An improved parameter extraction scheme for dumbbell shaped defected ground structure

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ABSTRACT

The present study proposes an improved scheme for parameter extraction of dumbbell shaped defected ground structure (DGS). A dumbbell DGS has similar kind of response like a single pole *LC*–Butterworth filter. There are *L–C*, *R–L–C*, π and quasi-static equivalent circuits available in literature for dumbbell DGS parameter extraction. In this article, the *R–L–C* equivalent of dumbbell DGS is re-formulated which provides better accuracy in parameter extraction and it is confirmed through simulation and experimental analysis. This improved model is finally used to extract *R–L–C* parameters against several structural variations of dumbbell DGS unit cell.

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1. Introduction

Microstrip transmission line was first developed by ITT laboratories [1] in 1952. Still now, continuous research is going on in the development of efficient and cost effective microstrip applications. Defected ground structure (DGS) [2–14] is a convenient way to remove spurious harmonics from microstrip circuit response and it also reduces the overall circuit size. It has different microwave applications like: filter, coupler, circulator, oscillator and many more. Park et al. [2] first proposed the concept of Defected Ground Structure in 1999. DGS is created by etching periodic structures at the metallic ground plane of microstrip a circuit and it has the same band gap property as found in photonic band gap (PBG) or electromagnetic band gap (EBG) structures. Initially, iteration based optimization was the only way to get desired performance from DGS slots. Dal Ahn et al. [3] in 2001 first proposed the equivalent circuit of a dumbbell shaped DGS which is a single pole *L–C*–Butterworth filter. By comparing the DGS response with the circuit response, effective inductance (*L*) and capacitance (*C*) were extracted. But, soon some more accurate equivalent circuits came into literature [4–10]. Apart from the *L–C* equivalent circuit, there are parallel *R–L–C* [5], π- network equivalent [6] and currently quasi static equivalent [8] circuits exist in literature. Due to increased number of variables present in π network, it is difficult to extract the DGS parameters using this technique and it is more

complex for quasi static equivalent circuit. Still, parallel *R–L–C* model [9,10] or *L–C* model [12,13] is preferred to extract the DGS electrical parameters. The *R–L–C* model is more appropriate than *L–C* model as it takes the radiation effect into consideration where an equivalent radiation resistance is placed parallel between *L* and *C*.

In this present study, a more accurate capacitance (*C*) and inductance (*L*) extraction formula is provided for equivalent *R–L–C* circuit of a dumbbell DGS unit cell. The *L* and *C* extraction formulas are derived from scattering parameter response of the parallel *R–L–C* circuit. A quantified analysis is performed for previous and currently proposed formulas to calculate and compare parameter extraction accuracy. It has been found that the newly derived capacitance (*C*) and inductance (*L*) formula is more accurate in terms of extraction efficiency. The simulations are carried out in HFSS (EM simulation) and MATLAB (circuit simulation) considering FR4 substrate having dielectric constant 4.4, substrate height 1.6 mm and loss tangent 0.004. A dumbbell DGS unit cell is fabricated using FR4 glass epoxy substrate and measured with Agilent N5221A vector network analyzer (VNA). The analysis performed with measured results shows good agreement with simulated results.

2. *R–L–C* parameter extraction method in literature

Figs. 1 and 2 show a dumbbell DGS and its equivalent *R–L–C* circuit. Capacitance (*C*), inductance (*L*) and resistance (*R*) can be extracted from Eqs. (1)–(3) as proposed in Ref. [5].

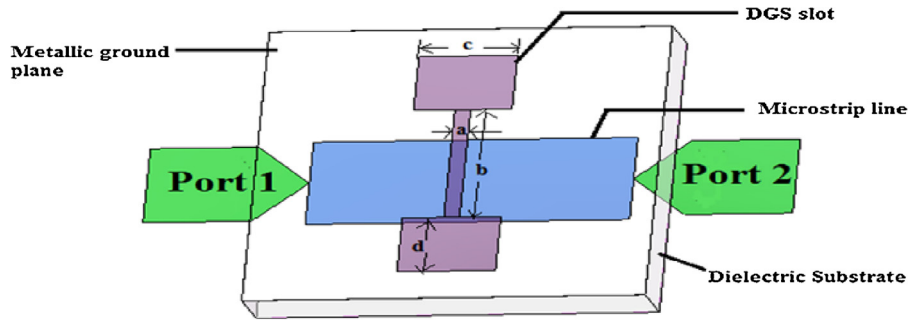$$C = \frac{\omega_c}{2Z_0(\omega_0^2 - \omega_c^2)} \tag{1}$$

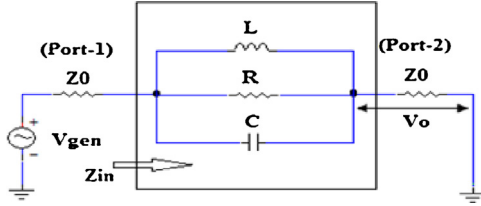**Fig. 1.** Dumbbell DGS cross section.
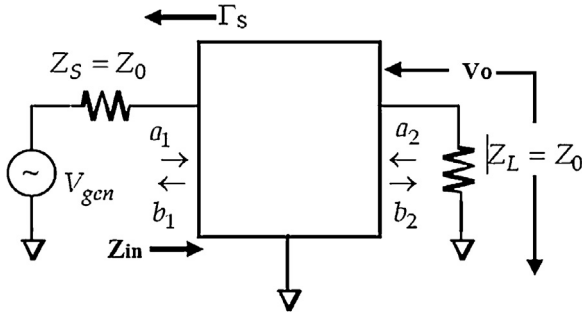

Fig. 2. Equivalent R–L–C circuit.

Fig. 3. Simplified 2 port network.

$$L = \frac{1}{(4\pi^2 f_0^2 C)} \tag{2}$$

$$R(\omega) = \frac{2Z_0}{\sqrt{1/\left\{|S_{11}(\omega)|^2 - \left(2Z_0\left(\omega C - \frac{1}{\omega L}\right)\right)^2 - 1\right\}}} \tag{3}$$

In the above equations Z_0 is port impedance, ω_0 and f_0 is angular and linear resonant frequency, ω_c is angular 3 dB upper/lower cut-off frequency and $S_{11}(\omega)$ is frequency dependent scattering parameter (return loss).

Furthermore, it was proposed in [10] that radiation resistance (R_{11}, R_{21}) can be calculated from both $S_{11}(\omega)$, $S_{21}(\omega)$ at resonant frequency (ω_0) using Eqs. (4) and (5). R_{11} and R_{21} are close but not equal and their average (R, Eq. (6)) represents the best approximation of radiation resistance.

$$R_{11} = \frac{2Z_0|S_{11}(\omega_0)|}{1 - |S_{11}(\omega_0)|} \tag{4}$$

$$R_{21} = 2Z_0 \frac{1 - |S_{21}(\omega_0)|}{|S_{21}(\omega_0)|} \tag{5}$$

$$R = \frac{R_{11} + R_{21}}{2} \tag{6}$$

Table 1
Arbitrary test input values of L, C and R.

Inductance (nH) (L)	Capacitance (pF) (C)	Resistance (Ω) (R)
0.7	0.5	800
0.85	0.64	500
0.75	0.55	1200
0.67	0.43	1500
0.87	0.52	2000

3. Proposed parameter extraction method

From Fig. 3, using transmission line techniques, scattering parameters (S_{21}, S_{11}) can be calculated as:

$$S_{21} = \frac{2V_o}{V_{\text{gen}}} \tag{7}$$

$$S_{11} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} \tag{8}$$

From Fig. 2, input impedance (Z_{in}) and scattering parameters ($S_{21}(\omega)$, $S_{11}(\omega)$) can be calculated as:

$$Z_{in} = Z_0 + \left(j\omega L || R || \frac{1}{j\omega C}\right) \tag{9}$$

$$S_{11}(\omega) = \frac{j\omega RL}{-\omega^2 (2RLCZ_0) + j\omega (2Z_0 L + RL) + 2Z_0 R} \tag{10}$$

$$S_{21}(\omega) = \frac{2 \times \left(-\omega^2 RLCZ_0 + j\omega LZ_0 + RZ_0\right)}{-\omega^2 (2RLCZ_0) + j\omega (2Z_0 L + RL) + 2Z_0 R} \tag{11}$$

Now, $S_{21}(\omega)$ of (11) can be rearranged as:

$$S_{21}(\omega) = \left(\frac{a^2 + b(b+c)}{a^2 + (b+c)^2}\right) + j\left(\frac{ab - a(b+c)}{a^2 + (b+c)^2}\right) \tag{12}$$

where $a = -\omega^2 RLCZ_0 + RZ_0$; $b = \omega LZ_0$; $c = \omega RL/2$.

At resonant frequency (ω_0), the imaginary part of Eq. (12) shall be zero and after solving this, the resonant frequency (ω_0) can be obtained as:

$$\omega_0 = \frac{1}{\sqrt{LC}} \tag{13}$$

$$L = \frac{1}{C\omega_0^2} \tag{14}$$

At 3 dB lower or upper cut-off frequency ($\omega = \omega$), magnitude of $S_{21}(\omega)$ becomes:

$$|S_{21}(\omega_c)| = \frac{1}{\sqrt{2}} \tag{15}$$

$$\sqrt{\left(\frac{a^2 + b(b+c)}{a^2 + (b+c)^2}\right)^2 + \left(\frac{ab - a(b+c)}{a^2 + (b+c)^2}\right)^2} = \frac{1}{\sqrt{2}} \tag{16}$$