Broadband Planar 5:1 Impedance Transformer

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***Abstract*—This paper presents a broadband Guanella-type planar impedance transformer that transforms 50 Ω to 10 Ω with a 10 dB bandwidth of 1–14 GHz. The transformer is designed on a flexible 50 μm thick polyimide substrate in microstrip and parallel-plate transmission line topologies, and is inspired by the traditional 4:1 Guanella transformer. Back-to-back transformers were designed and fabricated for characterization in a 50 Ω system. Simulated and measured results are in excellent agreement.**

***Index Terms*—Transformers, impedance matching, broadband, parallel-plate line.**

I. Introduction

TRANSMISSION line transformers (TLTs) are widely used as impedance matching networks in radio frequency applications [1]–[14]. TLTs, first implemented by Guanella in 1944 [2], can simultaneously exhibit an octave of bandwidth for discrete impedance transformation ratios, are compact, and attractive for use as broadband matching networks. A 4:1 Guanella-type TLT nominally exhibits frequency-independent characteristics when realized with a pair of appropriate impedance and equal-delay transmission lines. Fig. 1 (a) shows the schematic implementation of a 4:1 impedance transformer, where two delay lines of equal length are connected such that currents add in phase at the low-impedance end. As a result of the delay and line symmetry, the transformation is theoretically independent of line length at finite frequencies. A simpler and more common type of TLT is the Ruthroff transformer [3], which appears similar to the Guanella "equal delay" design, but differs in implementation with its use of a single transmission line delay. As a result, the Ruthroff transformer has a smaller footprint, however, its response is not frequency independent and its bandwidth is ultimately limited by the transmission line length.

Guanella TLT performance is a function of the transmission line impedance, delay, and symmetry. At UHF and low microwave frequencies, TLTs are implemented with coaxial lines; in order to decrease their low-frequency limit and suppress unbalanced currents, they are wound around ferrite cores [4]–[6]. The high-frequency response is limited by the transmission line interconnect junction parasitics; to avoid this limitation, compensation of the junctions and more precise fabrication is required to ensure suppression and control over parasitic reactances. A Guanella-type 4:1 transformer in wafer-scale micro-coaxial technology with an operating bandwidth of 2–24 GHz was recently implemented [7].

Most prior published planar TLTs are limited to the Ruthroff configuration. Examples have been implemented in monolithic microwave integrated circuits [4], [8], [9], several were implemented with coupled microstrip lines [10], [11], and one recent example for superconducting applications was implemented as a 6.25 Ω to 25 Ω transformer from 2–13 GHz [12].

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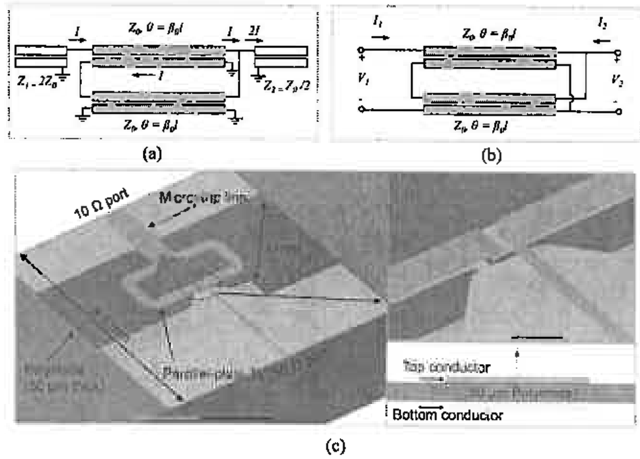

Fig. 1. (a) Circuit model of a conventional 4:1 Guanella impedance transformer. (b) Generalized circuit model of the two-transmission lines Guanella impedance transformer. (c) Rendered image of the transformer over a 10 mm × 4.5 mm × 4 mm cavity. The input and output transmission lines are microstrip for ease of integration with standard printed circuit board connectors. The bottom right inset shows the parallel-plate transmission line topology with one conductor on top and one conductor on the bottom used to implement the 5:1 planar transformer. The top right inset shows the connection at the high impedance (Z_2) end of the transformer; the parasitics associated with this series connection limit the high-frequency performance of the transformer.

In this paper we describe a planar Guanella-type 5:1 TLT that is easily integrable with common planar transmission line topologies, *e.g.*, microstrip. A primary goal of this study is to demonstrate a compact planar broadband matching circuit that is both cost effective and easily implementable with printed circuit boards (PCBs). Due to use of a durable and flexible substrate, this transformer can be used in extreme environments, including at cryogenic temperatures.

II. Transformer Design Procedure

A 4:1 Guanella transformer (Fig. 1 (a)) consists of two equal-length, equal characteristic impedance (Z_0) transmission lines that are connected in series at the high-impedance end (Z_1) and in parallel at the low-impedance end (Z_2). The impedance transformation is depicted in Fig. 1 (a); current flows in both conductors of a transmission line but in opposite directions. Starting from the high-impedance end, current I flows in the top conductor of the first transmission line (Z_0); an equal and opposite current flows in the bottom conductor of the same line. By adding a second transmission line with the same length and characteristic impedance, and then connecting the

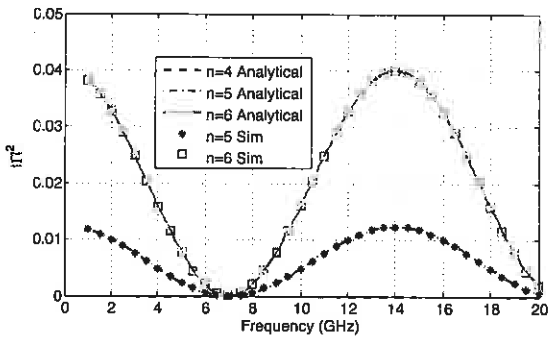

Fig. 2. Calculated and simulated reflection coefficient as a function of frequency for impedance ratios $n = 4$, 5 and 6 for the two-transmission line impedance transformer shown in Fig. 1 (b), where $Z_0 = (Z_1 Z_2)^{1/2}$, and $l = 10.714$ mm ($\lambda/4$ at 7 GHz in free-space). Simulation is performed using ideal circuit model in a circuit simulator.

two transmission lines in parallel and series at each end, a 4:1 impedance transformation ratio is achieved. The relationship between characteristic impedances is $Z_0 = (Z_1 Z_2)^{1/2}$ [1]. Due to the transmission lines' equal length and impedance, the differential mode is theoretically frequency independent. In practice, the structure's port isolation and match are interrelated, and the low frequency response is set by the length of the two transmission lines [14]. The high-frequency limit is optimized by minimizing the junction parasitics at the series and parallel transmission line connections.

A. 5:1 Impedance Transformer Design

A Guanella TLT can be realized for a discrete transformation ratio where the square root of the desired ratio n is equal to a rational number. If this condition is met, the ratio can be realized by connecting multiple transmission lines in series and parallel [1]. Therefore, in its simplest form, the Guanella topology does not support a 5:1 ratio. The closest realizable impedance ratios are $\left(\frac{7}{3}\right)^2$:1 or $\left(\frac{11}{5}\right)^2$:1, which are attainable with 5 and 7 transmission lines, respectively. The transmission lines are connected in parallel and series combination at the input and output. The use of multiple lines requires a larger footprint and more complex design with additional parasitics from multiple connections, which reduces the bandwidth and creates more resonances in the passband.

For a simplified realization of a 5:1 ratio, we used a similar topology as Guanella 4:1 transformer (Fig. 1 (b)). For the 5:1 transformer, due to deviation from the ideal 4:1 case, there is a length dependency where the phase cancels out completely at $l = \lambda/4$. In the absence of parasitics the odd mode input impedance looking into port 1 can be calculated using (1) [13]. The reflection coefficient Γ is calculated by (2). Fig. 2 shows the transmission line calculation and simulated $|\Gamma|^2$ as a function of frequency for ratios $n = 4$, 5, and 6, where $Z_0 = (Z_1 Z_2)^{1/2}$ and $l = 10.714$ mm ($\lambda/4$ at 7 GHz in free space).

$$Z_{\text{in}} = 2Z_0 \frac{2Z_2 + \jmath Z_0 \tan(\beta l)}{Z_0 + 2\jmath Z_2 \tan(\beta l)} \quad (1)$$

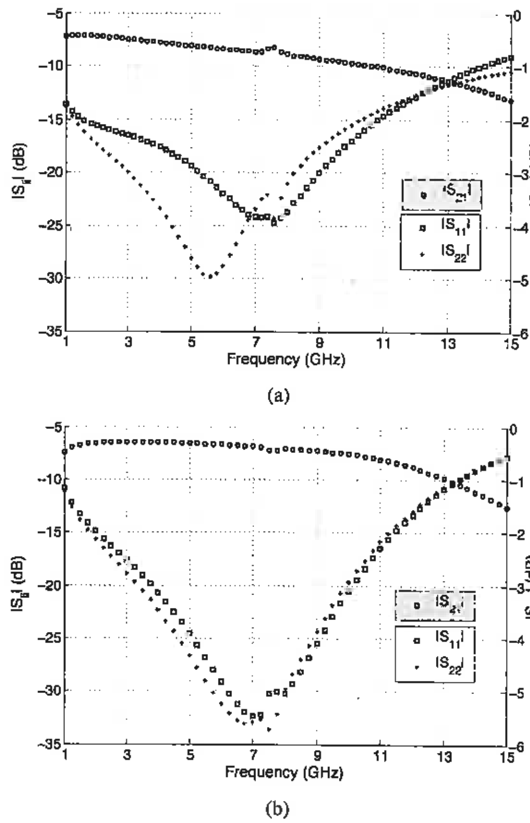

Fig. 3. (a) Simulated S-parameter results of the transformer in free space. (b) Simulated S-parameter results of the transformer above the 10 mm ×4.5 mm ×4 mm cavity. The transmission lines' phases cancel at 7 GHz ($l = \lambda/4$) resulting in the best match.

$$\Gamma = \frac{Z_{\text{in}} - Z_1}{Z_{\text{in}} + Z_1} \quad (2)$$

To implement the transformer in a planar topology, we chose a parallel-plate transmission line with one conductor atop the substrate and one conductor beneath the substrate for the two equal delays. The substrate is 50 µm thick polyimide with $\epsilon_r = 3.4$ and $\tan\delta = 2\times10^{-3}$ and the design impedances are $Z_0 = 22.3\,\Omega$, $Z_1 = 10\,\Omega$, and $Z_2 = 50\,\Omega$. The input and output transmission lines are microstrip, so that they can easily be integrated with other components on a PCB. Fig. 1 (c) shows a 3D rendering of the design above a 10 mm ×4.5 mm ×4 mm cavity. In this design a 10 Ω microstrip line transitions to a 10 Ω parallel-plate transmission line, then connects in parallel to a pair of 5.5 mm long 22 Ω parallel-plate transmission lines. The parallel-plate transmission lines then connect in series with two 75 µm diameter vias (Fig. 1 (c) inset), and finally connect to a 50 Ω microstrip line.

The transformer was first electromagnetically modeled using the finite element method (FEM) in an air-boundary box, Fig. 3 (a) shows the simulated S-parameter results of this model. This model has return loss better than 10 dB from 1–14 GHz with less than 1 dB transmission loss at 10 GHz.

Since the transformer needs to be packaged and integrated with external components, it is designed above a metallic 10 mm ×4.5 mm ×4 mm cavity. The cavity is required since the transformer cannot be attached to a conductive mount,

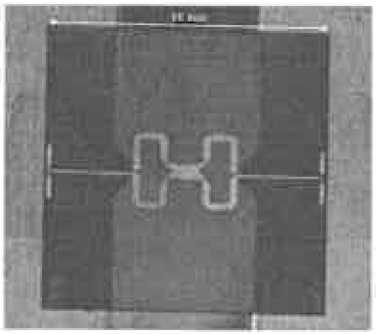

Fig. 4. Photograph of the fabricated back-to-back impedance transformer on a $50\,\mu$m thick polyimide substrate. The transformer was placed above the ground to prevent shorting of the bottom conductor of the parallel-plate transmission line.

as it would short the transformer. Additionally, because the transformer transmission lines are on a low dielectric constant material, the fringing fields at the edges of the conductors are significant and as a result there is significant radiation loss. By designing the transformer above an open cavity, the fields are confined and the overall loss of the device is reduced from 1 dB to 0.5 dB at 10 GHz. Further reductions are possible by enclosing the entire transmission line circuit in a non-resonant cavity. Fig. 3 (b) shows the simulated S-parameter results of the transformer placed above a cavity. The small resonant feature that appears at 7.5 GHz in both simulations is due to a small length difference of the two main transmission lines caused by the series interconnection at the high-impedance end. This phenomena creates a path for a weakly-coupled $\lambda/2$ shorted resonator within the transformer, allowing a small resonance feature to appear [7].

III. Characterization

To characterize the transformer in a $50\,\Omega$ system, we designed and fabricated a symmetric structure in which two transformers are connected at the $10\,\Omega$ port. Fig. 4 is a photograph of the back-to-back transformers recessed above a ground plane for characterization. A custom through-reflect-line (TRL) calibration set with two lines, an open, and a thru was used. Fig. 5 shows the simulated and measured S-parameter results of the back-to-back transformers over the range of 1–12 GHz. The simulation and measured results show very good agreement. The slight discrepancy between the measured and simulated $|S_{11}|$ is due to a manufacturing error; some of the transmission lines had small errors in their fabricated widths, resulting in slightly different characteristic impedances than specified in the design.

IV. Conclusion

In this paper we presented a planar Guanella-type impedance transformer with a ratio of 5:1. Based on these results, it is possible to employ the Guanella transformer topology beyond the 5:1 ratio presented here. As we deviate from the ideal 4:1 ratio in a Guenella-type transformer, the return loss will be degraded. In practice, the achievable upper frequency response is limited by the parasitics' reactance (*e.g.*, via inductance) and ones ability to compensate for these reactances. A Guanella-type transformer has been implemented in a planar circuit topology, using easily-fabricated transmission lines with two metallization layers. Some of the benefits of this transformer are its high impedance transformation ratio, compact size (1 cm×1 cm), low cost, and durability in extreme environments due to utilization of a flexible, thin, polyimide substrate. Compared to conventional broadband matching circuits such as tapers, they are an order of magnitude smaller in electrical length, resulting in less dielectric and metal loss.

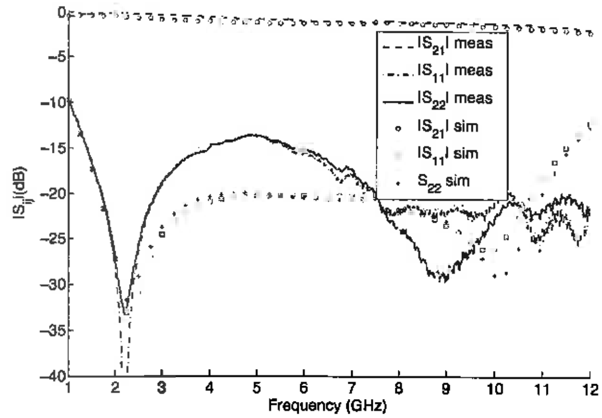

Fig. 5. Measured and simulated S-parameter results of the back-to-back transformers. The transformer is positioned above the ground plane to prevent the circuit from shorting. The phase cancellation (resonance) at 2.2 GHz and 9 GHz is due to the back-to-back measurement configuration.

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