

A generalized 90° out-of-phase Wilkinson power divider for dual port UHF CP RFID antennae with variable port distance

W. Akash Sovis*^{ID} and Manilka Jayasooriya^{†‡}

Abstract

In this paper a microstrip Wilkinson power divider with a 90° phase delay at one output port is proposed to obtain circular polarization to feed a dual port RFID antenna. The 90° phase delay was obtained by embedding an extra quarter wavelength at one port of the Wilkinson power divider. The feeding circuit is then mounted on the ground plane of the microstrip antenna feeding the radiating patch directly through the ground plane and dielectric layer thus reducing any fringing effect and resulting a mechanically compact unit. The proposed feeding method offers better expectation of antenna performance with minimal attenuation and coupling losses. The design process generalizes geometric parameters of the Wilkinson power divider for variable port distances. The paper considers both UK and US RFID center frequencies, 870 MHz and 915 MHz respectively. Numerically computed values for geometric design parameters for both frequencies are tabled as future design tools for port distances varying from 18 mm up to 34 mm at 870 MHz and 17 mm up to 32 mm at 915 MHz. Simulation results indicate a return loss (S_{11}) of -20 dB and -26 dB at 870 MHz and 915 MHz operational frequencies respectively at 270° angled quarter wavelength.

Key words: Wilkinson power divider, RFID, dual port feeder, port distance, circular polarization

1 Introduction

The field of Radio Frequency Identification (RFID) and its applications have been drastically expanding after the data revolution and coming age of automated human-free manufacturing industry: Industry 4.0. [1] [2] The technology is built around transportation, vehicle systems, logistics, healthcare, security, payment and detection in general. Microstrip antennae have been considered the most suitable for RFID applications due to its compact nature and performance. The RFID reader antenna plays a pivotal role in the overall

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efficiency of the system. For instance, when the RFID transponders are aligned with the antenna's polarization: a linearly polarized antenna will read further with a higher efficiency than a circularly polarized antenna. However, if the transponders are not aligned with the antenna's polarization the circularly polarized antenna will have the advantage over linearly polarized antennae. [3] [4] For certain detection systems where the transponders are mobile traveling in different orientations, a circularly polarized antenna suits best in avoiding blind spots as well as increasing the efficiency of the reader system. Blind or weak spots in an RFID system occur when two or more transponders are stacked together resulting in detuning due to mutual coupling of the transponder antennae. These blind spots are not necessarily distributed monotonically along the stack. Research has been performed in designing no surface dead zone antennae. [5] [6] [7] [8] The RFID reader transmits an electromagnetic signal of a full bandwidth less than 1 MHz hopping 75 or more channels 30 seconds by swiftly switching the carrier signals through the available ETSI frequency channels. Another solution is to use a circularly polarized antenna where a spiral beam rotating with time is emitted thus giving advantage for the reader to detect transponders in any orientation.

A circularly polarized beam can be obtained theoretically if the two orthogonal modes are excited with a 90° time phase difference is maintained between them. By doing so when the horizontal current flow is at its maximum the vertical current flow will be zero and vice versa in the other quarter cycle. Geometrically this can be achieved by one of the following methods. Truncating the diagonal edges of a square patch making the surface currents flow in a circular manner, using a single feed on an approximate square patch if the length were a bit less than the resonant length but a bit more than the height, perforating an asymmetric slot on the patch making the surface currents flow in a circular manner or by using two feeds one excited with a 90° time phase delay. [1] [9] [10] Using two feeds directly with a coaxial probe reduces the fringing effect compared to a stripline feeder. This method demands the power from the RFID reader to be split equally and fed to two coaxial ports with a 90° time phase delay.

T-junctions define poor isolation between the output ports many antenna feeding circuits prefer the equal split Wilkinson power divider. [11] A two port Wilkinson power divider comprises two quarter wavelength arms having a characteristic impedance of $50 \times \sqrt{2}$ (i.e. 70.7) ohms and a 100 ohm shunt resistor at the output ports thus giving the same 50 ohm port input impedance at the two output ports. [12] [13]

In this work an additional quarter wavelength with a characteristic impedance of 50 ohms is added to one output port of the Wilkinson divider in order to obtain the 90° time-phase difference. A stripline Wilkinson power divider with an embedded quarter wavelength arm is attached to be mounted on the reverse side of the antenna thus feeding the radiating patch via two coaxial probes through the ground plane and the dielectric layer. In this paper the design process is carried out with algebraic expressions for the purpose of generalization. A table is then generated to obtain design parameters for different port distances from which a feeding circuit could be designed for a dual port CP RFID antenna with a specific port distance. All calculations are done for both US and UK UHF RFID center frequencies. Numerical calculations were performed using Mathcad Prime 5.0 and RF models were simulated using CST Microwave studio suite.

2 Development of generalized Wilkinson divider with embedded quarter wavelength element

The general approach in designing a Wilkinson power divider with an extra quarter wavelength is to have a microstrip Wilkinson splitter with RG58/59 cables one having an extra quarter wavelength as shown in Figure 1

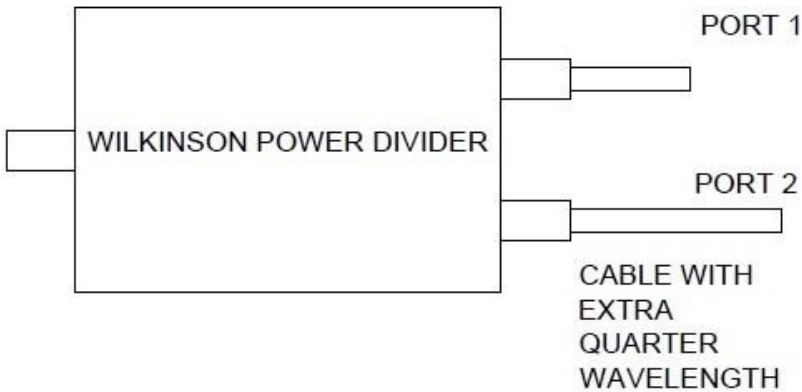


FIGURE 1: General Wilkinson divider with extra quarter wavelength as cable

The originality of the proposed design is the application of the extra quarter wavelength to the Wilkinson power divider itself and compute the port distance algebraically making it possible to numerically reverse engineer the antenna feeder's design parameters for different port distances. This makes possible to design a feeder circuit for a circular polarized dual port antenna constrained by the distance between the two ports. The feeder can be attached to the ground plane through which the coaxial pins travel to the radiating patch. The circuit designed in this work is fabricated on a standard FR4 PCB board having a thickness of 1.6 mm and a copper cladding of 35 microns. A block diagram is shown in Figure 2

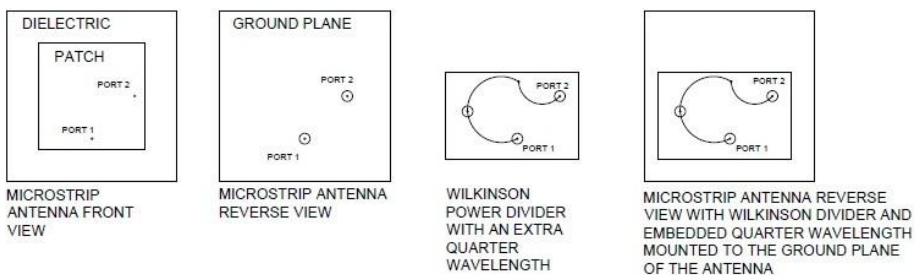


FIGURE 2: Block diagrams of antenna feeder mounting the microstrip patch antenna

The equal split 2 quarter wavelength arms having a characteristic impedance of 70.7 ohms can be constructed with 2 approximate semicircle arcs as shown in Figure 3 The gap between the two arcs are set such that a surface mount 100 ohm resistor could be soldered in between.

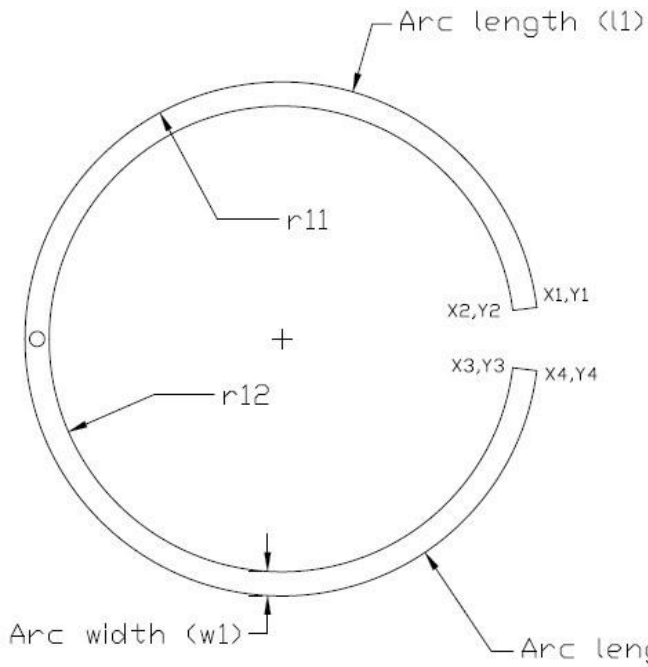


FIGURE 3: 70.7 ohm quarter wavelength arms of lengths $l1$ having stripline widths of $w1$ of the Wilkinson power divider

The extra quarter wavelength arm is attached to coordinates $(x1, y1)$ and $(x2, y2)$. The extra quarter wavelength is a stripline with a characteristic impedance of 50 ohms to match the impedance as shown in

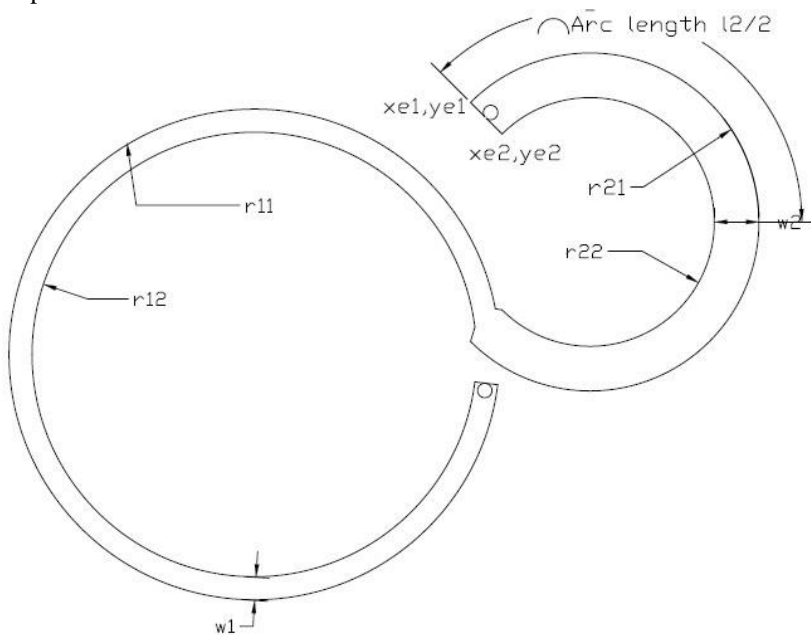


FIGURE 4: Additional quarter wavelength element added to Wilkinson divider for 90 degree time phase shift

Using Wheeler's equation [14] [3, 15] for characteristic impedance on a stripline fabricated on a FR4 PCB substrate having a dielectric constant of 4.3 ($\epsilon_r = 4.3$) and a thickness of 1.6mm ($h = 1.6$) and a copper thickness of 35 microns ($t = 0.035$), the effective width of the stripline for a width w is given by Equation (1)

$$w_{eff} = w + \frac{t}{\pi} \times \ln \left(\frac{4 \times e^1}{\sqrt{\left(\frac{t}{h}\right)^2 + \left(\frac{t}{w \times \pi + 1.1 \times t \times \pi}\right)^2}} \right) \times \frac{\epsilon_r + 1}{2 \times \epsilon_r} \quad (1)$$

If the following parameters $x1$ and $x2$ were defined as

$$x1 = 4 \times \left(\frac{14\epsilon_r + 8}{11\epsilon_r} \right) \times \left(\frac{h}{w_{eff}} \right)$$

$$x2 = \sqrt{16 \times \left(\frac{h}{w_{eff}} \right)^2 \times \left(\frac{14\epsilon_r + 8}{11\epsilon_r} \right)^2 + \left(\frac{\epsilon_r + 1}{2\epsilon_r} \right) \times \pi}$$

The characteristic impedance of the stripline is given by Equation (2)

$$Z0 = \frac{\eta}{2\pi \times \sqrt{2} \times \sqrt{\epsilon_r + 1}} \times \ln \left(1 + \frac{4 \times h \times (x1 + x2)}{w_{eff}} \right) \quad (2)$$

By using Equation (2) the stripline widths for 50 ohm and 70.7 ohms are calculated as 3.1 mm and 1.6 mm respectively, where η is the wave impedance in free space. (Characteristic impedance is independent from resonant frequency) [16] [17] [18]

A microstrip quarter wavelength can be calculated by Equation (3)

$$\frac{\lambda}{4} = \frac{c}{4f \sqrt{\epsilon_{eff}}} \quad (3)$$

Where f is the frequency, ϵ_{eff} is the effective permittivity and c the speed of light.

The effective permittivity can be calculated by Equation (4) and Equation (5)

When $w > h$

$$\epsilon_{reff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} * \left(1 + 12 * \frac{h}{w} \right)^{\frac{1}{2}} \quad (4)$$

And when $w \leq h$

$$\epsilon_{reff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \times \left(\frac{1}{\sqrt{\left(1 + 12 * \frac{h}{w} \right)}} + 0.04 \times \left(1 - \frac{w}{h} \right)^2 \right) \quad (5)$$

Values for $l1$ and $l2$ were obtained for UK and US RFID frequencies and are listed in Table 1

TABLE 1
 $l1$ and $l2$ values for UK and US RFID frequencies

	Z0	stripline width	at 870 MHz	at 915 MHz
l1	70.7 ohms	w1=1.6 mm	49 mm	46 mm
l2	50 ohms	w2=3.1 mm	48 mm	45 mm

The center of the second arc (x_2, y_2) is placed at 45 degrees (θ_3) as shown in Figure 5 from the end of first arc (x_1, y_1) Radii $r1$ and $r2$ are the averages of $r11, r12$ and $r21, r22$ respectively.

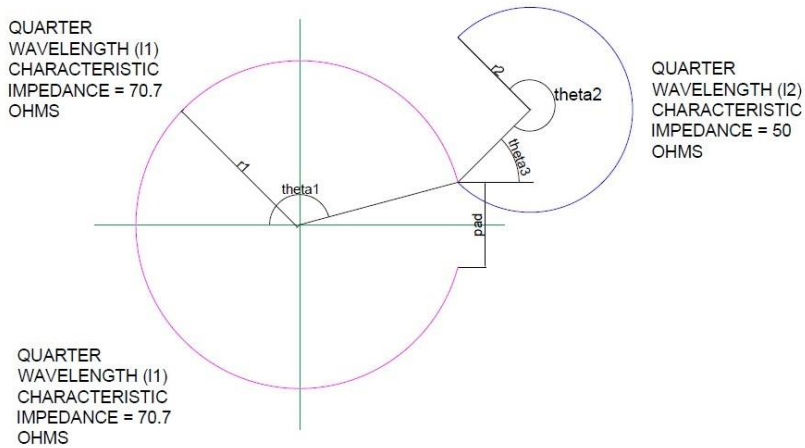


FIGURE 5: 2nd arc placed θ_2 angled from the end of arc 1

Distance between the two ports is defined M as depicted in Figure 6

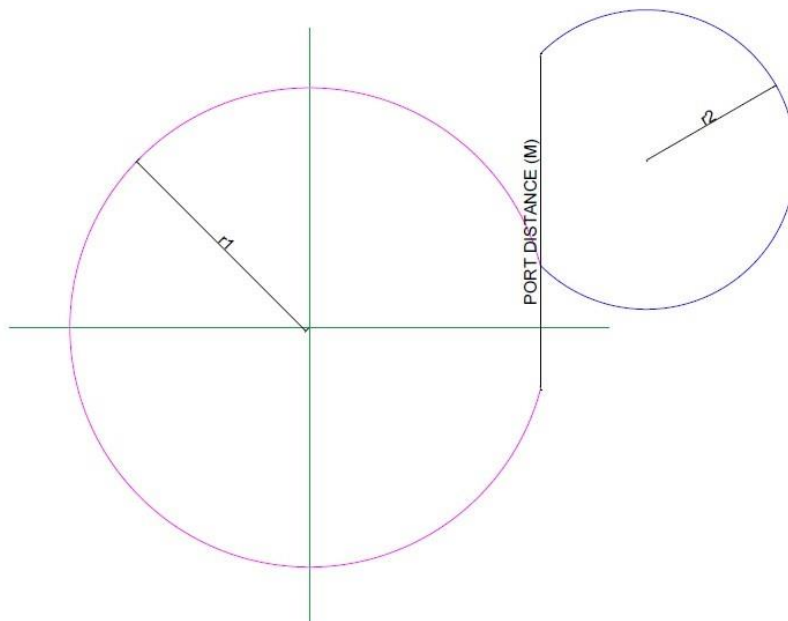


FIGURE 6: Defining port distance M

The 100 ohm shunt resistor is connected with a distance defined pad . Variable pad is set to 4 mm for 100 ohm SMD resistor. The angle between two ports of the Wilkinson divider without the quarter wavelength is $2 \times phi$ as defined in Figure 7

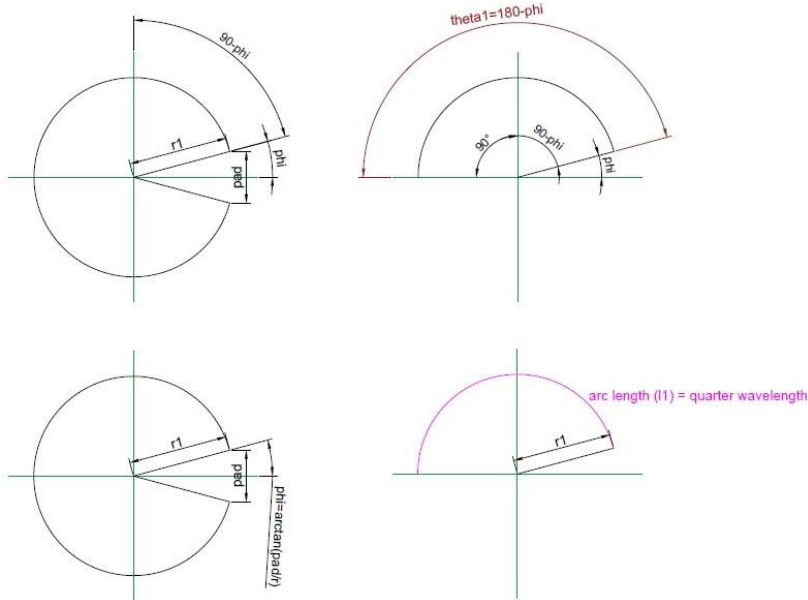


FIGURE 7: variables in the Wilkinson power divider without quarter wavelength element

By Figure 7 it can be written that $\phi = \sin^{-1}\left(\frac{pad}{r1}\right) = \sin^{-1}\left(\frac{pad}{2r1}\right)$

And $\theta_1 = \pi - \phi$

$r1$ can be calculated by solving Equation (6) numerically (Mathcad's symbolic engine was used in this design)

$$l1 = r1 \times \left(\pi - \sin^{-1}\left(\frac{pad}{2r1}\right) \right) \quad (6)$$

To generalize the distance between the two ports angle θ_2 is varied thus varying the position of the upper port and thereby the distance between them. The end point coordinate can then be calculated as a phasor with the magnitude of $r2$ and varying angle from the horizontal plane keeping $l2$ constant. This is depicted in Figure 8

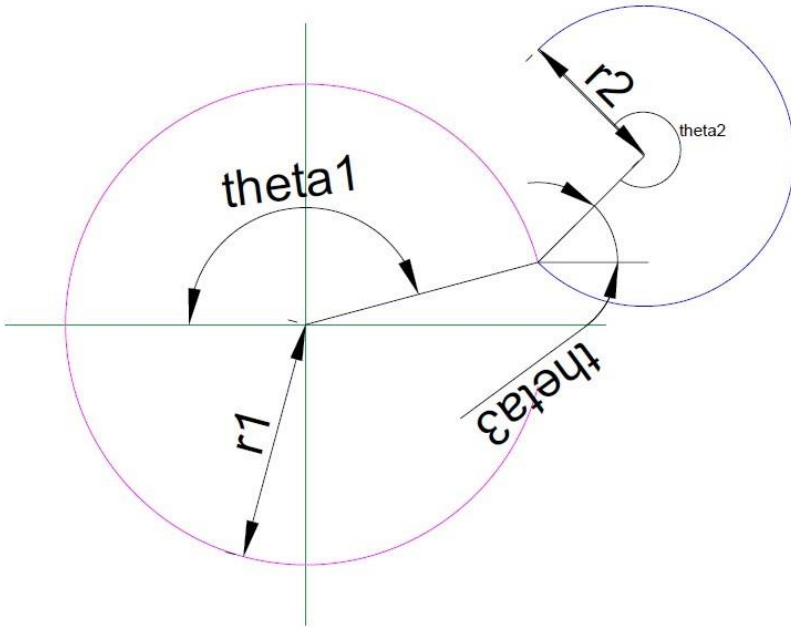


FIGURE 8: End point calculation in terms of θ_2 and r_2

The end points of Wilkinson arms as depicted in Figure 3 can be calculated by the following expressions

$$\begin{aligned}(x_1, y_1) &= [r_{11} \cos(\phi), r_{11} \sin(\phi)] \\(x_2, y_2) &= [r_{12} \cos(\phi), r_{12} \sin(\phi)] \\(x_3, y_3) &= [r_{12} \cos(\phi), -r_{12} \sin(\phi)] \\(x_4, y_4) &= [r_{11} \cos(\phi), -r_{11} \sin(\phi)]\end{aligned}$$

Where outer and inner arc's radii of first and second arc are given as

$$\begin{aligned}r_{11} &= r_1 + \frac{w_1}{2} \\r_{12} &= r_1 - \frac{w_1}{2} \\r_{21} &= r_2 + \frac{w_2}{2} \\r_{22} &= r_2 - \frac{w_2}{2}\end{aligned}$$

Center point of the second arc (x_r, y_r) can be calculated as

$$\begin{aligned}x_r &= r_1 \cos(\phi) + r_2 \cos(\theta_3) \\y_r &= r_1 \sin(\phi) + r_2 \sin(\theta_3)\end{aligned}$$

The location of the end point coordinates can be parameterized by changing θ_2 however, to compensate for the quarter wavelength (i.e. to keep second arc length l_2 constant) radius r_2 should be changed accordingly. Here r_2 and θ_2 are inversely proportional as shown by Equation (7)

$$l_2 = r_2 \times \theta_2 \quad (7)$$

A new angle θ_4 is defined to calculate the angle from horizontal x axis upwards to use as a variable. This is computed as

$$\begin{aligned}\theta_2 &= \left(\frac{\pi}{2} - \theta_3\right) + \frac{\pi}{2} + \theta_4 \\ \theta_2 &= \pi - \theta_3 + \theta_4 \\ \theta_4 &= \theta_2 - \pi + \theta_3\end{aligned}$$

The end point coordinates can be calculated using θ_4 as shown in Figure 9

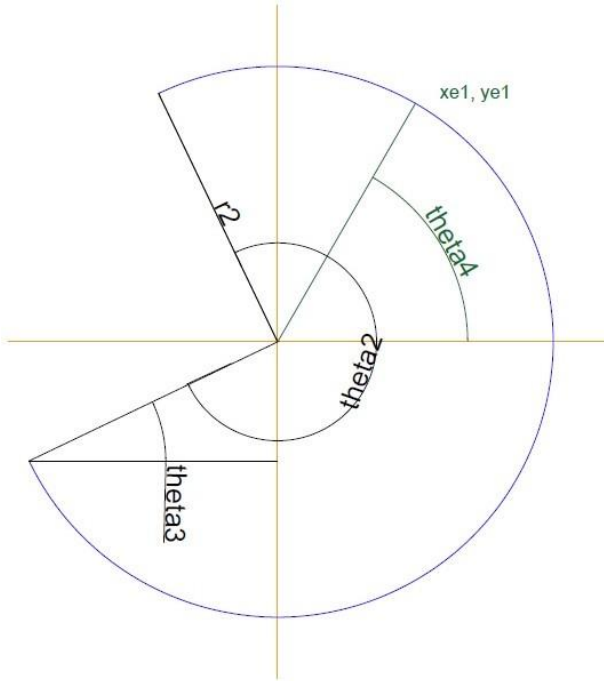


FIGURE 9: End point coordinates by variable θ_4 for generalization

The end point coordinate values can be written as:

$$\begin{aligned}xe1 &= xr + r21 \cos (\theta_4) \\ ye1 &= yr + r21 \sin (\theta_4) \\ xe2 &= xr + r22 \cos (\theta_4) \\ ye2 &= yr + r22 \sin (\theta_4)\end{aligned}$$

The distance between ports defined as M is between end points of the second arc and lower ending of the first arc. A mid-point is calculated from both ends and the diagonal distance M is calculated by Equation (8)

$$M = \sqrt{\left(\frac{ye1 + ye2}{2} - \frac{y3 + y4}{2}\right)^2 + \left(\frac{xe1 + xe2}{2} - \frac{x3 + x4}{2}\right)^2} \quad (8)$$

Using CST Microwave studio models were created for different angles of θ_2 as shown in Figure 10 varying r_2 accordingly using Equation (7) to maintain a quarter wavelength

(12) in the second arc.

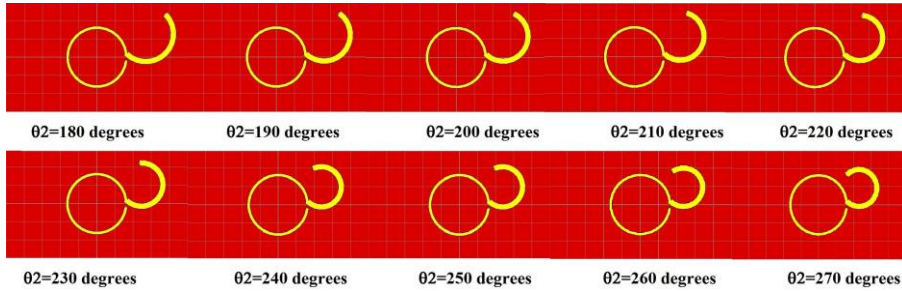


FIGURE 10: θ_2 and r_2 varying to obtain different port distance values (M)

3 Design parameters for UK RFID frequency 870 MHz

Considering origin as center of arcs 11, parameters which remain constant for Wilkinson power divider as described in Figure 4 for UK RFID frequency 870 MHz is listed in Table 2 (All dimensions measured in millimeters with a maximum error of 5 microns)

TABLE 2
Parameters remaining constant for UK 870 MHz RFID frequency

w1	w2	l1	l2	r1	pad	θ_1
1.6	3.1	49	48	16.24	4	172.92°

Using computations in section 2 of this paper Wilkinson power divider with embedded quarter wavelength arm at different θ_2 values could be carried out for 870MHz.

Design parameters for UK RFID (870 MHz) with varying θ_2 is listed in Table 3

TABLE 3
Design parameters for 870 MHz with varying θ_2 angle

θ_2°	r_2	xr	yr	$xe1$	$ye1$	$xe2$	$ye2$	M
180	15.28	26.92	12.80	38.82	24.70	36.62	25.51	22.51
190	14.47	26.35	12.24	35.54	25.36	33.76	22.82	22.82
200	13.75	25.84	11.72	32.30	25.59	30.99	22.78	22.78
210	13.10	25.37	11.26	29.16	25.41	28.36	22.41	22.41
220	12.50	24.95	10.84	26.18	25.84	25.91	21.75	21.75
230	11.96	24.57	10.46	23.39	23.91	23.66	20.82	20.82
240	11.46	24.21	10.10	20.85	22.67	21.65	19.67	19.67
250	11.00	23.89	9.78	18.59	21.15	19.90	18.34	18.34
260	10.58	23.59	9.48	16.64	19.41	18.41	16.81	16.87
270	10.19	23.31	9.20	15.02	17.50	17.21	15.31	15.31

Using Mathcad Prime 5.0 a graph is generated theoretically to calculate θ_2 from port distance M and is depicted in Figure 11

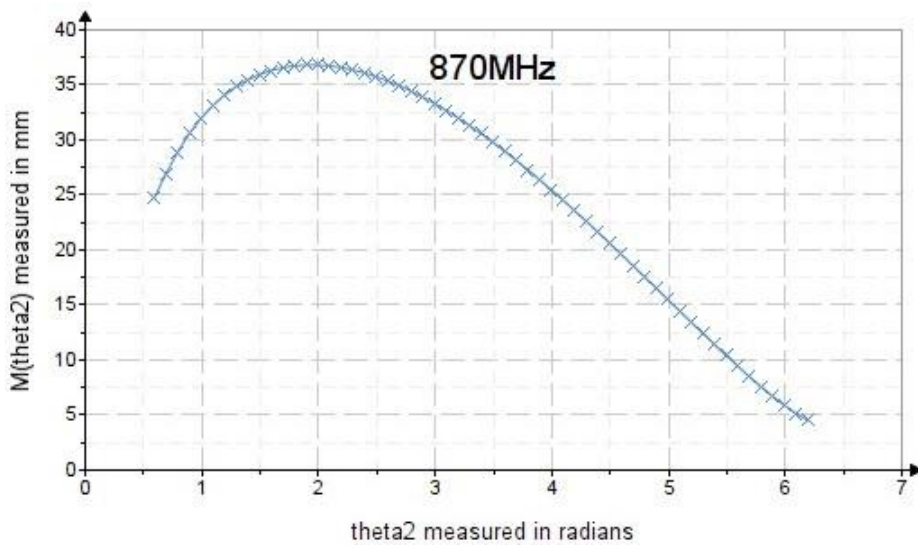


FIGURE 11: Relationship between port distance (M) and θ_2

For specific port distance (M) graph in Figure 11 can be used to find design parameters and are listed in Table 4

TABLE 4
Design parameters for 870 MHz with varying θ_2 angle

M	θ_2°	r_2	x_r	y_r	x_{e1}	y_{e1}	x_{e2}	y_{e2}
34	165	16.67	27.90	13.79	43.67	22.89	41.00	21.34
33	175	15.72	27.22	13.11	40.45	24.21	38.08	22.22
32	184	14.95	26.68	12.57	37.50	25.02	35.47	22.68
31	192	14.32	26.24	12.13	34.89	25.44	33.20	22.84
30	199	13.82	25.88	11.77	32.62	25.59	31.26	22.80
29	206	13.35	25.55	11.44	30.40	25.53	29.39	22.60
28	213	12.91	25.24	11.13	28.25	25.28	27.60	22.24
27	219	12.56	24.99	10.88	26.47	24.91	26.14	21.83
26	226	12.17	24.72	10.60	24.48	24.32	24.53	21.22
25	232	11.85	24.49	10.38	22.86	23.69	23.24	20.61
24	238	11.55	24.28	10.17	21.33	22.94	22.03	19.92
23	244	11.27	24.08	9.97	19.91	22.09	20.92	19.16
22	250	11.00	23.89	9.78	18.59	21.15	19.89	18.34
21	255	10.78	23.74	9.63	17.57	20.31	19.12	17.62
20	261	10.54	23.56	9.45	16.46	19.23	18.28	16.72
19	267	10.30	23.40	9.28	15.47	18.09	17.54	15.79
18	272	10.11	23.26	9.15	14.73	17.10	17.00	15.00

4 Design parameters for US RFID frequency 915 MHz

Considering origin as center of arcs l_1 , parameters which remain constant for Wilkinson power divider as described in Figure 4 for UK RFID frequency 870 MHz is listed in Table 5 (All dimensions measured in millimeters with a maximum error of 5 microns)

TABLE 5
Parameters remaining constant for 915 MHz RFID frequency

w1	w2	l1	l2	r1	pad	θ_1
1.6	3.1	46	45	15.28	4	172.48°

Using computations in section 2 of this paper Wilkinson power divider with embedded quarter wavelength arm at different θ_2 values could be carried out for 915 MHz.

Design parameters for US RFID (915 MHz) with varying θ_2 is listed in Table 6

TABLE 6
Design parameters for 915 MHz with varying θ_2 angle

θ_2°	r2	xr	yr	xe1	ye1	xe2	ye2	M
180	14.32	25.28	12.13	36.50	23.35	24.31	21.16	31.60
190	13.57	24.74	11.59	33.42	23.98	31.64	21.44	30.21
200	12.89	24.26	11.12	30.37	24.20	29.06	21.39	28.76
210	12.28	23.83	10.68	27.41	24.03	26.61	21.04	27.26
220	11.72	23.44	10.29	24.59	23.51	24.32	20.42	25.71
230	11.21	23.08	9.93	21.96	22.64	22.23	19.55	24.12
240	10.74	22.75	9.60	19.56	21.47	20.36	18.48	22.49
250	10.31	22.44	9.29	17.43	20.04	18.74	17.23	20.85
260	9.92	22.16	9.01	15.58	18.40	17.36	15.87	19.18
270	9.54	21.90	8.75	14.05	16.60	16.24	14.41	17.5

Using Mathcad Prime 5.0 a graph is generated theoretically to calculate θ_2 from port distance M and is depicted in Figure 12

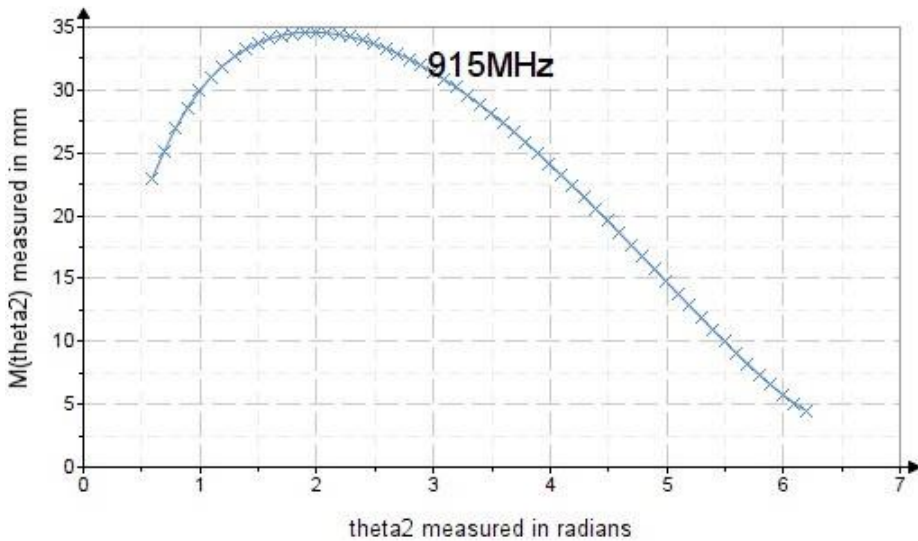


FIGURE 12: Relationship between port distance (M) and θ_2

For specific port distance (M) graph in Figure 12 can be used to find design parameters and are listed in Table 7

TABLE 7
Design parameters for 915 MHz with varying θ_2 angle

M	θ_2°	$r2$	xr	yr	$xe1$	$ye1$	$xe2$	$ye2$
32	166	15.53	26.13	12.98	40.77	21.78	38.12	20.18
31	176	14.65	25.51	12.36	37.73	22.99	35.39	20.95
30	185	13.94	25.00	11.84	24.96	23.72	32.97	21.34
29	194	13.29	24.55	11.40	32.19	24.12	30.59	21.46
28	202	12.76	24.17	11.03	29.77	24.20	28.58	21.35
27	209	12.34	23.87	10.72	27.70	24.07	26.85	21.09
26	216	11.94	23.59	10.44	25.70	23.76	25.21	20.07
25	223	11.56	23.32	10.18	23.78	23.28	23.67	20.18
24	230	11.21	23.08	9.93	21.96	22.64	22.23	19.55
23	236	10.93	22.87	9.73	20.49	21.97	21.07	18.93
22	242	10.65	22.68	9.53	19.11	21.20	20.02	18.24
21	249	10.35	22.47	9.32	17.63	20.20	18.89	17.37
20	255	10.11	22.30	9.15	16.47	19.25	18.02	16.56
19	262	9.84	22.11	8.96	15.25	18.06	17.12	15.58
18	267	9.66	21.98	8.83	14.48	17.16	16.55	14.85
17	273	9.44	21.83	8.63	13.66	16.03	15.96	13.96

5 Simulation Results

The proposed design was modelled with coaxial ports as depicted in Figure 13 on CST Microwave Studio Suite as shown in and simulated for both UHF RFID center frequencies.

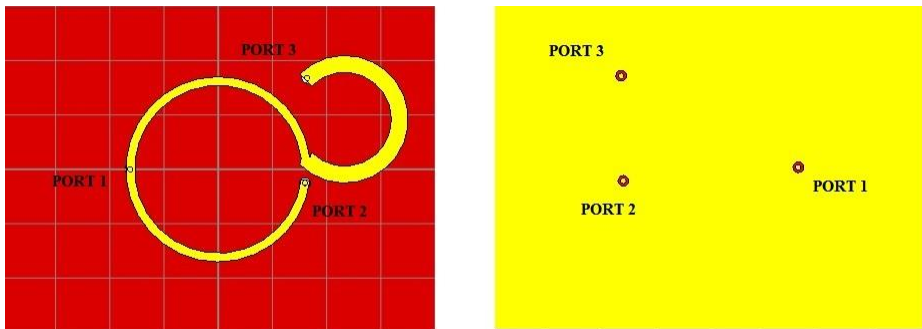


FIGURE 13: Coaxial ports on CST Model

S parameter simulation results for 870 MHz are shown in

Figure 14

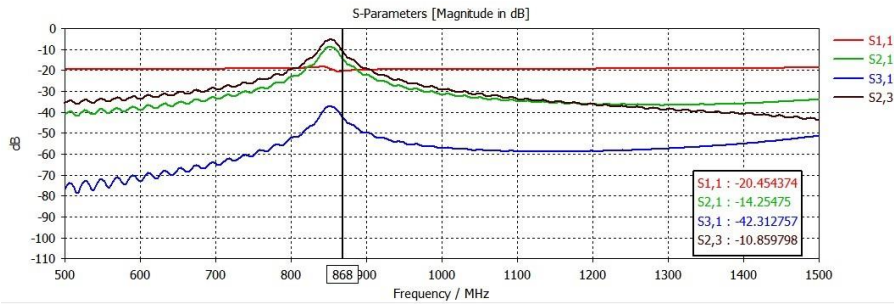


FIGURE 14: S11, S21, S31 and S12 at 870 MHz

S parameter simulation results for 915 MHz are shown in Figure 15

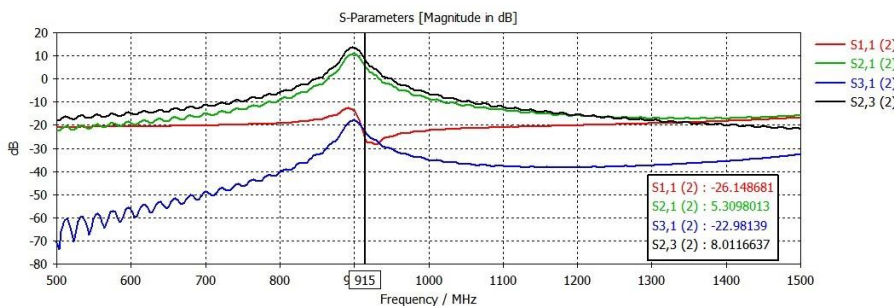


FIGURE 15: S11, S21, S31 and S12 at 915 MHz

Conclusion

In this paper a generalized antenna feeder for a dual port circular polarized RFID antenna is proposed. A Wilkinson power divider with an extra quarter wavelength arm embedded to one of the output ports provides a 90° phase difference suitable for a dual port CP RFID antenna. The design is first generalized algebraically and numerically computed tables are presented for design parameters for antenna feeders with different port distances varying from 18 mm to 34 mm and 17 mm to 32 mm at 870 MHz and 915 MHz respectively. These can be used as design guides in future work. Circular arcs are used for striplines. Generalizing is done by varying the angle of the quarter wavelength arm and radius thus keeping its length at the constant wavelength. Simulation results are obtained by modeling the proposed design on CST Microwave Studio Suite. Simulation results indicate a return loss of -20 dB and -26 dB at 870MHz and 915 MHz respectively at a 270° quarter wavelength angled arm.

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