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SLIDING BACKSHORTS FOR PLANAR CIRCUITS

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Abstract — **The Superconductor-Insulator-Superconductor (SIS) tunnel junction is an extremely sensitive heterodyne detector at millimeter and submillimeter wavelengths. The large inherent capacitance associated with this device results in a substantial impedance mismatch with typical antennas and, therefore, requires a tuning circuit for optimum results. At frequencies where waveguide dimensions are realizable, impedance matching can be accomplished by embedding the detector in a waveguide circuit with adjustable waveguide backshorts. At higher frequencies, where waveguide dimensions become prohibitively small, a planar transmission line embedding circuit provides a reasonable alternative. Typically, such planar circuits offer no post-fabrication adjustability, resulting in demanding materials and design requirements. An adjustable planar embedding circuit based on coplanar transmission lines with movable non-contacting shorting elements has been developed. The shorting elements each consist of a thin metallic plate with an optimized arrangement of rectangular holes, placed along the insulated metallic transmission line to provide a periodic variation of the line impedance. A scale model (1–5 GHz) has shown that a large reflection coefficient, $|s_{11}| \geq -0.5\,\mathrm{dB}$, can be achieved with these sliding elements. A low frequency tuning circuit incorporating these shorting elements has been tested to demonstrate practical tuning ranges.**

I. Introduction

Low noise, high conversion efficiency, and low local oscillator power requirements have made the superconductor-insulator-superconducter (SIS) tunnel junction the mixer of choice for a wide range of millimeter and submillimeter wave heterodyne receiver applications [1,2]. However, SIS junctions have a large inherent capacitance which results in an impedance mismatch when the detector is used with typical embedding circuits. It is therefore desirable to design a receiver which incorporates an adjustable impedance matching circuit to fully optimize the mixer.

SIS mixers with excellent performance have been demonstrated throughout the millimeter wave band using waveguide embedding circuits with one or two adjustable tuning elements [3,4]. At terahertz frequencies however, the very small physical dimensions of a waveguide become extremely difficult to fabricate accurately, and installing the SIS junction is a serious problem. If an array of mixers is desired, these difficulties are compounded. At this point, planar circuits are an attractive alternative. SIS junctions have a planar geometry and are readily integrated with planar transmission lines and antennas. Planar circuits, however, usually support tuning elements with only fixed dimensions. These are often in the form of open or short circuited tuning stubs. This lack of adjustability usually means that an effective circuit design requires an extremely well characterized SIS junction and substrate, as well as an extremely well controlled circuit fabrication process. This may not always be possible at these high frequencies [5,6].

An approach for a movable, non-contacting planar backshort which can be used to vary the electrical length of a coplanar strip transmission line tuning stub has been developed. This sliding circuit element allows for the design of an embedding circuit with a complex impedance that can be optimally adjusted to match the SIS device after the circuit has been fabricated.

II. The Planar Backshort

Placing a solid metallic plate across a coplanar strip transmission line, with a thin dielectric insulating layer in between, results in a reflection of RF power. Unfortunately, this reflection is not sufficient over any bandwidth for the plate to be used as a backshort in the design of a practical tuning circuit. This "sandwich" configuration does, however, result in a section of lower impedance transmission line. Quarter-wavelength sections of this line can be cascaded alternately with uncovered sections of "high" impedance line to create a series of impedance transformations which ultimately result in a very low impedance. When used as a backshort on a transmission line, this cascade produces an even larger reflection of RF power than the solid plate. We constructed and optimized such a planar backshort and it's geometry is illustrated in figure 1. It consists of a thin metal plate with rectangular holes of the proper dimension and spacing. A thin insulator keeps the plate from contacting the transmission line which reduces wear and allows the backshort to slide freely when pushed with a probe.

III. Model Measurements

We built a large scale model of the sliding backshort and measured the magnitude of the reflection coefficient with an HP 8510B network analyzer over a frequency range

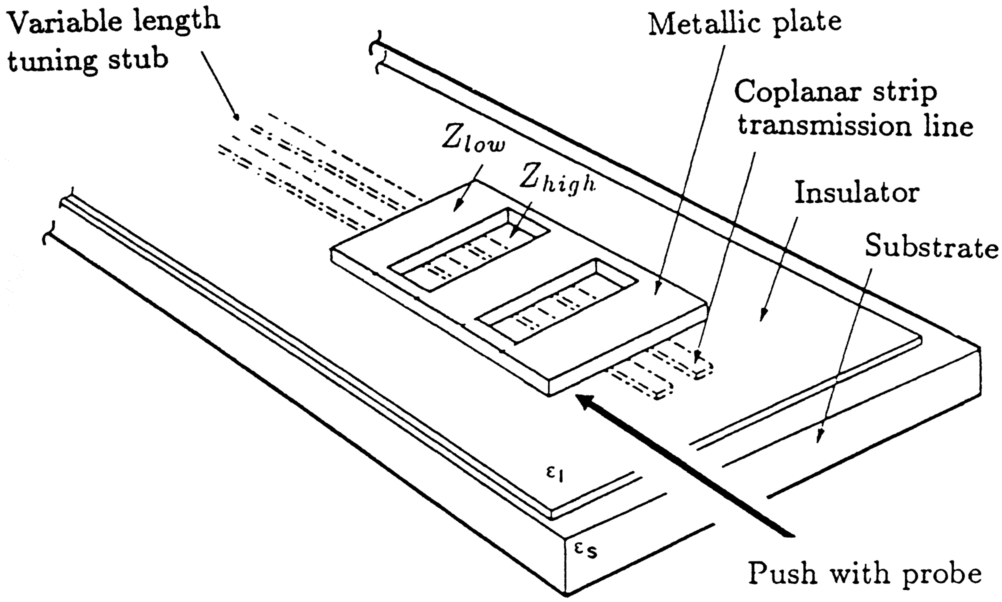

Figure 1. Design of sliding backshort on coplanar strip transmission line. The holes in the metal plate create a series of successive high and low impedance sections that produce a large reflection of RF power.

of 1 GHz to 5 GHz. This measurement required a transition between the coaxial input of the network analyzer and the coplanar line. Unfortunately, no standard transition with low VSWR is readily available. For this reason, we devised several distinct measurement techniques in order to obtain reliable results.

A natural choice of calibration technique for an HP8510B is the "Thru-Reflect-Line" (TRL) method which allows for measurements in nonstandard transmission media such as coplanar line. The transition between coaxial and coplanar line can, in principle, be accounted for in the calibration procedure. However, reproducible calibration standards in coplanar line are required. Reflections at connections between segments of coplanar line and uncertainties in the reflection standard lead to non-repeatable results and unacceptably high errors. As a result, this method was not pursued further.

Our second technique was a simple 1-port measurement of a circuit which employs a direct connection from coaxial to coplanar line. This measurement of $|s_{11}|$ was made over a 1.5 GHz to 2.5 GHz frequency range and figure 2(a) shows the test arrangement. The abrupt transition from coaxial line to coplanar line was formed at the edge of the stycast substrate with a flange mount SMA connector. The measurement was taken with the reference plane of the backshort adjusted to coincide with the SMA connector.

While this transition resulted in large unwanted reflections which increased the uncertainty of the measurement, the technique was useful because it allowed us to monitor the reflection coefficient for the backshort in real time as we optimized the dimensions of the backshort. The optimization was performed by systematically varying the

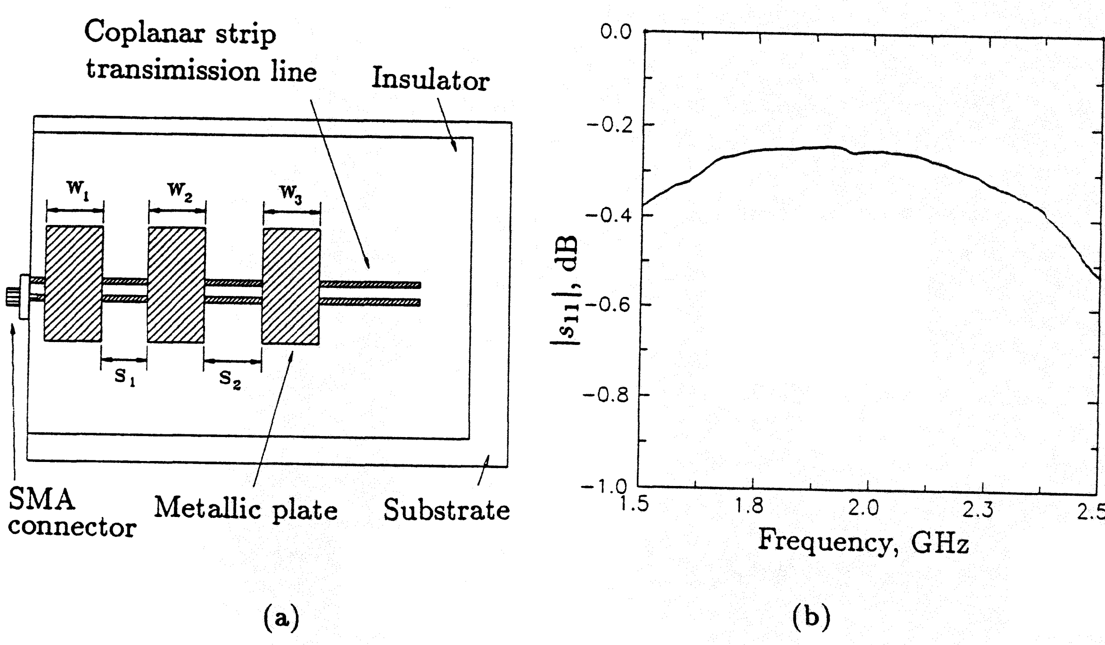

Figure 2. Test circuit used for optimization (a) and plot of reflection coefficient (b) for optimized sliding backshort at the SMA connector on 204-Ω transmission line.

length of the low and high impedance sections, the number of sections, the dimensions of the transmission line, and the thickness of the insulator in order to achieve the best reflection of RF power. Good performance was obtained for a coplanar transmission line with 2.1 mm wide copper strips, separated by a 5.2 mm gap and mounted on a 6 mm thick stycast substrate with a dielectric constant of 4. The characteristic impedance of this transmission line and its effective dielectric constant were determined to be 204 Ω and 2.3 respectively [7]. A 0.025 mm thick sheet of mylar was used to insulate the transmission line from the sliding shorting plate. This non-contacting, 76 mm wide, 6 mm thick aluminum shorting plate had two rectangular holes in it with dimensions and spacing of w_1 = 24.3 mm, s_1 = 19.8 mm, w_2 = 24.3 mm, s_2 = 23.5 mm, and w_3 = 24.3 mm. This resulted in uncovered high impedance sections, s_1 and s_2, and covered low impedance sections, w_1, w_2 and w_3, which were each approximately $\lambda_g/4$ long on the coplanar line.

A plot of $|s_{11}|$ versus frequency is shown in figure 2(b). This optimized planar backshort produced $|s_{11}|$ better than -0.3 dB over a 20% bandwidth. That is a reflection of more than 90% of the power in the incident wave. The center frequency was determined to be 2 GHz, which agrees exactly with the design frequency.

We noted that the front edge of the sliding metallic plate was very close to the discontinuity at the SMA connector and it may have interacted with fringing fields around the flange. In addition, the reflection coefficient decreased when the backshort was moved $\lambda_g/2$ from the SMA connector in order to reduce interactions with fringing fields. The error caused by this unwanted interference, along with the lack of a transmission measurement, motivated the design of a third measurement technique.

Figure 3 shows the system used for the third measurement technique. Here, the 204-Ω coplanar line was connected to the 50-Ω network analyzer inputs by means of two baluns of identical length. These baluns were made by gradually trimming the shield and teflon insulation from a semirigid, 3.5 mm wide coaxial line over approximately one wavelength at 2 GHz. This created a smooth transition to the coplanar line which minimized the power reflected at the connection. The return loss for these baluns is approximately −10 dB and the remaining undesired reflections from these transitions were gated out of the measurement using the low-pass time domain mode of the network analyzer. The frequency for this measurement was swept from 50 MHz to 20 GHz so that an accurate transformation between frequency and time could be made.

The full two-port scattering parameters for the system were measured under three different conditions. First, a reference measurement was made which would correspond to an ideal short. The baluns are identical in length and hence, the test model is symmetric about the midpoint of the coplanar line. Thus, the magnitude of the transmission measurement of this circuit with no short in place is equal to the reflection measurement with an ideal short at the midpoint. Reflection measurements for the sliding backshort were then made with the shorting plate arranged to reflect an incident wave from port 1 at the midpoint of the system and then, port 2. The values for $|s_{11}|$ from the first reflection measurement and $|s_{22}|$ from the second were normalized by dividing each by the values for $|s_{21}|$ and $|s_{12}|$ from the reference measurement, respectively. The two results were averaged to cancel the effect of any asymmetry in

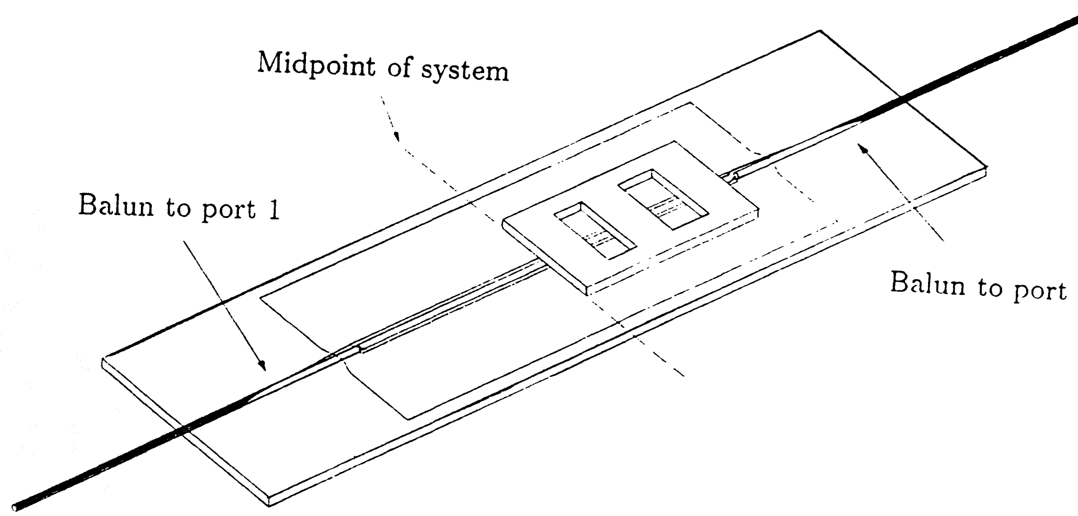

Figure 3. Test system used to measure two-port scattering parameters for sliding backshort. The coaxial line baluns are tapered to create a gradual transition between coaxial and coplanar line which reduces measurement uncertainty due to unwanted reflections at the transition.

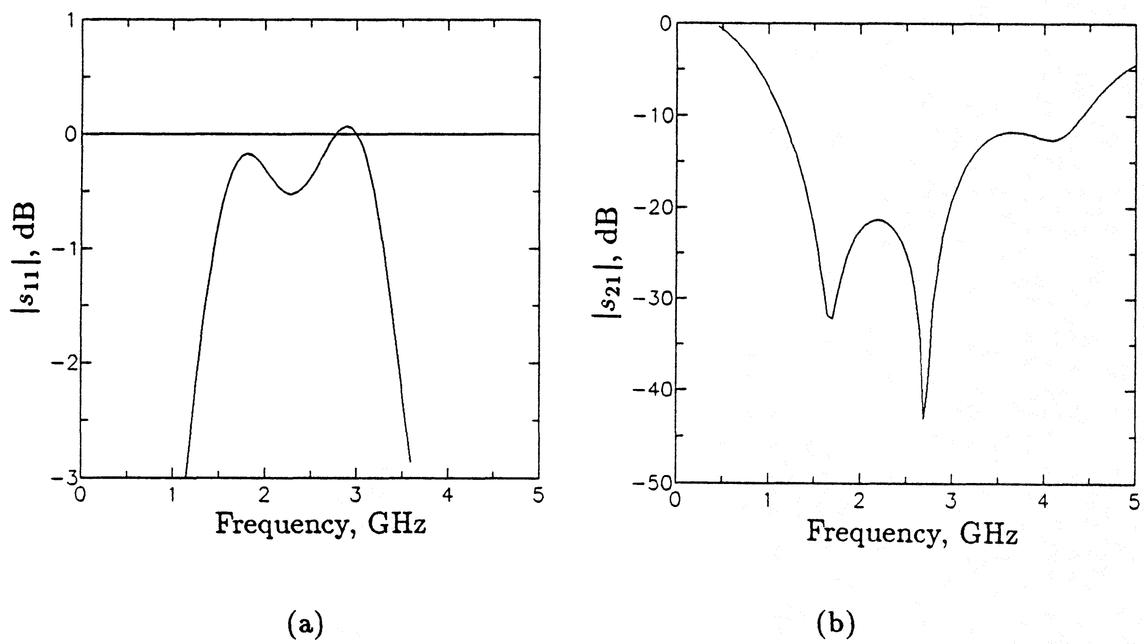

Figure 4. Plot of measured reflection coefficient (a) and transmission coefficient (b) for optimized sliding backshort. This measurement was made using baluns for transitions.

the system. Transmission measurements for the backshort were similarly normalized and averaged. The requirement of processing the measurement data, along with the large range of frequency, prohibited us form monitoring the 2 GHz reflection coefficient in real time for this measurement.

The results for the averaged, normalized $|s_{11}|$ and $|s_{21}|$ are shown in figure 4. The plot shows that $|s_{11}|$ was better than –0.5 dB over approximately an 80% bandwidth and the center frequency was slightly higher than 2 GHz. Over the peak located between 1.5 GHz and 2.5 GHz, $|s_{11}|$ is better than –0.3 dB over a 16% bandwidth and the center frequency is slightly lower than 2 GHz. This agrees well with the previous results shown in figure 2(b). Together, the plots of $|s_{11}|$ and $|s_{21}|$ appear to indicate that approximately 10% of the power for the incident wave is left unaccounted for, but the $\pm$ 0.2 dB uncertainty of our measurement is to large to verify this.

IV. Double Stub Tuner

In order to demonstrate the tuning range accessible in a planar circuit, we built a double-shunt stub tuner which incorporates two sliding backshorts. Figure 5(a) shows the equivalent circuit and figure 5(b) shows the circuit arrangement as measured. The characteristic impedance of the coplanar line is 90 Ω and the stub spacing is $\lambda_g/8$. A 90-Ω resistor was used to simulate a planar antenna impedance and a 3.5 mm wide semirigid coaxial probe was used to measure the range of impedance to which we could transform the resistor. The calibration reference plane for this measurement was set at end of the shield for the coaxial probe. Figure 6 is a Smith chart, normalized to

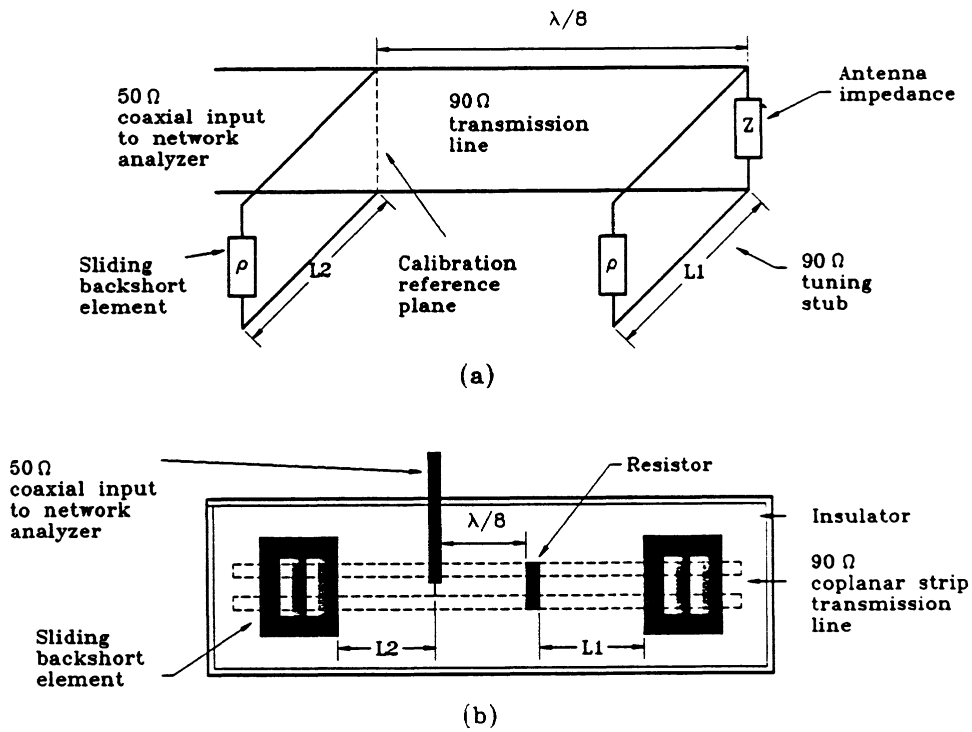

Figure 5. Schematic diagram for equivalent double-shunt stub tuner circuit (a) and circuit arrangement used for measurements (b).

the characteristic impedance of the line, which shows the accessible impedance region at 2 GHz. The overlap for the tuning region and the impedance region necessary for matching SIS devices implies that a circuit of this type might be useful for this purpose. Variations in the shape of this tuning region can be achieved by changing the spacing between the tuning stubs.

The solid and dashed boundary lines in figure 6 show a comparison between the measured impedance range of the double stub tuner and a computer simulation of the circuit using *Puff* [8], respectively. The simulation agrees well with the measured data. The reflection coefficient for the shorting element used in the computer model, a resistor, along with a phase shift at the probe tip to account for the calibration reference plane uncertainty, were fitted to match the measured Smith chart data. The resulting phase shift was 4 degrees and $|s_{11}|$ for the backshort model was –0.5 dB, which is slightly better than the –0.7 dB result that we measured for the same backshort by itself on a 90-Ω transmission line at 2 GHz.

V. Conclusion

We have demonstrated an approach for an adjustable planar backshort on coplanar strip transmission line. Results from a low-frequency model indicate that the backshort can be used to create tuning stubs whose electrical length can be varied after fabrication. This non-contacting backshort with cascaded high and low impedance sections should also work on slotline, coplanar waveguide and possibly microstrip line. By using advanced micro-machining techniques [9,10], it should be possible to create

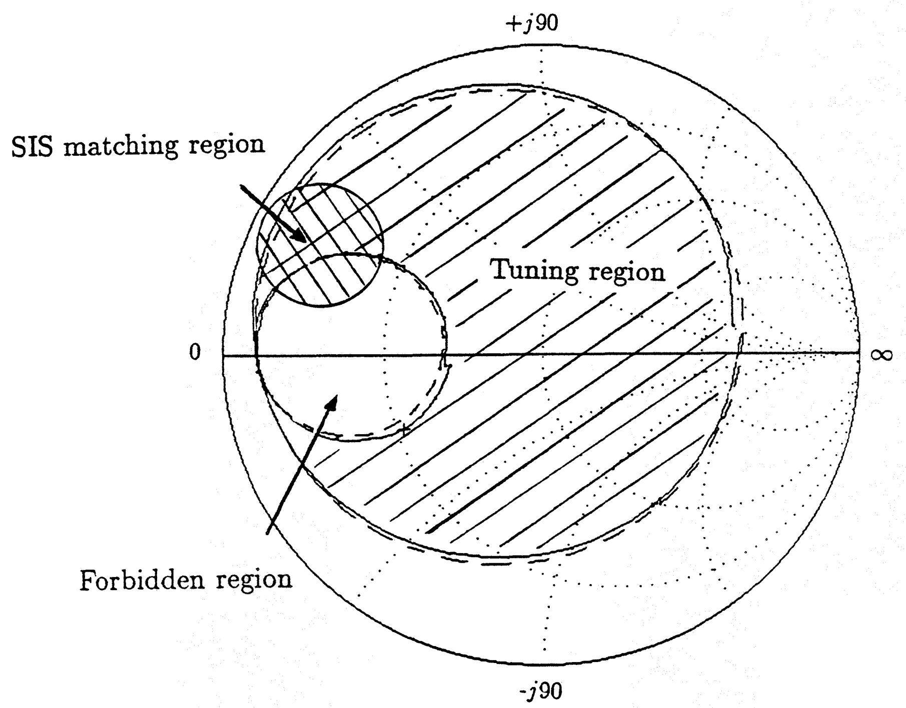

Figure 6. Smith chart showing measured (solid boundary) and fitted (dashed boundary) tuning region for double shunt stub tuner. The available tuning region covers the impedance region needed for matching to an SIS device.

adjustable impedance matching circuits at terahertz frequencies which would relax the design constraints for a wide range of integrated planar circuits.

VI. Acknowledgements

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