

## RESEARCH PAPER

# A linearity improved quad-band amplifier based on E-CRLH transmission line

RASOOL KESHAVARZ, ABBAS MOHAMMADI AND ABDOLALI ABDIPOUR

*This paper outlines a new class of quad-band amplifier (QBA) realized using extended composite right- and left-handed coupled lines. The design procedure to reach the QB CLs with predefined frequency operations and characteristic impedance of their ports is clearly presented. Designed and fabricated QBA operation frequencies are  $f_{b_1} = 2.9$  GHz and  $f_{b_2} = 4.3$  GHz (output port B)  $f_{f_1} = 3.8$  GHz and  $f_{f_2} = 5$  GHz (output port C) and the proposed structure exhibits a matching ( $-10$  dB) bandwidth of over 300 MHz in each operation frequency bands. In addition, the insertion loss is smaller than 1 dB. The length and the width of the proposed QBA are approximately 7.7 and 2.4 cm, respectively. To further demonstrate the idea, a QBA is successfully designed and fabricated. The simulated and measured results of the proposed QBA are presented to verify the proposed idea.*

**Keywords:** Coupled lines, E-CRLH, Quad-band amplifier, Metamaterial

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## I. INTRODUCTION

Radiofrequency (RF) amplifiers are one of the most critical components in almost all types of wireless applications. Each application has different requirements depending on frequency and type of input signal, etc. For each of specific purpose, they are expected to provide suitable output power and gain, while achieving acceptable efficiency under application-specific linearity considerations [1].

On the other hand, the request on wireless communication systems is forcing multi-band transceivers to deal with several services in the only terminal. Owing to the current demand on multi-frequency systems over several wireless communication bands, systems simultaneously working at some of the following bands (GSM, TETRA, DCS, PCS, IMT-2000, WLAN, or UWB) are currently being developed and are widely used in new portable wireless handsets [1].

Then, due to these drawbacks, the availability of multi-frequency high linearity and efficiency amplifiers would be welcome. Some attempts have been undertaken in order to achieve that goal. In this way, the use of switches [2], varying lumped components [3], or diplexers [4] have been proposed as potential techniques for achieving dual-frequency high-efficiency amplifiers.

During last years, the appearance of Composite Right/Left-Handed Transmission Lines (CRLH TLs) has opened new possibilities in the design of multi-band amplifiers [5].

Essentially CRLH structures, which are constituted by a left-handed (LH) series capacitance, shunt inductance in addition to the conventional right-handed (RH) series inductance, and shunt capacitance, along with their various microwave applications [6].

In [7], a distributed amplifier based on the active coupling of two artificial composed of RH/LH TLs has been presented. This reported work shows that it is possible to improve the isolation between bands by means of a non-uniform distribution of the devices' transconductance.

A novel extended composite RH/LH (E-CRLH) TL metamaterial structure, constituted by the combination of the conventional CRLH (C-CRLH) and the recently introduced dual CRLH (D-CRLH) prototypes, is proposed. Microwave devices by E-CRLH structure can be quad-band (QB)[8].

In this paper, a new type of quad-band-coupled lines (QB CLs) is presented. These with closely placing two E-CRLH lines in parallel are obtained that can operate with nearly 0 dB power coupling level. The proposed QB CLs operates at four frequencies; two frequencies in the backward direction and the other two frequencies in the forward direction.

Then, we use QB CLs for designing quad-band amplifier (QBA) with high linearity. The proposed structure consists of two QB CLs and two amplifiers. In the proposed QBA, four frequency bands come into the one of the proposed CLs as multiplexer and two dual-band output signals are amplified with two amplifiers. Finally, two amplified signals are demultiplexing and coming out by means of the other CLs. In this paper, the design process is clearly discussed and the analytical equations are derived.

The organization of this paper is as the following. At first, the principle of the E-CRLH TL and their property are presented. Then, the proposed QB E-CRLH CLs are introduced and design procedures are presented. Furthermore, the

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proposed linearity improved QBA is presented and is validated by simulation and measurement results.

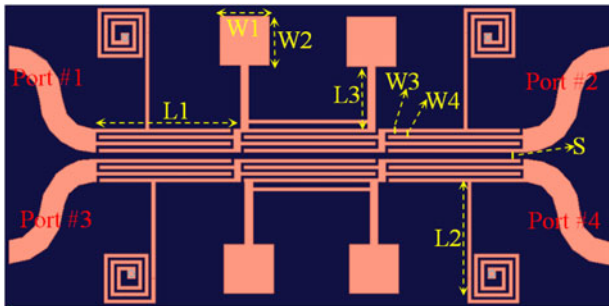
## II. QB E-CRLH-COUPLED LINES THEORY

Figure 1 shows the proposed E-CRLH CLs and its equivalent circuit model. As shown in Fig. 1(b),  $C_m$  represents the mutual capacitance between E-CRLH TLs. In addition, other lumped elements that presented in Fig. 1(b) are  $\Pi$  equivalent circuit model elements of E-CRLH TLs [8]. It should be stated that all of the parameters in the circuit model are per unit length quantities.

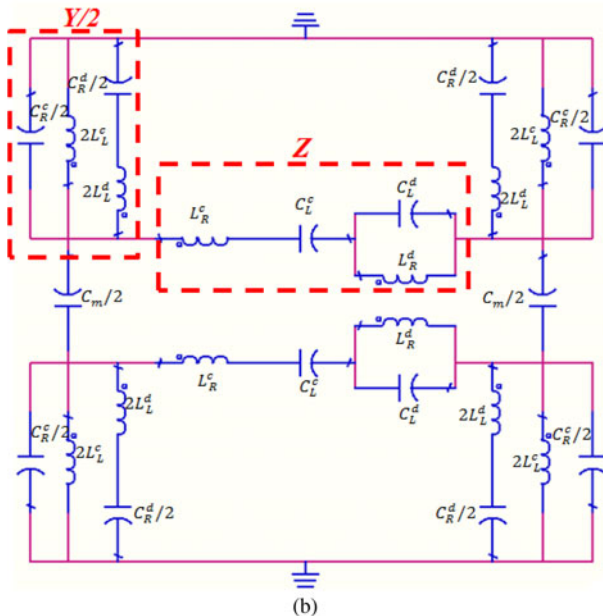
From Fig. 1, series impedance ( $Z$ ) and parallel admittance ( $Y$ ) of an E-CRLH unit cell can be derived as following [8]:

$$Z = j\omega L_R^c \left( 1 - \frac{\omega_{cs}^2}{\omega^2} \right) - \frac{j}{\omega C_L^d \left( 1 - \left( \frac{\omega_{dp}^2}{\omega^2} \right) \right)}, \quad (1)$$

$$Y = j\omega C_R^c \left( 1 - \frac{\omega_{cp}^2}{\omega^2} \right) - \frac{j}{\omega L_L^d \left( 1 - \left( \frac{\omega_{ds}^2}{\omega^2} \right) \right)}, \quad (2)$$



(a)



(b)

Fig. 1. (a) E-CRLH CLs, (b) equivalent circuit model of E-CRLH CLs.

where

$$\omega_{cs} = \frac{1}{\sqrt{L_R^c C_L^c}}, \quad \omega_{cp} = \frac{1}{\sqrt{L_L^c C_R^c}}, \quad (3)$$

$$\omega_{dp} = \frac{1}{\sqrt{L_R^d C_L^d}}, \quad \omega_{ds} = \frac{1}{\sqrt{L_L^d C_R^d}}, \quad (4)$$

$$\omega_{dscp} = \frac{1}{\sqrt{L_L^d C_R^c}}, \quad \omega_{csdp} = \frac{1}{\sqrt{L_R^c C_L^d}}, \quad (5)$$

$$\omega_{dp} = \omega_{ds}, \quad \omega_{cp} = \omega_{cs} \text{ (balance conditions)}. \quad (6)$$

Therefore, for an E-CRLH TL, the propagation constant is equal to [9]:

$$\cos(\beta l) = 1 + ZY/2, \quad (7)$$

where  $l$  is the physical length of the unit cell.

The Bloch impedance of a periodic transmission line is presented in [8] as:

$$Z_B = \sqrt{\frac{Z}{Y}} \sqrt{1 + \frac{ZY}{4}}. \quad (8)$$

When the dimension of the unit cell is very small  $\beta l \ll 1$  or  $l \ll \frac{\lambda}{2\pi}$ , then  $\cos(\beta l) \cong 1$  and according to equation (8):

$$ZY \cong 0 \rightarrow Z_B = \sqrt{\frac{Z}{Y}} \sqrt{1 + 0} = \sqrt{\frac{Z}{Y}}. \quad (9)$$

In this case, the Bloch impedance is approximately equal to the characteristic impedance  $Z_B \cong Z_c$ .

However, in the coupled-line couplers, input matching condition to terminations of impedance  $Z_c$  is obtained under the condition given by [10]:

$$Z_c = \sqrt{Z_{ce} Z_{co}}, \quad (10)$$

where  $Z_{ce}$  and  $Z_{co}$  are even- and odd-mode characteristic impedances of the CLs, respectively.

In design procedure,  $Z_{ce}$  and  $Z_{co}$  can be obtained from conventional expressions as [11]:

$$Z_{ce} \cong \sqrt{\frac{Z_e}{Y_e}}, \quad Z_{co} \cong \sqrt{\frac{Z_o}{Y_o}}, \quad (11)$$

where  $Z_e$  and  $Y_e$  are the series impedance and parallel admittance of an E-CRLH unit cell in the even mode, respectively, and are found as:

$$Z_e \cong Z, \quad Y_e \cong Y. \quad (12)$$

Also  $Z_o$  and  $Y_o$  are the series impedance and parallel admittance of an E-CRLH unit cell in the odd mode, respectively,

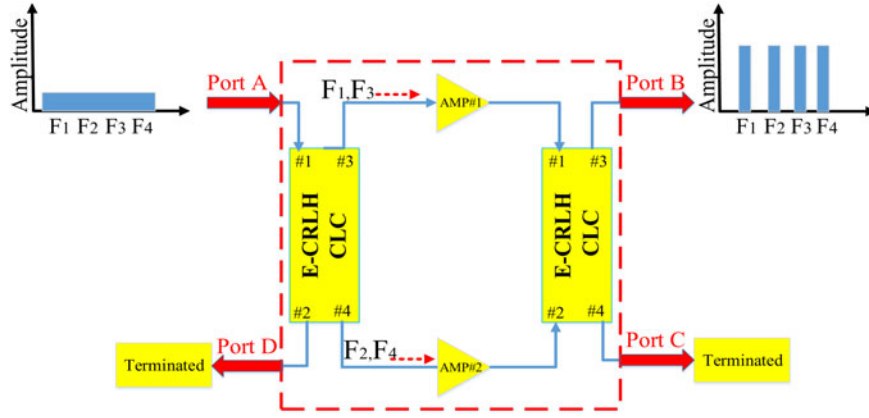


Fig. 2. The principle of dual-band duplexer ( $F_1 = f_{b_1}$ ,  $F_3 = f_{b_2}$ ,  $F_2 = f_{f_1}$ ,  $F_4 = f_{f_2}$ ).

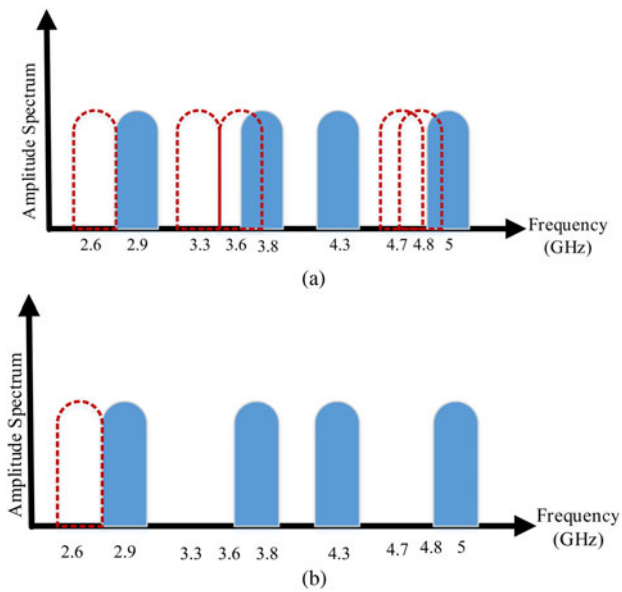


Fig. 3. Amplitude spectrum of the operational frequency bands ( $F_1 = 2.9$  GHz,  $F_2 = 3.8$  GHz,  $F_3 = 4.3$  GHz,  $F_4 = 5$  GHz) and their intermodulation components (dashed); (a) conventional amplifier, (b) proposed QBA.

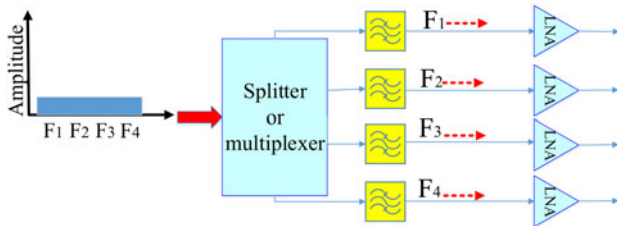


Fig. 4. Conventional QB receiver configuration ( $F_1 = f_{b_1}$ ,  $F_3 = f_{b_2}$ ,  $F_2 = f_{f_1}$ , and  $F_4 = f_{f_2}$ ).

and are found as:

$$Z_o \cong Z, \quad Y_o \cong Y + Y_m, \quad (13)$$

where  $Y_m$  is the influence of  $C_m$  that is added to the parallel

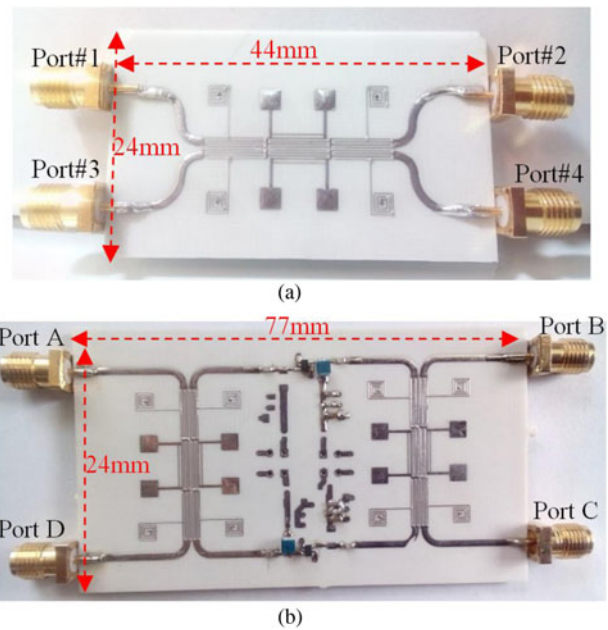


Fig. 5. Fabricated structures on the RO-4003 substrate,  $\epsilon_r = 3.38$ , thickness of 32 mil and  $\tan \delta = 0.003$ ; (a) QB CLs, (b) QBA.

admittance  $Y$  and is equal to:

$$Y_m = 2j\omega C_m \left( 1 - \frac{\omega_{cp}^2}{\omega^2} \right). \quad (14)$$

Furthermore, according to (12) and (13) the characteristic impedance in equation (10) could be rewritten as:

$$Z_c^4 = \frac{Z^2}{Y^2 + YY_m}. \quad (15)$$

Currently, we want to design a QB E-CRLH CLs that has two output ports. The signal is going out from port #4 (forward direction) at frequencies  $F_2$  and  $F_4$  and from port #3 (backward direction) at frequencies  $F_1$  and  $F_3$  (Fig. 1). In addition, port #2 is isolated. Therefore, these CLs exhibit such as QB couplers.

**Table 1.** Lumped element values of Fig. 1.

$C_R^c$	$L_L^c$	$L_L^d$	$C_R^d$	$L_R^c$	$C_L^c$	$C_L^d$	$L_R^d$
5.44 pF	1.07 nH	7.48 nH	0.67 pF	19.29 nH	2.7 pF	2.11 pF	2.38 nH

**Table 2.** Dimension of the proposed dual-band duplexer (Fig. 1).

L1 (mm)	L2 (mm)	L3 (mm)	W1 (mm)	W2 (mm)	W3 (mm)	W4 (mm)	S (mm)
8.5	7.2	4.3	3.1	3.1	0.2	0.2	0.2

In the E-CRLH CLs, similar to the conventional edge-coupled couplers, the backward wave ( $S_{31}$ ) and forward wave ( $S_{41}$ ) coupling coefficients are [11]:

$$S_{31} = \frac{jksin\theta}{\sqrt{1 - k^2\cos\theta + jsin\theta}}, \tag{16}$$

$$S_{41} = -je \frac{-j(\beta_e + \beta_o)l}{2} \sin\left[\frac{((\beta_e - \beta_o)l)}{2}\right], \tag{17}$$

where

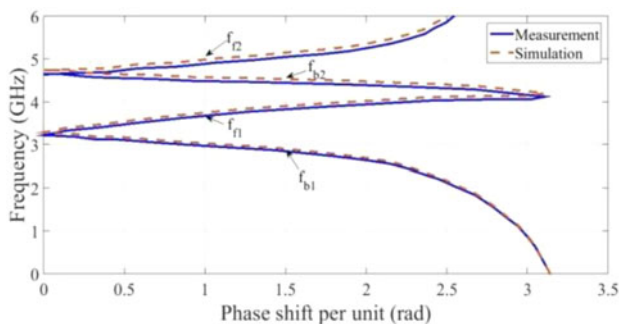
$$k = \frac{Z_{ce} - Z_{co}}{Z_{ce} + Z_{co}}, \theta = \beta l \tag{18}$$

and  $\beta_e$  and  $\beta_o$  are propagation constants of the even and odd modes, respectively, and  $l$  is the length of the structure. In addition, the direct coefficient is equal to:

$$S_{21} = \frac{\sqrt{1 - k^2}}{\sqrt{1 - k^2\cos\theta + jsin\theta}}. \tag{19}$$

Now, we discuss about the forward and backward directions coupling in the proposed QB CLs and the design equations will be derived.

*Forward direction coupling:* In the proposed E-CRLH CLs, we want the coupling coefficient in the forward wave equals to



**Fig. 6.** Dispersion curve of the E-CRLH unit cell (full-wave simulation and measurement results).

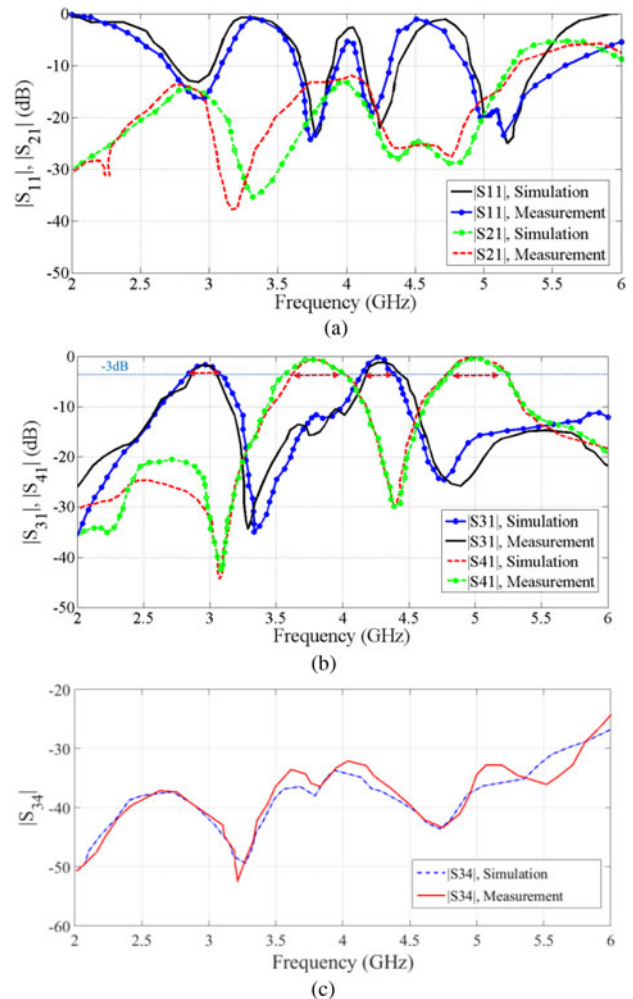
0 dB at frequencies  $F_2$  and  $F_4$ . So, according to (17):

$$l = \frac{\pi}{\beta_e - \beta_o} \rightarrow \theta_e - \theta_o = \pi \rightarrow \cos\theta_e = -\cos\theta_o. \tag{20}$$

Therefore, substituting (7), (12), and (13) into (20) result in:

$$(1 + Z_e Y_e/2) = -(1 + Z_o Y_o/2) \rightarrow \cos\theta_i = -ZY_m/4. \tag{21}$$

Equation (21) states two conditions for  $\omega_i = 2\pi F_i$ ,  $F_2 = f_{f1}$ ,  $F_4 = f_{f2}$ . The  $f_{f1}$  and  $f_{f2}$  are operation frequencies in the forward direction coupling.



**Fig. 7.** (a)  $|S_{11}|$  and  $|S_{21}|$ , (b)  $|S_{31}|$  and  $|S_{41}|$ , (c)  $|S_{34}|$  parameters of the proposed dual-band duplexer.

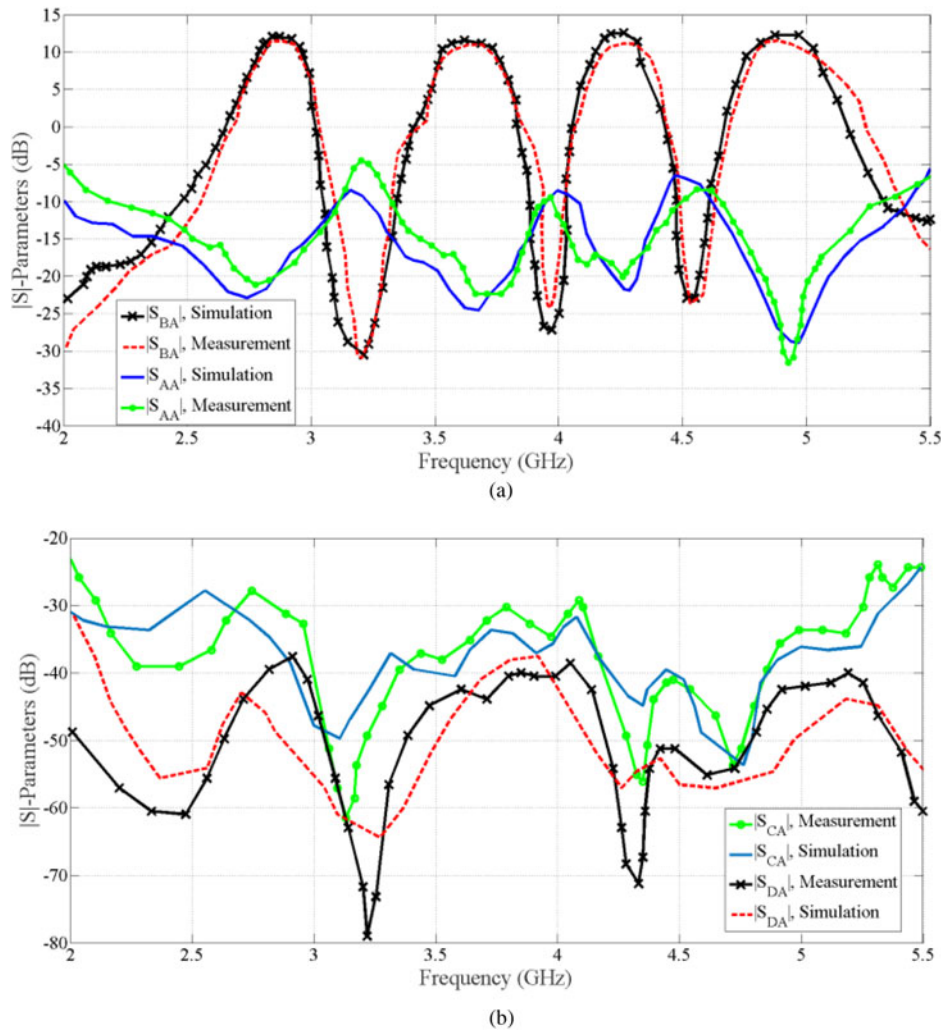


Fig. 8. (a)  $|S_{AA}|$  and  $|S_{BA}|$ , (b)  $|S_{CA}|$  and  $|S_{DA}|$ , parameters of the proposed QBA.

From (7), (15), and (21) the required condition for 0 dB forward coupling coefficient at frequency  $f_i$  is:

$$Y_m^2 = \frac{64(\cos\theta_i)^4}{Z_c^4(1 - (\cos\theta_i)^2)} = -C_i^2. \tag{22}$$

Then, (14) and (22) establish the following polynomial equation:

$$2\omega_i^2 x_1 - 2x_2 = C_i \omega_i, \tag{23}$$

where  $x_1 = C_m$  and  $x_2 = C_m \omega_{cp}^2$  are unknown parameters and  $\omega_i = 2\pi F_i$ ,  $F_i = f_{f1} f_{f2}$ .

**Backward direction coupling:** From (16), it is clear that complete power can be transferred between lines in backward

direction if  $\theta = 90^\circ$ , thus:

$$S_{31} = k = \frac{1 - \sqrt{(Y/Y_o)}}{1 + \sqrt{(Y/Y_o)}} = \frac{1 - \sqrt{(Y/(Y_m + Y))}}{1 + \sqrt{(Y/(Y_m + Y))}}. \tag{24}$$

According to (24), the maximum of  $k$  ( $k_{max} = 1$ ) is occurred when  $Y = 0$ . Setting (2) equals to zero produces below polynomial equation.

$$\omega_i^2 x_3 - x_4 = \omega_i^4, \tag{25}$$

where  $x_3 = (\omega_{dp}^2 + \omega_{cs}^2 + \omega_{csdp}^2)$  and  $x_4 = \omega_{dp}^2 \omega_{cs}^2$  are unknown parameters and  $\omega_i = 2\pi F_i$ ,  $F_i = f_{b1} f_{b2}$ . The  $f_{b1}$  and  $f_{b2}$  are operation frequencies in the backward direction coupling.

Then, according to (23) and (25), we have four equations and four unknown parameters and the equations set are as below:

$$[A].[X] = [D], \tag{26}$$

**Table 3.** Intermodulation components of the operational frequencies.

$f_i$ GHz	$2f_1 - f_i$ GHz	$2f_2 - f_i$ GHz	$2f_3 - f_i$ GHz	$2f_4 - f_i$ GHz
$f_1 = 2.9$	×	4.7	5.7	7.1
$f_2 = 3.8$	2	×	4.8	6.2
$f_3 = 4.3$	1.5	3.3	×	5.7
$f_4 = 5$	0.8	2.6	3.6	×

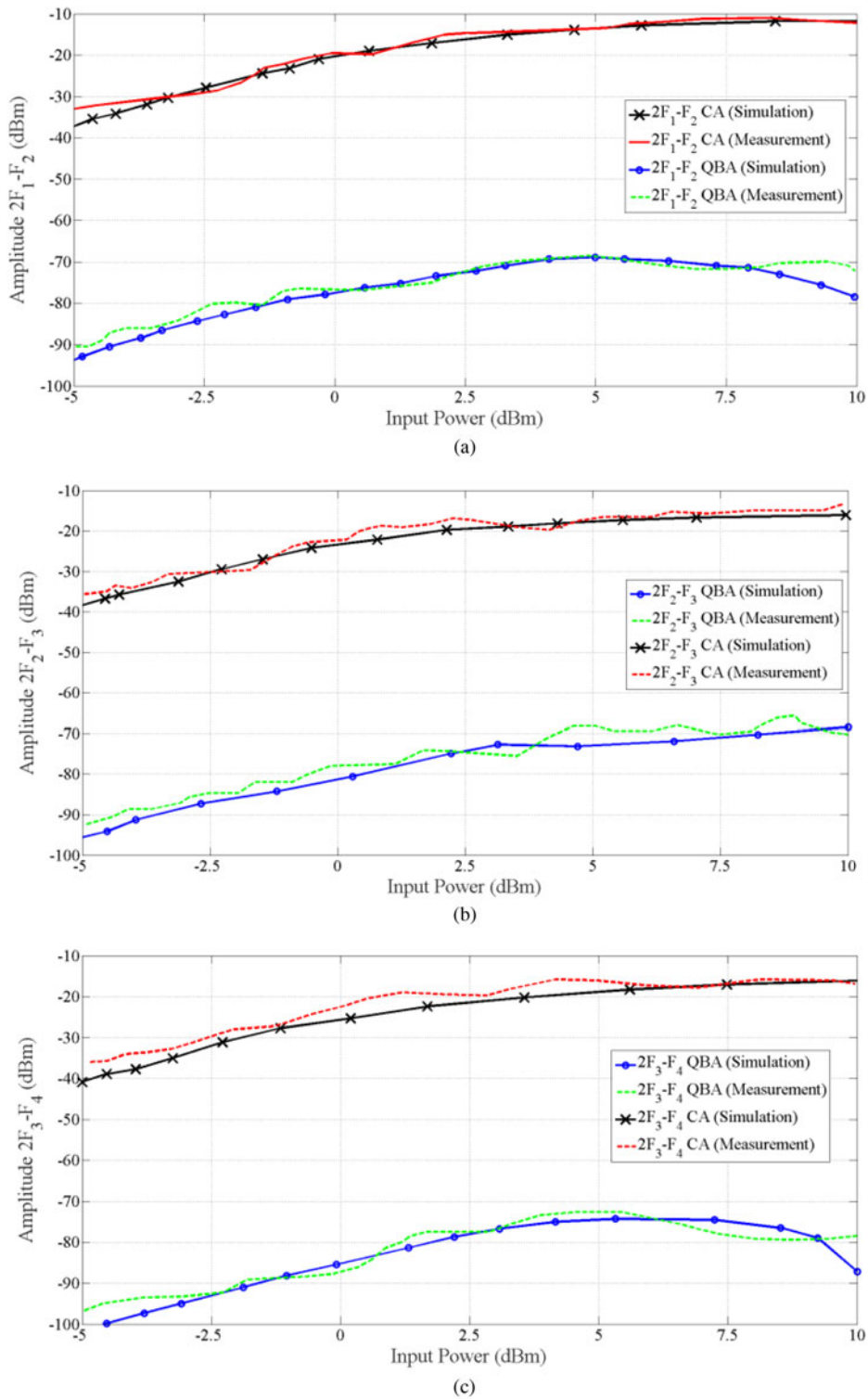


Fig. 9. (a) 2F<sub>1</sub> - F<sub>2</sub>, (b) 2F<sub>2</sub> - F<sub>3</sub>, (c) 2F<sub>3</sub> - F<sub>4</sub>, intermodulation components levels of the proposed QBA and conventional amplifier for F<sub>1</sub> = 2.9 GHz, F<sub>2</sub> = 3.8 GHz, F<sub>3</sub> = 4.3 GHz, and F<sub>4</sub> = 5 GHz (simulation and measurement results).

where

and

$$A = \begin{bmatrix} 2\omega_{f_1}^2 & -2 & 0 & 0 \\ 2\omega_{f_2}^2 & -2 & 0 & 0 \\ 0 & 0 & \omega_{b_1}^2 & -1 \\ 0 & 0 & \omega_{b_2}^2 & -1 \end{bmatrix}, \quad D = \begin{bmatrix} C_{f_1} \omega_{f_1} \\ C_{f_2} \omega_{f_2} \\ \omega_{b_1}^4 \\ \omega_{b_2}^4 \end{bmatrix} \quad (27)$$

$$[X] = \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{bmatrix} = \begin{bmatrix} C_m \\ C_m \omega_{cp}^2 \\ (\omega_{dp}^2 + \omega_{cp}^2 + \omega_{dscp}^2) \\ \omega_{dp}^2 \omega_{cs}^2 \end{bmatrix} \quad (28)$$

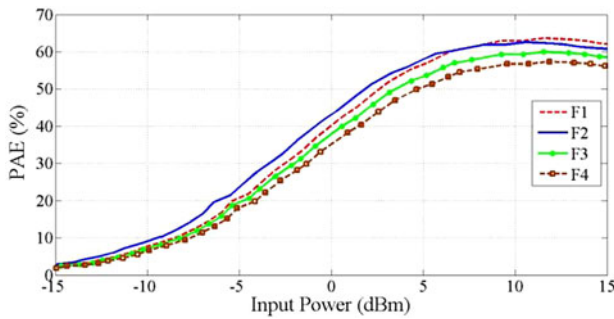


Fig. 10. Measured PAE of the proposed QBA for  $F_1 = 2.9$  GHz,  $F_2 = 3.8$  GHz,  $F_3 = 4.3$  GHz, and  $F_4 = 5$  GHz.

A set of four linear equations could be solved under balance conditions (6) simply as follows:

$$\begin{cases} C_m = x_1, \\ \omega_{cs}^2 = \omega_{cp}^2 = x_2/x_1, \\ \omega_{ds}^2 = \omega_{dp}^2 = (x_1 x_4)/x_2, \\ \omega_{dscp}^2 = x_3 - \omega_{dp}^2 - \omega_{cp}^2. \end{cases} \quad (29)$$

According to  $C_m$  the gap spacing ( $S$ ) between the E-CRLH-coupled lines is determined. Nevertheless, for extracting eight lumped elements of E-CRLH unit cell of Fig. 1 matching condition (15) is necessary at four operation frequencies  $f_{f_1}$ ,  $f_{f_2}$ ,  $f_{b_1}$ , and  $f_{b_2}$ .

### III. PROPOSED QBA THEORY

Equations Fig. 2 shows the proposed application of the QB CLs in the QBA. According to this figure when a wide band spectrum that including  $F_1$ ,  $F_2$ ,  $F_3$ , and  $F_4$  frequency bands incoming from port A then the output signal has QB spectrum and coming out from port B. Because first CLs operates as multiplexer and its dual-band output signals are amplified with two amplifiers. Finally, two amplified signals are demultiplexed by means of the other CLs and coming out from port B. In addition, ports C and D are loaded with  $50 \Omega$ .

This idea can be used in power amplifier or low-noise amplifier. Traditionally, the nonlinearity of an RF amplifier is described by third-order intermodulation point (IP<sub>3</sub>) or, equivalently, by the 1 dB compression point. However, the nonlinearity control is specified by the out-of-band power emission levels. Out-of-band emission is also called spectrum regrowth. Spectral regrowth is a condition that interferes with signals in adjacent frequency bands and can be reduced by using power amplifier linearization techniques [12].

In power amplifier applications instead of four frequency bands, only two frequency bands push into each amplifier. Therefore, intermodulation components decrease and linearity improve. Due to every other frequency bands are driven to each amplifier, so third-order intermodulation components are out of frequency bands. This concept is shown in Fig. 3. According to this figure, the third-order intermodulation components in the proposed QBA reduce to one component near the frequency bands. Whereas, the third-order

intermodulation components in the conventional amplifier distort the main operational frequency bands and we must use bank filter and isolated amplifiers for each frequency bands.

In low-noise amplifier applications, the proposed structure can improve the sensitivity up to 6 dB. As in the conventional receivers when four frequency bands need to receive simultaneously, we cannot use switches before the bank filter. As a result, splitter or multiplexer must be used before the bank filter (Fig. 4) [13, 14]. Using splitter leads to 6 dB reduction in the receiver sensitivity. Consequently, multiplexer is better solution for this concept. In the proposed structure, bank filter and multiplexer are combined with each other.

### IV. SIMULATION AND EXPERIMENTAL RESULTS

Citations to confirm the proposed structure, a QB CLs and a QBA based on the design procedure and presented expressions realized (Fig. 5). The structures has been fabricated on RO-4003 substrate with  $\epsilon_r = 3.38$ , thickness of 20 mil and  $\tan \delta = 0.003$ . The QBA amplifiers work in class-A with the PHEMT technology amplifier (MGA-82563). The amplifiers is biased in  $V_{GS} = 0$  V,  $V_{DS} = 3$  V, and  $I_{DS} = 80$  mA. The resistive feedback in this amplifier ensures stability ( $K > 1$  and  $|\Delta| < 1$ ) over all frequencies and temperatures.

The proposed structures operate at four frequencies:  $F_1 = 2.9$  GHz,  $F_2 = 3.8$  GHz,  $F_3 = 4.3$  GHz, and  $F_4 = 5$  GHz. According to these conditions, the lumped element components in Fig. 1 are calculated (Table 1). The dimensions of the proposed structure (Fig. 1) are presented in Table 2.

The absolute value of the dispersion curves of the proposed E-CRLH CLs unit cell (simulation and measurement) is presented in Fig. 6. Figure 7 presents the full-wave simulated (by using Agilent ADS software) and measured S-parameters for the QB CLs of Fig. 5(a). Using these figures, an amplitude balance of  $\pm 1$  dB over a bandwidth of 300 MHz, a matching ( $-10$  dB bandwidth) and an isolation at least  $-15$  dB and  $-20$  dB over bandwidth of operation in backward and forward direction are observed, respectively.

Furthermore, a QBA with at least 300 MHz bandwidth around the design frequencies is realized in the measured prototype (Fig. 8). As can be seen, a QBA around the design frequencies has been achieved. The proposed QBA nearly achieves 12 dB gain and return loss better than 15 dB in operation frequency bands. Excellent agreement can be observed between simulated and experimental results.

The intermodulation components of the four operation frequency bands are presented in Table 3. According to this table when four operational frequencies with 300 MHz bandwidth drive to the conventional amplifier there are some intermodulation components near to the operational frequencies. In band intermodulation components are marked in Table 3. In addition, Fig. 9 exhibits simulation and measurement results of the intermodulation components of the proposed QBA and conventional amplifier. According to this figure, intermodulation components in the proposed QBA decrease at least 50 dB proper to the conventional amplifier. Figure 10 shows the measured power-added efficiency (PAE) of the proposed QBA at four frequency bands. According to this figure, the PAE is better than 57% in pick power at all four frequency bands.

## V. CONCLUSION

In this paper, we present a new technique for realizing a QBA. New QBA consists of two QB E-CRLH coupled lines and two amplifiers. Designed and fabricated QBA at center frequencies  $F_1 = 2.9$  GHz,  $F_2 = 3.8$  GHz,  $F_3 = 4.3$  GHz, and  $F_4 = 5$  GHz exhibits a matching (15 dB) bandwidth of over 300 MHz and at least 12 dB gain operation frequency bands. The length and the width of the proposed structure are approximately 7.7 and 2.4 cm, respectively. The intermodulation components of the proposed QBA decreases at least 50 dB proper to the conventional amplifier and linearity improve. Due to the multiband, compact size, and linearity improvement, the proposed QBA is well suitable for microwave and millimeter-wave-integrated circuits, multi-standard communication systems and many kinds of transceivers.

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