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An Asymmetric 2.4 GHz Directional Coupler Using Electrical Balance

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***Abstract*— The concept of electrical balance is used to achieve wideband isolation in an asymmetric 2.4 GHz directional coupler. An intuitive approach using impedance transformation property of transmission lines is used to develop the theory for designing arbitrary-coupling-ratio directional couplers. A coupler based on this theory is designed and fabricated on a two-layered RO4003C PCB. The coupler achieves a return loss of better than 10 dB over a 1.7-2.7 GHz range. The transmission coefficients of the strongly and weakly coupled ports are around −3.5 dB and −5 dB respectively, at 2.4 GHz. The isolation is better than 25 dB over a 1.4-3 GHz range.**

***Index Terms*— Directional coupler, electrical balance, wideband isolation.**

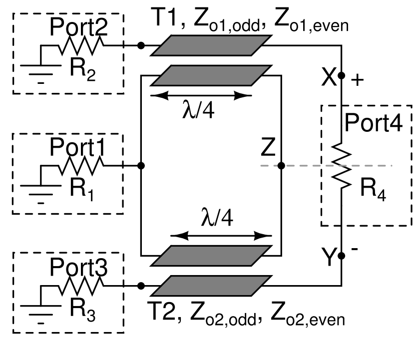

Fig. 1. Schematic of proposed directional coupler.

I. Introduction

DIRECTIONAL coupler is a passive, reciprocal four-port network used in microwave system design and measurement applications to process signals depending on their direction of flow. Conventional designs of directional coupler with transmission lines (TL) are branch-line hybrid, rat-race coupler, coupled line coupler and Lange coupler. All these couplers have multiple paths for signal to flow from incident port to other ports. Signals add at the coupled and through ports, whereas they cancel each other at the isolated port [1]. Different frequency responses of the multiple paths taken by the signal cause strong dependence of isolation on frequency.

Hybrid transformer is one of the earliest uses of electrical balance in a four port network to achieve wideband isolation [2]. Since it uses lumped elements, the maximum frequency of operation is limited due to self resonance of the winding. Moreover, capacitive coupling between primary and secondary of the transformer causes common-mode coupling to the balanced port [3]. A phase-reversal rat-race coupler was proposed in [4], with theoretically infinite isolation bandwidth using frequency independent phase shift circuits. In [5], Yang et al demonstrated a wideband coupler employing a modified Wilkinson bridge design which inherently used electrical balance to achieve frequency independent isolation. In [6] and [7], in-phase and out-of-phase power dividers are combined using ground slot lines to make a symmetrical structure with wideband operation.

In this work, an asymmetric directional coupler using electrical balance is proposed whose symmetric counterpart was presented earlier in [8]. The letter is organised as follows: Section II discusses the theory associated with the proposed directional coupler, and section III elaborates on the design and measurement results of the fabricated prototype. Section IV summarises the work.

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II. Theory of Proposed Directional Coupler for Arbitrary Coupling Ratios

In the proposed coupler, electrical balance is achieved using two signal paths whose frequency responses track each other. The two paths need not be identical, as is the case in earlier literature, enabling arbitrary coupling ratios other than −3dB each.

Fig. 1 shows the block diagram of the proposed directional coupler using electrical balance. T1 and T2 are quarter-wave TLs with odd-mode impedances of $Z_{o1,odd}$ and $Z_{o2,odd}$ respectively. Ports are numbered as per Fig. 1, and all ports other than Port 4 are single ended. For forthcoming analyses, all TLs are assumed to be lossless with infinite even-mode impedance. Differential characteristic impedance of a TL looking between its two lines, Z_{diff}, is equal to $2Z_{o,odd}$. All port impedances are assumed to be real and denoted by R_i for port i, and P_i refers to the power received at i^{th} port. The operation of the circuit for excitation at different ports is explained below.

A. Signal Applied at Port 1:

For Port 1 excitation, no current should flow in Port 4 to achieve isolation. This is achieved by ensuring

$$\frac{Z_{o1,odd}}{Z_{o2,odd}} = \frac{R_2}{R_3} \tag{1}$$

This results in equal voltages across pairs of terminals of T1 and T2 on the Port 4 side, and hence zero current in Port 4.

Open circuited lines T1 and T2 transform to short circuit at the other end, reducing the network to Port 1 feeding a parallel combination of Port 2 and Port 3. Matching at Port 1 now requires

$$R_1 = R_2||R_3 \tag{2}$$

When an input signal is applied at Port 1, its power splits between Port 2 and Port 3 in the following way:

$$\left.\frac{|P_2|}{|P_3|}\right|_{signal\ applied\ at\ Port1} = \frac{R_3}{R_2}$$

If matching is ensured at all the ports, the transmission coefficients can be written as

$$S_{12} = \sqrt{\frac{R_3}{R_2+R_3}}, \quad S_{13} = \sqrt{\frac{R_2}{R_2+R_3}} \tag{3}$$

B. Signal Applied at Port 2:

Even and odd-mode analysis can be utilized to understand this case. The even-mode equivalent circuit is similar to previous case with equal signals being applied at Port 2 and Port 3, instead of Port 1. In odd-mode, signals applied at Port 2 and Port 3 are equal and out of phase. The balanced nature of Port 4 causes signal ground to appear at its centre (line of symmetry as per Fig. 1). Therefore, Port 4 can be split into two equal halves, with the upper half being transformed by T1 and the lower half by T2. If matching is ensured at Port 2 and Port 3 for odd-mode excitation, then

$$R_4 = \left(\frac{2Z_{o1,odd}}{R_2}\right)^2 (R_2 + R_3) \tag{4}$$

It can also be shown that superposition of odd and even-modes results in isolation between Port 2 and Port 3.

C. Signal Excitation at Port 4:

In this case, no current flows in Port 1 as it is isolated from Port 4 due to electrical balance. Therefore, Port 4 sees a series combination of Port 2 and Port 3 transformed by T1 and T2 respectively. As the power absorbed at Port 2 and Port 3 has phase difference of 90^0 with respect to power incident at Port 4, the transmission coefficients can be written as

$$S_{24} = -j\frac{2Z_{o1,odd}}{\sqrt{R_2R_4}}, \ S_{34} = j\frac{2Z_{o2,odd}}{\sqrt{R_3R_4}} \tag{5}$$

Thus, scattering matrix (S) takes the following form:

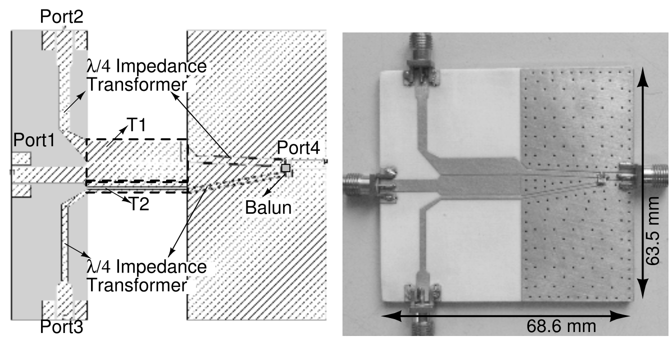
$$[S] = \begin{bmatrix} 0 & \sqrt{\frac{R_3}{R_2+R_3}} & \sqrt{\frac{R_2}{R_2+R_3}} & 0 \\ \sqrt{\frac{R_3}{R_2+R_3}} & 0 & 0 & -j\frac{2Z_{o1,odd}}{\sqrt{R_2R_4}} \\ \sqrt{\frac{R_2}{R_2+R_3}} & 0 & 0 & j\frac{2Z_{o2,odd}}{\sqrt{R_3R_4}} \\ 0 & -j\frac{2Z_{o1,odd}}{\sqrt{R_2R_4}} & j\frac{2Z_{o2,odd}}{\sqrt{R_3R_4}} & 0 \end{bmatrix} \tag{6}$$

When an input signal is applied at Port 1, equal voltages appear across pairs of terminals X-Z and Y-Z (see Fig. 1). If Port 4 has finite even-mode impedance, nodes X and Y stay at ground potential since even-mode currents cannot flow in T1 and T2. This causes the full even-mode signal swing to appear at node Z. Therefore, the proposed coupler is capable of lower common-mode coupling from Port 1 to Port 4.

It is worthwhile to mention that coupling ratio and port impedance cannot be set independently. Theoretically any coupling ratio is possible, but the practical range of characteristic impedance and the feasibility of matching network constrain the values of achievable coupling ratio. In a typical PCB process, ratio of maximum and minimum characteristic impedance possible for co-planar stripline is around 2:1. This translates to coupling ratios of 4.8 dB and 1.8 dB between weakly and strongly coupled ports respectively.

TABLE I
TRANSMISSION LINE PARAMETERS AT 2.4 GHz

TL	Line Width	Line Separation	$\|Z_{o,odd}\|$ (Ω)	$\|Z_{o,even}\|$ (Ω)	$\lambda_{odd}/4$ (mm)
T1	4.45 mm	0.25 mm	42.7	314.8	22.1
T2	0.76 mm	0.51 mm	70.7	466.5	21.0



(a) (b)

Fig. 2. (a) Layout of final design. Hatched region denotes top conductor and light shaded region denotes bottom conductor. Bottom conductor is etched under T1 and T2. (b) Photograph of fabricated board.

III. PCB DESIGN AND MEASUREMENT RESULTS

An asymmetric coupler centred at 2.4 GHz is designed based on the theory presented in the earlier section. This coupler is fabricated on RO4003C ($\epsilon_r = 3.38$) with a 1.52 mm thick substrate and double layer copper laminate. Design of transmission lines T1 and T2 is critical to achieve best coupler performance. The ground plane is etched underneath them to maximize even-mode impedance compared to odd-mode impedance (5 times), as seen in Table I. The sheet resistance of conductor is approximately 0.24 mΩ/ and the substrate has a loss tangent of 0.0021. The coupler is designed using Sonnet v13 EM simulation software.

As the coupled lines primarily carry odd-mode signals, fields are confined in the gap of the respective lines. The adjacent lines of T1 and T2 are merged together to get maximum area reduction. This changes the looking-in impedance at Port 4 by less than 5%, which is accommodated while designing the matching network at Port 4. Fig. 2a shows the final board layout with T1 and T2 merged to form a three-line structure. The open-circuit end of the merged line is made slightly shorter than $\lambda/4$ to compensate for extra fringing capacitances added by the open end.

Impedances of Ports 2 and 3 are chosen to be 81 Ω and 125 Ω respectively as per (1), with the additional constraint that their parallel combination result in a 50 Ω impedance at Port 1. Quarter-wave microstrip lines are used to transform Port 2 and Port 3 impedances to 50 Ω. Similarly quarter-wave co-planar waveguide (CPW) lines along with a 100:50 balun (Minicircuits NCS2-33) are used to transform Port 4 impedance to single-ended 50 Ω for measurement purposes.

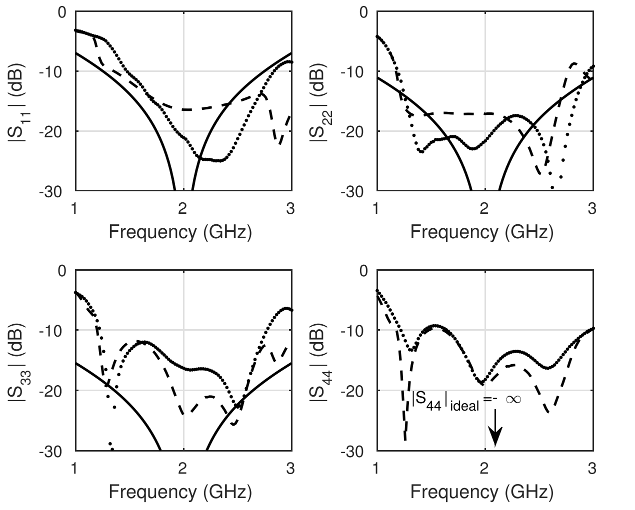


Fig. 3. Ideal (continuous), simulated (dashed) and measured (dotted) return loss of the coupler.

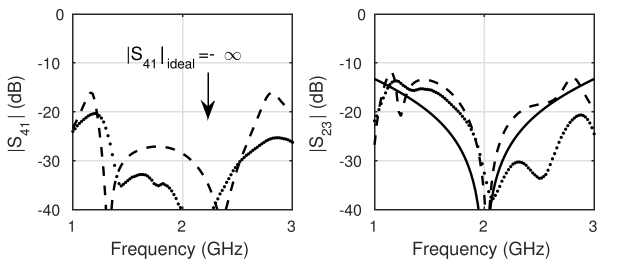


Fig. 4. Ideal (continuous), simulated (dashed) and measured (dotted) isolation of the coupler.

The fabricated coupler is shown in Fig. 2b. Measured and simulated return loss at all ports is better than 10 dB in the 1.7 GHz to 2.7 GHz frequency range (Fig. 3). At frequencies higher than the designed centre frequency of 2.4 GHz, transitions in transmission lines cause matching to degrade rapidly.

The coupler exhibits better than 25 dB measured isolation between electrically balanced Ports 1 and 4 in the 1.4 GHz to 3 GHz frequency range (Fig. 4). Limited and frequency dependent isolation between electrically balanced ports can be attributed to different frequency responses of transitions at T1 and T2. Balun imbalance and length mismatch in T1 and T2 also cause isolation degradation. As expected, isolation between Ports 2 and 3 is not as wideband as that between Ports 1 and 4. Difference between simulation and measurement results (Fig. 3–6) are due to deviation of balun frequency response from the expected nominal model available from the manufacturer.

Measured transmission coefficients of strongly coupled ports, S_{21} and S_{43}, vary from -3.5 dB to -3.8 dB and from -3.4 dB to -4.2 dB respectively, over the matching bandwidth. Similarly, for the weakly coupled ports, S_{31} and S_{42} vary between -4.7 dB to -5.5 dB and -5 dB to -5.5 dB respectively, as shown in Fig. 5. Coupling is lower in measurement than in simulation due to additional board and connector losses that could not be accounted for a priori.

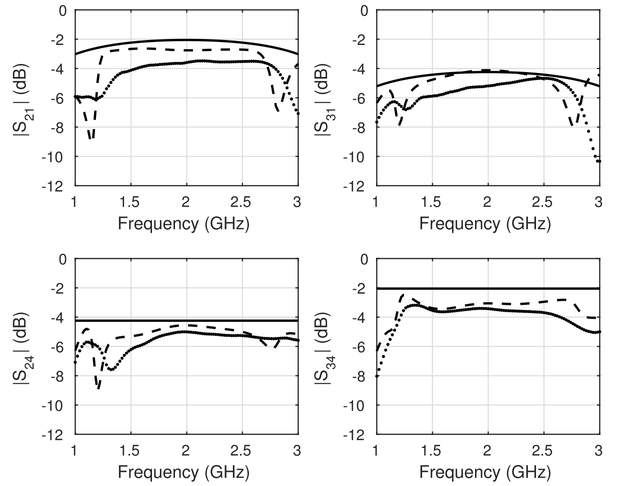


Fig. 5. Ideal (continuous), simulated (dashed) and measured (dotted) coupling of the coupler.

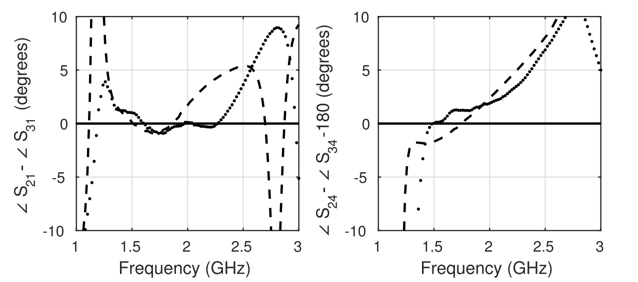


Fig. 6. Ideal (continuous), simulated (dashed) and measured (dotted) phase response of the coupler.

IV. Conclusion

The concept of electrical balance can be used to design directional couplers with wideband isolation. A theory has been developed to analyse and design an asymmetric transmission-line-based directional coupler using electrical balance for arbitrary coupling ratios. The theory developed has been successfully applied to the design of a 2.4 GHz coupler achieving better than 25 dB isolation in 1.4 to 3 GHz frequency range.

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