A Novel Wideband, Compact, Microstrip Coupled-Line Ring Hybrid for Arbitrarily High Power-Division Ratios

Hee-Ran Ahn, Senior Member, IEEE and Manos M. Tentzeris, Fellow, IEEE

Abstract—This paper introduces a novel compact and wideband coupled-line ring hybrid configuration that allows for the easy realization of arbitrarily high power division ratios featuring significant miniaturization and wider bandwidths compared to other commonly utilized implementations. The circumference of a proof-of-concept 12-dB power division ratio prototype is 209.3° long, which occupies 15 % of the conventional area, and the measured bandwidth of the 15-dB return loss is 78 %, an improvement by a factor around 1.8.

Index Terms—Coupled-line ring hybrids, compact ring hybrids, wideband ring hybrids, arbitrary power division ratios, high-power division ratios, symmetric equivalent circuits, rat-race couplers.

I. INTRODUCTION

The ring hybrids [1]-[9], which are commonly also called “rat race couplers”, have been extensively used for various applications, such as balanced amplifiers [10] and antenna arrays, typically necessitating the easy realization of arbitrary power division ratios [11]. Also, as numerous multifunctional wireless modules require a substantial reduction in mass and volume, the compactness of the ring hybrids has been of high interest. Typical ring hybrid topologies consist of three 90° and one 270° Transmission-Line sections (TL’s). If the power division ratios are not 0 dB for the same termination impedances, two TL’s should be identical and the other two should have the same characteristic impedances for an electrical length difference of 180°. If, for example, the power division ratio is 12 dB with the termination impedance of 50 Ω, the characteristic impedances of TL’s should be 205.24 and 51.56 Ω [9, eq. (1)], and all TL’s should be as short as possible due to compactness requirements. However, the TL’s with the characteristic impedance of 205.24 Ω do not seem to be easy to fabricate with a microstrip technology due to a required width around or below 40 μm for commonly used substrates (e.g. for the prototype discussed in this paper the required microstrip line width was 42μm for a substrate with a thickness of 20 mil and a dielectric constant 2.2), while the physical length of the low-impedance (51.56 Ω) TL’s should be reduced without a significant bandwidth compromise. Due to the above implementation difficulties, most of previously reported efforts have treated the compact ring hybrids for equal power divisions [1], [3], [4] or for power division ratios less than or equal to 6 dB [5], [6]. In [7], high power division ratios have been treated, but the size is not compact, and the demanded power division ratio is achieved only around the design frequency. In [8], the 12-dB power division ratio has been treated, and equivalent/artificial lumped-element transmission-line models have been suggested for high-impedance TL’s. However, the extraction method for the lumped-element values seems to be very complicated, and the bandwidth is much smaller than that of the conventional one, while the size is not compact. The power-division ratio of 13 dB has been investigated [9], but the size (320.58°) is not compact, as well.

In this paper, a novel compact wideband ring hybrid allowing for the realization of arbitrarily high power division ratios in a miniaturized form factor will be introduced. In the preliminary ring design, for a proof-of-concept 12 dB power division ratio prototype, we assume that it includes two high-impedance TL’s which effectively are 90° and 270° long, and two low-impedance 90° TL’s, and each TL should be as miniaturized as possible. However, the size reduction approach should be different for every TL, and conventional ways, such as T– and Π– types [12, Fig. 6 with N = 1], cannot be applied for this case, due to their inherent lack of a mechanism to reduce high-impedance values to easier-to-realize lower ones.

To overcome these issues, the high-impedance TL of the effective length of 270° can be reduced to a set of coupled TL’s which is physically less than 90° long but can also feature an additional inherent wideband 180° phase shift, while the high-impedance 90° TL is shortened by use of an L31 – section type configuration [9] that allows for the simultaneous reduction of the characteristic impedance value. The two low-impedance TL’s are miniaturized through the use of modified T- section types (MT-section types). Using the respective equivalent circuits of these three utilized topologies, a coupled-line ring hybrid prototype was easily fabricated at the design frequency of 1 GHz, and the total electrical length of the fabricated prototype was 209.3°, which occupies only 15 % of the area occupied by the conventional design. The measured bandwidth with 15-dB return loss is 78 %, a significantly higher value, compared to the conventional one of 44 %.

II. MINIATURIZATION APPROACH / EQUIVALENT CIRCUITS

The topology of a typical ring hybrid with the same termination impedance Z0 for all its ports is depicted for an arbitrary power division ratio in Fig. 1; it consists of three 90° and one 270° TL’s with the characteristic impedances of Z1 and Z2.
aggregate electrical lengths of 90° for the high- and the numbers of unit cells required to effectively realize the MT-section unit cells are typically smaller than 5 and represent the impedances of \((Z_\Theta)\) TL with an electrical length of 90° / angular frequency. The high-/low-impedance TL's with a total effective phase delay of \((90°/T_N)\) low-impedance \((Z_L)\) TL.

\[
Z_T = Z_2 \tan \frac{\pi}{4T} \tan \frac{\Theta_T}{2}
\]

where \(C\) is the target coupling coefficient and \(\omega_0\) is the design angular frequency.

III. VALIDATING RING HYBRID DESIGNS FOR POWER DIVISION RATIO OF 12 DB

The equations (1a)-(1b) imply that the values of \(Z_{\Theta C}, Z_{00}\) and \(S_c\) are determined by the electrical lengths of \(\Theta_c\) for a fixed value of \(C\). For the design of an \((L_{s1})\) unit block in Fig. 2(b), the characteristic impedance of \(Z_L\) should be sufficiently lower than \(Z_1\), so that it can be realized without any difficulty, and the inductance values of \(L\) should be easily fabricated with available off-the-shelf inductors (e.g. 0402 chip inductors) that typically require microstrip line widths of the \((Z_L)\) TL’s in the order of 500 um. The characteristic impedances of \(Z_T\) of the MT-unit blocks in Fig. 2(c) are inversely proportional to the electrical lengths of \(\Theta_T\). For the substrate (RT/Duroid 5880, \(\varepsilon_r = 2.2, H = 20\) mil) used in the proof-of-concept prototype, the required line width is around 450 um for the characteristic impedance value of 100 \(\Omega\). Therefore, the highest characteristic impedance values of \(Z_L\) and \(Z_T\) are set to around 100 \(\Omega\) to enable the easy soldering of the chip inductors in between two TL’s.

Numerous values of the characteristic impedances of \(Z_L\) and \(Z_T\) are calculated in Tables I and II by varying the aggregate electrical length \(N_\Theta\) of the \((L_{s1} -\) and MT-section types, that is kept constant for different number of utilized stages \(N_s\), where \(I\) is \(L\) or \(T\), used for the implementation of the high-/low-impedance TL’s with a total effective phase delay of
Table I shows that for the proof-of-concept prototype, characteristic impedance $Z_t$ values around 100 $\Omega$ can be obtained for the design value $N_t \theta_t = 45^\circ$. In this case, for $N_t = 2, 3$ and 4, the inductance values of $L$ are, based on (1c), calculated as 17.77 nH, 12.39 nH and 9.43 nH at 1 GHz, leading to the total inductance values of $N_t \cdot L$ being 35.54 nH, 37.17 nH and 37.74 nH, respectively. The values of $N_t \cdot L$ should be as small as possible, and therefore $N_t = 2$ is desirable. However, $L = 17.77$ nH is impossible to realize with off-the-shelf components the closest of which is 18 nH. Fixing $N_t = 2$ and $L = 18$ nH, the electrical length of $N_t \theta_t$ can be easily calculated as $N_t \theta_t = 44.27^\circ$ from equation (1c). From the data in Table II, it can be easily observed that values of the characteristic impedance $Z_t$ around 100 $\Omega$ can be obtained for design lengths $N_t \theta_t$ between 40$^\circ$ and 50$^\circ$.

Following the approach discussed in [2, Figs. 2 and 3], it can be easily shown that for the proof-of-concept prototype presented in this paper, the 90$^\circ$ and 270$^\circ$ high-impedance ($Z_1$) TL’s are impedance transformers that each convert an impedance of 842.45 $\Omega$ into the termination impedance of 50 $\Omega$, and the two low-impedance ($Z_2$) TL’s are transforming 53.16 into 50 $\Omega$. Based on the presented design concept, the equations (1a)-(1e) and the data in Tables I and II, the design parameters for the CII-, $L_{s1}$- and MT – section types are listed in Table III, IV and V, respectively. For the utilized CII-section types, the coupling coefficient is fixed at -10 dB, and the electrical lengths of $\theta_{sc}$ are varied, because the gap width is greater than 100 um for such a coupling coefficient on the same substrate and therefore can be fabricated without any difficulty. For the utilized $L_{s1}$ – section types, the total length of $N_t \theta_t$ is fixed at 44.27$^\circ$ and the number of $N_t$ is varied. For the utilized MT-section types, the total length of $N_t \theta_t$ is fixed at 45$^\circ$ and the number of $N_t$ is varied.
frequency of 1 GHz for the design data in Tables III-V and are plotted in Fig. 3 where the CTI- and \( L_{s1} \)-section types are terminated in 842.45 and 50 Ω, while the MT-section types are terminated in 53.155 and 50 Ω, and the susceptances of \( S_c \) and \( S_T \) were realized with capacitances at 1 GHz. The frequency responses of the CTI-section types in Fig. 3(a) feature a similar behavior, regardless of the electrical lengths of \( \Theta_{sc} \) between 65°-85°, that allow for even-mode characteristic impedance values of around or below 100 Ohm in very miniaturized implementations. The frequency responses of the \( L_{s1} \)-section types with \( N_L = 2 \) and \( 3 \) in Fig. 3(b) are about the same, while that of the \( L_{s1} \)-section type with \( N_L = 1 \) is slightly worse than the other two. For the MT-section types in Fig. 3(c), the performance for the one with \( N_T = 4 \) is the best.

In terms of compactness, minimum radiation loss, easy fabrication and broadband operability, the number of required unit blocks \( N_L \) and \( N_T \) as well as the required phase \( \Theta_{sc} \) were optimized for the fabrication of the prototype. The CTI-section type with \( \Theta_{sc} = 75° \) was chosen, because despite the fact it is not the shortest, the value of its respective \( S_c^{-1} \) can be easily fabricated with a TL and an available chip capacitor. The \( L_{s1} \)-section type with \( N_L = 2 \) was selected, because the required inductance at 1 GHz can be easily realizable with available off-the-shelf chip inductors of 18 nH as mentioned previously. The MT-section type with \( N_T = 4 \) features in Fig. 3(c) a slightly better performance than that with \( N_T = 3 \) but both have return loss better than 30 dB in the whole frequency range of interest. However, the individual unit block’s electrical length of \( \Theta_T \) with \( N_T = 4 \) is shorter than that with \( N_T = 3 \), which causes a significant difficulty to realize \( S_T^{-1} \) only with distributed elements. Due to the reason, the MT-section type with \( N_T = 3 \) was chosen.

**IV. Fabrication, Measurements and Comparison**

Based on the selected and optimized section types from the data in Tables III-V, a compact ring hybrid was fabricated on a commonly used substrate, and the resulting prototype is demonstrated in Fig. 4, where the 90° MT- and \( L_{s1} \)-section types are 45° and 44.27° long, the 270° CTI-section type is 75° long, each susceptance of \( S_c \) was realized with a TL and a chip capacitor with 0.5 pF similar to the form in [5, Fig. 13(a)], while the chip inductors with the inductance value of 18 nH (0402 Murata, LQW15AN18N980, Rdc; 0.13 Ω, tolerance: ±5%, SRF(min):5.2 GHz) were soldered and each \( S_T/2 \) was fabricated with an open-circuited high-low stepped impedance TL. The measured responses are compared with the predicted ones in Fig. 5. The measured bandwidth with 15-dB return loss at port \( \odot \) is 78 % (0.65-1.43 GHz) in Fig. 5(a), that with 15-dB isolation in Fig. 5(b) is 147 % (0-1.47 GHz), that with the power-division ratio of 11.5-12.5 dB is 78 % (0.75-1.19 GHz, 1.25-1.59 GHz) in Fig. 5(c), while the phase responses in Fig. 5(d) also show broadband performance. Given the fabrication errors, the measured results are in good agreement with the predicted ones, indicating that the usage of the chip inductors and capacitors is acceptable for this application even for typical tolerances and parasitic elements.

**Fig. 4.** Fabricated compact coupled-line ring hybrid prototype.

**Fig. 5.** Compared measured and predicted frequency responses. (a) \( |S_{11}| \). (b) \( |S_{31}| \). (c) Power divisions. (d) Phase responses.

The measured frequency responses of the fabricated prototype are compared in Fig. 6 with the simulated ones of a much larger conventional one for the same power division ratio of 12 dB that was simulated with ideal parameter values. The input matching, isolation, power division ratios and phase difference values are shown in Fig. 6(a), (b), (c) and (d), respectively. The
compared frequency responses are summarized in Table VI. The bandwidth with 15-dB return loss of the proposed one is 78 %, whereas that of the conventional one is 44 %. The bandwidth with the 15-dB isolation of the proposed one is 147 %, whereas that of the conventional one is 52 %. The bandwidths with the power division ratios of (12 ± 0.5) dB of the proposed and conventional ones are 56 and 21 %, respectively.

TABLE VI

<table>
<thead>
<tr>
<th>Frequency performance</th>
<th>Proposed (measured)</th>
<th>Conventional (simulated)</th>
</tr>
</thead>
<tbody>
<tr>
<td>15-dB return loss BW</td>
<td>78 %</td>
<td>44 %</td>
</tr>
<tr>
<td>15-dB isolation BW</td>
<td>147 %</td>
<td>52 %</td>
</tr>
<tr>
<td>power division ratios (12 ± 0.5) dB</td>
<td>78 %</td>
<td>21 %</td>
</tr>
<tr>
<td>out-of-phase difference</td>
<td>180°+ 10°</td>
<td>76 %</td>
</tr>
<tr>
<td>in-phase difference</td>
<td>0°+ 10°</td>
<td>70.4 %</td>
</tr>
</tbody>
</table>

The bandwidths of the absolute out-of- (180°) and in-phase (0°) differences of the proposed one are 76 and 70.4 %, while those of the conventional one are 60 and 52 %. In terms of size, the conventional one consists of transmission line sections with an aggregate length of 540°, whereas the circumference of the proposed one is only 209.27° long, leading to a drastic miniaturization to only 15% of the conventional area, while featuring a much more wideband performance.

V. CONCLUSIONS

In this paper, a novel, compact and wideband ring hybrid was introduced for the easy realization of arbitrarily high power division ratios. The proposed topology miniaturizes the size by utilizing CII-, \( L_{s1} \) – and MT-section topologies. The circumference of one fabricated proof-of-concept 12-dB power division ratio prototype is only 209.3° long, compared to 540° of conventional implementations verifying an area miniaturization down to 15 %, while the measured bandwidth of the 15-dB return loss is 78 % versus the conventional one's 44 %, an improvement by a factor 1.8. The topology introduced in this paper could be easily optimized for a wide range of power division ratios and could set the foundation for the implementation of similar ultra miniaturized hybrids to a variety of wearable, implantable, IoT (Internet of Things) and aerospace applications.

REFERENCES