0IO-Shape PCB Trace Negative Group-Delay Analysis
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ABSTRACT This paper elaborates a negative group delay (NGD) analysis of 0IO-shape printed circuit board (PCB) traces. This circuit topology is originally implemented with a tri-coupled line (3CL) six-port element with the lateral side connected through lossy transmission lines (TLs). After description of the electrical equivalent diagram, the S-matrix model is established. The group delay (GD) is formulated from the transmission coefficient as a function of the 0IO topological parameters. The effectiveness of the GD modelling is verified with a microstrip circuit proof-of-concept (POC). Simulations and measurements, which are in good agreement, confirm the dual-band bandpass NGD behavior of the 0IO POC. The fabricated prototype generates NGD levels better than \(-1\) ns at NGD center frequencies of about 2.2 GHz and 3 GHz. In addition, to this good NGD performance, the 0IO POC operates with a low insertion loss better than 2.5 dB and reflection losses better than 12 dB in the NGD bandwidths.

INDEX TERMS Microwave theory, distributed topology, negative group delay (NGD), tri-coupled line (3CL), modeling.

I. INTRODUCTION
The group delay (GD) is a key parameter in microwave electronic circuit systems. The GD was exploited for design electronic functions as phase shifting [1], antenna arrays [2] and feedforward amplifier [3] etc. To deal with the unexpected GD issue, the extra delay line is regularly used. The compensation consists generally in compensating the unbalanced GD in the RF and microwave components as in feedforward amplifier [4]. However, this classical solution may be penalizing by adding delay line. Furthermore, the delay line structure can also increase the circuit size greatly. To overcome this issue, recently, negative group delay (NGD) circuits are thought to be a good solution to equalize the microwave electronic system delay.

In [1], the NGD function was applied to phase shifter to generate at and constant phases over a broad bandwidth. A transmission lines loaded active NGD network is introduced in [2] to solve the beam-squinting problem of conventional series-fed antenna arrays. An NGD circuit is used to enhance feedforward amplifier efficiency by eliminating the delay element, which is one of the major sources of efficiency degradation, without affecting the linearization performance [3]. The NGD effect was also used to improve the flatness of power divider [5]. The NGD structure is constructed with a two-way microstrip line power divider with equal power division ratio. The power divider different transmission paths generate NGD effect despite the insertion losses. From above, it can be seen that the NGD circuits have very extensive applications on RF/microwave devices and systems.

The first NGD synthesizers, with microwave passive circuits were proposed in [6], [7] and with very low-frequency
active circuits were introduced in [8], [9]. The NGD passive circuits using lumped R, L and C elements only work at frequencies lower than 1 GHz [10], [11], limiting the available scope of microwave applications. Since NGD occurs at the range of frequencies where the absorption or the attenuation is maximum, the NGD generation is accompanied by excessive isolation loss more than 20 dB [6], [7]. Therefore, the cascading of active elements such as amplifiers is a practical compensation for these attenuations [12], [13]. Thus, in [13], a field effect transistor (FET) is applied to design an active NGD circuit with 2dB gain. However, it was found that these active NGD circuits would unavoidably suffer from design inflexibility restrictions about the fixed component values and increasing design difficulties in the microwave band. And also, it can increase the out-of-band noise as well as make the circuit more complicated. Moreover, such an active topology is rather complex to design and difficult to integrate because of the lossy lumped inductor.

As a result, simpler and low-loss passive topologies built with distributed transmission lines (TLs) were implemented [12]–[19]. To reduce the attenuation lower than 10 dB, a coupled line based NGD circuit is designed in [16] whose attenuation is decreased to 7.43 dB. A parallel interconnect line (PIL) NGD circuit is designed in [18]. This PIL NGD circuit is able to operate with attenuation of about 5 dB. An NGD circuit consists of the isolated- and coupled-accesses connected in a feedback loop as a coupling between “1” and “0” shape interconnect line which presents 2.4 dB insertion loss [19]. Such geometrical shape as “01O” sensitively with electromagnetic interference (EMI) can be found in the PCB traces.

These “01O” configurations can be found in PCB layouts constituted by copper plane hotspots, vias interconnects, serpentine and twirling paths (for length compensation, pads interconnects, symmetrical breakout etc.). It would be important to investigate the coupling effect modelling of this structure and the possibility of NGD effect generation. The motivation of this work is to properly assess the impact of NGD generation (quantifying and/or evaluating it) with respect to tri-coupled line EMC (electromagnetic compatibility) and EMI scenarios. When dealing with coupling scenarios involving “01O” structures, it is of utmost importance for the EMC designer to ensure the system functioning taking into account NGD effects. Compared with the existing single-band NGD circuits [12]–[19], [21]–[23], the design of dual-band NGD circuits [24]–[27] remains a challenging task. In order to enable the NGD circuit to operate in different frequency bands, few studies are focused on the design of dual-band NGD circuits. A dual-plane U-shaped defected structure is used to realize a dual band NGD circuit [25], whose first center frequency and second center frequency are determined by a defected microstrip structure (DMS) and a defected ground structure (DGS), respectively. However, the attenuation is worse than 44 dB. A compact dual-band NGD circuit composed of an open-circuited TL and two resistors connected by two TLs is proposed in [27]. The attenuation is worse than 16 dB. Therefore, it is especially important to design a low-loss dual-band NGD circuit.

For this reason, in this paper, a topology of dual band NGD passive topology based on fully distributed TLs with further challenge on TL loss and delay effect is developed. The challenging NGD topology presents an innovative shape similar to “01O” shape geometry. The topology is built with a tri-coupled line (3CL) six-port element with the lateral side connected through lossy transmission lines (TLs). The paper is mainly organized in four different sections as follows:

- The theoretical approach allowing to model the 01O topology S-matrix is described in Section II. After the GD modelling, the NGD analysis is introduced.
- The validity of the bandpass NGD functioning of the 01O topology is validated in Section III via excellent comparisons between calculation, simulation and measurement.
- Finally, Section IV concludes the paper.

II. NGD THEORETICAL INVESTIGATION ON “01O” TOPOLOGY

The present section is focused on the NGD theoretical approach of the 01O topology. The S-matrix model is established from wave power interactions between the topology constituting elements. The GD expression is derived from the transmission coefficient. Then, NGD analysis is introduced in function of the 01O parameters.

A. DESCRIPTION OF THE 01O TOPOLOGY

Fig. 2 sketches the topology of 01O structure under study. It is comprised mainly of a six-port tri-coupled line (3CL) denoted
CL($R_0, k$) having characteristic impedance $R_0 = 50$ Ω and the adjacent line coupling coefficient $k$. We suppose that the coupling between line TL$_{3-\cdots-6}$ and TL$_{3-\cdots-6}$ is negligible. The main access ports are constituted by $\odot$ and $\odot$.

The lateral lines are interconnected through different lossy TLs TL$_k(Z_k, a_k, \tau_k)$ for $k = \{1, 2\}$ which are supposed to be ideal with same characteristic impedances $Z_k = R_0$ and attenuation $a_k = a$. However, they present different propagation delay $\tau_1 < \tau_2$.

After the equivalent circuit introduction, the S-matrix modeling of the 0IO topology will be explored in the next paragraph.

**B. S-MATRIX EQUIVALENT MODEL OF 0IO STRUCTURE CIRCUIT**

To establish the S-matrix model of the 0IO topology, we propose to consider the equivalent diagram composed of TL$_1$, TL$_2$ and CL S-matrix black boxes $\{S\}_{TL_1}$, $\{S\}_{TL_2}$ and $\{S\}_{CL}$ shown in Fig. 2, respectively.

The modeling of our hexapole circuit is built with the consideration of equivalent S-parameters. The main constituting elements are a CL and the four pieces of terminal TLs introduced in Fig. 2. Based on the S-parameter theory, the general S-parameter of the 0IO topology can be written in function of the wave powers. The CL six-port S-matrix constituting the central element is linked to the input and output wave powers $a_m$ and $b_m$ ($m = \{1, 2, \ldots, 6\}$) by the relation:

$$[b_1 \ b_2 \ b_3 \ b_4 \ b_5 \ b_6] = [S]_{CL} \times \begin{bmatrix} a_1 \\ a_2 \\ a_3 \\ a_4 \\ a_5 \\ a_6 \end{bmatrix}.$$  

(1)

The main objective of the theoretical approach is to determine the overall equivalent two-dimension S-matrix $[S]_{0IO}$ by reducing this six-dimension matrix. Meanwhile, this total S-matrix can be determined from the constituting TL and CL ones. Substituting the expressions of wave powers $a_m$ and $b_m$ ($m = \{3, \ldots, 6\}$) into (1), the overall S-matrix is defined by the relation:

$$[b_1] = [S] \times [a_1 \ a_2] .$$  

(2)

1) S-MATRIX OF CL ELEMENT

According to the coupler theory, the CL S-matrix can be written in function of the two-neighboring lines coupling (between ports $\odot-\odot$, $\odot-\odot$, $\odot-\odot$ and $\odot-\odot$) $k$ and direct transmission (between ports $\odot-\odot$, $\odot-\odot$ and $\odot-\odot$):

$$k_0 = -j g = -j \sqrt{1 - k^2} ,$$  

(3)

2) S-MATRIX MODEL OF TL1,2

Following the topological description introduced in Fig. 3, the TL$_k$ for $k = \{1, 2\}$ S-matrices are linked to the wave powers by the relation:

$$[a_3 \ a_4 \ a_5] = [S]_{TL_1} \times [b_3 \ b_4 \ b_5] .$$  

(4)

By denoting $\omega$ is the angular frequency variable, let us take:

$$x(j\omega) = a \ e^{-j\omega \tau_k} .$$  

(5)

Therefore, the equivalent TL$_1$ and TL$_2$ S-matrices can be written as:

$$[S(j\omega)]_{TL_k} = \begin{bmatrix} 0 & x_k(j\omega) \\ x_k(j\omega) & 0 \end{bmatrix} .$$  

(6)

3) EXPRESSION OF 0IO STRUCTURE GLOBAL S-MATRIX

It can be derived from (4) and (6) that wave powers $a_m$ and $b_m$ ($m = \{3, \ldots, 6\}$) can be expressed in function of $a_1,2$ and...
$b_{1,2}$. Substituting the according equations into (1) combined with (4), the 0IO S-matrix introduced in (2) is written as:

$$[S(j\omega)] = \begin{bmatrix} 0 & S_{21}(j\omega) \\ S_{21}(j\omega) & 0 \end{bmatrix}$$

with the transmission coefficient:

$$S_{21}(j\omega) = \frac{x_1(j\omega) + x_2(j\omega)}{[1 + jg(x_1(j\omega) + x_2(j\omega) - 1]}. (9)$$

Based on this last expression, the NGD analysis is introduced in the next subsection.

C. ANALYSES OF S-MATRIX FREQUENCY RESPONSES

The transmission coefficient magnitude, $S_{21}(\omega) = |S_{21}(j\omega)|$, can be formulated as:

$$S_{21}(\omega) = \sqrt{\frac{\cos[\omega(\tau_1 + \tau_2)] + \sin[\omega(\tau_1 + \tau_2)]^2}{\cos[\omega(\tau_1 + \tau_2)] + \sin[\omega(\tau_1 + \tau_2)]^2 - a^2g^2 - a\{\sin[\omega(\tau_1 + \tau_2)] + \sin[\omega(\tau_1 + \tau_2)]\}^2}}. (10)$$

The transmission phase is defined by:

$$\varphi(\omega) = \angle S_{21}(j\omega) = \varphi_0(\omega) - \varphi_d(\omega)$$

with

$$\varphi_0(\omega) = \arctan\left\{\frac{a\{\sin[\omega(\tau_1 + \tau_2)] + \sin[\omega(\tau_1 + \tau_2)]\}}{a^2(1 + k^2) - \cos[\omega(\tau_1 + \tau_2)]}\right\}$$

$$\varphi_d(\omega) = \arctan\left\{\frac{a\{\sin[\omega(\tau_1 + \tau_2)] + \sin[\omega(\tau_1 + \tau_2)]\}}{a^2(1 + k^2) - \cos[\omega(\tau_1 + \tau_2)]}\right\}. (12)$$

D. NGD ANALYSIS OF 0IO STRUCTURE

The 0IO topology GD is defined from the transmission phase expressed in (11) with the equation:

$$\tau(\omega) = -\frac{\partial \varphi_d(\omega)}{\partial \omega}. (14)$$

Thanks to equations (12) and (13), we can rewrite this GD as follows:

$$\tau(\omega) = \tau_d(\omega) - \tau_n(\omega)$$

with:

$$\tau_n(\omega) = \frac{\partial \varphi_n(\omega)}{\partial \omega}. (16)$$

and

$$\tau_d(\omega) = \frac{\partial \varphi_d(\omega)}{\partial \omega}. (17)$$

In details, the phase numerator GD is given by:

$$\tau_n(\omega) = \frac{v_1 + [v_2 + v_3\cos(\omega\tau_1)]\sin(\omega\tau_2)}{\sin(2\omega\tau_1)}/2$$

$$\tau_n(\omega) = \frac{v_1 + [v_2 + v_3\cos(\omega\tau_1)]\sin(\omega\tau_2)}{\sin(2\omega\tau_1)}/2$$

with:

$$\xi_1 = a^2k^4\cos(\omega\tau_2)$$

$$\xi_2 = g^2 + a^2$$

$$\xi_3 = a\left\{2\tau_1 + \left[1 + a^2(1 + k^2)\right]\tau_2\right\}$$

$$\xi_4 = a\left\{1 + a^2(1 + k^2)\right\} \tau_1 + 2\tau_2$$

$$\xi_5 = a^2g^2(\tau_1 + \tau_2)$$

$$v_1 = 2a^2k^4\cos(\omega\tau_2)$$

$$v_2 = a^2 + g^2 + 1 + a^4(1 + k^2)$$

$$v_3 = 2ag(1 + a^4 + k^2)$$

$$v_4 = 2a^2g^2.$$  

The phase denominator GD is expressed as:

$$\tau_d(\omega) = \frac{1 + a^2g^2 + 2a\sin(\omega\tau_1)}{1 + a^2g^2 + 2a\sin(\omega\tau_2)}$$

with:

$$\chi_1 = (1 + a^2g^2)\tau_1 + 2\tau_2$$

$$\chi_2 = 2\tau_1 + (1 + a^2g^2)\tau_2.$$ (22)

Based on these analytical expressions, parametric analyses versus TL delays $\tau_{1,2}$ where performed to visualize the NGD effect. More importantly, simulations and experimental studies with a POC were also realized. The obtained validation results will be explored in the next section.

III. EXPERIMENTAL VALIDATIONS WITH 0IO PROTOTYPES

As application of the previous theory, a prototype of 0IO POC has been designed, simulated and fabricated. The obtained results will be discussed in the next paragraph.

A. DESCRIPTION OF CONSIDERED 0IO NGD PROTOTYPES

To verify the relevance of the previous theory, 0IO POCs were designed and simulated with ADS®. The designed and photographed two-port circuits based on:

- FR4-epoxy substrate are viewed in Fig. 3(a) and in Fig. 3(c),
- and based on Rogers substrate are shown in Fig. 3(b) and in Fig. 3(d), respectively.

The designed circuits were simulated in the ADS® environment. We can see in these designs the lateral delay lines.
and the middle CL. Additional access lines TL3 and TL4 are added in order to facilitate the measurement configuration. The pictures of the 0IO POC prototype are shown in Fig. 3(c) and Fig. 3(d). These prototypes are implemented in microstrip technology without use of lumped and lossy component. The physical parameters of the FR4 and Rogers substrate 0IO circuit prototypes are indicated in Tables 1 and 2, respectively. The conductor lines are Cu-metallized. The physical sizes of constituting distributed lines TL1, TL2 and CL are also indicated in Table 1. The associated electrical parameters as characteristic impedances, delay and coupling coefficient calculated from ADS™ LineCalc microwave circuit calculation tool are also given in this table.

Based on this prototype, validation studies were performed with parametric analyses with respect to the TL lengths. Furthermore, practical investigation with experimental testing will be discussed in the following subsection.

### B. PARAMETRIC ANALYSES

To get a predictive insight about the bandpass NGD behavior of the 0IO topology, parametric analyses with respect to TL1 and TL2 delays $\tau_1$ and $\tau_2$, by means of physical lengths $d_1$ and $d_2$ were computed with S-parameter simulations from 2 GHz to 3.4 GHz. The obtained results are explored in the following paragraphs.

1) INFLUENCE OF TL1 DELAY

The influence of TL1 delay is studied with parametric simulations by varying physical length $d_1$ from 35 mm to 45 mm. Fig. 4(a) exposes the GD map versus frequency and $d_1$. This result reveals that the 0IO topology behaves as a dual-band bandpass NGD function. Two NGD bandwidths (BWs) can be identified. The first NGD BW situated around 2.2 GHz is zoomed in Fig. 4(b). It shows an NGD function that is not sensitive to $d_1$ variation. It means that this NGD bandwidth should depend to the loop related to TL2 and therefore, it must be linked to delay $\tau_2$. The second NGD BW area, situated between 2.73 GHz and 3.33 GHz is displayed in Fig. 4(c). This NGD BW varies inversely with delay $\tau_1$. Therefore, we denote $f_1$ the NGD center frequency associated to this second NGD bandwidth.

The analytical relations between the two NGD center frequencies $f_1$ and $f_2$ and delays $\tau_1$ and $\tau_2$ can be understood more clearly with the graphical plots of Fig. 5(a). At these NGD center frequencies, NGDs are approximately estimated as GD($f_1$) varying between $-1.3$ ns and $-1.2$ ns, and GD($f_2$) $\approx -1.4$ ns. In addition to the GD analysis, the $S_{21}$ variation with respect to $d_1$ is displayed in Figs. 6. The low loss aspect with insertion

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### TABLE 1. Parameters of the 0IO prototype shown in Figs. 3(a) and 3(c).

| Components                  | Description            | Parameter | Value |
|-----------------------------|------------------------|-----------|-------|
| Dielectric substrate (FR4)  | Relative permittivity  | $\varepsilon_r$ | 4.4   |
|                             | Loss tangent           | $\tan(\delta)$ | 0.02  |
|                             | Thickness              | $h$       | 1.6 mm|
| Metallization (Cu)          | Conductor thickness    | $t$       | 35 $\mu$m|
|                             | Conductivity           | $\sigma$  | 58 MS/m|
| Feedback line TL1           | Physical length        | $d_1$     | 40 mm |
|                             | Attenuation loss       | $a_1$     | -0.2 dB|
|                             | Delay                  | $\tau_1$  | 0.24 ns|
|                             | Partial length         | $L_1$     | 3.515 mm|
| Feedback line TL2           | Length                 | $d_2$     | 40 mm |
|                             | Attenuation loss       | $a_2$     | -0.2 dB|
|                             | Delay                  | $\tau_2$  | 0.37 ns|
|                             | Partial length         | $L_2$     | 13.515 mm|
| Coupled CL                  | Width                  | $w$       | 3 mm  |
|                             | Interspace             | $s$       | 2 mm  |
|                             | Coupling coefficient   | $k$       | -21 dB|
| Access lines                | Width                  | $d_4$     | 10 mm |

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### TABLE 2. Parameters of the 0IO prototype shown in Fig. 3(b) and Fig. 3(d).

| Components                  | Description            | Parameter | Value |
|-----------------------------|------------------------|-----------|-------|
| Dielectric substrate (Rogers3210) | Relative permittivity  | $\varepsilon_r$ | 4.2   |
|                             | Loss tangent           | $\tan(\delta)$ | 0.0027|
|                             | Thickness              | $h$       | 1.27 mm|
| Metallization (Cu)          | Conductor thickness    | $t$       | 35 $\mu$m|
|                             | Conductivity           | $\sigma$  | 58 MS/m|
| Feedback line TL1           | Partial length         | $L_1$     | 40 mm |
|                             | Diameter               | $D_1$     | 5.39 mm|
| Feedback line TL2           | Partial length         | $L_2$     | 6 mm  |
|                             | Diameter               | $D_2$     | 8.85 mm|
| Coupled CL                  | Width                  | $w$       | 1.09 mm|
|                             | Interspace             | $s$       | 2.7 mm|
| Access lines                | Physical length        | $d_4$     | 6 mm  |
loss lower than 2.5 dB is explained by the wide frequency range mapping in Fig. 6(a). It can be underlined that the insertion loss presents a similar behavior as the GD shown in Figs. 4. The insertion loss in the first NGD bandwidth is insensitive to the variation of \( d_1 \).

2) INFLUENCE OF TL\(_2\) DELAY

Similar to the previous case of study, parametric analyses with respect to \( d_2 \) varied from 55 mm to 65 mm were realized to investigate the influence of TL\(_2\) delay \( \tau_2 \).

Fig. 7(a) displays the maps of GD from 2 GHz to 3.4 GHz. This map reveals that the first NGD BW with center frequency \( f_2 \) is sensitive to delay \( \tau_2 \).

As mapped in Fig. 7(b) and plotted in Fig. 8(a), the NGD center frequencies \( f_1 \approx 3 \) GHz and \( f_2 \) varies between 2 GHz and 2.4 GHz. As seen in Fig. 8(b), the associated GDs are approximately equal to \( \tau_1 \approx -1.25 \) ns and \( \tau_2 \approx -1.4 \) ns. Figs. 9 display the maps of transmission coefficient versus frequency and \( \tau_2 \). Once again, the map behaviors are similar to the GD and shows that only the first NGD BW is sensitive to \( \tau_2 \).

C. INVESTIGATION ON SIMULATED AND EXPERIMENTAL RESULTS

To complete the validation, practical analysis of 0IO NGD prototype was investigated via S-parameter measurement.
from 2 GHz to 3.4 GHz. The experimental setup configuration with vector network analyzer (VNA) is shown in Fig. 10.

The validation study was performed via comparisons of the plots of GDs, transmission and reflection coefficients from measurements (“meas.”), circuit schematic (“schem.”) and EM momentum (“mom.”) simulations from ADS®.

Two different “01O” interconnect prototypes from FR4- and Rogers-substrate based were tested to illustrate the NGD effect possibility.

1) DISCUSSION ON NGD TEST RESULTS OF FR4-BASED 01O PROTOTYPE

Figs. 11 present the comparison results in the wide frequency range which confirm the dual-band bandpass NGD behavior of FR4 substrate based 01O prototype. The simulated and measured GD and transmission coefficients of Figs. 11(a) and 11(b), respectively are in very good agreement. However, as shown in Fig. 11(c), the simulated and measured reflection coefficients are showing considerable differences of behaviors. These main differences are first of all linked to the low values of reflection coefficient under than –10 dB where the differences seem to be relatively significant but remain negligible in linear valuer.

### TABLE 3. Simulated and measured NGD specifications of the FR4 substrate based prototype shown in Fig. 3(a) and Fig. 3(c).

| Approach | $f_1$ (GHz) | $f_2$ (ns) | $BW$ (MHz) | $S_{11}(f_1)$ (dB) | $S_{21}(f_1)$ (dB) | $f_2$ (GHz) | $f_2$ (ns) | $BW$ (MHz) | $S_{11}(f_2)$ (dB) | $S_{21}(f_2)$ (dB) |
|----------|-------------|-------------|-------------|-------------------|-------------------|-------------|-------------|-------------|-------------------|-------------------|
| ADS®     | 2.194       | -1.44       | 31          | -1.9              | -17               | 3           | -1          | 45          | -2.3              | -13               |
| Mom.     | 2.219       | -2.13       | 26          | -2.6              | -13               | 2.988       | -1          | 43          | -2.3              | -12               |
| Meas.    | 2.209       | -1.43       | 35          | -2.4              | -14               | 2.941       | -0.76       | 53          | -2.5              | -13               |

FIGURE 9. $S_{12} = S_{21}$ parametric analysis with respect to the TL2 physical length $d_2$: (a) wideband, and around (b) the 1st and (c) 2nd NGD bandwidth.

FIGURE 10. S-parameter measurement experimental setup of the 01O prototype.

FIGURE 11. Comparisons of measured, ADS schematic, and momentum simulations of: (a) GD, (b) $S_{12} = S_{21}$ and (c) $S_{11} = S_{22}$ from the 01O prototype shown in Figs. 3(a) and 3(c).

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TABLE 4. Simulated and measured NGD specifications of the Rogers substrate based prototype shown in Fig. 3(b) and Fig. 3(d).

| Approach | $f_1$ (GHz) | $\tau(f_1)$ (ns) | $BW_1$ (MHz) | $S_{11}(f_1)$ (dB) | $S_{11}(f_2)$ (dB) | $f_2$ (GHz) | $\tau(f_2)$ (ns) | $BW_2$ (MHz) | $S_{11}(f_2)$ (dB) | $S_{11}(f_2)$ (dB) |
|----------|-------------|------------------|-------------|-------------------|-------------------|-------------|------------------|-------------|-------------------|-------------------|
| ADS®     | 2.716       | -3.34            | 20          | -2.33             | -12.82            | 3.58        | -3.9             | 30          | -2.68             | -12.89            |
| Meas.    | 2.725       | -3.65            | 21          | -3.17             | -10.53            | 3.568       | -1.86            | 34          | -2.81             | -14.93            |

TABLE 5. Performance comparison between the 0IO and other NGD circuits.

| References | $f_1$ (GHz) | $f_2$ (GHz) | $BW_1$ (MHz) | $BW_2$ (MHz) | $S_{11}$ (dB) | $S_{11}$ (dB) | Use of lossy resistor |
|------------|-------------|-------------|-------------|-------------|--------------|--------------|----------------------|
| [24]       | 2.1         | 3.5         | 180         | 180         | -3           | -3.1         | -34                  | -35         | -17               | -17               | Yes                  |
| [25]       | 3.5         | 5.15        | 62          | 59          | -4.5         | -4.2         | -27                  | -28         | N/A               | N/A               | Yes                  |
| [26]       | 3.5         | 5.2         | 200         | 400         | -5           | -5           | 13                   | 20          | N/A               | N/A               | Yes                  |
| [27]       | 0.66        | 1.39        | 446         | 198         | -1           | -1           | 16.9                 | 16.9        | -18               | -17               | Yes                  |
| This work with FR4-substrate | 2.2        | 2.9         | 35          | 43          | -1.5         | -0.9         | 2.2                  | 2.2         | -15               | -13               | No                   |
| This work with Rogers substrate | 2.72       | 3.56        | 21          | 34          | -3.65        | -1.86        | 3.1                  | -2.81       | -10.53            | -2.81             | No                   |

As indicated in comparative Table 3, the NGD center frequencies are localized around 2.2 GHz and 3 GHz. To highlight the NGD behaviors, the zoom in plot around the NGD narrow BWs are depicted in Figs. 12. It can be seen that the FR4 substrate based 0IO prototype presents an NGD BWs of about 35 MHz and 53 MHz. The slight shifts and differences between the NGD central frequency and values shift between the simulated and measurement results is due to the tolerance of the considered dielectric substrate effective permittivity. Moreover, similar to all microwave circuits, the 0IO NGD specifications must include the insertion and reflection losses. As seen in Fig. 11(b), the NGD 0IO prototype presents maximal attenuation of only 2.5 dB around the NGD center frequency. The S-parameters plotted in Fig. 11(c) confirm that the reflection losses are better than 12 dB in the NGD BWs.

2) DISCUSSION ON NGD TEST RESULTS OF ROGERS-BASED 0IO PROTOTYPE

A good correlation between simulated and measured results is realized with Rogers substrate based “0IO” prototype introduced in Fig. 3(b) and Fig. 3(d). Once again, it can be seen that the circuit generates the dual-band NGD effect with center frequencies of about $f_1 = 2.71$ GHz and $f_2 = 3.58$ GHz as shown in Figs. 13(a).

Despite the slight differences indicated in Table 4, the transmission and reflection coefficients as reported
in Fig. 13(b) and in Fig. 13(c), respectively are also in very good agreement notably in the NGD bandwidth. The same as in the results of previous case based on Figs. 11, the slight differences between simulated and measured results plotted in Figs. 13(c) are mainly caused by the imperfections of simulated and NGD circuit prototype fabrications.

To point out the innovative advantages of the 0IO structures, comparative study on NGD performance is addressed in the next subsection.

D. DISCUSSION ON NGD PERFORMANCES

The NGD passive performances of tested 0IO structures are, particularly interesting, compared to the existing dual-band NGD circuits proposed in [24]–[27]. Table 5 summarizes the comparison between the NGD, insertion and reflection loss performances including the dual-band NGD bandwidth. It can be pointed out that the 0IO-shape circuit allows to achieve a very good attenuation showing low-loss aspect under very good reflection coefficient. Moreover, compared to the NGD circuits designed in [24]–[27], the 0IO-shape circuit is merely built with a fully distributed circuit without using any lossy lumped element as resistor.

IV. CONCLUSION

An innovative NGD analysis of particular 0IO-shape PCB traces is investigated. The proposed topology consists originally of 3-CL with lateral side interconnect in feedback with lossy and delayed different TLs. The S-matrix model of this particular NGD topology is established. Then, the NGD analysis is presented by exploiting the GD through the transmission coefficient expression. The NGD theoretical approach was validated by designing fully distributed microstrip structures without lumped elements even a resistance. Therefore, an NGD bandpass behavior is realized. A very good agreement between the simulations and measured S-parameters and GDs are realized.

The investigated “0IO” topology is promising for future RF and microwave system applications with the possibility to operate with low attenuation and reflection losses. This solution is utmost important for the electronic systems designers, ensuring the system functioning jointly with PCB layout guidelines.

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F. Wan et al.: 0IO-Shape PCB Trace NGD Analysis

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