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Transversal Directional Filters for Channel Combining

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Abstract—A new concept for the design of power combiners based on matched directional filters is presented. The directional filters consist of individual balanced sections composed of hybrids and single resonators. Each of the sections corresponds to a pole of an all-pass function composed of the sum of S_{11} and S_{12} of the desired filter transfer function. A simple synthesis method is presented. The filter combiner has the advantage of ease of tunability because each pole is associated with a single resonator. Furthermore, no cross couplings are required to realize finite frequency transmission zeros. Experimental results for a prototype device are presented.

Index Terms— directional filters, combiner, power divider, general Chebyshev filter.

I. INTRODUCTION

Recent requirement to share sites for mobile communication base stations require the use of diplexers or multiplexers so that two or more bands may be transmitted on a single antenna simultaneously allowing service providers in the same vicinity to share antennae. This paper demonstrates a novel approach for designing combiners based on Directional Filters (DF's) which can be used in Long Term Evolution (LTE) base stations.

The concept of directional filtering has been presented in the literature over the past decades. Common designs use striplines [1]-[3] and waveguide [4] technologies. A DF is a matched four-port device shown in Fig.1a with an input at port 1. Power emerges at port 3 with the frequency response of a band-pass filter and the remaining power emerges at port 2 with the complementary response of a band-reject filter. Port 4 is isolated. In the design of DF, we use standard filter characteristics as illustrated in Fig 1b where S12 is the return loss of a general Chebyshev filter and S13 is its insertion loss with two transmission zeros.

Because of the transmission characteristics of DF's, they may be used for signal combining or multiplexing. As opposed to conventional techniques of designing combiners, this technique does not involve the design of channel filters or junctions. Instead, a single bandpass filter characteristic is used such that the insertion and return loss of this bandpass filter provides the

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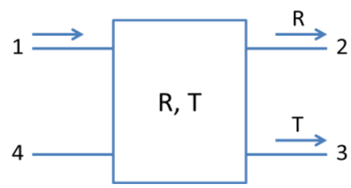
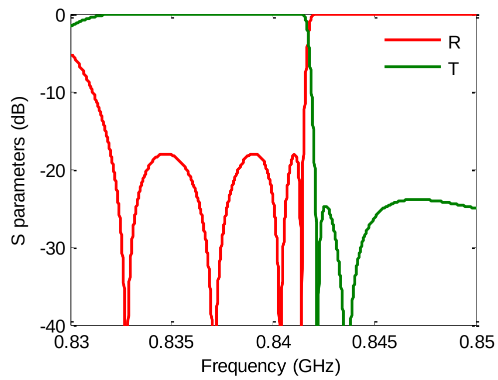
forward transmission characteristics of each band. Because the two passbands are formed by the insertion and return loss of a single filter characteristic, the two channels of the combiner have no interaction even when the two bands are very close to each other.



(a)



(b)

Fig. 1. A simplified diagram for a DF (a) and its response (b) that can be used for power combining.

The principles of operation of a DF used in this paper are an extension on the work presented in [6]. Section II presents the design theory for DFs. A synthesis method for cascaded DF's is presented in [7] and reviewed here. The design is based on a pole placement method where each single section of a DF is singly tuned to provide a pole of a bandpass filter. This may be realized by inserting two resonator networks between a pair of 90º hybrids. Cascading these matched four-port sections allows a multi-pole response equivalent to the characteristics an N^{th} degree filter. As in a transversal array, each section of the network corresponds to a pole of the filter's admittance parameters [8]; each cascaded section in a DF realizes a pole of its S parameters.

As the design theory provides a DF in its lowpass prototype, Section III provides its the circuit realization. Bandpass characteristics are achieved by standard transformations. Various approximations are used to derive a simplified and realizable equivalent circuit. The final circuit contains resonators and some internal couplings that can be implemented by standard filter technologies.

Section IV presents two design examples. First, the design concepts are validated with a design of a cellular combiner with specifications used in uplink 800 MHz LTE bands. The design is based on the characteristics of a 4^{th} order general Chebyshev filter and the required selectivity is achieved by two transmission zeros placed close to the passband. The cellular combiner was fabricated using coaxial resonators. The proposed solution achieves good isolation between the input ports, return loss and minimum in-band insertion loss. Dual or multi-bands responses can also be implemented with DF's. With the responses derived using the optimization presented in [9], the same synthesis method for DF's can be applied. The second example is given with two passband at 1.635-1.645GHz and 1.735-1.745GHz. Each passband is realized by a 2^{nd} response with one transmission zero in the upband. 30dB isolation is achieved within each band.

II. Design Theory

A. Single section DF's

The original design theory [6] was based on the DF configuration shown in Fig. 2. The DF consists of two identical filters and a pair of 3 dB hybrids. The scattering matrix of the filter network is given in (1). For this DF, the power incident at port 1 emerges at port 2 with the return loss of the filter network and at port 3 with the insertion loss. Port 1 is matched and port 4 is isolated. The response of the DF with R the power transferred to port 2 and T the power transferred to port 3 is thus given as in (2).

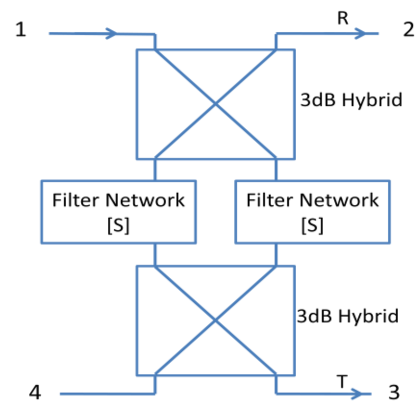


Fig. 2. Detailed diagram of DF with two identical filter networks inserted between a pair of 90° hybrids.

$$[S]=\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \tag{1}$$

$$\begin{aligned} R &= jS_{11} \\ T &= jS_{21} \end{aligned} \tag{2}$$

In our design, the filter network is a single resonator and higher order characteristics are achieved by cascading these single sections.

B. Cascaded directional filters

The single section of DF can be cascaded as shown in Fig. 3. The response of a single section is shown in (3). For a cascade of two sections, (3) is the input to the second section. P_1 input at port 1 provides P_1R_1 at port 2 and P_1T_1 at port 3. Because of the symmetry of the network, Q_1 input at port 4 provides Q_1R_1 at port 3 and Q_1T_1 at port 2. So the response of two sections is the combination of those outputs as in (4).

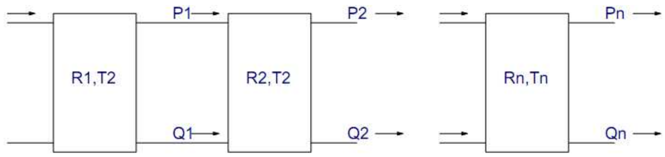


Fig.3. Cascading of n single section DF's.

$$\begin{aligned} P_1 &= R_1 \\ Q_1 &= T_1 \end{aligned} \tag{3}$$

$$\begin{aligned} P_2 &= R_1R_2 + T_1T_2 = P_1R_2 + Q_1T_2 \\ Q_2 &= R_1T_2 + T_1R_2 = P_1T_2 + Q_1R_2 \end{aligned} \tag{4}$$

With (3) being the initial values, a recurrence formula in (5) can be derived for the cascading of n sections. From the expression for P_n+Q_n, we can derive the equation in (6) which represents an all-pass response.

$$\begin{aligned} P_n &= P_{n-1}R_n + Q_{n-1}T_n \\ Q_n &= P_{n-1}T_n + Q_{n-1}R_n \end{aligned} \tag{5}$$

$$P_n + Q_n = \prod_{i=1}^{n}\left(R_i + T_i\right) \tag{6}$$

C. Decomposition of filter characteristics

The S parameters of a lossless filter network may be expressed as rational polynomials as in (7). E is the common denominator of S parameters whose degree is the degree of filter network. F11 is the numerator of the return loss which contains all the reflection zeros. P is the numerator of insertion loss and contains all the transmission zeros. The sum of these parameters is given in (8) which is also an all-pass response and n is the filter order, z_i represents a zero and p_i represents a pole.

$$S_{11} = \frac{F_{11}}{E}, \quad S_{21} = \frac{P}{E} \tag{7}$$

$$S_{11} + S_{21} = \frac{F_{11} + P}{E} = \frac{\prod_{i=1}^{n}\left(s - z_i\right)}{\prod_{i=1}^{n}\left(s - p_i\right)} \tag{8}$$

From power conservation, we can derive (9), which states that p_i is the same as either z_i or its conjugate.

$$E \cdot E^* = \left(F_{11} + P\right)\cdot\left(F_{11} + P\right)^* \tag{9}$$

By equating the two all-pass responses in (5) and (8), for a given filter characteristics, P_n+Q_n are known (10). By the decomposition of poles, R_i and T_i of each section can be derived according to (11) which states that R_i and T_i have the same pole

p_i and the zeros of the numerator of R_i+T_i are z_i. As a result, each pole of a filter characteristic may be realized by a single DF section and a higher order response may be formed by cascading DF's. The next problem is to find the network realisation with reflection coefficient R_i and transmission coefficient T_i.

$$P_n + Q_n = \frac{F_{11} + P}{E} \tag{10}$$

$$R_i + T_i = \frac{s - z_i}{s - p_i} \tag{11}$$

III. Circuit Realization of a DF Network

A. Realisation of Filter network

A simple resonator network as shown in Fig.4 is used to provide the characteristics of R_i and T_i. The network consists of a capacitance C and a frequency invariant reactance B which is included for complex poles. The transfer matrix of this network is given in (12a) and its S parameters can be derived as in (12b).

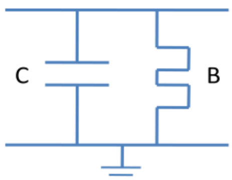


Fig. 4. Filter network to provide the required pole of a DF.

$$T = \begin{bmatrix} 1 & 0 \\ sC + jB & 1 \end{bmatrix} \tag{12a}$$

$$S_{11} = -\frac{s + jB/C}{s + jB/C + 2/C}$$
$$S_{21} = \frac{2/C}{s + jB/C + 2/C} \tag{12b}$$

According to the S parameters in (12b), we have a all pass response in (13a) whose magnitude should be equivalent to that in (11), as a result, a pole in (11) is shown in (13b). Then we obtain the values for the capacitance and frequency invariant reactance according to the values of poles as in (13). It should be noted that in the case when z_i is of the complex conjugate of p_i, a 180º phase shifter should be introduced after the i^{th} section.

$$S_{11} + S_{21} = -\frac{s + jB/C - 2/C}{s + jB/C + 2/C} \tag{13a}$$

$$p_i = jB/C + 2/C \tag{13b}$$

$$C_i = -\frac{2}{\mathrm{Re}(p_i)}$$
$$B_i = \frac{2\,\mathrm{Im}(p_i)}{\mathrm{Re}(p_i)} \tag{13c}$$

B. Node diagram

The complete admittances of a DF section are shown in Fig. 5 which is derived from Fig.2. Then number on the hybrid refers to the admittance of each 90º TL branch. The Y block refers to the filter network which is represented by its admittance matrix.

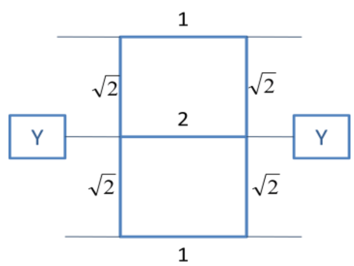
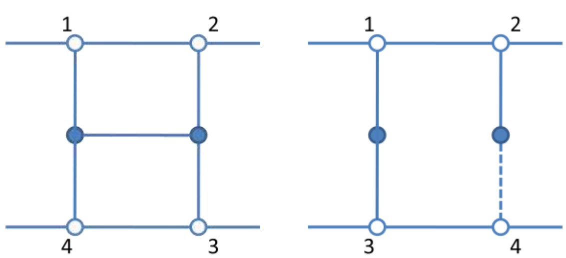


Fig. 5. Circuit diagram showing the admittances of a single section DF.

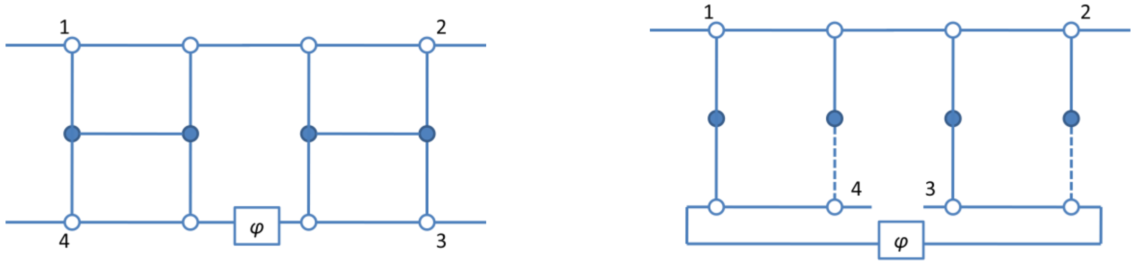
By replacing the 90º TL with invertors of different characteristic admittance, the network can be represented by a 4-port coupling matrix as illustrated by node diagram in Fig. 6a in which the empty dots represent non-resonating nodes. According to the node diagram in Fig.6, a coupling matrix that is usually used in the filter design can be found for a DF. The response of cascaded DFs can then be found by analysis of the response of a coupling matrix using the standard method given in [10]. In this way the response of DF networks are found by matrix analysis instead of circuit simulators. An alternative configuration for single section DF may be found as shown in Fig. 6b, while the dashed line represent negative couplings. This network is derived by scaling nodes in one of the hybrids and the two couplings in the center of the network cancel with each other. The response of this network remains the same except the port 3 and port 4 are exchanged.



(a) (b)

Fig. 6. Node diagram for single section DF (a) and its alternative (b). The empty node represents non-resonating nodes.

An N^{th} degree DF is derived by cascading N sections. Unlike a conventional filter, sections can be cascaded in any order. Fig. 7 shows the connection of two DF sections for the two configurations in Fig.6. The φ block represents the 180º phase shifter which should be added after the section when the corresponding z_i is the same as the conjugate of p_i.



(a) (b)

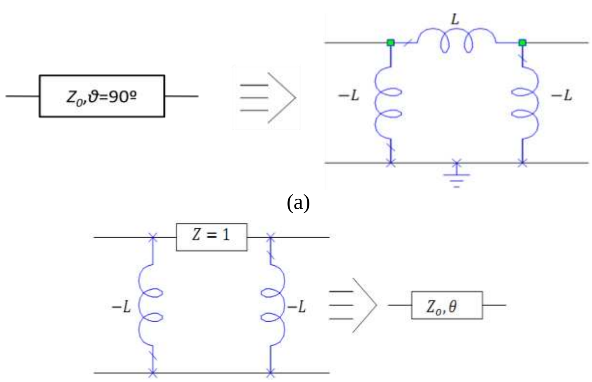
Fig. 7. Cascading of the single section DF (a) and its alternative(b).

C. Circuit simplification

The final circuit for a section may be derived after some scaling and simplifications. Standard lowpass to band-pass transformations are applied. The branches for the 90º hybrid are replaced by an equivalent π network of inductances as shown in Fig. 8(a). Elements of the inductances of the π network are then merged with the resonators. The 90º TL with unit Z_0 at the input and output is replaced by an equivalent TL as shown in Fig. 8(b).

TABLE I
ELEMENT VALUES FOR THE DF SECTIONS

Roots of F_{11}	Roots of E	Roots of F_{11}+P	C	B
-0.8366i	-0.7005-1.2885i	0.7005-1.2885i	0.9636	3.6787
0.9805i	-0.9531+0.2651i	-0.9531+0.2651i	0.7021	3.6787
0.7632i	-0.3062+0.9603i	0.3062+0.9603i	2.1762	-6.2716
0.0751i	-0.0537+1.0451i	-0.0537+1.0451i	2.1762	-38.9048



(a)

(b)

Fig. 8. Equivalent circuit for (a) a 90º TL by a π network of inductances and (b) a 90º TL with shunt inductances at the input and output by a TL with different characteristic impedance and phase length.

The final circuit for the single section DF is given in Fig.9. *L'* and *C'* can be realized by conventional resonators such as coaxial and dielectric resonators. The inductances L_v represents the input coupling to the resonators and L_m represents inter coupling between the resonators.

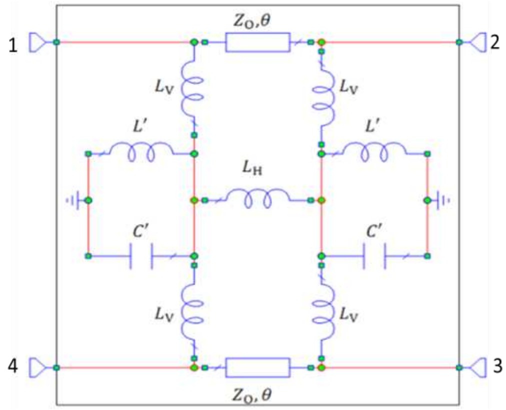


Fig. 9. Simplified circuit for a single section DF.

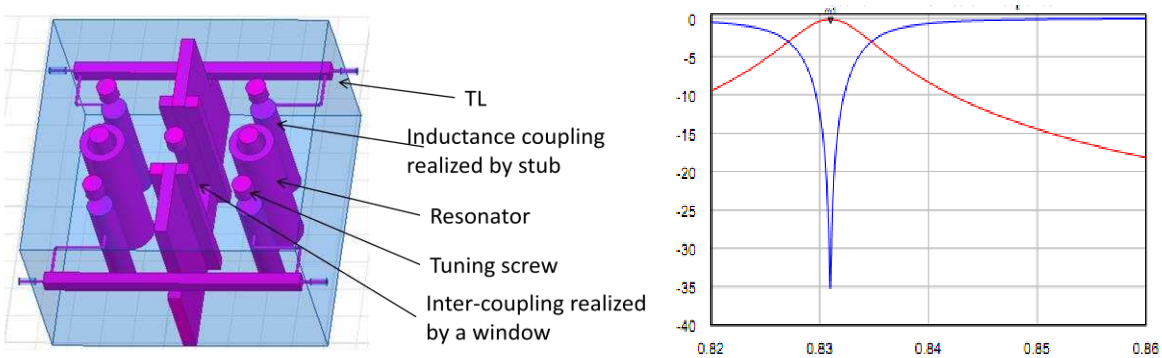
IV. Design Examples

A. A 4th order DF

With cascaded sections, high performance combiners may be designed. A combiner with two passbands of 832-841.5 MHz and 842.5-852MHz is used as an example to illustrate the design theory. The required passband return loss is >18dB and the insertion loss is <1dB.

A 4th order general Chebyshev filter is used. Because the two bands are close to each other, the filter characteristic should have a steep transition and is realized by two transmission zeros close to the passband. The transmission zeros are at 1.15j and 1.45j in the lowpass domain. After synthesizing the general Chebyshev response using the method given in [11], we have the polynomials F_{11} and E of the lowpass prototype and R_i and T_i of each section are found according to (11). Values of the capacitors and frequency invariant reactances are calculated according to (13) and are listed in Table I. A lowpass to bandpass transformation is applied to each resonator and the network is combined with the equivalent circuit of the 90º hybrid. Then the four DF sections are cascaded. The circuit model is simulated in ADS and the result is shown as the solid lines in Fig. 12 and it is the same as the synthesized Chebyshev response.

An EM model for the circuit in Fig. 9 using coaxial resonators is shown in Fig.10(a). The main branch of the hybrid is realized by a 50Ω line. The input couplings are realized by non-resonating stubs and the inter resonator coupling is realized by a window. As a lossless design, we used PEC for conductors. For this 4th order example, an EM model was built for each DF section in HFSS. As each DF section has a different resonant frequency and couplings, the EM model was tuned until the response as illustrated in Fig. 10(b) for the first section is consistent with that of the circuit model. The four-port S parameters of each DF section were then put in ADS and connected by TLs. The result of this combined EM and circuit simulation is given in Fig.11 as the dashed line responses.



(a) (b)

Fig. 10. EM model of a single section DF in HFSS (a) and its typical response (b).

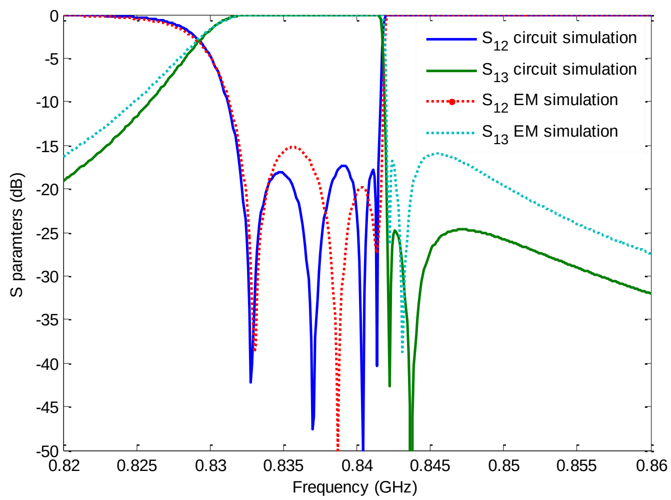


Fig. 11. Simulation result for the combiner (Solid line for the circuit simulation and dashed line for the combined EM/circuit simulation)

The EM model of the whole structure is shown in Fig.12. Because each DF section controls one pole of the filter characteristic, they may be tuned independently. No cross couplings are required when realizing transmission zeros and thus through tuning, different kinds of responses may be achieved by the same structure.

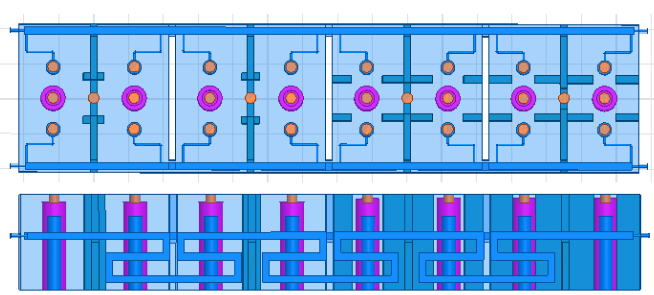


Fig. 12. EM model for the combiner (top view and side view).

Different Qs may be assigned to each section (800, 650, 4,000, 20,000). obtaining similar selectivity to a lossless filter and requiring a minimum number of high Q resonators. The insertion loss in the passband is compared between a lossless circuit and a lossy circuit with non-uniform Q in Fig.13.

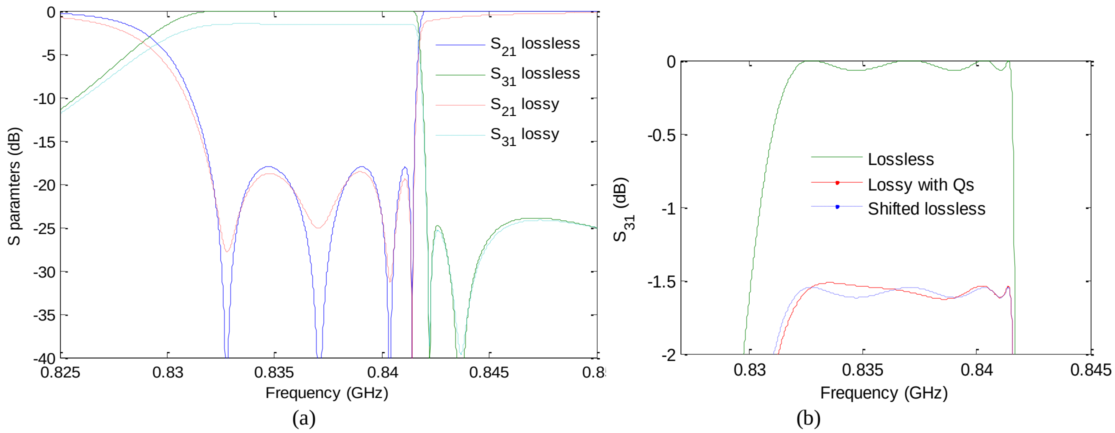


Fig. 13. S parameters of the lossy combiner compared with the lossless onse.

B. DF with Dual Response

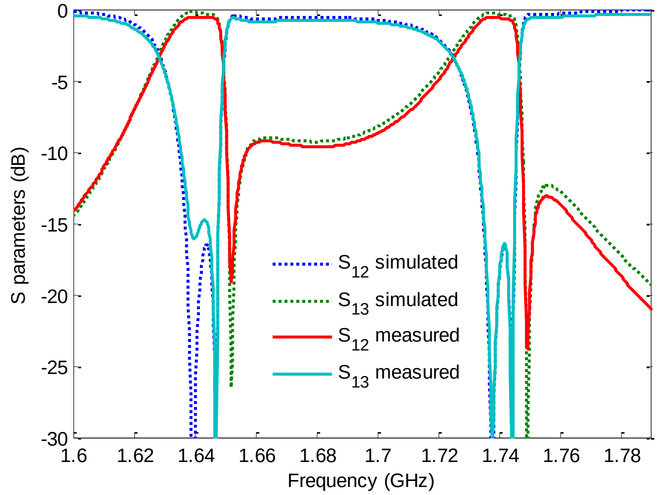
Dual-band or multi-band characteristics may be achieved with the same physical configuration by choosing the appropriate transfer function. With the dual or multi-band F_{11}, E and P polynomials derived by optimization method given in [9], the synthesis of DF sections remains the same.

The example here is a 4^{th} order dual-band combiner with two passband at 1.638-1.648GHz and 1.735-1.745GHz. For each band, a transmission zero is included in the upband to achieve a rejection of 10dB. The return loss level is 15dB. The element values of the circuit model of this DF are listed in Table II and the simulation results are shown in Fig. 16 as the dashed lines. From the S parameters, we can see that each passband corresponds to a 2^{nd} response.

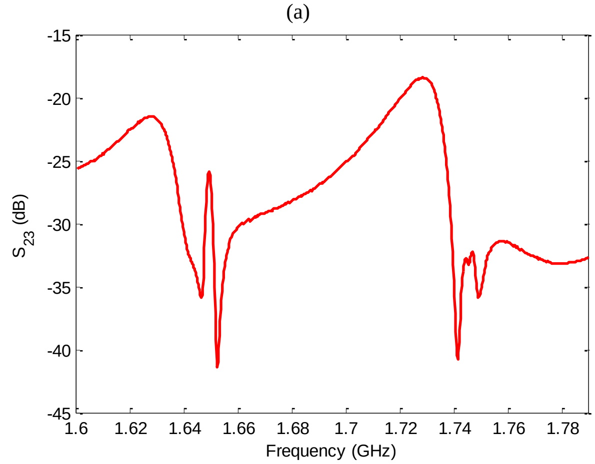
TABLE II
ELEMENT VALUES FOR THE DUAL-BAND DF SECTIONS

Roots of F_{11}	Roots of E	Roots of F_{11}+P	C	B
0.9848i	-0.0357+1.0277i	0.0357+1.0277i	56.0884	-57.6417
0.8641i	-0.2109+0.8373i	-0.2109+0.8373i	9.4847	-7.9411
-0.9725i	-0.2167-1.0415i	0.2167-1.0415i	9.2281	9.6112
-0.8308i	-0.0415-0.7778i	-0.0415-0.7778i	48.1663	37.4649

Measured transmissions are shown in Fig. 14(a) as the solid line. 30dB isolation is achieved within each band as shown in Fig. 14(b). For the two resonators in each band, Qus are 4400 and 3000 respectively. The 1dB channel bandwidths are 10MHz. A photo of the device is given in Fig. 15.



(a)



(b)

Fig. 14. Simulated and measured (a) transmissions and (b) isolation of the dual-band combiner.

Fig. 15. Photograph of the device made.

V. Conclusion

A new concept for designing channel combiners based on directional filters has been presented. Multi-section DF's are synthesized as a cascade of first-order all-pass networks. With cascaded DF's, high performance combiners may be designed. Because the two pass-bands are formed by the insertion and return loss of a single filter characteristic, the two channels of the combiner have little interaction even when the two bands are very close to each other. The combiner is simple to tune as each section is independent of the others and the structure requires no cross couplings. As in the case of transversal filters, non-uniform Q resonators may be used to great effect.

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