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# An All-Pass Topology to Design a 0-360° Continuous Phase Shifter with Low Insertion Loss and Constant Differential Phase Shift

Khaled Khoder, Marc Le Roy, André Pérennec

*Lab-STICC (UMR CNRS 6285), Université Européenne de Bretagne (UEB), Université de Bretagne Occidentale (UBO)  
6 avenue Le Gorgeu, CS93837, 29238 Brest cedex 3, France.*

Marc.LeRoy@univ-brest.fr

Kh.Khoder@gmail.com

**Abstract**— In this paper, an analog phase shifter is designed by using a novel all-pass topology. The phase shift can be continuously adjusted from 0 up to 380° by biasing varactor-diodes while maintaining the differential phase shift constant across the 6.7 GHz - 7.7 GHz band. This two-stage circuit is simple and compact with respectively insertion losses of  $2.9 \text{ dB} \pm 1.3 \text{ dB}$ , return losses better than 9.4 dB and a differential phase shift flatness of  $\pm 11^\circ$  in the worst case. With a 90.5°/dB Figure-of-Merit, this topology presents an interesting trade-off between low-cost, low loss, large phase-shift range, phase flatness and bandwidth. Measurements are discussed and carefully compared to current competing topologies.

## I. INTRODUCTION

Phase shifters (PS) are key components in RF and microwave front-end, particularly in phased-array transmitters and receivers to get electronic adaptive beam-forming and beam steering. Active and passive phase shifters can be distinguished. Active ones achieve 360° phase shift range by summing two amplified quadrature signals [1] as done in IQ modulator. However, power consumption and non-reciprocity may restrict their application domain.

Passive phase shifters can also be organized into two types: analog and digital.

This work focuses on analog passive phase shifters which prime interest is the possibility of accurate phase control with a continuously adjustable phase-shift, contrarily to digital ones that provide a discrete set of phase shift values. Moreover, analog phase shifters require a low control-complexity with a lower number of control voltages (preferably one) than for digital phase shifters. The number of digital to analog converter is thus minimized. If necessary, analog phase shifters may be easily transformed in discrete ones.

The initial choice of topology is crucial to satisfy phase shifter specifications or at least to obtain a satisfactory trade-off between these constraints. The classical design targets are: an accurate phase variation up to 360°, low loss, low cost, low complexity (for the control network and globally), compactness and compatibility with available implementation process. The phase-shift variation-range versus the active component tunability is a key parameter which also strongly depends on topology. If considering using phase shifters in a large phased-array, constant or flat differential phase-shift

over the signal bandwidth is required whatever the phase-shift value. This flat phase shift requirement also applies to broadband applications to avoid frequency beam squint.

We present, here, a topology based on distributed coupled lines and varactors. This technology meets the low cost and simplicity requirements (i.e. simple voltage control, easy implementation, no-packaging additional cost as for MEMS for example) together with a continuous tunable range, a rather small bias-voltage, low consumption and fast switching. Moreover, varactor-based passive phase shifters are CMOS compatible, even at 60 GHz [1]-[2]. Under this configuration, i.e. varactors and distributed networks, many analog passive phase shifters have been reported in the literature [1]-[11]. A comparison based on the Figure-of-Merit (FoM) of significant contributions of that type of PSs is proposed afterward in table II. Reflection-Type Phase-Shifters (RTPS) are mostly used [6]-[11]: they are mostly based on 90° branch-line couplers where two ports are terminated in reflective loads, varactors in most cases. The RTPS main issues are related i) to the coupler which is initially narrow band and ii) to the limited phase-shift variation for a single cell. Several improvement techniques to compensate for these drawbacks have been described [1], [8]-[11]; these results are also summarized in table II.

To illustrate the trade-off between PS goals introduced by our all-pass topology, it is compared in table I in a single stage configuration to a branch-line RTPS and also to another all-pass topology proposed by Hayashi [3]. This comparison is done for an identical capacitance control ratio ( $C_{\max}/C_{\min}$ ) at two different operating frequencies. Design rules are thus derived to improve the variation range in phase.

Then, two proof-of-concept analog PSs designed in low-cost hybrid technology are presented. They respectively reach a phase-control of 180° for the single stage PS in the 6-7 GHz band and 360° for the two-stage one in the 6.7-7.7 GHz band whereas the differential phase-shifts are kept constant over the phase-variation range. Measurements are presented and discussed for both circuits and benchmarked against current analog passive PSs, followed by conclusions and prospects.

## II. TOPOLOGY AND DESIGN

### A. Topology Principle

In order to obtain a maximum phase-agility with a constant phase shift over a wide frequency band, two approaches were investigated in parallel. In the first one, we tried different configuration to introduce agility in the well-known Schiffman topology while maintaining its interesting phase characteristic, i.e. a constant phase shift difference. The second one focused on all-pass topologies and particularly on improvements of an approach proposed by Hayashi [3] (Table I). Finally, the proposed approach (Fig. 1) can be considered as either a coupled-line section, as Schiffman PS, with an optimized placement of varactors to get a large phase-agility or as a configuration of Hayashi all-pass network folded to introduce a coupling-coefficient between the two lines.

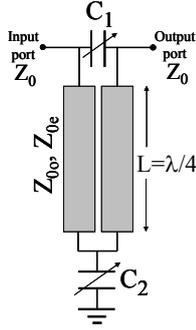


Fig. 1 Schematic of the proposed analog PS.

### B. Comparison between Single-Stage All-Pass PSs and RTPS

In Table I, the novel all-pass topology is compared to Hayashi PS and to a classic RTPS implemented with a branch-line coupler. This benchmark is performed in basic single stage configurations. For all these topologies, the matching conditions are optimal for quarter-wavelength lines at  $f_0$ .

For the proposed topology, the  $S_{11}$ -magnitude equation at  $f_0$  is given as follows:

$$S_{11}(f_0) = \frac{Z_{0e}^2 C_2 \omega_0 - 4Z_0^2 C_1 \omega_0}{Z_{0e}^2 C_2 \omega_0 + 4Z_0^2 C_1 \omega_0 + 2j(Z_{0e}^2 C_1 C_2 \omega_0^2 - 1)} \quad (1)$$

For ideal matching conditions, a simplified relationship between impedances and varactor-capacitance values is obtained:

$$C_2 = 4\left(\frac{Z_0}{Z_{0e}}\right)^2 C_1 \quad (2)$$

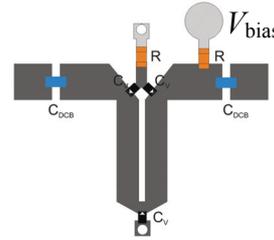
As previously detailed in [12], when the even-mode impedance  $Z_{0e}$  is set to  $50\Omega$ , it corresponds to an uncoupled line-section and thus to Hayashi PS with a ratio  $k = C_2/C_1 = 4$ . In other word, our topology can be considered as a generalized form of Hayashi one. In the coupled-lines configuration,  $k=2$  is the integer number closer to the optimal case and it relaxes technological constraint thanks to a

medium coupling coefficient. In that case,  $Z_{0e} = \sqrt{2} \cdot Z_0$ . In table I, the first column of results shows that, at the normalized frequency  $f_0$ , the three PSs are well-matched with a slightly narrower band for the RTPS and that the phase-shift ranges are quite similar for an identical varactor-capacitance variation. The optimum impedance values and  $C_2/C_1$  ratios are also recalled in Table I.

A wideband analysis shows that the proposed PS exhibits a maximum phase-shift range around  $f_0/2$  while becoming mismatched. As evidenced in the last column of table I, in that frequency band, optimizing the even- and odd-mode impedances allows getting simultaneously matched conditions over the whole varactor-capacitance variation and a wide phase-shift range (more than  $180^\circ$ ). At  $f_0/2$ , no-adjustment on RTPS (impedances, line length) provides such characteristics. In Hayashi cell, the return losses get higher than  $-10$  dB for high varactor-capacitance values which limit the phase-shift variation below  $85^\circ$ . Besides having intrinsically the smaller area among the 3 cases, our circuit has its size reduced by a factor 2 when operating at  $f_0/2$ .

### C. A Single-Stage $180^\circ$ PS Prototype

A single stage proof-of-concept  $180^\circ$  PS has been implemented in low-cost microstrip technology in the 4-7 GHz frequency band [12]. The MA46H120 varactor S-parameters were initially measured and the resulting capacitance-value ranges from 1.15 to 0.15 pF. The components' placement has been studied in order to require only a single bias-point. Fig. 2 shows the design schematic; measured S-parameters are presented in Fig.3 for the 4-7 GHz band. The phase-shift variation ranges of more than  $180^\circ$ . In the 6-7 GHz band, insertion losses and return losses are respectively better than 1.8 dB and 12 dB; the FoM reaches  $105^\circ/\text{dB}$  and the phase flatness is of  $\pm 8^\circ$ . These first results confirm the topology trade-off between wide phase-shift range, low-loss and bandwidth. Nevertheless, the phase-flatness could be improved.



$C_{DCB} = 47$  pF ;  $R = 3.3$  k $\Omega$   
 $Z_{0e} = 92\Omega$  ;  $Z_{0o} = 47\Omega$   
 $C_v$ : MA46H120  
 $V_{bias}$ : 0-18 V  
Area : 11.5x11.5 mm  
Substrate: RF35 ;  $h = 0.78$ mm;  
 $T = 17.5$   $\mu$ m  
 $\epsilon_r = 3.5$  ;  $\tan \delta = 0.0023$

Fig. 2 Layout of the single-stage  $180^\circ$  analog PS.

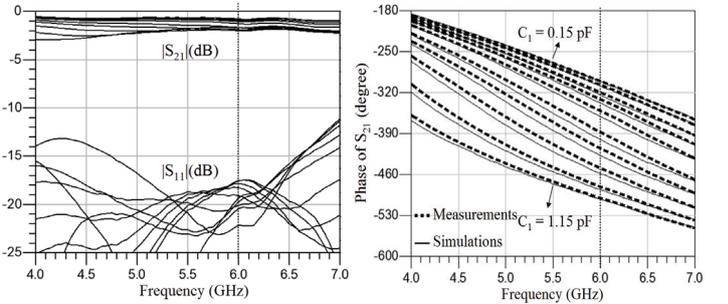


Fig. 3 Measured S-parameters magnitude and phase of  $S_{21}$ .

TABLE I  
PERFORMANCE COMPARISON OF SINGLE STAGE PS TOPOLOGIES

PS topology			$f_0$	$f_0/2$
RTPS (90° branch-line coupler)		Optimal case	$Z_0, Z_0 / \sqrt{2}, \lambda/4$	Unmatched (can be matched by impedance optimization but without significant phase-variation)
		$ S_{11} $ (dB)	perfectly matched (narrowband)	
		$\Delta C$ (pF)	$0.1 \rightarrow 1$	
		$\Delta\phi$ (°)	110	
Hayashi PS (all-pass network) [2]		Optimal case	$Z_C=Z_0, \lambda/4, C_2=4.C_1$	$ S_{11} _{dB} < -10$ dB
		$ S_{11} $ (dB)	perfectly matched (wideband)	
		$\Delta C$ (pF)	$0.1 \rightarrow 1$	
		$\Delta\phi$ (°)	110	
Proposed PS (all-pass network)		Optimal case	$Z_{0e} = \sqrt{2}.Z_0$ $L = \lambda/4, C_2=2.C_1$	$ S_{11} _{dB} < -11$ dB
		$ S_{11} $ (dB)	perfectly matched (wide band)	
		$\Delta C$ (pF)	$0.1 \rightarrow 1$	
		$\Delta\phi$ (°)	97	

### III. TWO-STAGE 360° ANALOG PHASE SHIFTER

Obtaining a 180° phase-shift together with low-loss in a basic-configuration single-stage PS means that this topology is a relevant 360° PS candidate. So, a two-stage PS has been designed using the same approach, i.e.  $k = C_2/C_1 = 2$  capacitance-ratio and coupled-section lengths close to  $\lambda/4$  at  $f_0$ , when the operating frequency is  $f_0/2$ . The all-pass topology still allows using a single bias-point. Finally, the odd- and even-mode impedances and the distributed-section lengths have been optimized while taking into account the measured varactor S-parameters in simulation and distributed-sections EM simulations. The components in use are identical to those of the single-stage PS. As illustrated by Fig. 4, the simulated phase-shift varies up to 385° at 7 GHz. The corresponding measured varactor-capacitance variation versus bias-voltage is also reported. Fig. 5 shows the photograph of the 360° PS implemented in microstrip technology. In Fig. 6, the corresponding measured S-parameters indicate that insertion losses are of  $2.9 \text{ dB} \pm 1.3 \text{ dB}$  and return losses are better than 9.4 dB in the 6.7-7.7 GHz frequency band. In Fig 9, the measured relative phase shift ranges from 0 to more 380° with a phase shift flatness of  $\pm 11^\circ$  in the worst case, i.e.  $V_{\text{bias}} = 5\text{V}$ . Measurements agree with simulation results (not presented here for clarity).

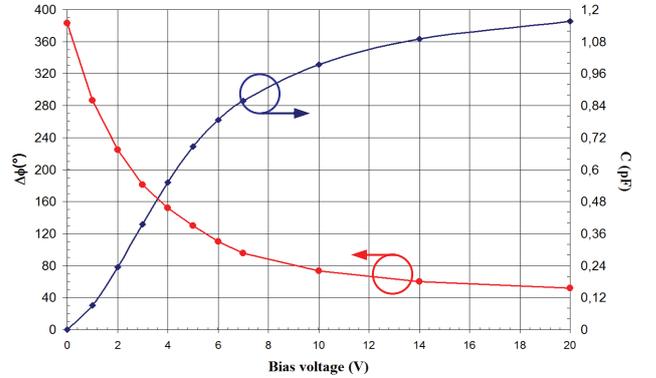
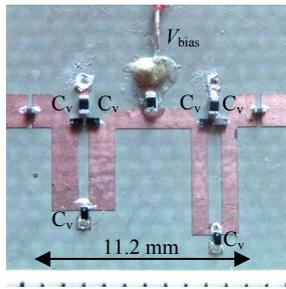


Fig. 4 Differential phase-shift and varactor-capacitance variation (MA46H120) versus bias-voltage at 7 GHz.

Even in this two-stage configuration and for that substrate, the device remains compact. The FoM is about 90°/dB. The performances of this analog PS are compared in table II with relevant contributions based on topologies with varactors and distributed lines. Several recent references get higher FoMs but they are based on improved loads in RTPS [9]-[11] and also at lower frequencies with narrower band [9]. A comparison of topologies in raw configuration [3]-[7] confirms that the proposed approach brings a new trade-off between FoM and bandwidth



$C_{DCB} = 47 \text{ pF}$   
 $R = 3.3 \text{ k}\Omega$   
 $C_V: \text{MA46H120}$   
 $V_{bias}: 0-18 \text{ V}$   
 Substrate: RF35  
 $h = 0.78 \text{ mm}$   
 $\epsilon_r = 3.5$   
 $\tan \delta = 0.0023$   
 $T = 17.5 \mu\text{m}$

Fig. 5 Photograph and components' details of the 360° PS.

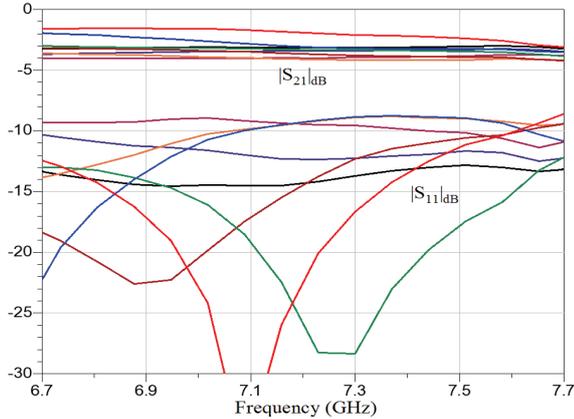


Fig. 6 Measured S-parameters.

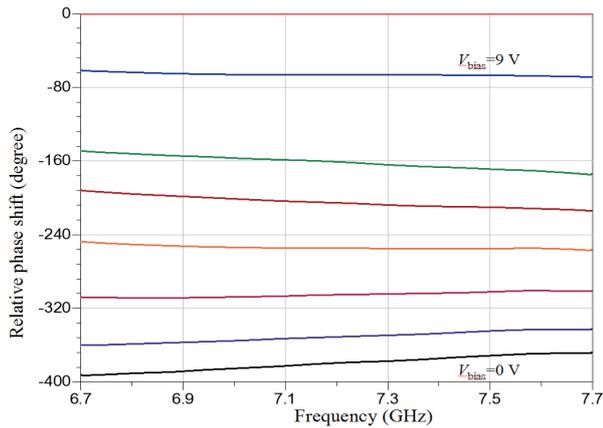


Fig. 7 Measured differential phase-shift.

#### IV. CONCLUSION

A novel analog PS topology based on an all-pass network has been presented. Compare to other single-cell classical topologies, an increase of the phase-shift range is expected. 180° and 360° continuously-adjustable PS prototypes have been implemented in low-cost technology in C-band. The resulting FoMs are respectively of 105°/dB and 90.5°/dB while keeping phase-shift flatness across the operating bandwidth. Those results obtained from a basic configuration show that the topology should be competitive to be used in phased-array networks implemented in CMOS technologies for X- and 60 GHz-bands. To that purpose, improvements could be done, e.g. substituting the varactors by optimized loads still with varactors as done in [9]-[11] for X-band and in [1]-[2] at 60 GHz.

TABLE II  
BENCHMARK OF VARACTOR-BASED ANALOG PHASE-SHIFTERS

Type	Freq. (GHz)	Phase control $\Delta\phi$ (°)	$IL_{max}$ (dB)	FoM* (°/dB)	Ref
All-pass network	12 - 14	180	4.7	38.2	[3]
Line loaded by varactors	5 - 6	360	5.7	53.7	[4]
LH line	4.4 - 5.6	180	4.5	40	[5]
RTPS	11 - 13	360	7.8	46.2	[6]
RTPS	10	380	5.3	68	[7]
Coupled-lines RTPS	2.2	360	9	40	[8]
RTPS with DGS	1.9 - 2.1	230	1.5	153	[9]
Improved RTPS	11.5-12.5	360	3.8	94	[10]
	30-31	315	5.4	58	
Improved RTPS	12	390	3	130	[11]
Improved RTPS and Lange-coupler	57.2-65.9	180	8	22.5	[1]
					[2]
<b>All-pass network</b>	<b>6-7</b>	<b>190</b>	<b>1.8</b>	<b>105</b>	<b>This work</b>
	<b>6.7 - 7.7</b>	<b>380</b>	<b>4.2</b>	<b>90.5</b>	

$$* FoM = \frac{\Delta\phi_{max}}{IL_{max}}$$

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