Design and Characterization of Modified Comb Patch Antennas

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ABSTRACT This work deals with the proposal of a novel type of microstrip antenna, called MCPA the modified comb patch antenna. The proposed antennas is composed of \( n \) parallel conductors, fed by a common microstrip. A dedicated mathematical framework, based on the multiconductors transmission line formalism, is proposed for antenna analysis and design. The analytical model is numerically validated with full-wave simulations, resulting in a 5% error in the predicted resonant patch length. A numerical study of antenna matching, size, radiation performance is carried out. The matching increases as the number of conductors increases, whilst gain of comb antennas made of \( n \) conductors are about half dB higher than the equivalent full patch counterpart. Then, an eighty conductors was realized and measured to assess the frequency response of the antenna, as well as its radiation performances. An error of 1% between the predicted and measured value resonance frequency was observed. A difference of about 0.67 dB was found for the measured maximum antenna gain, with respect to the simulated one. The proposed antenna design is appealing for printed electronics and wearable, on-textile applications.

INDEX TERMS Comb antenna, multi-conductor, wireless body area network.

I. INTRODUCTION The research interest in on-body wireless communication systems in recent years, is increasing due to the ever-growing relevance of personal electronic devices in everyday life and activities, such as personal communication [1], healthcare [2], identification, tracking and monitoring [3] of biometrics and biomedical parameters [4], [5]. In this framework, the electromagnetic engineering community is facing the challenge of designing and realizing innovative and low-cost antennas, which can be easily integrated in body area networks (BAN), or worn and integrated into wearable systems [6]. For these applications, microstrip antennas (e.g., patch) are mostly used, thanks to advantages such as ease of construction, low cost and a high adaptability. From an electromagnetic point of view, standard microstrip antennas have the relevant drawbacks of exhibiting low gain, reduced power capability and inherently narrowband [7]. To improve the performances of microstrip antennas, a pivotal requirement is to investigate new solutions, such as the use of multi-conductor antennas [8] or antenna arrays [9]. In this work, we propose a novel multi-conductor antenna. Therefore, when a new configuration is proposed, a robust, reliable and fast methodology for the study and optimization must be developed.
This work proposes a modified comb patch antenna (MCPA), i.e. a patch composed of \( n \) parallel conductors, fed by a common microstrip line, as shown in Fig. 1. The conductors branch out from the feeding line perpendicularly in one direction, being symmetric with respect to the feeding line. The configuration is proposed to ensure that all the conductors are all placed along a given direction, thus determining a uniform current density flowing along with them. In addition, these requirements guarantee that the antenna characteristic parameters could be derived easily and efficiently, while providing a pure polarization of the antenna.

Fig. 2 shows the novel configuration proposed. The MCPA is an antenna topology which strongly differs from previous microstrip comb antennas. Indeed, the most commonly investigated, designed and characterized multi-conductor antenna are comb antennas, having conductors branching out in to opposite directions from a common microstrip feeding line or with a scope of 30° or 45° angle with respect to the feeding line [10]–[12]. Recently, a comb-shaped microstrip patch Antenna (CSMPA) was proposed [13]. Flexible broadband monopoles in comb arrangement, working in the frequency range 1.7-2.68 GHz, were studied [14]. The size of the lateral conductors leads to an increase in bandwidth (BW) (up to a 44.75% BW at -10 dB) and allows to slightly reduce the size of the antenna. The comb geometry has also been used for the design of arrays working at 2.45 GHz, reaching a maximum gain of 12.85 dB [15]. Comb antennas are mainly design in microstrip tecnology, or developed as a planar inverted F antenna (PIFA) for the applications in the UHF band [10]. Furthermore, metasurface structures can be obtained with linear multi-conductors [18]. Aslo, single and double side comb antennas for radar applications were proposed [19]. In [20], an EBG comb structure was proposed for enhancing the performance of antennas in WiMax band. A summary of the main literature about comb antennas is provided in Tab. 1.

To summarize, the MCPA differs from previous comb antennas in the spatial arrangement of conductors. However, the difference between the proposed MCPA and the designs available in the literature is also the feeding strategy. Indeed, the use of a coaxial probe is the most used approach [14], [21]. The coaxial feed can present drawbacks from the design point of view and increases the technological criticalities for BAN applications (Fig. 1). The selection of the insertion point must be performed to ensure a homogeneous current flow in a single direction, thus raising difficulties in the design and matching procedures, hence calling for effective design strategies [10]–[14], [18]–[20]. Indeed, the traditional methods used for the design of microstrip patch antennas are not suitable for comb-shaped antennas. In this framework, comb antennas have been analyzed by developing models based on variational methods [22], modal analysis [23], or simple, but less accurate transmission line models (TLM) [24]–[26] or computationally costly, high-fidelity full-wave analysis (FW) [8], [27]–[33]. All these design strategies are often complemented by numerical optimization, heuristic or machine learning-based methods [34]–[38]. As a matter of fact, when designing the aforementioned comb antennas usually, the design strategy is mostly heuristic and driven by extensive numerical simulations [21], reflecting a poor theoretical effort [11], [13], thus highlighting the lack of a complete, exhaustive and simple model for the analysis of comb antennas. Therefore, there is a knowledge gap to be filed with a thorough and robust electromagnetics and antenna engineering analysis.

To the best of the authors’ knowledge, the potentialities of this kind of MCPA, multi-conductor structure, have never been investigated. Therefore, in this work, a specific, original mathematical framework for designing the MCPA, of the type shown in Fig. 2, is developed. The proposed MCPA has the advantage of being cost-effective solution for communications and BAN applications. Furthermore, with the proposed design

**TABLE 1. Resume of the state of the art of comb antennas.**

| Comb-shaped Antennas | Gain (dB) | Frequency (GHz) | \( S_{11} \) (dB) |
|----------------------|----------|-----------------|------------------|
| Comb array [13]      | 12.85    | 2.4             | -34.88           |
| PIFA [16]            | 3.3 dB   | 0.915           | -20              |
| Linear array [17]    | -        | 28              | -24              |
| Monopole [14]        | -        | 2.15            | -48              |
| Metasurface [18]     | 5 dB     | 2.56            | < -10            |
| Single comb [19]     | 6.05     | 5               | -30              |
| EBG structure [20]   | -        | 2.65 - 7        | -20              |

**FIGURE 1.** Concept, possible manufacturing approach and potential applications of the modified comb patch antenna MPCA.

**FIGURE 2.** Geometry of a multi-conductor patch antenna. The feed is a 50 \( \Omega \) microstrip line and the conductors are parallel microstrip segments having width \( w_c \), spaced of \( w_g \) with a length \( L_{\text{patch}} \), the current flows uniformly along \( l \) in the z-direction.
structure and the developed mathematical framework, the control on the intrinsic parasitic and matching capacitance makes this structure very flexible to be tuned.

The MCPA, shown in Fig. 2, is analyzed with a dedicated mathematical model aimed at the computation of the patch dimensions, to be matched and tuned in a given frequency range. The mutual coupling effects between the conductors, and the influence of the gaps are included in the thus allowing a realistic analytic model. The proposed model could be used to analyze and design MCPA printed or textile-based antennas (Fig. 1). The model is validated against numerical, full-wave simulations and experimental measurements. In this work, the selected frequency range is the GSM band (890-915 MHz). The proposed design and methodology is general and can be extended to other bands.

The paper is organized as follows, in Sect. II the set of equations for the design is presented. In Sect. III, both numerical and experimental validation are provided; then, in Sect. IV, the findings are presented and finally in Sect. V conclusion and discussion are given.

II. THE MODEL

The novel MCPA, shown in Fig. 2, can be analyzed as an $n$ conductor transmission line system, supporting quasi-TEM modes [39], since the current flows uniformly along the $l$-direction. By applying the transmission line equations [Eq. (22) in the Appendix], the spatial variation of the voltage and current vectors ($1 \times n$) can be calculated. The equations must be manipulated to include the patch length ($L_{\text{patch}}$) in the expression for deriving the mathematical framework for design and analysis. With straightforward algebraic derivation, the coupled telegraphers’ equations in the following second-order equation [40]

$$- \frac{d^2 I(l)}{dl^2} + \beta_k^2 L_{\text{patch}} C_n I(l) = 0$$

(1)

for the $k$-th conductor, the solution to the second-order differential equation is in the following form

$$I(l) = I_k e^{\beta_k l}$$

(2)

where $\beta_k$ are the eigenvalues of the matrix $\beta_0^2 L_{\text{patch}} C_n$ (with $k = 1, \ldots, n$). $L$ and $C$ are, respectively, the static, normalized inductance and capacitance matrices of the structure shown in Fig. 2. In other words, $C_n = \frac{1}{\varepsilon_0} \cdot C$ and $L_n = \frac{1}{\mu_0} \cdot L$, respectively, being $C$ and $L$ the capacitances and inductances of the system, whilst $\varepsilon_0$ is the vacuum permittivity, in (Fm$^{-1}$) and $\mu_0$ is the vacuum magnetic permeability, in (Hm$^{-1}$).

In this case, it follows that, the propagation constant can be written as

$$\beta_k = \beta_0 \sqrt{\lambda_k}$$

(3)

where $\lambda_k$ are the eigenvalues of the matrix $L_n$ and $C_n$, with size $n \times n$. The term $I_k$ indicates the eigenvector corresponding to the eigenvalue $\lambda_k$.

Then by rewriting the current in the form of a steady waves, we can impose the known current value for $l = L_{\text{patch}}$ and find that the voltage and current distribution as

$$V(l) = \int_0^{L_{\text{patch}}} \frac{C_n}{\varepsilon_0} \sum_{k=1}^n \left( \cos (\beta_0 \sqrt{\lambda_k} (l - L_{\text{patch}})) + j \sin (\beta_0 \sqrt{\lambda_k} (l - L_{\text{patch}})) \right) \frac{dS}{\lambda_k}$$

(4)

where $I_k^+ + I_k^- = I_{Ak}(L_{\text{patch}})$ and $I_k^- - I_k^+ = B_k$, being $I_k^+$ and $I_k^-$ unknown complex amplitudes.

Eqs. (4) are essential for deriving a procedure suitable for sizing the modified comb patch antenna, i.e. to find the values of $W_{\text{patch}}$ and $L_{\text{patch}}$ for the desired working frequency. To this aim, it is necessary to know the system capacitance and to further investigate the MCPA characteristic impedance.

A. DERIVATION OF L AND C PARAMETERS

In order to solve system (4) have to be derived the characteristic parameters $L$ and $C$ (see Fig. 2). The $n$ conductors are placed between air and a dielectric substrate. Firstly the computation of the inductances in the MCPA is required. Indeed, the evaluation of the inductance in the system is straightforward once $C$ is known [41]: $L$ can be derived as the inverse of $C$ in air. Therefore, the capacitance matrix is derived from the total electric energy ($W$) stored in the structure, by a FD computational method [42]

$$C = \int_0^{L_{\text{patch}}} \int_0^W \varepsilon |E|^2 dS$$

(5)

where $\varepsilon$ is the dielectric permittivity of the medium and $E$ is the electric field (Vm$^{-1}$).

To find the capacitance matrix of the system of $n$ conductors, in this work, the calculation is performed for simplicity moving from the analysis of a four-conductor structure. Only explicit structures with equal strips, equal external gaps, and internal gaps equal or different to external gaps were considered (with reference to Fig. 3(a)), while taking into account the mutual-coupling between the MCPA elements.

Once the energy $W$ is evaluated with the FD method, as previously described, it is found that,

$$W_k = \frac{1}{2} [v_k]^T [C] [v_k]$$

(6)

with $v_k$ the chosen voltage vector and $C$ is the capacity matrix [The Eq. (39) in the Appendix]. The Eq. (6) is a linear system of equations. Since it has six unknowns capacities, it requires six configurations. Therefore, we write a linear system of six equations in the six unknown capacities

$$A \cdot C = W$$

(7)
where $A$ is the matrix which has as columns the six independent voltage (eigen-)vectors ($v_k$, with $k = 1, \ldots, 6$). By using the Symbolic Toolbox of Matlab 2019b (The MathWork Inc., MA, USA) it is possible to derive a set of eigen-vectors which ensures that $A$ is well conditioned and symmetric. It is worth to point out that two of the eigenvectors terms are known, whilst the other two are symmetrical. The values of the four-line structure were then used as an approximation to obtain the matrix $C$ of a $n$-conductors system Fig. 3(b). The internal capacitances were approximated by the value of $C_i$ of the 4-conductors, and all mutual internal capacitances were approximated with $C_{mj}, C_{m2}$ and $C_{m3}$. Similarly, the two external capacitances were approximated with $C_m$, and the two mutual external capacitances were approximated by $C_{me}$. Therefore, the final matrix, for the $n$-conductor system is

$$
C_n = \begin{bmatrix}
C_{11} & 0 & \ldots & 0 \\
0 & C_{22} & \ldots & 0 \\
\vdots & \ldots & C_{33} & 0 \\
0 & \ldots & 0 & C_{44}
\end{bmatrix}
$$

When the coupling between conductors is neglected, the real-valued impedance matrix can be used to find the resonant $L_{\text{patch}}$ value. However, the operation under this assumption would lead to over/underestimated results. Therefore, in this work, a more in-depth analysis is performed.

Under the approximation of neglecting the coupling between the conductors, the results are limiting and less accurate for the analysis and design of the MCPA. Therefore, the proposed mathematical framework for the MCPA can be complicated by accounting for the coupling between conductors in the evaluation of the irradiation admittance. It is worth highlighting that, in this work, the radiation impedance is not calculated using the classic formula described by Eq. (11), but, instead, it is computed by comparing the total available power with the radiated field. Therefore, the radiation admittance $Y_{\text{irr}}$ is a full, complex matrix which accounts for the coupling between all conductors. Hence, the total power in the system can be written as follows [44], [45]

$$
P_c = \frac{1}{2} V_n^{*} Y V_n = \sum_{p,m=1}^{n} V_{n,p} V_{n,m}^{*} Y_{p,m}^{*}
$$

where $V_n$ is the voltage on the single conductor.
The electric field can be written as a piecewise constant function, as shown in Fig. 4, in the MCPA system, for \( n \) lines, having width \( T = \frac{W}{n} \) and with \( x_p = pT, x_m = mT \). Therefore, to ease the analytical derivation, we can perform the Fourier transform of the electric field and get

\[
\tilde{E} = iT \sum_{p=1}^{n} V_{n,p} \text{sinc}(\frac{T}{2} k_x) \text{sinc}(\frac{\Lambda T}{2} k_y) e^{ik_x x_p}
\]

\[
\tilde{E}^* = iT \sum_{p=1}^{n} V_{n,m}^* \text{sinc}(\frac{T}{2} k_x) \text{sinc}(\frac{\Lambda T}{2} k_y) e^{-ik_x x_m}
\]

where \( \hat{y} \) is the unit vector along the y-direction, as shown in Fig. 4, and \( k_x \) and \( k_y \) are the components of the propagation vectors along \( \hat{x} \) and \( \hat{y} \). Hence, by using Eq. (14), we can rewrite Eq. (13) as

\[
P_C = \frac{1}{8\pi^2 \eta} \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \left| k^2 \tilde{E} \right|^2 + \left| k^2 \tilde{E}^* \right|^2 \frac{dk_x d\eta}{k_x^2} \]

(15)

Where \( \eta \) is the free-space intrinsic impedance. To derive \( |\tilde{E}|^2 \), we multiply Eq. (14) together

\[
|\tilde{E}|^2 = T^2 \text{sinc}^2(\frac{T}{2} k_x) \text{sinc}^2(\Delta l^2 k_y) \sum_{p=1}^{n} V_{n,p} V_{n,m}^* e^{ik_x x_p}
\]

(16)

where \( x_d = x_p - x_m \); Whereas \( M(k_x, k_y) \) is defined as

\[
M(k_x, k_y) = \text{sinc}(\frac{T}{2} k_x) \text{sinc}(\Delta l^2 k_y)
\]

(17)

That is the spectrum of a 2D constant truncated spatial function (Fig. 4).

Finally, by substituting (16) into (15), we find

\[
P_C = \frac{1}{8\pi^2 \eta} \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \left\{ k_x^2 + |k_y|^2 \right\} \left\{ \cos k_x(x_d) + j \sin k_x(x_p - x_m) \right\} \frac{dk_x d\eta}{k_x^2}
\]

(18)

Finally, by comparing (13) and (18), we can find the complex admittance of the MCPA antennas as

\[
Y_{p,m} = \frac{1}{4\pi^2 \eta} \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \left\{ k_x^2 + |k_y|^2 \right\} M(k_x, k_y)
\]

(19)

As detailed in the Appendix, by manipulating Eq. (19), the radiation admittance can be divided into real and imaginary part to get the closed form expression for both the real part, so that

\[
R_e(Y_{p,m}) = \frac{T^2}{\pi^2 \eta} \int_{0}^{\infty} (k^2 - k_o^2)^2 M(k_x, k_y) \left\{ \cos k_x(x_d) + \int_{0}^{k} \frac{dk_x d\eta}{k_x^2} \right\}
\]

(20)

The imaginary part of the admittance is also found in closed-form by using the Rhodes’ relationships (see Appendix for details) [46]

\[
I_m(Y_{p,m}) = \frac{T^2}{\pi^2 \eta} \int_{0}^{k} (k^2 - k_o^2) \sin^2(\frac{\Delta l}{2} k_y)
\]

\[
\times \cos k_x(x_d) \left\{ \int_{0}^{\infty} \int_{-\infty}^{+\infty} \frac{dk_x d\eta}{\sqrt{k_x^2 - k_o^2}} \cos k_x(x_d) \right\}
\]

\[
\times \left\{ \int_{0}^{+\infty} \sin^2(\frac{\Delta l}{2} k_y) \frac{dk_x d\eta}{\sqrt{k_x^2 - k_o^2}} \right\}
\]

(21)

The analytical expressions for both real and imaginary parts of the total power is manipulated and solved by Mathematica v. 10.0 (Wolfram, GE). After having derived the closed-form expressions, the computation was carried out with Matlab Symbolic Toolbox and the impedance value of the single conductor element is found to fill the admittance matrix.

In order to design the MCPA antenna we need to find the set of length (\( L_{\text{patch}} \)) and width (\( W_{\text{patch}} \)) of the patch to work at the desired frequency range, under resonance conditions. To this aim, by imposing the imaginary part of the admittance to be zero and by using the bisection iterative method [47] the effective length of the patch which satisfies this condition can be found numerically.

### III. VALIDATION

#### A. NUMERICAL VALIDATION

In order to validate if the proposed model is suitable for designing a MCPA antenna, we compare our Matlab code with numerical, full-wave simulations performed in CST (Simulia, 3DS, GE). As shown in Fig. 5, we considered two multi-conductor comb patch antennas with \( n = 20 \) and \( n = 80 \) conductors. The substrate is a RT/Duroid 5880 with thickness \( h_{\text{sub}} = 1.5 \) mm and nominal \( \epsilon_r = 2.2 \). The proposed design is also compared to the standard rectangular patch antenna,
as control and reference case, as shown in Fig. 5. The antennas are fed by a microstrip line connected to a 223-CON-SMA-EDGE-S. The sizes of the patches for the validation are reported in Tab. 2.

The frequency range used in the simulation is 800-1000 MHz. We used as figures of merits the MCPA antenna size ($L_{\text{patch}} \times W_{\text{patch}}$), the characteristic impedance and the return loss ($|S_{11}|$). With this further comparison, the pros and cons of the MCPA antenna performance are critically analyzed to gain further insight into the full exploitation of its potential for telecommunications or on-body applications.

### IV. RESULTS

#### A. MODEL VALIDATION & ANALYSIS

The parameters used in the numerical simulations of the reference patch and the MCPA with a variable number of conductors are reported in Tab. 2. We have compared the resonance length computed by the proposed mathematical framework and that obtained by empirical refinement on CST. The results are shown in Fig. 3. The values of the relative percentage error calculated assuming as reference the CST simulations are reported in Tab. 3. It can be noticed that the error increase as the number of conductors increases. Given that the width of the patch is constant, from the geometry shown in Fig. 2, the size of conductors decreases. Therefore, the width of each comb line and the size of the gap halves when the number of conductors is doubled. In this framework, the accuracy in the evaluation of the model parameters presents a reduction, which results in a fourfold error on the patch length. However, it must be noticed that the error, with respect to the ground through of the full-wave simulation, is below 5%, which can be considered as a valuable result.

#### B. PROTOTYPE FABRICATION AND EXPERIMENTAL CHARACTERIZATION

After having analyzed the proposed antenna topology, we selected a case study to manufacture and characterize. The LPKF ProtoMat E44 (LPKF Laser & Electronics AG, GE) milling machine was used for manufacturing the $n = 80$ conductors MCPA.

The return loss of the MCPA was measured with Haelwtt-Packard 8720C vector network analyzer (VNA), as shown in Fig. 6. The antenna gain was measured with the three antenna method, by using the $n = 80$ conductors MCPA as antenna under test, together with the LP0410 PCB log-periodic directional antenna (National Instruments, with 5-6 dB of gain, working frequencies 400-1000 MHz) and the log-periodic USLP9142 (D-69250, Schwarzbeck, Schonau GE) antenna working between 800-1000 MHz, presenting a 3.7-6.8 dB gain [48]. The antennas were placed at a distance $d_A$ of 3.3 m, being $d_A \gg 2 \cdot W_{\text{patch}}^2 / \lambda$. The same VNA is used. The radiation patterns for $\phi = 0^\circ$ and $\phi = 90^\circ$ were acquired at $0^\circ$, $45^\circ$, $90^\circ$, $180^\circ$ angles.

Given the promising results, we have numerically investigated how the variation of the conductor numbers influences the antenna performances, in terms of matching, gain and directivity. The simulation results are reported in Tab. 4. The increase in $n$ results in a non-linear variation of the patch length, up to the asymptotic limit of a classic, full patch, as shown in Fig. 7. On the other hand, the matching of the MCPA decreases in a hyperbolic way (see Tab. 4). We have compared this figure of merit by observing the $|S_{11}|$ vs.

![Figure 5](https://example.com/figure5.png)

**FIGURE 5.** a) Rendered geometry of the reference patch. b) Rendered geometry of the 20 conductors comb patch antenna. c) Rendered geometry of the 80 conductors comb patch antenna.

![Figure 6](https://example.com/figure6.png)

**FIGURE 6.** a) Experimental setup for the return loss ($|S_{11}|$) measurement. b) Magnified image of the milled multi-conductor patch antenna.

| Parameter  | $n = 20$ | $n = 80$ | Full Patch |
|------------|----------|----------|------------|
| $W_{\text{patch}}$ (mm) | 150 | 150 | 131 |
| $L_{\text{patch}}$ (mm) | 106.8 | 110.6 | 108.1 |
| $W_2$ (mm) | 5 | 1.25 | - |
| $W_3$ (mm) | 2.6 | 0.65 | - |
| $h_{\text{sub}}$ (mm) | 1.575 | 1.575 | 1.575 |
| $L_{\text{ mA}}$ (mm) | 30 | 30 | 30 |

**TABLE 2.** Properties and geometries of the size of the reference patch and two MCPA. All dimensions are in millimeters.

| Patch | CST | Matlab | Error |
|-------|-----|--------|-------|
| $n = 20$ | 106.8 mm | 105 mm | 1.68% |
| $n = 80$ | 110.6 mm | 115.02 mm | 4.4% |

**TABLE 3.** Comparison of the calculated lengths for the three different patches.
TABLE 4. Comparison of antenna parameters for different number of conductors.

| Conductors | $|S_{11}|$ (dB) | -10 dB BW (%) | Gain (dB) | Directivity (dB) | Efficiency (%) | $L_{patch}$ (mm) |
|------------|----------------|---------------|-----------|-----------------|---------------|-----------------|
| $N = 4$    | -21.9          | 0.66          | 6.44      | 7.73            | 74            | 92.8            |
| $N = 12$   | -19.70         | 0.71          | 6.55      | 7.71            | 77            | 94              |
| $N = 20$   | -28.6          | 0.73          | 6.65      | 7.58            | 78            | 106.8           |
| $N = 40$   | -28.88         | 0.71          | 6.67      | 7.68            | 79.3          | 110             |
| $N = 80$   | -35.75         | 0.70          | 6.37      | 7.75            | 80            | 110.6           |
| $N = 160$  | -42.03         | 0.67          | 6.68      | 7.72            | 80            | 112             |
| $N \to \infty$ | -41.42 | 0.64 | 6.26 | 7.6 | 73 | 108.1 |

FIGURE 7. Simulated patch lengths (mm) as a function of conductor numbers (n).

FIGURE 8. Comparison of simulated $S_{11}$ parameter for the two MCPA and the reference patch. Dimensions are shown in Tab. 2.

B. EXPERIMENTAL CHARACTERIZATION

Among the designed MCPA, we selected the 80-conductors antenna as the best candidate for performing the experimental characterization. The tested prototype is shown in Fig. 6.b. The comparison of the measurements and the simulation results are reported in Fig. 10. From Fig. 10, we found a difference of 0.7% between the simulations and the measured performances, guaranteeing for the prototype a -10 dB BW of 1%. The resonance of the realized MCPA occurs at 910.25 MHz, instead of 901.3 MHz, with a 1.1% difference from the simulated value. To demonstrate that the frequency shift is not due to an error in the proposed mathematical model for the MCPA design, we performed an additional numerical simulation to elucidate the fact that the substrate batch used therein presented a large variability from the nominal value of the dielectric permittivity ($\varepsilon_r = 2.2$). Therefore, in Fig. 10, the simulated $|S_{11}|$ for the MCPA with 80 conductors varying $\varepsilon_r$ from 2.08 to 2.34, according to the variability...
reported in the datasheet. By comparing the measured return loss shown in Fig. 10 and the curves in Fig. 10, we can noticed that the actual relative dielectric permittivity of the substrate is 2.12. These findings are supported by the analysis of the manufactured antenna geometry, whose geometric features are respected well within the mechanical tolerances admissible in the GSM band, as shown in Fig. 11. Finally, the radiation performances of the manufactured MPCA were measured and the findings, for comparison with the simulations, are reported in Fig. 12. The measured maximum antenna gain is 5.75 dB, which differs by -0.63 dB from the simulated one. For angles higher than 0°, the maximum difference between the measured value and the simulated one is about 1.75 dB.

These findings indicate that our model can effectively be used to design a modified comb patch antenna with satisfactory performances in terms of tuning, matching and radiation.

V. CONCLUSION AND DISCUSSIONS
This work dealt with the design, realization and characterization of modified comb patch antennas. We developed and provided design equations in closed-form for sizing a patch composed of \( n \) parallel conductors fed by a common microstrip line. By comparison with full-wave numerical simulations, we found that our model can predict the resonance length of the patch with a maximum error of 4.4%. The proposed topology is a novel arrangement and present the advantage of adding degrees of freedom for tuning the antenna performances. We investigated numerically the effect of increasing the number of conductors on antenna behavior. We realized a prototype of a MCPA with 80 conductors and experimentally characterized its performances. Concerning predicted curves, a 1.1% shift in the resonance frequency was found. This effect can be ascribed to a 3.7% variation in the relative dielectric permittivity of the substrate.

Given that, compared to a standard patch antenna, the MCPA is low-cost, easy to manufacture, while presenting a reduced conductor area, it also demonstrated increased gain, slightly larger bandwidth, enhanced radiation efficiency and larger directivity, it is appealing and of potential interest for innovative applications. In particular, the possibility of obtaining matching, tuning and radiation performances comparable (or even higher) than a standard patch while using a lower amount of conductor material could be relevant for realizing communication systems in a sustainable, cost-effective way through printed electronics manufacturing techniques [49]. Furthermore, the MCPA configuration is interesting for designing textile-based antennas. Indeed, the antenna design shown in Fig. 2 is very appealing for BAN since the MCPA can be an electronic device easily integrable in garments. Indeed, we forecast that the proposed MCPA can be implemented by using conductive fibers in the arrangement shown in Fig. 1, ensuring good radiation performances in an on-body scenario. In fact, by using conductive textiles the MPCA geometry could be easily weaved. To date, several examples of textile, wearable antennas were proposed, such as the woven textiles antennas from [50], the electro-textile antennas made up of a mix of metallic and fabric filaments [51], the RFID tag manufactured with a sewing machine directly on clothes [52] or antennas synthesized with hydrophobic materials were studied [53], [54]. However, the complexity of the embroidery thread of the aforementioned antennas does not allow to perform a rigorous analysis...
(e.g., with TLM or FW approach) or an effective, easy and accurate design. The MCPA could be a cost-effective solution with a reasonable trade-off between performances and design complexity. Therefore, future works may deal with the design, realization and characterization of textile-based MCPA.

APPENDIX

A. DERIVATION OF MAIN EQUATIONS

The analysis and design equations of the MCPA are derived from the transmission lines equations:

\[-\frac{dV(l)}{dl} = jo\mu_0 L_n I(l)\]
\[-\frac{dI(l)}{dl} = jo\epsilon_0 C_n V(l)\]  \hspace{1cm} (22)

where the \( L \) and \( C \) are, respectively, the static, normalized inductance and capacitance matrix of the system shown in Fig. 2. In other words, \( C_n = \frac{1}{\epsilon_0} \cdot C \) and \( L_n = \frac{1}{\mu_0} \cdot L \), respectively, being \( C \) and \( L \) the capacitances and inductances of the system, whilst \( \epsilon_0 \) the vacuum permittivity, in \( \text{F/m} \), and \( \mu_0 \) the vacuum magnetic permeability, in \( \text{H/m} \). In the MCPA, the dielectric is not homogeneous, thus implying that quasi-TEM modes are supported [39]. Therefore, moving from the system (22), by taking the first derivative, with respect to the local space variable \( l \), of the second equation of the system (22), we obtain,

\[-\frac{d^2I(l)}{dl^2} = jo\epsilon_0 C_n \frac{dV(l)}{dl}\]  \hspace{1cm} (23)

By substituting Eq. (22) in Eq. (23), we get:

\[-\frac{d^2I(l)}{dl^2} + jo\epsilon_0 C_n L_n I(l) = 0\]  \hspace{1cm} (24)

Then, Eq. (24) can be written as follows:

\[-\frac{d^2I(l)}{dl^2} + \beta^2 I(l) = 0\]  \hspace{1cm} (25)

By introducing the following relationships, where \( f \) is the working frequency and \( c_0 \) is the speed of light in vacuum,

\[\begin{align*}
\sqrt{\epsilon_0 \mu_0} &= \frac{1}{c_0} \\
\omega &= 2\pi f \\
\beta_0 &= \frac{2\pi f}{c_0}
\end{align*}\]  \hspace{1cm} (26)

then Eq. (22) turns into,

\[\begin{align*}
V(l) &= -\frac{1}{\omega \epsilon_0} C_n \frac{dI(l)}{dl} \\
-\frac{d^2I(l)}{dl^2} + \beta_0^2 I(l) &= 0
\end{align*}\]  \hspace{1cm} (27)

The solution to the second-order differential equation is in the following form:

\[I(l) = I_k e^{\beta_0 l}\]  \hspace{1cm} (28)

where \( \beta_k \) are the eigenvalues of the matrix \( \beta_0^2 L_n C_n \) (with \( k = 1, \ldots, n \)). In this case, it follows that,

\[\beta_k = \beta_0 \sqrt{\lambda_k}\]  \hspace{1cm} (29)

where \( \lambda_k \) (with \( k \) ranging from 1 to \( n \)) are the eigenvalues of the matrix \( L_n \) and \( C_n \), with size \( n \times n \), and the term \( I_k \) indicates the eigenvector corresponding to the eigenvalue \( \lambda_k \).

Therefore, the current in the MCPA antenna can be written in the following form,

\[I(l) = \sum_{k=1}^{n} \left[ I_k^+ e^{-j\beta_0 l} + I_k^- e^{j\beta_0 l} \right] L_k\]  \hspace{1cm} (30)

where \( I_k^+ \) and \( I_k^- \) are scalar with suitable values. Now, by expanding the exponential terms of the current, it follows,

\[I(l) = \sum_{k=1}^{n} \left[ I_k^+ \cos (\beta_0 l) - I_k^- \sin (\beta_0 l) \right] + j \sum_{k=1}^{n} \left[ I_k^+ \sin (\beta_0 l) + I_k^- \cos (\beta_0 l) \right] L_k\]  \hspace{1cm} (31)

If we group the sine and cosine terms, we get,

\[I(l) = \sum_{k=1}^{n} \left[ (\cos (\beta_0 l) I_k^+ + I_k^-) + j \sin (\beta_0 l) (I_k^+ - I_k^-) \right] L_k\]  \hspace{1cm} (32)

By knowing the voltage and current in \( l = L_{\text{patch}} \), shown in Fig. 2, we can write:

\[I(l) = \sum_{k=1}^{n} \left[ \cos (\beta_0 \sqrt{\lambda_k} (l - L_{\text{patch}})) \cdot [I_{Ak(a)}(L_{\text{patch}})] + j \sin (\beta_0 \sqrt{\lambda_k} (l - L_{\text{patch}})) \cdot [B_k] \right] L_k\]  \hspace{1cm} (33)

where,

\[I_k^+ + I_k^- = I_{Ak(a)}(L_{\text{patch}})\]  \hspace{1cm} (34)
\[I_k^- - I_k^+ = B_k\]  \hspace{1cm} (35)

Eq. (31) can be re-written in the steady form (for the \( k \)-th eigenvalue and the \( n \)-th conductor, i.e.,

\[I(l) = \sum_{k=1}^{n} \left[ \cos (\beta_0 \sqrt{\lambda_k} (l - L_{\text{patch}})) \cdot [I_{Ak(a)}(L_{\text{patch}})] + j \sin (\beta_0 \sqrt{\lambda_k} (l - L_{\text{patch}})) \cdot [B_k] \right] L_k\]  \hspace{1cm} (36)

\[\cdot [I_{Ak(a)}(L_{\text{patch}})] + j \sin (\beta_0 \sqrt{\lambda_k} (l - L_{\text{patch}})) \cdot [B_k] \right] L_k\]
### B. Capacitance Calculation

1) **Capacity Matrix Representation**

The basic structure considered for the calculation of the capacitances consists of four conductors. The matrix of the capacities found for four conductors consists of six elements, these six elements or describe the external, internal and mutual capacities between the conductors. The capacity matrix can be constructed as a block matrix, composed of four blocks defined as follows:

\[
C_1 = \begin{bmatrix}
C_e + C_{me} + C_{m2} + C_{m3} & -C_{me} \\
-C_{me} & C_e + C_{me} + C_{m2} + C_{m3}
\end{bmatrix}
\]

\[
C_2 = \begin{bmatrix}
-C_{m2} - C_{mi} \\
-C_{m3} - C_{m2}
\end{bmatrix},
C_3 = \begin{bmatrix}
-C_{m2} - C_{mi} \\
-C_{m3} - C_{m2}
\end{bmatrix}
\]

(37)

(38)

Being a diagonal matrix, the blocks \(C_4\) are equivalent to the block \(C_1\) with the elements of the diagonal reversed.

\[
C = \begin{bmatrix}
C_1 & C_2 \\
C_3 & C_4
\end{bmatrix}
\]

(39)

The structure with four conductors can be generalized to a structure with \(n\) conductors. In this case the matrix \(C\) can be written as:

\[
C_n = \begin{bmatrix}
C_{11} & 0 & \ldots & 0 \\
0 & C_{22} & \ldots & 0 \\
\vdots & \vdots & \ddots & \vdots \\
0 & 0 & \ldots & C_{nn}
\end{bmatrix}
\]

(40)

2) **Complete Derivation of Real and Complex Impedances of the MCPA Antenna**

When the coupling between conductors is taken into account, the system admittance is a complex, dense matrix. To design the MCPA antenna, the length (\(L\)) and width (\(W\)) of the patch can be found numerically by imposing the imaginary part of the Eq. (19) to be equal to zero. Moving from Eq. (19), the two exponential contribution of the integrals can be highlighted,

\[
Y_{p,m}^* = \frac{1}{4\pi^2 k\eta} \int_{-\infty}^{+\infty} \left\{ k_z^2 + |k|_y^2 \right\} M(k_x, k_y) \left\{ \cos k_x(x_p - x_m) \right\} \frac{dk_x dk_y}{k_z^2}
\]

\[
-\frac{j}{4\pi^2 k\eta} \int_{-\infty}^{+\infty} \left\{ k_z^2 + |k|_y^2 \right\} M(k_x, k_y) \left\{ \sin k_x(x_p - x_m) \right\} \frac{dk_x dk_y}{k_z^2}
\]

The second integral is zero in all the integration domain, given that it is an odd \(\sin()\) function. Therefore, we get

\[
Y_{p,m}^* = \frac{1}{4\pi^2 k\eta} \int_{-\infty}^{+\infty} \left\{ k_z^2 + |k|_y^2 \right\} M(k_x, k_y) \left\{ \cos k_x(x_p - x_m) \right\} \frac{dk_x dk_y}{k_z^2}
\]

(41)

Given this simplification, we can focus on the first quadrant and solve the following integral

\[
Y_{p,m}^* = \frac{1}{4\pi^2 k\eta} \int_{0}^{+\infty} \left\{ k_z^2 + |k|_y^2 \right\} M(k_x, k_y) \left\{ \cos k_x(x_p - x_m) \right\} \frac{dk_x dk_y}{k_z^2}
\]

(43)

The integral can be further decomposed in three terms

\[
Y_{p,m}^* = \frac{1}{4\pi^2 k\eta} \int_{0}^{k} \int_{0}^{+\infty} (k^2 - k_y^2) (k_x, k_y) \left\{ \cos k_x(x_p - x_m) \right\} \frac{dk_x dk_y}{k_z^2}
\]

\[
+ \int_{k}^{+\infty} \int_{0}^{+\infty} (k^2 - k_y^2) M(k_x, k_y) \left\{ \cos k_x(x_p - x_m) \right\} \frac{dk_x dk_y}{k_z^2}
\]

(44)

Now, it is possible to easily separate the expression in its real part

\[
R_e(Y_{p,m}^*) = \frac{1}{\pi^2 k\eta} \int_{0}^{k} \int_{0}^{+\infty} (k^2 - k_y^2) M(k_x, k_y) \left\{ \cos k_x(x_p - x_m) \right\} \frac{dk_x dk_y}{k_z^2}
\]

and imaginary part,

\[
I_m(Y_{p,m}^*) = \int_{-\infty}^{+\infty} \int_{0}^{k} (k^2 - k_y^2) M(k_x, k_y) \left\{ \cos k_x(x_p - x_m) \right\} \frac{dk_x dk_y}{k_z^2}
\]

(45)

The real part of the irradiation admittance can be solved in closed form by writing

\[
R_e(Y_{p,m}^*) = \frac{T^2}{\pi^2 k\eta} \int_{0}^{k} (k^2 - k_y^2) \text{sinc}^2 \left( \frac{T}{2} k_x \right) \left\{ \cos k_x(x_p) \right\} \frac{dk_x}{k_z^2}
\]

(42)
The integral in curly braces can be expanded as

\[
\int_0^{\sqrt{k_1^2-k_2^2}} \sin^2\left(\frac{\Delta l}{2} k_s\right) \frac{dk_y}{\sqrt{k_1^2-k_2^2-k_y^2}} 
= \frac{\pi}{(k_1^2-k_2^2) \Delta l^2} \sum_{n=0}^{+\infty} \frac{(-1)^n (\Delta l (k_1^2-k_2^2))^2 (2n+1)}{(2n+2)(2n+1)2^{2n}(n!)^2}
\]

To get

\[
R_c(Y_{p,m}) = \frac{T^2}{\pi^2 k \eta} \int_0^{k_1^2-k_2^2} \sin^2\left(\frac{T}{2k_s}\right) \cos k_s(x_d) \left\{ \int_0^{+\infty} \frac{\Delta l}{2} k_y \frac{dk_y}{\sqrt{k_1^2-k_2^2-k_y^2}} \right\} dk_x + \frac{T^2}{\pi^2 k \eta} \left\{ \int_0^{+\infty} \frac{\Delta l}{2} k_y \frac{dk_y}{\sqrt{k_1^2-k_2^2-k_y^2}} \right\} dk_x
\]

This is the analytical expression which allows computing the real part of the admittance, which can be bound to be equal to the value of the feeding line.

As regards the imaginary part, Eq. 45 is made of three integrals, one of which is an improper integral, which nulls thanks to the odd sin() function. The integral can be simplified by using the relationships from Rhodes [46],

\[
I_m(Y_{p,m}) = \frac{T^2}{\pi^2 k \eta} \int_0^{k_1^2-k_2^2} \sin^2\left(\frac{T}{2k_s}\right) \cos k_s(x_d) \left\{ \int_0^{+\infty} \frac{\Delta l}{2} k_y \frac{dk_y}{\sqrt{k_1^2-k_2^2-k_y^2}} \right\} dk_x
\]

To speed up the computation of the imaginary part, the products in Eq. 21 are decomposed and solved separately.

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