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Impedance Bandwidth Improvement of a Planar Antenna Based on Metamaterial-Inspired T-Matching Network

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ABSTRACT **In this paper a metamaterial-inspired T-matching network is directly imbedded inside the feedline of a microstrip antenna to realize optimum power transfer between the front-end of a RF wireless transceiver and the antenna. The proposed T-matching network, which is composed of an arrangement of series capacitor, shunt inductor, series capacitor, exhibits left-handed metamaterial characteristics. The matching network is first theoretically modelled to gain insight of its limitations. It was then implemented directly in the 50-Ω feedline to a standard circular patch antenna, which is an unconventional methodology. The antenna's performance was verified through measurements. With the proposed technique there is 2.7 dBi improvement in the antenna's radiation gain and 12% increase in the efficiency, and this is achieved over a significantly wider frequency range by a factor of approximately twenty. Moreover, there is good correlation between the theoretical model, method of moments simulation, and the measurement results.**

INDEX TERMS ***T*-matching circuit, microstrip antenna, metamaterial, transmission-line, impedance matching.**

I. INTRODUCTION

Microstrip antennas have become popular for use in many wireless systems due to their planar profile, ease of fabrication and low manufacturing cost. Moreover, such antennas allow ease of integration with RF circuit components [1]. As a result, the use of planar based antenna configurations enables the integration of communication capabilities in a wide range of applications in the scope of the Internet of Things (IoT), such as context aware environments, Industry 4.0 or vehicular communications among others. This has been made possible by the design of flexible and compact transceivers especially in the sub-6 GHz band for wireless sensor networks, wireless local area networks, public land mobile networks and 5G communication systems. The vast number of applications which make use of wireless communication systems is leading to the need of employing different wireless systems in a cooperative way (i.e., Heterogeneous Network operation), as well as to an increase in the allocated channel bandwidth. Optimal performance of these wireless systems is given, among other factors, by adequate transceiver system operation. In relation with antenna sub-systems, impedance matching is one of the main factors to consider in order to reduce reflection-coefficient and hence, increase overall coverage range and minimize overall energy consumption [2-7].

The input impedance of an antenna is typically different from the impedance of the system to which it is connected, requiring impedance matching stages in order to minimize losses. In many cases, it is possible to realize impedance

matching by modifying the antenna geometry [8]. However, the main limitation of conventional microstrip antennas is their narrow impedance bandwidth (typically less than 3% for return-loss better than -10 dB). In addition, electrically thin edge-fed patch antennas usually suffer from high resonant input impedance of a microstrip patch, typically in a range of 300–500 Ω, which is challenging to directly match with a 50-Ω microstrip feedline. To maximize power transfer to and from the antenna, and therefore enhance the radiation efficiency of such antennas different matching approaches have been investigated previously [9-11].

In many wireless applications the matching circuit is not always designed for optimum power transfer between the antenna and transceiver, but consideration is given to maximizing the power handling capability, reducing non-linear distortion and impedance bandwidth. In [12] the matching and impedance bandwidth enhancement is achieved by stacking the radiation patch with a parasitic element. With this technique bandwidth extension of ~19% is possible. In [13], the feeding scheme employed provides an impedance bandwidth of ~28% however the design of the feeding scheme is complex to implement in practice. In [14], a capacitively coupled probe is used to excite a nonradiative mode close to the antenna's dominant radiation mode to realize a bandwidth of ~28%. In [15] multiple resonant modes are simultaneously excited to extend the bandwidth of a patch antenna by about 32%, however this technique is shown to adversely affect the antenna's broadside radiation pattern. In [16] the bandwidth is improved by 15.2% by exciting dual modes.

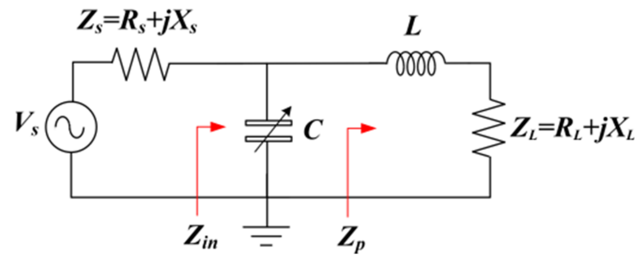
In this paper a simple and effective metamaterial-inspired wideband matching technique is proposed to enhance the radiation gain and efficiency performance of a planar antenna. The proposed matching technique uses a *T*-matching network consisting of series capacitor, shunt inductor, series capacitor configuration to realize metamaterial characteristics. The matching network is directly embedded inside the feedline of the antenna, which is an unconventional methodology. The technique is first theoretically characterized to gain an understanding of its effectiveness and is then validated through practical design and measurement.

II. IMPEDANCE MATCHED ANTENNA

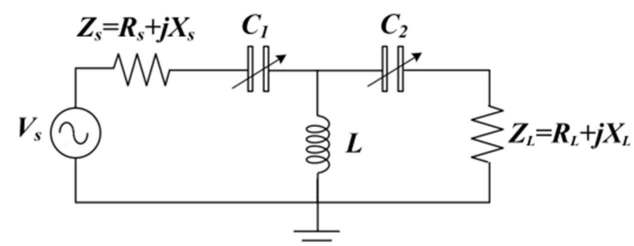
The three types of lumped element matching networks that are commonly used are based on the *L*-network, *π*-network, and *T*-network. The advantage of *L*-network is that it only has two reactive components whose values can be tuned easily for a given load impedance. However, its impedance matching capability is limited. On the other hand, the *π*- or *T*-networks have the capability of providing superior impedance matching flexibility than the *L*-network because they have three reactive components that can be tuned.

Fig.1 shows equivalent circuit of a typical *L*- and *T*-matching network. The reactance components can be either lumped elements or realized using transmission-lines based on microstrip integrated circuit technology. Capacitors can be interdigital or realized with low impedance microstrip-lines and vice versa for inductances. Reactance components realized using microstrip-lines however can result in a larger circuit.

The reactive components in the network control the return-loss between the antenna and matching network. In practice the capacitance values are adjusted or tuned until the return-loss of the antenna is lower than a specified threshold value, which is typically -10 dB, to (i) reduce system complexity, (ii) maintain the matching performance, and (iii) maximize transmission power to the antenna.



(a)



(b)

Fig.1. (a) Equivalent circuit of *L*-matching network, and (b) equivalent circuit of *T*-matching network.

To determine the effect of reactance components *C* and *L* on the input impedance Z_{in} of the *L*-network in Fig.1(a) the network is first represented in a simplified form in Fig.2, where the impedance Z_q accounts for inductance *L* and the load impedance $Z_L = R_L + jX_L$.

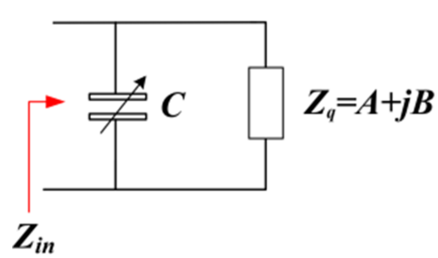


Fig.2. Impedance Z_{in} of the *L*-matching network in Fig.1.

$$Z_{in} = \frac{\frac{1}{\omega C}(B - jA)}{A + j\left(B - \frac{1}{\omega C}\right)} = R + jX \quad (1)$$

Where the real and imaginary impedances are:

$$R = A/[(\omega AC)^2 + (\omega BC - 1)^2] \quad (2)$$

$$X = [B - \omega C(A^2 + B^2)]/[(\omega AC)^2 + (\omega BC - 1)^2] \quad (3)$$

By dividing (2) by (3) it can be shown that

$$\omega C = (BR - AX)/R(A^2 + B^2) \quad (4)$$

Substituting (4) into (2) and (3), it can be shown that

$$[R - (A^2 + B^2)/2A]^2 + X^2 = [(A^2 + B^2)/2A]^2 \quad (5)$$

where $A = R_L$ and $B = \omega L + X_L$

Eqn. (5) describes a circle whose center and radius, i.e. the impedance position, is a function of the inductive and capacitive reactance and the load impedance. This equation reveals that the center and radius of the impedance circle can be changed by tweaking either one or both inductance and capacitance in the circuit to effect matching. Unlike the *L*-network a *T*-network can provide superior impedance matching because it has three reactive components that can be tuned. In the next section a *T*-matching circuit is investigated in the reduction of mismatch between the antenna and RF front-end circuit.

III. *T*-COMPENSATION CIRCUIT

In a real-world scenario, the matching network is connected to the antenna using a transmission-line. The proposed *T*-matching network in Fig.3 comprises an arrangement of series capacitor, shunt inductor, series capacitor that exhibits left-handed metamaterial characteristics [17]. The impedance Z_x seen by the signal from the transmitter is a combination of the impedance Z_1 of the transmission-line of length *l* and the impedance of the antenna load Z_L. Hence, to optimize power transfer between the transmitter to the antenna, we need to conjugately impedance match the impedance Z_{in} looking into the matching circuit to Z_x.

The impedance Z_x looking into the antenna load of impedance Z_L via the transmission-line of length *l* of impedance Z_1 is given by

$$Z_x = Z_1\left[\frac{Z_L + jZ_1 tan\beta l}{Z_1 + jZ_L tan\beta l}\right] \quad (6)$$

$$Z_y = \frac{\frac{L}{C_2} + j\omega L Z_x}{Z_x + j\left(\frac{\omega^2 LC_2 - 1}{\omega C_2}\right)} \quad (7)$$

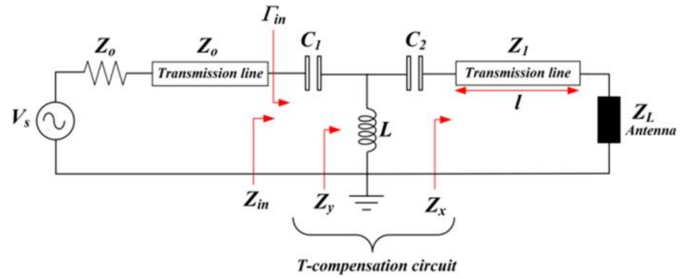


Fig.3. Equivalent circuit model of the *T*-matching network inserted between the source and the antenna.

Input impedance Z_{in} of the *T*-compensation circuit, shown in Fig.3, can be represented as $Z_{in} = R_{in} + jX_{in}$, where R_{in} and X_{in} can be determined to be given by

$$R_{in} = \frac{\frac{xL}{t}\left(\frac{1}{C_2} - \frac{\omega y}{t}\right) + \frac{\omega x L}{t}\left[\frac{y}{t} + \left(\frac{\omega^2 LC_2 - 1}{\omega C_2}\right)\right]}{\left(\frac{x}{t}\right)^2 + \left[\frac{y}{t} + \left(\frac{\omega^2 LC_2 - 1}{\omega C_2}\right)\right]^2} \quad (8)$$

$$X_{in} = \frac{\omega L\left(\frac{x}{t}\right)^2 - L\left(\frac{1}{C_2} - \frac{\omega y}{t}\right)\left[\frac{y}{t} + \left(\frac{\omega^2 LC_2 - 1}{\omega C_2}\right)\right]}{\left(\frac{x}{t}\right)^2 + \left[\frac{y}{t} + \left(\frac{\omega^2 LC_2 - 1}{\omega C_2}\right)\right]^2} - \omega C_1 \quad (9)$$

where

$$x = Z_L Z_1(1 + tan^2\beta l)$$

$$y = tan^2\beta l(Z_1^2 - Z_L^2)$$

$$t = Z_1^2 + Z_L^2 tan^2\beta l$$

Reflection-coefficient (Γ_{in}) at the input port of the antenna is a measure of how much power is reflected from the antenna. It determines how much power is radiated by the antenna and therefore the radiation efficiency of the antenna, which is defined as a ratio of radiated power to input power. The reflection-coefficient looking at the antenna via the *T*-compensation circuit is given by:

$$\Gamma_{in} = \frac{R_{in}^2 + X_{in}^2 - Z_o^2 - jR_{in}}{(R_{in} + Z_o)^2 + X_{in}^2} \quad (10)$$

where R_{in} and X_{in} are defined by Eqns. (8) and (9). The reflection-coefficient determines the maximum power transfer to the antenna. Eqn.(10) indicates that the center frequency and bandwidth of the matching system can be controlled by simply using appropriate values of the capacitances C_1 and C_2 and inductance L. For a *T*-matching circuit the relationship between bandwidth and reflection-coefficient is derived in [18] to be given by

$$BW \approx \frac{\sqrt{|\Gamma_{in}|}}{1-|\Gamma_{in}|} \qquad (11)$$

Eqn.(11) indicates the by appropriately selecting R_{in} and X_{in} the bandwidth of the matching network can be optimized.

The basic T-match circuit is shown in Fig.4. Like in π-match, T-matching circuit allows to set impedance transformation ratio and Q-factor/matching bandwidth of circuit independently.

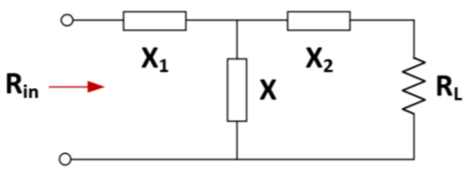


Fig.4. *T*-matching circuit.

$$Q_1 = \frac{X_1}{R_{in}} \qquad Q_2 = \frac{X_2}{R_L} \qquad (12)$$

The T-matching circuit can be decomposed to two L-match circuits, as shown in Fig.5. By series-to-parallel transformation as shown in From Fig.6,

$$Q_1 = \frac{R_I}{X_A} \qquad Q_2 = \frac{R_I}{X_B} \qquad (13)$$

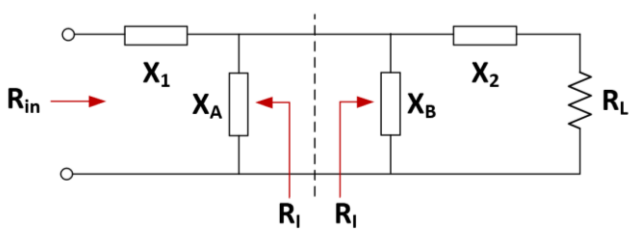


Fig.5. *T*-matching circuit decomposed to two *L*-match circuits.

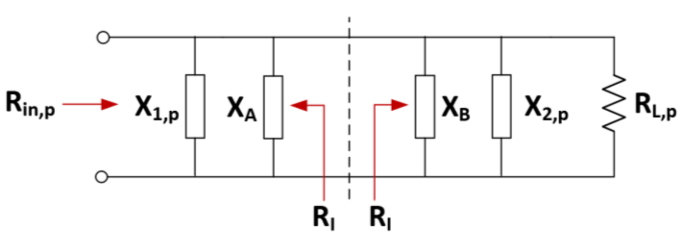


Fig.6. *T*-matching after series to parallel transformation.

$$R_{in,p} = R_I = R_{in}(1+Q_1^2) \quad R_{L,p} = R_I = R_L\left(1+\frac{1}{Q_2^2}\right) \qquad (14)$$

$$X_{1,p} = X_I\left(1+\frac{1}{Q_1^2}\right) \qquad X_{2,p} = X_I\left(1+\frac{1}{Q_2^2}\right) \qquad (15)$$

Rearranging Eqn.(14)

$$Q_1 = \sqrt{\frac{R_I}{R_{in}} - 1} \qquad Q_2 = \sqrt{\frac{R_I}{R_L} - 1} \qquad (16)$$

The total Q of the circuit is given by

$$Q = Q_1 + Q_2 = \sqrt{\frac{R_I}{R_{in}} - 1} + \sqrt{\frac{R_I}{R_{in}} - 1} \qquad (17)$$

At resonant frequency, $|X_A| = |X_{1,p}|$ and $|X_B| = |X_{2,p}|$.

Therefore, $\quad X_A = X_1\left(1+\frac{1}{Q_1^2}\right) \quad X_B = X_2\left(1+\frac{1}{Q_2^2}\right)$

The above relationships allow the design of the preliminary T-matching circuit. The design procedure involves: (i) Finding R_I using Eqn.(17) from given R_L, R_{in} and Q; (ii) Calculating Q_1 and Q_2 using Eqn.(16); (iii) Finding X_1 and X_2 using Eqn.(12), and (iv) Finding X_A and X_B using Eqn.(13).

In the next section it is shown that with proper selection of C_1, C_2, and L a high gain and high radiation efficiency can be obtained over a wide frequency band. However, this does introduce some complexity in the design as each variable capacitor needs to be controlled separately. Nevertheless, the proposed technique is amenable to electronic tuning hence allowing controllable matching capability according to system requirements.

IV. DESIGN EXAMPLE USING METAMATERIAL INSPIRED *T*-MATCHING NETWORK

A standard circular microstrip antenna was designed at 4 GHz using FR-4 lossy substrate with a relative permittivity of 4.3, thickness of 1.6 mm, and tanδ of 0.025. The radius of the circular patch is 10.9 mm. The patch antenna is excited using a microstrip feedline. The ground-plane on the backside of the substrate is truncated. There is no ground-plane in the region of the patch. Fig.7(a) shows the equivalent circuit model of the feedline and the patch antenna. The return-loss of the antenna predicted by theory and simulation in Fig.8 show it resonates at 4 GHz with a high-Q and hence has a narrow impedance bandwidth. The measured results show a bandwidth of 0.1 GHz for S_{11}≤-10 dB with an optimum return-loss value of -12 dB at 4 GHz.

Microstrip feedline length determines the input impedance of the antenna. Hence, a T-matching circuit is directly embedded in the feedline, which is an unconventional technique. Simply put, the proposed T-matching network is not an individual part from the antenna structure. The matching circuit used here is a metamaterial structure that exhibits left-handed properties of negative refractive index when interacts with electromagnetic signals [17]. The metamaterial characteristics are realized here with a T-network composed of series capacitance, shunt inductance, and series capacitance circuit, as per schematic diagram in Fig.7(b). The dispersion diagram of the antenna with the matching circuit in Fig.9 show the phase variation is virtually zero at 4 GHz.

The return-loss response in Fig.8 by theory, simulation, and measurement show that with the inclusion of the metamaterial matching network the antenna's impedance bandwidth is significantly improved. It is evident from Fig.8 that magnitudes of reactance components constituting the *T*-matching circuit has a great influence the Q-factor of the antenna. The results of the parametric study in Fig.10 show how the parameters constituting the *T*-matching circuit effect the return-loss response. The optimum values are: C_1 = 0.5462 pF, C_2 = 1.545 pF and L = 1.1586 pF. The fabricated antenna with the matching network is shown in Fig.11. Its measured return-loss in Fig.8 is better than -18 dB at 4 GHz. The magnitude of the geometrical parameters corresponding to the lumped element values, which are annotated in Fig.11, are given in Table I.

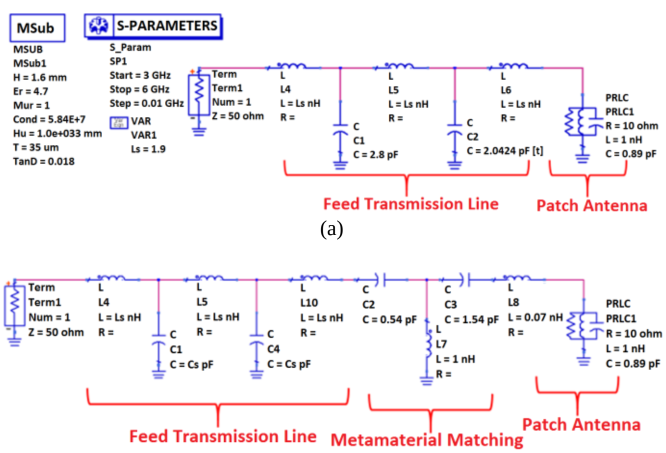


Fig.7. (a) Equivalent circuit model of feedline and patch antenna, and (b) circular patch antenna loaded with metamaterial matching network using Keysight Technologies' Advance Design System (ADS).

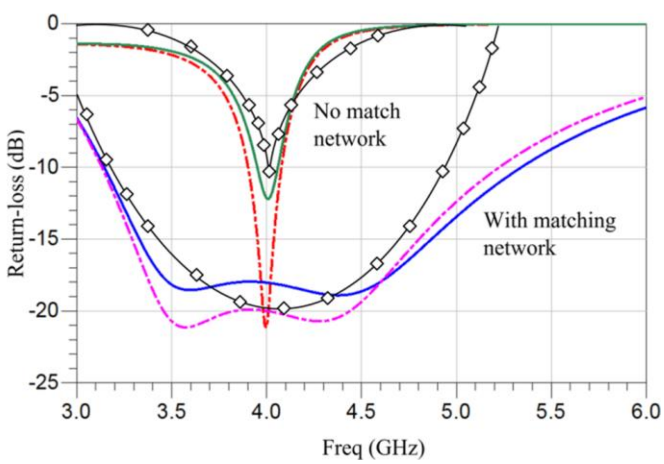


Fig.8. Theoretical, simulated and measured return-loss response of the antenna without and with the metamaterial-inspired matching network. Black diamond studded line is prediction by theory, dotted lines indicate the simulated result, and solid lines represent the measured result.

Simulated current distributions without and with the *T*-matching circuit at 4 GHz is shown in Fig.12. It is evident that the current distribution with the *T*-matching circuit the current intensity is concentrated at the shorting pin of the matching circuit in the feedline. The current however is much stronger over the patch compared to the case with no matching network.

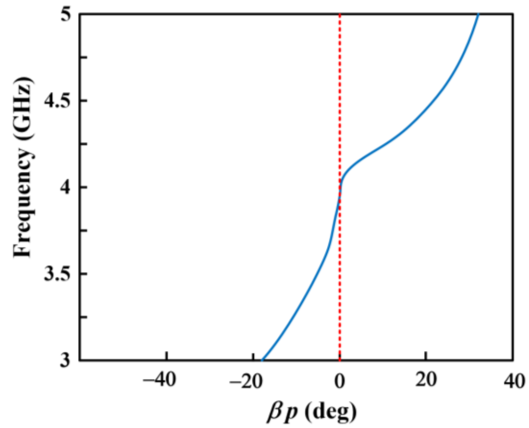


Fig.9. Dispersion diagram of the CRLH-TL feedline antenna.

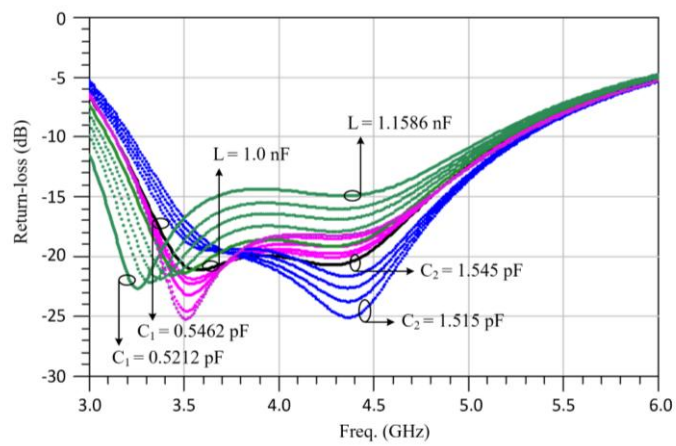


Fig.10. Effect on the return-loss response by the *T*-matching network parameters.

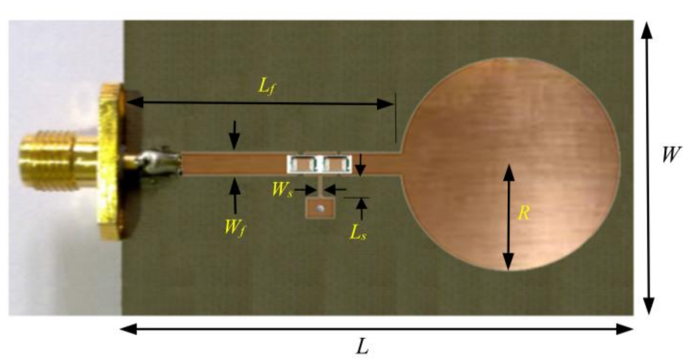


Fig.11. Photograph of the microstrip antenna incorporating the *T*-matching circuit.

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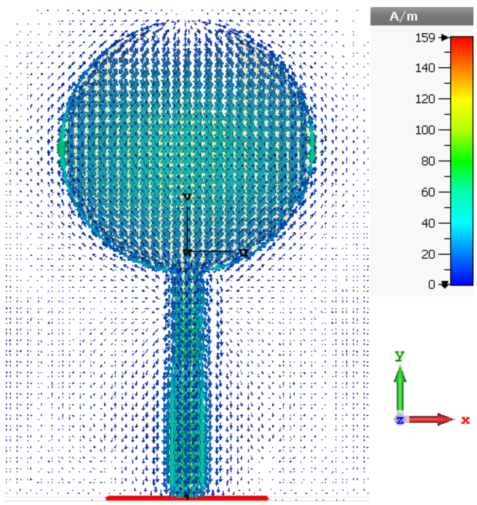
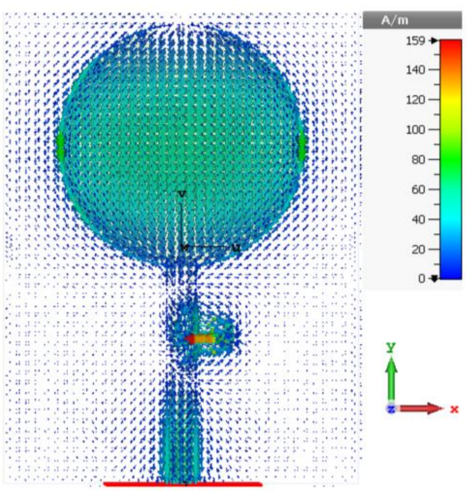
Table I. Antenna's geometrical parameters

Parameters	Size (mm)
L	54
W	29
R	11.2
L_f	28
W_f	2.47
L_s	2.3
W_s	0.5
Radius of shunt via-hole	0.5
Thickness of FR-4 substrate	1.6



(a)



(b)

Fig.12. Surface current distribution over the antenna at 4 GHz, (a) without the *T*-matching circuit, and (b) with the T-matching circuit.

The measured radiation gain and efficiency performance without and with the *T*-matching circuit are shown in Fig.13. It is evident from these results that there is substantial improvement in the radiation gain and efficiency by embedding the metamaterial inspired matching circuit. Measured results show without the matching circuit the antenna has a peak gain and efficiency of 5.5 dBi and 59.8%, respectively, at 4 GHz. However, with the matching circuit the peak gain and peak efficiency are 8.2 dBi and 71.8%, respectively. Moreover, the gain and efficiency extend over a significantly larger bandwidth. The gain ≥6 dBi extend from 3.35-5.25 GHz, and the efficiency ≥60% extend from 3.25-5.1GHz.

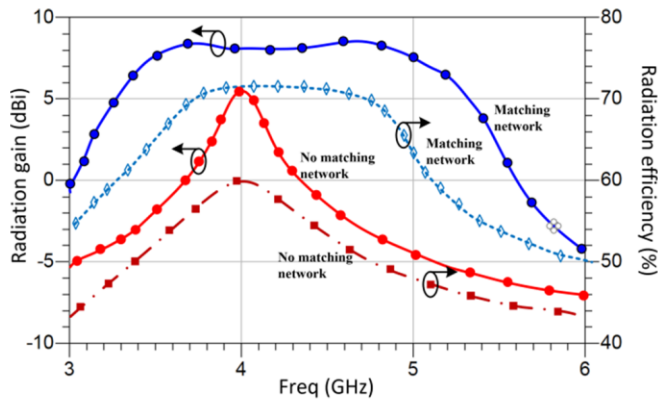
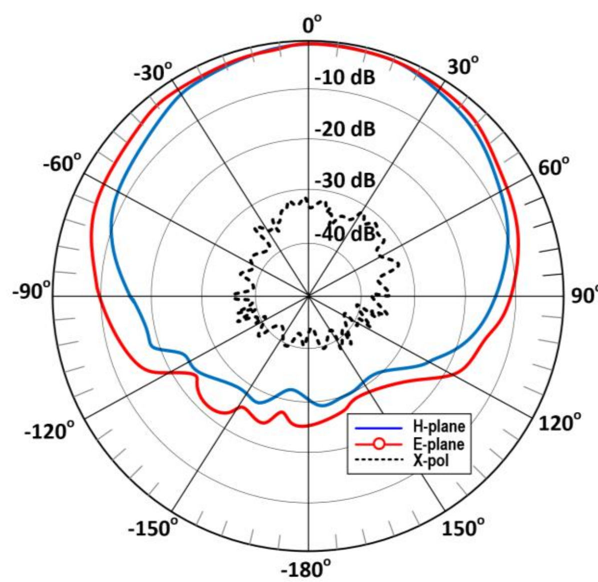
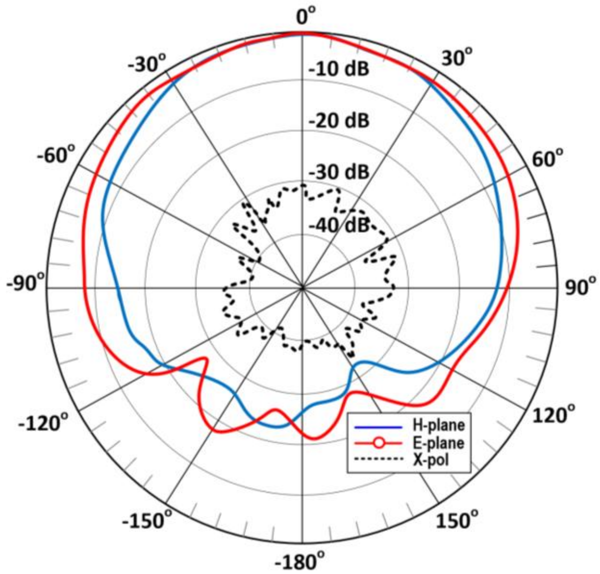


Fig.13. Measured radiation gain and efficiency response with and without the embedded *T*-matching circuit.

The radiation pattern in the E- and H-planes and the cross-polarization without and with the *T*-matching network at 4 GHz are shown in Fig. 14. The change in the radiation pattern with the *T*-matching network is marginal and, in both cases, the cross-polarization is less than -30 dB.



(a) With no *T*-matching circuit.



(b) With *T*-matching circuit.

Fig.14. The normalized measured radiation pattern in the E-plane and H-plane at 4 GHz, (a) without no matching circuit, and (b) with the embedded *T*-matching circuit.

Table II compares the present technique with some recently reported antenna bandwidth enhancement techniques. Impedance bandwidth enhancement of a microstrip patch antenna in [19] and [20] is achieved by loading the antenna with shorting pins and slots. In [21] the bandwidth is enhanced by simply curving the patch antenna in a partial cylindrical shape. In [22] one-dimensional electromagnetic bandgap ground structures and two-stage beam directors are employed in an inverted-L antenna topology to enhance the bandwidth. Even though the radiation efficiency of the proposed antenna is relatively low compared with the other techniques it is evident from the table that the proposed technique has a comparable gain and exhibits a substantially larger impedance bandwidth. Moreover, its size is relatively small with the exception of [22].

Table II. Comparison of the proposed matching technique of recently reported techniques.

Ref.	Technology	Antenna size	Fractional bandwidth (%)	Peak Gain (dBi)	Peak Eff. (%)
19	Shorting pins & slots	$1.14\lambda_o \times 1.27\lambda_o \times 0.009\lambda_o$	10	11.8	94
20	Shorting pins & slots	$1.02\lambda_o \times 1.47\lambda_o \times 0.045\lambda_o$	8.8 & 4.9	6 & 8	93 & 95
21	Microstrip	$0.58\lambda_o \times 0.58\lambda_o \times 0.016\lambda_o$	9	6.28	83.5
22	One-dimensional EM bandgap ground structures & two-stage beam directors	$0.21\lambda_o \times 0.32\lambda_o \times 0.009\lambda_o$	32.5	7	95
23	Differential microstrip	$0.8\lambda_o \times 0.8\lambda_o \times 0.013\lambda_o$	18.7	8.5	80
24	Magneto-electric dipole	$0.97\lambda_o \times 0.97\lambda_o \times 0.17\lambda_o$	28.2	9.0	-
25	Shorting pin & slot	$1.17\lambda_o \times 1.17\lambda_o \times 0.055\lambda_o$	32.2	6.5	93
26	Aperture-fed with shorting pins	$1.63\lambda_o \times 0.82\lambda_o \times 0.017\lambda_o$	15.2	6.8	-
This work	Metamaterial feedline	$0.67\lambda_o \times 0.38\lambda_o \times 0.021\lambda_o$	52.5	8.2	71.8

V. CONCLUSION

Impedance bandwidth of an antenna is determined by the matching conditions between the RF transceiver front-end and the antenna. We have demonstrated the effectiveness of employing an embedded metamaterial-based *T*-matching circuit directly implemented inside the feedline to enhance the antenna's impedance bandwidth. The measured results confirm significant improvement in the antenna's impedance bandwidth over its operating frequency range. Moreover, with the proposed technique improvement was observed in the radiation gain and efficiency too.

ACKNOWLEDGMENTS

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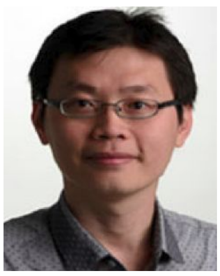

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Ernesto Limiti (S'87–M'92–SM'17) is a full professor of Electronics in the Engineering Faculty of the University of Roma Tor Vergata since 2002, after being research and teaching assistant (since 1991) and associate professor (since 1998) in the same University. Ernesto Limiti represents University of Roma Tor Vergata in the governing body of the MECSA (Microwave Engineering Center for Space Applications), an inter-universitary center among several Italian Universities. He has been elected to represent the Industrial Engineering sector in the Academic Senate of the University for the period 2007-2010 and 2010-2013. Ernesto Limiti is actually the president of the Consortium "Advanced research and Engineering for Space", ARES, formed between the University and two companies. Further, he is actually the president of the Laurea and Laurea Magistrale degrees in Electronic Engineering of the University of Roma Tor Vergata. The research activity of Ernesto Limiti is focused on three main lines, all of them belonging to the microwave and millimetre-wave electronics research area. The first one is related to characterisation and modelling for active and passive microwave and millimetre-wave devices. Regarding active devices, the research line is oriented to the small-signal, noise and large signal modelling. Regarding passive devices, equivalent-circuit models have been developed for interacting discontinuities in microstrip, for typical MMIC passive components (MIM capacitors) and to waveguide/coplanar waveguide transitions analysis and design. For active devices, new methodologies have been developed for the noise characterisation and the subsequent modelling, and equivalent-circuit modelling strategies have been implemented both for small and large-signal operating regimes for GaAs, GaN, SiC, Si, InP MESFET/HEMT devices. The second line is related to design methodologies and characterisation methods for low noise circuits. The main focus is on cryogenic amplifiers and devices. Collaborations are currently ongoing with the major radioastronomy institutes all around Europe within the frame of FP6 and FP7 programmes (RadioNet). Finally, the third line is in the analysis methods for nonlinear microwave circuits. In this line, novel analysis methods (Spectral Balance) are developed, together with the stability analysis of the solutions making use of traditional (harmonic balance) approaches. The above research lines have produced more than 250 publications on refereed international journals and presentations within international conferences. Ernesto Limiti acts as a referee of international journals of the microwave and millimetre wave electronics sector and is in the steering committee of international conferences and workshops. He is actively involved in research activities with many research groups, both European and Italian, and he is in tight collaborations with high-tech italian (Selex - SI, Thales Alenia Space, Rheinmetall, Elettronica S.p.A., Space Engineering …) and foreign (OMMIC, Siemens, UMS, …) companies. He contributed, as a researcher and/or as unit responsible, to several National (PRIN MIUR, Madess CNR, Agenzia Spaziale Italiana) and international (ESPRIT COSMIC, Manpower, Edge, Special Action MEPI, ESA, EUROPA, Korrigan, RadioNet FP6 and FP7 …) projects. Regarding teaching activities, Ernesto Limiti teaches, over his istitutional duties in the frame of the Corso di Laurea Magistrale in Ingegneria Elettronica, “Elettronica per lo Spazio” within the Master Course in Sistemi Avanzati di Comunicazione e Navigazione Satellitare. He is a member of the committee of the PhD program in Telecommunications and Microelectronics at the University of Roma Tor Vergata, tutoring an average of four PhD candidates per year.

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Original Article Title: "Impedance Bandwidth Improvement of a Planar Antenna Based Embedded Miniaturized T-Matching Network Inspired of Metamaterial Concept"

To: IEEE Access Editor

Re: Response to reviewers

Dear Dr. Raghvendra Kumar Chaudhary, and Reviewers,

Thank you for allowing us to resubmit the manuscript. We greatly appreciate your valuable time spending to read and evaluate our research work and providing us your constructive feedback to improve its quality.

We have carefully read the reviewers comments and addressed them in the revised manuscript. We have uploaded the following items: (i) our point-by-point response to the reviewer comments (see below); (ii) an updated manuscript with the changes highlighted in yellow; and (iii) an updated version of the manuscript without highlights in PDF format. We hope the changes are now satisfactory.

Best regards,
<Dr. Mohammad Alibakhshikenari> et al.

Reviewer#1

The author explains a T shaped matched network for enhancing the performance of circular patch antenna. However, minor corrections are needed for enhancing the quality of the manuscript.

Dear Reviewer #1, many thanks for your efforts to review our manuscript and given us your constructive feedback for its quality improvement. Hope the revised version including the required modifications can be acceptable for you.

Concern # 1: Explain how the gain and radiation efficiency are enhanced after placing the T-matching circuit. Explain with proper theory.

Authors response: Reflection-coefficient at the input port is a measure of how much power is reflected from the antenna. It determines how much power is radiated by the antenna and therefore determines the efficiency of the antenna which is a ratio of radiated power to input power. This is predicted by the reflection-coefficient Eqn.(10) that indicates that the center frequency and bandwidth of the matching system can be controlled by the serious capacitances C1 and C2 and shunt inductance L. We have employed an unconventional technique of embedding a T-matching circuit in the feedline of an antenna. The T-matching circuit, which consist of series capacitance, shunt inductance, and series capacitance circuit, equates to the electrical circuit of a metamaterial structure. The magnitude of the lumped elements in the matching circuit can be selected through simulation analysis to realize effective matching performance over a specified operating frequency. Empirical results presented in the paper demonstrate how the antenna's gain and radiation efficiency are improved by employing the proposed technique. The proposed technique is amenable to electronic tuning hence allowing matching capability over a wider bandwidth. Thank you.

Authors action: We updated the manuscript by carefully addressing the reviewer comment in the revised version.

Concern # 2: Radiation pattern is missing.

Authors response: We have now included the radiation pattern, accordingly. Thank you.

Authors action: We updated the manuscript with the radiation pattern.

Concern # 3: **Make a comparison table with similar kind of antennas and compare the performance of the presented and existing antenna in terms of antenna size and technology used.**

Authors response: The comparison table has now been updated as required, as requested. Thank you.

Authors action: We updated the manuscript accordingly.

Concern # 4: **Explain on what basis the antenna is metamaterial inspired.**

Authors response: The metamaterial characteristics are realized here with a T-network composed of series capacitance, shunt inductance, and series capacitance circuit (Fig.3). This circuit configuration represents a left-handed metamaterial circuit and the details analysis of this is given in [17] and in literature which is not repeated here. Thank you.

Authors action: We updated the manuscript accordingly.

Concern # 5: **Show the transmission and reflection performance of the T-shaped Matching Network.**

Authors response: We have shown the return loss of the antenna with and without the T-shaped matching network in Fig.5. return loss is directly related to reflection coefficient (Return loss = -20.log $|\Gamma|$). The antenna is a one port device therefore there is no transmission performance. Thank you.

Authors action: Return loss performance is included in the paper.

Concern # 6: **Explain theoretically, how the T-shaped matching network enhances the antenna bandwidth.**

Authors response: The theoretical analysis and derivation of the relationship between bandwidth and reflection-coefficient of a T-matching network in [17], which is not repeated here, is given by

$$BW \approx \frac{\sqrt{|\Gamma_{in}|}}{1-|\Gamma_{in}|} \qquad (11)$$

Which indicates that by appropriately selecting R_{in} and X_{in} in Γ_{in} of Eqn.(10) the bandwidth of the matching network can be optimized. Thank you.

Authors action: We updated the manuscript.

Concern # 7: **Minor corrections on English grammar.**

Authors response: We have now corrected the English grammar, accordingly. Thank you.

Authors action: We updated the manuscript.

Concern # 8: **Show the E-field distribution of the antenna with and without matching network.**

Authors response: We have now included the radiation patterns in the E-plane and H-plane without and with the matching circuit, as requested. Thank you.

Authors action: We updated the manuscript accordingly.

Concern # 9: Explains the new modes associated with the antenna with and without T shaped matching network.

Authors response: The dominant mode excited TM11 in the circular patch antenna is unaffected by the matching circuit because the feed point to the patch antenna is unaltered. Thank you.

Authors action: We updated the manuscript by with this explanation.

Concern # 10: Show the metamaterial properties such as E-field distribution and dispersion diagram of the antenna.

Authors response: We have now included E-field distribution and dispersion diagram of the antenna, accordingly. Thank you.

Authors action: We updated the manuscript by with this explanation.

Concern # 11: Add more number of latest relevant references in the introduction section and comparison table.

Authors response: We have added more references in the comparison Table 2, as requested. Thank you.

Authors action: We updated the manuscript.

Reviewer#2

Concern # 1: The idea behind the paper is good but lacking significant contents to be considered as a regular paper. The content has resemblances to a letter.

Authors response: Dear Reviewer #2, thank you so much for your attempts to review our article and given us your constructive comments for its quality improvement. We have added additional information in the revised version of the manuscript. Hope the revised version including the required modifications can be acceptable for you.

Authors action: We updated the manuscript accordingly.

Concern # 2: However, the more study is to be done for enhancing novelty of the work.

Authors response: The work is proof-of-concept study. As mentioned above we have added additional information in the revised version of the manuscript, accordingly. Thank you.

Authors action: We updated the manuscript accordingly.

Concern # 3: Please discuss T matching network design approach in more details and simulation approach in step by step manner.

Authors response: We have now included T-match theory to design it, as requested. Thank you.

Authors action: We updated the manuscript accordingly.

Concern # 4: **Please discuss effect of matching network on cross-polarization radiations of the antenna.**

Authors response: We have included the cross-polarization radiation patterns before and after applying the T-matching circuit as requested. Thank you.

Authors action: We updated the manuscript accordingly.

Reviewer#3

This paper presents an effective technique to enhance the bandwidth, and gain by employing T-matching network into the feed line. However paper suffers from the lack of evidences which should be provided in the revised version. I have mentioned some of the concerns with the papers as below:

Dear Reviewer #3, thank you so much for your efforts to review our article and providing us your constructive comments for its quality improvement. We have added additional information in the revised version of the manuscript. Hope the revised version including the required modifications can be acceptable for you.

Concern # 1: **The paper does not present any parametric analysis to study the effect of its capacitance on the resonant frequency.**

Authors response: We have now included parametric analysis in Fig.10 accordingly. Thank you.

Authors action: We updated the manuscript accordingly.

Concern # 2: **Authors should include the detailed schematic of the proposed antenna with detailed dimensions of the proposed antenna. This is necessary to know the electrical size of the proposed antenna, which is required to claim the miniaturization in the proposed structure.**

Authors response: We have added Table 1 the dimensions of the antenna and feedline as requested. Thank you.

Authors action: We updated the manuscript accordingly.

Concern # 3: **Comparison results show the information of the bandwidth, gain and efficiency however to compare the results it is very necessary to show the operating frequency for each of the compared results. It is obvious that high frequency offers larger bandwidth.**

Authors response: We have now shown the dimensions in terms of operating wavelength accordingly. Thank you.

Authors action: We updated the manuscript accordingly.

Concern # 4: **Apart from the S11 and Gain, radiation efficiency, plots no results show the contribution of the T- matching network. Authors should elaborate the contribution of the said network by comparing in terms of current distributions and E-Field Distributions.**

Authors response: We have added the current distributions and E-field plots before and after applying the *T*-matching network at operating frequency of 4 GHz, as requested. Thank you.

Authors action: We updated the manuscript accordingly.

Concern # 5: Why T- network is preferred as compared to pie and L-network? Explain.

Authors response: The metamaterial characteristics are realized here with a T-network composed of series capacitance, shunt inductance, and series capacitance circuit (Fig.3). This circuit configuration represents a left-handed metamaterial circuit and the details analysis of this is given in [17] and in literature which is not repeated here. Thank you.

Authors action: We updated the manuscript accordingly.

Concern # 6: Paper should be read out carefully for any sentence forming errors e.g. Proposed in this paper is a simple but effective metamaterial-inspired wideband matching technique to enhance power transfer between the RF front-end and the microstrip antenna.

Authors response: We have corrected the English language, accordingly. Thank you.

Authors action: We updated the manuscript accordingly.

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Impedance Bandwidth Improvement of a Planar Antenna Based on Metamaterial-Inspired T-Matching Network

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ABSTRACT **In this paper a metamaterial-inspired T-matching network is directly imbedded inside the feedline of a microstrip antenna to realize optimum power transfer between the front-end of a RF wireless transceiver and the antenna. The proposed T-matching network, which is composed of an arrangement of series capacitor, shunt inductor, series capacitor, exhibits left-handed metamaterial characteristics. The matching network is first theoretically modelled to gain insight of its limitations. It was then implemented directly in the 50-Ω feedline to a standard circular patch antenna, which is an unconventional methodology. The antenna's performance was verified through measurements. With the proposed technique there is 2.7 dBi improvement in the antenna's radiation gain and 12% increase in the efficiency, and this is achieved over a significantly wider frequency range by a factor of approximately twenty. Moreover, there is good correlation between the theoretical model, method of moments simulation, and the measurement results.**

INDEX TERMS ***T*-matching circuit, microstrip antenna, metamaterial, transmission-line, impedance matching.**

I. INTRODUCTION

Microstrip antennas have become popular for use in many wireless systems due to their planar profile, ease of fabrication and low manufacturing cost. Moreover, such antennas allow ease of integration with RF circuit components [1]. As a result, the use of planar based antenna configurations enables the integration of communication capabilities in a wide range of applications in the scope of the Internet of Things (IoT), such as context aware environments, Industry 4.0 or vehicular communications among others. This has been made possible by the design of flexible and compact transceivers especially in the sub-6 GHz band for wireless sensor networks, wireless local area networks, public land mobile networks and 5G communication systems. The vast number of applications which make use of wireless communication systems is leading to the need of employing different wireless systems in a cooperative way (i.e., Heterogeneous Network operation), as well as to an increase in the allocated channel bandwidth. Optimal performance of these wireless systems is given, among other factors, by adequate transceiver system operation. In relation with antenna sub-systems, impedance matching is one of the main factors to consider in order to reduce reflection-coefficient and hence, increase overall coverage range and minimize overall energy consumption [2-7].

The input impedance of an antenna is typically different from the impedance of the system to which it is connected, requiring impedance matching stages in order to minimize losses. In many cases, it is possible to realize impedance

matching by modifying the antenna geometry [8]. However, the main limitation of conventional microstrip antennas is their narrow impedance bandwidth (typically less than 3% for return-loss better than -10 dB). In addition, electrically thin edge-fed patch antennas usually suffer from high resonant input impedance of a microstrip patch, typically in a range of 300–500 Ω, which is challenging to directly match with a 50-Ω microstrip feedline. To maximize power transfer to and from the antenna, and therefore enhance the radiation efficiency of such antennas different matching approaches have been investigated previously [9-11].

In many wireless applications the matching circuit is not always designed for optimum power transfer between the antenna and transceiver, but consideration is given to maximizing the power handling capability, reducing non-linear distortion and impedance bandwidth. In [12] the matching and impedance bandwidth enhancement is achieved by stacking the radiation patch with a parasitic element. With this technique bandwidth extension of ~19% is possible. In [13], the feeding scheme employed provides an impedance bandwidth of ~28% however the design of the feeding scheme is complex to implement in practice. In [14], a capacitively coupled probe is used to excite a nonradiative mode close to the antenna's dominant radiation mode to realize a bandwidth of ~28%. In [15] multiple resonant modes are simultaneously excited to extend the bandwidth of a patch antenna by about 32%, however this technique is shown to adversely affect the antenna's broadside radiation pattern. In [16] the bandwidth is improved by 15.2% by exciting dual modes.

In this paper a simple and effective metamaterial-inspired wideband matching technique is proposed to enhance the radiation gain and efficiency performance of a planar antenna. The proposed matching technique uses a T-matching network consisting of series capacitor, shunt inductor, series capacitor configuration to realize metamaterial characteristics. The matching network is directly embedded inside the feedline of the antenna, which is an unconventional methodology. The technique is first theoretically characterized to gain an understanding of its effectiveness and is then validated through practical design and measurement.

II. IMPEDANCE MATCHED ANTENNA

The three types of lumped element matching networks that are commonly used are based on the L-network, π-network, and T-network. The advantage of L-network is that it only has two reactive components whose values can be tuned easily for a given load impedance. However, its impedance matching capability is limited. On the other hand, the π- or T-networks have the capability of providing superior impedance matching flexibility than the L-network because they have three reactive components that can be tuned.

Fig.1 shows equivalent circuit of a typical L- and T-matching network. The reactance components can be either lumped elements or realized using transmission-lines based on microstrip integrated circuit technology. Capacitors can be interdigital or realized with low impedance microstrip-lines and vice versa for inductances. Reactance components realized using microstrip-lines however can result in a larger circuit.

The reactive components in the network control the return-loss between the antenna and matching network. In practice the capacitance values are adjusted or tuned until the return-loss of the antenna is lower than a specified threshold value, which is typically -10 dB, to (i) reduce system complexity, (ii) maintain the matching performance, and (iii) maximize transmission power to the antenna.

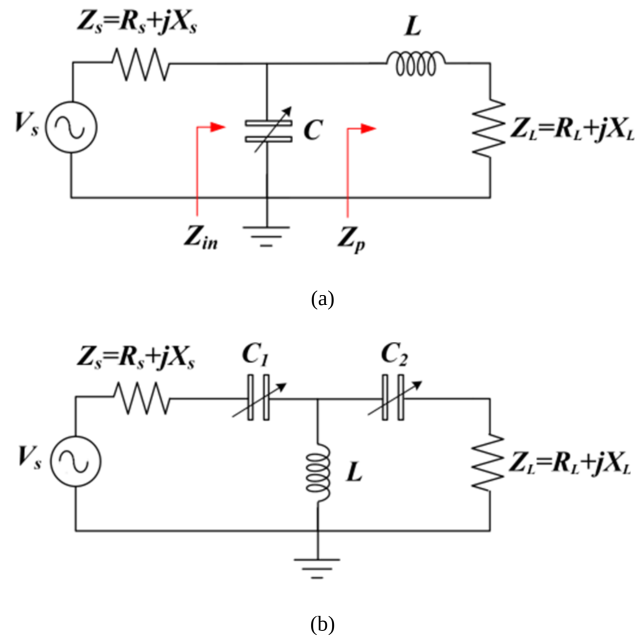

Fig.1. (a) Equivalent circuit of L-matching network, and (b) equivalent circuit of T-matching network.

To determine the effect of reactance components C and L on the input impedance Z_{in} of the L-network in Fig.1(a) the network is first represented in a simplified form in Fig.2, where the impedance Z_q accounts for inductance L and the load impedance $Z_L = R_L + jX_L$.

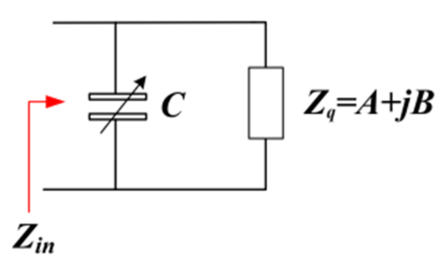

Fig.2. Impedance Z_{in} of the L-matching network in Fig.1.

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$$Z_{in} = \frac{\frac{1}{\omega C}(B-jA)}{A+j\left(B-\frac{1}{\omega C}\right)} = R + jX \qquad (1)$$

Where the real and imaginary impedances are:

$$R = A/[(\omega AC)^2 + (\omega BC - 1)^2] \qquad (2)$$

$$X = [B - \omega C(A^2 + B^2)]/[(\omega AC)^2 + (\omega BC - 1)^2] \qquad (3)$$

By dividing (2) by (3) it can be shown that

$$\omega C = (BR - AX)/R(A^2 + B^2) \qquad (4)$$

Substituting (4) into (2) and (3), it can be shown that

$$[R - (A^2 + B^2)/2A]^2 + X^2 = [(A^2 + B^2)/2A]^2 \qquad (5)$$

where $A = R_L$ and $B = \omega L + X_L$

Eqn. (5) describes a circle whose center and radius, i.e. the impedance position, is a function of the inductive and capacitive reactance and the load impedance. This equation reveals that the center and radius of the impedance circle can be changed by tweaking either one or both inductance and capacitance in the circuit to effect matching. Unlike the *L*-network a *T*-network can provide superior impedance matching because it has three reactive components that can be tuned. In the next section a *T*-matching circuit is investigated in the reduction of mismatch between the antenna and RF front-end circuit.

III. *T*-COMPENSATION CIRCUIT

In a real-world scenario, the matching network is connected to the antenna using a transmission-line. The proposed *T*-matching network in Fig.3 comprises an arrangement of series capacitor, shunt inductor, series capacitor that exhibits left-handed metamaterial characteristics [17]. The impedance Z_x seen by the signal from the transmitter is a combination of the impedance Z_1 of the transmission-line of length *l* and the impedance of the antenna load Z_L. Hence, to optimize power transfer between the transmitter to the antenna, we need to conjugately impedance match the impedance Z_{in} looking into the matching circuit to Z_x.

The impedance Z_x looking into the antenna load of impedance Z_L via the transmission-line of length *l* of impedance Z_1 is given by

$$Z_x = Z_1\left[\frac{Z_L + jZ_1 \tan\beta l}{Z_1 + jZ_L \tan\beta l}\right] \qquad (6)$$

$$Z_y = \frac{\frac{L}{C_2} + j\omega L Z_x}{Z_x + j\left(\frac{\omega^2 L C_2 - 1}{\omega C_2}\right)} \qquad (7)$$

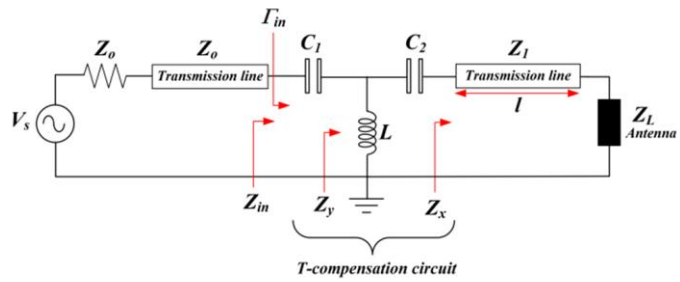

Fig.3. Equivalent circuit model of the *T*-matching network inserted between the source and the antenna.

Input impedance Z_{in} of the *T*-compensation circuit, shown in Fig.3, can be represented as $Z_{in} = R_{in} + jX_{in}$, where R_{in} and X_{in} can be determined to be given by

$$R_{in} = \frac{\frac{xL}{t}\left(\frac{1}{C_2} - \frac{\omega y}{t}\right) + \frac{\omega x L}{t}\left[\frac{y}{t} + \left(\frac{\omega^2 L C_2 - 1}{\omega C_2}\right)\right]}{\left(\frac{x}{t}\right)^2 + \left[\frac{y}{t} + \left(\frac{\omega^2 L C_2 - 1}{\omega C_2}\right)\right]^2} \qquad (8)$$

$$X_{in} = \frac{\omega L\left(\frac{x}{t}\right)^2 - L\left(\frac{1}{C_2} - \frac{\omega y}{t}\right)\left[\frac{y}{t} + \left(\frac{\omega^2 L C_2 - 1}{\omega C_2}\right)\right]}{\left(\frac{x}{t}\right)^2 + \left[\frac{y}{t} + \left(\frac{\omega^2 L C_2 - 1}{\omega C_2}\right)\right]^2} - \omega C_1 \qquad (9)$$

where

$$x = Z_L Z_1 (1 + \tan^2\beta l)$$

$$y = \tan^2\beta l (Z_1^2 - Z_L^2)$$

$$t = Z_1^2 + Z_L^2 \tan^2\beta l$$

Reflection-coefficient (Γ_{in}) at the input port of the antenna is a measure of how much power is reflected from the antenna. It determines how much power is radiated by the antenna and therefore the radiation efficiency of the antenna, which is defined as a ratio of radiated power to input power. The reflection-coefficient looking at the antenna via the *T*-compensation circuit is given by:

$$\Gamma_{in} = \frac{R_{in}^2 + X_{in}^2 - Z_o^2 - jR_{in}}{(R_{in} + Z_o)^2 + X_{in}^2} \qquad (10)$$

where R_{in} and X_{in} are defined by Eqns. (8) and (9). The reflection-coefficient determines the maximum power transfer to the antenna. Eqn.(10) indicates that the center frequency and bandwidth of the matching system can be controlled by simply using appropriate values of the capacitances C_1 and C_2 and inductance L. For a *T*-matching circuit the relationship between bandwidth and reflection-coefficient is derived in [18] to be given by

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$$BW \approx \frac{\sqrt{|\Gamma_{in}|}}{1-|\Gamma_{in}|} \qquad (11)$$

Eqn.(11) indicates the by appropriately selecting R_{in} and X_{in} the bandwidth of the matching network can be optimized.

The basic T-match circuit is shown in Fig.4. Like in π-match, T-matching circuit allows to set impedance transformation ratio and Q-factor/matching bandwidth of circuit independently.

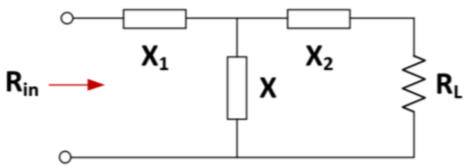

Fig.4. *T*-matching circuit.

$$Q_1 = \frac{X_1}{R_{in}} \qquad Q_2 = \frac{X_2}{R_L} \qquad (12)$$

The T-matching circuit can be decomposed to two L-match circuits, as shown in Fig.5. By series-to-parallel transformation as shown in From Fig.6,

$$Q_1 = \frac{R_I}{X_A} \qquad Q_2 = \frac{R_I}{X_B} \qquad (13)$$

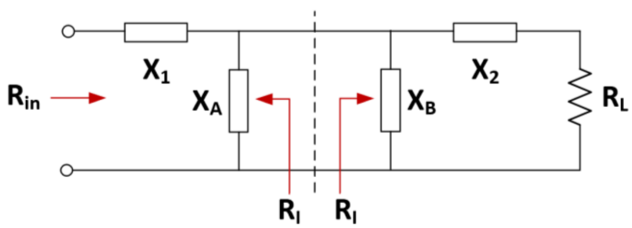

Fig.5. *T*-matching circuit decomposed to two *L*-match circuits.

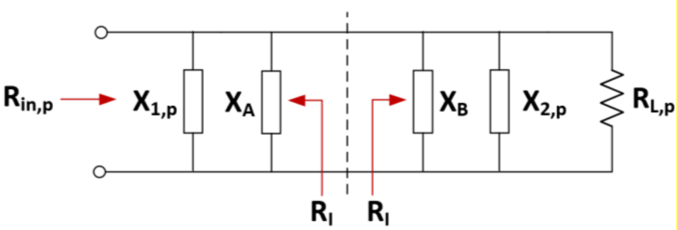

Fig.6. *T*-matching after series to parallel transformation.

$$R_{in,p} = R_I = R_{in}(1 + Q_1^2) \quad R_{L,p} = R_I = R_L\left(1 + \frac{1}{Q_2^2}\right) \qquad (14)$$

$$X_{1,p} = X_I\left(1 + \frac{1}{Q_1^2}\right) \qquad X_{2,p} = X_I\left(1 + \frac{1}{Q_2^2}\right) \qquad (15)$$

Rearranging Eqn.(14)

$$Q_1 = \sqrt{\frac{R_I}{R_{in}} - 1} \qquad Q_2 = \sqrt{\frac{R_I}{R_L} - 1} \qquad (16)$$

The total Q of the circuit is given by

$$Q = Q_1 + Q_2 = \sqrt{\frac{R_I}{R_{in}} - 1} + \sqrt{\frac{R_I}{R_{in}} - 1} \qquad (17)$$

At resonant frequency, $|X_A| = |X_{1,p}|$ and $|X_B| = |X_{2,p}|$.

Therefore, $X_A = X_1\left(1 + \frac{1}{Q_1^2}\right) \quad X_B = X_2\left(1 + \frac{1}{Q_2^2}\right)$

The above relationships allow the design of the preliminary T-matching circuit. The design procedure involves: (i) Finding R_I using Eqn.(17) from given R_L, R_{in} and Q; (ii) Calculating Q_1 and Q_2 using Eqn.(16); (iii) Finding X_1 and X_2 using Eqn.(12), and (iv) Finding X_A and X_B using Eqn.(13).

In the next section it is shown that with proper selection of C_1, C_2, and L a high gain and high radiation efficiency can be obtained over a wide frequency band. However, this does introduce some complexity in the design as each variable capacitor needs to be controlled separately. Nevertheless, the proposed technique is amenable to electronic tuning hence allowing controllable matching capability according to system requirements.

IV. DESIGN EXAMPLE USING METAMATERIAL INSPIRED *T*-MATCHING NETWORK

A standard circular microstrip antenna was designed at 4 GHz using FR-4 lossy substrate with a relative permittivity of 4.3, thickness of 1.6 mm, and tanδ of 0.025. The radius of the circular patch is 10.9 mm. The patch antenna is excited using a microstrip feedline. The ground-plane on the backside of the substrate is truncated. There is no ground-plane in the region of the patch. Fig.7(a) shows the equivalent circuit model of the feedline and the patch antenna. The return-loss of the antenna predicted by theory and simulation in Fig.8 show it resonates at 4 GHz with a high-Q and hence has a narrow impedance bandwidth. The measured results show a bandwidth of 0.1 GHz for S_{11}≤-10 dB with an optimum return-loss value of -12 dB at 4 GHz.

Microstrip feedline length determines the input impedance of the antenna. Hence, a T-matching circuit is directly embedded in the feedline, which is an unconventional technique. Simply put, the proposed T-matching network is not an individual part from the antenna structure. The matching circuit used here is a metamaterial structure that exhibits left-handed properties of negative refractive index when interacts with electromagnetic signals [17]. The metamaterial characteristics are realized here with a T-network composed of series capacitance, shunt inductance, and series capacitance circuit, as per schematic diagram in Fig.7(b). The

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dispersion diagram of the antenna with the matching circuit in Fig.9 show the phase variation is virtually zero at 4 GHz.

The return-loss response in Fig.8 by theory, simulation, and measurement show that with the inclusion of the metamaterial matching network the antenna's impedance bandwidth is significantly improved. It is evident from Fig.8 that magnitudes of reactance components constituting the *T*-matching circuit has a great influence the Q-factor of the antenna. The results of the parametric study in Fig.10 show how the parameters constituting the *T*-matching circuit effect the return-loss response. The optimum values are: C_1 = 0.5462 pF, C_2 = 1.545 pF and L = 1.1586 pF. The fabricated antenna with the matching network is shown in Fig.11. Its measured return-loss in Fig.8 is better than -18 dB at 4 GHz. The magnitude of the geometrical parameters corresponding to the lumped element values, which are annotated in Fig.11, are given in Table I.

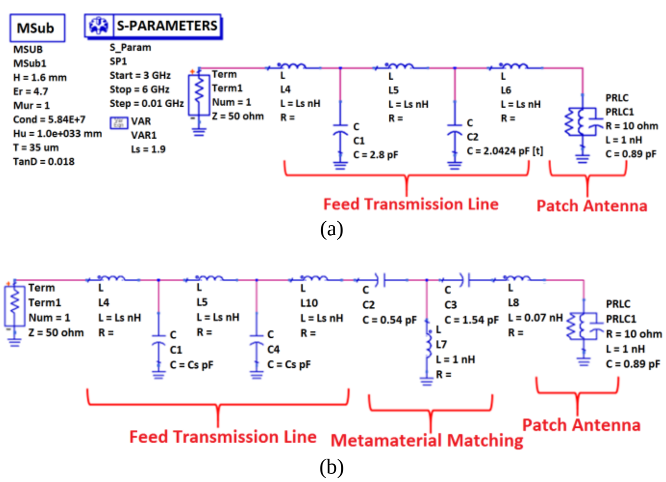


Fig.7. (a) Equivalent circuit model of feedline and patch antenna, and (b) circular patch antenna loaded with metamaterial matching network using Keysight Technologies' Advance Design System (ADS).

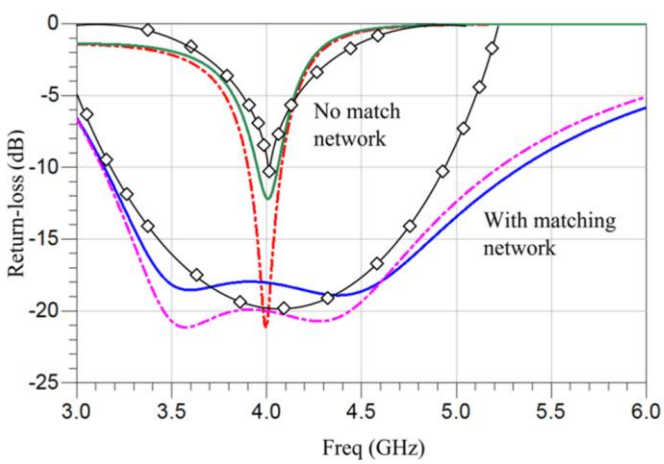


Fig.8. Theoretical, simulated and measured return-loss response of the antenna without and with the metamaterial-inspired matching network. Black diamond studded line is prediction by theory, dotted lines indicate the simulated result, and solid lines represent the measured result.

Simulated current distributions without and with the *T*-matching circuit at 4 GHz is shown in Fig.12. It is evident that the current distribution with the *T*-matching circuit the current intensity is concentrated at the shorting pin of the matching circuit in the feedline. The current however is much stronger over the patch compared to the case with no matching network.

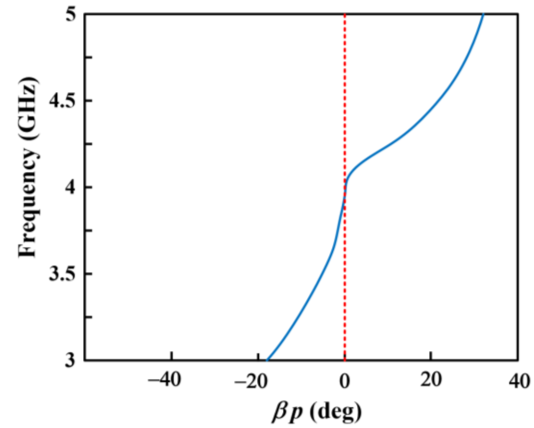


Fig.9. Dispersion diagram of the CRLH-TL feedline antenna.

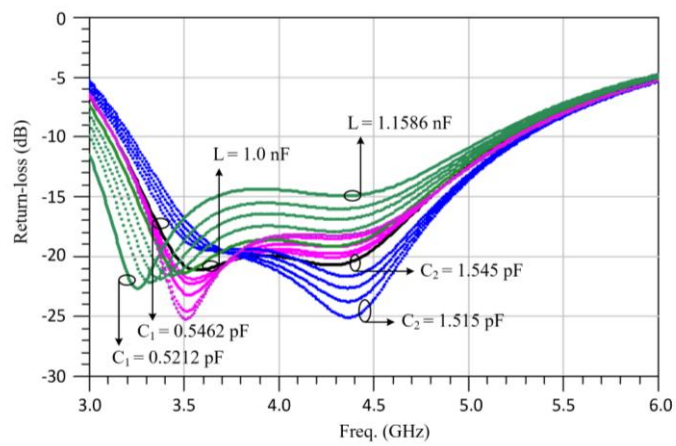


Fig.10. Effect on the return-loss response by the *T*-matching network parameters.

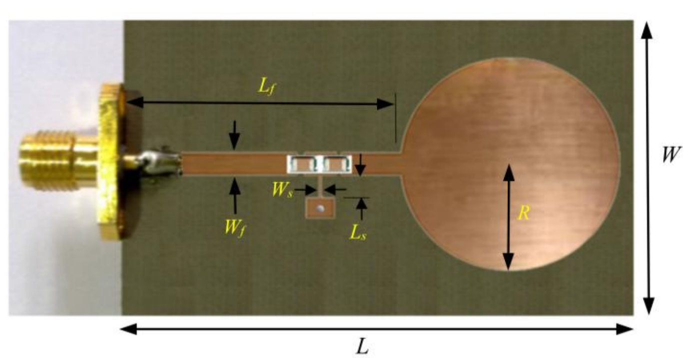
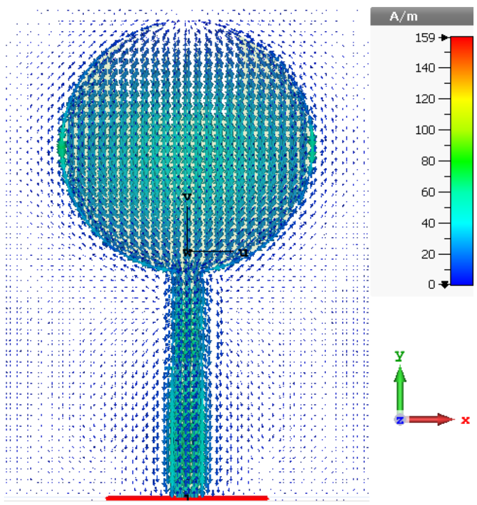


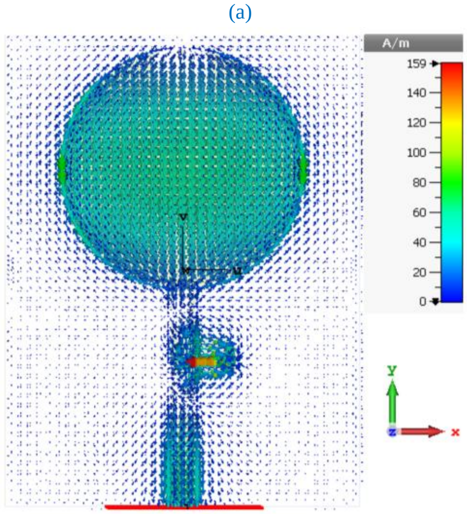
Fig.11. Photograph of the microstrip antenna incorporating the *T*-matching circuit.

Table I. Antenna's geometrical parameters

Parameters	Size (mm)
L	54
W	29
R	11.2
L_f	28
W_f	2.47
L_s	2.3
W_s	0.5
Radius of shunt via-hole	0.5
Thickness of FR-4 substrate	1.6



(a)



(b)

Fig.12. Surface current distribution over the antenna at 4 GHz, (a) without the *T*-matching circuit, and (b) with the T-matching circuit.

The measured radiation gain and efficiency performance without and with the *T*-matching circuit are shown in Fig.13. It is evident from these results that there is substantial improvement in the radiation gain and efficiency by embedding the metamaterial inspired matching circuit. Measured results show without the matching circuit the antenna has a peak gain and efficiency of 5.5 dBi and 59.8%, respectively, at 4 GHz. However, with the matching circuit the peak gain and peak efficiency are 8.2 dBi and 71.8%, respectively. Moreover, the gain and efficiency extend over a significantly larger bandwidth. The gain ≥6 dBi extend from 3.35-5.25 GHz, and the efficiency ≥60% extend from 3.25-5.1GHz.

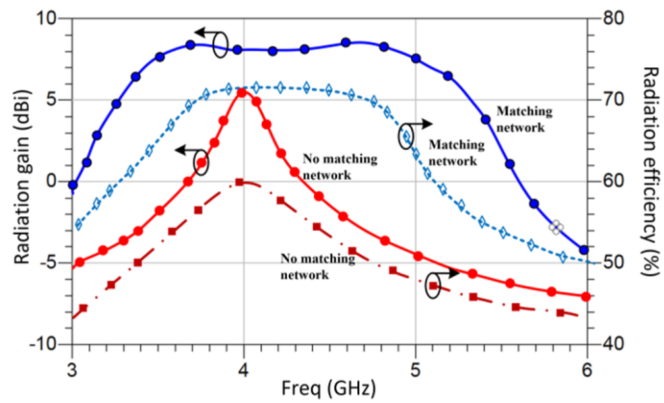
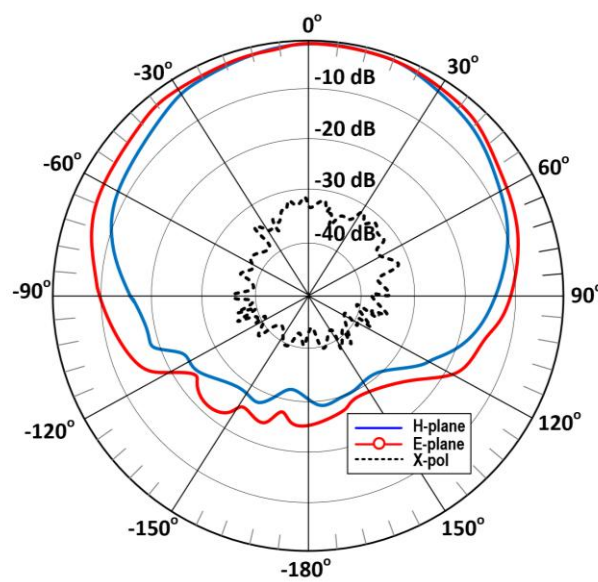


Fig.13. Measured radiation gain and efficiency response with and without the embedded *T*-matching circuit.

The radiation pattern in the E- and H-planes and the cross-polarization without and with the *T*-matching network at 4 GHz are shown in Fig. 14. The change in the radiation pattern with the *T*-matching network is marginal and, in both cases, the cross-polarization is less than -30 dB.



(a) With no *T*-matching circuit.

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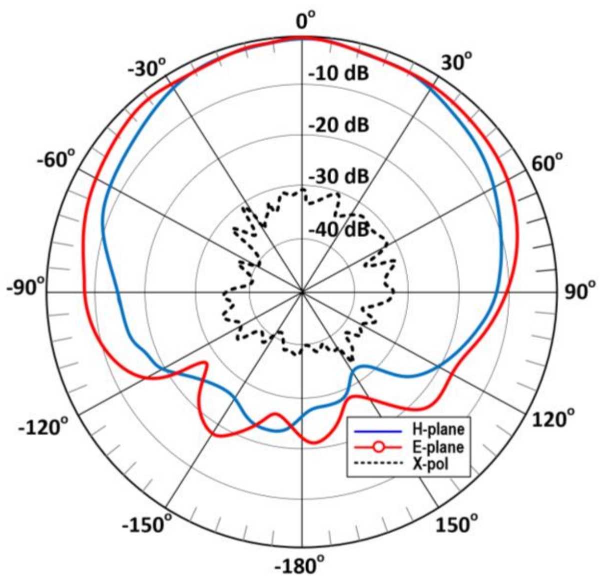

(b) With *T*-matching circuit.

Fig.14. The normalized measured radiation pattern in the E-plane and H-plane at 4 GHz, (a) without no matching circuit, and (b) with the embedded *T*-matching circuit.

Table II compares the present technique with some recently reported antenna bandwidth enhancement techniques. Impedance bandwidth enhancement of a microstrip patch antenna in [19] and [20] is achieved by loading the antenna with shorting pins and slots. In [21] the bandwidth is enhanced by simply curving the patch antenna in a partial cylindrical shape. In [22] one-dimensional electromagnetic bandgap ground structures and two-stage beam directors are employed in an inverted-L antenna topology to enhance the bandwidth. Even though the radiation efficiency of the proposed antenna is relatively low compared with the other techniques it is evident from the table that the proposed technique has a comparable gain and exhibits a substantially larger impedance bandwidth. Moreover, its size is relatively small with the exception of [22].

Table II. Comparison of the proposed matching technique of recently reported techniques.

Ref.	Technology	Antenna size	Fractional bandwidth (%)	Peak Gain (dBi)	Peak Eff. (%)
19	Shorting pins & slots	$1.14\lambda_o \times 1.27\lambda_o \times 0.009\lambda_o$	10	11.8	94
20	Shorting pins & slots	$1.02\lambda_o \times 1.47\lambda_o \times 0.045\lambda_o$	8.8 & 4.9	6 & 8	93 & 95
21	Microstrip	$0.58\lambda_o \times 0.58\lambda_o \times 0.016\lambda_o$	9	6.28	83.5
22	One-dimensional EM bandgap ground structures & two-stage beam directors	$0.21\lambda_o \times 0.32\lambda_o \times 0.009\lambda_o$	32.5	7	95
23	Differential microstrip	$0.8\lambda_o \times 0.8\lambda_o \times 0.013\lambda_o$	18.7	8.5	80
24	Magneto-electric dipole	$0.97\lambda_o \times 0.97\lambda_o \times 0.17\lambda_o$	28.2	9.0	-
25	Shorting pin & slot	$1.17\lambda_o \times 1.17\lambda_o \times 0.055\lambda_o$	32.2	6.5	93
26	Aperture-fed with shorting pins	$1.63\lambda_o \times 0.82\lambda_o \times 0.017\lambda_o$	15.2	6.8	-
This work	Metamaterial feedline	$0.67\lambda_o \times 0.38\lambda_o \times 0.021\lambda_o$	52.5	8.2	71.8

V. CONCLUSION

Impedance bandwidth of an antenna is determined by the matching conditions between the RF transceiver front-end and the antenna. We have demonstrated the effectiveness of employing an embedded metamaterial-based *T*-matching circuit directly implemented inside the feedline to enhance the antenna's impedance bandwidth. The measured results confirm significant improvement in the antenna's impedance bandwidth over its operating frequency range. Moreover, with the proposed technique improvement was observed in the radiation gain and efficiency too.

ACKNOWLEDGMENTS

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Impedance Bandwidth Improvement of a Planar Antenna Based on Metamaterial-Inspired T-Matching Network

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ABSTRACT In this paper a metamaterial-inspired T-matching network is directly imbedded inside the feedline of a microstrip antenna to realize optimum power transfer between the front-end of a RF wireless transceiver and the antenna. The proposed T-matching network, which is composed of an arrangement of series capacitor, shunt inductor, series capacitor, exhibits left-handed metamaterial characteristics. The matching network is first theoretically modelled to gain insight of its limitations. It was then implemented directly in the 50-Ω feedline to a standard circular patch antenna, which is an unconventional methodology. The antenna's performance was verified through measurements. With the proposed technique there is 2.7 dBi improvement in the antenna's radiation gain and 12% increase in the efficiency, and this is achieved over a significantly wider frequency range by a factor of approximately twenty. Moreover, there is good correlation between the theoretical model, method of moments simulation, and the measurement results.

INDEX TERMS *T*-matching circuit, microstrip antenna, metamaterial, transmission-line, impedance matching.

I. INTRODUCTION

Microstrip antennas have become popular for use in many wireless systems due to their planar profile, ease of fabrication and low manufacturing cost. Moreover, such antennas allow ease of integration with RF circuit components [1]. As a result, the use of planar based antenna configurations enables the integration of communication capabilities in a wide range of applications in the scope of the Internet of Things (IoT), such as context aware environments, Industry 4.0 or vehicular communications among others. This has been made possible by the design of flexible and compact transceivers especially in the sub-6 GHz band for wireless sensor networks, wireless local area networks, public land mobile networks and 5G communication systems. The vast number of applications which make use of wireless communication systems is leading to the need of employing different wireless systems in a cooperative way (i.e., Heterogeneous Network operation), as well as to an increase in the allocated channel bandwidth. Optimal performance of these wireless systems is given, among other factors, by adequate transceiver system operation. In relation with antenna sub-systems, impedance matching is one of the main factors to consider in order to reduce reflection-coefficient and hence, increase overall coverage range and minimize overall energy consumption [2-7].

The input impedance of an antenna is typically different from the impedance of the system to which it is connected, requiring impedance matching stages in order to minimize losses. In many cases, it is possible to realize impedance

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matching by modifying the antenna geometry [8]. However, the main limitation of conventional microstrip antennas is their narrow impedance bandwidth (typically less than 3% for return-loss better than -10 dB). In addition, electrically thin edge-fed patch antennas usually suffer from high resonant input impedance of a microstrip patch, typically in a range of 300–500 Ω, which is challenging to directly match with a 50-Ω microstrip feedline. To maximize power transfer to and from the antenna, and therefore enhance the radiation efficiency of such antennas different matching approaches have been investigated previously [9-11].

In many wireless applications the matching circuit is not always designed for optimum power transfer between the antenna and transceiver, but consideration is given to maximizing the power handling capability, reducing non-linear distortion and impedance bandwidth. In [12] the matching and impedance bandwidth enhancement is achieved by stacking the radiation patch with a parasitic element. With this technique bandwidth extension of ~19% is possible. In [13], the feeding scheme employed provides an impedance bandwidth of ~28% however the design of the feeding scheme is complex to implement in practice. In [14], a capacitively coupled probe is used to excite a nonradiative mode close to the antenna's dominant radiation mode to realize a bandwidth of ~28%. In [15] multiple resonant modes are simultaneously excited to extend the bandwidth of a patch antenna by about 32%, however this technique is shown to adversely affect the antenna's broadside radiation pattern. In [16] the bandwidth is improved by 15.2% by exciting dual modes.

In this paper a simple and effective metamaterial-inspired wideband matching technique is proposed to enhance the radiation gain and efficiency performance of a planar antenna. The proposed matching technique uses a *T*-matching network consisting of series capacitor, shunt inductor, series capacitor configuration to realize metamaterial characteristics. The matching network is directly embedded inside the feedline of the antenna, which is an unconventional methodology. The technique is first theoretically characterized to gain an understanding of its effectiveness and is then validated through practical design and measurement.

II. IMPEDANCE MATCHED ANTENNA

The three types of lumped element matching networks that are commonly used are based on the *L*-network, *π*-network, and *T*-network. The advantage of *L*-network is that it only has two reactive components whose values can be tuned easily for a given load impedance. However, its impedance matching capability is limited. On the other hand, the *π*- or *T*-networks have the capability of providing superior impedance matching flexibility than the *L*-network because they have three reactive components that can be tuned.

Fig.1 shows equivalent circuit of a typical *L*- and *T*-matching network. The reactance components can be either lumped elements or realized using transmission-lines based on microstrip integrated circuit technology. Capacitors can be interdigital or realized with low impedance microstrip-lines and vice versa for inductances. Reactance components realized using microstrip-lines however can result in a larger circuit.

The reactive components in the network control the return-loss between the antenna and matching network. In practice the capacitance values are adjusted or tuned until the return-loss of the antenna is lower than a specified threshold value, which is typically -10 dB, to (i) reduce system complexity, (ii) maintain the matching performance, and (iii) maximize transmission power to the antenna.

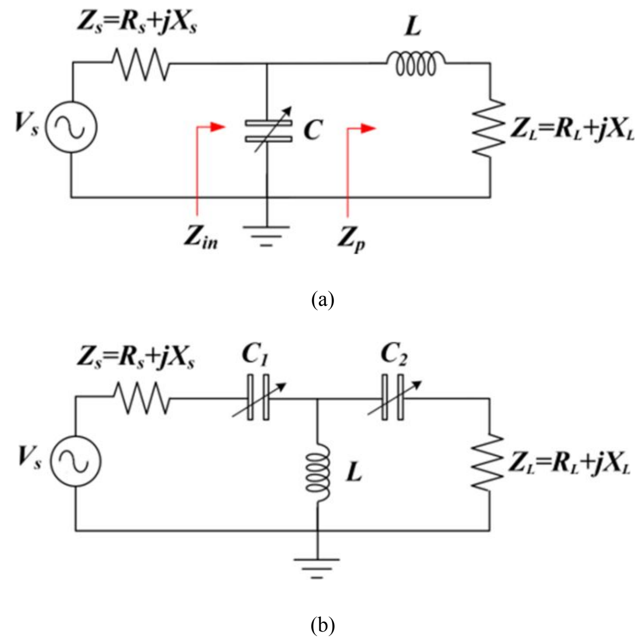

Fig.1. (a) Equivalent circuit of *L*-matching network, and (b) equivalent circuit of *T*-matching network.

To determine the effect of reactance components *C* and *L* on the input impedance Z_{in} of the *L*-network in Fig.1(a) the network is first represented in a simplified form in Fig.2, where the impedance Z_q accounts for inductance *L* and the load impedance $Z_L = R_L + jX_L$.

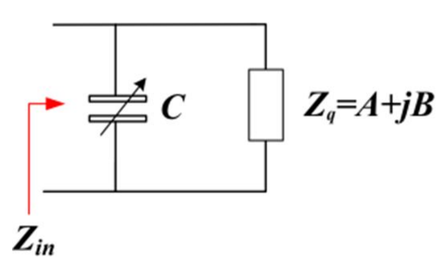

Fig.2. Impedance Z_{in} of the *L*-matching network in Fig.1.

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$$Z_{in} = \frac{\frac{1}{\omega C}(B - jA)}{A + j\left(B - \frac{1}{\omega C}\right)} = R + jX \qquad (1)$$

Where the real and imaginary impedances are:

$$R = {}^{A}/[(\omega AC)^2 + (\omega BC - 1)^2] \qquad (2)$$

$$X = [B - \omega C(A^2 + B^2)]/[(\omega AC)^2 + (\omega BC - 1)^2] \qquad (3)$$

By dividing (2) by (3) it can be shown that

$$\omega C = {}^{(BR - AX)}/{}_{R(A^2 + B^2)} \qquad (4)$$

Substituting (4) into (2) and (3), it can be shown that

$$[R - (A^2 + B^2)/2A]^2 + X^2 = [(A^2 + B^2)/2A]^2 \qquad (5)$$

where $A = R_L$ and $B = \omega L + X_L$

Eqn. (5) describes a circle whose center and radius, i.e. the impedance position, is a function of the inductive and capacitive reactance and the load impedance. This equation reveals that the center and radius of the impedance circle can be changed by tweaking either one or both inductance and capacitance in the circuit to effect matching. Unlike the *L*-network a *T*-network can provide superior impedance matching because it has three reactive components that can be tuned. In the next section a *T*-matching circuit is investigated in the reduction of mismatch between the antenna and RF front-end circuit.

III. *T*-COMPENSATION CIRCUIT

In a real-world scenario, the matching network is connected to the antenna using a transmission-line. The proposed *T*-matching network in Fig.3 comprises an arrangement of series capacitor, shunt inductor, series capacitor that exhibits left-handed metamaterial characteristics [17]. The impedance Z_x seen by the signal from the transmitter is a combination of the impedance Z_1 of the transmission-line of length l and the impedance of the antenna load Z_L. Hence, to optimize power transfer between the transmitter to the antenna, we need to conjugately impedance match the impedance Z_{in} looking into the matching circuit to Z_x.

The impedance Z_x looking into the antenna load of impedance Z_L via the transmission-line of length l of impedance Z_1 is given by

$$Z_x = Z_1\left[\frac{Z_L + jZ_1 tan\beta l}{Z_1 + jZ_L tan\beta l}\right] \qquad (6)$$

$$Z_y = \frac{\frac{L}{C_2} + j\omega L Z_x}{Z_x + j\left(\frac{\omega^2 LC_2 - 1}{\omega C_2}\right)} \qquad (7)$$

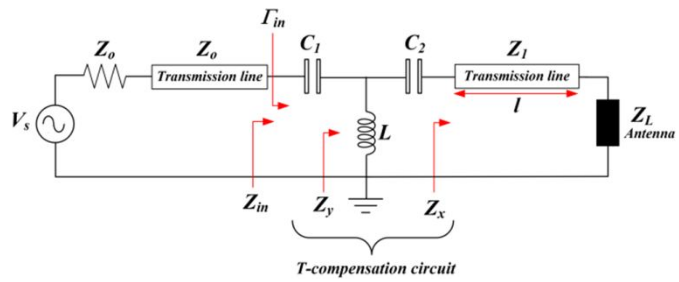


Fig.3. Equivalent circuit model of the *T*-matching network inserted between the source and the antenna.

Input impedance Z_{in} of the *T*-compensation circuit, shown in Fig.3, can be represented as $Z_{in} = R_{in} + jX_{in}$, where R_{in} and X_{in} can be determined to be given by

$$R_{in} = \frac{\frac{xL}{t}\left(\frac{1}{C_2} - \frac{\omega y}{t}\right) + \frac{\omega x L}{t}\left[\frac{y}{t} + \left(\frac{\omega^2 LC_2 - 1}{\omega C_2}\right)\right]}{\left(\frac{x}{t}\right)^2 + \left[\frac{y}{t} + \left(\frac{\omega^2 LC_2 - 1}{\omega C_2}\right)\right]^2} \qquad (8)$$

$$X_{in} = \frac{\omega L\left(\frac{x}{t}\right)^2 - L\left(\frac{1}{C_2} - \frac{\omega y}{t}\right)\left[\frac{y}{t} + \left(\frac{\omega^2 LC_2 - 1}{\omega C_2}\right)\right]}{\left(\frac{x}{t}\right)^2 + \left[\frac{y}{t} + \left(\frac{\omega^2 LC_2 - 1}{\omega C_2}\right)\right]^2} - \omega C_1 \qquad (9)$$

where

$$x = Z_L Z_1(1 + tan^2\beta l)$$

$$y = tan^2\beta l(Z_1^2 - Z_L^2)$$

$$t = Z_1^2 + Z_L^2 tan^2\beta l$$

Reflection-coefficient (Γ_{in}) at the input port of the antenna is a measure of how much power is reflected from the antenna. It determines how much power is radiated by the antenna and therefore the radiation efficiency of the antenna, which is defined as a ratio of radiated power to input power. The reflection-coefficient looking at the antenna via the *T*-compensation circuit is given by:

$$\Gamma_{in} = \frac{R_{in}^2 + X_{in}^2 - Z_o^2 - jR_{in}}{(R_{in} + Z_o)^2 + X_{in}^2} \qquad (10)$$

where R_{in} and X_{in} are defined by Eqns. (8) and (9). The reflection-coefficient determines the maximum power transfer to the antenna. Eqn.(10) indicates that the center frequency and bandwidth of the matching system can be controlled by simply using appropriate values of the capacitances C_1 and C_2 and inductance L. For a T-matching circuit the relationship between bandwidth and reflection-coefficient is derived in [18] to be given by

$$BW \approx \frac{\sqrt{|\Gamma_{in}|}}{1-|\Gamma_{in}|} \qquad (11)$$

Eqn.(11) indicates the by appropriately selecting R_{in} and X_{in} the bandwidth of the matching network can be optimized.

The basic T-match circuit is shown in Fig.4. Like in π-match, T-matching circuit allows to set impedance transformation ratio and Q-factor/matching bandwidth of circuit independently.

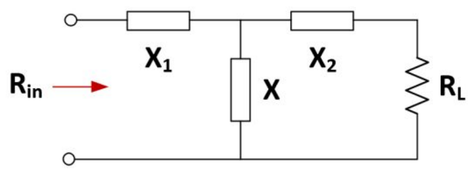


Fig.4. *T*-matching circuit.

$$Q_1 = \frac{X_1}{R_{in}} \qquad Q_2 = \frac{X_2}{R_L} \qquad (12)$$

The T-matching circuit can be decomposed to two L-match circuits, as shown in Fig.5. By series-to-parallel transformation as shown in From Fig.6,

$$Q_1 = \frac{R_I}{X_A} \qquad Q_2 = \frac{R_I}{X_B} \qquad (13)$$

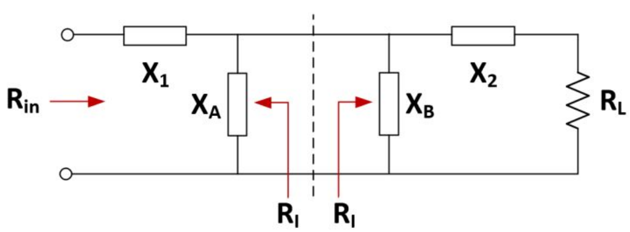


Fig.5. *T*-matching circuit decomposed to two *L*-match circuits.

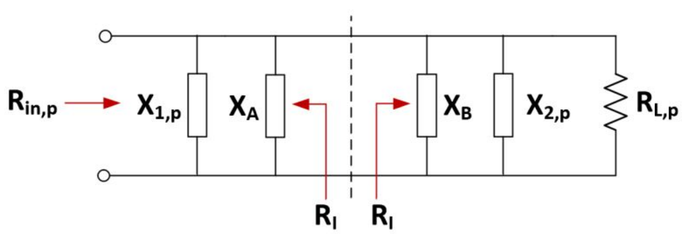


Fig.6. *T*-matching after series to parallel transformation.

$$R_{in,p} = R_I = R_{in}(1+Q_1^2) \qquad R_{L,p} = R_I = R_L\left(1+\frac{1}{Q_2^2}\right) \qquad (14)$$

$$X_{1,p} = X_I\left(1+\frac{1}{Q_1^2}\right) \qquad X_{2,p} = X_I\left(1+\frac{1}{Q_2^2}\right) \qquad (15)$$

Rearranging Eqn.(14)

$$Q_1 = \sqrt{\frac{R_I}{R_{in}} - 1} \qquad Q_2 = \sqrt{\frac{R_I}{R_L} - 1} \quad (16)$$

The total Q of the circuit is given by

$$Q = Q_1 + Q_2 = \sqrt{\frac{R_I}{R_{in}} - 1} + \sqrt{\frac{R_I}{R_{in}} - 1} \qquad (17)$$

At resonant frequency, $|X_A| = |X_{1,p}|$ and $|X_B| = |X_{2,p}|$.

Therefore, $X_A = X_1\left(1+\frac{1}{Q_1^2}\right) \qquad X_B = X_2\left(1+\frac{1}{Q_2^2}\right)$

The above relationships allow the design of the preliminary T-matching circuit. The design procedure involves: (i) Finding R_I using Eqn.(17) from given R_L, R_{in} and Q; (ii) Calculating Q_1 and Q_2 using Eqn.(16); (iii) Finding X_1 and X_2 using Eqn.(12), and (iv) Finding X_A and X_B using Eqn.(13).

In the next section it is shown that with proper selection of C_1, C_2, and L a high gain and high radiation efficiency can be obtained over a wide frequency band. However, this does introduce some complexity in the design as each variable capacitor needs to be controlled separately. Nevertheless, the proposed technique is amenable to electronic tuning hence allowing controllable matching capability according to system requirements.

IV. DESIGN EXAMPLE USING METAMATERIAL INSPIRED *T*-MATCHING NETWORK

A standard circular microstrip antenna was designed at 4 GHz using FR-4 lossy substrate with a relative permittivity of 4.3, thickness of 1.6 mm, and tanδ of 0.025. The radius of the circular patch is 10.9 mm. The patch antenna is excited using a microstrip feedline. The ground-plane on the backside of the substrate is truncated. There is no ground-plane in the region of the patch. Fig.7(a) shows the equivalent circuit model of the feedline and the patch antenna. The return-loss of the antenna predicted by theory and simulation in Fig.8 show it resonates at 4 GHz with a high-Q and hence has a narrow impedance bandwidth. The measured results show a bandwidth of 0.1 GHz for $S_{11} \leq$-10 dB with an optimum return-loss value of -12 dB at 4 GHz.

Microstrip feedline length determines the input impedance of the antenna. Hence, a T-matching circuit is directly embedded in the feedline, which is an unconventional technique. Simply put, the proposed T-matching network is not an individual part from the antenna structure. The matching circuit used here is a metamaterial structure that exhibits left-handed properties of negative refractive index when interacts with electromagnetic signals [17]. The metamaterial characteristics are realized here with a T-network composed of series capacitance, shunt inductance, and series capacitance circuit, as per schematic diagram in Fig.7(b). The dispersion diagram of the antenna with the matching circuit in Fig.9 show the phase variation is virtually zero at 4 GHz.

The return-loss response in Fig.8 by theory, simulation, and measurement show that with the inclusion of the metamaterial matching network the antenna's impedance bandwidth is significantly improved. It is evident from Fig.8 that magnitudes of reactance components constituting the *T*-matching circuit has a great influence the Q-factor of the antenna. The results of the parametric study in Fig.10 show how the parameters constituting the *T*-matching circuit effect the return-loss response. The optimum values are: C_1 = 0.5462 pF, C_2 = 1.545 pF and L = 1.1586 pF. The fabricated antenna with the matching network is shown in Fig.11. Its measured return-loss in Fig.8 is better than -18 dB at 4 GHz. The magnitude of the geometrical parameters corresponding to the lumped element values, which are annotated in Fig.11, are given in Table I.

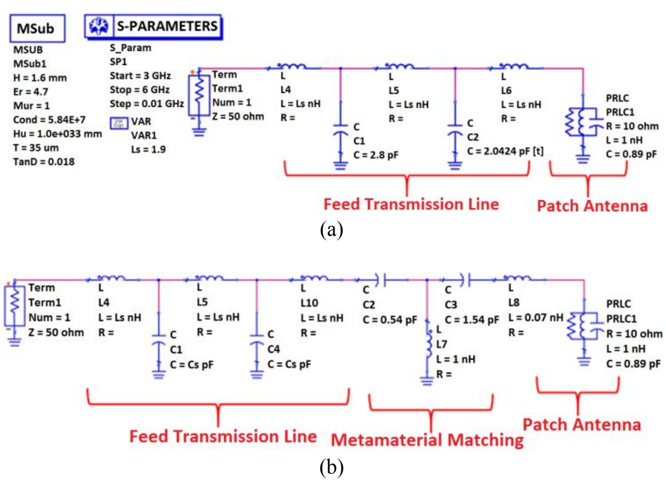


Fig.7. (a) Equivalent circuit model of feedline and patch antenna, and (b) circular patch antenna loaded with metamaterial matching network using Keysight Technologies' Advance Design System (ADS).

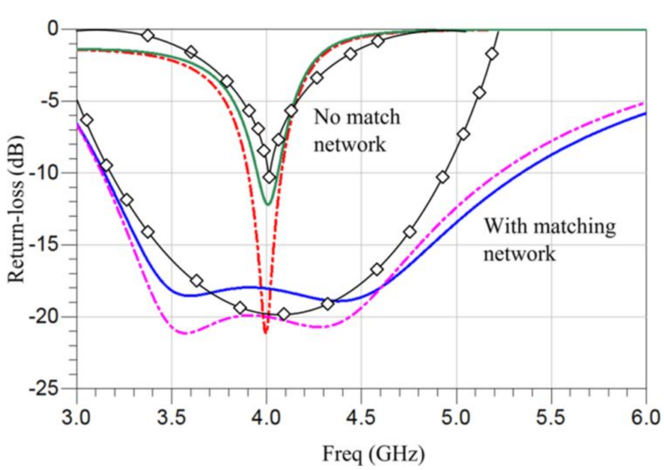


Fig.8. Theoretical, simulated and measured return-loss response of the antenna without and with the metamaterial-inspired matching network. Black diamond studded line is prediction by theory, dotted lines indicate the simulated result, and solid lines represent the measured result.

Simulated current distributions without and with the *T*-matching circuit at 4 GHz is shown in Fig.12. It is evident that the current distribution with the *T*-matching circuit the current intensity is concentrated at the shorting pin of the matching circuit in the feedline. The current however is much stronger over the patch compared to the case with no matching network.

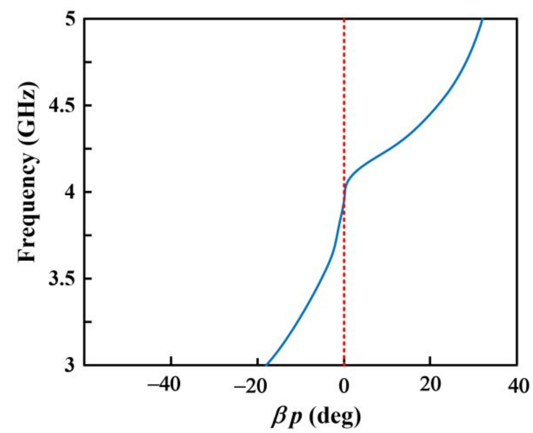


Fig.9. Dispersion diagram of the CRLH-TL feedline antenna.

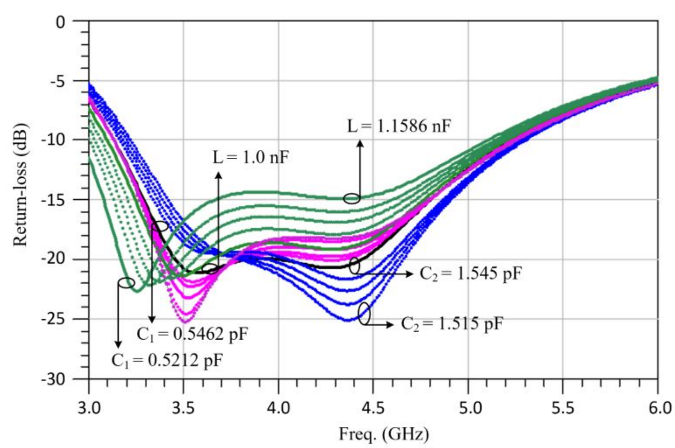


Fig.10. Effect on the return-loss response by the *T*-matching network parameters.

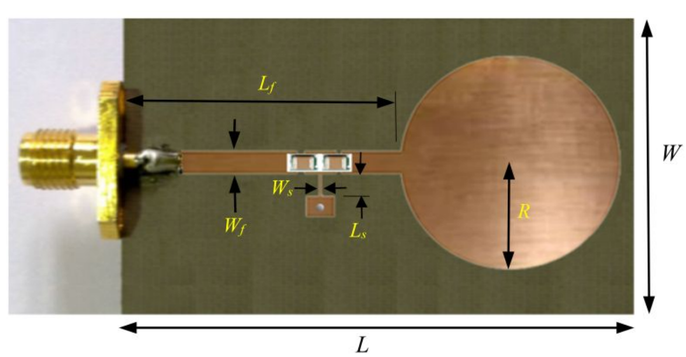
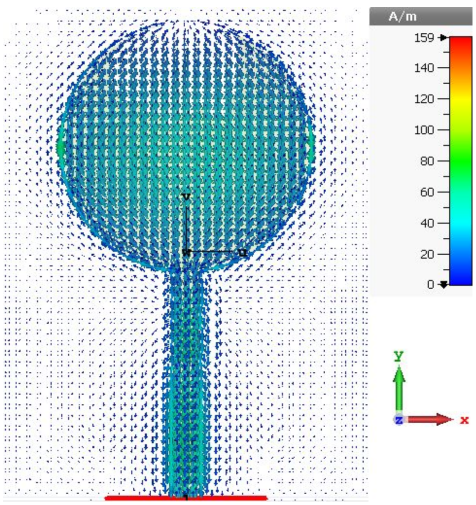


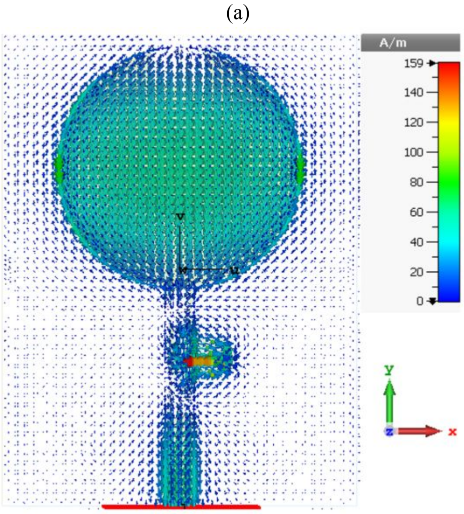
Fig.11. Photograph of the microstrip antenna incorporating the *T*-matching circuit.

Table I. Antenna's geometrical parameters

Parameters	Size (mm)
L	54
W	29
R	11.2
L_f	28
W_f	2.47
L_s	2.3
W_s	0.5
Radius of shunt via-hole	0.5
Thickness of FR-4 substrate	1.6



(a)



(b)

Fig.12. Surface current distribution over the antenna at 4 GHz, (a) without the *T*-matching circuit, and (b) with the T-matching circuit.

The measured radiation gain and efficiency performance without and with the *T*-matching circuit are shown in Fig.13. It is evident from these results that there is substantial improvement in the radiation gain and efficiency by embedding the metamaterial inspired matching circuit. Measured results show without the matching circuit the antenna has a peak gain and efficiency of 5.5 dBi and 59.8%, respectively, at 4 GHz. However, with the matching circuit the peak gain and peak efficiency are 8.2 dBi and 71.8%, respectively. Moreover, the gain and efficiency extend over a significantly larger bandwidth. The gain ≥6 dBi extend from 3.35-5.25 GHz, and the efficiency ≥60% extend from 3.25-5.1GHz.

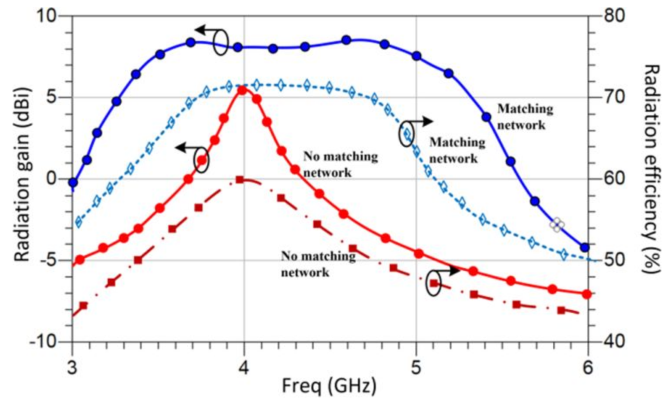
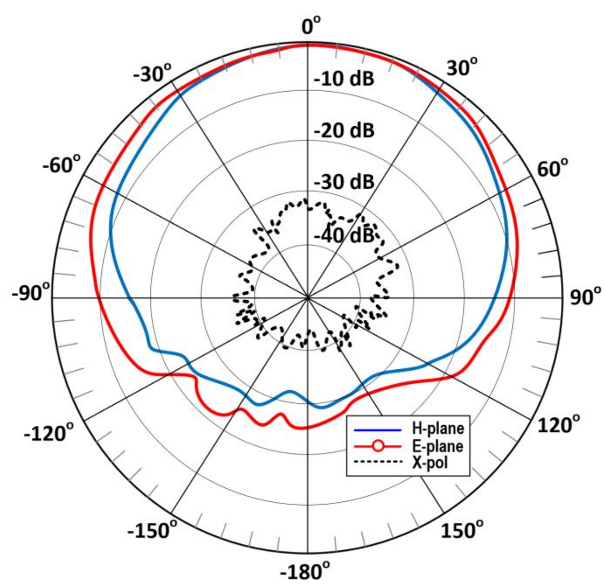
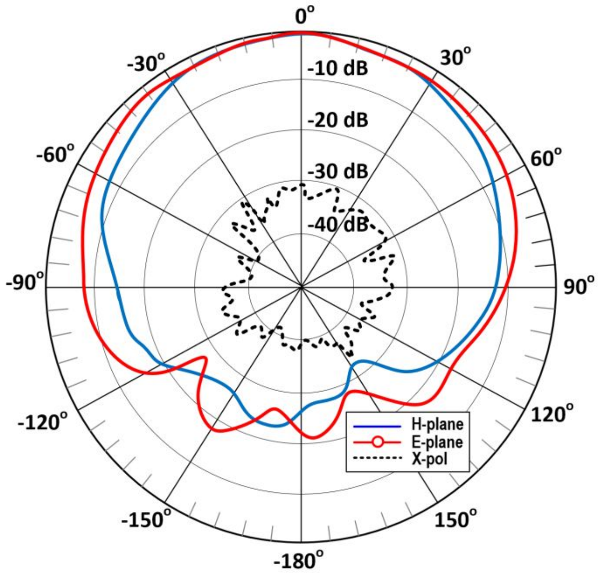


Fig.13. Measured radiation gain and efficiency response with and without the embedded *T*-matching circuit.

The radiation pattern in the E- and H-planes and the cross-polarization without and with the *T*-matching network at 4 GHz are shown in Fig. 14. The change in the radiation pattern with the *T*-matching network is marginal and, in both cases, the cross-polarization is less than -30 dB.



(a) With no *T*-matching circuit.



(b) With *T*-matching circuit.

Fig.14. The normalized measured radiation pattern in the E-plane and H-plane at 4 GHz, (a) without no matching circuit, and (b) with the embedded *T*-matching circuit.

Table II compares the present technique with some recently reported antenna bandwidth enhancement techniques. Impedance bandwidth enhancement of a microstrip patch antenna in [19] and [20] is achieved by loading the antenna with shorting pins and slots. In [21] the bandwidth is enhanced by simply curving the patch antenna in a partial cylindrical shape. In [22] one-dimensional electromagnetic bandgap ground structures and two-stage beam directors are employed in an inverted-L antenna topology to enhance the bandwidth. Even though the radiation efficiency of the proposed antenna is relatively low compared with the other techniques it is evident from the table that the proposed technique has a comparable gain and exhibits a substantially larger impedance bandwidth. Moreover, its size is relatively small with the exception of [22].

Table II. Comparison of the proposed matching technique of recently reported techniques.

Ref.	Technology	Antenna size	Fractional bandwidth (%)	Peak Gain (dBi)	Peak Eff. (%)
19	Shorting pins & slots	$1.14\lambda_o \times 1.27\lambda_o \times 0.009\lambda_o$	10	11.8	94
20	Shorting pins & slots	$1.02\lambda_o \times 1.47\lambda_o \times 0.045\lambda_o$	8.8 & 4.9	6 & 8	93 & 95
21	Microstrip	$0.58\lambda_o \times 0.58\lambda_o \times 0.016\lambda_o$	9	6.28	83.5
22	One-dimensional EM bandgap ground structures & two-stage beam directors	$0.21\lambda_o \times 0.32\lambda_o \times 0.009\lambda_o$	32.5	7	95
23	Differential microstrip	$0.8\lambda_o \times 0.8\lambda_o \times 0.013\lambda_o$	18.7	8.5	80
24	Magneto-electric dipole	$0.97\lambda_o \times 0.97\lambda_o \times 0.17\lambda_o$	28.2	9.0	-
25	Shorting pin & slot	$1.17\lambda_o \times 1.17\lambda_o \times 0.055\lambda_o$	32.2	6.5	93
26	Aperture-fed with shorting pins	$1.63\lambda_o \times 0.82\lambda_o \times 0.017\lambda_o$	15.2	6.8	-
This work	Metamaterial feedline	$0.67\lambda_o \times 0.38\lambda_o \times 0.021\lambda_o$	52.5	8.2	71.8

V. CONCLUSION

Impedance bandwidth of an antenna is determined by the matching conditions between the RF transceiver front-end and the antenna. We have demonstrated the effectiveness of employing an embedded metamaterial-based *T*-matching circuit directly implemented inside the feedline to enhance the antenna's impedance bandwidth. The measured results confirm significant improvement in the antenna's impedance bandwidth over its operating frequency range. Moreover, with the proposed technique improvement was observed in the radiation gain and efficiency too.

ACKNOWLEDGMENTS

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Impedance Bandwidth Improvement of a Planar Antenna Based on Metamaterial-Inspired T-Matching Network

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ABSTRACT **In this paper a metamaterial-inspired T-matching network is directly imbedded inside the feedline of a microstrip antenna to realize optimum power transfer between the front-end of a RF wireless transceiver and the antenna. The proposed T-matching network, which is composed of an arrangement of series capacitor, shunt inductor, series capacitor, exhibits left-handed metamaterial characteristics. The matching network is first theoretically modelled to gain insight of its limitations. It was then implemented directly in the 50-Ω feedline to a standard circular patch antenna, which is an unconventional methodology. The antenna's performance was verified through measurements. With the proposed technique there is 2.7 dBi improvement in the antenna's radiation gain and 12% increase in the efficiency, and this is achieved over a significantly wider frequency range by a factor of approximately twenty. Moreover, there is good correlation between the theoretical model, method of moments simulation, and the measurement results.**

INDEX TERMS ***T*-matching circuit, microstrip antenna, metamaterial, transmission-line, impedance matching.**

I. INTRODUCTION

Microstrip antennas have become popular for use in many wireless systems due to their planar profile, ease of fabrication and low manufacturing cost. Moreover, such antennas allow ease of integration with RF circuit components [1]. As a result, the use of planar based antenna configurations enables the integration of communication capabilities in a wide range of applications in the scope of the Internet of Things (IoT), such as context aware environments, Industry 4.0 or vehicular communications among others. This has been made possible by the design of flexible and compact transceivers especially in the sub-6 GHz band for wireless sensor networks, wireless local area networks, public land mobile networks and 5G communication systems. The vast number of applications which make use of wireless communication systems is leading to the need of employing different wireless systems in a cooperative way (i.e., Heterogeneous Network operation), as well as to an increase in the allocated channel bandwidth. Optimal performance of these wireless systems is given, among other factors, by adequate transceiver system operation. In relation with antenna sub-systems, impedance matching is one of the main factors to consider in order to reduce reflection-coefficient and hence, increase overall coverage range and minimize overall energy consumption [2-7].

The input impedance of an antenna is typically different from the impedance of the system to which it is connected, requiring impedance matching stages in order to minimize losses. In many cases, it is possible to realize impedance

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matching by modifying the antenna geometry [8]. However, the main limitation of conventional microstrip antennas is their narrow impedance bandwidth (typically less than 3% for return-loss better than -10 dB). In addition, electrically thin edge-fed patch antennas usually suffer from high resonant input impedance of a microstrip patch, typically in a range of 300–500 Ω, which is challenging to directly match with a 50-Ω microstrip feedline. To maximize power transfer to and from the antenna, and therefore enhance the radiation efficiency of such antennas different matching approaches have been investigated previously [9-11].

In many wireless applications the matching circuit is not always designed for optimum power transfer between the antenna and transceiver, but consideration is given to maximizing the power handling capability, reducing non-linear distortion and impedance bandwidth. In [12] the matching and impedance bandwidth enhancement is achieved by stacking the radiation patch with a parasitic element. With this technique bandwidth extension of ~19% is possible. In [13], the feeding scheme employed provides an impedance bandwidth of ~28% however the design of the feeding scheme is complex to implement in practice. In [14], a capacitively coupled probe is used to excite a nonradiative mode close to the antenna's dominant radiation mode to realize a bandwidth of ~28%. In [15] multiple resonant modes are simultaneously excited to extend the bandwidth of a patch antenna by about 32%, however this technique is shown to adversely affect the antenna's broadside radiation pattern. In [16] the bandwidth is improved by 15.2% by exciting dual modes.

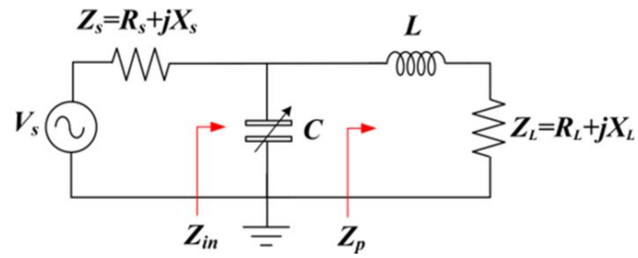
In this paper a simple and effective metamaterial-inspired wideband matching technique is proposed to enhance the radiation gain and efficiency performance of a planar antenna. The proposed matching technique uses a *T*-matching network consisting of series capacitor, shunt inductor, series capacitor configuration to realize metamaterial characteristics. The matching network is directly embedded inside the feedline of the antenna, which is an unconventional methodology. The technique is first theoretically characterized to gain an understanding of its effectiveness and is then validated through practical design and measurement.

II. IMPEDANCE MATCHED ANTENNA

The three types of lumped element matching networks that are commonly used are based on the *L*-network, *π*-network, and *T*-network. The advantage of *L*-network is that it only has two reactive components whose values can be tuned easily for a given load impedance. However, its impedance matching capability is limited. On the other hand, the *π*- or *T*-networks have the capability of providing superior impedance matching flexibility than the *L*-network because they have three reactive components that can be tuned.

Fig.1 shows equivalent circuit of a typical *L*- and *T*-matching network. The reactance components can be either lumped elements or realized using transmission-lines based on microstrip integrated circuit technology. Capacitors can be interdigital or realized with low impedance microstrip-lines and vice versa for inductances. Reactance components realized using microstrip-lines however can result in a larger circuit.

The reactive components in the network control the return-loss between the antenna and matching network. In practice the capacitance values are adjusted or tuned until the return-loss of the antenna is lower than a specified threshold value, which is typically -10 dB, to (i) reduce system complexity, (ii) maintain the matching performance, and (iii) maximize transmission power to the antenna.



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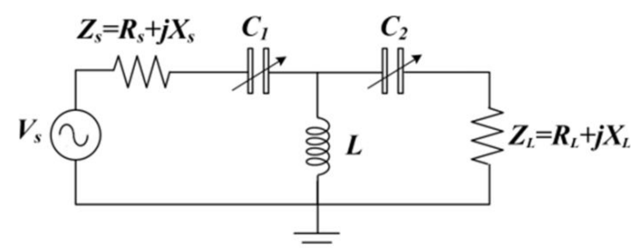

(b)

Fig.1. (a) Equivalent circuit of *L*-matching network, and (b) equivalent circuit of *T*-matching network.

To determine the effect of reactance components C and L on the input impedance Z_{in} of the *L*-network in Fig.1(a) the network is first represented in a simplified form in Fig.2, where the impedance Z_q accounts for inductance L and the load impedance $Z_L = R_L + jX_L$.

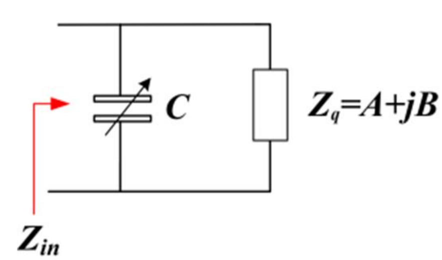

Fig.2. Impedance Z_{in} of the *L*-matching network in Fig.1.

$$Z_{in} = \frac{\frac{1}{\omega C}(B - jA)}{A + j\left(B - \frac{1}{\omega C}\right)} = R + jX \quad (1)$$

Where the real and imaginary impedances are:

$$R = {}^{A}/[(\omega AC)^2 + (\omega BC - 1)^2] \quad (2)$$

$$X = [B - \omega C(A^2 + B^2)]/[(\omega AC)^2 + (\omega BC - 1)^2] \quad (3)$$

By dividing (2) by (3) it can be shown that

$$\omega C = {}^{(BR - AX)}/R(A^2 + B^2) \quad (4)$$

Substituting (4) into (2) and (3), it can be shown that

$$[R - (A^2 + B^2)/2A]^2 + X^2 = [(A^2 + B^2)/2A]^2 \quad (5)$$

where $A = R_L$ and $B = \omega L + X_L$

Eqn. (5) describes a circle whose center and radius, i.e. the impedance position, is a function of the inductive and capacitive reactance and the load impedance. This equation reveals that the center and radius of the impedance circle can be changed by tweaking either one or both inductance and capacitance in the circuit to effect matching. Unlike the *L*-network a *T*-network can provide superior impedance matching because it has three reactive components that can be tuned. In the next section a *T*-matching circuit is investigated in the reduction of mismatch between the antenna and RF front-end circuit.

III. *T*-COMPENSATION CIRCUIT

In a real-world scenario, the matching network is connected to the antenna using a transmission-line. The proposed *T*-matching network in Fig.3 comprises an arrangement of series capacitor, shunt inductor, series capacitor that exhibits left-handed metamaterial characteristics [17]. The impedance Z_x seen by the signal from the transmitter is a combination of the impedance Z_1 of the transmission-line of length l and the impedance of the antenna load Z_L. Hence, to optimize power transfer between the transmitter to the antenna, we need to conjugately impedance match the impedance Z_{in} looking into the matching circuit to Z_x.

The impedance Z_x looking into the antenna load of impedance Z_L via the transmission-line of length l of impedance Z_1 is given by

$$Z_x = Z_1\left[\frac{Z_L + jZ_1 tan\beta l}{Z_1 + jZ_L tan\beta l}\right] \quad (6)$$

$$Z_y = \frac{\frac{L}{C_2} + j\omega L Z_x}{Z_x + j\left(\frac{\omega^2 L C_2 - 1}{\omega C_2}\right)} \quad (7)$$

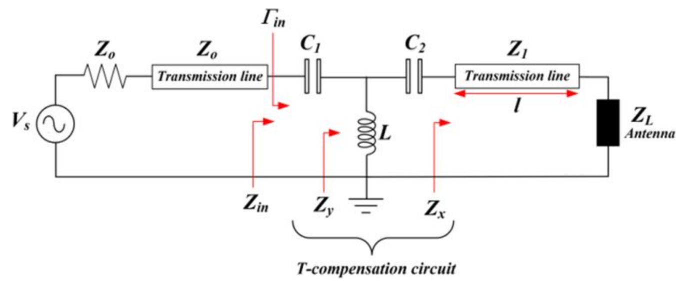

Fig.3. Equivalent circuit model of the *T*-matching network inserted between the source and the antenna.

Input impedance Z_{in} of the *T*-compensation circuit, shown in Fig.3, can be represented as $Z_{in} = R_{in} + jX_{in}$, where R_{in} and X_{in} can be determined to be given by

$$R_{in} = \frac{\frac{xL}{t}\left(\frac{1}{C_2} - \frac{\omega y}{t}\right) + \frac{\omega x L}{t}\left[\frac{y}{t} + \left(\frac{\omega^2 L C_2 - 1}{\omega C_2}\right)\right]}{\left(\frac{x}{t}\right)^2 + \left[\frac{y}{t} + \left(\frac{\omega^2 L C_2 - 1}{\omega C_2}\right)\right]^2} \quad (8)$$

$$X_{in} = \frac{\omega L\left(\frac{x}{t}\right)^2 - L\left(\frac{1}{C_2} - \frac{\omega y}{t}\right)\left[\frac{y}{t} + \left(\frac{\omega^2 L C_2 - 1}{\omega C_2}\right)\right]}{\left(\frac{x}{t}\right)^2 + \left[\frac{y}{t} + \left(\frac{\omega^2 L C_2 - 1}{\omega C_2}\right)\right]^2} - \omega C_1 \quad (9)$$

where

$$x = Z_L Z_1(1 + tan^2\beta l)$$

$$y = tan^2\beta l(Z_1^2 - Z_L^2)$$

$$t = Z_1^2 + Z_L^2 tan^2\beta l$$

Reflection-coefficient (Γ_{in}) at the input port of the antenna is a measure of how much power is reflected from the antenna. It determines how much power is radiated by the antenna and therefore the radiation efficiency of the antenna, which is defined as a ratio of radiated power to input power. The reflection-coefficient looking at the antenna via the *T*-compensation circuit is given by:

$$\Gamma_{in} = \frac{R_{in}^2 + X_{in}^2 - Z_o^2 - jR_{in}}{(R_{in} + Z_o)^2 + X_{in}^2} \quad (10)$$

where R_{in} and X_{in} are defined by Eqns. (8) and (9). The reflection-coefficient determines the maximum power transfer to the antenna. Eqn.(10) indicates that the center frequency and bandwidth of the matching system can be controlled by simply using appropriate values of the capacitances C_1 and C_2 and inductance L. For a *T*-matching circuit the relationship between bandwidth and reflection-coefficient is derived in [18] to be given by

$$BW \approx \frac{\sqrt{|\Gamma_{in}|}}{1-|\Gamma_{in}|} \qquad (11)$$

Eqn.(11) indicates the by appropriately selecting R_{in} and X_{in} the bandwidth of the matching network can be optimized.

The basic *T*-match circuit is shown in Fig.4. Like in π-match, *T*-matching circuit allows to set impedance transformation ratio and Q-factor/matching bandwidth of circuit independently.

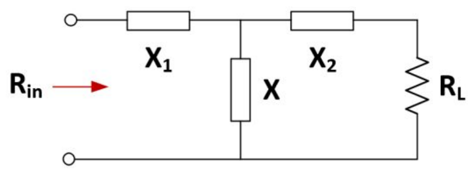

Fig.4. *T*-matching circuit.

$$Q_1 = \frac{X_1}{R_{in}} \qquad Q_2 = \frac{X_2}{R_L} \qquad (12)$$

The *T*-matching circuit can be decomposed to two *L*-match circuits, as shown in Fig.5. By series-to-parallel transformation as shown in From Fig.6,

$$Q_1 = \frac{R_I}{X_A} \qquad Q_2 = \frac{R_I}{X_B} \qquad (13)$$

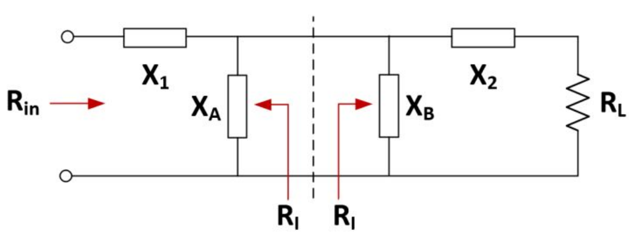

Fig.5. *T*-matching circuit decomposed to two *L*-match circuits.

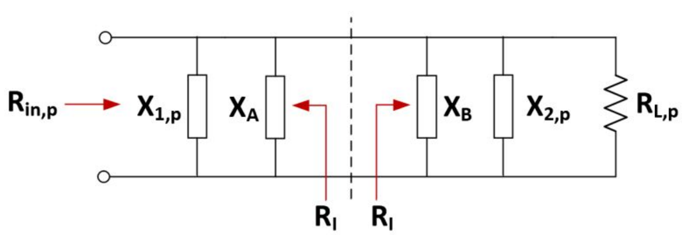

Fig.6. *T*-matching after series to parallel transformation.

$$R_{in,p} = R_I = R_{in}\left(1+Q_1^2\right) \qquad R_{L,p} = R_I = R_L\left(1+\frac{1}{Q_2^2}\right) \qquad (14)$$

$$X_{1,p} = X_I\left(1+\frac{1}{Q_1^2}\right) \qquad X_{2,p} = X_I\left(1+\frac{1}{Q_2^2}\right) \qquad (15)$$

Rearranging Eqn.(14)

$$Q_1 = \sqrt{\frac{R_I}{R_{in}}-1} \qquad Q_2 = \sqrt{\frac{R_I}{R_L}-1} \quad (16)$$

The total Q of the circuit is given by

$$Q = Q_1 + Q_2 = \sqrt{\frac{R_I}{R_{in}}-1} + \sqrt{\frac{R_I}{R_{in}}-1} \qquad (17)$$

At resonant frequency, $|X_A| = |X_{1,p}|$ and $|X_B| = |X_{2,p}|$.

Therefore, $X_A = X_1\left(1+\frac{1}{Q_1^2}\right) \qquad X_B = X_2\left(1+\frac{1}{Q_2^2}\right)$

The above relationships allow the design of the preliminary *T*-matching circuit. The design procedure involves: (i) Finding R_I using Eqn.(17) from given R_L, R_{in} and Q; (ii) Calculating Q_1 and Q_2 using Eqn.(16); (iii) Finding X_1 and X_2 using Eqn.(12), and (iv) Finding X_A and X_B using Eqn.(13).

In the next section it is shown that with proper selection of C_1, C_2, and L a high gain and high radiation efficiency can be obtained over a wide frequency band. However, this does introduce some complexity in the design as each variable capacitor needs to be controlled separately. Nevertheless, the proposed technique is amenable to electronic tuning hence allowing controllable matching capability according to system requirements.

IV. DESIGN EXAMPLE USING METAMATERIAL INSPIRED *T*-MATCHING NETWORK

A standard circular microstrip antenna was designed at 4 GHz using FR-4 lossy substrate with a relative permittivity of 4.3, thickness of 1.6 mm, and $\tan\delta$ of 0.025. The radius of the circular patch is 10.9 mm. The patch antenna is excited using a microstrip feedline. The ground-plane on the backside of the substrate is truncated. There is no ground-plane in the region of the patch. Fig.7(a) shows the equivalent circuit model of the feedline and the patch antenna. The return-loss of the antenna predicted by theory and simulation in Fig.8 show it resonates at 4 GHz with a high-Q and hence has a narrow impedance bandwidth. The measured results show a bandwidth of 0.1 GHz for $S_{11}\leq$-10 dB with an optimum return-loss value of -12 dB at 4 GHz.

Microstrip feedline length determines the input impedance of the antenna. Hence, a *T*-matching circuit is directly embedded in the feedline, which is an unconventional technique. Simply put, the proposed *T*-matching network is not an individual part from the antenna structure. The matching circuit used here is a metamaterial structure that exhibits left-handed properties of negative refractive index when interacts with electromagnetic signals [17]. The metamaterial characteristics are realized here with a *T*-network

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composed of series capacitance, shunt inductance, and series capacitance circuit, as per schematic diagram in Fig.7(b). The dispersion diagram of the antenna with the matching circuit in Fig.9 show the phase variation is virtually zero at 4 GHz.

The return-loss response in Fig.8 by theory, simulation, and measurement show that with the inclusion of the metamaterial matching network the antenna's impedance bandwidth is significantly improved. It is evident from Fig.8 that magnitudes of reactance components constituting the *T*-matching circuit has a great influence the Q-factor of the antenna. The results of the parametric study in Fig.10 show how the parameters constituting the *T*-matching circuit effect the return-loss response. The optimum values are: C_1 = 0.5462 pF, C_2 = 1.545 pF and L = 1.1586 pF. The fabricated antenna with the matching network is shown in Fig.11. Its measured return-loss in Fig.8 is better than -18 dB at 4 GHz. The magnitude of the geometrical parameters corresponding to the lumped element values, which are annotated in Fig.11, are given in Table I.

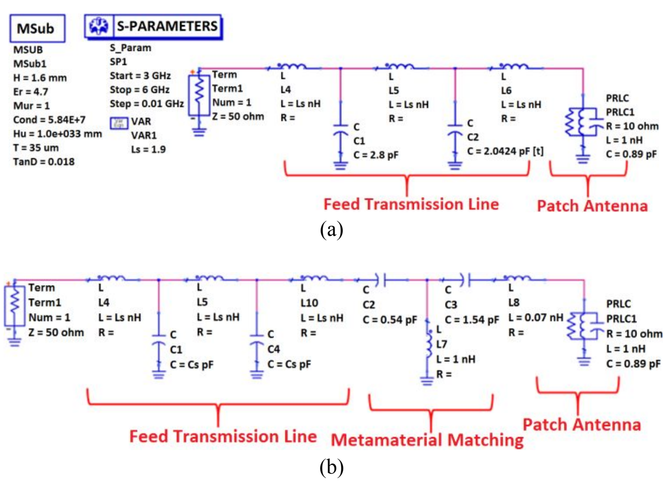

Fig.7. (a) Equivalent circuit model of feedline and patch antenna, and (b) circular patch antenna loaded with metamaterial matching network using Keysight Technologies' Advance Design System (ADS).

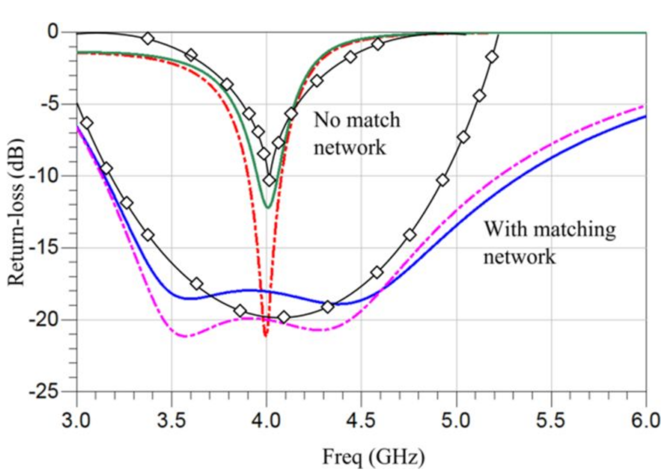

Fig.8. Theoretical, simulated and measured return-loss response of the antenna without and with the metamaterial-inspired matching network. Black diamond studded line is prediction by theory, dotted lines indicate the simulated result, and solid lines represent the measured result.

Simulated current distributions without and with the *T*-matching circuit at 4 GHz is shown in Fig.12. It is evident that the current distribution with the *T*-matching circuit the current intensity is concentrated at the shorting pin of the matching circuit in the feedline. The current however is much stronger over the patch compared to the case with no matching network.

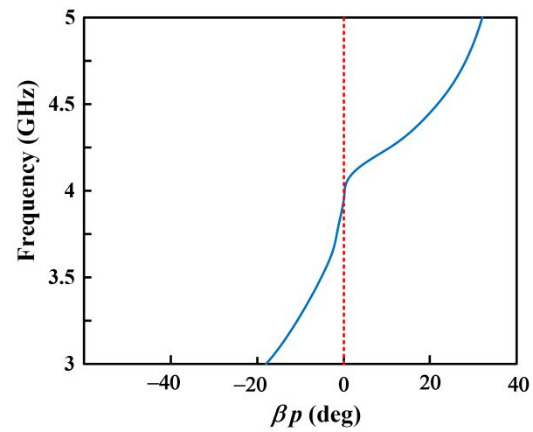

Fig.9. Dispersion diagram of the CRLH-TL feedline antenna.

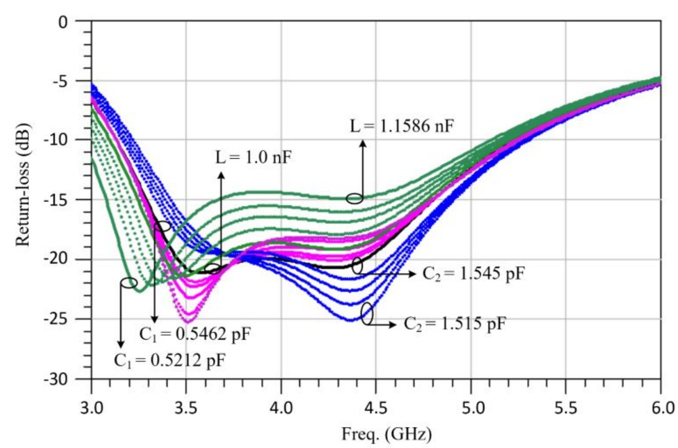

Fig.10. Effect on the return-loss response by the *T*-matching network parameters.

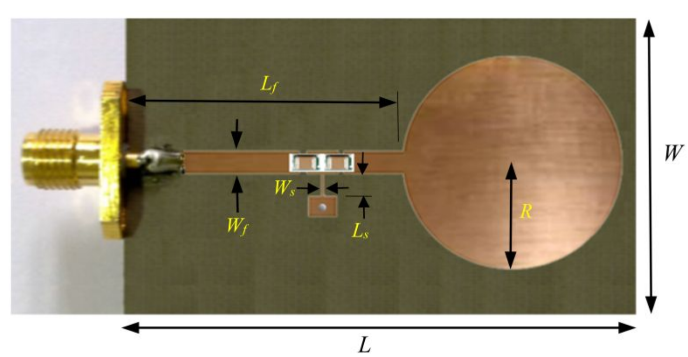

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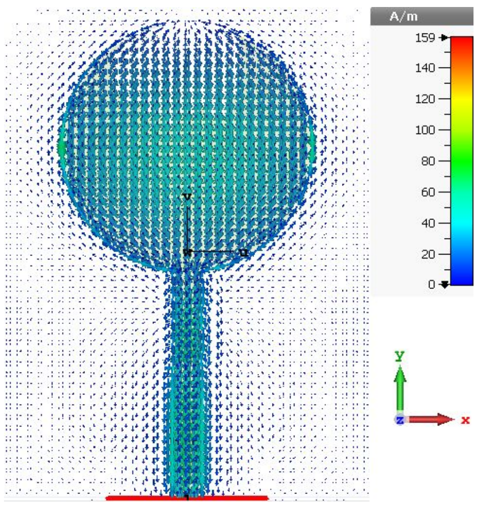
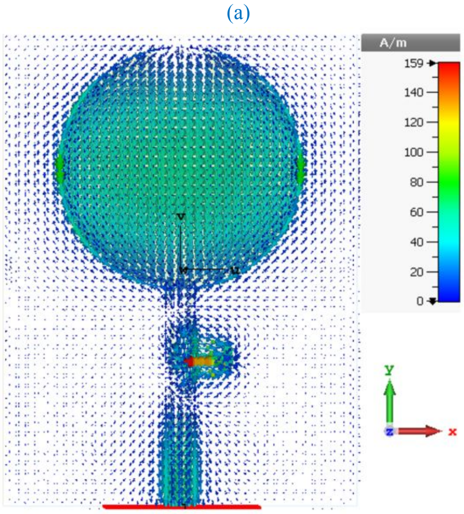
Fig.11. Photograph of the microstrip antenna incorporating the *T*-matching circuit.

Table I. Antenna's geometrical parameters

Parameters	Size (mm)
L	54
W	29
R	11.2
L_f	28
W_f	2.47
L_s	2.3
W_s	0.5
Radius of shunt via-hole	0.5
Thickness of FR-4 substrate	1.6



(a)



(b)

Fig.12. Surface current distribution over the antenna at 4 GHz, (a) without the *T*-matching circuit, and (b) with the T-matching circuit.

The measured radiation gain and efficiency performance without and with the *T*-matching circuit are shown in Fig.13. It is evident from these results that there is substantial improvement in the radiation gain and efficiency by embedding the metamaterial inspired matching circuit. Measured results show without the matching circuit the antenna has a peak gain and efficiency of 5.5 dBi and 59.8%, respectively, at 4 GHz. However, with the matching circuit the peak gain and peak efficiency are 8.2 dBi and 71.8%, respectively. Moreover, the gain and efficiency extend over a significantly larger bandwidth. The gain ≥6 dBi extend from 3.35-5.25 GHz, and the efficiency ≥60% extend from 3.25-5.1GHz.

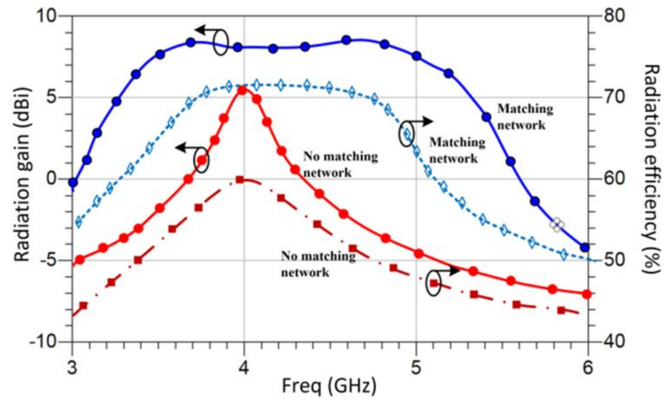
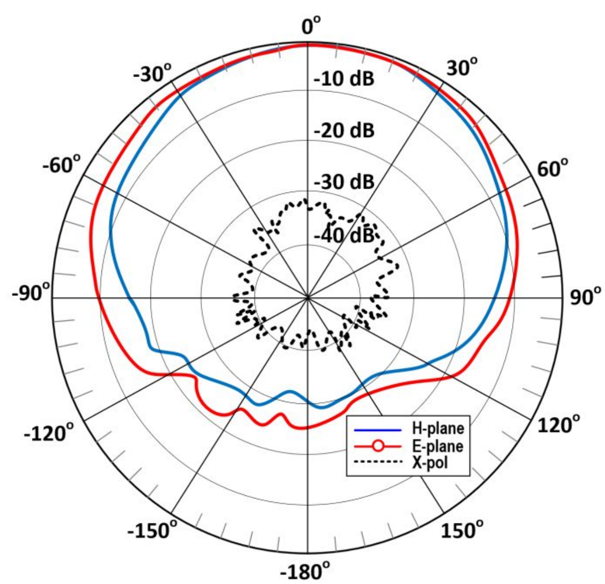
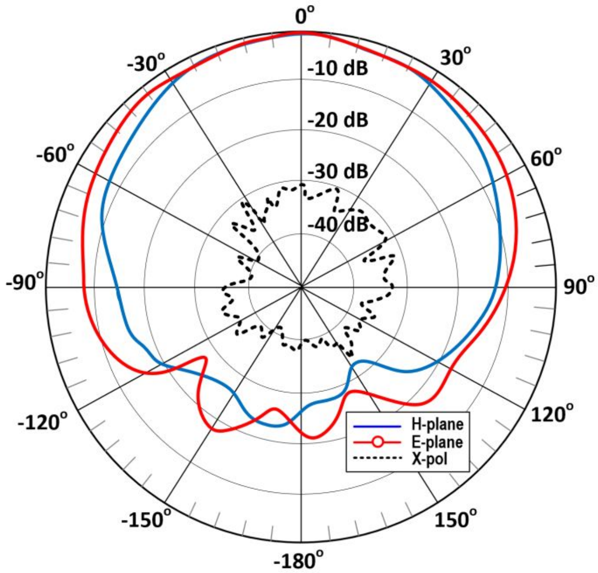


Fig.13. Measured radiation gain and efficiency response with and without the embedded *T*-matching circuit.

The radiation pattern in the E- and H-planes and the cross-polarization without and with the *T*-matching network at 4 GHz are shown in Fig. 14. The change in the radiation pattern with the *T*-matching network is marginal and, in both cases, the cross-polarization is less than -30 dB.



(a) With no *T*-matching circuit.



(b) With *T*-matching circuit.

Fig.14. The normalized measured radiation pattern in the E-plane and H-plane at 4 GHz, (a) without no matching circuit, and (b) with the embedded *T*-matching circuit.

Table II compares the present technique with some recently reported antenna bandwidth enhancement techniques. Impedance bandwidth enhancement of a microstrip patch antenna in [19] and [20] is achieved by loading the antenna with shorting pins and slots. In [21] the bandwidth is enhanced by simply curving the patch antenna in a partial cylindrical shape. In [22] one-dimensional electromagnetic bandgap ground structures and two-stage beam directors are employed in an inverted-L antenna topology to enhance the bandwidth. Even though the radiation efficiency of the proposed antenna is relatively low compared with the other techniques it is evident from the table that the proposed technique has a comparable gain and exhibits a substantially larger impedance bandwidth. Moreover, its size is relatively small with the exception of [22].

Table II. Comparison of the proposed matching technique of recently reported techniques.

Ref.	Technology	Antenna size	Fractional bandwidth (%)	Peak Gain (dBi)	Peak Eff. (%)
19	Shorting pins & slots	$1.14\lambda_o \times 1.27\lambda_o \times 0.009\lambda_o$	10	11.8	94
20	Shorting pins & slots	$1.02\lambda_o \times 1.47\lambda_o \times 0.045\lambda_o$	8.8 & 4.9	6 & 8	93 & 95
21	Microstrip	$0.58\lambda_o \times 0.58\lambda_o \times 0.016\lambda_o$	9	6.28	83.5
22	One-dimensional EM bandgap ground structures & two-stage beam directors	$0.21\lambda_o \times 0.32\lambda_o \times 0.009\lambda_o$	32.5	7	95
23	Differential microstrip	$0.8\lambda_o \times 0.8\lambda_o \times 0.013\lambda_o$	18.7	8.5	80
24	Magneto-electric dipole	$0.97\lambda_o \times 0.97\lambda_o \times 0.17\lambda_o$	28.2	9.0	-
25	Shorting pin & slot	$1.17\lambda_o \times 1.17\lambda_o \times 0.055\lambda_o$	32.2	6.5	93
26	Aperture-fed with shorting pins	$1.63\lambda_o \times 0.82\lambda_o \times 0.017\lambda_o$	15.2	6.8	-
This work	Metamaterial feedline	$0.67\lambda_o \times 0.38\lambda_o \times 0.021\lambda_o$	52.5	8.2	71.8

V. CONCLUSION

Impedance bandwidth of an antenna is determined by the matching conditions between the RF transceiver front-end and the antenna. We have demonstrated the effectiveness of employing an embedded metamaterial-based *T*-matching circuit directly implemented inside the feedline to enhance the antenna's impedance bandwidth. The measured results confirm significant improvement in the antenna's impedance bandwidth over its operating frequency range. Moreover, with the proposed technique improvement was observed in the radiation gain and efficiency too.

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