A NOVEL WILKINSON POWER DIVIDER WITH PREDICTABLE PERFORMANCE AT K AND Ka-BAND

Dimitrios Antsos, Rick Crist and Lin Sukamto

Jet Propulsion Laboratory
California Institute of Technology
Pasadena, California

Abstract

Multi-port microstrip power dividers, used in array antenna applications, frequently utilize the Wilkinson design. The standard asymmetrical Wilkinson design has problems at frequencies beyond X-band, due to coupling between the output branches. A novel, modified design is presented. Its operation can be easily and accurately modeled. Its successful operation is demonstrated using measurements from a 2:1 Ka-band power divider, and a 14-way K-band power combiner.

Introduction

Unequal power-split ratio, equi-phase power dividers and combiners are frequently employed in array antenna and distributed amplifier systems. Figure 1 shows a schematic representation of the popular Wilkinson power divider. The

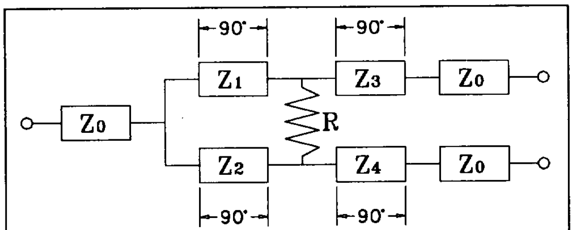

Figure 1 A schematic of the standard Wilkinson power divider.

latter provides matched input and matched, equal phase outputs with unequal power-split ratio and (theoretically perfect) isolation between the output ports. It consists of two branches, of different impedances, Z_1 and Z_2, that realize the desired power-split ratio. The two transmission lines that comprise the two branches have an electrical length of 90°, at the center frequency of the divider. A lumped resistor, R, is connected between the outputs of the two branches and provides the required isolation. Quarter-wave output transformers match the outputs of the divider to the normalizing characteristic impedance, Z_o. The ideal design parameters are given in reference [1].

Usually this Wilkinson design is used at X-band frequencies or below, especially in cases where the desired power-split ratio is significantly different from 1. For frequencies above X-band several problems arise. First, the chip resistors available have a resonance frequency comparable to the operating frequency, and the chip no longer behaves like a lumped element. Second, to achieve a high chip-resistor resonance frequency, the chip dimensions must be very small, of the order of 1×0.5 mm. This means that the two branches of the power divider must be placed very close to each other to be connected to the resistor. This gives rise to strong mutual coupling between the output lines, which destroys the desired power-split ratio. Also, wide low-impedance lines at the higher frequencies frequently often make it impossible to bend the branches in a semi-circle. The most common solution to this last problem is the use of branches of length $\frac{3\lambda}{4}$, rather than $\frac{\lambda}{4}$ [1]. Finally, there is a high translational degree of freedom for the exact placement of the chip-resistor between the two branches of the divider, where they converge at the optimum point. The response of the divider is strongly dependent on the position of the resistor and hence fabrication becomes difficult.

These problems are illustrated in figure 2. There is a high translational degree of freedom for the placement of the chip-resistor. This problem can be solved by making the gradual 180 degree bend of the output lines a sharp corner, at the expense of input return loss. Due to the proximity of the two branches there is substantial coupling between them. This decreases the isolation between the output ports and affects the power-split ratio.

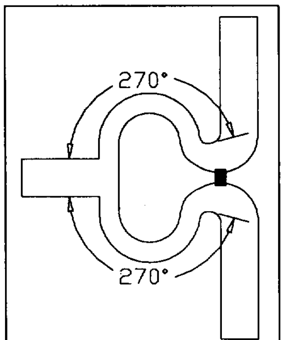

Figure 2 A traditional Wilkinson divider with elongated branches.

The Modified Wilkinson Power Divider

A novel modification to the Wilkinson design that solves the problems mentioned above, at the expense of the bandwidth of the device, is adopted in this paper. The idea is to present the correct impedance at the end of the branches in a way that does not require the close proximity of the latter.

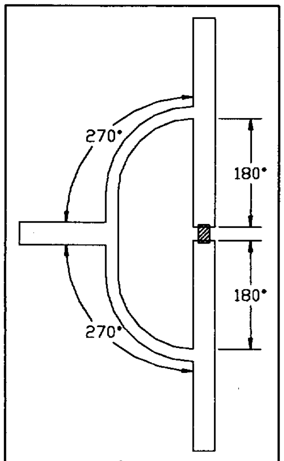

Figure 3 The modified Wilkinson geometry.

This is achieved by "opening up" the circle formed by the two branches into a semi-circle and connecting the resistor to the ends of the branches via $\lambda/2$ lengths of 50 Ω transmission line.

The new geometry is shown in figure 3. The $\lambda/2$ lengths of 50 Ω transmission line transform the impedance presented by the chip-resistor to the output branches at frequencies away from the design frequency of the device, and thus reduce the operating bandwidth. However, the output coupling problem is resolved and hence this design is optimum for highly asymmetric power-split ratios (not close to 1). Furthermore, the chip-resistor placement is easier and more a deterministic process than in the standard design.

Measurement Results of Microstrip Modified Wilkinson Power Dividers

Two examples of actual modified Wilkinson power divider circuits on microstrip that demonstrate good performance are presented below. They are both fabricated on a Thermoset Microwave Material [2] (TMM-3) substrate made by the microwave materials division of Rogers corporation. These substrates have relative dielectric constant 3.25, height 0.381 mm and metalization (copper) thickness 17.78 μm.

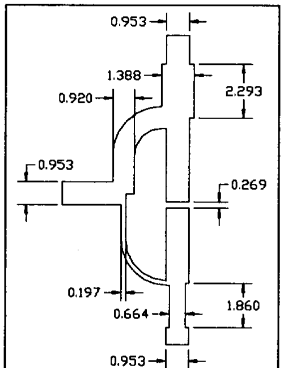

Figure 4 The layout of the 2:1 modified Wilkinson power divider. Dimensions are in mm.

1. A Ka-Band, 2:1 Power-Split Ratio, Modified Wilkinson Power Divider

The design frequency and power-split ratio are 30.5 GHz and 2:1 respectively. Figure 4 shows the layout of the copper on the TMM-3 substrate, with some key dimensions (in mm) indicated. The circuit is designed using EEsof's *Touchstone* and the lengths of the transmission lines are set to their optimum values for minimum integrated squared error between the desired and the calculated response. The quarter-wave transformers at the output arms are necessary to match the output port impedance to 50 Ω. The different line widths of the two branches create the impedance imbalance that provides the desired output power-split. The input T-junction can be further optimized, using a "notch" at the junction point, opposite the input line stub, but the effect is found to be minimal so this technique is not used.

Figures 5, 6 and 7 are plots of measured and calculated performance parameters of the device. Figure 5 is a plot of the return losses at all the ports (S11, S22 and S33) and the isolation between the output ports (S23).
The measured return losses are better than 20 dB at all ports, both at the design frequency and at the frequency of 30.4 GHz, at which the 2:1 power-split ratio is actually achieved (see figure 7). The isolation is better than 17 dB at both frequencies.

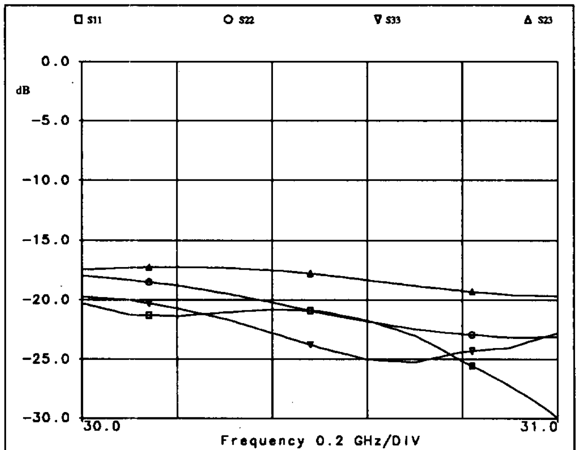

Figure 5 The measured return loss at the 3 ports and the isolation between the output ports.

Figure 6 shows the insertion losses at each of the two output ports. The ideal (due to power division) insertion losses are 1.76 and 4.77 dB.

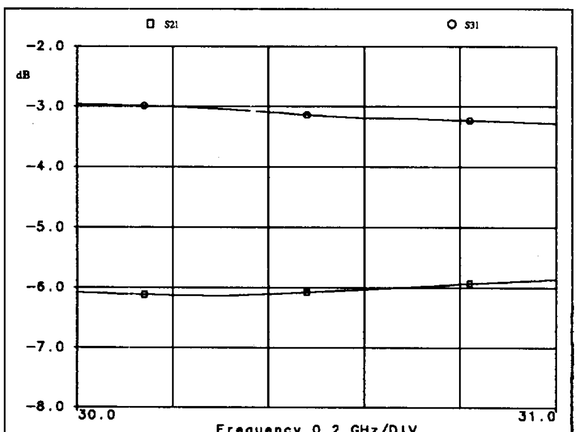

Figure 6 The measured insertion loss at the output ports.

The measured insertion loss is higher than the ideal, because it includes two kinds of losses: the (ideal) division loss and the excess ohmic and radiation losses. The dominant type of excess loss is, in this case, believed to be attributable to radiation because the substrate is small (i.e., there is a small ground plane) and the device is not packaged. Ohmic loss due to surface resistance would be on the order of 0.3 dB at this frequency.

Figure 7 shows the plots of three parameters derived from the measured S-parameters. On the left vertical scale are plotted the parts of the insertion losses in excess of the ideal division loss, at the two output ports. On the right vertical scale is plotted the deviation of the power-split ratio from the ideal (2:1 or 3.01 dB) in dB. The frequency at which the right curve crosses zero is the frequency at which the ideal power-split ratio of 2:1 is achieved.

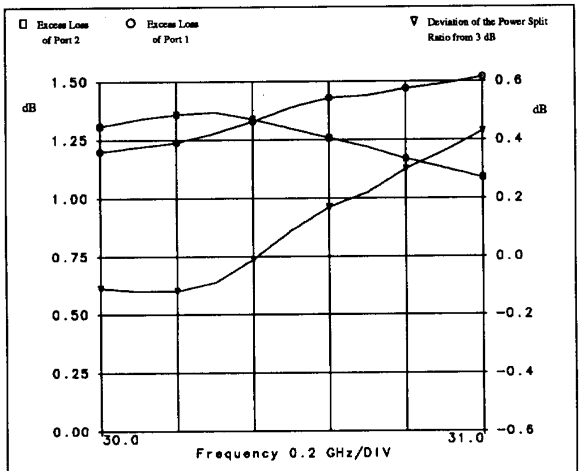

Figure 7 Excess insertion losses at the output ports and deviation of the output power-split ratio from the ideal.

The ideal output power-split ratio is attained at 30.4 GHz. At this frequency the excess insertion loss is 1.35 dB at each port..

2. A K-Band Equi-Power, Equi-Phase 14-Way Power Combiner

This power coupler is designed to operate as a power combiner. In this case there are two performance requirements: equal transmission *and* equal insertion phase at all ports. Figure 8 shows the layout of the combiner.

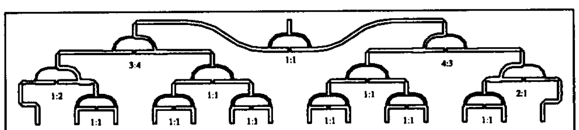

Figure 8 The layout of the K-band 14-way power combiner.

Figure 9 is a photograph of the device, mounted on a brass carrier, with 50 Ω Wiltron loads connected to 13 of the 14 output ports.

Figure 9 The 14-way power combiner.

The Measurement Method: A Millimeter-Wave Quick Release Test Fixture

N-way power combiner/divider networks with N>4 present a formidable measurement problem at millimeter-wave frequencies. Custom test fixtures are required which are easily assembled, reconfigurable, and can be accurately de-embedded. Previous experience with solid N-way test fixtures, i.e. custom test fixtures with N coaxial launchers was unsatisfactory.

Quick release test fixtures with non-soldered pressure contact microstrip launchers are commercially available in SMA technology. However, the SMA launcher is unsuitable at 20 GHz. A new quick release test fixture for frequencies up to 40 GHz has been developed using K-connector technology. The fixture, shown in figure 10, is constructed using the standard

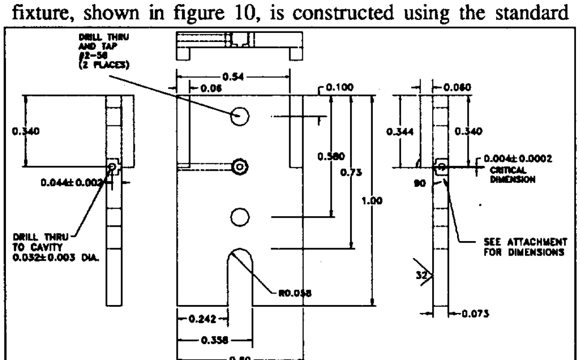

Figure 10 Mechanical drawing of the quick release test fixture.

assembly procedure for the K-connector. The U-shaped groove in the connector plate allows installation of the fixture over a screw attached to the carrier of the DUT. The 0.012" diameter center conductor of the K-connector is centered over the microstrip transmission line and the screw tightened while carefully applying downward pressure. Proper installation will result in an even mechanical interface between the microstrip and center conductor without any upward bending. Figure 9 shows the 14-way power divider with 14 test fixtures attached. Unused ports are loaded during the 2-port measurements with 50 ohm coaxial terminations.

The Measured Performance

The measured performance of the power combiner is shown in figures 11-14.

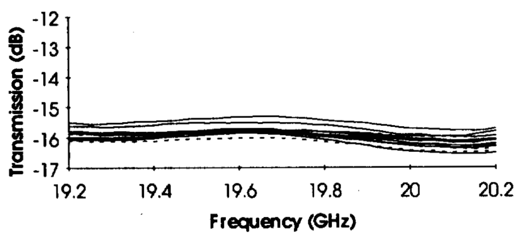

Figure 11 Transmission of the 14-way power combiner.

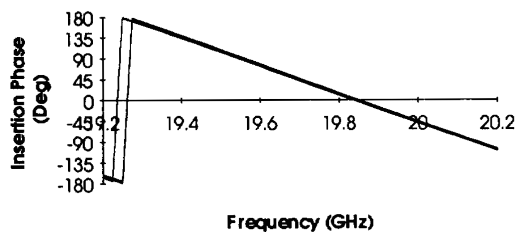

Figure 12 Insertion Phase of the 14-way power combiner.

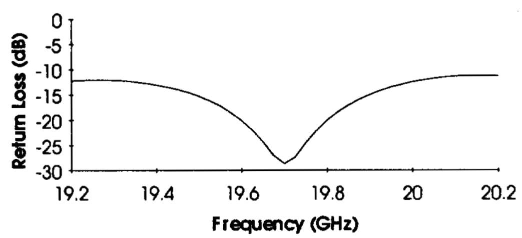

Figure 13 The return loss at the output port of the 14-way combiner.

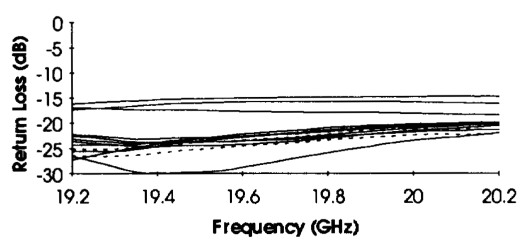

Figure 14 The return loss at the 14 input ports of the 14-way power combiner.

The measurements shown in figures 11-14 are those of a typical device. The latter exhibits good power flatness, with transmission variations of 0.6 dB in magnitude and 4 degrees in phase over all its ports. The best device, has transmission variations of 0.4 dB in magnitude and 2 degrees in phase over all its ports. The insertion loss, in excess of division loss, is 4.3 dB total, or 1.35 dB per level of dividers, verifying the numbers from measurements of the individual devices.

Conclusions

A new Wilkinson power divider design is presented. This design solves some of the problems of the traditional Wilkinson design at frequencies above X-band. The new design makes unequal power-split ratio dividers practical and permits accurate modeling of the latter at frequencies as high as Ka-band. Measurements of a three-level, 14-way, K-band power combiner, employing the novel Wilkinson design, show unusually good operating characteristics, although they exhibit high insertion loss (1.35 dB per level of dividers).

References

[1] Mazen Hamadallah, "Microstrip Power Dividers at mm-Wave Frequencies," Microwave Journal, Vol. 31, No. 7, pp. 115-127, July 1988.

[2] G. Robert Traut, "Thermoset Microwave Material - Enhances Microwave Hybrids," Applied Microwave, May 1989.

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