Development of Dielectric Resonator Antenna (DRA)

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- The DRA is an antenna that makes use of a radiating mode of a dielectric resonator (DR).
- It is a 3-dimensional device of any shape, e.g., hemispherical, cylindrical, rectangular, triangular, etc.
- Resonance frequency determined by the its dimensions and dielectric constant Er.

Some DRs :



Advantages of the DRA

- Low cost
- Low loss (no conductor loss)
- Small size and light weight
- Reasonable bandwidth (~10% for εr ~10)
- Easy of excitation
- High radiation efficiency (generally > 95%)

Excitation schemes

(i) Microstrip line feed



Excitation schemes

(ii) Aperture-couple feed



Excitation schemes

(iii) Coaxial feed



Coaxial feed



Top view

Bottom view

Aperture-coupled feed





Bottom view

Top view

Corporate feedline for DRA array



Slot-fed DRA array using corporate microstrip feed network

Conformal-Strip Method



Rectangular Dielectric Resonator Antennas

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Proposed Antenna Geometry



Analytical Solution

• Dielectric Waveguide Model (DWM)

Resonant frequency of $TE_{mnl}(y)$ mode

$$f_0 = \frac{c}{2\pi\sqrt{\varepsilon_r}}\sqrt{k_x^2 + k_y^2 + k_z^2}$$

$$k_x = \frac{m\pi}{a}, k_y = \frac{n\pi}{b}, k_z = \frac{l\pi}{2d}$$

$$k_x^2 + k_y^2 + k_z^2 = \varepsilon_r k_0^2$$

Numerical Solution

• Finite-Difference Time-Domain (FDTD) method

Advantages

- Very simple
- High modeling capability for general EM structures
- No spurious modes nor large matrix manipulation
- Provide a very wideband frequency response

Disadvantages

- Time consuming, powerful computer required

Source model and extraction of S parameters

Baseband Gaussian pulse

$$E_z = \exp[-(\Delta t \cdot n - 3T)^2/T^2]$$
 T : pulse width



Parameters

Uniform Cartesian grids

 $\Delta x = 0.715 \text{ mm}, \Delta y = 0.508 \text{ mm}, \Delta z = 0.5 \text{ mm}$

 $T = 0.083 \text{ns}, t_0 = 3T$

10-cell-thick PML with polynomial spatial scaling (m = 4 and $\kappa_{max} = 1$)

total grid size : $80\Delta x \times 110\Delta y \times 112\Delta z$

total time steps : 10000

Input Impedance/S₁₁



- Reasonable agreement.
- Wide Bandwidth of $\sim 43\%$.
- Dual resonant TE_{111}^{y} and TE_{113}^{y} modes are excited.

Comparison between Theory and Measurement

Resonant Modes	Measured resonant frequencies	Calculated resonant frequencies (FDTD)		Predicted resonant frequencies (DWM)	
	f _{mea} (GHz)	f _{FDTD} (GHz)	error (%)	f _{DWM} (GHz)	error (%)
TE_{111}^{y}	3.81	3.90	2.3	3.95	3.6
TE ^{<i>y</i>} ₁₁₂	N/A	N/A	N/A	4.26	N/A
TE_{113}^{y}	4.57	4.60	0.7	4.7	1.7

• Reasonable agreement.

Field Distribution --- TE₁₁₁^y





Imaged DRA (gound plane removed)

With gound plane

Field Distribution --- TE₁₁₂^y



Imaged DRA (gound plane removed)

Field Distribution --- TE₁₁₃^y



Imaged DRA (gound plane removed)

With gound plane

Radiation Patterns



f = 3.5 GHz

f = 4.3 GHz

- Broadside radiation patterns are observed.
- Measured E-plane crosspolarized fields mainly caused by finite ground plane diffraction.

III. Circularly Polarized Design using a Parasitic Strip

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Proposed Antenna Geometry



Input Impedance/S₁₁



- Reasonable agreement.
- Bandwidth $\sim 14\%$.
- Two nearly-degenerate $TE_{111}(y)$ modes are excited. \Rightarrow CP operation

Axial Ratio in the boresight direction



3-dB AR bandwidth is ~ 2.7%, which is a typical value for a singly-fed CP DRA.

The H field of the DRA without and with parasitic strip (Top view)



Radiation Patterns (f = 3.4GHz,)



- A broadside radiation mode is observed.
- For each radiation plane, the LHCP field is more than 20dB stronger than the RHCP field.
- The maximum gain is 5.7 dBic (not shown here).

Effects of feeding strip length l_1



- Input impedance changes substantially with l_1 .
- AR is almost unchanged for different l_1 .
- l_1 can be adjusted to match the impedance without changing AR.

II. Frequency Tuning Technique

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Backgruond

•The DRA for a paticular frequency may not be available from the comericial market.

•Fabrication tolerances cause errors between measured and calculated resonant frequencies.

Frequency tuning methods:
(i) loading-disk; and
(ii) parasitic slot.

Frequency Tuning Technique - using a loading disk

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The slot-coupled DRA with a conducting loading



•Hemispherical DRA: radius a = 12.5 mm, dielectric constant $\varepsilon_r = 9.5$. •Coupling slot : length L_s , width W_s

• Open-circuit stub: length L_t

cap

- •Grounded dielectric slab: $\varepsilon_{rs} = 2.33$, height d = 1.57 mm
- Microstrip feedline: width $W_f = 4.7 \text{ mm}$

Calculated and measured return losses $(L_s = 12 \text{ mm and } W_s = 1 \text{ mm})$



Resonance frequency:

- 3.52 GHz without any conducting cap ($\alpha = 0^0$), with $L_t = 4.42$ mm
- 3.25 GHz (α = 26.38° and L_t = 4.42 mm)
- 3.68 GHz (α = 52.8° and L_t = 13.6 mm)
Calculated and measured radiation patterns



• Reasonable agreement between theory and experiment.

3.25 GHz (α = 26.38° and L_t = 4.42 mm)



• The effect of loading cap on field pattern is not significant.

3.58 GHz (α = 52.8° and L_t = 13.6 mm)

Calculated α and L_t for having a good return loss (minimum $|S_{11}| < -20$ dB)



The resonant frequency can be tuned by varying α and L_t • α decreases from 26.38° to 0° (3.25 < f_r < 3.5 GHz) • α increases from 0° to 52.8° (3.5 < f_r < 3.78 GHz)

Impedance bandwidth



• The bandwidth decreases after a loading cap is added.

Frequency Tuning Technique - using a parasitic slot

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The annular-slot-excited cavity-backed DRA



(a) Side view



IV. Omnidirectional Circularly Polarized DRA

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Advantages of omnidirectional CP antenna

• Provide larger coverage.

CP DRAs concentrated on broadside-mode designs only.



Slotted omnidirectional CP DRA

Antenna configurations



 $\int_{x}^{z} \theta$ Slot $\int_{y}^{y} \frac{1}{y} \frac{1}{y}$

- Perspective view
- Dielectric cube with oblique slots (polarizer) fabricated on its four sidewalls.
- Centrally fed by a coaxial probe extended from a SMA connector, whose flange used as the small ground plane.



Photographs of the prototype

Prototype for 2.4 GHz WLAN design



Top face and sidewalls



Bottom face

Design parameters

 $\varepsilon_r = 15, a = b = 39.4 \text{ mm}, h = 33.4 \text{ mm}, w = 9.4 \text{ mm},$ $d = 14.4 \text{ mm}, r_1 = 0.63 \text{ mm}, l = 12.4 \text{ mm}, g = 12.7 \text{ mm}$

Simulated and measured results

Reflection coefficient

Axial ratio



Impedance bandwidth: Simulated: 20.3% (2.34-2.87 GHz) Measured: 24.4% (2.30-2.94 GHz) AR bandwidth: Simulated: 8.2% (2.34-2.54 GHz) Measured: 7.3% (2.39-2.57 GHz)

Simulated and measured radiation patterns



- Very good omnidirectional characteristic
- In the horizontal plane, LHCP fields > RHCP fields by ~20 dB.

Simulated and measured antenna gain





Wideband omnidirectional CP antenna with parasitic metallic strips

Antenna configurations

Hollow cylinder Parasitic strip

Perspective view

Front view

- Four parasitic metallic strips are embedded in the lateral slots to enhance the AR bandwidth.
- The hollow circular cylinder is introduced to enhance the impedance bandwidth.

Photographs of the prototype

Prototype for 3.4 GHz WiMAX design



Top face and sidewalls

Bottom face

Design parameters

 $\varepsilon_r = 15, a = b = 30 \text{ mm}, h = 25 \text{ mm}, r = 3 \text{ mm}, w = 7 \text{ mm}, d = 10.5 \text{ mm}$ $l_s = 30.5 \text{ mm}, w_s = 1 \text{ mm}, x_0 = 6.4 \text{ mm}, r_1 = 0.63 \text{ mm}, l = 19 \text{ mm}.$

Simulated and measured reflection coefficient and axial ratio



Impedance bandwidth: Simulated: 22.3% (3.11-3.89 GHz) Measured: 24.5% (3.08-3.94 GHz)

AR bandwidth: Simulated: 24.8% (3.11-3.99 GHz) Measured: 25.4% (3.16-4.08 GHz)

Overlapping bandwidth: 22.0%; bandwidth widened by ~3 times.

Simulated and measured results

Antenna gain

Radiation efficiency



> Measured gain: wider bandwidth.

➤ Measured antenna efficiency: 84-98% (3.1-3.9 GHz).

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Simulated and measured radiation patterns



LHCP fields > RHCP fields by more than 15 dB in horizontal plane.
Stable radiation patterns across the entire passband (3.1 – 3.9 GHz).

V. Dualband & Wideband DRAs

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(i) Rectangular DRA

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Background

- Dualband and wideband antennas are extensively used (e.g., WLAN)
- Multi-element DRA [1]
 - requiring more DR elements and space
- Hybrid slot-DRA [2]
 - coupling slot used as the feed and antenna
 - inflexible in matching the impedance
- [1] Petosa, N. Simons, R. Siushansian, A. Ittipiboon and C. Michel, "Design and analysis of multisegmentdielectric resonator antennas," *IEEE Trans. AP*, vol.48, pp.738-742, 2000.
- [2] Buerkle, K. Sarabandi, and H. Mosallaei, "Compact slot and dielectric resonator antenna with dual-resonance, broadband characteristics," *IEEE Trans. AP*, vol. 53, pp.1020-1027, 1983.

Use of higher-order DRA

- Wideband DRA [1]
- Dualband DRA [2]
- Trial-and-error approach is normally used
- Systematic design approach is desirable

[1] B. Li and K. W. Leung, "Strip-fed rectangular dielectric resonator antennas with/without a parasitic patch," *IEEE Trans. Antennas Propagat.*, vol.53, pp.2200-2207, Jul.2005.

[2] T. H. Chang and J. F. Kiang, "Dual-band split dielectric resonator antenna," *IEEE Trans. Antennas Propagat.*, vol.55, no.11, pp.3155-3162, Nov.2007.

Design Formulas for Dual-Mode rectangular DRA





- The E-field should vanish on the PEC and the TE_{112} mode cannot be excited properly.
- The TE_{111} mode and TE_{113} mode are used in the dualmode design.

Formula Derivation



The wavenumbers $k_{x1, x2}$ and $k_{z1, z2}$ can be written as follows:

$$k_{z2} = \frac{3\pi}{2d}$$
$$k_{z1} = \frac{\pi}{2d}$$
$$k_{x1} = k_{x2} = \frac{\pi}{a}$$

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From the DWM model, the frequencies f_1, f_2 are given by:

$$f_{1,2} = \frac{c}{2\pi\sqrt{\varepsilon_r}}\sqrt{k_{x1,x2}^2 + k_{y1,y2}^2 + k_{z1,z2}^2}$$

where

$$k_{y1,y2} = \sqrt{k_{1,2}^2 - k_{x1,x2}^2 - k_{z1,z2}^2}$$
 (*)

in which $k_{1,2} = 2\pi \sqrt{\varepsilon_r} f_{1,2} / c$ are wavenmubers in the dielectric, with *c* being the speed of light in vacuum.

Engineering Formulas for the DRA dimensions

$$a = \frac{10.32}{\sqrt{9k_1^2 - k_2^2}} + 10.32^{-(3.96 - \frac{f_2}{f_1})}$$

$$d = \pi \sqrt{\frac{2}{k_2^2 - k_1^2}} + \Delta d$$

$$b = 0.65b_1 + 0.35b_2$$

where

$$\Delta d = \left[0.1393 \left(\frac{f_2}{f_1} \right)^4 - 2.3209 \left(\frac{f_2}{f_1} \right)^3 + 11.4422 \left(\frac{f_2}{f_1} \right)^2 - 23.4984 \left(\frac{f_2}{f_1} \right) + 18.4437 \right] \times 10^{-3} \quad (m)$$
$$b_{1,2} = \frac{2}{k_{y1,y2}} \tan^{-1} \sqrt{\left(1 - \frac{1}{\varepsilon_r} \right) \left(\frac{k_{1,2}}{k_{y1,y2}} \right)^2 - 1}$$

Limit of frequency ratio f_2/f_1

From
$$a = \frac{10.32}{\sqrt{9k_1^2 - k_2^2}} + 10.32^{-(3.96 - \frac{f_2}{f_1})}$$

We have

$$9k_1^2 - k_2^2 d \ge 0 \quad \Rightarrow \quad 3k_1 > k_2 \text{ or } 3f_1 > f_2$$

giving
$$f_2/f_1 < 3$$

which is the theoretical limit that is not known before.

Error analysis



Compared with DWM results, errors of f_1 , f_2 are both less than 2.5% for $1 < f_2/f_1 \le 2.8$, $5 \le \varepsilon_r \le 70$.



Given: $f_1 = 3.47$ GHz (WiMax) $f_2 = 5.2$ GHz (WLAN), $\varepsilon_r = 10$

Using dual-mode formulas

a = 20.8 mm, *b* = 10.5 mm, and *d* = 18.5 mm.

Configuration of the dualband DRA



 $W = 2.6 \text{ mm}, L = 10.6 \text{ mm}, Ls = 7.2 \text{ mm}, W_{\text{f}} = 1.94 \text{ mm}, h = 0.762 \text{ mm}, \varepsilon_{\text{rs}} = 2.93$

Measured and simulated reflection coefficients



Measured bandwidths:

Lower band: 15% (3.25-3.78 GHz) covering WiMAX (3.4-3.7 GHz). Upper band: 8.3% (5.03-5.47 GHz) covering WLAN (5.15-5.35 GHZ).

COMPARISON OF DESIGN, SIMULATED, AND MEASURED RESONANCE FREQUENCIES OF TE₁₁₁^y AND TE₁₁₃^y MODES

Resonant Mea Mode frequ (G	Measured frequency	Design frequency		Simulated HFSS frequency	
	(GHz)	<i>f</i> _{1,2} (GHz)	Error (%)	f _{HFSS} (GHz)	Error (%)
TE ₁₁₁ ^y	3.40	3.47	2.05	3.47	2.05
TE ₁₁₃ y	5.18	5.30	2.32	5.24	1.15

Measured and simulated radiation patterns



- TE_{111}^{y} mode: measured (3.40 GHz), simulated (3.47 GHz).
- Broadside radiation patterns are observed for both planes.
- Co-polarized fields > cross-polarized fields by more than 20 dB in the boresight direction. 71

Measured and simulated radiation patterns



- TE_{113}^{y} mode: measured (5.18 GHz), simulated (5.24 GHz).
- Broadside radiation patterns are observed for both planes.
- Co-polarized fields > cross-polarized fields by more than 20 dB in the boresight direction. 72
Measured antenna gain



TE₁₁₁^y mode: Maximum gain of 4.02 dBi at 3.48 GHz.
TE₁₁₃^y mode: Maximum gain of 7.52 dBi at 5.13 GHz.
Electrically larger antenna has a higher antenna gain.

B. Example for Wideband DRA Design

Given: $f_1 = 1.98$ GHz (PCS) $f_2 = 2.48$ GHz (WLAN), $\varepsilon_r = 10$

> Using formulas for dual-mode rectangular DRA

a = 30.7 mm, b = 24.7 mm, and d = 47.7 mm.

Configuration of the wideband DRA



l = 17 mm, W = 1 mm

Measured and simulated reflection coefficients



Measured bandwidths : 30.9% (1.83-2.50 GHz) PCS (1.85-1.99 GHz), UMTS (1.99-2.20 GHz) & WLAN (2.4-2.48 GHz)

Measured and simulated radiation patterns



Measured (2.16 GHz), simulated (2.11 GHz).
Broadside radiation patterns are observed.
Co-polarized fields > cross-polarized fields by more than 20 dB in the boresight direction.

Measured and simulated radiation patterns



- Measured (2.41 GHz), simulated (2.46 GHz).
- Broadside radiation patterns are observed.

• Co-polarized fields > cross-polarized fields by more than 20 dB in the boresight direction.

Measured antenna gain



The maximum gain of 6.98 dBi at 2.47GHz.
TE₁₁₃^y -mode gain > TE₁₁₁^y -mode gain.

(ii) Cylindrical DRA

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Resonance frequency of the HEM_{mnr} mode of the cylindrical DRA



$$k_{\rho i}^{2} + k_{z i}^{2} = \varepsilon_{r} k_{0 i}^{2}$$
 (1)
 $i = 1, 2 \text{ for } f_{1}, f_{2}$

 f_1 : HEM₁₁₁ mode frequency f_2 : HEM₁₁₃ mode frequency

• $k_{\rho i} \& k_{z i}$: dielectric wavenumbers along the $\rho \& z$ directions

•
$$k_{0i} = 2\pi f_i/c$$
 : wavenumber in air

Resonance frequency of the HEM_{mnr} mode of the cylindrical DRA



For
$$k_{\rho}$$
:

$$\left(\frac{1}{k_{\beta}}\frac{J_{m}'(k_{\beta}a)}{J_{m}(k_{\beta}a)} + \frac{1}{k_{\beta}'}\frac{K_{m}'(k_{\beta}'a)}{K_{m}(k_{\beta}'a)}\right) \cdot \left(\frac{\varepsilon_{r}}{k_{\beta}}\frac{J_{m}'(k_{\beta}a)}{J_{m}(k_{\beta}a)} + \frac{1}{k_{\beta}'}\frac{K_{m}'(k_{\beta}'a)}{K_{m}(k_{\beta}'a)}\right) \\
= \frac{m^{2}(k_{\beta}^{2} + k_{\beta}'^{2})(k_{\beta}^{2} + \varepsilon_{r}k_{\beta}'^{2})}{(k_{\beta}k_{\beta}')^{4}a^{2}} \tag{2}$$

where

$$k_{\rho i}' = \sqrt{(\varepsilon_r - 1)k_{0i}^2 - k_{\rho i}^2}$$
(3)

is the radial wavenumber outside the dielectric rod

 $J_m(x)$: Bessel function of the first kind $K_m(x)$: modified Bessel function of the second kind.

Resonance frequency of cylindrical DRA



For k_z : approximated by the TM₀₁-mode wavenumber

$$\frac{hk_{zi}}{p_i} = \tan^{-1} \left(\frac{\varepsilon_r \sqrt{(\varepsilon_r - 1)k_{0i}^2 - k_{zi}^2}}{k_{zi}} \right)$$

$$(i = 1, 2 \text{ for } f_1, f_2)$$
 (4)

where $p_1 = 1$ and $p_2 = 3$ correspond to the HEM₁₁₁ and HEM₁₁₃ modes, respectively.

R. K. Mongia and P. Bhartia, "Dielectric resonator antennas- a review and general design relations for resonant frequency bandwidth," *International Journal of Microwave and Millimeter-Wave Computer-Aided Engineering*, vol. 4, no. 3, pp 230-247, 1994.

Design formula of ratio h/a for given f_1, f_2 , and ε_r

 f_1 : HEM₁₁₁ mode frequency (lower band) f_2 : HEM₁₁₃ mode frequency (upper band)

Using the covariance matrix adaptation evolutionary strategy again,



$$\frac{h}{a} = \frac{E_s}{\varepsilon_r^4} + \sum_{i=1}^4 \frac{1}{\varepsilon_r^{4-i}} \left(\frac{A_i}{\frac{B_i f_2}{f_1}} + D_i \right)$$
(1)

$$\begin{bmatrix} A_1 & B_1 & C_1 & D_1 & E_s \\ A_2 & B_2 & C_2 & D_2 & 0 \\ A_3 & B_3 & C_3 & D_3 & 0 \\ A_4 & B_4 & C_4 & D_4 & 0 \end{bmatrix} = \begin{bmatrix} 489.7 & 0.234 & -0.937 & -34800 & 116500 \\ 680.3 & -625.2 & -4.402 & 3682.7 & 0 \\ 36.15 & 1.511 & -4.713 & -160.2 & 0 \\ 19.23 & 1.162 & 3.982 & 1.996 & 0 \end{bmatrix}$$

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Design formula of radius *a*

Radius *a* can be found by inserting h/a into (2) below:

$$a = \frac{c}{2\pi\sqrt{\varepsilon_{r}}f_{1}} \left[\frac{E_{S}}{\varepsilon_{r}^{4}} + \sum_{i=1}^{4} \frac{1}{\varepsilon_{r}^{4-i}} \left(\frac{A_{i}}{\frac{B_{i}h}{e^{a}} + C_{i}} + D_{i} \right) \right]$$
(2)

$\int A_1$	B_1	C_1	D_1	E_s]=	[1.109	-1.751	0.00152	3107.8	-10932
A_2	B_2	C_2	D_2	0		-0.0571	-0.005	-0.9973	- 304.1	0
A_3	B_3	C_3	D_3	0		0.152	0.0368	-0.9764	17.814	0
$\lfloor A_4$	B_4	C_4	D_4	0_		4.429	5.659	6.114	0.057	0

After *a* is found, *h* can be determined from h/a.

Maximum error of *a*: 2.1% for $1 \le h/a \le 3.5$, $9 \le \varepsilon_r \le 27$ Maximum error of *h*: 3.0% for $1.28 \le h/a \le 1.85$, $9 \le \varepsilon_r \le 27$ A. Example for dualband cylindrical DRA design

Given: $f_1 = 1.71$ GHz (DCS:1.71-1.88 GHz) $f_2 = 2.4$ GHz (WLAN:2.4 - 2.48 GHz), $\varepsilon_r = 9.4$

- Using formulas (1) & (2)

a = 17.9 mm & *h* = 42.5 mm

Configuration of the dualband LP DRA



 $a = 18.7 \text{ mm}, h = 42.5 \text{ mm}, \epsilon_r = 9.4, l = 12.5 \text{ mm}, w = 1 \text{ mm}, Ls = 20 \text{ mm}, Ws = 1.5 \text{ mm}, and Ds = 12.75 \text{ mm}.$

• Radius *a* has been slightly increased to reduce the merging effect

Measured and Simulated Reflection coefficients



- •Reasonable agreement
- Lower band impedance bandwidth: 15.5% (1.70-2.00 GHz)
 Upper band impedance bandwidth: 3.7% (2.39-2.48 GHz)

Measured and simulated radiation patterns



HEM₁₁₁ mode: measured (1.8 GHz), simulated (1.8 GHz) HEM₁₁₃ mode: measured (2.42 GHz), simulated (2.45 GHz)

Broadside radiation patterns are observed.
 Co-polarized fields > cross-polarized fields by more than 20 dB in the boresight direction.

Measured and simulated gain



HEM₁₁₁ mode: Maximum measured gain of ~6 dBi (1.75 GHz)
HEM₁₁₃ mode: Maximum measured gain of ~ 8 dBi (2.43 GHz)

Dualband CP DRA



 $a = 18.7 \text{ mm}, h = 42.5 \text{ mm}, \varepsilon_r = 9.4, l = 12.5 \text{ mm}, w = 1 \text{ mm}, Ls = 21 \text{ mm}, Ws = 1.5 \text{ mm}, Ds = 12.75 \text{ mm}, L_1 = 26.9 \text{ mm}, L_2 = 26.5 \text{ mm}, L_3 = 56.65 \text{ mm}, W_0 = 4.66 \text{ mm}, W_1 = 7.3 \text{ mm}, W_2 = 4.44 \text{ mm}, \text{ and } W_3 = 0.46 \text{ mm}.$

Measured and simulated reflection coefficients



Reasonable agreement Lower band bandwidth:18.9% (1.58-1.91 GHz). Upper band bandwidth:7.8% (2.33-2.52 GHz).

Measured and simulated axial ratios (ARs)



•Reasonable agreement

- •Lower band AR bandwidth: 12.4% (1.67-1.89 GHz)
- •Upper band AR bandwidth: 7.4% (2.34-2.52GHz)

Measured and simulated radiation patterns



HEM₁₁₁ mode: measured (1.8 GHz), simulated (1.8 GHz) HEM₁₁₃ mode: measured (2.42 GHz), simulated (2.45 GHz)

Broadside radiation patterns are observed.
LHCP fields > RHCP fields by ~20 dB in the boresight direction.



Given: $f_1 = 2.90$ GHz, $f_2 = 3.72$ GHz, $\varepsilon_r = 9.4$

- Using formula (5) & (6)

a = 10.3 mm & h = 34.3 mm

Wideband LP cylindrical DRA

Configuration

Reflection coefficient



 $a = 10.3 \text{ mm}, h = 34.3 \text{ mm}, \varepsilon_r = 9.4,$ l = 12 mm, and w = 1 mm.

Good agreement Measured impedance bandwidth: 23.5% (3-3.8 GHz)

Measured and simulated gain



HEM₁₁₁ mode: Maximum measured gain of ~7 dBi (3.29 GHz)
HEM₁₁₃ mode: Maximum measured gain of ~10 dBi (3.83 GHz)

Wideband CP cylindrical DRA



Top view

Side view

 $a = 10.3 \text{ mm}, h = 34.3 \text{ mm}, \varepsilon_r = 9.4, l = 11.5 \text{ mm}, w = 1 \text{ mm}, L_1 = 14.67 \text{ mm}, W_0 = 1.94 \text{ mm}, \text{ and } W_1 = 3.21 \text{ mm}.$

Wideband CP DRA

Reflection coefficient Axial ratio Reflection Coefficient |S11| (dB) Axial ratio (dB) 8 0 6 -10 **HFSS Simulation** Measurement -20 **HFSS Simulation** Measurement -30 n 3.2 3.4 36 3.8 3 4 3.4 3.6 Frequency (GHz) 3 3.2 3.8 4 Frequency (GHz)

Measured impedance bandwidth: 25.5% (3.04-3.93 GHz).

Measured 3-dB AR bandwidth : 24.7% (3.05-3.91 GHz).

VI. Dualfunction DRAs

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Advantage

System size and cost can be reduced by using dualfunction DRAs.

Additional functions

- Packaging cover
- Oscillator



Packaging Cover



Antenna Configuration





Resonant frequency $f_0 = 2.4 \text{GHz}$

Parameters: • Hollow DRA: L=30mm, W=29mm, H=15mm, & $\epsilon_r = 12$

• Metallic Cavity: a = 15mm, b = 21.6mm, h = 5mm Top face : Duroid $\varepsilon_r = 2.94$ thickness 0.762mm Aperture: 0.2063 λ_e

Design Procedure (Simulation):

Step 1 Use the DWM to design a solid rectangular DRA at 2.4-GHz fundamental TE111 Mode.

Z

X

Step 2 Remove the lower center portion concentrically to form a notched DRA. As a result, the resonant frequency >2.4GHz

Step 3

Cover the two sides with the same material. Move the frequency back to 2.4GHz by increasing the thickness. (thickness $\uparrow \rightarrow f_0 \downarrow$)



- Hard-clad foam ($\epsilon_r \approx 1$) is used to form the container.
- ECCOSTOCK HiK Powder of ε_r =12 is used as the dielectric material.

Return Loss and Input Impedance (Passive hollow RDRA with a metallic cavity)



- •Good agreement.
- •Bandwidth $\sim 5.6\%$.

• Measured resonance frequency: 2.42GHz (error < 0.83‰)

Radiation Patterns (Passive hollow DRA with a metallic cavity)



- Broadside TE_{111}^{y} mode is observed.
- Co-polarized fields generally stronger than the crosspolarized fields by 20dB in the boresight direction. 108
Return Loss of the Active Integrated Antenna

- Integrated with Agilent AG302-86 low noise amplifier (LNA) (gain of 13.6dB at 2.4GHz)
- LNA prematched to 50Ω at the input.
- A small hole is drilled on the ground plane to supply the DC bias to the LNA.



Amplified Radiation Pattern



- Compared to the passive DRA, the active DRA has a gain of 7 - 12dB across the observation angle from -90° to 90°.
- The gain is less than the specification due to unavoidable impedance variations and imperfections in the measurement.

Distantia Deservator Automo

Dielectric Resonator Antenna Oscillator (DRAO)

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HEAR HOLENE



- The DRA is used as the oscillator load, named as DRAO.
- The reflection amplifier method is used to design the antenna oscillator.

DRAO Schematic Diagram



- Oscillate condition: $X_L + X_{in} = 0 \& R_L < |R_{in}|$
- DRA first replaced by a 50Ω load at 1.85GHz.

Antenna Configuration:



RF components not shown) Aperture X L M L M L M L S

Top view

Resonance frequency $f_0 = 1.85$ GHz at TE_{111}^y

Parameters: DRA L=52.2mm, W=42.4mm, H=26.1mm, $\epsilon_{r} = 6$.

Aperture $L_a = 0.3561\lambda_e, W_a = 2mm$ $L_s = 9.5 mm, L_m = 40 mm.$

Duroid substrate Ers=2.94, *d*=0.762mm

Return Loss and Input Impedance



- Good agreement.
- Bandwidth ~ 22.14%.
- Resonance frequency: Measured 1.86GHz

Simulated 1.83GHz (1.5% error).

Spectrum of the Free-running DRAO



- Transmitting power $P_t = 16.4$ dBm
- DC-RF efficiency: $\sim 13\%$ (2-25% in the literature).
- Phase noise: 103dBc/Hz at 5MHz offset
- Second harmonic < fundamental by 22dB

Radiation Pattern



- Broadside TE_{111}^{y} is observed.
- Co-polarized fields are generally 20dB stronger than the cross-polarized fields in the boresight direction.



DRA can be of any shape. Can it be made like a swan?

Yes!

DRA is simple made of dielectric. Can glass be used for the dielectric?

Yes!

It leads to probably the most beautiful antenna in the world

Glass-Swan DRA



Distinguished Lecture Transparent antennas: From 2D to 3D

Conclusion

- The DRA can be easily excited with various excitation schemes.
- Frequency tuning of the DRA can be achieved by using a loading-disk or parasitic slot.
- The dualband and wideband DRAs can be easily designed using higher-order modes.
- Compact omnidirectional CP DRAs have been presented
- Dualfuncton DRAs for packaging and oscillator designs have been demonstrated.



Thank you !



