Received 12 October 2021; revised 3 January 2022; accepted 5 January 2022. Date of publication 14 January 2022; date of current version 25 January 2022.

Digital Object Identifier 10.1109/OJAP.2022.3143170

Multi-Layered Coating Metasurfaces Enabling Frequency Reconfigurability in Wire Antenna

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This work was supported by the Italian Ministry of University and Research as a PRIN 2017 Project in the frame of the activities of the research contract MANTLES under Grant 2017BHFZKH.

ABSTRACT In this work, we introduce a conceptually new approach for designing frequency reconfigurable wire antennas based on the use of multi-layered wrapping metasurfaces. Specifically, we demonstrate that the complex-valued input impedance of a wire antenna can be tailored by engineering the electromagnetic characteristics of a coating metasurface and we discuss how this effect can be exploited for achieving wide-band frequency reconfigurability. We report the advantages and limitations of this approach – especially compared to conventional impedance matching techniques - and, as a relevant example, we discuss the design of a reconfigurable half-wavelength dipole. For this example, the coating metasurface consists of a three-layer capacitive structure loaded with varactor diodes. It is shown that the operative frequency band of the antenna can be dynamically and continuously shifted in a quite broad range of frequencies (2/3 octave bandwidth) while preserving the current distribution of the fundamental mode and the omnidirectional shape of its radiation pattern on the horizontal plane. The possibility to allocate the antenna service within continuous sub-bands of operation makes this solution particularly suited for cognitive radio systems.

INDEX TERMS Frequency reconfigurability, multi-band antenna, multi-layer metasurfaces, reconfigurable metasurfaces, varactors, cognitive radio, wire antennas.

I. INTRODUCTION

THE EXPONENTIALLY growing demand for wireless communication systems enabling multiple services has rapidly increased frequency usage, causing spectrum congestion [1], [2]. To mitigate this problem, the design of smart devices enabling maximum efficiency in the spectrum utilization through an intelligent and dynamic allocation of the services has been proposed [3], [4]. These systems, often referred to as cognitive radio systems [5], typically perform an electromagnetic scan of the surrounding environment in order to assess the actual spectrum occupancy and, consequently select an available sub-band for operation [6]. For this purpose, reconfigurable wireless communication devices combining different methodological paradigms such as signal processing, artificial intelligence, wideband/multiple antenna techniques, and information theory are usually required [7].

However, the physical reconfigurability of the antenna element remains a critical challenge, in particular for what concerns the possibility to reconfigure the frequency of operation almost continuously within a prescribed range.

Antenna reconfigurability can be achieved in several ways [8]–[10]. A common and simple solution relies on the use of external reconfigurable matching networks [11]. However, the implementation of such a strategy becomes rather complex for multiband operation or continuous tuning, and its adoption can significantly affect space occupancy. For these reasons, several electrical reconfiguration techniques exploiting switches to connect and disconnect various segments of the antenna metallic parts and redistribute the current flow have been proposed in the last decades [12], [13].

A completely different approach that makes use of extremely broadband antennas has been introduced by exploiting high-Q microwave filters able to alternatively select the desired frequency band [14]–[16]. However, the requirement of wideband antennas and bulky and expensive filters severely limits the feasibility of such a solution. Therefore, reconfigurable antennas with limited instantaneous bandwidth and switching or tuning capabilities are usually preferred [17]–[22].

Among the different radiating devices designed to operate in multiple bands, omnidirectional antennas play an important role. In fact, vertically polarized dipole/monopole antennas are widely employed in various mobile communications, broadcasting, networking, and sensing applications. To equip such antennas with unprecedented functionalities, one possibility is to surround them with properly designed metasurfaces [23]. For instance, the use of lightweight conformal metasurface coatings has been proposed and successfully demonstrated for the reduction of the blockage and mutual coupling effects arising between antennas placed in close proximity, allowing the design of extremely compact communication systems for both terrestrial [24]–[27] and satellite [28]–[30] applications. In [31]–[34], similar metasurface coatings have been exploited to design array systems with expanded radiating functionalities. For instance, a metasurface coat able to selectively hide a dipole antenna to a detecting radar while allowing communication with a base station has been introduced in [35], [36], paving the way to the design of antenna systems with both frequency- and time-domain selective properties [37]. More recently, the possibility to design an inductive coating metasurface able to hide dielectric support whilst efficiently radiating in a different frequency band has been proposed [38].

Despite these recent efforts to achieve mutual coupling and blockage reduction, to the authors' best knowledge, there are very few studies exploiting coating metasurfaces to tailor the impedance matching of wire antennas [39], [40]. Moreover, existing results are mainly focused on widening the antenna impedance bandwidth and, thus, they are not suitable for cognitive systems where a narrowband operation over a wide tuning range is required.

Inspired by these works, a new approach for tuning the electrical properties of vertically polarized antennas based on the use of multi-layered metasurface coats is proposed. Depending on the values of the surface impedance of the coating metasurface layers, independent resonances can be excited which can be dynamically and continuously shifted within a broad frequency range. These resonances give rise to current distributions similar to the one of the fundamental mode, thus preserving the vertical-axis symmetry of the radiated field. As a relevant example, the design of a coated dipole antenna where the surface impedance values are tailored through varactor diodes loading the multi-layered metasurface is illustrated.

The resulting lightweight circuit-loaded coatings are promising candidates for enabling advanced compact and low-cost reconfigurable antennas with omnidirectional patterns, and represents an innovative alternative route compared to conventional reconfigurable antenna technique, to be potentially adopted in cognitive radio systems operating in congested spectrum scenarios.

II. COATING METASURFACES FOR FREQUENCY RECONFIGURABILITY

In this Section, we show that coating metasurfaces can be used not only to tailor the scattering characteristic of wired antennas (as discussed in [24]–[34], [36]) but also to tune its input impedance and its matching to standard feeding lines at different frequencies. In other terms, we illustrate how a metasurface coat can be designed to behave as a generalized matching network implemented at a physical level, i.e., exploiting the electromagnetic coupling with the antenna rather than acting on the electrical waves flowing on the feeding network.

In order to introduce such a concept, a half-wavelength dipole antenna coated by an ideal conformal metasurface is considered for illustrative purposes (Fig. 1(a)). The antenna is designed to resonate at $f_0 = 1$ GHz, with a length $l_\mathrm{a} = \lambda_0/2.27$ and a diameter $a = \lambda_0/100$. The conformal metasurface consists of an ideal cylindrical reactive sheet described by a purely reactive scalar impedance X_s, with a diameter $a_\mathrm{c} = \gamma a$ and a length $l_\mathrm{m} = \lambda_0/1.5$.

As it can be appreciated from the magnitude of the reflection coefficient at the 50 Ω input port of the antenna, shown in Fig. 1, the matching frequency band of the antenna can be shifted within a quite broad range by tuning both the parameters γ and the surface reactance X_s.

In particular, the resonant frequency significantly moves towards higher values as $\gamma \rightarrow 1$ and the surface reactance X_s assumes small negative values (i.e., large capacitance). For instance, a value of $\gamma = 5$ and $X_s = -5\ \Omega/\mathrm{sq}$ corresponds to a shift in the resonance equal to $1.4 f_0$. Conversely, when both γ and $|X_\mathrm{s}|$ increase, the induced shift is quite small, and the matching frequency approaches f_0. As an example, the resonance is slightly shifted to $1.13 f_0$ when $\gamma = 13.3$ and $X_s = -49\ \Omega/\mathrm{sq}$. It is worthwhile to remark that although Fig. 1 only reports a selected set of parameter setups (for

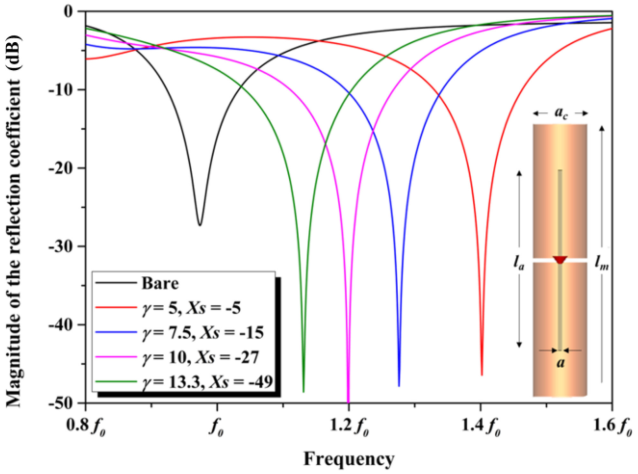

FIGURE 1. Magnitude of the reflection coefficient of the metasurface coated half-wavelength dipole antenna shown in the inset as a function of the metasurface parameters $\gamma = a_C/a$ and X_S [Ω/sq]. The black line refers to the case of the uncoated antenna. Full-wave simulation results when the dipole is excited by a standard 50 Ω source.

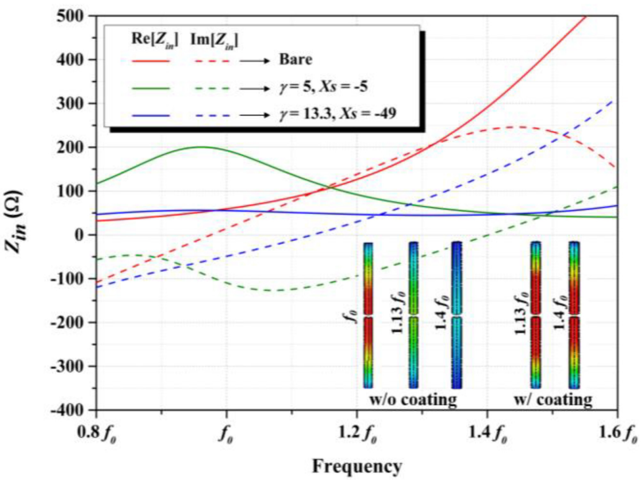

FIGURE 2. Complex input impedance of the metasurface coated antenna as a function of the metasurface parameters $\gamma = a_C/a$ and Xs [Ω]. In the insets, Current magnitude distributions on the antenna with (*w/*) and without (*w/o*) the coating metasurface at various frequencies. Full-wave simulations.

the sake of visual clarity), a continuous shift of the antenna resonance can be achieved, allowing for a smooth frequency reconfigurability within sub-bands of operation, through a proper combination of both γ and X_s, Furthermore, although only capacitive values of X_s have been reported, a more comprehensive study considering also inductive values has been performed. However, since inductive metasurfaces have a poor effect on the antenna resonance and cannot be efficiently exploited to control the antenna response, such results have been omitted.

To gain a better understanding of the operation principle, the input impedance (Z_{in}) of the coated antenna is shown in Fig. 2. Without the coating metasurface, the imaginary part of Z_{in} at the resonant frequency f_0 is almost zero, while the real part turns out to be around the standard 50 Ω value. By coating the antenna with the metasurface and reducing both the values of $|X_s|$ and γ, the antenna resonance frequency is progressively increased. At the same time, the real part of Z_{in} is flattened in a broad frequency range around 50 Ω, allowing to achieve resonance and impedance matching within a wide range of frequencies *without changing the length of the dipole*.

It is worth noting that this effect cannot be observed in a bare uncoated dipole. In fact, as well known and shown in Fig. 2 (green line), the real part of Z_{in} of a conventional dipole changes significantly as the frequency changes. Therefore, the coating metasurface allows to control not only the imaginary part of Z_{in} but *also* its real part, behaving as an equivalent complex matching transmission-line network that can be tuned by acting on its surface impedance and/or its diameter. Indeed, the described behavior is unprecedented and is enabled by the peculiar characteristics of electromagnetic cloaking when applied to antenna systems [34]–[36].

The current distribution in the coated cases for $\gamma = 5$; $X_s = -5$ Ω, and for $\gamma = 13.3$; $X_s = -49$ Ω evaluated at their own resonant frequencies ($1.4f_0$ and $1.13f_0$, respectively) are shown in the inset of Fig. 2, for completeness. These plots demonstrate that the fundamental mode of the bare antenna and the modes induced by the metasurface exhibit the same current distributions, even if the electrical length of the dipole is considerably different at the different frequencies. Such a result indicates that, unlike standard dipole antennas, an omnidirectional pattern analogous to the one of the uncoated case is expected within the entire tuning band in the coated scenario. From a physical point of view, this behavior is consistent with the fact that the coating metasurface reflects part of the radiated waves back to the antenna, inducing a secondary current that modifies the original current distribution and compensates for its variations. Depending on the values of X_{s} and γ, the interference between the currents can be tailored to achieve the desired complex input impedance Z_{in} and, thus, the desired resonance frequency.

The intriguing possibility of tuning both the imaginary and real parts of the input impedance is a *major advantage of the proposed approach* compared to conventional reconfigurable matching techniques based on reactive lumped elements, as shown in Fig. 3. To prove this point, the performances of the proposed approach are compared to the ones exhibited by a conventional capacitive antenna tuner. In particular, we compare the scenario of the coating metasurface with $\gamma = 5$ and $X_{\text{s}} = -5$ Ω/sq with the case of the same dipole loaded with a variable capacitor at its input port. As it can be appreciated, in the latter case (dashed curves), the obtained antenna frequency reconfigurability is very limited, due to the impossibility of engineering the *real part* of the dipole input impedance, since the reactive elements can compensate just the imaginary part of the impedance. Conversely, in the coated antenna scenario proposed in this work (orange continuous curve), a tuning range up to $1.4f_0$ is obtained.

This behavior confirms the nature of the coating metasurface as an advanced matching network implemented at

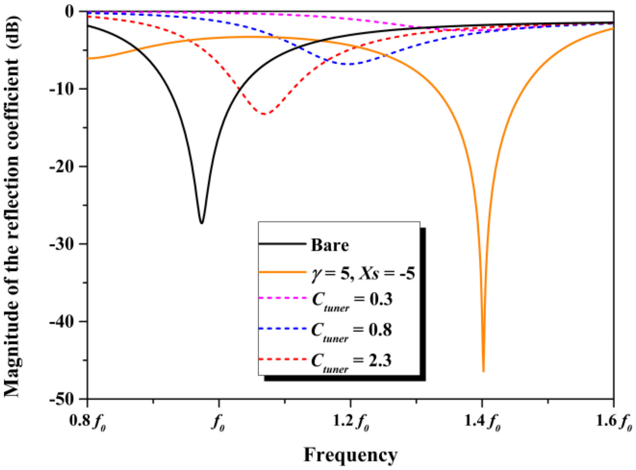

FIGURE 3. Magnitude of the reflection coefficient of the metasurface coated dipole antenna ($\gamma = 5$ and $Xs = -5\ \Omega/\text{sq}$) compared to the ones exhibited by a conventional capacitive antenna tuner. The black line refers to the case of the uncoated antenna. Full-wave simulations.

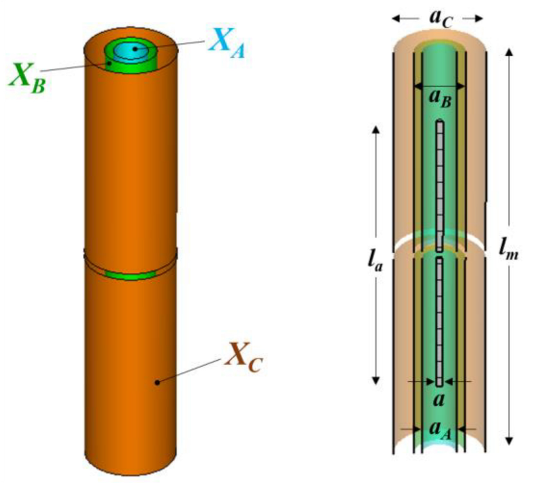

FIGURE 4. Half-wavelength dipole antenna coated by the multi-layered metasurface, composed of three different layers (*A*, *B*, *C*), characterized by different values of the surface reactance and radius (X_A, X_B, X_C and a_A, a_B, a_C).

a physical level. In principle, indeed, a similar huge resonant shift could be also achieved by implementing a complex matching network at the antenna input port but this would require the design of a complicated network able connecting/disconnecting matching stubs and reactive loads.

The above discussion suggests that the proposed metasurface should be able to change in real-time both its radius and its surface reactance to enabling tuning of the matching frequency of the coated dipole. While the latter quantity can be easily controlled by implementing an electric tuning mechanism, reconfiguring the radius of the metasurface would require a complex mechanical system. Therefore, the proposed approach to avoid this complication by replacing the single-layer metasurface with a multi-layered structure is shown next.

III. DESIGN OF VARACTOR-LOADED MULTI-LAYERED METASURFACES FOR FREQUENCY RECONFIGURABILITY

A realistic design of a frequency reconfigurable dipole antenna exploiting the operation principle discussed in the previous Section is shown in the following. For this purpose, let us consider an ideal multi-layered coating metasurface wrapped around a dipole antenna, like the one depicted in Fig. 4. The antenna has the same dimensions of the case discussed above, while the multi-layered metasurface is composed of three separate layers with diameters and surface reactance equal to (from inner to outer): $a_A = \gamma_A a$, $a_B = \gamma_B a$, $a_C = \gamma_C a$, and X_A, X_B, X_C. The idea is to tune the resonant frequency of the antenna through a proper modulation of the surface impedance of each layer, i.e., by varying the values of X_A, X_B, and X_C. The described geometry is compliant with the current fabrication techniques, which allow the realization of thin curved metasurfaces loaded with electronic elements [41].

As demonstrated above, only a specific combination of X_s and γ allows inducing a secondary current on the antenna able to properly interfere with the fundamental mode and, thus, to shift the resonance frequency. Therefore, only a specific value of X_s introduces the desired effect once the value of γ is defined.

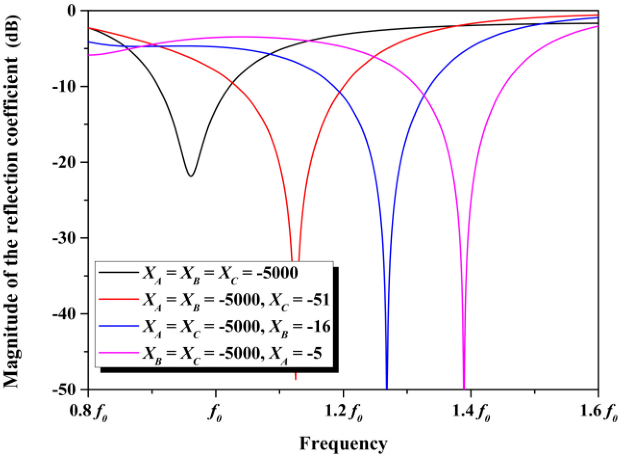

FIGURE 5. Magnitude of the reflection coefficient of the coated antenna as a function of different combinations of the layers surface reactance X_A, X_B, and X_C [Ω/sq]. Full-wave simulations results.

In Fig. 5, the amplitude of the reflection coefficient of the coated antenna for the cases with $\gamma_A = 5$, $\gamma_B = 7.5$, and $\gamma_C = 13.3$ is reported. Four distinct resonances can be noticed, depending on the combination of X_A, X_B, and X_C. Specifically, in order to excite the fundamental mode of the antenna at f_0, the surface impedances of all the layers should be characterized by an extremely large absolute value ($X_A = X_B = X_C = -5000\ \Omega/\text{sq}$), i.e., the metasurfaces behave as transparent layers. From an equivalent transmission-line point of view, the large impedances of the metasurfaces are connected in parallel to the antenna input impedance in this case, thus they do not affect the reflection coefficient at the input port [36].

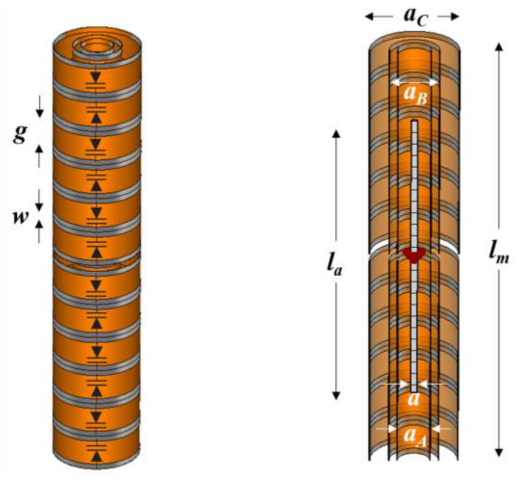

FIGURE 6. **Final design of the varactor-loaded metasurface coat for a dipole antenna. Slice and perspective view of the device.**

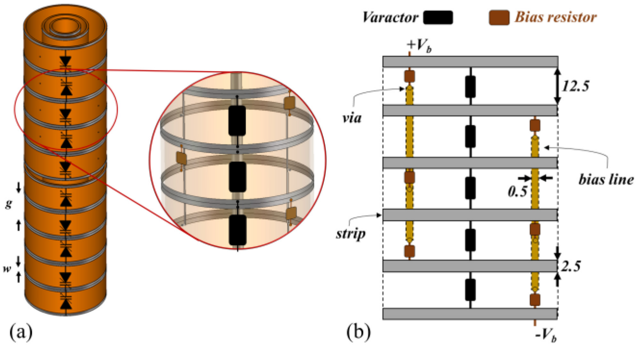

FIGURE 7. **(a) Detail of a section of the outer layer of the coating metasurface showing the structure of the biasing lines. (b) Sketch of a quarter section of an unwrapped metasurface layer. All lengths are expressed in millimeters.**

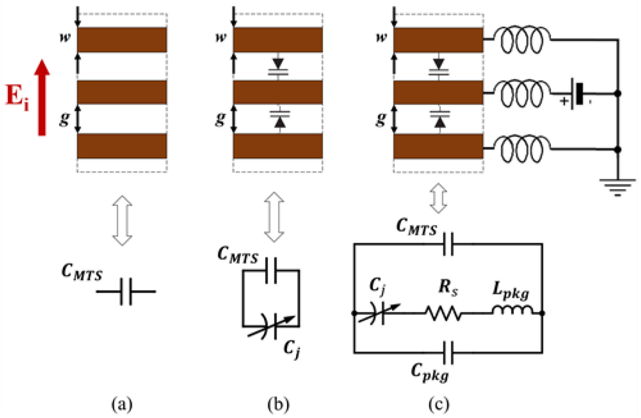

FIGURE 8. **Equivalent circuit model of the metasurface unit cell. (a) Unloaded scenario. (b) Loaded scenario considering a first-order varactor circuit model. (c) Loaded scenario also considering the varactor parasitic effects.**

Conversely, different resonant frequencies can be obtained when one of the metasurfaces is characterized by lower values of the surface reactance (*e.g.*, $X_C = -51\ \Omega/\text{sq}$). It is worth noting that the value of X_C required to induce the resonance at $1.13 f_0$ is slightly different compared to the value required in the single-layer coating metasurface ($X_s = -49\ \Omega/\text{sq}$) because of the mutual coupling effects arising between the metasurface layers. Once the desired values of X_A, X_B, and X_C are obtained from optimization, a practical implementation of the proposed multi-layered frequency reconfigurable antenna can be designed.

As shown in Fig. 6, simple horizontal metallic strips can be used to implement each metasurface layer. In fact, considering the TM_z field radiated by the antenna, the metallic rings can be used to synthesize capacitive metasurfaces with surface impedance values depending on the distance (g) and the width (w) of the strips [42], [43]. To take into account also the practical feasibility of the metasurfaces, each of the three layers is printed on a thin dielectric substrate ($t = 0.003\ \lambda_0$, $\varepsilon_r = 2.9$, $\tan\delta = 0.0025$).

In order to equip the system with reconfigurability capabilities, each of the metallic rings has been loaded with a set of three varactor diodes (C_A, C_B, C_C, from inner to outer), able to control the distributed equivalent capacitance of the rings themselves. As shown in Fig. 7, to properly set the voltage across them, a resistive bias network is used to connect the varactor in an anti-series configuration [44]. Since the DC current biasing the varactors is very low, the value of the resistors connecting the metallic rings to the biasing lines through vias is quite large to prevent the microwave-induced currents from flowing into the bias network ($R_{bias} = 4\ \text{k}\Omega$).

It is worth mentioning that a careful design of the network was required to minimize the interfering effect between the metallic strips and the biasing lines. In particular, a mirrored-rotated configuration of the $+/-$ biasing lines implemented through thin and short vertical metallic lines has been used, as it can be appreciated in Fig. 7(b). An alternating distribution of the biasing lines between the metasurface layers has been also introduced to further minimize possible interactions.

This configuration allows controlling the value of the surface impedance of each metasurface layer by a judicious variation of the varactor capacitance, i.e., by changing the DC voltage applied to the metasurface rings. From a circuital point of view the metasurface unit cell is in fact represented by an equivalent capacitance C_{MTS}, and its surface reactance is $X_{\text{MTS}} = -1/\omega C_{\text{MTS}}$, as can be seen from Fig. 8 (a). When introducing the loading varactor, its equivalent impedance appears in parallel to C_{MTS}. Since the varactor impedance can be modeled through a variable capacitance C_{j} in a first approximation (Fig. 8 (b)), the equivalent surface impedance of the loaded metasurface is given by $X_s = -1/\omega C_{\text{TOT}}$, where $C_{\text{TOT}} = C_{\text{MTS}} + C_{\text{j}}$. Thus, increasing the value of the C_{j} allows reducing the value of the X_{MTS} up to the target values.

Moreover, for the numerical simulations a more realistic model of the varactors taking into account also the varactor package parasitic effects have been considered. Specifically, the varactor diodes have been modeled through the equivalent circuit reported in Fig. 8 (c), where R_s is the series resistance of the varactors, L_{pkg} and C_{pkg} are the parasitic package reactance, while Cj is the variable junction capacitance.

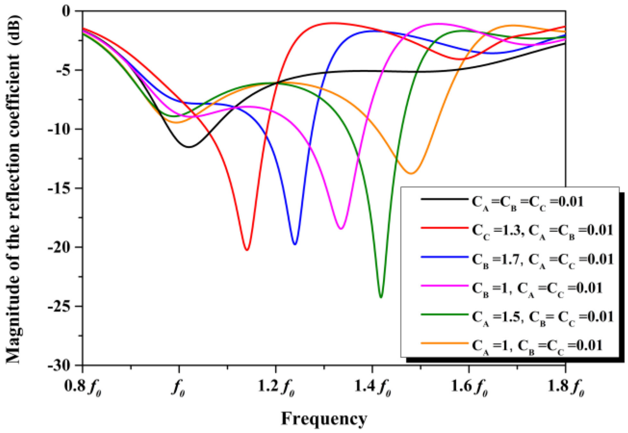

FIGURE 9. Magnitude of the reflection coefficient of the half-wavelength dipole antenna coated by the multi-layered metasurfaces in Fig. 5, for different combinations of the equivalent junction capacitances C_A, C_B, C_C of the varactors loading each metasurface layer. Junction capacitances are expressed in pF. Full-wave simulations results.

Assuming that the values of the parasitic reactances are small enough to be neglected, the expression for the equivalent surface impedance of the loaded metasurface can be derived as:

$$X_s^{tot} = -j\frac{\left(R_s - j/\omega C_j\right)}{C_{MTS}\left(R_s - j/\omega C_j - j/\omega C_{MTS}\right)\omega}$$

Thus, from the equation, the value of the junction capacitance C_j required to meet the targeted value of the equivalent surface impedance able to tune the antenna resonance can be evaluated.

As a first step of the design process, the horizontal strips have been designed to return a large equivalent reactance (i.e., $X_s = -5000\ \Omega/\text{sq}$). Towards this end, an analytical design procedure [23] and a full-wave numerical optimization through the commercial solver CST Microwave Studio aiming at considering the non-idealities due to the curvature of the strips have been exploited. It is worth pointing out that different combinations of g and w can be used to achieve the desired surface reactance. The combination ensuring the minimum number of rings (and, thus, of varactors) and, at the same time, geometrical values compatible with the standard electronic lumped elements dimensions have been selected. The optimized values turn out to be $g = \lambda_0/25$ and $w = \lambda_0/120$.

The different combinations of C_j needed to achieve the target values of X_A, X_B, and X_C and, thus, able to induce the distinct resonances, have been subsequently analytically derived through the model of Fig. 8 (c). Also, in this case, a numerical optimization procedure was required to take the non-idealities of the model into account, since four varactors are required for a correct homogenization of the distributed capacitance along the strips. Moreover, the values of C_j available from a commercial varactor diode (GC15006 Microsemi, with $R_s = 2.65\ \Omega$, $L_{pkg} = 0.4$ nH, and $C_{pkg} = 80$ fF) have been considered for the final choice of the optimal C_A, C_B, C_C.

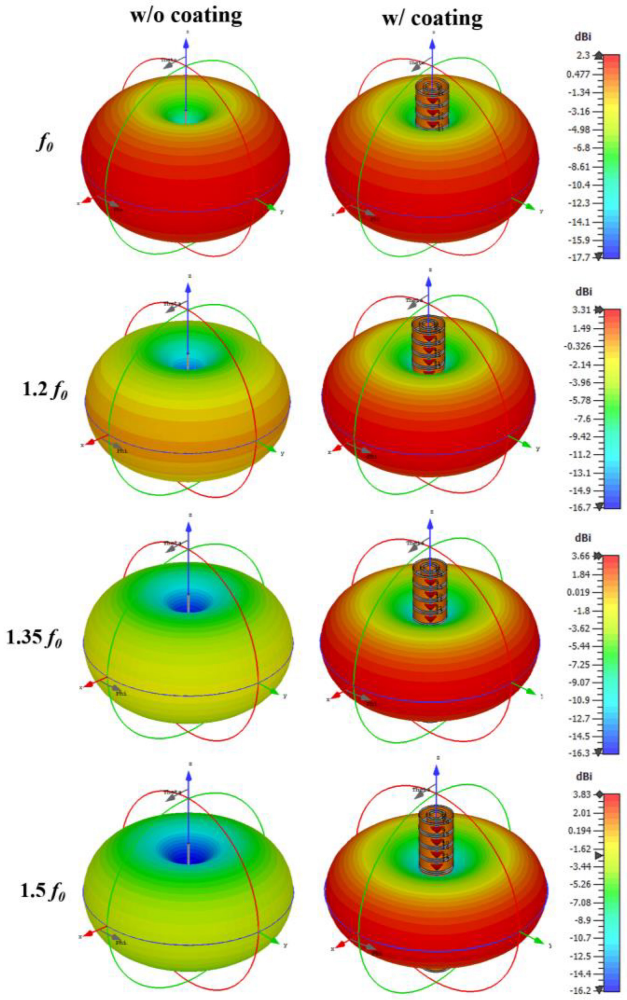

FIGURE 10. Realized gain radiation diagrams of the designed half-wavelength dipole antenna coated by a multi-layered metasurface evaluated at the four different resonant frequencies. For comparison, the left panel shows the radiation diagrams of the uncoated half-wavelength dipole. Full-wave simulations results.

As shown in Fig. 9, distinct resonances appear depending on the different combinations of C_A, C_B, C_C. In particular, the fundamental mode resonance at f_0 is excited when all the varactors on the three layers exhibit the lowest possible C_j value allowed from the varactor datasheet, i.e., C_A, C_B, $C_C = 0.01$ pF. A small resonance shift to $1.15 f_0$ appears instead when $C_C = 1.3$ pF and $C_A = C_B = 0.01$ pF. Whilst, for combinations of C_A, C_B, C_C ranging from 0.01 to 1.7 pF, the antenna resonance can be moved continuously up to $1.5 f_0$, for $C_A = 1.3$ pF and $C_B = C_C = 0.01$ pF.

Please note that the curves of the reflection coefficient magnitude are slightly different with respect to those in Fig. 5. This outcome is mostly due to the lowest value exhibited by the C_j of the varactors, which slightly alters the original surface reactance of the layers compared to their unloaded values. It is also worth mentioning that the resonances exhibited by the antenna allow covering adjacent frequency bands within a quite broad frequency spectrum (from f_0 up to $1.6 f_0$ – 50% fractional bandwidth), which is an essential feature in cognitive radio systems.

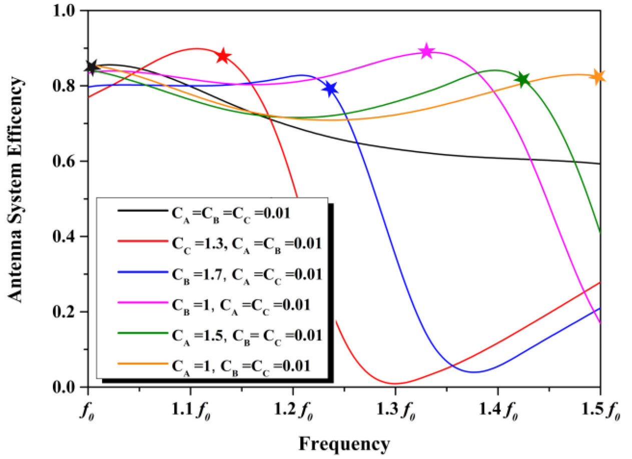

FIGURE 11. **Antenna system efficiency (i.e, ratio of the antenna radiated power to the power at the antenna input port) when considering the combination of C_A, C_B, C_C [pF] exciting the resonance curves in Fig. 10. The star symbols mark the antenna resonant frequency for the specific capacitance combinations. Full-wave simulations results.**

In Fig. 10, the realized gain radiation patterns at four different resonant frequencies are reported. As expect, because of the stable current distributions on the dipole antenna, the radiation diagrams at the different frequencies exhibit a uniform omnidirectional pattern. In Fig. 11, we report the antenna system efficiency (i.e., the ratio of the radiated power to the power at the antenna input port) for the same combinations of C_A, C_B, C_C used in Fig. 10. As it can be appreciated, the antenna efficiency keeps higher than 80% for all the considered scenarios. It is worth emphasizing that the presence of the biasing network and a realistic model for the varactors has been fully considered in these simulations. Indeed, the good efficiency performance is mainly related to the use of simple non-resonance metasurfaces, which support weak secondary currents, and to the minimization of the number of diodes required for the frequency reconfigurability.

Finally, it is worth remarking that the use of tunable varactor diodes to modify the values of the surface impedances of the metasurface layers offers dynamic spectrum management through the possibility of exciting neighboring bands of operation within a large frequency spectrum. In fact, Fig. 9 shows that the resonances excited can be slightly shifted in the frequency band $f_0 - 1.6f_0$ by a proper variation of C_A, C_B, and C_C within the varactor capacitive range, hence allowing an almost continuous allocation of the bandwidth of operation within the frequency range from f_0 up to $1.6f_0$.

IV. CONCLUSION

We have introduced a new approach to modify the resonance frequency of wire antennas exploiting flexible coating metasurfaces. By properly engineering the metasurface response, the complex-valued input impedance of the antenna can be tailored to match one of a standard 50 Ω feeding lines at different frequencies, enabling frequency reconfigurability.

This innovative design solution has been exploited to design a dipole antenna coated by multi-layered metasurface loaded by varactor diodes able to shift its resonant frequency within continuous and adjacent sub-bands of operation. At first, the radii of the metasurface layers have been numerically optimized to maximize the reconfigurable bandwidth and still guarantee continuous sub-bands of operation through feasible values of the surface impedances. Then, proper unit cells loaded with varactor diodes have been designed and the required values of the junction capacitances identified through analytical formulas. The final results show that a stable and efficient omnidirectional radiation pattern was ensured despite the presence of an integrated biasing line network and a realistic model for the varactor diodes, which have been accounted for in the full-wave numerical simulations.

In contrast to conventional matching networks, the proposed approach holds intrinsic advantages. In particular, it allows tuning both the real and the imaginary part of the complex input impedance, enabling, thus, wideband reconfigurability capabilities. The proposed approach does not require large values of the reactive and resistive loading or switches to achieve impedance matching. Moreover, since the designed coating material is based on conformal and lightweight multi-layered metasurfaces, it can be easily applied to preexisting antenna systems and, more in general, in all the scenarios where feasible reconfigurability is desired for cognitive radio purposes.

The use of reconfigurable electromagnetic structures, such as the metasurface considered here whose response is controlled through an integrated biasing network, further expands the field of engineered materials embedding "intelligence" at the physical layer, which will play a significant role for the next generation telecommunication systems.

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