# New Design Formulas for Impedance-Transforming 3-dB Marchand Baluns

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***Abstract*—New design formulas for impedance-transforming 3-dB Marchand baluns are proposed. They are about the even- and odd-mode impedances of the coupled transmission-line sections of the Marchand baluns and determined by coupling coefficients together with termination impedances. The particular property proposed in this paper is to choose the coupling coefficient arbitrarily, resulting in infinite sets of design formulas available. This is quite different from the conventional design approach in which only one coupling coefficient is possible. For the perfect isolation of the Marchand balun, an isolation circuit (IC) is needed, being composed of two 90° transmission-line sections and resistance(s). Sufficient area to build such a long IC is, however, inherently not available. For this, ways to reduce the IC size are also suggested. To validate them, a microstrip Marchand balun terminated in 130 and 70 Ω is designed at a design center frequency of 1.5 GHz and tested. The measured results are in good agreement with prediction, showing that power divisions are 3.57 and 3.262 dB, return losses at all ports are better than 21 dB, and the isolation is better than 20 dB around the design center frequency. The measured phase difference between two balanced signals is 180° ± 2° in about 50% bandwidth.**

***Index Terms*—Compact isolation circuits (ICs) of Marchand baluns, impedance-transforming 3-dB Marchand baluns, modified Π- and $T$-types of equivalent circuits of transmission-line sections, new design formulas for Marchand baluns.**

## I. Introduction

THE MARCHAND balun introduced in 1944 [1] has been realized in various transmission-line structures and utilized for various applications such as balanced mixers, balanced amplifiers, and frequency multipliers [2]–[23]. The Marchand balun has two sets of coupled transmission-line sections, and if the even- and odd-mode impedances of the two sets of transmission-line sections are identical, 180° phase difference between two balanced signals can be achieved theoretically in whole frequencies (except even multiples of the design center frequency). Due to such attractive performance, the Marchand balun has received substantial attention from microwave circuit designers. In the beginning of 2000, design equations were developed for impedance-transforming 3-dB Marchand baluns and have been used for more than one decade [2], [4], [6], [7], [9], [11], [15]–[18], [20]–[23]. Since the coupling coefficient of the coupled transmission-line sections suggested by the conventional design approach is, however, determined by the termination impedances, only one value is possible and it can be 0 or −3 dB. It is desirable that the coupling coefficients be determined by the coupling structures [24]–[28] and independent of the termination impedances of the Marchand baluns.

In this paper, new design formulas of the even- and odd-mode impedances are derived for the impedance-transforming 3-dB Marchand baluns. Since the coupling coefficient of the coupling transmission-line sections can be selected arbitrarily, infinite sets of design formulas of the even- and odd-mode impedances are available. It will be shown later that the conventional design equations [2], [4], [6], [7], [9], [11], [15]–[18], [20]–[23] are only one among the infinite sets of even- and odd-mode impedances proposed in this paper. Any impedance-transforming 3-dB balun may be equivalent to a circuit consisting of two impedance transformers and an isolation circuit (IC), which may be determined by the phase delay of the impedance transformers [24]. In the case of the Marchand balun, since the phase delay of the impedance transformers are 90° or −90°, the IC should be composed of two 90° transmission-line sections and resistance(s) [24]. Since the two output ports (balanced ports) of the Marchand baluns are, however, located close to each other, no sufficient area to construct such a long IC is available. The 90° transmission-line sections might be shortened by using the conventional Π- [26], [29] or $T$-type of equivalent circuit, but the resulting bandwidth may be reduced. To avoid the bandwidth reduction, a transmission-line section is divided by the number of $N$ and the corresponding design equations of Π- or $T$-type of equivalent circuit are derived. In the case of infinite number of $N$, the $T$-type equivalent circuit with $N$ is very similar to the artificial transmission-line approach [30], but different fundamentally with some number of $N$. To validate the design equations of the Marchand baluns and the modified equivalent circuits of a transmission-line section, a microstrip Marchand balun terminated in 130 and 70 Ω is fabricated and measured.

## II. Impedance-Transforming Marchand Baluns

### A. Description and Conventional Design Formulas

A Marchand balun is described in Fig. 1(a) where it consists of two sets of coupled transmission-line sections. It is terminated in $R_r$ at port ① and $R_L$ at ports ② and ③ for the impedance transforming, and an internal port ④ is located between two sets of coupled transmission-line sections. For the perfect matching at ports ② and ③ and the perfect isolation between the two ports, an IC is needed to be connected. For the analyses, a set of coupled transmission-line sections terminated

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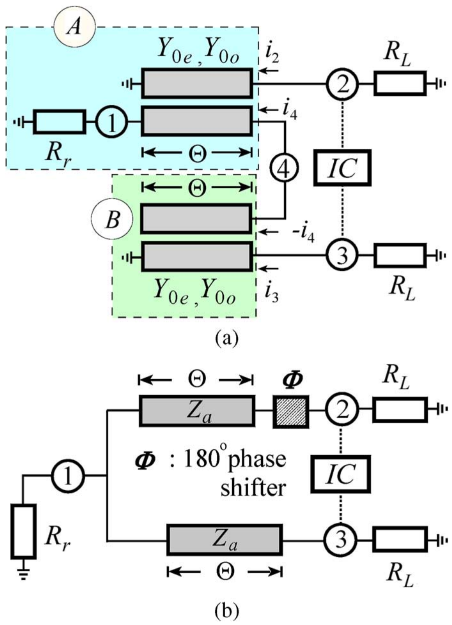


Fig. 1. Impedance-transforming baluns. (a) Marchand balun. (b) Equivalent circuit of the Marchand balun only at a design center frequency.

in $R_r$ at port ① is designated as $A$ and another as $B$, and an equivalent circuit being valid only at a design center frequency is introduced in Fig. 1(b).

The conventional design formulas of the even- and odd-mode impedances $Z_{0ec}, Z_{0oc}$ and its coupling coefficient $C_c$ [2], [4], [6], [7], [9], [11], [15]–[18], [20]–[23] are

$$Z_{0ec} = R_r \sqrt{\frac{1+C_c}{1-C_c}} \tag{1a}$$

$$Z_{0oc} = R_r \sqrt{\frac{1-C_c}{1+C_c}} \tag{1b}$$

$$C_c = \sqrt{\frac{R_r}{R_r + 2R_L}}. \tag{1c}$$

The conventional coupling coefficient in (1c) is a function of the termination impedances $R_r$ and $R_L$. It may be $-3$ dB when $2R_L = R_r$ and 0 dB when $R_r \gg 2R_L$. The coupling coefficient should be determined by the structure of the coupled transmission-line sections itself, not by the termination impedances [24]–[28]. Therefore, the design equations in (1) should be modified so that the coupling coefficients can be independent of the termination impedances. The two sets of coupled transmission-line sections will be synthesized for the design formulas.

### B. Design Formulas

If the currents at ports ②, ③, and ④ are denoted as $i_2$, $i_3$, and $i_4$, as shown in Fig. 1(a), and the voltages between ports ② and ④ and ground as $v_2$ and $v_4$, the relation between currents $i_2, i_4$ and voltages $v_2, v_4$ are

$$\begin{bmatrix} i_2 \\ i_4 \end{bmatrix} = \begin{bmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{bmatrix} \begin{bmatrix} v_2 \\ v_4 \end{bmatrix} \tag{2}$$

where

$$A_{11} = -j\frac{Y_{0e}+Y_{0o}}{2}\cot\theta + \frac{R_r(Y_{0o}-Y_{0e})^2\csc^2\theta}{4D_A} \tag{2a}$$

$$A_{12} = A_{21} = -j\frac{Y_{0o}-Y_{0e}}{2}\cot\theta + \frac{R_r\left(Y_{0o}^2-Y_{0e}^2\right)\csc^2\theta}{4D_A} \tag{2b}$$

$$A_{22} = -j\frac{Y_{0e}+Y_{0o}}{2}\cot\theta + \frac{R_r(Y_{0o}+Y_{0e})^2\csc^2\theta}{4D_A} \tag{2c}$$

with

$$D_A = 1 - jR_r\left(\frac{Y_{0e}+Y_{0o}}{2}\right)\cot\theta$$

where $A_{ij}$ is the admittance parameters contributed by ports ② and ④ of the $A$ circuit in Fig. 1(a).

If $v_3$ is the voltage between port ③ and ground, the admittance parameters of the coupled transmission-line sections $B$ contributed by ports ④ and ③ are

$$\begin{bmatrix} -i_4 \\ i_3 \end{bmatrix} = \begin{bmatrix} B_{11} & B_{12} \\ B_{21} & B_{22} \end{bmatrix} \begin{bmatrix} v_4 \\ v_3 \end{bmatrix} \tag{3}$$

where

$$B_{11} = j\frac{(Y_{0e}+Y_{0o})\tan\theta}{2} \tag{3a}$$

$$B_{12} = B_{21} = j\frac{(Y_{0o}-Y_{0e})\tan\theta}{2} \tag{3b}$$

$$B_{22} = -j\frac{Y_{0e}+Y_{0o}}{2}\cot\theta + j\frac{(Y_{0o}-Y_{0e})^2\csc 2\theta}{Y_{0e}+Y_{0o}} \tag{3c}$$

where $B_{ij}$ are two-port admittance parameters in terms of ports ④ and ③ of the $B$ circuit in Fig. 1(a).

From (2) and (3), with port ① terminated in $R_r$, the admittance parameters $Y_{ij}$ contributed by ports ② and ③ without the IC are obtained as

$$\begin{bmatrix} i_2 \\ i_3 \end{bmatrix} = \begin{bmatrix} Y_{11} & Y_{12} \\ Y_{12} & Y_{22} \end{bmatrix} \begin{bmatrix} v_2 \\ v_3 \end{bmatrix} \tag{4}$$

where

$$Y_{11} = A_{11} - \frac{A_{21}^2}{B_{11}+A_{22}} \tag{4a}$$

$$Y_{12} = Y_{21} = -\frac{A_{12}B_{12}}{B_{11}+A_{22}} \tag{4b}$$

$$Y_{22} = B_{22} - \frac{B_{12}^2}{B_{11}+A_{22}} \tag{4c}$$

When $\Theta = 90°$, the admittance parameters in (4) are simplified as

$$\begin{bmatrix} i_2 \\ i_3 \end{bmatrix} = R_r\left(\frac{Y_{0o}-Y_{0e}}{2}\right)^2 \begin{bmatrix} 1 & -1 \\ -1 & 1 \end{bmatrix} \begin{bmatrix} v_2 \\ v_3 \end{bmatrix} \tag{5}$$

As is well known, if the power is excited at port ① in Fig. 1(a), the power is divided between ports ② and ③ where the two waves are 180° out of phase at a design center frequency. To derive the design formulas of the complicated Marchand baluns easily, an equivalent is needed, as depicted in Fig. 1(b), consisting of two transmisson-line sections with the characteristic impedance of $Z_a$ and the electrical length of Θ, one 180° phase shifter, and IC. In this case, the electrical length of $\Theta$ is 90° at a design center frequency and the characteristic impedance of $Z_a$ is $\sqrt{2R_rR_L}$ [24], [31], [32]. Without the IC, the admittance parameters between ports ② and ③ of the equivalent circuit in Fig. 1(b) are

$$\begin{bmatrix} i_2 \\ i_3 \end{bmatrix} = \frac{1}{2R_L}\begin{bmatrix} 1 & -1 \\ -1 & 1 \end{bmatrix}\begin{bmatrix} v_2 \\ v_3 \end{bmatrix}. \tag{6}$$

For the two circuits in Fig. 1 to have the same frequency responses at the design center frequency, the admittance parameters in (5) and (6) should be equal to each other, from which the following relation holds:

$$\frac{Y_{0o} - Y_{0e}}{2} = \frac{1}{\sqrt{2R_rR_L}}. \tag{7}$$

The two sets of coupled transmission-line sections should be satisfied with the general definition of the coupling coefficient $C$ such as

$$Y_{0o} = Y_{0e}\frac{1+C}{1-C}. \tag{8}$$

From (7) and (8), the design formulas of the impedance-transforming 3-dB Marchand baluns are obtained as

$$Z_{0e} = \sqrt{2R_rR_L}\frac{C}{1-C} \tag{9a}$$

$$Z_{0o} = \sqrt{2R_rR_L}\frac{C}{1+C}. \tag{9b}$$

In the derived design formulas in (9), infinite sets of the even- and odd-mode impedances may be obtained by varying the coupling coefficient $C$, and the coupling coefficient is not the function of the termination impedances of $R_r$ and $R_L$. This is quite different from the conventional one in (1c). The conventional coupling coefficient in (1c) is one of infinitive sets in (9) and may be obtained by equating $Z_{0e} = Z_{0ec}$ or $Z_{0o} = Z_{0oc}$ in (1) and (9).

When the termination impedances are $R_r = 50\ \Omega$ and $R_L = 100\ \Omega$, the conventional coupling coefficient of $C_c$ is only one value of 0.4472 ($-6.9897$ dB), and the corresponding even- and odd-mode impedances are 80.9 and 30.9 Ω. The same values of the even- and odd-mode impedances are calculated by substituting $C = 0.4472$ into the equations in (9). In addition, there are so many other even- and odd-mode impedances available from the design (9).

### C. Frequency Responses of Marchand Baluns

With the termination impedances of $R_r = 50\ \Omega$ and $R_L = 100\ \Omega$, the even- and odd-mode impedances were calculated by varying the coupling coefficients and are written in Table I. When the coupling coefficient $C$ is $-3$ dB, the even-mode impedance of $Z_{0e}$ is 242.4 Ω, which is hard to be realized with a planar structure. When $C$ is $-5$ dB, $Z_{0e} = 128.5\ \Omega$ and $Z_{0o} = 36\ \Omega$, which may be realizable with microstrip technology. Based on the data in Table I, frequency responses were simulated at a design center frequency of 1 GHz, and the simulation results are plotted in Fig. 2.

Matching at port ①, power divisions of $|S_{21}|$ and $|S_{31}|$ and absolute phase difference between $S_{21}$ and $S_{31}$ are plotted in Fig. 2(a)–(c), respectively. Independently of the coupling coefficients, all are perfectly matched at the design center frequency of 1 GHz [see Fig. 2(a)]. As long as the coupling coefficient of the two coupled transmission-line sections is identical, the frequency responses of $S_{21}$ and $S_{31}$ are identical [see Fig. 2(b)], and the bandwidths increase with the coupling coefficient. The

TABLE I
EVEN- AND ODD-MODE IMPEDANCES OF MARCHAND BALUNS WITH $R_r = 50\ \Omega$ AND $R_L = 100\ \Omega$

| $C$(dB) | -3 | -5 | -7 | -9 | -11 |
|---|---|---|---|---|---|
| $Z_{0e}(\Omega)$ | 242.4 | 128.5 | 80.73 | 54.99 | 39.24 |
| $Z_{0o}(\Omega)$ | 41.45 | 36 | 30.88 | 26.19 | 21.98 |

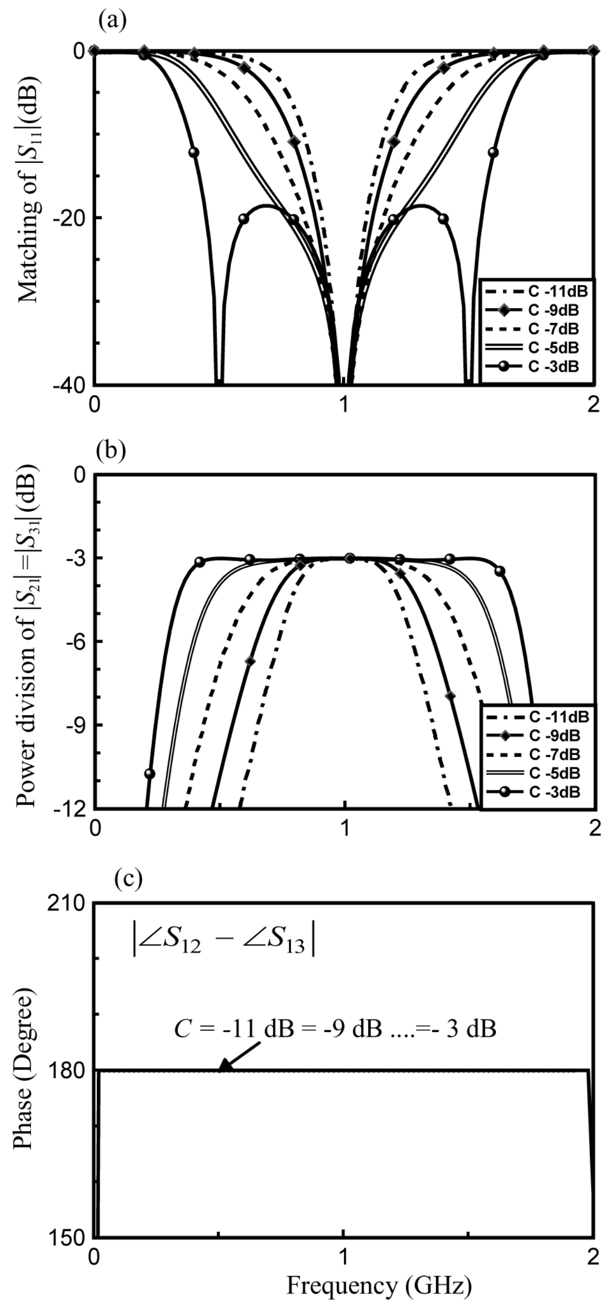


Fig. 2. Simulation results with $R_r = 50\ \Omega$, $R_L = 100\ \Omega$. (a) $|S_{11}|$. (b) $|S_{12}|$ and $|S_{13}|$. (c) Absolute phase difference between $S_{12}$ and $S_{13}$.

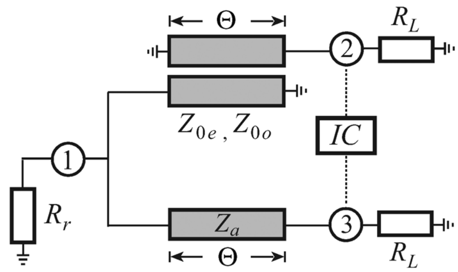


Fig. 3. One circuit being equivalent to that in Fig. 1(b) at a design center frequency.

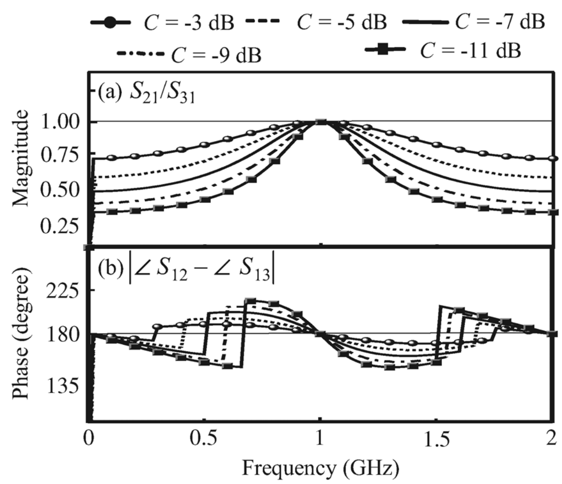


Fig. 4. Frequency responses. (a) Power division ratio of $S_{21}$ to $S_{31}$. (b) Absolute phase difference between $S_{21}$ and $S_{31}$.

absolute phase difference between $S_{21}$ and $S_{31}$ [see Fig. 2(c)] is 180° theoretically in whole frequencies, except even multiples of the design center frequency.

### D. Properties of Marchand Baluns

There are several ways to realize the equivalent circuit in Fig. 1(b) and one of them is depicted in Fig. 3, where the circuit consists of one set of coupled transmission-line sections with two short circuits in a diagonal direction and one transmission-line section. In this case, the design equations of the coupled transmission-line sections are the same as those in (9).

Under the same condition as the Marchand balun with $R_r = 50\,\Omega$ and $R_L = 100\,\Omega$ in Fig. 1(a) and Table I, the frequency responses of the equivalent circuit in Fig. 3 were simulated at the design center frequency of 1 GHz, and the simulation results are plotted in Fig. 4 where the magnitude of the ratio of $S_{21}$ to $S_{31}$ is in Fig. 4(a) and absolute phase difference between $S_{21}$ and $S_{31}$ is in Fig. 4(b). The ratio of $S_{21}$ to $S_{31}$ in Fig. 4(a) is unity at the design center frequency for all coupling coefficients, and the difference between $S_{21}$ to $S_{31}$ is gradually bigger with the coupling coefficient smaller and with the operating frequency farther from the center frequency. In contrast to this, the ratio of $S_{21}$ to $S_{31}$ in the Marchand balun is unity independently of the coupling coefficient and the operating frequency [see Fig. 2(b)]. The absolute phase difference [see Fig. 4(b)] between $S_{21}$ and $S_{31}$ of the equivalent circuit in Fig. 3 is 180° only at the design center frequency of 1 GHz for any coupling coefficient, and the deviation from the 180° phase difference is smaller with the higher coupling coefficient outside of the center frequency. For the Marchand balun in Fig. 2(c), the 180° phase difference between two output signals is achieved independently of the coupling coefficient and the operating frequencies. Due to the attractive properties of the Marchand baluns, they have received substantial attention from circuit designers and have been used for diverse applications for a long time.

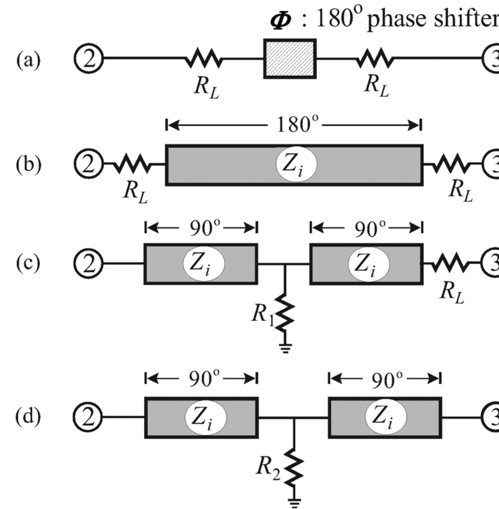


Fig. 5. ICs. (a) One 180° phase shifter and two series isolation resistances. (b) One 180° transmission-line section with two series resistances. (c) Two 90° transmission-line sections with one series and one shunt resistances. (d) Two 90° transmission-line sections with one shunt resistance.

## III. ICs

### A. ICs

Even without the IC (Fig. 1), perfect matching at port ① and perfect power divisions are achieved as shown in Fig. 2, but perfect matching at ports ② and ③ and perfect isolation between ports ② and ③ do not appear. For the perfect balun performance at all ports, the IC is needed. Since the phase delays of $S_{21}$ and $S_{31}$ are 90° and −90°, respectively, the admittance parameters of the IC between ports ② and ③ [see Fig. 1(a)] [24] are

$$\begin{bmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{bmatrix}_{IC} = \frac{1}{2R_L}\begin{bmatrix} 1 & 1 \\ 1 & 1 \end{bmatrix}. \tag{10}$$

A circuit with two series resistances of $R_L$ and an 180° phase shifter connected in cascade, as depicted in Fig. 5(a), satisfies the admittance parameters in (10). In addition, the three ICs in Fig. 5 are possible. To realize the 180° phase shifter, one 180° transmission-line section may be employed in Fig. 5(b). Moving one or two resistances of $R_L$ in Fig. 5(b) to the center of the 180° transmission-line section results in the circuit in Fig. 5(c) or (d), respectively. The relation among the characteristic impedance $Z_i$ of the transmission-line section, the output termination impedance of $R_L$ and the resistances of $R_1$ and $R_2$ [24] is

$$Z_i^2 = R_1 R_L \tag{11a}$$

$$R_2 = \frac{R_1}{2}. \tag{11b}$$

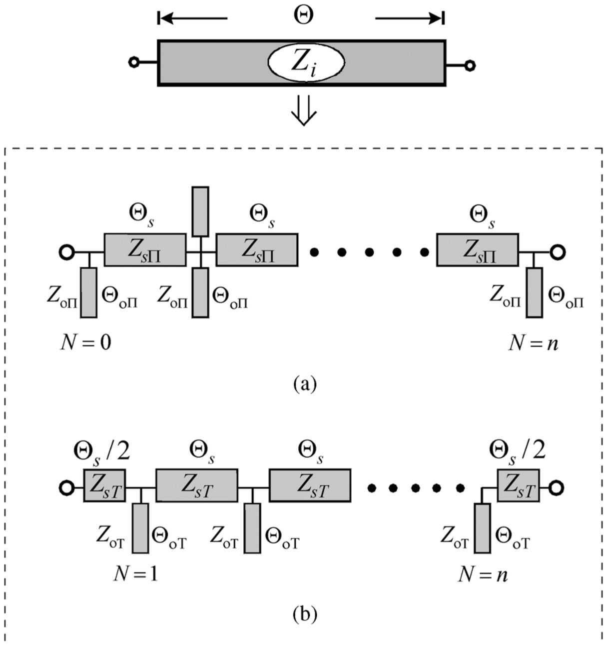


Fig. 6. Modified $\Pi$- and $T$-types of equivalent circuits. (a) $\Pi$-type. (b) $T$-type.

## *B. Compact ICs*

As shown in Fig. 1(a), since two output ports are connected to each other very closely, it is not easy to implement such long ICs in Fig. 5. The transmission-line sections should therefore be reduced for easy fabrication. To shorten the transmission-line sections without changing phase response, modified $\Pi$- and $T$-types of equivalent circuits may be utilized and detailed in Fig. 6 where the characteristic impedance and electrical length of the original transmission section are $Z_i$ and $\Theta$. The modified $\Pi$-type equivalent circuit [see Fig. 6(a)] consists of $N$ transmission-line sections with the characteristic impedance of $Z_{s\Pi}$ and $N+1$ open stubs. The first and last open stubs are the same and the others have two times susceptance of the first or last one. The number of $N$ is therefore equal to the number of the transmission-line sections with $Z_{s\Pi}$. The $T$-type equivalent circuit [see Fig. 6(b)] is a kind of dual network of the $\Pi$-type [see Fig. 6(a)] and composed of $N$ open stubs and $N+1$ transmission-line sections. The electrical length of the first and last transmission-line sections with the characteristic impedance of $Z_{sT}$ [see Fig. 6(b)] is half of the others and the others are therefore two times of the first or last one. The characteristic impedances of $Z_{o\Pi}$ and $Z_{oT}$ and the lengths of $\Theta_{o\Pi}$ and $\Theta_{oT}$ of the open stubs may be chosen arbitrarily as far as their susceptance values of $Y_{o\Pi}\cdot\tan\Theta_{o\Pi}$ and $Y_{oT}\cdot\tan\Theta_{oT}$ are kept the same, where $Y_{o\Pi}=Z_{o\Pi}^{-1}$ and $Y_{oT}=Z_{oT}^{-1}$.

The design equations of the equivalent circuits in Fig. 6 are

$$Z_{s\Pi}=Z_i\frac{\sin\dfrac{\Theta}{N}}{\sin\Theta_s} \tag{12a}$$

$$Y_{o\Pi}\cdot\tan\Theta_{o\Pi}=\frac{1}{Z_i}\left(\frac{\cos\Theta_s-\cos\dfrac{\Theta}{N}}{\sin\dfrac{\Theta}{N}}\right) \tag{12b}$$

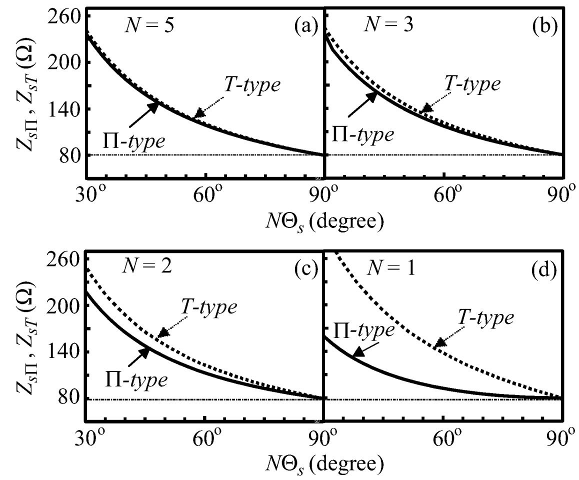


Fig. 7. Characteristic impedance of $Z_{s\Pi}$ and $Z_{sT}$ with $Z_i=80\ \Omega$. (a) $N=5$. (b) $N=3$. (c) $N=2$. (d) $N=1$.

TABLE II
DESIGN DATA OF $\Pi$- AND $T$-TYPES OF MODIFIED EQUIVALENT CIRCUITS WITH $Z_i=80\ \Omega$, $\theta=90°$, AND $N\theta_s=70°$

| | $\Pi$-type | | $T$-type | |
|---|---|---|---|---|
| $N$ | $Z_{s\Pi}$ ($\Omega$) | $S_{u\Pi}$ (Ʊ) | $Z_{sT}$ ($\Omega$) | $S_{uT}$ (Ʊ) |
| $N=5$ | 102.187 | 0.00078 | 103.195 | 0.00154 |
| $N=3$ | 100.990 | 0.00130 | 103.814 | 0.00254 |
| $N=2$ | 98.624 | 0.00198 | 105.097 | 0.00372 |
| $N=1$ | 85.134 | 0.00428 | 114.252 | 0.00637 |

$$Z_{sT}=Z_i\frac{\tan\dfrac{\Theta}{2N}}{\tan\dfrac{\Theta_s}{2}} \tag{13a}$$

$$Y_{oT}\cdot\tan\Theta_{oT}=\frac{2}{Z_{sT}}\cdot\frac{Z_{sT}-Z_i\cot\dfrac{\Theta}{2N}\tan\dfrac{\Theta_s}{2}}{Z_{sT}\tan\dfrac{\Theta_s}{2}+Z_i\cot\dfrac{\Theta}{2N}} \tag{13b}$$

where $\Theta>N\Theta_s, Y_{o\Pi}, Y_{oT}$ are characteristic admittances of open stubs, and $\Theta_s, \Theta_{o\Pi}, \Theta_{oT}$ are electrical lengths of transmission-line sections and open stubs in Fig. 6. The $\Pi$-type of the equivalent circuit with $N=1$ have been used for various applications [24], [26], [29], [33], but that with $N$ greater than 2 has not been discussed yet. For the use of the equivalent circuits, the characteristic impedances of $Z_{s\Pi}$ and $Z_{sT}$ should be realizable. For this, the relation among $Z_{s\Pi}$, $Z_{sT}$, and $N$ will be studied.

For a transmission-line section with the characteristic impedance of $Z_i=80\ \Omega$ and the electrical length of 90°, the characteristic impedances of $Z_{s\Pi}$ and $Z_{sT}$ were calculated with $N=5,3,2$, and 1, and the calculation results are plotted in Fig. 7 where $N\Theta_s$ is the total length of the resulting transmission-line section. When $N=5$, the characteristic impedances of $Z_{s\Pi}$ and $Z_{sT}$ are about the same with each other [see Fig. 7(a)]. With $N$ smaller, the difference between $Z_{s\Pi}$ and $Z_{sT}$ becomes bigger [see Fig. 7(d)]. In the case of $N\theta_s=70°$, $Z_{s\Pi}$, $S_{u\Pi}$, $Z_{sT}$, and $S_{uT}$ are calculated in Table II where $S_{u\Pi}=Y_{o\Pi}\cdot\tan\Theta_{o\Pi}$ and $S_{uT}=Y_{oT}\cdot\tan\Theta_{oT}$ in (12b) and (13b).

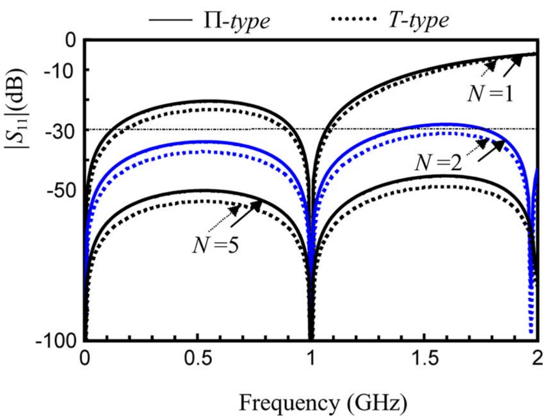


Fig. 8. Simulation results of $|S_{11}|$.

TABLE III
FABRICATION DATA OF A MICROSTIP MARCHAND BALUN

| | |
|---|---|
| Coupled transmission-line sections; $C$ = -10 dB | |
| $Z_{0e}$ = 62.39 Ω, $Z_{0o}$ = 32.41 Ω | $w$ = 1.915 mm, $s$ = 0.1655 mm |
| Isolation circuit in Fig. 5(d) ; $R_L$ = 70 Ω and $R_2$ = 51 Ω | |
| $Z_{sT}$ = 116.5 Ω, $\theta_s$ = 22° with $N$ = 3 | |
| Impedance transformers | |
| $R_r$ = 130 Ω -> 50 Ω | |
| CVT$_r$ : 50 Ω with 20° long and 97 Ω with 44.15°. | |
| $R_L$ = 70 Ω -> 50 Ω | |
| CVT$_L$ : 50 Ω with 30° long and 70.71 Ω with 28.56°. | |

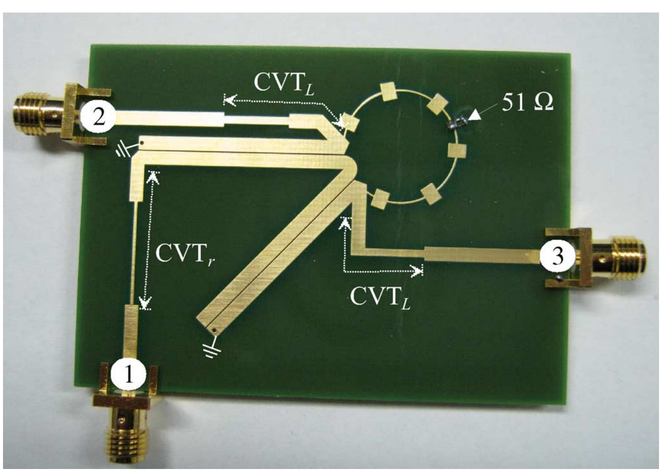


Fig. 9. Fabricated marchand balun.

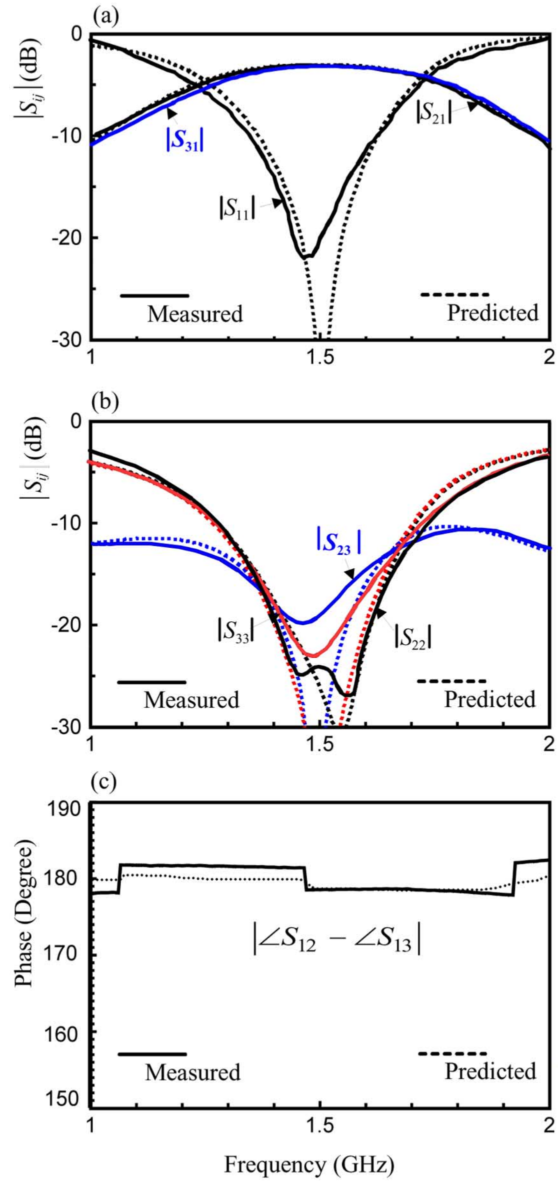


Fig. 10. Results measured and predicted are compared. (a) Scattering parameters from port ①. (b) Scattering parameters from ports ② and ③. (c) Absolute phase difference between $S_{12}$ and $S_{13}$.

Based on the data in Table II, the two types of equivalent circuits were simulated at the design center frequency of 1 GHz and ideal capacitances were used for the susceptances produced by the open stubs. The frequency responses of matching performance are plotted in Fig. 8 where the solid and dotted lines are those of the $\Pi$- and $T$-types of the modified equivalent circuits, respectively. When $N = 5$ (Fig. 8), the return loss is more than 50 dB in the entire frequency range of interest. With $N = 2$, the return loss with more than 30 dB is achieved, except several frequencies around 1.6 GHz, and with $N = 1$, poor response is shown in the frequencies higher than 1 GHz. For the use of the 90° transmission-line sections, the number of $N$ should be greater and equal to 2 for good matching performance.

## IV. MEASUREMENTS

To validate design formulas and modified equivalent circuits of a transmission-line section, a microstrip Marchand balun terminated in $R_r = 130\ \Omega$ and $R_L = 70\ \Omega$ was fabricated on a substrate (FR4, $\varepsilon_r = 4.2$, $H = 1$ mm). The dielectric constant of FR4 is found based on a third-order polynomial [34]. $C = -10$ dB and a design center frequency of 1.5 GHz were chosen. The corresponding even- and odd-mode impedances are $Z_{0e} = 62.39\ \Omega$ and $Z_{0o} = 32.42\ \Omega$, and width $w$ and gap size $s$ of the coupled microstrip transmission-line sections are 1.92 and 0.165 mm, respectively. For the conventional design equations in (1), the coupling coefficient of $C_c$ is $-3.174$ dB (0.6939) and the even- and odd-mode impedances should be $Z_{0e} = 305.8\ \Omega$ and $Z_{0o} = 55.26\ \Omega$, which are impossible to be implemented with a microstrip format on any substrate. The IC in Fig. 5(d) was employed and $Z_i = 84.5\ \Omega$ in Fig. 5(d) was calculated for the use of an available chip resistor of 51 Ω. Each 90° transmission-line section was reduced using the modified $T$-type equivalent circuit to have $N\Theta_s = 66°$ where $N = 3$. The susceptance of $S_{uT}$ produced by one open stub is 0.00280 ℧ and was realized with two open stubs connected in parallel. Each open stub is 4° long and the characteristic impedance is 50 Ω. The two termination impedances are not 50 Ω, and therefore, impedance transformers are needed. To transform $R_r = 130\ \Omega$ into 50 Ω, an asymmetric impedance transformer $\mathrm{CVT}_r$ [35], [36] was adopted, and to convert $R_L = 70\ \Omega$ into 50 Ω, $\mathrm{CVT}_L$ was employed. The detailed data are summed up in Table III. If the resistance value of $R_2$ is lowered in Fig. 5(d), then the characteristic impedance of $Z_i$ is lowered from (11), which leads to the size reduction more from the calculation results in Fig. 7.

The fabricated Marchand balun is displayed in Fig. 9, and the results measured and predicted are compared in Fig. 10 where the measured scattering parameters produced from port ① are in Fig. 10(a), those from ports ② and ③ in Fig. 10(b) and absolute phase difference between $S_{12}$ and $S_{13}$ in Fig. 10(c). Measured scattering parameters of the power divisions are $-3.57$ and $-3.62$ dB, measured return losses at all ports are better than about 21 dB, and the isolation is better than about 20 dB around the design center frequency of 1.5 GHz. The absolute phase difference between two balanced signals is $180 \pm 2°$ in about 50% bandwidth. The measured results are in good agreement with the prediction, as displayed in Fig. 10.

## V. Conclusions

Conventional design formulas of the even- and odd-mode impedances of the Marchand balun are determined by one coupling coefficient only, being a function of the termination impedances. The coupling coefficient may therefore be 0 or $-3$ dB, depending on the termination impedances. In general, the coupling coefficient should be determined by the coupling structure itself, not by the termination impedances. To solve the problem, new design formulas are derived by terminating input port with its matched termination impedance. Infinite sets of the even- and odd-mode impedances are possible by varying the coupling coefficient arbitrarily, which allows design flexibility and diverse applications. Since two output ports of the Marchand baluns are placed very closely with each other, no sufficient space for the long IC is available. For this, modified $\Pi$- and $T$-types of equivalent circuits are also proposed.

Using the presented design formulas of the Marchand baluns and the modified $\Pi$- and $T$-types of equivalent circuits, further applications can be expected.

## References


[1] N. Marchand, "Transmission line conversion transformers," *Electronics*, vol. 17, no. 12, pp. 142–145, Dec. 1944.

[2] L. Xu, Z. Wang, Q. Li, and J. Xia, "Modelling and design of a wideband Marchand balun," in *Asia–Pacific Electromagn. Compat. Symp. Dig.*, 2010, pp. 1374–1377.

[3] A. C. Chen, A.-V. Pham, and R. E. Leoni, III, "A novel broadband even-mode matching network for Marchand baluns," *IEEE Trans. Microw. Theory Tech.*, vol. 57, no. 12, pp. 2973–2980, Dec. 2009.

[4] C.-S. Lien, C.-H. Wang, C.-S. Lin, P.-S. Wu, K.-Y. Lin, and H. Wang, "Analysis and design of reduced-size Marchand rat-race hybrid for millimeter-wave compact balanced mixers in 130-nm CMOS process," *IEEE Trans. Microw. Theory Tech.*, vol. 57, no. 8, pp. 1966–1977, Aug. 2009.

[5] Z. Xu and L. NacEachern, "Optimum design of wideband compensated and uncompendated Marchand baluns with step transformers," *IEEE Trans. Microw. Theory Tech.*, vol. 57, no. 8, pp. 2064–2071, Aug. 2009.

[6] T. Johnsen and V. Krozer, "Analysis and design of lumped element Marchand baluns," in *17th Int. Microw., Radar, Wireless Commun. Conf. Dig.*, May 2008, pp. 1–4.

[7] S. Kumpang, R. Phromloungsri, and M. Chongcheawchawchamnan, "Design high performance Marchand balun with step-impedance transmission lines compensated parallel-coupled lines," in *Asia–Pacific Microw. Conf. Dig.*, 2008, 4 pp.

[8] C.-S. Lin, P.-S. Wu, M.-C. Yeh, J.-S. Fu, H.-Y. Chang, K.-Y. Lin, and H. Wang, "Analysis of multiconductor coupled-line Marchand baluns for miniature MMIC design," *IEEE Trans. Microw. Theory Tech.*, vol. 55, no. 6, pp. 1190–1199, Jun. 2007.

[9] R. Phromlungsri, M. Chongcheawchamnan, and I. D. Robertson, "Inductively compensated parallel coupled microstrip lines and their applications," *IEEE Trans. Microw. Theory Tech.*, vol. 54, no. 9, pp. 3571–3582, Sep. 2006.

[10] W. M. Fathelbab and M. B. Steer, "Tapped Marchand baluns for matching applications," *IEEE Trans. Microw. Theory Tech.*, vol. 54, no. 6, pp. 2543–2551, Jun. 2006.

[11] R. Phromloungsri, S. Srisathit, M. Chongcheawchamnan, and I. D. Robertson, "Novel technique for performance improvement in impedance-transforming planar Marchand bluns," in *Eur. Microw. Conf. Dig.*, 2005, 4 pp.

[12] Z.-Y. Zhang, Y.-X. Guo, L. C. Ong, and M. Y. W. Chia, "A new planar Marchand balun," in *IEEE MTT-S Int. Microw. Symp. Dig.*, 2005, pp. 1207–1210.

[13] W. M. Fathelbab and M. B. Steer, "New classes of miniaturized planar Marchand baluns," *IEEE Trans. Microw. Theory Tech.*, vol. 53, no. 4, pp. 1211–1220, Apr. 2005.

[14] H.-R. Ahn, "Comments on 'Converting baluns into broadband impedance-transforming 180° hybrids'," *IEEE Trans. Microw. Theory Tech.*, vol. 52, no. 1, pp. 228–230, Jan. 2004.

[15] K. S. Ang and Y. C. Leong, "Authors' reply "comments on "converting baluns into broadband impedance-transforming 180° hybrids"," *IEEE Trans. Microw. Theory Tech.*, vol. 52, no. 1, pp. 230–233, Jan. 2004.

[16] C. Y. Ng, M. Chongcheawchamnan, and I. D. Robertson, "Analysis and design of a high-performance planar Marchand balun," in *IEEE MTT-S Int. Microw. Symp. Dig.*, 2002, pp. 113–116.

[17] K. S. Ang and Y. C. Leong, "Converting baluns into broadband impedance-transforming 180° hybrids," *IEEE Trans. Microw. Theory Tech.*, vol. 50, no. 8, pp. 1990–1995, Aug. 2002.

[18] J.-L. Chen, S.-F. Chang, and B.-Y. Laue, "A 20–40 GHz monolithic doubly-balanced mixer using modified planar Marchand baluns," in *Asia–Pacific Microw. Conf. Dig.*, Dec. 2001, pp. 131–134.

[19] C.-W. Tang, J.-W. Sheen, and C.-Y. Chang, "Chip-type LTCC-MLC baluns using the stepped impedance method," *IEEE Trans. Microw. Theory Tech.*, vol. 49, no. 12, pp. 2342–2349, Dec. 2001.

[20] M. Chongcheawchamnan, C. Y. Ng, M. S. Aftanasar, I. D. Robertson, and J. Minalgiene, "Broadband CPW Marchand balun using photoimageable multilayer thick-film process," *Electron. Lett.*, vol. 37, no. 20, pp. 1228–1229, Sep. 2001.

[21] K. S. Ang and I. D. Robertson, "Analysis and design of impedance-transforming planar Marchand baluns," *IEEE Trans. Microw. Theory Tech.*, vol. 49, no. 2, pp. 402–406, Feb. 2001.

[22] M. Chongcheawchamnan, K. S. Ang, J. N. H. Wong, and I. D. Robertson, "A push–pull power amplifier using novel impedance-transforming baluns," in *Eur. Microw. Conf. Dig.*, 2000, 4 pp.

[23] K. S. Ang, I. D. Robertson, K. Elgaid, and I. G. Thayne, "40 to 90 GHz impedance-transforming CPW Marchand balun," in *IEEE MTT-S Int. Microw. Symp. Dig.*, 2000, pp. 1141–1144.

[24] H.-R. Ahn and T. Itoh, "New isolation circuits of compact impedance-transforming 3-dB baluns for theoretically perfect isolation and matching," *IEEE Trans. Microw. Theory Tech.*, vol. 58, no. 12, pp. 3892–3902, Dec. 2010.

[25] H. R. Ahn and T. Itoh, "Impedance-transforming symmetric and asymmetric DC blocks," *IEEE Trans. Microw. Theory Tech.*, vol. 58, no. 9, pp. 2463–2474, Sep. 2010.

[26] H.-R. Ahn and B. Kim, "Small wideband coupled-line ring hybrids with no restriction on coupling power," *IEEE Trans. Microw. Theory Tech.*, vol. 57, no. 7, pp. 1806–1817, Jul. 2009.

[27] H.-R. Ahn and B. Kim, "Transmission-line directional couplers for impedance transforming," *IEEE Microw. Wireless Compon. Lett.*, vol. 16, no. 10, pp. 537–539, Oct. 2006.

[28] H.-R. Ahn and B. Kim, "Toward integrated circuit size reduction," *IEEE Microw. Mag.*, vol. 9, pp. 65–75, Feb. 2008.

[29] T. Hirota, A. Minakawa, and M. Muraguchi, "Reduced-size branch-line and rat-race hybrids for uniplanar MMIC's," *IEEE Trans. Microw. Theory Tech.*, vol. 38, no. 3, pp. 270–275, Mar. 1990.

[30] K. W. Eccleston and S. H. M. Ong, "Compact planar microstripline branch-line and rat-race couplers," *IEEE Trans. Microw. Theory Tech.*, vol. 51, no. 10, pp. 2119–2125, Oct. 2003.

[31] H.-R. Ahn, I.-S. Chang, and S.-W. Yun, "Miniaturized 3-dB ring hybrid terminated by arbitrary impedances," *IEEE Trans. Microw. Theory Tech.*, vol. 42, no. 12, pp. 2216–2221, Dec. 1994.

[32] H.-R. Ahn, I. Wolff, and I.-S. Chang, "Arbitrary termination impedances, arbitrary power division, and small-sized ring hybrids," *IEEE Trans. Microw. Theory Tech.*, vol. 45, no. 12, pp. 2241–2247, Dec. 1997.

[33] M.-L. Chuang, "Miniaturized ring coupler of arbitrary reduced size," *IEEE Mircrow. Wireless Compon. Lett.*, vol. 15, no. 1, pp. 16–18, Jan. 2005.

[34] E. L. Holzman, "Wideband measurement of the dielectric constant of an FR4 substrate using a parallel-coupled microstrip resonators," *IEEE Trans. Microw. Theory Tech.*, vol. 54, no. 7, pp. 3127–3130, Jul. 2006.

[35] H.-R. Ahn and I. Wolff, "General design equations, small-sized impedance transformers, and their applications to small-sized three-port 3-dB power dividers," *IEEE Trans. Microw. Theory Tech*, vol. 49, no. 7, pp. 1277–1288, Jul. 2001.

[36] H.-R. Ahn, *Asymmetric Passive Components in Microwave Integrated Circuits*. New York: Wiley, 2006, ch. 11.


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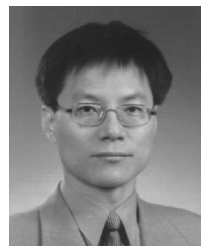

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