



Equivalent circuit model of a coaxial excited microstrip-fed quasi-lumped element resonator antenna array

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Abstract: An approximate lumped-parameter equivalent circuit model of the quasi-lumped element resonator antenna array is presented. The array consists of six series resonant circuits of frequency-independent elements, each with an interdigital finger capacitor (IDC), and shorted across by a parallel narrow strip inductor. Each resonator exhibits an aperture size of $0.207\lambda_g \times 0.2\lambda_g$ at a centre frequency of 5.8 GHz, with entire array size of $2.857\lambda_g \times 1.07\lambda_g$. The resulting design is fed by a microstrip line and excited by a coaxial feed probe. A simple equivalent circuit model representing the six resonant circuits as well as the coaxial feed probe circuit, using the magnetic-frill model approach but with the simplest uniform-current model is described to predict the characteristics of the proposed resonator (and by extension, the proposed array), the effects of the coaxial probe and the microstrip feed line. Simulated data obtained from a full wave analysis using finite integration technique commercial solver is compared with the Agilent advanced design system results of the equivalent circuit. The derived equivalent circuit model was validated both analytically and by numerical simulations.

1 Introduction

Advancement in wireless systems and subsequent increasing demand in wireless mobility has enabled the invention of the miniaturised wireless systems. Realising this fact, lightweight, conformable, unobtrusive and ubiquitous equipment, which can be integrated into everyday use, are now one of the crucial design considerations. These wireless communication systems contain several parts in which compact antennas are an essential subpart. However, many technical challenges arise in the design of robust, compact and efficient antennas that deliver the necessary performance to support the emerging applications. The reduction in antenna size presents problem to the system designers because of the performance limitations in antenna bandwidth and radiation efficiency. Nonetheless, the increasing needs for small antennas are because of the more and pressing demands for strait space requirement [1–3], low manufacturing cost, portability, light weight for easy mobility [1] and low power consumption design requirement seeing most of the terminals are battery-driven [4]. It is therefore expedient that the antenna be compact. Microstrip line short sections and stubs, whose physical lengths are smaller than a quarter of guided wavelength at which they operate are common approximate microwave realisation of lumped elements in microstrip circuit structures, and are otherwise termed quasi-lumped elements [5]. Owing to their capacity for size reduction and also to provide distinct benefit of reasonable bandwidth, they have found a wide application in filter designs and more

recently in antenna designs. Usually, an ideal lumped element is not realisable even at lower microwave frequencies because of the associated parasitic reactances as a result of fringing fields. Each component has associated electric and magnetic fields and finite dissipative loss at RF and microwave frequencies. Hence, such components store or release electric and magnetic energies across them and their resistance accounts for the dissipated power. The relative values of the C , L and R components in these elements depend on the intended use of the lumped element [6]. To adequately describe their electrical behaviour, equivalent circuit models for such components are commonly used. To this end, an equivalent model of the proposed antenna is investigated, and the resonator itself is characterised. When this is done, various antenna features such as input impedance, bandwidth, and return loss can easily be regulated by varying the antenna specifications and the feed mechanism. In general, the proposed resonator is readily applicably where size reduction is a premium such as in MIMO and medical implants.

2 Theoretical framework

2.1 Proposed antenna array

Fig. 1a is the photo of the array structure, whereas, Figs. 1b and c) are geometry and voltage allocation, respectively. The array is excited by a coaxial feed probe and fed by a microstrip line. In essence, the electromagnetic energy travels through the microstrip line and it is sufficient to

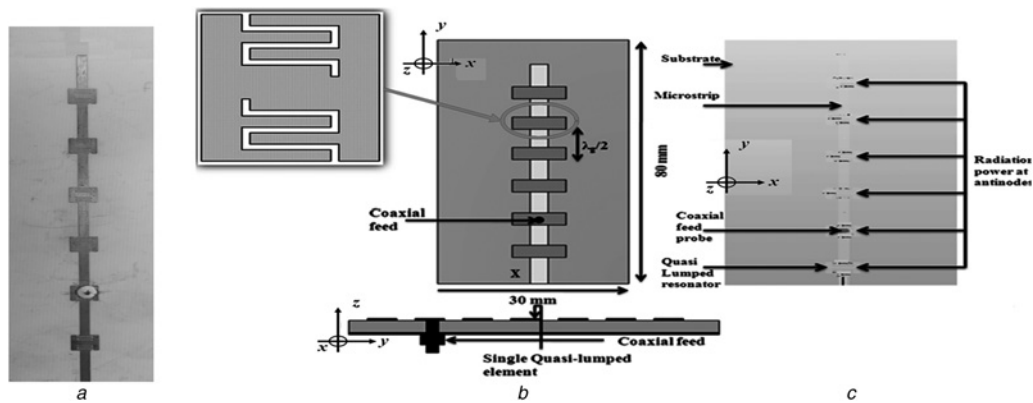


Fig. 1 Proposed antenna configurations

a Prototype
b Geometry
c Voltage allocation

excite each radiating element concurrently in such a magnitude that depends on the voltage allocation on an open ended microstrip line. When a transmission line is not matched to its load, some of the energy is absorbed by the load and some is reflected back down the microstrip line towards the source. The interference of incident wave and reflected waves creates the standing waves on the transmission lines. The ratio of the maximum to minimum magnitudes of voltage or current on a line having standing waves is called standing wave ratio. If the line is terminated in a load other than R_0 , the distribution of voltage at a point along a length of the line consists of maximum and minimum voltages. When a line is terminated with R_0 , the standing waves are absent and the line is thus referred to as a smooth line. Knowing full well that the voltage and current are in quadrature, thus it is obvious that the magnitude of the current along the line would be the same except at $\lambda/4$ position where there exist a shift between the maximum and minimum. The points along the line where voltage or current is zero are called nodes, whereas, the points along the line where voltage or current is maximum are called antinodes or loops. If the voltage or current is measured along the line of length (λ), such parameters will vary, depending on the load. Fig. 1c describes the voltage allocation over the length of the line when the load end of the line is shorted. The same impedance and voltage condition is periodic every half-wavelength down further the line from the load completion towards the generator. Since the antinodes will repeat itself at every half wavelength along the transmission line, the array elements were positioned every half wavelength in order to obtain the maximum electric field radiation of the antenna array as demonstrated in Fig. 1c.

2.2 Proposed equivalent circuit model

Each radiating element consists of an interdigital capacitor in parallel with a shunt narrow strip inductor as shown in Fig. 2 but better demonstrated in [7] with its dimensions fully explained. The pads connected at both ends of the structure acts as the capacitor to ground (C_{P1} & C_{P2}). Equations to determine the resonator circuit components, as well as the resonant frequency of the proposed resonator are equally stated. In Fig. 2 of [8], the equivalent circuit model of a coaxial feed probe was presented. Rigorous equations to determine the probe equivalent circuit components using simplest uniform-current model were presented. Of crucial importance are (8–12), (24) and (27). The transformer ratio ($1: R_0$) is the junction between the radial and coaxial junction of the coaxial probe, which is useful in determining the input impedance in one line given certain termination conditions in others, impedance of which may be the same if the electrical dimensions are very small [9, 10]. Usually at a short circuit end, the input impedance of a microstrip line is a low; hence the microstrip line behaves like a parallel resonant circuit. Fig. 3 demonstrates the input impedance for different wavelengths of short circuited line. For a quarter wavelength stub, the input impedance is highly resistive. Therefore the line turns to a parallel resonant circuit. Somewhere in between the two points, the input impedance is inductive. Similarly, in between a quarter wavelengths and half wavelength, the input impedance is capacitive as demonstrated in Fig. 3. According to Akhavan *et al.* [11]; the values of microstrip parallel lumped elements at antinodes can be computed following the below processes. First step is to determine the

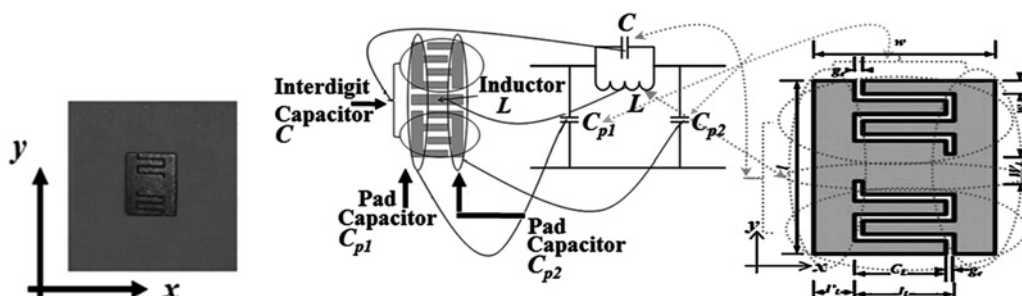


Fig. 2 Proposed resonator

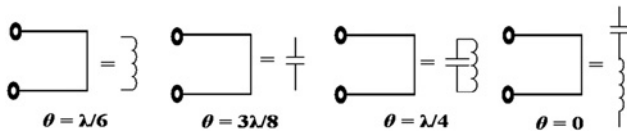


Fig. 3 Input impedance for different wavelengths of short circuited line [9]

equivalent radiation conductance G_{rm} expression given in (1) and reported in [10]

$$G_{rm} = \frac{160\pi^2 h^2}{Z_0^2 \lambda_0^2 \epsilon_{eff}} \quad (1)$$

where h is the substrate height, Z_0 is the characteristic impedance of the microstrip and ϵ_{eff} is the effective dielectric constant.

Second step is to obtain the microstrip line equivalent capacitance C_l ; (i.e. $B_m = \omega C_l$) because of the fringing fields where B_m is the susceptance. To do this, the equivalent length technique can be used. Assume that l_{eq} is the equivalent extra length of the microstrip line as a result of the fringing field effect, and using the expression for l_{eq} as given in [12]. Equation (2) describes the fringing field capacitance effect of the microstrip feed line as

$$C_l = \frac{l_{eq} c \sqrt{\epsilon_{eff}}}{Z_0} \quad (2)$$

where l_{eq} is the equivalent extra length of microstrip and c is the speed of light. Assuming that L_1 corresponds to the inductance of width W_s of the microstrip line [12], the value of which is stated as

$$L_1 = [Z_0 \sqrt{\epsilon_{eff}}] W_s / c \quad (3)$$

The subsequent procedures to realise the approximate model of the proposed array is documented in [13, 14] although for the dielectric resonator antenna array excited by a rectangular aperture coupling slots and fed by microstrip line. The equivalent circuit of the microstrip feed line is depicted as shown in Fig. 4 which is used to feed the array. Therefore integrating the proposed resonator equivalent circuit shown in Fig. 2 with the coaxial feed probe equivalent circuit of Fig. 2 in [7] and microstrip feed line equivalent circuit of Fig. 4 below demonstrates the equivalent circuit model of the single element microstrip-fed coaxially excited quasi-lumped element resonator antenna shown in Fig. 5. Equivalent circuit depicted in Fig. 5 is therefore re-designed with five more radiating element to catch the radiations occurring at every antinodes, and its consequently modelled using the Agilent advance design systems (ADS) as shown in Fig. 6.



Fig. 4 Equivalent circuit of a microstrip line

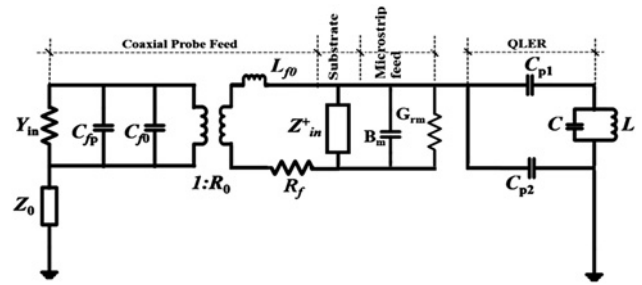


Fig. 5 Equivalent circuit of single coaxial-fed QLER antenna

3 Result and discussions

Table 3 reported in [7] summarised the design parameters of the single quasi-lumped element resonator. At 5.8 GHz resonant frequency, the guided wavelength was calculated to be 28 mm. The half wavelength was therefore determined to be equal to 14 mm, whereas the quarter wavelength was 7 mm. Since the antinodes will repeat itself at every half wavelength along the transmission line, the array elements were positioned at every half wavelength. The first element was positioned at distance X of 10 mm away from the generator in order to prevent further incoming antinodes along the feeder. Fig. 1c demonstrated the simulated voltage allocations at each antinode occurring over the entire microstrip line length. To accomplish the modelling of the proposed antenna array, the impedances of the coaxial feed probe, the substrate, the microstrip feed line and that of the proposed resonators were determined by various equations reported in [7, 8]. Their equivalent circuits were determined, integrated together and modelled using Agilent ADS as demonstrated in Fig. 6. Subsequently, the designed antenna array was simulated using CST microwave studio [15]. The simulated design was thereafter photo etched on a grounded Duroid RC4003C laminate microwave substrate of thickness of 0.813 mm, and relative permittivity (ϵ_r) of 3.38 and then measured. In essence, the simulated and measured results were compared with the derived equivalent circuit model in order to validate the model result both by analytical formulae through which the equivalent circuit components were obtained and numerical simulations. Fig. 7 shows the simulated and measured return loss of the six elements series array quasi-lumped element resonator antenna in CST, ADS and measurement, respectively. All the return loss results agreed sufficiently. In all, S_{11} was better than -10 dB. The simulation return loss results was 22.37 dB over a bandwidth of 5.77–5.98 GHz (210 MHz) whereas, it was 31.81 dB with a frequency pattern of 5.76–5.982 GHz (242 MHz) for the measurement. The modelled return loss was 16.97 dB with a bandwidth of 200 MHz. In Fig. 8, the CST simulated input impedance and the measured input impedance of the quasi-lumped antenna in form of Smith charts are presented. It was evident that impedance loops of both the measured and simulated input impedance agreed reasonably. In each result, there is a single loop which represent minimum S_{11} particularly in the measured loop. However, simulated input impedance differ slightly as it seemingly turns more than one indistinct loop. The simulated input impedance was $48.87 - j2.32 \Omega$ at 5.83 GHz whereas the measured input impedance was $47.98 - j1.21 \Omega$ at 5.79 GHz. The difference between simulated and measured input impedance may be attributed to the inability

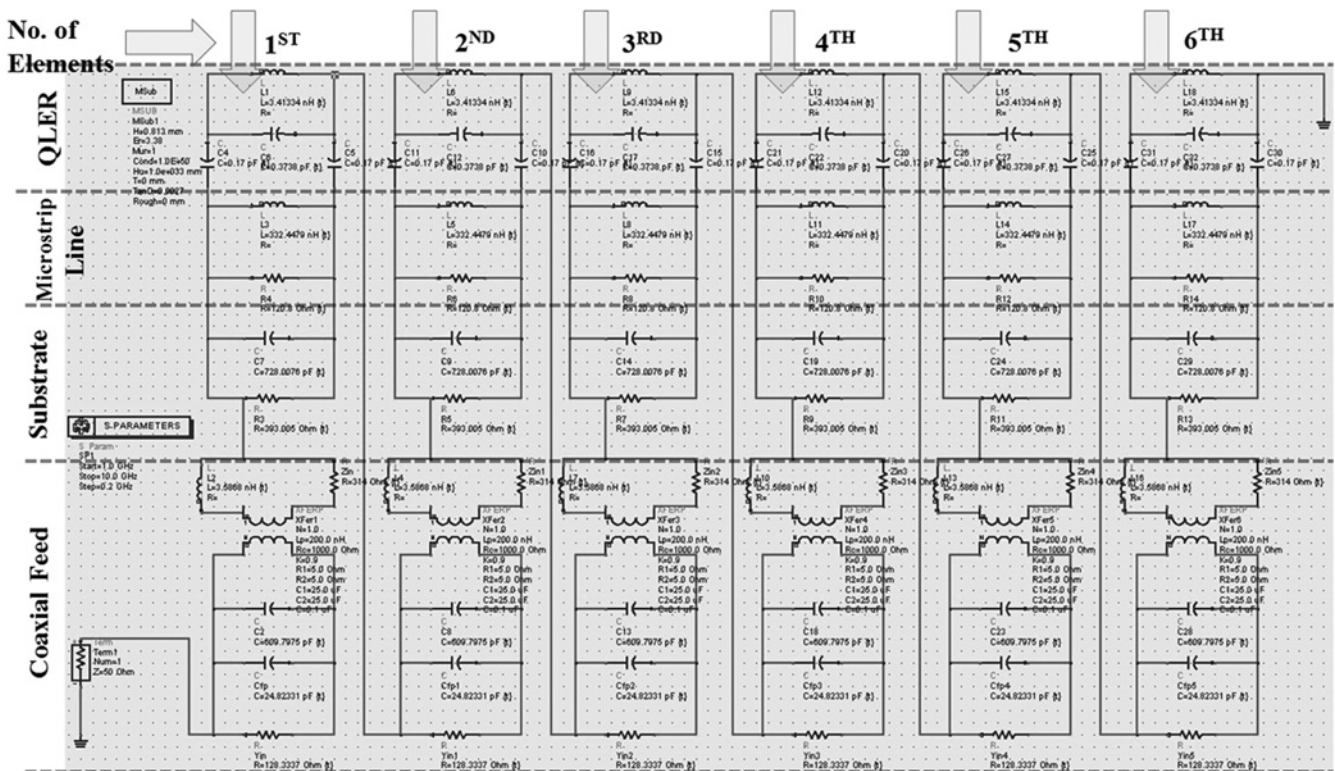


Fig. 6 Proposed model of the antenna array in ADS

to accurately locate the coaxial feed probe position during fabrication with respect to the simulation feed location. The modelled input impedance was determined to be $47.26 - j0.93 \Omega$ at 5.8 GHz using ADS and equations described in [8].

Fig. 9 shows the simulated and measured voltage standing wave ratio (VSWR). The simulated VSWR was 1.173:1 whereas the measured was 1.053:1 at the frequencies of 5.83 and 5.79 GHz, respectively. The modelled VSWR was 1.33:1 at a frequency of 5.8 GHz.

4 Validations

The proposed resonator consists of the interdigit capacitor as well as the centre inductor. The inductance increases steadily as both the strip inductor length and the width increase linearly from a position somewhere when both the length and the width are at about 1.5 mm and over as demonstrated in Figs. 10a and b. However, outside this range, the inductance becomes non-linear and unpredictable. To maintain compactness therefore, the length was chosen as 3.35 mm and the width as 0.35 mm. These values were chosen bearing in mind that the longer or wider these dimensions, the larger the resonator becomes. Coincidentally, it is at this length that the antenna was designed to resonate as evidenced by the antenna design parameters shown in Table 3 [7]. The capacitance increases as the narrow gap increases, although folded over to ensure compactness. It was also noted as shown in Figs. 10c and d that the interdigital capacitance increases linearly as the number of fingers (N) increases, and also, as the overlap of the interdigital capacitor fingers (C_L) increases. Therefore the proposed resonator was designed with capacitance of about 0.04 pF and interdigit fingers number of 8. The parasitic capacitance is the capacitance from the probe barrel to the unconnected plate because of

the higher-order modes, and it is only related to the geometry of the probe. It does not vary with the shapes and size of the quasi-lumped resonator antenna. For a thin substrate such as used in this work (of dimension 0.813 mm), C_p is too small to affect the first resonant frequency. Consequently, varying its value in the ADS modelling does not show any notable effect on the S_{11} model as the resonant frequency remains constant rather than shift as observed in the thick substrate. However, C_p increases as the substrate becomes thicker, and hence became too pronounced to be neglected as demonstrated in [8, 16–25]. By implication, the resonance frequency shifts in case of a thick substrate as a result of increased parasitic capacitance, C_p . This is the parasitic capacitance because of the zero-order mode and it's represented by C_0 . It largely depends on the radii of the probe and the substrate thickness. At a lower ranges of frequencies (of say less than 20 GHz), there was no notable variation effect of C_0 on the resonant frequency or the return loss. However, at higher range of frequencies (say over 20 GHz), the effect becomes

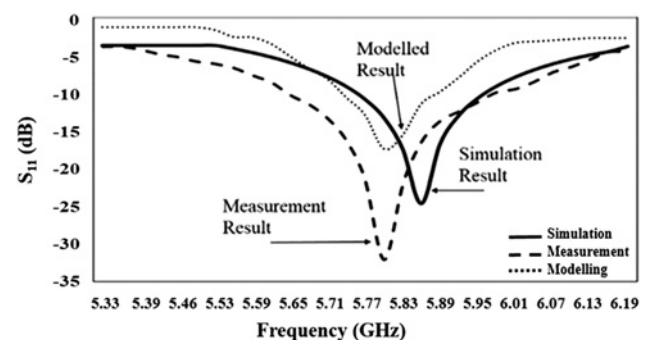


Fig. 7 Return loss

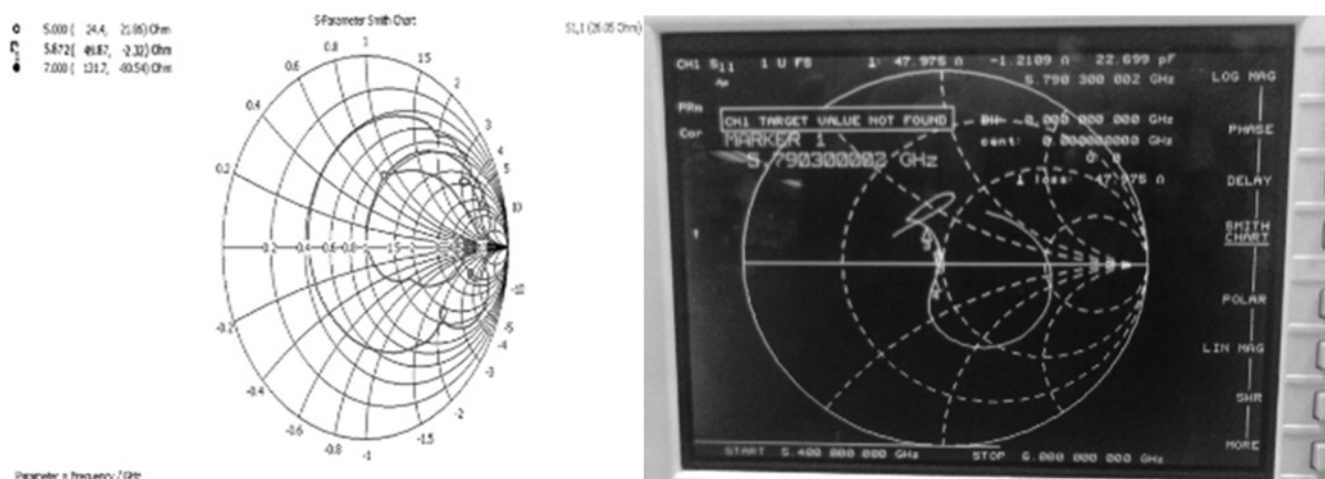


Fig. 8 Simulated and measured Smith chart plots

notable. The same scenario was observed for other parasitic such as the parasitic inductance because of zero-order mode L_0 . In particular, there was no noted effect of the transformer R_0 because of the variation of either the probe height or plate separation h until at higher frequency range when the transformer became significant, and therefore not negligible. In the case of the feed probe zero-order parasitic inductance L_0 , the effect of this inductance was prominent both on the return loss and the resonant frequency. Two major parameters are needed to properly examine its influences in a way to better understand the effect of this inductance. First is the effect of the outer radius (b), and the second is the larger probe height (h). The inductance L_0 increases linearly with the increase of probe height. Also, the longer probe height results in large C_p but small C_0 . The effect of this action was significant in this work. This scenario is effectively demonstrated in Fig. 11 where firstly the increase of the probe inductance from 1.79 to 5.34 nH improves the S_{11} from -10 dB to about -30 dB, both exploring the derived equivalent circuit via Agilent ADS as well as the finite integration technique (FIT) CST. Besides, the resonance frequency also shifts almost inversely proportionately from 9 to about 1 GHz. The increased values of L_0 , C_p and C_0 combined, improved the return loss tremendously. Simultaneously however, the longer the probe height, two crucial observations were noted. The first was that the return loss increases majorly both in the ADS modelling as well as the CST as earlier discussed. The second was that the radiation patterns suffer significantly, as the radiation pattern degradation was pronounced with a

major dip on the principal axis ($\theta = 0^\circ$ for broadside) of radiation when view in CST simulation. This is because increasing the probe height usually introduces more parasitic capacitance entry to the circuit, hence better impedance match, and return loss. It however introduces protrusion to the design and therefore make it bulky. Although, the effect of C_p for thin substrate as used in this work was expected to be negligible. However, it became significant as soon as the height of the probe increases beyond the substrate thickness. In such a case as this, the otherwise thin substrate inadvertently approximates to thick substrate, of which the assumption is valid.

More importantly, the resonant frequency shifts to the left and causes the resonance to occur at a frequency less than the designed frequency as a result of increase of C_p . Hu in [8] also reported the same observation and therefore corroborated this assertion. On the other hand, the larger outer radius (b) leads to large probe aperture. As a result of this, the return loss depreciates, but the bandwidth becomes wider. Further results to substantiate our position are presented in Table 1. To that extent, it was noted that with the increase of the probe aperture radius, C_p decreases but C_0 increases. Therefore there exist two extremes of influences. First is that the increase of the probe height increases the return loss and then deliver good input impedance, but exhibits narrower bandwidth. The second extreme, is that the increase of the aperture outer radius increases the bandwidth, but reduces both the return loss and the input impedance. Much more, the radiation pattern was better. The same observation was noted from the CST simulation. A compromise is necessary as a sort of trade-off to achieve the desired considerations. The influence of each component of the equivalent circuit, and the ADS simulation were thoroughly understudied, compared, and verified. For the 6-element array, the modelling was not too different from the single element except that it has six elements and a Microstrip feed line as earlier demonstrated in Fig. 6. The essence of the Microstrip line is to distribute the in-phase excitation potential across all the elements of the array, and concurrently. Therefore the properties explained for the single element is still tenable in this case; except for the Microstrip feed line. To this end, the width of the Microstrip feed line was determined using the impedance line calculator available in ADS software. This was done to ensure maximum power transfer because of proper match. Equation (4) states the equation for the width

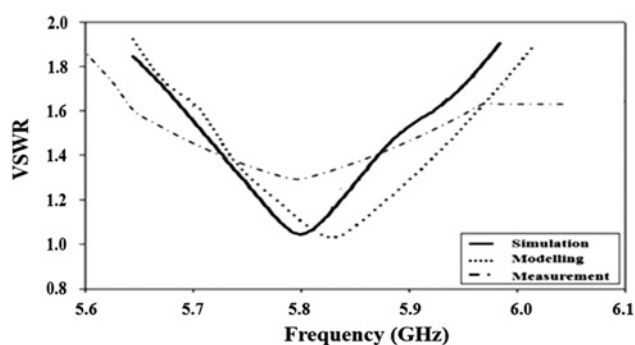


Fig. 9 Simulated and measured VSWR of the proposed antenna array

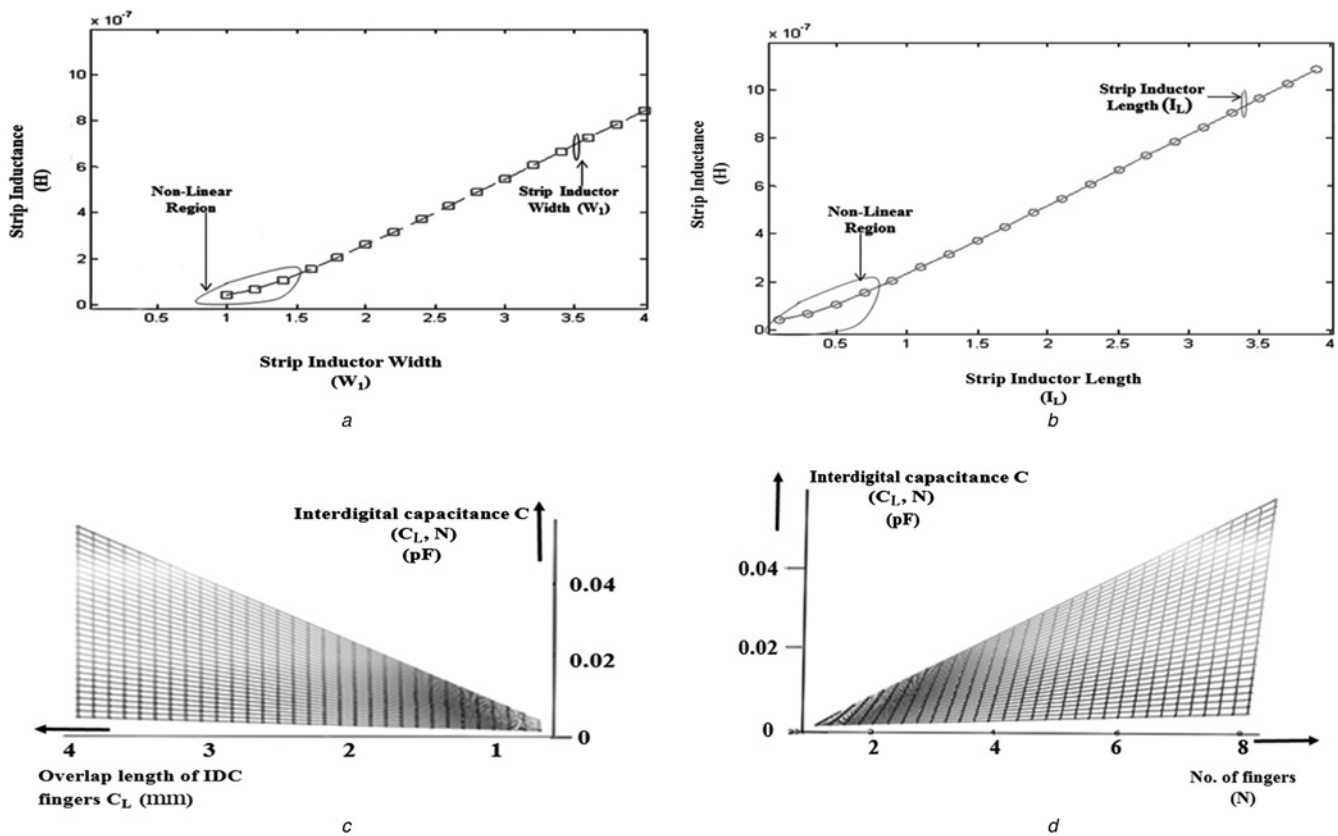


Fig. 10 Characterisation of the proposed resonator

- a Parallel strip inductance against inductor width
 b Parallel strip inductance against inductor length
 c Interdigital finger capacitor (IDC) against overlap length fingers
 d IDC against no. of fingers

of the Microstrip feed line with respect to its inductance

$$W_s = \frac{cL_1}{Z_0\sqrt{\epsilon_{\text{eff}}}} \quad (4)$$

where W_s is the Microstrip feed line width, L_s is the corresponding inductance as a result of the width, Z_0 is the characteristic impedance of the feed line set at 50Ω for proper impedance match, ϵ_{eff} is the effective permittivity of the strip, whereas c is the velocity of light. According to the equation, the width of the Microstrip line largely depends

on its permittivity, and its corresponding inductance. Usually, an antenna system is not completely dependent on the frequency response of the antenna elements themselves, but much more, on the frequency characteristics of the transmission line-antenna elements combined [27, 28].

This is because the two-media nature (substrate discontinuity) of the Microstrip line causes its dominant mode to be hybrid (Quasi-TEM) rather than TEM. The attendant effect of this is that the phase velocity, the characteristics impedance and the field variation all become frequency dependent marginally. Objectively, three essential parameters adequately define the microstrip lines namely:

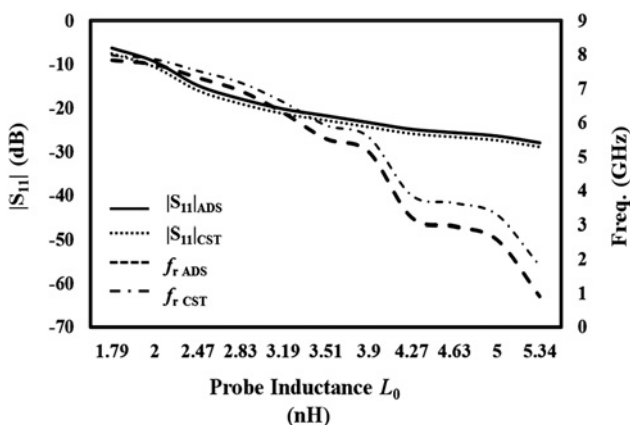


Fig. 11 Effect of probe height (L_0) on S_{11} the resonance freq

Table 1 Probe dimensions and its characteristic effect on the antenna

Probe height, mm	Probe inductance, nH	Resonant freq., GHz	S_{11} , dB	Bandwidth, GHz
		f_{QLER}	$S_{11\text{QLER}}$	BW_{QLER}
2.65	1.79	8.2	-9	0.00
2.70	2.00	7.8	-10	1.30
2.85	2.47	7.1	-13	1.40
2.90	2.83	6.7	-16	1.55
2.95	3.19	6.4	-21	1.70
3.00	3.51	6.2	-27	1.90
3.05	3.90	6.0	-30	1.20
3.10	4.27	5.8	-45	1.00
3.15	4.63	5.7	-47	0.80
3.20	5.00	5.6	-50	0.70
3.25	5.34	5.4	-63	0.64

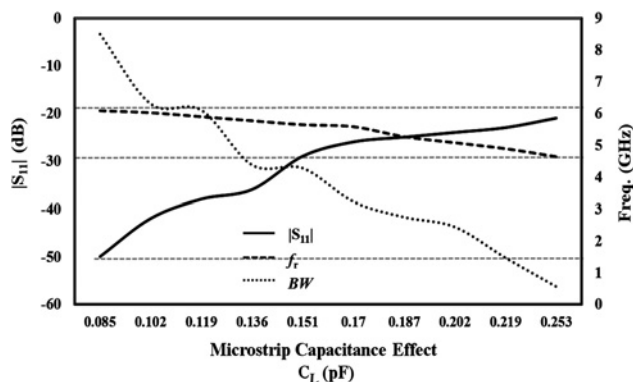


Fig. 12 Effect of microstrip line on S_{11} the resonance freq

Table 2 Microstrip line capacitance effect on the antenna

Microstrip capacitance effect, pF	S_{11} , dB	Resonant freq., GHz	Bandwidth, GHz
0.085	-50	6.0722	8.5
0.102	-42	6.0096	6.3
0.119	-38	5.8844	6.1
0.136	-36	5.7592	4.37
0.151	-29	5.634	4.27
0.17	-26	5.5714	3.22
0.187	-25	5.2584	2.74
0.202	-24	5.0706	2.43
0.219	-23	4.8828	1.46

the radiation conductance G_{rm} (that dependent on the substrate height, characteristic impedance, effective dielectric constants, the wavelength of propagation in free space all of which are constant as regards this design), the microstrip line susceptance B_m and finally the inductance effect as defined by (4). The susceptance largely depends on the line equivalent capacitance C_l because of the equivalent extra feed line as a result of the fringing field effect. Hence, adjusting the width inadvertently amount to varying this capacitance/inductance. Therefore decreasing the width of the Microstrip line indirectly increases the inductance, and subsequently improves the return loss, particularly from 25 dB to 30 dB as shown in Fig. 12. This also shifts the resonance frequency asides the designed 5.8 GHz from 6.9 to 4.8 GHz as shown in the same figure. Similarly, increasing the width depreciates the return loss and at the same time shifts the resonance frequency marginally above 5.8 GHz. Seeing from (4) that the Microstrip feed line width directly depends on its inductance, then increasing the width is synonymous to increasing the inductance. The microstrip line effects of the proposed equivalent circuit on bandwidth, S_{11} , and resonant frequency is presented in Table 2.

However, a compromise must be made otherwise, for very wide Microstrip line width; the fields are all almost in the substrate, while narrower width will present proportionally more field energy in the air. Table 3 essentially compares the equivalent circuit model with the CST simulation as well as the measurement. On a general note, the equivalent circuit model agreed reasonably with the CST prototype as well as the measured results. To this effect, the model represents the validity of the resonant frequency of the proposed equivalent circuit of microstrip line fed coaxially excited quasi-lumped element series array.

Table 3 Comparative assessment of the model

Parameter	Simulation	Modelling	Measurement
frequency	5.8 GHz	5.78 GHz	5.83 GHz
return loss	61.51 dB	19.94 dB	50.97 dB
VSWR	1.002:1	1.224:1	1.006:1
bandwidth	382 MHz	180	340 MHz
input impedances	$49.92 - j1.32 \Omega$	$49.73 + j0.09 \Omega$	$50.40 - j2.16 \Omega$

5 Conclusions

In this work considerable effort has been made to derive the equivalent circuit model of the microstrip fed coaxially excited six-element quasi-lumped resonator antenna. The proposed model was analytically deduced based on existing numerical explicit formulae, and then investigated using numerical solver such as the FIT as well as the Agilent ADS. The derived equivalent circuit model have been validated both by analytical formulae and numerical simulations. It has been established that not only the microstrip length that influence the response of the equivalent circuit, but also the feed (the height as well as the radii). Although the effect of excessive probe height leads to the radiation pattern degradation as evidenced in CST environment, its inherently able to reduce the proposed resonator's unnecessary harmonics (with attendant narrow bandwidth) being a periodic structure. Reasonable agreement between the simulated and measured was obtained.

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