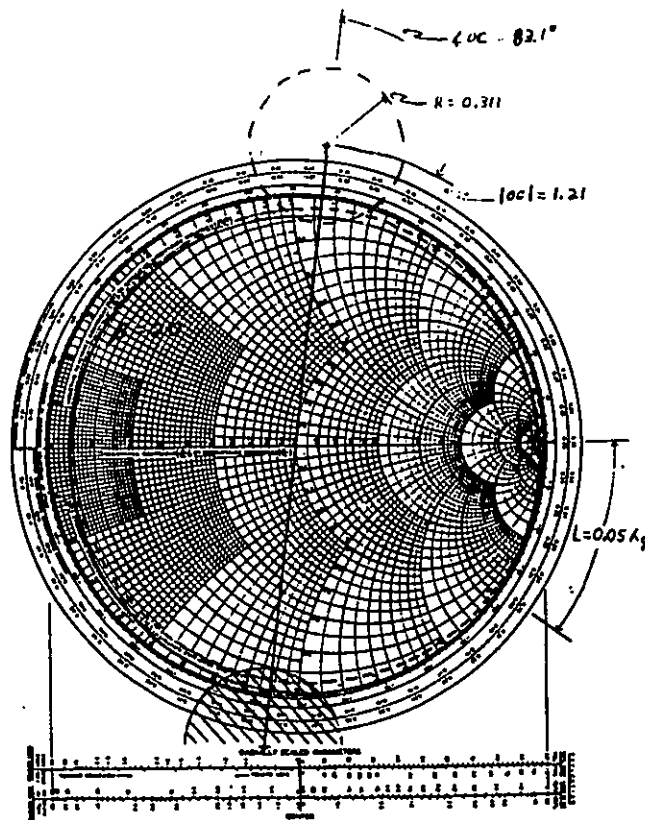


Figure 3.2.2.2.2 Graphical Solution for DR Position

3.2.2.3 The Realization of the Transmission-Line

The fin-line mutual spacing (S) and the fin-line gapwidth (W) are selected to achieve optimum E-field coupling between the fin-line and the FET. Refer to Figure 3.2.1.1 for the definition of S and W. The maximum gap width between the fins is determined by the physical distance between the leads of the transistor, as shown in Figure 3.2.2.3.1. The minimum gap width depends on the minimum etching resolution achievable. Hence:

$$W = 0.15 \text{ mm, and}$$

$$S = 0.65 \text{ mm.}$$

Additionally, for $\epsilon_r = 2.22$ and at 24.0 GHz, Appendix A, Section A-4 gives:

$$\lambda_g = 10.92 \text{ mm}$$

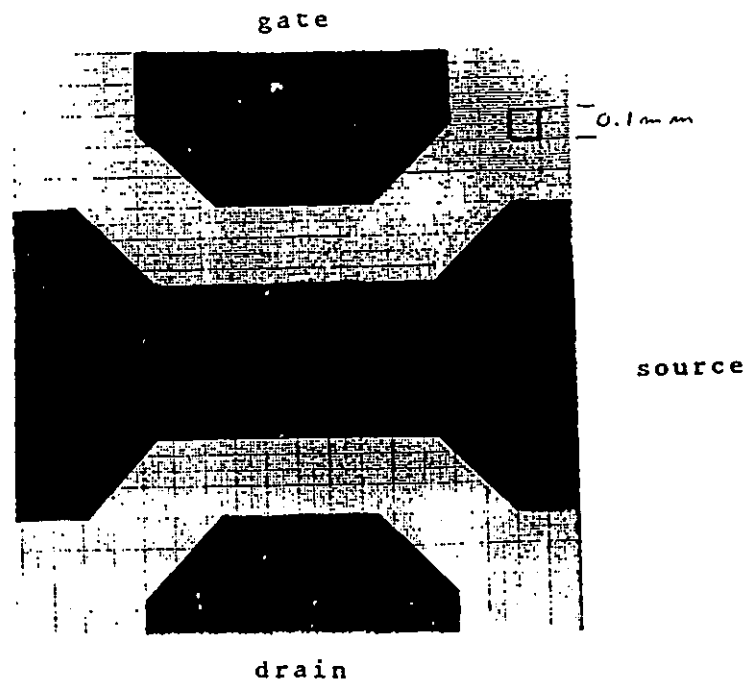


Figure 3.2.2.3.1 Close View of the back side of FET Showing the Leads (Metalization) in Black. Enlarged 40 times.

Figure 3.2.2.3.3a and b show the fin-line oscillator circuitry and the picture of the overall oscillator integrated to the mixer, respectively.

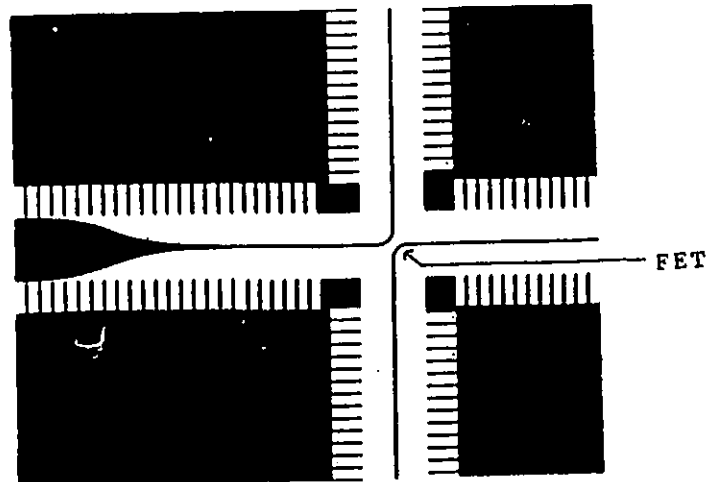


Figure 3.2.2.3.3(a) Fin-Line Oscillator Circuitry, Where Metalization is Represented By the White Area. Scale 2 to 1. Dimensions in mm.

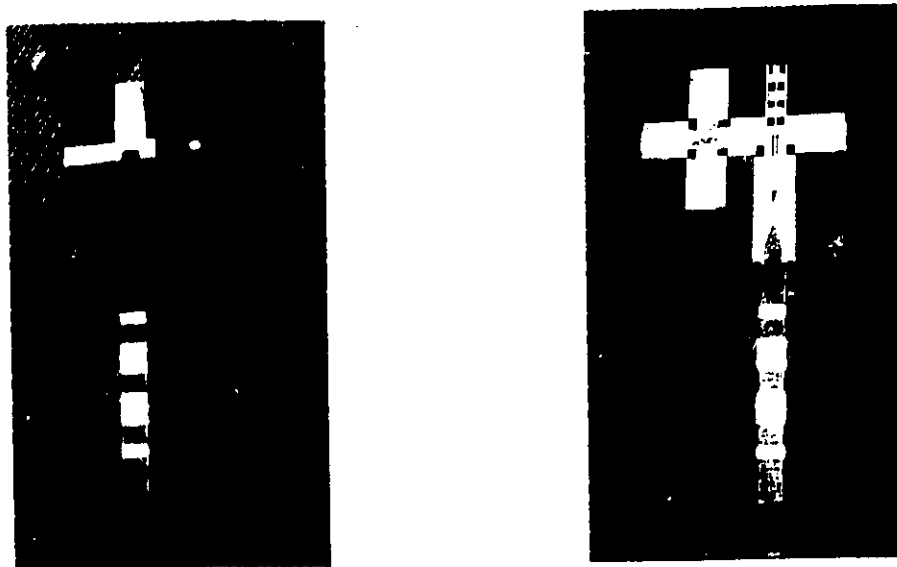


Figure 3.2.2.3.3(b) Pictures of Oscillator Substrate, Both Sides

3.2.3 Integration of the DRO with the Fin-Line Mixer

The electrical distance between the oscillator and the fin-line-to-coplanar coupling junction of the mixer is selected to present an open circuit (or a maximum impedance) to the latter for optimum LO power transfer from the DRO to the mixer. For proper operation, a movable short is placed at the transition allowing some fine-tuning of the LO input impedance. Figure 3.2.3.1 shows the known and unknown variables regarding the oscillator-to-mixer interconnection.

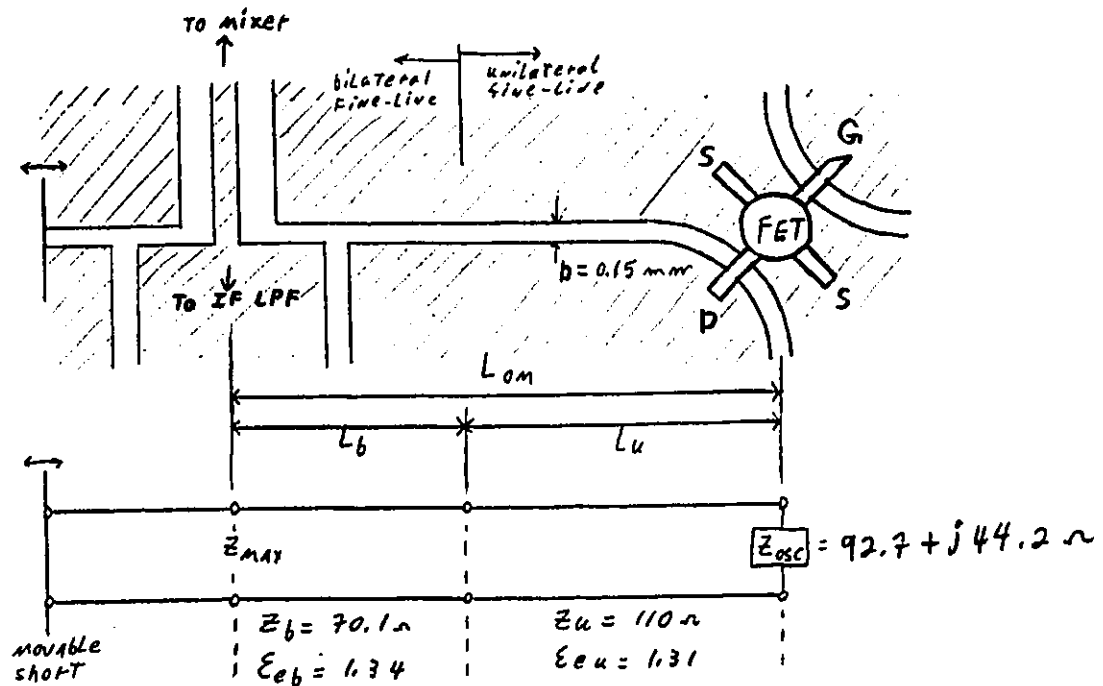


Figure 3.2.3.1 Oscillator-to-Mixer Interconnection Circuit

The input impedance of the oscillator has been measured separately using the HP8410 network analyzer at 23.5 and 24.5 GHz while the oscillator was turned on (generating signals at 24 GHz). The impedance at 24 GHz is assumed to be the average of the impedances measured at 23.5 and 24.5 GHz.

$$Z_{osc} = (92.7 + j44.2)\Omega$$

The transmission line between the DRO and the mixer consists of a combination of unilateral and bilateral fin-lines, where the bilateral fin-line is on the mixer side and the unilateral fin-line on the DRO side. The width of the fin-line gap (represented by D in Figure 3.2.3.1) is kept constant to 0.15 mm as defined in Section 3.2.2.3.

Given D , ϵ_r , and F_0 (0.15 mm, 2.22 mm, and 24 GHz respectively) the transmission line characteristics of the unilateral fin-line, Appendix A Section A-4 are:

$$\begin{aligned} Z_u &= 110 \Omega \\ \epsilon_{eu} &= 1.31. \end{aligned}$$

Similarly, for the bilateral fin-line, Appendix A, reference [10] gives:

$$\begin{aligned} Z_b &= 70.1 \Omega \\ \epsilon_{eb} &= 1.34. \end{aligned}$$

The length $L_u + L_b$ (unilateral + bilateral respectively) is selected to transform the impedance of the DRO $(92.7 + j44.2)\Omega$ into an open circuit (Z_{max}) at the mixer junction. By arbitrarily selecting L_b to be $0.5\lambda_{gb}$, L_u is determined as follows.

The normalized DRO input impedance is given by:

$$\begin{aligned}
z_{osc} &= Z_{osc}/Z_u \\
&= (92.7 + j44.2 \Omega)/110 \Omega \\
&= 0.84 + j0.40
\end{aligned}$$

which requires $0.363 \lambda_{gu}$ to present an open circuit. An additional $0.5 \lambda_{gu}$ is added for ease of realization.

Similarly, the λ_{gb} is restricted to $0.5 \lambda_{gb}$ for ease of realization. Therefore, using Appendix A;

$$\begin{aligned}
\text{Unilateral length } (L_u) &= 0.863 \lambda_{gu} \\
&0.863 \times 10.92 = 9.42 \text{ mm} \\
\text{Bilateral length } (L_b) &= 0.500 \lambda_{gb} \\
&= 0.500 \times 10.80 = 5.40 \text{ mm},
\end{aligned}$$

for a total spacing of:

$$L_u + L_b = 9.42 \text{ mm} + 5.40 \text{ mm} = 14.82 \text{ mm}.$$

The maximum impedance (Z_{max}) is given by;

$$Z_{max} = z_b \times Z_b$$

where z_b represents the unilateral to bilateral normalization at the unilateral to bilateral junction, and is given by

$$z_b = z_u \times Z_u/Z_b$$

where z_u at the unilateral to bilateral junction is 1.55,

$$z_b = 1.55 \times 110 \Omega / 70.1 \Omega = 2.43,$$

$$Z_{max} = 2.43 \times 70.1 \Omega = 170 \Omega.$$

3.3 Experimental Results for the Fin-Line Dielectric Resonator Oscillator

Figure 3.3.0.1 shows the equipment setup for the measurement of the oscillator performance.

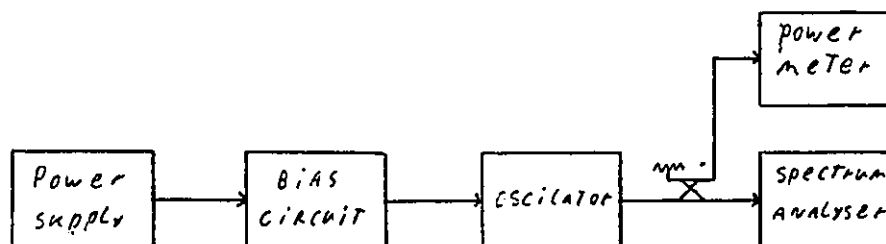


Figure 3.3.0.1 Oscillator Performance Measurement Block Diagram

All measurements have been performed with fixed short-circuit positions set for optimum output power at 24.0 GHz. Refer to Figure 3.2.1.1 for short-circuit designation.

$$\begin{aligned} L_{gs1} &= 14.8 \text{ mm} \\ L_{gs2} &= 16.2 \text{ mm} \\ L_{ds} &= 12.5 \text{ mm.} \end{aligned}$$

Note that for $\lambda_g = 10.92 \text{ mm}$,

$$L_{gs1} + L_{gs2} = 31.0 \text{ mm} \approx 3 \lambda_g$$

3.3.1 Output Power

Figure 3.3.1.1 and 3.3.2.1 show the frequency spectrum measured over a 5 MHz and a 500 KHz frequency span respectively. An output power of +13 dBm (20 mW) was achieved for the following bias conditions: $V_{ds} = 3.6 V_{dc}$ and $V_{gs} = -0.75 V_{dc}$.

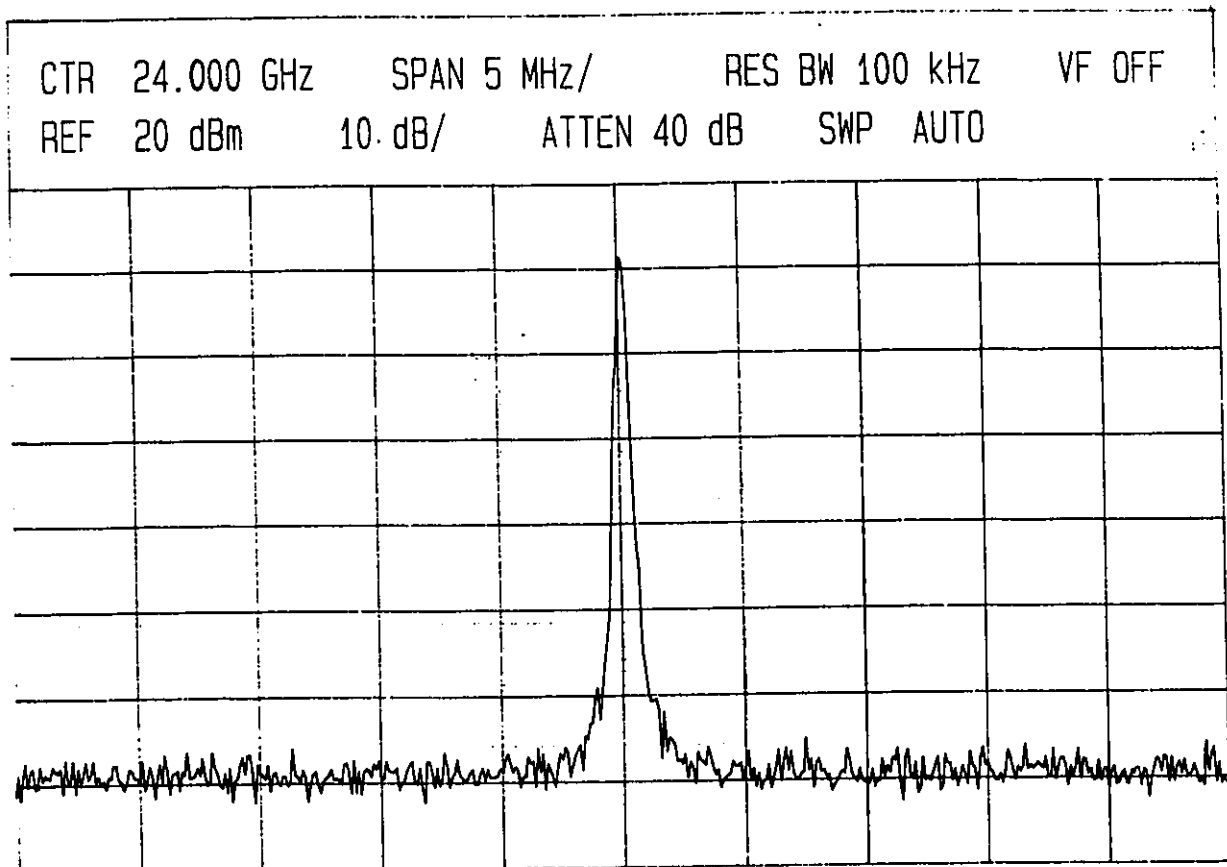


Figure 3.3.1.1 Oscillator Output Frequency Spectrum
(Resolution Bandwidth = 5 MHz)

3.3.2 Side-Band

Figure 3.3.2.1 shows spurious side-bands (at ± 1.1 MHz from the carrier) which are more than 50 dB down. The spurious response might be caused by the self resonance at 1.1 MHz of the DC bias circuitry.

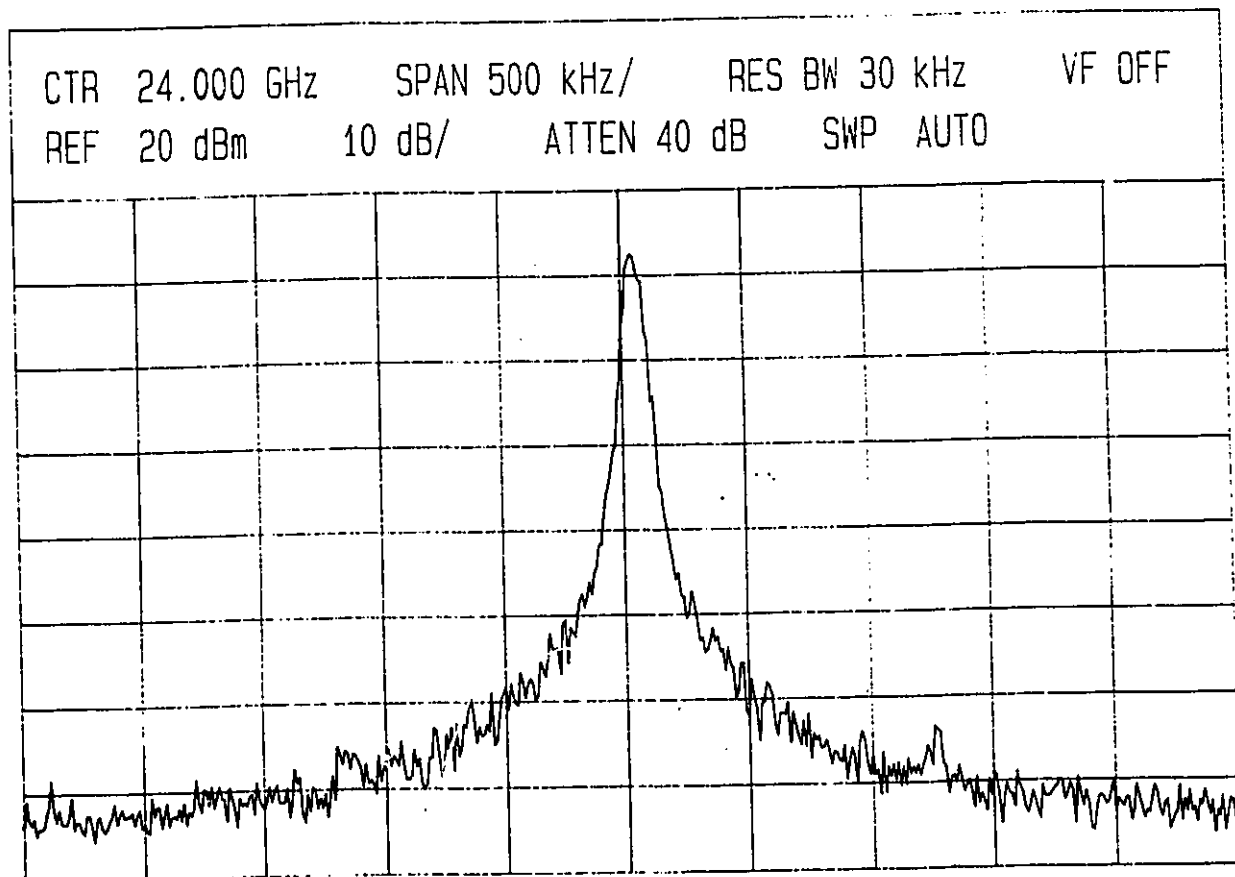


Figure 3.3.2.1 Oscillator Output Frequency Spectrum
(Resolution Bandwidth = 500 KHz)

3.3.3 Mechanical Frequency Tuning

The mechanical tuning of the resonant frequency is achieved by moving both short circuits, SC1 and SC2, represented in Figure 3.2.1.1. $\pm 1\%$ of tuning is obtained with less than 1 dB output power variation, whereas $\pm 5\%$ of tuning is possible with more than 5 dB output power variation.

3.3.4 Electrical Frequency Tuning

Electrical frequency tuning is achieved by adjusting the drain-to-source voltage (V_{ds}). $\pm 0.1\%$ of tuning is obtained with minor output power variations (less than 2 dB), whereas for more than $\pm 0.1\%$ of tuning, major output power degradation is observed.

3.3.5 Efficiency

An optimum efficiency of 23% is obtained at 24 GHz. Figure 3.3.5.1 shows the efficiency versus frequency for $\pm 1\%$ frequency deviation.

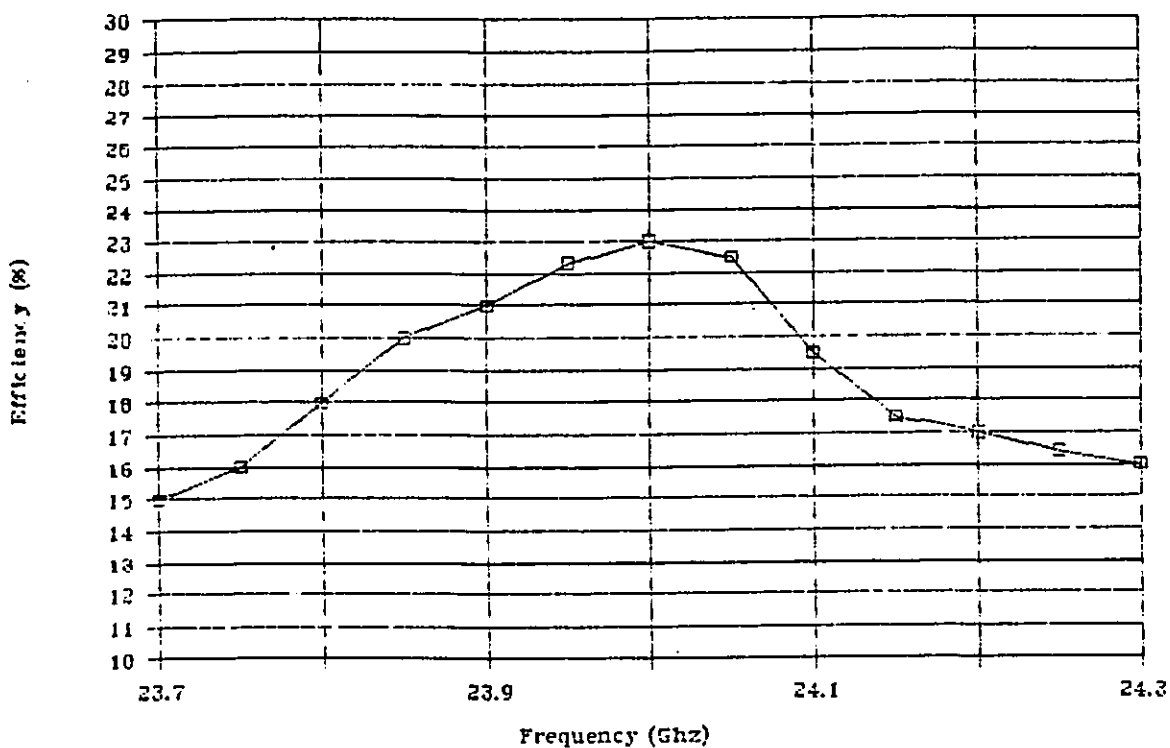


Figure 3.3.5.1 Oscillator Efficiency vs. Frequency

REFERENCES

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- [5] M. Dydyk, "Dielectric Resonators Add Q to MIC Filters," Microwaves, December, 1977.
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CHAPTER 4

RECEIVER FRONT END

The overall front end represents an integration of the mixer, the transmission-line, the filters and the oscillators, presented in the previous Chapters 3,4,5, and 6 respectively.

4.1 Overall Circuit Configuration

The planar circuit presented in Figure 4.1.1 has been produced by photolithographic means and suspended between two U-shaped waveguide channels. Starting from the RF input, the first component is a bandpass filter which rejects signals at the image frequency. The second component consists of a fin-line taper and a step transformer, followed by the balanced mixer and a traditional microstrip IF lowpass filter which rejects all spurious mixing products. The local oscillator is also realized in fin-line technology and coupled magnetically to the mixer. The coupling strength can be adjusted with a variable short.

Naturally, the integration of all components (including the biasing circuitry, not shown) on a single substrate reduces considerably the size of the front-end. The outside dimensions of the complete receiver are approximately 3 x 1.5 x 1 inches.

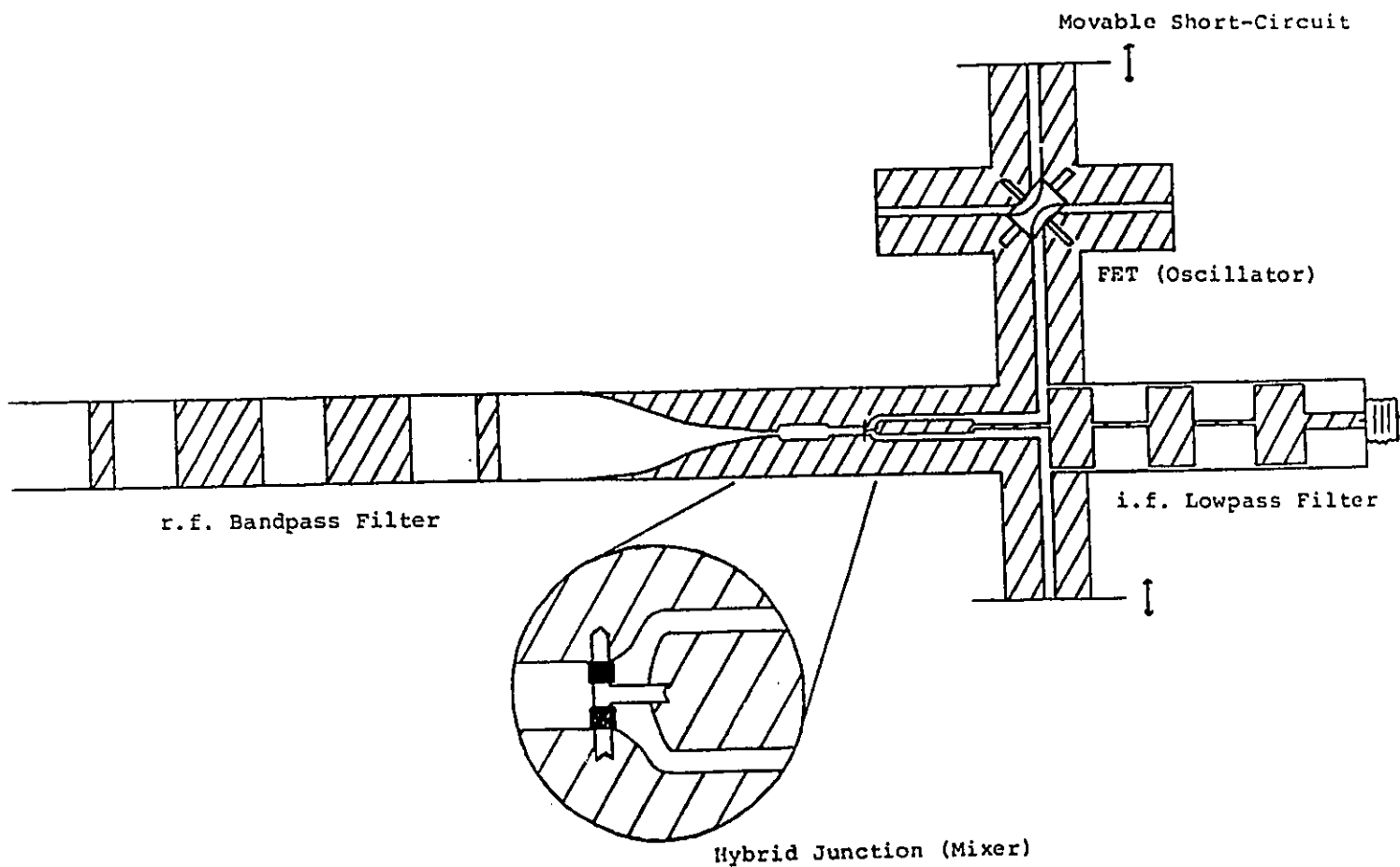


Figure 4.1.1 Integrated Receiver Front-End, the RF Input is on the Left Side. The Enlarged view Shows the Hybrid Junction and the Beam Lead Diode in a T-Configuration. Dielectric type 5880 from Roger was used with a $\epsilon_r = 2.22$.

4.2 Enclosure (Housing)

The housing incorporates a WR-42 waveguide RF input and a SMA-female coaxial¹ IF output. Figure 4.2.1 shows the assembled housing, whereas Figure 4.2.2 presents the same housing split in two pieces, so that the overall circuit can be seen.

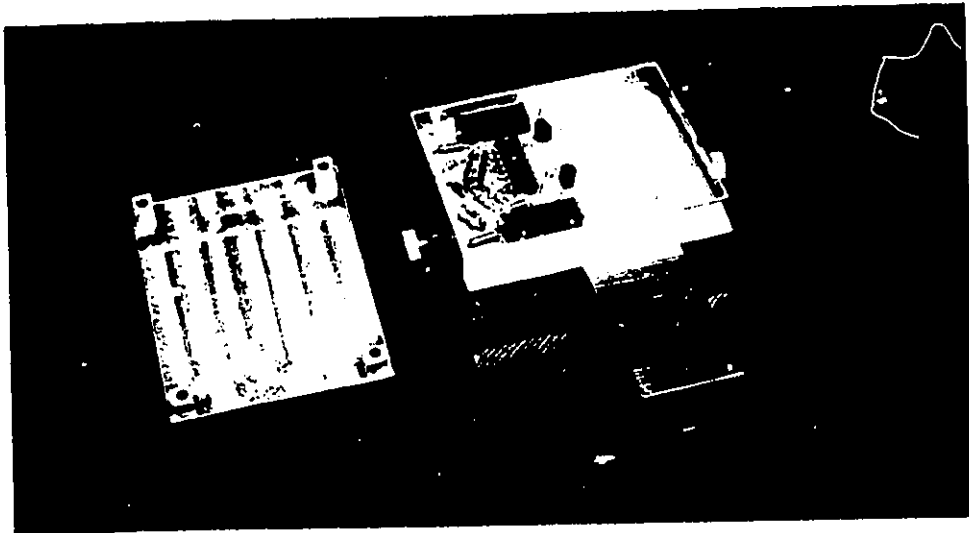


Figure 4.2.1 Picture of the Housing with Open Cover
Showing the DC Bias Circuitry

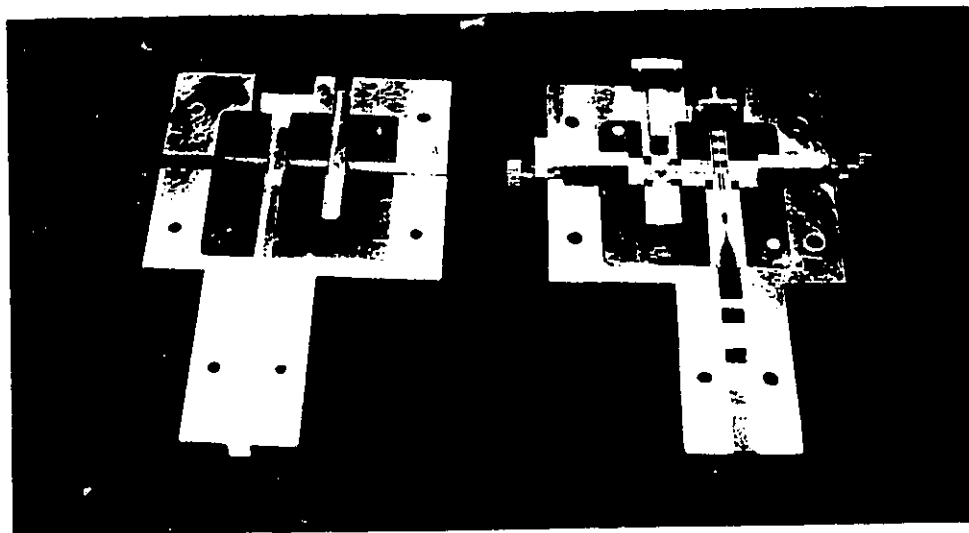


Figure 4.2.2 Picture of the Splitted Housing

The housing is made from aluminum, but ABS plastic with copper coating may be used for less weight. Microwave absorber is inserted into all resonant cavities to reduce the amount of RF leakage. The bias is applied directly to the fins via a low-pass-filter. The bias circuitry is enclosed in the housing for better noise performance. The mechanical dimensions of the housing are given in Figure 4.2.3

4.3 Measurement Results

The overall front-end performance is a function of all individual units described in the previous Chapters 1, 2, and 3. The measured performance of each individual unit was presented in the previous chapters. The overall front end performance at an RF of 24.4 GHz (with 500 MHz bandwidth) is summarized in Table 4.3.1.

Conversion Loss	6 dB
SSB Noise Figure	7 dB
RF Input VSWR	1:1.5
IF Output VSWR	1:1.9
RF-IF Isolation	40 dB
LO Radiation (@24 GHz)	-40 dBm
DC Power Supply	±15 Vdc

Table 4.3.1 Overall Front-End Performance

5.0 RECOMMENDATIONS

Some technical recommendations are presented and might be useful for any related future work.

- a) For better oscillator performance, the package type FET (NE67383) must be replaced by the chip type FET (NE67300).
- b) Measurements of temperature characteristics of the entire front-end should be done.

Before concluding, the author would like to propose some personal recommendations which might be useful to any student who attempts to do any similar research.

The recommendations are subdivided into four sections: study, design, construction, and test.

Study

Instead of trying to achieve state-of-the-art performance immediately, the student should get acquainted with standard industrial performance. Therefore, as a first step, the student must try to approach that performance level, and improve it later.

Design

If a model must be developed to simulate any effect, the student should make sure that all input parameters are known. As an example, the author had developed a non-linear analysis model to characterize the mixer. The model was unoperational since some critical input parameters (as diode ideality factor) were not available from the manufacturer, whereas their experimental evaluations required equipment that obviously was not accessible.

Construction

The important aspect related to the first construction of any device, is to predict and to allow enough flexibility in case some modifications are required. As an example, the housing was oversized to allow maximum flexibility if the circuit has to be modified.

Testing

Never believe the result obtained from equipment if you do not fully understand the utilization. Always read the utilization manual, and make simple measurements for which the expected results are known. Another point is to make sure that all the material required to perform the experimental measurement is available or at least accessible.

6.0 CONCLUSION

The realization and characteristics of an integrated receiver front end for 24.4 GHz has been discussed.

Naturally, since the circuit is realized in E-plane technology, the same configuration can be scaled down for higher frequencies, provided that appropriate FETs and mixer diodes are available. The mixer configuration has been shown to be viable at frequencies up to 140 GHz [1]. It is evident that fabrication tolerances are very critical at these frequencies, requiring excellent photolithographic facilities.

- [1] W. Menzel, "A 140 GHz Balanced Receiver for Fin-Line Integrated Circuits", 13th EuMC Nuernberg, West Germany, 1983, pp. 179-182

APPENDIX A

REVIEW OF FIN-LINE PROPERTIES

A.1 History

Although standard microstrip techniques may be applied to millimeter-wave circuits by scaling down the linear dimensions, several problems arise. Firstly, conductor and dielectric losses increase quickly with frequency, due to reduced size. If the size of elements is not reduced, the danger of multi-moding and cross-talk increases dramatically. Excessive miniaturization also leads to tolerance problems and manufacturing difficulties. This has led Meier [1] to propose the integrated "fin-line" as a new transmission line for millimetre-wave integrated circuits. Before Meier's proposal, fin-line was used only as a broadband coupling device [2]. In this basic form (Appendix A, Section A-1) the fins were not isolated from the waveguide housing. Therefore, active devices such as diodes (required for mixing or switching) and FET's (used as oscillators or amplifiers) could not be accommodated because of biasing problems.

With increasing activities in the millimetre-wave field, more attention has been paid to the fin-line technique, as an alternative to the traditional waveguide. (See Appendix A, Section A-2). Since the early 1970's, the fin-line medium has been combined with other planar media such as microstrip and coplanar line to form versatile integrated circuits mounted in the E-plane of a waveguide housing, hence the generic name of "E-plane circuits" for this family of transmission lines. A good example of the integration of multiple planar media is the balanced mixer which contains fin-line, coplanar line and microstrip, and which has been described in this thesis.

During the 1980's several civil and military systems in the U.S. and Europe [3-8] have been implemented with key components employing printed E-Plane techniques.

With the growing interest from the component designers' side, there has been an increasing activity concerning the theoretical foundations of the fin-line. The determination of the propagation coefficient (guided wavelength), field distribution and characteristic impedance is critical for device matching and circuit analysis. Calculation of discontinuities is required for accurate filter design. This new momentum will eventually result in very sophisticated computer aided design (CAD) of printed E-plane circuits.

A.2 Basic Configurations

E-plane integrated circuits must usually be compatible with standard rectangular waveguides. Thus, the integrated circuits will consist of a metal waveguide housing and a planar circuit. In addition to fin-line, several other printed transmission lines may be used as planar inserts. Figure A.2.1 (from left to right) shows three types of fin-lines (unilateral, bilateral, and antipodal fin-line), as well as dielectric loaded waveguide, ridge waveguide, coplanar line, shielded microstrip, and suspended stripline. All planar circuits are suspended in the E-plane of a waveguide housing.

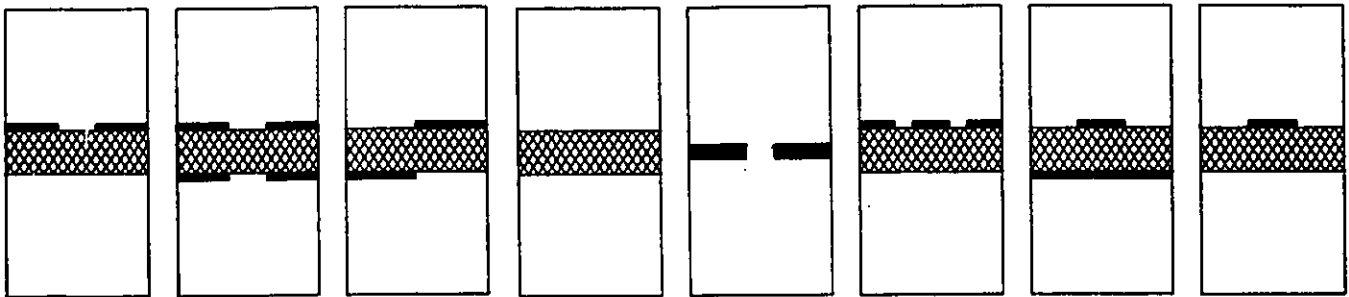


Figure A.2.1 Cross-Section of Various Transmission Lines Suspended in the E-Plane of a Waveguide

A.3 Advantages/Disadvantages of Fin-lines

Among the various structures presented in Fig. A.2.1, the unilateral and bilateral fin-line are of particular interest in this study. They are ideal transmission lines for

millimeter-wave circuits, because they avoid miniaturization and offer the potential for low-cost production through batch processing techniques. For aircraft applications, the use of ABS material for the waveguide with gold deposition on the inner walls makes the fin-line small and light weight. Fin-line is basically a shielded slotline where, in contrast to the conventional slotline, the concentration of the field is achieved by using a metal shield represented by the waveguide housing, rather than by using a thick high permittivity substrate, hence, lower substrate and radiation loss. If the dielectric is completely removed in a unilateral (or bilateral) fin-line circuit, the resulting waveguide is the well-understood ridge waveguide with very thin ridges. Both transmission lines have very similar characteristics. In summary, fin-line has a wide bandwidth for single mode operation (refer to Figure A.3.1), moderate attenuation and low dispersion in the frequency range of interest. These properties give it several advantages over microstrip in millimeter-wave applications.

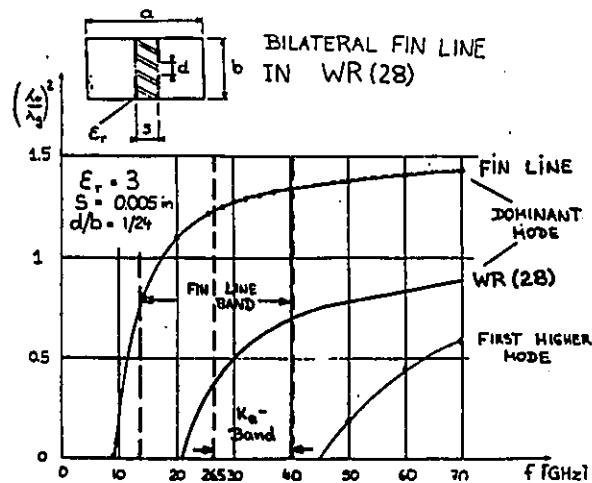


Figure A.3.1 Comparison of dispersion characteristics of waveguide and bilateral fin-line for K_a -band. Note the increased operating bandwidth of the dominant mode in the fin-line.

However, in spite of all the advantages of this novel transmission line, the basic problem faced by the designers is the lack of design information. Consequently, there remains a strong need for accurate and fast methods to compute the equivalent dielectric constant (or guided wavelength) and the characteristic impedance of fin-line. These methods will have to be accurate within $\pm 1\%$ over a wide range of fin-line physical dimensions and substrate dielectric constants. In order to design the front-end described in this thesis, empirical expressions for fin-line characteristics by Pramanick and Bhartia [9] have been used. For convenience, they are repeated in the following section A.4.

A.4 Unilateral Fin-line Empirical Expression

Previous expressions have been developed by Hoefer [10], [11]. These expressions were limited in range for b/d and s/a and restrict b/a to be $1/2$. Pramanick and Bhartia [9] have improved these expressions to accept a wider range of variables. The expressions are accurate within $\pm 0.6\%$. They have been derived using a stationary formula, and curve fitted to numerical results obtained with the spectral domain method. The expressions are valid over the range.

$$\begin{aligned} 0 &\leq b/a \leq 1 \\ 1/32 &\leq b/d \leq 3/4 \\ 1/32 &\leq s/a \leq 1/4 \\ 1 &\leq \epsilon_r \leq 3 \end{aligned}$$

where a, b, d , and s are shown in Figure A.4.1.

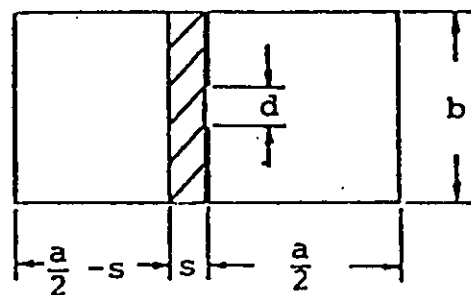


Figure A.4.1 Unilateral Fin-Line Parameters

Expressions For Guided-Wavelength

The guided wavelength in any wave guiding medium (such as fin-line) is defined as:

$$\lambda_g = \lambda / \sqrt{\epsilon_e} \quad (\text{A.1})$$

$$\text{where } \epsilon_e = k_e - (\lambda / \lambda_{ca})^2 \quad (\text{A.2})$$

$$\text{where [1], } \lambda_{ca} = 2a \left[1 + \frac{4b}{\pi d} (1 + 0.2 \sqrt{\frac{b}{a}}) \right] \text{Ln cosec} \left(\frac{\pi d}{2b} \right)^{1/2} \quad (\text{A.3})$$

$$\text{where } k_e = k_c F1$$

$$\text{where } k_c = (\lambda_{cf} / \lambda_{cd})^2$$

$$\text{where } \lambda_{cd} = 2a [1 + F2 (\epsilon_e - 1)]^{1/2}$$

$$\lambda_{cf} = \lambda_{ca} [1 + F2 (\epsilon_e - 1)]^{1/2}$$

$$\text{where [1], } F2 = \left[a1 \text{Ln cosec} \left(\frac{\pi d}{2b} \right) + b1 \right] \frac{s}{a}$$

$$\text{where } a1 = 0.4020974 [\text{Ln} [a/s]]^2$$

$$- 0.7684487 \text{Ln}[a/s]$$

$$+ 0.3932021$$

$$b1 = - 0.2234601 [\text{Ln} [a/s]]^2$$

$$+ 1.4549698 \text{Ln}[a/s]$$

$$+ 0.1050848$$

$$\text{where [2] for } \epsilon_r = 2.22$$

$$F1 = [1.0 + 0.43 (s/a) (d/b)]^p, \text{ for } 1/32 \leq s/a \leq 1/8$$

$$= [1.02 + 0.264 (s/a) (d/b)]^p, \text{ for } 1/8 \leq s/a \leq 1/4$$

$$\text{where } p = 0.096 [s/a] - 0.007$$

Expressions For Characteristic Impedance

The characteristic impedance in any wave guiding medium (such as fin-line) is defined as:

$$Z_0 = Z_{0\infty} / \sqrt{\epsilon_e} \quad (\text{A.4})$$

where ϵ_e is given by (A.2)

$$Z_{0\infty} = \frac{120\pi^2(b/\lambda_{ca})}{\frac{b}{d} \sin \frac{\pi s}{\lambda_{ca}} + \left[\frac{B_0}{Y_0} + \tan \frac{\pi(a-s)}{2\lambda_{ca}} \right] \cos \frac{\pi s}{\lambda_{ca}}}$$

$$\text{where [1],} \quad \frac{B_0}{Y_0} = \frac{2b}{\lambda_{ca}} \left\{ \ln \csc \left(\frac{\pi d}{2b} \right) \right\}$$

where λ_{ca} is given by (A.3)

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APPENDIX B

REVIEW OF RECEIVERS

The following list represents several types of receivers which are available for the detection and processing of microwave signals.

- crystal video receiver
- superheterodyne receiver
- IFM receiver
- microscan compressive receiver
- channelized receiver
- acousto-optical receiver.

B.1 Crystal Video Receiver

The typical wideband crystal video configuration consists of a RF bandpass filter with an optional low-noise RF amplifier, a square-law video detector and a log video amplifier as shown in Figure B.1.1.

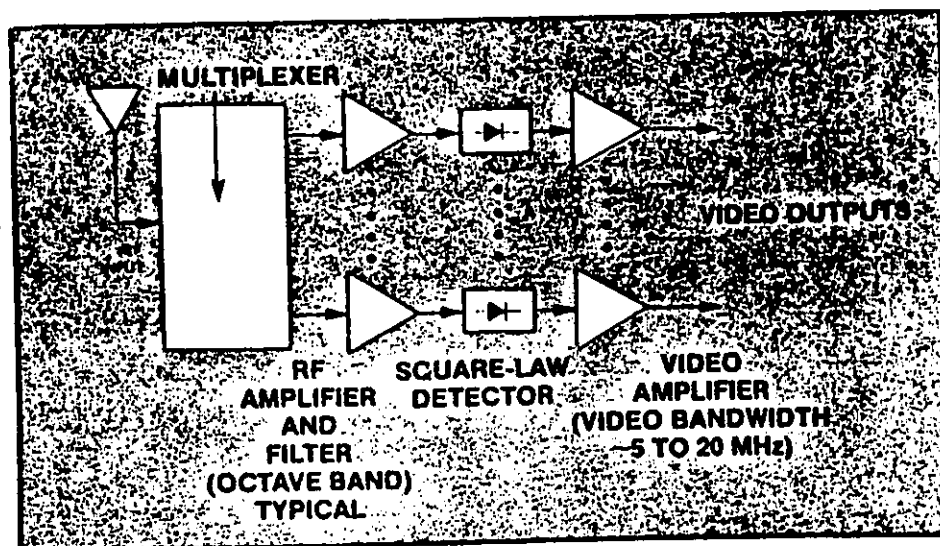


Figure B.1.1 Wideband Crystal Video Receiver [1]

B.2 Superheterodyne Receiver

Superhet receivers typically consist of a RF bandpass filter followed by a frequency conversion to IF as shown in Figure B.2.1. A narrowband IF amplifier and filter determines the processing bandwidth and is followed by any of a variety of detectors.

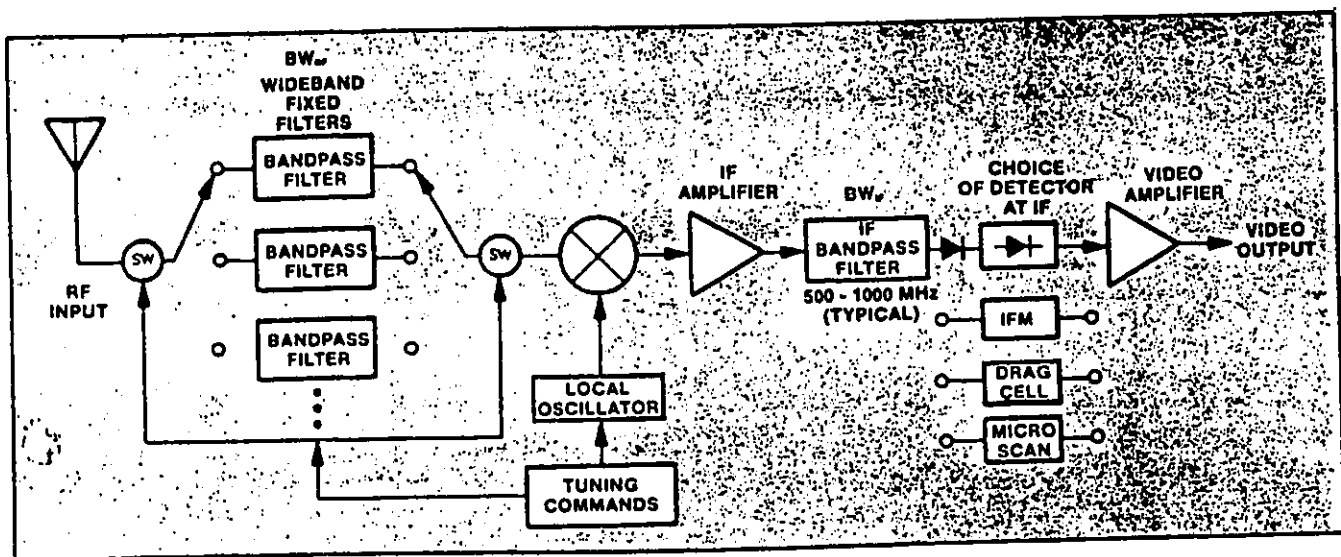


Figure B.2.1 Wideband Superheterodyne Receiver [1]

B.3 Instantaneous Frequency Measurement (IFM) Receiver

The IFM receiver [1] uses a delay line frequency discriminator to provide an instantaneous indication of signal frequency. In this receiver, a second signal path is used to generate a signal whose phase is offset by an amount proportional to the incident signal frequency. The signal frequency is then obtained from the output of a phase detector as shown in Figure B.3.1.

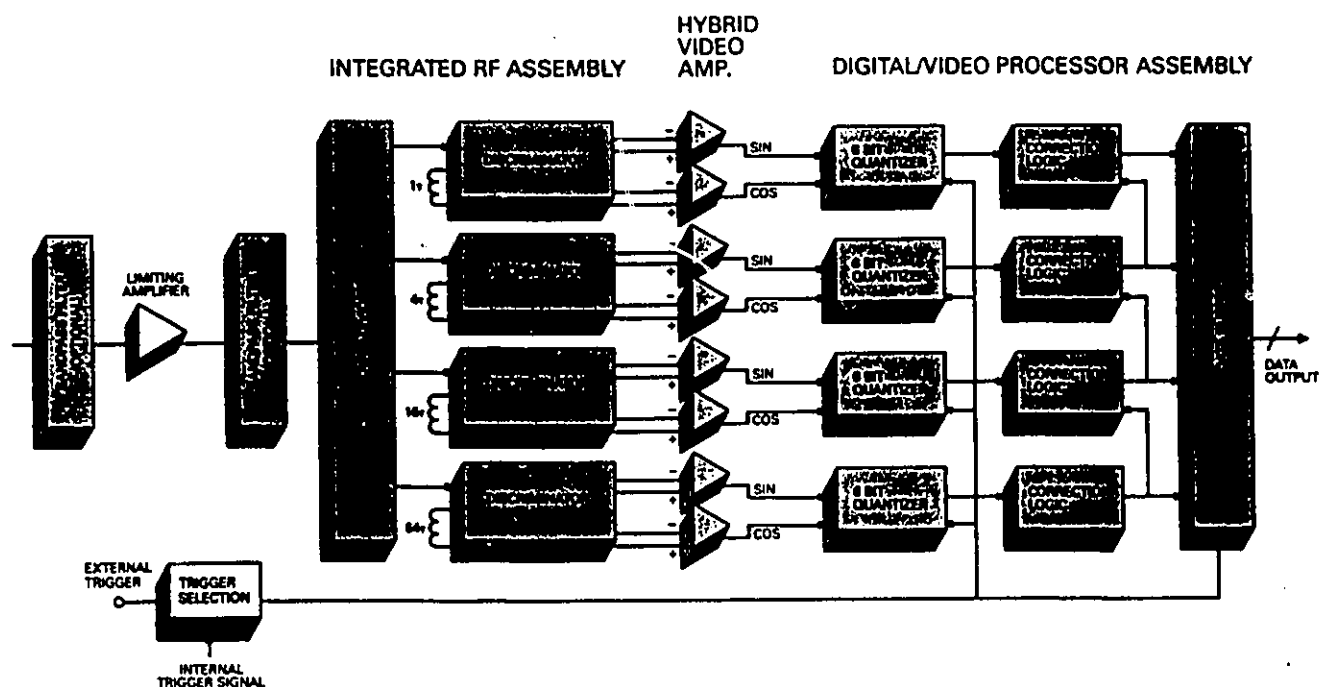


Figure B.3.1 Instantaneous Frequency Measurement Receiver [1]

B.4 Microscan Compressive Receiver

The microscan receiver shown in Figure B.4.1 is a type of superheterodyne receiver which consists of an RF bandpass filter followed by a swept local oscillator conversion. The IF is passed through a dispersive filter which maps the IF frequency into a time delay. This changing frequency signal then encounters a frequency-dependent delay which is arranged so that all the energy emerges from the filter output at one time, producing a sharp pulse whose onset time (measurement from the LO sweep start time) is proportional to the RF frequency.

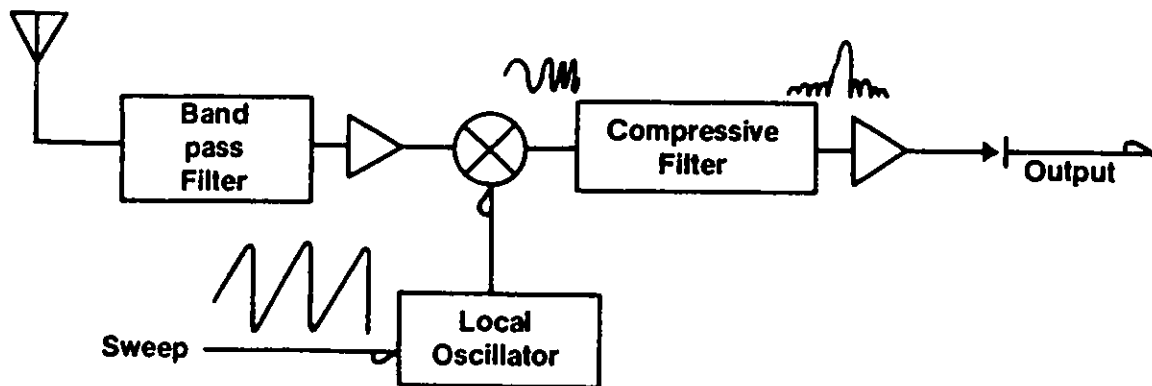


Figure B.4.1 Microscan Compressive Receiver [2]

B.5 Channelized Receiver

A channelized receiver [2] subdivides the RF spectrum into segments and simultaneously downconverts each channel to a common IF. The RF breakdown is accomplished by a series of contiguous filters, each mixed with a fixed frequency local oscillator. Each channel becomes, in effect, a fixed tuned superhet receiver. The basic architecture is shown in Figure B.5.1.

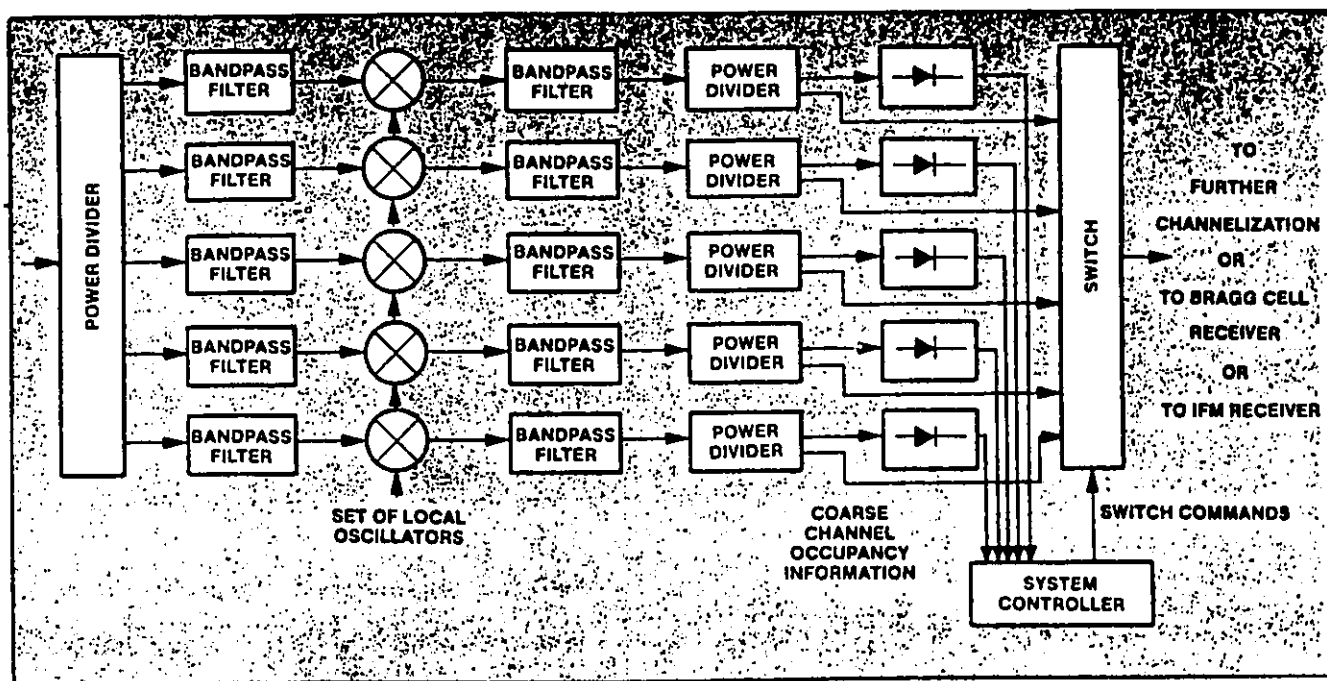


Figure B.5.1 Channelized Receiver [2]

B.6 Acousto-Optical Receiver

Acousto-optical receivers [3] make use of the interaction between a sound wave and light propagating through a crystal to provide spectral analysis of the signal. A laser beam is shone through the crystal, the light beam is scattered through an angle proportional to the ratio of the wavelength of light to sound. The sound wave is supplied by the RF stimulating the crystal through a piezoelectric transducer. The scattered angle corresponds to constructive interference in the reflection of light from successive wavefronts of the sonic wave. The acoustic wave affects the index of refraction such that the scattering angle is modulated by the RF. After passing through the crystal, the light is collected by a photo-sensitive array and read out electrically as represented in Figure B.6.1. B represents the deviation for a received signal at F_2 . A shows the deviation when no signals are present between F_1 and F_2 .

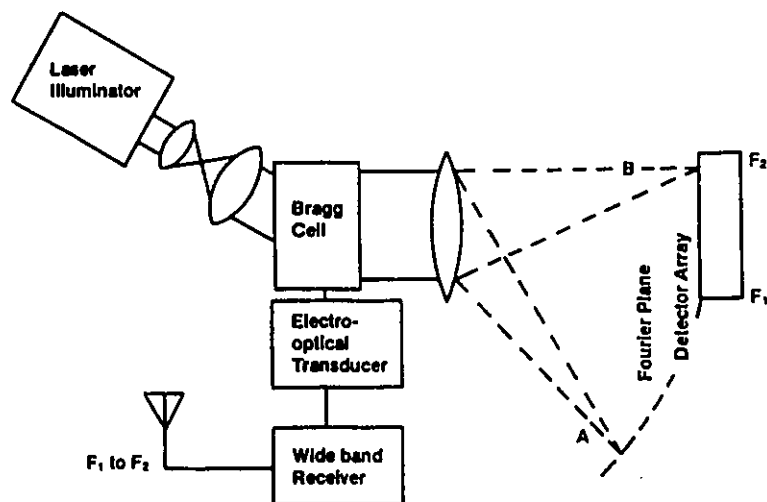


Figure B.6.1 Acousto-Optical Processor (Bragg Cell).
[3]

B.7 Receiver Type Comparison

Generally, a careful analysis must be carried out to determine which receiver is best suited for a given application. A summary of receiver types and their main characteristics is given in Table B.7.1. This table underlines the cost-effectiveness of the heterodyne receiver as compared with all other types.

spec. type	sensi- tivity	probability of intercept	resolution selectivity	reliabi- lity	physical	price
Crystal Video	poor	poor	fair	poor	small	cheap
Superhet	excellent	fair	excellent	excellent	moderate	Low
IFM	fair	good	good	good	moderate	moder- ate
Microscan	very good	excellent	good	poor	complex	expen- sive
Chanel- lized	excellent	excellent	excellent	excellent	complex	high
Acousto- optic	good	excellent	excellent	poor	complex	expen- sive

Table B.7.1 Type of Receiver

This comparison clearly demonstrates the advantage of the superheterodyne receiver for a majority of applications and illustrates the rationale for developing such a receiver for millimeter-wave communication and radar systems.

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APPENDIX C

REVIEW OF MIXERS IN GENERAL

C.1 Parameters for evaluation of Mixer Performance

There are many parameters which describe the performance of a mixer. A description of the most important of these follows.

Single Sideband Conversion Loss (SSB Cl)

Since a mixer converts power from one frequency to another, perhaps the most fundamental parameter is the measure of how efficiently the frequency is converted. This parameter is called the conversion loss, and is defined as the difference in dB between the received input RF power and the output IF power of the desired IF sideband. Both the up- and down-converted product, or sidebands, exit the IF-port. Since usually only one of these products is filtered, there will be an automatic 3 dB SSB (single sideband) conversion loss minimum. Conversion loss is a strong function of LO power, which affects mismatch between the system and mixer radically.

More power losses occur during frequency conversion because some of the down-converted power is also lost in the form of unwanted higher-order mixing products, heat due to the series resistance of the diodes, and mismatches at the mixer RF and IF port.

Mismatch loss (L1): This loss is expressed as a function of VSWR. S11 for RF and S22 for IF-port.

$$L1 \text{ (dB)} = 10 \left[\log \frac{(S11 + 1)^2}{4 S11} + \log \frac{(S22 + 1)^2}{4 S22} \right] \quad (C.1.1)$$

Parasitic loss (L₂): This loss is due to the parameters R_s and C_j. It reflects the ratio of incident RF power to the power effectively delivered at the diode junction.

$$L_2 \text{ (dB)} = 10 \log \left[1 + \frac{R_s}{R_j} + (C_j \omega)^2 R_j R_s \right] \quad (\text{C.1.2})$$

where R_j and C_j are obtained from (C.3.3) and (C.3.4) respectively. As represented in (C.3.3), R_j is mainly dependent on the detected current, consequently the function of L₀ power. L₂ is minimum for R_j = 1/ωC_j.

Intrinsic junction loss (L₃): This loss depends on the terminating conditions at the image frequency and nonlinearity of the diode conductance (g) which is related to the L₀ power.

$$L_3 \text{ (dB)} = 10 \log \left[\left[1 + \left(\frac{1 + \frac{g_2}{g_0} - 2 \left(\frac{g_1}{g_0} \right)^2}{1 + \frac{g_2}{g_0}} \right)^{1/2} \right]^2 \left[1 + \left(\frac{g_2}{g_0} \right) \right] \left(\frac{g_0}{g_1} \right)^2 \right] \quad (\text{C.1.3})$$

where, g₀, g₁, g₂ are the Fourier coefficients of the diode conductance, obtained by the non-linear analysis developed by Hicks [1], described in Section 1.1.4.

The optimum value for L₃ is in the order of 2 to 3 dB and is independent of frequency.

To resume, the conversion loss is expressed as:

$$\text{conversion loss (dB)} = L_1(\text{dB}) + L_2(\text{dB}) + L_3(\text{dB})$$

Voltage Standing Wave Ratio (VSWR)

VSWR is the measure of mismatch offered to the system by the mixer. It is usually specified over a given bandwidth as a function of LO power and temperature. It is calculated as follows:

$$\text{VSWR} = \frac{1 + |p|}{1 - |p|} \quad (\text{C.1.4})$$

where,

$$p = \frac{Z_L - Z_0}{Z_L + Z_0}$$

in which p is the reflection coefficient

Z_L is the mixer input impedance

Z_0 is the system characteristic impedance.

R, L, and I VSWR's are directly dependent on LO power, which defines the operating point of the diodes. Changing the LO power modifies the diode operating point resulting in a different impedance for all mixer ports, affecting the VSWR of all three ports.

When designing a mixer, the L-VSWR (VSWR at L-port) is first optimized by adjusting the L-port circuit to make sure that the LO power biases the diodes properly. Then, the RF- and IF- port circuit are adjusted to properly match the diodes to the RF input and IF output loads.

Isolation

Interport isolation is the measure of insertion loss between any two mixer ports. It is measured in dB and is usually specified over a given bandwidth as a function of LO power and temperature. Maximizing isolation between ports in a mixer is necessary because unwanted signals from RF- and LO-port feed through and can mask the desired IF output, as well as causing electromagnetic interference and coupling. Various factors such as diode match and circuit balance influence isolation in a mixer.

LO-to-IF isolation is more important to specify than RF-to-IF isolation since LO power is almost always 20 dB higher than RF power. RF-to-IF isolation is specified only when the relative power level of RF feedthrough and IF output power is critical. It is also required for mixers having broadband IF output, thus allowing the frequency of the RF feedthrough power to fall in the IF band. In this case no filtering can be used. When the RF-port has no buffer (an amplifier, for example) between the mixer and the receiving antenna, the LO feedthrough power can be radiated by the receiving antenna and cause feedback or interference in the system. Hence, LO-to-RF and LO-to-IF isolations are the most important. Generally, an isolation of 30 dB is typical.

Dynamic Range (DR)

Dynamic range is measured in dB and is the input RF power range over which the mixer is useful. The lower limit of the dynamic range is the noise floor, which depends on the mixer and system. The upper limit of the dynamic range is generally taken to be at the 1 dB compression point of the mixer. It is measured in dBm, and is the input RF power level at which conversion loss increases by 1 dB.

The 1 dB compression point is generally taken to be the top

of the dynamic range because the input RF power, which is not converted into desired IF output power, is instead converted into heat and higher-order intermodulation products. The intermodulation products that begin to appear when RF power is increased beyond the 1 dB compression point can begin to obscure the desired IF output. Generally, the 1 dB compression point is 5 to 10 dB lower than the LO input power.

Table C.1.1 shows the LO power levels generally associated with very high, high, medium and low level mixers. These power levels apply specifically to mixers using Schottky barrier diodes, but can also be applied in a more general way to mixers using other devices. The type and number of Schottky barrier diodes and resistor elements that may be used determine the level of LO input power.

Level	LO Power Range (dBm)
Very High	+27 to +20
High	+20 to +13
Medium	+13 to +10
Low	+10 to +6

Table C.1.1: Mixer LO Power Levels

Intermodulation Products (IM)

Intermodulation (IM) products are undesirable mixer-generated output products exiting the mixer from any port. Two types exist: single-tone and multiple-tone. Single-tone intermodulation products are composed of a single input RF signal mixing with the LO, and have the following frequencies:

$$f = \pm m f_R \pm n f_L \quad (C.1.5)$$

where,

$$m = 1, 2, 3, \dots$$

$$n = 1, 2, 3, \dots$$

Multi-tone intermodulation products are composed of two or more input RF signals mixing with the LO, and have the following frequencies:

$$f = (\pm m_1 f_{R1} \pm m_2 f_{R2} \pm m_3 f_{R3} \dots) \pm n f_L \quad (C.1.6)$$

where,

$$m_1, m_2, m_3, \dots = 0, 1, 2, 3, \dots$$

$$n = 0, 1, 2, 3, \dots$$

Multiple-tone intermodulation products for which all but one of the coefficients, m , are zero, resemble single-tone intermodulation products because their frequencies contain harmonics of the LO and harmonics of the one RF input signal that has the non-zero coefficient, m . Hence, single-tone intermodulation products can be present when multiple-input RF signals are incident at the RF-port, because output products can be generated that have frequencies given by equation (C.1.6). The level of output power of individual intermodulation products is very much affected by input LO and RF power levels and frequencies.

Charts exist that show trends in intermodulation suppression as a function of input power and frequency.

Notice that the even-by-even intermodulation signals, for which both m and n are even, are suppressed more than the odd-by-odd products. This is due to the circuit balance in double-balanced mixers. If diode match and mixer balance were perfect, only the odd-by-odd products would exit the IF-port, and other products would show infinite suppression.

Intercept Point (IIP,OIP)

The intercept point, measured in dBm, is a figure of merit for intermodulation product suppression. A high intercept point is desirable. Two types are commonly specified: input and output intercept point (IIP and OIP, respectively). The input intercept point is the level of input RF power at which the output power levels of the undesired intermodulation products and IF products would be equal; that is, intercept each other if the mixer did not compress. This output power level is the output intercept point, and equals the input intercept point minus conversion loss. As input RF power increases, the mixer compresses before the power level of the intermodulation products can increase to equal the IF output power. So, input and output intercept points are theoretical and are calculated by extrapolating the output power of the intermodulation and IF products past the 1-dB compression point until they equal each other. A high intercept point is desirable because it means the mixer can handle more input RF power before causing undesired products to rival the desired IF output product, and essentially means the mixer has a greater dynamic range. Dynamic range, 1-dB compression point, and intercept point are all interrelated.

The concept of intercept point can be applied to any intermodulation product; however, it normally refers to two-tone, third-order intermodulation products. If two input RF signals are incident at the mixer RF-port, they cause the mixer to generate the following two-tone intermodulation products:

$$(\pm m_1 f_{R1} \pm m_2 f_{R2}) \pm n f_L \quad (C.1.7)$$

where, $m_1, m_2, n = 0, 1, 2, 3, \dots$, m and n are integers and can assume any value.

Two-tone, third-order intermodulation products have the following frequencies:

$$(+2f_{R1} \pm f_{R2}) \pm f_L \quad (C.1.8a)$$

and

$$(\pm f_{R1} \pm 2f_{R2}) \pm f_L \quad (C.1.8b)$$

They are called third-order products because the coefficients of f_{R1} and f_{R2} sum up to 3.

The order of the intermodulation product is important because a 1 dB change in the power level of each input RF signal causes the power level of each intermodulation product to change by an amount of dB equal to its order.

The input intercept point is normally associated with two-tone, third-order intermodulation products because the third-order product is closest in frequency to the desired IF output product of any two-tone intermodulation product. The even-order, two-tone intermodulation products that exit from double and single-balanced mixers are suppressed far more than the odd-order products, due to mixer balance. Odd-order intermodulation products containing even-order LO harmonics are suppressed in double, but not in single-balanced mixers.

Both the suppression (in dBc) and the input RF power levels must be specified because intermodulation suppression varies as a function of input RF power.

The intercept point is normally presented as shown in Figure C.1.1. Input power is plotted along the horizontal axis, and output power is plotted along the vertical axis. Two lines are plotted: one relating IF output power to RF input power, and another relating intermodulation output power to RF input power. The point at which they intersect gives the input and output intercept points for the mixer at a particular set of input

frequencies for a given LO power level and temperature.

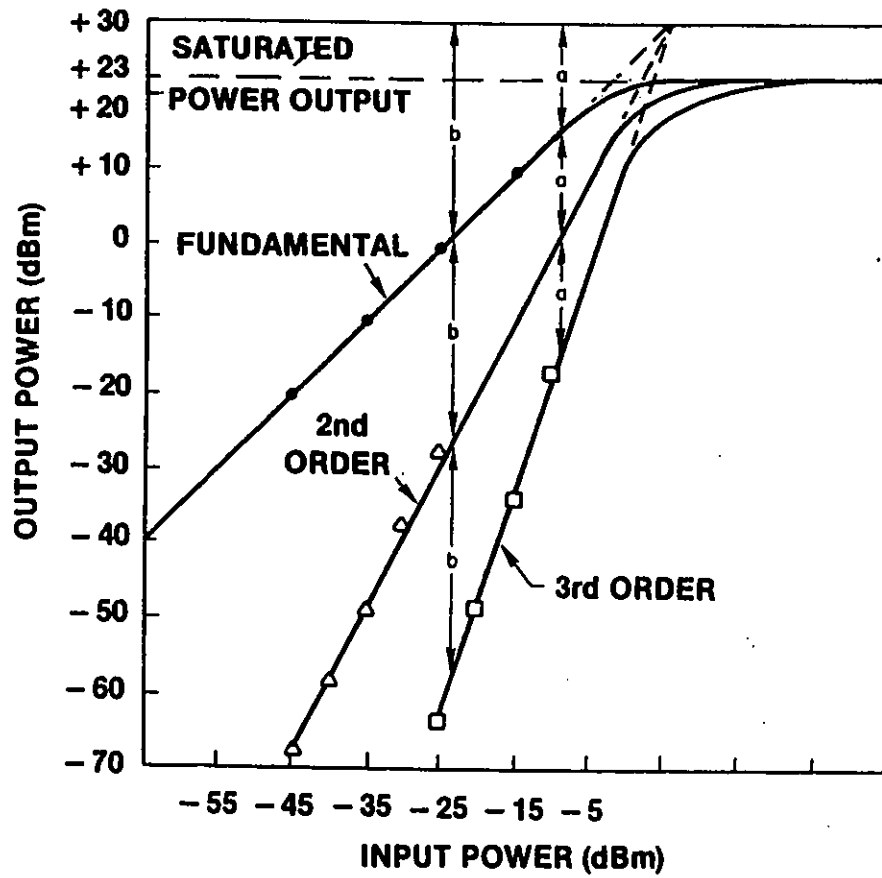


Figure C.1.1 Intercept Point Representation

A simple formula exists for calculating the input intercept point, given the level of intermodulation suppression, the order of the intermodulation, and the input RF power levels giving rise to this level of suppression.

$$\text{IIP} = \frac{\text{Intermodulation Suppression (dBc)}}{(\text{order} - 1)} + \text{input RF power (dBm)} \quad (\text{C.1.9})$$

Also, output and input intercept are related by the mixer conversion loss or gain (for active mixers):

$$\text{OIP (dBm)} = \text{IIP (dBm)} \left[\begin{array}{l} - \text{mixer conversion loss (dB)} \\ \text{or} \\ + \text{mixer conversion gain (dB)} \end{array} \right] \quad (\text{C.1.10})$$

Single Sideband Noise Figure (SSB NF)

Mixer SSB noise figure is measured in dB, and is the amount of noise added by the mixer to the converted signal plus the SSB conversion loss. Noise figure is the difference in dB between the input RF signal-to-noise ratio and the output IF signal-to-noise ratio.

$$\text{SSB NF (dB)} = 10 \log \frac{\text{RF PWR In}}{\text{Noise PWR In}} - 10 \log \frac{\text{IF PWR Out}}{\text{Noise PWR Out}} \quad (\text{C.1.11})$$

$$\begin{aligned} &= [\text{RF Power In (dBm)} - \text{IF Power Out (dBm)}] \\ &+ [\text{Noise Power Out (dBm)} - \text{Noise Power In (dBm)}] \end{aligned}$$

$$\begin{aligned} \text{SSB NF (dB)} &= [\text{SSB Conversion Loss (dB)} \\ &+ [\text{Output-to-Input Noise Ratio (dB)}] \quad (\text{C.1.12}) \end{aligned}$$

The output-to-input noise ratio of the mixer diode is defined as the ratio of available output noise power from the mixer to that of a resistor equal to the output resistance of the mixer, at room temperature. The contributions to the diode noise temperature ratio are: thermal, shot and flicker noise.

Thermal noise: The thermal noise of the series resistance R_s is due to the random motion of electrons in the resistance elements, and is given by the mean square current fluctuations as:

$$\overline{i_t^2} = 4 \frac{kTB}{R} \quad (\text{C.1.13a})$$

where B is the IF output bandwidth.

Shot noise: The shot noise of the rectifying barrier manifests itself as random fluctuations in the mean electron current. The mean square fluctuation is shown by the theory to be:

$$\overline{i_s^2} = 2 qIB \quad (C.1.13b)$$

where I is the mean current given by $(I + 2 I_s)$ for a metal-semiconductor device. I_s is the saturation current.

Flicker noise: The flicker noise (or $1/F$ noise) manifests itself as noise in excess of shot or thermal noise at low frequencies of the form:

$$\overline{i_f^2} = \frac{I^m}{f^n} \quad (C.1.13c)$$

where m and n are about 2 and 1 respectively.

At intermediate frequencies in the MHz range, the major contribution to the diode noise temperature ratio is usually the shot noise, but at lower frequencies the flicker noise predominates.

For minimizing the diode noise temperature ratio, the desirable features are: a large junction area, low reverse current leakage and low diode rectified current (or low LO power). Mainly depending on diode construction and LO drive, the diode noise temperature ratio is usually between 0.5 to 1.5 dB.

Like SSB conversion loss, the SSB noise figure is normally specified instead of the DSB (Double Sideband) noise figure, because the mixing process produces both up and down-converted IF products, and normally only one of these production is desired, so the other product is discarded. This causes half the input power

to be lost, making the SSB noise figure 3 dB higher than the DSB noise figure.

The single-balanced (SB) mixers have better LO-to-IF and LO-to-RF isolation (due to the balance of the balun and diodes match pairs) than the single ended (SE) mixer. In addition to this, SB mixers have better IM suppression. Half of the possible IM products exiting the I-port are suppressed because those with even harmonics of the RF are cancelled due to circuit balance and diode match. Of course, the balun and diode match are never perfect, hence, IM products with even harmonics of the RF are present, but they are attenuated.

Single-balanced mixers have twice the amount of diodes than SE mixers, so they require 3 dB more LO power. Since LO power is larger, the SB mixers will accept a larger dynamic range than SE mixers. If the filter arrangements are simple and the diodes are inexpensive, an SB mixer can be a very cost effective mixer.

C.2 Different Types of Mixer Circuitry

Single-Ended Mixer

Singe-ended (SE) mixers are the simplest type of mixer, since they use only one diode. Figure C.2.1 shows that the LO-, RF-, and IF- ports are electrically the same, being only separated by filters that provide interport isolation. The bandwidths of the filters must not overlap if high isolation is required. All the possible intermodulation products $f_{I,m,n}$ exit the IF-port of SE mixers.

$$f_{I,m,n} = \pm m f_R \pm n f_L \quad (C.2.1)$$

where $m, n = 0, 1, 2, 3, \dots$

Coefficients m and n are integers and can assume any value. SE mixers can operate with very low LO power because only one diode is used.

If the system is narrow-band and does not require large dynamic range, good IM suppression, and high isolation, a SE mixer may be the best choice, since it can be very inexpensive if the filter arrangements are simple. However, if a more broadband mixer that has better IM suppression is required, a balanced mixer is a better choice.

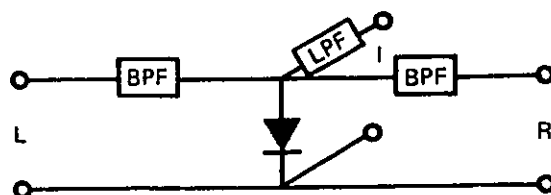


Figure C.2.1 Single-Ended Mixer [2]

Double-Balanced Mixer

Double-balanced (DB) mixers are composed of two SB mixers. Figures C.2.2a) and b) show that combining two SB mixers results in either a ring or a star DB mixer, depending on which type of SB mixers are used. DB mixers are so called because they use two baluns, whereas SB mixers use only one. LO-to-RF and LO-to-IF isolation in DB mixers is achieved in the same way as it is in SB mixers.

Double-balanced mixers theoretically generate only one quarter of the possible IM products; these have odd RF and odd LO harmonics. The other IM products are suppressed, the degree of which is a function of balun balance and diode match.

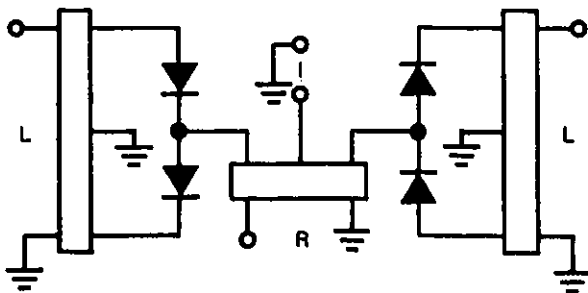


Figure C.2.2(a)
Ring Double Balanced Mixer [2]

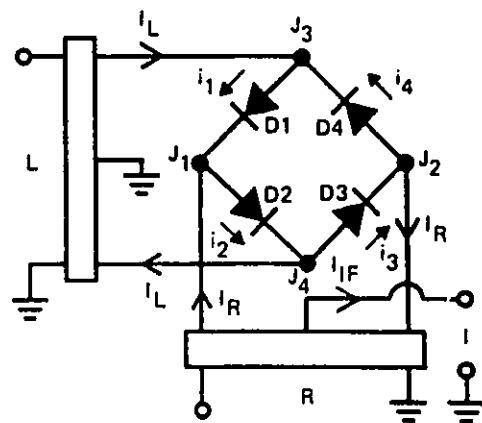


Figure C.2.2(b)
Star Doubled Balanced Mixer [2]

LO power for DB mixers is typically 3 dB higher than that for SB mixers because DB mixers use twice as many diodes as SB mixers. Hence, the 1-dB compression point of a DB mixer is higher than that of an SB mixer, causing correspondingly greater dynamic range and IM suppression.

For most microwave frequency applications, double balanced mixers are usually the best choice over SB and SE mixers. However, for millimeter-wave frequencies the fabrication of multiple well balanced baluns as well as matched diodes is very difficult to achieve. Thereby, the single balanced mixer is more appropriate for millimeter-waves.

Double-Double Balanced Mixer

Double-double balanced (DDB) mixers are so called because they are fabricated of two DB mixers, as shown in Figure C.2.3. DDB mixers usually offer greater dynamic range, better IM suppression and interport isolation, and broader IF-port bandwidth than DB mixers. The large LO power required allows very high intercept points. Thereby, the DDB mixers are good to be used as up-converters. But the disadvantages are the higher LO power requirement and greater cost for the extra four diodes and the intensive labour necessary to match eight diodes instead of four for the DB mixers or only two for the SB mixers.

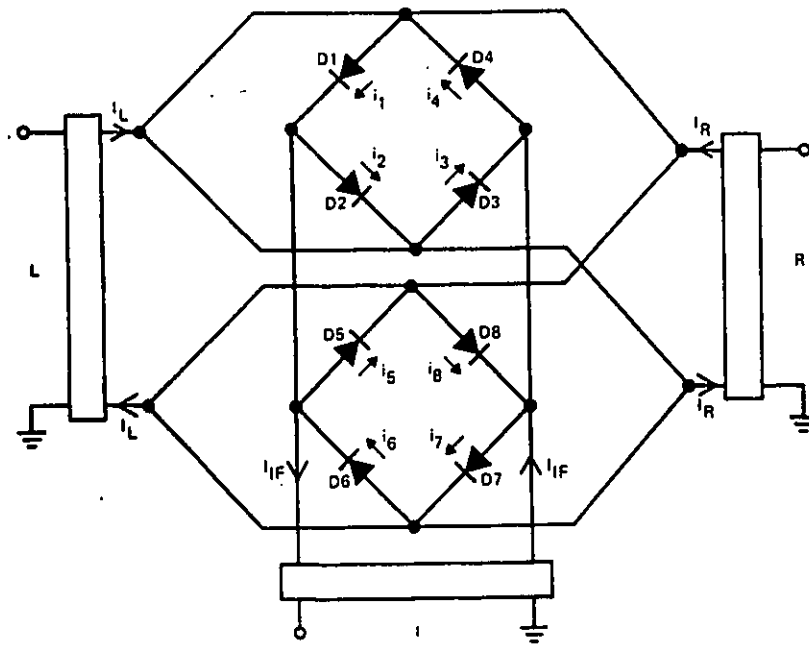


Figure C.2.3 Ring Double Double Balanced Mixer [2]

Harmonic Mixer

Harmonic mixers, as shown in Figure C.2.4, use the second harmonic of the LO to generate the desired IF signal. IF output occurs for IM products having odd f_R and even f_L harmonics, which include the desired $\pm 2f_L \pm f_R$ products. Harmonic mixing allows the LO to operate with half the normally required frequency. These mixers have higher conversion loss and a less stable IF output because frequency drift in the LO is doubled.

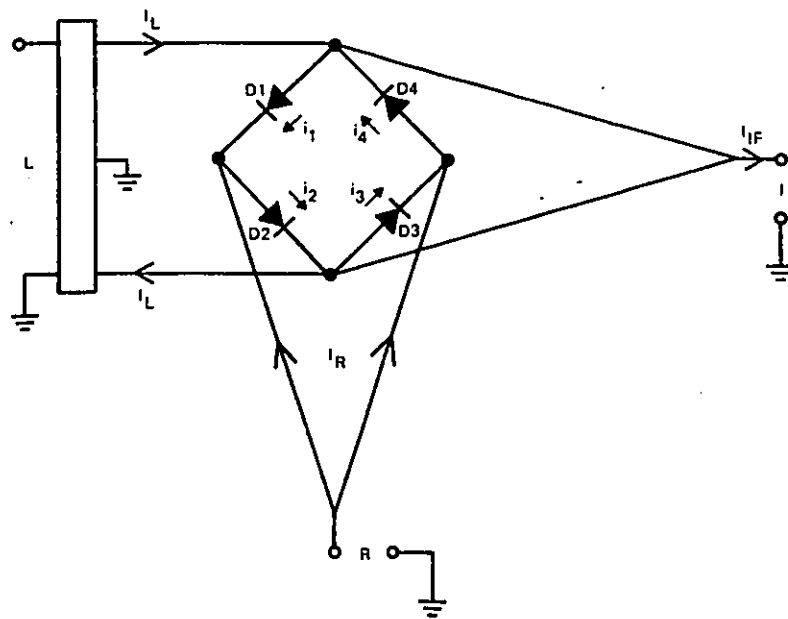


Figure C.2.4 Harmonic Mixer [2]

C.3 Different Types of Mixer Diode

The classical PN diode cannot be used for high frequencies since it is limited in frequency by the charge storage phenomena and by the diffusion capacity due to the presence of minority carriers. To avoid the use of minority carriers, the PN junction is replaced by metal-semiconductor junction. The majority of the diode parameter descriptions are listed at the end of this sub-appendix.

Metal-Semiconductor Junction Diode

The metal-semiconductor contact is described by the Schottky theory or rectification. The rectifying contact is based on majority carrier conduction and in normal operation exhibits virtually no storage of minority carriers.

The operation of a metal-semiconductor diode can be understood by referring to its appropriate electron energy diagrams shown in Figure C.3.1. These show the energies of free electrons in the semiconductor under various conditions of bias.

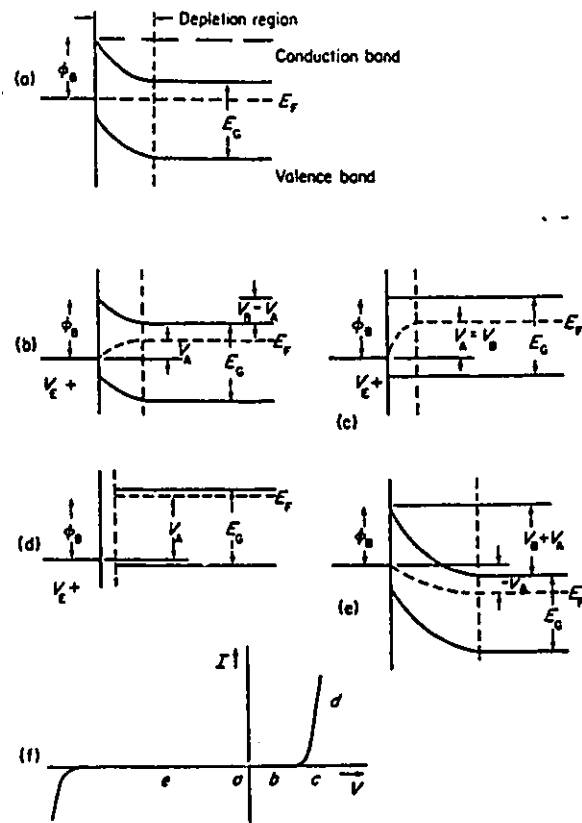


Figure C.3.1 Metal-Semiconductor Energy Band Diagrams
[2]

Zero bias: At zero bias, there is a constant interchange of electrons between the two materials.

Forward bias: Under forward bias conditions, the energy of the electrons in the conduction band of the semiconductor is increased, permitting them to overcome the diffusion potential and to be injected into the metal. The application of external bias V_E results in a lowering of the diffusion barrier by an amount V_A , that is, the height of the potential barrier is reduced, as seen by the electrons in the semiconductor looking towards the metal. ($V_A = V_E - IR_s$, represents a voltage drop across the semiconductor junctions and R_s is the diode series resistance.) This reduction in the barrier results in a current flow which is exponentially dependent on the applied voltage V_A .

Reverse bias: Under reverse bias conditions, the number of electrons in the conduction band of the semiconductor having sufficient energy to overcome the barrier is decreased. In the ideal case the reverse current will be a constant, $-I_s$. The diode theories follow an ideal I - V characteristic of the form shown in Figure C.3.2.

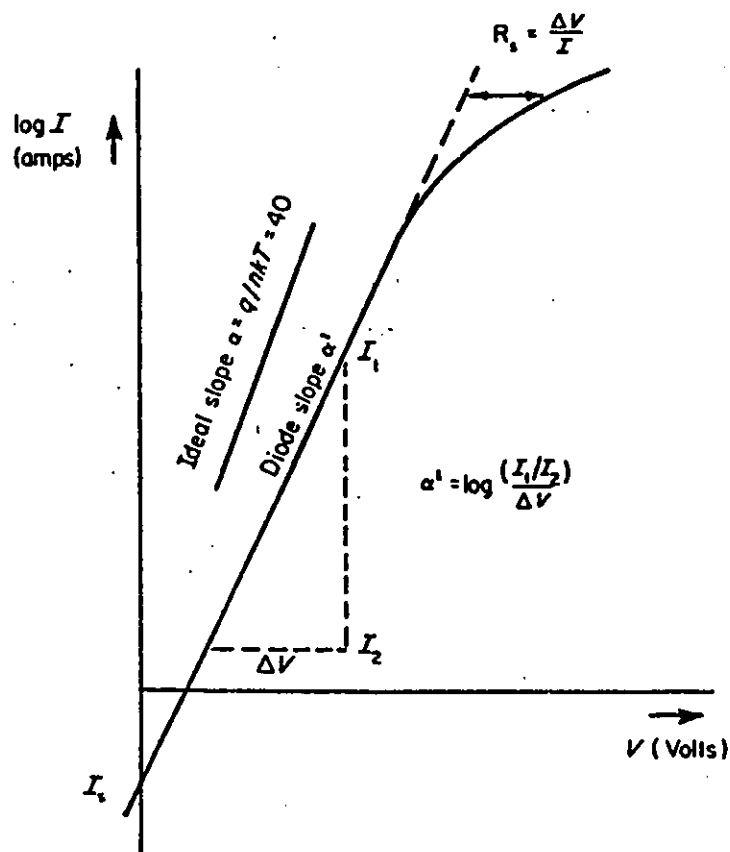


Figure C.3.2 Metal-Semiconductor I-V Characteristic [2]

The current in the direction metal to semiconductor is given by the relation

$$I = I_s (e^{\alpha V} - 1)$$

where

$$\alpha = \frac{q}{n k T}$$

where q is the electron charge, 1.6×10^{-19} coulomb
 n is the ideality factor, ideally $n = 1$
 k is Boltzmann's constant, 1.38×10^{-23} joule/k
 T is the absolute temperature in degrees Kelvin

$$\text{thus } I = I_s \left\{ \exp \left[\frac{q V}{n k T} \right] - 1 \right\} . \quad (\text{C.3.1.})$$

It should be noted that V is the voltage across the rectifier minus the voltage drop across the series resistance of the rectifier.

The saturation current (I_s) is a function of the barrier height (H_b) as expressed by Richardson's equation:

$$I_s = a A^* T^2 \exp \left[- \frac{q H_b}{k T} \right] \quad (\text{C.3.2})$$

where a is the diode area, and A^* is the effective Richardson's constant, which is a function of the effective mass of the semiconductor,

$$A^* = 96 \text{ A cm}^{-2} \text{ K}^{-2} \text{ for silicon [2]}$$

$$A^* = 4.4 \text{ cm}^{-2} \text{ K}^{-2} \text{ for GaAs [2].}$$

The junction dynamic resistance (R_j) is related to the derivation of equation (C.3.1) with respect to voltage as:

$$R_j = \frac{dV}{dI} = \frac{1}{(I+I_s)} \times \frac{nkT}{q}$$

and since

$I_s \ll I$, the junction dynamic resistance is expressed as:

$$R_j \approx \frac{n k T}{q I} = 0.028/I \text{ ohms} \quad (\text{C.3.3})$$

for $n = 1.07$, $T = 300\text{K}$, and I in Amps

The junction capacitance (C_j) is related to the voltage across the junction (V) as given by the relation:

$$C_j = C_j(0) / \sqrt{1 - \frac{V}{V_b - kT/q}} \quad (\text{C.3.4a})$$

and for most application

$$C_j \approx 1.5 C_j(0) \quad (\text{C.3.4b})$$

where, $C_j(0)$ is the zero bias junction capacity
 V_b is the metal-semiconductor potential barrier
 represented in Table C.3.1 [3].

METAL	V_b (VOLTS)	
	SILICON	GALLIUM ARSENIDE
Au	0.8	0.9
Ti	0.4	0.8
Ni	0.5	0.9
Rh	-	0.9
Pt	0.8	-
Pd	0.8	-
Au-Ge	-	0.25
Au-Ag	-	0.25
Cr-Au	0.25	-

Table C.3.1 Metal - Semiconductor Potential Barriers (V_b)

Other parameters are classified as parasitics. Figure C.3.3 identifies five of them: semiconductor layer resistance (R_{s1}), substrate layer resistance (R_{s2}), oxide layer capacitance (C_j), lead inductance (L_p) and the stray capacitance (C_s).

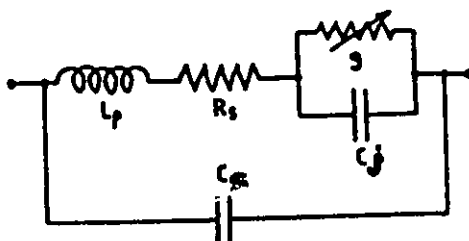


Figure C.3.3 Schottky Diode Equivalent Circuit

Point-Contact Diodes

The pressure contact is formed using a whisker wire to provide electrical connection in the diode, as shown in Figure C.3.4. The mixers realized with these diodes tend to be labour-intensive and unreliable. The electrical performance is sensitive to whisker contact pressure which leads to poor reproductivity and repeatability. These mixers are large in size and fragile to poor handling. For a long time the fragile point-contact diodes were used, they now tend to be replaced by planar Schottky barrier diodes.

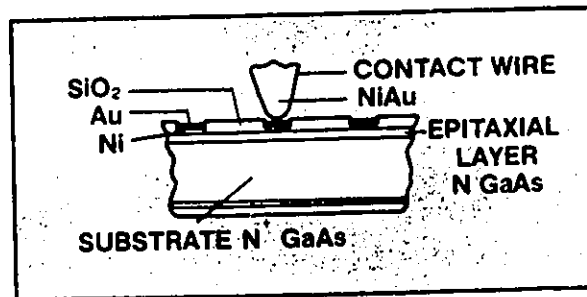


Figure C.3.4 Point Contact Diode [2]

Beam-Lead Diodes

The beam-lead diode approach for high frequency is attractive since it exhibits the smallest lead inductance. Briefly, the beam-lead technique implies the formation of two co-planar beams, one contacting the rectifying junction and the other providing the ohmic contact. A beam-lead structure is shown in Figure C.3.5(a) with the equivalent circuit shown in Figure C.3.5(b).

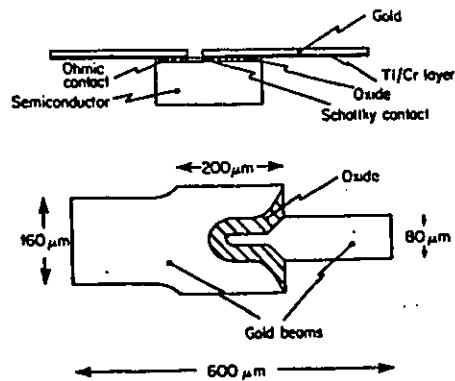


Figure C.3.5(a) Beam Lead Diode [2]

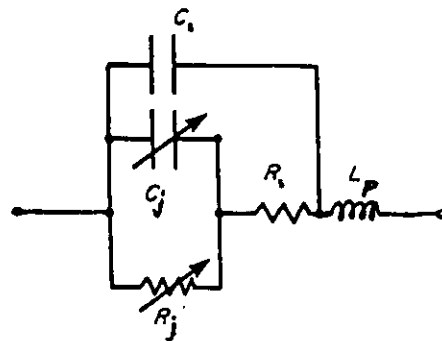


Figure C.3.5(b) Beam Lead Diode Equivalent Circuit [2]

The main design requirements of the beam-lead structure is the minimization and fabrication control of the stray capacitance (C_s) generated by the junction contacting beam overlaying the dielectric layer. This capacitance is effectively in parallel with the junction capacitance and series resistance, thus reducing the series resonant frequency. This parameter can be controlled by modifying the thickness of the dielectric layer and the beam geometry.

Gallium Arsenide beam-lead diodes have been designed with cut-off frequencies greater than 1000 GHz, and with self-resonant frequencies as high as 100 GHz [3]. Table C.3.2 summarizes the characteristics and performance of silicon and gallium arsenide diodes in beam-lead configuration [3].

	Si	GaAs	GaAs	GaAs
Design frequency (GHz)	12	12-32	26-50	60-90
Contact diameter (μm)	20	8	5	3
Total capacitance (pF)	0.25	0.08	0.055	0.045
Junction capacitance (pF)	0.2	0.03	0.025	0.015
Stray capacitance (pF)	0.05	0.05	0.03	0.03
Series resistance (Ω)	8	4	7	12
Series inductance (nH)	0.25	0.25	0.25	0.25
Series resonant freq. (GHz)	10.0	25	38	70
Cut-off frequency (GHz)	500	1000	1000	1000

Table C.3.2 Typical characteristic for MIC Schottky Barrier Devices

Diode Parameter Description

V_{BR} - Breakdown Voltage - The minimum reverse voltage at which avalanche breakdown may occur. It is always desirable to test for the maximum specified I_R (leakage) at the specified V_{BR} since a device driven into avalanche breakdown may be destroyed.

R_S - Series Resistance - The residual resistance of a semiconductor excluding the junction.

R_D - Dynamic Resistance - The combined residual resistance and the junction resistance (a variable which depends on the amount of bias) of a semi-conductor.

N_F - Noise Figure - Ratio of the input signal-to-noise ratio divided by the output signal-to-noise ratio. This is a direct measure of receiver sensitivity.

$$N_f = L_C + 10 \log_{10} [N_R + F_{IF} - 1]$$

Where:

N_F = Overall noise figure in dB

L_C = Conversion loss in dB

N_R = Noise ratio of diodes

F_{IF} = Noise ratio of IF amplifier

T_{SS} - Tangential Signal Sensitivity - T_{SS} is a direct measure of signal-to-noise voltage in a receiver circuit. T_{SS} is the input power level (referred to 1 mW) which produces a signal-to-noise ratio of 2.5

τ - Lifetime - A measure of the length of time a diode will store charge.

T_{RR} - Reverse Recovery Time - A measure of the switching speed of a diode.

Θ_{JC} - Thermal Resistance - A measure of the temperature rise of the junction above the ambient heat sink temperature. Thermal impedance is expressed in $^{\circ}\text{C}/\text{watt}$. Dissipated input power is limited by:

$$\frac{T_{MAX} - T_{AMB}}{\Theta_{JC}}$$

C_J - Junction Capacitance - The capacitance of the diode junction only.

C_T - Total Capacitance - The total capacitance is $C_{\text{junction}} + C_{\text{package}} + C_{\text{interconnect}}$, where $C_{\text{interconnect}}$ is typically .04 pf for a wire.

Z_{IF} - IF Impedance - The source impedance of a mixer diode at the derived output frequency.

VSWR - Voltage Standing Wave Ratio - A measurement of how well matched a device is to a given characteristic impedance, usually 50 ohms, in a specified fixture. This is a measurement of diode uniformity.

I_R - Reverse Leakage - A measure of the reverse current of a diode at a specified reverse voltage.

V_F - Forward Voltage - A measure of the forward voltage drop at a specified current level.

Maximum Rating - Maximum safe operating condition. Exceeding maximum rating may cause permanent change in diode characteristics.

C.4 Different Types of Mixer Construction

Waveguide Mixer

Balanced mixer designs in waveguide are well established. A widely used configuration utilizes a crossbar diode mounting arrangement, as shown in Figure C.4.1. RF and LO are applied orthogonally. Incident LO is in phase at the two diodes and the RF is applied in anti-phase (differing by 180 degrees). In order to ensure satisfactory performance, the two diodes must be very closely matched. The IF filter is in suspended substrate.

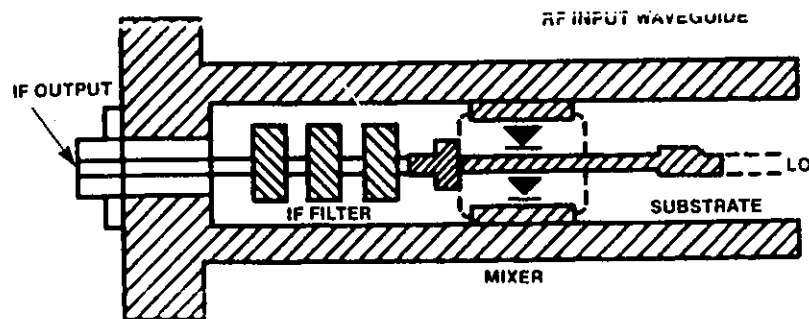


Figure C.4.1 Cross-Bar Mixer [2]

Fin-Line Mixer

There are several possible approaches to fin-line mixer design [4],[5]. These mixers readily lend themselves to integration; as the entire circuit can be produced using photolithographic techniques as has been described in this thesis. A general mixer arrangement is shown in Figure C.4.2. This is a hybridization of several transmission media that yields a balanced mixer configuration. This is the approach chosen by the author for this thesis.

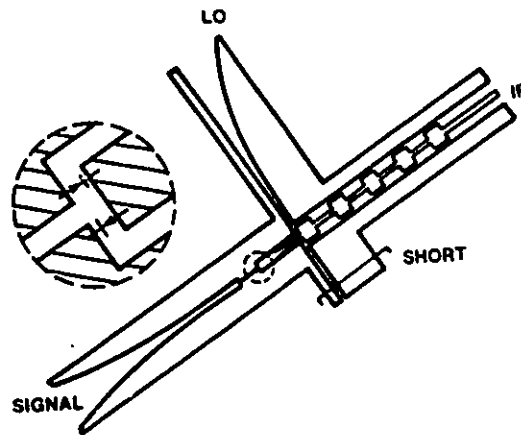


Figure C.4.2 Layout of Fin-Line Mixer, Menzel [8]

Microstrip Mixer

Microstrip circuits offer the ultimate miniaturization achievable in hybrid integrated circuits. Microstrip circuits offer a remarkable reduction in size and weight over the waveguide and fin-line circuits described above. These circuits offer the lowest cost in production. Although the future trend in millimeter-wave technology will be to develop such circuits up to 94 GHz (and even up to 140 GHz [8]), the major factor will be the limiting resolution available from various fabrication techniques. Conventional photolithography techniques tend to limit circuit development to around 40 GHz. Above 40 GHz, the yield becomes very poor and the cost becomes prohibitive with photolithography. Figure C.4.3 shows an hybrid ring arrangement. DC blocks are incorporated in the RF and LO feeds and the beam-lead mixer diodes are in an anti-parallel array at the remaining two ports of the hybrid. IF is extracted through a low-pass filter line.

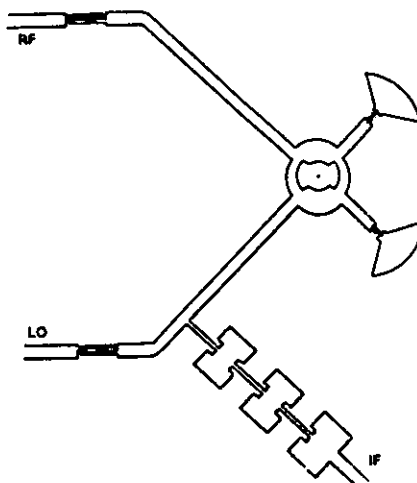


Figure C.4.3 Layout of a Balanced Microstripline Mixer
[2]

Monolithic Mixer

As GaAs technology advances, the monolithic integration of passive and active devices on GaAs substrates is becoming attractive at millimeterwave lengths. The advantages of monolithic circuits are considerable. They include improved control of parasitics and greater miniaturization than is possible in the hybrid integrated circuits discussed in previous section. One of the approaches involves the combination of GaAs microstripline substrates, monolithic GaAs IC and hybrid MIC. Only the circuits critical for the matching of active devices are monolithically integrated on a GaAs substrate. An approach of this type has been developed by Bastida [1] and is shown in Figure C.4.4.

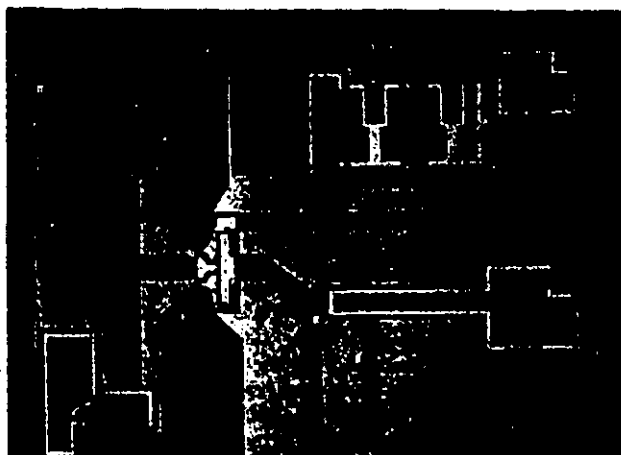


Figure C.4.4 A MMIC GaAs Mixer Chip [1]

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APPENDIX D

REVIEW OF COPLANAR LINE PROPERTIES

D.1 Basic Configuration

Coplanar line (also called coplanar waveguide or coupled slotlines) was proposed by Wen [1]. As shown in Figure D.1.1 it consists of a centre strip with two equidistant planes located parallel to and in the plane of the strip, (i.e. on the same surface of the dielectric slab). The coplanar line has two modes of propagation, the odd (also called unbalanced or non-symmetric) mode and even (also called balanced or symmetric) mode.

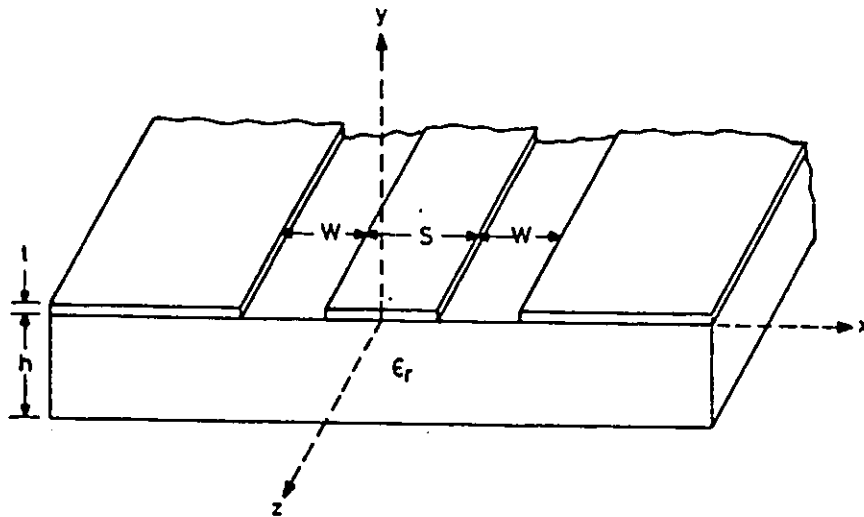


Figure D.1.1 Coplanar-Line Parameters Representation
[1]

D.2 Odd Mode

Odd Mode Field Propagation

The odd mode field configuration in coplanar line is represented in Figure D.2.1.

The quasi-static results for coplanar line presented by Wen [1] may be modified to give the characteristic impedance Z_0 and the effective dielectric constant ϵ_e .

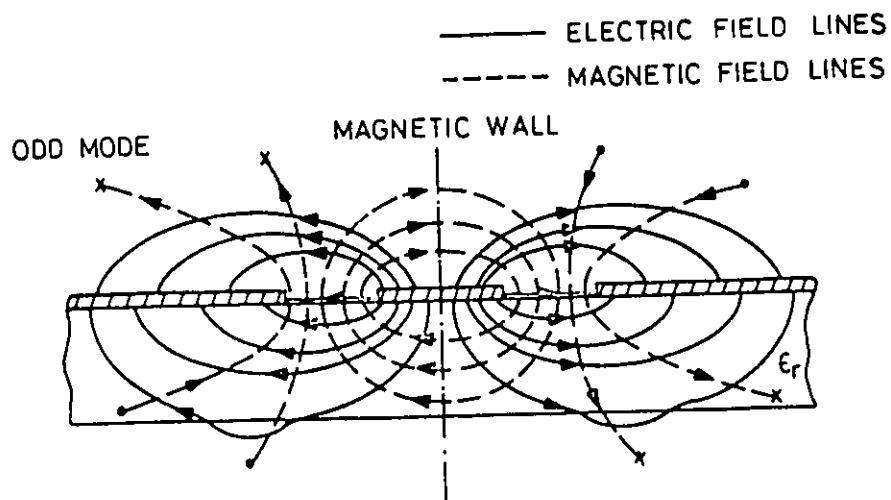


Figure D.2.1 Coplanar-Line Odd Mode Field Distribution
[1]

Expression For Guided Wavelength

The numerical results of Davis et al [2] have been curve-fitted [3] to obtain the following closed form expression for ϵ_e :

$$\epsilon_e = \frac{\epsilon_r + 1}{2} \left[\tanh \{ 1.785 \log (h/w) + 1.75 \} + \frac{kw}{h} \{ 0.04 - 0.7k + 0.01 (1 - 0.1\epsilon_r) (0.25 + k) \} \right] \quad (D.1)$$

Where w and h are defined in Figure D.1.1 and k is given by (D.3 c)

Expression For Characteristic Impedance

The characteristic impedance in any wave propagating medium (such as coplanar line) is defined as:

$$Z_0 = Z_{0\infty} / \sqrt{\epsilon_e} \quad (D.2)$$

Where; ϵ_e is given by D.1

$$Z_{0\infty} = 30 \ln \left[\frac{2(1+\sqrt{k'})}{1-\sqrt{k'}} \right] \quad \text{for } 0 \leq k \leq 0.707 \quad (D.3a)$$

$$= \frac{30\pi^2}{\ln \left[2 \left(\frac{1+\sqrt{k}}{1-\sqrt{k}} \right) \right]} \quad \text{for } 0.707 \leq k \leq 1 \quad (D.3b)$$

$$\begin{aligned} \text{where [4]; } k &= \frac{S}{S + 2W}, \\ k' &= \frac{(1-k^2)^{1/2}}{S+2W} \end{aligned} \quad (D.3c)$$

According to Dupuis and Campbell [5], to attain $\pm 10\%$ tolerance from the previous closed-form expressions it is necessary to employ a substrate thickness $h \geq 0.05$ " and an aspect ratio $k \geq 0.4$. For $h/w \geq 2$, the characteristic impedance Z_0 will be within 5% of the predicted value.

D.3 Even Mode

Even Mode Field Propagation

The even mode field configuration in coplanar-line is represented in Figure D.3.1.

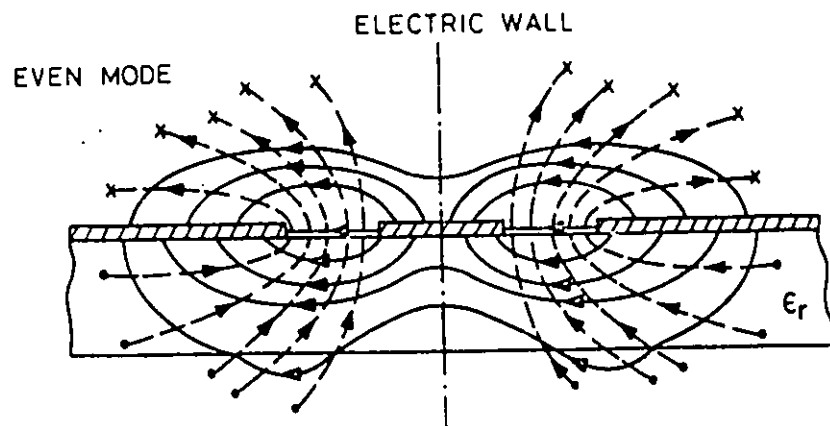


Figure D.3.1 Coplanar-Line Even Mode Field Distribution
[1]

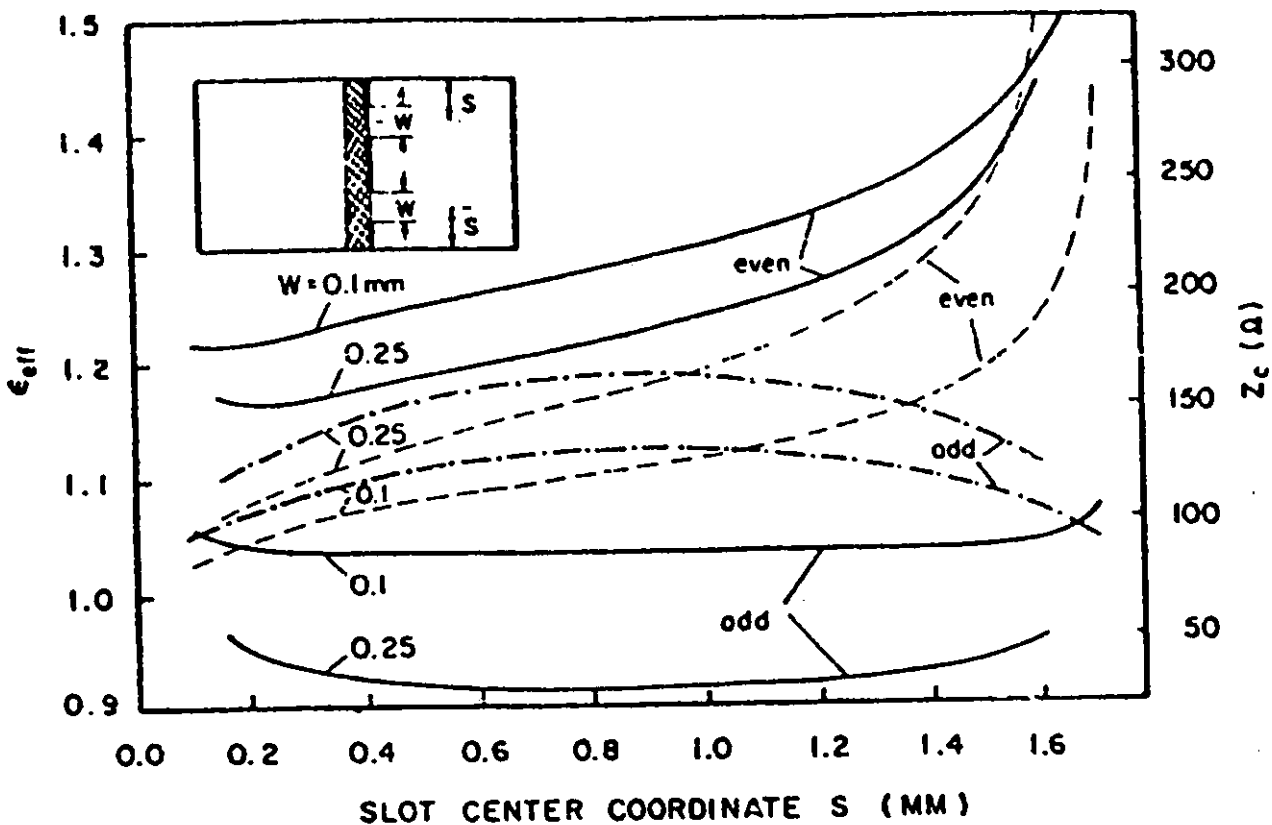


Figure D.3.2 Graphical Representation for the Guided Wavelength (λ_g) and Characteristic Impedance (Z_0) of the Even Mode [6]

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- [5] P.A.J. Dupuis, "Characteristics Impedance of Surface-Strip Coplanar Waveguides," Electronics Letters, Vol. 9, No. 12, pp. 354-355, 1973
- [6] L.P. Schmidt, T. Itoh, H. Hofmann, "Characteristics of Unilateral Fin-Line Structure with Arbitrarily Located Slots", IEEE Transactions on Microwave Theory and Technique, Vol. MTT-29, No. 4, April 1981, Page 352-355.

APPENDIX E

REVIEW OF FILTERS IN GENERAL

E-1 Parameters for Evaluation of the Filter Performance

There are many parameters which describe the performance of a filter. A description of parameters related to bandpass filters follows. Similar parameters describe low pass filter performance.

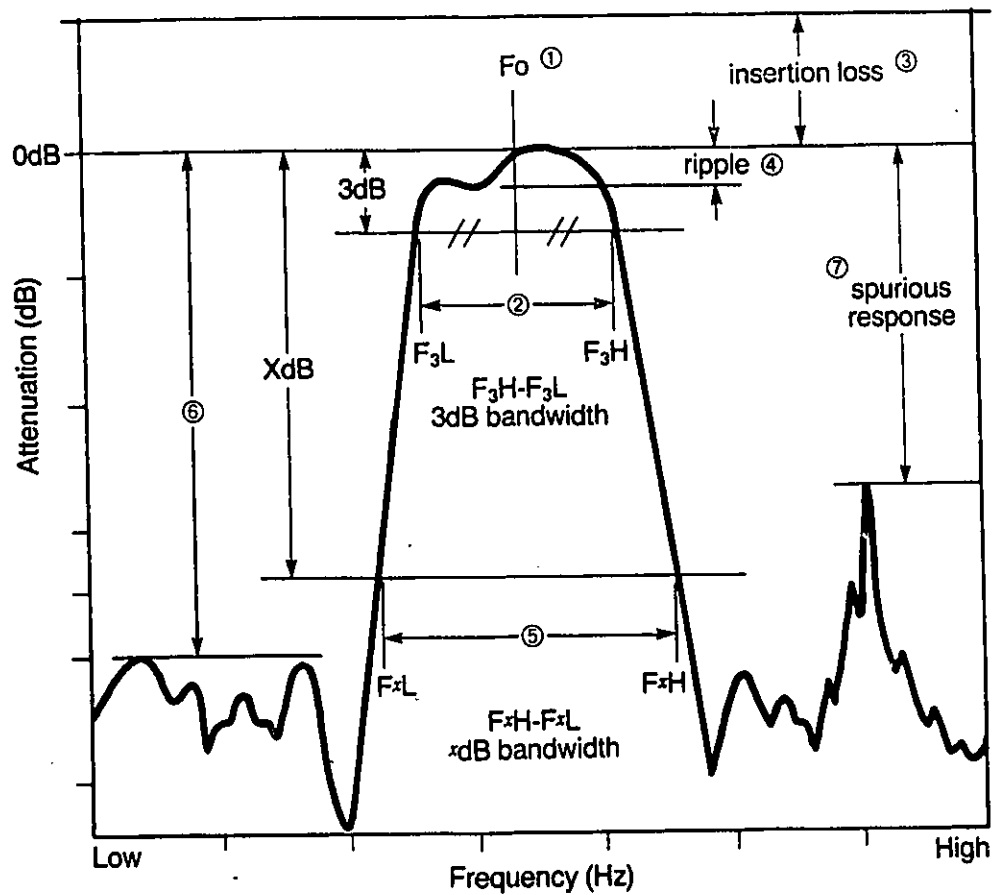


Figure E.1.1 Graphical Representation of Band Pass Filter [1]

Center Frequency (F_o)

This is the frequency in the center of the pass-bandwidth. However, the center frequency is sometimes expressed as the point where the loss is at its lowest. (Figure E.1.1, #1).

Pass-Bandwidth (3dB Bandwidth) (3dB BW)

This term signifies the difference between the two frequencies where the attenuation increase by 3dB above the level of the minimum loss point. (Figure E.1.1, #2).

Insertion Loss (IL)

This quantity is expressed as the input/output ratio at the point of minimum loss. (The insertion loss for some products is expressed as the input/output ratio at the center frequency). (Figure E.1.1, #3).

Ripple

If there are peaks and valleys in the pass-bandwidth, the ripple expresses the difference between the maximum peak and the minimum valley. (Figure E.1.1, #4).

Attenuation Bandwidth (dB Bandwidth) (20 dB BW)

This is the bandwidth at a specific level of attenuation. Attenuation may be expressed as the ratio of the input signal strength to the output signal strength in decibels. (Figure E.1.1, #5).

Stopband Attenuation

This is the level of signal strength at a specified frequency outside of the passband. (Figure E.1.1, #6).

Spurious Response (SR)

This is the difference in decibels between the insertion loss and the spurious signal in the stopband. (Figure E.1.1, #7).

E-2 Different Types of Filter Construction

The classification of electric wave filters can be done in several ways. They may be grouped in terms of the frequency spectrum against the realization (Figure E.2.1, [1]), or they may be grouped according to the elements that make up the filters.

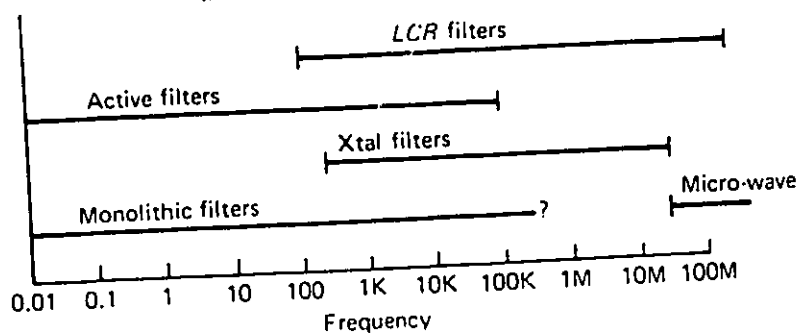


Figure E.2.1 Filter Frequency Guide [1]

E-3 Program Listing For Microstrip LPF

The program listing has been destroyed (permanently deleted) during a backup exercise. Therefore, this listing cannot be made available.

REFERENCES

- [1] "Electrical Engineer's Reference Book", 5th edition.
Edited by M.A. Langther and M.G. Say. pp. 36/3.
Butterworth & Co. Ltd, 1983.

APPENDIX F

REVIEW OF OSCILLATORS IN GENERAL

F-1 Parameters for Evaluation of Oscillator Performance

There are many parameters which describe the performance of an oscillator. A description of the most important follows.

Frequency Range (Freq.)

The mechanically or electronically tunable bandwidth over which the oscillator meets all specifications.

Output Power (P_{out})

The minimum and/or maximum output power at the output frequency under all specified conditions. Usually the specified conditions are temperature, load, VSWR, and supply voltage variations. It is typically expressed in dBm or Milliwatts (mW).

Frequency Stability with Temperature

A measure of the change in frequency over the specified operating temperature range. It is commonly expressed as parts-per-million per degree Celsius (ppm/C).

Mechanical Tuning

Maximum output frequency deviation which can be achieved without significantly affecting oscillator characteristics. For DRO's (dielectrics resonator oscillator) this is mainly achieved by adjusting the air gap spacing between the dielectric resonator and the tuning screw directly above the resonator. A typical mechanical tuning range is $\pm 1\%$ of the center frequency.

Electronic Tuning

The maximum output frequency deviation which can be achieved without significantly affecting oscillator characteristics. It is achieved by adjusting the bias voltages of a FET coupled to the dielectric resonator. Typical electronic tuning ranges are $\pm 0.1\%$.

Spurious Outputs

Spurious outputs are both harmonically and non-harmonically related signals. Their tolerance amplitude should be specified within and out of the frequency range of the oscillator. Typical values range from -60 dBc to -80 dBc.

Harmonic Signals

Signals which are coherently related to the output frequency. In general these signals are integer multiples of the output frequency. Typical values range from -20 dBc to -35 dBc.

FM Noise

The short term frequency variations in the frequency which appear as energy at frequencies other than the carrier. It is usually expressed in dBc or as a RMS frequency deviation in a specified video bandwidth at a specified frequency offset from the carrier.

Efficiency

The ratio of RF power generated over the DC bias power required.

F-2 Other Types of Oscillator Circuitry

DSO (Dielectrically Stabilized Oscillator)

For DSO (or often called, reflection oscillator), the initial design starts with either an unstable device or external feedback (low Q) to obtain negative resistance and reflection gain at the desired frequency. As shown in Figure F.2.1 a dielectric resonator is placed approximately one-half wavelength away from the FET device. In this configuration, the dielectric resonator acts as a weakly coupled bandstop filter with a very high external Q . Part of the output energy is reflected toward the device and such a self-injected oscillator will generate a signal at the resonant frequency of the dielectric resonator. However, because of the reflective mode of operation these designs are sensitive to load changes and require an output isolator or buffer amplifier.

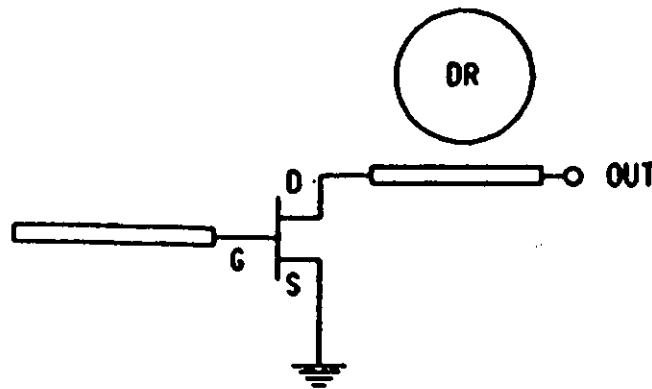


Figure F.2.1 Basic Configuration of DSO

DRO (Dielectric Resonator Oscillator) In Series Feedback

The series feedback DRO oscillator (illustrated in Figure F.2.2) consists of a high gain low noise FET, a 50 ohm transmission line connected to the FET gate (which is terminated with a 50 ohm resistor for out-of-band stability), a dielectric resonator coupled to the line and located at a specified distance from the gate, a shunt reactance connected to the FET source or drain, and matching output impedance. Critical to the performance of this circuit is the placement of the dielectric resonator on the gate port.

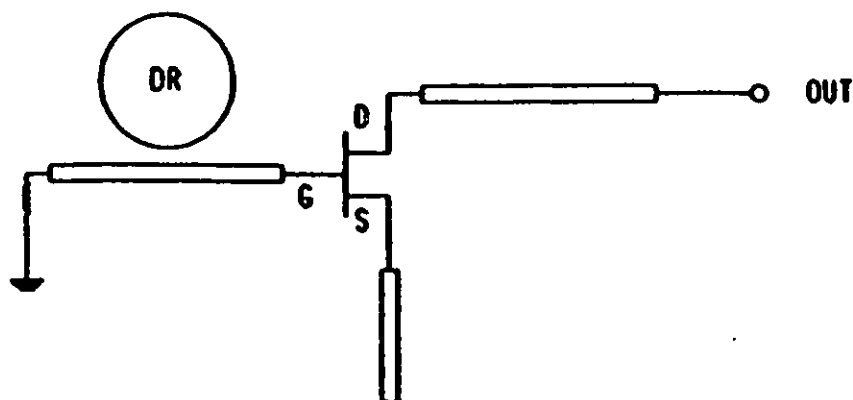


Figure F.2.2 Basic Configuration of Series Feedback DRO

APPENDIX G

COMPONENT SPECIFICATIONS

G-1 Matched-Pair Beam Lead Schottky Diode Specifications

Metelics Corporation part number MSS40, 241-B20 was used.

Preliminary Specifications @ 25°C

Model Outline	V_F Typ. @ 1 mA (Volts)	ΔV_F Typ. @ 1 mA (mV)	V_{BR} Min. @ 10 μ A (Volts)	C_J^1 Typ. @ 0 V (pF)	ΔC_J Max. @ 0 V (pF)	R_D^2 Typ. @ 5 mA (Ohms)	F_{CO}^3 Typ. (GHz)
MSS-30,242-B20 MSS-30,242-E35 MSS-30,242-H30	.29	10	2	.07	.03	18	175
MSS-30,248-B20 MSS-30,248-E30 MSS-30,248-H30	.27	10	3	.09	.03	15	180
MSS-30,254-B20 MSS-30,254-E30 MSS-30,254-H30	.26	10	3	.22	.03	8	240
MSS-40,244-B20 MSS-40,244-E30 MSS-40,244-H30	.44	10	3	.08	.03	24	100
MSS-40,248-B20 MSS-40,248-E30 MSS-40,248-H30	.44	10	3	.12	.03	15	130
MSS-50,248-B20 MSS-50,248-E30 MSS-50,248-H30	.53	10	3	.12	.03	15	130

¹ C_J is measured @ 1MHz across a single junction.

² $R_D = R_j + R_s$; $R_j \approx 28\Omega$ (1mA)

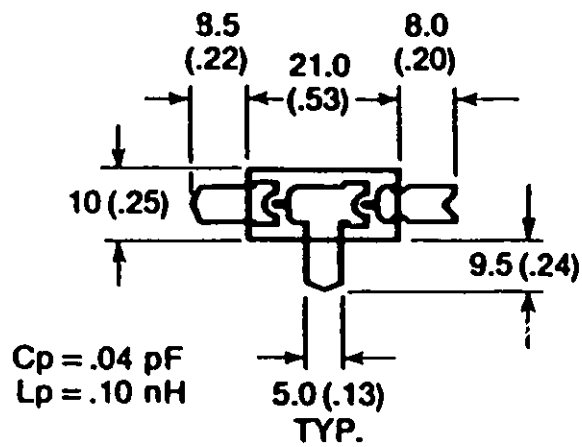
³ $F_{CO} = 1/(2\pi \cdot R_s \cdot C_J)$

Package Outlines
Dimensions are in Mils (mm).

POLARITY: CATHODE
DOT
CUT LEAD
POINTED BEAM

ANODE
CAP
PAD

B20



PRELIMINARY DATA SHEET

Low Noise Ku-K Band GaAs MESFET

FEATURES

- VERY HIGH $f_{\max}$
100GHz
- LOW NOISE FIGURE
0.4dB, $G_a = 14.5$ dB at 4.0GHz
0.8dB, $G_a = 11.5$ dB at 8.0GHz
1.4dB, $G_a = 10.0$ dB at 12.0GHz
1.9dB, $G_a = 8.0$ dB at 18.0GHz
3.3dB, $G_a = 6.0$ dB at 26.0GHz
- 0.3 MICRON RECESSED GATE
- N^+ CONTACT LAYER
(Triple Epitaxial Technology)
- PROVEN RELIABILITY AND STABILITY

DESCRIPTION AND APPLICATIONS

The NE673 features a super low noise figure and high associated gain thru K-band by employing a recessed 0.3 micron gate and triple epitaxial technology for industrial, military, and space applications.

The device is available as a chip (NE67300). The chip's gate and channel are glassivated with a thin layer of Si_3N_4 for mechanical protection only. The NE67383 is in a rugged hermetically sealed metal-ceramic stripline package selected for NF_{opt} performance at 12.0 GHz. The NE67383-4 is selected for NF_{opt} performance at 4.0 GHz.

PERFORMANCE SPECIFICATIONS ($T_a = 25^\circ C$)

NE PART NUMBER EIAJ ¹ REGISTERED NUMBER PACKAGE CODE			NE67300 CHIP			NE67383/NE67383-4 ² 2SK407 83		
SYMBOLS	PARAMETERS AND CONDITIONS	UNITS	MIN	TYP	MAX	MIN	TYP	MAX
$f_{\max}$	Maximum Frequency of Oscillation at $V_{DS} = 3V$, $I_{DS} = 30mA$	GHz		100			100	
MAG	Maximum Available Gain ³ at $V_{DS} = 3V$, $I_{DS} = 30mA$ $f = 8GHz$	dB		15			15	
	$f = 12GHz$	dB		12			12	
	$f = 18GHz$	dB		8.5			8.5	
NF_{opt}	Optimum Noise Figure ⁴ at $V_{DS} = 3V$, $I_{DS} = 10mA$ $f = 4GHz$, $\Gamma_{opt} = .64 \angle 69^\circ$, $R_n = .38$	dB		0.4			0.4	0.6 ²
	$f = 8GHz$, $\Gamma_{opt} = .55 \angle 115^\circ$, $R_n = .20$	dB		0.8			0.8	
	$f = 12GHz$, $\Gamma_{opt} = .48 \angle 155^\circ$, $R_n = .20$	dB		1.4	1.6		1.4	1.6
	$f = 18GHz$, $\Gamma_{opt} = .46 \angle -33^\circ$, $R_n = .40$	dB		1.9				
	$f = 26GHz$	dB		3.3				
G_a	Associated Gain at Optimum Noise Figure at $V_{DS} = 3V$, $I_{DS} = 10mA$ $f = 4GHz$	dB		14.5		12 ²	14.5	
	$f = 8GHz$	dB		11.5			11.5	
	$f = 12GHz$	dB	8.5	10.0		8.5	10.0	
	$f = 18GHz$	dB		8.0				
	$f = 26GHz$	dB		6.0				
P_{1dB}	Output Power at 1 dB Compression Point at $V_{DS} = 3V$, $I_{DS} = 30mA$ $f = 12GHz$	dBm		14.5			14.5	

SEE NOTES ON PAGE 3.

G-2 Field-Effect Transistor Specifications

NEC MESFET part number NE67383 was used.

ELECTRICAL CHARACTERISTICS($T_a=25^{\circ}\text{C}$)

NE PART NUMBER EIAJ ¹ REGISTERED NUMBER PACKAGE CODE			NE67300			NE67383/NE67383-4 ² 2SK407 83		
SYMBOLS	PARAMETERS AND CONDITIONS	UNITS	MIN	TYP	MAX	MIN	TYP	MAX
I_{DSS}	Drain Current at $V_{\text{DS}} = 3\text{V}$, $V_{\text{GS}} = 0$	mA	20	40	120	20	40	120
V_{p}	Pinch-off Voltage at $V_{\text{DS}} = 3\text{V}$, $I_{\text{DS}} = 0.1\text{mA}$	V	-0.5	-1.1	-3.5	-0.5	-1.1	-3.5
g_{m}	Transconductance at $V_{\text{DS}} = 3\text{V}$, $I_{\text{DS}} = 10\text{mA}$	m Ω	20	50	100	20	50	100
I_{GS}	Gate to Source Leakage Current at $V_{\text{GS}} = -5\text{V}$	μA		1.0	10		1.0	10
R_{th}	Thermal Resistance (ch-a)	$^{\circ}\text{C/W}$			190 ⁵			450
P_{T}	Total Power Dissipation	mW			400 ⁵			270

SEE NOTES ON PAGE 3.

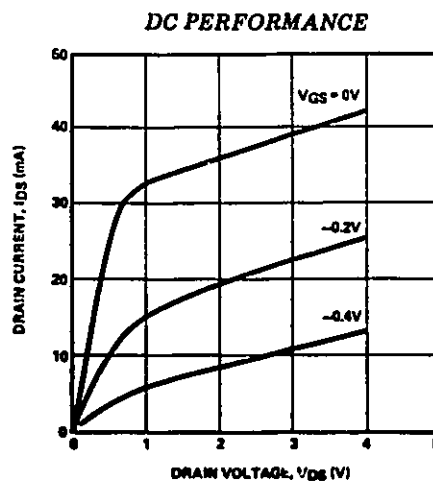
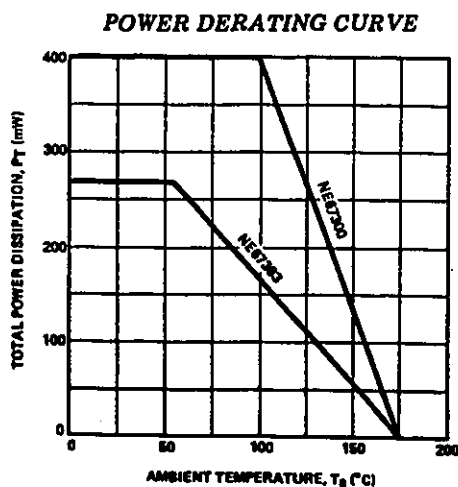
ABSOLUTE MAXIMUM RATINGS($T_a=25^{\circ}\text{C}$)

SYMBOLS	PARAMETERS	UNITS	RATINGS
V_{DS}	Drain to Source Voltage	V	5.0
V_{GS}	Gate to Source Voltage	V	-6.0
I_{DS}	Drain Current	mA	120
P_{in}	RF Input Power	mW	40
T_{ch}	Channel Temperature	$^{\circ}\text{C}$	175
T_{stg}	Storage Temperature	$^{\circ}\text{C}$	-65~+175

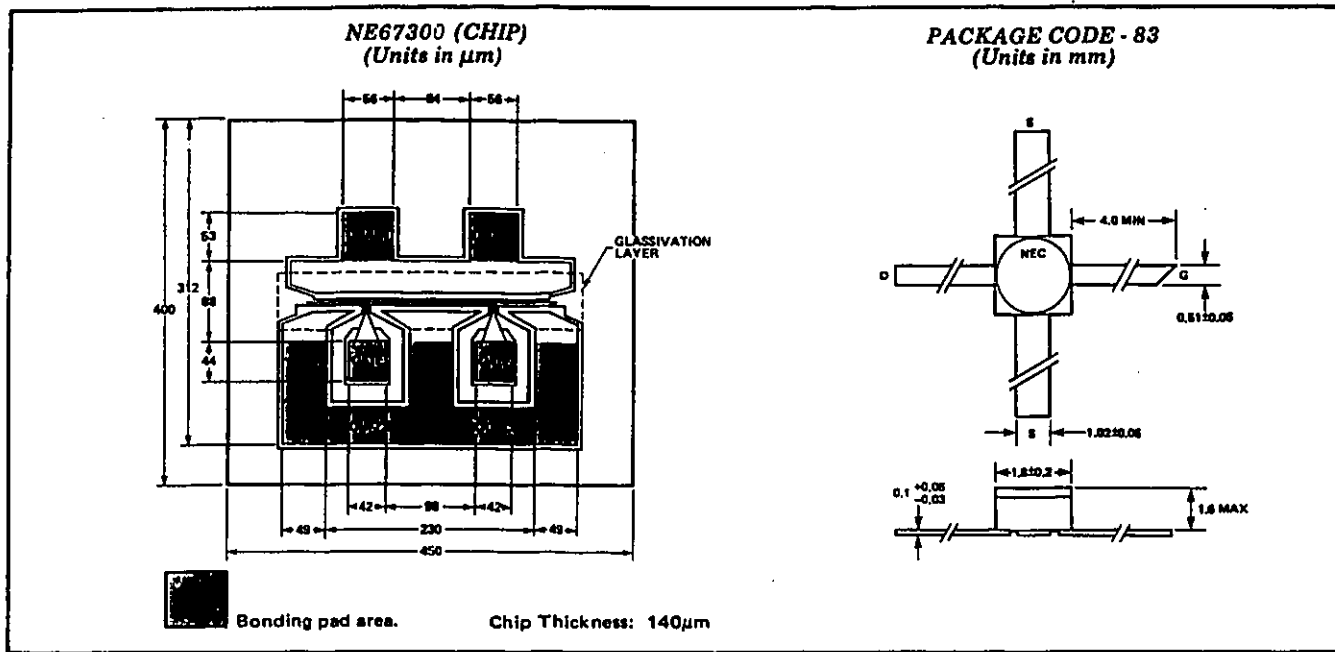
RELIABILITY SCREENING (HES-32325-05)

TEST	GRADE		
	C Military Avionics	CK Military General	D Industrial
Precap Visual Inspection	100%	100%	—
Vacuum Bake	100%	—	—
High Temperature Storage	100%	100%	100%
Temperature Cycling	100%	100%	—
Thermal Shock	100%	—	—
Mechanical Shock (Y only)	100%	—	—
Acceleration	100%	100%	—
Gross Leak Test	100%	100%	100%
Fine Leak Test	100%	100%	100%
Area of Safe Operation (power only)	100%	100%	100%
High Temp. Reverse Bias (HTRB)	Optional	Optional	—
Particle Impact Noise Detection (PIND)	Optional	Optional	—
Electrical (DC) Tests	100%	—	—
Power Burn-in (100 hrs)	100%	100%	—
Delta Calculation	Optional	—	—
Group A Screening	100%	100%	100%
Group A Data	Optional	—	—
External Visual	100%	100%	100%

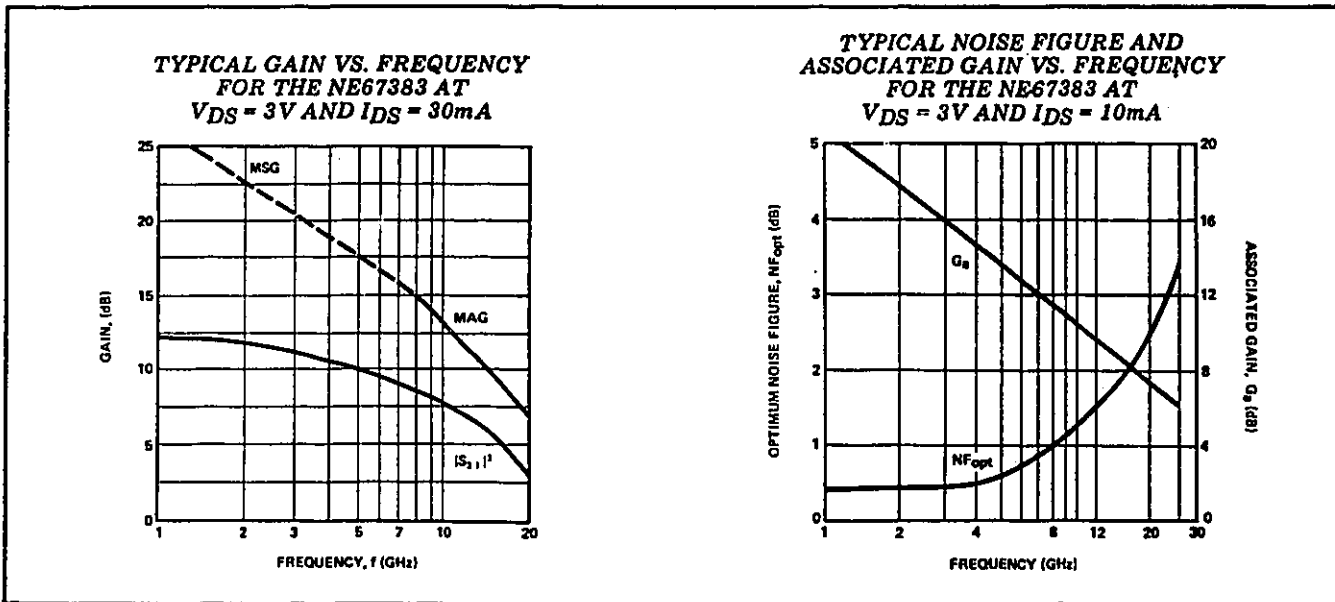
DEVICE CHARACTERISTICS($T_a=25^{\circ}\text{C}$)



PHYSICAL DIMENSIONS



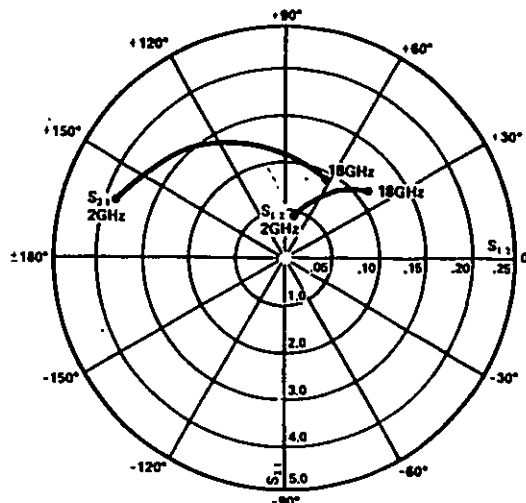
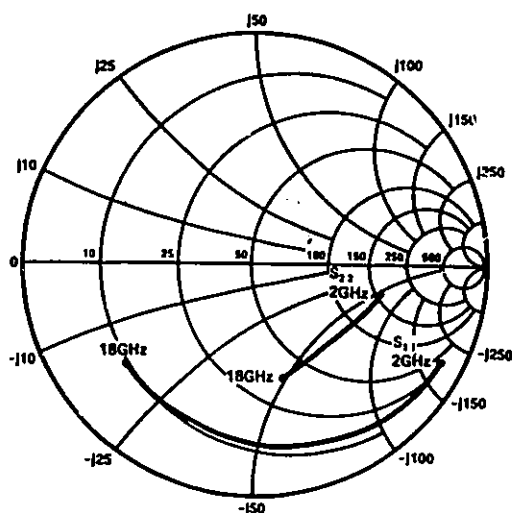
PERFORMANCE CHARACTERISTICS ($T_a = 25^\circ\text{C}$)



NOTES:

- Electronic Industries Association - Japan.
- NE67383-4 is tested for NF_{opt} at 4.0GHz. The standard NE67383 is tested at 12.0GHz.
- Gain Calculations: $MAG = \frac{|S_{21}|}{|S_{11}|} (k\sqrt{k^2 - 1})$, $k = \frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{12}||S_{21}|}$, $\Delta = S_{11}S_{22} - S_{21}S_{12}$
- Typical values of noise figures are those obtained when 50% of the devices from a large number of lots were individually measured in a circuit with the input individually tuned to obtain the minimum value. Maximum values are criteria established on the production line as a "go-no-go" screening test with the fixture tuned for the "generic" type but not for each specimen.
- R_{th} (channel to ambient) for chips mounted on a copper heatsink.

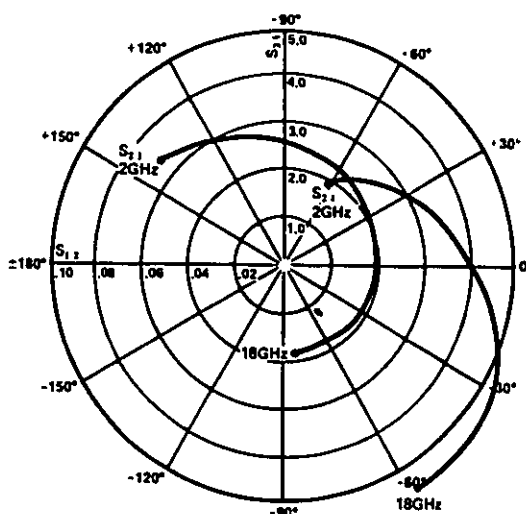
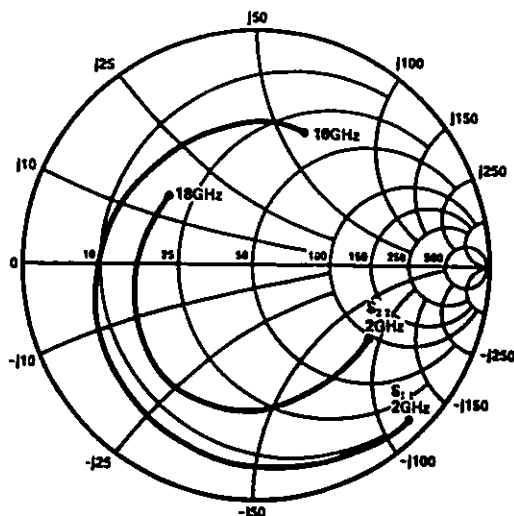
NE67300 COMMON SOURCE SCATTERING PARAMETERS



S-MAGN AND ANGLES: ($V_{DS} = 3V$, $I_{DS} = 10mA$)

FREQUENCY (MHz)	S11		S21		S12		S22	
2000	.95	-26	3.79	161	.04	79	.59	-13
4000	.89	-50	3.26	141	.06	66	.58	-24
6000	.82	-70	2.83	126	.08	56	.54	-33
8000	.78	-88	2.55	114	.09	51	.50	-42
10000	.73	-102	2.21	104	.10	48	.47	-48
12000	.71	-114	2.16	93	.10	43	.45	-55
14000	.71	-122	2.11	90	.11	44	.47	-62
16000	.67	-128	1.92	76	.11	43	.49	-64
18000	.66	-140	1.81	63	.11	40	.52	-70

NE67383 COMMON SOURCE SCATTERING PARAMETERS



S-MAGN AND ANGLES: ($V_{DS} = 3V$, $I_{DS} = 10mA$)

FREQUENCY (MHz)	S11		S21		S12		S22	
2000	.97	-43	3.39	140	.04	61	.63	-32
4000	.88	-79	2.85	107	.06	35	.61	-58
6000	.84	-103	2.57	81	.07	20	.62	-77
8000	.78	-126	2.33	56	.07	12	.62	-94
10000	.71	-146	2.11	38	.07	4	.61	-107
12000	.64	-174	1.98	12	.08	-7	.60	-123
14000	.59	157	1.95	-11	.10	-15	.57	-139
16000	.57	111	1.83	-42	.11	-31	.47	-169
18000	.62	64	1.74	-84	.11	-58	.43	137



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CALIFORNIA EASTERN LABORATORIES, INC. • Headquarters • 3280 Jay Street | • Santa Clara, CA 95050 • (408) 988-3500 • Telex 34-6393 or 171197

G-3 Dielectric Resonator Specifications

Trans-Tech dielectric resonator part number D8513.125.050 was used.

**D8500 Series
 Temperature Stable Resonator Materials**

**DIELECTRIC
 BULLETIN NO. 41-83**

Characteristics

Composition Type	Dielectric Constant $\epsilon' \pm 1.5\%$	Temp. Coef. of Res. Freq. $\tau_f \pm 0.5\text{ppm}/^\circ\text{C}$	Q Factor @ 4 GHz
BARIUM TETRATITANATE D-8512	38.6	+ 4 ppm/ $^\circ\text{C}$	> 10,000
Zr/Sn TITANATE D-8513	37.9	+ 6 ppm/ $^\circ\text{C}$	> 10,000
D-8514	37.7	+ 3 ppm/ $^\circ\text{C}$	> 10,000
D-8515	37.5	0 ppm/ $^\circ\text{C}$	> 10,000
D-8516	37.0	- 3 ppm/ $^\circ\text{C}$	> 10,000
D-8517	36.8	+ 9 ppm/ $^\circ\text{C}$	> 10,000

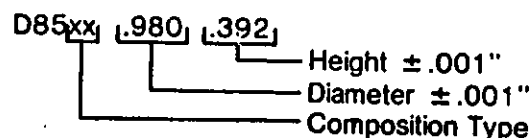
Frequency Range of Operation: 2 GHz to 45 GHz.

Temperature Range of Operation: -55 $^\circ\text{C}$ to +100 $^\circ\text{C}$.

Water Absorption: <0.10% As Measured by ASTM C373-56 Boiling Water Test.

D-8500 Series Standard Parts	Nominal Freq. of Use $f_o \pm \text{Approx. } 5\%$ (GHz)
D-85xx .980.392	2.2
D-85xx .875.350	2.6
D-85xx .750.300	3.0
D-85xx .625.250	3.6
D-85xx .500.200	4.5
D-85xx .375.150	6.0
D-85xx .312.124	7.2
D-85xx .250.100	9.0
D-85xx .220.088	10.2
D-85xx .187.074	12.0
D-85xx .165.066	13.3
D-85xx .125.050	17.9
D-85xx .1125.0450	20
D-85xx .0900.0360	25
D-85xx .0804.0321	28
D-85xx .0703.0281	32
D-85xx .0625.0250	36
D-85xx .0500.0200	45

Standard Parts Key



Note: For parts operating above 17.9 GHz, tolerance is held to $\pm .0005$ "

In addition to the standard parts listed above, the D-8500 series materials can be produced in any machined shape you may require, including tubes and substrates. Please contact your Alpha/Trans-Tech representative, or call the factory for commercial details.

Specifications subject to change without notice.
 Unless otherwise indicated, all data is nominal.

June 6, 1983