On the Phase Response and Radiation Efficiency of the Complementary Strip-Slot as an Array Element

Elena Abdo-Sánchez*, Teresa M. Martín-Guerrero*, Jaime Esteban†, and Carlos Camacho-Peñalosa*

*Dpto. Ingeniería de Comunicaciones, E.T.S.I. Telecomunicación, Universidad de Málaga, Andalucía Tech, E-29071 Málaga, Spain, Email: elenaabdo@ic.uma.es, teresa@ic.uma.es and ccp@ic.uma.es
†Dpto. Señales, Sistemas y Radiocomunicaciones, E.T.S.I. Telecomunicación, Universidad Politécnica de Madrid, E-28040 Madrid, Spain, Email: jesteban@etc.upm.es

***Abstract*—The complementary strip-slot element is a broadly-matched microstrip radiator that has been used to design innovative series-fed arrays. It consists of a microstrip series-fed slot that have its complementary stub on the layer of the microstrip and aligned to the slot. In this contribution, the influence of the strip and slot geometry on its performance is studied through the analysis of four different designs. The obtained results highlight the possibility of controlling the radiation efficiency or the phase response, without compromising the broad matching. Therefore, potential series-fed arrays built with this element can exploit this feature to set the magnitude and phase of the excitations with certain flexibility.**

***Index Terms*—array, microstrip, series-fed, strip-slot.**

I. Motivation

Recently, the authors proposed a novel planar radiating element based on a microstrip-centred-fed slot modified by adding its complementary stub on the microstrip layer and aligned to the slot (a prototype is shown in Fig. 1) [1]. An schematic of the radiating element is shown in Fig. 2, where the geometric variables are defined. This structure overcomes the narrowband response inherent in the resonant nature of planar radiators and has a very broadband impedance matching, whereas it behaves as a conventional microstrip-fed slot in terms of radiation.

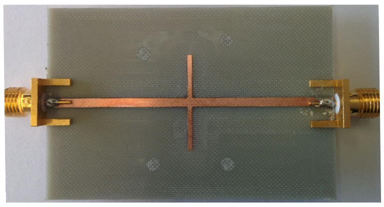
Due to its series feeding and broad matching, the complementary strip-slot element has been used to build travelling-wave arrays with interesting features, such as linear arrays with full frequency scanning in several bands [2], log-periodic arrays with size reduction [3] and multiband sequentially-rotated arrays with circular polarization [4]. These arrays have been manufactured as proofs of concept, without any particular specifications. However, if arrays with a better control of the radiation characteristics, such as specifications of the side-lobe level, are pursued, it is necessary to analyse how the phase response and the power radiated by the element can be controlled. Therefore, this paper deals with a study of these parameters as a function of the element geometry.

II. Theory

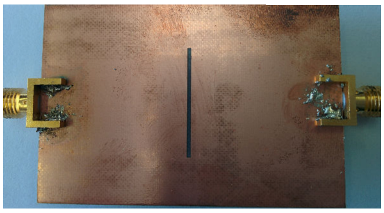
The resulting coupled structure formed by the strip and the slot has three conductors and, since it is symmetrical, an even and an odd mode propagate. Fig. 3 shows the electric field distributions of both modes in a cross section. This resulting coupled microstrip-slotline has been properly modeled by a lattice network [1], and the resulting image impedance can be obtained as

$$Z_{im} = \frac{1}{2}\sqrt{Z_{0e}Z_{0o}\cot\theta_e\tan\theta_o}, \tag{1}$$

where Z_{0e} and Z_{0o} are the characteristic impedances, and θ_e and θ_o, the electrical lengths of the even and odd modes, respectively.



(a) Strip Side



(b) Slot Side

Fig. 1. Prototype of the complementary strip-slot.

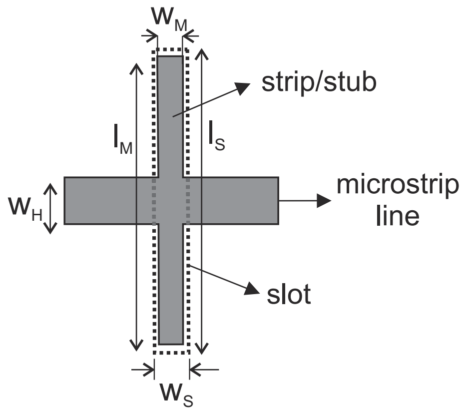
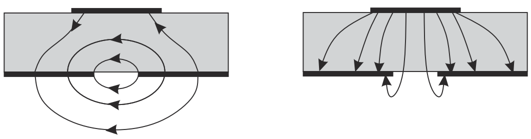


Fig. 2. Schematic of the complementary strip-slot.



(a) Odd mode (b) Even mode

Fig. 3. Electric field distributions for the modes of the microstrip-slotline coupling section.

Theoretical perfect matching can be achieved under the following conditions:

$$\theta_e = \theta_o = \theta \tag{2a}$$

$$\frac{1}{2}\sqrt{Z_{0e}Z_{0o}} = Z_0 \,, \tag{2b}$$

where Z_0 stands for the characteristic impedance of the feed microstrip line.

A distinctive feature of the microstrip-slotline coupling is that the even mode is mainly determined by the dimensions of the microstrip and the odd mode by the geometry of the slot [5]:

$$Z_{0e} \approx 2Z_M; \qquad \epsilon_{eff,e} \approx \epsilon_{eff,M} \tag{3a}$$

$$Z_{0o} \approx \frac{Z_S}{2}; \qquad \epsilon_{eff,o} \approx \epsilon_{eff,S} \,. \tag{3b}$$

where Z_M and Z_S stand for the characteristic impedances and $\epsilon_{eff,M}$ and $\epsilon_{eff,S}$, for the effective permittivities of such a microstrip line and a slotline, respectively.

Therefore, in the strip-slot element, the parameters of the modes can be controlled independently by adjusting the widths and the lengths of the strip and the slot. In this way, condition (2b) can be fulfilled by adjusting the widths and condition (2a), by adjusting the lengths of the slot l_S and the strip l_M so that $\sqrt{\epsilon_{eff,M}} \cdot l_M = \sqrt{\epsilon_{eff,S}} \cdot l_S$.

When condition (2a) is fulfilled, the phase factor (i.e., the phase of S_{21} provided the circuit is loaded by Z_{im}) is given by

$$\cos\phi = \frac{\frac{Z_{0e}}{Z_{0o}} - \tan^2\theta}{\frac{Z_{0e}}{Z_{0o}} + \tan^2\theta}. \tag{4}$$

Note that when $Z_{0e} = Z_{0o}$, $\phi = 2\theta$, and the phase factor ϕ is linear with frequency.

III. Study and Results

The geometric parameters to attain the design are four: w_M, w_S, l_M and l_S, defined in Fig. 2. Given a substrate, the parameters of the modes (Z_{0e}, Z_{0o}, $\epsilon_{eff,e}$ and $\epsilon_{eff,o}$) are determined by the widths w_M and w_S. Infinite pairs of element widths (w_M, w_S) can fulfill (2a). Therefore, there are extra degrees of freedom in the design while keeping the matching. In order to illustrate this, four different designs of strip-slot elements that guarantee the matching are used to study the phase response and the radiation efficiency. The dimensions of these designs are shown in Table I. All of them are designed to present $\theta = \pi/2$ (l_S=l_M=$\lambda/2$) at 5.4 GHz, in order to make the slot and the strip resonate at the same frequency, thus fulfilling condition (2a). Therefore, according to (4), the phase factor at 5.4 GHz is fixed at π. However, all the designs achieve these conditions with different pairs of widths. That means that $Z_{0e}Z_{0o}$ is constant for all of them, but Z_{0e}/Z_{0o}, and then the coupling coefficient $k = |\frac{Z_{0e}-Z_{0o}}{Z_{0e}+Z_{0o}}|$, is different for the four cases. For availability reasons, the chosen substrate was GIL 1032, with ϵ_r=3.2 and h=30mil.

TABLE I
Dimensions of different designs for comparison.

	w_M	w_S	l_M	l_S
Design 1	0.22 mm	0.10 mm	9.13 mm	10.45 mm
Design 2	0.61 mm	0.47 mm	9.18 mm	10.76 mm
Design 3	0.90 mm	0.80 mm	9.04 mm	10.86 mm
Design 4	1.90 mm	1.8 mm	8.72 mm	9.99 mm

The parameters of the even and odd modes in the strip-slot structure were obtained from the ANSYS HFSS electromagnetic simulator. Parametrization of the widths w_M and w_S was done, in order to extract conclusions about how these geometric variables influence the performance. No coupling between elements in a potential array has been taken into account, since the array designs that have been already carried out showed good results when analysing the arrays with the isolated element model.

The isolines of the image impedance of the coupled microstrip-slotline section at 5.4 GHz, when θ_e=θ_o is assumed, are shown in Fig. 4 as a function of w_S and w_M. The feed microstrip line was designed to have $50\,\Omega$-impedance level (i.e., w_H=1.83 mm). Therefore, the line of $50\,\Omega$ represents perfect matching. For this substrate, the results show that good matching is achieved when the slot and the strip have similar widths. It is also clear that a wide range of widths leads to a reasonably good matching for this substrate, since, for example, an image impedance of $40\,\Omega$ corresponds to return losses of around 20 dB. This fact means that the structure is quite robust with regard to dimensioning, an important advantage for design and fabrication. In addition, it can be observed that the four designs (marked with black dots) present very good matching.

Fig. 5 shows the isolines of the coupling coefficient, defined as $k = \left|\frac{Z_{0e}-Z_{0o}}{Z_{0e}+Z_{0o}}\right|$, of the coupled microstrip-slotline section as a function of the slot and strip widths, w_S and w_M, respectively. It can be observed that the four designs, which point out the line of good return losses, have very different coupling coefficients. In fact, for an isoline of 50Ω image impedance, the coupling coefficient can vary between 0 and 1. The wider the elements are, the lower the coupling coefficient is.

The coupling coefficient is closely related to the phase response in (4), since both parameters depend on the ratio between the mode characteristic impedances: $\frac{Z_{0e}}{Z_{0o}}$. Then, the phase response can be controlled independently of the return losses. Since the four studied designs have been chosen to have the same phase at 5.4 GHz, θ must be the same for the four designs at that frequency. Moreover, if low frequency

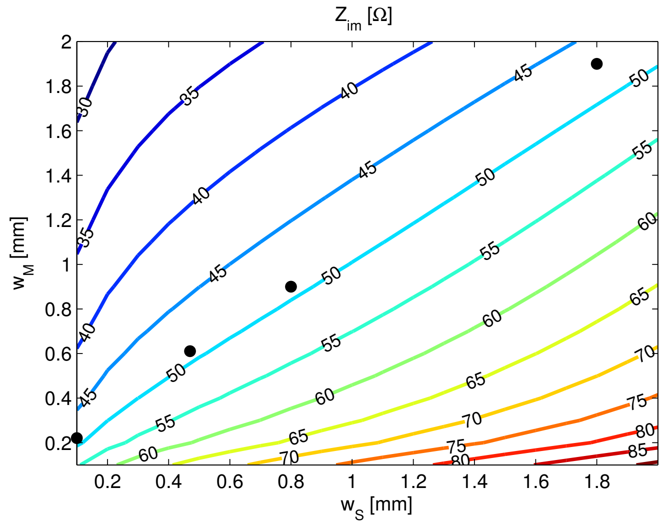


Fig. 4. Isolines of the image impedance of the coupled section at 5.4 GHz. The designs of Table I are marked with black dots.

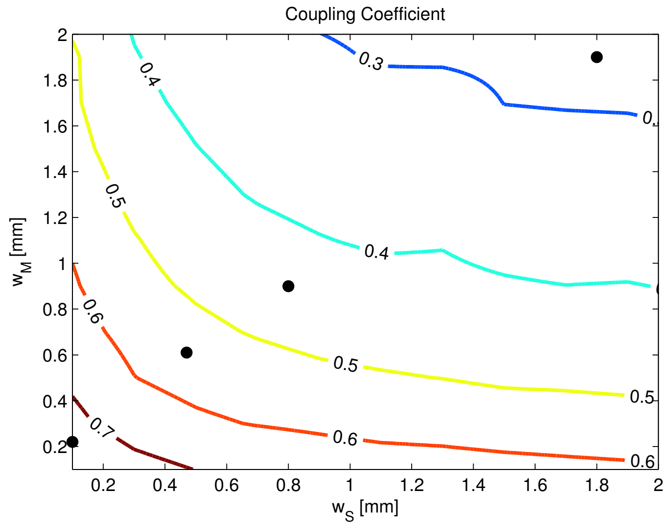


Fig. 5. Isolines of the coupling coefficient k of the coupled section at 5.4 GHz. The designs of Table I are marked with black dots.

dispersion in the effective permittivities of the modes is assumed, the only parameter that is different in the phase of the four designs at any frequency is $\frac{Z_{0e}}{Z_{0o}}$. According to (4), the phase response becomes independent of $\frac{Z_{0e}}{Z_{0o}}$ when $\theta = n\pi$. Therefore, it is expected that the phase factor of the four designs coincides at those frequencies at which θ=$n\pi/2$ with a value of ϕ=$n\pi$ and differs out of them according to the coupling coefficient or $\frac{Z_{0e}}{Z_{0o}}$.

Fig. 6 shows the comparison of the phase factor for the four studied cases. Measurements are only available for the Design 2. As predicted, the phase response becomes more linear for lower values of the coupling coefficient. This degree of freedom in the linearity between the points of phase factor ϕ=$n\pi$ allows certain control to set the required phase shifts between elements in a potential series-fed array at two different frequencies, useful for multiband applications. It must be clarified that for the Design 4, the resonance frequency f_r has not been obtained at 5.4 GHz, that is why there is a shift with respect to the rest of the curves at ϕ=$n\pi$. The strip and slot are made resonant at f_r by making l_S=l_M=$\lambda/4$. When the widths increase, the need of a length correction, due to the slot and strip end and junction effects, is higher. However, the perfect tuning of the resonance frequency has not been the purpose here. For that reason, this correction has not been carried out and is the main cause of the shift among the curves.

On the other hand, the change in the dimensions of the structure modifies the amount of radiation losses. Fig. 7 shows the ratio of the power radiated by and dissipated in the element and the input power for the four designs. This ratio is a good approximation of the radiation efficiency. The element starts radiating more efficiently from the resonance frequency of the slot (frequency at which $l_S = \lambda/2$). At lower frequencies, all the elements radiate practically the same low amount. As the widths of the strip and slot are made higher, an interesting increase in the radiation efficiency can be observed, but only

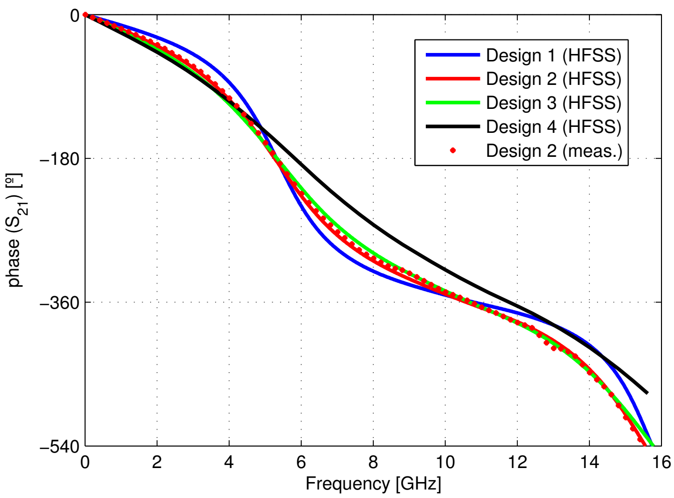


Fig. 6. Comparison of the phase response for the four designs of Table I.

between the two frequencies at which the slot length is $\lambda/2$ and $3\lambda/2$, which are also the frequencies at which the phase factor is fixed for all the designs. The undesired variation in the measurement of Design 2 is mainly due to the connector effects, since a TRL calibration kit was not built. Anyway, it is evident that the radiation efficiency between two slot resonances can be control to some extent by the widths of the elements. In this case, a variation between 15 and 40 % of radiation efficiency has been found between the four designs over the frequency band in which $l_S \in [\lambda/2, 3\lambda/2]$. In a series-fed array, the control of the radiation efficiency at each element will determine the excitation amplitudes, and thus the radiation pattern and the side-lobe level.

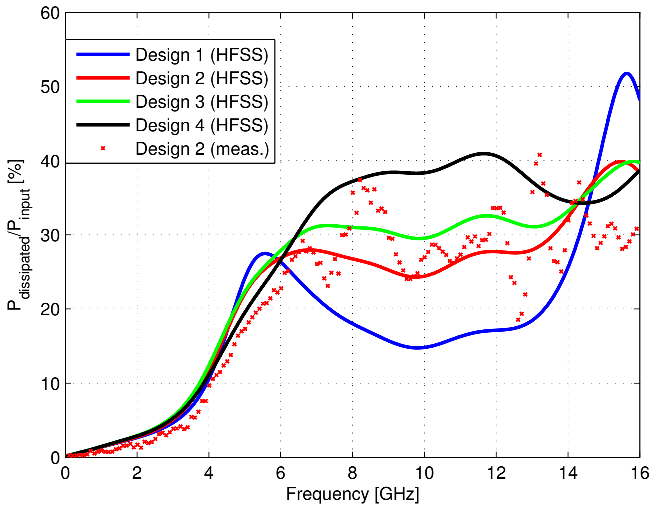


Fig. 7. Ratio of the power radiated by, and dissipated in, the element and the input power for the four designs of Table I.

IV. Conclusion

Four different designs of the complementary strip-slot radiating element have been studied, in order to show the influence of the geometric parameters in the matching, phase response and radiation efficiency. The four structures were properly designed to be broadly matched and have the same phase factor at a certain frequency (same slot resonance frequency). However, the widths were chosen very different, in order to highlight that the complementary strip-slot element is not only able to provide broad matching but also allows the control of the radiation efficiency or the phase response.

For good matching, the strip and slot widths must be chosen to make the microstrip-slotline coupling structure have an impedance level of $50\,\Omega$, as the feed microstrip line. Moreover, the strip and slot lengths must be designed so that they have the same electrical length, in order to cancel out the resonant behaviour of the image impedance. With these two conditions, ultra broad impedance matching is achieved. However, four variables are available to satisfy broad matching: w_M, w_S, l_M and l_S, therefore two extra degrees of freedom are available. The width pair also determines the linearity of the phase response or the radiation efficiency in the region between two slot resonances. The length pair sets the value of the structure phase factor at a certain frequency. Therefore, by taking advantage of these two degrees of freedom, some control in the phase and radiation efficiency is possible. For example, if it is desired to build an array with a low number of elements, the widths of the elements must be chosen large, in order to radiate all the power with few elements. On the contrary, if the directivity is to be maximised, the element must be chosen very thin, in order to radiate low power amount and make the antenna aperture longer by placing many elements. In this way, this design freedom in the complementary strip-slot can be exploited for the construction of competitive broadly-matched series-fed arrays in which certain flexibility in the magnitude and the phase of the excitations is essential.

Acknowledgment

This work was supported by the Spanish Ministerio de Ciencia e Innovación (Programa Consolider-Ingenio 2010) under Grant CSD2008-00066, EMET, and by the Junta de Andalucía (Spain) under Grant P10-TIC-6883.

References

[1] E. Abdo-Sánchez, J.E. Page, T.M. Martín-Guerrero, J. Esteban, and C. Camacho-Peñalosa, "Planar Broadband Slot Radiating Element Based on Microstrip-Slot Coupling for Series-fed Arrays," *IEEE Trans. Antennas Propag.*, vol. 60, no. 12, pp. 6037–6042, Dec. 2012.

[2] E. Abdo-Sánchez, J. Esteban, T.M. Martín-Guerrero, J.E. Page, and C. Camacho-Peñalosa, "Microstrip series-fed array based on the strip-slot element," in *Proc. 7th Eur. Conf. Antennas and Propagation*, Gothenburg, Sweden, Apr. 2013, pp. 1447–1450.

[3] E. Abdo-Sánchez, J. Esteban, T.M. Martín-Guerrero, C. Camacho-Peñalosa, and P.S. Hall, "A Novel Planar Log-Periodic Array Based on the Wideband Complementary Strip-Slot Element," *IEEE Trans. Antennas Propag.*, vol. 62, no. 11, pp. 1–9, Nov. 2014.

[4] E. Abdo-Sánchez, C. Camacho-Peñalosa, T.M. Martín-Guerrero, J. Esteban, and J.E. Page, "Multiband circularly-polarized planar array based on the complementary strip-slot element," in *Proc. Eur. Microwave Conference*, Nuremberg, Germany, Oct. 2013, pp. 652–655.

[5] R. Hoffmann and J. Siegl, "Microstrip-Slot Coupler Design-Part I: S-Parameters of Uncompensated and Compensated Couplers," *IEEE Trans. Microwave Theory Tech.*, vol. 30, no. 8, pp. 1205–1210, Aug. 1982.