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Differential Microstrip Lines with Common-Mode Suppression based on Electromagnetic Bandgaps (EBGs)

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***Abstract*— A technique for the suppression of the common-mode in differential (balanced) microstrip lines, based on electromagnetic bandgaps (EBGs), is presented in this paper. It is demonstrated that by periodically modulating the common-mode characteristic impedance of the line and simultaneously forcing the differential-mode impedance to be uniform (and equal to the reference impedance of the differential ports), the common-mode can be efficiently suppressed over a certain frequency band, whilst the line is transparent for the differential-mode. The main advantage of EBGs, as compared to other approaches for common-mode suppression in differential microstrip lines, is the fact that the ground plane is kept unaltered. Moreover, the design of the differential line is straightforward since the required level of common-mode suppression and bandwidth are given by simple approximate analytical expressions. As a design example, we report a 4-stage common-mode suppressed differential line with 68% fractional bandwidth for the common-mode stopband centered at 2.4GHz, and maximum common-mode rejection ratio (CMRR) of 19dB at that frequency. Furthermore, we have designed and fabricated a 6-stage double-tuned common-mode suppressed differential line in order to enhance the stopband bandwidth for the common mode around 2.4GHz.**

***Index Terms*– Electromagnetic bandgaps (EBGs), periodic structures, differential transmission lines, common-mode noise suppression.**

I. Introduction

Differential (balanced) lines are of foremost interest for high-speed interconnects and high-speed digital circuits due to their high immunity to noise, electromagnetic interference and crosstalk. In these lines, common-mode noise rejection in the region of interest for the differential signals is necessary to prevent common-mode noise radiation and electromagnetic interference. Therefore, the design of differential lines able to transmit the differential signals and simultaneously suppress the common-mode over a certain (predefined) frequency band has attracted the attention of microwave engineers in recent years.

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Several approaches for the implementation of balanced lines with common-mode noise suppression have been reported. In [1], dumbbell shaped periodic patterns etched in the ground plane, underneath the differential microstrip lines, were used to suppress the even mode by opening the return current path through the ground plane. In [2], the authors achieved a wide stop-band for the common-mode by using U-shaped and H-shaped coupled resonators symmetrically etched in the ground plane. In [3], the common-mode was suppressed by etching complementary split ring resonators (CSRRs) aligned with the symmetry plane of the line. An efficient approach for the suppression of the common-mode over broad frequency bands was reported in [4],[5], where the pair of coupled microstrip lines was loaded with a periodic distribution of centered conductor patches connected to the ground plane by means of narrow (high impedance) strip lines. The structure (unit cell) is described by a circuit that resembles the canonical model of a quasi-elliptic low pass filter. Finally, other approaches, based on multilayer structures, are reported in [6],[7].

In the previous implementations, either the common-mode suppressed balanced lines are complex (including several metal levels and via holes), or they are etched with slots in the ground plane (i.e., they belong to the category of defected ground structures – DGSs). DGSs prevent from back side isolation and make the fabrication process more complex. As an alternative, we propose in this paper a novel and simple approach for the implementation of common-mode suppressed balanced lines. Only two metal levels are required, and the ground plane is kept unaltered. The common-mode is suppressed by periodically modulating the characteristic impedance for that mode and simultaneously maintaining the differential-mode impedance uniform along the line. Due to the well-known Bragg effect, the differential line acts as a reflector for the common-mode, opening a bandgap in the vicinity of the Bragg frequency for that mode. However, as long as the differential-mode impedance is uniform along the line and equal to the reference impedance of the differential ports, the line is transparent for the differential-mode, as will be shown later.

Periodic structures able to inhibit wave propagation at certain frequencies and/or directions due to periodicity and operative at microwave frequencies were designated as electromagnetic bandgaps (EBGs) in the nineties. In planar technology, EBGs implemented by drilling holes in the ground plane were applied to the design of reflectors and high-Q resonators [8]-[10]. It was also demonstrated that wideband stop band filters are possible by periodically modulating the width of a microstrip line, and these structures were applied to the implementation of microstrip bandpass filters with spurious suppression [11],[12]. By periodically loading a line with capacitive elements (lumped or semi-lumped) a combined Bragg and slow wave effect arise [13], and stop band rejection and miniaturization are simultaneously possible. This combined effect was exploited in [14] for the implementation of compact bandpass filters with spurious suppression. Finally, EBGs have been recently applied to the design of coupled line directional couplers with enhanced coupling factor [15]-[17]. By properly modulating the common-mode and differential mode characteristic impedances it is possible to achieve contra-phase reflection coefficients for the even and odd modes and, consequently, redirect the reflected signal to the coupled port.

To the best of our knowledge, the application of EBG-based structures for the suppression of the common-mode in differential lines, as it is proposed in this paper, has never been explored so far.

II. Principle for Common-Mode Suppression and Design

Single-ended non-uniform transmission lines with periodically modulated characteristic impedance (or transverse dimensions) inhibit wave propagation in the vicinity of the Bragg frequency, given by

$$f_{\max} = \frac{c}{2l\sqrt{\varepsilon_{eff}}} \quad , \tag{1}$$

and, eventually, in the vicinity of its harmonics. In (1), c is the speed of light in vacuum, l is the period of the line, and ε_{eff} is an averaged effective dielectric constant. Indeed, according to the coupled mode theory [18][19], the reflection characteristics of the non-uniform periodic transmission line are dictated by the coupling coefficient between the forward and backward travelling waves associated to the operation mode. This coupling coefficient is also a periodic function, given by:

$$K(z) = \frac{1}{2Z_o(z)} \frac{dZ_o(z)}{dz} \tag{2}$$

where $Z_o(z)$ is the point-to-point characteristic impedance of the line and z indicates the position along the line.

Specifically, the coupled mode theory determines that the rejection bands are given by the harmonic content of the coupling coefficient. Namely, if the coupling coefficient is a sinusoidal function, only a bandgap centered at the Bragg frequency is expected. This is the most convenient periodic function for $K(z)$ in order to achieve maximum attenuation at f_{max} for a given perturbation amplitude. The reason is that the weighting coefficients of the Fourier expansion of $K(z)$ are all null, except the first one (associated to the period of the structure), that is maximized.

Let us denote by K_n the weighting coefficients of the Fourier expansion of $K(z)$, each one giving a rejection band that can be characterized by the order, n-th. The maximum attenuation and bandwidth (delimited by the first reflection zeros around the maximum reflectivity) for each stop band are approximately given by [19]:

$$\left|S_{21}\right|_{\min,n} = \operatorname{sech}\left(\left|K_n\right| \cdot L\right) \tag{3}$$

$$BW_n = \frac{c\left|K_n\right|}{\pi\sqrt{\varepsilon_{eff}}} \sqrt{1 + \left(\frac{\pi}{\left|K_n\right| \cdot L}\right)^2} \tag{4}$$

where L is the length of the structure (i.e., $L = l \cdot m$, m being the number of cells), and the averaged effective dielectric constant can be calculated according to [19]:

$$\varepsilon_{eff} = \left(\frac{1}{l}\int_0^l \sqrt{\varepsilon_{eff}(z)} \cdot dz\right)^2 \tag{5}$$

Notice that in (3) and (4) the dependence of the phase constant and effective dielectric constant with z is neglected since an averaged value of ε_{eff}, given by (5), is considered.

Using expressions (1)-(5), it is possible to design periodic transmission lines with specific value of central stop band frequency (or frequencies, if $K(z)$ has harmonic content), rejection level, and bandwidth [16]. Using the same expressions, it is possible to generate an even-mode characteristic impedance profile useful to obtain a bandgap (or bandgaps) for the common-mode in differential lines. The degrees of freedom in differential microstrip lines are enough to simultaneously achieve line impedance modulation for the common-mode and a uniform characteristic impedance for the differential mode. This is essential to achieve common-mode noise suppression and keep the differential signals unaltered. However, since the differential-mode impedance is always smaller than the common-mode impedance, and the line must be matched to the ports for the differential mode (i.e., the differential mode impedance must be $Z_{oo}(z)$ = 50 Ω), it follows that the common-mode impedance must be periodically modulated, satisfying $Z_{oe}(z) \geq Z_{oo}(z)$ = 50 Ω. If the coupling coefficient is sinusoidal, and the common-mode impedance is forced to be 50 Ω at the extremes of the EBG structure (i.e., $Z_{oe}(0) = Z_{oe}(L)$ = 50 Ω), integration of (2) with these boundary conditions gives:

$$Z_{oe}(z) = Z_{oe}(0) \cdot e^{-\frac{2|K_1| \cdot l}{\pi}\left[\cos\left(\frac{2\pi}{l}z\right) - 1\right]} \tag{6}$$

Notice that expression (6) satisfies $Z_{oe}(z) \geq Z_{oe}(0) = Z_{oe}(L) = Z_{oo}(z)$ = 50 Ω, as required.

III. Results

We have determined the line parameters in order to achieve a maximum rejection level (common-mode) of 19dB in the vicinity of f_{max} = 2.4GHz. This frequency gives a period (using 1) of l = 2.38 cm. Considering 4 cells (m = 4), L = 9.53 cm and, using (3), the weighting factor is found to be K_1 = 0.305 cm^{-1}. Applying (4), the common-mode rejection bandwidth is 1.63GHz (i.e., 68%). The parameters of the *Rogers RO3010* substrate with dielectric constant ε_r = 10.2 and thickness h = 1.27 mm have been considered.

Using (6), the common-mode characteristic impedance varies between 50 Ω and 131.5 Ω, which are implementable values. The transverse geometry has been determined in order to achieve a common-mode impedance given by (6) and, simultaneously, a differential mode impedance of $Z_{oo}(z)$ = 50Ω along the line. To this end, the transmission line calculator *LineCalc*, integrated in *Agilent ADS*, has been used.

Since the common-mode impedance is a continuously varying function, in practice we have calculated the transverse geometry at 40 discrete points along the period, and we have then connected the corresponding extremes of the strips by straight lines, resulting actually in a linear piecewise function.

The photograph of the fabricated line is depicted in Fig. 1. The differential and common-mode insertion and return loss are depicted in Fig. 2. The agreement between the electromagnetic simulation (inferred from *Agilent Momentum*) and measurement (obtained by means of the *Agilent E8364B PNA* vector networks analyzer) is reasonable (discrepancies are attributed to tolerances in the fabrication and dielectric constant). The maximum simulated rejection level for the common-mode, obtained at the design frequency f_{max}, i.e., 19 dB, is in agreement with the design value given above, and the bandwidth (62.5 %) is reasonably predicted by expression (4). The differential insertion loss is better than 0.7dB (with insertion loss equal to 0.59 dB at desired design frequency f_0) in the whole common-mode rejection band. The designed line is roughly transparent to the differential-mode, as desired. Some level of common-mode rejection at the first harmonic can also be appreciated. This can be attributed to the fact that a perfect sinusoidal coupling coefficient is never obtained in practice. Nevertheless, this does not significantly affect the rejection level at the design frequency.

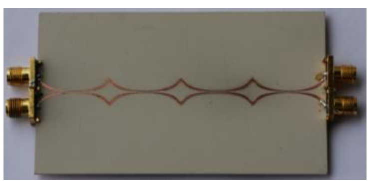

Fig. 1. Photograph of the fabricated common-mode suppressed differential transmission line. Line width W and separation S at the planes of maximum (131.5 Ω) and minimum (50 Ω) value of Z_{oe} are W = 0.16 mm, S = 0.32 mm, and W =1.06 mm, S = 6.4 mm, respectively.

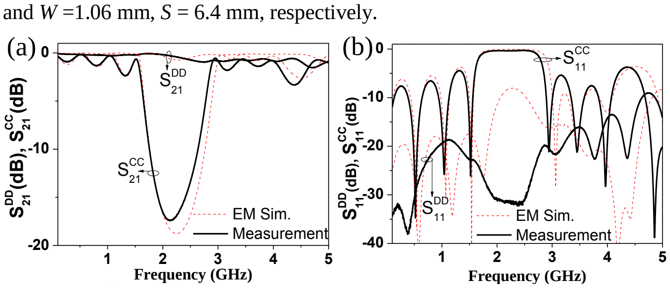

Fig. 2. Differential and common-mode insertion (a) and return (b) loss corresponding to the designed and fabricated 4-cell non-uniform common-mode suppressed differential line of Fig. 1.

IV. DISCUSSION ON BANDWIDTH LIMITATIONS

One clear advantage of the proposed approach for common-mode noise suppression in differential microstrip lines is the easy control of rejection level and bandwidth, given by expressions (3) and (4), respectively. The maximum rejection level only depends on the product $|K_1|\cdot L$, whereas the bandwidth depends on both $|K_1|\cdot L$ and $|K_1|$. According to (4), increasing the maximum rejection level (i.e., $|K_1|\cdot L$) has the effect of reducing the bandwidth. However, once the maximum rejection level is set to a certain value (this is typically a design parameter), the bandwidth can be enhanced by increasing $|K_1|$. This increase in $|K_1|$ is at the expense of reducing L in order to preserve the product $|K_1|\cdot L$. Reduction of L is in favor of miniaturization; however, the increase of $|K_1|$ is limited by the maximum implementable characteristic impedance for the common-mode. Notice that, according to expression (6), the maximum common-mode impedance, given by

$$Z_{oe,\max} = Z_{oe}(0)\cdot e^{\frac{4|K_1|\cdot l}{\pi}}\ , \tag{7}$$

increases exponentially with $|K_1|$. Figure (3) depicts the dependence of bandwidth (BW) and $Z_{oe,\max}$ on $|K_1|$, considering $|K_1|\cdot L$ = 4 and l = 2.38 cm (corresponding to the period of the designed structure). As can be seen in the values of line width W for different values of $|K_1|$ (and hence $Z_{oe,max}$) given in Fig. 3, W = 160 μm (close to the typical critical dimension in PCB technology) when $|K_1|$ is around 0.3 cm^{-1}, corresponding to a bandwidth of roughly 1.5 GHz. Thus, bandwidth cannot be enhanced significantly beyond this value by considering this rejection level and substrate. In order to enhance the bandwidth with such rejection level, it is necessary to consider another substrate with a smaller dielectric constant. Note that this reduces the averaged effective dielectric constant (present in the denominator of 4), contributing to improve the bandwidth, and provides wider line width. Though line separation is reduced for substrates with smaller dielectric constant, this is not the critical parameter.

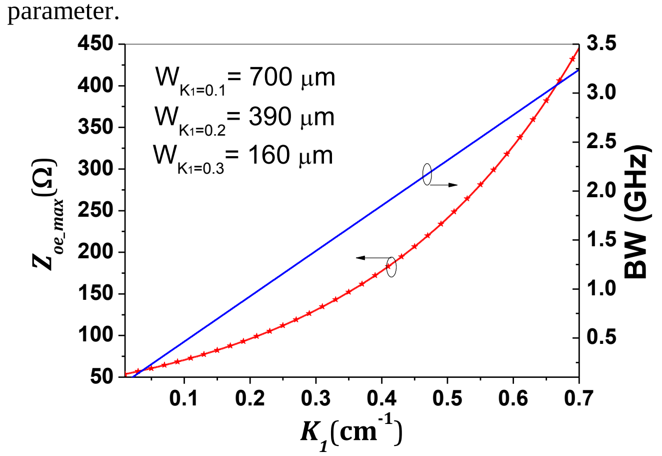

Fig. 3. Dependence of bandwidth (BW) and $Z_{oe,\max}$ on $|K_1|$, considering $|K_1|\cdot L$ = 4 and l = 2.38 cm. W is the line width for $Z_{oe,\max}$ (3 cases are indicated).

An alternative to improve the bandwidth is to cascade more than one EBG structure with different periods, or to implement impedance profiles (for the common-mode) corresponding to the superposition of two or more coupling coefficients (multi-tuned structures [20]). To illustrate this possibility, we have designed a common-mode suppressed differential line with two cascaded EBG structures (with $f_{max,1}$ = 2 GHz and $f_{max,2}$ = 2.8 GHz). Both have been designed to exhibit a maximum attenuation of 13 dB with a bandwidth of 66% ($f_{max,1}$) and 93.6% ($f_{max,2}$) considering 3 cells. Using the previous equations we have determined the parameters of both EBG structures, i.e., l_1= 2.85 cm, m_1 = 3, K_1 = 0.254 cm^{-1}, and l_2= 2.04 cm, m_2 = 3 and K_2 = 0.356 cm^{-1}, where now the sub-index denotes the EBG, rather than the index of the weighting coefficient of the series expansion of the coupling coefficient. Using the previous procedure, we have determined the layout of both EBG-based differential lines and, after cascading them, we have obtained the electromagnetic simulation, which is compared to measurements in Fig. 4. In this case, the differential insertion loss is better than 0.9 dB in the common-mode rejection band (with a value of 0.7 dB at f_0). The common-mode insertion loss (S_{21}^{CC}) of a designed single-tuned EBG-based line with $m = m_1+m_2 = 6$ and comparable rejection level (30dB) at 2.4GHz is also included for comparison purposes. It can be appreciated that bandwidth can be notably improved by using two cascaded EBGs, without a penalty in device size (both are comparable).

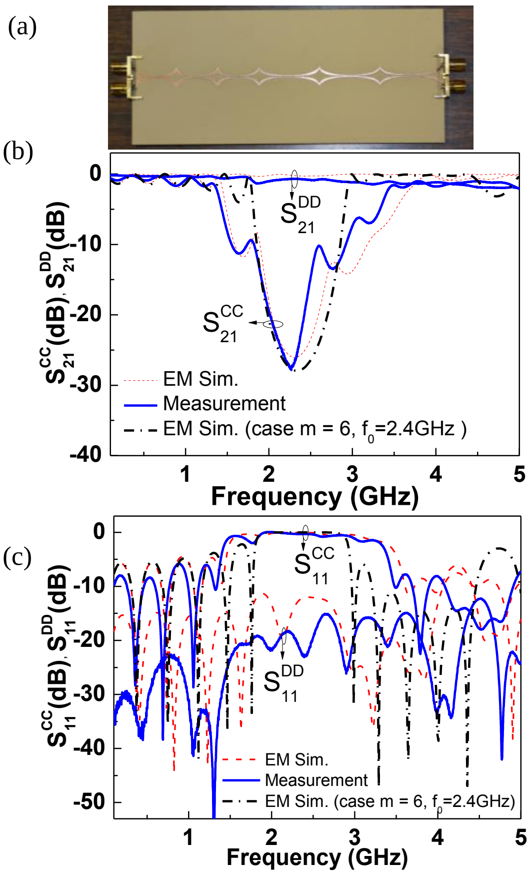

Fig. 4. Photograph (a), differential and common-mode insertion (b) and return (c) loss of the designed 6-cells non-uniform transmission line based on EBGs with different periods. The considered substrate is the *Rogers RO3010* with dielectric constant ε_r = 10.2 and thickness h = 1.27 mm.

V. CONCLUSION

A new strategy for common-mode suppression in microstrip differential lines, based on the modulation of the common-mode characteristic impedance, has been presented. This EBG-like approach has been validated by considering a single-tuned and a double-tuned (providing larger bandwidth) EBG differential line. In both cases, a sinusoidal coupling coefficient for the common-mode has been considered. Through this approach the ground plane is kept unaltered, and the designed structures are transparent for the differential signals.

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