Tunnel diode integrated two-layer microstrip patch antenna

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Theoretical investigations were carried out on the GaAs and Ge tunnel-diode-loaded patch with one parasitic element. It has been found that the radiated power increases due to mutual coupling between the patches typically by 1.95 dB in case of GaAs tunnel-diode-loaded patch and 3.95 dB in case of Ge tunnel-diode-loaded patch. It is also observed that the antenna can be operated over a range of frequency from 0.9975 GHz to 1.0002 GHz for GaAs tunnel-diode-loaded patch and from 56.694 GHz to 64.729 GHz for Ge tunnel-diode-loaded patch just by varying the bias voltage.

1 Introduction

One of the major drawbacks of microstrip antenna is its low efficiency, i.e., only a small portion of the input power is radiated and the rest resonates within the boundary of the antenna. The technique extensively used to increase the power and bandwidth is by stacking. By placing a parasitic (unexcited) patch over an excited rectangular patch one can obtain electromagnetic coupling between the patches. A fair amount of experimental work has been carried out in this context by different workers. The excited element induces current in the parasitic patch and the radiated fields from the parasitic patch affect the current distribution in the excited patch. The characteristics of an element change when some other element comes to its proximity. These characteristics are known as 'in place' characteristics of the element. In this endeavor the lower patch is excited through a tunnel diode. By integrating a tunnel diode with the patch the antenna can be operated over a range of frequencies only by changing the bias voltage. The effect of electromagnetically coupled patches is theoretically investigated for GaAs tunnel diode at low frequencies (0.9975-1.0002 GHz) and Ge tunnel-diode-loaded patch at high frequencies (56.694-64.729 GHz).

2 Theoretical analysis

A large part of the voltage-current characteristic curve of the tunnel diode shows negative resistance, i.e., current decreases with increasing bias voltage. In the negative resistance region of the curve, power is transferred from supply to the external circuit which makes tunnel diode useful for oscillator circuits. Equivalent circuit of tunnel diode shows the resistance $R_s$ and inductance $L_s$ in series, and negative resistance $-R_0$ and junction capacitance $C_D$ in parallel.

The operating frequency range of a tunnel diode oscillator is given by the self resonance frequency $f_s$ and resistive cut-off frequency $f_r$ as follows:

$$f_s = \frac{1}{2\pi C_D R_0} \sqrt{\frac{C_D R_0^2 - L_s}{L_s}}$$

and

$$f_r = \frac{1}{2\pi C_D R_0} \sqrt{\frac{R_0 - R_s}{R_s}}$$

Resistive cut-off frequency, $f_r$, is the frequency at which the negative resistance of the circuit goes to zero and oscillation ceases. Therefore, the operating frequency should always be below $f_r$. In a tunnel diode integrated microstrip patch, the operating frequency varies with the applied bias voltage.

When two or more patches are mutually coupled, the driving point impedance of each antenna depends on its input impedance and, in addition, upon the mutual impedance between itself and the other antenna. Let $A_1$ be the excited patch and $B_1$ the parasitic patch as shown in Fig. 1. The whole system can be represented by a general four terminal network as shown in Fig. 2, where $Z_1$ is the input impedance of the excited patch and $Z_2$ is that of the parasitic patch. The parameter $Z_{11}$ is the impedance measured at terminal 1 with 2 open and $Z_{22}$ is the impedance measured at terminal 2 with 1 open. The parameter $Z_{12}$ is the mutual impedance between the two patches and is defined as the ratio of the voltage induced at the terminals of one antenna (by the current in the other) to the input current in the second antenna. Thus,
respectively. Applying field equivalence principle\(^{10}\) on cavity with magnetic sources only, the integration is carried out along the edge of one patch and, therefore,

\[
Z_{12} = \frac{h}{I_1 I_2} \int \frac{H^y J^x}{I_1 I_2} dl
\]

or,

\[
Z_{12} = h \int_{y=0}^{L_y} \left\{ (H_{YAB} + H_{YBC} + H_{YCD} + H_{YAD})_{z=W} \right\} J^y dx
\]

\[
+ (H_{YAB} + H_{YBC} + H_{YCD} + H_{YAD})_{z=2W} \} J^y dy
\]

where,

\[H_{YAB}, H_{XAB} = y \text{ and } x \text{ component of magnetic field due to non-radiating edge AB of patch } A_1\]

\[H_{YCD}, H_{XCD} = y \text{ and } x \text{ component of magnetic field due to non-radiating edge CD of patch } A_1\]

\[H_{YAD}, H_{XAD} = y \text{ and } x \text{ component of magnetic field due to radiating edge AD of patch } A_1\]

\[H_{YBC}, H_{XBC} = y \text{ and } x \text{ component of magnetic field due to radiating edge BC of patch } A_1\]

and are given as\(^5\),

\[
H_{YAB} = \frac{I_0}{j 4 \pi \mu_0} \left\{ \left( L_y - y \right) \left( j k R + 1 \right) e^{-j R} \right\} \left[ \frac{k \sin(kL)}{r} - \frac{y \cos(kL)}{r^2} \left( j k r + 1 \right) e^{-j k r} \right]
\]

\[\text{... (4)}\]

\[
H_{XAB} = \frac{-I_0}{j 4 \pi \mu_0} \left\{ \left( L_y - y \right) \left( j k R + 1 \right) \frac{1}{R} e^{-j k r} \right\}
\]

\[\text{... (5)}\]

\[
H_{YCD} = \frac{-I_0}{j 4 \pi \mu_0} \left\{ \left( L_y - y \right) \left( j k R' + 1 \right) e^{-j R'} \right\} \left[ \frac{k \sin(kL)}{r} - \frac{y \cos(kL)}{r^2} \left( j k r' + 1 \right) e^{-j k r'} \right]
\]

\[\text{... (6)}\]
The magnetic current on the edges of the patch are given by

\[ J_m^x = 2j \frac{W}{\cos(k(y-L))} \]

\[ J_m^y = -2j \frac{W}{\sin(kL)} \cos(k(x-L)) \]

where,

\[ l_0 = 2j \frac{w}{\cos(kL)} \]

\[ R = \sqrt{(y-L)^2 + x^2} \]

\[ R' = \sqrt{(y-L)^2 + (x-W)^2} \]

\[ r = \sqrt{y^2 + x^2} \]

\[ r' = \sqrt{y^2 + (x-W)^2} \]

\[ S = (x-h')^2 + y^2 \]

\[ S' = (x-h')^2 + (y-L)^2 \]

\[ \omega = 2\pi f \]

\[ k = \frac{2\pi}{\lambda_c} = k_0 \sqrt{\varepsilon_r} \quad \text{Propagation constant} \]

\[ \lambda_c = \frac{c}{f} \]

\[ c = \text{Velocity of light} \]

\[ f = \text{Design frequency of the antenna} \]

\[ \mu = \mu_0 \mu_r \quad \text{Permittivity of the antenna} \]

\[ \varepsilon_d = \varepsilon_0 \varepsilon_r \quad \text{Dielectric constant} \]

\[ L = \text{Length of the patch} \]

\[ W = \text{Width of the patch} \]

\[ w = \text{Width of the feed line} \]

\[ h = \text{Thickness of the substrate} \]

\[ d = \text{Distance between the two patches} \]

\[ V = \text{Radiating edge voltage} \]

3 Effect of mutual coupling on radiation pattern

Let \( E_1 \) and \( E_2 \) be the fields due to elements A and B, respectively, then

\[ E_z = -\frac{Z_{12}}{Z_{22}} E_1 \]

or, \( E_z = \frac{Z_{12}}{Z_{22}} E_1 e^{\alpha} \)

where,

\[ \alpha = \tan^{-1} \left( \frac{Z_{12}}{Z_{22}} \right) \]

Coupling coefficient for \( E \)-plane radiation pattern is as follows:

\[ F(\theta) = 1 - \frac{Z_{12}}{Z_{22}} e^{\alpha} \]

Radiation pattern for a single patch is given by

\[ F_0 = -\frac{Z_{12}}{Z_{22}} \left[ \cos(k_0 h \cos \theta) \right] \]

\[ \times \left[ \sin \left( \frac{k_0 W}{2} \sin \theta \sin \phi \right) \right] \]

\[ \times \left[ \left( \frac{k_0 W}{2} \sin \theta \sin \phi \right) \right] \]

\[ \times \left[ \cos \left( \frac{k_0 L}{2} \sin \theta \cos \phi \right) \right] \cos \phi; \quad (0 \leq \theta \leq \pi / 2) \]

and

\[ F_0 = -\frac{Z_{12}}{Z_{22}} \left[ \cos(k_0 h \cos \theta) \right] \]

\[ \times \left[ \sin \left( \frac{k_0 W}{2} \sin \theta \sin \phi \right) \right] \]

\[ \times \left[ \left( \frac{k_0 W}{2} \sin \theta \sin \phi \right) \right] \]

\[ \times \left[ \cos \left( \frac{k_0 L}{2} \sin \theta \cos \phi \right) \right] \cos \theta \sin \phi; \quad (0 \leq \theta \leq \pi / 2) \]
The $E$-plane radiation pattern is given by

$$F'(\theta) = |F_x(x)F'(\theta)|$$

... (16)

4 Effect of mutual coupling on input impedance

The equivalent impedance of the circuit shown in Fig.2 is given by

$$\frac{1}{Z_{\text{eq}}} = \frac{1}{Z_1} + \frac{1}{Z_2 + Z_{12}}$$

$$Z_{\text{eq}} = \frac{Z_1(Z_2 + Z_{12})}{Z_1 + Z_2 + Z_{12}}$$

... (17)

5 Effect of mutual coupling on VSWR and return loss

Reflection coefficient $\rho$ is calculated as follows:

$$\rho = \frac{|Z_{\text{eq}} - Z_0|}{|Z_{\text{eq}} + Z_0|}$$

where, $Z_0$ = Impedance of the coaxial feed = 50 $\Omega$

and, $\text{VSWR} = \frac{1 + \rho}{1 - \rho}$

... (19)

The return loss ($RL$) is further calculated as:

$$RL = -10 \log \left( \frac{1}{\rho^2} \right)$$

... (20)

6 Design details

The design details are given as follows:

Design Details

Relative permittivity of the substrate ($\varepsilon_r$) = 4.78
Loss tangent (tan$\delta$) = 0.048
Thickness of the dielectric substrate ($h$) = 1.6 mm
Distance between the two patches ($d$) = 1.6 mm

For GaAs tunnel-diode-loaded patch:

Design frequency ($f$) = 1 GHz
Bias voltage range ($V$) = 110-550 mV
Frequency range = 0.9975-1.0002 GHz
Length of the patch ($L$) = 68.441 mm
Width of the patch ($W$) = 88.235 mm
Resistance of the patch ($R$) = 40.028 $\Omega$
Inductance of the patch ($L$) = 63.38 pH
Capacitance of the patch ($C$) = 400.071 pF
Series resistance of the tunnel diode ($R_s$) = 1.5 $\Omega$
Series inductance of the tunnel diode ($L_s$) = 6 nH
Negative resistance ($R_o$) = 152 $\Omega$
Junction capacitance ($C_{j0}$) = 7 pF
Self resonance frequency ($f_s$) = 0.76244 GHz
Resistive cut-off frequency ($f_0$) = 1.49907 GHz

For Ge tunnel-diode-loaded patch:

Design frequency ($f$) = 50 GHz
Bias voltage range ($V$) = 60-350 mV
Frequency range = 56.694-64.429 GHz
Length of the patch ($L$) = 0.5048 mm
Width of the patch ($W$) = 1.765 mm
Resistance of the patch ($R$) = 492 $\Omega$
Inductance of the patch ($L$) = 0.9015 nH
Capacitance of the patch ($C$) = 0.01125 pF
Series resistance of the tunnel diode ($R_s$) = 4.5 $\Omega$
7 Calculations

The values of magnetic fields at the edges of the antenna are calculated using Eqs [(4)-(11)] for a single frequency (1 GHz) for GaAs tunnel-diode-loaded patch, and for a frequency range of 56.694-64.729 GHz for Ge tunnel-diode-loaded patch. This frequency range depends on the bias voltage. In Eqs [(4)-(11)], real and imaginary parts are separated and integration is done using Simpson’s 3/8th rule. The value of $Z_n$ is further calculated using Eq. (3) to find $Z_{in}$ of the stacked antenna. The frequency varies from 64.729 GHz to 56.694 GHz as the bias voltage is varied from 60 mV to 350 mV for Ge-tunnel-diode loaded patch. The equivalent input impedance for

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Series inductance of the tunnel diode ($L_s$)</td>
<td>0.169 nH</td>
</tr>
<tr>
<td>Negative resistance of the tunnel diode ($R_0$)</td>
<td>-106 Ω</td>
</tr>
<tr>
<td>Junction capacitance of the tunnel diode ($C_0$)</td>
<td>0.03550 pF</td>
</tr>
<tr>
<td>Self resonance frequency ($f_s$)</td>
<td>49.3528 GHz</td>
</tr>
<tr>
<td>Resistive cutoff frequency ($f_c$)</td>
<td>200.9708 GHz</td>
</tr>
</tbody>
</table>

Fig. 3 — Variation of $Z_n$ with bias voltage
GaAs and Ge tunnel-diode-loaded patch is calculated for the equivalent circuit using Eq.(17) and is shown in Table 1 as well as in Fig. 3. Comparative data for VSWR and return loss are also calculated by using Eqs (19) and (20) which are shown in Table 1, as well as in Figs 4 and 5. The radiation pattern is calculated using Eqs [(13)-(16)] and is shown in Figs 6 and 7.

8 Discussion of results

Variation of real and imaginary parts of input impedance with bias voltage is shown in Fig. 3 for Ge tunnel-diode-loaded patch with one parasitic element. It is observed that both resistance and reactance decrease with bias voltage for the case of coupling and non-coupling between the patches. However, the value of real and imaginary parts of the impedance is slightly higher with coupling as compared to that without coupling. It is further observed that the coupling between the patches improves between 150 mV and 250 mV (Fig. 4) bias voltage although there

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Without coupling</th>
<th>With coupling</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Z_0$ (ohms)</td>
<td>10.1525+17.8057</td>
<td>11.6924+10.0567</td>
</tr>
<tr>
<td>VSWR</td>
<td>2.44</td>
<td>3.24</td>
</tr>
<tr>
<td>Return loss (dB)</td>
<td>-7.565</td>
<td>-5.539</td>
</tr>
<tr>
<td>3 dB beamwidth (deg.)</td>
<td>88</td>
<td>90</td>
</tr>
</tbody>
</table>

![Fig. 4 — Variation of VSWR with bias voltage](image1)

![Fig. 5 — Variation of return loss with bias voltage](image2)

![Fig. 6 — Radiation pattern of GaAs tunnel-diode-loaded patch without coupling and with coupling](image3)
is not much deviation in the matching condition (VSWR) below 150 mV and above 250 mV between coupling and without coupling condition. The improvement in the VSWR within the bias range 150 mV-250 mV is attributed to the fact that $|Z_m|$ assumes value between 52.48 $\Omega$ and 48.16 $\Omega$ in case of mutual coupling which provides better matching condition with the 50 $\Omega$ feed line. The variation of return loss with bias voltage is shown in Fig. 5. It is observed that minimum of the return loss changes to higher bias voltage for the case of coupling between the patches and the value of return loss remains within acceptable limits. This shows that the antenna can be operated for the entire range of bias voltage without any significant change in the performance of the antenna. In case of GaAs tunnel-diode-loaded patch $Z_m$ decreases resulting in enhanced VSWR due to mutual coupling.

The radiation pattern of GaAs tunnel-diode-loaded patch with parasitic element is shown in Fig. 6. In the case of coupling, the radiated power increased by 1.95 dB as compared to that without coupling. Similarly, the radiated power for Ge tunnel-diode-loaded patch with parasitic element shows enhanced power (5.95 dB) in case of coupling as compared to the case without coupling (Fig. 7). The lower enhancement in the radiated power for GaAs tunnel-diode-loaded patch is due to the fact that the coupling causes increase in the VSWR, whereas in the case of Ge tunnel-diode-loaded patch the coupling improves VSWR, particularly, in the bias range of 150-250 mV which causes considerable enhancement (5.95 dB) in the radiated power. It is further observed that the radiated power depends on biasing voltage and is found to be maximum for the lowest bias voltage both for the case of with and without coupling. There is a slight increase in bandwidth too as shown in Fig. 8.

It may be concluded that the radiation characteristics of the GaAs and Ge tunnel-diode-loaded patch antenna with parasitic element are affected by biasing voltage. The coupling between the elements of the antenna enhances the radiated power, as the loading of the active antenna with parasitic element improves the matching with feed line, and coupling between active and parasitic element makes the overall input impedance of the antenna near to 50 $\Omega$ and hence VSWR around 1.

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References