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K(A) BAND MICROWAVE INTEGRATED CIRCUIT SPDT SWITCH DEVELOPMENT.(U)
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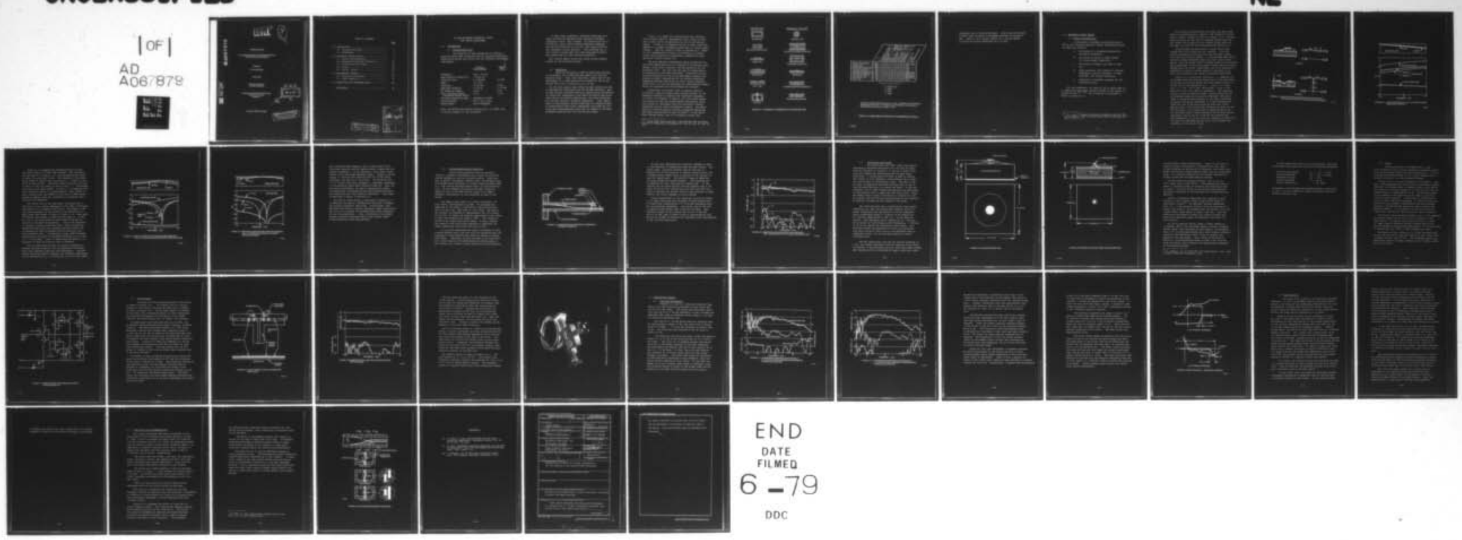

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INTERIM REPORT

K_A BAND MICROWAVE INTEGRATED CIRCUIT SPDT SWITCH DEVELOPMENT

Prepared by

Dr. Charles Buntschuh

January 1979

Microwave Associates, Inc.
Burlington, MA 01803

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K_A-BAND MICROWAVE INTEGRATED CIRCUIT SPDT SWITCH DEVELOPMENT

1.0 INTRODUCTION

1.1 Program Objectives

The objective of this program was to develop a practical broad band, integrated circuit K_a-band single-pole-double-throw switch with driver with the following performance requirements:

	Design Requirements	Design Goal
Frequency:	26.5 - 40 GHz	
Power Handling Capability:	2 watts CW	
Insertion Loss:	2.5 dB max.	2.0 dB
VSWR:	1.7 max.	
Isolation:	30 dB min.	40 dB
Amplitude Tracking:	$\pm$ 0.5 dB	$\pm$ 0.2 dB
RF Transition Speed:	10 ns	5 ns
Total Switching Speed: (including driver delay)	20 ns	15 ns
Switching Repetition Rate:	Variable to 2 MHz	
Driver Control:	T^2L, 1 or 2 bits	
Input/Output:	WR-28 Waveguide	

Also, the switch must survive looking into a 2:1 VSWR, and it must be capable of "hot switching".

To meet these objectives, Microwave Associates proposed to build two switches using different microwave integrated circuit (MIC) formats. One switch would be in microstrip, using a Duroid teflon-fiberglass substrate, and the other would be in fin-line, also using Duroid as the integrated circuit material. Both switches are to use a high-Q, plated heat sink diode developed by Microwave Associates Ltd, M/A's English subsidiary.

This Interim Report covers the design and performance results on the microstrip switch.

1.2 Background

There are currently no SPDT K_a-band diode switches on the market; and no SPST switches which cover the full waveguide band. Thus, there has been a great need for the development of a broad band switch suitable for use in new systems which will be operating up to 40 GHz.

For the past several years there has been growing activity in extending MIC techniques into the mm-wave spectrum. Much of this activity has been devoted to exploring the suitability of various MIC transmission line media--including some quite novel forms--for mm-wave applications, and their potentials for low-cost reliable microwave components. Also, improvments in semiconductor device technology are yielding diodes of exceptionally high cutoff frequencies, such that MIC circuits using unencapsulated chips will be capable of low-loss broadband operation well into the mm-wave region.

Figure 1.2.1 shows the cross-sections and configurations of most types of transmission lines, either in wide use or under active investigation for use in the mm-wave region. Figure 1.2.2 compares several of these lines with respect to a number of properties pertinent to various applications. They are listed in order of decreasing Q. These Q's are representative values to be expected in K_A-band; actual values could vary considerably, depending on size, materials, surface finish, etc.

The most important transmission medium properties for the switch application are its suitability for mounting chip diodes, intercircuit isolation, and the potential for low production cost. Low line loss and ease of transitioning to waveguide are also quite desirable. Microstrip and finline were selected for the switch development because they possess favorable combinations of the desired properties. Suspended substrate stripline was also a strong contender, except that it is not as convenient for shunt-mounted diodes, which are preferred over series diodes for switches.

Conventional microstrip, that is alumina substrate microstrip, loses its appeal above K_u-band because, as the substrate thickness is reduced with increasing frequency to control its tendency to radiate, the line widths become very narrow and the line losses become unacceptably large. This problem is alleviated using low dielectric constant substrates such as teflon fiberglass or quartz. The applicability of Duroid material through K_a-band and beyond has been amply demonstrated by Rubin and Saul[1] of the Naval Ocean Systems Center, and it is gradually coming into

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(1) David Rubin and David Saul, "mm-wave MICs Use Low Value Dielectric Substrates", Microwave Jnl. Vol 19, No. 11, Nov 76, P35.

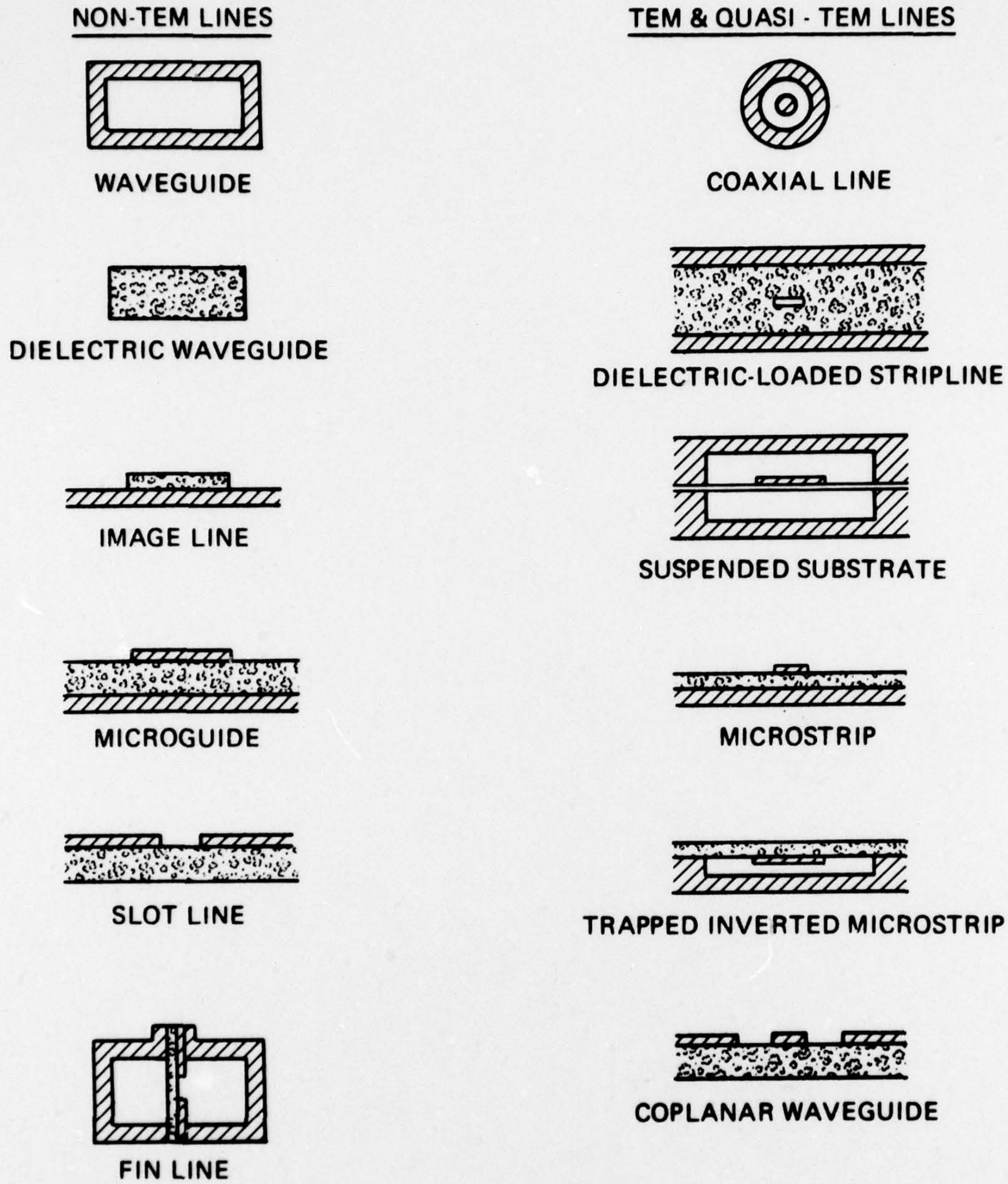

FIGURE 1.2.1 MICROWAVE TRANSMISSION LINE CROSS-SECTIONS

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	Q	TRANSITION TO WG	ISOLATION	SEMI-COND. DEVICE MOUNTING: PKG'D	SEMI-COND. DEVICE MOUNTING: CHIP	SEMI-COND. DEVICE MOUNTING: BEAM LEAD.	SUITABILITY FOR: FILTER	SUITABILITY FOR: 3 dB HYBRID	SUITABILITY FOR: 20 dB CPLR.	SUITABILITY FOR: GUNN RES.	RADIATION FROM DISCONTINUITIES
METAL WAVEGUIDE	3000	–	4	4	1	1	4	3	3	4	VERY LOW
IMAGE LINE	2500	3	1	2	1	1	1	1	2	1	HIGH
SUSPENDED STRIPLINE	600	3	3	2	2	1	3	3	4	1	LOW
FIN LINE (DUROID)	500	4	4	1	2	3	4	1	3	2	LOW
MICROSTRIP (DUROID)	300	3	2	2	3	4	2	3	4	2	MODERATE
TIM LINE (ALUMINA)	300	1	2	2	3	4	2	3	4	2	MODERATE
CO PLANAR LINE	150	1	2	1	3	4	1	1	2	2	MODERATE
SLOT LINE	100	3	1	1	3	4	1	3	3	2	HIGH

4 = EXCELLENT
3 = GOOD
2 = FAIR
1 = POOR

(ADAPTED FROM PRESENTATION BY U. GYSEL, FORMERLY OF STANFORD RESEARCH INSTITUTE, AT THE SOLID STATE MM-WAVE TECHNOLOGY WORKSHOP, SAN DIEGO, CA, JUNE 21, 1977.)

FIGURE 1.2.2 COMPARISON OF MM-WAVE MIC TRANSMISSION LINE MEDIA

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widespread use in the mm-wave bands. Also Duroid microstrip is being used routinely in broadband switches up through 18 GHz. Thus, it is a logical choice for a circuit medium to extend MIC switch technology up to 40 GHz.

2.0 MICROSTRIP SWITCH DESIGN

2.1 Design Considerations

In developing a broadband K_a-band microstrip switch, the following important design considerations must be taken into account:

a) The design of a broadband waveguide-to-microstrip transition.

b) The choice of series or shunt diodes, or a series-shunt combination.

c. The number of diodes to be used in each arm.

d. Diode selection, with respect to junction capacitance, series resistance, I-layer thickness, and thermal resistance.

e. Construction techniques suitable for low production costs.

For the transitions, we chose to use a cosine taper to a ridge guide microstrip launcher similar to the design reported by Saul.[(1), (2)] The transition is discussed in detail in Section 2.2

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(2) D. Saul, "Wideband Microstrip Components and the IFM Discriminator", Proc 1974 Millimeter Wave Techniques Cont," Vol 2, p D4-1, 12.

The mounting configurations for series and shunt chip diodes in microstrip with a nominal .008" dielectric thickness are shown, roughly to scale, in Figure 2.1.1. The equivalent circuits for the zero or reversed biased state are also shown; the circuits for the forward biased state are the same, but with the junction capacitance shorted out.

Series diodes are much easier to mount since holes through the circuit board are not required. Also the series inductance can be much less than for shunt diodes since only one strap is needed. A series-diode SPDT switch has much broader bandwidth capability than a shunt diode switch, since the first diode is located right at the junction. The disadvantage of the series diode is that switch isolation depends upon the diodes' capacitance, and it is essentially impossible to obtain a useful amount of isolation at K_a-band from diodes with the lowest practicable capacitance.

Figure 2.1.2 illustrates this point. It compares the calculated insertion loss and return loss in the pass state, and isolation of single series and shunt diodes from 20 to 40 GHz in a 50 ohm system for C_j = .03 pF, R = 1 ohm, and L = .12 nH. (At 33 GHz, mid K_a-band, .03 pF is 160 ohms and .12 nH is 25 ohms of reactance.) The insertion losses of both are about the same, limited essentially by the mismatch caused by the series inductance; the series diode is slightly better because of the lower inductance. Increasing the series resistance to 3 ohms increases the insertion loss of the shunt diode by less than .02 dB whereas that of the series diode increases by about 0.2 dB. The isolation, on the other hand is approximately 30 dB for the shunt diode, limited primarily by the diode series resistance, and as low as 3.5 dB for the series diode. Even reducing C_j to .01 pF, which is probably not practical, at least for high-speed switch, would only increase the isolation to 11.5 dB at 40 GHz.

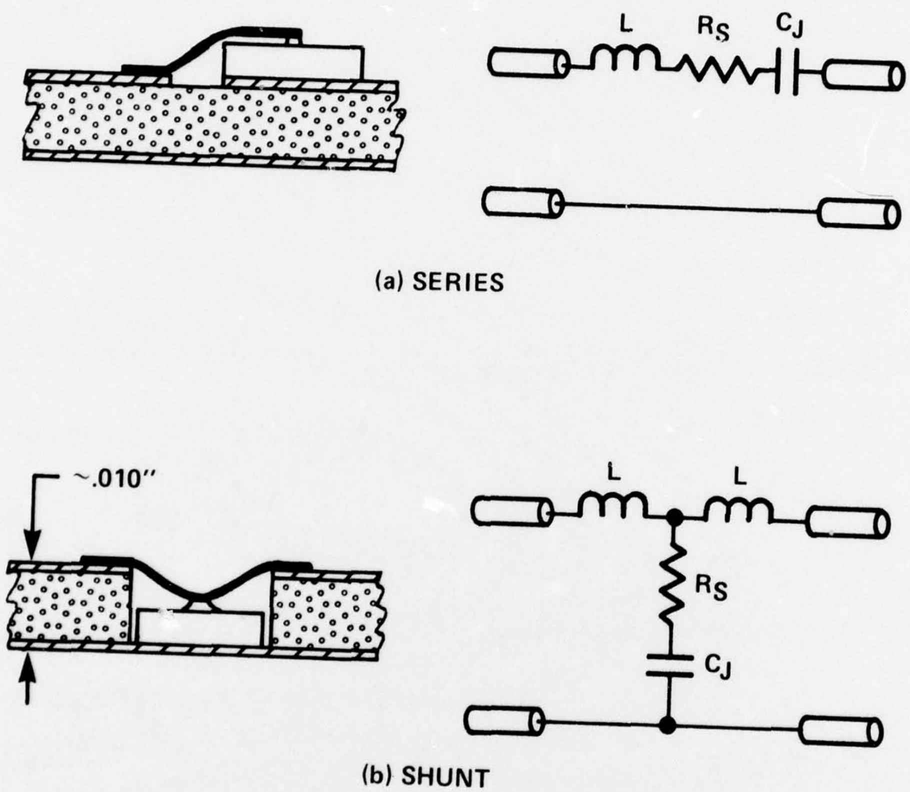

FIGURE 2.1.1 CONFIGURATIONS AND EQUIVALENT CIRCUITS FOR SERIES & SHUNT MOUNTED CHIP DIODES IN MICROSTRIP.

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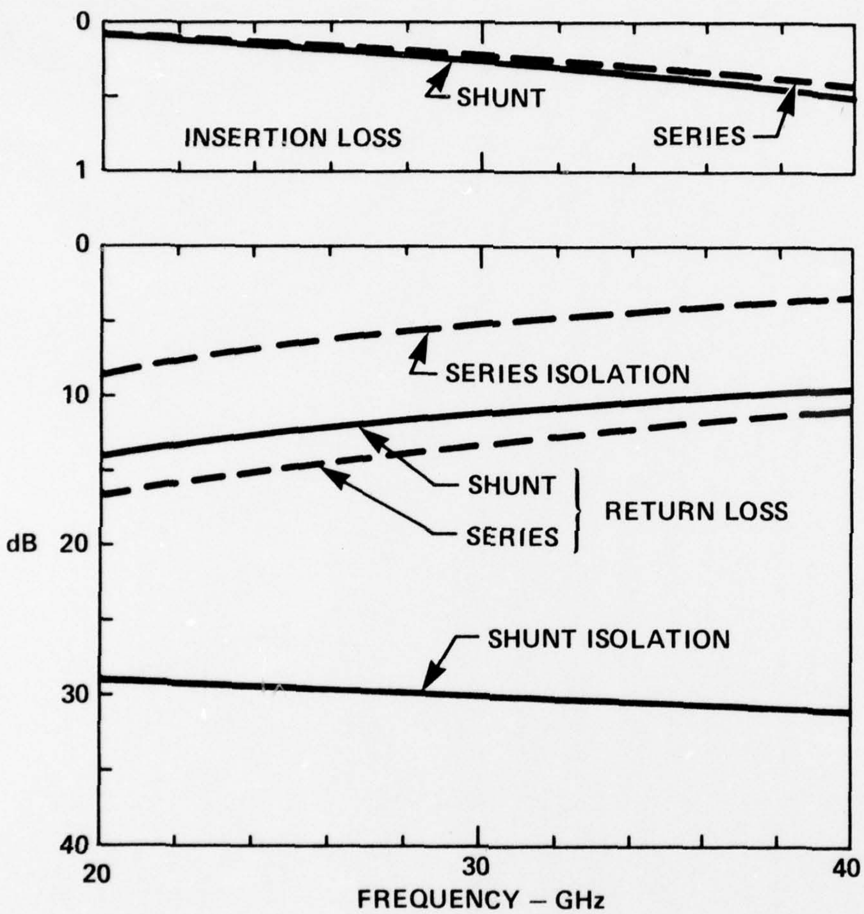

FIGURE 2.1.2 CHARACTERISTICS OF .03 pF, 1Ω DIODE IN SERIES AND SHUNT MOUNTING

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Figure 2.1.3 compares the characteristics of series and shunt diodes in 2-diode SPST switches. The parameters are the same as in Figure 2.1.2, and the spacing was chosen to match the pairs near 33 GHz. The isolation of the shunt diode switch is about 65 dB; even with R_s = 3 ohms, it will still be about 45 dB. The series diode isolation runs from 22 down to 7 dB; its insertion loss is also .1 - .2 dB higher than that of the shunt switch. Clearly, it is essential that shunt diodes be used for a K_a-band switch. Also, these results indicate that two diodes in each arm of a SPDT switch should be adequate from the standpoints of insertion loss, isolation and bandwidth.

The return loss at the band edges, calculated for a SPST with 0.12 nH series inductance is only about 1 dB higher than desired for a complete SPDT switch. The effect of the wire bond inductance on the insertion loss and match of a shunt diode SPST switch is shown in Figure 2.1.4. Here again the spacings were chosen to match the pair near 30 - 33 GHz. For .04 nH, the match would be better than 40 dB and the insertion loss would be negligible. Such a low inductance is, however, beyond a practical limit. Small capacitance diodes have very small top contacts--typically under .001" in diameter--so that the bonding must be done with fine, high inductance wire. Consequently, using .0008" diameter bonding wire, we expect the minimum inductance to be in the .10 - .15 mH range. Also, in fabricating the switch it is important to keep the leads as short as possible.

For any shunt diode switch, low series resistance is desired for high isolation, and low capacitance for broadband, high frequency insertion loss performance. The preceding calculated results indicate that excellent performance can be obtained with $C_j \leq .03$ pF and R_s in the 1-3 ohm range.

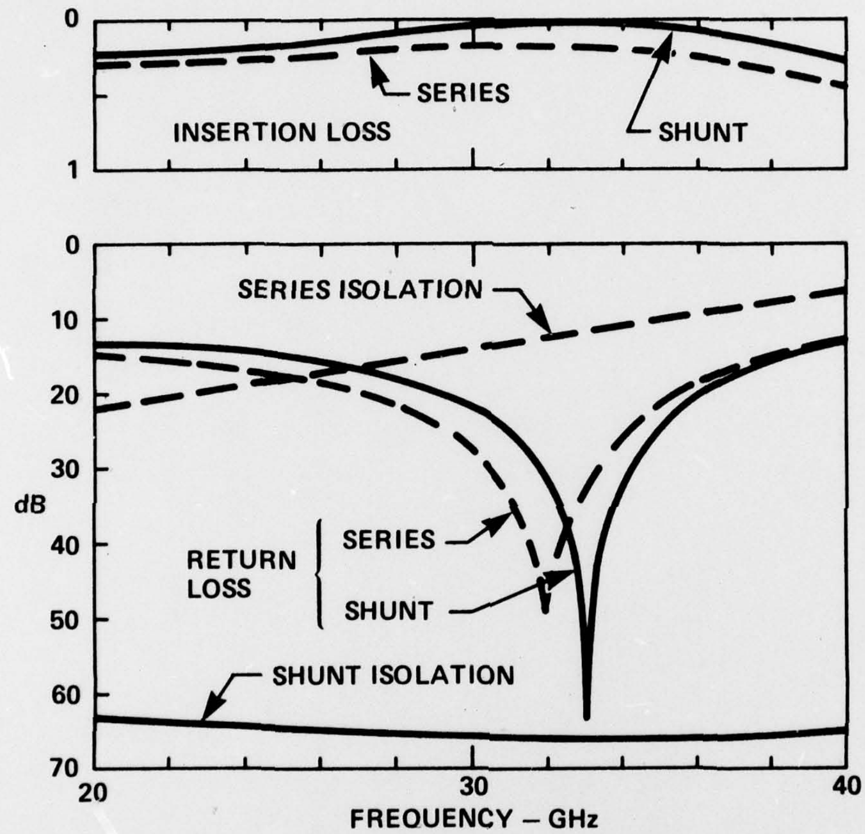

FIGURE 2.1.3 CHARACTERISTICS OF 2-DIODE SPST SWITCHES WITH .03 pF, 1Ω DIODES IN SERIES AND SHUNT CONFIGURATIONS

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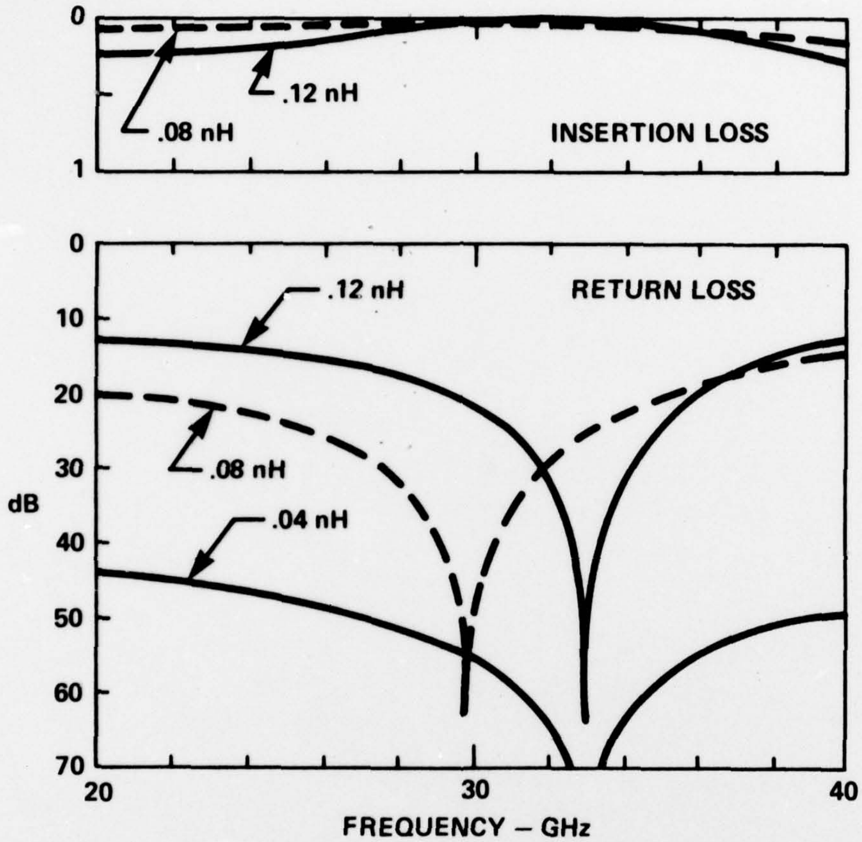

FIGURE 2.1.4 INSERTION LOSS & RETURN LOSS FOR TWO SHUNT .03 pF, 1Ω DIODES FOR SEVERAL VALUES OF SERIES INDUCTANCE

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Fast switching speed demands a thin I-layer diode, while high power handling ability requires thick I-layers and low thermal resistance. At the specified 2 watt level, diode burnout is not the power-limiting effect. Rather it is the change in isolation and insertion loss due to non-linear resistance and heating of the diodes. We estimated that a 5 μm I-layer diode would represent a reasonable compromise in the switching speed--power handling trade off. Thus, for the switch development, we used an exceptionally high-Q, low capacitance, plated-heat-sink diode developed by Microwave Associates Ltd, M/A's English subsidiary. This diode is described in detail in Section 2.3.

The overall switch design is described in Section 2.4. In designing the switch, consideration was given to keeping the ultimate production costs down. In particular, this motivated the choice of plastic substrate, rather than quartz, and it led us to design the waveguide-microstrip transition for low cost production. Also, in Section 4.0 we suggest another transition technique which could further reduce volume production costs substantially.

2.2. Waveguide-Microstrip Transition

Single-ridge waveguide forms an excellent, "natural" launcher to microstrip line, since the field and current patterns of the two are quite similar. Rubin and Saul at NOSC(1) developed a tapered ridge, transforming from full height WR-28 waveguide to a single ridge gap which mated to a .010" Duroid microstrip board with a 60 ohm microstrip line. We have adapted and modified the NOSC transition design for use in the SPDT switch; Figure 2.2.1 shows a schematic cross-section of the transition as used in the switch.

The shape of the taper is a cosine curve about 1.5" long--about 3 wavelengths at K_a-band. We increased the amplitude of the cosine to mate with a .008" Duroid board with a 50 ohm line. The ridge is .040" thick. The nose of the launcher is chamfered to approximately the .023" width of the line, to reduce the fringing capacitance of the corner. To match the transition we added a .005" - .010" length of Duroid in the ridge guide gap as shown. In fact, varying this length is the only tuning required to compensate for slight dimensional variations from unit to unit.

In our early experiments we found it necessary to have a narrow channel with a low roof over the microstrip near the launchers to avoid mismatch and radiation effects of higher order modes. Therefore, we have included in the transition unit a short section of reduced height and width guide at the launcher end, forming a small bridge over the microstrip.

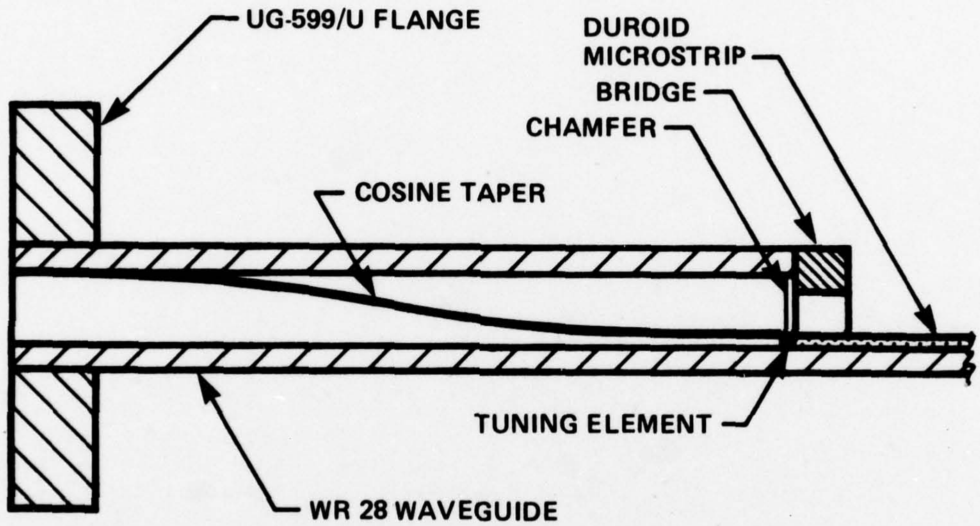

FIGURE 2.2.1 CROSS SECTION OF WAVEGUIDE–MICROSTRIP TAPERED TRANSITION

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We have also redesigned the transition assembly to make it less expensive to manufacture than the NOSC version. The cosine taper, including indexing tabs on each end, is machined out of flat stock on an optical tracing miller, and a standard WR-28 copper guide is slotted in the top wall to take the tapered part. The UG-599/U flange is added and the three pieces are brazed together. The waveguide end is machined off flat, removing one indexing tab. At the launcher end the guide top wall and the indexing tab on the taper are milled off, leaving the side walls and floor as a guide to position the bridge, which is then soldered in place. Finally, the entire assembly is silver plated.

Typical performance of the transition is given in Figure 2.2.2, which shows the loss and match of back-to-back pairs of transitions, separated by 0.5" of 50 ohm microstrip line, for two different assemblies. The return loss of the pair is typically much better than 20 dB over the band. The insertion losses of the circuits ran 0.5 to 0.7 dB from 26.5 to 40 GHz, most of which is due to the microstrip line. We did not attempt to evaluate the individual loss contributions of the line and transitions.

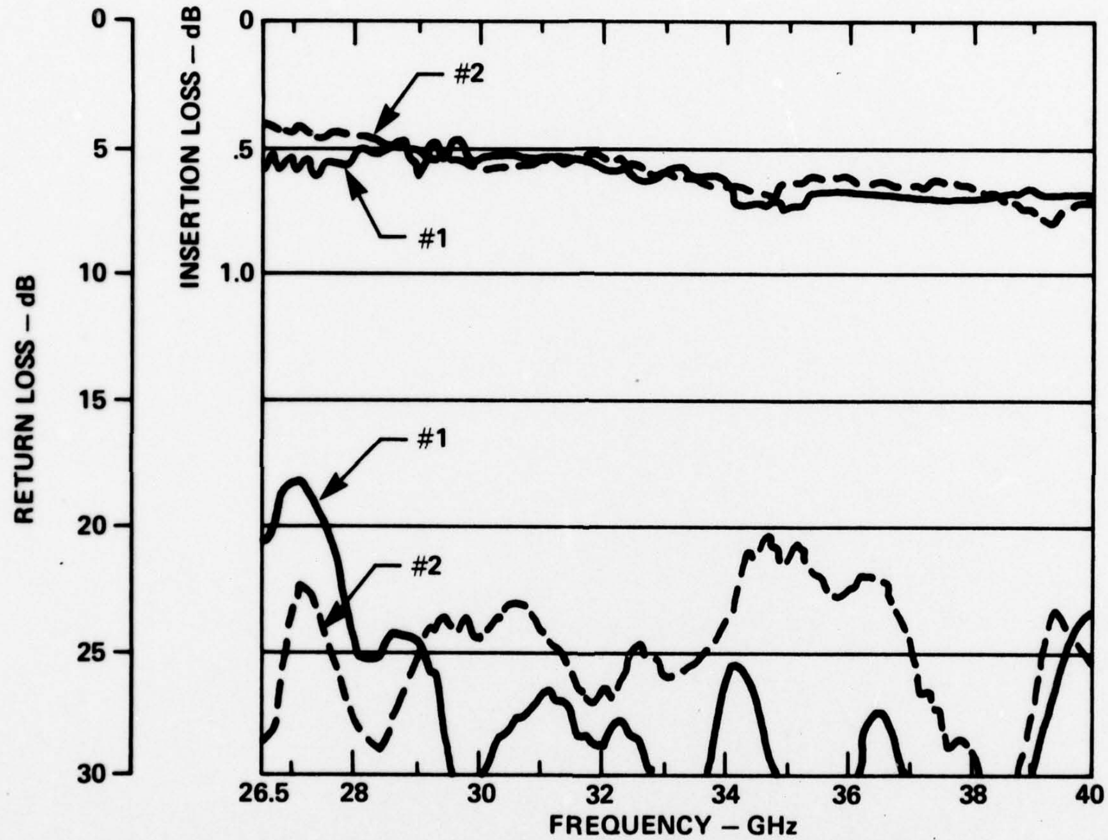

FIGURE 2.2.2 INSERTION & RETURN LOSS OF TWO MICROSTRIP TEST FIXTURES WITH 0.5" OF DUROID MICROSTRIP LINE

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2.3 Plated Heat Sink Diode

The plated heat sink (PHS) diode construction developed at Microwave Associates Ltd. can best be understood by reference to Figures 2.3.1 and 2.3.2 which illustrate and compare the conventional and PHS mesa diode structures. A conventional PIN diode is made by growing the I-layer on an N+ substrate, and diffusing the P+ layer into the surface of the epitaxial layer. The back of the N+ layer is completely metalized and small dots on the P+ side are metalized prior to etching the mesas. In the PHS technique, the P+ side is metalized entirely and then plated up with several mils of copper and then gold plated. Next, the unmetalized N+ layer is thinned to a fraction of a mil by etching, and then the dots are metalized on the remaining N+ material, followed by the etching of the mesas.

The resulting diode has a significantly lower series resistance, R_s, than the conventional diode by virtue of almost eliminating the contribution of the N+ substrate. Series resistances around 1.5 ohms at 10 mA forward bias for junction capacitances in the .02 - .05 pF range are typical. Thus switching cutoff frequencies of $>$ 2000 GHz are attained at 10 mA whereas cutoff frequencies $<$ 1000 GHz at 10 mA are typical for conventional mesa diodes. Conventional diodes can sometimes be driven into saturation with about 100 mA of bias to bring its resistance down to comparable values, but the switching speed will suffer correspondingly.

The PHS construction also has the obvious advantage of a very low thermal resistance from the junction to the mounting surface in the microwave circuits, since the normal thermal path through silicon is replaced by copper, which has almost

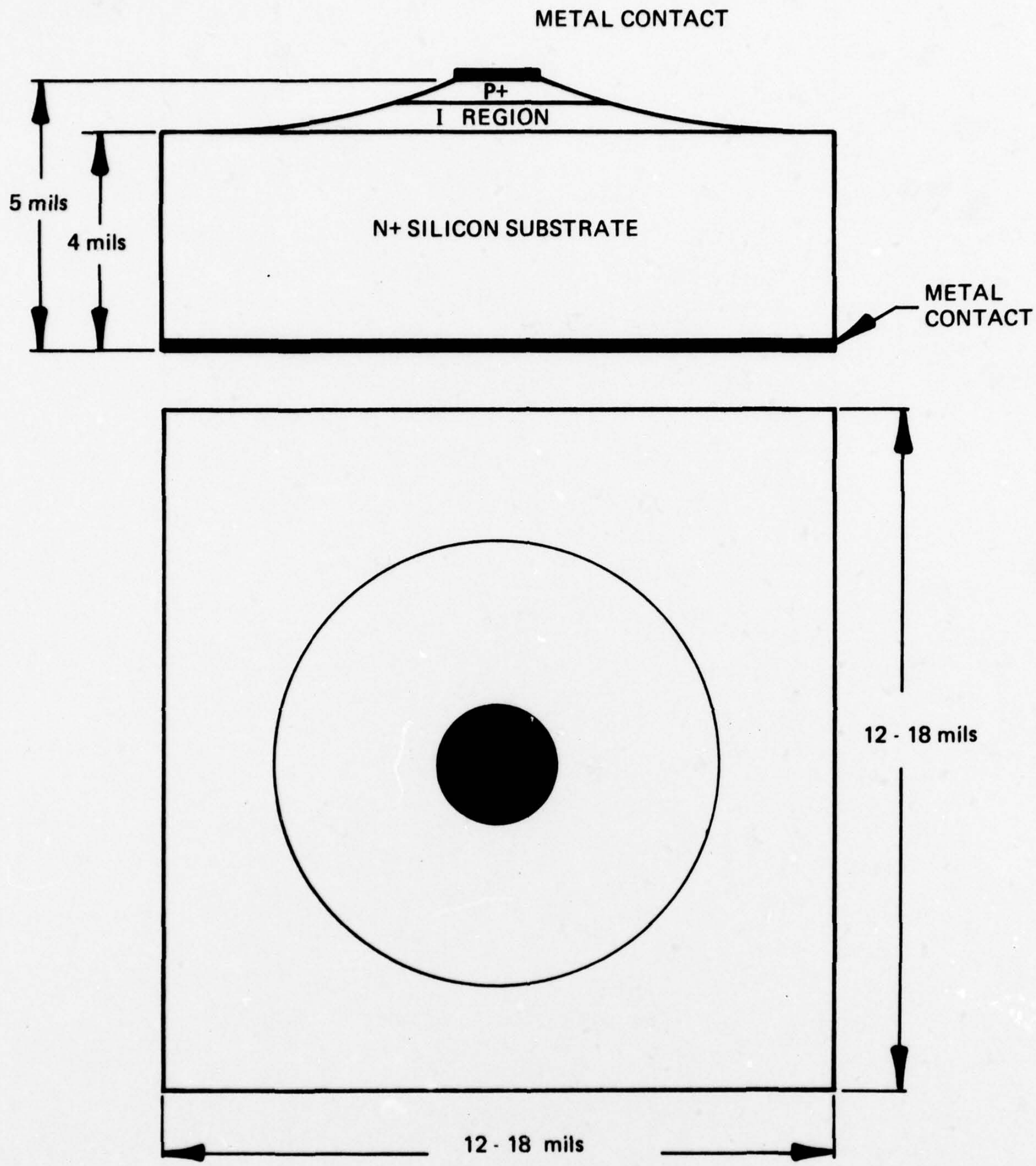

FIGURE 2.3.1 SILICON PIN DIODE CHIP

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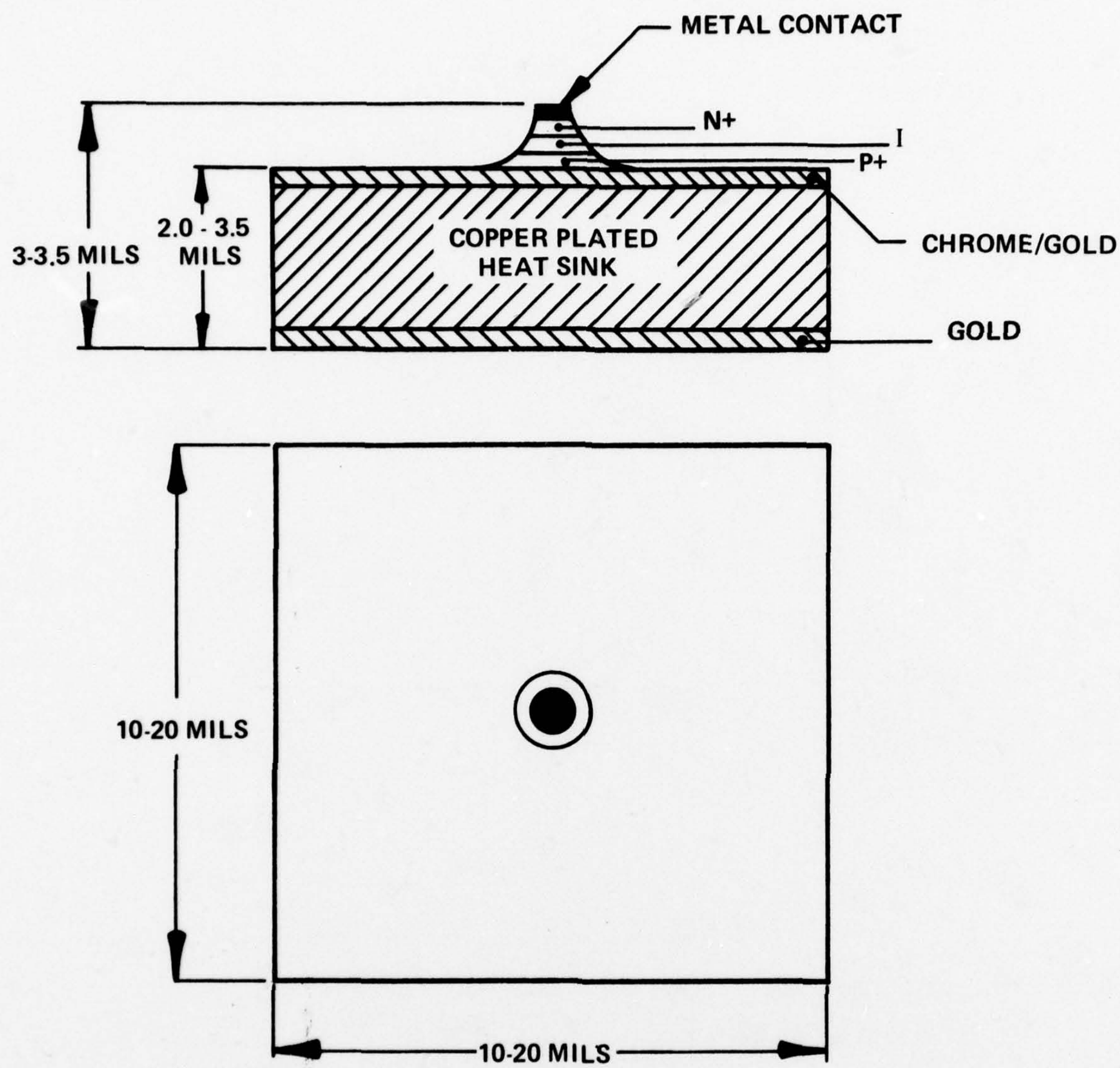

FIGURE 2.3.2 SILICON PLATED HEAT SINK (PHS) NIP DIODE CHIP

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5 times higher thermal conductivity. Thus, for circuits in which the diode is mounted on an adequate heat sink, such as shunt mounting in microstrip or stripline, the circuit will handle substantially higher average power than the conventional diode. In a typical installation the thermal resistance is roughly halved and the power handling doubled.

The PHS diode has reversed polarity, that is it is a "NIP" rather than a "PIN" diode. This necessitates using a reversed polarity driver or isolating the diode mount from dc ground. Since the driver we use provides both positive and negative outputs, we have employed the reversed polarity approach.

Pafford of Microwave Associates Ltd. reported on SPST switches in the 55-75 GHz range, using the PHS diode in a ridge-guide circuit.[3] He achieved 5 percent bandwidths with 1.0 - 1.2 dB insertion loss and 23-25 dB isolation in a single diode switch. He estimated that the switch, with 3 μm I-layer diode, would handle at least 1 watt CW. Also the diodes were gold-doped to reduce the carrier lifetime and enhance the switching speed; RF transition time under 1 ns were achieved.

For the K_a-band MIC switch, where a 2-watt capability is desired, with a 5 ns switching speed, we chose a slightly thicker diode, and eliminated the gold doping. We estimated that a 4-5 μm diode would handle the 2 watts without suffering insertion loss degradation with 5 volts of reverse bias applied in the pass state. Also, since the driver will switch a 8 μm non-gold-doped diode in under 5 ns, we could forgo the doping, which improves the diode series resistance at low forward currents.

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(3) A. Pafford, "An "O" Band Fast PIN Diode Switch", Proc. 1976 European Microwave Conference, p674.

The PHS diodes delivered for this program were specified by Microwave Associates Ltd to have the following parameters:

Junction Capacitance:	C_j	$= .02 - .03$ pF
Series Resistance:	R_s	$= 1.0 - 1.5$ ohms
Breakdown Voltage:	V_B	≥ 90 V
Carrier Lifetime:	τ	≥ 90 ns
Thermal Resistance:	Θ	≈ 40 ^{o}C/W

Furthermore, these diodes were punched-through at zero bias, so that it is not necessary to apply any reverse voltage to minimize insertion loss

2.4 Driver

The switch driver is the MA8417-203, a thin film integrated circuit driver developed at Microwave Associates for high speed, high clock rate applications. It is in an hermetically-sealed dual in-line ceramic package and is readily integrable into the microwave package.

Figure 2.4.1 is a circuit schematic of the MA8417. Note that it has complimentary inverting and non-inverting outputs. Thus with the outputs wired to opposite sides of the switch, one side will be reverse biased to 5V, and the other will be forward biased to approximately 30 mA. The forward current can be reduced to any desired value by adding an external current limiting resistor. Each output is divided into a resistive and a capacitive terminal. If the two output terminals are tied together, approximately 150 mA of spiking current is fed through the capacitive lead to obtain fast switching speeds. In our switch, we used the resistive output only, and did not add an external limiting resistor.

This driver may be biased by either $\pm$ 5V or $\pm$ 12V supplies; we provided leads only to the $\pm$ 5 Volt terminals on the final switch. It operates from 2-bit TTL logic. A logic "1" input yields a positive voltage (reverse bias on our PHS NIP diodes) on the non-inverting output. It is also rated to operate over the $-55^{\circ}C$ to $+125^{\circ}C$ temperature range.

The MA8417-203 will switch an 8 μm thick PIN diode from 30 mA to 5V with an RF rise time under 5 ns and a total time, including delay of under 10 ns. The switching rate is conservatively rated at 15 MHz, although it will operate acceptably up to 20 MHz.

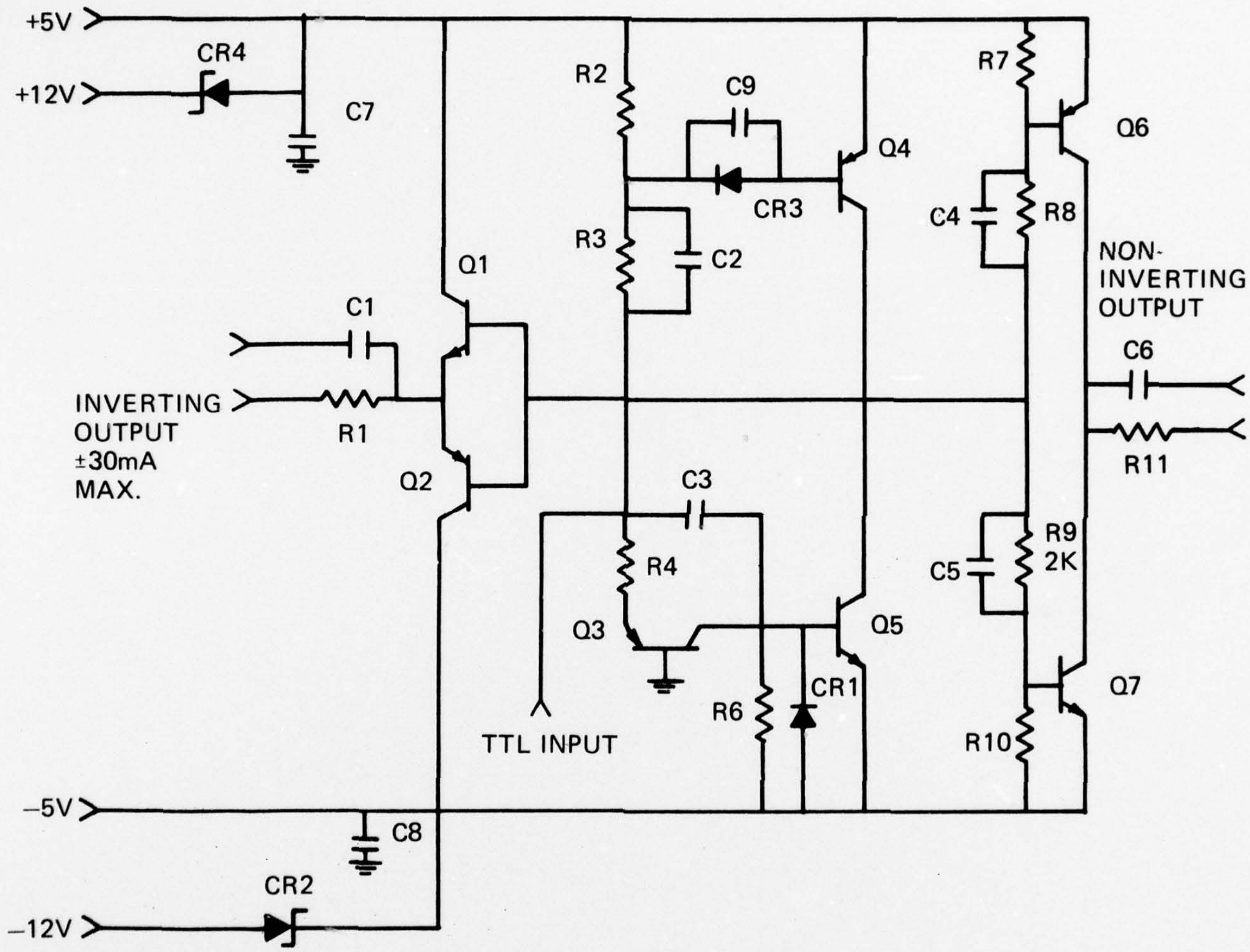

FIGURE 2.4.1 MA8417-203 SPDT COMPLIMENTARY OUTPUT DRIVER SCHEMATIC

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2.5 Switch Design

The layout of the microstrip portion of the switch is shown in Figure 2.5.1. The circuit fits in a narrow channel in the housing, and each arm mates with a launcher as shown in Figure 2.2.1. As one might expect, the design is very much like a microstrip switch at lower frequencies. The principal differences are the details of the diode mounting, the bias network, and, of course, the launchers.

We experienced some difficulty with the dc block, and we explored numerous possibilities before finding a suitable solution. The design selected consists of a .010" thick metalized titanium oxide ceramic chip, mounted on edge in a .010" gap in the line as shown in the cross-sectional view. These blocks have very low loss and, although they do tend to mismatch the line slightly, they can be readily tuned with short stubs. Figure 2.5.2 shows the response of a 50 ohm line between two transistions with four dc blocks in place (with additional tuning). The loss penalty for the four blocks is roughly 0.1 dB, and the return loss is better than 18 dB across the band.

The diodes in each arm should be electrically spaced 90° apart and the first diode should be 90° from the T-junction, at the center frequency of 33 GHz. The electrical lengths of the microstrip lines, and the physical spacings must be adjusted to compensate for the effective lengths of the diode bond wires and the blocking capacitors adjacent to the tee. We determined the spacing empirically. The final values are consistent with the optimum spacing calculated for a bond wire inductance of 0.12 nH and a negligible line length for the dc block.

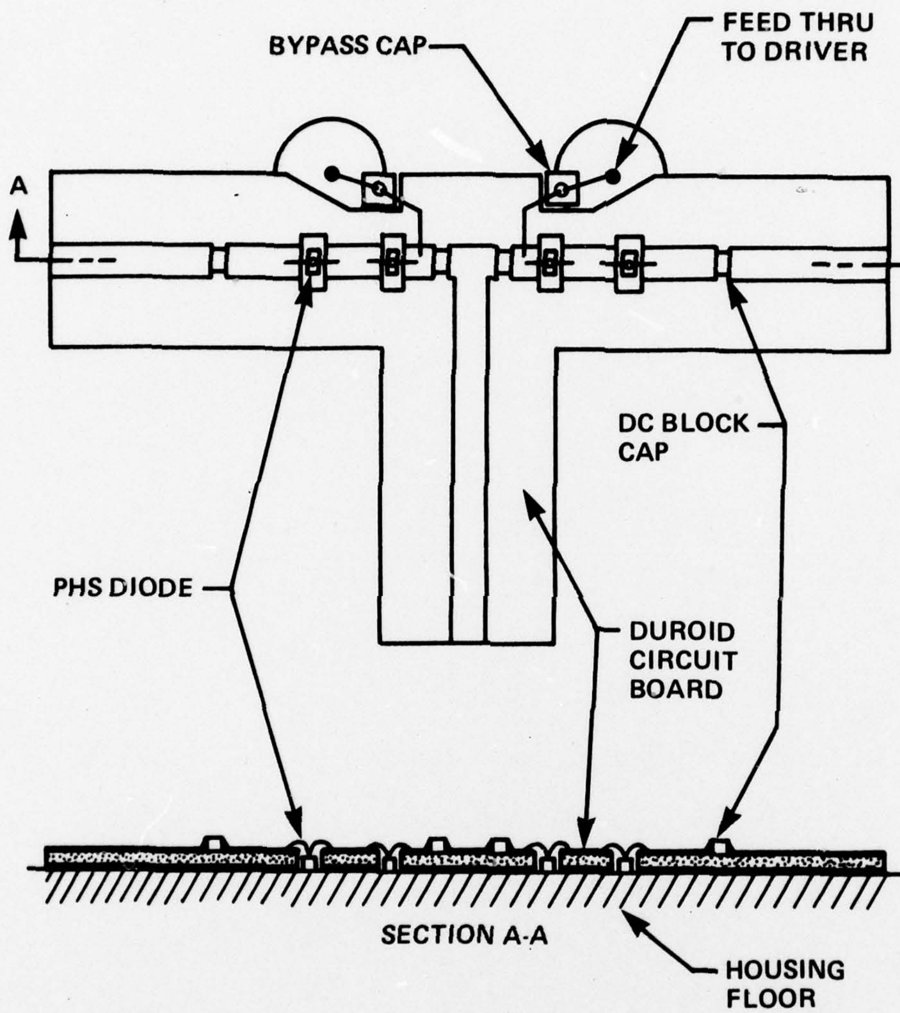

FIGURE 2.5.1 CIRCUIT BOARD LAYOUT OF MICROSTRIP SPDT SWITCH

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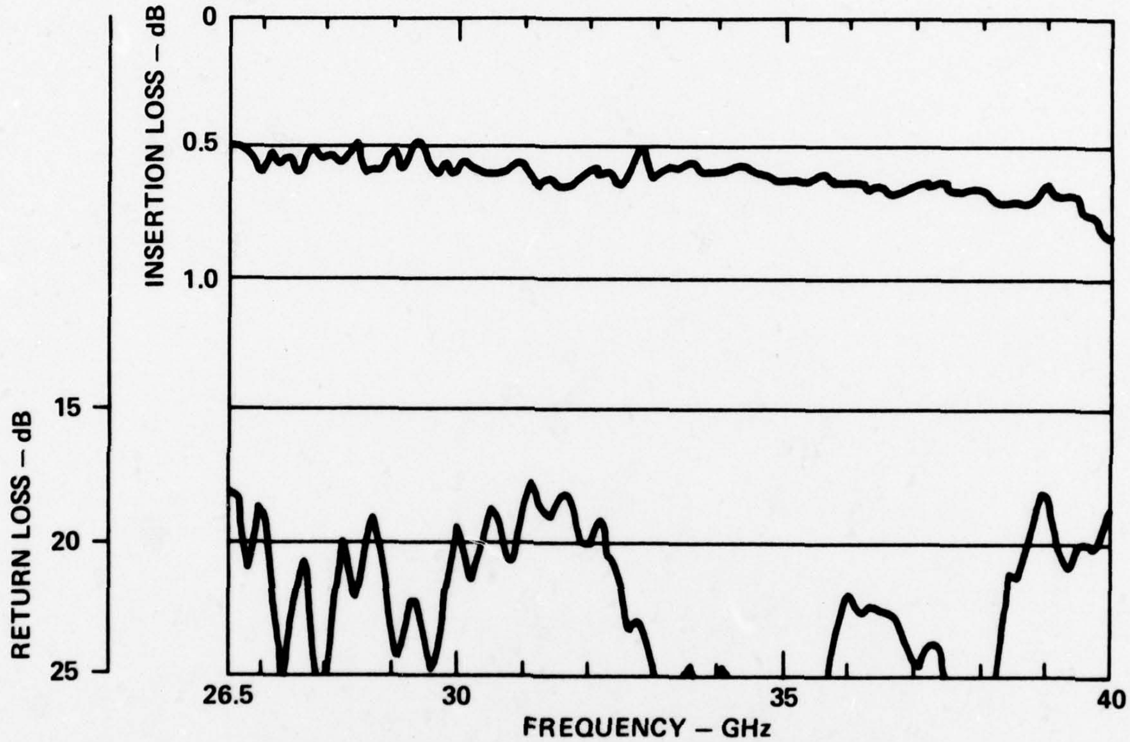

FIGURE 2.5.2 RESPONSE OF 50Ω LINE WITH DC BLOCKS IN FINAL SWITCH HOUSING

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The PHS diodes are about 1/3 the thickness of the circuit board. To help minimize the bond wire length and to make the diodes more assessible for bonding, they are mounted on shims which raise the ground plane in the holes by 3-4 mils. Because the area around the diode mesa is copper and not SiO_2 as on conventional chips, extra care must be taken to insure that a bond wire does not touch this base and short out. Thus, the diodes are best mounted 3-4 mils below the top of the microstrip line. Contact is made to the diodes and the microstrip line with .0008" gold wire.

The bias networks consist of .0008" gold wires approximately a quarter wavelength long from the microstrip line to 20 pF alumina chip by-pass capacitors and to the fee-thru terminals located at the edge of the board. The feed thrus lead through the housing floor to the driver housing below. Broadband, low frequency switches use miniature multiturn coils for bias chokes, however these cannot be used here since K_a-band is well above the self-resonant frequency of any such coil.

The complete switch is shown in Figure 2.5.3. The main housing is 1.5 x 1.2 x 0.44". The overall size, including launchers, flanges, and connector is 3.64 x 2.47 x 0.75, and it weighs about 3 ounces. The waveguide inputs are humidity sealed with teflon-figerglass windows.

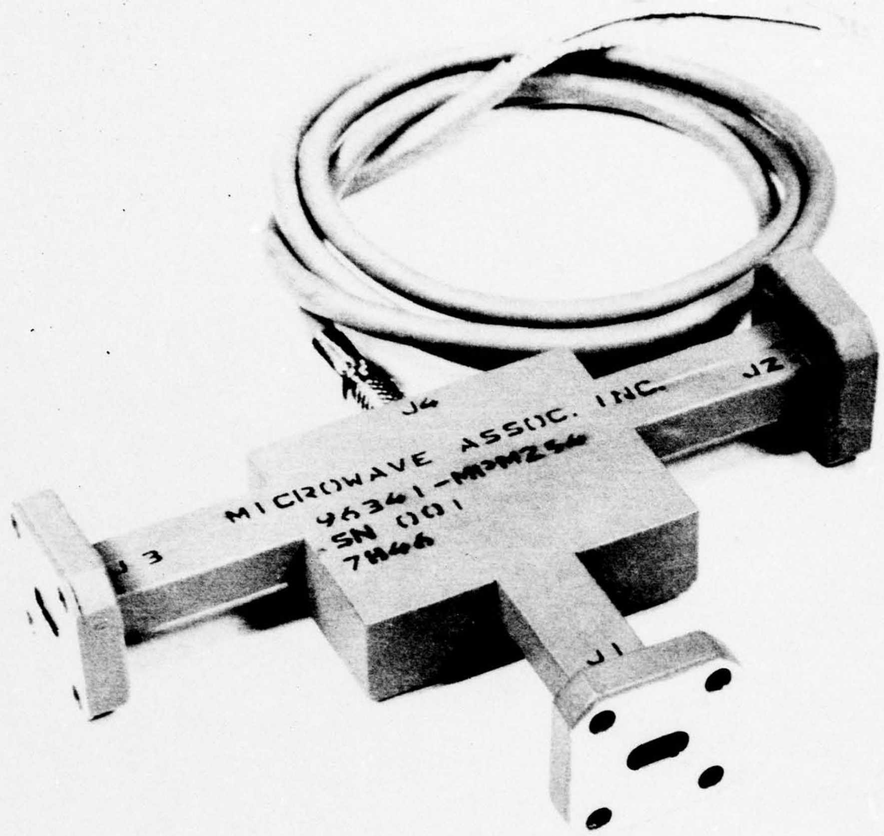

FIGURE 2.5.3 K_A-BAND SPDT MICROSTRIP SWITCH

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3.0 EXPERIMENTAL RESULTS

3.1 Switching Performance

Figures 3.1.1 and 3.1.2 show the insertion loss, return loss, and isolation performance in each state over the 22 to 40 GHz range. The switch was tuned while observing only the 26.5 to 40 GHz responses; we then recorded the responses down close to waveguide cutoff simply to see what the resulting performance was.

Over the 26.5 - 40 GHz design band the insertion loss for both throws is under 2.5 dB, the return loss is better than 14 dB (VSWR $<$ 1.5) and the isolation is greater than 30 dB. The insertion loss tracking of the two arms is within 0.4 dB.

The insertion loss rolls off above 30 GHz from approximately 1.4 to 2.3 dB in one direction and 1.6 to 2.4 dB in the other. This roll off is somewhat greater than should occur due to the normal increase in loss with frequency. We would expect the maximum loss due to this cause to be in the 1.9 - 2.1 dB range. In fact, the J1-J2 arm exhibits such behavior up to 38 GHz, above which the loss rather abruptly increases. The excess loss cannot be attributed to high bond wire inductance or other mismatch, since the match actually improves in this region. It also does not appear to be due to higher order modes or radiation, since the structure should be cutoff to the higher modes and the behavior is the same with and without the cover plate.

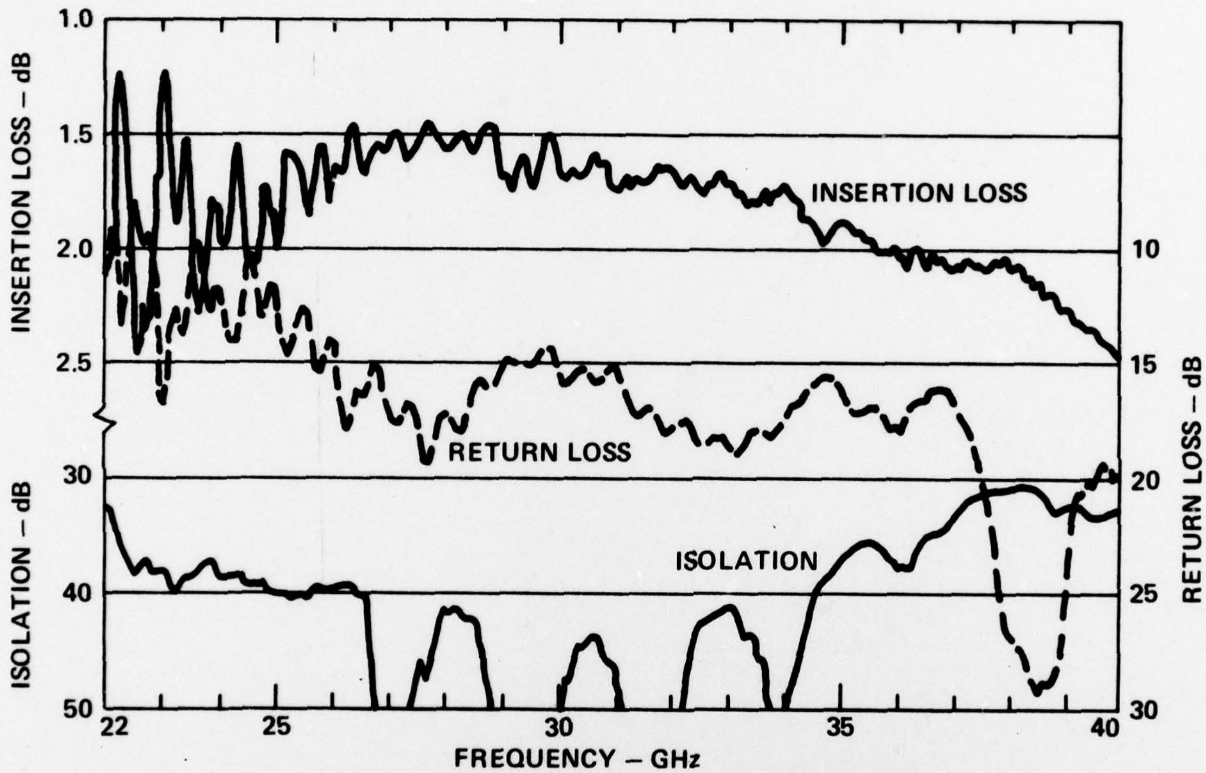

FIGURE 3.1.1 TRANSMISSION RESPONSES OF MIC SWITCH #1, J1 TO J2. REFLECTION RESPONSE CORRESPONDING TO INSERTION LOSS CASE.

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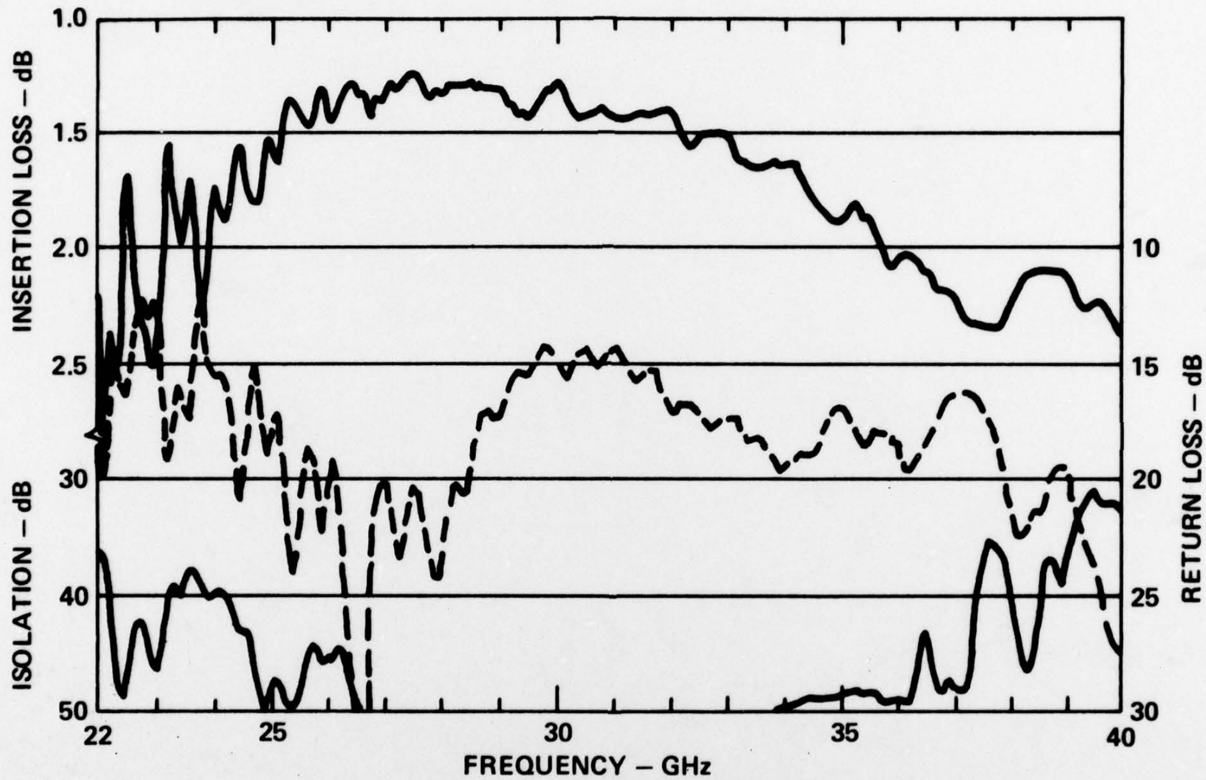

FIGURE 3.1.2 TRANSMISSION RESPONSES OF MIC SWITCH #1, J1 TO J3. REFLECTION RESPONSE (CORRESPONDING TO INSERTION LOSS CASE.)

D-17016

We were not successful in determining the cause of this excess loss on this program, but we suspect that there is a low-Q lossy resonance associated with the launcher-bridge area. Clearly, eliminating this loss contribution could improve the insertion loss at the high end by 0.3 to 0.4 dB, bringing it close to, or within, the 2.0 dB design goal.

The maximum isolation was well over 50 dB, (system sensitivity was approximately 55 dB) which implies that the diode series resistance is, in fact, under 2 ohms as expected. There are several spurious responses between 27 and 34 GHz in the J1-J2 arm, but the isolation remains above 41 dB. The isolation of both arms degrade below 50 dB below 26.5 and above 34 GHz, with some spots close to 30 dB. The reason for these spurious responses was not explored. One possibility is that the second *harmonic* output of the sweeper source is much less attenuated by the switch than the fundamental, due to finite inductance in series with each diode.

Below 26.5 GHz, the performance is somewhat obscured by sharp fluctutations in the return loss due to interference between widely spaced mismatches as the WR-28 waveguide cutoff is approached. There is also a general decrease in the return loss, especially when switched to transmit through the J1-J2 arm. Nevertheless, it appears that the microstrip switch still performs acceptably down to 22 GHz, and that it would only be necessary to extend the bandwidth of the launcher to extend the bandwidth of the entire switch. To cover the full 18-40 GHz range, however, a transition from ridged waveguide or coaxial line, and a transformer section in the T-junction would be required.

The switching transients are shown in Figure 3.1.3 which is traced from sampling oscilloscope photos. The RF turn-off transient--the diodes going from reverse to forward bias--is 1.6 ns for 10 to 90 percent transmission while the turn-on transient takes 3.4 ns. These are both well under the 5 ns design goal, and do not use the current spiking capability of the driver. The bias current was approximately 20 mA.

The driver delays shown are measured from the driver input pin and do not include the delay through the cable. The input pulse to the driver from the HP 8082A pulse generator had a 5^V amplitude and less than 1 ns risetime when fed to a 50 ohm matched load. The driver loads the pulser such that the rise and fall times are several nanoseconds long, and on the rise, it levels off at 3 volts for 6-8 ns before continuing up to 5 volts. The total switching times shown of 8.4 ns and 18 ns are measured arbitrarily from the 2.5 volt level, i.e. "50% TTL", which is within the required 20 ns speed. Current spiking would shorten the turn-on delay several nanoseconds.

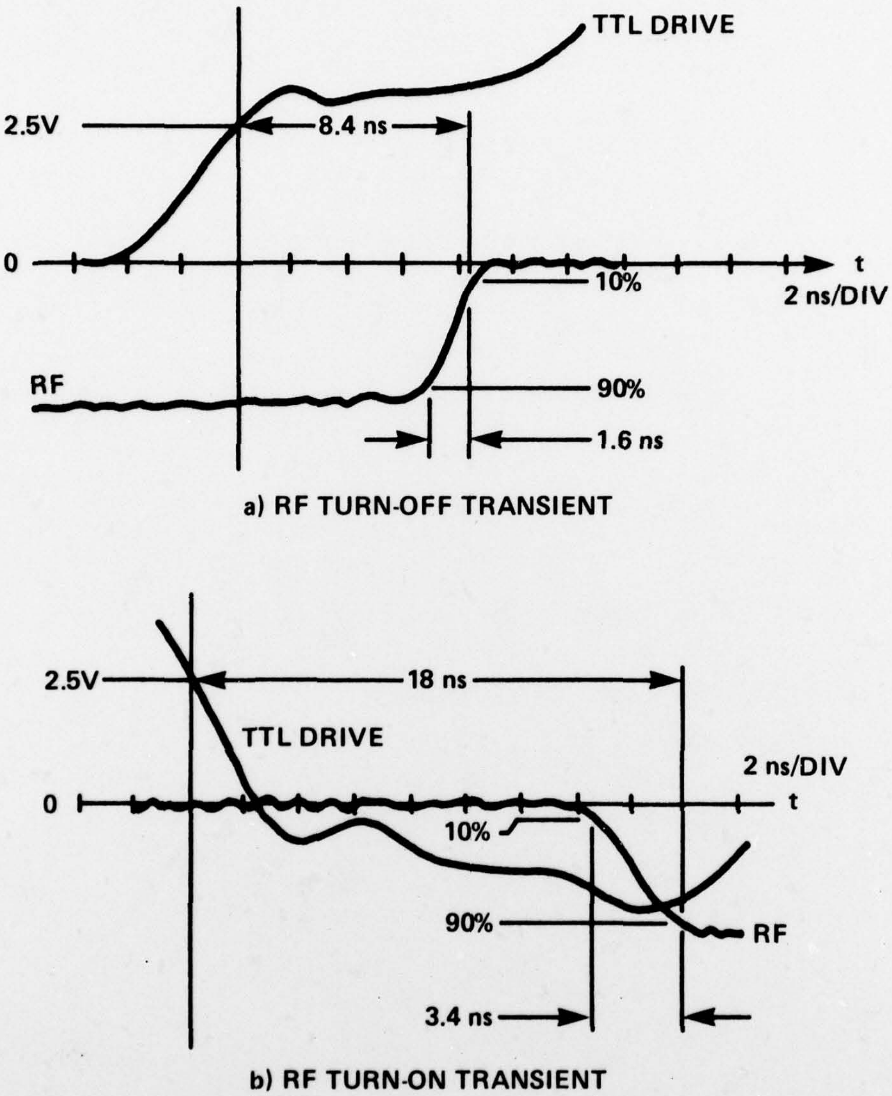

FIGURE 3.1.3 SPDT MIC SWITCH – SWITCHING TRANSIENT

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3.2 Power Handling

Two watts CW at K_a-band is not generally available. Therefore, in lieu of high power testing the K_a-band switch directly, we evaluated its performance indirectly at 17.5 GHz. We assembled a 2-diode SPST switch with coaxial input and output, using the PHS diodes mounted as in the K_a-band switch, but with the diode spacing appropriate to K_u-band.

Burnout of the PIN diode is not a problem at 2 watts CW, even if the switch were terminated in a short circuit. PIN diodes are safe from burnout in the reverse biased state if the junction temperature remains below 125-150°C and the RF voltage remains below the breakdown voltage. Forward biased diode are safe up to more than 200°C and are relatively immune to burnout. For 2 watts into a switch with 2.5 dB of insertion loss, only 0.9 watts will be dissipated when the output is matched. Roughly one third of this is dissipated in the circuit. The other two thirds is fairly evenly divided among the two reverse biased diodes and the forward biased diode nearest the T-junctions, i.e. about 0.2 watts per diode Thus, with a thermal resistance as high as 80°C/watt--twice the rated value for the diode--the temperature rise will be only 16°C, well within the safe region even for case temperatures of 70-80°C in a typical military environment.

At 2 watts, the diode RF voltage will be about 15 volts in a matched line and it could be about 30 volts in a shorted line. This is well below the breakdown voltage of 90 volts and is therefore also in the safe region.

At the 2 watt level, the potential high power problem is possible degradation of insertion loss or isolation due to nonlinear effects in the diodes. In the reverse biased state, carriers are injected into the I-layer during the forward bias half of the RF cycle. At high RF voltages this carrier injection makes the I-region lossy, increasing the effective series resistance of the diode as the RF voltage increases. In the forward bias case, when the RF current approaches the magnitude of the dc bias current, partial carrier sweep out can occur, increasing the series resistance and reducing the isolation. These nonlinear effects become less efficient as the RF frequency increases since the RF period become steadily shorter relative to the carrier relaxation or response times and the carriers become less able to follow the RF voltage. Consequently, we expect high power degradation at K_a-band to be less than any observed at K_u-band.

In the SPST test switch, at 17.5 GHz, we measured the variation in insertion loss and isolation with input power up to 2.5 watts input for bias levels of 10V, 0V, 10 mA, and 40 mA. There was no change in insertion loss (i.e. less than 0.1 dB) at 10V and 0V. At 2 watts the isolation decreased by 0.9 dB at 10 mA and 0.5 dB at 40 mA; at 2.5 watts it decreased by 1.9 dB and 0.7 dB at 10 and 40 mA. Thus we expect that the K_a-band switch will suffer a negligible decrease in isolation at least to the 2 watt level with 10 mA or more bias.

The diodes also survived hot switching up to 2.5 watts. Hot switching was done simply by making and breaking contact with the bias lead, so that the switching speed was governed by the carrier lifetime, about 40 ns in this case, and was therefore much slower than it would be with a fast driver.

Note that the diode voltage at 2.5 watts input is equivalent to the maximum voltage that could occur at 2 watts with a 2:1 output VSWR, in the absence of losses. A similar statement holds for the diode current. Thus, the experiment at 2.5 watts with a matched load represents a slightly more severe test than 2 watts with a 2:1 output mismatch in terms of the electrical stresses on the diodes.

4.0 CONCLUSIONS AND RECOMMENDATIONS

The K_a-band microstrip SPDT switch developed on this program met all of the design requirements set forth on page 1. Also, it met or exceeded the design goals for insertion loss and isolation from 26.5 to 34 GHz, for amplitude tracking over the entire K_a-band, and for the RF transition speed. As experience is gained in building future units, we expect to improve the insertion loss and isolation above 34 GHz to 2.1 dB max. and 40 dB min., respectively.

The plated heat sink diodes, with their low capacitance, high Q, and thin I-layers, are superbly suited to this high speed, millimeter wave switch application. They were the key to the switch performance attained on this program.

The switch will readily handle 2.5 watts of CW power with, at most, a slight ($<$ 1 dB) decrease in isolation above 1.8 watts. We believe that with higher bias current, 40-50 mA, acceptable performance will be achieved in the 3 to 4 watt range.

There are three particular areas in which future development work on this switch should be undertaken.

The first is to ascertain the causes for the high frequency rolloffs in insertion loss and isolation. Elimination or reduction of these effects will result in a K_a-band switch with performance comparable to X-band switches using conventional diodes.

Second is to broadband the switch to cover the 18-40 GHz frequency range. This involves two separate aspects, the microstrip circuit and the transitions. The circuit work would be addressed to improving the diode mountings to reduce the series inductance, and to adding a broadbanding transformer in the T-junction. The broadband microstrip portion alone would then be suitable for integrated circuit package, since transitions to waveguide would not be necessary.

The design of a broadband launcher will depend, of course, on the type of feed guide to be used. A broadband waveguide feed would no doubt be either single or double ridge guide; the design of new tapers to a ridge guide microstrip launcher should be relatively straightforward.

The third area is to look at additional methods to reduce production costs. The waveguide-microstrip transition deserves particular consideration in this respect. A transition, etched on the same circuit board as the microstrip line, as illustrated in Figure 4.0.1, has been developed at NOSC.* Such a transition would not only reduce materials costs and labor costs, since no manual tuning would be required, it would also result in a much smaller, more compact package.

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* D. Rubin, D. Saul, Naval Ocean Systems Center, San Diego, CA, private communication.

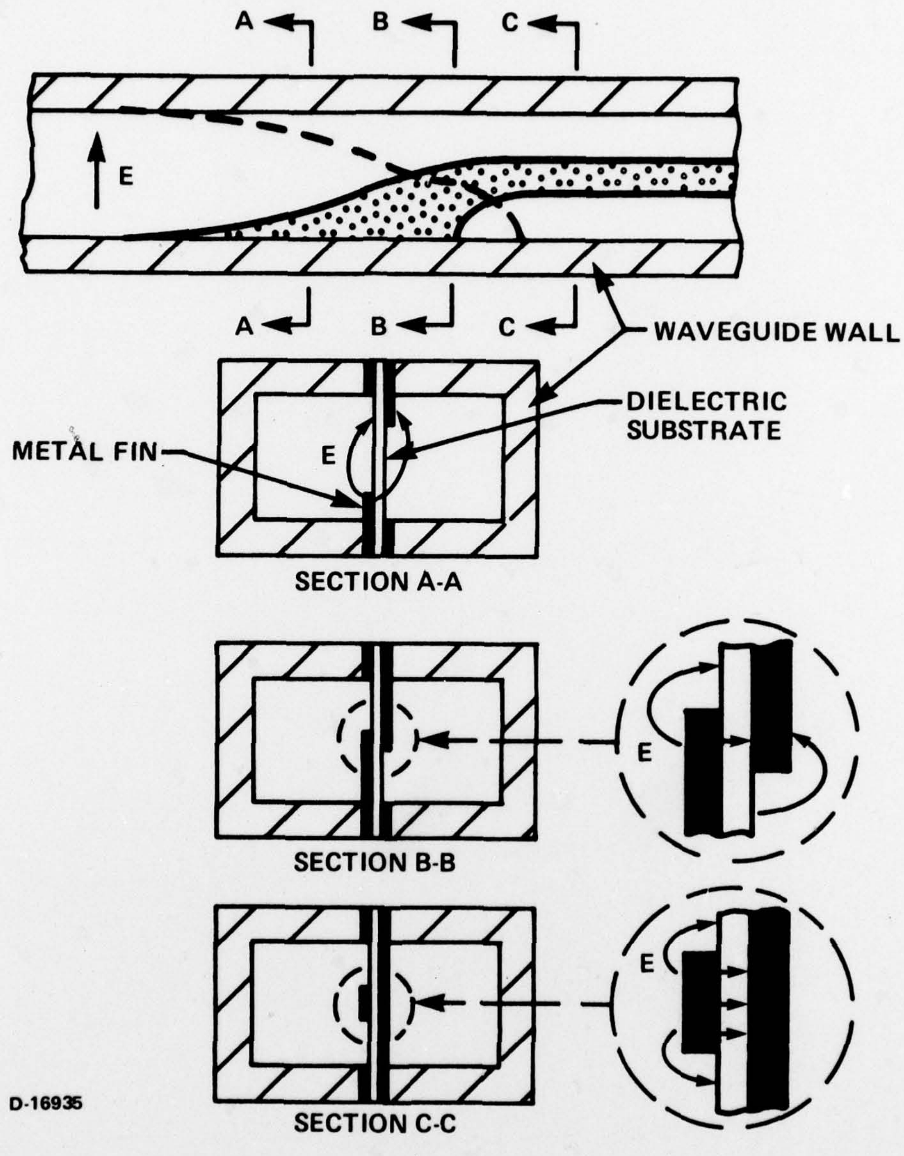

FIGURE 4.0.1 WAVEGUIDE-MICROSTRIP TRANSITION

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The report describes the design and development of a K_a-band (26.5 - 40 GHz) microstrip microwave integrated circuit SPDT switch with driver.

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The report describes the plated heat sink PIN diodes and the waveguide to microstrip transitions used in the design. Final performance data are presented and discussed.