A Novel T-Fed 4-Element Quasi-Lumped Resonator Antenna Array

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Abstract. *In this paper, electrically small corporately T-fed quasi-lumped element resonator antenna array is investigated. The radiating element, a quasi-lumped element resonator is excited by a novel semi hybrid ring-like T-shaped corporate feed network. The characteristics losses due to Ohmic and discontinuities along the feed line which invariably constitutes complex feed structures are mitigated at the instance of the proposed antenna. Technique to implement the compact array with the intent to enhance the gain is presented.*

The operation dynamics of the feed along with its theoretical explanation is also reported. Findings indicates that the measured gain is 10.97 dBi for antenna of an aperture size of about $0.677\lambda_0 \times 1.257\lambda_0$ sq. mm. Valuable insight to the optimum design in terms of compactness, good gain, and ease of fabrication is documented.

Keywords

Corporate array, T-fed, power divider, quasi-lumped element, quarter wave transformer.

1. Introduction

Array configuration has been the most popular way to enhance the gain among the antenna designers [1]. Besides, beam forming, scanning, or steering capability there is a possibility when discrete radiators are combined to form arrays. The array may be series or parallel, otherwise known as corporate array. While the series array has the advantages of compact network due to less transmission lengths, fewer junction, low insertion loss; it however suffers from narrow bandwidth, and inherent phase differences caused by the difference in the lengths of the feed lines [2].

Alternatively, corporate feed network that offers promising enhanced performance in terms of relatively wide bandwidth due to reduced mismatch losses [3], great design flexibility, ease of vertical integration to form a 2-D array [4], ease of control to each radiating element both in terms of magnitude and phase, equal excitation of the discrete radiating elements - is however notorious for large antenna real estate cum large ohmic loss in the feed structure of the arrays, and therefore unsuitable for applications requiring compact antennas. Yet, the need for small and efficient antenna is increasingly becoming critical and pressing, 1) partly because of insufficient space to fit conventional antennas as often times, the platform onto which the antenna is to be mounted or interfaced, is itself physically small; 2) due largely to ever increasing demand for mobile handheld devices at different frequency band of applications; and finally, to the fact that these mobile handheld devices are battery driven.

However, the reduction in antenna size presents problem to the system designers due to the performance limitations in antenna bandwidth and the radiation efficiency. The intent of this paper therefore is to explore an alternative corporate feed solution with compact capability as a goal in order to utilize its inherent advantages. To truly achieve overall systemic reduction, it is expected that both the radiating element and the proposed feed must be compact. To that extent, 1) a quasi-lumped element resonator antenna is presented as the radiating element, whereas 2) a novel modified T-shaped corporate feed network is introduced as the feed. The radiating element is a digit-like periodic pattern of parallel in-plane electrode in an interdigital geometry.

It basically consists of an interdigital capacitor (IDC) in parallel with a strip inductor. The inductor is the centre finger and shorted across the IDC. Being a lumped element device, it is characterized by physical dimensions smaller than a quarter wavelengths at which they operate. On the other hand, the feed is a quasi corporate network feed but configured in the modified T-shaped geometry. Deploying this novel technique will firstly reduce the complexity in the feed structure of the proposed, reduce the discontinuities, and automatically neutralizes losses associated with the conventional corporate feed network. Importantly, a slim Duroid substrate and reduced metal thickness was used to further eliminate these losses. Holistically, it is expected that the proposed antenna will of necessity be efficient & compact, and secondly, exhibit enhanced gain.

2. The Proposed Resonator

The proposed resonator, a quasi-lumped element resonator is a finger-like periodic pattern of parallel in-plane conductors used to build up the capacitance associated with the electric fields. Intrinsically, the structure shown in Fig. 1 is made up of microstrip line sections and shorts. It consists of interdigital capacitor (IDC) in parallel with a strip inductor. The pad capacitances, which occur when the pads are connected at both ends of the structure, can be adjusted to tune the resonant frequency of the proposed resonator. These pads also act as the capacitor to the ground. Being lumped element, they are basically defined as passive components whose size across any dimension is much smaller than the operating wavelength so that there is no appreciable phase shift between the input and output terminal.

Usually, the maximum dimension of less than $\lambda_g/2$ is a good approximation of the operating wavelength λ_g where λ_g is the guided wavelength. Essentially, it employs the capacitance occurring across the narrow gap between the conductors. These gaps are essentially very long and folded in order to use a small amount of area and consequently, its attendant relevance as a lumped element. The capacitance increases as the 1) gap decreases, 2) overlapping width of the interdigital finger C_L increases, and 3) number of finger increases. However, increasing the number or C_L will invariably increase the antenna size. Relevant equations to determine the proposed resonator design parameters, dimensions, and resonant frequency are stated in [5].

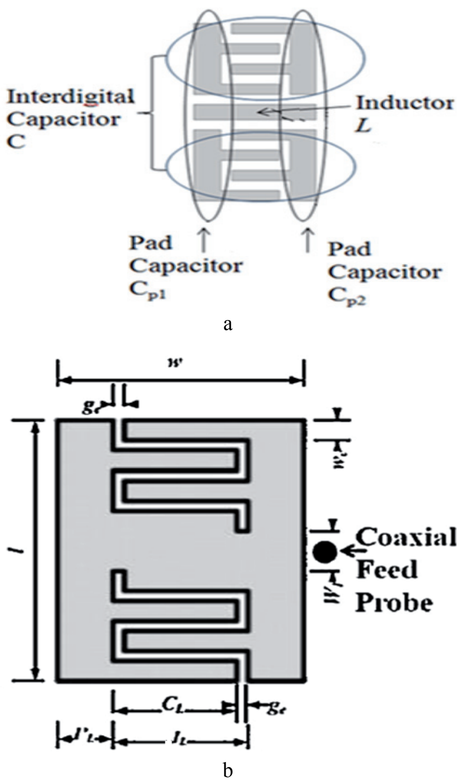

Fig. 1. The proposed radiating element. (a) Subcomponent, (b) Design dimensions.

3. The Proposed Feed

The hybrid-ring coupler is an appealing choice in basic power division among radiating elements. It is very useful in beam-forming networks for printed-circuit array antennas. This attribute is primarily based on two reasons namely; 1) the input impedance of a hybrid-ring coupler can be matched when the other arms are terminated by matched impedances, and 2) the output arms can be isolated from each other [6]. The proposed feed used the same principle but differs to an extent. Instead of the full hybrid-ring, a quasi semi hybrid-ring liked T-shaped feed was deployed. The semi hybrid-ring like shape was formed by splitting the full hybrid-ring four-port network along its plane of symmetry into equivalent two-port networks as shown in Fig. 2(b), thus the proposed array relies on a hybrid-ring topology but however, it is a T-shape in principle as the ring is cut in which case, not all properties of the hybrid ring topology are simply applicable.

It has been identified that surface waves are generated at any discontinuity of the microstrip line [7], and so it is for bends. The surface wave, once generated, can travel, propagate, and excite other modes other than that of the radiating element. Besides, the traditional corporate feed usually employs a conventional T-shaped λ/4 wavelength transformer as a feed line between the excitation source signal and the radiating elements. This T-shape comprises of two 90° bends.

Characteristically, the current distributions of the excitation signals that flow through each bend are symmetrical along the edges as they travel from the λ/4 wavelength stub toward the junction. However, both the magnitude and phase of this current distribution are asymmetrical and decrease monotonically toward the junction [8]. In effect, there exists an excess capacitance at the square corner junction of the bend. Hence, because of the existence of excess capacitance, the characteristic impedance of the junction will be lower than that of the uniform connecting lines.

Existing literatures has proved that the traditional T-shaped feed system has been inadequate because of inherent losses at the discontinuities, which inadvertently increase with the number of bends. Therefore, compensation techniques such as increased inductance or decreased capacitance technique are often recommended [9]. Experimental observations have shown that a decrease in input reflection coefficients can be achieved if the corners are mitered or curved, where curving tends to be better in terms of performance [10-12].

It is on the basis of this that we consider a curve-like T-shaped corporate feed as a mean of excitation. It is expected that the feed will default the unnecessary losses associated with these discontinuities. Fig. 2(b) shows the arm of the semi hybrid-ring like geometry which is analogous to Fig. 2(a) but re-configured in such a way that the structure was rather turn outward but with the same input feed location at port 1. Consider an input signal at port 1 that splits into two equal waves travelling in opposite directions around the semi ring.

At port 2 and 3, the 2 waves have traveled an equal distance of λ/4 each but in opposite direction. Thus the two waves are in phase at port 2 and 3. Port 1 discretely shown in Fig. 2(d) is a quarter wave transformer, which directly transforms the characteristic impedance Z_1 and then split the line impedance Z_2 into twice Z_3. Fig. 2(c) shows a practical realization of balanced coupling as demonstrated by Iveland [13]. It is in reality constituting a simple balun.

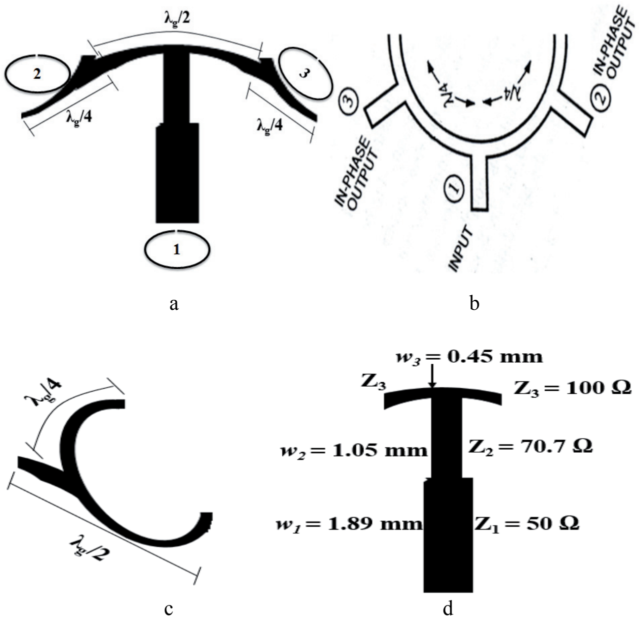

Fig. 2. The proposed feed geometry.(a) Modified hybrid-ring T-shaped. (b) Typical semi hybrid-ring. (c) Practical realization of balanced coupler-like shape. (d) Quarter wave transformer/power divider.

The 180° phase shift is obtained by means of power splitter and hence, a path difference of λ/2 between the two signal paths. The arm with an impedance of Z_3 is subsequently transformed to impedance Z_1 via a quarter wave transformer with an impedance Z_2.

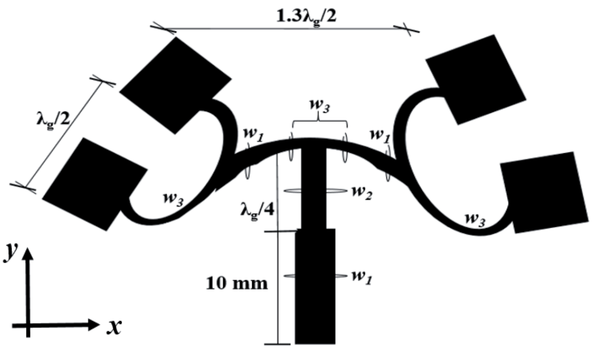

Fig. 3. The proposed antenna array.

Hence, Z_1 can further split power into Z_3 of equal magnitude but with 180° out of phase as current moves toward the load. This feature is responsible for the inherent circular polarization (CP) potential of the proposed feed.

4. Antenna Geometry

The proposed resonator design was depicted in Fig. 1(b). The length l of the radiating element was 5.8 mm whereas, its breadth w was 5.6 mm, with $N = 8$, where N is the number of interdigital fingers. The width of each finger w_1 was 0.35 mm, the length C_L of 3.05 mm, and inter-finger width g_e of 0.3 mm. The width of the centre finger, otherwise known as the narrow strip inductor W was 1.2 mm which is of the same length with the interdigit fingers. The pad width I'_L was calculated to be 1.125 mm. The design equations to determine these values with respect to its resonance frequency were stated explicitly in [5].

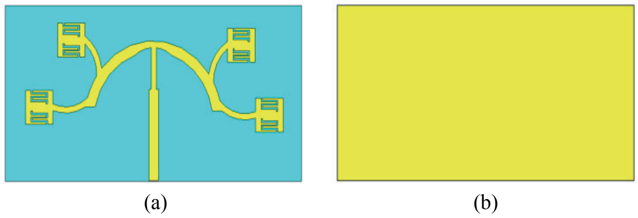

Fig. 4. The proposed antenna array. (a) CST model, (b) Ground plane.

The excitation signal was applied to the feed (port 1) with characteristic impedance Z_1 equal to 50 Ω and a width w_1 of 1.89 mm. The modified Wilkinson power divider-like shown in Fig. 2(d) further splits the 50 Ω into 2 × 100 Ω of width w_3 equal to 0.46 mm via a quarter wave transformer with impedance of Z_2 equal to 70.7 Ω and width w_2 of 1.05 mm. These transmission line widths w_1, w_2 and w_3 are dependent on the characteristic impedances at each branch and are determined by equations (7) and (8) stated in [14] for $w/h \geq 1$.

The entire design was modeled using full wave spectral analysis finite integral technique (FIT) commercial solver CST microwave studio [15]. The resulting simulation was photo etched on a laminate Duroid microwave board RO4003C with dielectric permittivity ε_S equal to 3.38, substrate thickness h of 0.813 mm, loss tangent δ of 0.0027, and a metal thickness t of 0.035 mm.

In Fig. 3, the proposed antenna was depicted with its CST capture as shown in Fig. 4. Each radiating element was positioned at an inter-element spacing of $\lambda_g/2$ on each side of the arm, and 1.3 $\lambda_g/2$ between the two divides where λ_g is the guided wavelength estimated to be 28 mm at 5.8 GHz resonance frequency. The element spacing of $\lambda_g/2$ was chosen to minimize the side lobe level [16] whereas the spacing between the two arms was set at 1.3 $\lambda_g/2$ in order to 1) effectively accommodate the corporate feed arrangement vis-à-vis its curve geometry, and 2) sufficiently provide enough spacing to overcome the mutual coupling effect between any two junctions [17].

Fig. 5 shows the current distribution over the proposed feed line connected to the radiating elements. For an input reference power of about 5 mW, the excitation current as seen from Fig. 5 was significant on all radiating elements. The surface current density values are between 145 A/m to 109 A/m. The peak current density of 145.351 A/m occurred at coordinate of (-15.24, 24.73, 0.85) with a phase of 292.5° as shown in the figure. In all, the radiating elements can be said to be optimally excited. Each radiating element was coupled at the non-radiating edge to deliver better performance [18].

Fig. 5. The current distribution of the proposed antenna array.

The proposed antenna naturally defrays losses particularly associated with radiation loss as discontinuities were eliminated to the barest minimum except at the quarter wavelength junction toward the source. Therefore, the strip corners were inherently metered sufficiently to secure a large bend in order to satisfy a scenario whereby the radiating elements, the transmission lines metal thickness t, and the grounded plane are assumed to be at least three to four skin depths thick. Usually, radiation loss increases with $(h\sqrt{\varepsilon_r}/\lambda_0)^2$.

Hence, a thin RT/Duroid RO4003C microwave substrate of dielectric permittivity of 3.38, a thickness of 0.813 mm, and a loss tangent (tan δ) of 0.00027 at λ_0 equal to 51.72 mm was used. Doing this reduces radiation loss sufficiently. As a matter of fact, the dielectric loss does not depend on the transmission line geometry but rather on the loss tangent (tan δ) of the substrate material. Hence, using a loss tangent of 0.00027 is sufficient to achieve low dielectric loss.

It has been identified in [19] that the radiation loss (two largest contributors to radiation among the various circuit discontinuities are the open circuit and the right-angle bend (RHB)) of a microstrip transmission line with a given thickness and characteristic impedance depends on the line length L, and that this dependency is oscillatory in nature. Hence, the losses grow with $(L/\lambda_0)^2$ in the range $0 < L > \lambda_0$, and insensitive otherwise. Knowing that losses are proportional to $1/Z_C$, both the radiation power (P_r) and surface-wave excitation (P_S) are loosely dependent on Z_C. However, the ohmic and dielectric losses are significant particularly in the dominant microstrip mode. Thus, it was observed that the dielectric attenuation loss factor differs not much irrespective of their width, nor do the dielectric loss depend on the transmission line geometry, but rather on the loss tangent (tan δ) of the substrate material. Therefore, the substrate with such loss tangent is adequate to reduce the loss.

It has also been noted that the RHB have high voltage standing wave ratio (VSWR) due to increased inductance, and as such, a decrease in the input reflection coefficient can be achieved if the corner is chamfered (mitered) via an instrumentality of increased inductance or decreased capacitance techniques. This is the basis for the proposed design where open circuit transmission lines are eliminated, and also, both left-angle bend (LHB) and RHB in the form of T-bend are curved in order to mitigate the losses. The ohmic attenuation losses α_o were calculated for the entire length (of width $w_1 = 0.189$ cm, $w_2 = 0.105$ cm, and $w_3 = 0.046$ cm translating to the characteristic impedances of 50 Ω, 70.7 Ω, and 100 Ω respectively) to be equal to 0.28 dB/cm with a corresponding ohmic losses of 0.63 dB.

By this it is evident that the proposed feed reduced the ohmic loss to less than 1.0. Besides, a metal thickness t proportional to three or four times the skin depth was used. Conscious effort was made to reduce the surface wave by using low impedances such as 50 and 70 Ω (except in few unavoidable cases where 100 Ω was expected to be used), in order to minimize surface excitation and hence its loss, as surface wave is loosely dependent of $1/Z_C$.

5. Results and Discussions

Fig. 6 shows the simulated and measured return loss results of the proposed antenna. Mode analysis of the antenna array feeder via CST microwave studio indicates that proposed feed excites the TM_{11} mode of the quasi-lumped element resonators, and hence, there exists only one resonance. By so doing, the impossibility of surface wave excitation was ascertained. The simulated frequency pattern was 5.37 – 5.84 GHz with a return loss better than 10 dB and precisely put at 26.19 dB.

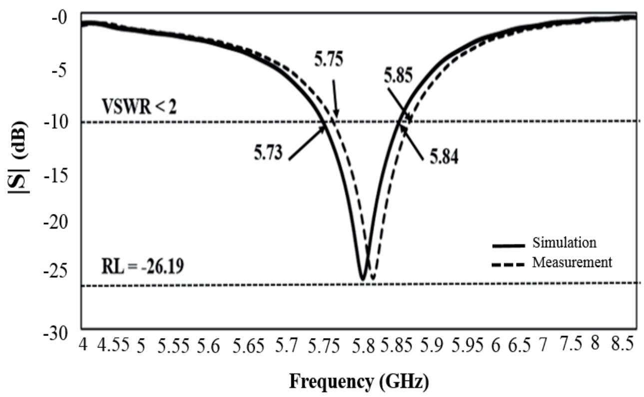

Fig. 6. The simulated and measured return loss.

The measured return loss result was not too different but with a frequency pattern of 5.75 – 5.85 GHz at the same return loss. Both the simulated and measured results agreed considerably. The simulated input impedance was

50 – j0.02 Ω whereas the measured impedance was 49 – j1.03 Ω. Both the simulated and measured input impedances are sufficient and adequate for proper impedance match, and hence, the VSWR ≤ 2:1 which is a measure of how well a transmission line is matched to the load, indicates a good match.

Fig. 7 shows the simulated and measured radiation patterns in *xz*- and *yz*-plane. The co-polar and cross polar fields are obtained by using Ludwig's (Ludwig III) third definition [20], where each pattern cut begins at θ = 0° and polarization axis is used to set the pattern cut angle. Using the same coordinate system and antenna alignment (i.e. antenna main polarization is aligned with *y*-axis, and *z*-axis is the roll axis), the reference and cross polarization directions can be expressed in terms of rectangular and spherical coordinate systems. A pattern cut is measured by rotating the support in azimuth. Then the polarization axes are aligned to next pattern cut angle. Both simulated and measured result is presented in Fig. 7(a) and (b) for co-polarization and cross-polarization measurements.

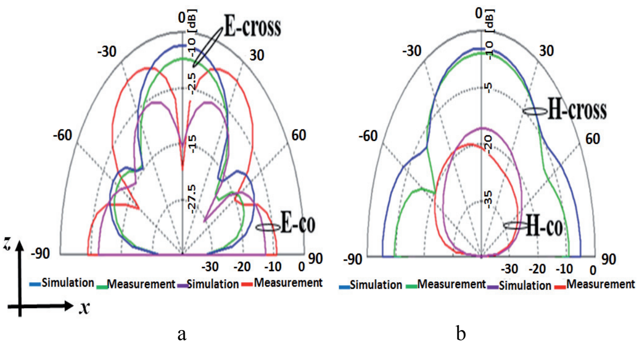

Fig. 7. The simulated and radiation patterns. (a) *xz*-plane, (b) *yz*-plane.

In both cases, the cross-polarization component was larger than the co-polarization thus indicating the left hand circular polarization (LHCP) peculiarity of the proposed antenna. In Fig. 7(a), the simulated main lobe magnitude was 9.8 dB with a main lobe oriented in 0° direction.

The beam width was 40° with a side lobe level of -5.0 dB. The measured result was similar except with a main lobe magnitude of 9.6 dB and a side lobe level of -7.2 dB. In *yz*-plane shown in Fig 7(b), the simulation indicated a main lobe magnitude of 9.5 dB oriented in 0° direction with a beam width of 51° and a side lobe level of -10.4 dB. The same scenario repeats itself in the measured radiation pattern with the main lobe magnitude of 9.57 dB oriented similarly but with a beam width of 48° and a side lobe level of -11.6 dB.

The simulated and measured gain against frequency was shown in Fig. 8. The simulated gain was 11.18 dBi whereas the measured gain was 10.97 dBi. It is noteworthy therefore that the proposed antenna of an aperture size of $0.406\lambda_0 \times 1.16\lambda_0$ sq. mm exhibits such a considerable gain. It then further demonstrates both the robustness and the efficiency of the proposed antenna.

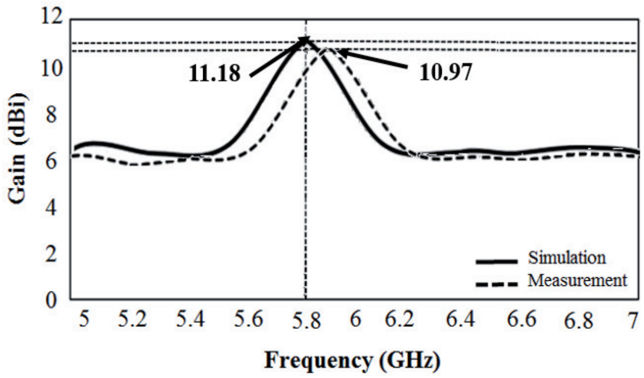

Fig. 8. The simulated and measured gain.

In [21], application of simulated annealing to design a serial feed sequentially rotated 2 × 2 antenna array was reported at a resonant frequency of 5.8 GHz with antenna real estate of $0.74\lambda_0 \times 0.74\lambda_0$, and a gain of 12.4 dBi. In a similar scenario, circularly polarized array antenna with corporate feed network and series-fed elements was reported by Ko & The-Nan, [22] where the array element inter-spacing was $0.85\lambda_0 \times 0.85\lambda_0$ and a gain of 15.3 dBic approximating 13 dBi. It was a 2 × 2-element antenna array operating at a centre frequency of 11 GHz.

Not only that, a dual-CP antenna using a new traveling-wave feed concept was reported by Kum et al., [23] with antenna dimensions of 65 × 65 sq. mm at a frequency of 15 GHz. The 8-element design exhibited a meager measured gain of 7.8 dBi notwithstanding its number of elements. Finally, two single-ring array with four (R = 17.3 mm) and eight (R = 31 mm) elements were developed by Min & Free [24]. The single-ring four-element array demonstrated a bigger estate area of 2π(R = 17.3) mm and a gain of 13 dB at a frequency of about 10 GHz.

Comparing these existing works with ours of antenna estate area of $0.406\lambda_0 \times 1.16\lambda_0$ sq. mm, and a simulated/measured gain of 11.18/10.97 dBi respectively indicates that our design is more superior particularly in terms of size advantage. Hence, it could be more preferable where size is a premium. Much more, the gain is reasonable when compared with these works with respect to their aperture size. Take for instance the proposed antenna demonstrated a gain advantage of 40.64% over the work reported in [23].

6. Conclusion

Experimental procedure to design a modified semi hybrid-ring directional coupler T-shaped fed 4-element quasi-lumped element antenna is presented. The hybrid-ring directional coupler T-type was divided along its plane of symmetry, and turned outward in order to achieve the modified semi hybrid-ring directional T-type corporate feed. It was observed that the proposed feed provides better performance compared with the conventional corporate feed network in terms of gain, as well as the compact antenna real estate capability.

Also, the usual losses associated with the conventional corporate feed were adequately reduced by 1) the inherent curve geometry of the modified hybrid-ring T-shaped which automatically limits the number of junctions, and where they are unavoidably present, were naturally metered sufficiently to secure a large bend; and 2) the feed was optimized using a thin substrate with a metal thickness of 0.035 mm. In general, both the simulated results and measured results agreed significantly. The proposed antenna is simple, easy to fabricate and cost effective.

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