A Varactor Tuned Branch-Line Hybrid Coupler

Ernest A. Fardin, Kamran Ghorbani and Anthony S. Holland

School of Electrical and Computer Engineering

RMIT University

GPO Box 2476V, Melbourne, 3001, Australia

Email: efardin@ieee.org

Telephone: +61 3 9925 3250

Fax: +61 3 9925 2007

Abstract— This paper introduces a novel branch-line 90° hybrid coupler incorporating varactor diodes which allow tuning of the frequency response. A design covering the DCS, PCS and IMT2000 cellular frequency bands (1710-2170 MHz) is presented. Given a varactor tunability of 2.5:1, simulations suggest 20 dB return loss and 3±1 dB coupling is achievable across each transmit and receive sub-band by tuning the varactors. These results offer an improvement over a conventional single-section branch-line hybrid centred at 1950 MHz, and some miniaturisation is also achieved due to the capacitive loading. A prototype is constructed using commercially available varactor diodes, and reasonable agreement between the measured and simulated results is achieved.

I. INTRODUCTION

With the proliferation of wireless applications over the past decade, there is a need to incorporate radios operating at different frequency bands into a single hardware device. Separate transceivers for each band lead to an increased component count and hence cost [1]. This means there is an incentive to employ frequency agile components in future RF front-end designs which must cover several frequency bands. Possible alternatives would be to use broad band components, or to switch between multiple narrow band components. However, these solutions often require more circuit real-estate and result in higher cost.

Varactor diodes have been used for many years to electronically tune microwave circuits. They have previously been integrated into a microstrip directional coupler for the 4 GHz band [2]. The purpose of the varactors in [2] was to vary the coupling level from 4 dB to 20 dB. More recently, a varactor tuned LC resonator was used to extend the operating frequency range of a microstrip directional coupler [3]. In the resonator-based tunable coupler, the regions of flat coupling, high return loss and high directivity are narrow-band, but the return loss and directivity nulls can be tuned in frequency by the varactor. This paper applies the varactor tuning technique to a 3 dB hybrid coupler.

A schematic diagram of the proposed hybrid coupler is shown in Fig. 1. The design is based on the branch-line 90° hybrid coupler, a standard component in many microwave systems. The conventional design employs two 35 Ω series arms and two 50 Ω shunt arms, 1/4-wavelength long at the centre frequency, to provide 3 dB power split with 90° phase offset between the coupled ports [4]. In this design, four
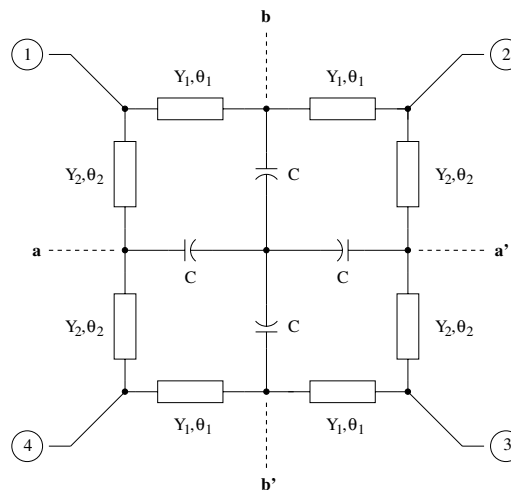

Fig. 1. Schematic diagram of the varactor tuned hybrid coupler.

varactors are employed to extend the operating frequency range of the conventional hybrid, while maintaining a 3 dB coupling level. Changing the varactor capacitance, C, modifies the electrical length of the series and shunt transmission line sections, and therefore the frequency at which 3 dB coupling is achieved.

II. ANALYSIS OF VARACTOR TUNED HYBRID

The equations for the S-parameters of the varactor tuned hybrid can be determined by the even-odd mode technique [5]. With aa′ and bb′ in Fig. 1 alternately open circuit and short circuit, and by considering the fourfold symmetry of the structure, the four subnetworks shown in Fig. 2 are obtained. The capacitor only appears in (b) and (c), since it is open circuit and short circuit in (a) and (d), respectively. Starting with the circuit of Fig. 2(a),

$$Y_{in,1a} = jY_1 \tan\theta_1 \qquad (1)$$

$$Y_{in,2a} = jY_2 \tan\theta_2 \qquad (2)$$

where $Y_{in,1a}$ and $Y_{in,2a}$ are the input admittances of the microstrip lines with characteristic admittance Y_1 and Y_2 and electrical length θ_1 and θ_2, respectively. In this case, both microstrip lines are terminated in an open circuit. Therefore,
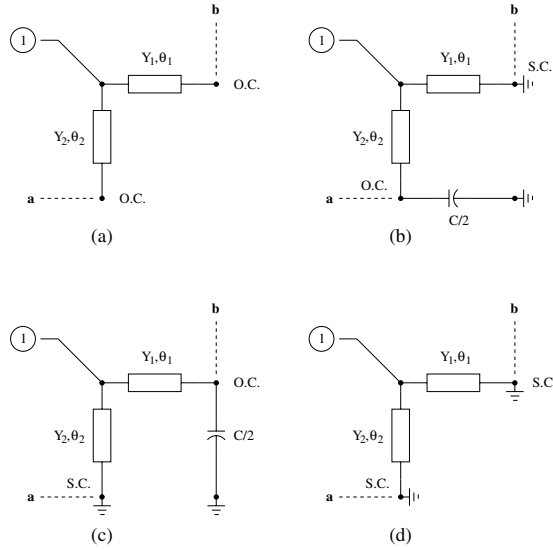
Fig. 2. Even-odd mode subnetworks (a) aa′ and bb′ open circuit (b) aa′ open circuit, bb′ short circuit (c) aa′ short circuit, bb′ open circuit (d) aa′ and bb′ short circuit.

the reflection coefficient at port 1 is

$$\Gamma_a = \frac{Y_o - (Y_{in,1a} + Y_{in,2a})}{Y_o + Y_{in,1a} + Y_{in,2a}} \tag{3}$$

where Y_o is the characteristic admittance. Similarly,

$$Y_{in,1b} = \frac{Y_1}{j \tan \theta_1} \tag{4}$$

$$Y_{in,2b} = \frac{\frac{1}{Y_2} + (\frac{2}{\omega C}) \tan \theta_2}{(\frac{1}{Y_2}) \cdot (\frac{2}{j\omega C} + j\frac{1}{Y_1} \tan \theta_2)} \tag{5}$$

$$Y_{in,1c} = \frac{\frac{1}{Y_1} + (\frac{2}{\omega C}) \tan \theta_1}{(\frac{1}{Y_1}) \cdot (\frac{2}{j\omega C} + j\frac{1}{Y_1} \tan \theta_1)} \tag{6}$$

$$Y_{in,2c} = \frac{Y_2}{j \tan \theta_2} \tag{7}$$

$$Y_{in,1d} = Y_{in,1b} \tag{8}$$

$$Y_{in,2d} = Y_{in,2c} \tag{9}$$

Finally, by superposition, the S-parameters can be calculated using

$$S_{11} = \frac{1}{4}(\Gamma_a + \Gamma_b + \Gamma_c + \Gamma_d) \tag{10}$$

$$S_{21} = \frac{1}{4}(\Gamma_a - \Gamma_b + \Gamma_c - \Gamma_d) \tag{11}$$

$$S_{31} = \frac{1}{4}(\Gamma_a - \Gamma_b - \Gamma_c + \Gamma_d) \tag{12}$$

$$S_{41} = \frac{1}{4}(\Gamma_a + \Gamma_b - \Gamma_c - \Gamma_d) \tag{13}$$

where Γ_a, Γ_b, Γ_c and Γ_d are the reflection coefficients at port 1 under the four different conditions illustrated in Fig. 2(a)-(d). In order to simulate the circuit, the values of Y_o and Y_2 are set
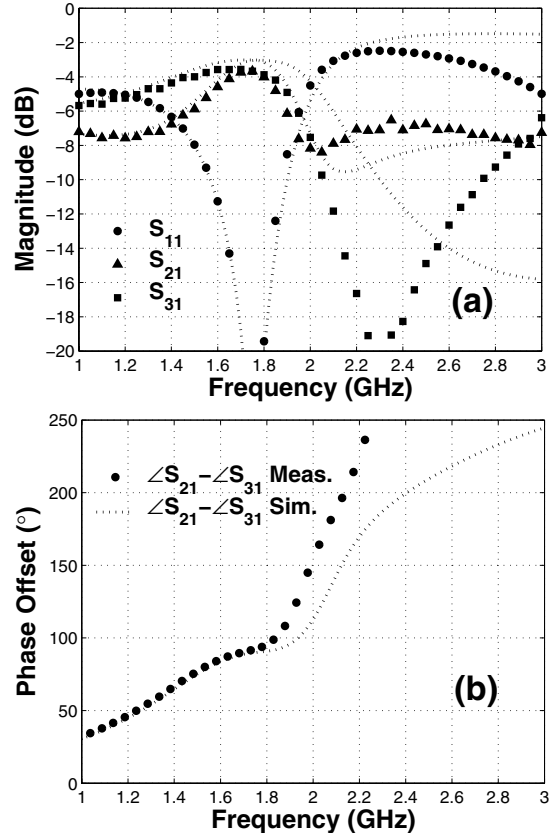

Fig. 3. Coupler characteristics at DCS uplink band (a) measured and simulated S-parameters (b) measured and simulated phase offset between coupled ports. The measured and simulated results are presented as point series and dashed lines, respectively.

to $1/50 \ \Omega^{-1}$, and Y_1 is set to $1/35 \ \Omega^{-1}$, as in the conventional design [4]. The electrical length of the microstrip lines can be calculated using

$$\theta = \frac{\omega \ell \sqrt{\epsilon_{r,eff}}}{c} \tag{14}$$

where ℓ is the physical length of the line, $\epsilon_{r,eff}$ is the effective dielectric constant of the substrate material, and c is the speed of light.

III. SIMULATED AND MEASURED RESULTS

In order to validate the design prior to fabrication, the circuit was simulated using both the analytical solutions (10) to (13) and Agilent Advanced Design System 2004A (ADS) software. The substrate material is 0.508 mm Rogers 4003C ($\epsilon_r = 3.38$). A circuit model identical to that of Fig. 1 was simulated using ADS, with series and shunt branch line lengths of 18 mm. The simulation results, shown as dashed lines in Figs. 3 and 4, were obtained using (10) to (13). In Figs. 3 and 4, the varactor capacitance is set to 2.0 pF and 0.8 pF, respectively. Adjusting the varactor capacitance over this range allows the tunable hybrid to cover 1710 to 2170 MHz with better than 20 dB return loss. The required varactor tuning range of 2.5:1 is well with in the capability of commercially available varactor

Frequency Band	Band Edge (MHz)	Coupling (S_{21}, dB)	Return Loss (dB)	Coupled Port Phase Offset (°)	V_{bias} (V)
DCS uplink	1710	3.7	21.1	89.8	2.1
	1785	3.8	22.4	92.8	
DCS downlink	1805	3.5	26.8	90.7	2.7
	1850	3.6	25.9	92.0	
PCS uplink	1850	3.5	24.6	90.9	3.2
	1910	3.5	25.0	92.2	
PCS downlink	1930	3.4	24.7	91.0	4.1
	1990	3.5	26.1	92.2	
IMT2000 uplink	1920	3.5	24.7	90.9	4.0
	1980	3.5	26.2	92.0	
IMT2000 downlink	2110	3.3	27.4	91.6	8.0
	2170	3.4	22.3	92.1	

diodes. As expected, the results of the ADS simulation were identical to the analytical values. Note that ideal varactors are assumed for the simulated results.

A prototype circuit was constructed using Skyworks SMV1231-079 varactors in place of the ideal components. DC bias was applied to port 1 via a bias tee, and DC blocking capacitors were placed at ports 2, 3 and 4. A 10 nH chip inductor was placed between the common node of the varactor diodes and ground, in order to establish a bias voltage across the diodes.

The measured results are presented as marker series in Figs. 3 and 4. Bias voltages of 2.1 and 8.0 V were applied to tune the hybrid to the DCS uplink (Fig. 3) and IMT2000 downlink (Fig. 4), respectively. It is evident from these results that the coupler bandwidth decreases as the varactor capacitance increases. The bandwidth should still be sufficient to cover a single cellular uplink or downlink channel with acceptable coupling variation. There is good agreement between measured and simulated results in Fig. 4. However, the analytical model loses accuracy beyond 1.8 GHz in Fig. 3. This could be attributed to the series resistance and lead inductance of the varactor diodes. Also, a short length of high impedance transmission line must be placed between the anode of each diode and the common node, for layout purposes. The effect of this transmission line was not modelled.

A summary of measured results for each uplink and downlink channel in the DCS, PCS and IMT2000 frequency bands is presented in Table I. The coupling level at ports 2 and 3 is within 3 ± 1 dB and the return loss better than 20 dB in each case. A coupled port phase offset of $90\pm3°$ is achieved. In comparison, the S_{11} response of a conventional single-section branch-line hybrid designed for 1950 MHz has a 20 dB return loss bandwidth of ~200 MHz. Each of the four branch lines in the conventional design [4] is 23.3 mm long, while the varactor tuned design uses 18 mm branches. Some useful miniaturisation is therefore achieved by the use of varactors.
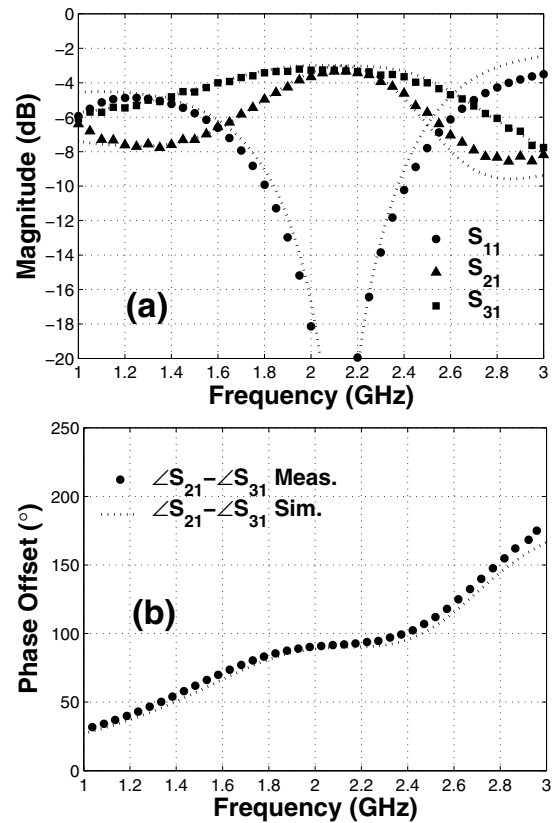

Fig. 4.   Coupler characteristics at IMT2000 downlink band (a) measured and simulated S-parameters (b) measured and simulated phase offset between coupled ports. The measured and simulated results are presented as point series and dashed lines, respectively.

IV. Conclusions

This paper has presented a novel branch-line coupler configuration which uses varactors to extend the bandwidth of the conventional 3 dB branch-line hybrid design. Equations for the S-parameters of the device have been derived by the even-odd mode analysis technique. The suitability for implementation was demonstrated by even-odd mode analysis and ADS simulations. A single-section hybrid covering 1710 to 2170 MHz with 3±1 dB coupling and 20 dB minimum return loss was fabricated. This design approach may provide a lower cost alternative to implementing several conventional 90° hybrids in a multiband transceiver.

Acknowledgment

References

[1] A. R. Rofougaran, M. Rofougaran, and A. Behzad, "Radios for the next-generation wireless networks," *IEEE Microwave*, vol. 6, pp. 38–43, 2005.

[2] S. Toyoda, "Variable coupling directional couplers using varactor diodes," in *IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 82, 1982, pp. 419–421.

[3] C.-S. Kim, C.-S. Yoon, J.-S. Park, D. Ahn, J.-B. Lim, and S.-I. Yang, "Design of the novel varactor tuned directional coupler," in *IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 4, 1999, pp. 1725–1728.

[4] D. M. Pozar, *Microwave Engineering*, 2nd ed.   New York: John Wiley, 1998.

[5] R. E. Collin, *Foundations for Microwave Engineering*.   New York: McGraw Hill, 1992.