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KU-BAND MINIATURE MODULATORS

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16. Abstract

Ku-Band Microminiature Modulators have been designed to convert a 10 mW signal at 400 MHz to a 10 mW signal at 15 GHz. The designs incorporate gallium arsenide Schottky barrier varactors used in upper-sideband up-converters and include the use of Ku-Band microstrip circulators and hairpin resonator bandpass filters at 2.1 GHz and 2.5 GHz. The evaluation compares the design and fabrication of a single up-conversion unit with a double up-conversion unit. Various filter configurations are studies, and the use of both alumina and quartz substrates are considered. The various impedance matching networks are evaluated using computer aided design techniques.

The output of a double conversion unit fabricated on alumina substrates is about +3.4 dBm, while the output of a single-conversion unit fabricated on a quartz substrate is about +8.5 dBm. Recommendations for the selection of frequencies and the choice of substrates for optimum up-converter performance are given.

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PREFACE

The objective of this program is the design, development, and fabrication of five models of a Ku-Band Microminiature Modulator which converts a signal at 400 MHZ to a signal at 15 GHZ using state-of-the-art microwave integrated circuit (MIC) techniques. A two-stage up-converter requiring pumps at 2.190 GHZ and 12.410 GHZ was developed first in conformance with the original goals. Power output and bandwidth performance were low. Based upon the technology learned during the design phase of this unit, a new design utilizing a single pump at 14.6 GHZ and having one-fifth the size and weight of the two stage unit was proposed. This unit was then developed and fabricated, and produced 5 dB greater output and much greater bandwidth than the original two-stage unit. The delivered units consisted of two two-stage units having a midband output of about +3.4 dBm and three single-stage units having an output of +8.5 dBm.

It is recommended that upper sideband up-converters having an output at Ku-band frequencies utilize 10 mil thick quartz substrates. A high ratio of output-to-input frequencies is very desirable to simplify circuit design and layout.

TABLE OF CONTENTS

LIST OF ILLUSTRATIONS

LIST OF TABLES

Section 1

INTRODUCTION

The object of this program is the development of five units of a Ku-Band Microminiature Modulator. The original specifications called for the following items:

a. Upper sideband up-converter, complete with low-pass input filter and bandpass output filter, and capable of converting a 400 MHz ±50 MHz IF signal to a 2,500 MHz ±50 MHz signal with a minimum loss in the conversion process. Carrier and image signals will be suppressed to at least 30 dB below the 2,500 MHz ±50 MHz signal. The up-converter will operate with the following power levels:

   10 milliwatts input power at 400 MHz ±50 MHz.

   50 milliwatts input power at 2,100 MHz.

   10 milliwatts output power at 2,500 MHz ±50 MHz.

b. Upper sideband up-converter, complete with low-pass input filter and bandpass output filter, and capable of converting the 2,500 MHz ±50 MHz signal from the first converter to a signal at a frequency of 15,000 MHz ±50 MHz. This second up-converter will operate with the following power levels:

   10 milliwatts input power at 2,500 MHz ±50 MHz.

   50 milliwatts input power at 12,500 MHz.

   10 milliwatts output power at 15,000 MHz ±50 MHz.

c. Microcircuit circulator operating at a center frequency of 12,500 MHz with sufficient bandwidth, extremely low insertion loss (≈0.25 dB) and a power handling capability well over the required 50 milliwatts.

d. Microcircuit circulator operating at a center frequency of 15,000 MHz with sufficient bandwidth, extremely low insertion loss (≈0.25 dB) and a power handling capability well over the required 100 milliwatts.

All above four elements of the modulator will be mounted within a single rectangular container whose maximum outside dimensions are 1 inch x 1.5 inch x 2 inches. OSM type coaxial fittings will be used for all input and output connections.

Both upper sideband up-converters will be constructed using gallium arsenide Schottky barrier type diodes. All elements will be of the microminiature type construction on sapphire substrate material.

These modules required the development of filters utilizing hairpin resonators to achieve the most efficient utilization of substrate area. At 2.5 GHz, the following performance of a two-section filter has been achieved:

Midband Insertion Loss:	1 dB
Midband VSWR:	1.2
0.5 dB bandwidth:	100 MHz
Rejection at 2.1 GHz:	27 dB
Area Required:	450 mils x 240 mils

All garnet microstrip circulators at Ku-band were developed. Over the band of 10.5 to 18 GHz, the insertion loss is less than 1 dB and the isolation greater than 12 dB. The substrate is 255 mils in diameter and 25 mils thick.

Designing a two stage up-converter as specified resulted in a circuit which is very complicated and critical. After tuning, an output of +3.4 dBM and a 3 dB bandwidth of about 100 MHz was achieved.

A new design was developed utilizing a single-stage up-converter fabricated on a 1 inch by $\frac{1}{2}$ inch quartz substrate. This design achieved an output of +8.5 dBM and had a 1 dB bandwidth of greater than 200 MHz. This circuit required little tuning, was simple, and not critical.

Section 2

VARACTOR DIODE SELECTION AND CHARACTERIZATION

The parameters of the varactor diodes chosen for the two frequency converters must permit operation at high efficiency for the power levels and frequencies involved. Once the diodes have been selected, they must be properly characterized to determine the operating impedances. This information is then used to design the up-converter circuits.

2.1 DIODE PARAMETER SELECTION

To satisify the requirements of the parametric up-converters, the diodes must be chosen according to the following criteria:

a. The diode must have low loss. A suitable measure of diode loss is the cut-off frequency, f_c. At any given voltage, V, the cut-off frequency is related to the diode series resistance, R_s, and junction capacitance, C_j, by the expression

$$f_c = \frac{1}{2\pi R_s C_j} \tag{1}$$

Many diode manufacturers have standardized upon a bias of 6 volts in specifying cut-off frequency. If the cut-off frequency is high relative to the operating frequencies, low loss is expected. Equation 1 shows that this quantity is raised by decreasing R_s and C_j.

b. The diode must present reasonable impedances to the circuit to facilitate matching, especially if broad bandwidth is required. Since input and output impedance levels of MIC's are typically 50 ohms, it is perferred to have the diode present real and reactive impedances in the range of 5 ohms to 500 ohms. Using the subscripts s, p, and u to respectively denote the signal, pump, and upper sideband frequencies, the input and load resistances which should be presented to the varactor are given by (Ref. 1):

$$R_{in_s} = R_s + R_s \left(f_c/f_s\right)\left(m_p\, m_u/m_s\right) \tag{2}$$

$$R_{in_p} = R_s + R_s \left(f_c/f_p\right)\left(m_s\, m_u/m_p\right) \tag{3}$$

$$R_u = -R_s + R_s \left(f_c/f_u\right)\left(m_s\, m_p/m_u\right) \tag{4}$$

The terms m_s, m_p, and m_u are the diode modulation ratios, and are a measure of the capacitance swing resulting from the various voltage components. These expressions show that R_s and f_c must be properly chosen to produce realistic impedance levels.

The reactance of the diode is given by the familiar relation:

$$X_n = 1/2\pi f_n C_j \tag{5}$$

where the subscript n can be either s, p, or u. This shows that selecting too low a capacitance level can result in too great a reactance, especially at the lowest frequency, f_s.

c. The diode must have sufficient power handling capability. This capability is proportional to the term

$$P_{norm} = \frac{(\phi + V_B)^2}{R_s} \tag{6}$$

where P_{norm} is called the normalization power, ϕ is the diode contact potential, and V_B is the diode breakdown voltage. In an up-converter, the powers supplied to the diode, P_s and P_p, and the power delivered to the output, P_u, are given by:

$$P_s = 8\,P_{norm}\left(\frac{f_s}{f_c}\right)^2\left(\frac{f_c}{f_s}\,m_s\,m_p\,m_u + m_s^2\right) \tag{7}$$

$$P_p = 8\,P_{norm}\left(\frac{f_p}{f_c}\right)^2\left(\frac{f_c}{f_p}\,m_s\,m_p\,m_u + m_p^2\right) \tag{8}$$

$$P_u = 8\,P_{norm}\left(\frac{f_u}{f_c}\right)^2\left(\frac{f_c}{f_u}\,m_s\,m_p\,m_u - m_u^2\right) \tag{9}$$

It can be seen that power handling capability for a given R_s is proportional to the square of the breakdown voltage (assume $\phi \ll V_B$). Hence greater power can be handled in a diode with greater breadown voltage.

The selection of the diode parameters R_s, C_j, and V_B cannot be arbitrarily made since all three parameters are dependent upon the diode doping density. In general, as the doping density is increased, R_s decreases, C_j increases, and V_B decreases. Therefore, the choice of any diode must be based upon its predicted performance in an upconverter circuit.

2.2 PREDICTED CIRCUIT PERFORMANCE

The proper selection of a varactor diode is verified by checking its predicted circuit performance. This means that once a diode having a specified R_s, C_j, and V_B are chosen, the predicted values for the impedances and power levels based upon the equations given above, must be checked for reasonable circuit performance.

Circuit performance is also determined by the chosen values of the diode modulation ratios, m_s, m_p, and m_u. These must be selected to give optimum values for power levels, and overall efficiency. The three ratios are limited by the approximation:

$$m_s + m_p + m_u \leq 0.25 \tag{10}$$

For a given selected set of values for the modulation ratios, the impedances presented to the diode are determined from Equations 2, 3, 4, and 5.

2.3 UP-CONVERTER OPTIMIZATIONS

The up-converter can be optimized by selecting the values of m_s, m_p, and m_u properly. For example, the up-converter can be optimized for: a. maximum pump efficiency; b. maximum total efficiency; or c. maximum power output. These conditions will now be considered separately.

2.3.1 Maximum Pump Efficiency

Maximum pump efficiency is achieved by maximizing the ratio:

$$n_p = \frac{P_u}{P_p} . \tag{11}$$

Mathematically, the solution for this case becomes the trivial condition

$$m_s = 0.25, \quad m_p = 0 ; \; m_u = 0. \tag{12}$$

Physically this means that the diode is being pumped by the signal, and that the pump and output levels are now small signals. The pump efficiency for this case is given by:

$$n_p = \frac{\left(\frac{m_s f_c}{f_p}\right)^2}{\left(1 + \sqrt{1 + \frac{m_s^2 f_c^2}{f_p f_u}}\right)^2} . \tag{13}$$

Total efficiency, pump power, and output power are all essentially zero for this case.

2.3.2 Maximum Total Efficiency

Maximum total efficiency is achieved by maximizing the ratio:

$$n_t = \frac{P_u}{P_s + P_p} . \tag{14}$$

A condition which approximates this goal is given by (Ref 2)

$$w_s m_s^2 = w_p m_p^2 = w_u m_u^2 \tag{15}$$

Combining this with Equation 10 results in the following expressions:

$$m_s = \frac{0.25}{1 + \left(\frac{f_s}{f_p}\right)^{\frac{1}{2}} + \left(\frac{f_s}{f_u}\right)^{\frac{1}{2}}} \tag{16}$$

$$m_p = m_s \sqrt{\frac{f_s}{f_p}} \tag{17}$$

$$m_u = m_s \sqrt{\frac{f_s}{f_u}} . \tag{18}$$

2.3.3 Maximum Output Power

The values for the modulation ratios which maximize the output power, given by Equation 9, are given by the expressions:

$$m_s = \frac{1}{24} - \frac{2}{3}\frac{f_u}{f_c} + \sqrt{\frac{4}{9}\left(\frac{f_u}{f_c}\right)^2 + \frac{1}{9}\left(\frac{f_u}{f_c}\right) + \frac{1}{576}} \qquad (19)$$

$$m_p = m_s \qquad (20)$$

$$m_u = \frac{m_s^{2}}{m_1 + 2\frac{f_u}{f_c}} . \qquad (21)$$

2.4 COMPARISON OF OPTIMIZATIONS

These various optimizations have been programmed on a time-shared computer to facilitate comparison. Commercially available Gallium arsenide Schottky barrier diodes have been selected to provide input parameters for the various calculations. The parameters of the selected diodes are summarized in Table 2-1.

Table 2-1. Varactor Diode Parameters

Application:	First Up-converter	Second Up-converter
Manufacturer:	Parametric Industries	Parametric Industries
Model No.:	PF 4500 C	PF 4500 G
Cut-off Frequency (-6V):	300 GHz	500 GHz
C_j (-6V):	0.2 - 0.45 pF	0.3 - 0.45 pF
Breakdown Voltage:	-6 V	-6 V

The predicted performances for the various optimizations are shown in Table 2-2 for the first up-converter and in Table 2-3 for the second up-converter.

Table 2-2. Comparison of First Up-Converter Optimizations

(Diode: Parametric Industries PF 4500C)

Optimization	Max. Pump Eff.	Max. Total Eff.	Max. Power Out.
m_s	0.250	0.136	0.088
m_p	0	0.059	0.088
m_u	0	0.055	0.074
Gain (P_u/P_s)	0	7.47 dB	7.52 dB
Pump Eff.	112.0	106.4%	100.1%
Total Eff.	0	89.4%	85.0%
Signal Power	0.03 mW	0.18 mW	0.22 mW
Pump Power	0	0.93 mW	1.25 mW
Output Power	0	0.98 mW	1.25 mW
R_{in_s}	1.26 Ω	23.8 Ω	71.4 Ω
R_{in_p}	41.4 Ω	23.8 Ω	14.6 Ω
R_u	41.4 Ω	21.3 Ω	14.6 Ω
X_{c_s}	←	473.7 Ω	→
X_{c_p}	←	90.2 Ω	→
X_{c_u}	←	75.8 Ω	→

Table 2-3. Comparison of Second Up-Converter Optimizations

(Diode: Parametric Industries PF 4500G)

Optimization	Max. Pump Eff.	Max. Total Eff.	Max. Power Out.
m_s	0.250	0.135	0.096
m_p	0	0.060	0.096
m_u	0	0.055	0.058
Gain (P_u/P_s)	0	5.99 dB	6.50 dB
Pump Eff.	96.4%	79.5%	68.0%
Total Eff.	0	66.2%	59.0%
Signal Power	0.39 mW	0.67 mW	0.72 mW
Pump Power	0	3.33 mW	4.75 mW
Output Power	0	2.65 mW	3.23 mW
R_{in_s}	1.44 Ω	8.6 Ω	18.5 Ω
R_{in_p}	13.3 Ω	8.6 Ω	4.8 Ω
R_u	13.3 Ω	5.7 Ω	6.1 Ω
X_{c_s}	←	144.7 Ω	→
X_{c_p}	←	28.9 Ω	→
X_{c_u}	←	24.1 Ω	→

In comparing data in Tables 2-2 and 2-3, it must be realized that the varactors are usually slightly overdriven and that actual power levels can be as much as ten times greater than the levels indicated. These tables show clearly that the solution of maximum pump efficiency is not practical. The other two solutions are comparable. The maximum power out condition is used for the designs in this program. This condition results in greatest output power with a small loss in total efficiency and more practical resistance levels.

2.5 DIODE PACKAGE TRANSFORMATION

The impedance of the varactor diode, determined by the maximum output power optimization, is modified by the package parasitic inductance and capacitance as shown in Figure 2-1. The impedances resulting from this transformation are given in Table 2-4 for the first up-converter and in Table 2-5 for the second up-converter.

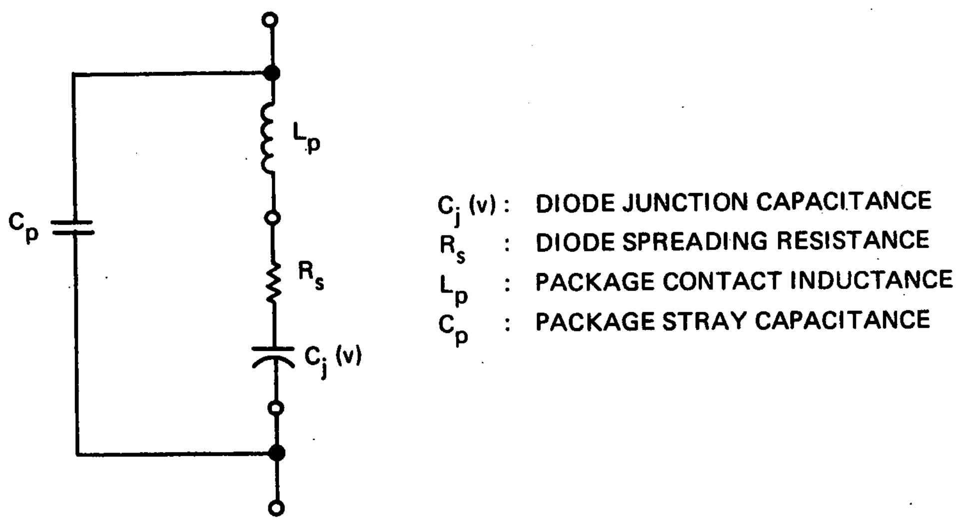

Figure 2-1. Equivalent Circuit of a Packaged Varactor Diode

These impedance values are the ones to be used in designing the up-converter circuitry. Sufficient data has been tabulated to permit the determination of input and output VSWR's and resulting mismatch losses across the frequency bands of interest.

Table 2-4. Packaged Diode Impedances For The First Up-Converter

(C_v = 0.85pF, C_p = 0.15pF, L_p = 0.1 nH)

Frequency (GHz)	Z_{Total} (Ohms)	R_v (Ohms)
0.350	59.9-j 455.9	81.5
0.375	55.0-j 425.5	76.1
0.400	51.5-j 398.9	71.4
0.425	48.5-j 375.4	67.2
0.450	45.8-j 354.5	63.4
2.100	10.6-j 75.1	14.6
2.450	10.8-j 64.2	14.9
2.500	10.6-j 62.8	14.6
2.550	10.4-j 61.6	14.4

Table 2-5. Packaged Diode Impedances For the Second Up-Converter

(C_v = 0.45 pF, C_p = 0.15 pF, L_p = 0.1 nH)

Frequency (GHz)	Z_{Total} (Ohms)	R_v (Ohms)
2.45	10.6-j 107.7	18.8
2.50	10.4-j 105.6	18.5
2.55	10.2-j 103.4	18.1
12.50	3.1-j 16.6	4.9
14.95	4.2-j 12.2	6.1
15.00	4.2-j 12.1	6.1
15.05	4.2-j 12.0	6.0

Section 3

DESIGN OF AUXILIARY CIRCUITS

Many different circuits have been designed to meet the specifications of the Ku-band Microminiature Modulator. The design criterion of typical circuits approach will be briefly discussed together with a tabulation of predicted performance and a listing of advantages and disadvantages. The recommendations for the best way to satisfy system specifications will be described later.

All of the considered approaches require filter circuits and circulators. Suitable filter and circulator configurations with measured performance follow.

3.1 FILTER DESIGNS

3.1.1 2.1 GHz and 2.5 GHz Bandpass Filters

Bandpass filters normally incorporate half-wavelength coupled resonators. At 2.1 GHz, this corresponds to a length of approximately 1.1 inches. The overall filter designed in the conventional edge-coupled straight resonator approach would be almost 1-3/4 inches long. Because an excessive substrate area would be consumed by such a design, new filter configurations have been tried instead (Ref. 3, 4).

One configuration uses quarter-wavelength resonators which are short circuited at one end and open circuited at the other. Input and output lines are directly tapped into the resonators as shown in Figure 3-1. In principle, the impedance presented to the input and output can be controlled by the location of the tap-point. Lower impedances are presented by approaching the short circuit end, while higher impedances result as the tap is moved closer to the open-circuited end.

A filter of this type has been designed for 2.5 GHz, fabricated, and tested. The performance, plotted in Figure 3-2, can be summarized as follows:

Midband Insertion Loss	1.5 dB
Midband VSWR	1.6
0.5 dB bandwidth	110 MHz
Rejection at 2.1 GHz	25 dB
Area required	455 mils x 155 mils

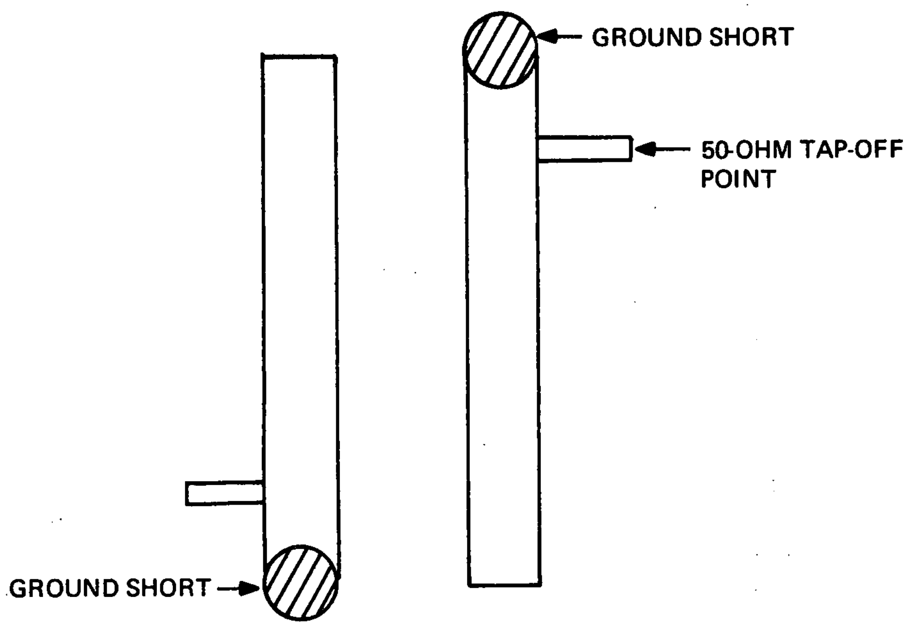

Figure 3-1. Tapped Bandpass Filter Configuration

This configuration has the advantages of giving good performance in a compact area. However, reproducible results are difficult to obtain because the location of the short circuit through the substrate is extremely critical even at this comparatively low frequency. Furthermore, the quality of the short circuit has a direct effect on the insertion loss of the filter.

An alternate band pass filter configuration recently described in the literature is the hairpin resonator filter. This filter uses resonators bent back upon themselves at their midpoints and coupled along half their length as shown in Figure 3-3. Filters of this type have been fabricated at 2.5 GHz. The performance is plotted in Figure 3-4 and is summarized as follows:

Midband Insertion Loss	1. dB
Midband VSWR	1.2
0.5 dB Bandwidth	100 MHz
Rejection at 2.1 GHz	27 dB
Area required	450 mils x 240 mils

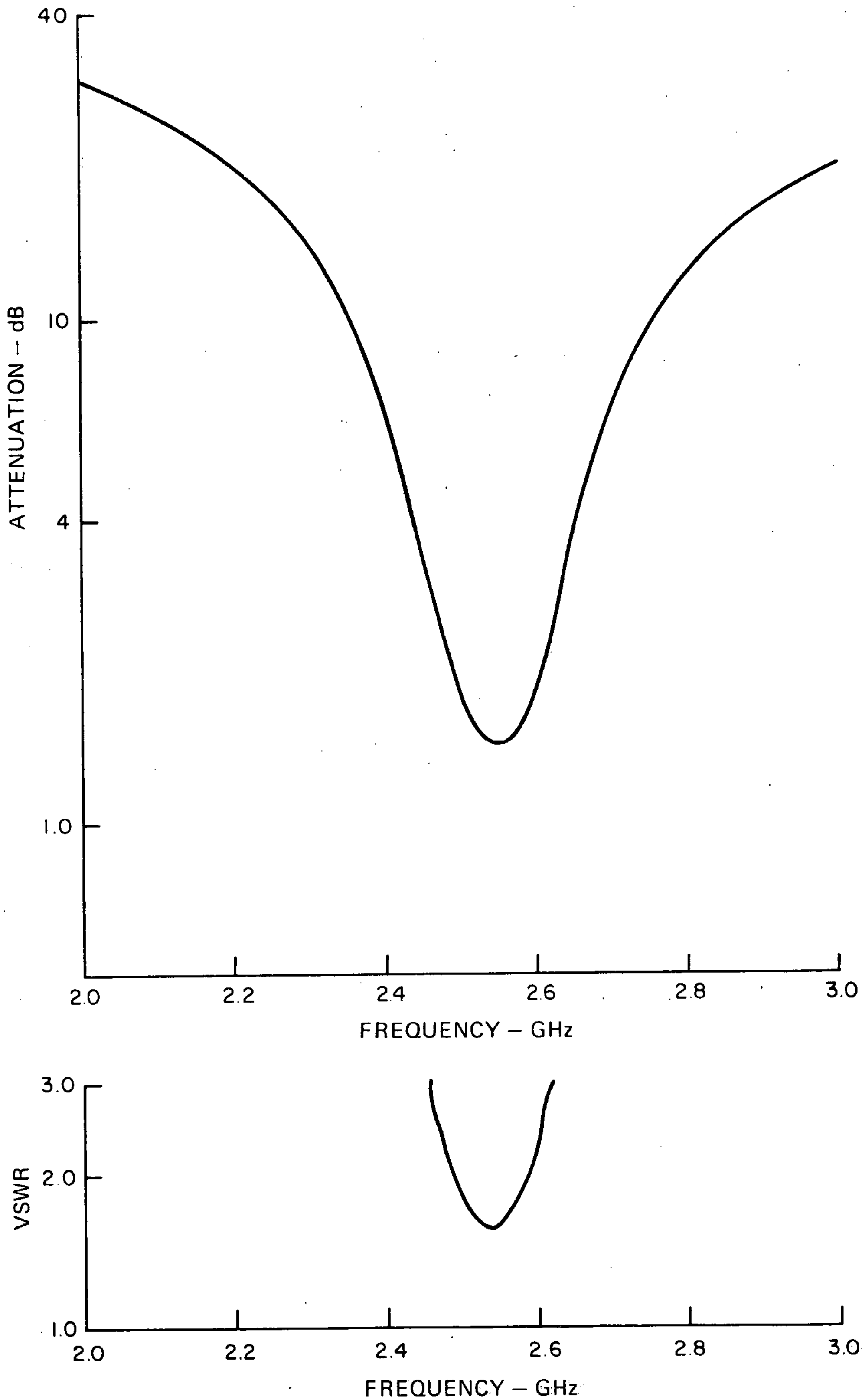

Figure 3-2. Performance of Tapped 2.5 GHz Bandpass Filter

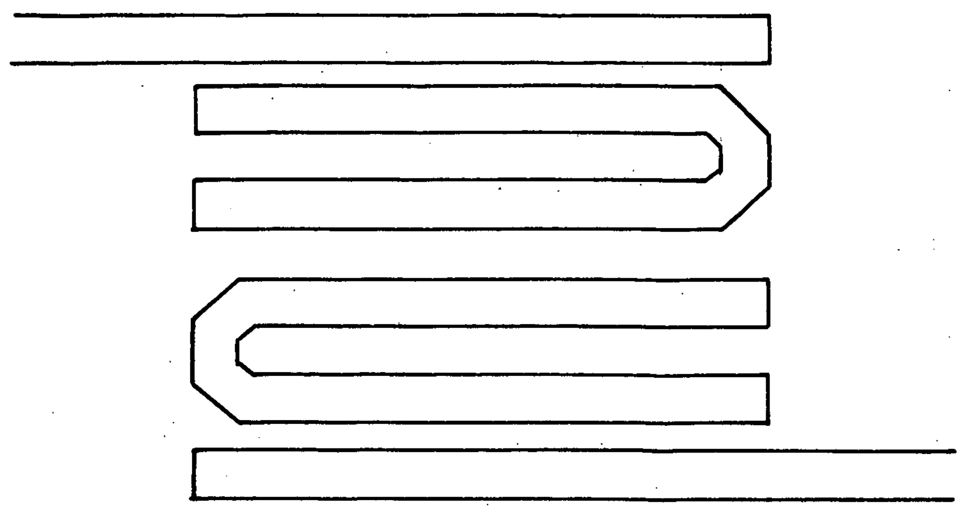

Figure 3-3. Hairpin Resonator Bandpass Filter Configuration

This type of filter has several advantages. The drilling of holes in the substrate is not needed and the resulting filter circuit is highly reproducible. Although the required substrate area is slightly larger, the improvement in filter performance justifies this increase.

Another filter configuration which might find application in frequency converter circuits is the directional filter. Basically, this filter utilizes a full wavelength resonator coupled into a four port configuration as shown in Figure 3-5. This structure has the property of a bandpass filter between ports 1 and 3 and of a band reject filter between ports 1 and 2. The measured performance of a filter designed to operate at 2.1 GHz is given in Figure 3-6 and is summarized as follows:

Ports	1-2	1-3
Midband Loss	11.6 dB	2.4 dB
VSWR	1.15	1.15
Loss at 2.5 GHz	0.1 dB	22.4 dB
Area required	600 mils x 380 mils	

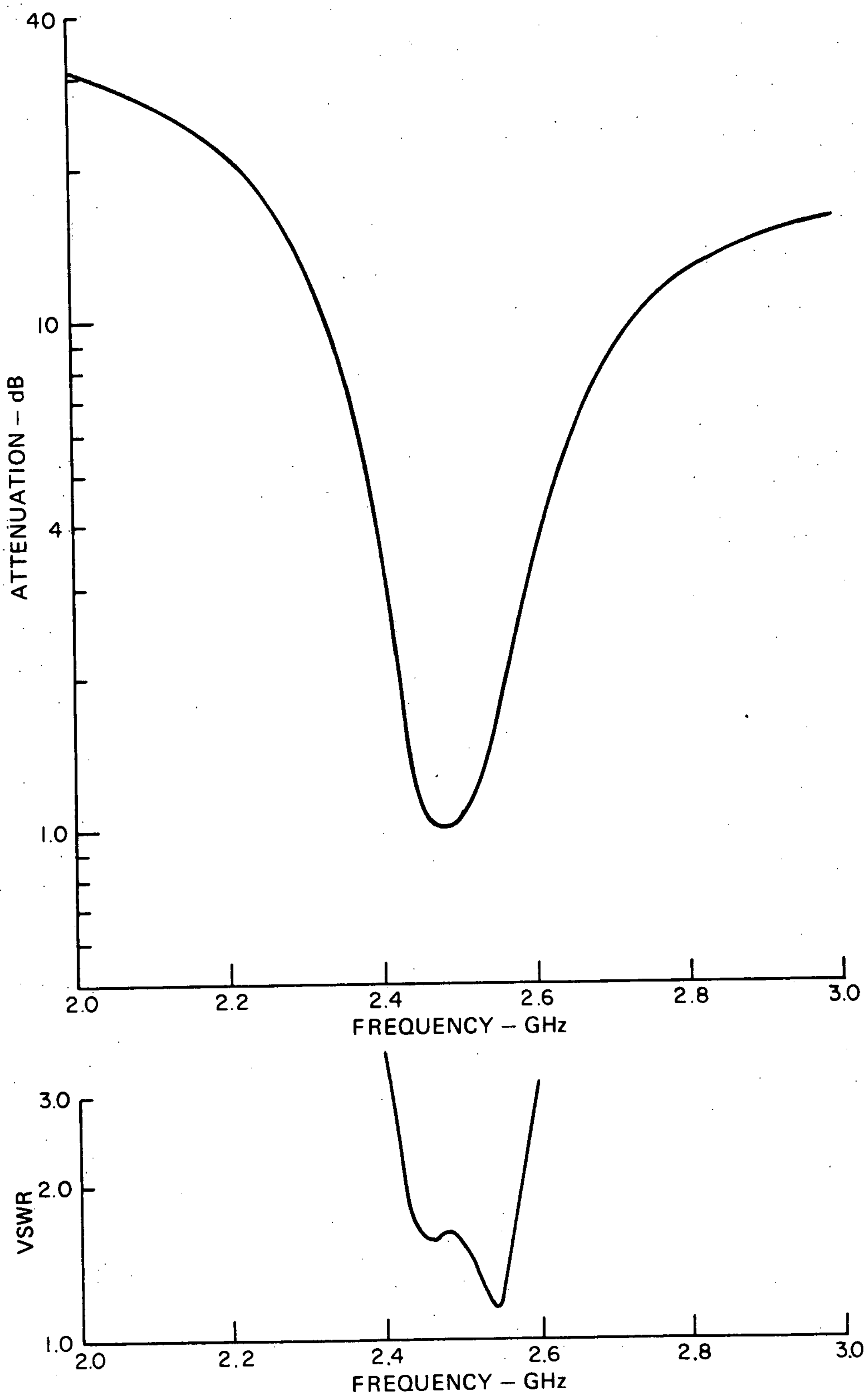

Figure 3-4. Performance of Hairpin Resonator 2.5 GHz Bandpass Filter

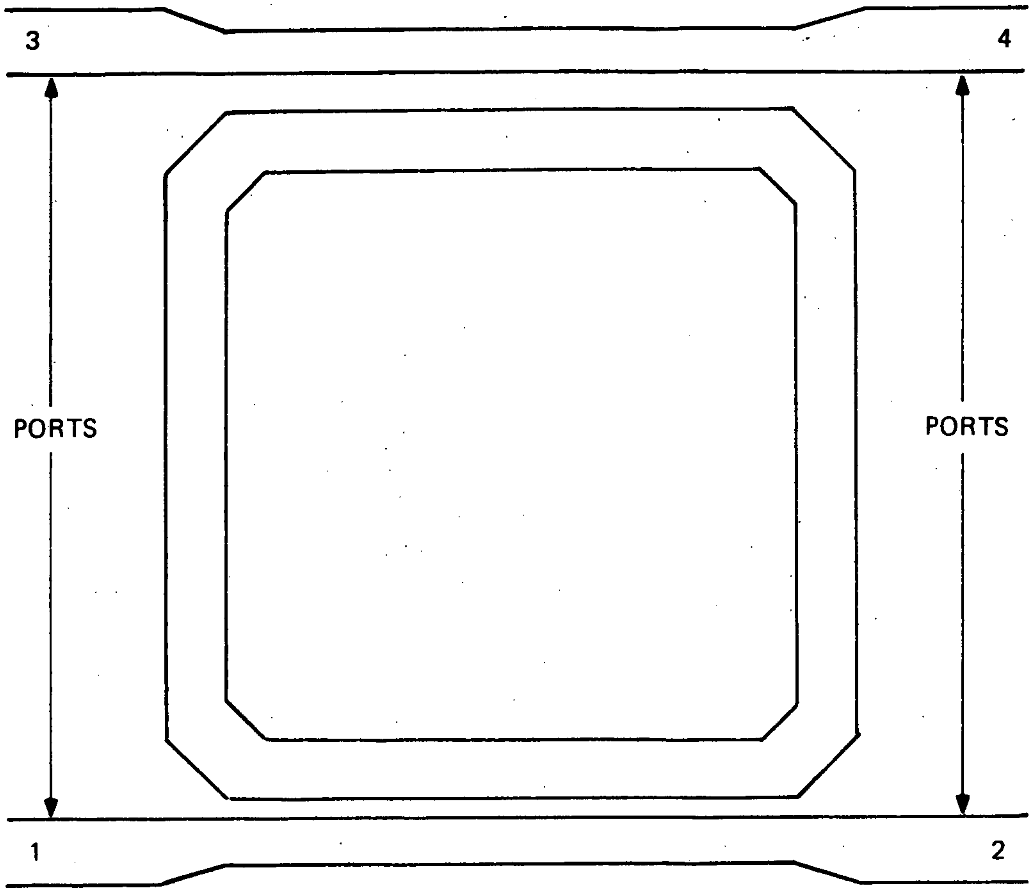

Figure 3-5. Directional Filter Configuration

3.1.2 Ku-Band Filters

For the second stage up-converter, the frequencies of interest (12.41 GHz and 15 GHz) are so high that filters designed by standardized techniques are very small. These bandpass filters are designed by using half wavelength resonators coupled along the edges as shown in Figure 3-7. The measured properties of these bandpass filters is summarized as follows:

Design Frequency	12.41 GHz	15 GHz
Insertion Loss at Design Frequency	1 dB	2 dB
Rejection (Frequency)	21 dB (15 GHz)	18 dB (12.41 GHz)
Area Required	300 mils x 70 mils	225 mils x 70 mils

3.2 KU-BAND CIRCULATORS

Ferrite microstrip circulators have been successfully designed and fabricated at RCA over the frequency range of 2 to 12 GHz. This program requires circulators at 12.5 GHz and 15 GHz. It has been decided to satisfy both requirements with a single

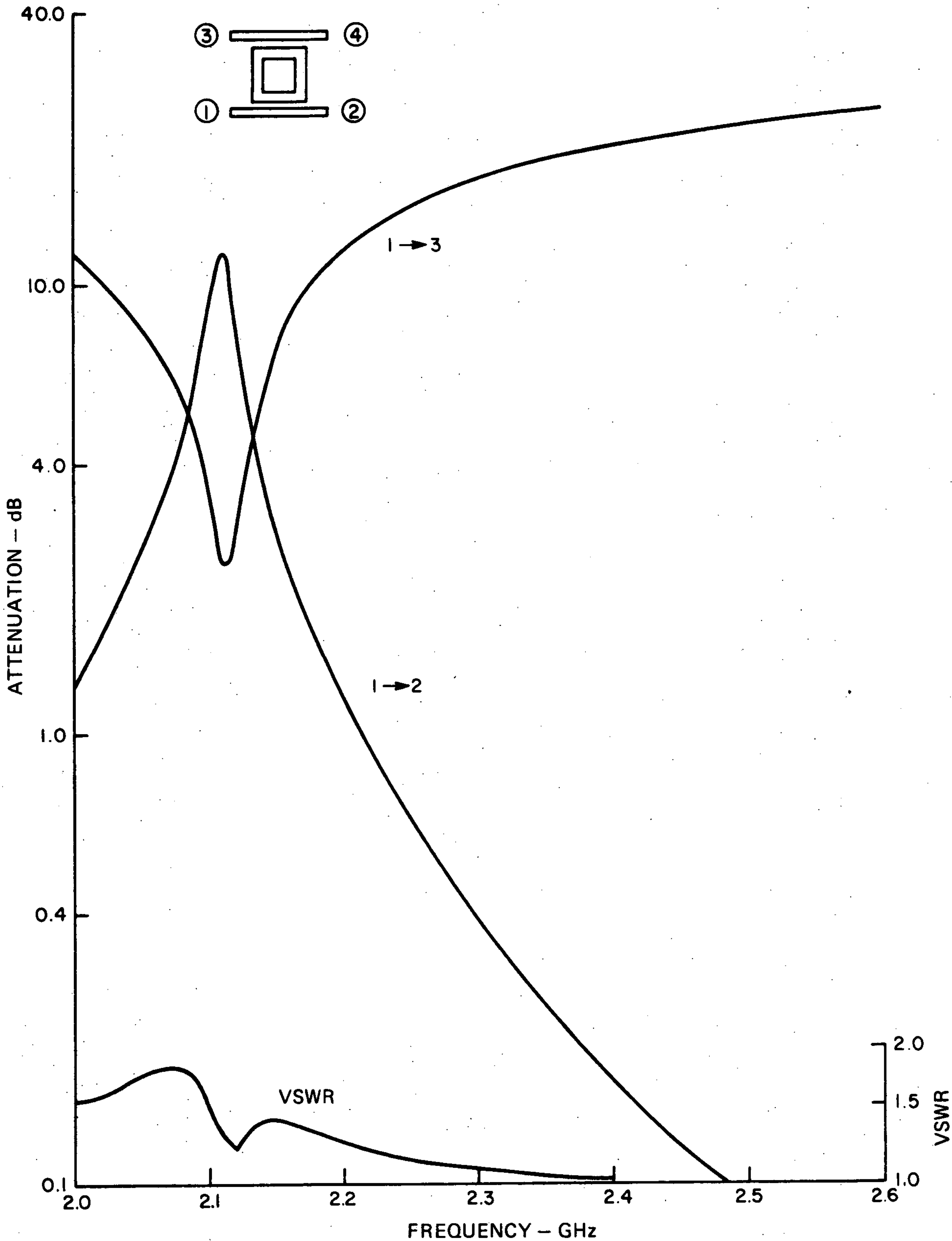

Figure 3-6. Performance of 2.1 GHz Directional Filter

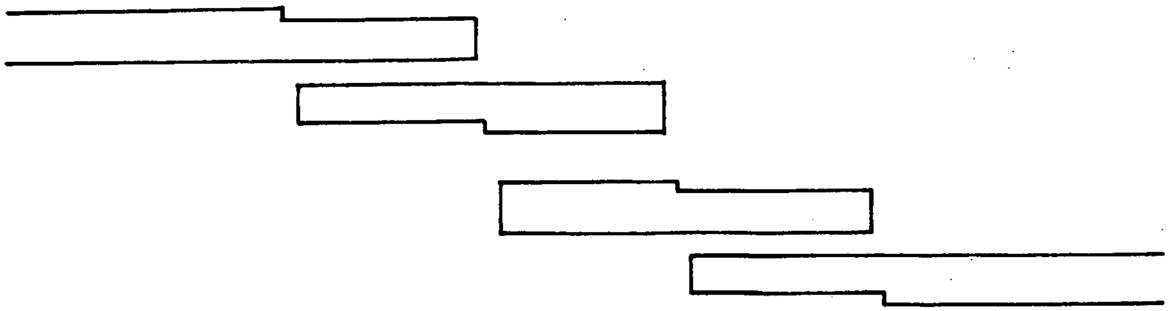

Figure 3-7. Edge Coupled Bandpass Filter Configuration

circulator having sufficient bandwidth for both frequencies. By scaling a successful low frequency design (Ref. 5), a circulator having a ferrite diameter of 255 mils and a thickness of 25 mils has been designed and tested. The measured performance of this unit is shown in Figure 3-8. The circulator covers the band of 10.5 to 18 GHz to the 1-dB insertion loss points and the 12-dB isolation points. It can be seen that this performance is more than adequate for this program.

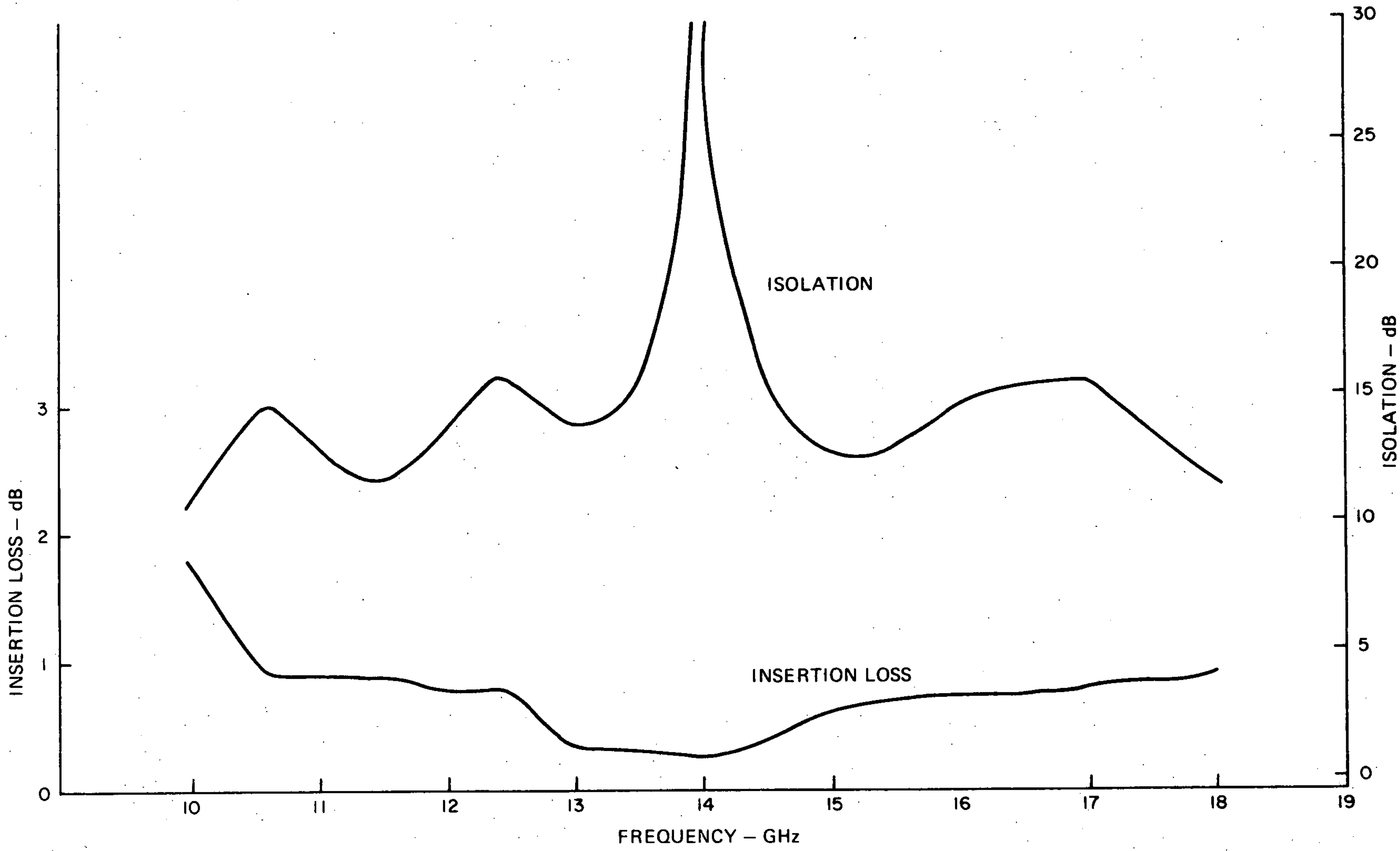

Figure 3-8. Ferrite Circulator Performance

Section 4

UP-CONVERTER CIRCUIT DESIGNS

Any up-converter circuit must be designed to achieve two goals: 1. present the desired impedances to the varactor diode at the frequencies of interest, and 2. separate the various frequency components to different connectors. Circuits designed to achieve these criteria are presented here.

4.1 EARLY TWO-STAGE DESIGNS

One of the first approaches to up-converter design uses two transmission lines branching from the diode. One line brings the signal component to the diode while the other line serves the dual function of bringing in the pump and extracting the upper sideband output.

The first designs use series mounted diodes. This results in simpler fabrication techniques because no holes are required through the substrate.

The design of the low frequency circuit using two 1-inch by 1-inch substrates is shown in Figure 4-1. The circuit to the left of the diode must look like a tuning inductance at the pump and output frequencies while presenting a reasonable match across the 350 to 450 MHz frequency band. The circuit to the right side of the diode must tune out the reactive part of the diode impedance at the 400 MHz input frequency while presenting the desired transformed real part impedance at the pump and output frequencies.

This circuit has been analyzed in detail by computer. Table 4-1 summarizes the performance of the matching networks. The meanings of the terminology "generator impedance", "input impedance", "output impedance", and "load impedance" is illustrated in Figure 4-2. The VSWR and resulting mismatch loss are shown at the frequencies of significance in Table 4-1.

A study of this data provides insight into the difficulty of designing the up-converter. The following problems become apparent: 1. Matching across the 25% bandwidth of the input frequency range is virtually impossible for the reactive device; 2. The loss of microstrip lines can result in a very large real part impedance component and, in turn, greatly change the impedance looking into the diode from the other matching network; and 3. The pump frequency and the output frequency are too far apart to match with a common impedance transformer network.

The same design technique has also been used for the second stage up-converter. It is hoped that because the percentage bandwidth is smaller in this case, namely 4% of the input frequency range, that this technique might succeed. The circuit is shown in Figure 4-3 and the computer analysis results shown in Table 4-2. By studying this Table, it is seen that the same difficulties are encountered in this circuit as in the low frequency design.

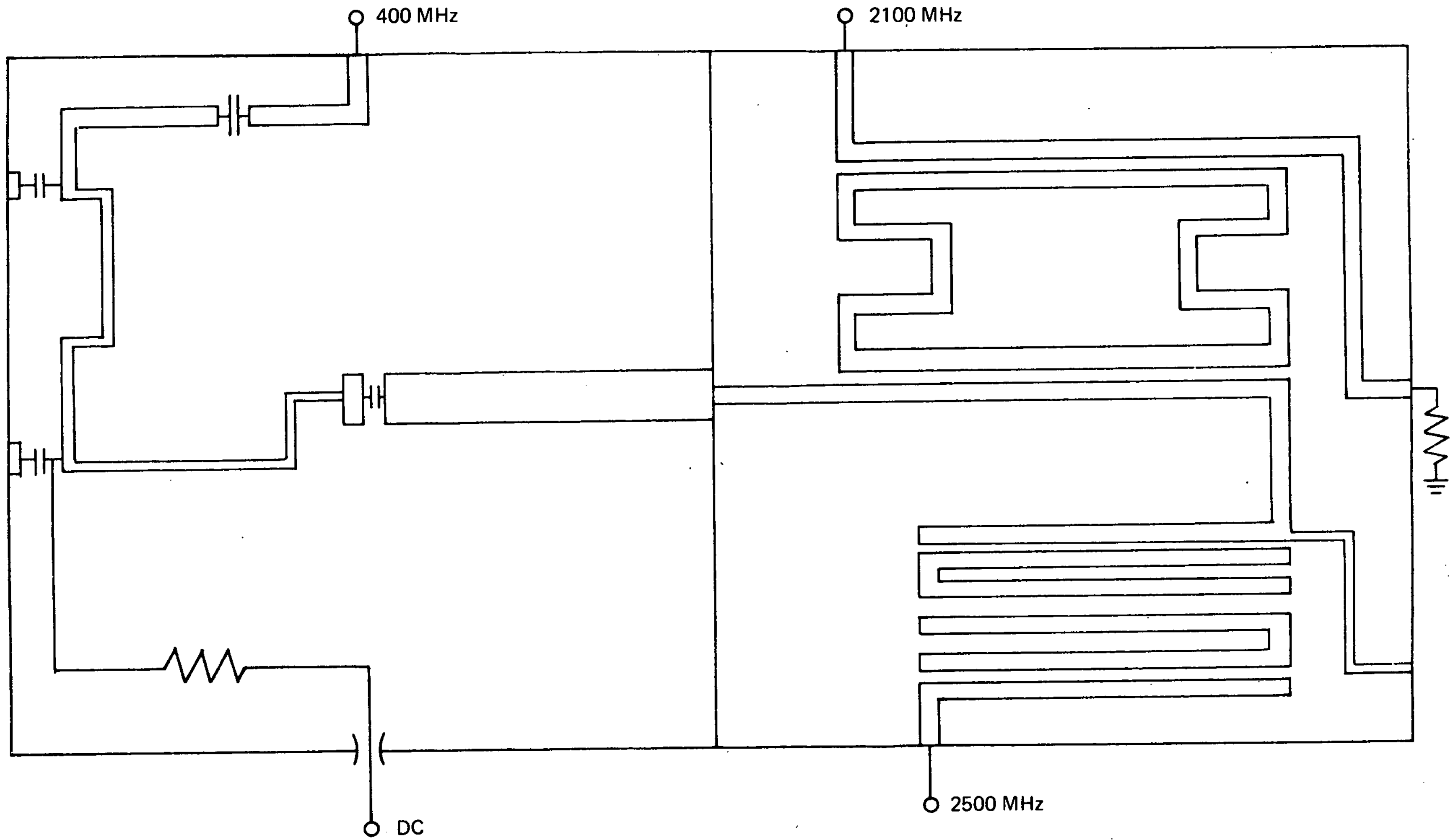

Figure 4-1. Layout of First Low Frequency Up-Converter

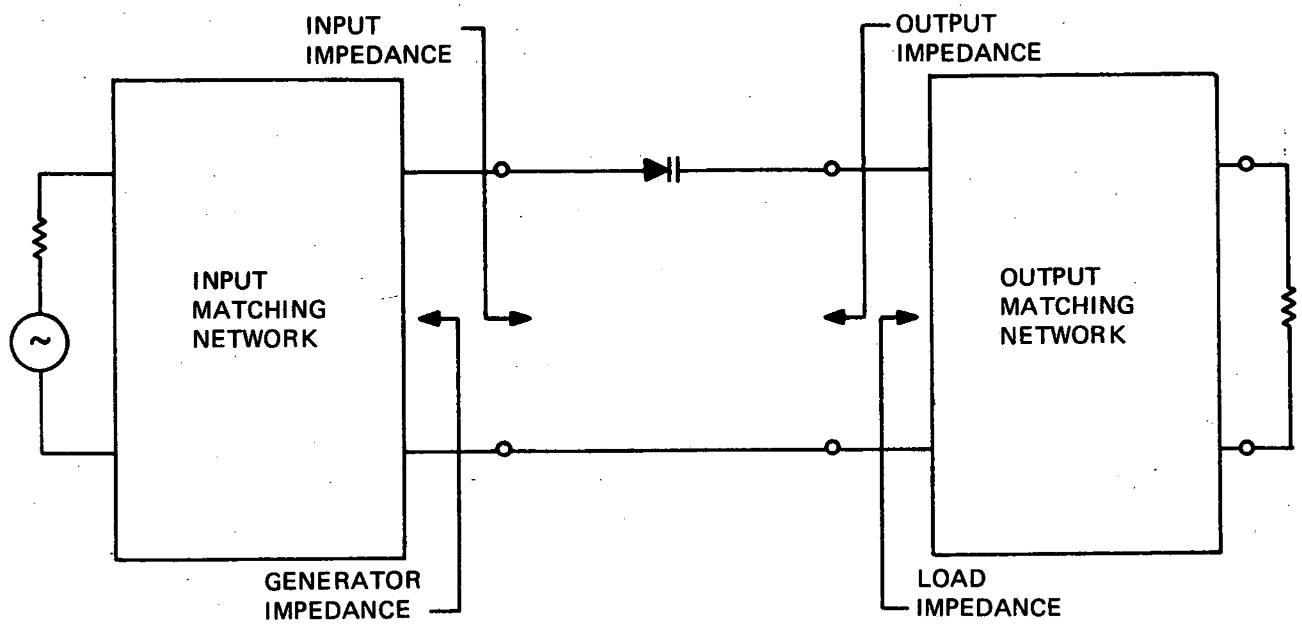

Figure 4-2. Illustration of Impedance Terminology

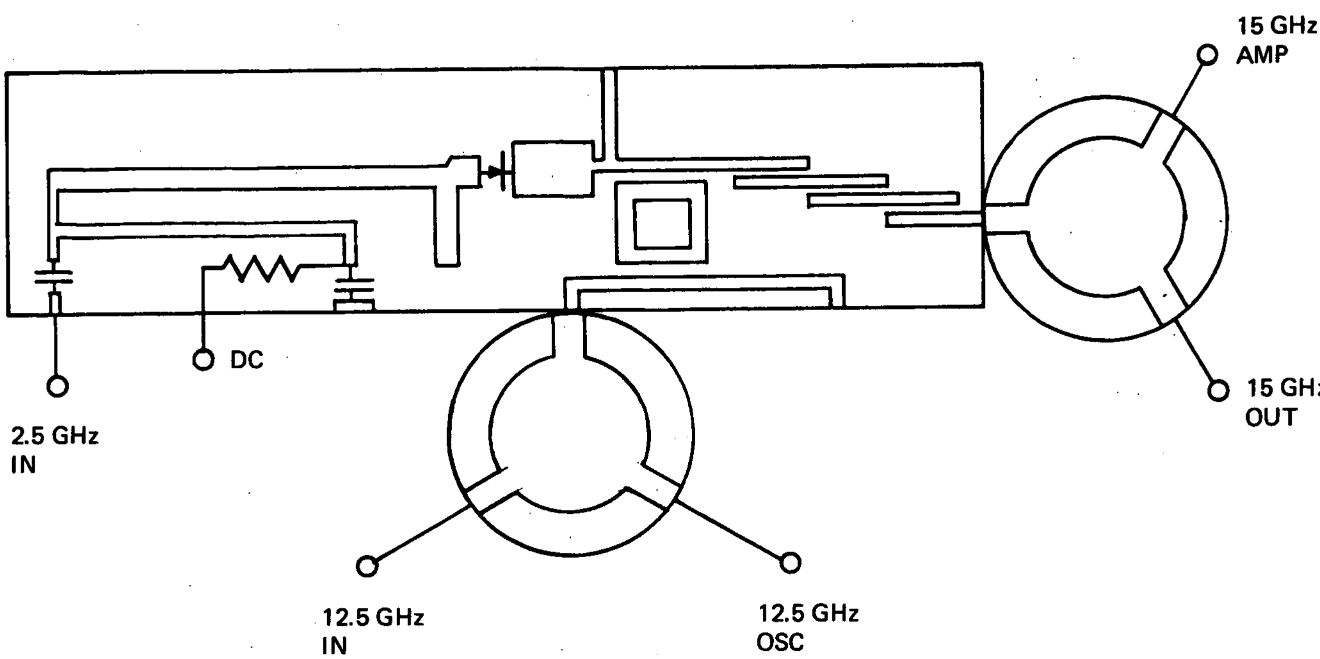

Figure 4-3. Layout of First High Frequency Up-Converter

Table 4-1. Computed Matching Network Performance for First Low Frequency Up-Converter Design

Frequency GHz	Generator Impedance (Ohms) Goal	Generator Impedance (Ohms) Calculated	Input Impedance Ohms	VSWR	Mismatch Loss (dB)
0.350	17.5 + j0	19.5 - j4.26	24.8 - j499	528	21
0.400	17.5 + j0	18.7 - j0.9	111 - j24	6.2	3.2
0.450	17.5 + j0	19.0 + j2.2	58 - j865	676	22
2.1	-	0.8 + j80.9	34 - j111		
2.3	0 + j97	0.9 + j97.6	30 - j98		
2.45	-	1.2 + j112.5	27 - j89		
2.50	-	1.3 + j118.1	27 - j86		
2.55	-	1.4 + j124.1	28 - j83		
	Load Impedance (Ohms) Goal	**Load Impedance (Ohms) Calculated**	**Output Impedance Ohms**		
0.350		7.3 + j150	36.9 - j654		
0.400	0 + j567	93.6 + j543	36.2 - j569		
0.450		40.6 - j361	36.5 - j503		
2.1	17.5 + j0	16.8 - j3.7	18.2 - j26.0	4.67	2.4
2.3	15 + j0	15.1 - j0.6	15.9 + j0.2	1.06	0
2.45	12.5 + j0	15.0 + j2.2	13.7 + j21.0	4.45	2.2
2.50	12.5 + j0	15.2 + j3.1	13.7 + j28.7	6.67	3.4
2.55	12,5 + j0	15.5 + J3.9	13.8 + j36.5	9.48	4.6

Table 4-2. Computed Matching Network Performance for First High Frequency Up-Converter Design

Frequency GHz	Generator Impedance Goal	Generator Impedance Calculated	Input Impedance	VSWR	Mismatch Loss (dB)
2.45	10.7 + j0	10.9 - j0.6	18.8 - j18.8	3.9	1.9
2.5	10.7 + j0	11.2 + j0.1	22.4 + j8.7	2.4	0.8
2.55	10.7 + j0	11.4 + j0.9	28.9 + j47.9	10.0	4.8
12.5		1.1 + j11.4	17.6 - J23.0		
13.75	0 + j17.1	0.2 + j18.9	14.3 - j16.7		
14.95		11.9 + j37.1	11.5 - j11.4		
15.0		18.3 + j26.7	11.5 - j11.2		
15.05		9.6 + j20.9	11.5 - j11.0		
	Load Impedance Goal	**Load Impedance Calculated**	**Output Impedance**	**VSWR**	**Mismatch Loss (dB)**
2.45		8.1 + j124	21.7 - j143.4		
2.5	0 + j140	1.7 + j149	21.9 - j139.8		
2.55		18.2 + j185	22.1 - j136.2		
12.5	10.7 + j0	6.9 - j2.5	11.8 - j8.9	3.7	1.7
13.75	7.1 + j0	7.2 + j0.3	7.3 + j1.9	1.3	0.1
14.97	3.5 + j0	7.9 + j2.8	15.4 + j22.9	7.7	3.9
15.0	3.5 + j0	8.0 + j2.9	21.9 + j12.5	4.2	2.1
15.05	3.5 + j0	8.0 + j3.0	13.0 + j6.8	2.8	1.1

Several variations of these circuits, some incorporating discrete components, have been designed and analyzed. In each instance, the predicted performance is poor and the bandwidth very deficient. As a result, designs using a common matching network for both pump and output frequencies have been abandoned. Instead, the simplicity of circuits using two branches from the diode is sacrificed for the complexity of three branch networks.

4.2 FINAL TWO-STAGE DESIGNS

The use of three branch networks requires the use of shunt mounted diodes in which one side of the diode is grounded. Each branch of the network is then designed to present the desired impedance at one of the three frequency bands while looking like an open circuit at the two out of band frequencies.

From the insight gained from the first set of designs, and from a search of published articles, especially Reference 6, it became apparent that it is not realistic to try to conjugately match the first diode over the 25% bandwidth at 400 MHz. Reference 3 shows that broadband impedance transforming networks must incur high loss. A better method is to use a detuned circuit technique which produces a mismatch loss decreasing with increasing frequency to compensate for the anticipated variation of gain in the varactor diode. In Reference 6, this technique results in a bandwidth of 40% at the input frequency of 300 MHz.

The final design for the first stage up-converter is shown in Figure 4-4. Although the diode is in a shunt mounted configuration, it is not necessary to drill a hole through the substrate to reach ground. Instead the diode is mounted above the substrate and a chip capacitor is used for grounding to the circuit frame which encloses the substrates. The impedance matching performance of the three branches is summarized in Table 4-3.

From this tabulation, it can be seen that each of the three circuits does an adequate job of achieving high impedances at the correct frequencies and of matching the diode at the pump and output frequencies. When the overall circuit is analyzed by the computer, it is found that the mismatch loss at the input frequency is about 9 dB while the VSWR at 2.1 GHz and 2.5 GHz is better than 2:1.

The high frequency up-converter design uses the same design technique of three different networks branching from a common shunt mounted diode as shown in Figure 4-5. Each of the three networks is designed to match to one of the three bands of interest while presenting a high impedance to the other two bands. The computed performance of these networks is given in Table 4-4.

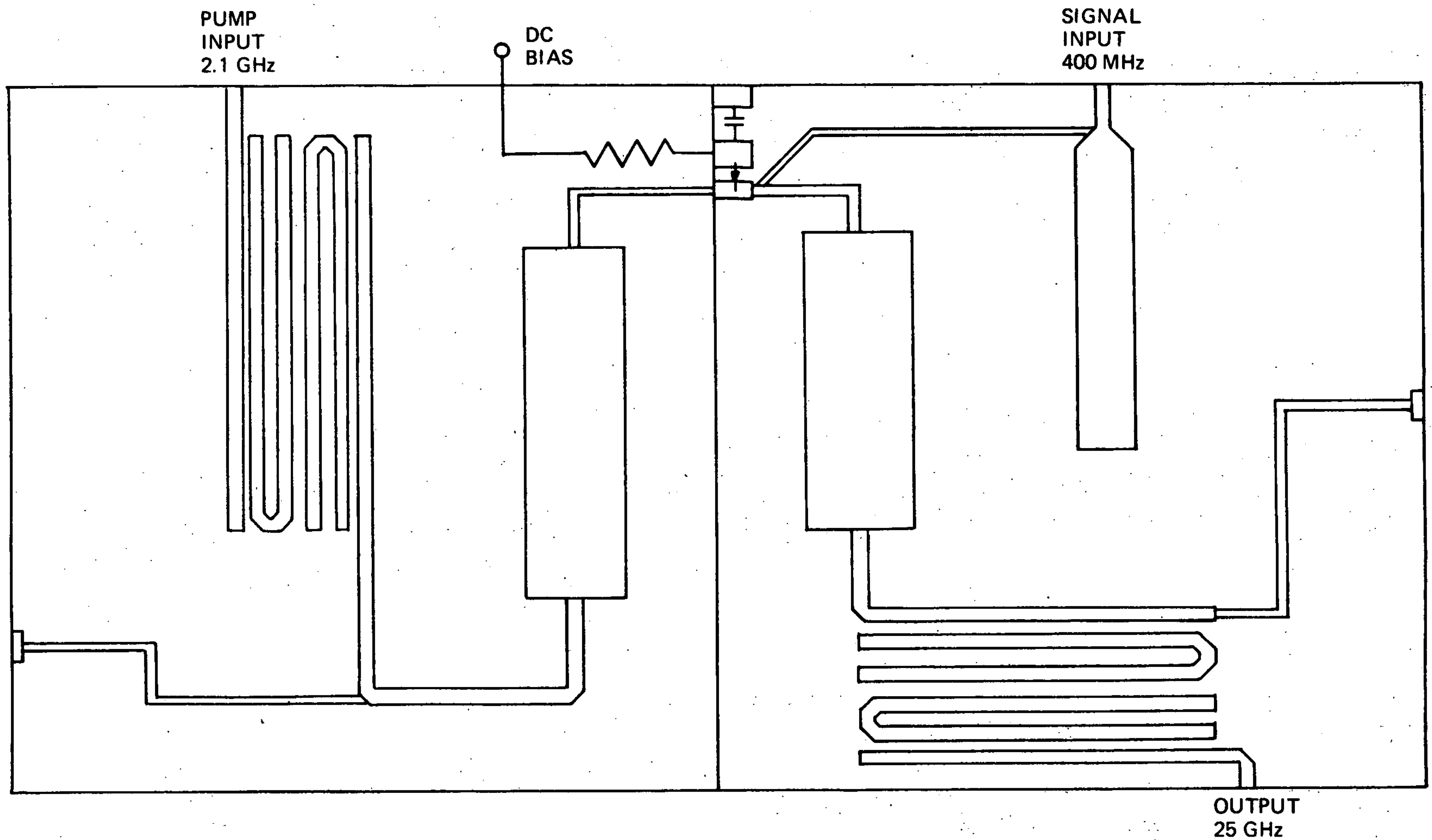

Figure 4-4. Layout of Final Low Frequency Up-Converter

Table 4-3. Impedance Transformations in the Final Low Frequency Up-Converter

Frequency (GHz)	Generator Impedance (Ohms)		VSWR
	Goal	Calculated	
A. Signal Input Circuit			
.35	-	34.7 - j4.2	
.40	-	31.8 - j2.4	
.45	-	29.1 - j0.1	
2.10	inf.	28.8 + j522	
2.30	inf.	5170 - j5600	
2.45	inf.	43.1 - j623	
2.50	inf.	28.0 - j473	
2.55	inf.	20.7 - j380	
B. Pump Input Circuit			
.35	inf.	7.5 + j145	
.40	inf.	2600 + j304	
.45	inf.	8.4 - j137	
2.10	10.6 + j75.1	10.5 + j75.0	1.011
2.45	inf.	6.8 + j173	
2.50	inf.	30.6 + j275	
2.55	inf.	301 - j369	
C. Output Circuit			
.35	inf.	2.6 + j94	
.40	inf.	11.2 + j199	
.45	inf.	1910 - j1770	
2.1	inf.	673 - j280	
2.45	10.8 + j64.2	10.4 + j60.0	1.477
2.50	10.6 + j62.8	10.6 + j62.7	1.013
2.55	10.4 + j61.6	10.9 + j65.5	1.451

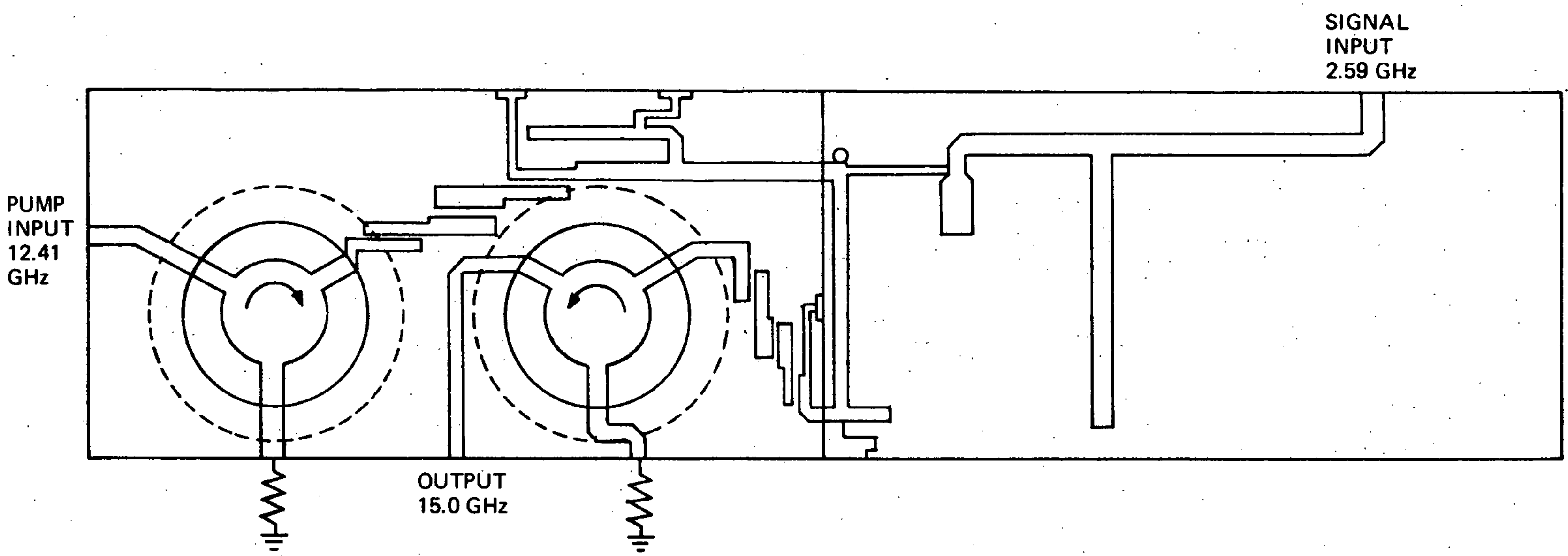

Figure 4-5. Layout of Final High Frequency Up-Converter

Table 4-4. Impedance Transformations in the Final High Frequency Up-Converter

Frequency (GHz)	Generator Impedance (Ohms) Goal	Generator Impedance (Ohms) Calculated	VSWR
A. Signal Input Circuit			
2.59	10.4 + j106	12.8 + j103	1.12
12.41	inf.	25.9 + j438	
13.7	inf.	8990 - j10100	
15.0	inf.	9.5 - j410	
B. Pump Input Circuit			
2.59	inf.	16.3 - j266	
12.41	3.1 + j16.6	3.7 + j16.0	1.26
15.0	inf.	113.3 - j747	
C. Output Circuit			
2.59	inf.	300 - j1590	
12.41	inf.	124 + j560	
15.0	4.2 + j12.1	3.4 + j12.5	1.25

Of course, many different element values had been used in the computer model before the final values could be determined. In this process it becomes apparent that this circuit is far more critical than the low frequency up-converter.

4.3 FREQUENCY SELECTION CRITERIA

When the prototype of the first low frequency up-converter was fabricated and tested, it became apparent that the selection of output frequency was not optimum. In-band signals were being generated in the absence of a pump. What was happening was that harmonics of the input signal were being generated in the output passband with significant amplitude.

To appreciate this phenomena, consider the original specifications in which the input band covers the range of 350 MHz to 450 MHz and the output band 2450 MHz to 2550 MHz. The sixth harmonic of a signal at 420 MHz is 2520 MHz. This coincides with up-converted signal from 420 MHz to 2520 MHz. The seventh harmonic of a signal at 350 MHz is 2450 MHz, which is identical to the sum of 350 MHz and 2100 MHz.

An even more severe problem is anticipated in the second up-converter. In this case, a midband component at 2500 MHz has a sixth harmonic of 15,000 MHz, which is in the center of the output passband.

A study was done to determine the optimum frequencies for the system, given that the input must be centered about 400 MHz and the overall output at 15,000 MHz. It was determined that the minimum interference would occur if the first pump were at 2190 MHz and the second pump at 12410 MHz.

This choice was determined by considering a reduced bandwidth case, namely 400 MHz ±20 MHz as the input signal. For this case no harmonics of low order coincide with the output passbands. The desired signal at 2590 MHz ±20 MHz is then equally spaced from a sixth harmonic component at 2520 MHz and a seventh harmonic component at 2660 MHz. For the required ±50 MHz band, some interference from harmonic components will still occur in the first up-converter, but this will be over a minimum frequency range.

These new frequencies have been incorporated in the design of the second up-converter as described earlier. For the first converter, tuning of the matching networks could be done simply on the prototype and the changes incorporated in the final models. The bandpass filters are simply scaled to the desired frequencies.

4.4 SINGLE STAGE DESIGN

During the design of the two-stage up-converter specified for this contract, many shortcomings became apparent. Some of these are: 1. The interference from harmonics of the input signal; 2. The low ratio of output frequency to input frequency in each stage, approximately six-to-one, does not permit the separation of frequencies with simple capacitors or inductors; 3. The pump and output frequencies are not close enough to permit matching with a single simple transformer network; 4. The resulting circuits have branches which are designed to achieve three different impedance conditions at three different frequencies simultaneously. The conditions cannot be achieved independently. Any changes to improve one condition will upset the impedances at the other frequencies; and 5. The matching circuits use very long overall transmission line lengths which are critically tuned. As a result, computer analysis shows the circuits to be very sensitive to minor changes such as the change of substrate thickness from 24 to 26 mils and the change in substrate relative dielectric constant over the range of 9.6 to 10.4. It is, therefore, anticipated that circuits fabricated with this design will require extensive trimming to achieve a reasonable performance.

As a result of these problems encountered during the design phase, it is realized that a much simpler circuit which should be smaller, lighter, much less critical, and more broadband can be designed. Such a circuit utilizes a single pump at 14.6 GHz which up-converts the input at 400 MHz to the desired output at 15 GHz. This circuit uses a 10 mil thick quartz substrate because a longer wavelength is desirable at Ku-band frequencies. When 25 mil thick alumina substrates are used, short transmission line elements result which increases the sensitivity to normal tolerance variation. Furthermore, the use of wide transmission lines is limited to lines whose widths are less than one-quarter wavelength; otherwise the behaviour of the transmission line elements becomes unpredictable.

Various circuit configurations using both series and shunt mounted diodes were analyzed on computer. The best arrangement found is shown in Figure 4-6. The pump and output VSWR are about 1.5:1 while the input mismatch loss is less than 10 dB.

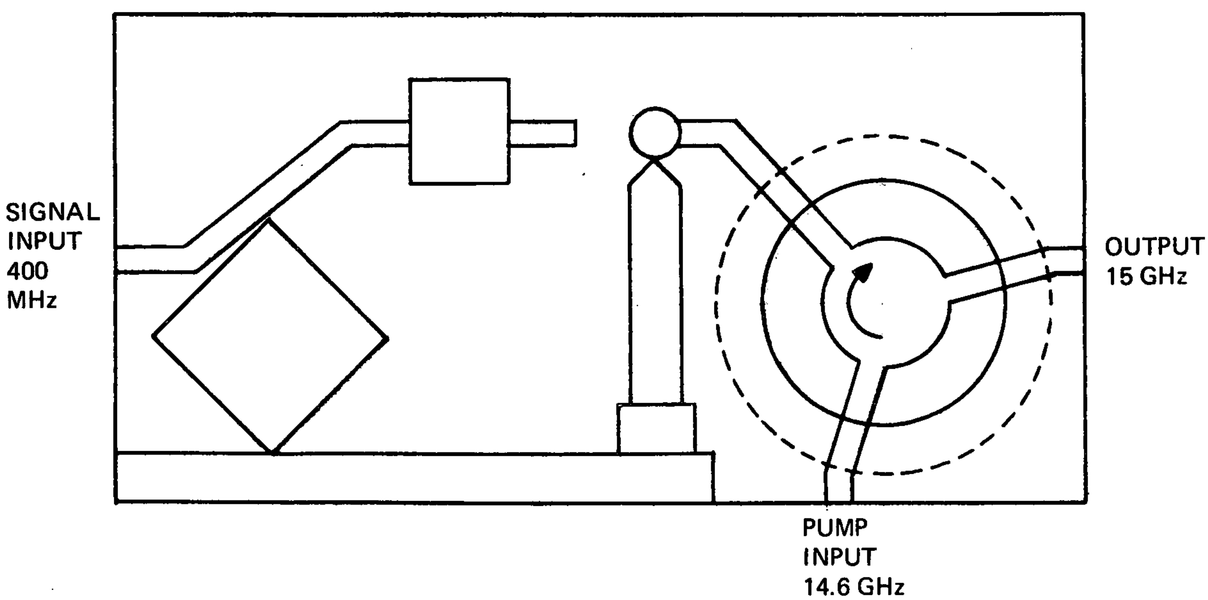

Figure 4-6. Layout of Single Stage Ku-Band Modulator

Section 5

MEASURED PERFORMANCE

A photograph of the two stage Ku-band Microminiature Modulator is shown in Figure 5-1. The first stage is fabricated on two 1" by 1" alumina substrates, while the second stage is on two 1" by 5/8" alumina substrates. The circuit frame is made in two sections which can be separated to permit testing of each stage without having to remount the circuits. Each section provides a continuous and direct ground current path from the SMA connector to the ground metalization of the substrates because it is machined from a single piece of brass. This insures good VSWR's from the connectors and low radiation even at Ku-band frequencies.

The diode in the second stage is placed above the substrate and the upper end returned to ground through a metallic strap. At Ku-band frequencies, it is found that the impedance of this strap is too high to provide a good ground return. This is provided by a separate metallic stub with one end resting on the top of the diode and the other end on the substrate. The length and position of this stub is varied to tune this diode and provide maximum output power. This strap is held in position by epoxies. It is found that better performance is achieved with this method rather than bringing the diode to ground through a hole in the substrate.

Tuning of the circuit is done with strips of indium. The size and location of the strips are adjusted to give optimum performance and then secured in position with epoxy. The resulting performance of two modulators of this configuration is given in Table 5-1. It can be seen that the midband output is about 3.4 dBm, and that the output falls about 2 dBm over the 100 MHz band.

The single stage modulator is shown in the photograph in Figure 5-2. It is fabricated on a 10 mil thick quartz substrate which is 1 in. by 1/2 in. The circuit frame also uses a single piece of brass to provide a continuous direct ground current path from connector to substrate. The only adjustment needed in this circuit is the length and position of the metallic strip attached to the top of the diode. Other tuning is unnecessary because of the circuit simplicity, the independence of input and output circuits, and the use of quartz substrates. The performance of three modulators built in this configuration is also shown in Table 5-1. It can be seen that the typical midband output is about +8.5 dBm and that the output drops only 1 dB over the 100 MHz band.

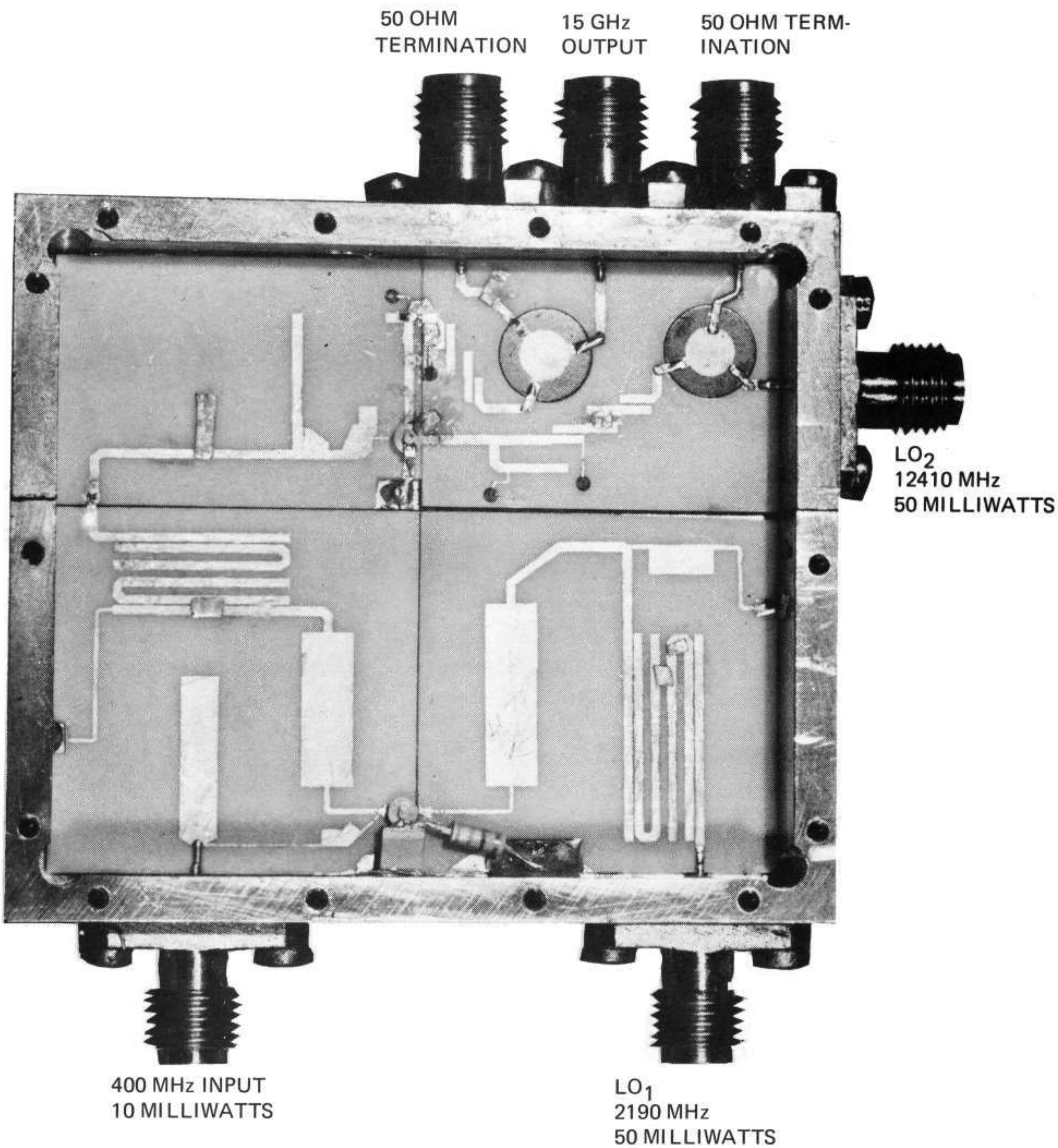

Figure 5-1. Two-Stage Ku-Band Microminiature Modulator

Table 5-1. Power Output as a Function of Input Frequency for the Five Delivered Modulators

Freq.	Unit 1	Unit 2	Unit 3	Unit 4	Unit 5
275			8.7	7.7	8.5
300		-0.6	9.0	7.9	8.6
325	-1.6	0.9	9.0	8.0	8.7
350	1.5	2.2	8.8	8.0	8.7
375	3.4	3.4	8.7	7.9	8.7
400	3.4	3.3	8.8	8.0	8.7
425	2.4	2.2	8.2	7.7	8.6
450	1.3	0.9	7.8	7.0	8.2
475	-3.0	-0.3	7.6	7.0	8.1
500		-1.9	7.2	6.7	7.8
525			6.2	5.9	7.0
550			5.9	5.3	6.7
575			5.6	4.9	6.0
600			4.6	4.0	5.0
	Two-Stage		Single Stage		

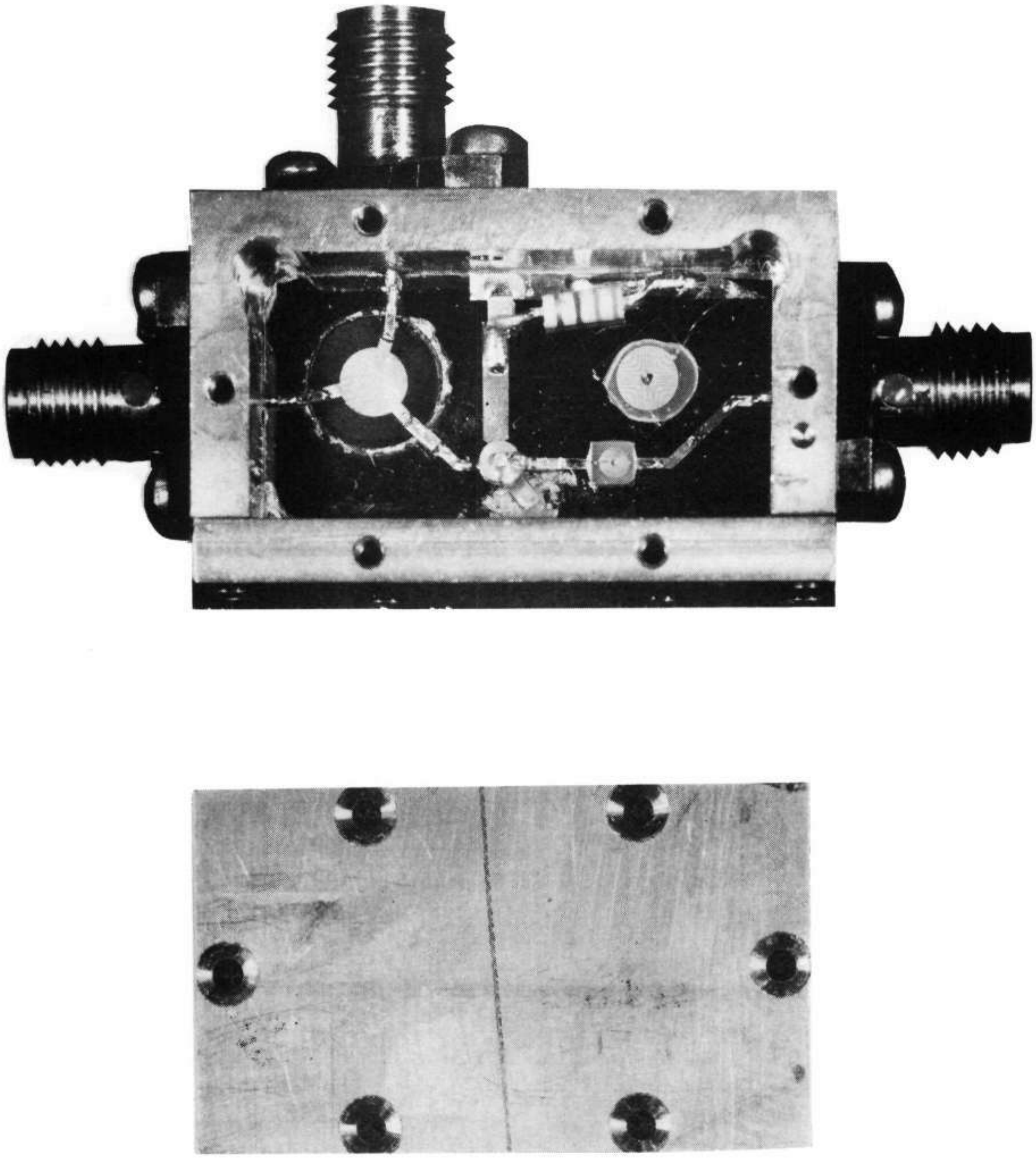

Figure 5-2. Single Stage Ku-Band Microminiature Modulator

Section 6

NEW TECHNOLOGY

Ferrite circulators utilizing Microstrip techniques were developed for operation at Ku-band frequencies. They were 255 mils in diameter, 25 mils thick, and had a 1 dB insertion loss, 12 dB isolation band of 10.5 GHz to 18 GHz.

MIC bandpass filters utilizing hairpin resonators have been developed at 2.1 GHz and at 2.5 GHz to achieve maximum utilization of substrate area. These filters have a 1 dB insertion loss and a 0.5 dB bandwidth of 100 MHz.

Fused quarty substrates have been utilized at Ku-band frequencies to achieve better circuit performance. A wider range of characteristic impedance is available and transmission line lengths are not as critical.

Computer aided design (CAD) techniques have been modified to improve correlation with measured performance at Ku-band frequencies. The model for microstrip transmission lines now includes a built in correction for dispersion effects.

Circuit frames have been developed to improve the VSWR at the connectors and to reduce radiation effects at Ku-band. Previously, base blocks and surrounding frames were separate pieces held together by two screws. In the worst case, ground currents would be forced through a long path including these screws to get from the connector to the substrate. Now a single piece of brass is used to form the complete base block and to minimize the ground current path length.

Section 7

CONCLUSIONS AND RECOMMENDATIONS

The output power as a function of frequency for the five modulators is shown in Figure 7-1. It can be seen that the single stage units have about 5 dB greater output and much greater bandwidth than the double stage modulators. In addition, these units are only 20% of the size and weight of the larger units.

As a result of this program, the following recommendations can be made when designing Ku-band varactor diode up-converters.

a. The Ku-band circuitry should be on 10 mil thick quartz substrates. This results in reasonable dimensions for transmission line elements and permits a wider range of characteristic impedances to be used.

b. The ratio of output frequency to input frequency should be higher than 10 to 1. This results in much simpler circuitry. If any tuning is necessary, it can usually be done without upsetting other frequency components because simple capacitors, inductors, or stubs can be used to eliminate interaction. Another advantage is that any multiples of the input frequency that fall within the output passband will be low in level because of the high order of multiplication required.

c. The up-conversion should be done in as few stages as possible. This eliminates losses in interstage matching networks and completely avoids the problem of parametric oscillations frequently encountered when two or more stages are cascaded. The resulting circuits have smaller size and weight.

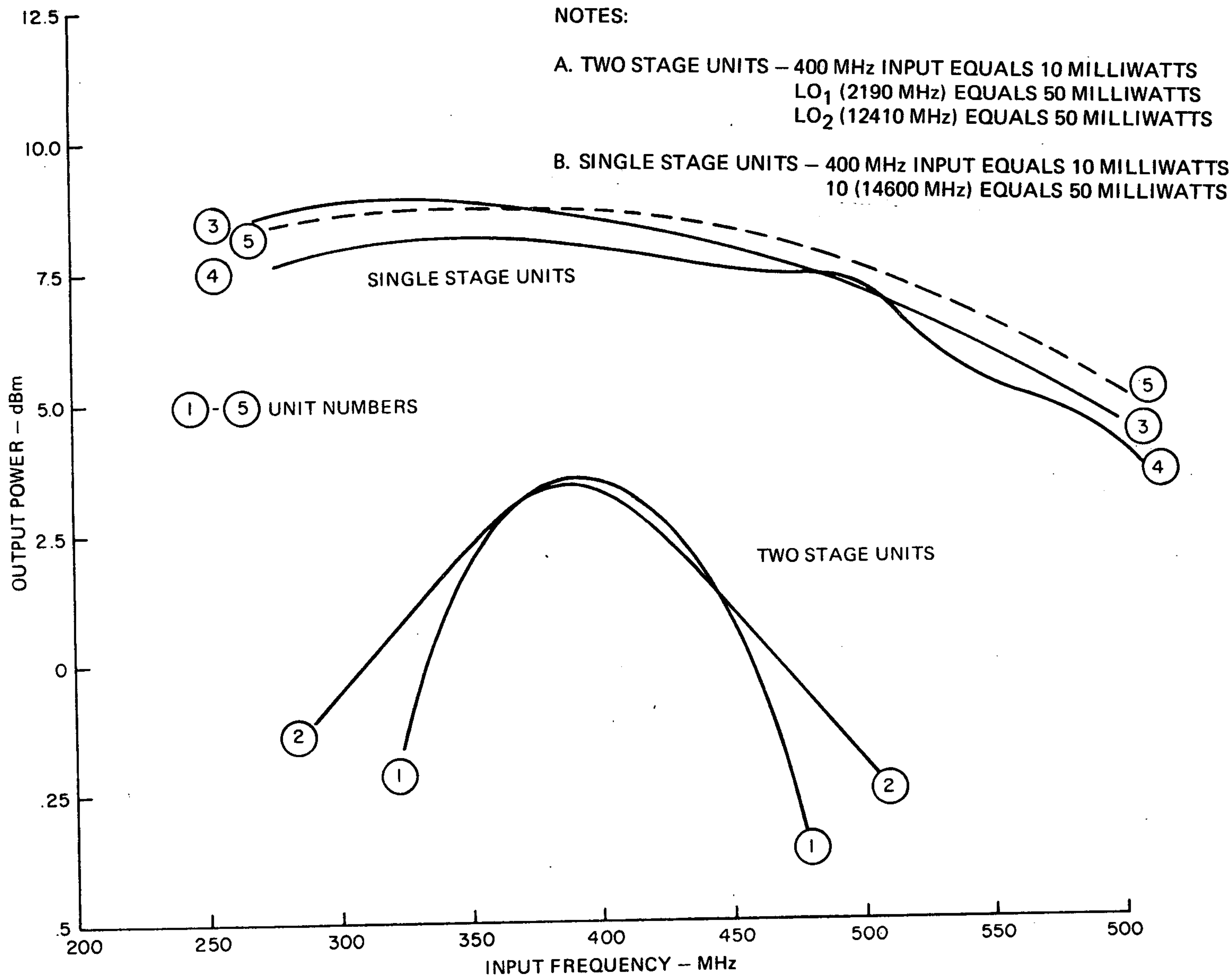

Figure 7-1. Output Power Performance of Five Ku-Band Microminiature Modulators

Appendix

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