Design of a Novel Harmonic-Suppressed Microstrip Low-Pass Filter

Jongkuk Park, *Member, IEEE*, Jeong-Phill Kim, *Member, IEEE*, and Sangwook Nam, *Member, IEEE*

***Abstract*—A novel harmonic-suppressed microstrip low-pass filter (LPF) is proposed which is composed of defected ground structures and stepped-impedance shunt stubs as series L and shunt C components in the pass-band. Due to the attenuation poles of two kinds of resonators, it is found that harmonic responses are not only suppressed effectively, but the rejection in the stop-band is also deep and wide. Besides, the proposed LPF can be compactly implemented since both kinds of resonators have slow-wave characteristics.**

***Index Terms*—Defected ground structure (DGS), harmonic-suppression, low-pass filter (LPF), step-impedance shunt stub (SISS).**

I. Introduction

RECENTLY, a harmonic-suppressed low-pass filter (LPF) has been highly required in many communication systems to reject the spurious responses caused by power amplifiers, mixers, and oscillators. For this purpose, a lumped element such as a chip capacitor [1] or a sheet resistor [2] has been incorporated in the distributed line circuits in order to break their periodicity with respect to frequency. Another approach is to employ a periodic bandgap (PBG) structure [3] or a defected ground structure (DGS) [4]. Especially, since a DGS has a simple equivalent circuit model and yields a low-pass property with a wide stop-band, many research activities have been performed in order to apply it to the LPF design [4]–[7]. However, most of them are not focused on the suppression of harmonics, or their design procedures are so dependent upon full-wave electromagnetic (EM) optimization that it is difficult to apply the conventional LPF design method. In this letter, a novel harmonic-suppressed microstrip LPF and its design procedures are proposed. Conventional dumbbell-shaped DGSs and stepped-impedance shunt stubs (SISSs) are employed as series and shunt elements of the LPF, respectively. It is noted that they have simple equivalent circuits dual to each other and both of them provide attenuation poles within the stop-band of the LPF. By adjusting their resonant frequencies properly without affecting the original low-pass characteristics, the proposed structure is shown to be able to suppress the harmonic responses efficiently and provide the deep and wide stop-band. Due to the slow-wave effects of both kinds of resonators, the proposed LPF has a shorter physical length than conventional ones and it is helpful to the compact design.

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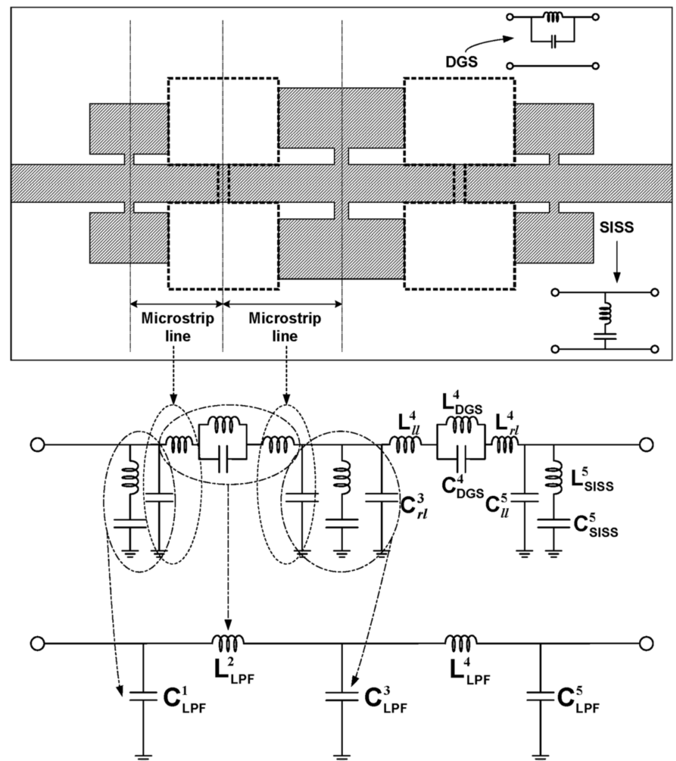


Fig. 1. Equivalent circuits for designing an LPF. (Equivalent circuits of DGSs, SISSs, and short microstrip line sections, and grouping scheme for relating them to the L and C values obtained from a prototype filter.)

II. Design Procedures

As shown in Fig. 1, the equivalent circuits of a unit DGS and a unit SISS are simply represented by a series-connected parallel L–C and a shunt-connected series L–C resonators, respectively [4], [8]. Note that the unit SISS consists of two identical arms by dividing a rectangular patch providing capacitance into two smaller ones. In this figure, the parallel L–C resonator of a unit DGS behaves like a simple series inductance and the series L–C resonator of a SISS looks like a simple shunt capacitance in the low frequency region, and thus they can be used as elements of an LPF. According to the approach proposed by [4], their circuit values are made equal to the corresponding element values obtained by frequency mapping and impedance scaling of prototype values at the cutoff frequency. Then, the required low-pass response can be easily obtained.

In [4], the effect of a microstrip line section between two neighboring elements has been ignored since its length is electrically short enough below the cutoff frequency. However, in our case, such a line section between a SISS and a DGS should be necessarily considered since it is still electrically short but it is found to make a remarkable effect on the pass-band characteristics and the cutoff frequency. For this purpose, the line section is modeled as an L-shaped network which consists of a series L and a shunt C as shown in Fig. 1. As usual, a T- or Π-network model is used for its exact equivalent two-port network. However, the following approximation of ABCD parameters shows that the short line section can be replaced with this L-shaped network

$$\begin{pmatrix} \cos\beta_c l & jZ_0\sin\beta_c l \\ j/Z_0\sin\beta_c l & \cos\beta_c l \end{pmatrix} \approx \begin{pmatrix} 1 & jZ_0\beta_c l \\ j/Z_0\beta_c l & 1 \end{pmatrix} \sim \begin{pmatrix} 1 & j\omega_c L_{ll} \\ j\omega_c C_{rl} & 1-\omega_c^2 L_{ll}C_{rl} \end{pmatrix} \text{ or } \begin{pmatrix} 1-\omega_c^2 L_{rl}C_{ll} & j\omega_c L_{rl} \\ j\omega_c C_{ll} & 1 \end{pmatrix} \quad (1)$$

where Z_0 and l is the characteristic impedance and the length of the line section, respectively. Also, β_c and ω_c are the propagation constant and angular frequency at the cutoff of the given LPF. From the above equation, the equivalent inductance and capacitance are given by $L_{l(r)l} = Z_0 l/v_c$ and $C_{l(r)l} = l/(Z_0 v_c)$ where v_c is the corresponding phase velocity.

Based on this approximation, the same procedures as in [4] are applied by using the line parameters and the equivalent circuits of DGSs and SISSs as shown in Fig. 1. Mathematically, these relations can be represented as follows:

$$\frac{1}{\omega_c\left(L^i_{\mathrm{LPF}} - L^i_{ll} - L^i_{rl}\right)} = -\left(\omega_c C^i_{\mathrm{DGS}} - \frac{1}{\omega_c L^i_{\mathrm{DGS}}}\right) \quad (2)$$

$$\frac{1}{\omega_c\left(C^j_{\mathrm{LPF}} - C^j_{ll} - C^j_{rl}\right)} = -\left(\omega_c L^j_{\mathrm{SISS}} - \frac{1}{\omega_c C^j_{\mathrm{SISS}}}\right) \quad (3)$$

where L^i_{LPF} and C^j_{LPF} in the left sides of the equations represent the ith inductance and the jth capacitance values obtained from a prototype filter, respectively. Also, $L^i_{l(r)l}$ and $C^j_{l(r)l}$ are the inductance and the capacitance for the line sections in the left (right) side of the ith DGS and the jth SISS, respectively. It is noted that C^1_{ll} (or L^1_{ll}) in the first resonator and C^N_{rl} (or L^N_{rl}) in the last resonator should be excluded in the above equation. On the other hand, $L^{i(j)}_{DGS(SISS)}$ and $C^{i(j)}_{DGS(SISS)}$ in the right sides are unknowns which should be determined as the equivalent inductance and capacitance values of the ith DGS (the jth SISS), respectively. Since two unknowns per equation should be determined even if the line parameters are given, there exist a lot of solutions. However, if the resonant frequencies of all the resonators are also given, the unique values can be obtained from (2) and (3). It is worthy to note that these two variables or the line lengths and the resonant frequencies make enormous effects on the rejection characteristics and they can be arbitrarily selected if only the corresponding structures could be physically implemented.

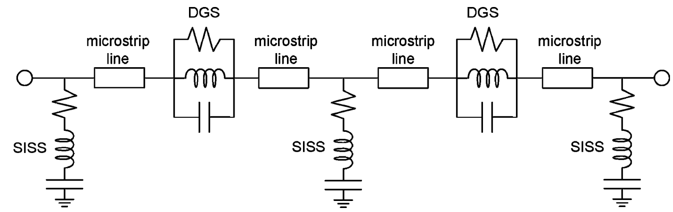


Fig. 2. Modified equivalent circuit for estimating the whole-band characteristics.

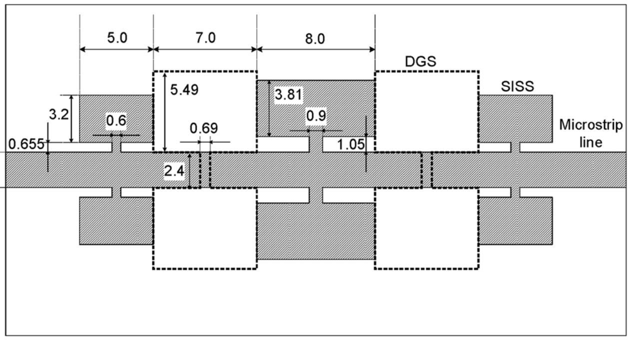
Therefore, the goal is to find out their optimal values which provide the best rejection performance. First, the initial values can be determined by considering the typical size of the resonators and the harmonic frequencies of a conventional LPF. After then, $L^{i(j)}_{DGS(SISS)}$ and $C^{i(j)}_{DGS(SISS)}$ are calculated from the (2) and (3). With these equivalent circuit parameters, the frequency responses are easily obtained from the circuit simulation. However, the equivalent circuit shown in Fig. 1 is valid only near the pass-band region because of the approximation of the line sections.

Hence, it should be modified to such an equivalent circuit as shown in Fig. 2 in order to include the stop-band responses. Instead of lumped elements for line sections, they are replaced with distributed microstrip elements. Also, the loss term of each resonator is included in its equivalent circuit and approximated to the typical value calculated from the S-parameters at its resonant frequency. Once the rejection performance is obtained with the initial parameters, the above procedures are repeated with new parameters given by an arbitrary optimizing technique.

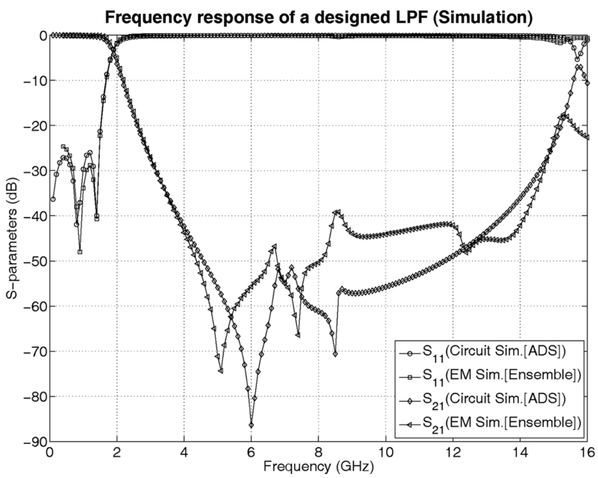
After all these procedures are completed, the remaining one is to determine the physical dimensions of respective resonators providing the optimized circuit values. For a dumbbell-shaped DGS, its initial dimensions can be estimated by means of a quasi-static model [9]. In case of a unit SISS, there exist various analytical models which make it easy to find out the initial values. However, the accurate dimensions should be obtained from the same procedures as in [5] or given by several tuning procedures through EM simulations.

III. DESIGN AND MEASUREMENT RESULTS OF A FIVE-POLE LPF

As an example of the proposed method, a five-pole LPF is designed at the cutoff frequency of 1.5 GHz with 0.01-dB ripple level. The substrate is RT Duorid5880 with the thickness of 31 mil and the dielectric constant of 2.2. Since the LPF is symmetric, the resonant frequencies of three resonators and the lengths of two line sections should be optimized for the best performance according to the above procedures. In this design, the resonant frequencies of a DGS and two SISSs are determined first by considering the locations of conventional harmonics and the skirt characteristics, and they are set to 7.0 GHz, 6.0 GHz, and 8.5 GHz, respectively. On this condition, two line lengths are so optimized as to provide the deep and wide rejection band. However, the fact is found that the line inductance and capacitance given by (1) are somewhat overestimated and the



(a)



(b)

Fig. 3. (a) Configuration of the designed LPF with its dimensions (unit: mm) and (b) the simulated frequency responses.

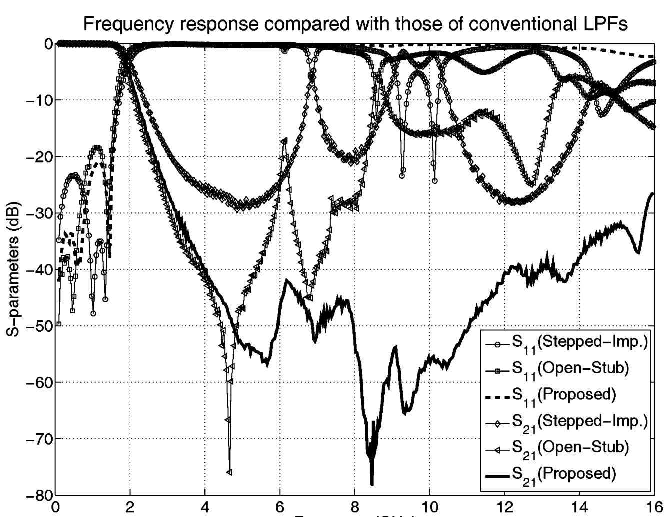


Fig. 4. Measured S-parameters of the designed LPF and conventional ones.

cutoff frequency increases slightly at the circuit simulation in Fig. 2. Hence, they are compensated with a constant, or modified to $L_{l(r)l} = 0.8Z_0 l/v_c$ and $C_{l(r)l} = 0.8l/(Z_0 v_c)$, and this relation is shown to be usually valid in other cases.

Fig. 3(a) depicts the configuration of the designed LPF with its physical dimensions, and Fig. 3(b) represents the simulated frequency responses. Two results are compared between circuit and full-wave EM simulations and they are in good agreement. In order to verify the simulated results, the proposed LPF is fabricated together with conventional five-pole stepped-impedance and open-stub LPFs [8] with the same specification for comparison. As expected, the measured result in Fig. 4 shows that the proposed LPF provides a superior rejection characteristic to those of conventional ones. In the stop-band of 4 to 12 GHz, the transmission is limited to less than about -40 dB. All the spurious responses are suppressed below -30 dB up to more than 15 GHz, which corresponds to about 10× the cutoff frequency. It is obvious that the rate of cutoff is sharper than or similar to those of the others. Also, the insertion losses in the pass-band are measured to less than about 0.2 dB in all cases.

IV. Conclusion

A harmonic-suppressed microstrip LPF with a deep and wide stop-band is proposed and its brief design procedure is described. By using two kinds of resonators with dual equivalent circuits and adjusting their attenuation poles to the proper frequency points, conventional harmonic responses are shown to be effectively suppressed without the change of the original low-pass characteristic. The measured result shows that all the harmonic responses are suppressed below -30 dB up to the point which is about 10× the cutoff frequency. Besides, since it has a sharp rejection slope and a relatively small size, the proposed structure is expected to be useful to design a compact LPF with good rejection property.

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